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MÉTHODES D'ESTIMATION DE CANAL ET DE DÉTECTION ITERATIVE POUR LES COMMUNICATIONS CDMA

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Résumé

L'industrie des communications sans fil a abouti à d'importantes conclusions, dont l'utilisation de la division du code à séquence directe pour un accès multiple (*Direct Sequence Code Division Multiple Access* - DS-CDMA) comme concept adéquat d'accès multiple pour garantir les transmissions de données à débit multiple où des services hétérogènes incluant la voix, l'image et le multimédia sont véhiculés dans le même canal physique pour fournir simultanément une large variété de services de communication pour chaque individu, partout et à tout moment.

Cette technologie est communément connue sous le nom de la troisième génération (3G) du système cellulaire qui adopte le DS-CDMA comme concept d'accès. Le code est conçu de manière à assurer l'orthogonalité parmi les différents usagers. L'orthogonalité assure que, dans un canal additif Gaussien, les informations d'un usager donné peuvent être récupérées facilement grâce à un simple filtre apparié. Malheureusement, le canal est loin d'être aussi simple vu du récepteur. Les codes ne sont plus orthogonaux à cause des imperfections du canal. Les signaux des différents usagers subissent plusieurs retards et atténuations aléatoires. De plus, plusieurs versions du signal d'un même usager sont induites à cause de l'effet de trajets multiples.

Ainsi, l'interférence due à l'accès multiple (MAI – *Multiple Access Interference*) et à l'interférence inter symbole (ISI – *Intersybole Interference*) sont les imperfections du canal auxquelles le concepteur d'un système de communication est confronté. Le filtre apparié et le récepteur RAKE sont utilisés comme solutions dans la seconde génération (2G) des systèmes DS-CDMA pour la faible complexité.

Malheureusement, ces récepteurs sont limités à cause de l'interférence due à l'accès multiple et à l'interférence inter symbole, ce qui motive à chercher des solutions plus performantes. Verdú a été parmi les pionniers quant à ses suggestions sur ce qui est connu comme la détection à usagers multiples. Récemment, la combinaison de la détection d'usagers multiples et le décodage a fait l'objet d'une attention considérable, vu son potentiel à améliorer la performance du système dans le but d'approcher celle d'un même système à un seul usager.

Le récepteur optimal pour les systèmes DS-CDMA employant un procédé de codage combine les treillis du détecteur à usagers multiples et du décodeur. La complexité de ce récepteur varie exponentiellement en fonction du produit du nombre d'usagers et de la contrainte (*code constraint length*) du code. Cette complexité fait que l'utilisation de ce détecteur optimal est prohibée même pour les petits systèmes.

Le principal objectif de notre travail consiste à proposer une classe d'algorithmes, de faible complexité, pour la détection et le décodage (détection Turbo) multi-usagers, pour les systèmes DS-CDMA qui adoptent une modulation BPSK (*Binary Phase Shift Keying*). Cet objectif inclut la proposition d'un nouvel estimateur des paramètres du canal intrinsèque à l'atteinte des performances du récepteur. Précisément, deux récepteurs de détection turbo des signaux codés, utilisant une modulation BPSK, à travers un canal multi trajet sont développés:

- Le premier récepteur est basé sur une élimination souple (*soft*) des interférences (IC – *Interference Canceller*), suivie d'un filtre « MMSE » (*Minimum Mean Square Error*) d'une entrée et sortie souple (SISO – *Soft Input Soft Output*), dont les coefficients sont conçus pour être la solution d'un problème d'optimisation MMSE, basée sur des valeurs (forcées) réelles de la sortie, plutôt qu'une sortie à valeur complexe, comme dans le cas d'un récepteur MMSE conventionnel [AHM05c].
- Dans le second récepteur, le problème d'optimisation avec contrainte pour la conception d'un filtre est reformulé par la mise en évidence des valeurs réelles de la sortie du filtre, avec des contraintes à valeurs réelles. Ainsi, le détecteur SISO est suivi d'une banque de décodeurs à maximum de vraisemblance (MAP – *Maximum A-Posteriori*). Les deux étages, détection et décodage, sont séparés par un module d'entrelacement (*interleaver*) et de dé-entrelacement (*deinterleaver*), de manière à leur permettre d'inter changer des informations souples [AHM05c].

Le premier récepteur proposé produit un gain moyen de 2dB, comparativement au récepteur turbo développé par Wang et Poor, avec moins d'itérations [AHM05c]. Le second récepteur, d'une complexité sensiblement réduite, peut atteindre des performances comparables à celle produites par le soft IC-MMSE conventionnel. L'évaluation des performances est réalisée à travers une multitude de simulations riches et variées, incluant un canal multitrajet asynchrone, le problème « *near far* », un canal variant dans le temps, des systèmes à débit multiple et en respectant dans la plupart des cas les standards de l'industrie.

D'autre part, une autre classe d'algorithmes, à faible complexité, pour l'estimation de canal reste un important objectif à réaliser, puisque d'une manière ou d'une autre, les performances de la plupart des détecteurs dépendent directement des estimées du canal. Cet objectif visé concerne particulièrement la problématique liée à l'acquisition des retards, et la poursuite (*tracking*) d'un canal multitrajet variant dans le temps, dans un système DS-CDMA. Dans cette perspective, une structure LMS (*Least Mean Square*) multi-usagers accompagnée d'un filtre de lissage/prédiction (*smoothing/prediction*) est suggérée, afin d'améliorer la qualité de la poursuite du canal [AHM05b]. L'adoption de cette famille d'adaptation est motivée d'une part par la faible complexité de calcul, et d'autre part, par sa structure modulaire appropriée à une implantation VLSI efficace, où entre autres, le parallélisme et le pipeline peuvent s'appliquer facilement. À cet avantage s'ajoute la répartition équitable de calcul. Un modèle stochastique autorégressif utilisé pour la modélisation des processus d'évanouissements corrélés d'atténuation Rayleigh (*correlated Rayleigh fading process*), est incorporé indirectement dans la structure. L'une des propriétés les plus attrayantes de la structure est la possibilité de maintenir les coefficients du filtre constants, pour une large gamme de vitesses du mobile, et une multitude de modèles de canal.

Une analyse des performances versus complexité est aussitôt élaborée, à la fois, dans un environnement cdma2000 et WCDMA, à travers des scénarios de simulations d'intérêts, incluant une multitude de valeurs de débits de données, modèles de canal (*Pedestrian* et *Vehicular*) et vitesses du mobile. L'impact des imperfections du contrôle de puissance est aussi analysé par des simulations. La structure proposée, offre une amélioration substantielle, en comparaison avec la méthode conventionnelle du corrélateur.

Summary

The wireless industry has come into important conclusions among which, the use of the direct sequence code division multiple access (DS-CDMA) as a suitable multiple access scheme to support multi-rate data transmissions, higher data rates where heterogeneous services including voice, video and multimedia are to be vehiculed on the same physical channel to seamlessly provide these wide variety of communication services to anybody, anywhere and at any time. The technology available is popularly known as the third generation (3G) cellular systems which adopted DS-CDMA as its ultimate access scheme.

In DS-CDMA system, each user is assigned a unique code (sequence) that will differentiate it among other active users. The code is designed to ensure orthogonality among the different users. The related users' information bits are spread at the transmitter side using their appropriate codes, and then dispread at the receiver side. The orthogonality ensures that, in an additive white gaussian noise (AWGN) channel, a given user information sequence can be retrieved easily by a simple matched filtering.

Nevertheless, the channel is far to be that simple so that at the receiver side the codes are no longer orthogonal due to the channel impairments. Different users' signals undergo different random attenuations and delays. More than that multiple versions of the same user's signal are induced due to the multipath effect. Hence multiple access interference

(MAI) and inter-symbol interference (ISI) are the major channel impairments the system designer is faced with.

Matched filter and Rake receivers are used as solutions for the second generation (2G) DS-CDMA system due to the simple complexity they offer. Unfortunately, these receivers are MAI and ISI limited, the reason that pushes the research community to look for powerful solutions. Verdú was among the pioneers through his suggestions for what is known as multi-user detection (MUD).

Recently, combined multiuser detection and decoding has received considerable attention because of its potential to improve the performance of a multiuser system to mesh that of a single user system. It was shown that the optimum receiver for a CDMA system employing forward error control (FEC) coding combines the trellis of both the multi-user detector and the FEC code. The complexity of this receiver is exponential in the product of the number of users and the constraint length of the code. This complexity makes the use of the optimal detector prohibitive for even small systems.

The main objective of our work is to suggest low complexity algorithms for iterative multiuser detection and decoding (turbo detection) for DS-CDMA systems utilizing BPSK (Binary Phase Shift Keying) modulation [AHM04], [AHM05a], [AHM05c], [AHM05d]. Namely, two low complexity turbo detection receivers for coded DS-CDMA signals utilizing BPSK as a modulation scheme in multipath channels are derived. The first scheme is based on a soft Interference Canceller (IC) followed by a Soft Input Soft Output (SISO) MMSE filter, whose coefficients are thought to be the solution of an MMSE-optimization problem based on a (forced) real valued filter output rather than a complex one as in conventional MMSE receivers. In the second scheme, we reformulate the traditional constrained filter design by considering a real valued filter output along with real valued constraints. Following the SISO detectors is the bank of a full codeword MAP decoder (SISO channel decoder). Both stages are separated by interleavers and deinterleavers so that they iteratively exchange soft information of a "likelihood" nature. The proposed soft receiver provides an average 2dB gain with fewer iterations when compared with the novel conventional soft IC-MMSE developed by Wang and Poor. The second scheme, of a much lower complexity, provides performances comparable to the conventional soft IC-MMSE. Simulation results for performance evaluation are conducted under highly interesting scenarios including asynchronous multipath channels, near far problem, time varying channels, and multirate systems.

In a secondary objective, a low complexity algorithm for channel estimation is, nevertheless, of a major concern, since most of MUDs performances depend on one way or another on the channel estimates. This objective addresses the problem of asynchronous multiuser delays acquisition and time varying channel tracking for multipath channels in DS-CDMA systems. A multiuser-LMS-like (*Least Mean Square*) structure along with smoothing/prediction filters to improve tracking quality is suggested [AHM05b]. The choice for such adaptation family stems from its low computational complexity and its regular structure suitable for an efficient VLSI implementation where parallelism and wave pipelining, among other techniques, are easily applied. It is computationally effective due to the even distribution of the computational load and no extra heavy computation is required at the end of the processing window or preamble. An autoregressive stochastic model of correlated Rayleigh fading processes is used and indirectly embedded into the design. The attractive property of the proposed structure is the unique filter settings used

for a large range of mobile speeds and channel types for all users accessing the system. A performance versus complexity analysis is conducted, in both cdma2000 and WCDMA environments, over highly interesting settings including different data rates, channel types (pedestrian and vehicular) and mobile speeds. Power control imperfections impact is, also, analyzed through simulations. The proposed multiuser LMS structure offers substantial improvements compared with the Correlator method.

Sommaire

1. La course vers une plus grande capacité et un plus haut débit

Le nombre de clients ayant recours à des services cellulaires, augmente exponentiellement, et il existe une demande visant l'intégration de services variés de multimédia. Ces éléments constituent les motifs qui ont amené l'industrie sans fil à en venir à des conclusions importantes telles, l'utilisation d'accès multiple par division de code (CDMA – *Code Division Multiple Access*) comme un concept d'accès multiple approprié, pour supporter des hauts débits de transmissions, où des services hétérogènes incluant la voix, la vidéo et le multimédia doivent être véhiculée sur le même canal physique, et cela, afin de fournir un large éventail et variété de services de communication d'une manière équitable à tout le monde, n'importe où, et à n'importe quel moment. Ainsi, le débit requis pour ces services varie largement entre quelques kilo bit par seconde (kbps) pour une simple pagette jusqu'à plusieurs Méga bit par seconde (Mbps) pour une transmission vidéo. En conséquence, le fait de supporter un aussi large choix de débits, tout en maintenant la flexibilité dans la gestion de la mobilité, augmente considérablement la complexité du réseau. Cette technologie largement en vogue, connue comme la troisième génération (3G) des systèmes cellulaires, a adopté le CDMA comme l'ultime concept d'accès [FUK96].

CDMA est une modulation numérique et un système d'accès radio qui emploie des codes de signature (plutôt que des intervalles de temps ou des bandes de fréquences) pour arranger un accès radio simultané et continu à des utilisateurs multiples. La contribution à l'interférence dans le canal radio des communications mobiles résulte de l'accès multiple des utilisateurs, la propagation radio à trajets multiples, et l'interférence de canaux adjacents entre autres.

La performance des systèmes à étalement de spectre est relativement immunisée contre l'interférence radio. La sectorisation de la cellule et l'activité vocale dans des schèmes radio CDMA fournissent une capacité additionnelle comparativement aux systèmes radio FDMA et TDMA. Cependant, le CDMA a toujours quelques inconvénients, le principal étant que la capacité (le nombre d'utilisateurs actifs à n'importe quel instant) reste limitée par l'interférence d'accès. En outre, l'effet «near-far» exige une implantation précise du contrôle de puissance. Le premier système radio CDMA cellulaire a été construit conformément aux standards IS-95, et est maintenant connu commercialement sous le nom de cdmaOne.

Le système CDMA IS-95 est un système radio à bande étroite. La largeur de bande est limitée à 1.25 MHz avec un débit de bribe (*chip*) de 1.2288 Mcps. Le système est destiné à fournir un service de voix et de données à faible débit en utilisant des techniques de commutation de circuit (*circuit-swicthing*). Le débit de données varie de 1.2 Kbps à

9.6 Kbps. Les structures de liaison descendante (de la station de base au mobile – *downlink*) et montante (du mobile à la station de base – *uplink*) sont différentes et chacune a une capacité distincte. La liaison descendante est cohérente et synchrone tandis que la liaison montante est asynchrone. La « canalisation » est réalisée dans chaque lien, en utilisant des codes orthogonaux à 64 bribes, avec la possibilité d'inclure le pilote, les signaux de synchronisation, de pagination et d'accès au réseau. Par conséquent, le nombre d'utilisateurs actifs capables d'avoir simultanément l'accès au réseau est limité par le niveau d'interférence, des dispositions du service et le nombre de canaux disponibles. Dans IS-95B, un mobile actif a toujours un canal fondamental à 9.6 Kbps et quand une transmission à grand débit est exigée, la station de base assigne au mobile jusqu'à 7 canaux supplémentaires. Ainsi, le débit maximal de données est de 76.8 Kbps. Le débit de données est contrôlé à la station de base et transmit au mobile via un canal supplémentaire par le message d'attribution (ou d'assignation).

Le système CDMA à large bande (le WCDMA – *Wideband CDMA*) est l'un des principaux standards dans la 3G. Le WCDMA supporte une transmission de haut débit, soit 384 kbps pour une couverture d'un large secteur et 2 Mbps pour une couverture locale de services multimédia. Ainsi le WCDMA est capable d'offrir une transmission de voix, de texte, de données, d'image (l'image fixe) et de vidéo avec une "simple" plate-forme. Cependant, en plus des inconvénients résultant de l'environnement mobile et de l'interférence due à l'accès multiple, la transmission de haut débit cause l'interférence inter-symbole (ISI – *Interference Intersymbole*). L'ISI doit donc être prise en considération durant la transmission et/ou la réception.

L'importante caractéristique du canal sans fil, est qu'il est aléatoirement variant dans le temps. Il existe de multiples trajectoires de propagation, émanant de l'émetteur et allant vers le récepteur, et celles-ci ont des amplitudes et des délais aléatoirement variant dans le temps, dépendamment de l'emplacement géographique et la mobilité du récepteur par rapport à l'émetteur. Cette variation temporelle des caractéristiques du signal est d'habitude connue comme évanouissement (*Fading*), et mène vers une dégradation significative de la performance, comparativement à la performance dans un canal déterministe traditionnel, avec un bruit Gaussian blanc additif (AWGN – *Additive White Gaussien Noise*) [PRO95]. La diversité dans la signalisation est une technique puissante pour contrer l'évanouissement. Cette technique fournit au récepteur des répliques (copies) du même signal d'information, mais avec des évanouissements indépendants et réduit ainsi significativement la probabilité que toutes les répliques du signal souffrent du même évanouissement simultanément. La fréquence, le temps et l'espace (antennes) sont les « dimensions » généralement utilisées de la diversité [PRO95]. Dans la diversité fréquentielle et temporelle, la même transmission est répétée dans plusieurs bandes de fréquences et à des instants différents du temps, respectivement. Idem pour ce qui est des systèmes de diversité spatiale où plusieurs copies du signal sont obtenues en utilisant de multiples antennes de réception et/ou de transmission.

La réalisation matérielle du réseau (*Arrays*) d'antennes au niveau du récepteur et/ou au niveau de l'émetteur, constitue une solution qui présente quelques inconvénients de calibrage et de coûts qui empêcheront sa réalisation pour les quelques années à venir. Cependant, les techniques avancées de traitement numérique de signal pour l'acquisition (estimation du délai), la poursuite des variations d'atténuation des trajets multiples variant dans le temps et les techniques de détection à usagers multiples, semblent être les candidats favoris de l'industrie du sans fil, pour augmenter l'ensemble des performances du système.

Le traitement numérique de signal à usagers multiples réfère au traitement conjoint des signaux de tous les usagers dans le système. Ce traitement conjoint est très efficace (vu sa capacité à fournir une amélioration significative des performances) comparativement au traitement à utilisateur unique traditionnel utilisé en pratique, dans lequel les performances des systèmes CDMA sont encore plus limitées par les interférences, [VIT95], [PRO95], [PIC82]. L'intérêt pour le traitement de signal à usagers multiples dans les systèmes CDMA provient des travaux pionniers de Verdú [VER86], dans lesquels il a proposé et analysé le détecteur optimal à usagers multiples (*Maximum Likelihood Detector*). Malheureusement, ce détecteur possède une complexité exponentielle avec le nombre d'usagers. Ainsi, au cours de la dernière décennie, la recherche dans ce secteur s'est focalisée sur plusieurs solutions sous optimales [LUP89], [LUP90], [VAR90], [XIE90], [MAD94], [ZVO96] qui possèdent une complexité linéaire au nombre d'usagers et qui semblent plus faciles à mettre en oeuvre. Par ailleurs, d'autres algorithmes de traitement de signal au niveau du récepteur, tels l'estimation de canal et le décodage, ont aussi été étudiés récemment dans la perspective d'usagers multiples. L'estimation à usagers multiples du canal est essentielle pour l'obtention d'estimations précises des retards et des atténuations qui peuvent être plus tard utilisées par les détecteurs. Le décodage est nécessaire dès qu'un système est codé dans le but de contrôler les erreurs. Des algorithmes optimaux et sous-optimaux de détection et décodage (détection turbo) à usagers multiples ont été proposés dans [GIA96a], [GIA96b], [FAW96], [ALE99], [WAN99], [MOH98] tandis que d'autres algorithmes pour l'estimation de canal ont été proposés dans [BEN96], [BEN98], [STR96], [MAD97], [SEN98a], [SEN98b], [CAM96], [MAN98], [MAN00].

Dans ce qui va suivre, nous présenterons notre problématique, méthodologie et nos objectifs qui constitueront le sujet des chapitres 3 et 4.

2. Problématique

Traditionnellement, dans le système de DS-CDMA, un code (séquence) unique est assigné ou attribué à chaque usager, qui le différenciera des autres usagers actifs. Le code est conçu pour assurer l'orthogonalité des différents usagers. Les informations relatives à chaque usager sont étalées au niveau de l'émetteur, en utilisant leurs codes appropriés et ensuite désétalées au niveau du récepteur. L'orthogonalité assure une récupération facile, par un simple filtre apparié, d'une séquence d'information d'un usager donnée dans un canal de bruit Gaussien [PRO95].

Cependant, le canal est loin d'être aussi simple, ce qui fait qu'à la réception les codes ne sont plus orthogonaux en raison des imperfections du canal. Les différents signaux des usagers subissent différents retards et atténuations aléatoires [RAP96]. En plus de cela, des versions multiples du signal du même utilisateur sont induites en raison de l'effet multitrajets. L'interférence due à l'accès multiples (MAI – *Multiple Access Interference*) et l'interférence inter-symbole (ISI) sont les principales imperfections du canal auxquelles fait face le concepteur du système.

Le filtre apparié et le récepteur à râteau (récepteur Rake) sont utilisés comme solutions pour la deuxième génération (2G) des systèmes DS-CDMA, vu la complexité simple qu'ils exhibent. Malheureusement, les performances de ces récepteurs sont limitées à cause des MAI et ISI, ce qui pousse la communauté des chercheurs à proposer des solutions plus puissantes. Verdú [VER98] était parmi les pionniers de par ses suggestions sur ce que l'on connaît comme la détection à usagers multiples (MUD – *MultiUser Detector*). Depuis lors, plusieurs détecteurs à usagers multiples ont été proposés, nous pouvons citer à titre illustratif: le récepteur à minimum des moindres carrés (MMSE – *Minimum Mean Square Error*), le décorrelateur (avec ou sans retour de décision), l'annulateur d'interférence parallèle (PIC – *Parallel Interference Canceller*), l'annulateur d'interférence sériel (SIC – *Serial Interference Canceller*), le récepteur à rejet d'interférence sous-espace (ISR – *Interference Subspace Rejection*) [AFF02] [VER98]. Ces récepteurs exploitent la structure d'interférence dans le signal reçu en fonction de la signature des usagers et des paramètres du canal. Ils diffèrent dans leur performance et leur complexité en fonction de la manière dont cette structure est traitée.

Récemment, la détection combinée d'usagers multiples avec le décodage a fait l'objet d'une attention considérable, vu leur potentiel à améliorer la performance d'un système à usagers multiples afin de se rapprocher de celle d'un système avec un seul usager. Il a été démontré, dans [GIA96a] et [HAG97], que le récepteur optimal pour un système CDMA, employant le codage canal sans retour (FEC – *Forward Error Correction*), combine simultanément le treillis du détecteur à usagers multiples et du codage canal. La complexité de ce récepteur est exponentielle dans le produit du nombre d'utilisateurs et la longueur de la contrainte du code (*code constraint length*). Cette complexité rend prohibitive l'utilisation du détecteur optimal, même pour de petits systèmes. Dans [GIA96b], on étudie quelques récepteurs à complexité réduite, qui réalisent la détection et le décodage de symboles à usagers multiples séparément ou conjointement. Des concepts de décodage itératifs pour les systèmes CDMA, convolutionnelle ou turbo codé, sont proposés dans [REE98] et [WAN99]. La différence principale dans la structure de ces concepts est le type du détecteur utilisé.

Un autre type de détection itératif est basé sur l'annulation d'interférence souple (*Soft Decision*) [ALE98] [WAN99] où les estimations souples des bits de code sont représentées par un nombre réel plutôt que par des valeurs de décision dure (*Hard Decision*), et sont combinées pour produire les estimations du LLRs (*Log Likelihood Ratio*) des bits de code.

Ces récepteurs exploitent la structure du signal, en utilisant simultanément l'information des usagers dans leur forme souple, pour exploiter de manière efficace tout le gain potentiel, du traitement de signal, qui peut être réalisé. En plus de MAI, ISI et de l'effet des trajets multiples; le problème *near-far*, et les variations temporelles causées par « l'effet Doppler » sont relativement de la même importance pour une conception efficace du système.

Les récepteurs peuvent contrer l'interférence due aux accès multiples, en utilisant l'estimation de canal à usagers multiples [BEN96], [MOO94], [STR96], [MAD97], [SEN98a] et [SEN98b]. Ces algorithmes sont développés pour des systèmes CDMA avec des codes d'étalement courts, qui se répètent à chaque symbole. Cependant, les codes d'étalement utilisés dans les systèmes pratiques ont une période beaucoup plus grande que la durée du symbole et sont appelés codes d'étalement long. Donc, la plupart des algorithmes existants sont inapplicables ou ont besoin de ressources de calcul prohibitives.

La conception d'un algorithme efficace d'acquisition et de poursuite à usagers multiples du canal multitrajets variant dans le temps, est devenu un « goulot d'étranglement ». Les algorithmes actuels pour l'estimation du canal comptent principalement sur une sorte de moyenne, assumant bien sûr que les coefficients du canal sont au moins constants, pendant la période d'intérêt (période d'apprentissage, fenêtre prédéfinie). Cela amène un manque dans la qualité de la poursuite nécessaire des canaux variants dans le temps. L'estimation du canal basée sur la méthode de maximum de vraisemblance (ML – *Maximum Likelihood*) [BHA02] qui opère sur une décision statistique prise à partir de la moyenne des sorties de filtres appariés de tous les usagers simultanément, n'est qu'un exemple.

La complexité de la mise en œuvre matérielle est et a toujours été le facteur décisif pour l'industrie dans le choix d'un algorithme d'estimation de canal tout en respectant des objectifs de performances élevés. Le Correlator, en raison de la simple complexité qu'il exhibe, a été le candidat pour l'industrie de la 2G. De par son accès matériel, l'industrie de la 3G a conservé le Correlator comme technique d'estimation de canal dans les déploiements initiaux, mais compte tenu de ses performances limitées, surtout en transmission des données, de nouvelles techniques à faible complexité doivent être proposées. Pour améliorer la précision des estimés fournis du canal, par le Correlator, la réponse impulsionnelle obtenue du canal est en plus traitée en employant un filtre passe-bas appelé filtre d'estimation du canal [LIG93].

3. Objectifs

L'objectif principal de notre travail est de suggérer une nouvelle classe d'algorithmes pour la détection et le décodage itératifs à usagers multiples, qui exploite la modulation BPSK (*Binary Phase Shift Keying*). Cet objectif de base est divisé en des sous objectifs qui impliquent:

- la conception de nouvelles classes d'algorithmes à usagers multiples (MUD) avec entrées douces et sorties douces (SISO), et;
- la conception d'un algorithme pour l'estimation de canal puisque les performances de la plupart des MUDs en dépendent d'une façon ou d'une autre.

Ces algorithmes sont conçus pour la liaison montante DS-CDMA, où l'accès simultané aux informations des usagers est possible, et les exigences en terme de la consommation énergétique, ne constituent pas des contraintes.

4. Méthodologie

Nous abordons brièvement dans cette section le travail relatif aux techniques d'estimation de canal et de détection itérative (turbo). Cependant, il serait pertinent de souligner qu'une compréhension appropriée de la nature du canal sans fil est requise pour une conception et une analyse efficaces des systèmes de communication. Une des caractéristiques principales du canal sans fil réside dans le fait qu'il est aléatoirement variable dans le temps. Il existe des trajets de propagation multiples, possédant des amplitudes et des délais variables dans le temps [AHM01]. En fait, on peut réaliser une détection fiable en exploitant l'estimation du canal proposée dans [AHM02] pour les systèmes TDMA, et dans [BEN96], [SEN98a] et [SEN98b] pour les systèmes CDMA, dans lesquels sont utilisés les codes d'étalement courts. Dans [THO99] ET [WEI99], les techniques proposées sont basées sur la connaissance des séquences d'étalement, l'estimation de canal et les informations des usagers interférents, et l'utilisation de la technique d'annulation des interférences et l'approche MMSE, respectivement. D'autre part, la technique du maximum de vraisemblance pour l'estimation de canal (Maximum Likelihood, ML) [BHA02], se base sur une décision statistique obtenue à partir d'une moyenne des sorties successives des filtres appareillés de tous les usagers.

Étant donné sa simple complexité, le corrélateur (*Correlator*) est un bon candidat pour une implantation de faible complexité. Afin d'améliorer la précision des estimations de canal fournies par le corrélateur, on utilise un filtre passe-bas, qu'on appelle le filtre d'estimation de canal (FEC) [LIG93]. Dans le sens MMSE, il est admis que le filtre de Wiener est optimal comme un FEC dans un canal stationnaire [LIG93]. Cependant, pour concevoir un filtre de Wiener, il est essentiel de connaître la densité spectrale du canal et du bruit, soit deux informations qui ne peuvent être obtenus en temps réel. Il est préférable pour cela d'utiliser un filtre passe-bas comme un FEC, dont la fréquence de coupure est égale au maximum de la fréquence du Doppler [LIG93], [SCH98a] et [SCH98b].

Dans notre travail, nous suggérons une structure LMS multi-usager avec des filtres de prédiction et de lissage pour améliorer la qualité de poursuite. Le choix d'une telle famille d'adaptation est motivé par sa faible complexité de calcul et sa structure régulière. Tout comme dans les méthodes d'estimation de canaux basées sur le filtre de Kalman, nous utilisons un modèle stochastique autorégressif du processus d'évanouissement de type

Rayleigh corrélé [BAD01], qui est cependant indirectement intégré dans la conception. Le filtre unique qui est utilisé pour un large éventail de vitesses des mobiles, constitue le principal attrait de la structure proposée. Le processus de conception est basé la méthodologie suivante : premièrement, un modèle stochastique autorégressif d'ordre p est conçu suivant un profil de Doppler moyen ; deuxièmement, le modèle résultant ainsi que la valeur admissible du « parameter-drift-to-noise-ratio » (υ) [LIN01] sont utilisés pour le calcul des coefficients du filtre de lissage/prédiction selon la théorie de Wiener LMS [LIN01] ; finalement, la structure LMS multi-usager est améliorée avec des filtres de lissage/prédiction.

En pratique, l'emploi du codage correcteur d'erreur et de l'interlacement (interleaver) par la plupart des systèmes CDMA, a amené de nombreux travaux sur la détection à usagers multiples pour les systèmes codés [GIA96b]. D'autre part, des détecteurs itératifs pour les systèmes CDMA, convolutionnelles ou turbo codés sont proposés dans [REE98] et [WAN99]. Cependant, la différence entre ces propositions réside dans le type de détecteur à usagers multiples avec entrée et sortie douces (soft-input/soft-output, SISO). Dans [REE98] et [MOH98], un détecteur itératif optimal est proposé pour un système CDMA synchrone et à code convolutionnelle avec une complexité de calcul d'ordre de $O(2^K)$ (où K est le nombre total des usagers). Basé sur le concept du code convolutionnel variant dans le temps, des mécanismes récursifs similaires à ceux qu'on trouve dans l'algorithme de BCJR [BAH74] sont proposés pour la conception d'un détecteur à usagers multiples. Un autre type de détecteurs itératifs est basé sur l'annulation d'interférences souples [HAG96b], [ALE98], [WAN99]. [WAN99] avec une structure qui utilise le critère MMSE pour l'annulation souple des interférences construites en utilisant des informations *a priori* fournies par un décodeur.

Il est à souligner que l'exploitation de la modulation BPSK dans la conception des récepteurs linéaires a été déjà suggérée dans [TUL01], [BUZ01a], [HER05]. Cependant, ceci n'a pas été évoqué dans la conception des détecteurs itératifs (turbo). Notre contribution peut, donc, être perçue comme l'extension des résultats dans [GAM00] et [AFF02] pour les systèmes utilisant la modulation BPSK.

Dans notre travail, deux récepteurs itératifs (turbo) à faible complexité sont proposés. Nous considérons un système CDMA convolutionnellement codé et utilisant une modulation BPSK dans le contexte d'un canal à trajets multiples. Le premier récepteur est basé sur un bloc d'annulation d'interférences souples suivi d'un filtre MMSE. Les coefficients de ce dernier sont les solutions d'un problème d'optimisation basé sur le critère de MMSE ou la valeur réelle de la sortie du filtre est explicitement prise en considération contrairement à l'approche conventionnelle dans [WAN99]. Quant au deuxième récepteur, nous reformulons le problème de conception de filtre avec contraintes où nous considérons explicitement la valeur réelle de la sortie du filtre avec contraintes réelles.

Il est primordial de mentionner qu'une plate-forme de simulation DS-CDMA précise et rapide, est exigée pour valider, analyser et tirer des conclusions sur nos algorithmes, en les comparant à différentes techniques d'estimation des canaux et de détection de référence, dans un environnement DS-CDMA asynchrone. Une telle plate-forme est créée dans l'environnement MATLAB et C avec notre partenaire industriel Axiocom inc..

Par la suite, des algorithmes pris de la littérature, appropriés tant pour l'estimation du canal que pour la détection itérative, sont implémentés. Leurs performances respectives, ainsi que l'analyse de la complexité, sont établies et commentées.

En se basant sur les dernières conclusions de performance/complexité faites ci-dessus, de nouvelles techniques sont suggérées. Le but à atteindre, comme mentionné ci-dessus dans nos contributions, est d'augmenter la performance du système en gardant une complexité de mise en oeuvre autant réduite que possible ou au moins justifier la complexité additionnelle par un gain de performance jugé suffisant.

Mes remerciements les plus vifs à Axiocom Inc qui a mis à ma disposition ses plates-formes de WCDMA "MONARK©" et de cdma2000 "MACHAON©", pour les simulations. Le tableau 1 récapitule en cinq étapes la méthodologie de la thèse.

Tableau 1. Méthodologie en cinq étapes

Étape 1.	Relevé de la littérature sur les systèmes DS-CDMA et MUD en général, *Soft* MUD, codage canal et estimation de canaux en particulier.
Étape 2.	Évaluation de plusieurs algorithmes proposés dans la littérature, de détection à usagers multiples, d'estimation de canaux et de détection itérative (turbo) sur les mêmes conditions concernant le canal à trajets multiples, MAI et effet *near-far*, etc. Cette étape permettra l'identification des méthodes de références pour comparaison.
Étape 3.	Proposition d'un algorithme à usagers multiples pour acquisition et poursuite des canaux mobiles à trajets multiples.

Étape 4.	Proposition d'une classe d'algorithmes de détection itérative à usagers multiples dans des systèmes DS-CDMA utilisant la modulation BPSK.
Étape 5.	Validation et évaluation de performance des algorithmes proposés dans différents scénarios utilisant aussi bien la plate-forme WCDMA, et la plate-forme cdma2000 pour l'estimateur de canal, qu'une plate-forme DS-CDMA pour algorithmes de détection itératives.

5. Organisation de la thèse

Ce rapport est organisé comme suit; dans le chapitre 2, un modèle de signal DS-CDMA est présenté. Deux modèles duels sont présentés, le premier est approprié pour la détection tandis que le deuxième est plus adéquat pour l'estimation de canal. Dans le chapitre 3, une approche multi-usager pour l'acquisition et la poursuite du canal est proposée. Le chapitre 4 est dédié à la dérivation d'une nouvelle classe de détection itérative à usager multiple pour des systèmes de DS-CDMA codés et utilisant la modulation BPSK. Finalement, une conclusion est tirée et les travaux futurs seront présentés.

6. Contributions de la thèse

Nous retrouvons deux contributions majeures dans cette thèse:

- **Acquisition et poursuite des canaux de communication variant dans le temps**

Dans notre travail, une structure LMS (*Least Mean Square*) à usagers multiples avec filtres pour lissage et prédiction est proposée afin d'améliorer la qualité de poursuite du canal [AHM05b]. Le choix d'une telle famille d'adaptation, émane de son faible niveau de complexité arithmétique et de sa structure régulière aux communications locales propice à sa mise en œuvre efficace en technologie VLSI (*Very Large Scale Integration*), où le parallélisme comme le pipeline fin, parmi d'autres techniques, est facilement applicable [MOR99], [MOZ99], [SAK98] et [MAS98]. En plus de cela, cette structure est efficace du point de vue de la complexité, vu la distribution de la charge de calcul sur chaque durée de symbole et qu'aucun calcul supplémentaire n'est exigé à la fin de la fenêtre de traitement ou de préambule.

En résumé:

- La structure proposée LMS multi-usager prend en considération les contributions de tous les usagers, et elle produit à chaque symbole (symbole pilot) « une réponse impulsionnelle composite du canal ». La réponse impulsionnelle composite du canal est définie comme étant au moins un vecteur de $(N+1)K$ éléments dont le contenu représente, simultanément, les informations sur les retards et les atténuations variant dans le temps (K est le nombre des usagers et N est le gain de traitement, ou la longueur du code d'étalement du symbole pilot).

- Un filtre de lissage/prédiction unique est conçu selon un modèle AR d'ordre p considérant un profil Doppler moyen sur un intervalle de vitesses des mobiles.

- Le pas d'adaptation est dynamiquement ajusté à chaque itération.

Tout comme dans [BAH02],

- Une distribution équitable de la complexité de calcul à la longueur de préambule, et
- une structure régulière pour une implémentation VLSI efficace,

constituent les motivations pour un tel choix de structure LMS multi-usager.

- **Exploitation de la modulation BPSK dans la détection itérative (Turbo)**

Soucieux de la question de la mise en oeuvre, nous proposons une structure de détection itérative à complexité réduite, qui est proche de par sa complexité du récepteur I-Rake, où I est le nombre d'itérations. Désireux d'améliorer la performance tout en gardant la complexité réduite, nous proposons une nouvelle structure itérative [AHM05a] basée sur la conception d'un filtre en avant/en arrière (*feed-forward/feed-back*) dans le sens de MMSE. Il est important de mentionner que l'exploitation de la modulation BPSK dans la conception des récepteurs linéaires n'est pas un nouveau défi [TUL01], [BUZ01a], [HER05] mais l'idée de le prendre en considération dans des détecteurs turbo n'a pas encore été étudiée. Notons que cette contribution peut être considérée comme une extension de [WAN99] pour les signaux BPSK, visant aussi, à étudier le gain potentiel additionnel réalisé en exploitant ce type de modulation [AHM05c], [AHM05d].

En résumé:

Deux récepteurs itératifs (turbo) sont proposés. Nous prenons on considération un système DS-CDMA utilisant une modulation BPSK dans un contexte de canaux à trajets multiples.

- Le premier détecteur est basé sur un bloc d'annulation d'interférences douces suivi d'un filtre MMSE. Les coefficients de ce dernier constituent la solution du problème d'optimisation selon un critère MMSE où la valeur réelle de la sortie du filtre est explicitement prise en considération contrairement à l'approche conventionnelle proposée dans [WAN99].

- Une réduction de la complexité de calcul est réalisée en redéfinissant le critère d'optimisation de telle façon que le problème ne nécessite que l'inversion d'une matrice réelle au lieu d'une matrice complexe. L'utilisation de la technique du gradient pour résoudre un système d'équations constitue une autre alternative attrayante.

- Dans une seconde proposition, nous reformulons le problème de conception de filtre avec contraintes, en prenant en considération la valeur réelle de la sortie du filtre ainsi que les valeurs réelles des contraintes qui y sont associées.

- Ces mêmes idées sont exploitées dans [AHM05a] pour une structure de récepteur avec retour de décisions.

Davantage d'explications et de commentaires sont fournis dans les sections 3.7 et 4.5 à propos des performances et de la complexité relatives des algorithmes proposés. Dans un souci de concision, nous préférons discuter de certaines perspectives futures.

Les travaux futurs impliquent plusieurs issues; quant à l'algorithme multi-usager Wiener LMS, la stratégie de conception du filtre de lissage/prédiction évoque plusieurs itérations pour obtenir les valeurs des coefficients quasi-optimaux. Cette étape nécessite plus d'attention pour élaborer une stratégie systématique simple et efficace. En même temps, il sera intéressant d'analyser la performance de l'algorithme du point de vu d'excès MSE et de la constante de temps. Cette étude nous permettra de comprendre comment les paramètres de l'algorithme (nombre d'usagers, coefficients de filtre, pas d'adaptation, etc.) affectent les performances. Cette même étude nous aidera à raffiner la méthodologie de conception.

