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COMME EXIGENCE PARTIELLE
DE LA MAÎTRISE EN GÉNIE ÉLECTRIQUE

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Résumé substantiel en français

Contexte

De nos jours, la communication sans fil est largement répandue et le nombre d'utilisateurs augmente sans cesse. Le coût d'opération de ces réseaux, leur consommation d'énergie et leur taux d'erreurs (BER) sont cependant des problèmes cruciaux pour cette nouvelle technologie émergente.

Depuis que le premier système numérique commercial (GSM, 2G) a été lance en Finlande en 1991 [1][2], des ingénieurs ont développé sans cesse la technologie au niveau de la performance. Depuis ce temps, les technologies sans fil ont passé de la deuxième génération à la troisième génération et nous pourrions voir une quatrième (4G) ainsi qu'une cinquième génération (5G) prochainement. La technologie sans fil évolue donc très rapidement. Par exemple, dans un système 2G typique, les données se transmettent à une vitesse de 56Kbits par secondes alors que dans la technologie GSM, GPRS la vitesse passait à 114 Kbits par seconde. Avec le développement de la technologie EDGE, une évolution de la rapidité pour GSM, apparue en 2003, il était désormais possible d'atteindre le 236.8 Kbit/s.

Dans le but de satisfaire la forte demande en rapidité de transfert et en bande passante, une troisième génération (3G) a été développée [4]. Cette technologie a révolutionné l'Industrie avec un 14.4Mbits par seconde en réception et 5,8Mbits en envoie. Basé sur un haut débit de donnée, le réseau 3G a permis le développement à grande échelle de la téléphonie numérique sans fil, des appels vidéo ainsi que la technologie mobile à large bande.

Le terme 4G (quatrième technologie) sert présentement à décrire la prochaine évolution complète en matière de technologie sans fil. Le système 4G sera en mesure de fournir une solution IP ou la voix, les données et du streaming multimedia pourraient être fournis a l'utilisateur sur un «*n'importe où, n'importe quand*». Il n'y a cependant aucune définition concrète de ce que sera le 4G. Toutefois, on peut déjà dire que le 4G sera un système ayant complètement intégré la technologie IP. Le 4G permettra d'atteindre une vitesse entre 100Mbits/sec et 1Gbits/s tant à l'intérieur qu'à l'extérieur, et ce, avec une très haute qualité et sécurité [3].

Problèmes

MIMO-CDMA est une technologie intégrée qui a principalement été conçue pour le 4G et même le 5G. Nous continuerons d'utiliser la base «codage-décodage» mais quand nous ferons un travail de transmission nous utiliserons une antenne MIMO pour émettre et recevoir les signaux. Comme pour dans la technologie traditionnelle CDMA, l'estimation des canaux est toujours un élément important dans l'ensemble du réseau.

La plupart des recherches et des expériences se sont concentrées sur un seul algorithme ou sur une version améliorée d'un seul. Dans ces cas, nous pouvons seulement voir la performance de cet algorithme. De plus, toutes ces recherches ont été conduites dans des environnements différents au niveau de l'encodage et des canaux. En d'autres mots, elles n'utilisent généralement pas la même plateforme. Dans cette situation, nous ne pouvons pas voir la comparaison des différents algorithmes clairement. Nous n'avons donc aucune idée si un algorithme est meilleur qu'un autre. Il est donc très important de construire une telle plateforme qui mettra les différentes estimations de canaux dans un même environnement.

Méthodologie

Avant de commencer une recherche dans ce domaine, nous avons consulté plusieurs références à la bibliothèque et sur Internet. Nous étudierons les algorithmes d'estimation du canal pour les systèmes MIMO. Ensuite, nous appliquerons aux systèmes MIMO-CDMA. Après, nous avons construit une plateforme de simulation. Nous assumons la situation multiutilisateur. La plateforme de simulation devra comprendre une station de base et plusieurs utilisateurs. Pour chaque utilisateur/ station, il y aura plusieurs antennes. Nous devrons aussi considérer le modèle du signal pour rendre la situation le plus proche possible de la réalité.

Lorsque la plateforme de simulation sera complétée, nous étudierons plusieurs algorithmes d'estimation de canaux. Nous tenterons aussi d'intégrer ces algorithmes sur la plateforme construite précédemment.

Ensuite, nous ferons la simulation de tous ces algorithmes afin d'obtenir des résultats et enfin, nous comparerons les résultats.

Système MIMO-CDMA

CDMA

Dans un système radiophonique, il y a deux bases que l'on doit prendre en compte soit la fréquence et le temps. Le système est divisé et une fréquence du spectre est alors allouée en tout temps à chaque paire de communicateurs. On appelle cela l'accès multiple par répartition en fréquence (ou FDMA en anglais). Les systèmes sont divisés en temps et chaque pair de communicateurs alloue tout (ou du moins une grande partie) du spectre pour la partie du temps appelé multiplexage temporel ou plus couramment *Time division multiple access (TDMA)* en anglais. En CDMA, tout le spectre sera alloué aux communicateurs en tout temps.

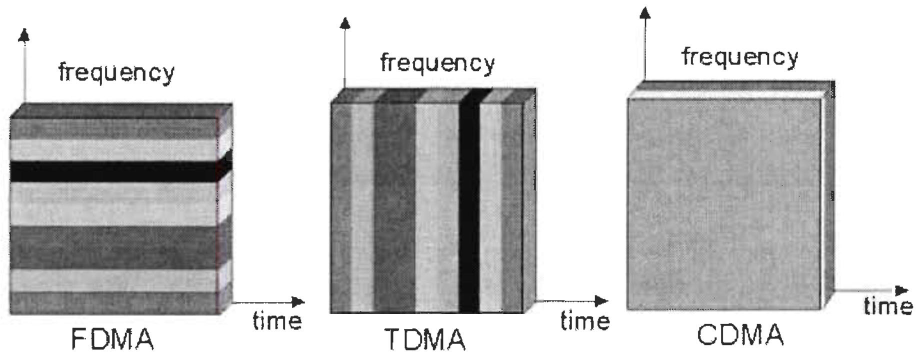


Figure 2.1 Comparaison de différent accès.

Le CDMA utilise des codes pour identifier les connexions. C'est une forme d'étalement de spectre à séquence directe (DSSS) qui est utilisée en communication militaire depuis longtemps. En réponse à une demande mondiale sans cesse croissante en communication mobile et sociale, la technologie à large spectre numérique a accompli une bien meilleure efficacité en utilisation de bande passante par chaque allocation de spectre et par le même fait, a pu répondre à bien plus d'utilisateurs qu'en technologie analogique et même les autres technologies numériques . Comme auparavant lors de son implantation chez les militaires, les réseaux à large spectre ont accompli beaucoup en incorporant un grand nombre de fonctions uniques. En plus d'augmenter l'efficience du spectre, cela élimine aussi la tâche de planifier différentes allocations de fréquences pour les utilisateurs avoisinants. Beaucoup d'options sur les systèmes à accès multiples sont rendues possibles par cette fréquence universelle réutilisée par des terminaux employant des ondes du style sonore (*noise-like signal waveforms*). L'avantage le plus important est la rapidité et la précision de l'ajustement de la puissance, ce qui assure un haut niveau de transmission de qualité pendant qu'on règle le problème de la distance en maintenant un bas niveau de puissance pour chaque terminal et par le même fait, un bas niveau d'interférence aux autres terminaux.

MIMO

Multiple Input Multiple Output (MIMO) est une technologie d'antenne dans laquelle plusieurs antennes sont utilisées pour à la fois les transmetteurs et les émetteurs. C'est une forme d'antenne intelligente. Sa plus ancienne mention sur le terrain remonte à A.R Kaye et D.A George en 1970 et W. van Etten en 1975 et 1976. Après cela, plusieurs recherches sur le *beamforming* ont mentionné des applications en 1984 et 1986 publiées par Jack Winters et Jack Salz au laboratoire Bell. Jusqu'en 1993, Arogyaswami Paulraj et Thomas Kailath ont proposé le concept du *Spatial Multiplexing* utilisant MIMO. C'était la première fois que cette technologie était utilisée pour une diffusion sans fil. Dans le domaine commercial, Iospan Wireless Inc. a développé le premier système en 2001 qui utilisait la technologie MIMO-OFDMA. Ce système supportait un encodage diversifié ainsi que le *Spacial Multiplexing*. En utilisant la technologie MIMO, nous pouvons minimiser les erreurs et optimiser la vitesse des données. En même temps, cela ne nécessitera pas de bandes passantes additionnelles ou de puissance de transmission.

Étant une antenne intelligente, MIMO s'est développé à partir de SISO (Single Input Single Output), SIMO (Single Input Multiple Output) et MISO (Multiple Input Single Output).

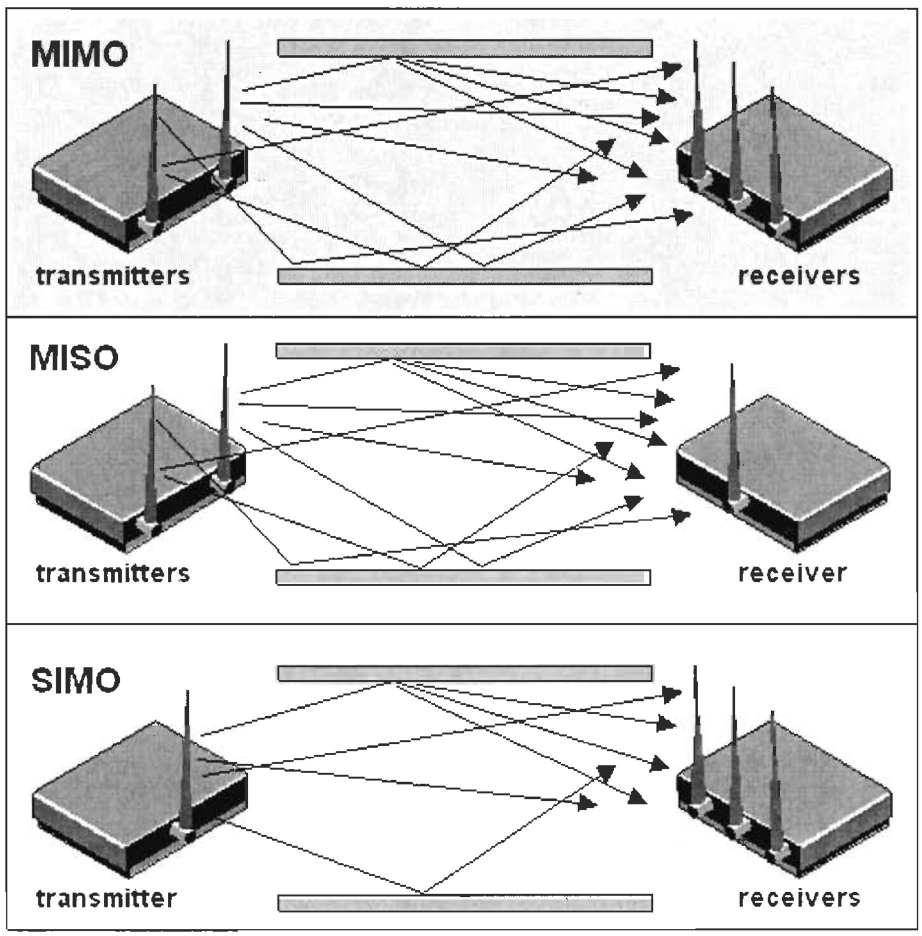


Figure 2.5 SISO à MIMO

Dans la télécommunication traditionnelle, nous retrouvons une antenne de transmission ainsi qu'une autre qui reçoit. La plupart des systèmes fonctionnent sous cette structure. Cette situation à changé avec l'apparition de la technologie SIMO. Avec une seule antenne, le SIMO avait la même rapidité de transmission mais avec un bien plus grand périmètre. La technologie MISO fonctionne sur une autre structure. En comparaison avec le SISO, le système MISO a la même vitesse de transmission et le même périmètre mais est beaucoup plus fiable. Le système MIMO a quant à lui améliorer la vitesse et le périmètre.

Algorithme d'estimation de canaux

Les canaux de radio dans les systèmes mobiles ont souvent causé les interférences inter-symbole (ISI) dans le signal reçu. Pour enlever ces ISI du signal, plusieurs sortes d'égaliseurs peuvent être utilisés. Les algorithmes de détection basés sur les recherches de Trellis (MLSE et MAP) offrent une bonne performance de réception en plus de ne pas être trop complexes informatiquement. Ces algorithmes sont donc devenus très populaires.

Cependant, ces détecteurs demandent une bonne connaissance sur la réponse impulsionnelle (CIR), ce qui est fourni séparément par un bon estimateur de canaux. Normalement l'estimation des canaux est basée sur une séquence connue de bits, qui est unique pour certains transmetteurs et qui se répète à chaque rafale de transmission. L'estimateur de canaux est donc capable d'estimer le CIR pour chaque rafale séparément en exploitant les bits transmis et en les comparant aux échantillons correspondants.

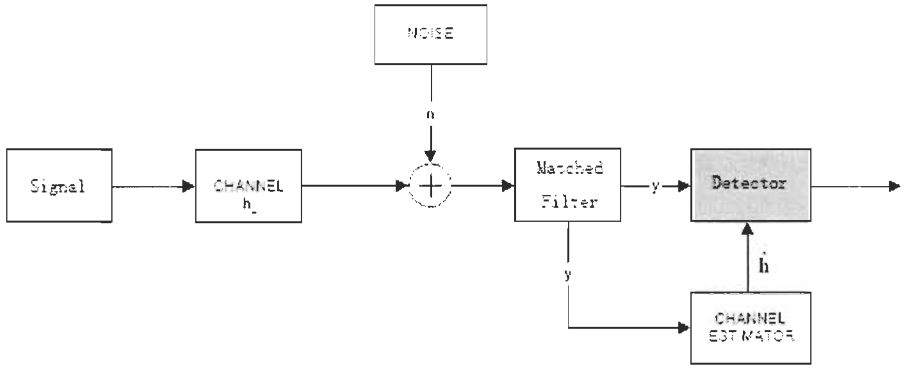


Figure 3.1 Plan d'un estimateur de canaux.

La figure 3.1 ci-dessus est le plan d'un estimateur dans un système CDMA. Nous pouvons y voir le signal transmis dans un canal multiple atténuant (*fading multiple channel*). Des bruits thermiques sont générés d'un côté d'un récepteur. Un détecteur est aussi utilisé pour détecter le signal original reçu. Simultanément, le détecteur a aussi besoin du canal estimé $\hat{h}$ d'un estimateur de canaux spécifiques. Le signal y reçu peut être exprimé comme suit:

$$y = Mh + n \tag{3.1}$$

Où la réponse d'impulsion complexe de canal h du signal voulu est exprimée comme suit:

$$h = [h_0 \quad h_1 \quad \cdots \quad h_L]^T \tag{3.2}$$

n exprime les échantillons de bruit

Généralement, on utilise quatre méthodes en algorithme d'évaluation de canal. Soit : l'évaluation sans visibilité (*Blind Channel Estimation*), la technique indirecte comme celle de la matrice inversée, et l'évaluation mixe (comme démontré dans la figure 4.1, le détecteur et l'estimateur fonctionnent ensemble). Dans cet article, nous nous concentrerons sur la technique directe et la technique indirecte. Nous assumerons donc l'hypothèse que nous connaissons déjà le signal original du côté du récepteur. En utilisant cette information, nous faisons l'estimation pour avoir le $\hat{h}$.

Techniques indirectes

Estimation des LS (Least Square)

L'estimation des moindres carrés peut être interprétée comme une méthode pour adapter les données. La meilleure valeur dans les valeurs carrées sont l'instance dans laquelle le modèle pour lequel la somme des carrés résiduels est la plus petite valeur et par résiduel, on veut dire la différence entre la valeur observe et la valeur donnée par le modèle. La méthode a été décrite originalement par Carl Friedrich Gauss autour de 1974.

Selon (3.1) et (3.2), nous pouvons obtenir :

$$\hat{h} = \arg\min_h \|y - Mh\|^2 \tag{3.3}$$

Supposant que nous avons le bruit, nous obtenons:

$$\hat{h}_{LS} = (M^H M)^{-1} M^H y \tag{3.4}$$

Où $(\,)^H$ et $(\,)^{-1}$ dénote les matrices hermitiennes et inverses respectivement.

1 Estimation de canal LMS (Least-Mean-Squares)

L'algorithme LMS est un important membre de la famille stochastique gradient. Le terme stochastique gradient algorithme est prévu pour distinguer l'algorithme LMS de la méthode de la descente rapide (*steepest descent)*, qui emploie un gradient déterministe dans le calcul récursif de Wierner pour les entrées stochastiques. Un avantage significatif de l'algorithme LMS est sa simplicité. D'ailleurs, il n'exige pas de mesure de corrélation ni de matrice inversée

L'algorithme LMS est un algorithme linéaire de filtrage adaptatif qui se compose généralement de deux processus de base:

1. Un procédé de filtration, qui implique (a) un calcul de rendement linéaire en réponse à un signal d'entrée et (b) la production d'une erreur d'estimation comparant ce rendement à une réponse désirée.
2. Un processus adaptatif qui comporte l'ajustement automatique des paramètres du filtre selon l'erreur d'évaluation.

La combinaison de ces deux processus fonctionnant ensemble constitue une boucle de contre-réaction montrée ci-dessous:

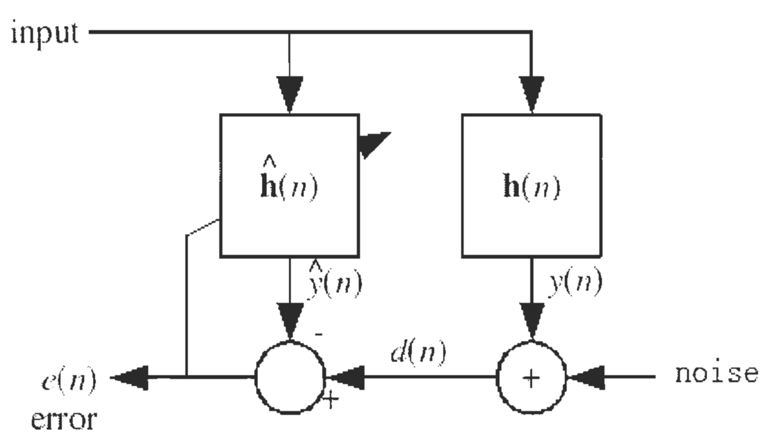


Figure 3.4 Diagramme Block de filtrage adaptée

Dans cette recherche, nous considérons, la situation par trajets multiples dans le système de SISO et MIMO. Nous assumons aussi que le délai sera de 6 ce qui signifie que l'ordre de filtrage de tous les filtres adaptatifs qui sont utilisés sera de 6.

