Studies on Reliable Public Safety Wireless Communication Systems Employing Software Defined Radio and Cognitive Radio

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Abstract

Public safety mobile wireless communication systems (PMCSs) are widely used by public safety personnel, such as firefighters and police, as well as local governments. PMCSs are crucial to protect safety and security of communities. Conventional PMCSs effectively cover underpopulated areas as well as urban areas by employing long-zone scheme. Since the PMCSs can cover areas that are not covered by commercial cellule systems, they play the important role as the only communication tool.

Moreover, the conventional PMCSs have enhanced robustness and reliability. The conventional PMCSs can keep their services even if backbone lines are cut off. In contrast, short-zone scheme systems cannot offer stable and wide service area without backbone line connection. For example, the Great East Japan Earthquake in Japan, police mobile communication systems had kept their functions while cellular phones became disabled. PMCSs are required to be quite high robustness and reliability in order to save human life.

Recently, conventional PMCSs are required to realize further expansion of service areas and high speed transmission although they have stably provided users with wide service areas so far. Nowadays, in order to solve complicated public affair quickly, more stable service areas and broadband communication are required. Compared with conventional PMCSs in urban areas, commercial wireless mobile communication systems (CWMCSs) such as cellular systems supply stable service areas and broadband communication in times of peace. In accordance with development of wireless technology, PMCSs need to keep pace with CWMCSs.

However, conventional PMCSs can hardly realize further stable service areas and high speed transmission because of large-zone scheme.

In terms of realization of further stable service areas, no-service areas cannot be eliminated easily. This is because no-service areas are mostly attributed to shadowing; in large-zone scheme, a no-service area that must essentially be covered by a certain base station is seldom covered by other neighboring base stations. Although new allocation of base stations is fundamental answer to solve no-service area problem in PMCSs, building new base stations of PMCSs that are not used for a commercial purpose is restricted by national and local budget.

Realization of high speed transmission of PMCSs is also difficult because of large-zone scheme. To realize high speed transmission, increase of transmit power or shrinking of service area coverage is required to compensate Signal to Noise Power Ratio (SNR) deterioration caused by expanding bandwidth. Increase of transmission power of mobile station used in large-zone scheme systems is almost impossible because transmission power of mobile station is originally high. Thus, shrinking of service areas is necessary for high speed communication.

Currently, to realize high speed transmission, next generation broadband PMCSs (BPMCSs) employing short- or middle-zone scheme are being developed. In the 3GPP, it is considered that the Long Term Evolution (LTE) is used for communication of public safety. In Japan, National Institute of Information and Communications Technology (NICT) has researched and developed Public Broadband Wireless Communication System (PBWCS), which employs 200MHz as carrier frequency. The PBWCS has already been equipped in national police agency in Japan.

However, we consider that the conventional narrowband PMCSs (NPMCSs) are not replaced with the BPMCSs completely. This is because the BPMCSs cannot cover all the areas that the conventional NPMCSs have covered. Moreover, there are problems of robustness and reliability when accidents happen. Hence, users of PMCSs will utilize both of NPMCSs and BPMCSs in accordance with the situation. In this case, users equipping several terminals feel inconvenient and also radio resources are not used effectively.

The best solution to realize optimal PMCSs is employing heterogeneous cognitive radio (HCR) for PMCSs. By applying the HCR to PMCSs, service areas expansion and high speed transmission in PMCSs will be realized effectively. We propose an integrated system combining NPMCSs with CWMCSs and BPMCSs to make communication quality of the PMCSs improve. The proposed HCR recognizes communication conditions of several systems and then provides PMCS's users with optimal communication quality. Although software defined radio techniques (SDR) are ideal to operate cognitive radio more flexibly, we deal with HCR mainly to realize combined systems in this thesis. We study advantages, problems, and their solution to realize the HCR for PMCSs.

Firstly, we research service area expansion of NPMCSs using HCR. The proposed HCR is utilized for stabilization of NPMCS's service area. If communication quality of a NPMCS deteriorates owing to shadowing, the proposed HCR terminal obtains a part of NPMCS's data called subsidiary information (SI) from CWMCSs or BPMCSs. The proposed HCR terminal can improve PMCS's bit error rate (BER) performance by combining the SI with received signals of the NPMCS and then decoding the combined signals using forward error correction (FEC). Since convolutional codes are often used in FEC of NPMCSs, we consider BER improvement methods of the convolutional code. We derive modified Viterbi algorithm from maximum likelihood sequence estimation (MLSE) of the combined signals. Moreover, we introduce the distance spectrum to evaluate characteristics of the convolutional codes. The distance spectrum is used for estimating improvement of BER performances.

Next, we consider synchronization methods to realize the proposed HCR. In the HCR, there are two types of synchronization method; one is the self-synchronization method to synchronize each system itself. The other is the co-synchronization method to combine different systems. In this thesis, we consider self-synchronization methods of NPMCSs mainly. This is because the HCR aims to improve communication quality of NPMCSs equipping conventional self-synchronization methods that are not probably available in low SNR environments. In this environment, since NPMCSs can hardly obtain their self-synchronization alone, powerful self-synchronization methods using HCR techniques are required.

We propose two synchronization methods that are utilization of global portioning system (GPS) signals and utilization of the SI, respectively.

The synchronization methods utilizing GPS signals can acquire timing synchronization. To obtain timing synchronization, the proposed HCR acquires accurate time and own location using the GPS signals. The HCR also gets the location of base stations and the frame timing by making the SI convey their information. Since the HCR can know accurate time and distance between the base station and the HCR, synchronization timing can be calculated. However, in GPS based method, preciseness of timing synchronization may be deteriorated by measurement error of GPS signals, diffraction caused by mountains, and propagation delay caused by reflection. For this reason, we consider a mitigation method of the timing error and then evaluate BER performances using computer simulation.

Moreover, we propose a SI based synchronization method that can obtain timing synchronization without GPS signals. The proposed method is employed when a NPMCS uses differential coded $\pi/4$ shift QPSK as the modulation scheme. The notable feature of the proposed method is to convey the phase rotation of the $\pi/4$ shift QPSK as the SI. The HCR can forecast PMCS's envelopes from the obtained SI and then obtain the timing synchronization by correlating the forecasted envelopes with real received envelopes. Since the proposed method can also be used for co-synchronization and BER improvement, CWMCS's resource consumption to convey the SI is suppressed.

Finally, we consider HCRs combining several PMCSs. In this thesis, the combination of NPMCSs and the combination of a NPMCS and a BPMCS are researched. In the combination of NPMCSs, we consider that several PMCSs are integrated by SDR. In the combination of a NPMCS and a BPMCS, we propose site diversity based on HCR to improve uplink communication quality of the BPMCS. In this diversity, since uplink interference must be avoided, we employ combination of the adaptive array and HCR techniques. Moreover, we propose information compression methods for narrow band backbone lines so that received data can be conveyed to head office with little BER deterioration.

PMCSs will have played an important role to ensure social safety. In the thesis, we consider the one of the next generation PMCSs employing SDR and HCR. Using this research, we can obtain a direction of optimal PMCSs. The next step that we need to perform is to apply our proposed method to actual radio systems. We must continue this research so that high reliable and compact PMCSs can be realized.

Acknowledgements

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Chapter 1

Introduction

In the Introduction, background and motivation of the research are shown in section 1.1. In section1.2, we explain features of conventional narrowband public safety wireless communication systems (NPMCSs). Then, problems of the NPM-CSs are explained in section 1.3. In sub-section 1.4, we give brief explanation of broadband public safety wireless communication systems (BPMCSs) and their problems. To further improve NPMCSs and BPMCSs, we introduce software defined radio (SDR) and heterogeneous cognitive radio (HCR) in section 1.5. In the end of the Chpater, organization of this thesis is explained in section 1.6.

1.1 Background and Motivation of the Study

Public safety wireless communication systems (PMCSs) are widely used for public safety applications involving firefighters, the police, local government, etc. PM-CSs are crucial to protect secure and safe such as maintenance of police order and rescue operation of disasters. Conventional PMCSs are narrow band systems (NPMCSs) and used for voice communication.