Comme nous avons brièvement suggéré dans le chapitre 4, des travaux futurs peuvent être élargis pour effectuer une étude analytique des performances des récepteurs turbo proposés en utilisant « EXIT charts ». Il est à noter que si la modulation n'est plus de type BPSK, toutes les méthodes turbo peuvent être équivalentes.

L'extension des algorithmes proposés aux systèmes DS-CDMA avec antennes multiples reste une avenue à explorer, dans le but de savoir si les mêmes gains de performance observés avec une seule antenne peuvent être maintenus et à quelle complexité [AHM05b].

Finalement, nous présentons l'analyse et l'étude des performances des détecteurs turbo et ce, en présence d'un estimateur de canal [AHM05b]-[AHM05d], des imperfections dues au contrôle de puissance, des interférences inter cellules et/ou des effets de quantifications.

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List of Acronyms

2G/3G	Second/Third generation
3GPP/3GPP2	Third generation project partner/ Third generation project partner 2
AWGN	Additive white Gaussian noise
BPSK	Binary phase shift keying
CDMA	Code division multiple access
cdmaOne	Trade mark of IS-95 system
CEF	Channel estimation filter
CN	Core network
DFT	Discrete Fourier transform
EDGE	Enhanced data rate for GSM evolution
EGPRS	Enhanced GPRS
EXIT	Extrinsic information transfer
FDMA	Frequency division multiple access
FEC	Forward error control
GGSN	Gateway GPRS support node
GMSK	Gaussian minimum shift keying
GPRS	General packet radio service
GSM	Global system for mobile communications
GTP	GPRS tunnel protocol
IC	Interference cancellation
IDFT	Inverse discrete Fourier transform
IMT-2000	International mobile telecommunications
IS-95/IS-95B	Interim standard 95/95B
ISI	Intersymbol interference
ISR	Interference subspace rejection

LLR	Log-likelihood ratio
LMS	Least mean square
LSSI	Laboratory of signal and system integration
MAI	Multiple access interference
MAP	Maximum a posterior
MC-CDMA	Multi-carrier CDMA
MC-CDMA	Multi-carrier CDMA
MCM	Multi-carrier modulation
MIMO	Multiple input-multiple output
ML	Maximum likelihood
MMSE	Minimum mean square error
MUD	Multiuser detector
OFDM	Orthogonal frequency division multiplexing
PC	Power control
PIC	Parallel interference canceller
QPSK	Quaternary phase shift keying
RLS	Recursive least square
SGSN	Serving GPRS support node
SIC	Successive (serial) interference canceller
SISO	Soft input-soft output
TDMA	Time division multiple access
UMTS	Universal mobile telecommunication systems
UTRAN	UMTS terrestrial radio access network
VLSI	Very large scale integration
WCDMA	Wideband CDMA
WLMS	Wiener LMS

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1.

INTRODUCTION

1.1 The Race Toward Higher Capacity and Higher Data Rates

The number of customers requiring cellular services is increasing exponentially, and there is a demand for integration of a variety of multimedia services. These were the motives that pushed the wireless industry to come into important conclusions among which, the use of the code division multiple access (CDMA) as a suitable multiple access scheme to support higher data rates transmissions where heterogeneous services including voice, video and multimedia are to be vehiculed on the same physical channel to seamlessly provide these wide variety of communication services to anybody, anywhere and at any time. Consequently, the bit rate required for the services varies largely from just 1.2 kbps for paging up to several Mbps for video transmission. Furthermore, supporting such a wide range of data rates with flexible mobility management increases network complexity dramatically. The technology available is popularly known as the third generation (3G) cellular systems that adopted CDMA as its ultimate access scheme [FUK96]. CDMA is a digital modulation and radio access system that employs signature codes (rather than time slots or frequency bands) to arrange simultaneous and continuous access to a radio network by multiple users. Contribution to the radio channel interference in mobile communications arises from multiple user access, multipath radio propagation, adjacent channel radiation and radio jamming.

The spread spectrum system's performance is relatively immune to radio interference. Cell sectorisation and voice activity used in CDMA radio schemes provide additional capacity compared to FDMA and TDMA. However, CDMA still has a few drawbacks, the main one being that capacity (number of active users at any instant of time) is limited by the access interference. Furthermore, Near-far effect requires an accurate and fast power control scheme. The first cellular CDMA radio system has been constructed in conformity with IS-95 specifications and is now known commercially as cdmaOne.

1.1-1 Packet data over GSM, GPRS, and EDGE

There is increasing demand for data traffic over mobile radio. The mobile radio industry has to evolve the current radio infrastructures to accommodate the expected data traffic with the efficient provision of high-speed voice traffic. The General Packet Radio Service (GPRS) is being introduced to efficiently support high-rate data over GSM. GPRS signaling and data do not travel through GSM network. The GPRS operation is supported by new protocols and new network nodes: Serving GPRS support node (SGSN) and Gateway GPRS support node (GGSN). One prominent protocol used to tunnel data through IP backbone network is the GPRS tunnel protocol (GTP). GPRS obtains user profile data using location register database of GSM network. GPRS supports quality of service and peak data rate of up to 171.2 kbps with GPRS using all 8 timeslots at the same time. GPRS uses the same modulation as that used in GSM, that is Gaussian Minimum Shift Keying (GMSK) with 4 coding schemes. GPRS packetises the user data and transports it over 1 to 8 radio channel timeslots using IP backbone network.

The Enhanced Data Rates for GSM Evolution (EDGE) employs an Enhanced GPRS (EGPRS) to supports data rate up to 384 kbps through optimised modulation. EGPRS support 2 modulation schemes, namely GMSK with 4 coding schemes and 8-PSK with 5 coding schemes. Unlike GPRS where header and data are encoded together, headers are encoded separately in EGPRS.

1.1-2 CDMA and MC-CDMA

The IS-95 CDMA system is a narrow band radio system. Bandwidth is limited to 1.25 MHz and a chip rate of 1.2288 Mcps. The system is intended to provide voice and low bit rate data service using circuit-switching techniques. Data rate varies from 1.2 kbps to 9.6 kbps. Forward (base station to mobile) and reverse (mobile to base station) link structures are different and each is capable of distinctive capacity. Forward transmission is coherent and synchronous while the reverse link is asynchronous. The "chanellisation" in each link is achieved by using 64-chip orthogonal codes, including provision for pilot, synchronization, paging, and network access. Consequently, the number of active users able to simultaneously access the network is limited by the level of interference, service provisions and the number of "channels" available. In IS-95B, an active mobile always has a fundamental code channel at 9.6 kbps and when high data rate is required, the base station assign the mobile up to 7 supplementary code channels. Thus peak data rate is up to 76.8 kbps. Data rate is controlled at the base station and conveyed to mobile through the supplementary channel assignment message.

The Wideband CDMA (WCDMA) system is the major standard in the next-generation Global Mobile Telecommunications standard suite IMT-2000. The WCDMA supports high data rate transmission, typically 384 kbps for wide area coverage and 2 Mbps for local coverage for multimedia services. Thus WCDMA is capable of offering the transmission of voice, text, data, picture (still image) and video over a single platform. However, in addition to the drawbacks arising from the mobile environment and multiple access interference, high bit rate transmission causes Inter-symbol interference (ISI) to occur. The ISI therefore has to be taken into account during transmission.

The 3GPP architecture of the Universal Mobile Telecommunications System (UMTS) is composed of IP-based core network (CN) connected to the user equipment through UMTS Terrestrial Radio Access Network (UTRAN). The UTRAN consists of a set of radio network subsystem comprising a radio controller and one or more node base station. The network controller is responsible for the handover decisions that require signaling to the user equipment. Each subsystem is responsible for the resources of its set of cells and each node B has one or more cells.

3GPP Release 2000 on the architecture for an all IP mobile network proposed 2 reference architectures. The first option is based on packet technologies and IP telephony for simultaneous real time and non real time wireless mobile services. The second option support the IP based services and also supports the release 99 circuit switched terminals.

Multi-carrier modulation (MCM) is a data transmission technique where several subcarriers are employed to transport the user's data stream signal. Originally this technique was implemented using a bank of analogue Nyquist filters that provide a set of continuous-time orthogonal basis functions. Today using very fast and cost effective digital signal processors, multi-carrier modulation can be implemented using Discrete Fourier Transform (DFT) as the set of orthogonal subcarriers. This makes the technique very attractive.

Multi-Carrier Modulation (MCM) improves system capacity by making transmission more robust to frequency selective fading and enhances user spectral efficiency. The main drawbacks are mainly the difficult subcarriers' synchronisation in fading transmissions, the sensitivity to frequency offset is more pronounced than for a single carrier, and the sensitivity to non-linear amplification (peak factor problem). To gain the advantages of both schemes (CDMA & MCM), a combination known as multi-carrier CDMA (MC-CDMA) was proposed in 1993 taking after both CDMA & MCM schemes [Ref. MC-CDMA].

An MC-CDMA transmitter spreads the original data stream in the frequency domain over different subcarriers using a given spreading code. In this system the subcarriers convey the same information at one time. The MC-CDMA offers better frequency diversity to combat frequency selective fading.

The simplicity of the multi-carrier system is an important aspect in a cellular system especially for the down link receiver (mobile station). The modulation-demodulation is done by IDFT/DFT. A wavelet-based system can be used instead of DFT for the multi-carrier modulation. Wavelet transform has a property of time-frequency multi resolution. By choosing the right wavelet function and scaling function, the system can achieve the optimum resolution according to need.

An important characteristic of the wireless channel is that it is a randomly time-varying channel. There are multiple propagation paths from the transmitter to the receiver and these paths have randomly time-varying amplitudes and delays depending on the geographical location and mobility of the receiver relative to the transmitter. This time-variation in signal characteristics is usually referred to as fading and leads to significant performance degradation compared to the performance for a traditional deterministic channel with Additive White Gaussian Noise (AWGN) [PRO95]. Diversity signaling is a powerful technique to combat fading. This technique provides several independently faded replicas of the same information signal at the receiver and significantly reduces the probability that all the signal replicas suffer from amplitude fading at the same time. Frequency, time and space (antenna) diversity are some commonly used forms of diversity [PRO95]. In frequency and time diversity, the same transmission is repeated at multiple frequencies or at multiple time instants. Similarly, in space diversity systems, multiple copies of the signal are obtained using multiple receive and/or transmit antennas.

Implementing antenna arrays at the receiver and/or transmitter presents some calibration drawbacks and costs that may make such solution not feasible for the next few years. However, advanced signal processing techniques for delay acquisition, time varying channel tracking and multiuser detection techniques seem to be the mobile wireless industry's favorite candidates to enhance the overall system performances.

Multiuser signal processing refers to the joint processing of the signals of all the users in the system. This joint processing is very effective, since CDMA systems are interference-limited, and provides significant performance improvement compared to traditional single-user processing [VIT95], [PRO95], [PIC82] used in practice. The interest in multiuser signal processing for CDMA stemmed from Verdú's seminal work in [VER86], where he proposed and analyzed the optimal multiuser detector, or the Maximum Likelihood (ML) sequence detector. Unfortunately, this detector has a complexity exponential in the number of users. Therefore, over the last two decades, research in this area has focused on several sub-optimal solutions [LUP89], [LUP90], [VAR90], [XIE90], [MAD94], [ZVO96] which have linear complexity in the number of users and are more feasible to implement. Furthermore, other signal processing algorithms at the receiver, such as channel estimation and decoding, have also been studied recently from the multiuser perspective. Multiuser channel estimation is essential to obtain, for all the users, accurate estimates of the delays and attenuations that can later be used by the multiuser detectors. Multiuser decoding is required since every practical system is coded for error control. Optimal and sub-optimal multiuser algorithms for decoding have been proposed in [GIA96a], [GIA96b], [FAW96], [ALE99], [WAN99], [MOH98] while those for channel estimation have been proposed in [BEN96], [BEN98], [STR96], [MAD97], [SEN98a], [SEN98b], [CAM96], [MAN98], [MAN00].

In the sequel, we will describe the problematic and set our objectives that will make the subject of Chapter 3 and 4.

1.2 Problem Set up

Traditionally, in DS-CDMA system each user is assigned a unique code (sequence) that will differentiate it among other active users. The code is designed to ensure orthogonality among the different users. The related users' information bits are spread at the transmitter side using their appropriate codes, and then de-spread at the receiver side. The orthogonality ensures that, in an AWGN channel, a given user information sequence can be retrieved easily by simple matched filtering [PRO95].

Nevertheless, the channel is far to be that simple so that at the receiver side the codes are no longer orthogonal due to the channel impairments. Different users' signals undergo different random attenuations and delays [RAP96]. More than that multiple versions of the same user's signal are induced due to the multipath effect. Hence Multiple Access Interference (MAI) and Inter-Symbol Interference (ISI) are the major channel impairments the system designer is faced with.

Matched filter and Rake receivers are used as solutions for the second generation (2G) DS-CDMA system for the simple complexity they exhibit. Unfortunately, these receivers are MAI and ISI limited, the reason that pushes the research community to look for powerful solutions. Verdú [VER98] was among the pioneers through his suggestions for what is known as multi-user detection (MUD). Since then, many MUDs have been proposed, we may list for a matter of illustration, the minimum mean square error (MMSE), the decorrelator (the decision feedback decorrelator), the Parallel Interference Cancellor

(PIC), the Serial Interference Cancellor (SIC), the Interference Subspace Rejection (ISR) [AFF02] receivers and more [VER98]. These receivers exploit the interference structure of the received signal as function of the users' signatures and the channel parameters. They differ in performance and complexity due to the way this structure is processed.

Recently, combined multiuser detection and decoding has received considerable attention with its potential to improve the performance of a multiuser system to mesh that of a single user system. In [GIA96a], and [HAG97] it was shown that the optimum receiver for a CDMA system employing Forward Error Control (FEC) coding combines the trellis of both the multiuser detector and the FEC code. The complexity of this receiver is exponential in the product of the number of users and the constraint length of the code. This complexity makes the use of the optimal detector prohibitive for even small systems. In [GIA96b], some low-complexity receivers, which perform multiuser symbol detection and decoding either separately or jointly, are studied. Iterative decoding schemes for convolutionally or turbo coded CDMA systems were proposed in [REE98] and [WAN99]. The main difference on the structure of those schemes is the type of the Soft input Soft Output (SISO) multiuser detector used. In [REE98] and [MOH98], a full complexity SISO multiuser detector was proposed for convolutional coded synchronous CDAM systems, resulting in a computational complexity of $O(2^K)$ for the multiuser detector. Where K is the number of users.

Another type of iterative decoding is based on the soft interference cancellation [HAG96], [ALE98] and [WAN99] where the soft estimates of the code bits are represented by real number rather than hard-decision values and are combined to produce estimates of the LLR's (Log Likelihood Ratio) of the code bits.

These receivers exploit the signal structure along with the soft information on the users' information to effectively exploit all the potential signal processing gain that can be achieved using this signal structure. Beside MAI, ISI and multipath effects; near far problem[1], and time variations due to Doppler effect are of major concern for efficient system design.

From the other side, receivers can combat MAI by using multiuser channel estimation in [BEN96], [MOO94], [STR96], [MAD97], [SEN98a] and [SEN98b]. These algorithms are developed for CDMA systems with short spreading codes for various users that repeat every symbol. However, spreading codes used in practical CDMA systems have a period much larger than the symbol duration and are called long spreading codes. Therefore, most of the existing algorithms for channel estimation[2] are either inapplicable or need prohibitive computational resources.

Designing an efficient multiuser channel estimation and tracking for time varying multipath channel became a bottleneck. The current algorithms for channel estimation mainly rely on some kind of averaging assuming of course the channel coefficients constant at least during the period of interest (training period, defined window) and hence lack the needed tracking capability. Maximum Likelihood (ML) [BHA02] channel estimation operates on an averaged decision statistic over successive (windowed) matched-filters'

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[1] Near far problem refers to the situation where the receiver sees the different users contribution having large difference in their respective received powers.

[2] Channel estimation, mostly, refers to both delay acquisition/tracking and the relative complex attenuation estimation.

outputs for all users. Even though it provides satisfactory performance in slow channel variation, it is still not efficient in case of fast time varying channels.

Implementation complexity was and still is the driving factor, to favor one channel estimation algorithm from another, as far as the performances are satisfactory but not necessarily outstanding. The Correlator, due to the simple complexity it exhibits, is a good candidate. In order to improve the accuracy of the channel estimates provided by the Correlator, the obtained channel impulse response is further processed by employing a low pass filter called Channel Estimation Filter (CEF) [LIG93] (Correlator-CEF).

1.3 Objectives

The main objective of our work is to suggest new class of algorithms for iterative multi-user detection and decoding that exploit BPSK (Binary Phase Shift Keying) modulation. The emphasis on BPSK modulation stems from the fact that the current 3rd generation DS-CDMA systems, such as WCDMA, has adopted BPSK modulation technique. This "root" objective is split into sub-objectives that involve the design of new classes of SISO multiuser detectors. A new class of algorithms for multiuser channel estimation is of a major concern, since most of MUDs performances depend in one way or another on the channel estimates. These classes of algorithms are designed for the uplink DS-CDMA where the access to the users' information simultaneously is possible and the computational and power consumption aspects, as far as the real time constraints are satisfied, are not tight.

- **Acquisition and Tracking of Time Varying Channel**

In our work, a multiuser LMS-like (Least Mean Square) structure along with smoothing and prediction filters to improve tracking quality is suggested. The choice for such adaptation family stems from its low computational complexity and its regular structure favourable for an efficient Very Large Scale Integration (VLSI) implementation where parallelism and wave pipelining, among other techniques, are easily applied [MOR99], [MOZ99], [SAK98] and [MAS98]. They are computationally effective due to the even distribution of the computation load over each symbol duration and no extra computation is required at the end of the processing window or preamble.

- **Exploiting BPSK Modulation in Iterative (Turbo) Detection**

Driven by implementation issues, we suggested in [AHM04] a low complexity iterative detection structure at a computational load close to I-Rake, where I is the number of iterations. Triggered by performance improvement while keeping the complexity low, we developed in [AHM05a] a new iterative structure based on the MMSE design of a feed-forward and feed-back filters that can preserve the potential gains.

It is worth to mention that exploiting the BPSK modulation in designing linear receiver is not a new issue [TUL01], [BUZ01a], [HER05] but the idea of taking this into account when designing turbo detectors is not covered yet. One can notice that our contribution can be viewed as an extension of [WAN99] for BPSK signals. The contribution investigates, also, the additional potential gain that can be achieved when exploiting the BPSK modulation.

As a second objective, low complexity turbo detection receivers for joint detection and decoding for coded DS-CDMA systems utilizing BPSK modulation in multipath channels are proposed.

1.4 Methodology

In this section we briefly review (steps 1-3 in table 1.1) the related works on channel estimation and iterative (turbo) detection techniques. Furthermore, it is worth mentioning that an understanding of the wireless channel nature is required for an efficient communication system design and analysis. An important characteristic of the wireless channel is that it is a randomly time-varying channel. There are multiple propagation paths from the transmitter to the receiver and these paths have randomly time-varying amplitudes and delays [AHM01]. Basically, reliable detection can be performed by using multiuser channel estimation in [AHM02] for TDMA systems, and [BEN96], [SEN98a] and [SEN98b] wherein short spreading codes are used in the CDMA context. Some channel estimation algorithms have recently been proposed in [CAM96], [MAN98], [THO99], [WEI99], [XUZ00], [TOR97] for long code systems. The techniques in [THO99] and [WEI99] are based on the knowledge of the spreading sequences, channel estimates and bits of the interfering users, and use the interference cancellation and the MMSE approach, respectively. On the other hand, a maximum likelihood (ML) [BHA02] channel estimation operates on an averaged decision statistic over successive (windowed) matched-filters' outputs for all users.

The Correlator, because of its simple complexity, is a good candidate for a low complexity implementation. To improve the accuracy of the channel estimates provided by the Correlator, the channel impulse response obtained is further processed by employing a low pass filter, called a channel estimation filter (CEF) [LIG93] (Correlator-CEF). It is known that the Wiener filter is optimal as the CEF in a stationary channel in the MMSE sense [LIG93]. To design a Wiener filter, however, it is essential to know the power spectrum of the channel and noise, which may not be obtainable in real time. Thus it may be desirable to employ a brick-wall type lowpass filter such as the CEF, whose cut-off frequency is equal to the maximum Doppler frequency of the channel [LIG93], [SCH98a] and [SCH98b].

In our work, a multiuser-LMS-like structure along with smoothing and prediction filters to improve tracking quality is suggested. The choice for such an adaptation family stems from its low computational complexity and its regular structure. Like Kalman-based channel estimation methods, an autoregressive stochastic model of correlated Rayleigh fading processes is used but indirectly embedded into the design. The attractive property of the proposed structure is the unique filter settings used for a large range of mobile speeds. The design process is based on the following methodology: first, a p-th order autoregressive stochastic model [BAD01] is designed for an average Doppler profile; second, the resulting model along with the admissible parameter-drift-to-noise floor ratio (υ) [LIN01] are used to design the smoothing/prediction coefficients using Wiener LMS design methodology [LIN01]; finally, a multiuser-LMS structure is augmented with an extra smoothing/prediction procedure.

Since, in practice, most CDMA systems employ error control coding and interleaving, recent works in this area have addressed multiuser detection for coded CDMA systems. In [GIA96b], some low-complexity receivers that perform multiuser symbol detection and decoding either separately or jointly are studied. Iterative decoding schemes for convolutionally or turbo coded CDMA systems are proposed in [REE98] and [WAN99]. The main difference in the structure of those schemes is the type of the soft-input/soft-output (SISO) multiuser detector used. In [REE98] and [MOH98], a full complexity SISO multiuser detector is proposed for convolutional coded synchronous CDMA systems, resulting in a computational complexity of $O(2^K)$ for the multiuser detector, where K is the number of users. Based on the concept of a time varying convolutional code, forward and backward recursions similar to those of the BCJR algorithm [BAH74] were proposed for the SISO multiuser detector. Another type of iterative decoding is based on the soft interference cancellation [HAG96b], [ALE98], [WAN99]. [WAN99] suggests a low complexity SISO multiuser detector based on the soft instantaneous MMSE interference cancellation/suppression, based, in turn, on the a priori likelihood ratios of the code bits of all users provided by the SISO channel decoder from the previous stages.

It is worth mentioning that exploiting the BPSK modulation for the design of a linear receiver is not a new issue [TUL01], [BUZ01a], [HER05]; it has not yet, however, been considered in the design of turbo detectors. Our contribution may, therefore, be viewed as a development of the results in [WAN99], [GAM00] and [AFF02] for BPSK signals. The contribution investigates, also, the additional potential gain that can be achieved when exploiting the BPSK modulation.

In the present work two low complexity turbo detection receivers for convolutionally coded DS-CDMA systems utilizing BPSK modulation in multipath channels are presented. The first receiver is based on a soft Interference Canceller (IC) followed by a MMSE filters, whose coefficients are thought to be the solution of an MMSE-optimization problem based on a (forced) real valued filter output rather than a complex one as in conventional MMSE receivers used in [WAN99]. In the second structure, we reformulate the traditional constrained filter design by considering a real valued filter output along with real valued constraints.

It is worth to mention that an accurate and fast DS-CDMA simulation platform is required to conduct our design, analysis and draw conclusions regarding the different estimation and detection techniques implemented over a DS-CDMA environment. Such a platform is implemented in MATLAB and C to take advantage of the coding ease and fast simulation these two languages may offer.

Next, relevant algorithms for both channel estimation and iterative detection, from the literature, are implemented. Their respective performance and complexity analysis is conducted and commented.

Up on the last performance/complexity conclusions made above, new techniques are devised. The target to reach is, as mentioned above in our objectives, to enhance the system performance while keeping the implementation complexity as low as possible, or at least the additional gains may justify the added computational burden to the system.

Many thanks to Axiocom Inc. who made their WCDMA platform "MONARK©" and cdma2000 platform "MACHAON©" available for simulations. The table 1.1 summarizes the steps.

1.5 Thesis Organization

This thesis is organized as follows; in Chapter 2, a DS-CDMA signal model is presented. Two dual models are presented; the first one is suitable for detection while the second one is convenient for channel estimation. In Chapter 3, base on WLMS, a channel estimation and tracking algorithm, from a multiuser approach, is presented. Chapter 4 is dedicated to the derivation of a class of iterative (turbo) multiuser detection for coded DS-CDMA systems. Finally, the conclusions are drawn and open problems for future researches are presented.

Table 1. Methodology in five steps

Step 1.	Bibliographical search on the related materials on DS-CDMA system in general and MUD, soft MUD, channel coding and estimation in particular.
Step 2.	Evaluation of the known algorithms for multiuser detection, channel estimation and iterative detection and decoding in a common framework regarding multipath channel, MAI and near far situation. This step will help to identify the methods for reference.
Step 3.	Proposition of an algorithm for multiuser channel acquisition and tracking.
Step 4.	Proposition of a set of algorithms for iterative (turbo) multiuser detection in DS-CDMA systems utilizing BPSK modulation scheme.
Step 5.	Validation and performance evaluation of the proposed algorithms in different scenarios using a WCDMA platform, a cdma2000 platform for channel estimation as well as a DS-CDMA platform for iterative detection algorithms.

2.

DS-CDMA SIGNAL MODEL

In this section we will present a system model that efficiently incorporates the effect of multipaths in the multi-antenna context for DS-CDMA systems. As the algorithms developed in the next chapters assume a single antenna at the receiver, the multi-sensor model is reduced and some modifications are incorporated to account for long codes and construct two compact models suitable for (1) channel estimation, described in Chapter 3, and (2) iterative (Turbo) detection, to be used in Chapter 4.

The same reduced single sensor model is modified to account for long spreading codes. The later is developed in Chapter 3 for convenience.

2.1 Multisensor DS-CDMA Signal Model

Consider a convolutionally coded DS-CDMA system with K users, using BPSK modulation and signaling over their respective multipath channels with AWGN. The binary information $\{d_k(m)\}$ for user $k = 1, 2, ..., K$, are convolutionally encoded with rate R_k. An interleaver is used to reduce error burst effect at the input of the decoder. The interleaved code-bits of the kth user are mapped into BPSK symbol resulting in data symbols $b_k(i)$. Each data symbol $b_k(i)$ is then modulated by a spreading waveform $c_k(t)$ of duration T. The complex base band representation of the kth user's transmitted signal is given by

$$s_k(t) = \sqrt{E_k}\sum_i b_k(i)c_k(t-iT) \tag{2.1}$$

where E_k is the transmitted power, $b_k(i) \in \{-1,+1\}$ is the ith transmitted symbol and $c_k(t)$ is a normalized spreading waveform given by

$$c_k(t) = \sum_{n=0}^{N-1} c_{k,n}\Pi(t-nT_c) \tag{2.2}$$

where $c_{k,n} \in \left\{-1/\sqrt{N}, +1/\sqrt{N}\right\}$ and the chip pulse $\Pi(t)$ is a rectangular pulse of duration T_c where $\frac{T}{T_c} = N$ is an integer. The rectangular pulse shaping waveform is used in the sequel to develop simple models, while a square root raised cosine is used for simulations.

The front end of the receiver consists of an array of M antennas. The corresponding array response vector, which is the response to a propagating plane wave impinging on the array at an angle θ, with respect to the plane perpendicular to the plane of the array

$$\left[\rho^{(1)}(\theta),\cdots,\rho^{(\mathrm{m})}(\theta),\cdots,\rho^{(\mathrm{M})}(\theta)\right]^{\mathrm{T}}, \tag{2.3}$$

and is determined by the geometry of the array. We may assume that the time taken by the signal to traverse the physical array is much smaller than the inverse of the message bandwidth and hence, the envelope characteristics of the signal do not vary across the array.

We assume that the channel, for each user consists k, of L_k distinct and resolvable propagation paths [BHA02]. The impulse response $h_k(t)$ of the channel seen by user k is given by

$$h_k(t) = \sum_{\ell=1}^{L_k} w_{k,\ell}\delta(t - \tau_{k,\ell}), \tag{2.4}$$

where $w_{k,\ell}$ is the complex amplitude with which the ℓth path of the kth user is received and includes contributions from the channel attenuation and the phase offset and $\tau_{k,\ell}$ is the relative delay with respect to a reference at the receiver. The delay spread of each user is assumed to be less than half a symbol period. The delay is not constrained to be multiples of a chip. The system model assumes that the delay consists of an integer part in multiples of a chip as well as a fractional part which is a fraction of a chip.

Consequently, the received signal at the base station is a superposition of multiple copies of attenuated and delayed signals transmitted by all the K users. Therefore, the signal at the mth sensor is given by

$$r^{(\mathrm{m})}(t) = \sum_{k=1}^{K}\sum_{\ell=1}^{L_k} w_{k,\ell}\rho^{(\mathrm{m})}(\theta_{k,\ell})s_k(t - \tau_{k,\ell}) + v^{(\mathrm{m})}(t) \tag{2.5}$$

where $\theta_{k,\ell}$ is the direction of arrival of the ℓth path of the kth user with respect to an axis in the plane of the array and the additive noise $v^{(\mathrm{m})}(t)$ is assumed to be Gaussian.

The continuous time signal at each sensor is then discretized by sampling the output of a chip-matched filter, at the chip rate. The chip-matched filtering is a simple integrate and dump operation, over a time interval equal to the chip period

$$r^{(\mathrm{m})}[n]=\frac{1}{T_c}\int_{nT_c}^{(n+1)T_c} r^{(\mathrm{m})}(t)\,dt \tag{2.6}$$

The observation vectors $\mathbf{r}_i^{(\mathrm{m})}$ are formed by collecting together N successive $r^{(\mathrm{m})}[n]$. Each observation vector corresponds to a time interval equal to the (data or pilot) bit period. Since each spreading code is periodic with period N, $r^{(\mathrm{m})}[n]$ are wide-sense cyclostationary, and the observation vectors $\mathbf{r}_i^{(\mathrm{m})}$ are wide-sense stationary. The observation vector $\mathbf{r}_i^{(\mathrm{m})} \in \mathbb{C}^{\mathrm{N}\times 1}$ at time i is given by

$$\mathbf{r}_i^{(\mathrm{m})}=\left[r^{(\mathrm{m})}[iN], r^{(\mathrm{m})}[iN+1], \cdots, r^{(\mathrm{m})}[iN+N-1]\right]^{\mathrm{T}}. \tag{2.7}$$

The system is asynchronous and the receiver has an arbitrary timing reference that will not be aligned to actual transmitted bit boundaries. Hence, each observation vector can be viewed as a linear combination of $2K$ signal vectors- two components due to the past and current bits as shown in Figure 2.1.

In Figure 2.1, the received vector $\mathbf{r}_i^{(\mathrm{m})}$ overlaps the two adjacent transmitted bits b_{i-1} and b_i. So the contribution of the spreading code of the user to vector $\mathbf{r}_i^{(\mathrm{m})}$ appears in two parts—the right part (shown as R) of the bit b_{i-1} and the left part (shown as L) of the bit b_i.

Expressing $\mathbf{r}_i^{(\mathrm{m})}$ in terms of its various components

$$\mathbf{r}_i^{(\mathrm{m})}=\mathbf{A}^{(\mathrm{m})}\mathbf{b}_i+\mathbf{v}_i^{(\mathrm{m})}, \quad \mathbf{v}_i^{(\mathrm{m})} \sim N\left(\mathbf{0},\mathbf{K}^{(\mathrm{m})}\right) \tag{2.8}$$

The $2K$ length vector $\mathbf{b}_i$ is of the form: $\mathbf{b}_i=\left[b_{1,i-1}, b_{1,i}, \cdots, b_{k,i-1}, b_{k,i}, \cdots, b_{K,i-1}, b_{K,i}\right]^{\mathrm{T}}$ where $b_{k,i}$ is the ith bit of the kth user.

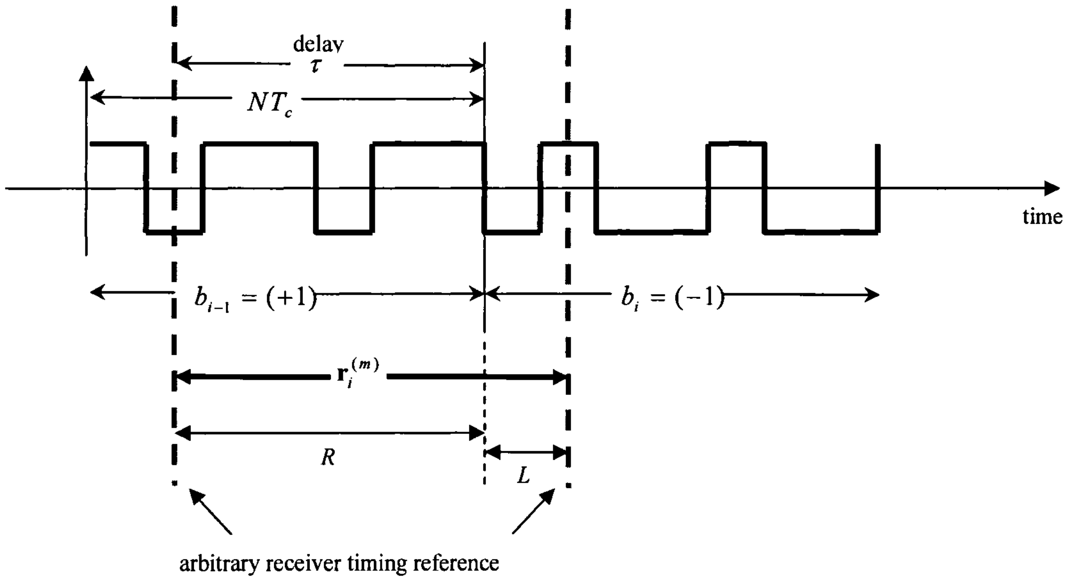


Figure 2.1. Arbitrary timing reference at the receiver

Let us consider the columns of the matrix $\mathbf{A}^{(m)}$. As explained in Figure 2.1, the contribution of the spreading code of the user to vector $\mathbf{r}_i^{(m)}$ appears in two parts – R and L. The delays of the users with respect to the receiver reference are assumed to be such that each observation vector can be considered to have contributions from two bits of each user only. This assumption has been adopted for easy mathematical development. However, this assumption may be removed by adding more than two columns for each user. For example, for the case of the delay spread less than half of the symbol period, each user can contribute one, two or three vectors to $\mathbf{A}^{(m)}$. Hence, matrix $\mathbf{A}^{(m)}$ has columns corresponding to these two parts, denoted by the superscripts R and L

$$\mathbf{A}^{(m)} = \left[\mathbf{a}_1^{\mathrm{R}} \quad \mathbf{a}_1^{\mathrm{L}} \quad \cdots \quad \mathbf{a}_K^{\mathrm{R}} \quad \mathbf{a}_K^{\mathrm{L}}\right]. \tag{2.9}$$

Each of the columns $\mathbf{a}_k^{\mathrm{R}}$ and $\mathbf{a}_k^{\mathrm{L}}$ for each user k are functions of the corresponding delays $\left(\tau_{k,1},\cdots,\tau_{k,L_k}\right)$, attenuation factors $\left(w_{k,1},\cdots,w_{k,L_k}\right)$, and array responses

$\left(\rho_{k,1}^{(\mathrm{m})},\cdots,\rho_{k,L_k}^{(\mathrm{m})}\right)$. Here, $\rho_{k,\ell}^{(\mathrm{m})}=\rho^{(\mathrm{m})}\left(\theta_{k,\ell}\right)$. The superscripts R and L refer to Right and Left shifts respectively.

