行政院國家科學委員會專題研究計畫成果報告

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IMT-2000 軟體無線電架構之研究(III) – 總計畫
Software Radio Structure for Phase III IMT-2000

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IMT-2000 軟體無線電架構之研究 - 總計畫

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中文摘要

關鍵詞： 軟體無線電，IMT-2000，寬頻分碼多工

本群體研究計畫期能提供寬頻分碼多工 IMT-2000 技術，並針對

- 更有效之語音/影像編碼
- 二維空-時 RAKE 接收器及天線
- 同步技術
- 錯誤控制技術
- HoS 偵測技術與等化
- 多碼多速率分碼多工技術
- 系統性能評估
- 軟體整合技術

不僅進行理論研究，更將以共同的軟體平台發展實際數位訊號處理演算法及實際驗證。更將進行初期整合工作以為下期實作準備。

Abstract

Key words: software radio, IMT-2000, wide-band CDMA

This research group project intends to provide technology solutions for wide-band CDMA IMT-2000. It aims at the following working items.

- ❑ More effective speech/video coding
- ❑ 2-D antenna array and RAKE receiver
- ❑ New synchronization techniques to enhance link performance
- ❑ New error control technology
- ❑ HOS to enhance detection performance
- ❑ Multicode, multirate CDMA for multimedia/Internet applications
- ❑ System performance
- ❑ Channel Estimation and Software Integration

In addition to theoretical study and evaluation, we shall use a common platform to verify developed DSP algorithms in each part. In the third year, we shall start initial integration of software for the preparation of real implementation of IMT-2000 software radio in the second stage (4^{th} and 5^{th} years) toward practical industrial applications.

研究計畫之背景及目的

Since the first introduction of analog cellular (the first generation) AMPS in 1983, cellular communication has become as one of the fastest growing industry. While migrating from the analog cellular systems into the second generation digital systems such as GSM in Europe and IS-54 in North America using TDMA technology, and IS-95 applying CDMA technology, all of them are regional standard and technologies. In light the need of world wide common standard for mobile users, ITU-R (International Telecommunication Union – Radio) started the effort of the third generation wireless communication systems, initially known as FPLMTS (Future Public Land Mobile Telecommunication System). Later on, it has been changed to a new name, International Mobile Telecommunications – 2000 (IMT-2000). The meaning of 2000 is two-fold, with frequency range around 2000 MHz and being realized around year 2000. The ultimate goal of this ITU-R effort is to create a world-wide common system on a world-wide common frequency band to fully realize the no-boundary mobiie communications. Although the official deadline for IMT-2000 proposal is June 1998 and it will not be finalized prior to the end of 1999. It suffices to summarize the important air-interface technologies that are possibly adopted in IMT-2000.

■ IMT-2000 Requirements Related to Air-Interface

IMT-2000 targets at common network signaling and numbering, in addition to international roaming. For the purpose of worldwide common frequency band to ensure this goal, the following frequency bands have been assigned in WARC'92 and WARC'95. IMT-2000 will use 1885-2025 MHz and 2110-2200 MHz while 1980-2010 MHz and 2170-2200 MHz can be used for satellite components.

IMT-2000 targets not only for conventional voice wireless communications but also for multimedia wireless communications services. It is expected that a good portion of IMT-2000 traffic will be of data/multimedia form. According to the consensus, Table 1 summarizes the minimum service rates for IMT-2000.

Application Scenario	**Minimum Service Rate**
Indoor	2M bps
Pedestrians	384k bps
Vehicles	144k bps
Satellite Mobile	9.6k bps

Table 1 Minimum Service Rate Requirements for IMT-2000

- <u>Wide-band CDMA</u>

Due to the success of narrow-band CDMA cellular to demonstrate advantages in high system capacity and other networking management considerations, adopting CDMA technology for IMT-2000 has been seriously evaluated in all major players in the world. However, some inherent difficulties for IS-95 narrow-band CDMA exist

- ❑ Based on IS-95 1.2288 Mcps and 8 Kbps basic transmission rate, many high-bandwidth applications required by IMT-2000 are not practically feasible even by using multi-code CDMA.
- ❑ IMT-2000 operates for wide range applications in higher frequency band and in much more complicated environments such as indoor and other severe fading channels. 1.2288 Mcps might not suffice to support high-quality wireless transport simply due to channel bandwidth and more receiving techniques are needed to meet basic link performance.

Consequently, encouraging by early investigations and experiments in broadband CDMA, several proposed system architectures based on wide-band CDMA haven been seriously considered in the North America, Europe, and Asia. The fundamental frequency bandwidth is likely 5 MHz and its multiples primarily due to US PCS band assignment.

■ Basic Features of IS-95

In spite of so many detailed documentation and papers about IS-95, we are going to discuss some important characteristics that have strong impacts on various IMT-2000 designs.

- ❑ IS-95 applies pilot signals in the forward link to result in coherent demodulation. However, pilot signals can be not applied in the reverse link due to the limitation of conventional communication theory.

- IS-95 uses RAKE receiver structure to combat multipath fading.
- IS-95B uses hard handoff for high-speed data transmission, which sets up the initial trial on multimedia wireless communications in wide area.

In the IMT-2000, there are several technologies that can be applied to enhance IS-95 for wide-band CDMA serving multimedia traffic:

- More effective speech/video coding
- 2-D antenna array and RAKE receiver
- new synchronization techniques to enhance link performance
- new error control technology
- HOS to enhance detection performance
- Multicode, multirate CDMA for multimedia/Internet applications

In above, we shall further understand the complicated operating environments and their channel modeling. Furthermore, we shall understand system performance of different approach. This is the core target for this project.

- Family of Standards for IMT-2000 and Need of Software Radio

For the purpose of smooth migration, IMT-2000 is considering a family of air-interface definitions conforming to the same technology.

How to develop a software radio that can be flexibly adjusting to operating environments is of highest interest for IMT-2000. Under the efforts of 3GPP and 3GPP2, programmable or software radio has much more need to be realized. This project therefore concentrates on developing technology to solve this challenge. Since it involves a wide range of expertise, professors with good research credential in related areas must work together to develop this group project and can easily justify the need of grouping effort.

This project is expected to play one of the leading roles in this area. Recent activities from group members in IEEE Communications Magazine and IEEE Personal Communications Magazine, organization of 2001 IEEE Workshop on Signal Processing Advances in Wireless Communications, and much more publications, demonstrate this group's active role in software radio research.

研究方法

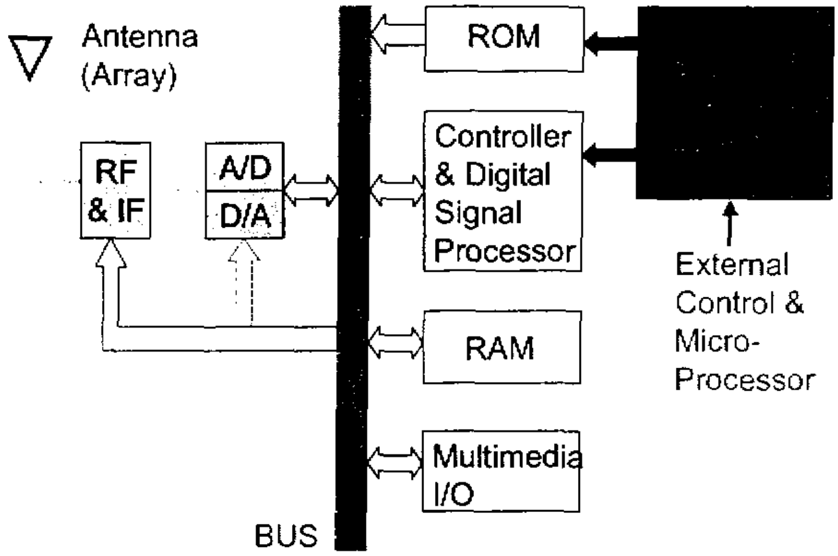
在第三代行動通訊系統 IMT-2000 中，寬頻分碼多工(wideband CDMA)預計將是重要的系統架構，而基於系統的平順演進及彈性系統運作，軟體無線電預期是 IMT-2000 中最重要的研究之一。然而軟體無線電不僅是一項整合型科技並且至今尚未成熟。為完成 IMT-2000 為國家型電信計畫目標，發展出適用的軟體無線電架構，整合性的研究以克服許多大不相同的問題仍是必需的。

下列課題為軟體無線電有關通訊及數位訊號處理之重要研究:

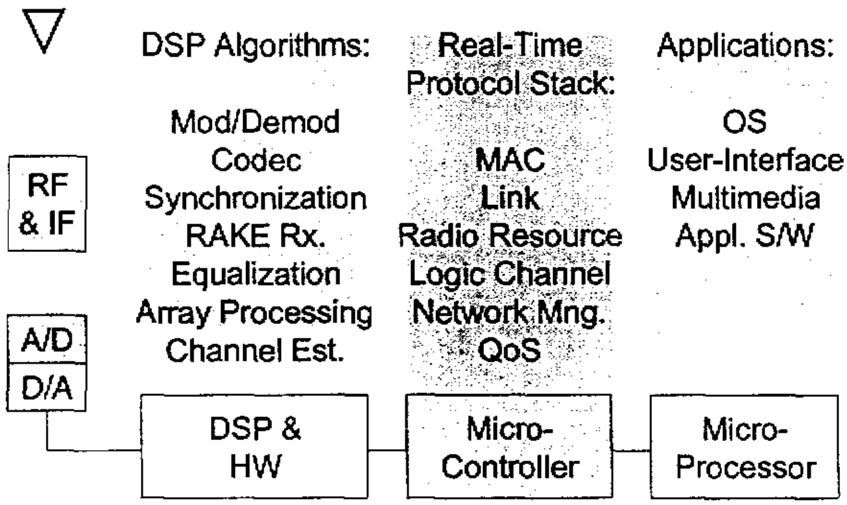
- Modulation and error control techniques
- Adaptive array
- HOS techniques
- Multi-carrier CDMA
- Multi-code CDMA
- Synchronization
- Speech/video coding
- System performance
- Radio resource for CDMA multimedia communications
- Equalization
- RAKE receiver
- Channel Estimation
- Hardware and software architecture
- Software integration

我們以下列的硬體架構及軟體架構做為軟體無線電各部分之研究設計參考，充分整合計劃研究。

Proposed Radio Architecture



Proposed Software Architecture



重點說明:

1. 軟體無線電技術是目前無線通訊中最主要的 technology 之一，不論民用軍用均有極大的應用價值。國內外的研究機構，系統製造商，軍事系統業者，行動電話服務業均為此先進研究成果之使用者。

2. IMT-2000 軟體無線電是所明列研究重點，已於上年度通過審查。

3. 本計畫目前已明列出軟硬體架構，並針對 IMT-2000 Wide band CDMA 技術做研究及 DSP 的發展，並有多家廠商及研究單位參與技術討論交流，並與世界先進研究單位交流，本計畫的研究資訊及成果經驗均可對所用者有明顯之貢獻，本計畫並希望在國際上 IMT-2000 研究及 Software radio 研究佔有一席之地。(主持人之一已在 ICC98，ICC99 發表 IMT-2000 tutorial 及 IEEE PC Magazine 的 Software Radio 的專刊主編，並在 IEEE Communication Magazine Nov. 1999 發表 Software radio architecture 論文，證明已有好的開始。)

4. 本計畫運用 DSP 技術去實現無線通訊系統，不論在學理,智財,實用上對電信均有相當之貢獻。

進行步驟

Based on above discussion, there are 7 projects to cover our research items and major tasks:

- ❑ More effective speech/video coding
- ❑ 2-D antenna array and RAKE receiver
- ❑ New synchronization techniques to enhance link performance
- ❑ New error control technology
- ❑ HOS to enhance detection performance
- ❑ Multicode, multirate CDMA for multimedia/Internet applications
- ❑ Channel measurement and modeling

Although the software radio is still not a clearly defined term, it is clear that we must transform all communication implementation into digital signal processing so that a programmable platform can be built to carry out the communication transmitter and receiver functions. Wide-band CDMA for IMT-2000 is a very complicated system and it requires a huge amount of processing power. Fortunately, TI recently developed a new generation DSP engine, C6X digital signal processor, to deliver 1600 MIPS that seems to be able to meet our need. Therefore, we select it as our platform. Since every sub-project in this group project needs certain degree of DSP research, we thus ask 5 TI C6x development tools in 5 universities. This is a critical part of our successful research and we do need this funding. In the mean time, we need to deal with complicated simulations, analytical verification, and design process. Each sub-project needs a workstation and it is also a must. We spread the need into 2 or 3 years for financial purpose.

The next challenge is how to select a direction for research since there are quite a few proposals. Luckily, the principle investigator attended several recent TIA 45.5 CDMA cellular standard meetings and recent ITU-R TG8/1 meeting. After extensive study and discussion among professors, we note that the technology difference of various approaches is not significant at all. The principle investigator will describe this conclusion in details in IEEE ICC tutorial and an IEEE invited journal submission. That is why we come out above working list and project list. Especially, software radio must provide flexibility and we are sure to work in the right direction. However, we also realize the importance to catch state-of-the-art technology change. This project will cooperate with major companies in standard efforts in north America and Europe.