L'algorithme LMS pour un ordre de pth peut être résumé ainsi:

$$\hat{h}(0) = 0 \tag{3.5}$$

$$x(n) = [x(n), x(n-1), \ldots, x(n-p+1)]^T \tag{3.6}$$

$$e(n) = d(n) - \hat{h}^H(n)x(n) \tag{3.7}$$

$$\hat{h}(n+1) = \hat{h}(n) + \mu e^*(n)x(n) \tag{3.8}$$

Algorithme de RLS (Recursive Least Squares)

L'algorithme RLS est employé dans des filtres adaptatifs pour trouver les coefficients de filtrage qui se rapportent à produire périodiquement le moins de carrés du signal d'erreur. Cela est contraire aux autres algorithmes qui visent à réduire l'erreur de moyenne carrée. La différence est que les filtres RLS dépendent des signaux eux-mêmes, tandis que les filtres MSE dépendent de leurs statistiques. Si ces statistiques sont connues, un filtre MSE avec des coefficients fixes peut être construit. Un dispositif important de ce filtre est que son taux de convergence est typiquement un ordre de grandeur plus rapide que celui du filtre LMS étant donné que le filtre RLS blanchit les données d'entrée en employant la matrice de corrélation inverse des données, assumées pour être de zéro. Cette amélioration en performance est cependant réalisée aux dépens d'une augmentation de la complexité informatique du filtre RLS.

L'algorithme RLS pour un filtre RLS de l'ordre *p*th peut être résumé ainsi:

$$w_n = 0 \tag{3.9}$$

$$P(0) = \delta^{-1} I \tag{3.10}$$

$$x(n) = \begin{bmatrix} x(n) \\ x(n-1) \\ \vdots \\ x(n-p) \end{bmatrix} \tag{3.11}$$

$$\alpha(n) = d(n) - w(n-1)^T x(n) \tag{3.12}$$

$$g(n) = P(n-1)x^*(n)\{\lambda + x^T(n)P(n-1)x^*(n)\}^{-1} \tag{3.13}$$

$$P(n) = \lambda^{-1}P(n-1) - g(n)x^T(n)\lambda^{-1}P(n-1) \tag{3.14}$$

$$w(n) = w(n-1) + \alpha(n)g(n) \tag{3.15}$$

Où p est un ordre de filtre, λ est un facteur d'oublie, δ est la valeur pour initialiser $P(0)$.

Algorithme MMSE (Minimum Mean Square Error)

L'estimation de canal MMSE se résume à l'équation suivante:

$$\hat{\mathrm{h}}_{\mathrm{n,MMSE}} = [\mathrm{S}^{\mathrm{H}}\mathrm{R}_{\mathrm{nn}}^{-1}\mathrm{S} + \mathrm{R}_{\mathrm{hh}}^{-1}]^{-1}\mathrm{S}^{\mathrm{H}}\mathrm{R}_{\mathrm{nn}}^{-1}\mathrm{x}_{\mathrm{n}} \tag{3.16}$$

Où $R_{hh} = E\{h_n \quad h_n^H\}$ est la matrice de corrélation de la réponse d'impulsion du canal du vecteur h_n. Sur le côté du récepteur, la matrice de corrélation du canal R_{hh} est inconnue.

$$R_{hh} = E\{h_n \quad h_n^H\} \tag{3.17}$$

$$= diag\left(\sigma_{n,1}^2(0), \dots, \sigma_{n,1}^2(L), \sigma_{n,2}^2(0), \dots, \sigma_{n,N_T}^2(L)\right)$$

Considérant la normalisation, nous obtenons:

$$E\left\{\left|h_{n,m}(k)\right|^2\right\} = \sigma_{n,m}^2(k) = \sigma_h^2 = \frac{1}{L+1} \qquad \forall n, m, k \tag{3.18}$$

L'estimation de canal MMSE est donnée par:

$$\hat{h}_{n,MMSE} = \left[S^H S + \frac{\sigma_n^2}{\sigma_h^2} I\right]^{-1} S^H x_n \tag{3.19}$$

L'expression $\left[S^H S + \frac{\sigma_n^2}{\sigma_h^2} I\right]^{-1} S^H$ peut être calculée un temps d'avance, car elle dépend des symboles connus et la puissance du bruit σ_n^2.

Algorithme sans visibilité (Blind Channel Estimation)

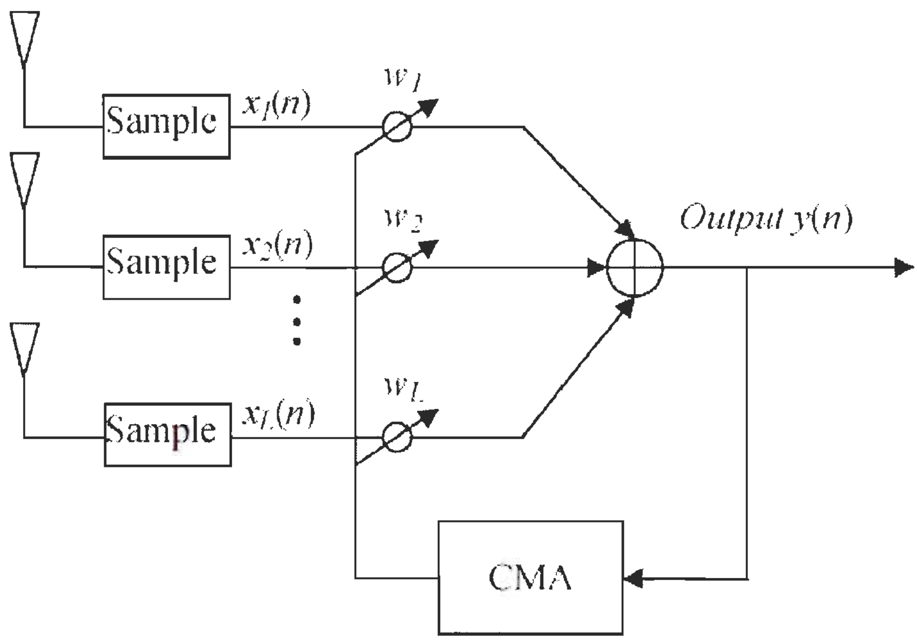


Figure 3.7 Structure du CMA

A Constant Modulus Algorithm (CMA) de tableau est indiqué ci-dessus [35]. Les signaux sont reçus par un réseau d'antennes.

$$y(n) = \sum_{l=1}^{L} w_l^* x_l(n) \quad (3.189)$$

Le but du tableau d'adaptation est d'extraire le signal désiré en trouvant un vecteur de poids approprié. Il est bien connu que l'algorithme d'adaptation basé sur l'algorithme RLS ont un taux de convergence plus rapide. Donc, nous obtenons l'algorithme RLS-CMA comme ci-dessous:

$$w(0) = \left[1, 0_{1\times(L-1)}\right]^T \quad (3.190)$$

$$C(0) = \delta^{-1} I_{L\times L}, \delta = \text{petite valeur positive} \quad (3.191)$$

$$z(n) = x(n)x^H(n)w(n-1)|x^H(n)w(n-1)|^{p-2} \tag{3.192}$$

$$h(n) = z^H(n)C(n-1) \tag{3.193}$$

$$g(n) = C(n-1)z(n)/(\lambda + h(n)z(n)) \tag{3.194}$$

$$C(n) = \frac{C(n-1)-g(n)h(n)}{\lambda} \tag{3.195}$$

$$e(n) = w^H(n-1)z(n) - 1 \tag{3.196}$$

$$w(n) = w(n-1) + g(n)e^*(n) \tag{3.197}$$

Simulations

Simulation d'un SISO-CDMA

Dans la simulation, nous commencerons à partir de la plate-forme SISO-CDM parce que l'antenne SISO est moins compliquée que le système d'antenne MIMO. En attendant, nous pouvons comparer les résultats de la simulation SISO aux résultats de la simulation MIMO. Nous pouvons voir les différences du même algorithme d'évaluation de canal entre deux systèmes différents d'antenne. Nous pouvons voir comment les éléments de canal affectent les résultats d'évaluation.

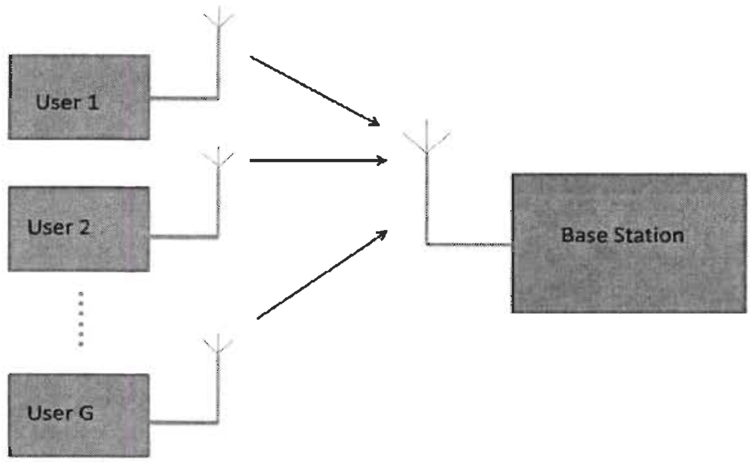


Figure 4.1 **SISO-CDMA System Block Diagram**

Dans cette simulation, nous supposons l'environnement comme suit : du côté de l'utilisateur, nous avons deux utilisateurs qui ont chacun une antenne de transmission. Du côté de la station de base, il a une antenne de réception. Chaque longueur de signal est de 10000 et la longueur de diffusion est de 31. Le nombre de multitrajet est de 6. Pour l'algorithme NLMS, nous savons qu'il est l'algorithme normalise du LMS. Afin de rivaliser avec l'algorithme LMS, nous assumons la même taille d'étape soit $\mu = 0.001$. Pour l'algorithme RLS, nous assumons un facteur d'oublie$\lambda = 1$, un coefficient initial $\delta = 0.001$. Nous savons que pour le coefficientλ, plus petit est le λ , plus petite sera sa contribution des échantillons précédents.

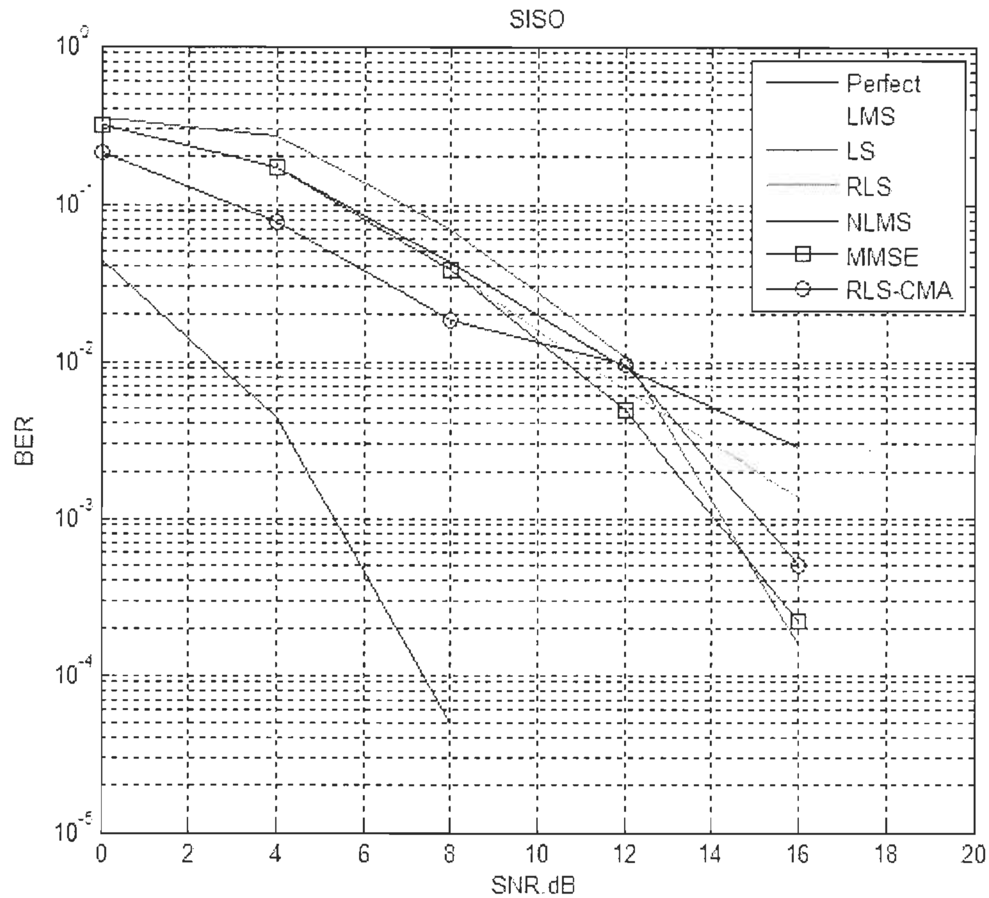


Figure 4.2 Résultats des simulations en environnement SISO-CDMA (1)

Nous pouvons voir que dans un environnement SISO-CDMA, lorsqu'un SNR est entre 0dM et10dB, les algorithmes adaptatifs ont des performances semblables avec l'algorithme RLS-CMA. D'une façon générale, en cet état de SNR, l'algorithme adaptatif et RLS-CMA est un peu meilleur que l'algorithme LS. Dans la même gamme de SNR, les algorithmes MMSE et algorithmes adaptatifs n'ont pas beaucoup de différence. Quand le SNR est entre 15dB et 20 dB, nous pouvons voir que l'algorithme LS devient bien mieux que tous les autres algorithmes.

Simulation du cas MIMO-CDMA

Dans l'environnement de simulation MIMO, nous supposons que nous avons de multiples utilisateurs et le nombre d'utilisateurs est G. Pour chaque utilisateur, nous avons plusieurs antennes de transmission, le nombre d'antennes de transmission est Nt. Sur le côté de la station de base, le nombre d'antennes est Nr. Nous avons le schéma MIMO-CDMA dans le graphique suivant :

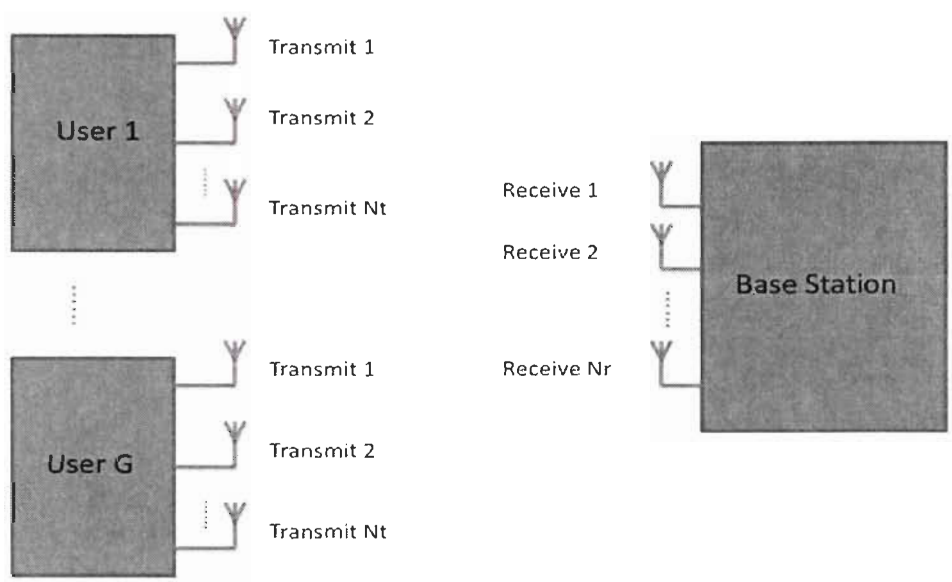


Figure 4.3 MIMO System Block Diagram

Afin de s'approcher de l'environnement le plus réaliste possible, nous devrions également considérer le canal par trajets multiples dans ce cas-ci.

Dans cette simulation, nous supposons que nous avons deux utilisateurs. Chaque utilisateur a deux antennes de transmission. Sur le côté du récepteur, le nombre d'antennes de réception est de quatre. Chaque paire de canaux de signal (une antenne de transmission et une antenne de réception) a un canal d'effacement de 6 chemins (*fading channel*).

Pour l'algorithme d'évaluation LMS, nous supposons la taille $\mu = 0.1$. Nous savons que plus la valeur de la taille d'étape est grande, plus la convergence d'algorithme sera rapide. Vu la complexité du système d'antenne MIMO, nous espérons que ce sera rapide

Pour l'algorithme NLMS, nous choisissons la même taille que le LMS soit : $\mu = 0.1$.

Pour l'algorithme RLS, nous assumons un facteur d'oublis $\lambda = 1$, un coefficient initial de $\delta = 0.001$. Nous savons que pour le coefficient λ, plus petit est λ is, plus petit sera la contribution des échantillons précédents. Dans le système d'Antenne MIMO, l'état des canaux est très compliqué.

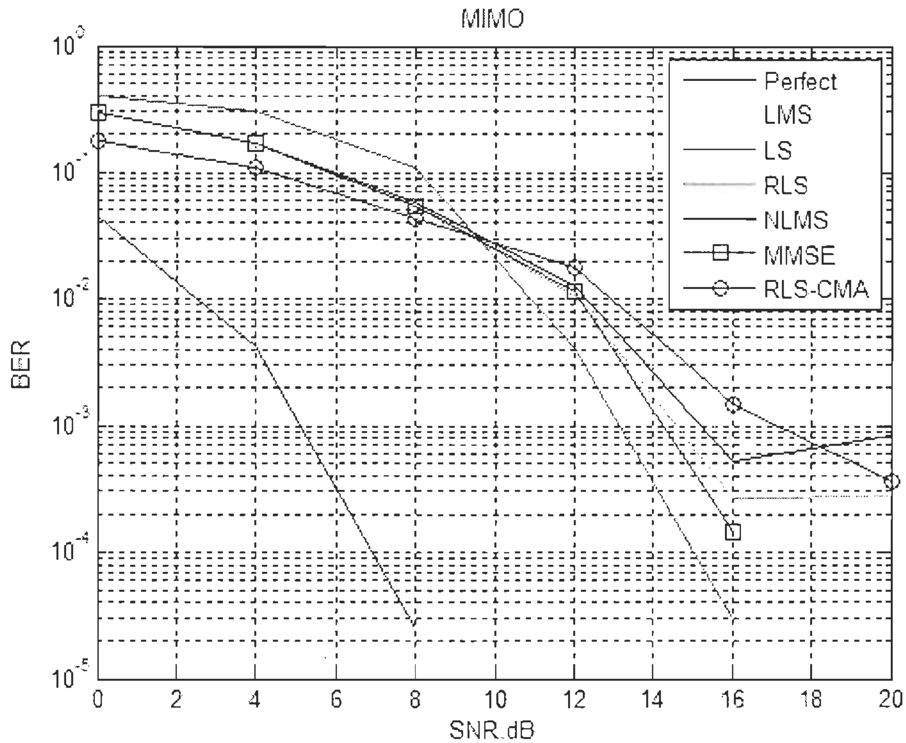


Figure 4.4 Résultats de la simulation en MIMO-CDMA environnement

Nous pouvons voir dans la figure 4.4 que dans l'environnement MIMO-CDMA, lorsque le SNR est entre 0dB et 10dB, tous les autres algorithmes sont meilleurs que l'Algorithme LS. Par contre, la performance de l'algorithme RLS-CMA est à peine meilleure que la performance du algorithme adaptatif dans cette gamme de SNR. Dans ce cas-ci, la différence entre l'algorithme LMS et NLMS est visible. Normaliser aide l'algorithme NLMS mais généralement, tous les algorithmes adaptés ont des performances similaires. Nous notons que quand le SNR est entre 10dB et 20dB, l'algorithme LS devient mieux que pour les algorithmes adaptatifs. Dans l'ensemble de l'algorithme adaptatif, l'algorithme MMSE a une meilleure performance que tous les autres algorithmes.