Recently, to deal with getting more complex public activities, more stable and higher speed transmitting communication is required. To realize high speed transmission, broadband PMCSs (BPMCSs) have been discussed as next generation PMCSs. However, the conventional NPMCSs are still required for public safety activity because of demands of large service areas and robustness. In this reason, we study application of SDR and HCR combining NPMCSs, BPMCSs, and commercial mobile wireless communication systems (CMWCSs) to realize more reliable and higher speed communication. We believe that our research will be useful for future development of PMCSs.

1.2 Features of conventional narrow band PMCSs (NPMCSs)

Since utilization purpose of the PMCSs differs from that of CMWCS such as cellular systems, NPMCSs have special features that are not equipped in CMWCSs. The Association of Radio Industries and Businesses (ARIB) has standardized several PMCSs, e.g., Narrow Band Digital Telecommunication System (RCR STD-39) and Digital Mobile Telecommunication System for Local Government(RCR STD-T79) in Japan [1]. In Europe and the U.S. narrow-band digital radio standards TETRA [2] and P25 [3] respectively are employed in public safety radio applications. Main features of the NPMCSs are as follows.

Multicasting Service Using a feature of the multicasting service, the information that one user has sent is concurrently delivered to all users to accomplish public affairs effectively. In public safety activities, the multicasting service is necessary to command many officers such as policemen and firefighters

Long-range Communication Systems NPMCSs employ large-zone scheme using very high frequency (VHF) for carrier waves. The large-zone scheme systems are very important for public safety activities. NPMCSs can cover its required service areas effectively. NPMCSs cover not only crowded areas but also underpopulated areas, e,g., mountainous areas, the sea, so that people involved in alpine or marine accidents can be rescued.

Direct Communication Service PMCSs must have a direct communication service function that allows PMCS's land mobile stations (MLs) to communicate other MLs without base stations (BSs). This function is very useful when signals from BSs cannot reach the MLs or if BSs are not worked by accidents.

1.3 Problems of NPMCSs

Although PMCSs play very important role as nerves systems of communications to assist public safety activities, NPMCSs have several problems to deal with public affairs effectively. In this section, We clarify two problems of NPMCSs by comparing commercial systems.

1.3.1 High Speed Transmission

Many PMCS's users are eager for realization of high speed communication. However, so far, realization of high speed communication is very difficult because the NPMCSs employ large-zone scheme. When employing large-zone scheme, PMCSs must select low carrier frequency such as VHF band whose distance decay is smaller than that of ultra-high frequency (UHF) band. In this case, high speed transmission can hardly be realized because of shortage of bands in VHF band and other reasons. For instance, in VHF band, MIMO application is difficult owing to long antennas and long antenna spacing. Equipment of many antennas and realization of low antenna correlation are difficult in a small ML. To make matter worse, since transmission power of a ML is high (more than 1W) in large-zone scheme; OFDM is not easily applied owing to problems that are high power consumption and heat radiation.

1.3.2 Improvement of Service Coverage

The other problem is the improvement of coverage in urban areas. Owing to the low density of their BSs (relative to cellular systems employing small-zone scheme), the large-zone scheme systems are often restricted by shadowing[4] which makes radio waves weaken. The BSs employing the large-zone scheme can hardly cover neighbor BSs' areas suffering from shadowing. NPMCSs have several no-service areas due to low density BSs. On the other hand, since CWMCSs allocate so many base stations in urban areas, stabilized service areas are realized. Especially in urban areas, CWMCSs provide users with more reliable service areas than PMCSs because BSs are located with high density and then can cover neighbor BS's areas each other. From this reason, The subscribers of the NPMCS require the service coverage areas of the NPMCS comparable to that of CWMCSs.

1.4 Broadband PMCSs (BPMCSs)

To solve the problems of NPMCSs, next generation PMCSs are being discussed. In this thesis, we describe the Long Term Evolution (LTE) for public safety and the public broadband wireless communication system (PBWCS).

1.4.1 Long Term Evolution (LTE) for the Public Safety

First, we describe the LTE for public safety. In the 3GPP [5], many customizations have been discussed to apply LTE [5] to public safety's communication systems. The typical features are the group communication and direct communication. By introducing LTE into the public safety communication, high speed transmission will be realized. However, we are worried whether underpopulated areas, which should effectively be covered by large-zone scheme, can be covered by the LTE system that is essentially used in small-zone scheme.

1.4.2 Public Broadband Wireless Communication System (PBWCS)

In Japan, public broadband wireless communication system (PBWCS) [6] is proposed as one of the next generation PMCSs. In Japan, digitalization of terrestrial TV broadcasting has made a new spectrum. 32.5 MHz of the new bandwidths can be allocated for the public safety communication systems. The PBWCS utilizes this allocated band to realize high speed transmission. In the police department, this system has already used for transmitting videos. However, currently, this system is large and heavy. So to make this system smaller and enhance the performance, diversity technique, high efficient ARQ and low power consumption have been considered. This system will be candidates of the Next-generation PMCS.

1.4.3 Problems of BPMCSs

Although several BPMCSs have been proposed, we must solve some problems. In the LTE for public safety, we must consider how underpopulated areas are covered on LTE systems effectively. In the case of the PBWCS, although underpopulated areas may be covered by using VHF band and large-zone scheme, MLs are too big and heavy so far. Hence, users of PMCSs need to utilize both of NPMCSs and BPMCSs in accordance with the situation.

1.5 Application of SDR and HCR

In this thesis, we propose application of SDR and HCR combining PMCSs, BPMCSs, and CMWCSs. As mentioned above, for the public safety activity, users need to equip NPMCSs, BPMCSs, and CMWCSs terminals. Having many terminals are quite inconvenient in accidents or disaster areas. Moreover, spectrum is probably wasted when several systems are operated independently.

By employing SDR and HCR, PMCSs will offer reliable and comfortable communication to users. And also frequency utilization efficiency will be improved. The proposed HCR recognizes wave radio condition of PMCSs and CMWCSs and then provides users with the best communication by taking user needs and the radio condition into account. In the proposed systems, the SDR is utilized for realization of the HCR by equipping several different radios on one software radio. The application is useful not only for conventional NPMCSs but also for BPMCSs. In this section, we explain advantages of the application and applying methods. Finally, problems solved in this thesis and the future are shown for the proposed HCR.

1.5.1 Advantages of Applying HCR

By employing HCR, four advantages are obtained as follows.

1.5.1.1 High Speed Transmission

When conventional NPMCSs are combined with BPMCSs or CMWCSs, NPMCS's users will be provided with high speed transmission.

1.5.1.2 Expansion of Service Coverage

The PMCSs will obtain new service areas when employing HCR. The HCR provides PMCS's users not only with original areas of a NPMCS but also with new areas covered by the combined systems. Especially, in the urban area, stabilized areas will be realized by CMWCSs. For LTE for public safety, small areas restricted by small-zone scheme are expanded by the conventional NPMCSs employing long-zone schme.

1.5.1.3 Increase of Reliability

Since several systems are available, system reliability is increased. If one system breaks down, other systems included in HCR help users continue communication.

1.5.1.4 Economic Efficiency

The HCR will be inexpensive because each system has already existed. Although the HCR must be constructed by combining existing systems, we consider that the cost of the HCR is cheaper than that of developing new systems.

1.5.2 Applying Methods

When applying the HCR to PMCSs, we consider three applying methods in this thesis. Those are a method of combining NPMCSs with CWMCSs, a method of combining NPMCSs, and a method of combining NPMCSs with BPMCSs. The three methods need to be studied to realize further improvement of next generations PMCSs.

1.5.3 Problems of Proposed HCR

To realize our HCR, several problems must be solved. The considered problems are as follows.

- BER improvement (service area expansion) methods
- synchronization methods
- network administration.
- delay control
- power consumption

Although several problems must be considered to realize our HCR, BER improvement methods of NPMCSs and BPMCSs, self-synchronization of NPMCSs, and co-synchronization between PMCSs and CWMCSs are discussed in the thesis. The other problems will be considered in future work.

1.6 Organization of the Thesis

The thesis summarizes our research works on reliable public safety mobile wireless communication systems employing SDR and HCR. The thesis consists of six chapters as follows.

Chapter 1 This chapter describes the research background, requirements and research motivation and objectives.