This project will start from theoretical research, then DSP algorithm development, and finally real verification. We wish in the second stage of this project (4^{th} and 5^{th} years) that we can work with local industry to deliver our technologies into real product R&D. In the first stage, based on earlier successful research experience by all 7 professors in this group project, we are confident in the initial achievement. We also have a close tie among all sub-projects and shall proceed careful and detailed mutual discussion very frequently. We are planning to organize 2001 IEEE Workshop on Signal Processing Advances in Wireless Communications to get together world-wide experts into an open forum to promote this related research. We believe that the national program shall provide world-level leadership. Professor Jin-Fu Chang will be the first workshop chair and world-wide experts are in the contact.

Another major challenge is the integration of different DSP efforts from each part. We initially request a major support of a post-doctor position in the second and third year to work on integrated architecture of DSP communication system for IMT-2000 software radio technologies in this Project. In the third year, another assistant to help integration is asked. Not to limit our scope, we believe that it is the minimum support for this project for practical success.

We shall also coordinate with another national program proposal with antenna technology implementation from National Taiwan University to approach a total solution of IMT-2000 wide-band CDMA software radio.

Up to this point, we already comes out preliminary design for our W-CDMA software radio. The following is the **UNPUBLISHED** abstract of our design (**please keep confidential** due to intellectual property from this project):

■ Software Realization of W-CDMA Radio[1]

Introduction

Software radio provides a wide range of interests for future wireless communications:

- ❑ flexibility to fit multi-standards in a common platform

- flexibility to fit operation in different frequency bands
- ability to normalize to a family of chip rates from a single clock
- flexibility to serve variety of applications
- ease of upgrades and updates.

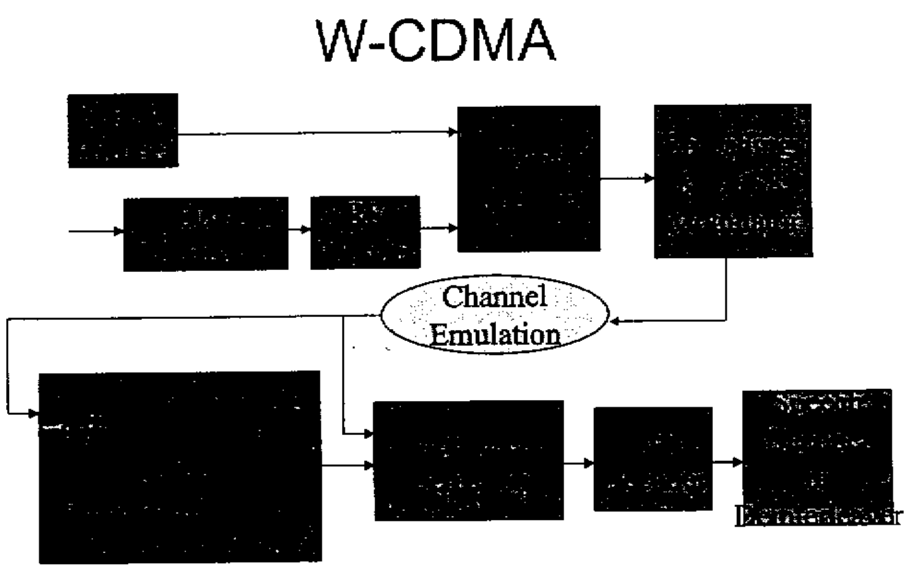
However, the realization of software radio is not an easy task by changing traditional hardware realization into a hardware platform running effective software to detect signal. The difficulties include

- Effective algorithms for pure DSP functions: We have to turn all traditional hardware design into digital signal processing algorithms, at least for IF and baseband functions. These algorithms result in huge amount of processing and our estimation is more than thousands of MIPS.
- High-speed ADC with large dynamic range: The near-far ratio could be more than 90 dB.
- Effective hardware platform to support tremendous processing power at least thousands of MIPS:
- High battery power consumption due to high clock frequencies and large amount of DSP MIPS. An efficient power management scheme may be necessary.

As W-CDMA is going to be the base for the third generation wireless systems, our initial goal here is to develop the design of hardware platform and software structure based on this platform, for W-CDMA. IN other words, we shall develop the programmable design of a W-CDMA radio to demonstrate feasibility of future software radio.

■ Target W-CDMA Systems

The following figure illustrates our target system:



■ Speech coding

The possible speech coding standards, including TIA/EIA-IS-96A, IS-127, IS-733, G.729, and RPE-LTP, from Europe, North America, and Asia, for IMT-2000 wideband CDMA are studied. The bitrate ranges from less than 4 to over 13 kbps. The error protection and error concealment of speech coding are very important in the environment of wireless channels. We have studied IS-96 and perform simulations in fading channel conditions [chang1]. Uneven error protection and selective concealment can efficiently improve the speech quality in wireless communications.

■ Video coding

We study video coding for IMT-2000 from both the standard-compatible and advanced technique approaches. The proposed bitrates for video coding in IMT-2000 ranges from 2.048 Mbps for asymmetric transmission for indoor or in slow moving scenarios to 64 Kbps for satellite communications. Both MPEG and H.263 could be used in this environment. Because of the error propagation property, error tracking and error concealment are crucial to wireless communications. We have developed a precise error tracking algorithm [chang2] for H.263. It improves the video quality significantly. For the new technique part, zero-tree wavelet coding is studied. The error resilience of wavelet coding is also enhanced by interleaving, re-synchronization, and concealment [chang3].

■ Error Control Codes and Decoding

Turbo codes will be considered for use in the target W-CDMA systems for high-quality high-rate traffics, and convolutional codes for the rest traffics. The main goal of our research is towards designing a flexible and efficient decoder architecture. We have developed a unified structure for various trellis-based soft-output decoding algorithms, including the BCJR algorithm, RTMEP algorithm, SOVA, modified SOVA, and several new algorithms, suitable for use in the turbo decoders [WWC00]. For different soft-output algorithms, the proposed structure has the same set of storage, but differ in the ways how the stored quantities are updated and how the soft outputs (reliability estimates) are computed from these quantities. Detailed complexity and performance comparisons between structures for different soft-output algorithms have also been conducted. We are currently studying efficient implementation of the interleaver (and deinterleaver) used in the turbo encoders and decoders. Other work in progress includes investigation of stability and convergence properties of iterative (turbo) decoding.

■ 2-D Rake Receiver

With the help from Prof. T.S. Lee, [WuC99] proposed a general 2-D RAKE receiver in general multi-path fading channel and ways to implement it in linear complexity.

■ Synchronization (frequency offset estimation & phase recovery & timing recovery)

For multiuser timing recovery, we have developed the theoretical framework [CH00] from multiuser detection to show that acquisition and tracking are two necessary steps toward effective synchronization. Tracking can be achieved by traditional descend search algorithms. Acquisition is generally a NP hard problem. Effective linear complexity acquisition schemes have been developed and pilot-aided scheme is further suggested [CHM99]. Joint estimation of amplitude, phase and timing in fading channels has been demonstrated realizable with linear complexity.

In the down-link, the block diagram of the proposed synchronization subsystem is shown as in Fig. A. A cell search based upon passive correlation and non-coherent integration is performed jointly with a coarse frequency offset estimation.

The received signal, after frequency offset compensation, is used for code group identification and long code synchronization, as specified in the 3gpp specification. Coherent open loop fine code tracking (path tracker) is proposed in this synchronizer, because the traditional close-loop tracking may not be workable in a multipath environment. The received signal is then de-spread with a fine code phase estimation, and the residual frequency offset is estimated accurately with diversity combining, due to the fact that the frequency offset is the same in all the paths (fingers) of the channel. Finally, after fine frequency compensation, the channel can be estimated easily with the maximum likelihood criterion, and RAKE combining follows.

■ Equalization (including interference rejection)

To increase system capacity with acceptable performance in the presence of multipath, multi-input multi-output (MIMO) equalization processing and channel estimation prior to multiuser detection are needed by W-CDMA systems for effectively removing effects of the multipath channel for each user and rejecting co-channel interference in the mean time. The design of the equalizer for multiuser W-CDMA systems basically includes nonblind equalization and blind equalization. The former requires the channel information that can be estimated through regular training procedure at the expense of system resources. The latter performs equalization only with the received signal through more complicated signal processing procedures. Effective equalization is crucial to improving the performance of the following multiuser detection.

Nonblind equalizers such as inverse filters, LMMSE equalizers, successive cancellation type equalizers, multistage type equalizers etc., and blind equalizers such as successive cancellation type equalizers, inverse filter type equalizers and minimum output energy type equalizers etc. need be implemented through software real-time running on a hardware platform with sufficient computing power for their practical use in software radio. Practical software implementation of equalization processing will become feasible by simplifying equalization algorithms on one hand and using DSP processors with higher computing capability on the other hand. The former is attainable through adaptive filtering or new simple algorithms and the latter depends on the progress of IC and VLSI technologies. We shall realize general equalization in WCDMA software radio.

■ Multuser Detection (including multu-rate multi-code detection)

We have proceeded analysis on de-correlating receiver and other linear multiuser detection receivers in the general multi-path environments based on mutli-rate and multi-code WCDMA. It turns out that these linear multiuser detection schemes can still function normally except minor difference. For multimedia traffic, the strategy for power assignment and power allocation has been developed for multi-rate multi-code W-CDMA. [ChuC99] proposed strategy for multiuser receiver and its power allocation strategy for WCDMA.

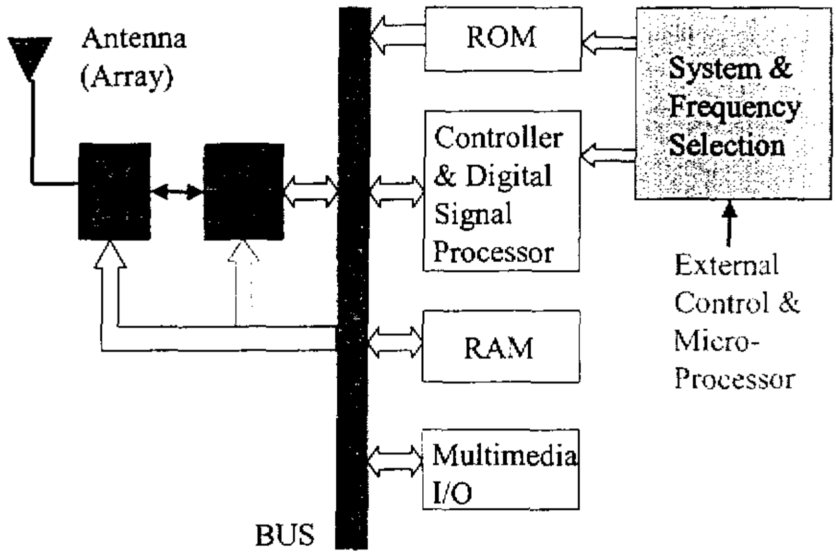
■ Channel Estimation and Software Integration

This is a newly added research subject in this project. We shall integrate channel estimation and RAKE receiver into our software radio implementation to complete design. In the mean time, this part plays a natural role in software integration for the whole design, due to its wide interaction with various parts of system.

■ Implementation

The proposed programmable radio architecture is as follows.

Proposed Radio Architecture

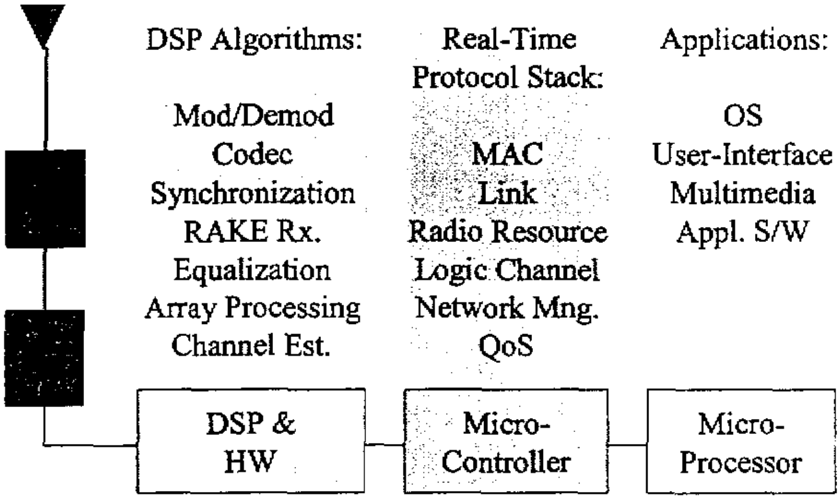


The major challenges lie in the requirement of multiple digital signal processors and thus the effective architecture to handle multi-tasking among these digital signal processors.

The proposed software subroutines can be found in the following. At this stage, we only concentrate on the DSP part and very minimum link control function.