CONTENTS

1. Introduction

1.1 Background

Nowadays, wireless communication develops quickly. The number of users becomes more and more. Cost of network operation, power consumption, number of users and low bit error rate (BER) are the main issues of emerging wireless technologies. Since 1991, the first GSM (Global System for Mobile communications) network was launched in Finland [1]. Telecommunication system comes through several generations' development.

1.1.1 2G Network

In 1991, Finland launched the first GSM network. Since then, engineers are always trying to improve the performance of telecommunication systems. GSM is considered as a typical 2G network technology. It can provide voice call service, text messaging service and so on. Because of the limitation of bandwidth, a traditional GSM network cannot support high-speed data transform. Which means it's impossible to use a GSM cell phone to browser website via GSM network. That's why GPRS (General Packet Radio Service) comes out. Compare with "pure" GSM network, GPRS network can provide data rate from 56 up to 114 kbit/s. It's approaching early-age internet dialing access rate on computer. We often describe GPRS as a 2.5G technology, which between 2G and 3G. In 2003, a new technology called EDGE (Enhanced Data rates for GSM Evolution) deployed on GSM network [2]. It is also a technology which enhanced traditional GSM network. It can carry data rate up to 236.8 kbit/s for 4 timeslots. Which means it's 4 times faster than a standard GPRS network. In fact, EDGE can match part of the requirements of 3G telecommunication system. So sometimes we sort it into 3G. But most frequently we consider it as a 2.75G technology, which exactly indicated its position. It is based on GSM network. We don't need to change any core equipments to upgrade to EDGE. Only some modifications are needed in base station. That's why EDGE becomes more and more popular in worldwide. It is the easiest and most economy resolution for all mobile operators in the world.

1.1.2 3G Network

Although GSM network is the most widely-used network on the earth, meanwhile we also have new technology such as EDGE to improve the performance of GSM network, we still need faster data rate and more bandwidth. 3G (Third Generation) system is exactly designed for this. 3G network enable network operators to offer users a wider range of more advanced services while achieving greater network capacity through improved spectral efficiency. Services include wide-area wireless voice telephony, video calls, and broadband wireless data, all in a mobile environment. Additional features also include HSPA data transmission capabilities able to deliver speeds up to 14.4 Mbit/s on the downlink and 5.8 Mbit/s on the uplink. Right now we have 3 mainly standards for 3G network, which are CDMA2000, WCDMA and TD-SCDMA. These technologies are based on CDMA.

1.1.3 4G Network

4G (Fourth Generation) telecommunication system is a term used to describe the next complete evolution in wireless communications [3]. A 4G system will be able to provide a comprehensive IP solution where voice, data and streamed multimedia can be given to users on an "Anytime, Anywhere" basis, and at higher data rates than previous generations. There is no formal definition for what 4G is. However, there are certain objectives that are projected for 4G. These objectives include: that 4G will be a fully IP-based integrated system. 4G will be capable of providing between 100 Mbit/s and 1 Gbit/s speeds both indoors and outdoors, with premium quality and high security.

In order to achieve this aim, we bring several new concepts in. IPv6 is one of the most important concepts. Because 4G will be based on packet switching only. This will require low-latency data transmission. By the time that 4G is deployed, the process of IPv4 address exhaustion is expected to be in its final stages. Therefore, in the context of 4G, IPv6 support is essential in order to support a large number of wireless-enabled devices. Another important new concept for 4G networks is advanced antenna system. The performance of radio communications obviously depends on the advances of an antenna system, refer to smart or intelligent antenna. This increases the data rate into multiple folds with the number equal to minimum of the number of transmit and receive antennas, which is called MIMO (as a branch of intelligent antenna). Apart from this, the reliability in transmitting high speed data in the fading channel can be improved by using more antennas at the transmitter or at the receiver. This paper is exactly focus on this subject. A MIMO model will be built on CDMA basis.

1.2 Problems

Code Division Multiple Access (CDMA) is a well-known channel access method technology which is used in commercial telecommunication for a long time. CDMA uses spread-spectrum technology and a special coding scheme to allow multiple users to be multiplexed over the same physical channel.

Multiple Input Multiple Output (MIMO) is actually a kind of smart antenna technology. Compare with traditional antenna system, such as SISO (Single Input Single Output), MIMO system uses more than one antenna in both transmits and receives side which can improve the capacity of the signal channel. So, we can get the better performance of the system. In this work, MIMO-CDMA systems are considered combining multiple access scheme capability of CDMA with MIMO performance.

We have lots of papers and researches about channel estimation in traditional CDMA. But for MIMO-CDMA, there is not as much research as SISO-CDMA. Most of the research and paper only focus on one kind of algorithm or improve from one kind of algorithm. In these cases, we can only see the performances of one algorithm. Besides, all these researches are in different environment, for example the different channel and the different coding system. In other words, they usually do not use the same platform. In this situation, we cannot see the comparison of different algorithms' performances clearly. It becomes difficult to compare their performances. So it is very important to build such a platform which makes all different channel estimation algorithms work in the same environment.

1.3 Objective

The main objective of this work is to compare channel estimation techniques for MIMO-CDMA systems within the same simulation environment.

1.3 Methodology

Before doing research in this field, we collect lots of research paper form library and internet. We know that there are not many papers in MIMO-CDMA channel estimation field. So we try to study the other channel estimation algorithm in SISO DS-CDMA system also. Then we can try to extend the SISO technology into MIMO filed.

After then, we build a simulation platform. We assume the multi-users situation. The simulation platform should contain a base-station and several users. For each user/station, we assume multiple antennas situation. We also should consider the signal channel model to make it close to real.

When simulation platform is done, we study different kind of channel estimation algorithms. Try to implement these algorithms on the platform we built previously. Then, we do the simulation of all these algorithms.

1.4 Organization of the papers

In Chapter 2, we will introduce the MIMO-CDMA system. We study from the single input single output system to multiple input multiple output system. In Chapter 3, we study several kinds of channel estimation algorithms, which include indirect technologies and direct technologies. In Chapter 4, we study the simulation progress. We do simulation on both SISO and MIMO platform by using different kind of channel estimation algorithms. In Chapter 5, we make the conclusion.

2. MIMO-CDMA System

2.1 CDMA

2.1.1 Background

For radio systems there are two resources, frequency and time. System which division by frequency, and each pair of communicators is allocated part of the spectrum for all of the time, is called Frequency Division Multiple Access (FDMA). System division by time, and each pair of communicators is allocated all (or at least a large part) of the spectrum for part of the time is called Time Division Multiple Access (TDMA). In Code Division Multiple Access (CDMA), every communicator will be allocated the entire spectrum all of the time.

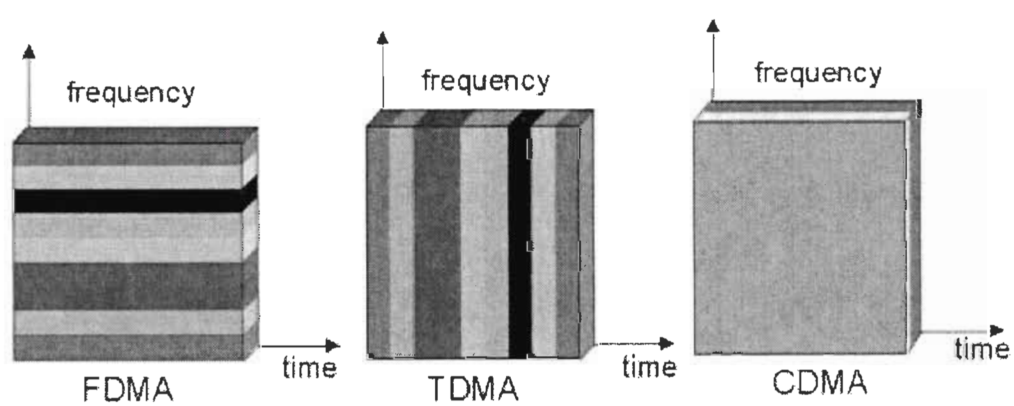


Figure 2.1 Comparison of different access [22]

CDMA uses codes to identify connections. It is a form of Direct Sequence Spread Spectrum communication technology, which has been used in military communications for a long time. In response to an ever-accelerating worldwide demand for mobile and personal portable communications, spread spectrum digital technology has achieved much higher bandwidth efficiency for given wireless spectrum allocation, and hence serves a far large population of multiple access users, than analog or other digital technologies. Like its implementation in military predecessors, the spread spectrum wireless network achieves efficiency improvements by incorporating a number of unique features made possible by the benign noise-like characteristics of the signal waveform. Chief among these is universal frequency reuse. Besides increasing the efficiency of spectrum usage, this also eliminates the chore of planning for different frequency allocation for neighboring users or cells. Many other important multiple access system features are made possible through this universal frequency reuse by terminals employing wideband (spread) noise-like signal waveforms. Most important one is fast and accurate power control, which ensure a high level of transmission quality while overcoming the distance problem by maintaining a low transmitted power level for each terminal, and hence a low level of interference to other user terminals. Another is mitigation of faded transmission through the use of a Rake receiver, which constructively combines multipath components rather than allowing them to destructively combine as in narrowband transmission. A third major benefit is soft handoff among multiple cell base station, which provides improved cell-boundary performance and d prevents dropped calls.

2.1.2 Principle of Spread Code Communication

CDMA uses unique spreading codes to spread the baseband data before transmission. The signal is transmitted in a channel, which is below noise level. The receiver then uses correlators to dispreads the wanted signal, which is passed through a narrow band pass filter. Unwanted signals will not be despread and will not pass through the filter. Codes are in the form of a carefully designed binary (one/zero) sequence produced at a much higher rate than that of the baseband data. The rate of a spreading code is referred to as chip rate rather than bit rate. Usually, in CDMA a spread code runs at a much higher rate than the data to be transmitted. Data for transmission is simply logically XOR added with the faster code.

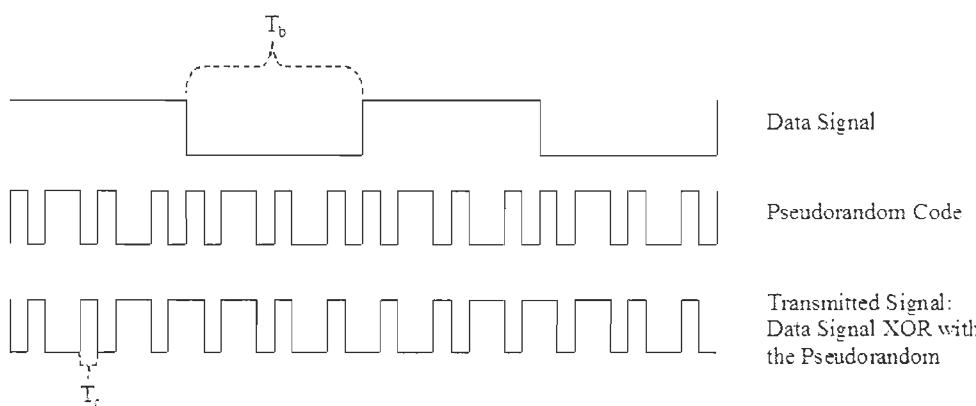


Figure 2.2 Generate spread spectrum signal [23]

The figure 2.2 shows how spread spectrum signal is generated. The data signal with pulse duration of T_b is XOR added with the code signal with pulse duration of T_c. Therefore, the bandwidth of the data signal is $1/T_b$ and the bandwidth of the spread spectrum signal is $1/T_c$. Since T_c is much smaller than T_b, the bandwidth of the spread spectrum signal is much larger than the bandwidth of the original signal. Each user in a CDMA system uses a different code to modulate their signal. Choosing the codes used to modulate the signal is very important in the performance of CDMA systems. The best performance will occur when there is good separation between the signal of a desired user and the signals of other users. The separation of the signals is made by correlating the received signal with the locally generated code of the desired user. If the signal matches the desired user's code then the correlation function will be high and the system can extract that signal. If the desired user's code has nothing in common with the signal the correlation should be as close to zero as possible (thus eliminating the signal); this is referred to as cross correlation. If the code is correlated with the signal at any time offset other than zero, the correlation should be as close to zero as possible. This is referred to as auto-correlation and is used to reject multi-path interference.

In general, CDMA belongs to two basic categories: synchronous (orthogonal codes) and asynchronous (pseudorandom codes). In this paper, we will discuss the asynchronous CDMA.

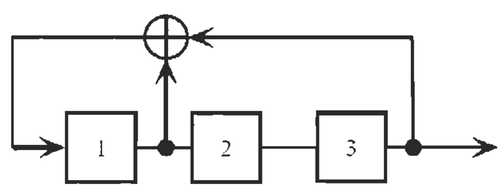
2.1.2.1 m-sequence in CDMA

For CDMA spreading code, we need a random sequence that passes certain "quality" criterion for randomness. These criterions are

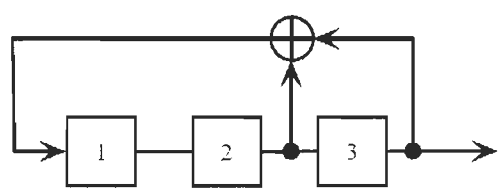
- The number of runs of 0's and 1's is equal. We want equal number of two 0's and 1's, a length of three 0's and 1's and four 0's and 1's etc. This property gives us a perfectly random sequence.

- There are equal number of runs of 0's and 1's. This ensures that the sequence is balanced.
- The periodic autocorrelation function (ACF) is nearly two valued with peaks at 0 shift and is zero elsewhere. This allows us to encrypt the signal effectively and using the ACF peak to demodulate quickly.

Maximum length sequence (m-sequence) meets all these three requirements strictly. We can generate this sequence by using a linear feedback registers. Which structure is shown in Figure 2.3.



3 stage LFSR generating m-sequence of period 7., using taps 1 and 3.



Another 3 stage LFSR generating m-sequence of period 7. using taps 2 and 3

Figure 2.3 The structure of linear feedback registers (LFSR) [24]

Each configuration of N registers produces one sequence of length 2^N. If taps are changed, a new sequence is produced in the same length. There are only a limited number of m-sequences of a particular size.

The cross correlation between an m-sequences and noise is low which is very useful in filtering out noise at the receiver. The cross correlation between any two different m-sequences is also low and is useful in providing both encryption and spreading. The low amount of cross-correlation is used by the receiver to discriminate among user signals generated by different m-sequences.

Think of m-sequence as a code applied to each message. Each letter (bit) of the message is changed by the code sequence. The spreading quality of the sequence is an added dimensionality and benefit in CDMA systems.

2.1.2.2 Gold Codes in CDMA

In this paper we discuss the asynchronous CDMA systems. So we need to learn the Gold code, which is used in this kind of system.

A Gold Code is a set of random sequences which was discovered in 1967, often called pseudo-noise (PN) sequences, which are statistically uncorrelated. Gold codes have three-valued autocorrelation and cross-correlation function with values $\{-1, -t(m), t(m)-2\}$, where

$$t(m) = \begin{cases} 2^{(m+1)/2} + 1 & for\ odd\ m \\ 2^{(m+2)/2} + 1 & for\ even\ m \end{cases} \tag{2.1}$$

All pairs of m-sequences do not yield Gold codes and those which yield Gold codes are called preferred pairs. Moreover, such preferred pairs can be used to construct a whole family of codes that have the same period as well as the same correlation property.

Gold codes are used in non-orthogonal CDMA as chipping sequences that allow several users to use the same frequency, resulting in less interference and better utilization of the available bandwidth.

We usually combine two m-sequence to get the Gold codes.

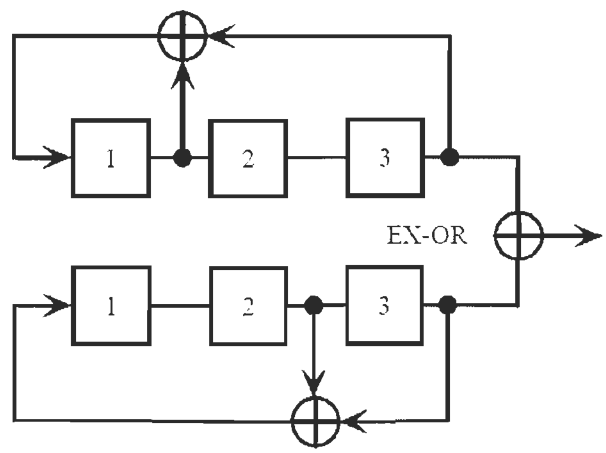


Figure 2.4 Generating Gold codes by combining two preferred pairs of m-sequences [25]

2.1.2.3 Long Code

In IS-95 and IS-2000 system, there is another kind of m-sequence which called long code. Its length is $2^{42} - 1$ bits. We can create it from a LFSR of 42 registers. It runs at 1.2288 Mb/s. The time it takes to recycle this length of code at this speed is 41.2 days.

Long code is used to both spread the signal and to encrypt it. A cyclically shifted version of the long code is generated by the cell phone during call setup. The shift is called the Long Code Mask and is unique to each phone call. CDMA networks have a security protocol called CAVE that requires a 64-bit authentication key, called A-key and the unique Electronic Serial Number (ESN). The network uses both of these to create a random number that is then used to create a mask for the long code used to encrypt and spread each phone call. This number, the long code mask is not fixed but changes each time a connection is created.

Here are another two concepts, public long code and private long code. The Public long code is used by the mobile to communicate with the base during the call setup phase. The private long code is one generated for each call then abandoned after the call is completed.

2.1.2.4 Short Code

Short code in CDMA is based on a m-sequence which length is $2^{15} - 1$. These codes are used for synchronization in the forward and reverse links and for cell/base station identification in the forward link. The short code repeats every 26.666 milliseconds. The sequences repeat 75 times in every 2 seconds. We want this sequence to be fairly short because during call setup, the mobile is looking for a short code and needs to be able find it fairly quickly. Two seconds is the maximum time that a mobile will need to find a base station.

A mobile is assigned a short code PN offset by the base station to which it is transmitting. The mobile adds the short code at the specified PN offset to its traffic message, so that the base station in the region knows that the particular message is meant for it. This is the way the primary base station is identified in a phone call. The base station keeps a list of nearby base stations and during handoff the mobile is notified of the change in the short code.