Let $\tau_{k,\ell}/T_c=q_{k,\ell}+\gamma_{k,\ell}$, $q_{k,\ell}\in\{0,1,\ldots,N-1\}$, $\gamma_{k,\ell}\in[0\,1)$, we have [BHA02]

$$\begin{aligned}\mathbf{a}_k^{\mathrm{R}}&=\sum_{\ell=1}^{L_k}\left[w_{k,\ell}\left\{\left(1-\gamma_{k,\ell}\right)\mathbf{c}_k^{\mathrm{R}}\left[q_{k,\ell}\right]+\left(\gamma_{k,\ell}\right)\mathbf{c}_k^{\mathrm{R}}\left[q_{k,\ell}+1\right]\right\}\rho_{k,\ell}^{(\mathrm{m})}\right]\\\mathbf{a}_k^{\mathrm{L}}&=\sum_{\ell=1}^{L_k}\left[w_{k,\ell}\left\{\left(1-\gamma_{k,\ell}\right)\mathbf{c}_k^{\mathrm{L}}\left[q_{k,\ell}\right]+\left(\gamma_{k,\ell}\right)\mathbf{c}_k^{\mathrm{L}}\left[q_{k,\ell}+1\right]\right\}\rho_{k,\ell}^{(\mathrm{m})}\right]\end{aligned}\tag{2.10}$$

where $\mathbf{c}_k^{\mathrm{R}}[q]$ and $\mathbf{c}_k^{\mathrm{L}}[q]$ are the spreading codes shifted by an integer (multiples of chips) delays. Note that $\mathbf{c}_k^{\mathrm{R}}[N]$ is an all-zero vector when the observation vector is aligned to a symbol boundary

$$\begin{aligned}\mathbf{c}_k^{\mathrm{R}}[q]&=\begin{bmatrix}c_{k,N-q}&\cdots&c_{k,N-1}&0&\cdots&0\end{bmatrix}^{\mathrm{T}}\\\mathbf{c}_k^{\mathrm{L}}[q]&=\begin{bmatrix}0&\cdots&0&c_{k,0}&\cdots&c_{k,N-q-1}\end{bmatrix}^{\mathrm{T}}\end{aligned}\tag{2.11}$$

Note that (2.9) and (2.10) show that the columns of $\mathbf{A}^{(\mathrm{m})}$ are linear combinations of $\mathbf{c}_k[q]$'s. Let $\mathbf{U}_k^{\mathrm{R}}\in\mathbb{C}^{\mathrm{N}\times\mathrm{N}}$ and $\mathbf{U}_k^{\mathrm{L}}\in\mathbb{C}^{\mathrm{N}\times\mathrm{N}}$ be matrices formed from the spreading codes, delayed by all possible integer delays. Let $\mathbf{z}_k^{(\mathrm{m})}\in\mathbb{C}^{\mathrm{N}\times 1}$ be the composite channel impulse response vector to the mth sensor, consisting of the weight vectors in the linear combinations of the $\mathbf{c}_k[q]$'s, expressed in (2.10). Hence, we can write $\mathbf{U}_k^{\mathrm{R}}$, $\mathbf{U}_k^{\mathrm{L}}$ and $\mathbf{z}_k^{(\mathrm{m})}$ as

$$\begin{aligned}\mathbf{U}_k^{\mathrm{R}}&=\left[\mathbf{c}_k^{\mathrm{R}}[0]\cdots\mathbf{c}_k^{\mathrm{R}}[N-1]\right]\\\mathbf{U}_k^{\mathrm{L}}&=\left[\mathbf{c}_k^{\mathrm{L}}[0]\cdots\mathbf{c}_k^{\mathrm{L}}[N-1]\right]\end{aligned}\tag{2.12}$$

$$\mathbf{z}_k^{(\mathrm{m})}=\mathbf{p}_k^{(\mathrm{m})}\circ\mathbf{h}_k\tag{2.13}$$

where $\mathbf{h}_k$ and $\mathbf{p}_k^{(\mathrm{m})}$ are given by

$$\mathbf{h}_k = \left[\begin{array}{c} 0 \\ \vdots \\ w_{k,1}\left(1-\gamma_{k,1}\right) \\ w_{k,1}\gamma_{k,1} \\ 0 \\ \vdots \\ w_{k,2}\left(1-\gamma_{k,2}\right) \\ w_{k,2}\gamma_{k,2} \\ \vdots \\ w_{k,L_k}\left(1-\gamma_{k,L_k}\right) \\ w_{k,L_k}\gamma_{k,L_k} \\ 0 \\ \vdots \\ 0 \end{array}\right] \begin{array}{l} \\ \\ \leftarrow \quad q_{k,1} \\ \leftarrow q_{k,1}+1 \\ \\ \\ \leftarrow \quad q_{k,2} \\ \leftarrow q_{k,2}+1 \\ \\ \leftarrow \quad q_{k,L_k} \\ \leftarrow q_{k,L_k}+1 \\ \\ \\ \\ \end{array} \text{ and } \mathbf{p}_k^{(\mathrm{m})} = \left[\begin{array}{c} 0 \\ \vdots \\ \rho_{k,1}^{(\mathrm{m})} \\ \rho_{k,1}^{(\mathrm{m})} \\ 0 \\ \vdots \\ \rho_{k,2}^{(\mathrm{m})} \\ \rho_{k,2}^{(\mathrm{m})} \\ \vdots \\ \rho_{k,L_k}^{(\mathrm{m})} \\ \rho_{k,L_k}^{(\mathrm{m})} \\ 0 \\ \vdots \\ 0 \end{array}\right] \begin{array}{l} \\ \\ \leftarrow \quad q_{k,1} \\ \leftarrow q_{k,1}+1 \\ \\ \\ \leftarrow \quad q_{k,2} \\ \leftarrow q_{k,2}+1 \\ \\ \leftarrow \quad q_{k,P} \\ \leftarrow q_{k,P}+1 \\ \\ \\ \\ \end{array} \tag{2.14 and 2.15}$$

Here the operator "$\circ$" denotes the element-wise multiplication of the two vector operands. Now, the columns of $\mathbf{A}^{(\mathrm{m})}$ can be expressed in terms of $\mathbf{U}_k^{\mathrm{R}}$, $\mathbf{U}_k^{\mathrm{L}}$ and $\mathbf{z}_k^{(\mathrm{m})}$ as

$$\begin{aligned} \mathbf{a}_k^{\mathrm{R}} &= \mathbf{U}_k^{\mathrm{R}}\mathbf{z}_k^{(\mathrm{m})} \\ \mathbf{a}_k^{\mathrm{L}} &= \mathbf{U}_k^{\mathrm{L}}\mathbf{z}_k^{(\mathrm{m})}. \end{aligned} \tag{2.16}$$

We can now rewrite matrix $\mathbf{A}^{(\mathrm{m})}$ as

$$\mathbf{A}^{(\mathrm{m})} = \left[\mathbf{U}_1^{\mathrm{R}}\mathbf{z}_1^{(\mathrm{m})}\ \mathbf{U}_1^{\mathrm{L}}\mathbf{z}_1^{(\mathrm{m})} \cdots \mathbf{U}_K^{\mathrm{R}}\mathbf{z}_K^{(\mathrm{m})}\ \mathbf{U}_K^{\mathrm{L}}\mathbf{z}_K^{(\mathrm{m})}\right] \tag{2.17}$$

hence, the observation vector $\mathbf{r}_i$ of length MN, contains signals from all antennas $\mathbf{r}_i = \left[\mathbf{r}_i^{(1)^{\mathrm{T}}}, \cdots, \mathbf{r}_i^{(\mathrm{m})^{\mathrm{T}}}, \cdots, \mathbf{r}_i^{(\mathrm{M})^{\mathrm{T}}}\right]^{\mathrm{T}}$ is

$$\mathbf{r}_i = \begin{bmatrix} \mathbf{U}_1^{\mathrm{R}}\mathbf{z}_1^{(1)} & \mathbf{U}_1^{\mathrm{L}}\mathbf{z}_1^{(1)} & \cdots & \mathbf{U}_K^{\mathrm{R}}\mathbf{z}_K^{(1)} & \mathbf{U}_K^{\mathrm{L}}\mathbf{z}_K^{(1)} \\ & & \vdots & & \\ \mathbf{U}_1^{\mathrm{R}}\mathbf{z}_1^{(\mathrm{M})} & \mathbf{U}_1^{\mathrm{L}}\mathbf{z}_1^{(\mathrm{M})} & \cdots & \mathbf{U}_K^{\mathrm{R}}\mathbf{z}_K^{(\mathrm{M})} & \mathbf{U}_K^{\mathrm{L}}\mathbf{z}_K^{(\mathrm{M})} \end{bmatrix} \mathbf{b}_i + \mathbf{v}_i \tag{2.18}$$

where the noise $\mathbf{v}_i$ is formed from the components $\mathbf{v}_i^{(\mathrm{m})}$ and is assumed to have a covariance of $\mathbf{K} = \sigma^2 \mathbf{I}_{MN}$.

Let $\mathbf{C}_k^{\mathrm{R}} \in \mathbb{C}^{\mathrm{MN} \times \mathrm{MN}}$ and $\mathbf{C}_k^{\mathrm{L}} \in \mathbb{C}^{\mathrm{MN} \times \mathrm{MN}}$ be matrices defined as follows

$$\mathbf{C}_k^{\mathrm{R}} = \mathbf{I}_M \otimes \mathbf{U}_k^{\mathrm{R}}, \quad \mathbf{C}_k^{\mathrm{L}} = \mathbf{I}_M \otimes \mathbf{U}_k^{\mathrm{L}}. \tag{2.19}$$

Here the operator "$\otimes$" represents the Kronecker product of two matrices. Also, let $\mathbf{z}_k \in \mathbb{C}^{\mathrm{MN} \times 1}$ be a vector defined as

$$\mathbf{z}_k = \left[\mathbf{z}_k^{(1)^{\mathrm{T}}}, \mathbf{z}_k^{(2)^{\mathrm{T}}}, \cdots, \mathbf{z}_k^{(\mathrm{m})^{\mathrm{T}}}, \cdots, \mathbf{z}_k^{(\mathrm{M})^{\mathrm{T}}} \right]^{\mathrm{T}}. \tag{2.20}$$

Using these matrices defined above, the columns of combined code and channel matrix, corresponding to user k, in the expression for $\mathbf{r}_i$ (2.18) can be written as $\mathbf{C}_k^{\mathrm{R}} \mathbf{z}_k$ and $\mathbf{C}_k^{\mathrm{L}} \mathbf{z}_k$. This helps us to rewrite the expression for $\mathbf{r}_i$ as

$$\mathbf{r}_i = \left[\mathbf{C}_1^{\mathrm{R}} \mathbf{z}_1 \ \mathbf{C}_1^{\mathrm{L}} \mathbf{z}_1 \quad \cdots \quad \mathbf{C}_K^{\mathrm{R}} \mathbf{z}_K \ \mathbf{C}_K^{\mathrm{L}} \mathbf{z}_K \right] \mathbf{b}_i + \mathbf{v}_i = \mathbf{C}\mathbf{Z}\mathbf{b}_i + \mathbf{v}_i \tag{2.21}$$

where $\mathbf{C} = \left[\mathbf{C}_1^{\mathrm{R}} \ \mathbf{C}_1^{\mathrm{L}} \quad \cdots \quad \mathbf{C}_K^{\mathrm{R}} \ \mathbf{C}_K^{\mathrm{L}} \right]$ and

$$\mathbf{Z} = \begin{bmatrix} \mathbf{z}_1 & \mathbf{0} & \mathbf{0} & & \cdots & & \mathbf{0} & \mathbf{0} \\ \mathbf{0} & \mathbf{z}_1 & & & & & & \mathbf{0} \\ \mathbf{0} & & \ddots & & \vdots & & & \\ & & & \mathbf{z}_k & & & & \vdots \\ \vdots & & & & \mathbf{z}_k & & & \vdots \\ \vdots & & & \vdots & & \ddots & & \\ & & & & & & \mathbf{z}_K & \mathbf{0} \\ \mathbf{0} & \mathbf{0} & & \cdots & \cdots & & \mathbf{0} & \mathbf{z}_K \end{bmatrix}$$

The advantage to be gained from expressing $\mathbf{r}_i$ as in (2.21) is that it allows for an easy modeling of multiple propagation paths without increasing the size of any of the matrices involved. Increasing the sizes of the matrices involved is undesirable because it is directly related to an increase in the computational load of the algorithms that use the model. An alternative method to model multipaths is to treat the multiple paths as different virtual users and have two additional columns in matrix $\mathbf{A}$ for each extra path of each user. Hence, as the number of paths increases the size of matrix $\mathbf{A}$ also increases. However, in (2.21), the size of matrices $\mathbf{C}$ and $\mathbf{Z}$ does not increase as L increases; instead matrix $\mathbf{Z}$ becomes more dense. In effect, we discretize the multipath delay axis of the impulse response [SEN98a] into N equal time delay segments, each corresponding to a chip interval. Any number of multipath signals received within the nth bin is represented by a single resolvable multipath component.

2.2 Single Antenna Signal Model

A special case can easily be deduced by setting $M=1$, hence, the observation vector of length N at time instant i, $\mathbf{r}_i=\left[r(iN+N)\ r(iN+N-1)\ \ldots\ r(iN+1)\right]^{\mathrm{T}}$ with $\mathbf{r}_i\in\mathbb{C}^{\mathrm{N}\times 1}$, is [BEN98], [ERT98]

$$\mathbf{r}_i=\mathbf{CHb}_i+\mathbf{v}_i \tag{2.22}$$

In this case we assume that $\mathbf{v}_i\sim N\left(\mathbf{0},\sigma^2\mathbf{I}_{N\times N}\right)$, $\mathbf{b}_i=\left[b_{1,i-1}\,b_{1,i}\,b_{2,i-1}\,b_{2,i}\ldots b_{K,i-1}\,b_{K,i}\right]^{\mathrm{T}}$, $\mathbf{C}=\left[\mathbf{C}_1^{\mathrm{R}},\mathbf{C}_1^{\mathrm{L}},\cdots,\mathbf{C}_K^{\mathrm{R}},\mathbf{C}_K^{\mathrm{L}}\right]$ with

$$\mathbf{C}_k^{\mathrm{R}}=\begin{bmatrix} 0 & c_{k,N-1} & c_{k,N-2} & \cdots\cdots & \cdots & c_{k,2} & c_{k,1} \\ 0 & 0 & c_{k,N-1} & \vdots\ \vdots & \vdots & c_{k,3} & \vdots \\ \vdots & \vdots & \vdots & \vdots\ \vdots & \vdots & \vdots & c_{k,N-1} \\ 0 & 0 & \cdots & \cdots\cdots & \cdots & 0 & 0 \end{bmatrix},$$

$$\mathbf{C}_k^{\mathrm{L}} = \begin{bmatrix} c_{k,0} & 0 & 0 & \cdots\cdots & \cdots & 0 & 0 \\ c_{k,1} & c_{k,0} & 0 & \vdots\;\vdots & \vdots & \vdots & \vdots \\ \vdots & \vdots & \vdots & \vdots\;\vdots & \vdots & c_{k,0} & 0 \\ c_{k,N-1} & c_{k,N-2} & c_{k,N-3} & \cdots\cdots & \cdots & c_{k,1} & c_{k,0} \end{bmatrix}$$

and

$$\mathbf{H} = \begin{bmatrix} \mathbf{h}_1 & 0 & 0 & \cdots & \cdots & \cdots & 0 & 0 \\ 0 & \mathbf{h}_1 & 0 & & & & \vdots & 0 \\ 0 & 0 & \mathbf{h}_2 & 0 & & & \vdots & \vdots \\ \vdots & 0 & 0 & \mathbf{h}_2 & & & \vdots & \vdots \\ \vdots & & & 0 & \ddots & & \vdots & \vdots \\ \vdots & & \vdots & \vdots & & \ddots & 0 & 0 \\ 0 & & & & & & \mathbf{h}_K & 0 \\ 0 & 0 & \cdots & \cdots & \cdots & 0 & 0 & \mathbf{h}_K \end{bmatrix},$$

where $\mathbf{h}_k$ is given in (2.14).

The asynchronous model presented above is introduced in [BEN98] and extended in [ERT98] and [WAN99] to a more general case. The asynchronous model in (2.22) as stated is not useful for detection as it may be noticed that at any instant i the observations $\mathbf{r}_i$ does partially contain the information about the users' bits $b_{k,i}$, $k = 1, \ldots, K$, $i = 1, \ldots$, hence the observations $\mathbf{r}_{i+1}$ at $i+1$ are needed. The extended signal model will be

$$\begin{bmatrix} \mathbf{r}_i \\ \mathbf{r}_{i+1} \end{bmatrix} = \begin{bmatrix} [\mathbf{CH}]_0 & [\mathbf{CH}]_1 & \mathbf{0}_{N\times K} \\ \mathbf{0}_{N\times K} & [\mathbf{CH}]_0 & [\mathbf{CH}]_1 \end{bmatrix} \mathbf{b}_{i/i+1} + \mathbf{v}_{i/i+1} = \mathbf{r}_{i/i+1} \tag{2.23}$$

where $[\mathbf{CH}]_0$ and $[\mathbf{CH}]_1$ are matrices constructed from the odd and even numbered columns of $\mathbf{CH}$, respectively, $\mathbf{b}_{i/i+1} = \left[b_{1,i-1}, b_{2,i-1}, \cdots b_{k,i-1}, \cdots, b_{K,i-1},\ b_{1,i}, b_{2,i}, \cdots b_{k,i}, \cdots, b_{K,i}, b_{1,i+1}, b_{2,i+1}, \cdots b_{k,i+1}, \cdots, b_{K,i+1}\right]^{\mathrm{T}}$ and $\mathbf{v}_{i/i+1} = [\mathbf{v}_i^{\mathrm{T}}\ \mathbf{v}_{i+1}^{\mathrm{T}}]^{\mathrm{T}}$. In the sequel, let $\mathbf{A} = \begin{bmatrix} [\mathbf{CH}]_0 & [\mathbf{CH}]_1 & \mathbf{0}_{N\times K} \\ \mathbf{0}_{N\times K} & [\mathbf{CH}]_0 & [\mathbf{CH}]_1 \end{bmatrix}$ for brevity.

While equation (2.23) is used to represent the received vector for detection, we rewrite the received vector for channel estimation as

$$\mathbf{r}_i = \mathbf{C}_i \mathbf{B}_i \mathbf{z} + \mathbf{v}_i \tag{2.24}$$

where $\mathbf{z} = [\mathbf{z}_1^{\mathrm{T}}\ \mathbf{z}_2^{\mathrm{T}} \cdots \mathbf{z}_K^{\mathrm{T}}]^{\mathrm{T}}$ is a $KN \times 1$ channel response vector, $\mathbf{z}_k = \mathbf{h}_k$ and $\mathbf{B}_i$ is a $2KN \times KN$ matrix defined as

$$\mathbf{B}_i = \begin{bmatrix} b_{1,i-1} & 0 & 0 & \cdots & 0 & 0 \\ b_{1,i} & 0 & 0 & \cdots & 0 & 0 \\ 0 & b_{2,i-1} & 0 & \cdots & 0 & 0 \\ 0 & b_{2,i} & 0 & \cdots & 0 & 0 \\ \vdots & \vdots & & & & \vdots \\ \vdots & \vdots & & & & \vdots \\ 0 & 0 & 0 & \cdots & 0 & b_{K,i-1} \\ 0 & 0 & 0 & \cdots & 0 & b_{K,i} \end{bmatrix} \otimes \mathbf{I}_N$$

where $\otimes$ denotes the Kronecker product and $\mathbf{I}_N$ is the identity matrix of rank N. Thus we estimate N channel parameters for each user. This effective channel response accounts for all the paths under the assumption that their delays are within one symbol duration. The number of non-zero coefficients in this effective channel response vector is determined by the number of paths and delays.

3. Multipath Channel Estimation and Tracking Algorithm for DS-CDMA systems: Methodology and Design

This chapter addresses the problem of asynchronous multiuser delays acquisition and time varying channel tracking for multipath channels in DS-CDMA systems. A multiuser-LMS-like structure along with smoothing/prediction filters to improve tracking quality is suggested [AHM05b]. The choice for such adaptation family stems from its low computational complexity and its regular structure suitable for an efficient VLSI implementation. The attractive property of the proposed structure is the unique filter settings used for a large range of mobile speeds and channel types for all users accessing the system. A performance versus complexity analysis is conducted, in both cdma2000 and WCDMA environments, over highly interesting settings including different data rates, channel types (pedestrian and vehicular) and mobile speeds. The impact of the power control imperfections is analyzed through simulations. The proposed multiuser LMS structure offers substantial improvements compared to the Correlator method at a cost of 2 to 3 times more computational resources [AHM05b].

3.1 Related Works

CDMA systems are inherently interference limited. Receivers can combat MAI by using multiuser channel estimation in [BEN96], [MOO94], [STR96], [MAD97], [SEN98a] and [SEN98b]. These algorithms are developed for CDMA systems with short spreading that repeats every symbol. However, spreading codes used in practical CDMA systems have a period much larger than the symbol duration and are called long spreading codes. Most of the existing algorithms are therefore either inapplicable or require prohibitive computational resources.

Some channel estimation algorithms have recently been proposed in [CAM96], [MAN98], [THO99], [WEI99], [XUZ00], [TOR97] for long code systems. The techniques in [THO99] and [WEI99] are based on a knowledge of the spreading sequences, channel estimates and bits of the interfering users, and use the interference cancellation and the MMSE approach, respectively. In [MAN98], an acquisition scheme for a single user entering the system is developed using the knowledge of the spreading sequence and delays of the interfering users, who have already been acquired, without using their bit decisions. This leads to an estimator similar in complexity to the linear decorrelating detector. Blind estimation on the complex channel amplitudes is studied in [XUZ00] and [TOR97] assuming knowledge of the delays of the various propagation paths for the interferers and [WEI99] develops channel estimation algorithms for synchronous downlink channels.

Designing efficient multiuser channel estimation and tracking for time varying multipath channels is a major concern. The current algorithms for channel estimation rely mainly on some type of averaging, assuming, of course, that the channel coefficients remain constant at least during the period of interest (training period, predefined window) and therefore lack the needed tracking capability. A maximum likelihood (ML) [BHA02] channel estimation operates on an averaged decision statistic over successive (windowed) matched-filters' outputs for all users. With respect to implementation, Bhashyam and Aazhang in [BHA02] designed an ML approach for long codes, applying gradient-based methods to approximate the ML solution and evenly distribute the computational burden over each sample, and thereby offering good tracking capabilities for slow channel variations. This method can be viewed as an iterative search for the *composite channel impulse response* of all users that minimizes a gradient with an "identity implementation law."

Implementation complexity was and still is the driving factor for preferring one channel estimation algorithm over another, as long as performances are satisfactory but not necessarily outstanding. The Correlator, because of its simple complexity, is a good candidate for implementation. To improve the accuracy of the channel estimates provided by the Correlator, the channel impulse response obtained is further processed by employing a low pass filter, called a channel estimation filter (CEF) [LIG93] (Correlator-CEF). It is known that the Wiener filter is optimal as the CEF in a stationary channel in the MMSE sense [LIG93]. To design a Wiener filter, however, it is essential to know the power spectrum of the channel and noise, which may not be obtainable in real time. Moreover, a large implementation complexity is required. The Doppler spectrum is usually spread to the maximum Doppler frequency of an experiencing channel. Thus it may be desirable to employ a brick-wall type lowpass filter such as the CEF, whose cut-off frequency is equal to the maximum Doppler frequency of the channel [LIG93], [SCH98a] and [SCH98b]. Such a CEF may not be practical, however, owing to the difficulty of implementation using a small number of filter taps. As a result, the CEF is realized in the form of a conventional lowpass filter like the Finite Impulse Response (FIR) filter or Infinite Impulse Response (IIR) filter [OJA98], [KAA98] and [BAR97]. Such lowpass filters can provide relatively good channel estimation performance if they are appropriately designed according to the channel condition [BET97], [KAA98] and [BAR97].

The improvement introduced by using an appropriate CEF at the output of the Correlator is highly appreciated, and we used this Correlator (Correlator-CEF) version for comparison throughout the simulation.

In this chapter, a multiuser-LMS-like structure along with smoothing and prediction filters to improve tracking quality is suggested. The choice for such an adaptation family stems from its low computational complexity and its regular structure, which is favourable for an efficient VLSI implementation where parallelism and wave pipelining, among other techniques, are easily applied [MOZ99], [MAS98], [YUE00] and [CHE03]. They are computationally effective due to the even distribution of the computation load over each symbol duration; no extra computation is required at the end of the processing window or preamble. Like Kalman-based channel estimation methods, an autoregressive stochastic model of correlated Rayleigh fading processes is used but indirectly embedded into the design. The attractive property of the proposed structure is the unique filter settings used for a large range of mobile speeds and for all users accessing the system. The design process is based on the following methodology: first, a p-th order autoregressive stochastic model [BAD01] is designed for an average Doppler profile; second, the resulting model along with the admissible parameter-drift-to-noise floor ratio (υ) [LIN01] are used to design the smoothing/prediction coefficients using Wiener LMS design methodology [LIN01]; finally, a multiuser-LMS structure is augmented with an extra smoothing/prediction procedure.

To sum up:

- The LMS structure takes into account all users' contributions simultaneously and delivers a composite channel impulse response, as in [BHA02], at each symbol. The composite channel impulse response is defined to be at least an $(N+1)K$ column vector, whose content provides the information about the multipath delays and time varying attenuations simultaneously; K represents the number of users and N the pilot spreading factor (in case of cdma2000 and WCDMA).
- A unique smoothing/prediction filter is designed based on a single p-order AR model over an averaged Doppler profile for all users.
- The adaptation step is dynamically examined at each iteration using the same approach as in steepest descent based methods.

As in [BHA02], our motives for choosing the multiuser-LMS structure are:

- an even distribution of the computational burden over a training window. or a preamble,
- a regular structure suitable for an efficient VLSI implementation

The chapter is organized as follows. In Section 3.2, a parametric DS-CDMA model, which gathers in a single column vector of length $K(N+1)$ all the users' parameters (named composite channel impulse response [BHA02]), is defined. Section 3.3 discusses the Wiener-LMS adaptive algorithm design and methodology, Section 3.4 describes the proposed composite channel impulse response estimation and tracking method. In Section 3.5 a complexity analysis, in terms of number of arithmetic operation counts, is performed regarding the optimized Correlator-CEF [OJA98], [KAA98] and [BAR97] and the proposed method. Section 3.6 provides some simulation results in cdma2000 and WCDMA environments with discussions. Finally, a conclusion is drawn in Section 3.7.

3.2 DS-CDMA Signal Model

For convenience a similar DS-CDMA model developed in the previous chapter is repeated here where the emphasis is put on considering long spreading codes and time varying complex channel attenuations.

As in Chapter 2, consider a DS-CDMA system with K users, using binary phase shift keying (BPSK) modulation and signaling over their respective multipath channels with additive white Gaussian noise. The binary information $b_k(i)$ (or simply $b_{k,i}$) (this information represents the code multiplexed pilot symbols in the case of cdma2000 and WCDMA systems for coherent estimation) is modulated by a spreading waveform $c_k(t)$

(this represents the long/short scrambling code for user k). The spreading sequence corresponding to $b_{k,i}$, the ith bit of the kth user, is denoted by $c_{k,i}(t)$ and consists of N chips, where N is the spreading gain. The corresponding discrete chip sequence is denoted by $\left[c_{k,i}[1]\, c_{k,i}[2] \cdots c_{k,i}[N]\right]$. The transmitted signal of the kth user corresponding to an information sequence of length M is given in baseband format by

$$s_k(t) = \sqrt{E_k} \sum_{i=1}^{M} b_{k,i}\, c_{k,i}(t - iT) \tag{3.1}$$

where T is the bit duration and E_k is the transmitted power of the user.

Let the channel be a multipath channel with L_k paths for the kth user and let the complex attenuation and delay with respect to the timing reference at the receiver of the ℓth path of the kth user be denoted by $w_{k,\ell}$ and $\tau_{k,\ell}$ respectively. The received signal can be represented as

$$r(t) = \sum_{k=1}^{K} \sum_{\ell=1}^{L_k} w_{k,\ell}(t)\, s_k\left(t - \tau_{k,\ell}\right) + n(t), \tag{3.2}$$

where $n(t)$ is the additive white Gaussian noise. As the maximum likelihood channel estimation technique, our schemes provide an estimate of the composite channel impulse response and not the estimates of the individual attenuations and delays. Therefore, the information about the number of paths of each user is not used in the derivation of our scheme.

The received signal is discretized at the receiver by sampling the output of a chip-matched filter at the chip rate [BEN96], [MAD97], [SEN99]. The observation vectors are formed by collecting N successive outputs of the chip-matched filter $r[n]$. The observation vectors correspond to a time interval equal to one symbol period and start at an arbitrary timing reference at the receiver. If we assume that all the paths of all users are within one symbol period from the arbitrary timing reference[3], we will have only two symbols of each user in each observation window, and we can develop a representation similar to that in [SEN99]. This model can be easily extended to include more general situations for the delays without affecting the derivation of the channel estimation algorithms [ERT98]. The discrete received vector model is given by

$$\mathbf{r}_i = \mathbf{C}_i \mathbf{Z}_i \mathbf{b}_i + \mathbf{n}_i, \tag{3.3}$$

where $\mathbf{r}_i$ is the ith $N \times 1$ observation vector, $\mathbf{C}_i$ is $N \times 2K(N+1)$ spreading matrix, $\mathbf{Z}_i$ is a $2K(N+1) \times 2K$ channel response matrix, $\mathbf{b}_i$ is a $2K \times 1$ symbol vector and $\mathbf{n}_i$ is a $N \times 1$ complex Gaussian zero-mean random vector with independent elements each of variance σ^2. In particular, the spreading matrix $\mathbf{C}_i$, is constructed using the shifted versions of the spreading codes corresponding to the ith and $i-1$th symbols of each user in the observation window. Thus, $\mathbf{C}_i$ is of the form $\left[\mathbf{C}_{1,i-1}^{\mathrm{R}}\ \mathbf{C}_{1,i}^{\mathrm{L}}\ \mathbf{C}_{2,i-1}^{\mathrm{R}}\ \mathbf{C}_{2,i}^{\mathrm{L}} \cdots \mathbf{C}_{K,i-1}^{\mathrm{R}}\ \mathbf{C}_{K,i}^{\mathrm{L}}\right]$ where

$$\mathbf{C}_{k,i-1}^{\mathrm{R}} = \begin{bmatrix} 0 & c_{k,i-1}[N] & \cdots & c_{k,i-1}[2] & c_{k,i-1}[1] \\ 0 & 0 & \cdots & c_{k,i-1}[3] & c_{k,i-1}[2] \\ \vdots & \vdots & & \vdots & \vdots \\ 0 & 0 & \cdots & c_{k,i-1}[N] & c_{k,i-1}[N-1] \\ 0 & 0 & \cdots & 0 & c_{k,i-1}[N] \end{bmatrix} \tag{3.4.a}$$

is constructed with the right part of the spreading code of the kth user corresponding to the $i-1$th symbol and

[3] The assumption is realistic in WCDMA and cdma2000 context as the pilot processing gain is on the order of 128 and 256 chips respectively. The equivalent round trip delay of $256T_c$ or $128T_c$ can be mapped to a cell radius of 20 and 10Km respectively.

$$\mathbf{C}_{k,i}^{\mathrm{L}} = \begin{bmatrix} c_{k,i}[1] & 0 & 0 & \cdots & 0 \\ c_{k,i}[2] & c_{k,i}[1] & 0 & \cdots & 0 \\ \vdots & \vdots & \vdots & & \vdots \\ c_{k,i}[N-1] & \cdots & c_{k,i}[1] & & 0 \\ c_{k,i}[N] & \cdots & c_{k,i}[2] & c_{k,i}[1] & 0 \end{bmatrix} \tag{3.4.b}$$

is constructed with the left part of the spreading code of the kth user corresponding to the ith symbol. Since the spreading codes change from symbol to symbol, the last columns of $\mathbf{C}_{k,i-1}^{\mathrm{R}}$ and $\mathbf{C}_{k,i}^{\mathrm{L}}$ are used additionally as compared to the short code case. The channel response matrix $\mathbf{Z}_i$ is of the form $\mathrm{diag}\left(\mathbf{z}_{1,i}, \mathbf{z}_{1,i}, \mathbf{z}_{2,i}, \mathbf{z}_{2,i}, \ldots, \mathbf{z}_{K,i}, \mathbf{z}_{K,i}\right)$ where $\mathbf{z}_{k,i}$ is the $(N+1)\times 1$ channel response vector for the kth user. When rectangular chip waveforms of duration T_c are used, the $\left(q_{k,\ell}\right)$th and $\left(q_{k,\ell}+1\right)$th elements of $\mathbf{z}_{k,i}$ have a contribution of $\left(1-\gamma_{k,\ell}\right)w_{k,\ell}(i)$ and $\gamma_{k,\ell}w_{k,\ell}(i)$ from the ℓth path of the kth user, where $\tau_{k,\ell} = \left(q_{k,\ell}+\gamma_{k,\ell}\right)T_c$. For example, when user k has only one path at delay $\tau_{k,1}$

$$\mathbf{z}_{k,i} = \left[0, \;\cdots\; 0, \;\left(1-\gamma_{k,1}\right)w_{k,1}(i), \;\gamma_{k,1}w_{k,1}(i), \;0, \;\cdots, \;0\right]^{\mathrm{T}} \tag{3.5}$$

where the non-zero elements are at the $\left(q_{k,1}\right)$th and $\left(q_{k,1}+1\right)$th positions. The symbol vector $\mathbf{b}_i = \left[b_{1,i-1}\; b_{1,i}\; b_{2,i-1}\; b_{2,i} \cdots b_{K,i-1}\; b_{K,i}\right]^{\mathrm{T}}$ has two symbols (chosen to be binary information bits ± 1 in this paper) corresponding to each user.

As in the previous chapter, we need to rewrite the received observation vector for channel estimation as

$$\mathbf{r}_i = \mathbf{C}_i\mathbf{B}_i\mathbf{z} + \mathbf{n}_i \tag{3.6}$$

where $\mathbf{z} = [\mathbf{z}_1^{\mathrm{T}}\ \mathbf{z}_2^{\mathrm{T}} \cdots \mathbf{z}_K^{\mathrm{T}}]^{\mathrm{T}}$ is a $K(N+1)\times 1$ channel response vector and $\mathbf{B}_i$ is a $2K(N+1)\times K(N+1)$ matrix defined as

$$\mathbf{B}_i = \begin{bmatrix} b_{1,i-1} & 0 & 0 & \cdots & 0 & 0 \\ b_{1,i} & 0 & 0 & \cdots & 0 & 0 \\ 0 & b_{2,i-1} & 0 & \cdots & 0 & 0 \\ 0 & b_{2,i} & 0 & \cdots & 0 & 0 \\ \vdots & \vdots & & & & \vdots \\ \vdots & \vdots & & & & \vdots \\ 0 & 0 & 0 & \cdots & 0 & b_{K,i-1} \\ 0 & 0 & 0 & \cdots & 0 & b_{K,i} \end{bmatrix} \otimes \mathbf{I}_{N+1}$$

where $\otimes$ denotes the Kronecker product and $\mathbf{I}_{N+1}$ is the identity matrix of rank $N+1$. Thus we estimate $N+1$ channel parameters for each user. This effective channel response accounts for all the paths under the assumption that their delays are within one symbol duration. Let us, for the sake of brevity, set $\mathbf{X}_i^{\mathrm{H}} = \mathbf{C}_i\mathbf{B}_i$.

In the following sections, we will discuss the Wiener LMS algorithm and its application to channel estimation and tracking problem. The following section 3.3 exposes the WLMS theory that is inspired from [LIN01].

3.3 Wiener LMS Theory

It has been argued that various averaging methods appropriate for slowly time-varying dynamics [KUS84], [KUB91], are not applicable to fast time-variations [LIN01]. Toward this end, time varying Kalman filter for estimating linear models' parameters [AND79] is used for channel tracking [DAV98]. Unfortunately, its computational complexity often prevents its use [LIN93].

Recently, Wiener LMS algorithm, wherein the mean squared tracking error is minimized, has been developed [LIN01]. Wiener LMS exhibits interesting practical features, that renders it attractive for channel estimation and tracking in communication systems where performance-complexity trade-off is an unavoidable issue.

Within this class of algorithms, our multiuser LMS like structure design focuses on the variant of lowest complexity based on Wiener LMS algorithm (WLMS). At close to LMS complexity, it offers a significant performance improvement compared to the Correlator-CEF.

For notational convenience, we follow the same notation as in [LIN01] to introduce some relevant concepts. Define a backward shift operator $q^{-1}\left(q^{-1}x_i = x_{i-1}\right)$ upon which the following polynomial is adopted

$$P\left(q^{-1}\right) = p_0 + p_1 q^{-1} + \ldots + p_{n_p} q^{-n_p}. \tag{3.7}$$

Defined also is the conjugate polynomial as $P_*(q) = p_0^* + p_1^* q \ + \ldots + p_{n_p}^* q^{n_p}$ where p^* represents the complex conjugate of p. This same definition is extended to rational matrices, or transfer function matrices, $\wp\left(q^{-1}\right)$.

Now, consider a $N_r \times 1$ sequence of measurement vectors $\mathbf{r}_i, i = 1,2,\ldots$ to be generated by

$$\mathbf{r}_i = \mathbf{X}_i^{\mathrm{H}} \mathbf{z}_i + \mathbf{n}_i. \tag{3.8}$$

where $\mathbf{n}_i$ represents noise and the $N_r \times N_z$ matrix sequence $\mathbf{X}_i^{\mathrm{H}}, i = 1,2,\ldots,$ is complex valued and known. Note that the linear model in (3.8) is in-agreement with the channel model (2.23) outlined in chapter 2. Of importance for Wiener LMS design, $\mathbf{X}_i^{\mathrm{H}}, i = 1,2,\ldots$ is usually assumed stationary with zero mean and invertible covariance matrix $\mathbf{R} = E\left\{\mathbf{X}_i \mathbf{X}_i^{\mathrm{H}}\right\}$.

In the other hand, as is the case in different types of models for fading channel, the time-variations are in general statistical in nature. This means that the parameter vector $\mathbf{z}_i$ is modeled by a stochastic process, with known statistical properties. Therefore, $\mathbf{z}_i$, in the Wiener LMS context, is modeled as linear time-invariant stochastic hypermodels [LIN01]

$$\mathbf{z}_i = \wp\left(q^{-1}\right)\mathbf{e}_i \tag{3.9}$$

where $\mathbf{e}_i$ is a complex valued white noise vector and $\wp$ is a $N_z \times N_z$ transfer function matrix. For a simpler model, $\wp\left(q^{-1}\right)$ is assumed to be diagonal, with equal elements along the diagonal so that

$$\mathbf{z}_i = \wp\left(q^{-1}\right)\mathbf{e}_i = \frac{C\left(q^{-1}\right)}{D\left(q^{-1}\right)}\mathbf{I}_{N_z \times N_z}\mathbf{e}_i \tag{3.10}$$

To ease the design process a special model, for which explicit solutions to the design equations are available, can be used. Toward this end, (3.10) can be further customized with

$$C\left(q^{-1}\right) = 1,\ D\left(q^{-1}\right) = 1 + a_1 q^{-1} + a_2 q^{-2} \tag{3.11}$$

One special case, that some adaptation schemes have indirectly incorporated and frequently used in simulations is the random walk model $(a_1 = -1, a_2 = 0)$. Another special case, namely, the filtered random walk, is obtained by setting $a_1 = -\left(1 + a_2\right)$. Unless no prior information- about the dynamics of $\mathbf{z}_i$, are available, then (3.11), with a_2 chosen in the interval $[0.90\ \ 0.999]$, will often be a reasonable and appropriate first choice as a hypermodel.

To address the tracking problem, the tracking performance is measured by

$$\operatorname{tr}\left(\mathbf{P}_k\right) \equiv \lim_{i\to\infty} \operatorname{tr}\left(E\left\{\tilde{\mathbf{z}}_{i+k}\tilde{\mathbf{z}}_{i+k}^{*}\right\}\right) = \lim_{i\to\infty} \sum_{v=1}^{N_z} E\left\{\left|\mathbf{z}_{v,i+k} - \hat{\mathbf{z}}_{v,i+k|i}\right|^2\right\}, \tag{3.12}$$

where the expectation is taken with respect to $\mathbf{e}_i$ in (3.9) and $\mathbf{n}_i$ in (3.8). $\mathbf{z}_{v,i+k}$ and $\hat{\mathbf{z}}_{v,i+k|i}$ are the vth elements of $\mathbf{z}_{i+k}$ and $\hat{\mathbf{z}}_{i+k|i}$ respectively. The prediction error is defined as $\tilde{\mathbf{z}}_{i+k} = \mathbf{z}_{i+k} - \hat{\mathbf{z}}_{i+k|i}$ where $\hat{\mathbf{z}}_{i+k|i}$ is the estimate of $\mathbf{z}_{i+k}$ at time i. Wiener LMS scheme is derived by minimizing (3.12) with the low-complexity LMS adaptation scheme by introducing two modifications into the basic LMS structure. The main modifications to migrate from LMS to Wiener LMS are made possible by, first, pre-multiplying the term $\mathbf{X}_i\boldsymbol{\varepsilon}_i$, in the LMS' update equation, by $\mathbf{R}^{-1}$. Second, to obtain the smoothing and prediction estimates of $\hat{\mathbf{z}}_i$ for arbitrary time-lags k, a smoothing/prediction filter $\Im_k\left(q^{-1}\right)$ is used as in (3.15).

Accordingly, the Wiener LMS is given by

$$\boldsymbol{\varepsilon}_i = \mathbf{r}_i - \mathbf{X}_i^{*}\hat{\mathbf{z}}_{i|i-1} \tag{3.13}$$

$$\hat{\mathbf{z}}_i = \hat{\mathbf{z}}_{i|i-1} + \mu\mathbf{R}^{-1}\mathbf{X}_i\boldsymbol{\varepsilon}_i \tag{3.14}$$

$$\hat{\mathbf{z}}_{i+k|i} = \Im_k\left(q^{-1}\right)\hat{\mathbf{z}}_i \tag{3.15}$$

The choice of $\Im_k\left(q^{-1}\right)$ will depend on the dynamics of $\mathbf{z}_i$, the signal-to-noise ratio (SNR), the number of users K and the processing gain N (in case of multiuser detection problem).

Usually, in wireless fading channels, the channel coefficients undergo similar fading-types; they mainly differ in their relative energies. Unless otherwise specified, the coefficients are statistically independent. Furthermore, $\mathbf{R}$ can be assumed to be diagonal which is a good approximation in WCDMA and cdma2000 system wherein the regressor matrix is constructed from gold-like sequences. From the above arguments, $\Im_k\left(q^{-1}\right)$ is constrained to an equal-element-diagonal matrix as well.