Again, the major challenge is the parallel processing of different algorithms and effective arrangement of computations.

Proposed Software Architecture



研究成果與討論

Based on the programmable transceiver architecture designed for three categories of OFDM-CDMA systems, a general linear-complexity near-far resistant programmable multiuser detection structure using an antenna array is proposed. Various system scenarios, including MC-CDMA, MC-DS-CDMA, and MT-CDMA, can be realized by adjusting system parameters for the intended multiuser detection scheme. To enhance multiuser detector, two schemes to estimate channel parameters and array responses are proposed. By taking complexity into consideration, we further propose a joint detection strategy to reduce the complexity. From simulations, we verify satisfactory performance in multipath channels. The details are listed in the appendix (附錄), while for the details of other subprojects readers are suggested to refer to their individual reports.

附錄

Abstract

Based on the programmable transceiver architecture designed for three categories of OFDM-CDMA systems[9], a general linear-complexity near-far resistant programmable multiuser detection structure using an antenna array is proposed. Various system scenarios, including MC-CDMA [5], MC-DS-CDMA [6], and MT-CDMA [7], can be realized by adjusting system parameters for the intended multiuser detection scheme. To enhance multiuser detector, two schemes to estimate channel parameters and array responses are proposed. By taking complexity into consideration, we further propose a joint detection strategy to reduce the complexity. From simulations, we verify satisfactory performance in multipath channels.

I. Introduction

Code division multiple access (CDMA) adopting power control and good maintenance of the orthogonality of codes is proved to provide desirable features for cellular systems [12]. In practical CDMA applications, however, there are two technical challenges. First, due to the demands of higher data rates and quality of service, an effective transmission to combat potential frequency selective multipath fading at wider signal bandwidth is desired. Second, a near-far resistant receiver is needed to remove multiaccess interference (MAI) so that system capacity can be greatly enhanced.

Combing CDMA with orthogonal frequency division multiplexing (OFDM) techniques [3] has been proposed to solve the first issue as well as increase bandwidth efficiency. Three schemes [4] well known as multicarrier CDMA (MC-CDMA) [5], multicarrier direct sequence CDMA (MC-DS-CDMA) [6] and multitone CDMA (MT-CDMA) [7] attract lots of attention in the area of wideband CDMA, which greatly reduce the complexity of equalization or RAKE receiver structure. On the other hand, CDMA combined with OFDM facilitates finer partition of radio resources in time domain, frequency domain, and code domain, so that more effective radio resources allocation might be possible provided an effective bandwidth allocation mechanism is available [8] [9]. As those schemes have their advantages in different environments, a programmable transceiver architecture for general OFDM-CDMA [9] was developed based on the unified framework known as OFCDMA [10] so that various system scenarios can be realized by adjusting system parameters.

Using multiuser detection techniques [11] to solve the near-far problem has been considered a desirable feature in the third and future generations of CDMA systems. The optimum multiuser detector has computational complexity of exponential order with respect to the number of users which is difficult in practical applications. Therefore, a family of suboptimum multiuser detection schemes of acceptable complexity have been studied (e.g. [13] and [11]), while a general multiuser detection structure for OFDM-CDMA systems is still not available but is desired for future cellular systems.

In addition, the smart antenna, processing signals in both space and time domains with an antenna array, can introduce more degrees of freedom such that the system performance is significantly improved. Recently, it has been proposed for cellar systems to enhance the overall capacity, including spectrum efficiency, channel capacity, and coverage range, and has attracted lots of attention thanks to fast innovations of digital signal processors (DSPs). The increase in spectrum efficiency is due to the antenna array's capability of producing virtual channels. Those virtual channels provide an alternative to discriminate signals among users if their directions of arrivals (DOAs) differ and hence may help mitigating MAI.

The programmable OFDM-CDMA system is highly desired because of its adaptability to multisystems and capability of efficiently managing radio resources especially for the upcoming multimedia communication age. However, it has seldom been studied. By taking all above issues into account, we are going to present a thorough study of programmable OFDM-CDMA multiuser detection. This paper is organized as follows. We derive a multiuser receiver structure with an antenna array subject to multiray Rayleigh fading in section III after introducing the programmable OFDM-CDMA transceiver architecture in section II. This proposed receiver provides not only a general space-time-frequency processing structure for CDMA-based communication systems, but also a prototype for other multiple access approaches such as orthogonal division multiple access (OFDMA), and space division multiple access (SDMA). Then, a family of linear complexity programmable space-time multiuser detection schemes are presented, and the programmability are verified with illustrations in section VI. Since the channel parameters and array responses are necessary for those schemes, two approaches to estimate them are illustrated in section V. Then the computational complexity of the proposed structure is considered, and a joint detection strategy based on interference cancellation is further proposed to reduce the complexity in section VI. The performance about the proposed detection structure is investigated, and finally some simulation examples are shown with practical channel considerations in section VIII followed by conclusion.

II. The programmable OFDM-CDMA transceiver architecture

It is well known that three different types of systems apply OFDM techniques to conventional CDMA: multicarrier-CDMA (MC-CDMA) [5], multicarrier direct sequence-CDMA (MC-DS-CDMA) [6], and multitone-CDMA (MT-CDMA) [7]. Those OFDM-CDMA systems can be described by a unified framework known as OFCDMA proposed in [10]. The k^{th} user's transmitter structure is shown in Figure 1 [10]. The original data stream $b_k(t)$ with period T' is serial-to-parallel (S/P) converted to J groups of data streams $b_{kj}(t)$ with period T. For each group, M identical branches of $b_{kj}(t)$ spread by signature waveforms $c_{km}(t)$ respectively modulate the subcarriers $cos(w_{jm}t)$ for $m = 1, 2, ...M$. Let $w_o = 2\pi/T_o$ be the frequency separation between adjacent subcarriers. The relation between two subcarriers is given by

$$w_{ab} - w_{cd} = \frac{2\pi}{T_o}(a-c) + \frac{2J\pi}{T_o}(b-d).$$

Figure 1 can be accommodated to those three multicarrier CDMA systems. We get the MC-CDMA by setting $c_{km}(t) = c_k^m$ with $(c_k^1 c_k^2 ... c_k^M)$ constituting a spreading code, $T = JT'$, and $T_o = T$; the MC-DS-CDMA by setting $c_{km}(t) = c_k(t) = (c_k^1 c_k^2 ... c_k^{N_{MD}})$, $T = MJT'$, and $T_o = T/N_{MD}$; the MT-CDMA by setting $M = 1$, $c_{k1}(t) = c_k(t) = (c_k^1 c_k^2 ... c_k^{N_{MD}})$, $T = JT'$, and $T_o = T$. The details and the differences among the three are discussed in [4].

The digital transmitter of the OFCDMA, known as programmable OFDM-CDMA [9], is depicted in Figure 2(a) [9]. M identical branches of data streams at rate of $f' = 1/T'$ are spread by the sequences $(c_{km}^1 ... c_{km}^N)$ for $m = 1, 2, ...M$, and then block-multiplexed (Block Mux) to J groups. The MJ samples are zero-padded and then a GH-point inverse fast Fourier transform (IFFT) is operated, where $G = \lceil (MJ-1)/H + 1 \rceil$, $H = T_o/T_c$, $\lceil x \rceil$ is the smallest integer no less than x, and T_c is the chip duration of the signature waveforms. After IFFT, only the $\left(\frac{n}{N}\right)^{\text{th}}$ fraction of the GH samples is retained for $n = 1, 2, ...N$, where the operation is called shift windowing. Finally they are passed to an analog low-pass filter

(LPF) with bandwidth

$$BW = \frac{MJ - 1 + \frac{T_o}{T_c}}{2T_o}.$$

The i^{th} user's RAKE receiver is depicted in Figure 2(b) [9], and the details of the n^{th} finger is shown in Figure 2(c) [9]. The synchronized signal is sampled at rate of $f_s = G/T_c$. After each S/P conversion the GH/N samples are shift zero padded to GH ones, by which we mean that the GH/N samples are zero-padded to GH samples and end-roundly shifted by nGH/N samples for $n = 0, 1, ...N - 1$. A GH-point FFT is next in operation and then only the first MJ samples are retained after windowing. At last, the MJ parallel branches are block-multiplexed to the M tapped-delay-lines for despreading. The final outputs are the sufficient statistics for detection. The programming scenarios of the OFDM-CDMA transceiver is described in details in [9].

III. Optimum multiuser detection

Consider the uplink case for an asynchronous OFDM-CDMA communication system. In the following, all signals are presented by their low-pass equivalent complex envelopes. The k^{th} user's transmitted signal is

$$s_k(t) = \sqrt{2a_k/JM} \sum_{j=1}^{J} \sum_{m=1}^{M} \sum_{p=-P}^{P} b_{kj}(t) c_{km}(t - pT) e^{\imath w_{jm} t}, \tag{1}$$

where a_k is the transmission power, $b_{kj}(t)$ is the data stream at the subcarrier $e^{\imath w_{jm} t}$; $\imath \equiv \sqrt{-1}$, and $c_{km}(t)$ is the corresponding T-durationed signature waveform. The j and m are called index of independent and dependent subcarriers respectively because $e^{\imath w_{jm} t}$ carries independent information for different j while identical information for all m. The number of subcarriers and transmitted symbols on each subchannel are JM and $2P + 1$ respectively.

An X-element antenna array is deployed at the receiver. At the x^{th} antenna, L-ray multipath fading environment is assumed in the $(jm)^{\text{th}}$ subchannel for the k^{th} user with channel impulse response:

$$h_{kjmx}(t) = \sum_{l=1}^{L} u_{kjmlx} g_{kjml} \delta(t - t_{kl}). \tag{2}$$

Where g_{kjml} and u_{kjmlx} are the fading coefficients modeled by zero-mean complex Gaussian random variables (r.v.s) and antenna responses respectively for the l^{th} path of the k^{th} user's signal in the $(jm)^{\text{th}}$ subchannel, and t_{kl} is the corresponding ray delay. Let θ_{kl} denote the DOA of the k^{th} user's signal along the l^{th} path with respect to the half-wavelength-spaced linear antenna array; then the antenna response is given by $u_{kjmlx} = e^{i\pi(1+w_{jm}/w)(x-1)\sin(\theta_{kl})}$, where w is the RF carrier frequency. We may assume that u_{kjmlx} are identical for all j and m when $w_{jm}/w << 1$. It is reasonable to assume that g_{kjml} are independent for different k and l. And if the transmission bandwidth of each subchannel is smaller than the channel coherent bandwidth, we may assume g_{kjml} independent for different j and m; otherwise, some issues must be carefully considered and will be discussed in section VII.

Let $b_{kj}[p]$ represent the p^{th} symbol of $b_{kj}(t)$. The received signal at the x^{th} antenna is

$$v_x(t) = \sum_{k=1}^{K}\sum_{j=1}^{J}\sum_{m=1}^{M}\sum_{l=1}^{L}\sum_{p=-P}^{P} \beta_{kjmlx} b_{kj}[p] c_{km}(t - pT - \tau_k - t_{kl}) e^{iw_{jm}t} + \eta_x(t),$$

where $\beta_{kjmlx} \equiv \sqrt{2a_k/JM}\, u_{kjmlx} g_{kjml}$, and τ_k is k^{th} user's propagation delay. Without loss of generality, we consider the phase shift $-w_{jm}(\tau_k + t_{kl})$ into g_{kjml}; $\eta_x(t)$ are independent and identically distributed (i.i.d.) complex AWGN processes with variance σ^2 for all x; K is the number of users. We introduce a block matrix construction method to facilitate equation formulation, in which we use a variable with less subscripts to represent a block vector (a bold-faced lowercase letter) or a block matrix (a bold-faced capital letter) constructed from the same variable with more subscripts by sequencing the additional ones. For example, $\boldsymbol{x}_{km}$ is a larger dimensional vector constructed from a smaller dimensional vector $\boldsymbol{x}_{kjml}$ by

$$\boldsymbol{x}_{km} \equiv [\boldsymbol{x}_{k1m1}^T, \boldsymbol{x}_{k2m1}^T, \dots, \boldsymbol{x}_{kJm1}^T, \boldsymbol{x}_{k1m2}^T, \dots, \boldsymbol{x}_{kJm2}^T, \dots, \boldsymbol{x}_{kJmL}^T]^T;$$

$\boldsymbol{U}_{l,n}$ is constructed from $\boldsymbol{U}_{kml,isn}$ by

$$\boldsymbol{U}_{l,n} \equiv \begin{pmatrix} \boldsymbol{U}_{11l,11n} & \cdots & \boldsymbol{U}_{11l,K1n} & \cdots & \boldsymbol{U}_{11l,KMn} \\ \vdots & \ddots & & \ddots & \vdots \\ \boldsymbol{U}_{K1l,11n} & & \boldsymbol{U}_{K1l,K1n} & & \boldsymbol{U}_{K1l,KMn} \\ \vdots & \ddots & & \ddots & \vdots \\ \boldsymbol{U}_{KMl,11n} & \cdots & \boldsymbol{U}_{KMl,K1n} & \cdots & \boldsymbol{U}_{KMl,KMn} \end{pmatrix}.$$