2.2 MIMO

2.2.1 Background

Multiple Input Multiple Output (MIMO) is an antenna technology, in which multiple antennas are used in both transmitter and receiver [5]. It is one of forms of smart antenna. The earliest idea in this field was mentioned by A.R. Kaye and D.A. George in 1970 and W. van Etten in 1975 and 1976. After then, several papers on beamforming related applications in 1984 and 1986 which were published by Jack Winters and Jack Salz at bell laboratories. Until 1993, Arogyaswami Paulraj and Thomas Kailath proposed the concept of Spatial Multiplexing using MIMO. It is the first time that this technology is used in wireless broadcast. In commercial arena, Iospan Wireless Inc. developed the first commercial system in 2001 which used MIMO-OFDMA technology. It supports both diversity coding and spatial multiplexing. By using MIMO technology we can minimize the errors and optimize data speed. At the same time, it does not need additional bandwidth or transmit power.

As a part of smart antenna technology, MIMO develop from SISO (Single Input Single Output), SIMO (Single Input Multiple Output), MISO (Multiple Input Single Output).

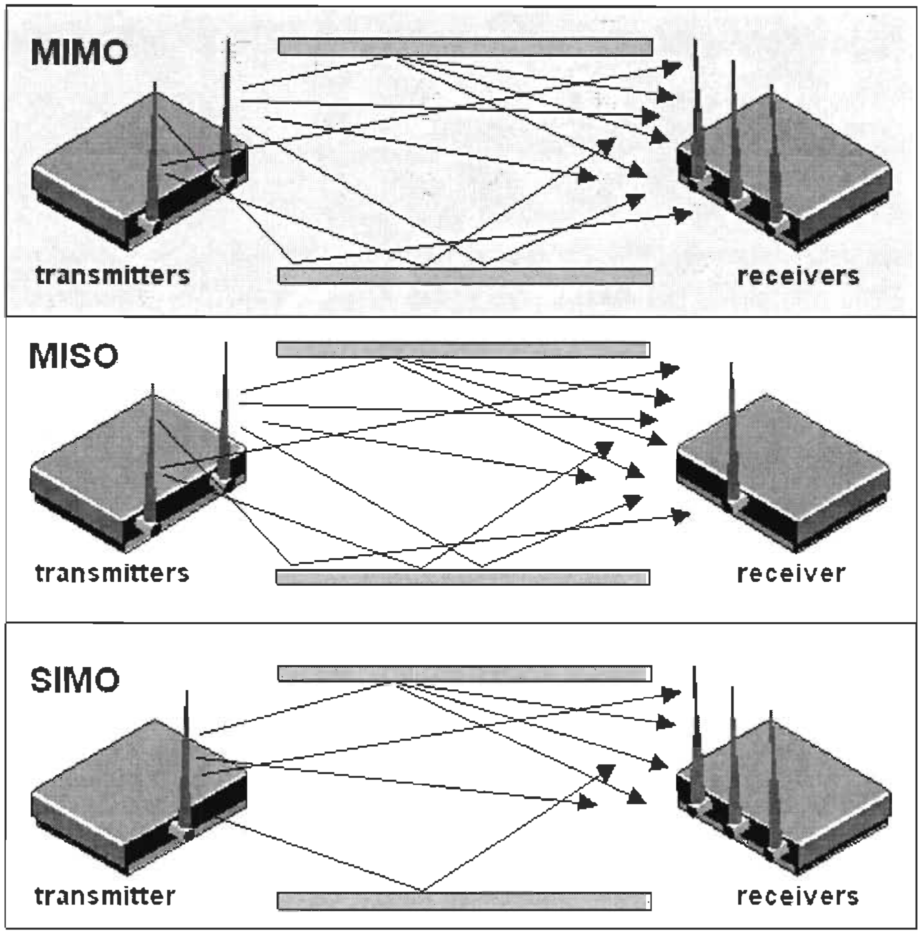


Figure 2.5 SISO to MIMO [26]

In traditional telecommunication system, we have 1 transmit antenna and 1 receive antenna. Most of the telecommunication system now is using this structure. Then, SIMO technology appears. Compare with single antenna technology, SIMO has the same data transmit rate but greater range. MISO is another kind of structure. Compare with SISO system, MISO system has the same data rate and same transmit range but more reliable. And in MIMO system, we can get greater data rate and greater range.

2.2.2 Benefits of MIMO

The benefits of MIMO technology that help achieve such significant gains as below [5]

- **Array gain**

Array gain is the increase in receive SNR that results from a coherent combing effect of the wireless signals at a receiver. The coherent combing may be realized through spatial processing at the receive antenna array or spatial pre-processing at the transmit antenna array. Array gain improves resistance to noise, thereby improving the coverage and the range of wireless network.

- **Spatial diversity gain**

Spatial diversity gain mitigates fading and is realized by providing the receiver with multiple copies of the transmitted signal in space, frequency or time. With an increasing number of independent copies, assume that at least one of the copies is not experiencing a deep fade increases, thereby improving the quality and reliability of reception. A MIMO channel with Nt transmitters and Nr receivers potentially offer NtNr independently fading links, and hence a spatial diversity order of NtNr.

- **Spatial multiplexing gain**

Using spatial multiplexing in MIMO system can increase data rate. For example, transmitting multiple, independently data streams within the bandwidth of operation. Under suitable channel conditions, in this kind of environment the receiver can separate the data streams. Furthermore, each data stream experiences at least the same channel quality that would be experienced by a SISO system, effectively enhancing the capacity by a multiplicative factor equal to the number of streams. In general, the number of data streams that can be reliably supported by a MIMO channel equals the minimum of the number of transmit antennas and the number of receive antenna. The spatial multiplexing gain increases the capacity of a wireless network.

- **Interference reduction and avoidance**

Interference in wireless networks results from multiple users sharing time and frequency resources. Interference may be mitigated in MIMO systems by exploiting the spatial dimension to increase the separation between users. For instance, in the presence of interference, array gain increase the tolerance to noise as well as the interference power, hence improving the SINR. Additionally, the spatial dimension may be leveraged for the purposes of interference avoidance. For example, directing signal energy forwards the intended used and minimizing interference to other users. Interference reduction and avoidance improve the coverage and range of wireless network.

In general, it may not be possible to exploit simultaneously all the benefits described above due to conflicting demands on the spatial degrees of freedom. However, using some combination of the benefits across a wireless network will result in improved capacity, coverage and reliability.

2.2.3 Capacity of MIMO

As we mentioned before, one of the benefits of MIMO system is the capacity increases [6]. When the system is SISO status, we could estimate its capacity by using *Shannon's Capacity Formula*:

$$C = B \cdot \log_2\left(1 + \frac{P}{N_0 B}\right) \tag{2.2}$$

In which B is bandwidth, P is transmitted signal power and N_o is single noise spectrum. Assumes the channel is White Gaussian. This formula gives an upper limit for the achieved error-free SISO transmission rate. If the transmission rate is less than C bits/sec(bps), then an appropriate coding scheme exits that could lead to reliable and error-free transmission. On the contrary, if the transmission rate is more than C bps, then the received signal, regardless of the robustness of employed code, will involve bit errors.

When we discuss the case in MIMO system, we can extend Shannon's Capacity Formula for MIMO. We consider an antenna array with n_t elements at the side of transmitter and an antenna array with n_r elements at the side of receiver. The impulse of the channel between the *j*th transmitter element and the *i*th receiver element is denoted as $h_{i,j}(\tau, t)$. The MIMO channel can then be described as below:

$$H(\tau, t) = \begin{bmatrix} h_{1,1}(\tau,t) & h_{1,2}(\tau,t) & \cdots & h_{1,n_t}(\tau,t) \\ h_{2,1}(\tau,t) & h_{2,2}(\tau,t) & & h_{2,n_t}(\tau,t) \\ \vdots & & \ddots & \vdots \\ h_{n_r,1}(\tau,t) & h_{n_r,2}(\tau,t) & \cdots & h_{M_R,n_t}(\tau,t) \end{bmatrix} \tag{2.3}$$

The matrix elements are complex numbers that correspond to the attenuation and phase shift that the wireless channel introduces to the signal reaching the receiver with delay τ. The input-output notation of the MIMO system can now be expressed by the following equation:

$$y(t) = H(\tau, t) \otimes s(t) + u(t) \tag{2.4}$$

Where $\otimes$ denotes convolution, s(t) is corresponding to the n_t transmitted signals, y(t) is corresponding to the n_r received signals and u(t) is the additive white noise. If the transmitted signal bandwidth is narrow enough that the channel response can be seen as flat across frequency, then the discrete time description corresponding to equation (2.3) is

$$r_\tau = Hs_\tau + u_\tau \tag{2.5}$$

Then the capacity of MIMO channel can be estimated by the following equation

$$C = \max_{tr(R_{SS}) \leq p} log_2[det(I + HR_{SS}H^H)] \tag{2.6}$$

Where H is the channel matrix, R_{SS} is the covariance matrix of the transmitted vector s, H^H is the transpose conjugate of the H matrix and p is the maximum normalized transmit power. C is the estimation of capacity for MIMO channel.

2.2.4 MIMO Channel

MIMO channels arise in wireless communications environment where multiple transmit and receive antennas are used. MIMO channels work best in highly scattering transmission environment, where multiple paths exist between transmitters and receivers. Multipath is the propagation phenomenon which causes by atmospheric ducting, ionosphere reflection and refraction, and reflection from terrestrial objects, such as mountains and buildings.

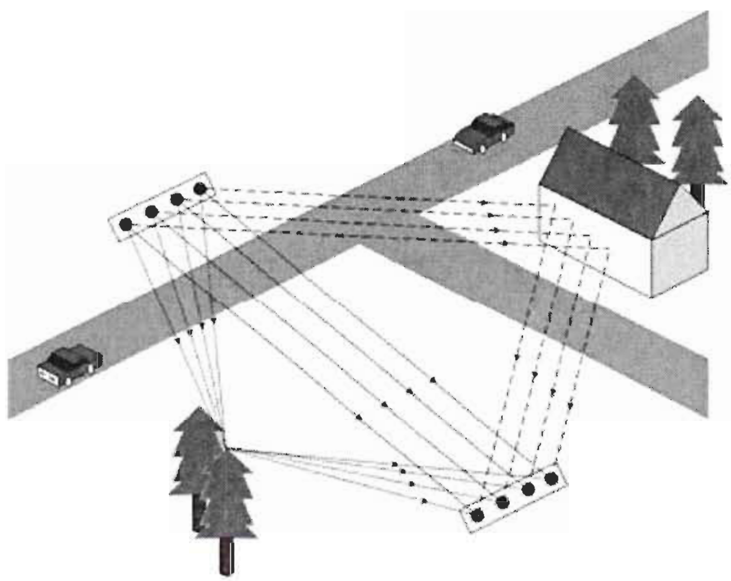


Figure 2.6 MIMO channels [27]

In engineering, we usually describe MIMO system as follow

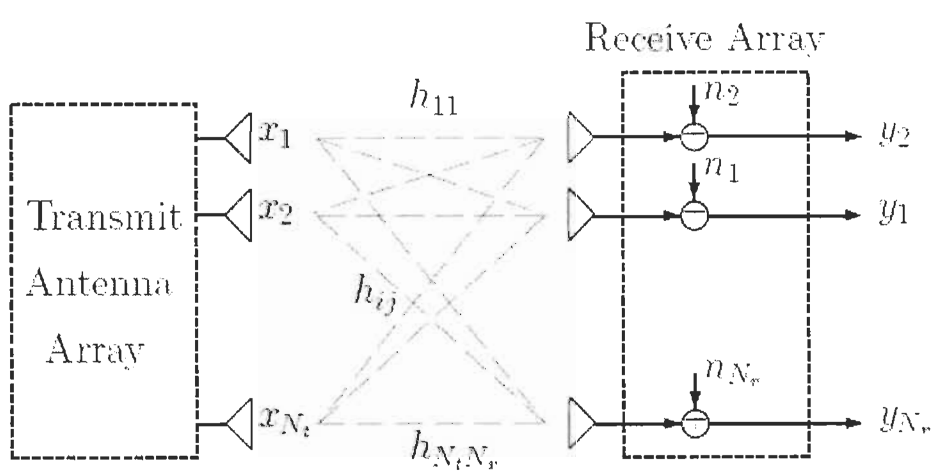


Figure 2.7 Block Diagram of MIMO channel [28]

In a MIMO system, the effects of multipath include constructive and destructive interference, and phase shifting of the signal. It may cause ISI which will affect signal transmission quality.

2.3 Rake Receiver

At the side of the receiver, we have to use some technique to correct the ISI. It is very important for the whole communication system. Rake receive is one of the choices which uses in CDMA based system. A Rake receiver is a radio receiver designed to counter the effects of multipath fading. It does this by using several "sub-receivers" called fingers, that is, several correlators each assigned to a different multipath component. Each finger independently decodes a single multipath component; at a later stage the contribution of all fingers are combined in order to make the most use of the different transmission characteristics of each transmission path. It could well result in higher Signal to Noise Ratio (SNR) environment.

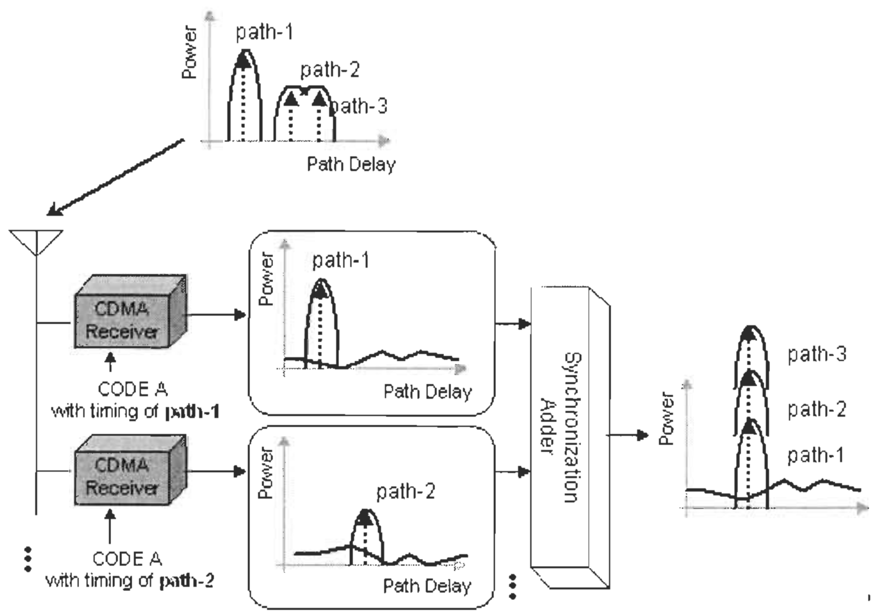


Figure 2.8 Rake receiver [29].

The multipath channel through which a signal transmits can be seen as transmitting the original signal plus a number of multipath components. Multipath components are delayed copies of the original transmitted signal traveling through a different echo path, each with a different magnitude and time-of-arrival at the receiver. Since each component contains the original information, after the process of channel estimation all the components can be added coherently to improve the information reliability.

3. Channel Estimation Algorithms

3.1 introductions

The radio channel in mobile radio systems are usually causing intersymbol interference (ISI) and enhances multiple access interferences (MAI) in the received signal [9]. In order to remove ISI and MAI from the signal, many kinds of equalizers can be used. Indirect detection algorithms require knowledge of the channel impulse response (CIR), which can be provided by a separate channel estimator. Usually the channel estimation is based on the known sequence of bits, which is unique for a certain transmitter and which is repeated in every transmission burst. Thus, the channel estimator is able to estimate CIR for each burst separately by exploiting the known transmitted bits and the corresponding received samples.

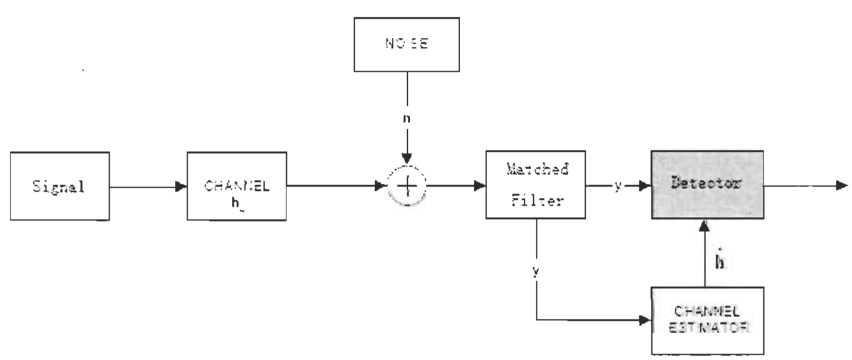


Figure 3.1 Layout of the channel estimation [30]

The figure 3.1 above is the layout of the channel estimation in a CDMA system. We can see the signal is transmitted over a channel, which we assume as a fading multiple channel. Thermal noise is generated at the side of receiver. A detector is used to detect the original signal from the received signal. At the same time, the detector also needs the channel estimate $\hat{h}$ from a specific channel estimator. The received signal y can be expressed as follows

$$y = Mh + n \tag{3.1}$$

Where the complex channel impulse response h of the wanted signal is expressed as below

$$h = [h_0 \quad h_1 \quad \cdots \quad h_L]^T \tag{3.2}$$

n denotes the noise samples.

Generally speaking, four kind of methods are used in channel estimation algorithm. Which are Blind Channel Estimation, Direct Technique (Such as Adaptive Channel Estimation), Indirect Technique (Such as Matrix Inverse method) and Mixed Estimation (Such as the layout showed in Fig. 3.1, detector and estimator work together). In this article, we will focus on the Direct Technique and the Indirect Technique. This means that we assume we have already known the original signal at the side of the receiver. By using this information we do channel estimation to get the $\hat{h}$.

3.2 Indirect Techniques

3.2.1 Least-squares (LS) channel estimation

Least squares can be interpreted as a method of fitting data. The best fit in the least-squares sense is that instance of the model for which the sum of squared residuals has its least value, a residual being the difference between an observed value and the value given by the model. The method was first described by Carl Friedrich Gauss around 1794.

According to the equation (3.1) and (3.2), we can get that

$$\hat{h} = \arg\min_h ||y - Mh||^2 \quad (3.3)$$

Assuming we've got the noise, we get

$$\hat{h}_{LS} = (M^H M)^{-1} M^H y \quad (3.4)$$

Where $()^H$ and $()^{-1}$ denote the Hermitian and inverse matrices, respectively.

3.2.2 Blind Channel Estimation for MIMO-CDMA using first-order statistics

Blind channel estimation algorithm is another important estimation algorithm which receives considerable interest recently [10]. People believe that this kind of algorithm has potential to increase the system throughput significantly.