To summarize, the time-varying parameters $\mathbf{z}_i$ ($\mathbf{z}_i = \wp\left(q^{-1}\right)\mathbf{e}_i$), using the observations $\mathbf{r}_i = \mathbf{X}_i^{\mathrm{H}}\mathbf{z}_i + \mathbf{n}_i$, are to be estimated by minimizing the criterion (3.12) within the class of tracking algorithms represented by (3.13)-(3.15) with $\Im_k\left(q^{-1}\right)$ being diagonal, and assuming $\mathbf{X}_i^{\mathrm{H}}$, $\mathbf{R}$, $\Re\left(q^{-1}\right)$and $\mathbf{R}_e = E\left\{\mathbf{e}_i\mathbf{e}_i^{\mathrm{H}}\right\}$ to be accessible.

Further more, to simplify the design to a single scalar design equation, the hypermodel (3.10) is adopted. The choice of using (3.10) is also supported by the arguments we provided above regarding the similarities in the statistical properties of the channel coefficients and the independence assumptions.

[LIN01] argued that the optimal constrained filter $\Im_k\left(q^{-1}\right)$ minimizing (3.12) is unique. To implement (3.13)-(3.15) the unique solution for $\Im_k\left(q^{-1}\right)$ can be sought of as

$$\Im_k\left(q^{-1}\right)=\frac{Q_k\left(q^{-1}\right)}{Q_0\left(q^{-1}\right)}\mathbf{I}_{N_z\times N_z} \tag{3.16}$$

where the causal polynomial $Q_k\left(q^{-1}\right)$ ($L_{k*}(q)$ being the non-causal part) is the unique solution to the scalar Diophantine equation [LIN01]

$$q^k \upsilon CC_* = rQ_k\beta_* + qDL_{k*} \tag{3.17}$$

Here, the polynomial $\beta\left(q^{-1}\right)$ is the stable and monic solution to the polynomial spectral factorization of

$$r\beta\beta_* = \upsilon CC_* + DD_* \tag{3.18}$$

with r being a real valued positive scalar and υ is the parameter-drift-to-noise ratio defined by $\upsilon \triangleq \mathrm{tr}\left(\mathbf{R}_\mathbf{e}\right)/\mathrm{tr}\left(\mathbf{R}^{-1}\mathbf{R}_\boldsymbol{\eta}\mathbf{R}^{-1}\right)$ where $\boldsymbol{\eta}_i=\left(\mathbf{X}_i\mathbf{X}_i^*-\mathbf{R}\right)\left(\hat{\mathbf{z}}_{i|i-1}-\mathbf{z}_i\right)+\mathbf{X}_i\mathbf{n}_i$ represents the gradient noise [LIN01]. As a summary of the WLMS design steps, one need to:

1. Determine the spectral polynomial β and the scalar r from (3.18);
2. for any time lag *k*, compute Q_k from (3.17). Notice that Q_1 ($Q_1=q\left(\beta-D\right)$ and $Q_0=\beta-D/r$) is always required as the one-step-prediction estimates are required in (3.13) and (3.15)
3. set $\mu=1-\left(1/r\right)$ and $\Im_k=\dfrac{Q_k}{Q_0}\mathbf{I}_{N_z\times N_z}$.

Detailed discussion of Wiener LMS can be found in [LIN01]. It is beyond the scope of this thesis to highlight every detail.

3.4 Composite Channel Impulse Response Estimation and Tracking: Methodology and Algorithm

It was first mentioned in [KAM84] that the estimation of channel coefficients is equivalent to carrier phase tracking, and more work [HAE88], [HAE89] was devoted to applying the Kalman filter to channel estimation. In most of these studies the fading channel was modeled as an autoregressive (AR) process in order to apply the Kalman filter. One can verify that a simple second order AR process (AR2) can approximate the Jakes' model [JAK74] and can be used as a hyper model embedded into a Kalman filter. It was observed in [LIN93] that the spectral peak frequency of the AR2 process should be adjusted by a factor of $\sqrt{2}$ from the maximum Doppler frequency. In [JAK74], Jakes proposed modeling a Rayleigh fading process $w(t)$ by a number of oscillators with different phases and frequencies which reflect the Doppler spread. The Jakes model was further modified in [DEN93] to satisfy the desired properties of a fading channel, i.e., the in-phase and quadrature components are uncorrelated and their variances are identical. The modified model of the Rayleigh fading process is [DEN93]

$$w(t) = \sqrt{\frac{2}{N_{Os}}} \sum_{n=1}^{N_{Os}} e^{j\phi_n} \cos\left(2\pi f_d t \cos(\alpha_n)\right) \tag{3.19}$$

where N_{Os} is the number of distinct oscillators (the total number of oscillators is $4N_{Os}$), ϕ_n the phase of the nth oscillator and $\alpha_n = 2\pi \frac{n-0.5}{4N_{Os}}$.

If we consider the fading coefficients modeled by an AR2 process, we can write[4]

$$w(i) = -a_1 w(i-1) - a_2 w(i-2) + e(i) \tag{3.20}$$

where $e(i)$, the driving noise of the fading process, is a complex zero-mean white Gaussian process. The AR2 process parameters a_1 and a_2 are determined by the location of the poles of the transfer function on the unite circle. These parameters are closely related to the physical parameters of the underlying fading process by [YUW98]

$$a_1 = -2 r_d \cos\left(2\pi \bar{f}_d T\right) \tag{3.21}$$

$$a_2 = r_d^2 \tag{3.22}$$

where $\bar{f}_d$ is the spectral peak frequency, T is the symbol period, and r_d is the pole radius which corresponds to the steepness of the peaks of the power spectrum. [YUW98] demonstrates that $\bar{f}_d = \frac{f_d}{\sqrt{2}}$, where, f_d is the maximum Doppler frequency.

The above AR2 model was motivated by the fact that the design based on Kalman filtering would not be complex insofar as a higher order is not used. Unfortunately, the AR2 is far from being a good approximation nor for channel modeling [BAD01] neither for channel tracking [LIN01]. The reader is referred to the paper by Baddour and Beaulieu [BAD01], which shows that an AR model order of at least 100 is sometimes required for channel modeling. Fortunately for channel estimation purpose low-order models can be considered [LIN01]. We followed the methodology outlined in [BAD01] to compute the

[4] The AR model is used only for the receiver (channel estimation block) design and not for channel attenuation generation. The later is generated using the modified Jakes model.

AR coefficients. Unlike [BAD01] an averaged autocorrelation sequence over the Doppler range of interest (c.f. (3.24)) is used to solve,

$$\mathbf{R}_{ww}\mathbf{a} = -\mathbf{r}_{ww} \tag{3.23}$$

where $\mathbf{r}_{ww} = \begin{bmatrix} r_{ww}[1] & r_{ww}[2] & \cdots & r_{ww}[p] \end{bmatrix}^{\mathrm{T}}$, $\mathbf{a} = \begin{bmatrix} a_1 & a_2 & \cdots & a_p \end{bmatrix}^{\mathrm{T}}$ and

$$\mathbf{R}_{ww} = \begin{bmatrix} r_{ww}[0] & r_{ww}[-1] & \cdots & r_{ww}[-p+1] \\ r_{ww}[1] & r_{ww}[0] & \cdots & r_{ww}[-p+2] \\ \vdots & \vdots & \ddots & \\ r_{ww}[p-1] & r_{ww}[p-2] & \cdots & r_{ww}[0] \end{bmatrix}.$$

We define the averaged autocorrelation sequence $r_{ww}[n]$ for the absolute time lag *n*, to be given by

$$r_{ww}[n] = \sum_{s=1}^{S} J_0\left(2\pi f_m^{s} |n|\right) \Pr\left(f_m^{s}\right) \tag{3.24}$$

where $J_0(\bullet)$ is the zeroth-order Bessel function of the first kind and $\Pr\left(f_m^{s}\right)$ is the probability density function, intuitively chosen, for the normalized maximum Doppler frequency $\left(f_m^{s} = \left(f_d^{s}T\right) = \left(v_s f_c T/c\right)\right)$ corresponding to the mobile speed v_s in meter per second. The weighting profile $\Pr\left(f_m^{s}\right)$ may take a range of profiles ranging from a simple uniform distribution to a bell shaped profile over the desired speed range delimited by $v_{s=1}$ and $v_{s=S}$. The numerical problem encountered in solving (3.23) is addressed in [BAD01], so that a solution exists for a given order p.

Now that, the ARp is designed, one can use $\mathbf{a} = \begin{bmatrix} a_1 & a_2 & \cdots & a_p \end{bmatrix}^{\mathrm{T}}$ to design the filter coefficient using the procedure developed in [LIN01] and summarized below for convenience.

From equation (3.10) in section 3, define the following rational polynomial $\dfrac{C(q^{-1})}{D(q^{-1})}$ based on $\mathbf{a} = \begin{bmatrix} a_1 & a_2 & \cdots & a_p \end{bmatrix}^{\mathrm{T}}$ as

$$\frac{C(q^{-1})}{D(q^{-1})} = \frac{1}{a_1 + a_2 q^{-1} + \ldots + a_p q^{-p+1}} \tag{3.25}$$

Define the following rational polynomial $\dfrac{Q_1(q^{-1})}{Q_0(q^{-1})}$ as

$$\frac{Q_1(q^{-1})}{Q_0(q^{-1})} = \frac{\sum_{n=0}^{N_s-1} \zeta_n^{\mathrm{s}} q^{-n}}{1 - \sum_{n=1}^{N_p-1} \zeta_n^{\mathrm{p}} q^{-n}} \tag{3.26}$$

Wiener LMS theory developed in [LIN01] can be applied to solve for ζ_n^{s}'s and ζ_n^{p}'s, using the definition (3.25), as

1. Set a value for $\upsilon \in]0 \quad \infty[$ then
2. Calculate the spectral factor $\beta(q^{-1})$ and the real scalar r. The polynomial $\beta(q^{-1})$ is the stable and monic solution to the polynomial spectral factorization $r\beta(q^{-1})\beta_*(q) = \upsilon + D(q^{-1})D_*(q)$.

3. Let $Q_1(q^{-1}) = q(\beta(q^{-1}) - D(q^{-1}))$ and $Q_0(q^{-1}) = \beta(q^{-1}) - D(q^{-1})/r$.

4. Match $Q_1(q^{-1})$ and $Q_0(q^{-1})$ with the rational polynomial in (3.26) to extract the smoothing/prediction FIR coefficients ζ_n^{s}'s and ζ_n^{p}'s and set $\mu = 1 - \frac{1}{r}$.

Using the signal model (3.6) highlighting the composite impulse response for all users, the Multiuser WLMS algorithm is summarized below

For $i = 1, 2, \ldots, M$, as soon as the observation $N \times 1$ vector $\mathbf{r}_i$ is received,

1. Compute the error $\boldsymbol{\varepsilon}_i$ as

$$\boldsymbol{\varepsilon}_i = \mathbf{r}_i - \mathbf{X}_i^{\mathrm{H}} \hat{\mathbf{z}}_{i|i-1} \tag{3.27}$$

where $\mathbf{X}_i^{\mathrm{H}} = \mathbf{C}_i \mathbf{B}_i$.

2. Compute the smoothed version of $\hat{\mathbf{z}}_i$ as

$$\hat{\mathbf{z}}_i = \hat{\mathbf{z}}_{i|i-1} + \mu \mathbf{X}_i \boldsymbol{\varepsilon}_i \tag{3.28}$$

The term $\mathbf{X}_i \boldsymbol{\varepsilon}_i$ can be viewed as performing the correlation over the prediction error $\boldsymbol{\varepsilon}_i$ from (3.27). The adaptive step size μ can be fixed or computed at each iteration i, μ_i (c.f. (3.12))

3. Compute the one step prediction[5] $\hat{\mathbf{z}}_{i+1/i}$ as

$$\hat{\mathbf{z}}_{i+1/i} = \hat{\mathbf{z}}_i^{\mathrm{p}} + \hat{\mathbf{z}}_i^{\mathrm{s}} \tag{3.29}$$

where, the smoothing FIR implements

[5] This step implements "one step prediction" instead of "identity law" $\hat{\mathbf{z}}_{i+1/i} = \hat{\mathbf{z}}_i$ as in a standard LMS algorithm.

$$\hat{\mathbf{z}}_i^{\mathrm{s}} = \sum_{n=0}^{N_s-1} \zeta_n^{\mathrm{s}} \hat{\mathbf{z}}_{i-n} \tag{3.30}$$

while the prediction FIR implements

$$\hat{\mathbf{z}}_i^{\mathrm{p}} = \sum_{n=1}^{N_p-1} \zeta_n^{\mathrm{p}} \hat{\mathbf{z}}_{i-n+1/i-n} \tag{3.31}$$

In a standard LMS structure, (3.29) is implemented using an identity law $\hat{\mathbf{z}}_{i+1/i} = \hat{\mathbf{z}}_i$. Using the procedure outlined in [LIN01] after a proper choice of the AR model order p and υ, the coefficients ζ_n^{s} and ζ_n^{p} and the optimal adaptive step size $\mu^{(\mathrm{opt})}$ are computed. The choice of the AR order is driven by complexity constraints: the larger the order is, the more accurate the AR model is at the expense of a complex design.

4. Compute the $K(N+1)\times 1$ vector variance

$$\mathbf{V}_{\hat{z}}(i) = \frac{i-1}{i}\mathbf{V}_{\hat{z}}(i-1) + \frac{1}{i}\boldsymbol{\rho}_i \tag{3.32}$$

where $\boldsymbol{\rho}_i = \left[\left|\hat{z}_{i,1}\right|^2, \left|\hat{z}_{i,2}\right|^2, \cdots, \left|\hat{z}_{i,K(N+1)}\right|^2\right]^{\mathrm{T}}$, with $\boldsymbol{\rho}_v = \mathbf{0}_{K(N+1)\times 1}$ if $\mathrm{mod}(v, M) = 0$ and $z_{i,j}$ presents the j th element of the vector $\mathbf{z}_i$.

5. For delay detection, examine $\mathbf{V}_{\hat{z}}(M)$ by segments for each user k, beginning at position $(k-1)(N+1)+1$ and terminating at position $k(N+1)$ to select the largest components ($\ell = 1,2,\ldots,L_k$) to be considered as the correct (most significant) path position (delay) for which the path attenuation, $\{\hat{w}_{k,\ell}\}$, is deduced from $\hat{\mathbf{z}}_i$ at the same element position ($\{\hat{\tau}_{k,\ell}\}$).

In the sequel, define an optimal value, $\mu^{(\mathrm{opt})}$, of the adaptive step size for a given system/channel conditions (c.f. Tables 3.1 and 3.2) as an appropriate step size that minimizes the Bit Error Rate (BER). This optimal value dependents on the system/channel conditions (number of users, signal to noise ratio, data rate, pilot-to-data power ratio and others). The algorithm as outlined above needs, for optimum BER performance, to set μ to its optimal value as the real-world system/channel conditions certainly differ from the ones used (c.f. tables 3.1 and 3.2). If we assume that the nominal value of μ (c.f. (3.33)) is close to the optimal value and that the filter taps are less sensitive to slow changes in υ, one can devise a recursive search method to determine μ. At an additional computational load, one way to search dynamically for μ is to compute its value at each recursion $i = 1, 2, 3, \ldots$, using

$$\mu_i = \frac{\left\|\mathbf{X}_i \boldsymbol{\varepsilon}_i\right\|^2}{\left\|\mathbf{X}_i^H \mathbf{X}_i \boldsymbol{\varepsilon}_i\right\|^2} \tag{3.33}$$

Equation (3.33) is derived as follows; first define the smoothing error term

$$\hat{\boldsymbol{\varepsilon}}_i = \mathbf{r}_i - \mathbf{X}_i^{\mathrm{H}} \hat{\mathbf{z}}_i \tag{3.34}$$

This error term is related to the prediction error in (3.27) by substituting (3.28) into (3.27) and rearranging the terms to yield

$$\begin{aligned} \hat{\boldsymbol{\varepsilon}}_i &= \mathbf{r}_i - \left[\mathbf{X}_i^{\mathrm{H}} \hat{\mathbf{z}}_{i/i-1} + \mu \mathbf{X}_i^{\mathrm{H}} \mathbf{X}_i \boldsymbol{\varepsilon}_i\right] \\ &= \boldsymbol{\varepsilon}_i - \mu \mathbf{X}_i^{\mathrm{H}} \mathbf{X}_i \boldsymbol{\varepsilon}_i \\ &= \boldsymbol{\varepsilon}_i - \mu \mathbf{R}_i \boldsymbol{\varepsilon}_i \end{aligned} \tag{3.35}$$

where $\mathbf{R}_i = \mathbf{X}_i^{\mathrm{H}} \mathbf{X}_i$.

μ_i, in (3.33), is chosen as to minimize the squared error function given by

$$\phi(\hat{\mathbf{z}}_i) = \hat{\boldsymbol{\varepsilon}}_i^{\mathrm{H}} \hat{\boldsymbol{\varepsilon}}_i \tag{3.36}$$

The optimal value of μ_i is obtained as

$$\begin{aligned}
\mu_i &= \arg\min_{\mu} \phi(\hat{\mathbf{z}}_i) \\
&= \arg\min_{\mu} \left[\left(\boldsymbol{\varepsilon}_i^{\mathrm{H}} - \mu \boldsymbol{\varepsilon}_i^{\mathrm{H}} \mathbf{R}_i^{\mathrm{H}} \right) \left(\boldsymbol{\varepsilon}_i - \mu \mathbf{R}_i \boldsymbol{\varepsilon}_i \right) \right] \\
&= \arg\min_{\mu} \left[\boldsymbol{\varepsilon}_i^{\mathrm{H}} \boldsymbol{\varepsilon}_i - \mu \boldsymbol{\varepsilon}_i^{\mathrm{H}} \mathbf{R}_i^{\mathrm{H}} \boldsymbol{\varepsilon}_i - \mu \boldsymbol{\varepsilon}_i^{\mathrm{H}} \mathbf{R}_i \boldsymbol{\varepsilon}_i + \mu^2 \boldsymbol{\varepsilon}_i^{\mathrm{H}} \mathbf{R}_i^{\mathrm{H}} \mathbf{R}_i \boldsymbol{\varepsilon}_i \right]
\end{aligned} \tag{3.37}$$

Notice that the second and the third terms are of the form $a^* + a$, which can be written as $2\Re\{a\}$, where $\mathrm{Re}(\bullet)$ represents the real value operator. Observe further that $a = \left(\mathbf{X}_i \boldsymbol{\varepsilon}_i\right)^{\mathrm{H}} \mathbf{X}_i \boldsymbol{\varepsilon}_i = \left\|\mathbf{X}_i \boldsymbol{\varepsilon}_i\right\|^2$ is a real quantity, hence $2\Re\{a\} = 2\left\|\mathbf{X}_i \boldsymbol{\varepsilon}_i\right\|^2$. The same idea can be extended to the last term, which can be replaced by $\left\|\mathbf{X}_i^{\mathrm{H}} \mathbf{X}_i \boldsymbol{\varepsilon}_i\right\|^2$. Hence equation (3.33). Since the term $\mathbf{X}_i \boldsymbol{\varepsilon}_i$ is already computed from step 2 of the algorithm, the additional complexity added by (3.33) is a matrix-vector multiplication and one division. We refer to the algorithm that uses (3.33) as Multiuser Steepest WLMS (Multiuser S-WLMS).

At an additional computational cost the proposed Multiuser Steepest Wiener LMS has shown a stable and suitable solution for setting μ. A computer based search methodology has been devised offline: given a system/channel condition (c.f. Tables 3.1 and 3.2),

(a) the model order p is set to a given desired value (let us say 6);

(b) the parameter-drift-to-noise ratio (a tuning knob), υ, is chosen from $]0 \quad \infty[$;

(c) as (3.23) is solved, the filters' smoothing/prediction coefficients are deduced using Wiener LMS design theory; and finally,

(d) the entire design is used along with a detector (e.g. Rake, PIC).

The BER performance will dictate the appropriate range of υ. This can be done through simulations at targeted SNR, system load (number of user K), data rates and channel type. We may fine tune the value of υ along with the filter taps till we are confident that the BER has reached a desirable value[6]. We can use this value to make the appropriate final design settings. The process is iterated a few times to determine the desired values for the smoothing/prediction filter taps and the optimal adaptive step size $\mu^{(\mathrm{opt})}$. This methodology proved appropriate as we experimented with the simulation in both cdma2000 and WCDMA environments.

Unlike [LIN01], the proposed methodology and algorithm are outlined to estimate the composite channel impulse response for all users simultaneously, rather than single user attenuation, by assuming perfect acquisition. Unlike [LIN01], the design is less dependent on speed variations and does not assume any Doppler estimation, as it uses the average autocorrelation function over the entire range of Doppler frequency of interest.

3.5 Computation Complexity

3.5-1 Operations counts

Before we conclude the discussion on the multiuser WLMS and multiuser S-WLMS based channel estimation, we will evaluate the computational complexity of the proposed schemes. In both the proposed WLMS and S-WLMS channel estimation and tracking, the

[6] For WCDMA and cdma2000 a target SNR at a desired BER is dictated by the system's minimum performance specifications.

computations (additions and multiplications) during the i th processing sample for the four step update outlined above are

1. Computation of $\mathbf{X}_i^{\mathrm{H}} = \mathbf{C}_i\mathbf{B}_i$ requires $8N(N+1)K^2 - 2N(N+1)K$ additions and no multiplications (c.f. (3.27))
2. Computation of $\boldsymbol{\varepsilon}_i = \mathbf{r}_i - \mathbf{X}_i^{\mathrm{H}}\hat{\mathbf{z}}_{i|i-1}$ requires $4N^2K$ addition and no multiplications[7] (c.f. (3.27))
3. Compute the update $\hat{\mathbf{z}}_i = \hat{\mathbf{z}}_{i|i-1} + \mu\mathbf{X}_i\boldsymbol{\varepsilon}_i$ requires $4N(N+1)K$ additions and $2(N+1)K$ multiplications (c.f. (3.28))
4. Compute the one step prediction (for AR2 design model) $\hat{\mathbf{z}}_{i+1|i} = -\xi_1^{\mathrm{p}}\hat{\mathbf{z}}_{i|i-1} - \xi_0^{\mathrm{s}}\hat{\mathbf{z}}_i - \xi_1^{\mathrm{s}}\hat{\mathbf{z}}_{i-1}$ requires $3(N+1)K$ additions and $6(N+1)K$ multiplications ((c.f. (3.29)-(3.31)))

The most complex step is the computation of $\mathbf{X}_i^{\mathrm{H}} = \mathbf{C}_i\mathbf{B}_i$ which has the complexity on the order of $O(N^2K^2)$. However, this matrix multiplication involves only multiplications of ± 1 's and this can be taken advantage of to speed up practical implementations [RAJ00].

3.5-2 Reduced size channel estimation

In the estimation methods proposed above, the channel could have any number of paths with delays lying within one symbol period. All of these paths will be captured in the channel response vector $\mathbf{z}$. However, in some practical scenarios where the number of paths is small and rectangular chip waveforms are used, we may not need the whole

[7] This step involves no multiplication as the current setting assumes the regressor matrix $\mathbf{X}_i^{\mathrm{H}}$ having $\pm 1 \pm j$ elements, where a multiplication by $\pm 1 \pm j$ will be accounted for two sign changes.

$\mathbf{z}$ vector. For example, when there are just 2 paths for a user and the chip waveform is rectangular, the number of non-zero elements in $\mathbf{z}$ corresponding to that user is at most 4. For other non-rectangular chip waveforms, more coefficients might be non-zero based on the autocorrelation of the pulse waveform used and the delays of the paths. For the rectangular pulse, the support of the autocorrelation function is only over the interval $[-T_c \quad T_c]$. If such information about the pulse shape and paths are available at the receiver, the iterative estimate obtained earlier can be further improved by using this knowledge. This information can be used to reduce the size of the estimated channel response vector $\hat{\mathbf{z}}$. One simple ad-hoc method to reduce the size of the estimated channel vector $\hat{\mathbf{z}}$ is to choose a few large coefficients of $\hat{\mathbf{z}}$. In particular, we choose a few large coefficients, say L, for each user. Thus, we are left with a smaller vector of size LK. If the elements that were truly zero were dropped by this procedure, the error in estimation of the zero elements would be made zero and the total squared error in the estimate will be lower. Once the LK significant elements are chosen, the error in these LK elements can be improved by repeating the estimation schemes with a new reduced model of the discrete received signal. Other complex statistical tests to choose the significant coefficients from the ML estimate can be derived using the ideas in [WAL43], [RAO66].

3.5-3 Computation complexity of the gradient descent ML

Gradient descent ML technique is summarized in appendix A for brevity. In the proposed iterative ML channel estimation schemes, the computations (additions and multiplications) during the ith processing window for the three-step update outlined above are

1. Computation of $\mathbf{R}_i$: $2N(N+1)^2 K^2 + (N+1)^2 K^2$ (c.f. (A-4))
2. Computation of $\mathbf{y}_i$: $2N(N+1)K + (N+1)K$ (c.f. (A-5))
3. Update of $\mathbf{z}$: $2(N+1)^2 K^2 + 3(N+1)K$ (c.f. (A-6)) for the simple gradient descent method and $4(N+1)^2 K^2 + 7(N+1)K$ for the steepest descent method (due to additional computation to obtain μ (c.f. (A-7))).

The most complex step is the computation of $\mathbf{R}_i$ which has the complexity of $O(N^3K^2)$. However, like multiuse S-WLMS method, this matrix multiplication involves only multiplications of ± 1's and this can be taken advantage of to speed up practical implementations [RAJ00].

3.6 Results and Comments

It is worth to mention that gradient descent ML channel estimation's performance is not included in this section as we attempt to provide a good alternative solution (performance/complexity trade off) to the Correlator based channel estimator. Nevertheless, some preliminary simulation results are shown in appendix A. Notice that the gradient descent ML algorithm and the proposed Multiuser S-WLMS show very similar performances.

The AR order is chosen to provide an average gain of 2 to 3 dB with a complexity increase of no more than a factor of 4 as compared to the Correlator-CEF. After few design iterations, we concluded that the order 6 is suitable for most of the scenarios in both WCDMA and cdma2000. The choice for such an order was dictated by: (1) keeping the model order as small as possible and hence keeping the computational burden low, and (2) satisfying the cdma2000 and WCDMA minimum system performance requirements. $\Pr\left(f_m^s\right)$ in (3.24) is chosen to be bell-shaped centered at 60 km/h. For the Correlator, an appropriate CEF is optimized [CHO04].

The simulations were conducted in both cdma2000 and WCDMA environments. Because of the huge simulation time, tasks were distributed, so that in cdma2000 our intention was mainly to check performances at different channel simulator configurations [3GP02], realistic channels (namely vehicular A (VA) and pedestrian A (PA)), mixed voice/data rates (9.6kb/s and 38.6kb/s) and under the effect of power control transfer error rate (PC TER). In the WCDMA platform, we were interested in assessing the improvement at different data rates ranging from voice rate (12.2kb/s) to higher data rates (384kb/s) using different detectors (Rake, Multistage Hard PIC (Hard IC) and Multistage Soft PIC (Soft IC)) [YUE00], [CHE03] and different users' mobile speeds uniformly distributed between 0 and 120km/h. For cdma2000 simulations, both frame error rate (FER) and bit error rate (BER) were computed, while only raw BER was considered for WCDMA simulations (due to simulation time constraints).

3.6-1 cdma2000 system

Consider a cdma2000 system with one frequency band and one receiving antenna. The data rate is 9.6kb/s. 14 users are assumed to asynchronously access the system. The channel types are from 3GPP2's minimum performance specifications ([3GP02], Table 6.4.1.1-1 of 3GPP2 C.S0011-B). We use the term "channel simulator configuration 1 (CSC1)" for the channel with two paths whose relative delays in microseconds are 0 and 2, and whose relative powers in dB's are 0 and 0; "CSC2" refers to the channel with three paths whose relative delays in microseconds are 0, 2 and 14.5, and whose relative powers in dB's are 0, 0 and -3. Mobile speeds of 8km/h and 30km/h in CSC1 and 100km/h in CSC2 were simulated. Perfect power control (PC) was performed (0% power control transfer error rate) with a PC step size of 1dB, a PC dynamic range of 72dB and a PC frequency of 800Hz.

A comparison was performed between the proposed Multiuser S-WLMS and Correlator-CEF [CHO04], which we sometimes call, simply, the Correlator. A Rake receiver was used and the performance was measured in terms of FER and BER. Table 3.1 represents the performance evaluation in terms of BER for adaptive step size values fixed to optimal value, $\mu^{(\mathrm{opt})}$, or dynamically adjusted using equation (3.33). This evaluation is done for a given cdma2000 system/channel set-up. The simulation, whose results are provided in the figures below, is conducted using (3.33) instead of $\mu^{(\mathrm{opt})}$. As we argued the adaptive step size μ is, only, optimal for the mentioned system/channel condition (Table 3.1). The last row in Table 3.1 provides the corresponding BER with perfect channel estimates. Notice the BER-loss compared to the perfect channel estimate situation. As we can further read from the figures below (where (3.33) is used) the BER degradation is not that high so that the Multiuser S-WLMS method, still, keeps substantial gains compared to the Correlator.

Table 3.1: BER performance based on the optimal values of μ, $\mu^{(\text{opt})}$, and the dynamically adjusted value, μ_i, using (3.33) for a given system/channel set-up in cdma2000 system.

K (Users)	14	14	14	10	10	4	4
Channel Type	CSC1	CSC1	CSC2	PA	VA	CSC1	CSC2
Speed (km/h)	8	30	100	3	90	30	60
E_b/N_0 (dB)	10	10	10	10	10	10	10
Data Rates (kb/s)	9.6	9.6	9.6	9.6	9.6	9.6/2×38.6	9.6/2×38.6
$\mu^{(\text{opt})}$	0.123	0.131	0.141	0.183	0.225	0.206	0.214
BER using $\mu^{(\text{opt})}$	0.0030	0.0032	0.0025	0.031	0.0097	0.0061/0.131	0.0003/0.093
BER using (3.33)	0.012	0.0047	0.0039	0.034	0.011	0.0075/0.152	0.0005/0.107
BER, perfect channel	0.0020	0.0022	0.0018	0.021	0.0069	0.0045/0.083	0.0002/0.066

- **3GPP2 C-S0011-B test channels, Pedestrian A and Vehicular A channels**

In Figures 3.1, 3.2 and 3.3, 14 users signaling at a data rate of 9.6kb/s in CSC1 (8km/h, 30km/h) and CSC2 (100km/h) were simulated. The simulation results show the substantial gains of Multiuser S-WLMS compared with those of the Correlator. Depending on type of channel and mobile speeds, the gains may even exceed 3dB. The filter taps were optimized for speed ranges from 0 to 150km/h. This optimization explains the differences in gains as a function of the speed. At 8km/h (30km/h) over CSC1 channel, Multiuser S-WLMS outperforms the Correlator by more than 4 dB (3 dB) at 2% BER. At 100km/h over CSC2 channel type, we recorded a gain of 1.5 dB at 1% BER.

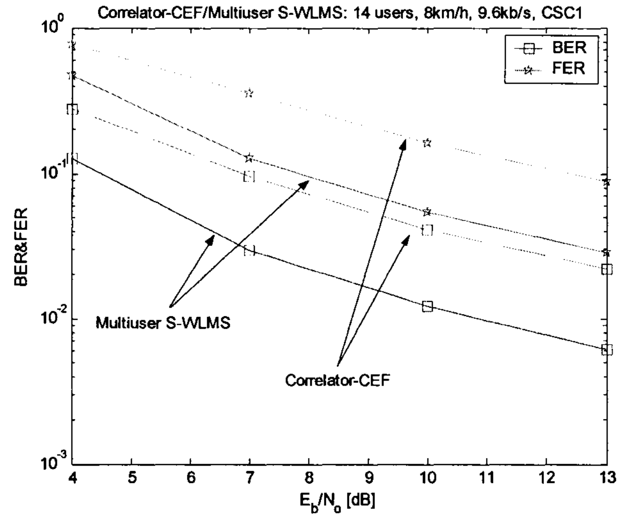


Figure 3.1. Multiuser S-WLMS versus Correlator, FER & BER performances, 14 users, 9.6kb/s, 8km/h in CSC1. RAKE receiver is used.

The differences in gains are due to the fact that both methods were optimized for relatively high speed centered at 60km/h using a bell shaped speed profile (c.f. (3.24)).

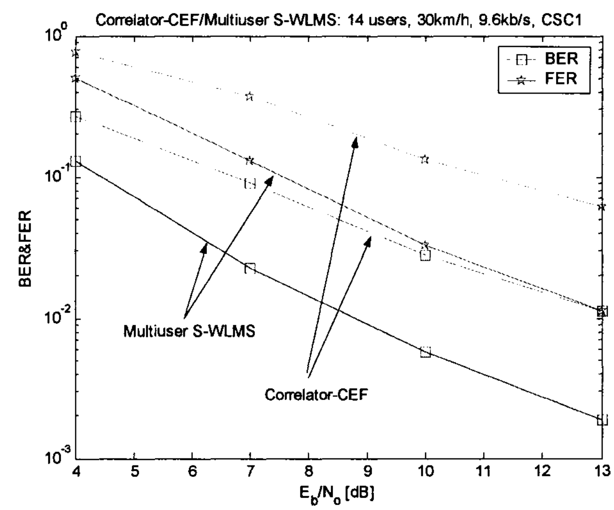


Figure 3.2. Multiuser S-WLMS versus Correlator, FER & BER performances, 14 users, 9.6kb/s, 30km/h in CSC1. RAKE receiver is used.

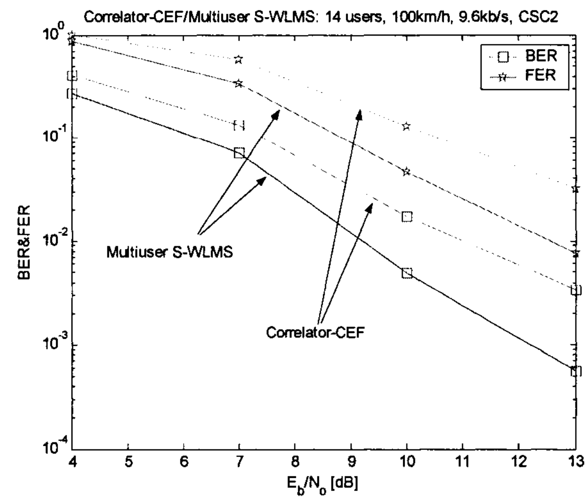


Figure 3.3. Multiuser S-WLMS versus Correlator, FER & BER performances, 14 users, 9.6kb/s, 100km/h in CSC2. RAKE receiver is used.

For more realistic channel settings, Figure 3.4 depicts the performance in pedestrian A channel at 3km/h and vehicular A channel at 90km/h. Note that the gains are as high as 2dB. It is worth mentioning that 10 users signaling at a data rate of 9.6kb/s were simulated.

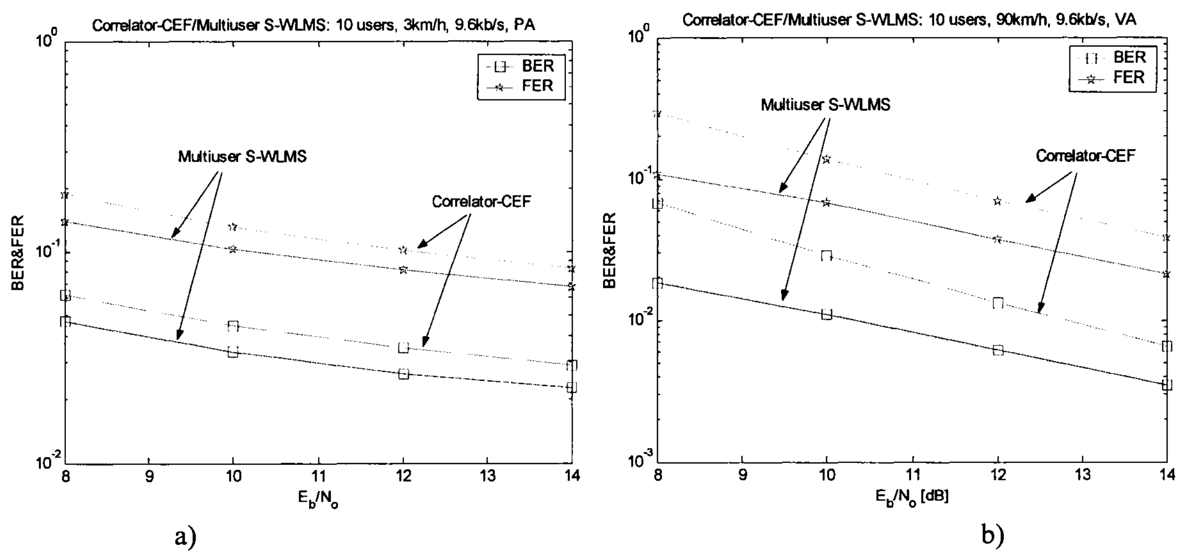


Figure 3.4. Multiuser S-WLMS versus Correlator, FER & BER performances, 10 users, 9.6kb/s, a) 3km/h in Pedestrian A (PA) channel, b) 90km/h in Vehicular A (VA) channel. RAKE receiver is used.

- **Mixed rate configuration**

As for the mixed rate settings, the mixed rate signaling configuration used for simulations (many configurations are possible, of course) is as follows: 4 users are simulated in CSC1 (30km/h) (Figure 3.5.a) and CSC2 (60km/h) (Figure 3.5.b). Each user signals at three data rates, namely, 9.6kb/s, 38.6kb/s and 38.6kb/s on the reverse forward channel (R-FCH), reverse primary supplemental channel (R-SCH1) and reverse secondary supplemental channel (R-SCH2) respectively.