Assume that all the subchannels are subject to slow fading. Let $\boldsymbol{B}$, the amplitude matrix, be a $KJMLX \times KJMLX$ diagonal matrix with the $(kjmlx)^{\text{th}}$ diagonal term corresponding to β_{kjmlx}. The jointly optimum multiuser detector given $\boldsymbol{b}'$ transmitted is [1][2] $\arg\max_{\boldsymbol{b}} Pr(\,\boldsymbol{v}(t)|\boldsymbol{b}', \boldsymbol{B})$, where $\boldsymbol{b}$ is constructed from $\boldsymbol{b}[p]$ and $\boldsymbol{v}(t)$ from $v_x(t)$. Assuming that the maximum delay between users is less than PT and that $\boldsymbol{b}'$ are equally probable, it is equivalent to selecting $\boldsymbol{b}$ that minimizes

$$\int_{-\infty}^{\infty} \sum_{x=1}^{X} |v_x(t) - \sum_{k=1}^{K}\sum_{j=1}^{J}\sum_{m=1}^{M}\sum_{l=1}^{L}\sum_{p=-P}^{P} \beta_{kjmlx} b_{kj}[p] c_{km}(t - pT - \tau_k - t_{kl}) e^{\imath w_{jm} t}|^2 dt,$$

or maximizes

$$\Omega(\,\boldsymbol{b}) = 2Re\left\{ \sum_{p=-P}^{P} \boldsymbol{b}^H[p]\ \boldsymbol{J}^H \boldsymbol{B}^H \boldsymbol{v}[p] \right\} - \sum_{p,q=-P}^{P} \boldsymbol{b}^H[p] \boldsymbol{J}^H\ \boldsymbol{B}^H \boldsymbol{R}[p-q] \boldsymbol{B} \boldsymbol{J} \boldsymbol{b}[q]. \quad (3)$$

where $\boldsymbol{J}$, the diversity matrix, is a $KJMLX \times KJ$ matrix constructed from $KJ \times KJ$ identity matrices $\boldsymbol{I}$ by $\boldsymbol{J} = [\boldsymbol{I}, \boldsymbol{I}, ..., \boldsymbol{I}]^T$, $(.)^H$denotes the conjugate transpose operation of a matrix, and $Re\{.\}$ denotes the real part of a complex number. The $KJMLX \times 1$ observation vector $\boldsymbol{v}[p\ ]$ is constructed from

$$v_{kjmlx}[p] = \int_{-\infty}^{\infty} v_x(t) c^*_{km}(t - pT - \tau_k - t_{kl}) e^{-\imath w_{jm} t} dt, \quad (4)$$

which can be produced via the programmable OFDM-CDMA receiver shown in Figure 2(b); $\{\}^*$ denotes the complex conjugate. Here we use k and i to represent the indexes for users; j and r for independent subcarriers; m and s for dependent subcarriers; l and n for rays; x and y for antennas. The $KJMLX \times KJMLX$ correlation matrix $\boldsymbol{R}[p-q]$ is constructed from the asynchronous code correlation function

$$R_{kjmlx,irsny}[p-q] = \delta_{x,y} \int_{-\infty}^{\infty} c^*_{km}(t - pT - \tau_k - t_{kl}) c_{is}(t - qT - \tau_i - t_{in}) e^{-\imath \Delta w_{jm,rs} t} dt, \quad (5)$$

where $\Delta w_{jm,rs} = w_{jm} - w_{rs}$; $\delta_{x,y}$ is the Kronecker delta function, which is introduced due to the independent assumption of noise processes at different antennas. Note that $\boldsymbol{R}_{x,x}$ are identical for all x and $\boldsymbol{R}_{x,y} = \boldsymbol{0}$ if $x \neq y$. Because $R^*_{irsny,kjmlx}[q-p] = R_{kjmlx,irsny}[p-q]$; we have $\boldsymbol{R}[p-q] = \boldsymbol{R}^H[q-p]$. $\boldsymbol{v}[p]$ may be expressed as

$$\boldsymbol{v}[p] = \sum_{q=-P}^{P} \boldsymbol{R}[q-p] \boldsymbol{B} \boldsymbol{J} \boldsymbol{b}[q] + \boldsymbol{\eta}[p]. \quad (6)$$

where $\boldsymbol{\eta}[p]$ is a $KJMLX \times 1$ vector constructed from the noise components

$$\eta_{kjmlx}[p] = \int_{-\infty}^{\infty} \eta_x(t) c_{km}^*(t - pT - \tau_k - t_{kl}) e^{-\imath w_{jm} t} dt.$$

We see that

$$E[\boldsymbol{\eta}[p]\boldsymbol{\eta}[q]^H] = \sigma^2 \ \boldsymbol{R}[p-q],$$

because

$$\begin{aligned}
& E[\eta_{kjmlx}[p]\eta_{irsny}^*[q]] \\
= & \int_{-\infty}^{\infty}\int_{-\infty}^{\infty} E[\eta_x(t)\eta_y^*(u)] c_{km}^*(t - pT - \tau_k - t_{kl}) c_{is}(u - qT - \tau_i - t_{in}) e^{-\imath w_{jm} t} e^{\imath w_{rs} u} dt du \\
= & \sigma^2 R_{kjmlx,irsny}[p-q].
\end{aligned}$$

The general signal model shown in (6) may degenerate to some special cases if some of the $\{K, J, M, L, X\}$ are set to 1. For example, let $J = M = L = X = 1$, and then (6) is reduced

$$\boldsymbol{v}[p] = \sum_{q=-P}^{P} \boldsymbol{R}[q-p]\boldsymbol{B}\boldsymbol{b}[q] + \boldsymbol{\eta}[p].$$

This special case is the signal model for the asynchronous single carrier CDMA in a flat-fading channel. Letting $J = M = 1$, we have from (6)

$$\begin{aligned}
\boldsymbol{v}_{lx}[p] &= \sum_{n=1}^{L}\sum_{y=1}^{X}\sum_{q=-P}^{P} \boldsymbol{R}_{lx,ny}[q-p\ ]\boldsymbol{B}_{ny}\boldsymbol{b}[q] + \boldsymbol{\eta}_{lx}\ [p] \\
&= \sum_{n=1}^{L}\sum_{q=-P}^{P} \boldsymbol{R}_{lx,nx}[q-p\ ]\boldsymbol{B}_{nx}\boldsymbol{b}[q] + \boldsymbol{\eta}_{lx}\ [p], \qquad (7)
\end{aligned}$$

where $\boldsymbol{v}_{lx}[p]$ =$[v_{1lx}[p],\ v_{2lx}[p], \ldots,\ v_{Klx}[p]]^T$ are the receiver outputs locked to the l^{th} path at the x^{th} antenna for users $k = 1, 2, \ldots K$. This special case of (7) is just the space-time signal model in a multipath CDMA channel proposed in [14]. Now let $X = 1$, we get from (6)

$$\boldsymbol{v}_{ml}[p] = \sum_{s=1}^{M}\sum_{n=1}^{L}\sum_{q=-P}^{P} \boldsymbol{R}_{ml,sn}[q-p]\boldsymbol{B}_{sn}\boldsymbol{b}[q] + \boldsymbol{\eta}_{ml}[p].$$

where $\boldsymbol{v}_{ml}[p\ ] = [v_{11ml}^T[p],\ v_{K1ml}^T[p], \ldots,\ v_{KJml}^T[p]]^T$ is the matched filter outputs locked to the l^{th} path at the m^{th} dependent subcarrier for independent subcarriers $j = 1, 2, \ldots J$ and users $k = 1, 2, ...K$. This special case is the signal model for general OFDM-CDMA with only one receiving antenna.

IV. Programmable multiuser detection

The jointly optimum multiuser detection for the OFDM-CDMA is just the maximum likelihood sequence detection, which searches for $\boldsymbol{b}$ belonging to some finite set (for example, $\{\pm 1\}^{KJ(2P+1)}$ for binary phase shift keying (BPSK)) such that (3) is maximized. This can be implemented by Viterbi Algorithm [2]. Since the complexity grows exponentially with the size of $\boldsymbol{b}$, suboptimum joint detection methods are preferable. We observe that the signal model for OFDM-CDMA shown in (6) has the correspondence to that for CDMA. Therefore, we can apply multiuser detection techniques in CDMA directly to OFDM-CDMA possibly by some modifications. Moreover, all that needed for the proposed multiuser detection can be acquired using the programmable OFDM-CDMA transceiver. We then demonstrate the programmability of the structure.

Let $\widetilde{\boldsymbol{R}}$ and $\widetilde{\boldsymbol{B}}$ be $KJMLX(2P+1) \times KJMLX(2P+1)$ matrices with $\widetilde{\boldsymbol{R}}_{p,q} = \boldsymbol{R}[p-q]$, $\widetilde{\boldsymbol{B}}_{p,q} = \delta_{p,q}\boldsymbol{B}$; $\widetilde{\boldsymbol{\eta}}$, $\widetilde{\boldsymbol{v}}$ be $KJMLX(2P+1) \times 1$ vectors with $\widetilde{\boldsymbol{\eta}}_p = \boldsymbol{\eta}[p]$ and $\widetilde{\boldsymbol{v}}_p = \boldsymbol{v}[p]$; $\widetilde{\boldsymbol{J}}$ be a $KJMLX(2P+1) \times KJ(2P+1)$ matrix with $\widetilde{\boldsymbol{J}}_{p,q} = \delta_{p,q}\boldsymbol{J}$. $\widetilde{\boldsymbol{b}}$ be a $KJ(2P+1) \times 1$ vector with $\widetilde{\boldsymbol{b}}_p = \boldsymbol{b}[p]$. Then (6) can be expressed as

$$\widetilde{\boldsymbol{v}} = \widetilde{\boldsymbol{R}}\widetilde{\boldsymbol{B}}\widetilde{\boldsymbol{J}}\widetilde{\boldsymbol{b}} + \widetilde{\boldsymbol{\eta}}.$$

For convenience, we ignore the tildes above those symbols, that is

$$\boldsymbol{v} = \boldsymbol{RBJb} + \boldsymbol{\eta}. \tag{8}$$

We use the two-step mechanism for detection: a soft signal estimator $\widehat{\boldsymbol{b}}$ is derived by some criterion first, and then passed to a hard limiter. For example, the single user maximum ration combining (MRC) estimator is given by

$$\widehat{\boldsymbol{b}} = \boldsymbol{J}^H \boldsymbol{B}^H \boldsymbol{v}, \tag{9}$$

and the hard limiter for BPSK is operated by $sign(Re(\hat{\boldsymbol{b}}))$, where $sign\{.\}$ is the signum function. And the single user equal gain combining (EGC) coherent estimator is given by

$$\widehat{\boldsymbol{b}} = \boldsymbol{J}^H \boldsymbol{\Phi}^H \boldsymbol{v}, \tag{10}$$

where $\boldsymbol{\Phi}$ is the phase of $\boldsymbol{B}$. Let $\boldsymbol{A}$ be a linear operator such that $\hat{\boldsymbol{b}} = \boldsymbol{Av}$ is the linear estimator for $\boldsymbol{b}$. The linear minimum mean square error (LMMSE) estimator can be derived by minimizing $E[\|\boldsymbol{b} - \boldsymbol{Av}\|^2]$, where the expectation is taken with respect to $\boldsymbol{b}$ and $\boldsymbol{\eta}$. Consider the covariance matrix of the error $\boldsymbol{b} - \boldsymbol{Av}$:

$$\begin{aligned} E[(\boldsymbol{b} - \boldsymbol{Av})(\boldsymbol{b} - \boldsymbol{Av})^H] &= E[\boldsymbol{bb}^H] - \boldsymbol{A}E[\boldsymbol{vb}^H] - E[\boldsymbol{bv}^H]\boldsymbol{A}^H + \boldsymbol{A}E[\boldsymbol{vv}^H]\boldsymbol{A}^H \\ &= \boldsymbol{I} + \boldsymbol{A}(\boldsymbol{RBJJ}^H\boldsymbol{B}^H\boldsymbol{R} + \sigma^2\boldsymbol{R})\boldsymbol{A}^H - \boldsymbol{J}^H\boldsymbol{B}^H\boldsymbol{RA}^H - \boldsymbol{ARBJ} \\ &= (\boldsymbol{I} + \sigma^{-2}\boldsymbol{J}^H\boldsymbol{B}^H\boldsymbol{RBJ})^{-1} + (\boldsymbol{A} - \boldsymbol{A}_{MS})E[\boldsymbol{vv}^H](\boldsymbol{A} - \boldsymbol{A}_{MS})^H, \end{aligned} \tag{11}$$

where

$$\boldsymbol{A}_{MS} = (\boldsymbol{J}^H\boldsymbol{B}^H\boldsymbol{RBJ} + \sigma^2\boldsymbol{I})^{-1}\boldsymbol{J}^H\boldsymbol{B}^H, \tag{12}$$

and $\boldsymbol{B}$ is assumed nonsingular. The identity in (11) can be checked by

$$\boldsymbol{A}_{MS}(\boldsymbol{RBJJ}^H\boldsymbol{B}^H\boldsymbol{R} + \sigma^2\boldsymbol{R}) = \boldsymbol{J}^H\boldsymbol{B}^H\boldsymbol{R},$$

and

$$(\boldsymbol{I} - \boldsymbol{J}^H\boldsymbol{B}^H\boldsymbol{RA}_{MS}^H)(\boldsymbol{I} + \sigma^{-2}\boldsymbol{J}^H\boldsymbol{B}^H\boldsymbol{RBJ})^{-1} = \boldsymbol{I}.$$

Because the correlation matrix $E[\boldsymbol{vv}^H]$ is nonnegative definite, the trace of the second term in (11) is always nonnegative, with equality when $\boldsymbol{A} = \boldsymbol{A}_{MS}$. Since $E[\|\boldsymbol{b} - \boldsymbol{Av}\|^2] = trace\{E[(\boldsymbol{b} - \boldsymbol{Av})(\boldsymbol{b} - \boldsymbol{Av})^H]\}$, $\boldsymbol{A}_{MS}$ is the desired solution.