We assume that we have a Nt transmit antenna and Nr receive antenna system. At the side of transmitter, each symbol is spread by aperiodic code vector $c_i(k)$ with spreading gain P, followed by a chip pulse-shaping filter. At the side of receiver, signals are passed through a chip-matched filter and sampled at the chip rate. So we can get channel model $h_{ji}(k), k = 1, \ldots L_{ji}$,

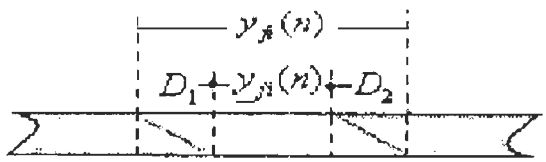


Figure 3.2 Channel response [31]

We get the received signal as below,

$$y_{ji}(n) = \begin{bmatrix} c_i(nT+1) & & \\ \vdots & c_i(nT+1) & \\ c_i(nT+P) & \vdots & c_i(nT+1) \\ & c_i(nT+P) & \vdots \\ & & c_i(nT+P) \end{bmatrix} \begin{bmatrix} h_{ji}(l) \\ \vdots \\ h_{ji}(L_{ji}) \end{bmatrix} b_i(n) \tag{3.5}$$

From figure 3.2 we can see that D1 and D2 are overlapped zones corrupted by data from the previous and following symbol. We consider only the part without intersymbol interference, we get

$$\underline{y}_{ji}(n) = \begin{bmatrix} c_i(nT+L_{ji}) & \cdots & c_i(nT+1) \\ c_i(nT+L_{ji}+1) & & \\ \vdots & \ddots & \vdots \\ c_i(nT+P) & \cdots & c_i(nT+P-L_{ji}+1) \end{bmatrix} \begin{bmatrix} h_{ji}(l) \\ \vdots \\ h_{ji}(L_{ji}) \end{bmatrix} b_i(n) =$$

$$C_i(n) h_{ji} b_i(n) \tag{3.6}$$

So we can write the received signals as following

$$y_j(n) = \sum_{i=1}^{Nt} C_i(n) h_{ji} b_i(n) = C(n) \cdot diag(h_{j,i}, \cdots, h_{j,Nt}) \cdot b(n) = C(n) H_j b(n) \tag{3.7}$$

$$C(n) \triangleq [C_1(n), \cdots C_{Nt}(n)] \tag{3.8}$$

$$b(n) \triangleq [b_1(n), \cdots b_{Nt}(n)]^T \tag{3.9}$$

Where matrix H_j is a block diagonal with h_{ji} as the ith block. We can get the signal at the side of receiver which is

$$Y_j = diag\big(C(n), \cdots, C(n+m-1)\big) \cdot \big(I_m \otimes H_j\big)b + w = C\underline{H_j}\,\mathrm{b}+w_j$$

$$b = [b^T(n), \cdots, b^T(n+m-1)]^T \tag{3.10}$$

Where w_j represents the additive Gaussian noise. C is block diagonal matrix with C(n) as its diagonal block. $I_m \otimes H_j$ is the Kronecker product. When we consider about Nr receive antenna. We get

$$Y = \begin{bmatrix} Y_1 \\ \vdots \\ Y_{Nr} \end{bmatrix} = \begin{bmatrix} C & & \\ & \ddots & \\ & & C \end{bmatrix} \begin{bmatrix} \underline{H_1} \\ \vdots \\ \underline{H_{Nr}} \end{bmatrix} b + w = \mathsf{C}\mathcal{H}b + w \tag{3.11}$$

Then, we can get output of the matched filter $\mathsf{C}^{\dagger}$ as

$$t = \mathsf{C}^{\dagger}Y = \mathcal{H}b + n = \begin{bmatrix} \underline{H_1} \\ \vdots \\ \underline{H_{Nr}} \end{bmatrix} b + v \tag{3.12}$$

Where $v = \mathsf{C}^{\dagger}w$ is the colored noise vector. $\mathsf{C} = (I_{Nr} \otimes C)$, so we have

$$\mathsf{C}^{\dagger} = (I_{Nr} \otimes C)^{\dagger} = ({I^{-1}}_{Nr} \otimes C) = I_{Nr} \otimes C^{\dagger} \tag{3.13}$$

Where

$$t_{jn} = H_j b(n) + v_{jn} \quad j = 1, \cdots, Nr. \;\; n = 1, \cdots, m \tag{3.14}$$

$$t_{ji}^{(n)} = b_i(n) \cdot h_{ji} + v_{ji}(n) \quad i = 1, \cdots, Nt \tag{3.15}$$

Because $b_i(n)$ is generated from the finite-alphabet set {-1,+1} and $v_{ji}(n)$ is zero-mean additive noise, the set of t_{jn} can be classified into two sets as below

$$S_1\colon\{h_{ji} + v_{ji}(n)\} \tag{3.16}$$

$$S_2\colon\{-h_{ji} + v_{ji}(n)\} \tag{3.17}$$

The centers of the two sets can be found by using such as K-means clustering.

$$c_1 = center(S_1) \tag{3.18}$$

$$c_2 = center(S_2) \tag{3.19}$$

Assume that for m synbols there are p points belonging to S_1 and q points belonging to S_2,

We have

$$\hat{h}_{ji} = \frac{p \cdot c_1 - q \cdot c_2}{m} \tag{3.20}$$

3.2.3 Blind Multipath Estimation with Toeplitz displacement for long code DS-CDMA

According to [11], the baseband representation of the received signal after coherent reception is given by

$$x(t) = \sum_{n=-\infty}^{\infty} \sum_{K=1}^{K} A_k c_k^{-n}(t - nT - \tau_k) b_k(n) + w(t) \tag{3.21}$$

Where w(t) is the additive and circularly symmetric Gaussian noise process with variance σ_w^2, and A_k is the amplitude of the signal for user k.

$$c_k^n(t) = \sum_{l=1}^{N} c_k^n(l) \psi(t - lT_c) \tag{3.22}$$

Where $\psi(t)$ is the shape of the chip with a duration T_c .

We use M matched filters per received symbol to fully exploit the properties of the DS-CDMA signals.

$$y(n) = S_1(n)x(n) = S_1(n)\left(\sum_{k=1}^{K} A_p C_k(n) H_k b_k(n)\right) + S_1(n)w(n) \tag{3.23}$$

Where the matched filtering matrix $S_1(n)$ is presented by

$$S_1^T(n) = \begin{bmatrix} C_{1,M}^1(n+N) & \cdots & 0 \\ C_{1,M}^2(n+N) & & \\ \vdots & \ddots & \vdots \\ 0 & \cdots & C_{1,M}^1(n+aN) \\ & & C_{1,M}^2(n+aN) \end{bmatrix} \tag{3.24}$$

The matrix $S_1^T(n)$ and $C_1(n)$ are related by

$$C_1(n) = \begin{bmatrix} C_{1,M}^2(n) & \vdots & 0 \\ 0 & S_1^T(n) & 0 \\ 0 & \vdots & \tilde{C}_{1,M}^1(n+(a+1)N) \end{bmatrix} \tag{3.25}$$

Let us consider the $aM \times 1$ matched filter output vector $y(n)$, the covariance matrix of this observation vector is obtained as

$$R_y(n) = E_{b,w}[y(n)y^H(n)] = \sigma_1^2 S_1(n)C_1(n)H_1H_1^H(n)S_1^T(n) + R_I(n) + R_w(n) \tag{3.26}$$

Where $\sigma_1^2 = A_1^2 E\{b_1^2(n)\}$, $R_w(n) = \sigma_w^2 S_1(n)S_1^T(n)$ is noise autocorrelation matrix, and the contribution of other user's interference is

$$\mathrm{R_I(n)} = \sum_{\mathrm{k}=1}^{\mathrm{K}} \sigma_{\mathrm{k}}^2 \mathrm{S_1(n)C_k(n)H_kH_k^H C_k^H(n)S_1^T(n)} \tag{3.27}$$

We introduce the Toeplitz displacement method here to remove the effects of the channel noise and interference. Let us define

$$SC_1 = \lim_{N_s \to \infty} \frac{1}{N_S} \sum_{n=1}^{N_S} S_1(n) C_1(n) \tag{3.28}$$

Where N_S is the number of transmitted symbols. Then we can get

$$R_h = R_y(2{:}aM, 2{:}aM) - R_y(1{:}aM-1, 1{:}aM-1) = R_y^+ - R_y^- =$$

$$\sigma_1^2 SC_1^+ H_1 H_1^H SC_1^{+H} - \sigma_1^2 SC_1^- H_1 H_1^H SC_1^{-H} \tag{3.29}$$

SC_1^+ and SC_1^- are formed by removing the first and the last row of SC1.

Then we can get

$$\hat{R}_h(n) = \hat{R}_y^+(n) - \hat{R}_y^-(n) = \hat{R}_y(n)(2{:}aM, 2{:}aM) - \hat{R}_y(n)(1{:}aM-1, 1{:}aM-1)$$

(3.30)

We present the estimation error matrix to be

$$E_h(n) = R_h(n) - \hat{R}_h(n) = \sigma_1^2 SC_1^+ H_1 H_1^H SC_1^{+H} - \sigma_1^2 SC_1^- H_1 H_1^H SC_1^{-H} - \hat{R}_h(n) \tag{3.31}$$

The estimation error can be defined with squared Frobenius norm of E_h

$$J(n) = \|E_h(n)\|_F^2 = tr[E_h(n) E_h^H(n)] \tag{3.32}$$

The cost function can be built as the cumulative error

$$J = \frac{1}{N_s} \sum_{n=1}^{N_s} J(n) = \frac{1}{N_s} \sum_{n=1}^{N_s} tr[E_h(n) E_h^H(n)] = \frac{1}{N_s} \sum_{n=1}^{N_s} vec^H[E_h(n)] vec[E_h(n)]$$

(3.33)

The channel parameters can be obtained by minimizing this cost function. In practice, the average correlation matrix $\hat{R}_y$ is sampled and formed by

$$\hat{R}_y = \frac{1}{N_s} \sum_{n=1}^{N_s} \hat{R}_y = \frac{1}{N_s} \sum_{n=1}^{N_s} y(n) y^H(n) \tag{3.34}$$

The estimated $\hat{R}_h$ can be formed. We define new unknown by

$$D_1 = \sigma_1^2 H_1 H_1^H \tag{3.35}$$

The error matrix (12) becomes

$$E_h(n) = SC_1^+ D_1 SC_1^{+H} - SC_1^- D_1 SC_1^{-H} - \hat{R}_h(n) \tag{3.36}$$

And

$$vec\big(E_h(n)\big) = (SC_1^{+*} \otimes SC_1^+ - SC_1^{-*} \otimes SC_1^-)vec(D_1) - vec(\hat{R}_h(n)) \tag{3.37}$$

Let

$$d_1 = vec(D_1) \tag{3.38}$$

$$Q = (SC_1^{+*} \otimes SC_1^+ - SC_1^{-*} \otimes SC_1^-) \tag{3.39}$$

We have

$$J(n) = \{Qd_1 - vec(\hat{R}_h(n))\}^H \{Qd_1 - vec(\hat{R}_h(n))\} \tag{3.40}$$

Therefore, our cost function becomes

$$J = \frac{1}{N_s}\sum_{n=1}^{N_s}\{Qd_1 - vec(\hat{R}_h(n))\}^H \{Qd_1 - vec(\hat{R}_h(n))\} \tag{3.41}$$

The following LMS type recursion can be formulated for d_1 with step size μ

$$d_1^{(n+1)} = d_1^{(n)} - \mu\nabla_{\mathrm{d}_1^{\mathrm{H}}}\mathrm{J(n)} \tag{3.42}$$

Where $\mu\nabla_{\mathrm{d}_1^{\mathrm{H}}}\mathrm{J(n)} = Q^H Q d_1^{(n)} - Q^H vec\big[\hat{R}_h(n)\big]$ (3.43)

So we can get

$$d_1^{(n+1)} = d_1^{(n)} - \mu Q^H Q d_1^{(n)} + \mu Q^H vec\big[\hat{R}_h(n)\big] \tag{3.44}$$

3.2.4 EM Vector Channel Estimation

In [12], the system consists of G users. There are Nt antennas in the transmitters of wach mobile user and Nr antennas in the receiver of the base station through a correlated multipath Rayleigh fading channel, so that the total number of transmitting antennas from all users is K=G*Nt. The considered MIMO-CDMA system is over a correlated multipath fading channel. At each time interval [lT, (l+1)T], the ith antenna of gth user generates a symbol $b_{g,j}(l)$ from the symbol set {+1.-1}th equal probability. The each symbol is spread by the aperiodic spreading sequence

$$s_{l,g}(t) = \sum_{n=0}^{N-1} \frac{1}{\sqrt{N}} c_{lN+n,g} \psi(t - lT - nT_c) \tag{3.45}$$

Where $c_{lN+n,g}$ here is the quadratic-phase random signature sequence for user g, and $\psi(t)$ is the chip waveform with duration $[0, T_c]$. The spreading waveform are normalized. The transmitted signal of the gth user's the ith antenna is

$$v_g(t) = \sum_{l=-\infty}^{\infty} b_{g,i}(l) s_{l,g}(t) \tag{3.46}$$

Where $b_{g,i}(l)$ is the data symbol of the gth user's the ith antenna at symbol index l.

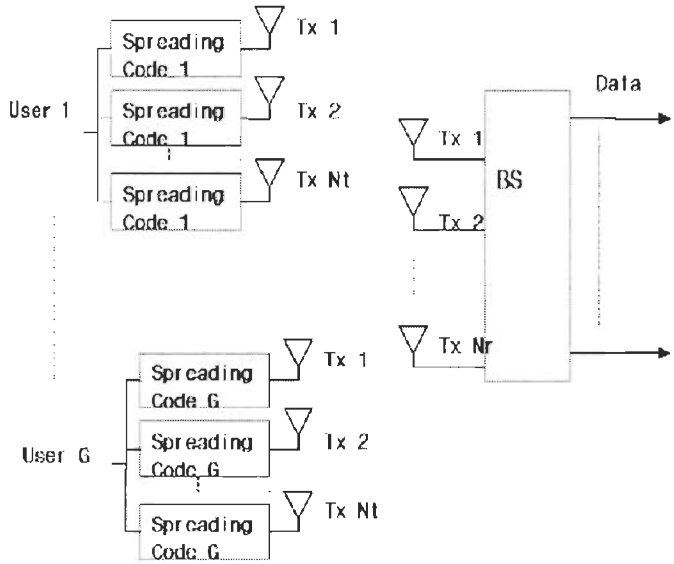


Figure 3.3 Multiuser MIMO-CDMA system block diagram [32]

$$r_j(t) = \sum_{i=1}^{Nt} \sum_{m=1}^{M} \sum_{g=1}^{G} \sum_{l=-\infty}^{\infty} b_{g,i}(l) h_{g,j,i}^{(m)}(l) s_{l,g}\left(t - \tau_{m,g}\right) + n(t) \qquad (3.47)$$

M is the number of the paths. N(t) is additive white noise.

We define the channel vector for the gth user's ith transmitter antenna in the jth received antenna as

$$h(l) = (h^{(1)}(l) \quad h^{(2)}(l) \quad \cdots \quad h^{(M)}(l))^T \qquad (3.48)$$

a First-Order Vector AutoRegressive (FOVAR) model is used to approximate the vector channel process and includes the state equation

$$h(l) = \Phi h(l-1) + w \qquad (3.49)$$

$$\tilde{y}(l) = b(l)h(l) + n(l) \qquad (3.50)$$

$\tilde{y}$ is the generalized MMSE multiuser detector output, Φ is the autoregressive matrix, w is the FOVAR (First-Order Vector AutoRegressive) model noise which is assumed to be a zero-mean complex white Gaussian process with covariance matrix

$$Q = E\{w \quad w^H\} \tag{3.51}$$

The measurement noise vector

$$n(l) = \left(n^{(1)}(l) \quad n^{(2)}(l) \quad \cdots \quad n^{(M)}(l)\right)^T \tag{3.52}$$

We can get GM-dimensional vector

$$y_j(l) = \left(y_{1,j}(l), \dots, y_{G,j}(l)\right)^T = R(l)d_j(l) + n_j(l) \tag{3.53}$$

Then we can get at jth receiver

$$y_{g,j}(l) = \left(y_{g,j}^{(1)}(l) \quad y_{g,j}^{(2)}(l) \quad \cdots \quad y_{g,j}^{(M)}(l)\right)^T \tag{3.54}$$

Where

$$y_{g,j}^{(m)}(l) = \int_{lT+\tau_m}^{(l+1)T+\tau_m} r_j(t)s_g^*(t-\tau_m)dt \tag{3.55}$$

Where R(l) is the cross-correlation matrix.

$R =$

$$\begin{bmatrix} \rho_{11,11} & \rho_{12,11} & \rho_{13,11} & & & \rho_{G(M-1),11} & \rho_{GM,11} \\ \rho_{11,12} & \rho_{12,12} & \rho_{13,12} & \cdots & & & \rho_{GM,12} \\ \rho_{11,13} & \rho_{12,13} & \rho_{13,13} & & & & \\ & \vdots & & \rho_{gm,g'm'} & & \vdots & \\ & & & & \rho_{G(M-2),G(M-2)} & \rho_{G(M-1),G(M-2)} & \rho_{GM,G(M-2)} \\ \rho_{11,G(M-1)} & & & \cdots & \rho_{G(M-2),G(M-1)} & \rho_{G(M-1),G(M-1)} & \rho_{GM,G(M-1)} \\ \rho_{11,GM} & \rho_{12,GM} & \cdots & & \rho_{G(M-2),GM} & \rho_{G(M-1),GM} & \rho_{GM,GM} \end{bmatrix}$$

(3.56)

$\rho_{gm,g'm'}$ is the cross-correlation of the symbols of the gth user, the mth path and the g'th user, the m'th path.

$$\rho_{gm,g'm'} = \int_0^T S_g(t - m\tau_{m,g})S_{g'}^*(t - m'\tau_{m',g'})dt \quad (3.57)$$

$\tau_{m,g}$ is the path delay of gth user, the mth path.

$$d_j = \left[d_{1,j}^{(1)}, d_{1,j}^{(2)}, \dots, d_{1,j}^{(M)}, \dots, d_{G,j}^{(1)}, d_{G,j}^{(2)}, \dots, d_{G,j}^{(M)}\right]^T \quad (3.58)$$

Which denote the MG-dimensional column vector.

$d_{g,j}^{(m)} = \sum_{i=1}^{Nt} h_{g,j,i}^{(m)} b_{g,i}$ represents the all of transmitter antenna for the gth user by the mth path at jth receiver.

n_j is MG-dimensional AWGN noise vector with variance matrix $\sigma^2 I$.

So we use MMSE for all users' data and MMSE output are denote by MG-dimensional column vector

$$\tilde{y}_j = \left(R + \frac{1}{SNR} I_{MG\times MG}\right)^{-1} y_j \quad (3.59)$$

Where SNR is signal to noise ratio.