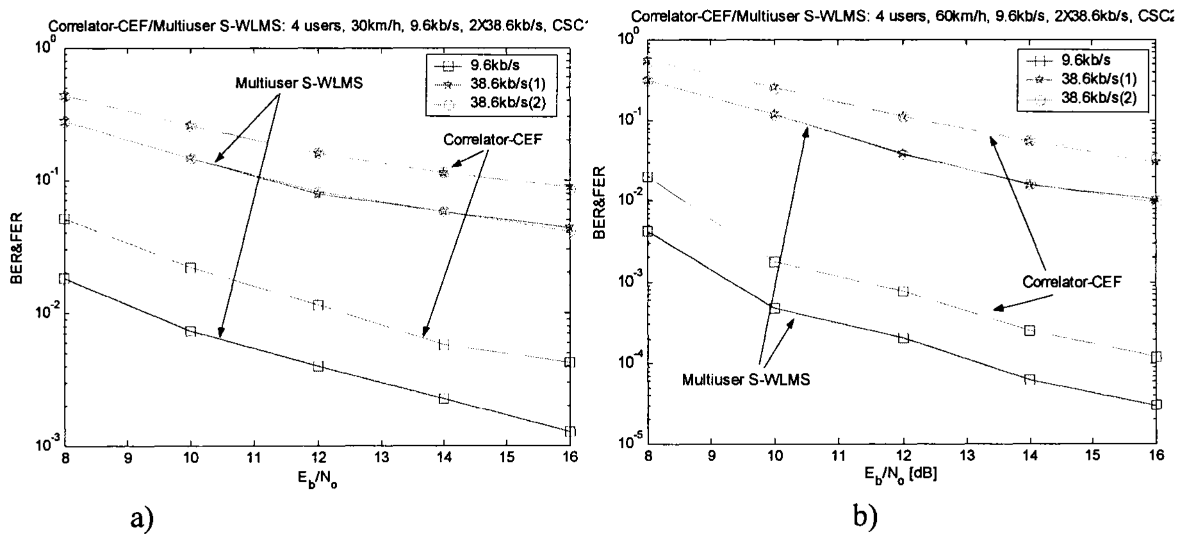
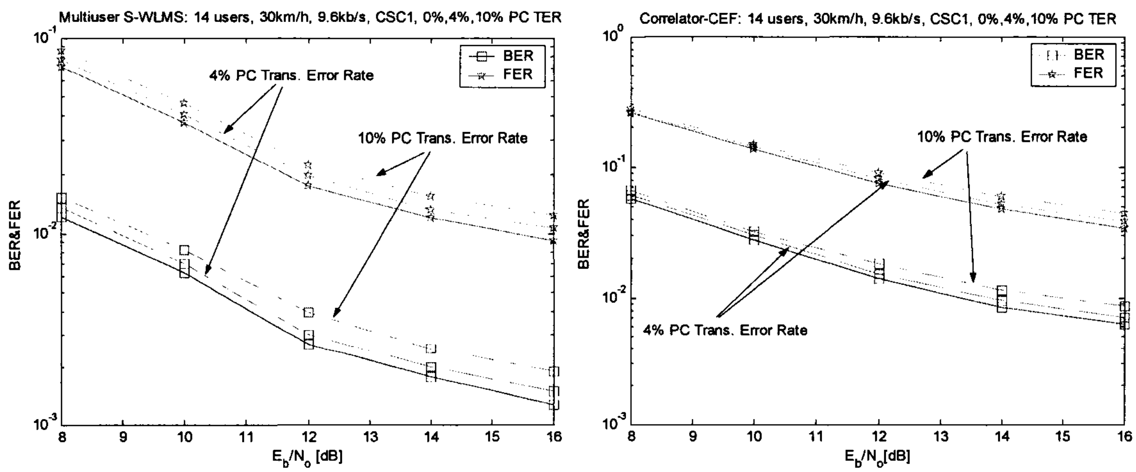


Figure 3.5. Mixed Rate: Multiuser S-WLMS versus Correlator, FER & BER performances, 4 users signalling each 9.6kb/s voice information over primary channel, 38.6kb/s data rate over supplemental channel 1 and 38.6kb/s over supplemental channel 2. a) 30km/h in CSC1 channel, b) 60km/h CSC2 channel. RAKE receiver is used.

The FER curves, depicted in Figure 3.5, show a 4-users mixed rate simulation results in CSC1 and CSC2 channels at mobile speeds of 30km/h and 60km/h respectively. The proposed multiuser S-WLMS maintains a superior performance at all data rates. A gain close to 2dB can be recognized. Note that gains are relatively low at 60km/h because of the optimized Correlator-CEF design.

- **Power control imperfection impact**

A power control mechanism is highly recommended and essential for DS-CDMA systems using a Correlator and Rake receiver for channel estimation and data detection respectively. Analytically analysing the impact of power control transfer error rate is mathematically overwhelming. The impact of power control imperfection must be analyzed through simulations for two main reasons. First, if a channel estimation method demonstrates less sensitivity to power control imperfection, one can solve the problem of designing a tight power control procedure. Second the designed channel estimation must maintain its performance superiority, even with power control imperfections. 0%, 4% and 10% power control transfer error rates (PC TER) are considered (Figure 3.6). The Multiuser S-WLMS appears slightly less sensitive to PC TER compared with the Correlator-CEF. This is obvious from the flat BER/FER curves exhibited by the Correlator-CEF. At a BER of 1%, the loss due to the 10% PC TER is less than 1dB using the proposed Multiuser S-WLMS compared with the 1.5dB loss exhibited by the Correlator-CEF.



a) b)

Figure 3.6. Effect of power control transfer error rate: FER & BER performances, 14 users, 9.6kb/s, 30km/h in CSC1 channel at 0, 4 and 10% PC transfer error rates a) Multiuser S-WLMS, b) Correlator-CEF. RAKE receiver is used.

3.6-2 WCDMA system

Consider a WCDMA system with one receiving antenna for 12.2kb/s, 64kb/s and 384kb/s data rates in compliance with 3GPP. The users are assumed to asynchronously access the system. Pedestrian A channel at 3km/h mobile speeds and Vehicular A channel, with uniformly distributed mobile speeds between 0km/h and 120km/h as depicted in figure 3.7, are considered. Perfect power control is implemented (0% power control transfer error rate) with a PC step size of 1dB, a PC dynamic range of 80dB and a PC frequency of 1500Hz.

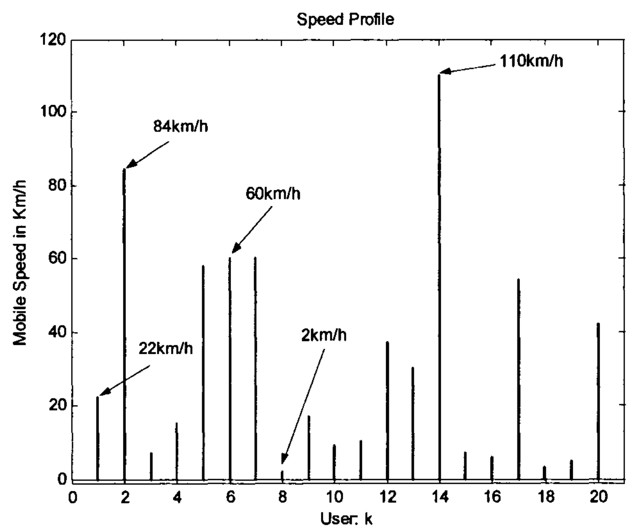


Figure 3.7. Uniformly distributed mobile speeds (U~[0 120]Km/h) used for simulations in vehicular A channel.

The comparison was performed between Multiuser S-WLMS and Correlator-CEF along with different detectors, namely, the Rake receiver, multistage hard PIC (Hard IC) and multistage soft PIC (Soft IC) [YUE00] and [CHE03].

The data rates considered were 12.2kb/s, 64kb/s and 384kb/s with 20, 10 and 3 users respectively. The first simulation results depict the performance of the proposed Multiuser S-WLMS in pedestrian A at 3km/h (Figure 3.8). The same results were produced using a vehicular A type of channel with mobile speeds uniformly distributed (Figure 3.7) over 0km/h to 120km/h interval (Figure 3.9).

Both long and short code WCDMA systems were considered. For the Rake receiver, long and short code WCDMA systems were simulated, whereas only short codes were implemented for other detection methods. The legend in Figures 3.8 and 3.9 indicates the respective detectors where a given detector performance is represented by two BER curves. A continuous line represents the detector's performances where the pilot contribution was removed from the received signal prior to any data detection, while a dotted line represents the performances without pilot extraction prior to data detection process.

For reference purposes, the Rake receiver with perfect channel estimates using long codes is represented by a continuous solid line with no symbol. Using either a Correlator-CEF or a multiuser S-WLMS channel estimator, the Rake receiver with long codes is represented by a continuous solid line with a "circle" symbol, whereas a dotted line with a "circle" symbol represents the Rake performance using short codes and pilot extraction prior to data detection. As in cdma2000 context, the last rows, from table 3.2, represent the respective detectors' BER for a specified system/channel condition using $\mu^{(opt)}$ and (3.33), and the detectors' BER using perfect channel estimates. Likewise, equation (3.33) is used to conduct the simulations. The results are ploted in figures 3.8 and 3.9.

Table 3.2: BER performance based on the optimal values of μ, $\mu^{(\mathrm{opt})}$, and the dynamically adjusted value, μ_i, using (3.33) for a given system/channel set-up in WCDMA system.

K (Users)		20	10	3	20	10	3
Channel Type		PA	PA	PA	VA	VA	VA
Speed (km/h)		3	3	3	60	60	60
E_b/N_0 (dB)		10	10	10	10	10	10
Data Rates (kb/s)		12.2	64	384	12.2	64	384
$\mu^{(\mathrm{opt})}$		0.231	0.210	0.192	0.284	0.263	0.237
BER using $\mu^{(\mathrm{opt})}$	Rake	0.041	0.067	0.045	0.042	0.070	0.061
	Hard IC	0.015	0.023	0.017	0.017	0.027	0.025
	Soft IC	0.013	0.011	0.0092	0.014	0.016	0.0097
BER using (3.33)	Rake	0.046	0.070	0.059	0.047	0.077	0.066
	Hard IC	0.019	0.031	0.028	0.021	0.033	0.031
	Soft IC	0.016	0.017	0.013	0.018	0.019	0.012
BER using perfect channel	Rake	0.039	0.065	0.040	0.035	0.065	0.058
	Hard IC	0.0065	0.011	0.0074	0.0068	0.014	0.01
	Soft IC	0.0057	0.0048	0.001	0.0061	0.006	0.002

Interesting conclusions can be drawn from Figures 3.8 and 3.9:

- For the 12.2kb/s voice rate, the Rake receiver in long code WCDMA system shows acceptable performances even when compared with the Soft IC. This observation caused the WCDMA base station (BS) manufacturers to back away from, or at least delay, the idea of introducing multiuser detectors. As can be seen, a 1dB's Rake loss compared with a Soft IC might not motivate the introduction of an advanced detection methods (such as Soft IC) as the latter represents a substantial additional computational cost that is unsuitable for the current DSP-based BS.

  From Figures 3.8.a and 3.8.b (3.9.a and 3.9.b), it is apparent that performance improvement can be obtained using a superior channel estimator. From these figures, the proposed multiuser S-WLMS shows a close to 2dB (4dB) gain at a complexity increase of 2 to 3 times that of the Correlator.

- For the 64kb/s and 384kb/s data rates, Figures 3.8.c,d,e,f (3.9.c,d,e,f) demonstrate the importance of using multiuser detectors instead of the Rake receiver to reach a BER as low as 5%. The multiuser detectors show performance gains that can reach even 4dB's over the Rake receiver using a Correlator-CEF; in addition, extra 2 to 4dB's are attainable when our proposed channel estimator is used.

This observation is apparent from Figure 3.9.c,d,e,f, when mobiles are allowed to move at different speeds. The simulation scenarios in Figures 3.9.c,d,e,f show the importance of utilizing a more accurate channel estimation than the Correlator. In these scenarios, the correlator-CEF tends to penalize the multiuser detectors, since it can be seen that the BER curve barely reaches the 5% BER. When it is easily attained using multiuser S-WLMS, gains of more than 4dB's are recorded.

In addition to the implementation advantages mentioned in the introduction, two important strategical reasons for using the multiuser S-WLMS with respect to system performance are:

1. The 3G mobile wireless industry will continue to enjoy using the Rake receiver at higher performances by introducing the proposed multiuser S-WLMS.
2. If the 3G mobile wireless industry is willing to introduce multiuser detectors, especially to respond to the high data rate demand, the proposed Multiuser S-WLMS will ensure that the desired performance is maintained and the performance will not be compromised.

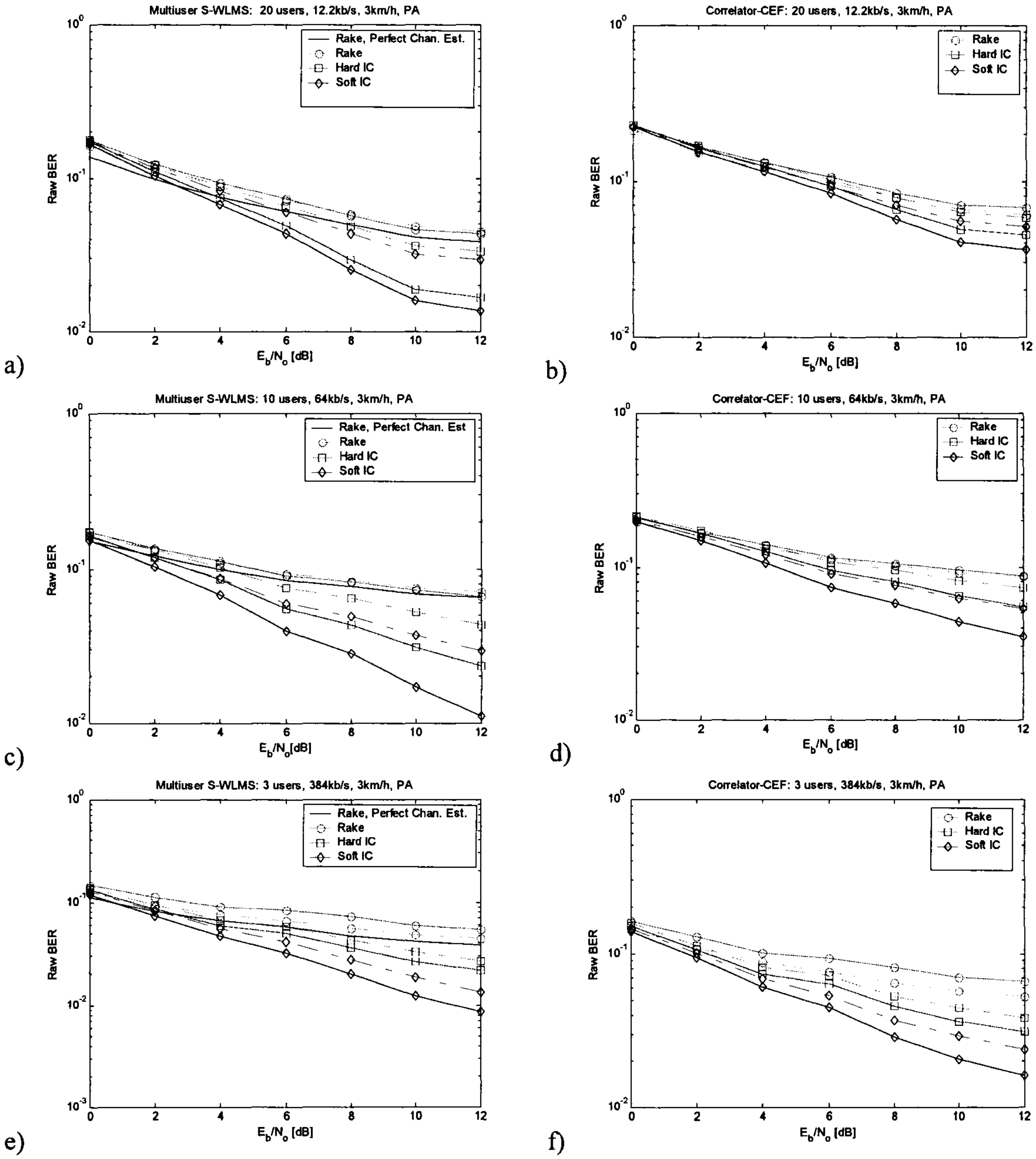


Figure 3.8. Raw BER performance in Pedestrian A (PA) channel at 3km/h, a) Multiuser S-WLMS, 20 users, 12.2kb/s, b) Correlator-CEF, 20 users, 12.2kb/s, c) Multiuser S-WLMS, 10 users, 64kb/s, d) Correlator-CEF, 10 users, 64kb/s, e) Multiuser S-WLMS, 3 users, 384kb/s, f) Correlator-CEF, 3 users, 384kb/s.

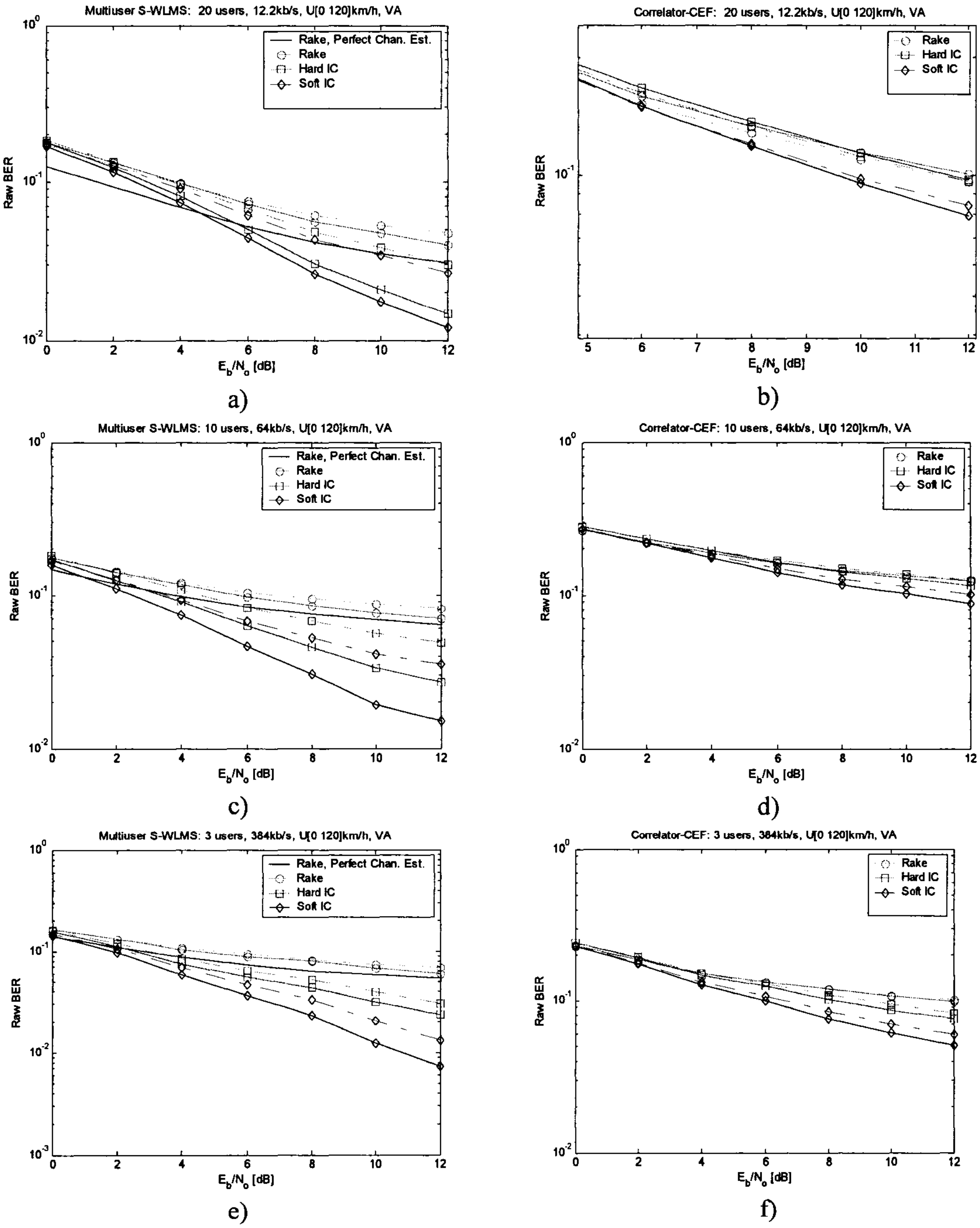


Figure 3.9. Raw BER performance in Vehicular A (VA) channel at mobile speeds uniformly distributed in [0 120]Km/h, a) Multiuser S-WLMS, 20 users, 12.2kb/s, b) Correlator-CEF, 20 users, 12.2kb/s, c) Multiuser S-WLMS, 10 users, 64kb/s, d) Correlator-CEF, 10 users, 64kb/s, e) Multiuser S-WLMS, 3 users, 384kb/s, f) Correlator-CEF, 3 users, 384kb/s.

The last remark stems from the relatively low dB-gains at high data rates for the proposed multiuser S-WLMS over the Correlator. This is explained by the fact that a small number of users was considered in the simulations. Under such conditions, the correlation based channel acquisition is as good, even, as the maximum likelihood (ML) based method for a small number of mobile users.

We conclude that for the uplink DS-CDMA system, all channel estimators, from the less complex (Correlator) to the complex one (ML), provide almost similar performances when a very low number of users is used. The reader is invited to check [SEN98a] and [SEN98b] for the probability of acquisition curves at small number of users.

3.6-3 Complexity evaluation[8]

The complexity counts are based on one power control group worth of processing time according to 3GPP2 (1.25 ms). Radio configuration 3 (RC3) was considered. The required number of additions and multiplications are depicted in Figure 3.10 as a function of the number of users. The complexity analysis was conducted using the following settings, with the same setting used to produce the simulation results in subsection 3.6:

1. One radio frame is 20ms.
2. One Power Control Group (PCG) is 1.25ms (equivalent to 1536 chips).
3. Oversampling rate is 8 samples/chip to construct the received signal.
4. For one user, the maximum multipath delay spread is 18 chips (corresponding to 3GPP2 3 path channel, CSC2).

[8] Only the cdma2000 system was considered for the implementation complexity analysis as the same figures and conclusions can be produced for WCDMA. The complexity analysis includes the ML in [BHA02] for comparison.

5. The delay spread among all the users in chips is defined as the time difference between the first path of the closest user to the last path of the farthest user. In the simulation, it was set to 128 chips.
6. Only the computations for the channel identification have been taken into account. The conditional check will be counted as 1 addition for the simplification. One subtraction is considered as one addition when considering the computation complexity.
7. Finger management complexity is not counted because it usually involves sorting every 4 frames (step 4 in the proposed Multiuser S-WLMS algorithm.)
8. Search resolution is 0.5 chips.
9. Number of Correlators in parallel is 128.
10. Correlation period is 384 chips.
11. Keep track of 4 most possible delays.

The following remarks are observed from Figure 3.10:

1. Complexity in terms of the number multiplications is negligible for the three methods as compared with the number of additions.
2. The gradient descent ML [BHA02] has a complexity that is approximatly 100 times that of the Correlator.
3. The Multiuser S-WLMS possesses a complexity on the order of 2 to 3 compared with the Correlator.

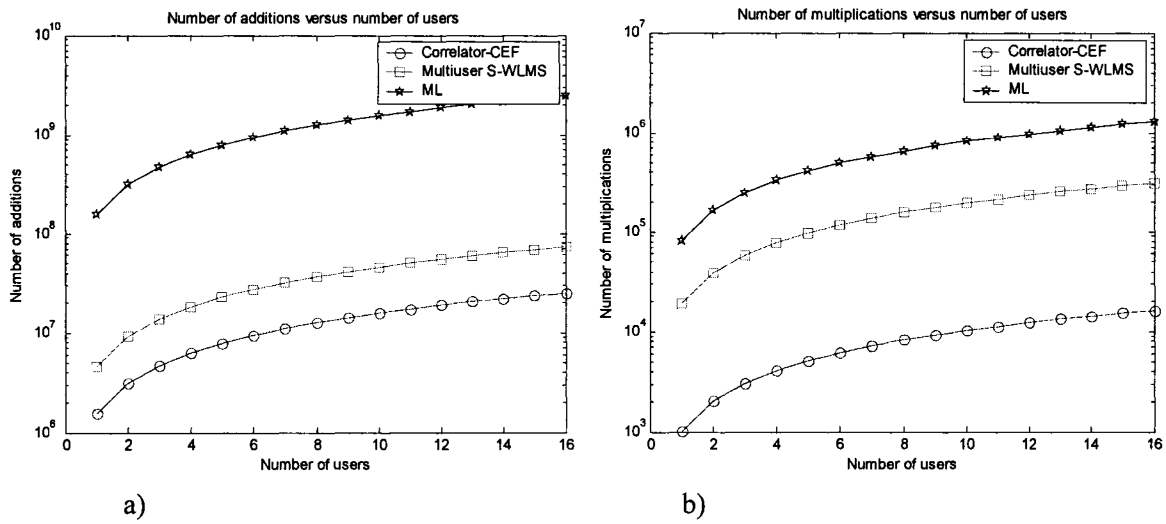


Figure 3.10. Computational complexity in 1.25ms PCG worth of time for Multiuser S-WLMS, Correlator and gradient descent ML, a) Number of additions, b) Number of multiplications.

3.7 Conclusion

Usually driven by the algorithm complexity, the choice of an adaptation family (LMS, RLS, KALMAN, etc...) constitutes the key for an efficient architecture implementation. The proposed methodology and design based on a multiuser LMS structure augmented with a smoothing/prediction FIR proved through simulations to be effective in highly interesting multipath channels, Doppler shifts and data rate scenarios [AHM05b]. Another key aspect of the design lies in the use of the average Doppler profile, where the averaging is done over a continuum speed interval that covers a wide range while performing a multiuser parameter estimation. The Multiuser S-WLMS offers a superior performance compared with the Correlator-CEF method. As regards implementation, the dominant LMS/FIR structure renders it attractive for throughput/area VLSI optimization techniques by exploiting the intrinsic parallelism. Nevertheless, the very small number of multiplications compared with the number of additions makes it feasible for general purpose DSP implementation.

With regard to system performance, the Multiuser S-WLMS is just 2 to 3 times more complex than the Correlator-CEF. The Multiuser S-WLMS offers a certain flexibility for the 3G mobile wireless industry, which includes (1) updating and thereby enhancing the current Rake receiver based base stations and, (2) introducing multiuser detectors and/or new schemes such as beam formation for smart antennas and adaptive coding where a "good" channel estimate is usually a requirement for the new generation of 3G mobile wireless systems. Unlike the Correlator-CEF, the proposed Multiuser S-WLMS will not represent a bottleneck for new generations.

The Multiuser S-WLMS is uniquely designed for both cdma2000 and WCDMA systems. Ongoing work in the future will focus on the design of efficient low power VLSI architectures; in the meantime, side tasks deal with optimized DSP implementations.

4. Iterative (Turbo) Multiuser Detection Algorithms for Coded DS-CDMA Systems Utilizing BPSK Modulation

In this chapter, two low complexity turbo detection receivers for coded DS-CDMA signals utilizing BPSK as a modulation scheme in multipath channels are derived [AHM05c]. When compared with the novel conventional soft IC-MMSE developed by Wang and Poor, the first receiver provides an average 2dB gain with fewer iterations. The second scheme, of a much lower complexity, provides performances comparable to the conventional IC-MMSE. Simulation results for performance evaluation are conducted under highly interesting scenarios including asynchronous multipath channels, near far problem, time varying channels, and multirate systems. Finally, we briefly end up the discussion by investigating the convergence behavior, using EXIT charts, of the first proposed Soft detector [AHM05c].

4.1 Related Works

Many works on multiuser detection focused on uncoded CDMA systems, i.e., the demodulation of multiuser signals. Since, in practice, most CDMA systems employ error control coding and interleaving, recent works in this area have addressed multiuser detection for coded CDMA systems. In [GIA96a], it is shown that the optimal decoding scheme for an asynchronous convolutionally coded CDMA system combines the trellises of both the asynchronous multiuser detector and the convolutional code, resulting in a prohibitive computational complexity of $O(2^{\nu K})$, where K is the number of users in the channel, and ν is the code constraint length. In [GIA96b], some low-complexity receivers that perform multiuser symbol detection and decoding either separately or jointly are studied. Iterative decoding schemes for convolutionally or turbo coded CDMA systems are proposed in [REE98] and [WAN99]. The main difference in the structure of those schemes is the type of SISO multiuser detector used. In [REE98] and [MOH98], a full complexity SISO multiuser detector is proposed for convolutional coded synchronous CDMA systems, resulting in a computational complexity of $O(2^K)$ for the multiuser detector. The work was then extended to a coded asynchronous CDMA system [ALE99], where the asynchronous random CDMA channel was viewed as a time varying convolutional code. Based on the concept of a time varying convolutional code, forward and backward recursions similar to those of the BCJR algorithm [BAH74] were proposed for the SISO multiuser detector, and the M-algorithm was presented to reduce the computational complexity. In [NAS98], a stepwise difference calculation is presented to determine all the 2^K possible likelihood values for synchronous CDMA systems. In addition, a reduced-complexity multiuser detector was proposed to keep a variable number of likelihood values at each step as determined by comparing likelihood values with a threshold. The max-log-MAP algorithm [HAG96a] is then used to obtain the LLR for each code bit. Another type of iterative decoding is based on the soft interference cancellation [HAG96b], [ALE98] and [WAN99], where the soft estimates of the code bits are represented by real numbers rather than hard-decision values and are combined to produce estimates of the LLR's of the code bits. The LLR's of the code bits are then fed to the SISO channel decoders for the next decoding iteration. [WAN99] suggests a low complexity SISO multiuser detector based on the soft instantaneous MMSE interference cancellation/suppression, based, in turn, on the a priori likelihood ratios of the code bits of all users provided by the SISO channel decoder from the previous stages. The same idea was exploited in [GAM00] for AWGN channels by designing filters in the MMSE sense for each user using a priori information provided by SISO channel decoder implemented by SOVA. Unfortunately, the receiver does not provide outputs in terms of LLR's. [BOU00], in a unified framework, summarizes some current low complexity turbo detectors.

Motivated by implementation issues, we suggest in [AHM04] a low complexity iterative detection structure at a computational load close to I-Rake, where I is the number of iterations. Encouraged by performance improvement when the complexity was kept low, we developed in [AHM05a] a reduced complexity iterative structure based on the design of feed-forward and feed-back filters in the MMSE sense.

Recent advances highlight the use of EXIT charts to evaluate the iterative (turbo) processing due to the information exchange in terms of LLRs between the modules involved in the soft detection and/or decoding. EXIT charts are introduced to investigate the convergence aspect of turbo codes [BRI01]. Recently, they are used to evaluate the performance of the turbo receivers in MIMO systems [HER03]. It is apparent that EXIT charts offer more than just a convergence analysis, they depict in a simple chart the speed of convergence and the BER can be estimated accordingly [BRI01]. We will use, in addition to the traditional SNR-BER curve, EXIT charts to analyze the performance of the proposed soft detector.

It is worth mentioning that exploiting the BPSK modulation for the design of a linear receiver is not a new issue [TUL01], [BUZ01a], [HER05]; it has not yet, however, been considered in the design of turbo detectors. Our contribution may, therefore, be viewed as a development of the results in [WAN99], [GAM00] and [AFF02] for BPSK signals. The contribution investigates, also, the additional potential gain that can be achieved when exploiting the BPSK modulation.

In the present work two low complexity turbo detection receivers for convolutionally coded DS-CDMA systems utilizing BPSK modulation in multipath channels are presented. The first receiver is based on a soft Interference Canceller (IC) followed by a MMSE filters, whose coefficients are thought to be the solution of an MMSE-optimization problem based on a (forced) real valued filter output rather than a complex one as in conventional MMSE receivers used in [WAN99]. The "raw" solution will lead to a computational complexity due to the inverse of a Hermitian complex valued matrix. Such complexity can be overcome by properly redefining the filter output that leads to an inverse of a real valued symmetrical matrix. Further complexity reduction is achieved by using the steepest descent method. In the second structure we reformulate the traditional constrained filter design by considering a real valued filter output along with real valued constraints.

In section 4.2 of this chapter, a convolutionally coded DS-CDMA model for multipath channel is presented, followed in section 4.3 by the proposed SISO low complexity detectors. Simulation results along with the EXIT charts are presented in section 4.4 with discussions, followed by a conclusion in section 4.5.

4.2 Convolutionally Coded DS-CDMA Model

For the sake of brevity we will focus on the single rate model described in Chapter 2; the multirate system can be drawn easily [BUZ01b] and is only considered in the simulations.

Recall from Chapter 2 that the received signal represented in $2N \times 1$ observation vector $\mathbf{r}_{i/i+1}$ is given below

$$\begin{bmatrix} \mathbf{r}_i \\ \mathbf{r}_{i+1} \end{bmatrix} = \begin{bmatrix} [\mathbf{CH}]_0 & [\mathbf{CH}]_1 & \mathbf{0}_{N\times K} \\ \mathbf{0}_{N\times K} & [\mathbf{CH}]_0 & [\mathbf{CH}]_1 \end{bmatrix} \mathbf{b}_{i/i+1} + \mathbf{v}_{i/i+1} = \mathbf{r}_{i/i+1} \tag{4.1}$$

where $[\mathbf{CH}]_0$ and $[\mathbf{CH}]_1$ are matrices constructed from the odd and even numbered columns of $\mathbf{CH}$, respectively, $\mathbf{v}_{i/i+1} = [\mathbf{v}_i^{\mathrm{T}}\ \mathbf{v}_{i+1}^{\mathrm{T}}]^{\mathrm{T}}$ and $\mathbf{b}_{i/i+1} = \left[b_{1,i-1}, b_{2,i-1}, \cdots b_{k,i-1}, \cdots, b_{K,i-1},\ b_{1,i}, b_{2,i}, \cdots b_{k,i}, \cdots, b_{K,i}, b_{1,i+1}, b_{2,i+1}, \cdots b_{k,i+1}, \cdots, b_{K,i+1}\right]^{\mathrm{T}}$. In the sequel, let $\mathbf{A} = \begin{bmatrix} [\mathbf{CH}]_0 & [\mathbf{CH}]_1 & \mathbf{0}_{N\times K} \\ \mathbf{0}_{N\times K} & [\mathbf{CH}]_0 & [\mathbf{CH}]_1 \end{bmatrix}$. Figure 4.1.a depicts the convolutionally coded DS-CDMA model described in Chapter 2.

4.3 Iterative Turbo Detector

4.3-1 Iterative detector

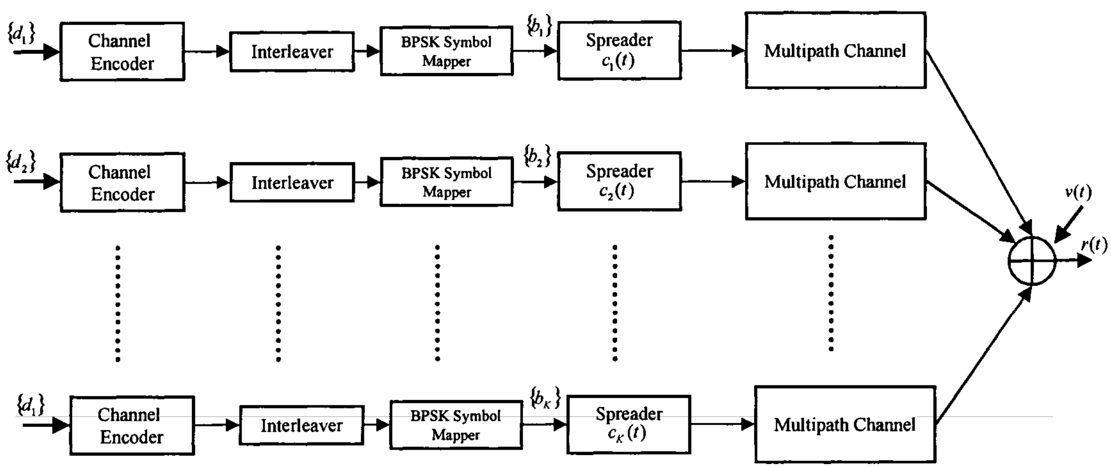
The iterative receiver shown in Figure 4.1.b consists mainly of two stages: a cascade of the SISO detector and a bank of the K-parallel full codeword SISO channel decoder [WAN99]. In the overall receiver structure in Figure 4.1.b, our contribution focuses on the SISO detector design.

Before deriving the new SISO detector, it may be noted that any linear receiver makes a decision as to the bit $b_k(i)$ according to

$$\hat{b}_k(i) = \mathrm{sgn}\left(\Re\left(\mathbf{w}_k^{\mathrm{H}}(i)\mathbf{r}_{i/i+1}\right)\right) \tag{4.2}$$

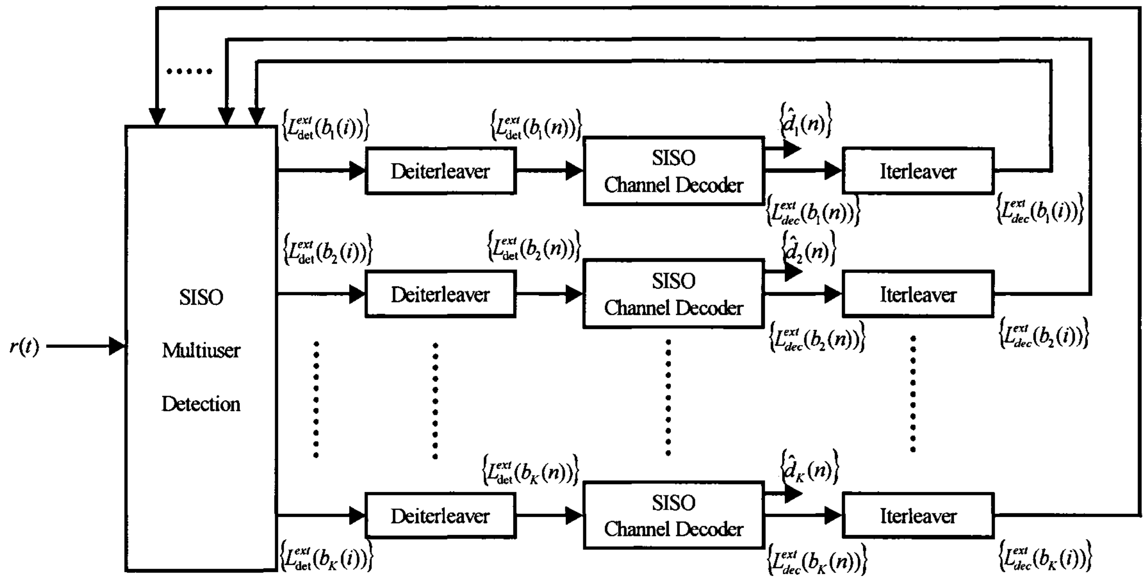
where $\mathrm{sgn}(\bullet)$ denotes the signum function, $\Re(\bullet)$ denotes the real part, and $(\bullet)^{\mathrm{H}}$ denotes the conjugate-transpose. The conventional MMSE detector[9] selects $\mathbf{w}_k(i)$ according to

$$\mathbf{w}_k(i) = \arg\min_{\mathbf{w}\in\mathbb{C}^{2N\times1}} E\left\{\left|\mathbf{w}^{\mathrm{H}}(i)\mathbf{r}_{i/i+1} - b_k(i)\right|^2\right\} \tag{4.3}$$



a)

[9] We mean by conventional MMSE detector the receiver developed in [MAD94] and used in [WAN99].



b)

Figure 4.1. Iterative multiuser detection for convolutionally coded DS-CDMA, a) Transmitter configuration, b) Receiver's structure

On the other hand, since

$$\min_{\mathbf{w}\in\mathbb{C}^{2N\times 1}} E\left\{\left|\mathbf{w}^{H}(i)\mathbf{r}_{i/i+1}-b_k(i)\right|^2\right\} \geq \min_{\mathbf{w}\in\mathbb{C}^{2N\times 1}} E\left\{\left(\Re\left(\mathbf{w}^{H}(i)\mathbf{r}_{i/i+1}\right)-b_k(i)\right)^2\right\} \tag{4.4}$$

it is understood that defining [BOUZ01a]

$$\mathbf{w}_k(i)=\arg\min_{\mathbf{w}\in\mathbb{C}^{2N\times 1}} E\left\{\left(\Re\left(\mathbf{w}^{H}(i)\mathbf{r}_{i/i+1}\right)-b_k(i)\right)^2\right\} \tag{4.5}$$

the receiver performance based on decision rule (4.5) is necessarily not inferior to the that of the conventional MMSE[10]. The solution to (4.5) can not be found by applying the usual orthogonality principle, i.e., (4.5) is not a conventional MMSE problem [BUZ01a].