Chapter 2 This chapter presents the overview SDR and HCR. In explanation of the HCR, two types, spectrum sharing and heterogeneous, are given in this chapter.

Chapter 3 This chapter presents a BER improvement method for NPMCSs employing HCR. We explain the method in which CWMCSs assist NPMCSs to improve communication quality of the NPMCSs. By introducing integrated receive codewords (IRC) consisting of received signals of a NPMCS and subsidiary information (SI) delivered by CWMCSs, BER of NPMCSs can be improved.

Chapter 4 Synchronization methods are proposed in this chapter. We consider self-synchronization of NPMCSs and co-synchronization between NPMCSs and CWMCSs. We propose two methods; a GPS based method and a SI based method.

Chapter 5 This chapter presents the SDR integrating conventional PMCSs. Then the HCR combining a NPMCS with a BPMCS is shown. The HCR can be used for enhancing the BPMCS uplink performances.

Chapter 6 This chapter summarizes the research contribution of this thesis and explores the future works..

Chapter 2

Software Defined Radio and Cognitive Radio

2.1 Software Defined Radio

Nowadays, wireless systems have been developed significantly. Hence, we can utilize many wireless systems such as wireless LAN, cellular systems, WiMAX, GPS, and RFID. However, wireless systems that can freely change their functions are not realized. At present, user needs to prepare devices of several systems when they want to communicate other systems. To defeat such present conditions, software defined radio (SDR) is considered. SDR technique can change a system into another system by installing software to SDR[7][8][9]. Moreover, SDR can integrate several systems to one device. In this sub-section, we explain the concept and problem of SDR. In ITU-R, SDR is defined as follows[23].

"*Software-defined radio (SDR)*: A radio transmitter and/or receiver employing a technology that allows the RF operating parameters including, but not limited to, frequency range, modulation type, or output power to be set or altered by software, excluding changes to operating parameters which occur during the normal pre-installed and predetermined operation of a radio according to a system specification or standard."

2.1.1 Background of SDR Development

Although wireless communication systems are crucial for our information society, a specific issue needs to be solved because of characteristics of radio waves. That is mutual interference occurrence when radio waves are transmitted without any restriction. Since space where radio waves pass and frequency of radio waves are common resources for all users of wireless systems, utilization of radio waves is strictly restricted. Once a wireless system is fixed, changing frequency and modulation method of its system is quite difficult. To fix wireless communication schemes refuses new effective radio techniques, that is, technical choice is narrowed down and flexibility is lost. To solve this situation, SDR is planned as one candidate of future radio systems. Although SDR techniques have been originally developed for the military, these techniques are also used for commercial systems. SDR, which can freely change frequency, modulation methods, multipule access method, etc. by installing software, is greatly expected in terms of effective frequency utilization.

2.1.2 Concept of SDR

To clearly concept of SDR, differences between conventional digital radios and a SDR as are shown in Fig.2.1.

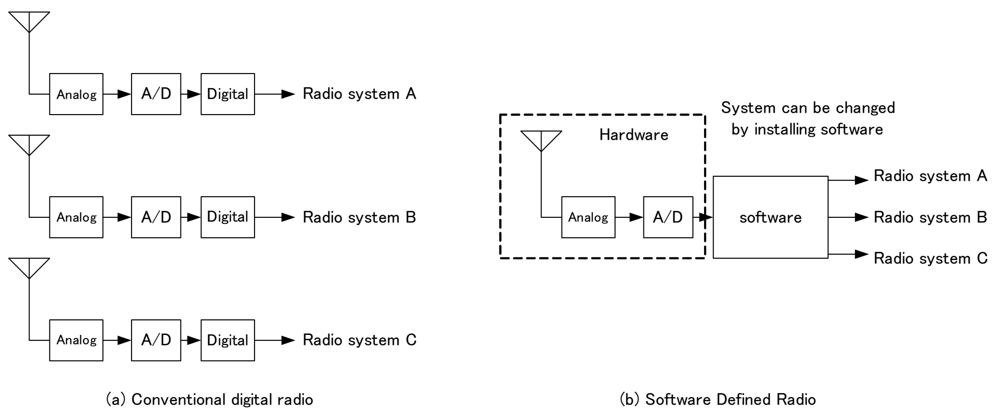

Figure 2.1: Conventional Digital Radios and Software Defined Radio(SDR).

Functions of conventional digital radios as shown in Fig. 2.1.(a) depend on hardware when they are manufactured. Hence, when using several radio systems, we must have each device. At present, although a radio including several systems has been developed such that one device can be used for several systems, it is not called the SDR. This is because this kind of radio is not changed by software. It is called multi-mode radio[7]. By installing software, SDR can freely change frequency, filter, modulation method coding method, etc.. Definition of SDR is summarized as follows.[7]

- Basic functions(sampling rate, filter, modulation, coding, etc.) are not fixed.
- Functions can be changed and add by installing software.
- Several systems can be realized on one device simultaneously.
- Programs installed in a SDR can be rewritten from the outside.
- Bug-fix and version up can be done by changing programs.

2.1.3 Problems of Radio Act

In the case of transmitting radio wave without restriction, interference is unavoidable. Hence, in order to realize environments where harmful interference does not occur, transmitting of radio waves must be restricted by considering physical property of radio waves. [7]. Acceptable use policy of radio waves is defined in detail in the world. In Japan, SDR is probably positioned as ”radio equipment” prescribed in the Radio Act and obligated to obtain ”technical standards conformity certification” by a public institution as conventional radio devices are required. Since modification of the radio equipment is not allowed after obtaining technical standards conformity certification, SDR that can freely change own system by software is not perhaps allowed as the radio equipment that can satisfy the Radio Act. Hence, we think that the Radio Act needs to be revised to allow commercial applications of SDR after some functions of SDR are restricted.

2.1.4 Technical Problems for Realization of SDR

2.1.4.1 Antenna

SDR requires broadband antennas so that the system can be changed and frequency can freely be used. Moreover, in the case of MLs, it also requires small antennas.

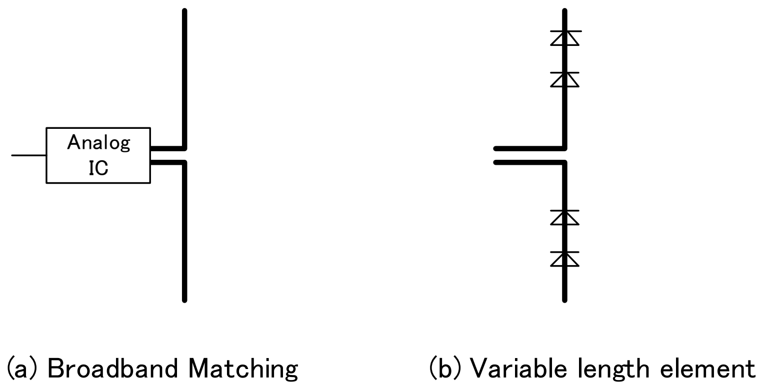

Figure 2.2: Broadband of antenna [7].

Figure 2.2 shows example of antennas that can be broadband. Figure 2.2(a) shows concept of a broadband antenna equipping a broadband matching circuit of analog IC. In general, input impedance characteristic of small antennas is narrowband. In this antenna, by using active characteristics of the analog IC, the narrowband characteristic of the antenna is compensated. Figure 2.2(b) shows concept of changing antenna length depending on using frequency while paying attention to a resonance phenomenon of antennas. It is one candidate of realizing plural resonance point. In this example, the antennas can be resonated by turning on or off diodes included in the antenna so that length of the antenna can be changed.

2.1.4.2 A/D Converter (receiver)

Ideal SDR is a digital system that can receive signals of the entire bandwidth and convert analog signals to digital signals. However, in general, to directly convert signals received by an antenna to digital signals is difficult because of problems of low sampling rate and resolution.

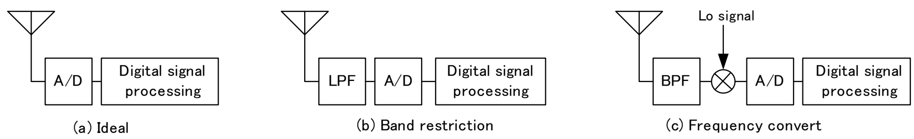

Figure 2.3: Ideal SDR and realistic SDR.