3.2.5 Reduced-Rank Space-time Channel Estimation for DS-CDMA

In [13], consider a asynchronous DS-CDMA system with K-users. The transmitted baseband signal model can be written as follow

$$x_k(t) = A_k \sum c_k(j)\varphi(t - jT_c) \quad (3.60)$$

Where T_c is the chip interval. $\{c_k(j)\}$ is the spreading sequence of user k. $\varphi(t)$ is normalized chip wave form of duration T_c. Then, the baseband multipath channel between the transmitter of user k and the receiver of the base-station can be written as a SISO channel as follow

$$\underline{h}(t) = \sum_{l=1}^{L_p} \underline{\phi}(\theta_{kl})\beta_{kl}\delta(t-\tau_{kl}) \quad (3.61)$$

Where L_p is the number of paths in each user's channel.

Let's consider the single user case first.

$$y(n) = Hx(n) + n(n) \quad (3.62)$$

Where $H = [h(0), h(T_c), \dots, h((L_c-1)T_c)]$ is the $L_r \times L_c$ space-time channel matrix. And n(n) is temporal white zero-mean Gaussian noise with covariance R_n.

$$H = \left[\underline{\phi}(\theta_1) \cdots \underline{\phi}\left(\theta_{L_p}\right)\right]\begin{bmatrix}\beta_1 & \cdots & 0\\ \vdots & \ddots & \vdots \\ 0 & \cdots & \beta_{L_p}\end{bmatrix}\begin{bmatrix} g^T(\tau_1) \\ \vdots \\ g^T(\tau_{L_p})\end{bmatrix} = \Phi\left(\underline{\phi}\right) diag\left(\underline{\beta}\right) G^T(\underline{\tau}) \quad (3.63)$$

After collect N samples, we get

$$Y = HX + N \quad (3.64)$$

The unconstrained maximum likelihood estimate (MLE) of the channel is found to be

$$\widehat{H}_{ML} = R_{yx}R_{xx}^{-1} \quad (3.65)$$

Where

$$R_{yx} = \frac{1}{N}\sum_{n=1}^{N} y(n)x^H(n) = \frac{1}{N}YX^H \quad (3.66)$$

$$R_{xx} = \frac{1}{N}\sum_{n=1}^{N} x(n)x^H(n) = \frac{1}{N}XX^H \quad (3.67)$$

Let $\tilde{y} = vec(Y), \tilde{h} = vec(H), and\ \tilde{n} = vec(N)$.

Where vec(·) denotes the operator of stacking columns.

$$\Delta\, \tilde{h}_{ML} = \mathrm{vec}\left(\widehat{H}_{ML} - H\right) = \frac{1}{N}((R_{xx}^{-1}X)^{*} \otimes I_{L_r})\tilde{n} \tag{3.68}$$

Where X^{*} denotes the conjugate of X. $\otimes$ is Kronecker product.

We can get

$$\mathrm{Cov}\{\Delta\, \tilde{h}_{ML}\} = \frac{1}{N} R_{xx}^{*-1} \otimes R_n \tag{3.69}$$

The MSE is given by

$$\mathrm{MSE}_{ML} = \frac{1}{N}\mathrm{tr}\{R_{xx}^{-1}\}\mathrm{tr}\{R_n\} \tag{3.70}$$

Let $\hat{y}(n) = \widehat{H}x(n)$ be an estimate of y(n) from x(n).

The correlation matrix of the error vector $z(n) = y(n) - \hat{y}(n)$ can be represented as

$$R_{zz} = E\{z(n)z(n)^{H}\} = R_{yy} - R_{yx}\widehat{H}^{H} - \widehat{H}R_{yx}^{H} + \widehat{H}R_{xx}\widehat{H}^{H} \tag{3.71}$$

We assume that R_{yy} and R_{xx} are nonsingular. The optimum choice of the estimate $\widehat{H}$ depends on the measure applied to R_{zz}.

The Reduced-Rank Minimum Mean Square Error (RRMMSE) estimate of H, denotes by $\widehat{H}_{RRMMSE}$.

$$\mathrm{tr}\{R_{zz}\} = \mathrm{tr}\left\{\left(AB^{H}R_{xx}B - R_{yx}B\right)(B^{H}R_{xx}B)^{-1}(AB^{H}R_{xx}B - R_{yx}B)^{H}\right\} + f(B) =$$
$$\mathrm{tr}\left\{R_{xx}^{-1}\left(R_{xx}BA^{H}A - R_{yx}^{H}A\right)(A^{H}A)^{-1}(R_{xx}BA^{H}A - R_{yx}^{H}A)^{H}\right\} + g(A) \tag{3.72}$$

where

$$f(B) = \mathrm{tr}\{R_{yy} - R_{yx}B(B^{H}R_{xx}B)^{-1}B^{H}R_{yx}^{H}\} \tag{3.73}$$

$$g(A) = \mathrm{tr}\{R_{yy} - R_{xx}^{-1}R_{yx}^{H}A(A^{H}A)^{-1}A^{H}R_{yx}\} \tag{3.74}$$

according to the two equations above, we can minimize $tr\{R_{zz}\}$ with respect to A and B alternately, and we can get

$$A(i+1) = R_{yx}B(i)(B^H R_{xx} B(i))^{-1} \tag{3.75}$$

$$B(i+1) = R_{xx}^{-1} R_{yx}^H A(i+1)(A(i+1)^H A(i+1))^{-1} \tag{3.76}$$

Let define $R_{tr} \triangleq R_{yx} R_{xx}^{-H/2}$, which has the singular value decomposition as

$$R_{tr} = \sum_{i=1}^{\min(L_r, L_t)} \sigma_i u_i v_i^H = U_{tr,1}\Sigma_{tr,1}V_{tr,1}^H + U_{tr,2}\Sigma_{tr,2}V_{tr,2}^H \tag{3.77}$$

Then we can get

$$A(i)B(i)^H \to R_{yx} R_{xx}^{-H/2} V_{tr,1} V_{tr,1}^H R_{xx}^{-1/2} = U_{tr,1} U_{tr,1}^H R_{yx} R_{xx}^{-1} \tag{3.78}$$

Upon convergence, we can get

$$A(i) = R_{yx} R_{xx}^{-H/2} V_{tr,1} Q \tag{3.79}$$

$$B(i) = R_{xx}^{-H/2} V_{tr,1} Q^{-H} \tag{3.80}$$

Let's define

$$R_{det} \triangleq R_{yy}^{-1/2} R_{yx} R_{xx}^{-H/2} \tag{3.81}$$

Then we can get

$$\hat{H}_{RRML} = R_{yx} R_{xx}^{-H/2} V_{det,1} V_{det,1}^H R_{xx}^{-1/2} = R_{yy}^{1/2} U_{det,1} U_{det,1}^H R_{yy}^{-1/2} R_{yx} R_{xx}^{-1} \tag{3.82}$$

3.2.6 Linear MMSE channel estimation

The linear MMSE algorithm [14] compute a matrix W, which is chosen to minimize the mean square error $E\{\|h - W^{*}r\|^{2}\}$.

$$W_{MMSE} = arg\,\min_{W} E\{\|h - W^{*}r\|^{2}\} = R^{-1}\Phi \tag{3.83}$$

$$R = E[rr^{*}] = E[(Ah+n)(Ah+n)^{*}] = APA^{*} + N_{0}I \tag{3.84}$$

$$\Phi = E[rh^{*}] = E[(Ah+n)h^{*}] = AP \tag{3.85}$$

$$P = E[hh^{*}] = diag(P_{1,1}, \cdots, P_{k,l}, \cdots, P_{K,L_K}) \tag{3.86}$$

$$\hat{h}^{LMMSE} = W_{MMSE}^{*} r = \Phi^{*}R^{-1}r = P^{*}\hat{A}^{*}(\hat{A}P\hat{A}^{*} + N_{0}I)^{-1}r \tag{3.87}$$

3.3 Direct Techniques

3.3.1 Least-Mean-Squares (LMS) Channel Estimation

The LMS algorithm is an important member of the family of stochastic gradient algorithms [15]. The term stochastic gradient is intended to distinguish the LMS algorithm from the method of steepest descent, which uses a deterministic gradient in a recursive computation of the Wiener filter for stochastic inputs. A significant feature of the LMS algorithm is its simplicity. Moreover, it does not require measurements of the pertinent correlation functions, nor does it require matrix inversion.

LMS algorithm is a linear adaptive filtering algorithm which generally consists of two basic processes:

1. A filtering process, which involves (a) computing the output of a linear filter in response to an input signal and (b) generating an estimation error by comparing this output with a desired response.
2. An adaptive process, which involves the automatic adjustment of the parameters of the filter in accordance with the estimation error.

The combination of these two processes working together constitutes a feedback loop, as shown below.

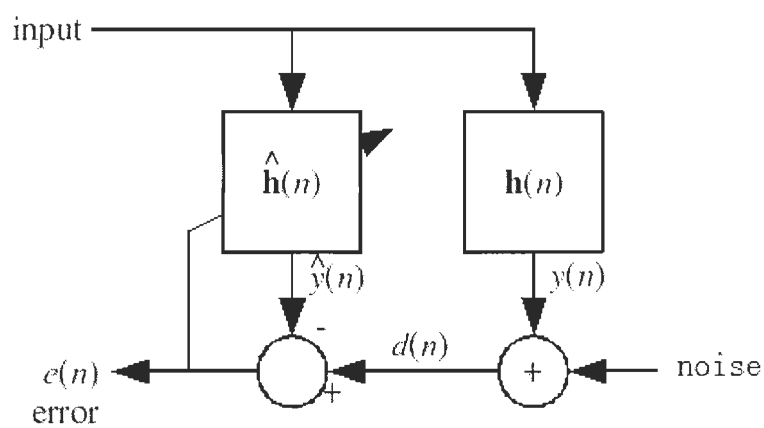


Figure 3.4 Block Diagram of adaptive filter [33]

In this paper, we consider multipath situation in both SISO and MIMO system. And we assume the delay taps as 6. Which means the filter order of all adaptive filters which are used in this case will be 6.

The LMS algorithm for a *p*th order algorithm can be summarized as

$$\hat{h}(0) = 0 \tag{3.88}$$

$$x(n) = [x(n), x(n-1), \dots, x(n-p+1)]^T \tag{3.89}$$

$$e(n) = d(n) - \hat{h}^H(n)x(n) \tag{3.90}$$

$$\hat{h}(n+1) = \hat{h}(n) + \mu e^*(n)x(n) \tag{3.91}$$

Where $\hat{h}^H(n)$ denotes the Hermitian transpose of $\hat{h}(n)$.

3.3.2 Recursive Least Squares (RLS) Channel Estimation

Recursive least squares (RLS) algorithm is used in adaptive filters to find the filter coefficients that relate to recursively producing the least squares of the error signal [16]. This is contrast to other algorithms that aim to reduce the mean square error. The difference is that RLS filters are dependent on the signals themselves, whereas MSE filters are dependent on their statistics. If these statistics are known, an MSE filter with fixed coefficients can be built. An important feature of this filter is that its rate of convergence is typically an order of magnitude faster than that of the simple LMS filter, due to the fact that the RLS filter whitens the input data by using the inverse correlation matrix of the data, assumed to be of zero mean. This improvement in performance, however, is achieved at the expense of an increase in computational complexity of the RLS filter.

The RLS algorithm for a pth order RLS filter can be summarized as

$$w_n = 0 \tag{3.92}$$

$$P(0) = \delta^{-1}I \tag{3.93}$$

$$x(n) = \begin{bmatrix} x(n) \\ x(n-1) \\ \vdots \\ x(n-p) \end{bmatrix} \tag{3.94}$$

$$\alpha(n) = d(n) - w(n-1)^T x(n) \tag{3.95}$$

$$g(n) = P(n-1)x^*(n)\{\lambda + x^T(n)P(n-1)x^*(n)\}^{-1} \tag{3.96}$$

$$P(n) = \lambda^{-1}P(n-1) - g(n)x^T(n)\lambda^{-1}P(n-1) \tag{3.97}$$

$$w(n) = w(n-1) + \alpha(n)g(n) \tag{3.98}$$

Where p is filter order, λ is forgetting factor, δ is value to initialize $P(0)$.

3.3.3 Iterative Channel Estimation for Turbo Receivers in DS-CDMA

Here we introduce another kind of direct technique of channel estimation [17].

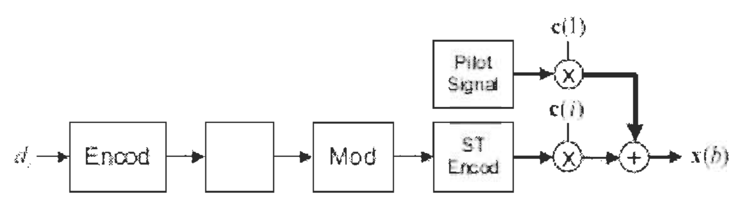


Figure 3.5 Transmission System [34]

We consider a transmission system (see Fig. 3.5) in which the binary information bits are first encoded via a turbo or a convolution encoder, interleaved and then modulated to symbols using a complex constellation. We will denote this set of bits as a coding block. In the modulation stage, the bits to symbols mapping must be done in such a form that the in-phase and in-quadrature components of each symbol can be treated as independent. Afterwards, the modulated symbols are encoded with a ST encoder, spreaded with a spreading factor F and transmitted along with the pilot signal

by the M transmit antennas. Since the ST encoding process is linear the transmitted signal can be decomposed in B ST blocks that depend linearly on the in-phase and in-quadrature components of the modulated symbols. Each ST block is transmitted during T symbol periods using the Linear Dispersion Codes formulation [2].

In this case the chip-level transmitted signal for the b' th block can be represented by the vector $\bar{x}(b) \in \mathcal{C}^{MTF\times 1}$ equal to

$$\bar{x}(b) = \bar{D}s(b) + \bar{p}(b) \qquad b = 1,\cdots,B \tag{3.99}$$

Where $\bar{p}(b)$ is the spreaded pilot signal, $\bar{D}$ is a $MTF \times 2Q$ matrix representing the ST encoding and spreading processes and s(b) is the vector formed by the in-phase and in-quadrature components of the Q complex symbols corresponding to bth ST block.

The MIMO physical channel is modeled as a set of linear filters with L taps. Each tap $\bar{H}^{(l)} \in \mathcal{C}^{N\times M}$ is a complex time varying matrix containing the gains of all the possible paths between transmit and receive antennas. The received signal corresponding to the bth transmitted ST block, denoted by the vector $\bar{y}(b) \in \mathcal{C}^{NTF\times 1}$ is given then by

$$\bar{y}(b) = \bar{H}(b)\big(\bar{D}s(b) + \bar{p}(b)\big) + IBI + \bar{w}(b) \tag{3.100}$$

Where $\bar{H}(b)$ is the $NTF \times MTF$ convolution matrix of the MIMO channel, the term IBI denotes the interblock interference and $\bar{w}(b)$ is the additive spatially and temporally white Gaussian noise vector.

$$\bar{h}(b) = [\bar{H}^{(1)}(1,:)^T \quad \cdots \quad \bar{H}^{(1)}(N,:)^T \quad \cdots \quad \bar{H}^{(L)}(N,:)^T]^T \tag{3.101}$$

Neglecting the IBI for the moment, received signal can be written as below

$$\bar{y}(b) = \bar{H}(b)\bar{D}s(b) + \bar{G}p(b)h(b) + \bar{w}(b) \tag{3.102}$$

$$\bar{G}_p(b)h(b) = \bar{H}(b)\bar{p}(b) \tag{3.103}$$

$$y(b) = G_s(b)s(b) + G_p(b)h(b) + w(b) \tag{3.104}$$

As the Bayesian estimate of the channel vector h(b) by assuming it to be Gaussian distributed with mean η_h and covariance matrix C_{hh}. We also consider symbol vector s(b) as Gaussian with mean η_s and covariance matrix C_{SS}. So we can get

$$\hat{h}(b) = \mathcal{E}\{h(b)|y(b)\} = \eta_h(b-1) + C_{hh}(b-1)G_p(b)^T C_{yy}(b)^{-1}(y -$$

$$G_s(b)\eta_s(b) - G_p(b)\eta_h(b-1) \quad (3.105)$$

Where

$$\eta_h(b) = \hat{h}(b) \tag{3.106}$$

$$C_{hh}(b) = C_{hh}(b-1)\big(I - G_p^T(b)C_{yy}^{-1}(b)G_p(b)C_{hh}(b-1)\big) \tag{3.107}$$

3.3.4 Blind Adaptive MIMO Channel Estimation Algorithm

Here we will study a kind of combine technique [18].

We consider a DS-CDMA system which has K users. The received signal is sampled once per spreading chip.

$$\underline{X} = \begin{bmatrix}\underline{x}_1^T & \underline{x}_2^T & \cdots & \underline{x}_A^T\end{bmatrix}^T = \bar{C}\underline{\phi} + \underline{n} \tag{3.108}$$

$\underline{n}$ is noise vector, which can be defined as

$$\underline{n} = \begin{bmatrix} \underline{n}_1^T & \underline{n}_2^T & \cdots & \underline{n}_A^T \end{bmatrix}^T \tag{3.109}$$

$$\underline{\phi} = \begin{bmatrix} \underline{\phi}_1^T & \underline{\phi}_2^T & \cdots & \underline{\phi}_K^T \end{bmatrix}^T \tag{3.110}$$

Define $\underline{c}_{ak} = [c_{ak}(0) \quad c_{ak}(1) \quad \cdots \quad c_{ak}(M-1)]^T$ (3.111)

$$C_{ak} = \begin{bmatrix} c_{ak}(0) & 0 & & \cdots & 0 \\ c_{ak}(1) & c_{ak}(0) & & \ddots & \vdots \\ \vdots & c_{ak}(1) & & \ddots & \vdots \\ c_{ak}(M-1) & \vdots & & \ddots & \vdots \\ 0 & c_{ak}(M-1) & & \ddots & \vdots \\ \vdots & \ddots & & \ddots & \vdots \\ \vdots & \ddots & \ddots & & c_{ak}(0) \\ \vdots & \ddots & \ddots & & c_{ak}(1) \\ \vdots & \ddots & \ddots & & \vdots \\ 0 & 0 & \ddots & & c_{ak}(M-1) \end{bmatrix} \tag{3.112}$$

C represents MIMO channel:

$$C = \begin{bmatrix} C_{11} & C_{12} & \cdots & C_{1K} \\ \vdots & \vdots & \ddots & \vdots \\ \vdots & \vdots & \ddots & \vdots \\ C_{A1} & C_{A2} & \cdots & C_{AK} \end{bmatrix} = [C_1 \quad C_2 \quad \cdots \quad C_K] \tag{3.113}$$

The spreading signature code matrix S is defined as follow

$$S = diag\{\underline{s}_1 \quad \underline{s}_2 \quad \cdots \quad \underline{s}_K\} \tag{3.114}$$

Data vector is defined as below

$$\underline{b} = [b_1 \quad b_2 \quad \cdots \quad b_K]^T \tag{3.115}$$

We know that for multiuser signals detection, the decision is made in every symbol interval. The channel information is updated in every chip interval. So we get

$$x_a(n) = \underline{c}_a^T \underline{\Psi}(n) + n_a(n) \tag{3.116}$$

Where the channel vector for the ath antenna

$$\underline{c}_a = [\underline{c}_{a1}^T \quad \underline{c}_{a2}^T \quad \cdots \quad \underline{c}_{aK}^T]^T \tag{3.117}$$

And the input vector

$$\underline{\Psi}(n) = [\underline{\Psi}_1^T(n) \quad \underline{\Psi}_2^T(n) \quad \cdots \quad \underline{\Psi}_K^T(n)]^T \tag{3.118}$$

Using the Linear Minimum Mean Square Estimation (LMMSE) technique, our objective is to estimate the MIMO channels by minimizing the Mean Square Error (MSE) between the received signal and the estimator,

$$J_{MMSE}(n) = \sum_{a=1}^{A} J_a(n) = \sum_{a=1}^{A} E|x_a(n) - \hat{x}_a(n)|^2 \tag{3.119}$$

$$\hat{x}_a(n) = \underline{\hat{c}}_a^T \underline{v}(n) \tag{3.120}$$

Where $\underline{v}(n)$ is the respread chip sequence between the (n-M+1)th and the nth chip intervals. $\underline{\hat{c}}_a$ is the estimator of channel vector.

$$\underline{\hat{c}}_a^T(n+1) = \underline{\hat{c}}_a^T(n) + \mu_{ch}\left(x_a(n) - \underline{\hat{c}}_a^T(n)\underline{v}^c(n)\right)\underline{v}^c(n)^H \tag{3.121}$$

Where μ_{ch} is the step size for interation.