To solve (4.5), first notice that [BUZ01a]

[10] i.e., the performance of the receiver based on (4.5) is at least as good as that of a receiver based on (4.3).

$$\begin{aligned}\Re\left(\mathbf{w}^{\mathrm{H}}(i)\mathbf{r}_{i/i+1}\right) &= \frac{1}{2}\left(\mathbf{w}^{\mathrm{H}}(i)\mathbf{r}_{i/i+1} + \mathbf{w}^{\mathrm{T}}(i)\mathbf{r}_{i/i+1}{}^{*}\right)\\ &= \frac{1}{2}\begin{bmatrix}\mathbf{w}(i)\\ \mathbf{w}^{*}(i)\end{bmatrix}^{\mathrm{H}}\begin{bmatrix}\mathbf{r}_{i/i+1}\\ \mathbf{r}_{i/i+1}{}^{*}\end{bmatrix}\\ &= \mathbf{w}_a^{\mathrm{H}}(i)\mathbf{r}_{a,i/i+1}\end{aligned} \tag{4.6}$$

As a result, solving (4.5) is equivalent to solving

$$\mathbf{w}_{a,k}(i) = \arg\min_{\mathbf{w}_a \in \mathbb{C}_a^{4\mathrm{N}\times 1}} E\left\{\left(\mathbf{w}_a^{\mathrm{H}}(i)\mathbf{r}_{a,i/i+1} - b_k(i)\right)^2\right\} \tag{4.7}$$

In (4.7), $\mathbb{C}_a^{4\mathrm{N}\times 1}$ is a vector space in field $\mathbb{C}$. Its elements are the augmented $4N$-dimensional complex vector whose first $2N$ entries are the complex conjugate of the last $2N$. The internal operation is the usual component-wise vector sum in $\mathbb{C}^{4\mathrm{N}\times 1}$, and the external operation is $\times : \alpha \in \mathbb{C}, \mathbf{w}_a \in \mathbb{C}_a^{4\mathrm{N}\times 1} \to \alpha\mathbf{w}_a = \begin{bmatrix}\alpha\mathbf{w}\\ \alpha^{*}\mathbf{w}^{*}\end{bmatrix}$ [BUZ01a]. We will return to this new family of MMSE [BUZ01a], apply it at the output of SISO IC and present it below.

As it can be expected, departing from (4.6) will lead to solving problems involving inversion of complex valued matrices (c.f. 4.15). This problem can be overcome by reformulating (4.6) in such a way an inverse of real valued matrix will be required. An extension, we will be soon developing as a way of reducing the computational complexity without any sacrifice in the performances.

Based on the a priori LLR of the code bits of all users, $L_{\mathrm{dec}}(b_k(i))$, $1 \le k \le K, 0 \le i \le M-1$, provided by the SISO channel decoder from the previous stage, soft estimate of the user codes bits, as $\tilde{b}_k(i) = \frac{e^{L_{\mathrm{dec}}(b_k(i))}-1}{e^{L_{\mathrm{dec}}(b_k(i))}+1} = \tilde{b}_{k,i}$, are formed.

Denote

$$\tilde{\mathbf{b}}_{i/i+1} = \left[\tilde{b}_{1,i-1}, \tilde{b}_{2,i-1}, \cdots \tilde{b}_{k,i-1}, \cdots, \tilde{b}_{K,i-1},\ \tilde{b}_{1,i}, \tilde{b}_{2,i}, \cdots \tilde{b}_{k,i}, \cdots, \tilde{b}_{K,i}, \tilde{b}_{1,i+1}, \tilde{b}_{2,i+1}, \cdots \tilde{b}_{k,i+1}, \cdots, \tilde{b}_{K,i+1}\right]^{\mathrm{T}} \tag{4.8}$$

and

$$\begin{aligned}\tilde{\mathbf{b}}_{k,i/i+1} &= \tilde{\mathbf{b}}_{i/i+1} - \tilde{b}_{k,i}\mathbf{e}_{K+k} \\ &= \begin{bmatrix} \tilde{b}_{1,i-1}, \tilde{b}_{2,i-1}, \cdots \tilde{b}_{k,i-1}, \cdots, \tilde{b}_{K,i-1}, \tilde{b}_{1,i}, \tilde{b}_{2,i}, \cdots, \tilde{b}_{k-1,i}, 0, \\ \qquad \tilde{b}_{k+1,i}, \cdots, \tilde{b}_{K,i}, \tilde{b}_{1,i+1}, \tilde{b}_{2,i+1}, \cdots \tilde{b}_{k,i+1}, \cdots, \tilde{b}_{K,i+1} \end{bmatrix}^{\mathrm{T}}\end{aligned} \tag{4.9}$$

where $\mathbf{e}_u$ denotes the u th unit vector from $\mathbb{R}^{3K}$.

Using (4.9), a soft interference cancellation is performed on the received discrete time signal $\mathbf{r}_{i/i+1}$ to obtain

$$\mathbf{r}_{k,i/i+1} = \mathbf{r}_{i/i+1} - \mathbf{A}\tilde{\mathbf{b}}_{k,i/i+1} \tag{4.10}$$

Now the new linear Conjugate[11] MMSE (CMMSE) filter developed in (4.5), (4.6) and (4.7) is applied to $\mathbf{r}_{k,i/i+1}$ to obtain

$$y_k(i) = \mathbf{w}_{a,k}^{\mathrm{H}}(i)\mathbf{r}_{a,k,i/i+1} \tag{4.11}$$

where the filter $\mathbf{w}_{a,k} \in \mathbb{C}^{4\mathrm{N}\times 1}$ is chosen to minimize the mean square error between the bit $b_k(i)$ and the real valued quantity $y_k(i)$ in (4.11), i.e.,

$$\begin{aligned}\mathbf{w}_{a,k}(i) &= \arg\min_{\mathbf{w}_a \in \mathbb{C}_a^{4\mathrm{N}\times 1}} E\left\{\left(\mathbf{w}_a^{\mathrm{H}}\mathbf{r}_{a,k,i/i+1} - b_k(i)\right)^2\right\} \\ &= \arg\min_{\mathbf{w}_a \in \mathbb{C}_a^{4\mathrm{N}\times 1}} \mathbf{w}_a^{\mathrm{H}} E\left\{\mathbf{r}_{a,k,i/i+1}\mathbf{r}_{a,k,i/i+1}^{\mathrm{H}}\right\}\mathbf{w}_a - \mathbf{w}_a^{\mathrm{H}} E\left\{b_k(i)\mathbf{r}_{a,k,i/i+1}\right\} - E\left\{b_k(i)\mathbf{r}_{a,k,i/i+1}\right\}^{\mathrm{H}}\mathbf{w}_a\end{aligned} \tag{4.12}$$

where

[11] The term *Conjugate* is due to the fact that, $\mathbf{r}_{k,i/i+1}$ and its *conjugate* $\mathbf{r}_{k,i/i+1}^{*}$ are used.

$$
\begin{aligned}
E\left\{\mathbf{r}_{a,k,i/i+1}\mathbf{r}_{a,k,i/i+1}{}^{\mathrm{H}}\right\} &= \begin{bmatrix} E\left\{\mathbf{r}_{k,i/i+1}\mathbf{r}_{k,i/i+1}{}^{\mathrm{H}}\right\} & E\left\{\mathbf{r}_{k,i/i+1}\mathbf{r}_{k,i/i+1}{}^{\mathrm{T}}\right\} \\ E\left\{\mathbf{r}_{k,i/i+1}{}^{*}\mathbf{r}_{k,i/i+1}{}^{\mathrm{H}}\right\} & E\left\{\mathbf{r}_{k,i/i+1}{}^{*}\mathbf{r}_{k,i/i+1}{}^{\mathrm{T}}\right\} \end{bmatrix} \\
&= \begin{bmatrix} \mathbf{A}\boldsymbol{\Lambda}_k(i)\mathbf{A}^{\mathrm{H}} + \sigma^2\mathbf{I}_{2N\times 2N} & \mathbf{A}\boldsymbol{\Lambda}_k(i)\mathbf{A}^{\mathrm{T}} \\ \mathbf{A}^{*}\boldsymbol{\Lambda}_k(i)\mathbf{A}^{\mathrm{H}} & \mathbf{A}^{*}\boldsymbol{\Lambda}_k(i)\mathbf{A}^{\mathrm{T}} + \sigma^2\mathbf{I}_{2N\times 2N} \end{bmatrix} \\
&= \boldsymbol{\Xi}_{a,k}(i)
\end{aligned}
\tag{4.13}
$$

$$
E\left\{b_k(i)\mathbf{r}_{a,k,i/i+1}\right\} = \begin{bmatrix} \mathbf{A}^{\mathrm{T}} & \mathbf{A}^{\mathrm{H}} \end{bmatrix}^{\mathrm{T}} \mathbf{e}_{K+k} = \mathbf{B}\mathbf{e}_{K+k} = \mathbf{B}_{K+k}
\tag{4.14}
$$

where

$$
\boldsymbol{\Lambda}_k(i) = \begin{bmatrix}
1-\tilde{b}_{1,i-1}{}^2 & 0 & 0 & 0 & 0 & & & 0 & & 0 & & 0 & & 0 & 0 \\
0 & 1-\tilde{b}_{2,i-1}{}^2 & 0 & \vdots & \vdots & & & & & & & & & & 0 \\
0 & 0 & \ddots & 0 & 0 & & & & & & & & & \vdots & \vdots \\
0 & 0 & 0 & 1-\tilde{b}_{K,i-1}{}^2 & 0 & & & & & & & & & & \\
0 & 0 & 0 & 0 & 1-\tilde{b}_{1,i}{}^2 & 0 & & & & & & & & & \\
 & 0 & 0 & 0 & & \ddots & 0 & & & & & & & & \\
 & & 0 & 0 & & & 1-\tilde{b}_{k-1,i}{}^2 & 0 & & & & & & & \\
 & & & 0 & & & 0 & 1 & 0 & \vdots & & \vdots & & \vdots & \vdots \\
 & & & & & & & & 1-\tilde{b}_{k+1,i}{}^2 & 0 & & & & & \\
 & & & & & & & & & \ddots & 0 & & & & \\
 & & & & & & \vdots & & & & 1-\tilde{b}_{K,i}{}^2 & 0 & & & \\
\vdots & \vdots & & \vdots & & & & & & & & 1-\tilde{b}_{1,i+1}{}^2 & 0 & 0 & 0 \\
 & & & & & & & & & & & 0 & 1-\tilde{b}_{2,i+1}{}^2 & 0 & 0 \\
0 & & & & & & & & & & & \vdots & \vdots & \ddots & 0 \\
0 & 0 & \cdots & 0 & & 0 & \cdots & & 0 & & 0 & 0 & 0 & 0 & 1-\tilde{b}_{K,i+1}{}^2
\end{bmatrix}
$$

and $\mathbf{r}_{a,k,i/i+1} = \begin{bmatrix} \mathbf{r}_{k,i/i+1} \\ \mathbf{r}_{k,i/i+1}^{*} \end{bmatrix}$.

The solution to (4.12) is given by

$$
\mathbf{w}_{a,k}(i) = \boldsymbol{\Xi}_{a,k}^{-1}(i)\mathbf{B}_{K+k}
\tag{4.15}
$$

Notice that, unlike the conventional MMSE receiver [WAN99], which is uniquely determined by data covariance matrix "$\mathbf{A}\boldsymbol{\Lambda}_k(i)\mathbf{A}^{\mathrm{H}} + \sigma^2\mathbf{I}_{2N\times 2N}$", the solution (4.15) depends, also, on "$\mathbf{A}\boldsymbol{\Lambda}_k(i)\mathbf{A}^{\mathrm{T}}$".

Unlike [BUZ01a], the expectation (4.12) is applied with respect to both the additive white Gaussian noise and the information sequence, and unlike [WAN99], the optimized cost function in (4.12) uses the right hand side of (4.4) instead of the left hand side.

In the first iteration, we set $\mathbf{\Lambda}_k(i) = \mathbf{I}_{3K\times 3K}$, which is equivalent to assuming that the code bits are uniformly distributed and equiprobable. Then the MMSE solution is equivalent to the conventional non-iterative CMMSE solution [BUZ01a]. At each iteration, $\widetilde{\mathbf{b}}_{k,i/i+1}$ is calculated using the SISO full codeword soft information in the form of LLR's.

Necessary to the turbo processing, the SISO detector proposed here should be amounted to provide LLR's instead of soft decisions $y_k(i)$. One way of doing this was to assume that the output of the soft MMSE filters to be Gaussian [WAN99]. In what follows, we assume that the output of the soft CMMSE detector $y_k(i)$ in (4.11) represents the output of an equivalent additive white Gaussian noise channel having $b_k(i)$ as its input symbol. This equivalent channel can be represented as [WAN99]

$$y_k(i) = \lambda_k(i) b_k(i) + z_k(i) \tag{4.16}$$

where $\lambda_k(i)$ is the equivalent amplitude at instant i for the kth user SISO filter output, and $z_k(i) \sim N\left(0, \rho_k(i)^2\right)$ is a Gaussian noise sample [WAN99]. Using (4.1) and (4.16), the parameters $\lambda_k(i)$ and $\rho_k(i)^2$ can be computed as follows, where the expectation is taken with respect to the code bits and $\mathbf{v}_{i/i+1}$ in (4.1).

$$\begin{aligned}\lambda_k(i) &= E\{y_k(i)b_k(i)\} = E\left\{\mathbf{w}_{a,k}(i)^{\mathrm{H}}\,\mathbf{r}_{a,k,i/i+1}b_k(i)\right\}\\ &= \mathbf{w}_{a,k}(i)^{\mathrm{H}}\,\mathbf{B}\,E\left\{\left(\mathbf{b}_{i/i+1}-\tilde{\mathbf{b}}_{k,i/i+1}\right)b_k(i)\right\}\\ &= \mathbf{B}_{K+k}{}^{\mathrm{H}}\mathbf{w}_{a,k}(i)\end{aligned} \tag{4.17}$$

$$\begin{aligned}\rho_k(i)^2 &= E\left\{\left|y_k(i)\right|^2\right\}-\lambda_k(i)^2\\ &= \mathbf{w}_{a,k}(i)^{\mathrm{H}}\,E\left\{\mathbf{r}_{a,k,i/i+1}\mathbf{r}_{a,k,i/i+1}{}^{\mathrm{H}}\right\}\mathbf{w}_{a,k}(i)-\lambda_k(i)^2\\ &= \lambda_k(i)-\lambda_k(i)^2\end{aligned} \tag{4.18}$$

Therefore the extrinsic information $L_{\text{det}}^{\text{ext}}\left(b_k(i)\right)$ is derived by the SISO detector as

$$L_{\text{det}}^{\text{ext}}\left(b_k(i)\right)=\left\{-\frac{\left|y_k(i)-\lambda_k(i)\right|^2}{2\rho_k(i)^2}\right\}+\left\{-\frac{\left|y_k(i)+\lambda_k(i)\right|^2}{2\rho_k(i)^2}\right\}=\frac{2\Re(\lambda_k(i)\,y_k(i))}{\rho_k(i)^2} \tag{4.19}$$

The complexity of the proposed algorithm is more than that of [WAN99] by a factor of at least 2^3. The computational burden stems mainly from the inverse of a matrix in (4.15). To reduce the computational complexity one might consider the following strategies [AHM05c]:

1. Reformulate (4.6) in such away the problem solving will amount to a $4N\times 4N$ real valued matrix inversion as opposed to complex valued matrix of the same dimension, and/or
2. Use a steepest descent algorithm to solve $\mathbf{\Xi}_{a,k}(i)\mathbf{w}_{a,k}(i)=\mathbf{B}_{K+k}$, as far as the number of iteration is low enough, or
3. Design a low complexity SISO detector using different optimization criteria based on the design of a filter whose coefficients are thought to be a solution of a real valued constrained problem.

- **Complexity reduction by redefining (4.6)**

Equation (4.6) will lead to solve a problem involving inverse of complex valued matrices (c.f. 4.15). This problem can be overcome by reformulating (4.6) in such a way an inverse of, only, a real valued matrix will be required.

First notice that $\Re\left(\mathbf{w}^{\mathrm{H}}(i)\mathbf{r}_{k,i/i+1}\right)$ can, instead, be written as

$$\begin{aligned}\Re\left(\mathbf{w}^{\mathrm{H}}(i)\mathbf{r}_{k,i/i+1}\right) &= \begin{bmatrix}\Re\left(\mathbf{w}(i)\right)\\ \Im\left(\mathbf{w}(i)\right)\end{bmatrix}^{\mathrm{T}}\begin{bmatrix}\Re\left(\mathbf{r}_{k,i/i+1}\right)\\ \Im\left(\mathbf{r}_{k,i/i+1}\right)\end{bmatrix}\\ &= \mathbf{w}_a^{\mathrm{T}}(i)\mathbf{r}_{a,k,i/i+1}\end{aligned} \tag{4.20}$$

where in this case, $\mathbf{w}_a(i)=\begin{bmatrix}\Re(\mathbf{w}(i))\\ \Im(\mathbf{w}(i))\end{bmatrix}$ and $\mathbf{r}_{a,k,i/i+1}=\begin{bmatrix}\Re\left(\mathbf{r}_{k,i/i+1}\right)\\ \Im\left(\mathbf{r}_{k,i/i+1}\right)\end{bmatrix}$ are two $4N\times 1$ real valued column vectors, hence (4.12) reduces to

$$\begin{aligned}\mathbf{w}_{a,k}(i) &= \arg\min_{\mathbf{w}_a\in\mathbb{R}^{4N\times 1}} E\left\{\left(\mathbf{w}_a^{\mathrm{T}}\mathbf{r}_{a,k,i/i+1}-b_k(i)\right)^2\right\}\\ &= \arg\min_{\mathbf{w}_a\in\mathbb{R}^{4N\times 1}} \mathbf{w}_a^{\mathrm{T}}E\left\{\mathbf{r}_{a,k,i/i+1}\mathbf{r}_{a,k,i/i+1}{}^{\mathrm{T}}\right\}\mathbf{w}_a-\mathbf{w}_a^{\mathrm{T}}E\left\{b_k(i)\mathbf{r}_{a,k,i/i+1}\right\}-E\left\{b_k(i)\mathbf{r}_{a,k,i/i+1}\right\}^{\mathrm{T}}\mathbf{w}_a\end{aligned} \tag{4.21}$$

with

$$E\left\{\mathbf{r}_{a,k,i/i+1}\mathbf{r}_{a,k,i/i+1}{}^{\mathrm{T}}\right\}=\begin{bmatrix}E\left\{\Re\left(\mathbf{r}_{k,i/i+1}\right)\Re\left(\mathbf{r}_{k,i/i+1}\right)^{\mathrm{T}}\right\} & E\left\{\Re\left(\mathbf{r}_{k,i/i+1}\right)\Im\left(\mathbf{r}_{k,i/i+1}\right)^{\mathrm{T}}\right\}\\ E\left\{\Im\left(\mathbf{r}_{k,i/i+1}\right)\Re\left(\mathbf{r}_{k,i/i+1}\right)^{\mathrm{T}}\right\} & E\left\{\Im\left(\mathbf{r}_{k,i/i+1}\right)\Im\left(\mathbf{r}_{k,i/i+1}\right)^{\mathrm{T}}\right\}\end{bmatrix}=\Xi_{a,k}(i) \tag{4.22}$$

The term $E\left\{\Re\left(\mathbf{r}_{k,i/i+1}\right)\Re\left(\mathbf{r}_{k,i/i+1}\right)^{\mathrm{T}}\right\}$ can be expended as

$$\begin{aligned}E\left\{\Re\left(\mathbf{r}_{k,i/i+1}\right)\Re\left(\mathbf{r}_{k,i/i+1}\right)^{\mathrm{T}}\right\} &= E\left\{\frac{1}{2}\left(\mathbf{r}_{k,i/i+1}+\mathbf{r}_{k,i/i+1}^{*}\right)\cdot\frac{1}{2}\left(\mathbf{r}_{k,i/i+1}^{\mathrm{T}}+\mathbf{r}_{k,i/i+1}^{\mathrm{H}}\right)\right\}\\ &= \frac{1}{4}\left\{E\left\{\mathbf{r}_{k,i/i+1}\mathbf{r}_{k,i/i+1}^{\mathrm{T}}\right\}+E\left\{\mathbf{r}_{k,i/i+1}\mathbf{r}_{k,i/i+1}^{\mathrm{H}}\right\}+E\left\{\mathbf{r}_{k,i/i+1}^{*}\mathbf{r}_{k,i/i+1}^{\mathrm{T}}\right\}+E\left\{\mathbf{r}_{k,i/i+1}^{*}\mathbf{r}_{k,i/i+1}^{\mathrm{H}}\right\}\right\}\end{aligned} \tag{4.23}$$

where

$$\begin{aligned} E\left\{\mathbf{r}_{k,i/i+1}\mathbf{r}_{k,i/i+1}^{\mathrm{T}}\right\} &= \mathbf{A}\ \mathbf{\Lambda}_k(i)\mathbf{A}^{\mathrm{T}} = \mathbf{X}_{k,i} \\ E\left\{\mathbf{r}_{k,i/i+1}\mathbf{r}_{k,i/i+1}^{\mathrm{H}}\right\} &= \mathbf{A}\ \mathbf{\Lambda}_k(i)\mathbf{A}^{\mathrm{H}} + \sigma^2\mathbf{I}_{2N\times 2N} = \mathbf{Y}_{k,i} \\ E\left\{\mathbf{r}_{k,i/i+1}^{*}\mathbf{r}_{k,i/i+1}^{\mathrm{T}}\right\} &= \mathbf{A}^{*}\mathbf{\Lambda}_k(i)\mathbf{A}^{\mathrm{T}} + \sigma^2\mathbf{I}_{2N\times 2N} = \mathbf{Y}_{k,i}^{*} \\ E\left\{\mathbf{r}_{k,i/i+1}^{*}\mathbf{r}_{k,i/i+1}^{\mathrm{H}}\right\} &= \mathbf{A}^{*}\mathbf{\Lambda}_k(i)\mathbf{A}^{\mathrm{H}} = \mathbf{X}_{k,i}^{*} \end{aligned} \tag{4.24}$$

The same arrangements can be done for the remaining terms, $E\left\{\Re\left(\mathbf{r}_{k,i/i+1}\right)\Im\left(\mathbf{r}_{k,i/i+1}\right)^{\mathrm{T}}\right\}$, $E\left\{\Im\left(\mathbf{r}_{k,i/i+1}\right)\Re\left(\mathbf{r}_{k,i/i+1}\right)^{\mathrm{T}}\right\}$ and $E\left\{\Im\left(\mathbf{r}_{k,i/i+1}\right)\Im\left(\mathbf{r}_{k,i/i+1}\right)^{\mathrm{T}}\right\}$, using the fact that $\Im\left(\mathbf{r}_{k,i/i+1}\right) = \frac{1}{2}\left(\mathbf{r}_{k,i/i+1} - \mathbf{r}_{k,i/i+1}^{*}\right)$ so that

$$\begin{aligned} E\left\{\Re\left(\mathbf{r}_{k,i/i+1}\right)\Re\left(\mathbf{r}_{k,i/i+1}\right)^{\mathrm{T}}\right\} &= \frac{1}{4}\left\{\mathbf{X}_{k,i} + \mathbf{Y}_{k,i} + \mathbf{Y}_{k,i}^{*} + \mathbf{X}_{k,i}^{*}\right\} \\ E\left\{\Re\left(\mathbf{r}_{k,i/i+1}\right)\Im\left(\mathbf{r}_{k,i/i+1}\right)^{\mathrm{T}}\right\} &= \frac{1}{4}\left\{\mathbf{X}_{k,i} - \mathbf{Y}_{k,i} + \mathbf{Y}_{k,i}^{*} - \mathbf{X}_{k,i}^{*}\right\} \\ E\left\{\Im\left(\mathbf{r}_{k,i/i+1}\right)\Re\left(\mathbf{r}_{k,i/i+1}\right)^{\mathrm{T}}\right\} &= \frac{1}{4}\left\{\mathbf{X}_{k,i} + \mathbf{Y}_{k,i} - \mathbf{Y}_{k,i}^{*} - \mathbf{X}_{k,i}^{*}\right\} \\ E\left\{\Im\left(\mathbf{r}_{k,i/i+1}\right)\Im\left(\mathbf{r}_{k,i/i+1}\right)^{\mathrm{T}}\right\} &= \frac{1}{4}\left\{\mathbf{X}_{k,i} - \mathbf{Y}_{k,i} - \mathbf{Y}_{k,i}^{*} + \mathbf{X}_{k,i}^{*}\right\} \end{aligned} \tag{4.25}$$

Hence, $\Xi_{a,k}(i)$ will be a $4N\times 4N$ real valued matrix. Notice that $\Xi_{a,k}(i)$ is symmetrical, therefore, it admits real valued Eigen values. Notice that using (4.22) will substantially reduce the complexity burden by a factor of 2 at no performance cost as (4.6) and (4.20) are equivalent representations.

Using either (4.6) or equivalently (4.20), Equation (4.15) can be viewed as solving the following linear equation $\Xi_{a,k}(i)\mathbf{w}_{a,k}(i) = \mathbf{B}_{K+k}$. Steepest descent algorithm, among other techniques, can be used.

- **Complexity reduction using a steepest descent algorithm**

The complexity burden of the proposed algorithm is obvious from (4.15), where the inverse of a $4N \times 4N$ square matrix $\boldsymbol{\Xi}_{a,k}(i)$ at each instant i (data sample) is performed with a complexity of $O\left((4N)^3\right)$[12]. We can reduce the computational complexity to $O\left(P_{iteration}(4N)^2\right)$ using the following steepest descent algorithm:

$$\mathbf{w}_{a,k}^{(p)}(i) = \left(\mathbf{I}_{4N\times 4N} - \mu\boldsymbol{\Xi}_{a,k}(i)\right)\mathbf{w}_{a,k}^{(p-1)}(i) + \mu\mathbf{B}_{K+k}, \quad p = 1,2,\ldots,P_{iteration} \tag{4.26}$$

The complexity saving is obvious if at least $P_{iteration} < 4N$. The simulation results in practically all cases show that the steepest descent solution (4.26) is as good as the direct matrix inverse (4.15) for $P_{iteration}$ less than 4 using a proper choice of adaptation step μ such that $\mu < \frac{2}{\lambda_{\max}}$, where $\lambda_{\max}$ is the maximum eigen value of $\boldsymbol{\Xi}_{a,k}(i)$.

The convergence aspect of (4.26) is addressed as follows; let $\mathbf{w}_{a,k}(i)$ be the steady state optimal solution (4.15) and $\mathbf{e}_k^{(p)}(i) = \mathbf{w}_{a,k}^{(p)}(i) - \mathbf{w}_{a,k}(i)$ the error at iteration p. Furthermore assume that $\mathbf{e}_k^{(p)}(i) = \sum_{n=1}^{4N} g_n^{(p)}\mathbf{V}_n(i)$, where $\mathbf{V}_n(i)$ is the n-th column of the matrix $\mathbf{V}(i)$ such that $\boldsymbol{\Xi}_{a,k}(i) = \mathbf{V}(i)\boldsymbol{\Pi}(i)\mathbf{V}^{H}(i)$ is the eigen decomposition of $\boldsymbol{\Xi}_{a,k}(i)$[13]. With $g_n^{(p)} \in \mathbb{R}$.

---

[12] Notice that this complexity can further be reduced if we assume that the channel is constant over the processed data block so that $\bar{\boldsymbol{\Lambda}}_k = \frac{1}{M}\sum_{i=1}^{M}\boldsymbol{\Lambda}_k(i)$ is used instead of $\boldsymbol{\Lambda}_k(i)$ and hence one matrix inversion is required per data block of M information bits (ergodicity assumption).

[13] Notice that $\boldsymbol{\Xi}_{a,k}(i)$ is a symitrical matrix which admits real valued eigen values.

It can be shown that if, λ_n, $n = 1,2,\ldots,4N$ such that $0 < \lambda_1 < \lambda_2 < \cdots < \lambda_{4N}$ are the real valued eigen values of $\Xi_{a,k}(i)$ then

$$\lim_{p\to\infty} \mathbf{w}_{a,k}^{(p)}(i) = \mathbf{w}_{a,k}(i) \tag{4.27}$$

if and only if

$$\mu < \frac{2}{\lambda_{4N}} \tag{4.28}$$

Proof

First notice that

$$\begin{aligned} \mathbf{e}_k^{(p+1)}(i) &= \mathbf{w}_{a,k}^{(p+1)}(i) - \mathbf{w}_{a,k}(i) = \mathbf{w}_{a,k}^{(p)}(i) + \mu\left(\mathbf{B}_{K+k} - \Xi_{a,k}(i)\mathbf{w}_{a,k}^{(p)}(i)\right) - \mathbf{w}_{a,k}(i) \\ &= \mathbf{e}_k^{(p)}(i) + \mu\left(\mathbf{B}_{K+k} - \Xi_{a,k}(i)\mathbf{w}_{a,k}^{(p)}(i)\right) \end{aligned} \tag{4.29}$$

Now let $\mathbf{e}_k^{(p)}(i) = \sum_{n=1}^{4N} g_n^{(p)}\mathbf{V}_n(i) = \mathbf{V}(i)\mathbf{g}^{(p)}$, where $\mathbf{g}^{(p)} = \left[g_1^{(p)}, g_2^{(p)}, \cdots, g_{4N}^{(p)}\right]^{\mathrm{T}}$, hence (4.29) can be written as

$$\mathbf{V}(i)\mathbf{g}^{(p+1)} = \mathbf{V}(i)\mathbf{g}^{(p)} + \mu\left(\mathbf{B}_{K+k} - \Xi_{a,k}(i)\mathbf{w}_{a,k}^{(p)}(i)\right) \tag{4.30}$$

Using the fact that $\mathbf{B}_{K+k} = \Xi_{a,k}(i)\mathbf{w}_{a,k}(i)$, (4.30) will become

$$\begin{aligned} \mathbf{V}(i)\mathbf{g}^{(p+1)} &= \mathbf{V}(i)\mathbf{g}^{(p)} - \mu\,\Xi_{a,k}(i)\mathbf{e}^{(p)} \\ &= \mathbf{V}(i)\mathbf{g}^{(p)} - \mu\,\mathbf{V}(i)\mathbf{\Pi}(i)\mathbf{V}^{\mathrm{H}}(i)\mathbf{e}^{(p)} \\ &= \mathbf{V}(i)\mathbf{g}^{(p)} - \mu\,\mathbf{V}(i)\mathbf{\Pi}(i)\mathbf{V}^{\mathrm{H}}(i)\mathbf{V}(i)\mathbf{g}^{(p)} \\ &= \mathbf{V}(i)\mathbf{g}^{(p)} - \mu\,\mathbf{V}(i)\mathbf{\Pi}(i)\mathbf{g}^{(p)} \\ &= \mathbf{V}(i)\left(1 - \mu\,\mathbf{\Pi}(i)\right)\mathbf{g}^{(p)} \end{aligned} \tag{4.31}$$

or simply

$$\mathbf{g}^{(p+1)} = \left(1 - \mu\,\mathbf{\Pi}(i)\right)\mathbf{g}^{(p)} \tag{4.32}$$

Hence, if $\mu < \frac{2}{\lambda_{4N}}$, we can make sure that $\lim_{p\to\infty} \mathbf{w}_{a,k}^{(p)}(i) = \mathbf{w}_{a,k}(i)$ will be guaranteed.

Another important point to consider is the eigen spread of $\Xi_{a,k}(i)$ at each turbo iteration p_{turbo}. Toward this end, one can view $\Xi_{a,k}(i)$, using the exact $\Lambda_k(i)$, as the exact covariance matrix from which $\mathbf{w}_{a,k}(i)$ is correctly computed. Now that $\Lambda_k(i)$ estimated at every iteration $p_{turbo}, \Lambda_k^{p_{turbo}}(i)$, is to be used for the $(p_{turbo}+1)$th iteration, it can be seen that the problem now turns out to solving a different linear equation

$$\left(\Xi_{a,k}(i) + \Delta\Xi_{a,k}(i)\right)\left(\mathbf{w}_{a,k}(i) + \Delta\mathbf{w}_{a,k}(i)\right) = \mathbf{B}_{K+k} \tag{4.33}$$

We have quantified the departure from the exact $\Xi_{a,k}(i)$ by a relative error of $\Delta\Xi_{a,k}(i)$ and a relative error $\Delta\mathbf{w}_{a,k}(i)$ in the estimate of $\mathbf{w}_{a,k}(i)$. This error is bouneded by, suppose $\Delta\Xi_{a,k}(i) = \varepsilon\mathbf{E}_{a,k}(i)$, where $\mathbf{E}_{a,k}(i)$ is an arbitrary matrix,

$$\frac{\left\|\Delta\mathbf{w}_{a,k}(i)\right\|}{\left\|\mathbf{w}_{a,k}(i)\right\|} \le \kappa \frac{\left\|\Delta\Xi_{a,k}(i)\right\|}{\left\|\Xi_{a,k}(i)\right\|} + O\left(\varepsilon^2\right) \tag{4.34}$$

where κ is the condition number of $\Xi_{a,k}(i)$. ■

From (4.34), we can see that as long as the condition number of $\Xi_{a,k}(i)$ is low, the relative error in $\mathbf{w}_{a,k}(i)$, denoted by $\Delta\mathbf{w}_{a,k}(i)$, is bounded by a reasonably small multiple of the relative error in $\Xi_{a,k}(i)$. In our simulation example shown in Figure 4.2, the condition number of $\Xi_{a,k}(i)$ over the iterations changes from 6 (5.8) to almost 4.2 (2.4) at the second iteration and steadily remains constant at 3.8 (2.01) for the rest of the iterations for E_b/N_0 of 8dB (0dB).

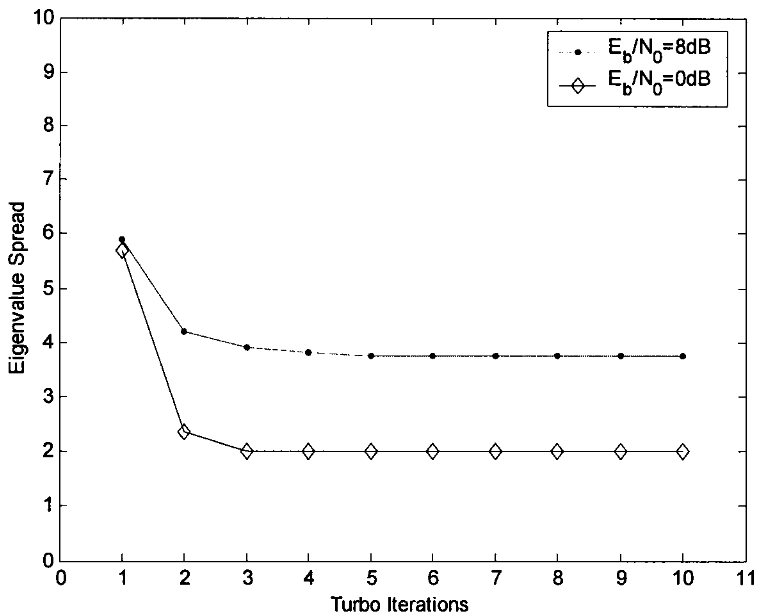


Figure 4.2. Eigen spread evolution of $\Xi_{a,k}(i)$ over 10 turbo iterations. The settings correspond to the simulation scenario 1 in section 4.4

- **Complexity reduction from a real valued constrained optimization**

As noted above, the SISO detector delivers soft information in terms of LLR's for every user k. At any instant 'i' an instantaneous filter $\mathbf{w}_k(i)$ is designed to operate over $\mathbf{r}_{i/i+1}$ in (4.1) as

$$y_k(i) = \mathbf{w}_k(i)^{\mathrm{H}} \mathbf{r}_{i/i+1} \tag{4.35}$$

under the constraints

$$\begin{cases} \mathbf{w}_k(i)^{\mathrm{H}} \mathbf{A}_{K+k} = 1 \\ \mathbf{w}_k(i)^{\mathrm{H}} \tilde{\mathbf{I}}_k(i) = 0 \end{cases} \tag{4.36}$$

where $\mathbf{A}_{K+k}$ is the $(K+k)$th column of $\mathbf{A}$ and

$$\tilde{\mathbf{I}}_k(i) = E\{b_k(i-1)\}\mathbf{A}_k + E\{b_k(i+1)\}\mathbf{A}_{2K+k} + \sum_{j\neq k}\left(E\{b_j(i-1)\}\mathbf{A}_j + E\{b_j(i)\}\mathbf{A}_{K+j} + E\{b_j(i+1)\}\mathbf{A}_{2K+j}\right)$$

denotes the Soft Interference Subspace (SIS).

Note that unlike [AFF02], the structure (4.36) takes into account a soft information sequence from the previous SISO decoding iteration. A suboptimal solution is derived by assuming that after few iterations (usually no more than four), $E\{b_k(i)\} \cong b_k(i)$, so that the soft interference subspace is the "same" as the exact interference subspace $\mathbf{I}_k(i)$, where

$$\mathbf{I}_k(i) = b_k(i-1)\mathbf{A}_k + b_k(i+1)\mathbf{A}_{2K+k} + \sum_{j\neq k}\left(b_j(i-1)\mathbf{A}_j + b_j(i)\mathbf{A}_{K+j} + b_j(i+1)\mathbf{A}_{2K+j}\right) \tag{4.37}$$

The solution to the constrained problem is

$$\mathbf{w}_k(i) = \frac{\tilde{\mathbf{P}}_k(i)\mathbf{A}_{K+k}}{\mathbf{A}_{K+k}{}^{\mathrm{H}}\tilde{\mathbf{P}}_k(i)\mathbf{A}_{K+k}} \tag{4.38}$$

where

$$\tilde{\mathbf{P}}_k(i) = \mathbf{I}_{2N\times 2N} - \tilde{\mathbf{I}}_k(i)\left(\tilde{\mathbf{I}}_k(i)^{\mathrm{H}}\tilde{\mathbf{I}}_k(i)\right)^{-1}\tilde{\mathbf{I}}_k(i)^{\mathrm{H}} \tag{4.39}$$

Hence, the output of the Soft Interference Subspace Canceller (or Conventional Soft ISC) is computed using (4.35).