According to the sampling theorem of Nyquist, an analog signal needs to be sampled by two times frequency of the highest frequency of the signal in order to sample the signal without distortion. Figure 2.4 shows the extent of spectrum overlapping when sampling rate is changed. In the case of Fig. 2.4(c), spectrum overlapping, which is called aliasing[20], occurs. In this case, the original analog signal cannot be restored from the sampled signal. If frequency of a signal received by an antenna is A GHz, the sampling rate of an analog-to-digital converter (A/D) needs to be more than 2A GHz. However, realization of an A/D that can sample signals whose highest frequency is GHz is difficult at present.[7]

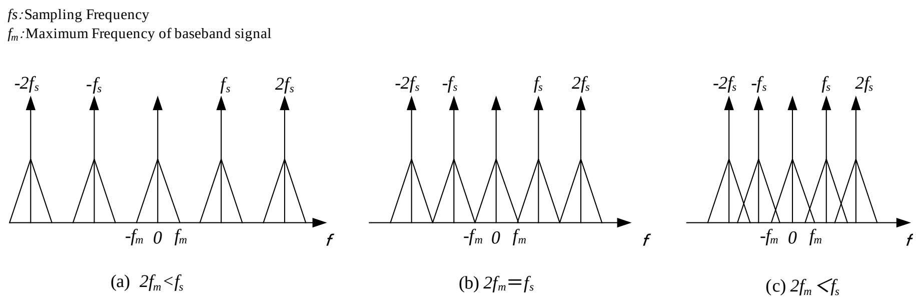

Figure 2.4: Sampling theorem of Nyquist.

Moreover, when signals of radio frequency are directly sampled, resolution of A/D is too low at current technologies. In the case of mobile phone systems, if a signal is converted to a base band signal, the resolution needs to be 5 to 6 bit. However, if a signal of radio frequency is directly sampled, the resolution needs to be about 20 bit.

2.1.4.3 Frequency Converter (receiver)

To reduce the sampling frequency, we consider that radio frequency is converted to lower frequency and then the converted signal is sample. By converting RF to IF, the signal can be sampled and avoid aliasing. Moreover, converting IF is effective for reducing jitter. This is because the jitter becomes larger as frequency of signals increases. Figures 2.5, 2.6, and 2.7 show the methods that convert frequency of signals and then sample the signals. Well-known methods for frequency convert are super-heterodyne, IF sampling, and direct conversion.

- super-heterodyne
  Since a super-heterodyne method shown in Fig. 2.5 has several local frequency multiplexing and analog filter to convert frequency effectively, it is usually used for conventional radio systems. Hence, firstly, we consider that super-heterodyne is applied to SDR. However, in the case of SDR employing super-heterodyne, limited frequency is converted. Hence, small SDRs are not probably realized because many devices for the super-heterodyne are necessary for supporting multi systems. The SDR needs a number of analog devices in proportion to systems that the SDR realizes. When new system is installed, new analog devices may need to be equipped, that is, the SDR cannot be changed to another system by only re-installing software.

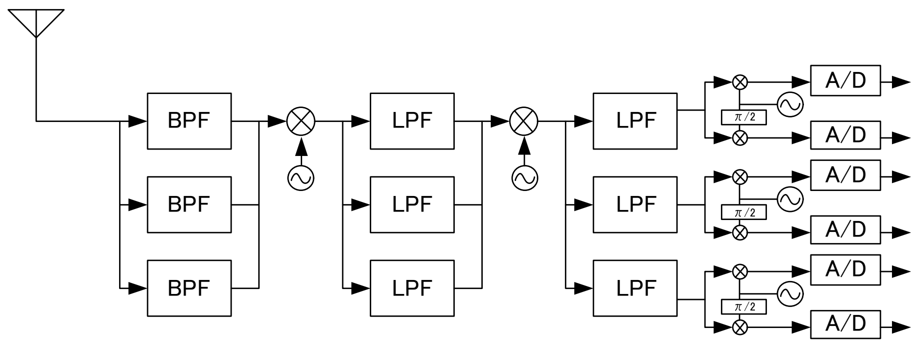

Figure 2.5: Super-heterodyne method.

- IF sampling
  Figure 2.6 shows an IF sampling method. This method samples IF signals. By down-converting RF signals to IF signals, speed of sampling rate can be reduced. Although the IF sampling method can easily deal with digital processing, to make small SDRs is difficult as well as the super-heterodyne method. This is because many analog devices are required.

- Direct Conversion
  The direct conversion samples RF signals directly to base band (BB) signals as shown in Fig. 2.7. The method is useful for making small SDRs because of reducing analog components. However, the method requires RF band-path filters that have high Q to prevent DC offset and back radiation

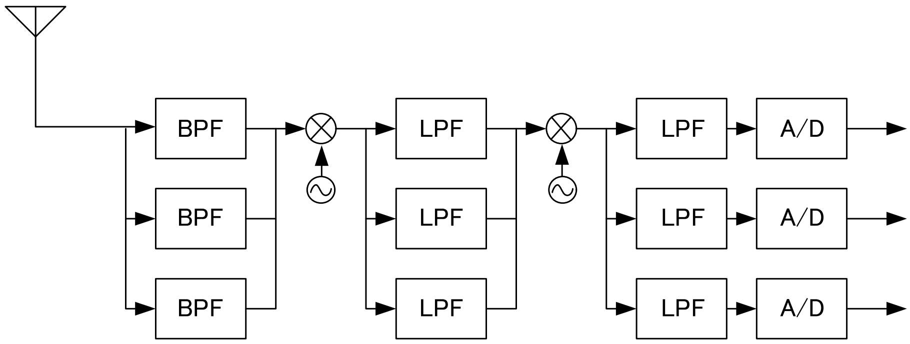

Figure 2.6: IF Sampling method.

because of adverse currents of local signals. Nevertheless, since the direct conversion method can reduce analog components and deal with broadband signals, application to SDR such as USRP [25] is widely researched.

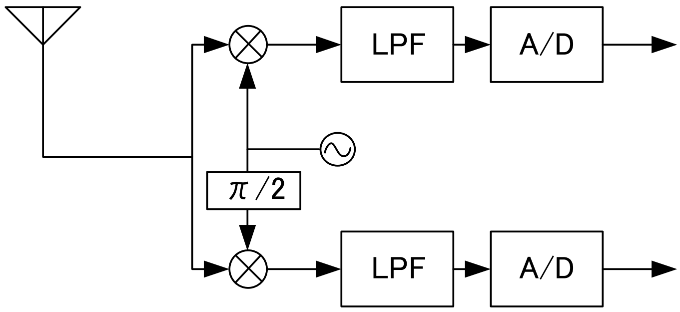

Figure 2.7: Direct conversion method.

2.1.4.4 Frequency Converter (transmitter)

The direct conversion method is the best in terms of miniaturization and easily system change in transmitters as well as receivers. Figure 2.8 shows the transmitter of the direct conversion.

Since the transmitters of the direct conversion have problems that are local leak and image output, the researches for reduction of the problems are underway[7]. Figure 2.9 shows basic structure of the Digital-to-RF direct converter.

In the Digital-to-RF direct converter, the most important device is a digital-to-analog converter (D/A). When a D/A is used for converting RF signals, following problem must be solved.[7]

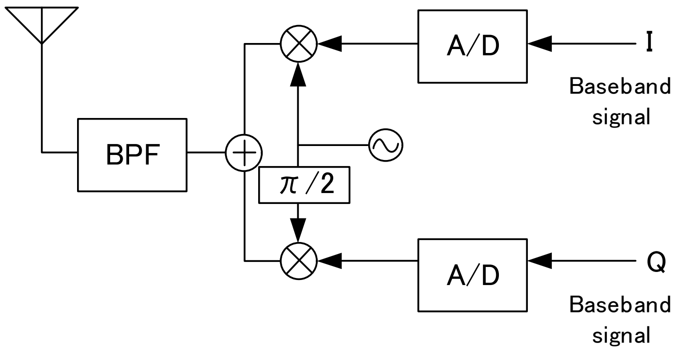

Figure 2.8: Direct Conversion method for transmitters.