3.3.5 Novel Channel Estimation Algorithm Using Kalman Filter

Assume K-user asynchronous DS-CDMA system has J paths [19].

$$x_k(t) = \sum_m b_k(m) s_k(t - mT_s) \tag{3.122}$$

Spreading waveform $s_k(t)$ is given by

$$s_k(t) = \sum_{n=1}^{N} c_k(n) p(t - nT_c) \tag{3.123}$$

Where $T_c = T_s/N$ is the chip duration, $c_k(n)$ is the spreading sequence of kth user, p(t) is the chip waveform and N is the processing gain.

Assume path delay is D, we can set multipath channel as a vector $h_k(m)$ pf order q=D+1.

We define spreading code as follow

$$\bar{c}_k(m) = C_k h_k(m) \tag{3.124}$$

Where C_k is $(N + D) \times D$ matrix as

$$C_k = \begin{bmatrix} c_k(0) & 0 & \cdots & 0 \\ \vdots & c_k(0) & \cdots & \vdots \\ c_k(N-1) & \vdots & \ddots & 0 \\ 0 & c_k(N-1) & \cdots & c_k(0) \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & c_k(N-1) \end{bmatrix} \tag{3.125}$$

Then, the received signal is

$$r(m) = [r(0) \quad r(1) \quad \cdots \quad r(N+D-1)]^T = \sum_{k=1}^{K} C_k h_k(m) b_k(m) + u(m) =$$

$$\sum_{k=1}^{K} \bar{c}_k(m) b_k(m) + u(m) \tag{3.126}$$

Where u includes the multiple access interference (MAI) signal and inter-symbol interference (ISI) signals.

Then the channel model is

$$h_k(m+1) = a h_k(m) + v(m) \tag{3.127}$$

Where $a = \exp(-2\pi f_d T_s)$ and v is a zero-mean white Gaussian variable with variance σ_v^2.

The received signal can be written as

$$R_r(m) = E\{r(m)r^T(m)\} = \sum_{k=1}^{K} C_k h_k(m) h_k^T(m) C_k + R_i(m) \tag{3.128}$$

Where $R_i(m)$ is the autocorrelation matrix of the interference signals.

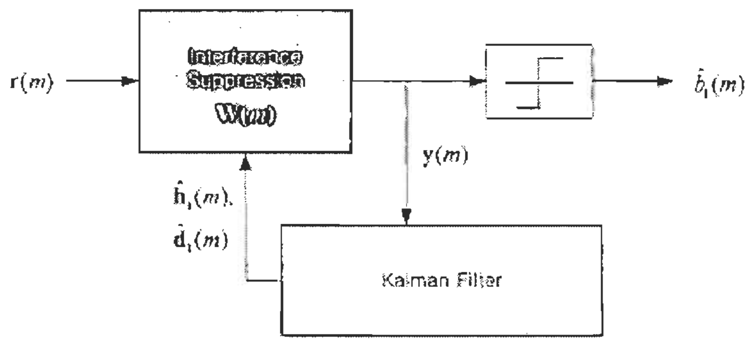


Figure 3.6 System block diagram of channel estimation using Kalman filter without bit estimation [35]

The state vector at the time m is defined not to use the estimated bit symbol using operation as follow

$$d_1(m) = vec(h_1(m)h_1^T(m)) \tag{3.129}$$

$$d_1(m+1) = vec\big(h_1(m+1)h_1^T(m+1)\big) = a^2 d_1(m) + v_2(m) \tag{3.130}$$

Where

$$v_2(m) = avec(h_1(m)v^T(m) + v(m)h_1^T) + vec(v(m)v^T(m)) \tag{3.131}$$

$$R_p(m) = E\{v_2(m)v_2^T(m)\} \tag{3.132}$$

The inside equation can be written as follow

$v_2(m)v_2^T(m) =$

$a^2 vec\big(h_1(m)v^T(m)\big)\{vec\big(h_1(m)v^T(m)\big)\}^T +$

$$a^2vec\big(v(m)h_1^T(m)\big)\big\{vec\big(v(m)h_1^T(m)\big)\big\}^T +$$

$$vec\big(v(m)v^T(m)\big)\big\{vec\big(v(m)v^T(m)\big)\big\}^T +$$

$$a^2vec(h_1(m)v^T(m))\{vec(v(m)h_1^T(m))\}^T +$$

$$avec\big(h_1(m)v^T(m)\big)\big\{vec\big(v(m)v^T(m)\big)\big\}^T +$$

$$a^2vec\big(v(m)h_1^T(m)\big)\big\{vec\big(h_1(m)v^T(m)\big)\big\}^T +$$

$$avec\big(v(m)h_1^T(m)\big)\big\{vec\big(v(m)v^T(m)\big)\big\}^T +$$

$$avec\big(v(m)v^T(m)\big)\big\{vec\big(h_1(m)v^T(m)\big)\big\}^T +$$

$$avec\big(v(m)v^T(m)\big)\big\{vec\big(v(m)h_1^T(m)\big)\big\}^T \tag{3.133}$$

The expectation of each term in the right side is solved as

$$E\{a^2vec(h_1(m)v^T(m))\{vec(h_1(m)v^T(m))\}^T\} = a^2\sigma_v^2 I_J \otimes (h_1(m)h_1^T(m)) \tag{3.134}$$

$$E\{a^2vec(v(m)h_1^T(m))\{vec(v(m)h_1^T(m))\}^T\} = a^2\sigma_v^2 (h_1(m)h_1^T(m)) \otimes I_J \tag{3.135}$$

$$E\{vec(v(m)v^T(m))\{vec(v(m)v^T(m))\}^T\} = \sigma_v^4 I_{J\times J} \tag{3.136}$$

$$E\{a^2vec(h_1(m)v^T(m))\{vec(v(m)h_1^T(m))\}^T\} = a^2\sigma_v^2 h_1^T \otimes (I_J \otimes h_1(m)) \tag{3.137}$$

$$E\{avec(h_1(m)v^T(m))\{vec(v(m)v^T(m))\}^T\} = 0_{J\times J} \tag{3.138}$$

$$E\{a^2vec(v(m)h_1^T(m))\{vec(h_1(m)v^T(m))\}^T\} = a^2\sigma_v^2 h_1(m) \otimes (I_J \otimes h_1^T(m))$$

(3.139)

$$E\{avec(v(m)h_1^T(m))\{vec(v(m)v^T(m))\}^T\} = 0_{J\times J} \tag{3.140}$$

$$E\{avec(v(m)v^T(m))\{vec(v(m)v^T(m))\}^T\} = 0_{J\times J} \tag{3.141}$$

$$E\{avec(v(m)v^T(m))\{vec(v(m)h_1^T(m))\}^T\} = 0_{J\times J} \tag{3.142}$$

The measurement vector y(m) is defined as follow

$$y(m) = vec\{(W^T(m)r(m))(W^T(m)r(m))^T\} \quad (3.143)$$

Where

$W(m) =$

$$\min_{W(m)} E\{(W^T(m)r(m))^T(W^T(m)r(m))\} = R_r^{-1}(m)C_1(C_1^T R_r^{-1}(m)C_1)^{-1} \quad (3.144)$$

$$\hat{R}_r(m) \equiv E\{r(m)r^T(m)\} = C_1\hat{h}_1(m)\hat{h}_1^T(m)C_1 + \hat{R}_{in}(m) \quad (3.145)$$

Where $\hat{R}_{in}(m)$ is the estimated autocorrelation matrix of MAI and ISI

$$\hat{R}_{in} = \frac{m-1}{m}\hat{R}_{in}(m-1) + \frac{1}{m}\{r(m)r^T(m) - C_1\hat{h}_1(m)\hat{h}_1^T(m)C_1\} \quad (3.146)$$

Assume that the decorrelating filter suppress interference signals sufficiently, the measurement equation is constructed using the output of CMVM filter as follow

$$y(m) = vec\{(W^T(m)C_1h_1(m)b_1(m) + W^T(m)v(m))(W^T(m)C_1h_1(m)b_1(m) + W^T(m)v(m))^T\} \equiv Q_1(m)d_1(m) + v_w(m) \quad (3.147)$$

Where

$$Q_1(m) = (W^T(m)C_1) \otimes (W^T(m)C_1) \quad (3.148)$$

$$v_w(m) = b_1(m)(W^T(m) \otimes W^T(m))vec(C_1h_1(m)v(m)) + b_1(m)(W^T(m) \otimes W^T(m))vec(v(m)h_1^T(m)C_1^T) + (W^T(m) \otimes W^T(m))vec(v(m)v^T(m)) \quad (3.149)$$

$$vec(ABC) = (C^T \otimes A)vec(B) \quad (3.150)$$

The correlation matrix of the measurement noise vector is defined as

$$E\{v_w(m)v_w^T(m)\} = W_2(m)E\{vec(v(m)v^T(m)vec(v(m)v(m))\}W_2^T(m) = W_2(m)W_2^T(m)\sigma_n^4 \mathrm{I}_{(N+D)\times(N+D)} \quad (3.151)$$

From the equations above, Kalman filter can be written as follow

$$\Gamma(m) = a^2\Lambda(m, m-1)Q_1^T(m) \cdot [Q_1\Lambda(m, m-1)Q_1^T(m) + R_m(m)]^{-1}$$

(3.152)

$$\alpha(m) = y(m) - Q_1(m)\hat{d}_1(m)$$

(3.153)

$$\hat{d}_1(m+1) = a^2\hat{d}_1(m) + \Gamma(m)\alpha(m) \quad (3.154)$$

$$\mathrm{K(m)} = \Lambda(\mathrm{m}, \mathrm{m}-1) - \mathrm{a}^2\Gamma(\mathrm{m})\mathrm{Q}_1(\mathrm{m})\Lambda(\mathrm{m}, \mathrm{m}-1) \quad (3.155)$$

$$\Lambda(\mathrm{m}, \mathrm{m}-1) = \alpha^4\mathrm{K(m)} + \mathrm{R_p(m)} \quad (3.156)$$

Where $\Gamma(m)$ in Kalman filter gain at the time m, $\alpha(m)$ is the innovations vector at time m, $\Lambda(m)$ is the correlation matrix of the error in $\hat{d}_1(m+1)$, and K(m) is the correlation matrix of error in $\hat{d}_1(m)$

Finally, we can get $\hat{h}_1(m)$ from $\hat{d}_1(m)$.

3.3.6 Joint Channel Estimation

We assume that we have K users in the channel [20].

$$\mathrm{s_k(t)} = \sum_{n=0}^{N-1} \mathrm{c_k(n)P_{T_c}(t - nT_c)}, \quad 0 \le \mathrm{t} \le \mathrm{T} \quad (3.157)$$

The baseband equivalent received signal from K users plus the additive noise can be written as

$$\mathrm{y(t)} = \sum_{i=0}^{M-1} \sum_{k=1}^{K} \mathrm{A_k x_k(i)} \sum_{l=1}^{L} \mathrm{a_{kl}(0_{kl}) g_{kl} s_k(t - iT - \tau_{kl}) + n(t)} \quad (3.158)$$

Where M is the number of data symbols per user in the data frame of interest, T is the symbol interval, and L is the number of paths in each user's channel.

At the receiver, when signal passed matched-filter, linear prediction is applied for each lth path of the receiver to estimate the column vector subspace of the channel matrix using only the spreading code of the desired user.

The received the signal can be written as

$$y(i) = \left[y_0(i), y_1(i), \dots, y_{\overline{N}-1}(i)\right]^T \tag{3.159}$$

Then we can get the signal vector as follow

$$y_n(i) = \int_{iT+\tau_{kj}+nT_c}^{iT+\tau_{kj}+(n+1)T_c} r(t)P_{T_c}(t - iT - \tau_{kl} - nT_c)dt \tag{3.160}$$

Then, we can get

$$y(i) = A_k x_k(i) \sum_{l=1}^{L} a_{kl}(O_{kl}) g_{kl} s_{kl}^{[0]} + \bar{i}(i) + n(i) \tag{3.161}$$

Where $n(i) = \sigma I_{\overline{N}}$ is the white Gaussian noise vector of the receiver with variance of σ^2, and $\bar{i}(i)$ consists of the interfering signals

We can get

$$\bar{i}(i) = \sum_{j=-d}^{d} A_k x_k(i+j) \sum_{l=1}^{L} a_{kl}(O_{kl}) g_{kl} s_{kl}^{[j]} + \sum_{j=-d}^{d} \sum_{k' \neq k} A_{k'} x_{k'}(i + j) \sum_{l=1}^{L} a_{k'l} g_{k'l} s_{k'l}^{[j]} \tag{3.162}$$

Where

$$s_{k'l}^{[j]}(n) = \int_{\tau_{kl}+nT_c}^{\tau_{kl}+(n+1)T_c} s_{k'l}(t - jT - \tau_{k'l}) P_{T_c}(t - \tau_{kl} - nT_c)dt \tag{3.163}$$

We construct following linear prediction problem

$$\varepsilon(i) = y^{[0]}(i) - P_{y^{[else]}}(i) \tag{3.164}$$

Where $y^{[0]}(i)$ is corresponding to $s_{kl}^{[0]}$, $y^{[else]}(i)$ is corresponding to $s_{kl}^{[j]}$.

Assume that the symbols $x_k(i)$ are uncorrelated in time. We define

$$y^{[0]}(i) = H_0 x_k(i) + \widetilde{H}_0 \tilde{x}_k(i') \tag{3.165}$$

Where $\tilde{x}_k(i')$ contains all symbol components in $[x_1^H(i), \cdots, x_K^H(i)]^H$ except for $x_k(i)$. H_0 is the column vector of the channel matrix $[a_{kl}(O_{kl})g_{kl}s_k]$ corresponding to $x_k(i)$, where all other columns of $[a_{kl}(O_{kl})g_{kl}s_k]$ comprise $\widetilde{H}_0$.

The optimal linear prediction matrix P gives

$$\varepsilon_1(i) = H_0 x_k(i) \tag{3.166}$$

Let

$$\varepsilon_2(i) = y^{[0]}(i) - \varepsilon_1(i) \tag{3.167}$$

Then we have

$$\varepsilon_2(i) = \widetilde{H}_0 \tilde{x}_k(i') \tag{3.168}$$

The solution of P and the linear prediction error can also be represented by the data correlation. Then we get the prediction equation

$$\varepsilon_1(i) = [I = P] \begin{bmatrix} y^{[0]}(i) \\ y^{[else]}(i) \end{bmatrix} \tag{3.169}$$

And let

$$R = E\left\{ \begin{bmatrix} y^{[0]}(i) \\ y^{[else]}(i) \end{bmatrix} [y^{[0]}(i) \quad y^{[else]}(i)] \right\} = \begin{bmatrix} R_{11} & R_{12} \\ R_{21} & R_{22} \end{bmatrix} \tag{3.170}$$

$$P = R_{12} R_{22}^{+} \tag{3.171}$$

$$E\{\varepsilon_1(i)\varepsilon_1^H(i)\} = A_k H_0 H_0^H = R_{11} - R_{12} R_{22}^{+} R_{21} \tag{3.172}$$

Where R_{22}^{+} denotes the pseudo inverse of R_{22}.

Using above linear prediction algorithm we can estimate the channel response H_0 corresponding to $x_k(i)$ for the receiver.

3.3.7 Modified Leaky LMS Algorithm

In DS-CDMA system, after the RAKE receiver, the received signal can be written as follow [21]

$$x(k) = h(k) + n(k) \tag{3.173}$$

Where h(k) is channel parameter and n(k) is zero-mean additive white noise.