Next if we treat $\boldsymbol{b}$ as an unknown constant, we are just dealing with a Gaussian linear model, in which the minimum variance unbiased estimator exists and is identical to the best linear unbiased estimator (BLUE)

$$\boldsymbol{A}_{BL} = (\boldsymbol{J}^H\boldsymbol{B}^H\boldsymbol{RBJ})^{-1}\boldsymbol{J}^H\boldsymbol{B}^H. \tag{13}$$

Similar to using EGC instead of MRC when the amplitude matrix is unavailable, (13) can be modified to

$$\boldsymbol{A}'_{BL} = (\boldsymbol{J}^H\boldsymbol{\Phi}^H\boldsymbol{R\Phi J})^{-1}\boldsymbol{J}^H\boldsymbol{\Phi}^H. \tag{14}$$

In addition, based on the purpose of the decorrelating detector in CDMA, which demodulates the received signal without error in the absence of background noise, the one in OFDM-CDMA is modified to

$$\boldsymbol{A}_{DC} = \boldsymbol{J}^H \boldsymbol{\Phi}^H \boldsymbol{R}^{-1}. \tag{15}$$

When $\boldsymbol{R}$ is singular, the inverse is replaced by the Moore-Penrose generalized inverse. Note that $\boldsymbol{A}_{BL}$, $\boldsymbol{A}'_{BL}$, $\boldsymbol{A}_{DC}$ all degenerate to the decorrelating operator in the single carrier, single antenna case.

For the interference cancellation based estimator, the general form can be expressed as

$$\hat{\boldsymbol{b}} = \boldsymbol{M}(\boldsymbol{v} - \boldsymbol{f}(\widetilde{\boldsymbol{b}})), \tag{16}$$

where $\hat{\boldsymbol{b}}$ is the updated estimator for the old one $\widetilde{\boldsymbol{b}}$. We get some typical types by setting $\boldsymbol{M}$ and $\boldsymbol{f}(\widetilde{\boldsymbol{b}})$. For example, for the decision feedback estimator, we set $\boldsymbol{f}(\widetilde{\boldsymbol{b}}) = (\boldsymbol{R} - \boldsymbol{I})\boldsymbol{B}\boldsymbol{J}sign(\widetilde{\boldsymbol{b}})$ and $\boldsymbol{M} = \boldsymbol{J}^H\boldsymbol{B}^H$; for the parallel interference cancellation (PIC) estimator, $\boldsymbol{f}(\widetilde{\boldsymbol{b}}) = (\boldsymbol{R} - \boldsymbol{I})\boldsymbol{J}\widetilde{\boldsymbol{b}}$, and $\boldsymbol{M} = \boldsymbol{J}^H$; and for the successive interference cancellation (SIC) estimator, $\boldsymbol{f}(\widetilde{\boldsymbol{b}}) = \boldsymbol{L}^H\widetilde{\boldsymbol{b}}$, $\boldsymbol{M} = \boldsymbol{J}^H(\boldsymbol{I} + \boldsymbol{L})^{-1}$, where $\boldsymbol{L}$ is such that $\boldsymbol{R} = \boldsymbol{I} + \boldsymbol{L} + \boldsymbol{L}^H$. It is straightforward to design multistage detection structures by combining those discussed above.

Observe that the estimators described above involve the correlation matrix $\boldsymbol{R}$, and some involve the amplitude matrix $\boldsymbol{B}$ or its phase $\boldsymbol{\Phi}$. As will be shown in the next section, the methodology to estimate $\boldsymbol{B}$ is the same as that to estimate $\boldsymbol{b}$. Therefore, if the process for getting $\boldsymbol{R}$ is programmable, all the multiuser detection schemes described above can be programmable. Therefore, the remaining difficulty is how to get $\boldsymbol{R}$ without losing programmability. From equations (4) and (5), if we replace $v_x(t)$ in (4) with $c_{is}(t - qT - \tau_i - t_{in})e^{\imath w_{rs} t}$ then we get (5). In addition, $c_{is}(t - qT - \tau_i - t_{in})e^{\imath w_{rs} t}$ is just a part of the i^{th} user's transmitted signal. Therefore, if the timings τ_i and t_{in} are available, known or estimated, the correlation matrix $\boldsymbol{R}_{x,x}$ can be produced from the programmable transceiver. The structure for calculating $\boldsymbol{R}_{x,x}$ is shown in Figure 3(a). The details are shown in Figure 3(b), in which the "GH points semi-IFFT" differs just a little from the conventional IFFT, and the comparison is shown in Figure 3(c). Since $\boldsymbol{R}_{x,x}$ are identical for all x, we ignore the subscripts "x, x" in those figures.

The programmable multiuser detection structure is depicted in Figure 4. The received signals $v_x(t)$ for $x = 1, 2, \ldots X$ are preprocessed by the programmable OFDM-CDMA receiver (referring to Figure 2(c) for details) and then the sufficient statistics $\boldsymbol{v}$ is generated. The correlation matrix $\widehat{\boldsymbol{R}}$ is computed in terms of timings $\widehat{\boldsymbol{t}}$, $\widehat{\boldsymbol{\tau}}$, and spreading codes c_{km}^n (referring to Figure 3(b) for details). The programmable estimator estimates both the signal $\boldsymbol{b}$ and the amplitude vector $\boldsymbol{\beta}$ constructed from β_{kjmlx}, while those two operations may be duplexed in time by scheduling or in frequency by pilot-added signaling. To compute the linear operator $\boldsymbol{A}$ and $\boldsymbol{A}'$ requires $\widehat{\boldsymbol{R}}$, $\widehat{\boldsymbol{B}}$ and $\widehat{\boldsymbol{R}}$, $\widehat{\boldsymbol{D}}$ respectively, where the reference matrix $\widehat{\boldsymbol{D}}$ relevant to training sequences is defined in the next section. Those operations can be implemented by DSP subroutines. The estimated noise variance $\widehat{\sigma^2}$ is necessary in LMMSE estimators.

We now demonstrate how the multiuser detection structure is programmed to implement the MC-CDMA, MC-DS-CDMA, and MT-CDMA respectively. Those systems can be realized by adjusting system parameters where the parameter changes are summarized inside Figure 4.

A. Programming to MC-CDMA

For the MC-CDMA, at the receiver end, the inverse of adjacent subcarriers separation is set to $T_o = T_c = T$, and the data sampling rate is set to $f_s = \frac{MJ}{T_o} = \frac{M}{T'}$. The two blocks "shift zero padding" and "windowing" are not needed for $H = 1$ and $GH = MJ$. The tapped-delay-lines regress to only one tap (that is $N = 1$), that is $c_{km}^1 = c_k^m$ for $m = 1, 2, \ldots M$. The correlation matrix generator produces $\widehat{\boldsymbol{R}}$ with

$$\widehat{R}_{kjmlx,irsnx}[p-q] = (c_k^m)^* c_i^s \int_{-\infty}^{\infty} \Pi(t - pT - \widehat{\tau}_k - \widehat{t}_{kl}) \Pi(t - qT - \widehat{\tau}_i - \widehat{t}_{in}) e^{\imath \frac{2\pi}{T}((j-r)+J(m-s))t} dt. \tag{17}$$

where $\Pi(t) \equiv 1$ for $t \in [0, T]$ and $\Pi(t) \equiv 0$ otherwise.

B. Programming to MC-DS-CDMA

For the MC-DS-CDMA, at the receiver end, we set $T_o = T_c = \frac{T}{N_{MD}}$ and $f_s = \frac{MJ}{T_c} = \frac{N_{MD}}{T'}$. The "shift zero padding" and "windowing" are not needed because $H = 1$ and $GH = MJ$. The weighting coefficients are identical for all tapped-delay-lines, that is c_{km}^n

$= c_k^n$, $n = 1, 2, ...N_{MD}$. The correlation matrix generator produces $\widehat{\boldsymbol{R}}$ with

$$\widehat{R}_{kjmlx,irsnx}[p-q] = \int_{-\infty}^{\infty} c_k^*(t-pT-\widehat{\tau}_k-\widehat{t}_{kl})c_i(t-qT-\widehat{\tau}_i-\widehat{t}_{in})e^{\imath\frac{2\pi N_{MD}}{T}((j-r)+J(m-s))t}dt. \quad (18)$$

C. Programming to MT-CDMA

For the MT-CDMA in which $M = 1$, at the receiver end, we let $f_s = \frac{G}{T_c} = \frac{GN_{MT}}{T}$. We set $c_{k1}^n = c_k^n$ for $n = 1, 2, ...N_{MT}$. The correlation matrix generator produces $\widehat{\boldsymbol{R}}$ with

$$\widehat{R}_{kjlx,irnx}[p-q] = \int_{-\infty}^{\infty} c_k^*(t-pT-\widehat{\tau}_k-\widehat{t}_{kl})c_i(t-qT-\widehat{\tau}_i-\widehat{t}_{in})e^{\imath(\frac{2\pi}{T}(j-r)t}dt. \quad (19)$$

V. The amplitude estimator

Those signal estimators described in the previous section require information about the amplitude matrix $\boldsymbol{B}$. We now propose two estimators for $\boldsymbol{B}$, the LMMSE estimator and the BLUE. Let $d_{kj}[p]$ be the training sequences for $p = -N_t, ..., 0, ...N_t$. The training time interval τ_t should be designed short enough ($\tau_t << 1/f_D$, where f_D is the Doppler frequency) such that $\boldsymbol{B}$ can be regarded constant during training and detection.

We get from (6) the signal model

$$\boldsymbol{v}[p] = \sum_{q=-N_t}^{N_t} \boldsymbol{R}[q-p]\boldsymbol{D}[q]\boldsymbol{\beta} + \boldsymbol{\eta}[p].$$

The reference matrix $\boldsymbol{D}[q] = diag(\ \boldsymbol{D}''[q], \boldsymbol{D}''[q], ..., \boldsymbol{D}''[q])$ is a $KJMLX \times KJMLX$ diagonal matrix in which $\boldsymbol{D}''[q]$ is a $KJ \times KJ$ diagonal matrix constructed from the training sequences $d_{kj}[q]$; $\boldsymbol{\beta}$ is a $KJMLX \times 1$ vector constructed from β_{kjmlx}. Let $\boldsymbol{v}'$ be constructed from $\boldsymbol{v}[p]$ and vice versa, we get $\boldsymbol{v}' = \boldsymbol{R}'\boldsymbol{D}'\boldsymbol{J}'\boldsymbol{\beta}' + \boldsymbol{\eta}'$, where $\boldsymbol{J}' = [I', I', ...I']^T$ is a $KJMLX(2N_t+1) \times KJMLX$ matrix constructed from the $KJMLX \times KJMLX$ identity matrix I'. By the same approaches as those in the previous section, the LMMSE operator can be verified to be

$$\boldsymbol{A}_{MS} = (\boldsymbol{J}'^H\boldsymbol{D}'^H\boldsymbol{R}'\boldsymbol{D}'\boldsymbol{J}' + \sigma^2\boldsymbol{C}^{-1})^{-1}\boldsymbol{J}'^H\ \boldsymbol{D}'^H,$$

where $\boldsymbol{C} = E[\boldsymbol{\beta}'\boldsymbol{\beta}'^H]$. And the BLUE is

$$\boldsymbol{A}'_{BL} = (\boldsymbol{J}'^H\ \boldsymbol{D}'^H\boldsymbol{R}'\boldsymbol{D}'\ \boldsymbol{J}')^{-1}\boldsymbol{J}'^H\boldsymbol{D}'^H.$$

In fact, the methodology of acquiring $\boldsymbol{\beta}$ is the same as that of acquiring $\boldsymbol{b}$. Therefore, other multiuser detection schemes, such as PIC and SIC detection schemes from (5), can be applied to estimating $\boldsymbol{\beta}$.