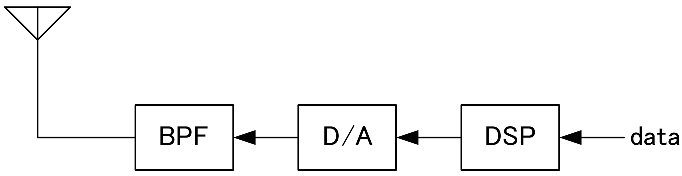

Figure 2.9: Digital-to-RF direct converter.

- Nonlinearity caused by stray capacitance and other factors
- Frequency characteristic depending on amplitude of input signals
- Deterioration of signals in the case of amplitude change

As frequency of RF signals, to produce amplitude change is more difficult in the structure of Fig. 2.8. To solve this, new types of D/A, pulse density modulation, pulse width modulation, and other methods, are researched and proposed.

2.1.4.5 Digital Signal Processing

Field programmed gate array (FPGA), digital signal processor (DSP), and application specific integrated circuit (ASIC) are used for digital signal processing devices of SDR. These devices can change own functions by altering programs.

- FPGA
  FPGA is one of the programmable logic devices (PLD) that can construct arbitrary logic circuit by programing. FPGA can directly use programing

programming language to construct real circuits such as combinational circuits or sequential circuits. An advantage of FPGA is high speed signal processing that can work with more than several hundred MHz in spite of programmable circuits. A disadvantage of FPGA is to take a lot of time when a system is changed to another system.

- DSP
  DSP is a digital signal processing device that is programed by C language and assembly language and then compiled. DSP can quickly change own functions by providing target devices constructed in advance with run commands of software. However, so far, if necessary functions of SDR are realized by employing only DSPs, many DSPs may be required and power consumption is large. After this, high-performance and low power consumption are important issues.

- ASIC
  Basically, ASIC is not suitable for SDR because the functions are not changed by programs. However, if SDR can be constructed by ASIC, high-speed processing and low power consumption attract us. Hence, some institutes research SDR employing ASIC that can change functions by altering parameter of ASIC.

2.1.4.6 Software

Software of SDR affects not only characteristics of radio systems but also hardware structure, re-construction time, development time, cost, etc. (Simultaneously, hardware structure sometimes affects software.) Hence, establishment of efficient and effective software design methods is importance issues for SDR. [2]

- Ensuring Compatibility
  To diffuse SDR, standards of programming methods, interface, etc. are required, that is, development of software that does not depend on hardware is necessary. To realize this, standards for compatibility should be decided in software and both hardware companies and software programming methods should be also standardized.

- Design Methods of software
  Software framework standardized in Software Communication Architecture (SCA) as open architecture is used for software design method of SDR. The software framework of the SCA is developed based on Common Object Request Broker Architecture (CORBA) included in Object Management Group (OMG). Each Module (application) can be used as components. Moreover, the reuse and addition of modules can be easily realized because API is regulated and hardware is abstracted to design the software. However, a problem is to require a large overhead because of using CORBA.[5]

- Avoidance of Running Malicious Programs
  Since alteration and illegal copy of programs can easily be done, protecting techniques are very important to prevent it. Moreover, SDR must be constructed so that malicious programs cannot be installed or can be rejected. Prevention techniques for rejecting malicious programs are the most important issue in SDR techniques.

2.1.5 Summary

In the sub-section, we explain structure and its problems. At present, to realize cellular systems by employing SDR has many problems such as low digital processing device performance, large radio components, high power consumption, and high costs. However, we think that these problems will gradually be solved as devices characteristics and software programing methods are developed.

2.2 Cognitive radio (CR)

The concept of CR was proposed by Mitola [22]. At present, since frequency allocation of radio systems is very tight, cognitive radio (CR) that can effectively communicate by reconstructing frequency allocation and radio systems is widely researched. The CR can select the optimal radio systems or decide radio communication functions such as using frequency and modulation methods by sensing radio wave environments around. The CR can select the optimal radio systems or can decide radio-communication function such as frequency employments and modulation methods by sensing radio wave environments around. In ITU-R, the CR is defined as follows.[23]

"*Cognitive radio system (CRS)*: A radio system employing technology that allows the system to obtain knowledge of its operational and geographical environment, established policies and its internal state; to dynamically and autonomously adjust its operational parameters and protocols according to its obtained knowledge in order to achieve predefined objectives; and to learn from the results obtained."

2.2.1 Relationship between SDR and CR

CR is closely related to SDR. However, strictly speaking, CR and SDR are not the same. CR is positively considered as an important technique to utilize radio resource effectively and developed actively by using SDR. SDR is a radio that can process all signals by software. SDR is used for realization of CR. CR employing SDR recognizes radio wave environments and then comprehends the results, and then can coexist in the environments. CR can hardly be realized without SDR techniques, and therefore CR and SDR are researched simultaneously[10][11]. In practical research, since SDR is mainly developed for realization of CR, CR and SDR cannot be researched separately.

2.2.2 Classification of CR

CR is classified broadly into two types that are called the white space and the heterogeneous. In the white space type, white space means frequency bands that are not used geographically or temporally. The white space type temporarily employs white space that is originally used by other systems. Heterogeneous type means that CR selects the optimal systems or combines several systems.

2.2.3 Cognitive Radio (CR) for White Space

Utilization of radio systems is usually restricted by licenses and regulations defined in the radio act. However, the licensed frequency bands are not geographically or temporally used 100%. In some cases, spectral of certain systems may be very inefficient. Hence, the vacant frequency bands should be used by other systems to improve spectral efficiency. However, to deal with rapid increase of radio frequency demand is difficult because change or reorganization of frequency allocation usually needs very long time more than several years. To solve this problem, white space, frequency bands that is not used geographically or temporally by original systems, attracts attention. In particular, TV white space (TCWS) that is white space of terrestrial TV broadcast bands is frequently researched. To employ white space, intelligent communication systems need to recognize radio frequency environments around and then choose vacant frequency for communication. One candidate of the intelligent communication systems is CR. CR must comprehend radio wave conditions and then change employing frequency and communication parameters. The CR using white space suppresses interference to original licensed systems by technical methods. Technical standards of white space CR are provided in the world.

2.2.4 Heterogeneous Cognitive Radio (HCR)

Heterogeneous cognitive radio (HCR) can recognize several radio systems such as cellular systems, wireless LAN, and WiMAX simultaneously and then inform the results to user or select the optimal systems automatically by considering high-speed communication, stable communication, and cheaper communication fee. HCR can use surplus radio resource of other systems effectively and select radio systems in according with user demand. To realize the HCR, the system needs to sense radio systems and then select the optimal system or multiplex several systems to get required bandwidth. First, the system obtains sensing information of other user from database on networks and then adds the information to own sensing information. Next, the system constructs the optimal systems, considering the added information and information that user wants to transmit.[24]. For example, ER2（End-to-End Configurability）[26] aims to realize the system combining various existing radio network taking users' needs into account. The HCR reconstructs user terminals, IP networks, and services dynamically in the point of view of end-to-end.

According to [27], following three steps are considered for the spread of HCR. As the first step, HCR terminals recognize several systems and then offer the optimal radio resource to users by considering users' needs and frequency utilization efficiency. The HCR terminals communicate seamlessly between heterogeneous systems. In the second step, networks begin to equip intelligence. By gathering user information in the networks, the optimal communication is offered. In the third step, communication method of base-stations can also be changed. By selecting modulation method, frequency, and transmitting timing, further improvement of frequency utilization efficiency can be realized.

In this thesis, we mainly focus on HCR combining PMCSs with CWMCSs.

2.3 Chapter Summary

In NPMCS, radio systems such as STD-39 and T-79 do not have compatibility. SDR and CR can realize integration of PMCSs and utilize several systems simultaneously. Additionally, once SDR that can change own systems by altering software is employed to PMCSs, new hardware may not be required to introduce a new system. This shows a possibility that the cost concerning maintenance of a new system can be suppressed. Moreover, by employing HCR, more intelligent PMCSs can be realized. From these advantages, application of SDR and HCR is expected in future PMCS.