In what follows, we assume that the output of the Conventional Soft ISC $y_k(i)$ in (4.35) represents the output of an equivalent additive white Gaussian noise channel having $b_k(i)$ as its input symbol. This equivalent channel can be represented as

$$y_k(i) = \lambda_k(i)b_k(i) + z_k(i) \tag{4.40}$$

where $\lambda_k(i)$ is the equivalent amplitude at instant i for the kth user at the Conventional Soft ISC output, and $z_k(i) \sim N\left(0, \rho_k(i)^2\right)$ is a Gaussian noise sample. Using (4.1) and (4.40), the parameters $\lambda_k(i)$ and $\rho_k(i)^2$ can be computed as follows, where the expectation is taken with respect to the code bits and $\mathbf{v}_{i/i+1}$

$$\begin{aligned}\lambda_k(i) &= E\{y_k(i)b_k(i)\} = E\left\{\left(\mathbf{w}_k(i)^{\mathrm{H}}\mathbf{r}_{i/i+1}\right)b_k(i)\right\} \\ &= E\left\{\mathbf{w}_k(i)^{\mathrm{H}}(\mathbf{A}\mathbf{b}_{i/i+1} + \mathbf{v}_{i/i+1})b_k(i)\right\} \\ &= \mathbf{w}_k(i)^{\mathrm{H}}\mathbf{A}E\{\mathbf{b}_{i/i+1}b_k(i)\} = \mathbf{w}_k(i)^{\mathrm{H}}\mathbf{A}\tilde{\mathbf{b}}^{(\mathrm{k})}_{i/i+1}\end{aligned} \quad (4.41)$$

where $\tilde{\mathbf{b}}^{(\mathrm{k})}_{i/i+1} = E\{\mathbf{b}_{i/i+1}\}\cdot E\{b_k(i)\} + \left(1 - E\{b_k(i)\}^2\right)\mathbf{e}_{K+k}$, and

$$\begin{aligned}\rho_k(i)^2 &= E\left\{\left|y_k(i)\right|^2\right\} - \lambda_k(i)^2 = \mathbf{w}_k(i)^{\mathrm{H}}E\left\{\mathbf{r}_{i/i+1}\mathbf{r}_{i/i+1}{}^{\mathrm{H}}\right\}\mathbf{w}_k(i) - \lambda_k(i)^2 \\ &= \mathbf{w}_k(i)^{\mathrm{H}}(\mathbf{A}\mathbf{\Lambda}_{i/i+1}\mathbf{A}^{\mathrm{H}} + \sigma^2\mathbf{I}_{2N\times 2N})\mathbf{w}_k(i) - \lambda_k(i)^2\end{aligned} \quad (4.42)$$

where

$$\begin{aligned}\mathbf{\Lambda}_{i/i+1} &= E\{\mathbf{b}_{i/i+1}\}\cdot E\{\mathbf{b}_{i/i+1}\}^{\mathrm{T}} + \sum_j\left(1 - E\{\mathbf{b}_{i/i+1}(j)\}^2\right)\mathbf{e}_j\mathbf{e}_j{}^{\mathrm{T}} + \sum_j\left(1 - E\{\mathbf{b}_{i/i+1}(K+j)\}^2\right)\mathbf{e}_{K+j}\mathbf{e}_{K+j}{}^{\mathrm{T}} \\ &\quad + \sum_j\left(1 - E\{\mathbf{b}_{i/i+1}(2K+j)\}^2\right)\mathbf{e}_{2K+j}\mathbf{e}_{2K+j}{}^{\mathrm{T}}\end{aligned}.$$

Therefore the extrinsic information $L^{\mathrm{ext}}_{\mathrm{det}}(b_k(i))$, delivered by the Conventional Soft ISC, is

$$L^{\mathrm{ext}}_{\mathrm{det}}(b_k(i)) = \frac{2\Re\left(\lambda_k(i)y_k(i)\right)}{\rho_k(i)^2} \quad (4.43)$$

The design problem formulated in (4.35) and (4.36) can be modified into the design of a filter $\mathbf{w}_k(i)$ so that

$$y_k(i) = \Re\left(\mathbf{w}_k(i)^{\mathrm{H}} \mathbf{r}_{i/i+1}\right) \tag{4.44}$$

subject to

$$\begin{cases} \Re\left(\mathbf{w}_k(i)^{\mathrm{H}} \mathbf{A}_{K+k}\right) = 1 \\ \Re\left(\mathbf{w}_k(i)^{\mathrm{H}} \tilde{\mathbf{I}}_k(i)\right) = 0 \end{cases} \Leftrightarrow \begin{cases} \mathbf{w}_{a,k}(i)^{\mathrm{H}} \mathbf{A}_{K+k}^{\mathrm{a}} = 1 \\ \mathbf{w}_{a,k}(i)^{\mathrm{H}} \tilde{\mathbf{I}}_k^{\mathrm{a}}(i) = 0 \end{cases} \tag{4.45}$$

with $\mathbf{w}_{a,k}(i)^{\mathrm{H}} = \begin{bmatrix} \mathbf{w}_k(i) \\ \mathbf{w}_k^*(i) \end{bmatrix}$, $\mathbf{A}_{K+k}^{\mathrm{a}} = \begin{bmatrix} \mathbf{A}_{K+k} \\ \mathbf{A}_{K+k}^* \end{bmatrix}$, and $\tilde{\mathbf{I}}_k^{\mathrm{a}}(i) = \begin{bmatrix} \tilde{\mathbf{I}}_k(i) \\ \tilde{\mathbf{I}}_k^*(i) \end{bmatrix}$. The reason for such a constraint follows from the choice of (4.44), where we force the filter output to be real. (4.44) can be viewed as[14]

$$y_k(i) = \Re\left(\mathbf{w}_k(i)^{\mathrm{H}} \mathbf{r}_{i/i+1}\right) = \frac{1}{2} \mathbf{w}_k^{\mathrm{a}}(i)^{\mathrm{H}} \mathbf{r}_{i/i+1}^{a} \tag{4.46}$$

where $\mathbf{r}_{i/i+1}^{\mathrm{a}} = \begin{bmatrix} \mathbf{r}_{i/i+1} \\ \mathbf{r}_{i/i+1}^* \end{bmatrix}$.

If we consider the filter design in (4.46) (equivalent to (4.44)), the associated constraints (4.45) will follow. The structure (4.45)-(4.46) considers the real valued part of the filter output with modified real valued constraints.

Solving (4.45) and (4.46) yields the SISO Soft Conjugate Interference Subspace Canceller (Soft Conjugate ISC) below:

$$\mathbf{w}_k^{\mathrm{a}}(i) = \frac{\tilde{\mathbf{P}}_k^{\mathrm{a}}(i) \mathbf{A}_{K+k}^{\mathrm{a}}}{\mathbf{A}_{K+k}^{\mathrm{a}\,H} \tilde{\mathbf{P}}_k^{\mathrm{a}}(i) \mathbf{A}_{K+k}^{\mathrm{a}}} \tag{4.47}$$

[14] (4.20) can, instead, be used for a reduced complexity implementation.

where

$$\tilde{\mathbf{P}}_k^{\mathrm{a}}(i) = \mathbf{I}_{4N\times 4N} - \tilde{\mathbf{I}}_k^{\mathrm{a}}(i)\left(\tilde{\mathbf{I}}_k^{\mathrm{a}}(i)^{\mathrm{H}}\tilde{\mathbf{I}}_k^{\mathrm{a}}(i)\right)^{-1}\tilde{\mathbf{I}}_k^{\mathrm{a}}(i)^{\mathrm{H}} \tag{4.48}$$

In the same way, equations equivalent to (4.41), (4.42) and (4.43) can be easily derived.

- **Other structure (Appendix B)**

Equation (4.6) can be applied to other MMSE based structures using decision feedback [GAM00]. Appendix B briefly outlines the derivations [AHM05a].

4.3-2 SISO channel decoder

The input to the kth SISO channel decoder are the de-interleaved LLR's $L_{\mathrm{det}}^{\mathrm{ext}}\left(b_k(n)\right)$ (or $L_{\mathrm{det}}\left(b_k(n)\right)$) of the code bits of the kth user provided by SISO detector just described above. The SISO decoder produces an update of the LLR's of the code bits $b_k(n)$, as well as the LLR's of the information bits $d_k(m)$, $m = 1,2,\ldots$ based on the code trellis. The SISO decoder used here is a modified version of the BCJR algorithm [BAH74], an idea exploited in [WAN99] as well.

Without loss of generalities we consider a binary rate $R_k = R = 1/\delta, k = 1,2,\cdots,K$ convolutional encoder of constraint length v. The input to the encoder at time $n, n = 1,2,\cdots,D$, is $d(n)$ and the corresponding output is $\{b(n)\} = \{b^1(n), b^2(n),\cdots,b^\delta(n)\}$. If $S_n = s$ and $S_{n-1} = s'$ denote the states at time n and $n-1$ taking values s and s' respectively, we define the forward and backward recursions as follows

$$\alpha_n(s)=\sum_{s'}\alpha_{n-1}(s')P\left(\{b(n)\}_{s'\to s}\right),\quad n=1,2,\cdots,(D+v) \tag{4.49}$$

$$\beta_n(s)=\sum_{s'}\beta_{n+1}(s')P\left(\{b(n)\}_{s\to s'}\right),\quad n=(D+v-1),\cdots,1,0 \tag{4.50}$$

where $P\left(\{b(n)\}_{s\to s'}\right)=P\left(\{b(n)\}=\underline{b}(s',s)\right)$ and $\underline{b}(s',s)$ is the hypothesis output of the encoder as the state changes from s to s', or in simple words the branch sequences defined by the states s and s'.

It is worth to state that the input stream $\{d_k(m)\}$ of length D is followed by v zeroes to terminate the trellis so that the boundary conditions: $\alpha_0(s=0)=1$, $\alpha_0(s\neq 0)=0$; and $\beta_{D+v}(s=0)=1$, $\beta_{D+v}(s\neq 0)=0$ are forced.

Implementation of (4.49) and (4.50) is numerically unstable, since both $\alpha_n(s)$ and $\beta_n(s)$ exponentially drop to zero [WAN99]. In order to obtain a numerically stable algorithm we use the same idea developed in [WAN99]. Let $\tilde{\alpha}_n(s)$ denote the scaled version of $\alpha_n(s)$. Initially, $\alpha_1(s)$ is computed according to (4.49), and then we set $\hat{\alpha}_1(s)=\alpha_1(s)$ and $\tilde{\alpha}_1(s)=\zeta_1\hat{\alpha}_1(s)$ with $\zeta_1=\frac{1}{\sum_s\hat{\alpha}_1(s)}$, for each $n\geq 2$, we compute $\tilde{\alpha}_n(s)$ according to

$$\hat{\alpha}_n(s)=\sum_{s'}\alpha_{n-1}(s')P\left(\{b(n)\}_{s'\to s}\right) \tag{4.51}$$

$$\alpha_n(s)=\zeta_n\hat{\alpha}_n(s),\ \zeta_n=\frac{1}{\sum_s\hat{\alpha}_n(s)} \tag{4.52}$$

It can, by induction, be shown that the scaling factor is cancelling out and that each $\alpha_n(s)$ is effectively scaled by $\frac{1}{\sum_s \alpha_n(s)}$. Likewise, we overcome the numerical instability in computing $\beta_n(s)$ in (4.50).

Let $S_j^+, j=1,2,\cdots,\delta$ be the set of state pairs (s',s) such that the jth bit of the codeword $\underline{b}(s',s)=\left\{b^1_{s'\to s}(n), b^2_{s'\to s}(n), \cdots, b^{j-1}_{s'\to s}(n), 1, b^{j+1}_{s'\to s}(n), \cdots, b^{\delta}_{s'\to s}(n)\right\}$ is $+1$, similarly we define $S_j^-, j=1,2,\cdots,\delta$. Assuming that the code bits are statistically independent, the a posteriori LLR of the code bit $b^j(n)$ at the output of the SISO decoder is

$$\begin{aligned} L_{\text{dec}}\left(b^j(n)\right) &= \log \frac{p\left(b^j(n)=+1 \middle| \left\{L_{\text{det}}(b(n))\right\},\ n=1,2,\cdots,M=\delta(D+v)\right)}{p\left(b^j(n)=-1 \middle| \left\{L_{\text{det}}(b(n))\right\},\ n=1,2,\cdots,M=\delta(D+v)\right)} \\ &= \log \frac{\sum_{S_j^+} \tilde{\alpha}_{n-1}(s')\tilde{\beta}_n(s)\prod_{l\neq j}^{\delta} P\left(b^l_{s'\to s}(n)\right)\cdot P\left(b^j_{s'\to s}(n)=+1\right)}{\sum_{S_j^-} \tilde{\alpha}_{n-1}(s')\tilde{\beta}_n(s)\prod_{l\neq j}^{\delta} P\left(b^l_{s'\to s}(n)\right)\cdot P\left(b^j_{s'\to s}(n)=-1\right)} \end{aligned} \tag{4.53}$$

Equation (4.53) uses the a posteriori information produced by the SISO detector, from which we derive $P\left(b^l_{s'\to s}(n)\right)$ for $n=1,2,\cdots,D+v$ and $l=1,2,\cdots,\delta$ according to

$$P\left(b^l_{s'\to s}(n)=+1\right)=1-P\left(b^l_{s'\to s}(n)=-1\right)=\frac{e^{L_{\text{det}}(b(i))}}{e^{L_{\text{det}}(b(i))}+1},\quad \begin{cases} l=1,2,\cdots,\delta \\ n=1,2,\cdots,D+v \\ i=\delta(n-1)+l \end{cases} \tag{4.54}$$

It is at the last iteration that the LLR's of the information bits are computed using equation (4.53), where the summation is done over D^+ (the set of state pairs (s',s) such that the information bit $d(n)=+1$ at instant n causes the transition from s' to s) and D^- as bellow

$$L_{\text{dec}}(d(n)) = \log \frac{p\left(d(n)=+1 \middle| \left\{L_{\text{det}}(b(n))\right\},\ \ n=1,2,\cdots,M=\delta(D+v)\right)}{p\left(d(n)=-1 \middle| \left\{L_{\text{det}}(b(n))\right\},\ \ n=1,2,\cdots,M=\delta(D+v)\right)}$$

$$= \log \frac{\sum_{D^+} \tilde{\alpha}_{n-1}(s')\tilde{\beta}_n(s)\prod_{l=1}^{\delta} P\left(b_{s'\to s}^{l}(n)\right)}{\sum_{D^-} \tilde{\alpha}_{n-1}(s')\tilde{\beta}_n(s)\prod_{l=1}^{\delta} P\left(b_{s'\to s}^{l}(n)\right)} \tag{4.55}$$

Finally the information bits $d(n)$ are decoded according to

$$d(n) = \operatorname{sgn}\left(L_{\text{dec}}(d(n))\right) \tag{4.56}$$

4.4 Simulation Results

4.4-1 BER-SNR performance

To test the performance of the proposed soft IC-CMMSE[15] ($P_{iteration}=3$ and $\mu=0.1$) against the conventional soft IC-MMSE developed in [WAN99] and the Soft Conjugate ISC (4.47) against the conventional Soft ISC (4.38) as well as the Approximate ML [AHM04], simulations were conducted over different scenarios. Unless otherwise stated, we consider an overloaded system with processing gain 7 (N) (using gold sequences of length 7) and 9 (K) users asynchronously accessing the system with a multipath channel of 3 paths. The delays are uniformly distributed over $[0 \quad NT_c)$. The complex attenuations are

[15] The simulation results show the reduced complexity algorithm performance using steepest descent. Direct matrix inversion provides similar (if not identical) performances. To make the figures more readable, the performance curves using direct matrix inversion are omitted.

Gaussian distributed (Figure 4.3). All users have the same power. A ½-rate convolutional encoder of constraint length 4 and generation polynomial $(13, 15)_8$ is used. For brevity, we refer to this convolutional code as $CC(13,15)_8$. Blocks of 300 symbols are transmitted.

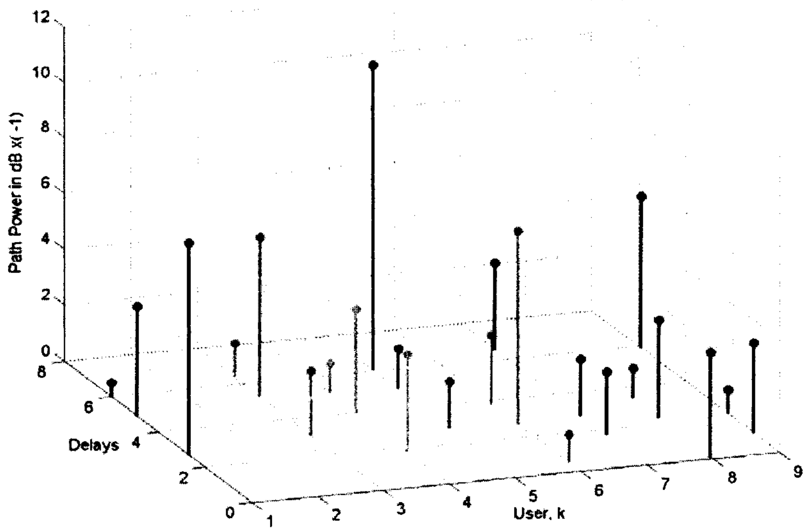


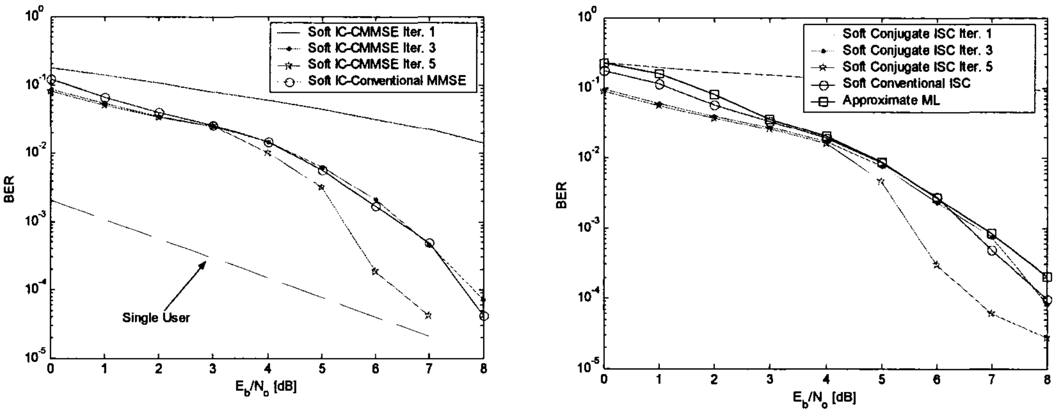
Figure 4.3. $10\log_{10}\left(\left\|w_{k,\ell}\right\|^2\right)$, $k=1,\cdots, K, \ell=1,\cdots,L_k$[16].

Scenario 1: Single rate system with a static channel (Figure 4.4)

In a simple configuration described above, Figure 4.4 depicts the relative performances of the proposed soft detectors. For reference, the single user performance is added.

Both the proposed soft IC-CMMSE and the conventional soft IC-MMSE perform equally well at low E_b/N_o less than 3dB for the single rate system. The soft IC-CMMSE, however, outperforms the conventional Soft IC-MMSE at high E_b/N_o, (Figure 4.4.a).

[16] Refer to the definition of $\mathbf{h}_k$ in Chapter 2.



a) b)

Figure 4.4. BER performance of the a) soft IC-CMMSE vs. Conventional soft IC-MMSE, b) Soft Conjugate ISC vs. conventional soft ISC and Approximate ML: N=7 and K=9

Meanwhile, the Conventional ISC and Approximate ML show the same performance at high and low E_b/N_0. We note a superior performance of the proposed Soft Conjugate ISC at high E_b/N_0, (Figure 4.4.b). Observe that, for a comparable performance, few iterations (3 iterations) are required for the soft IC-CMMSE receiver.

Scenario 2: Near far situation.

In the flowing scenario, users' powers are uniformly distributed over $\begin{bmatrix}0 & 6\end{bmatrix} dB$, as depicted in Figure 4.5, a situation that may raise for users signaling at unequal distances from a receiving base station or due to power control failures.

In the current setting, user 2 is the weakest user while user 5 is at least 6dB's stronger. Without a doubt, the soft IC-CMMSE outperforms the conventional soft IC-MMSE, on average, by 1.5dB (Figure 4.6.a). The weak and strong users' performances do show the same relative gains compared with those of their conventional counterparts (Figs 4.6.b, c).

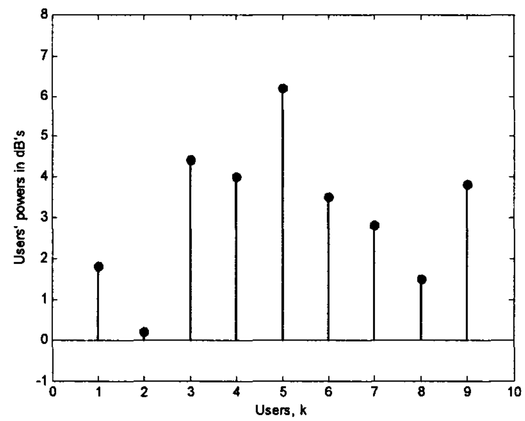


Figure 4.5. Users' Powers. User 2 is the weakest user, user 5 is the strongest user.

We observe, as in [ALE98], that the performance of the weak user is better than that of the strong user. We agree, based on the same arguments invoked in [ALE98] that low power users benefit more from the reliable soft information of the strong users.

As depicted in Figures 4.6.d, e, f, all three other methods have, on average, the same performances at high E_b/N_0, while differing from each other at low E_b/N_0. Note that the soft conjugate ISC slightly outperforms the conventional soft IC-MMSE (Figures 4.6.a, d).

Scenario 3: Time varying channel

In this configuration, the channel is time varying; we consider the pedestrian speed of 3 km/h and high speed of 100 km/h. Other parameters are left the same as in Scenario 1. The 3 path powers in dB's are 0, -3, -9 dB respectively for both speeds. The carrier frequency is 900 MHz and the chip rate is 1.246 Mcps (equivalent to a symbol duration of $T = 1.246\ 10^6 / \mathrm{N}$ sec). At mobile speeds as low as 3km/h, the proposed soft IC-CMMSE outperforms the conventional soft IC-MMSE by more than 2dB, and at mobile speeds as high as 100km/h the gain exceeds even 2.5dB.

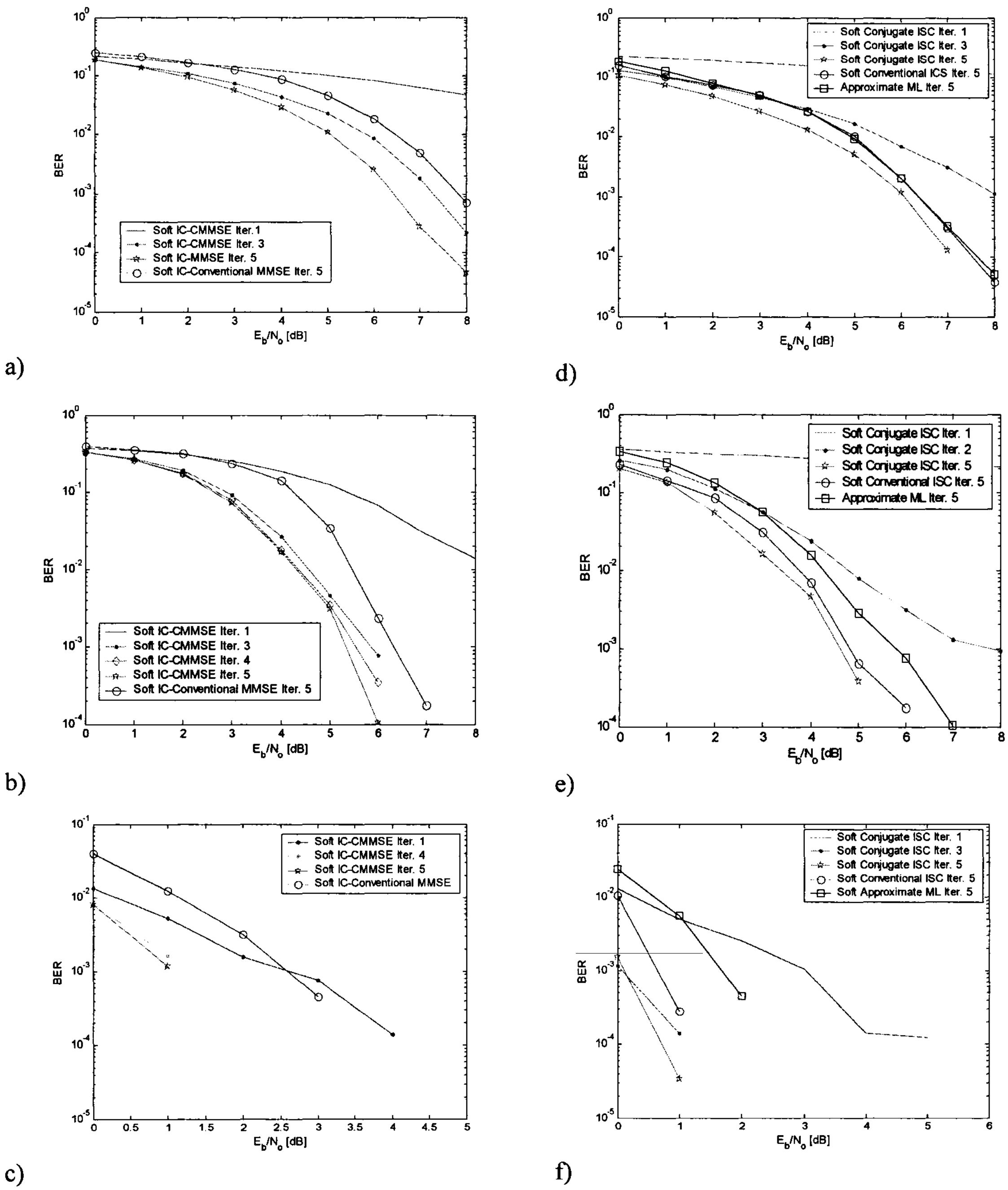


Figure 4.6. Near far situation: N=7, K=9 and MAI=6dB, a) Average performance over all users, b) Weak user performance, c) Strong user performance, for soft IC-CMMSE and conventional soft IC-MMSE, and d) Average performance over all users, e) Weak user performance, f) Strong user performance, for Soft Conjugate ISC, Conventional Soft ISC and Approximate ML.

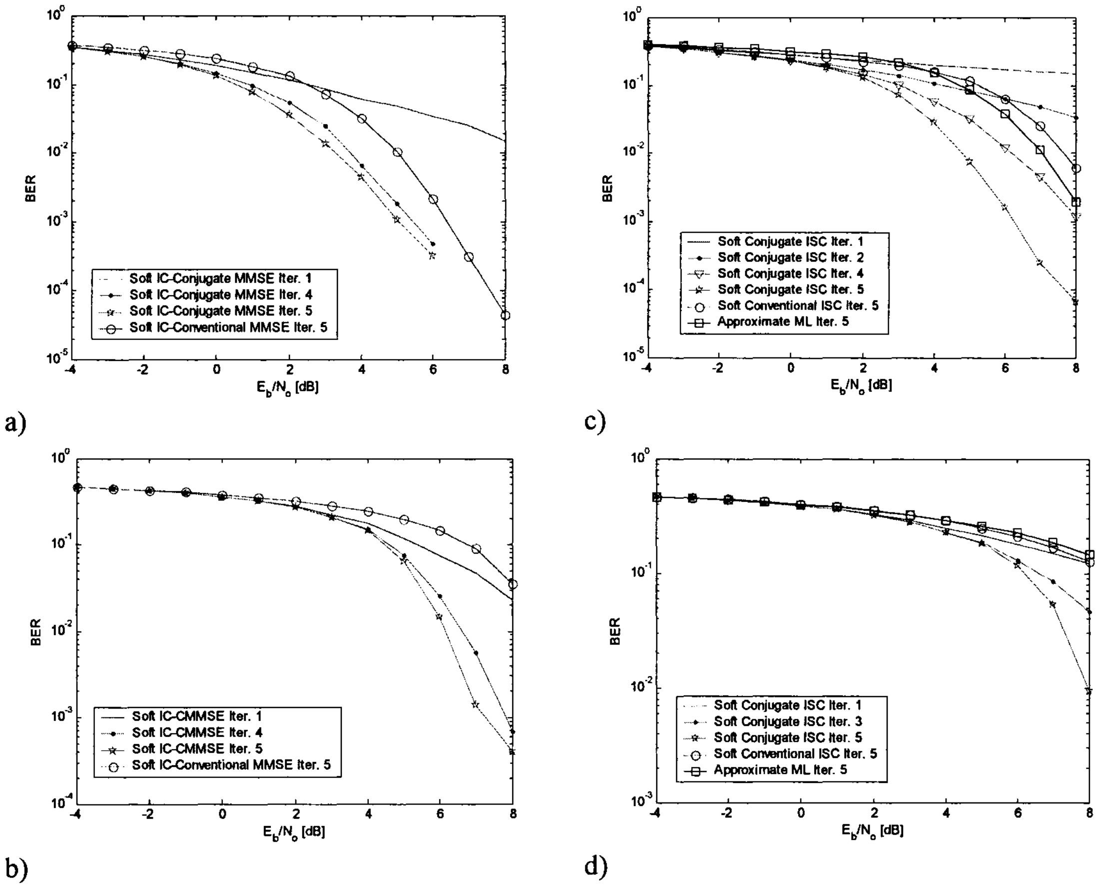


Figure 4.7. Time varying channel: N=7 and K=9, a) 3Km/h, b) 100Km/h for soft IC-CMMSE and conventional soft IC-MMSE, and c) 3Km/h, d) 100Km/h for Soft Conjugate ISC, Conventional Soft ISC and Approximate ML.

The proposed Soft Conjugate ISC outperforms the Conventional ISC as well as the Approximate ML at low and high mobile speeds. It demonstrates a good robustness to channel variations (at 100km/h) compared with the traditional Soft ISC, which has almost the same performance as the approximate ML. We can record gains of more than 2dB in slow time varying channels, and more than 4dB in fast time varying channels. Once again, notice that the soft conjugate ISC slightly outperforms the soft IC-MMSE.

Scenario 4: Multirate system

Current and next generation wireless systems adopt signaling configurations in which heterogeneous services, including voice and data, are provided. Therefore, it will be interesting to investigate the performances of the proposed methods in a multirate context. For this purpose, we consider two classes of users, High Rate Users Class (HRUC) and Low Rate Users Class (LRUC). The first class consists of 9 users with a spreading factor of 7 (178kb/s) from the gold sequence of length 7, and the second class consists of 16 users with a spreading factor of 14 (89kb/s) from the subset of the Gold sequence of length 15 with the 15th chip of each sequence being omitted.

As depicted in Figures 4.8.a and b, gains as much as 2dB for high rate users and 3dB for low rate users are recorded. Low rate users (analogous to weak users for a single rate system in a near far situation) perform better than high rate users (analogous to strong users in a single rate system.)

Note that the Conventional ISC is poor for both low and high rate users. The approximate ML appears to maintain a good performance. The proposed Soft Conjugate ISC shows superior performance for both rate classes, and gains of more than 1dB are observed. In addition, it shows performances equivalent to the conventional soft IC-MMSE in [WAN99].

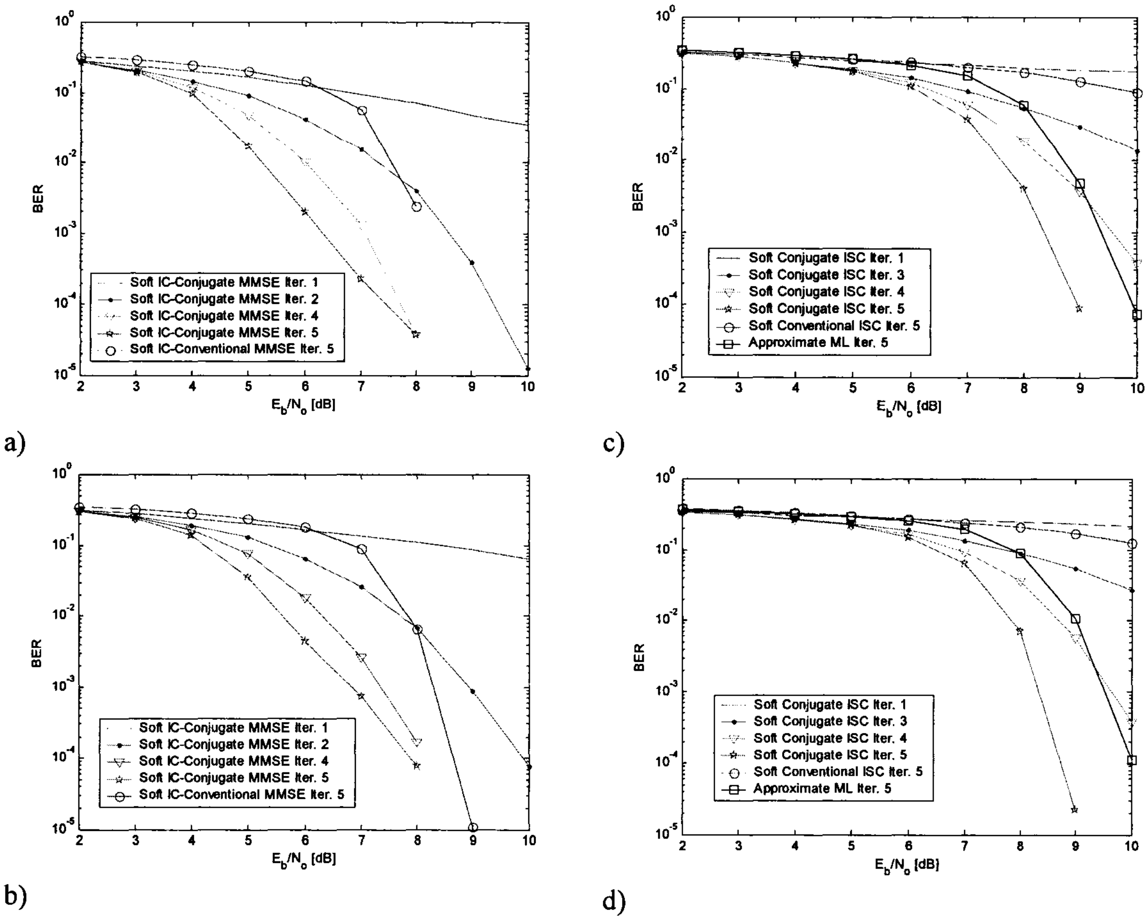


Figure 4.8. Multirate system, a) Low rate users, b) High rate users for soft IC-CMMSE and conventional soft IC-MMSE, and c) Low rate users, d) High rate users for Soft Conjugate ISC, Conventional Soft ISC and Approximate ML.

4.4-2 EXIT chart analysis for Soft IC-CMMSE

Analysis of convergence of the Turbo decoder using Gaussian approximation based on the signal-to-noise ration (SNR) is presented in [GAM01]. Instead of using SNR, [BRI99] and [BRI01] introduced the EXIT chart using mutual information as an indicator for the analysis of trellis code modulation and turbo codes. This same technique is applied to the analysis in [TUC02a]. Based on the EXIT charts of different iterative detection techniques, the exchange of extrinsic information between the SISO multiuser detectors and the SISO decoders is perceived as a decoding trajectory. In other words, the idea behind the EXIT chart is to summarize the description of the whole iterative (turbo) process to the exchange of one parameter, called mutual information (MI) defined between the information symbols (bits) and their Log-Likelihood Ratios (LLR) at the input and the output of the SISO iterative detector/decoder. The EXIT chart technique explained in details in [BRI02] is used to investigate the exchange of information between two devices, namely the Soft detector and the Soft decoder in the MI-two parameter plane. For this purpose, mutual information $I(L;x)$ is obtained assuming that the input and the output LLRs of each device are i.i.d. random variable defined by their conditional pdfs $p_{L|x}(\zeta|x=b)$ on the transmitted symbols (bits) $b=-1,+1$ [BRI01],

$$I(L;x)=\frac{1}{2}\sum_{b=-1,1}\int_{-\infty}^{\infty}p_{L|x}(\zeta|x=b)\log_2\frac{p_{L|x}(\zeta|x=b)}{p_{L|x}(\zeta|x=-1)+p_{L|x}(\zeta|x=1)}d\zeta \tag{4.57}$$

Since the LLR's pdf is not known, it has to be estimated from the data samples. One can use a simple expression introduced in [TUC02a]

$$I(L;x)\cong 1-\frac{1}{M}\sum_{m=1}^{M}\log_2\left(1+e^{-x(m)L(m)}\right) \tag{4.58}$$

where M is the number of bits transmitted during the simulation and $L(m)$ is the LLR associated with the bit $x(m)$. It is worth noticing that by using the mutual information as measurements, we do not require that the distribution of the output extrinsic information to be Gaussian, since the mutual information is a function of the entire pdf instead of the first and second order parameters of the extrinsic information (c.f. (4.57)), which is quite different from the SNR or variance based analysis.

Each device transforming LLRs is modeled by a transfer function mapping the MI defined for the input and the output. Let $I_i^{det} = I(L_i;x)$ and $I_o^{det} = I(L_o;x)$ represent the input and output MI of the soft detector associated with their respective input and output LLRs. A similar notation is adopted for the soft decoder using the upper script "dec", instead. The soft detector and decoder are, then, described by their EXIT function as $I_o^{det} = f_{MI}^{\det}(I_i^{det})$ and $I_o^{dec} = f_{MI}^{\mathrm{dec}}(I_i^{dec})$ with $I_i^{dec} = I_o^{\det}$.

The numerical procedure for obtaining the EXIT transfer function can be designed as follows [TUC02b]. First, randomly generate input bits $x(m) \in \{-1,1\}$ and their corresponding LLRs, $L(m)$, according to a Gaussian distribution with variance σ_L^2 and mean $x(m)\sigma_L^2/2$. Second, apply a soft device (detector or decoder) to the sequence $\{L(m)\}$. From the resulting output LLRs, calculate the output MI using (4.58), or (4.57), after estimating the pdf. The input MI can be computed using (4.57) as far as the generated input LLRs exibit a known Gaussian distribution. Repeating the above procedure for different values of σ_L^2, a set of couples (I_i^{det}, I_o^{det}) and (I_i^{dec}, I_o^{dec}) for both soft devices is constructed to constitutes the knots for MI transfer functions $f_{MI}^{det}(\bullet)$ and $f_{MI}^{dec}(\bullet)$ respectively. Note that to determine $f_{MI}^{det}(\bullet)$, a special setting regarding the system parameters has to be simulated. Such parameters' set includes the number of users, the multipath channel type (synchronous/asynchronous), the fading rate, the SNR and spreading sequences (and processing gains).

Next, we investigate some interesting cases to describe how EXIT charts can be used to analyze the soft multiuser detector performances including, convergence analysis, SNR effect and user capacity determination, and we may conclude by some ideas on how such chart can be used for system design. The coded-BER contours accompanying the EXIT charts are plotted according to the procedure in [BRI01, eq (26)-(31)].

Case 1: Convergence and SNR effect

EXIT chart of the soft IC-CMMSE and the soft IC-MMSE over asynchronous multipath channel is depicted in Figure 4.9. We consider a 12 users system signal with a periodic spreading code of processing gain 15 over a multipath fading channel. The EXIT chart in figure 4.9 depicts the MI transfer function as a function of SNR. It is apparent that the soft detector exhibits pinch-off (the shrinkage of the area between the soft detector and decoder MI curves) as SNR decreases. At an SNR of 2dB, if the vertices constituting the soft detection/decoding trajectory falls within the detector/decoder MI transfer curve, the convergence will lead to a coded-BER less than 0.1%. At low SNR (0dB) the tunnel (the area between the soft detector and decoder MI curves) is closing and the MI transfer curves of the detector and the decoder crosses before even reaching a coded-BER of 20%. Hence the EXIT chart can be used in case of soft IC-CMMSE to deduce the required SNR for a targeted coded-BER. In addition, the EXIT charts suggest that exploiting BPSK modulation improves performance.