VI. Reduced complexity joint detection

Complexity and performance are two important benchmarks in multiuser detector design. For complexity, there are two concerns that need to be carefully taken into consideration: complexity introduced by programmability and complexity introduced by number of subcarriers. In this section, we discuss the complexity of the programmable multiuser detector in terms of two parts, preprocessing and postprocessing. And then a joint detection strategy based on interference cancellation is further proposed to reduce the complexity. For the performance, detailed discussions are in next section.

Since the methodology to estimate $\boldsymbol{B}$ is the same as that to estimate $\boldsymbol{b}$, we only consider the complexity of the latter. Refer to Figure 4, the whole detection processes may be separated into two parts, namely preprocessing and postprocessing. The former generates the sufficient statistics $\boldsymbol{v}$ from the received signal $\boldsymbol{v}(t)$ via the programmable OFDM-CDMA receiver; the latter estimates the transmitted bits $\boldsymbol{b}$ in terms of $\boldsymbol{v}$ under some optimization criterion. The complexity of preprocessing is system predominant, that is the computational complexity only depends on the specific system chosen, and is independent of the multiuser detection scheme chosen. Although the complexity of postprocessing is also system predominant, the order of computational complexity is dominated by the multiuser detection scheme chosen. Since the computational time in DSPs is dominated by multiplication operations. We just analyze the complexity of complex-valued multiplication per user per bit.

A. *Complexity of preprocessing*

We see in Figure 2(c) that the whole operations of preprocessing are dominated by the "GH points FFT" and the "tapped-delay-lines". The former requires $NGH \log GH$ operations, where log is the base 2 logarithm, $G = \lceil (MJ-1)/H + 1 \rceil$, $H = T_o/T_c$, and N is the length of the signature sequences in time domain. The latter requires MJN operations. Summing up the operations with respect to all paths, antennas and users, it requires totally

$$LX(NGH\log GH + MJN)/J = o(LXNM\log MJ)$$

operations per user per bit. The complexity order of the preprocessing for different systems is summarized in Table I.

In practical applications, if the programmable OFDM-CDMA is capable of adapting to the three multicarrier CDMA systems, the processing speed should be fast enough to deal with the maximum computational complexity among the three. In addition, the ratio of complexity of OFDM-CDMA to CDMA is

$$\frac{L'XNM\log MJ}{LXN} = \frac{L'M\log MJ}{L},$$

where L', L are the number of paths in OFDM-CDMA and CDMA respectively. Please do not ignore the advantage in processing non-frequency-selective fading channel in OFDM-CDMA, compared to CDMA counterpart.

B. Complexity of postprocessing

Now we consider the complexity of postprocessing. Let $2P+1=1$ for ease of illustration. Refer to (13), it requires $o[(KJ)^2(ML)^2X/(KJ)] = o[KJXM^2L^2]$ operations to do $\boldsymbol{J}^H\boldsymbol{B}^H\boldsymbol{RBJ}$, $o[(KJ)^3/(KJ)] = o[K^2J^2]$ operations to compute its inverse, and $o[KJMLX/KJ] = o[MLX]$ operations to do $\boldsymbol{J}^H\boldsymbol{B}^H\boldsymbol{v}$. Hence the total complexity for the BLUE is $o[K^2J^2 + KJXM^2L^2]$. If iterative interference cancellation algorithms, such as the Gauss-Seidel algorithm or the Jacobi's method [15], are adopted instead of direct matrix inversion, the complexity is reduced to $o[KJ(S+XM^2L^2)]$, where S, usually ≤ 10, is the total number of iterations. Please note that our programmability does not introduce any extra order of complexity (see table).

C. Reducing complexity

First of all, let us consider the case when $J=M=L=X=2P+1=1$. If we substitute $\widetilde{\boldsymbol{R}}$ for $\boldsymbol{R}$ with $\widetilde{R}_{k,i} \equiv \delta_{k,i}R_{k,i}$, then (8) is reduced to $v_k = R_{k,k}\beta_k b_k + \eta_k$, for $k=1,2,...,K$, in which the optimum multiuser estimator and the conventional single user one coincide. Based on the fact that detection made jointly between two users effectively suppresses their mutual interferences, we can control the factors to be detected jointly or not, by setting some $R_{kjml,irsn} = 0$ in (12),(13),(15) and (16) according to the significance of interference types, such as MAI or interchannel interference (ICI). For example, when MAI is the most serious, the detection is made jointly only with respect to users to reduce complexity by setting

$$\widetilde{\boldsymbol{R}}_{jml,rsn} = \delta_{j,r}\delta_{m,s}\delta_{l,n}\boldsymbol{R}_{jml,rsn}. \tag{20}$$

Then the term $\boldsymbol{J}^H\boldsymbol{B}^H\boldsymbol{R}\boldsymbol{B}\boldsymbol{J}$ in (12) and (13) is reduced to

$$\sum_{m=1}^{M}\sum_{l=1}^{L}\sum_{x=1}^{X}\boldsymbol{B}_{mlx}^H\boldsymbol{R}_{mlx,mlx}\ \boldsymbol{B}_{mlx},$$

and the complexity is further reduced to $o[K(S+XML)]$ for iterative algorithms. When ICI is the most serious, the detection is made jointly only with respect to all subcarriers by setting

$$\widetilde{\boldsymbol{R}}_{kl,in} = \delta_{l,n}\delta_{k,i}\boldsymbol{R}_{kl,in}.$$

Another approach is combining the identical bits after they are estimated by some criterion. For example, let $\widehat{\boldsymbol{b}}_k = \sum_{m=1}^{M}\widehat{\boldsymbol{b}}_{k;m}$, where $\widehat{\boldsymbol{b}}_{k;m}$ is the estimator of $\boldsymbol{b}_k$ in terms of $\boldsymbol{v}_{km}$. Since

$$\boldsymbol{v}_{km} = [\boldsymbol{R}\boldsymbol{B}\boldsymbol{J}\boldsymbol{b}]_{km} + \boldsymbol{\eta}_{km} = \sum_{i=1}^{K}\sum_{s=1}^{M}\boldsymbol{R}_{km,is}\boldsymbol{B}_{is}\boldsymbol{J}\boldsymbol{b}_i + \ \boldsymbol{\eta}_{km},$$

where $\boldsymbol{J}$ here is a $JLX \times J$ matrix obtained by setting $K = M = 1$ in the $\boldsymbol{J}$ defined in (3). Let $\boldsymbol{A}_{km}$ be a linear operator such that $\widehat{\boldsymbol{b}}_{k;m} = \boldsymbol{A}_{km}\boldsymbol{v}_{km}$. If the LMMSE criterion is considered, by the orthogonality principle [1], we have

$$\begin{aligned}\boldsymbol{A}_{km} &= E[\boldsymbol{v}_{km}\boldsymbol{b}_k^H]^H(E[\boldsymbol{v}_{km}\boldsymbol{v}_{km}^H]^{-1})^H \\ &= (\sum_{s=1}^{M}\boldsymbol{J}^H\boldsymbol{B}_{ks}^H\boldsymbol{R}_{ks,km})(\sum_{i=1}^{K}\sum_{s',s=1}^{M}\boldsymbol{R}_{km,is}\boldsymbol{B}_{is}\boldsymbol{J}\boldsymbol{J}^H\boldsymbol{B}_{is'}^H\ \boldsymbol{R}_{is',km} + \sigma^2\ \boldsymbol{R}_{km,km})^{-1},\end{aligned}$$

which degenerates to the single user LMMSE estimator proposed in [14] if we let $J = M = 1$ and one-shot detection is adopted. The order of postprocessing complexity for several types of multiuser detectors is summarized in Table II. The proposed method significantly reduces the complexity.

Theoretically, to some subchannel, its more neighboring subchannels produce more interference. Moreover, major terms of insignificant interference (usually ICI) may still be as important as minor terms of significant interference (usually MAI). Hence, a systematic approach to make a detection jointly with only some neighboring subcarriers rather than all subcarriers is more desirable. Assume that the most efficient way is to detect jointly with the most N_J neighboring independent subcarriers, the most N_M neighboring dependent subcarriers. We now simplify $\boldsymbol{R}$ to $\widetilde{\boldsymbol{R}}$ by setting $\widetilde{R}_{kjml,irsn} = 0$ for $|j-r| > N_J$, $|m-s| > N_M$. Because the matrix $\widetilde{\boldsymbol{R}}$ simplified in this may become very sparse and that $\widetilde{\boldsymbol{R}}$ itself has a very particular structure so an efficient algorithm may probably exist for those detection schemes.

VII. Channel model consideration and Simulation results

Those complex Gaussian r.v.s g_{kjml} in (2) may be assumed independent for different k and l. As the number of carriers grows, the number of resolvable paths reduces under fixed transmission bandwidth; the diversity is transformed from time domain to frequency domain. However, it is unreasonable to assume g_{kjml} independent for different jm anymore if the separation between adjacent subcarriers is less than the channel coherent bandwidth.

We first discuss how to model the statistical relations of those g_{kjml} in practical applications. Assume that the system undergoes a frequency selective fading channel with maximum delay spread τ_m. Let us begin with the single carrier case, where we add an additional number '1' to the subscripts of corresponding symbols. The impulse response of the multiray fading channel can be modeled by

$$h(t) = \sum_{l=0}^{L_1-1} \gamma_l \delta(t-\tau_l) \tag{21}$$

where γ_l are i.i.d. zero-mean complex Gaussian r.v.s with variance σ_l^2, and $\tau_l = (l+\Delta_l)T_{c1}$ with Δ_l uniformly distributed on $[0,1]$. L_1 is the number of resolvable paths given by

$$L_1 = \lceil \tau_m / T_{c1} \rceil \,. \tag{22}$$

In general OFDM-CDMA systems, it is more useful to characterize the behaviors in individual subchannels. Assume that there are totally D subcarriers with carrier separation

$2/T_o$ Hz and subbandwidths $2/T_c = 2/(NT)$, where T_c is the chip duration and T is the symbol duration and N is the length of the time-domain spreading codes. The frequency response of $h(t)$ is

$$H(w) = \int_{-\infty}^{\infty} h_1(t)e^{-jwt}dt = \sum_{l=0}^{L_1-1} \gamma_l e^{-jw\tau_l} = \sum_{l=0}^{L_1-1} \gamma_l e^{-jw(l+\Delta_l)T_{c1}} = \sum_{l=0}^{L_1-1} \gamma_l e^{-jwlT_{c1}},$$

where we consider the random phases $e^{-jw\Delta_l T_{c1}}$ into γ_l without loss of generality. $H(w)$ is a periodic function of w and one period of $H(w)$ equivalently characterize the frequency response of the channel. Theoretically, we need only L_1 samples of $H(w)$ to maintain the statistical information. Let $M = \frac{2/T_{c1}}{1/T_o} = \frac{2T_o}{T_{c1}}$, which is the ratio of carrier separation and the total transmission bandwidth. M uniformly separated samples of one period of $H(w)$ also maintains the statistical information if $M \geqslant L_1$. Let $w_n = \frac{2\pi n}{MT_{c1}}$, and define Γ_n as the samples of $H(w)$, we have

$$\Gamma_n = H(w_n) = \sum_{l=0}^{L_1-1} \gamma_l (e^{-j2\pi/M})^{nl}.$$

We observe that Γ_n is the M-point discrete Fourier transform of γ_l . On the d^{th} subchannel, the related frequency samples are Γ_{d+n-1} for $n = 0, 1, ..., \lceil \frac{2NT_o}{T} \rceil - 1$. For MC-CDMA and MC-DS-CDMA, $\lceil \frac{2NT_o}{T} \rceil - 1 = 1$, therefore only two samples Γ_{d-1}, Γ_d characterize the d^{th} subchannel. When the number of carriers M is so large that each subchannel undergoes flat fading, it is more convenient to characterize the fading coefficient in the d^{th} subchannel by $\Gamma_d/\sqrt{D}$ only, where $1/\sqrt{D}$ is introduced to make total power conserved.

For other types such as MT-CDMA, in which the subchannels are subject to frequency selective fading, the impulse response $h_d(t)$ can be verified to be related to the inverse discrete Fourier transform of Γ_{d+n-1} for $n = 0, 1, ..., L' - 1$, where $L' = \lceil \frac{2NT_o}{T} \rceil$, that is

$$h'_d(t) = \sum_{l=0}^{L'-1} \gamma'_{dl} \delta(t - \tau'_l),$$

where

$$\gamma'_{dl} = \frac{1}{\sqrt{DL'}} \sum_{n=0}^{L'-1} \Gamma_{d+n-1}(e^{j2\pi/L'})^{nl},$$

and $\tau'_l = (l + \Delta_l)T_{c1}$ with Δ_l uniformly distributed on $[0, 1]$. Because the number of resolvable paths on the subchannels is $L = \lceil \tau_m / T_{c1} \rceil$. An equivalent model is

$$h_d(t) = \sum_{l=0}^{L-1} \gamma_{dl} \delta(t - \tau_l''),$$

where and $\tau_l'' = (l + \Delta_l) T_c$ with Δ_l uniformly distributed on $[0, 1]$ and L samples of γ_{dl} is generated from L' samples of $\sqrt{\frac{L'}{L}} \gamma_{dl}'$ by interpolation. When the spread subspectral overlap very tightly we may assume that all subchannels experience the same fading statistics.