Chapter 3

BER Improvement method of Narrowband PMCSs using Heterogeneous Cognitive Radio

In this section, We propose a BER improvement method for NPMCSs. In urban areas, the communication quality of the NPMCS is sometimes inferior to that of CMWCSs, e.g., cellular communication systems, which employ small zone schemes. CMWCSs provide reliable service in urban areas owing to high-density base station deployment; however, NPMCSs have several no-service areas because of low-density base station deployment. In a large-zone scheme, no-service areas due to shadowing [28] [29] [30] are often not covered by adjacent base stations. In urban areas, the NPMCS service areas are smaller than that of cellular systems.

By applying CWMCSs, expansion of service area and improvement of liability will be expected. In the above advantages, although realization of high rate transmission will be realized, expansion of service areas is mainly discussed in this thesis.

To improve the reliability and service coverage areas of a NPMCS in urban areas, we have proposed combining a conventional NPMCS with a CMWCS using HCR techniques. Fig 3.1 shows the system model with the HCR. When the communication quality of the NPMCS is adversely affected by shadowing, the HCR acquires additional information from the CMWCS to improve the communication quality of the NPMCS. We refer to this information as subsidiary information (SI). The HCR recognizes communication quality of the NPMCS. If the communication quality of NPMCSs is deteriorated, the HCR acquires the SI from the CMWCS. The amount of the SI is changed depending on degree of the deterioration.

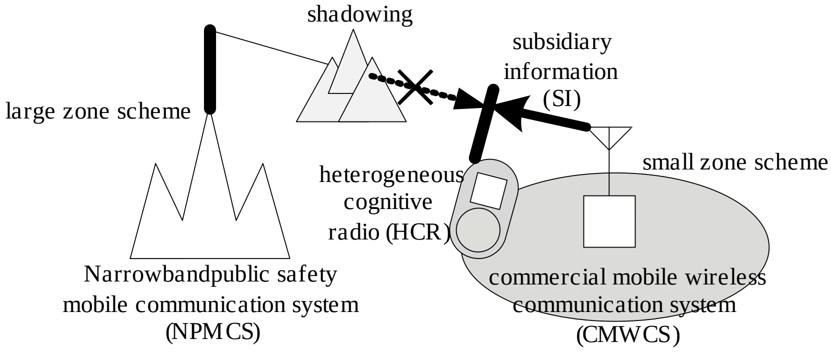

Figure 3.1: Schematic of the proposed HCR system.

The BER improvement method employs the subsidiary information (SI) conveyed by CWMCSs for decoding of the convolutional code which is used in the NPMCS for forward error correction (FEC). The SI is combined with original convolutional codewords in decoding process for BER improvement. We refer to the combined codewords as integrated received codewords (IRC). The proposed HCR realizes system diversity to improve NPMCS's communication quality. So far, no one evaluate the MLSE for signals combining several codewords that have mutually different credibleness. In this research, we derive the method that can evaluate BER improvement without simulation.

To evaluate BER performance, We derive maximum likelihood sequence estimation (MLSE) for the IRC and calculate distance spectrum which represents capability of the convolutional code and theoretical BER performance. In this section, We verify that the theoretical BER closely approximates BER determined by computer simulation.

In Section 3.1, requirements of the proposed system is introduced. In Section 3.2, the structure of the proposed HCR is shown. In Section 3.3, I present BER improvement method. To decode the proposed method, the maximum likelihood sequence estimation (MLSE) and the Viterbi algorithm are derived for IRC. Moreover, for calculating the theoretical BER performance of the proposed system, we introduce distance spectrum customized for the proposed systems, which indicates the performance of convolutional codes. Then, we show that the

BER performance calculated from the distance spectrum closely approximates the performance determined by computer simulation. Furthermore, we also provide an analysis on the BER performance and discuss that how much the proposed system can improve the performance. Finally we conclude this chapterr in the end of the section.

3.1 Requirements of the Proposed System

3.1.1 Ability to Operate without Commercial Systems

Assuming that the received signal power from NPMCSs are high enough, the proposed HCR must be used even if any commercial system is not located around. Otherwise, the proposed system cannot be used in mountainous area or the sea area where the commercial systems are not available. The object of the HCR is to improve BER performances in the urban areas.

3.1.2 Saving Resource of CMWCS

Radio resource consumption of the commercial systems needs to be minimized. Since the SI that is originally transmitted by the NPMCS must be transmitted by the commercial systems; the burden of the NPMCS moves to the commercial systems, the HCR should reduce the resource consumption of the commercial systems as possible. Hence, the HCR assesses the communication quality and then makes the commercial system send the minimum subsidiary information that can be used to decrease BER until required level.

3.1.3 Avoidance of Eavesdropping

In the proposed HCR system, the significant information should not be restored only by the SI. That is, as PMCSs need to strictly protect communicative secrecy in some cases, divulging significant information from the commercial systems should be avoided. In this thesis, the HCR must be constructed so that any eavesdropper cannot understand the received information even if the commercial systems can be eavesdropped.

3.1.4 Utilization of Machine to Machine (M2M)

We assume that CMWCS' receivers included in the HCR are extant products such as M2M (Machine to Machine) module. This is because structures of recent CMWCS are so complex that we cannot modify the CMWCS' receiver for the proposed HCR. In the case of the extant module, the HCR gains correct hard decision values as the SI after CRC (cyclic redundancy check) and ARQ (automatic repeat-request) perform in the M2M. Although we suppose that these hard decision values are always correct in this thesis, in the future work, we must consider situations where the HCR does not gain correct values.

3.2 System Structure

Figure 3.2 shows a construction of the proposed HCR that is designed to overcome above three constraint conditions.

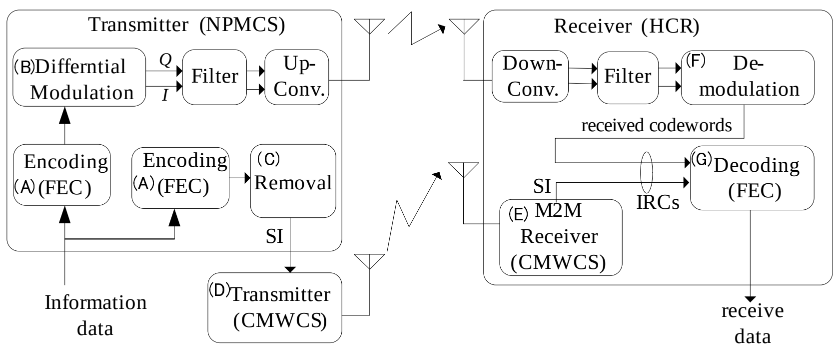

Figure 3.2: HCR combining a NPMCS and a CMWCS.

In the transmitter of the NPMCS, send data (digital voice data) is split into two parts and coded into separate codewords by using separate encoders. In the NPMCS transmitter, the send data is coded by a forward error correction (FEC) encoder (see module (A) in Fig. 3.2). Convolutional code is often used for FEC of the NPMCS. The two codewords are input into two modules. One module is the differential modulator (Fig. 3.2, module (B)). The codewords are modulated, filtered, up-converted, and transmitted by an antenna as NPMCS's transmitting signals. The other module is Removal (Fig. 3.2, module (C)) where the codewords are periodically removed to reduce the amount of information arriving at the CMWCS. The removal rate can be changed depending on the level of deterioration of the NPMCS's communication quality. The output signals of the module constitute the SI, which is transmitted by the CMWCS (Fig. 3.2, module (D)).

In the HCR, signals transmitted by the NPMCS are down-converted, filtered, demodulated, (Fig. 3.2, module (F)) and, finally, combined with the SI from the CMWCS (Fig. 3.2, module (E)); we refer to these combined signals as integrated received codewords (IRCs). The IRCs are decoded and restored to "receive data" (Fig. 3.2, module (G)). By adding the SI, BER performance is improved because of the increase in the free distance [35][36] of the IRCs compared with that of conventional codewords of the NPMCS.

3.2.1 Verification of Requirements

Four requirements can be satisfied by the proposed system as follows. The first requirement can be realized using the convolutional decoder, which can be decoded without subsidiary information. The next two and three requirements can be realized by partially removing the codewords before handed to the commercial systems. Especially for the last requirement, the security can be improved by making coding rate of the subsidiary information exceed 1. This is because unique information cannot be determined from the codewodes whose code rating is over 1. The rest requirements are achieved by employing M2M module.