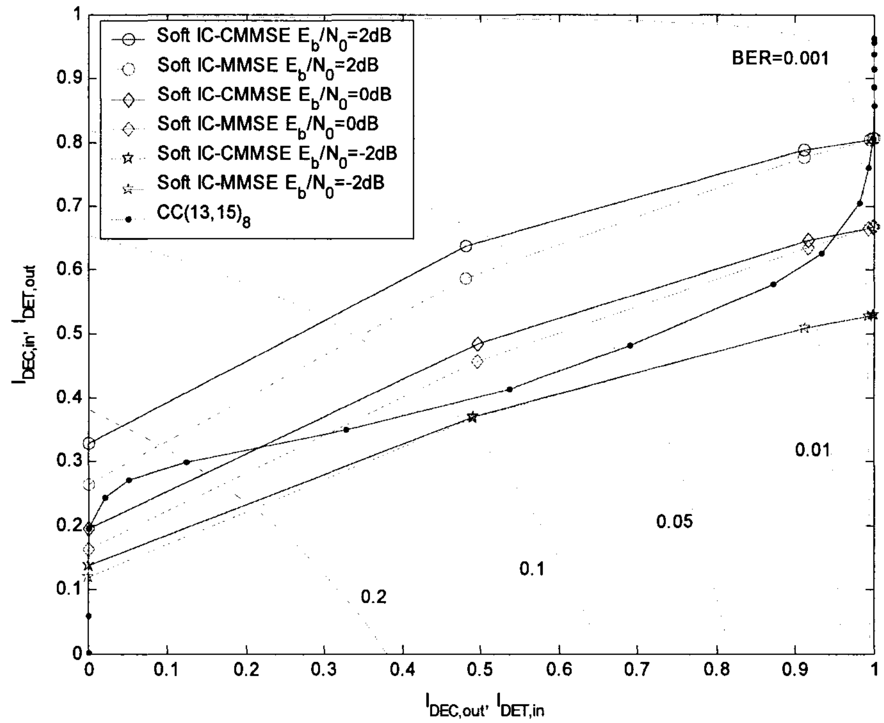


Figure 4.9. EXIT chart for Soft IC-Conjugate MMSE for different SNR values.

Case 2: User capacity determination

Traditionally, the total number of users that can be decoded given a certain SNR and a target coded-BER (usually in the order of 10^{-2} to 10^{-5}, depending on the system's requirements) is determined through lengthy BER versus SNR simulations. EXIT chart can be used to determine, given the system settings, the total number of users. Figure 4.10 below shows that the tunnel (the area between the soft detector and decoder MI curves gap) shrinks as the number of users increases. Along with the coded-BER contours, we can easily read that the system capacity is, for instance, less than 25 users for a reliable communication.

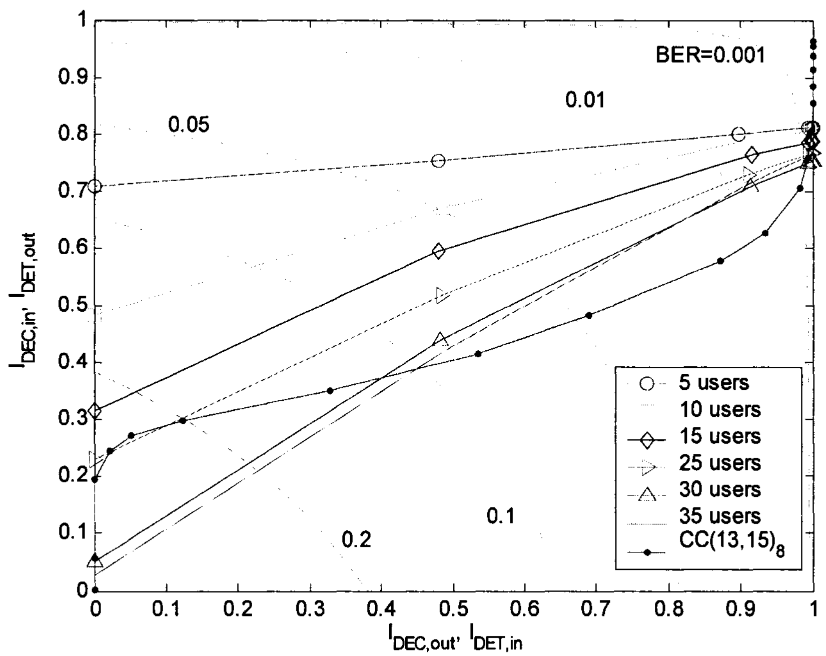


Figure 4.10. EXIT chart for Soft IC-Conjugate MMSE user capacity. SNR=4dB.

The above scenario can be used to determine channel coding and/or spreading strategies. The same idea of adaptive modulation/coding can be extended to systems utilizing iterative detection/decoding. The idea is simple, as the number of users grows, the soft detector MI transfer curve approaches from above the soft decoder MI transfer curve, hence reducing the convergence gap. Two measures can be devised; the first is, increasing the processing gain (adaptive spreading/rate) for the users suffering from deteriorated BER, due to unfavorable channel and MAI conditions, and/or devise a coding strategy (adaptive coding) that exhibits a MI transfer curve that moves downward toward the horizontal axes hence restoring the convergence gap. This strategy can be performed for fading channels at a rate analogous to the power control mechanism in most commercial CDMA systems.

4.5 Conclusion

To offer an interesting trade-off between performance and complexity, new Soft Input Soft Output iterative (turbo) multiuser detectors have been proposed for BPSK DS-CDMA systems [AHM05c]. When complexity increase by a factor of 2^3 is affordable, the soft IC-CMMSE offers, over all scenarios, average performance gains of 2dB. The second proposed scheme, namely the soft Conjugate ISC, offers - at a lower complexity - performances that are as good as those of the method developed by Wang and Poor. For the soft IC-CMMSE structure, note that, unlike the new MMSE family structure developed by Buzzi and Lops, the expectation (4.12) is applied with respect to both the additive white Gaussian noise and the information sequence and that, unlike the novel algorithm developed by Wang and Poor, the optimized cost function (4.12) uses the right hand side of (4.4) instead of the left hand side. Complexity reduction is achieved using (4.20) instead of (4.6) to define the real valued filter output. Further complexity reduction can be achieved using the steepest descent algorithm as long as the number of iterations does not exceed N. As for the second structure of the soft Conjugate ISC, the structure (4.45)-(4.46) considers the real valued part of the filter output with modified constraints coupled with the soft information sequence from the previous soft decoding iteration to construct, for a given user, the appropriate soft interference subspace.

Simulation results demonstrate that, over almost all scenarios, the proposed soft IC-CMMSE is the best candidate when the complexity issue is not addressed [AHM05c]. One may assess the complexity to be on the order of $O\left(2^3(2N)^3\right)$, while that of the conventional one is on the order of $O\left((2N)^3\right)$. A reduced complexity design would call for the use of

- The solution to (4.21), where a real valued matrix inversion will be required, and/or
- The steepest descent algorithm $O\left(2^2 P_{iteration} (2N)^2\right)$, as far as the number of iterations is low enough, or
- The soft Conjugate ISC $O\left(2^2 (2N)^2\right)$.

5.

CONCLUSIONS AND FUTURE WORKS

The success behind signal processing algorithms applied to communication systems witnesses the idea of exploiting any relevant information or structure inherent in the received signal. Channel estimation algorithms that use information regarding the time variations dynamics along with the multiuser approach proved to be more powerful compared to the traditional methods, yet it costs some extra signal processing to estimate the statistics that might be used to define these time variations and hence refine the channel estimates. The proposed multiuser-WLMS is an instant of a low complexity channel estimation and tracking algorithm that performs as good as the most popular method, Maximum Likelihood, and yet has maintained the computational complexity close to the LMS algorithm. The simulation results show the good performances of the proposed multiuser-WLMS technique in both WCDMA and cdma2000 environments.

Contribution I: Acquisition and Tracking of Time Varying Channel

In our work [AHM05b], a multiuser LMS-like structure along with smoothing and prediction filters to improve tracking quality is suggested. The choice for such adaptation family stems from its low computational complexity and its regular structure favourable for an efficient VLSI implementation where parallelism and wave pipelining, among other techniques, are easily applied [MOR99], [MOZ99], [SAK98] and [MAS98]. They are computationally effective due to the even distribution of the computation load over each symbol duration and no extra computation is required at the end of the processing window or preamble.

To summarize:

- The proposed multiuser LMS structure takes into account all users contributions simultaneously and delivers a composite channel impulse response, as in [BHA02], at each symbol (pilot symbol). The composite channel impulse response is defined to be at least an $(N+1)K$ column vector, whose content provides, simultaneously, information about the multipath delays and time varying attenuations, where K represents the number of users and N the pilot (in case of cdma2000 and WCDMA) spreading factor.

- A unique smoothing/prediction filter is designed based on a single p-order AR model over an averaged Doppler profile for all users.

- The adaptation step is dynamically searched at each iteration.

As in [BHA02],

- An even distribution of the computational burden over a training window or a preamble,

- A regular structure for an efficient VLSI implementation,

constitute the motives for our choice of a multiuser-LMS structure.

Iterative exchange of soft information (extrinsic LLRs) between the two constituent blocks of the receiver, namely the detector and the channel decoder, systematically improves the whole receiver performance. This idea along with redefining a new cost function which takes into account, as it may seem obvious, the idea of comparing real with real, for cases[17] where BPSK is the preferred DS-CDMA system's modulation technique, provides gains as high as 2dB and more. These same ideas gave birth to the new soft IC-CMMSE method which outperforms the traditional soft IC-MMSE. Further complexity reduction can be achieved by avoiding direct matrix inverse or by redefining a new cost function based on filter design with real constrains that will lead to matrix-inversion free type of algorithms.

Contribution II: Exploiting BPSK Modulation in Iterative (Turbo) Detection

Driven by implementation issues, we suggested in [AHM04] a low complexity iterative detection structure at a computational load close to I-Rake, where I is the number of iterations. Triggered by performance improvement while keeping the complexity low, we

[17] BPSK is the preferred modulation technique in current W-CDMA systems.

developed in [AHM05a] a new iterative structure based on the MMSE design of a feed-forward and feed-back filters that can preserve the potential gains even in multirate signals.

It is worth to mention that exploiting the BPSK modulation in designing linear receiver is not a new issue [TUL01], [BUZ01a], but the idea of taking this into account in design turbo detectors is not covered yet. One can notice that our contribution can be viewed as an extension of [WAN99] for BPSK signals. The contribution investigates, also, the additional potential gain that can be achieved when exploiting the modulation type [AHM05c], [AHM05d].

To summarize:

Two low complexity turbo detection receivers for joint detection and decoding for coded DS-CDMA systems utilizing BPSK modulation in multipath channels are presented.

- The first scheme is based on soft Interference Canceller (IC) followed by a MMSE filter whose coefficients are designed to be the solution of an MMSE-optimization problem based on a (forced) real valued filter output rather than a complex one as in conventional MMSE receivers used in [WAN99].

- A complexity reduction is achieved by redefining the filter output in such away a real valued matrix will, instead, be required as opposed to a complex valued one. Using steepest descent method, among others, to solve a large linear problem sounds attractive as well.

- In a second scheme, we reformulated the traditional constrained filter design by considering a real valued filter output along with real valued constrains.

- The work is extended in [AHM05a] to a more advanced receiver structure using a decision feedback scheme.

Further comments on the proposed algorithms' performances and computational complexity are provided respectively in sections 3.7 and 4.5. As we attempt to make this conclusion short, we prefer rather discuss some future perspectives.

Future works involve many issues; as for the multiuser-WLMS algorithm, the design strategy involves many iterations to come at end with quasi-optimal filters' coefficient design, this step needs extra work to systematically devise a simple and straightforward design strategy. Meanwhile, it will be interesting to analyze the multiuser-WLMS performance from the excess MSE and time constant perspectives, which answer the question; how do the algorithm parameters (number of users, filter taps, adaptation step, etc) affect performances? Such side information will help refine the design methodology.

Future works may, also, be extended to an analytic performance analysis of the proposed turbo detectors, as we have attempted using EXIT charts in Chapter 4, to explain the discrepancies among the different proposed iterative detection algorithms' performances. Finally, it is worth to mention that if the preferred modulation technique is no longer BPSK, QPSK for instance, all the methods may be equivalent.

An extension of the proposed algorithms to DS-CDMA system with multiple antennas [AHM05b] is still an open issue to explore, and answer to the questions like; are the performance gains maintained for multiple antenna systems? And how does the algorithms' complexity change accordingly?

Finally, it will be interesting to investigate the different turbo detectors performances in the presence of channel estimates [AHM05b]-[AHM05d], power control imperfections and inter-cell interference, as well as quantification effect, as we may attempt at first step, before any prompt VLSI implementation.

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APPENDIX A

A ON THE PERFORMANCE OF STEEPEST DESCENT ML VERSUS MULTIUSER S-WLMS CHANNEL ESTIMATORS

In this appendix, we outline the maximum likelihood (ML) estimate of the channel response of all the users [BHA02] ($\mathbf{z}$ considering a time invariant channel) using the knowledge of their spreading codes and transmitted bits. These known bits could be available either as a preamble before the data or as bits in a separate pilot channel. The joint conditional distribution of M received observation vectors $\mathbf{r}_1, \mathbf{r}_2, \ldots, \mathbf{r}_M$, given the knowledge of the spreading sequences, channel and the bits is given by

$$p\left(\mathbf{r}_1, \mathbf{r}_2, \ldots, \mathbf{r}_M / \mathbf{C}_1, \mathbf{C}_2, \ldots, \mathbf{C}_M, \mathbf{B}_1, \mathbf{B}_2, \ldots, \mathbf{B}_M, \mathbf{z}\right) = \frac{1}{\left(\pi\sigma^2\right)^{NM}} exp\left\{\frac{1}{\sigma^2}\sum_{i=1}^{M}\left(\mathbf{r}_i - \mathbf{C}_i\mathbf{B}_i\mathbf{z}\right)^{\mathrm{H}}\left(\mathbf{r}_i - \mathbf{C}_i\mathbf{B}_i\mathbf{z}\right)\right\}$$

the estimate $\hat{\mathbf{z}}_{ML}(M)$ that uniquely maximizes this likelihood function is the ML estimate and it satisfies the equation

$$\left\{\sum_{i=1}^{M}(\mathbf{C}_i\mathbf{B}_i)^{\mathrm{H}}(\mathbf{C}_i\mathbf{B}_i)\right\}\hat{\mathbf{z}}_{ML}(M) = \sum_{i=1}^{M}(\mathbf{C}_i\mathbf{B}_i)^{\mathrm{H}}\mathbf{r}_i \qquad \text{(A-1)}$$

For simplicity, we will denote $\frac{1}{M}\sum_{i=1}^{M}(\mathbf{C}_i\mathbf{B}_i)^{\mathrm{H}}(\mathbf{C}_i\mathbf{B}_i)$ by $\mathbf{R}_M$, a $(N+1)K\times(N+1)K$ matrix and $\frac{1}{M}\sum_{i=1}^{M}(\mathbf{C}_i\mathbf{B}_i)^{\mathrm{H}}\mathbf{r}_i$ by $\mathbf{y}_M$ a $(N+1)K\times 1$. The rank of $\mathbf{R}_M$ increases by N with each additional term $(\mathbf{C}_i\mathbf{B}_i)^{\mathrm{H}}(\mathbf{C}_i\mathbf{B}_i)$ in the summation. This is based on the assumption that random spreading codes are used and the spreading codes over this duration are linearly independent. Therefore, for $\mathbf{R}_M$ to be full rank, M should be at least equal to $K+\lceil K/N\rceil$. The current and next generation standards provide enough preamble or pilot resources to easily satisfy this condition. Therefore, assuming that $\mathbf{R}_M$ is full rank, we can write

$$\hat{\mathbf{z}}_{ML}(M)=\mathbf{R}_M^{-1}\mathbf{y}_M \tag{A-2}$$

Since $\mathbf{r}_i$ is jointly Gaussian random vector with mean $\mathbf{C}_i\mathbf{B}_i\mathbf{z}$ and covariance matrix $\sigma^2\mathbf{I}$, any linear transformation $\mathbf{Tr}_i$ of $\mathbf{r}_i$ is also jointly Gaussian random vector with mean $\mathbf{TC}_i\mathbf{B}_i\mathbf{z}$ and covariance matrix $\sigma^2\mathbf{TT}^{\mathrm{H}}$. Using this priority of Gaussian random vectors, it can be shown that $\hat{\mathbf{z}}_{ML}(M)$ is also jointly Gaussian with mean $\mathbf{z}$ and covariance matrix $\sigma^2\mathbf{R}_M^{-1}/M$ [BHA02].

For comparison purposes the single user channel estimate given by

$$\hat{\mathbf{z}}_{SU}=\frac{1}{NM}\sum_{i=1}^{M}(\mathbf{C}_i\mathbf{B}_i)^{\mathrm{H}}\mathbf{r}_i \tag{A-3}$$

A. Iterative channel estimation

In this section, the maximum likelihood estimate is approximated using iterative algorithms developed using gradient-based adaptation [BHA02]. Iterative algorithms based on the true gradient or an estimated stochastic gradient have been used earlier for various adaptive filtering and detection problems [HAY96], [HON95]. [BHA02] applied gradient-based adaptation techniques using the exact gradient in the multiuser channel estimation problem. A direct computation of the exact ML channel estimate involves the computation of the correlation matrix $\mathbf{R}_M$ and then the computation of $\mathbf{R}_M^{-1}\mathbf{y}_M$ at the end of the preamble. The direct computation of the inverse of the correlation matrix at the end of the preamble is computationally intense and could delay the channel estimation process beyond the preamble duration and limit the information rate.

B. Gradient descent method

The simplest gradient descent algorithm performs the following computations during the ith bit duration [BHA02].

1. Compute

$$\mathbf{R}_i = \frac{i-1}{i}\mathbf{R}_{i-1} + \frac{1}{i}\left(\mathbf{C}_i\mathbf{B}_i\right)^{\mathrm{H}}\left(\mathbf{C}_i\mathbf{B}_i\right) \tag{A-4}$$

2. Compute

$$\mathbf{y}_i = \frac{i-1}{i}\mathbf{y}_{i-1} + \frac{1}{i}\left(\mathbf{C}_i\mathbf{B}_i\right)^{\mathrm{H}}\mathbf{r}_i \tag{A-5}$$

3. Update the estimate $\hat{\mathbf{z}}$ via

$$\hat{\mathbf{z}}^{(\mathrm{i})} = \hat{\mathbf{z}}^{(\mathrm{i-1})} - \mu\left(\mathbf{R}_i\hat{\mathbf{z}}^{(\mathrm{i-1})} - \mathbf{y}_i\right) \tag{A-6}$$

where $\mathbf{R}_i\hat{\mathbf{z}}^{(\mathrm{i-1})} - \mathbf{y}_i$ is the gradient of the squared error surface (corresponding to the exponent in the likelihood function that needs to be minimized) and μ should be chosen to ensure convergence and to control speed of convergence. In each iteration, the estimate of the channel is updated by taking a step along the gradient vector.

In this algorithm, the ML estimate for a preamble of length i is approximated as soon as the ith bit is received. In fact, the updating step (step 3) can be repeated to improve accuracy. It can be repeated as many times as allowed by the available computational resources. In our report, we will assume that this updating is done only once per bit. Therefore, the number of iterations is equal to the preamble length.

C. Steepest descent method

In the simple gradient descent algorithm for channel estimation, the step size is chosen to be constant for all the iterations. To speed up convergence, the step size can be chosen optimally for each iteration to minimize the squared error achieved by the updating step (step 3 which updates the channel estimate along the direction opposite to the gradient). Using at each iteration i the step size given by [BHA02]

$$\mu^{(i)} = \frac{\mathbf{e}^{(i)H}\mathbf{e}^{(i)}}{\mathbf{e}^{(i)H}\mathbf{R}_i\mathbf{e}^{(i)}} \tag{A-7}$$

the optimal step size can be easily calculated using the knowledge of $\mathbf{R}_i$ and the gradient. Therefore, the steepest descent algorithm can be implemented with the same information needed for the constant step size algorithm. Further speed up in convergence can be achieved by choosing the search directions in addition to choosing the step size for each iteration. This can be done by the conjugate gradient algorithm [GOL96]. In the conjugate gradient algorithm, the search direction in any iteration is chosen to be orthogonal to the search directions used in the previous iterations. The steepest descent algorithm does not ensure this since it uses the gradient directly as the search direction. However, the implementation of the conjugate gradient algorithm would require significant additional computation to obtain the search directions.

D. Tracking time-varying channels

The iterative channel estimation algorithm scheme can be easily extended to track time variations in the channel after the preamble. The channel is assumed to be approximately constant over the preamble duration and the tracking is performed by sliding the estimation window and using data decisions instead of training sequences.

In the tracking scheme, the correlation matrix $\mathbf{R}_M$ and the matched filter outputs $\mathbf{y}_M$ are averaged over a sliding window of length M. The tracking is done as follows:

1. Detect bits using multishot multistage detection with previous channel estimate;
2. Compute new correlation matrix and matched filter vector: if $\mathbf{R}_M^{\text{old}}$ corresponds to the old window over the time indices $T+1, T+2, \ldots, T+M$ and D bits for user are detected using multistage detection, then

$$\mathbf{R}_M^{\text{new}} = \mathbf{R}_M^{\text{old}} + \sum_{i=T+M+1}^{T+M+D} (\mathbf{C}_i\mathbf{B}_i)^{\text{H}} (\mathbf{C}_i\mathbf{B}_i) - \sum_{i=T+1}^{T+D} (\mathbf{C}_i\mathbf{B}_i)^{\text{H}} (\mathbf{C}_i\mathbf{B}_i) \tag{A-8}$$

$$\mathbf{y}_M^{\text{new}} = \mathbf{y}_M^{\text{old}} + \sum_{i=T+M+1}^{T+M+D} (\mathbf{C}_i\mathbf{B}_i)^{\text{H}} \mathbf{r}_i - \sum_{i=T+1}^{T+D} (\mathbf{C}_i\mathbf{B}_i)^{\text{H}} \mathbf{r}_i \tag{A-9}$$

3. Update channel estimate

$$\hat{\mathbf{z}}^{\text{new}} = \hat{\mathbf{z}}^{\text{old}} - \mu\left(\mathbf{R}_i^{\text{new}}\hat{\mathbf{z}}^{\text{old}} - \mathbf{y}_i^{\text{new}}\right) \tag{A-10}$$

As discussed for the estimation scheme, the updating step can be repeated to improve estimation accuracy. Since the channel is assumed to be roughly constant over the window length, the ML channel estimate for the new window should be close to the previous ML estimate. Therefore, in practice, we notice that one iteration per bit is sufficient, i. e., the channel estimate from the previous window is a good initialization for both the simple gradient descent with constant step size and the steepest descent algorithm to estimate the new channel estimate.

We briefly describe some of the preliminary simulations that we conducted to evaluate the performance of the proposed estimators against the Steepest Decent Maximum Likelihood (SD-ML) and the single user (SU) estimators. A pilot processing gain of $N_p = 256$ was used. The delays of all the users were assumed uniformly distributed in $[1 \quad N_p)$ chips. The defaults values of the system parameters, unless otherwise varied along the x-axis: the number of observations, of $N_p \times 1$ dimension, is $M = 150$ per frame, the signal-to-noise ratio is $SNR = 8\,dB$ and the number of users is $K = 10$, signaling at a rate of $N_d = 16$ chip/bit, with a chip rate of 3.84Mchip/sec at a carrier frequency of 2GHz. The number of paths is $L_k = L = 2,\ k = 1,2,\cdots,K$, with relative powers of 0 and -3dB respectively.

Since a large number of the interdependent parameters are being estimated, it is not very revealing to determine or calculate the estimation error for each individual parameter.

It is rather more appealing to look at the *loss* in *dB* calculated as follows, for a given SNR and the system/channel parameters (K, N_p, N_d, L_k, V, etc...)

$$\mathrm{Loss}\left(SNR, K, N_p, N_d, L_k, V\right) = \frac{1}{M}\sum_{i=1}^{M} E\left[\left(\frac{\mathbf{z}_i}{\|\mathbf{z}_i\|}\right)^{\mathrm{H}}\left(\frac{\hat{\mathbf{z}}_i}{\|\hat{\mathbf{z}}_i\|}\right)\right] \tag{A-11}$$

The Figures A-1.a and A-1.b depict the loss at mobile speeds of 3km/h and 50km/h which corresponds to the maximum Doppler shifts of 5.5Hz and 93Hz respectively.

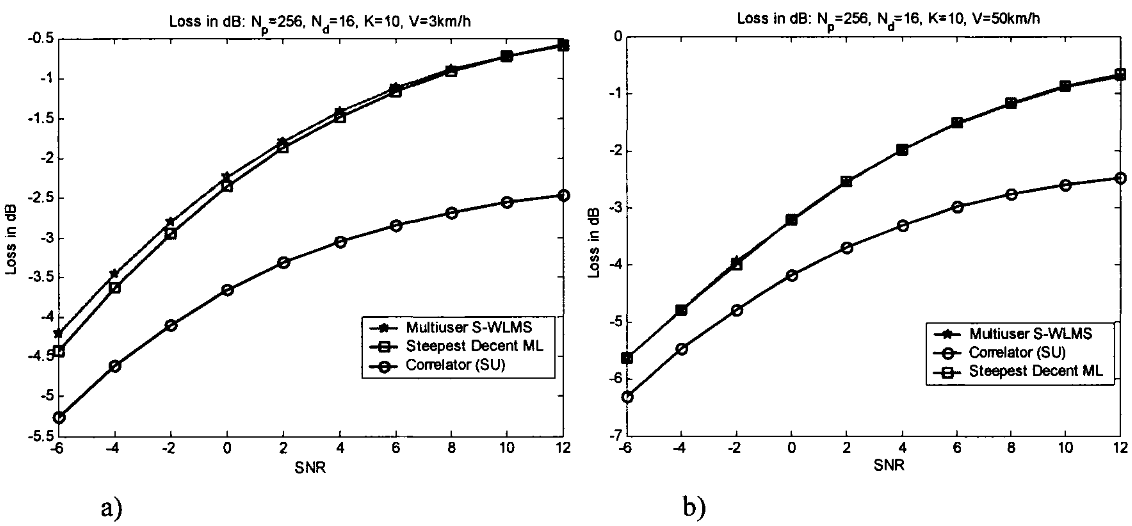

Figure A-1. The loss in dB's (A-11) for $N_p = 256$, $N_d = 16$, $K = 10$, a) $V = 3km/h$ and b) $V = 50km/h$.

We may in a separate figure (Figure A-2) take a close look at the tracking quality using Multiuser S-WLMS for one of the taps, others look almost the same.

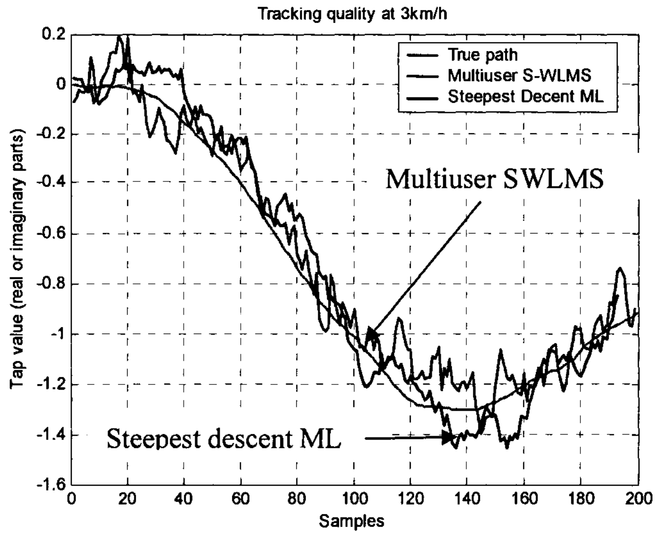

Figure A-2. The tracking quality for $N_p = 256$, $N_d = 16$, $K = 10$, at $V = 3$ Km/h.

APPENDIX B

B ITERATIVE (TURBO) DF-MMSE EXPLOITING BPSK SIGNALING

This appendix discusses the extension of the newly defined MMSE problem in (4.7) by defining the output $y_k(i)$ of a decision feedback MMSE structure [AHM05a] (used in [GAM00] as well) as

$$y_k(i) = \Re\left\{\mathbf{w}_{fk}^{\mathrm{H}}(i)\mathbf{r}_{i/i+1} + \mathbf{w}_{bk}^{\mathrm{H}}(i)\tilde{\mathbf{b}}_{i/i+1}^{(k)}\right\} \tag{B-1}$$

where $\mathbf{w}_{fk}(i)$ is a $2N \times 1$ optimized feed forward vector, $\mathbf{w}_{bk}(i)$ is a $(3K-1)\times 1$ optimized feedback vector, and $\tilde{\mathbf{b}}_{i/i+1}^{(k)}$ is obtained from $E\{\mathbf{b}_{i/i+1}\}$ by eliminating the $(K+k)$th element.

To ease the design let

$$\xi_k(i) = \Re\left(\mathbf{w}_{bk}^{\mathrm{H}}(i)\tilde{\mathbf{b}}_{i/i+1}^{(k)}\right) \tag{B-2}$$

then (B-1) becomes [AHM05a]

$$y_k(i) = \mathbf{w}_{a,fk}^{\mathrm{H}}(i)\mathbf{r}_{a,i/i+1} + \xi_k(i) \tag{B-3}$$

Finding $\mathbf{w}_{a,fk}(i)$ and $\xi_k(i)$ consists in solving the following optimization problem:

$$\begin{aligned}\left\{\mathbf{w}_{a,fk}(i),\xi_k(i)\right\} &= \min_{\substack{\mathbf{w}_a(i)\in\mathbb{C}^{4N\times1},\\ \xi(i)\in\mathbb{R},\\ E\{\mathbf{b}_{i/i+1}\}\neq 0}} E\left\{\left(y_k(i)-b_k(i)\right)^2\right\} \\ &= \min_{\substack{\mathbf{w}_a(i)\in\mathbb{C}^{4N\times1},\\ \xi(i)\in\mathbb{R},\\ E\{\mathbf{b}_{i/i+1}\}\neq 0}} E\left\{\left(\mathbf{w}_a^{\mathrm{H}}(i)\left(b_k(i)\mathbf{B}_{K+k}+\mathbf{B}^{(\mathrm{k})}\mathbf{b}_{i/i+1}^{(\mathrm{k})}+\mathbf{v}_{a,i/i+1}\right)+\xi(i)-b_k(i)\right)^2\right\}\end{aligned} \tag{B-4}$$

where $\mathbf{B}_{K+k}$ is the $(K+k)$th column of the $4N\times 3K$ matrix $\mathbf{B}$, $\mathbf{B}^{(\mathrm{k})}$ is obtained from $\mathbf{B}$ by eliminating the $(K+k)$th column, and $\mathbf{b}_{i/i+1}^{(\mathrm{k})}$ is obtained from the vector $\mathbf{b}_{i/i+1}$ by omitting the $(K+k)$th element.

The MMSE optimization problem is reduced to a solution of the following equations [AHM05a]

$$E\left\{\mathbf{b}_{i/i+1}^{(\mathrm{k})}\right\}^{\mathrm{T}}\mathbf{B}^{(\mathrm{k})\mathrm{H}}\mathbf{w}_a(i)+\xi\ (i)=0 \tag{B-5}$$

$$\left\{\mathbf{B}_{K+k}\mathbf{B}_{K+k}{}^{\mathrm{H}}+\mathbf{B}^{(\mathrm{k})}E\left\{\mathbf{b}_{i/i+1}^{(\mathrm{k})}\mathbf{b}_{i/i+1}^{(\mathrm{k})}{}^{\mathrm{T}}\right\}\mathbf{B}^{(\mathrm{k})\mathrm{H}}+\sigma^2\mathbf{I}_{4N\times 4N}\right\}\times\mathbf{w}_a(i)+\mathbf{B}^{(\mathrm{k})}E\left\{\mathbf{b}_{i/i+1}^{(\mathrm{k})}\right\}\xi(i)=\mathbf{B}_{K+k} \tag{B-6}$$

Solving (B-5) and (B-6) leads to [AHM05a]

$$\mathbf{w}_{a,fk}(i)=\left(\boldsymbol{\psi}_k+\boldsymbol{\varphi}_k(i)+\sigma^2\mathbf{I}_{4N\times 4N}-\boldsymbol{\chi}_k(i)\boldsymbol{\chi}_k^{\mathrm{H}}(i)\right)^{-1}\mathbf{B}_{K+k} \tag{B-7}$$

$$\xi_k(i)=-\boldsymbol{\chi}_k^{\mathrm{H}}(i)\mathbf{w}_{a,fk}(i) \tag{B-8}$$

Using (B-2),

$$\mathbf{w}_{bk}^{\mathrm{H}}(i)=\xi_k(i)\left(\tilde{\mathbf{b}}_{i/i+1}^{(\mathrm{k})}\tilde{\mathbf{b}}_{i/i+1}^{(\mathrm{k})}{}^{\mathrm{T}}\right)^{-1} \tag{B-9}$$

with $\boldsymbol{\psi}_k=\mathbf{B}_{K+k}\mathbf{B}_{K+k}^{\mathrm{H}}$, $\boldsymbol{\varphi}_k(i)=\mathbf{B}^{(\mathrm{k})}\left(\tilde{\mathbf{b}}_{i/i+1}^{(\mathrm{k})}\tilde{\mathbf{b}}_{i/i+1}^{(\mathrm{k})}{}^{\mathrm{T}}+\boldsymbol{\Delta}_k(i)\right)\mathbf{B}^{(\mathrm{k})\mathrm{H}}$ and $\boldsymbol{\chi}_k(i)=\mathbf{B}^{(\mathrm{k})}\tilde{\mathbf{b}}_{i/i+1}^{(\mathrm{k})}$, where

$$\Delta_k(i) = \sum_j \left(1 - E\{\mathbf{b}_{i/i+1}(j)\}^2\right)\mathbf{e}_j\mathbf{e}_j^{\mathrm{T}} + \sum_{j \neq k}\left(1 - E\{\mathbf{b}_{i/i+1}(K+j)\}\right)^2 \mathbf{e}_{K+j}\mathbf{e}_{K+j}^{\mathrm{T}}$$
$$+ \sum_j \left(1 - E\{\mathbf{b}_{i/i+1}(2K+j)\}^2\right)\mathbf{e}_{2K+j}\mathbf{e}_{2K+j}^{\mathrm{T}}$$

and $\mathbf{e}_l$ denotes a $(3K-1)\times 1$ all-zeros vector with "1" at the l-th element. In the first iteration, we set $\tilde{\mathbf{b}}_{i/i+1}^{(\mathrm{k})} = \mathbf{0}_{(3K-1)\times 1}$, which is equivalent to assuming that the code bits are uniformly distributed and equiprobable. At each iteration, $\tilde{\mathbf{b}}_{i/i+1}^{(\mathrm{k})}$ is calculated using the soft information in the form of LLR obtained from the decoder [AHM05a].

Essential to the turbo processing, the SISO detector proposed here should be amounted so that it provides LLR's instead of soft decisions $y_k(i)$. To do this, we assume that the output of the soft MMSE $y_k(i)$ in (B-1) represents the output of an equivalent additive white Gaussian noise channel having $b_k(i)$ as its input symbol [WAN99]

$$y_k(i) = \lambda_k(i)\, b_k(i) + z_k(i) \tag{B-10}$$

where $\lambda_k(i)$ is the equivalent amplitude at instant i for the kth user DF-MMSE filter output, and $z_k(i) \sim N\left(0, \rho_k(i)^2\right)$ is a Gaussian noise [WAN99]. Therefore, the extrinsic information is given by

$$L_{\mathrm{det}}^{\mathrm{ext}}\left(b_k(i)\right) = \frac{2\Re\left(\lambda_k(i)\, y_k(i)\right)}{\rho_k(i)^2} \tag{B-11}$$

Using (4.1) and (B-10), the parameters $\lambda_k(i)$ and $\rho_k(i)^2$ can be derived by respectively computing the mean and the variance of $y_k(i)$ where the expectation is taken with respect to the code bits and $\mathbf{v}_{i/i+1}$

$$\lambda_k(i) = \mathbf{w}_{a,fk}(i)^{\mathrm{H}}\mathbf{B}\hat{\mathbf{b}}_{i/i+1}{}^{(\mathrm{k})} + \xi_k(i)E\{b_k(i)\} \tag{B-12}$$

where $\hat{\mathbf{b}}_{i/i+1}{}^{(\mathrm{k})} = E\{\mathbf{b}_{i/i+1}\}E\{b_k(i)\} + \left(1 - E\{b_k(i)\}^2\right)\mathbf{e}_{K+k}$

$$\begin{aligned}\rho_k(i)^2 = {} & \mathbf{w}_{a,fk}(i)^{\mathrm{H}} E\{\mathbf{r}_{a,i/i+1}\mathbf{r}_{a,i/i+1}{}^{\mathrm{H}}\}\mathbf{w}_{a,fk}(i) + \\ & \xi_k(i)\xi_k^*(i) + 2\Re\left(\xi_k(i)E\{\mathbf{b}_{i/i+1}\}^{\mathrm{T}}\mathbf{B}^{\mathrm{H}}\mathbf{w}_{a,fk}(i)\right) - \lambda_k(i)^2\end{aligned} \tag{B-13}$$

Note that the algorithm complexity obvious from (B-7) where the inverse of a $4N\times 4N$ matrix, $\left(\boldsymbol{\psi}_k + \boldsymbol{\varphi}_k(i) + \sigma^2\mathbf{I}_{4N\times 4N} - \boldsymbol{\chi}_k(i)\boldsymbol{\chi}_k^{\mathrm{H}}(i)\right)$, is performed at each instant i, a complexity burden $O\left(2^3(2N)^3\right)$ as compared with $O\left((2N)^3\right)$ of [GAM00]. We can reduce the computational complexity to the order of $O\left(P_{iteration}(4N)^2\right)$ using the following steepest descent algorithm.

At instant i, compute iteratively (locally) $P_{iteration}$ times,

$$\begin{aligned}&\mathbf{w}_{a,k}^{\mathrm{p}}(i) = \left(\mathbf{I}_{4N\times 4N} - \mu\left(\boldsymbol{\psi}_k + \boldsymbol{\varphi}_k(i) + \sigma^2\mathbf{I}_{4N\times 4N} - \boldsymbol{\chi}_k(i)\boldsymbol{\chi}_k^{\mathrm{H}}(i)\right)\right)\mathbf{w}_{a,k}^{\mathrm{p-1}}(i) + \mu\mathbf{B}_{K+k},\\ &p = 1,2,\ldots,P_{iteration}\end{aligned} \tag{B-14}$$

The complexity saving is obvious if at least $P_{iteration} < 4N$. The simulation results demonstrate that for almost all cases the steepest descent solution is as good as the direct inversion for a $P_{iteration}$ of 2 to 3 using a proper choice of adaptation step μ.

A summary of the simulation results is provided in [AHM05a]. The newly defined iterative DF-MMSE demonstrates a superior performance relative to its conventional Iterative DF-MMSE [GAM00].