Next, we consider the performance of our proposed multiuser detector. The first issue we concern about is whether the proposed multiuser detection schemes outperform the conventional single user detection one, which is certainly true in the single carrier case. Let us assume $L = X = 1$ for ease of illustration. Observing (5), we can regard $\boldsymbol{R}$ as a new crosscorrelation matrix for imaginary users $k' = 1, 2, \ldots, KJM$ with their signature waveforms being $c_{k'}(t) = c_k(t) e^{w_{jm} t}$. Therefore, we can regard a K-user OFDM-CDMA system as a special case of a KJM-user CDMA system (This proposition also provides an opportunity to implement multirate transmission via an OFDM-CDMA structure). Therefore, all characteristics of multiuser detection in CDMA are inherited to OFDM-CDMA. Another reason to justify the performance is the optimization process during derivation of the multiuser detector. For linear multiuser detection, we make linear combinations of the statistics of all users. Due to some optimization criterions, for example the LMMSE and BLUE, the performance cannot be worse than the single user detection for which we use the statistic of the intended user only.

We also want to make sure no performance degradation in a programmable detector compared to a specific multicarrier CDMA, for example, a programmable LMMSE multiuser detector realized to MC-CDMA compared to a multiuser detector specifically designed for a specific multicarrier CDMA. Observe that those derived programmable multiuser detection schemes are operated in terms of the observation vector $\boldsymbol{v}$, for example, $\widehat{\boldsymbol{b}} = \boldsymbol{A}_{MS} \boldsymbol{v}$ for the LMMSE detector. Since $\boldsymbol{v}$ is a sufficient statistics for $\boldsymbol{b}$ by (3), there is no loss of information if the detection is made according to $\boldsymbol{v}$ rather than the received waveform vector $\boldsymbol{v}(t)$. Fatherhood, those multiuser detector schemes, formula forms shown in (12)-(16), are not constrained to be programmable during derivation. Programmability is actually from producing $\boldsymbol{v}$ and the other operations required for multiuser detection, such as getting $\boldsymbol{R}$ and estimating $\boldsymbol{B}$ while being implemented via the original programmable transceiver architecture. Since there is no difference in mathematical formulation between programmable and nonprogrammable multiuser detector, we conclude that there is no performance degradation to achieve programmability.

Finally, we investigate the programmable detector performances in MC-CDMA and MC-DS-CDMA by simulation. Beginning with (21), we assume that in the single carrier case, the transmission bandwidth is such that the number of resolvable paths is $L_1 = 4$ with unit-energy uniform multipath power profile. The statistical relations of g_{kjml} in (2) are generated according to the channel consideration discussed above. Propagation delays are assumed to be i.i.d. uniformly distributed on $[0, T']$ where T' is the symbol period in the single carrier case. The number of transmission symbols is $2P + 1 = 1$, and the number of subcarriers is designed such that each subchannel undergoes frequency-flat fading ($L = 1$). In $X > 1$ cases, the DOAs are randomly generated according to uniform distribution.

The following programmable detection schemes are evaluated: the LMMSE and BLUE detection shown in (12) and (13), the corresponding reduced complexity detection RLMMSE and RBLUE by (20), the decorrelating (DC) detection shown in (15) and the modified BLUE (BLUE1) shown in (14). In addition, the single user MRC and EGC detection shown in (9) and (10) are also investigated for comparison.

Case 1: Programming to MC-CDMA

The number of independent and dependent subcarriers are set to $J = 1$ and $M = 16$ respectively. Figure 5(a) shows the BER versus SNR plot for $K = 8$ users using Hadamard codes of length 16 with one antenna, that is $X = 1$. Figure 5(b) shows the case for $X = 4$.

Case 2: Programming to MC-DS-CDMA

Since in asynchronous MC-DS-CDMA, bandwidths are spread in the time domain, the codes with good crosscorrelation properties are preferred. We use Gold codes of length 31 in this case. The number of independent and dependent subcarriers are $J = 1$ and $M = 4$ respectively; the number of users is $K = 31$. The BER versus SNR plot for single antenna and four antennas are shown in Figure 6(a) and Figure 6(b) respectively.

For the two cases, it is obvious that the proposed multiuser detection approaches outperform the single user detection approach. In addition, we see that the performance of

RLMMSE and RBLUE is satisfactory although they are still interference limited. The decorrelating detector (DC) is suitable for MC-DS-CDMA but very poor for MC-CDMA. The reason is that the correlation matrix $\boldsymbol{R}$ in MC-CDMA is usually near singular so that the elements of $\boldsymbol{R}^{-1}$ is usually very large resulting in tremendous noise enhancement. The BLUE1 is preferable for MC-DS-CDMA. Therefore, the amplitude $\boldsymbol{B}$ is more critical in MC-CDMA than in MC-DS-CDMA. In addition, it is obvious from those results that using antenna array also provides significant gains. The design of multiuser detection schemes and the number of antennas deployed trade off between performance and complexity.

VIII. Conclusion

The programmable OFDM-CDMA systems are highly desirable because of its adaptability to multisystems and capability of efficiently managing radio resources, especially for the upcoming wireless multimedia communication age. We propose a multiuser detection structure to enhance system performance while maintaining its programmability. Because the theoretical formulations related to the OFDM-CDMA receiver are similar to those related to the conventional CDMA, the detection schemes for the OFDM-CDMA may simply follow those in CDMA. The BLUE and LMMSE estimator for the necessary amplitude matrix $\boldsymbol{B}$ is also derived, and other estimators may simply follow the multiuser detectors. The complexity of the proposed detection structure are considered and the proposed strategy to design an efficient detector, that is good enough in performance and as low as possible in complexity, significantly reduce the complexity. The performance of the programmable multiuser detection structure in the OFDM-CDMA is guaranteed because we may consider the OFDM-CDMA as a special case of CDMA with more imaginary users. Moreover, there is no performance loss to achieve programmability theoretically, because the proposed detection structure processes identical sufficient statistics $\boldsymbol{v}$. The simulation results justify that the proposed programmable multiuser detection structures including reduced complexity schemes have effective performance.

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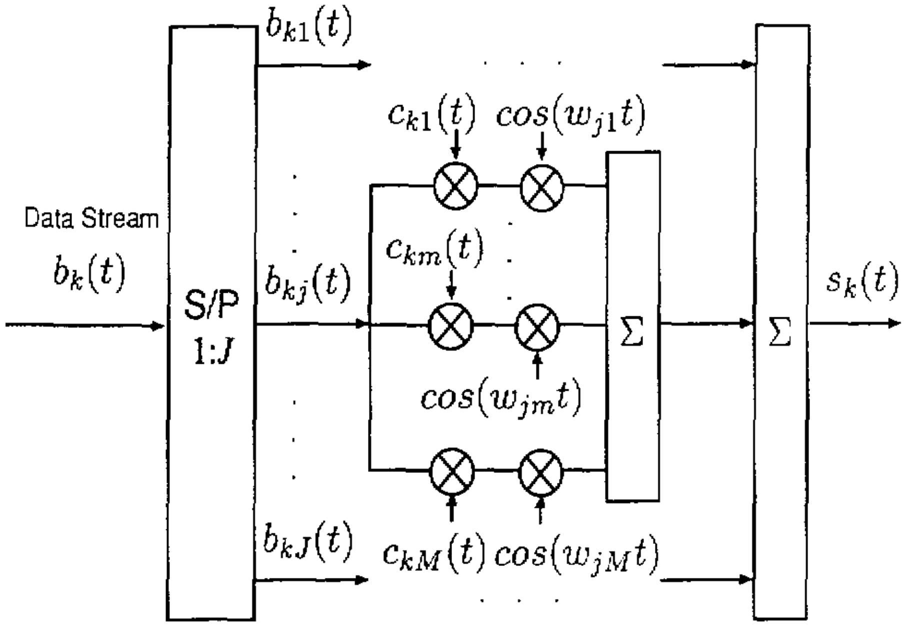


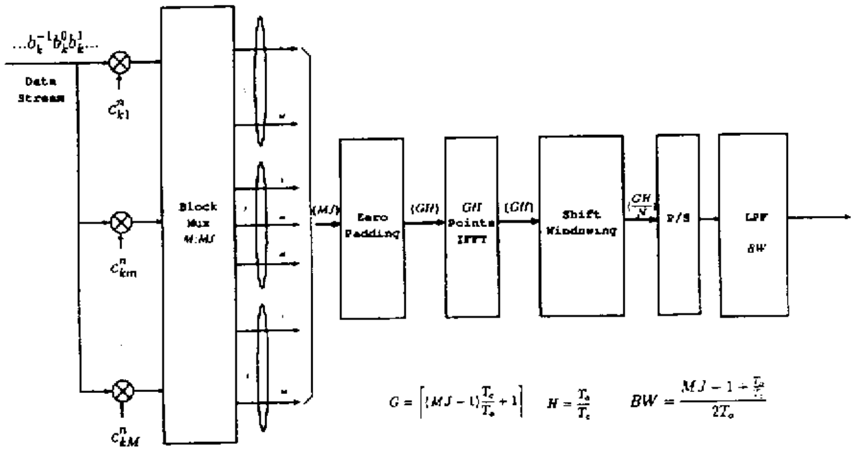
$$w_{ab} - w_{cd} = 2\pi(a-c)/T_o - 2\pi(b-d)J/T_o$$

Fig. 1. The k^{th} user's transmitter structure of the OFCDMA.

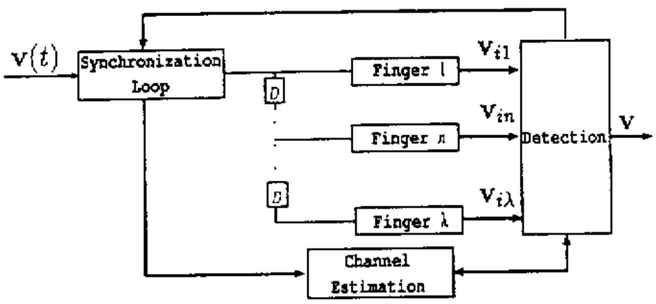
TABLE I

COMPLEXITY ORDER OF THE PREPROCESSING FOR PROGRAMMABLE OFDM-CDMA, MC-CDMA, MC-DS-CDMA, MT-CDMA, AND CDMA. NOTE THAT $M = 1$ IN MT-CDMA, AND THEREFORE THE COMPLEXITY DOES NOT INVOLE M.

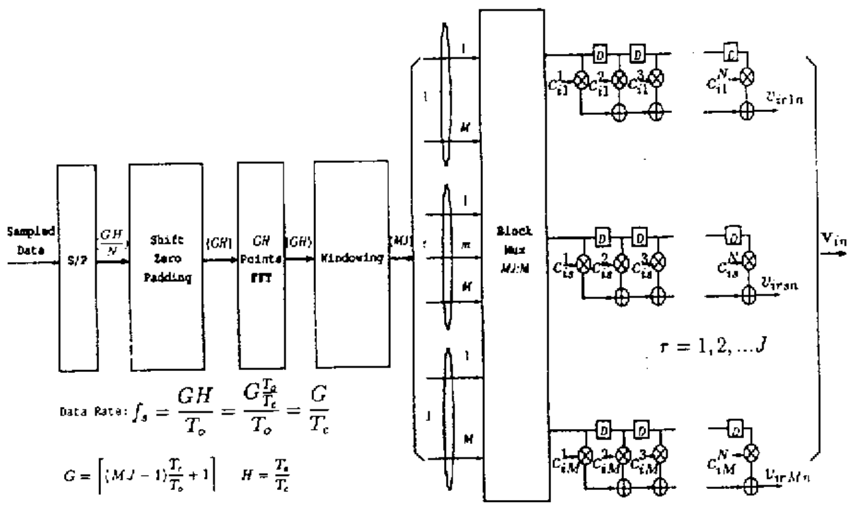
	OFDM-CDMA	MC-CDMA	MC-DS-CDMA	MT-CDMA	CDMA
Complexity	$LXNM \log MJ$	$LXM^2 \log MJ$	$LXNM \log MJ$	$LXN \log J$	LXN



(a) The k^{th} user's transmitter.