3.2.2 Consideration of SI

Using the M2M module of the commercial systems, the HCR obtains correct hard-decision values (CHDV) as the SI after a cyclic redundancy check (CRC) and an automatic repeat-request (ARQ) are performed in the M2M module.

However, in the general heterogeneous cognitive networks, Soft Decision Values (SDV) is often used for exchange of information between heterogeneous systems. Hence, in this chapter, we consider both CHDV and SDV so that the proposed method can be applied to other HCR. Although CHDV can be defined as SDV whose SNR is infinite, in this chapter, we distinguish CHDV from SDV because the proposed HCR employs CHDV.

When CHDV that can passed CRC are employed, effect of FEC is much larger than that of SDV. However, in the case of failure of CRC and ARQ, SI would be lost. To solve this problem, two solutions can be considered as follows. First, we can consider the erasure correction code for restoring the SI. Second, we can consider interleaving for scattering no data duration on the SI. However, since the consideration of these solutions makes the proposed system too complex, we suppose that the SI of hard decision values is always correct when hard decision values are employed. In future work, We must consider situations where the HCR does not obtain correct values.

3.3 BER Improvement Method

3.3.1 Employing Convolutional Code and Viterbi Algorithm

In this thesis, we employ convolutional code as Forward Error Correction (FEC) and modify the Viterbi algorithm, although strong FEC codes such as turbo code[31] and LDPC[32] have been developed recently. This is because NPMCSs are used for voice communication in most case; voice communication requires few delay time for stresses conversations. If we employ recursive FEC codes, turbo code and LDPC, large delay time cannot be prevented owing to large size interleaver for iteration. In particularly, since transmission rate of NPMCSs is very slow (several kbps), equipping of large size interleaver is difficult. We therefore employ convolutional code as NPMCS's FEC code.

3.3.2 Utilization Methods of SI

We propose two types of SI utilization method. We name the two utilization method "replacement method" and "addition method".

3.3.2.1 Replacement Method

Figure 3.3 shows the decoding structure of the replacement method. In this figure, coding rate of a NPMCS is 1/2. In the replacement method, received signals of a NPMCS are periodically replaced with SI conveyed by CMWCSs. The advantage of the method is that we do not need to change a Viterbi decoder that is originally employed in a NPMCS. By replacing NPMCS's received signals with SI that does not have error information, BER performance can be improved.

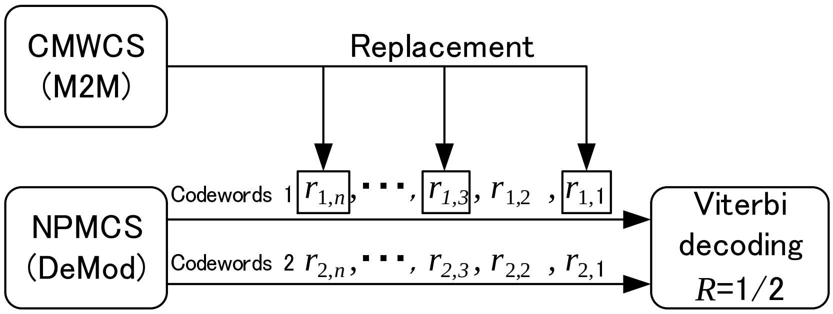

Figure 3.3: Replace method.

3.3.2.2 Addition Method

Figure 3.4 show the decoding structure of the addition method. Coding rate of a NPMCS is also 1/2. In the addition method, the SI is periodically added to the received signals of a NPMCS. The SI may be regarded as new codewords of the NPMCS in this case. We can freely select generating polynomial employed for producing SI in the addition method, although SI employed in the replacement method must be produced from the generating polynomial of the NPMCS. BER performances of the addition method are superior to those of the replacement method. This is because coding rate of the IRC in the addition method can be reduced owing to adding SI to received signals, while the replacement method cannot reduce the coding rate of IRC. The IRC of the addition method looks like punctured code[33][34] whose coding rate was originally 1/3. The addition method is not discard any received signals of a NPMCS. However, the decoder of the addition method is more complex than that of the replacement method because the addition method must change the coding rate of the decoder.

3.3.3 Decoding of IRCs and its Performance

For design of the decoder, we consider above two methods for utilization of the SI. Although we employ the addition method described in this thesis, the replacement method can easily be evaluated by slightly changing the addition method.

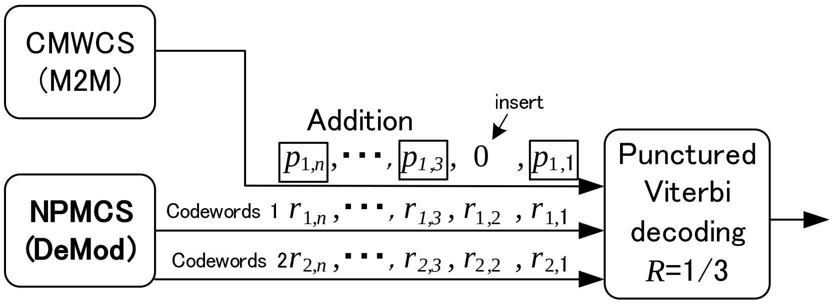

Figure 3.4: Addition method.

Figure 3.5 shows an example of the addition method. The coding rate of the original NPMCS is 1/2 and that of the commercial system is over 1. In this method, both the original signals from the NPMCS and the SI are decoded simultaneously. As shown in Fig. 3.5 , although the original coding rate is 1/2, the coding rate of the decoder employing the addition method becomes lower than 1/2, resulting in punctured code of 1/3.

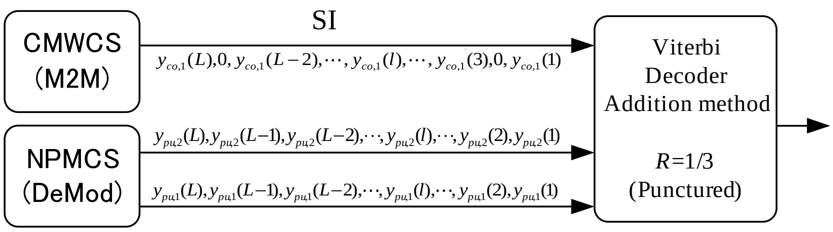

Figure 3.5: The proposed decoding method (addition method).

Firstly, we formulate the information bits, the codewords, the received signals and the noise signals, where the L bits of the information bits are denoted by $\boldsymbol{a} = [a_1, a_2, ..., a_L]$ by using indices $(1 \leqq l \leqq L)$. The convolutional encoder in the NPMCS outputs M words per one information bit (M : natural number $(1 \leqq m \leqq M)$. In Fig. 3.5, M=2). Coding rates of the convolutional codes transmitted by the NPMCS, R_{pu}, is $1/M$. The lth codewords which are transmitted by the NPMCS are denoted by $\boldsymbol{x}_{a,pu}(l) = [x_{a,pu,1}(l), x_{a,pu,2}(l), ..., x_{a,pu,M}(l)]^T$, where is $x_{a,pu,m}(l)$ represent either 1 or -1 indicating lth and mth codeword. The received

$$\boldsymbol{y}_{pu}(l) = r_{pu}\boldsymbol{x}_{a,pu}(l) + \boldsymbol{\omega}_{pu}(l), \tag{3.1}$$

where $\boldsymbol{\omega}_{pu}(l) = [\omega_{pu,1}(l), \omega_{pu,2(l)}, ..., \omega_{pu,M}(l)]^T$ represents noise signals whose power is normalized by power of the receive signals. We assume that $\mathrm{E}[\omega^2_{pu,m}(l)]$ is a constant σ^2_{pu}.