Suppose that an estimation of h(k) is given by

$$\hat{h}(k) = w^H x(k) \tag{3.174}$$

Where w is an N-dimensional tap weight vector

$$x(k) = [x(k), x(k-1), \cdots, x(k-N+1)]^T \tag{3.175}$$

The optimal tap weight w_0 minimizing $E\left[\left|h(k) - \hat{h}(k)\right|^2\right]$ satisfies the Wiener-Hopf equation

$$Rw_0 = p_{xh} \tag{3.176}$$

Where $R = E[x(k)x^H(k)]$ (3.177)

$$p_{xh} = E[x(k)h^*(k)] \tag{3.178}$$

We can write p_{xh} as follow

$$p_{xh} = p_{xx} - [\sigma^2, 0, \cdots, 0]^T \tag{3.179}$$

Where

$$p_{xx} = E[x(k)x^*(k)] \tag{3.180}$$

σ^2 is the variance of n(k)

Then we can get

$$(R + A)w_0 = p_{xx} \tag{3.181}$$

The parameter α_0 is given by

$$\alpha_0 = \frac{\sigma^2}{w_{0,0}} \tag{3.182}$$

The steepest descent algorithm for obtaining w_0 is given by

$$w(k+1) = w(k) - \mu E[-x(k)x^*(k) + (x(k)x^H(k) + A)w] \tag{3.183}$$

After removing the expectation E[.], we can get

$$w(k+1) = (I - \mu A)w(k) + \mu x(k)\{x(k) - w^H(k)x(k)\}^* \tag{3.184}$$

So

$$\alpha(k) = \frac{\sigma^2}{w_0(k)} \tag{3.185}$$

Where $w_0(k)$ is the first element of w(k)

The proposed channel estimation algorithm is summarized as follow:

- Channel estimation

$$\hat{h}(k) = w^H(k)x(k) \tag{3.186}$$

- Estimation error

$$e(k) = x(k) - \hat{h}(k) \tag{3.187}$$

- Tap-weight adaptation

$$w(k+1) = \left(I - \mu\hat{A}(k)\right)w(k) + \mu x(k)e^*(k) \tag{3.188}$$

3.3.8 Recursive Least Square Constant Modulus Algorithm for Blind Adaptive Array

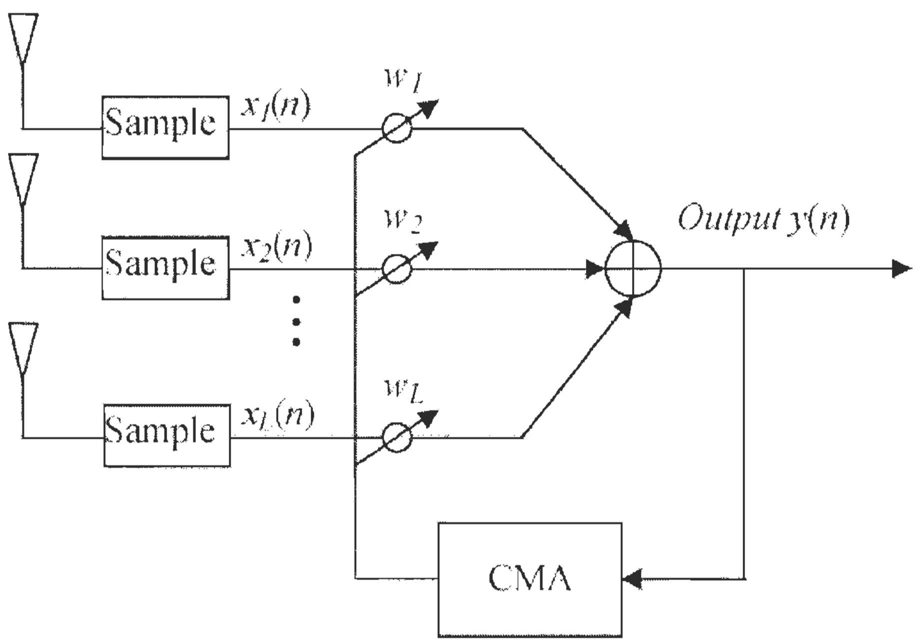


Figure 3.7 CMA structure

A Constant Modulus Algorithm (CMA) array is shown above [35]. The signals are received by an antenna array which has L antennas. The received signal from each antenna are scaled by complex weight $w_n(n)$, array output is given by:

$$y(n) = \sum_{l=1}^{L} w_l^* x_l(n) \tag{3.189}$$

Where the superscript * denotes complex conjugate.

The purpose of the adaptive array is to extract the desired signal by finding a suitable weight vector. It is well known that adaptation algorithm based on the RLS algorithm have a faster convergence rate. So we get RLS-CMA algorithm as below:

$$w(0) = \left[1, 0_{1\times(L-1)}\right]^T \tag{3.190}$$

$$C(0) = \delta^{-1} I_{L\times L}, \delta = small\ positive\ constant \tag{3.191}$$

$$z(n) = x(n)x^H(n)w(n-1)|x^H(n)w(n-1)|^{p-2} \tag{3.192}$$

$$h(n) = z^H(n)C(n-1) \tag{3.193}$$

$$g(n) = C(n-1)z(n)/(\lambda + h(n)z(n)) \tag{3.194}$$

$$C(n) = \frac{C(n-1) - g(n)h(n)}{\lambda} \tag{3.195}$$

$$e(n) = w^H(n-1)z(n) - 1 \tag{3.196}$$

$$w(n) = w(n-1) + g(n)e^*(n) \tag{3.197}$$

4. Simulations

4.1 Simulation of SISO-CDMA

4.1.1 SISO-CDMA Platform

In simulation, we will start from the SISO-CDMA platform. Because of the SISO antenna system is less complicated than the MIMO antenna system. Meanwhile, we can compare the SISO's simulation results with the MIMO's simulation results. We can see the differences of the same channel estimation algorithm between two different antenna systems. We can see how the channel elements affect the estimation results.

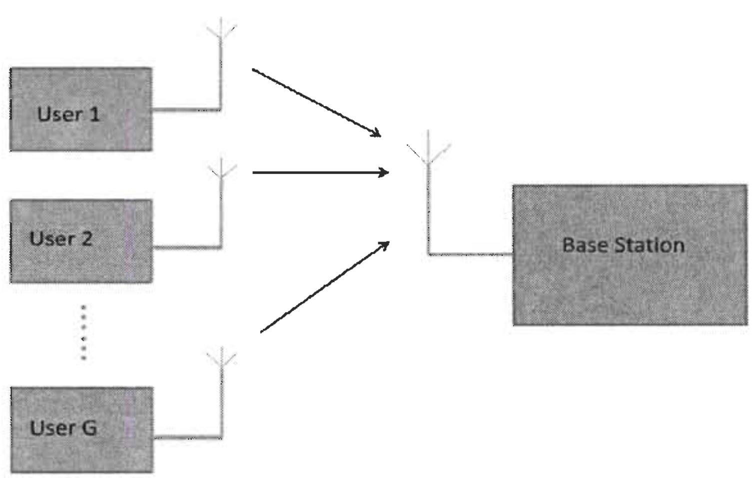


Figure 4.1 SISO-CDMA System Block Diagram

As shown in Figure 5.1, we assume this simulation platform is composed by multiple users and 1 base station. Between each user and base station, we will consider the multi-path channels. In realistic environment we have many multipaths. In this simulation environment, considering the computation complexity, we assume that there are 6 multipaths. It will cause delay during signal transmission.

The platform we build here is actually a short-code DS-CDMA. At the side of the transmitter, we use Gold code to spread the original signal. We get

$$x = B \times S \qquad (4.1)$$

Where B denotes the original signals, S denotes spread code, in this case which is Gold code.

In order to approaching the real environment, we consider the signal channel as a multipath channel. So we get the channel as below:

$$H = [h_0 \quad h_1 \quad \cdots \quad h_L]' \qquad (4.2)$$

Where L depends on the number of tap we assume in the multipath delay. In this case, L=6.

At the side of receiver, we describe the received signal as below:

$$y = Hx + n \qquad (4.3)$$

Where y is received signals, H is the signal channel coefficient, x is the spreading code, n denotes noise sample.

In real, a complete telecommunication receive side always have detector and estimator. We also know that it is impossible to know the x. We usually get this x from a detector. In this case, we only discuss the estimation problem, so we assume we have already known the x and the received signal y. Here we assume that we have a "perfect" detector.

Because we considered the multipath situation, we use Rake receiver to remove the multipath inference.

Take the LS algorithm for example. We can get estimation coefficient $\hat{H}_{LS}$ as below:

$$\hat{H}_{LS} = (x^H x)^{-1} x^H y \tag{4.4}$$

And then, we compare it with the ideal channel coefficient H so we can get BER of the LS algorithm.

4.1.2 Simulations

In this simulation, we assume that the simulation environment as follow: At the side of user, although the platform is design for N users, considering computation complexity, especially for the following MIMO-CDMA platform, we assume we have 2 users, each user has 1 transmit antenna. At the side of base station, there is 1 receive antenna. Each signal length is 10000 and spread code length is 31. The number of multipath is 6.

For LMS estimation algorithm, we assume that the step size $\mu = 0.001$. We know that the value of step size the bigger the algorithm convergence the faster. We will choose different step size to compare the simulation results.

For NLMS algorithm, we know that the NLMS algorithm is normalized LMS algorithm. In order to compare with LMS algorithm, we hereby assume the step size the same as the LMS algorithm, step size $\mu = 0.001$.

For the RLS algorithm, we assume forgetting factor $\lambda = 1$, initial coefficient $\delta = 0.001$. We know that for the coefficient λ, the smaller the λ is, the smaller contribution of previous samples.

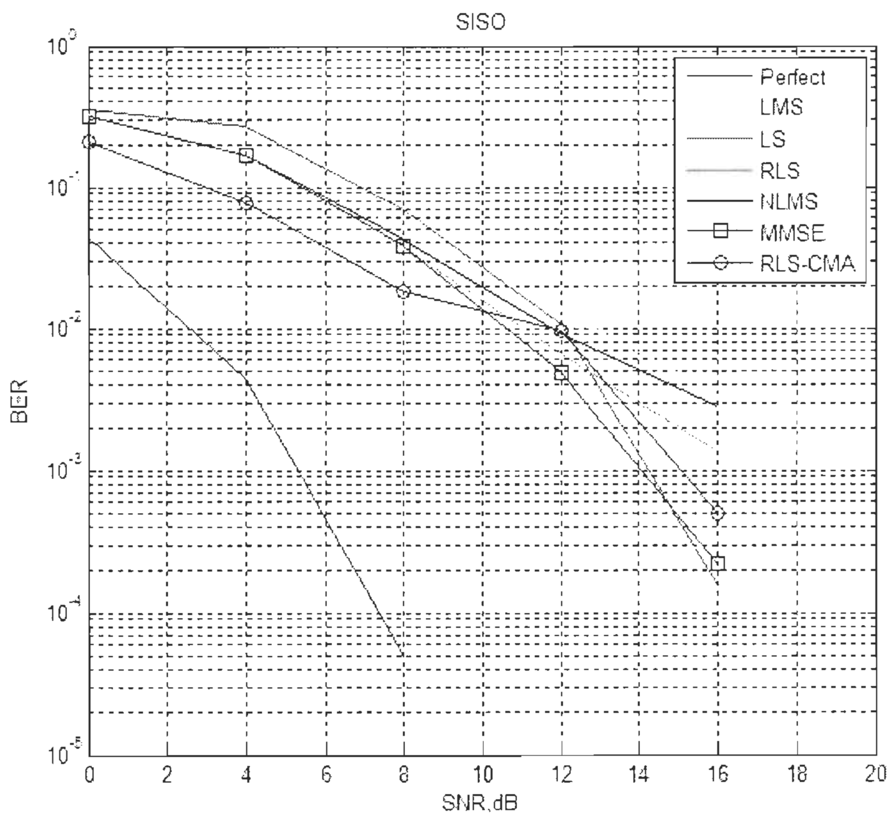


Figure 4.2 Simulation Results in SISO-CDMA environment

We can see that in SISO-CDMA environment. When SNR is between around 0dB and 10dB, adaptive filter algorithms have the similar performance with the MMSE algorithm. RLS-CMA (Blind) algorithm has the best performance. Generally speaking, in this SNR condition, MMSE algorithm's performance is a little bit better than adaptive algorithms'. When SNR 12dB and above, adaptive algorithms become appear low efficience. Especially when the SNR reaches 15dB, we can see that LS algorithm becomes better than any other algorithms.

In low SNR environment, we can see that blind algorithm has the best performance. In high SNR environment, MMSE has its advantage.

4.2 Simulation of MIMO-CDMA

4.2 .1MIMO-CDMA Platform

In MIMO simulation environment, we assume that we have multiple users and the user number is G. For each user, we have multiple transmit antennas, the number of transmit antenna is Nt. At base station side, receive antenna number is Nr. We have the MIMO-CDMA system block diagram is as follow

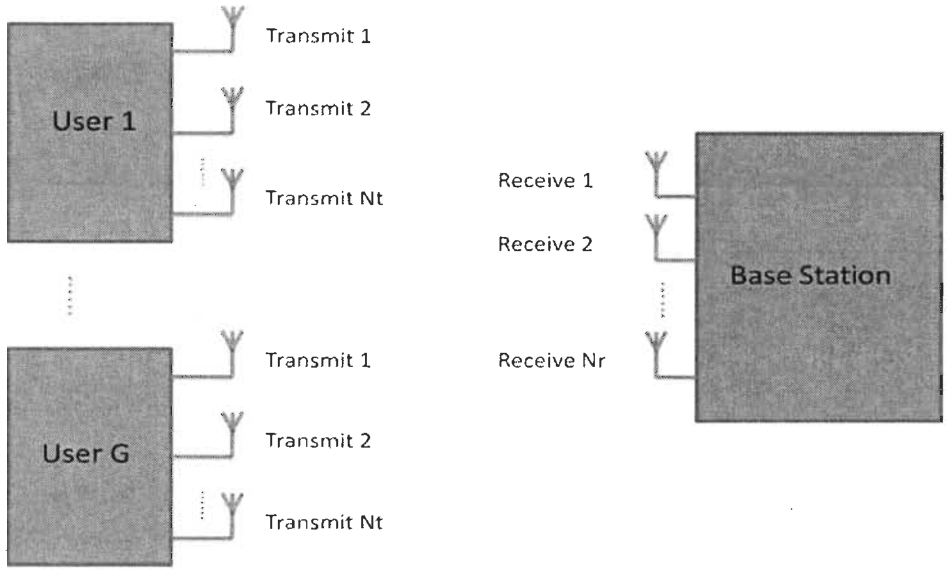


Figure 4.3 MIMO System Block Diagram

In order to approach the realistic environment, we should also consider the multipath channel in this case.

Same as the SISO-CDMA platform, we use Gold code as the spread code. In the simulation platform, the same user use the same spread code on different transmit antenna. The transmitted signal of the gth user's the *i*th antenna is:

$$x_g(t) = \sum_{l=-\infty}^{\infty} b_{g,i}(l) s_{l,g}(t) \tag{4.5}$$

Where $b_{g,i}(l)$ is the data symbol of the gth user's the ith antenna at symbol index l.

The received signal by the jth antenna is

$$y_j(t) = \sum_{i=1}^{Nt} \sum_{L=1}^{L} \sum_{g=1}^{G} \sum_{l=-\infty}^{\infty} b_{g,i}(l) h_{g,j,i}^{(m)}(l) s_{l,g}(t - \tau_{m,g}) + n(t) \tag{4.6}$$

Where L denotes the number of multipath. $h_{g,j,i}^{(m)}(l)$ is the ith transmit antenna of the gth user's fading coefficient of the Lth path at the jth receive antenna at symbol index l, and the delay $\tau_{m,g}$ is the delay for the Lth multipath of the gth user. n(t) denotes thermal noise.

We get the channel model as

$$H(l) = \left(h^{(1)}(l)\ h^{(2)}(l)\ \cdots\ h^{(L)}(l)\right)' \tag{4.7}$$

So, the received signal should be as

$$y(l) = x(l)H(l) + n(l) \tag{4.8}$$

4.2.2 Simulations

In this simulation, we assume that we have 2 users. Each user has 2 transmit antennas. At the side of receiver, receiving antenna number is 4. Each pair of signal channel (1 transmit antenna and 1 receive antenna) has 6-path fading channel.

For LMS estimation algorithm, we assume that the step size $\mu = 0.1$. We know that the value of step size the bigger the algorithm convergence the faster. Considering the complexity of MIMO antenna system, we hope the convergence time the shorter the better.

For NLMS algorithm, we choose the step size the same with LMS, which $\mu = 0.1$.

For the RLS algorithm, we assume forgetting factor $\lambda = 1$, initial coefficient $\delta = 0.001$. We know that for the coefficient λ, the smaller the λ is, the smaller contribution of previous samples. In MIMO antenna system, the channel condition is very complicated.

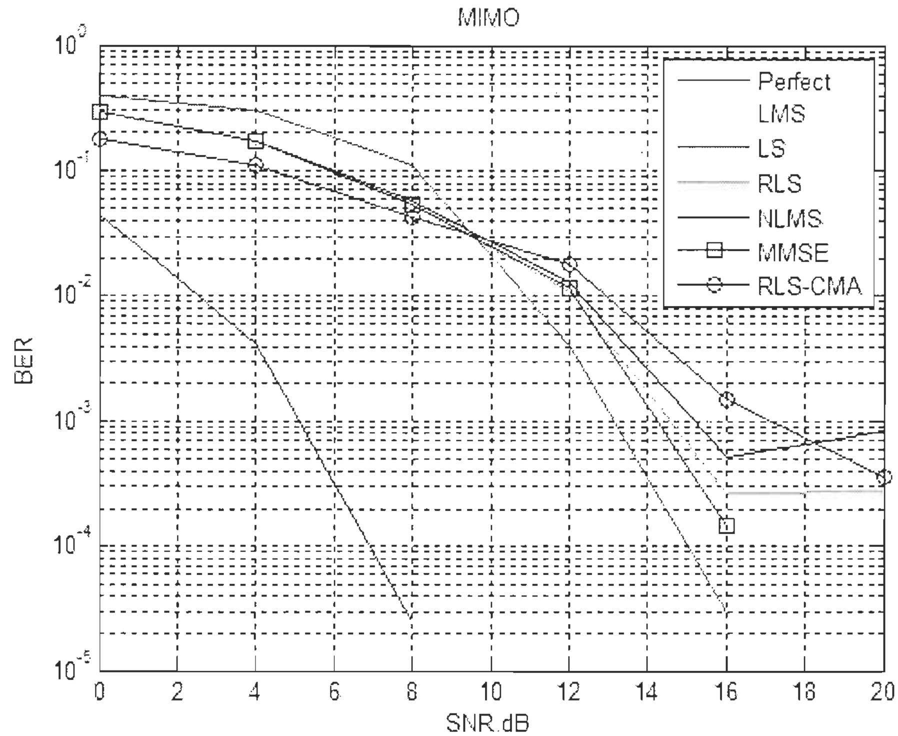


Figure 4.4 Simulation Results in MIMO-CDMA environment

From Figure 4.4, we can find that in low SNR condition (between 0dB and 10dB), adaptive estimation algorithms, MMSE and blind algorithm are better than LS algorithm. Blind algorithm has the best performance in this area. All adaptive algorithm have the similar performance. When in high SNR condition (10dB and above), blind algorithm (RLS-CMA) become low efficient. LS algorithm has the best performance in this condition. The performance of each algorithm can be written as below, LS>MMSE>RLS>NLMS>LMS>RLS-CMA (Blind).

5. Conclusion

From the simulation results above, because of that all the algorithms are simulated on the same platforms, we can compare the results directly. We can find out that in both SISO and MIMO antenna system environment, when SNR is relatively low, which means in the high noise environment, blind algorithm and adaptive techniques have better performance. Although in SISO-CDMA system, different technique's advantage in low SNR environment is not distinctly. When SNR is rising, indirect technique such as LS algorithm becomes much better than others. We can see that generally speaking blind algorithm is better for low SNR environment, while indirect technique is better for high SNR environment. Compare with SISO platform, MIMO platform has more complicated channel environment and noise interference, direct technique's performance in MIMO environment is worse than in SISO environment. For indirect technique, the performance is more stable than direct technique; the complicated channel does not affect indirect technique as much as direct technique does.

We should also notice that because of the signal resource we use in this simulation platform is generated randomly by computer, sometimes the simulation will have some different results. But the nexus of each algorithm's performances are comparatively stable.

When we choose estimation algorithm, we should consider different SNR environment. Like what we see in the simulations above, different estimations in different SNR environment may have their own advantage. Also, we have to consider the processing time and computation complexity.

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