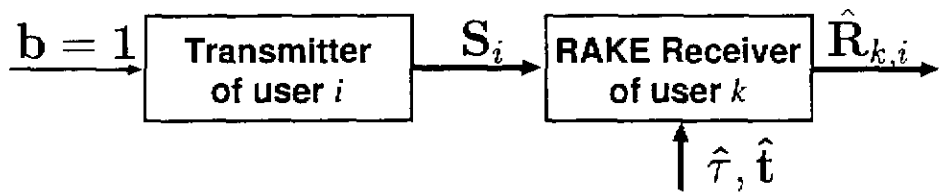


(b) The i^{th} user's RAKE receiver

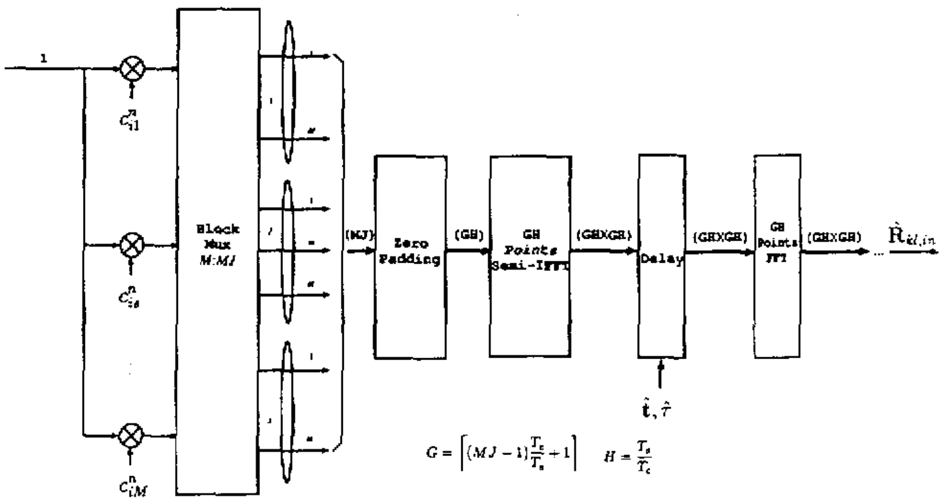


(c) The details for the n^{th} finger.

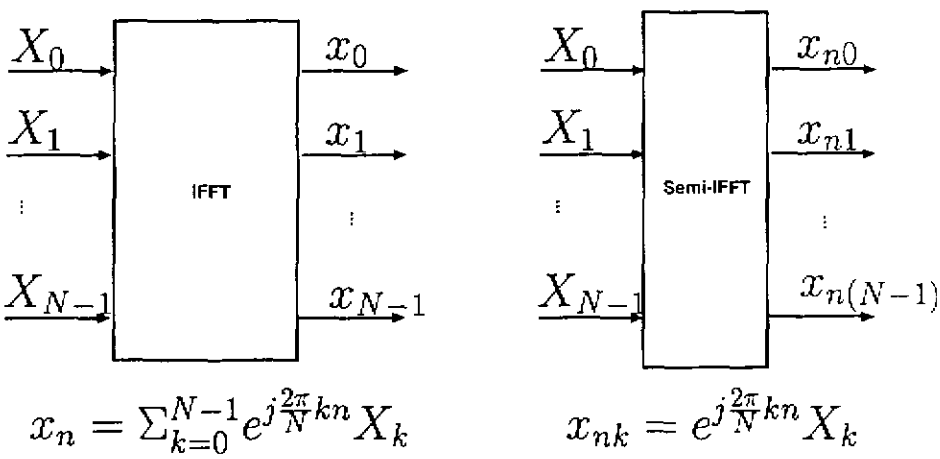
Fig. 2. The transceiver for the programmable OFDM-CDMA.



(a) The block diagram.



(b) The details for getting the correlation matrix.



(c) The comparison between IFFT and semi-IFFT.

Fig. 3. The structure for the correlation matrix generator.

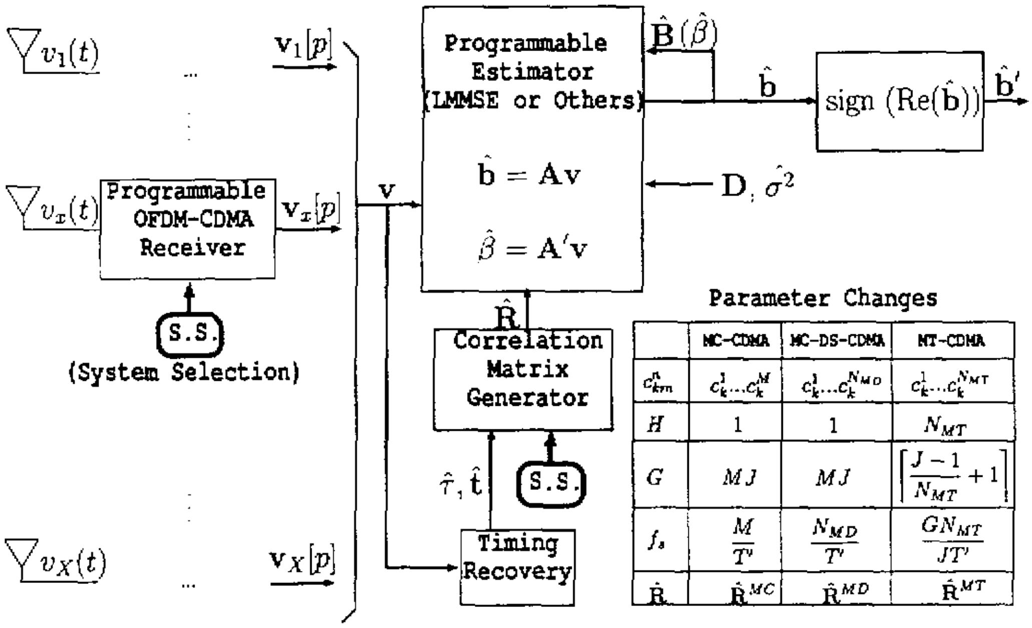
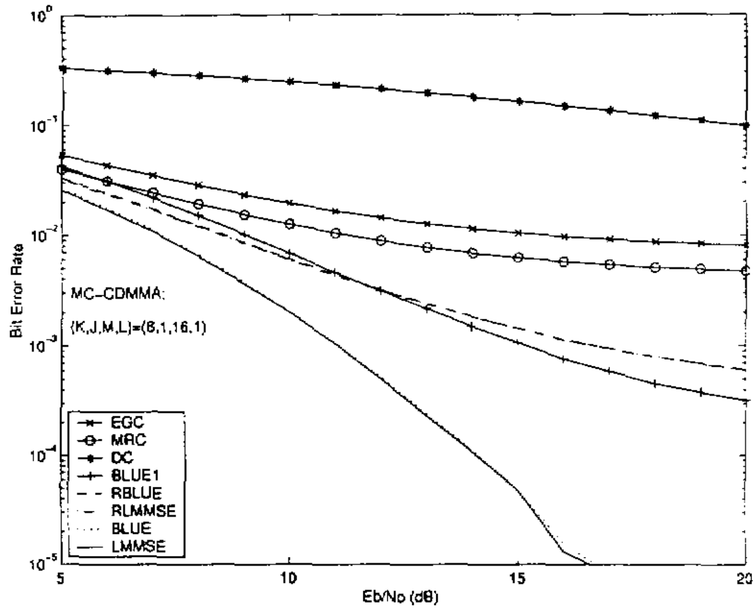


Fig. 4. The linear programmable multiuser detector, in which the programmable OFDM-CDMA receiver and correlation matrix generator is shown in Figure 2 and Figure 3 respectively.

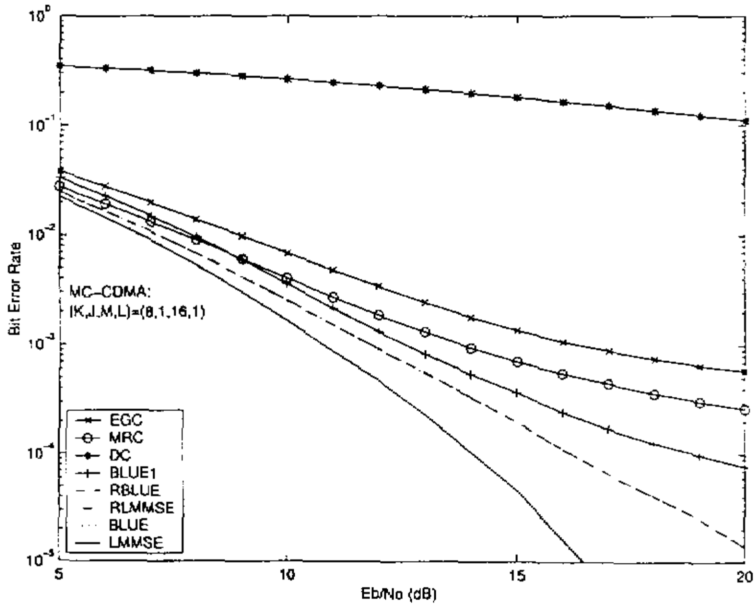
TABLE II

COMPLEXITY ORDER OF THE POSTPROCESSING FOR SEVERAL TYPES OF MULTIUSER DETECTORS: LMMSE, BLUE, REDUCED LMMSE (RLMMSE), REDUCED BLUE (RBLUE), AND DECORRELATING (DC). THE RLMMSE AND RBLUE, ARE REDUCED FROM LMMSE AND BLUE ACCORDING TO (20). S IS THE NUMBER OF ITERATIONS OF THE ALGORITHM.

	Direct Matrix Inversion	Iterative Algorithm
LMMSE and BLUE	$K^2J^2 + KJXM^2L^2$	$KJ(S + XM^2L^2)$
RLMMSE and RBLUE	$K^2 + KXML$	$K(S + XML)$
DC	$K^2J^2XM^3L^3$	$SKJXM^2L^2$

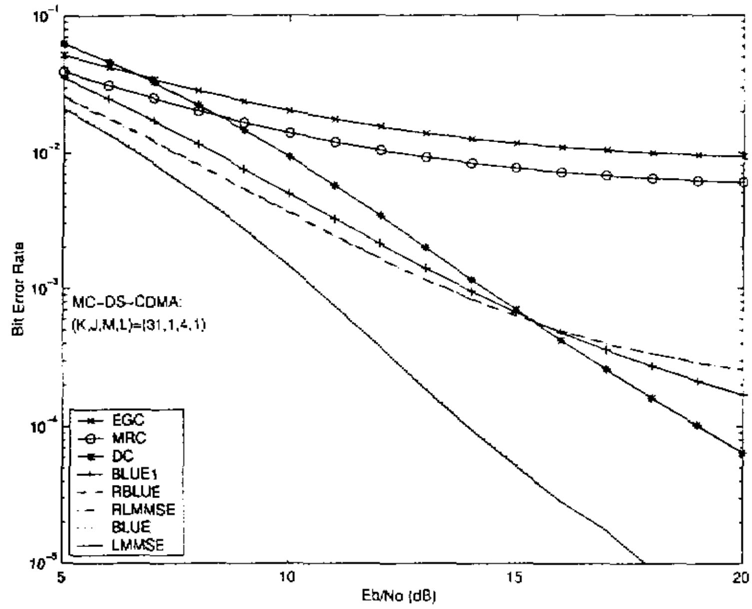


(a) $X = 1$

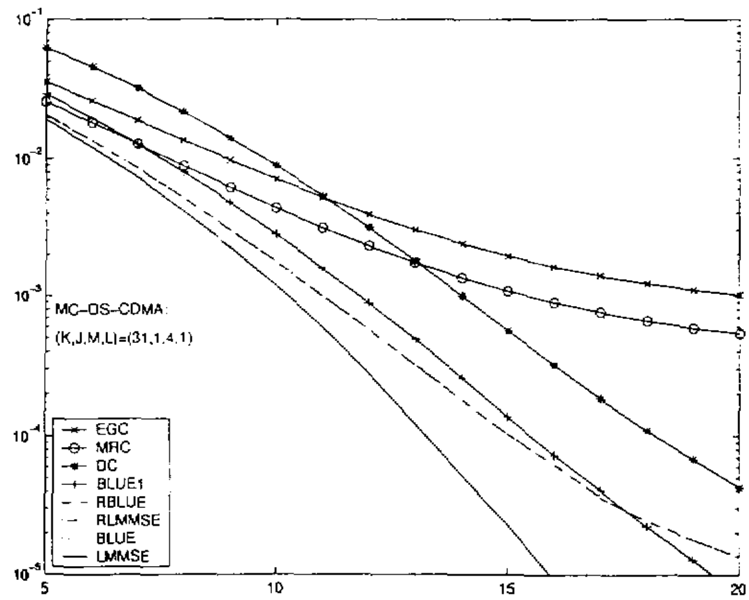


(b) $X = 4$

Fig. 5. The BER versus SNR for the programmable multiuser detection structure programmed to MC-CDMA system. Several detection schemes are shown for comparision. 8 users, Hadamard code, code length = 16.



(a) $X = 1$



(b) $X = 4$

Fig. 6. The BER versus SNR for the programmable multiuser detection structure programmed to MC-DS-CDMA system. Several detection schemes are shown for comparision. 31 users, Gold code, code length = 31.