In the same manner, the signals from the commercial systems can be formulated. The convolution encoder which produces codewords transmitted from the CMWCS's outputs N (N : natural number) words per one information bit, where the codewords are denoted by $\boldsymbol{x}_{a,co}(l) = [x_{a,co,1}(l), x_{a,co,2}(l), ..., x_{a,co,N}(l)]^T$ ($x_{a,co,n}(l) = 1$ or -1) and coding rates of the convolutional codes transmitted by the CMWCS, R_{co}, is $1/N$. Moreover, the received signals, the noise signals and the noise power are denoted $\boldsymbol{y}_{co}(l) = [y_{co,1}(l), y_{co,2}(l), ..., y_{co,N}(l)]^T$, $\boldsymbol{\omega}_{co}(l) = [\omega_{co,1}(l), \omega_{co,2}(l), ..., \omega_{co,N}(l)]^T$ and σ^2_{co} ,respectively. If $x_{a,co}(l)$ is removed periodically for saving the radio resources and avoiding eavesdropping, the received removal codewords are filled with 0 in the receiver.

In this study, we consider both cases that $y_{co}(l)$ is soft decision (SD) values and correct hard decision (CHD) values.

3.3.3.1 Derivation of MLSE and Modified Viterbi Algorithm

In this section, I derive maximum likelihood sequence estimation (MLSE) for the convolutional code decoding procedure using the addition method. Assuming additive white Gaussian noise (AWGN), the probability $P(\boldsymbol{y}_{pu}, \boldsymbol{y}_{co}|\boldsymbol{a})$ can be determined by

$$\begin{aligned} P(\boldsymbol{y}_{pu}, \boldsymbol{y}_{co}|\boldsymbol{a}) &= \left(\frac{1}{2\pi\sigma^2_{pu}\sigma^2_{co}}\right)^L \prod_{l=1}^{L} \exp\left(-\frac{|\boldsymbol{y}_{pu}(l) - r_{pu}\boldsymbol{x}_{a,pu}(l)|^2}{2\sigma^2_{pu}}\right) \times \\ &\quad \exp\left(-\frac{|\boldsymbol{y}_{co}(l) - r_{co}\boldsymbol{x}_{a,co}(l)|^2}{2\sigma^2_{co}}\right) \\ &= U \exp\left(-\frac{r_{pu}\boldsymbol{y}_{pu}(l)^T \cdot \boldsymbol{x}_{a,pu}(l)}{\sigma^2_{pu}}\right) \exp\left(-\frac{r_{co}\boldsymbol{y}_{co}(l)^T \cdot \boldsymbol{x}_{a,co}(l)}{\sigma^2_{pu}}\right), \end{aligned} \tag{3.2}$$

where U is a constant value that is independent from $x_{a,pu}(l)$ or $x_{a,co}(l)$, and $|x_{a,pu,m}(l)|^2 = |x_{a,co,m}(l)|^2 2 = 1$. Estimation of amlse by MLSE can be obtained by

$$\boldsymbol{a}_{mlse} = \max_{\forall \boldsymbol{a}} P(\boldsymbol{y}_{pu}, \boldsymbol{y}_{co}|\boldsymbol{a}). \tag{3.3}$$

Then, we can simplify Eq.3.2 for reducing the computational complexity as follows. By taking logarithm of both sides in Eq.3.2 and defining a constant value U', we can obtain

$$LLF(\boldsymbol{y}_{pu}, \boldsymbol{y}_{co}|\boldsymbol{a}) = \sum_{l=1}^{L} \left(\frac{r_{pu}\boldsymbol{y}_{pu}^{T}(l) \cdot \boldsymbol{x}_{a,pu}(l)}{\sigma_{pu}^{2}} + \frac{r_{co}\boldsymbol{y}_{co}^{T}(l) \cdot \boldsymbol{x}_{a,co}(l)}{\sigma_{co}^{2}} \right) + U', \tag{3.4}$$

where LLF stands for Log likelihood function.

Since $P(\boldsymbol{y}_{pu}, \boldsymbol{y}_{co}|\boldsymbol{a})$ in Eq.3.2 is monotonically increasing function with regard to $y_{pu}^{T}(l) \cdot x_{a,pu}(l)$ and $y_{co}^{T}(l) \cdot x_{a,co}(l)$, we can infer that $LLF(\boldsymbol{y}_{pu}, \boldsymbol{y}_{co}|\boldsymbol{a})$ is also monotonically increasing function. Hence, the largest $LLF(\boldsymbol{y}_{pu}, \boldsymbol{y}_{co}|\boldsymbol{a}_{mlse})$ is defined as MLSE output for an arbitrary $\boldsymbol{a}$.

Next, Viterbi algorithm is derived from Eq.3.4. The part inside the summation in Eq.3.4 can be defined as branch metric BR.

$$BR(l) = \frac{r_{pu}\boldsymbol{y}_{pu}^{T}(l) \cdot \boldsymbol{x}_{a,pu}(l)}{\sigma_{pu}^{2}} + \frac{r_{co}\boldsymbol{y}_{co}^{T}(l) \cdot \boldsymbol{x}_{a,co}(l)}{\sigma_{co}^{2}}. \tag{3.5}$$

Since the summation of Eq.3.5 results in LLF, Viterbi algorithm can be implemented by defining accumulation of BR as metric J, where the survivor path for Viterbi algorithm is the path that has largest J.

The inner products $y_{pu}^{T}(l) \cdot x_{a,pu}(l)$ and $y_{co}^{T}(l) \cdot x_{a,co}(l)$ decrease as the noise power σ_{pu}^{2} and σ_{co}^{2} increase, respectively. This is because $y_{pu}(l)$ and $y_{co}(l)$ include noise signal $\omega_{pu}(l)$ and $\omega_{co}(l)$ whose average power are σ_{pu}^{2} and σ_{co}^{2}, respectively. Moreover, the inner products are divided by the noise power for calculating Eq.3.5. The noise power affects the calculation of the branch metric twice in Eq.3.5; the branch metric decreases exponentially as noise power increases, while the branch metric of the conventional Viterbi algorithm is not divided by the noise power σ_{pu}^{2} and σ_{co}^{2} in Eq.3.5.

3.3.3.2 Error Occurrence and its Probability

To search the case of an error occurrence, I shows a concrete example. Firstly, R_{pu} and R_{co}, are defined as 1/2 ($M = 1/R_{pu} = 2$) and 1 ($N = 1/R_{co} = 1$), respectively. The constrain length K is 3. We consider initiating all the information in $\boldsymbol{a}_0$ to 0, and accordingly, all the codewords are 0, i.e., all the transmitted signals are -1. That is, the codewords transmitted by the NPMCS and the commercial system are $x_{a0,pu}(l) = [-1, -1]^T$ and $x_{a0,co}(l) = -1 (1 \leqq l \leqq L)$, respectively. Figure 3.6 shows the trellis diagram of this assumption.

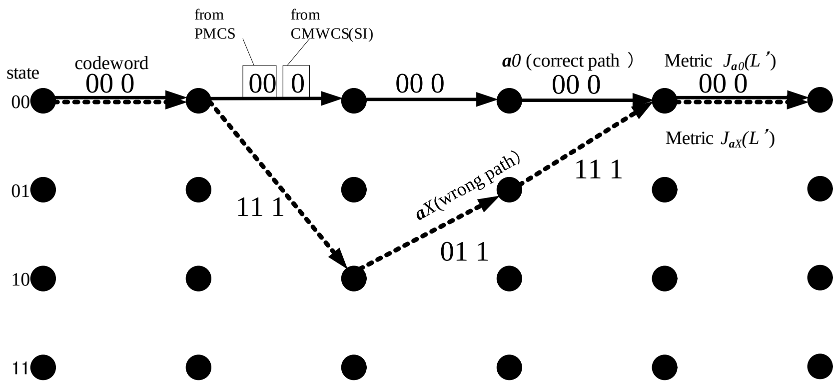

Figure 3.6: A trellis diagram of this assumption.

Then, we consider that MLSE applying the Viterbi algorithm for the proposed system. In the Viterbi algorithm, if a wrong path is chosen as the survivor pass, decoding error occurs. In the case of Fig 3.6, the wrong path transitions from 00 state to any other state and then goes back to 00 state after several transitions. When the path with all 0 codewords does not survive, decoding error occurs. The metric used to determine the survivor path at time L' is obtained by Eq.3.4. The metrics of the correct path $\boldsymbol{a}_0$ and the wrong path $\boldsymbol{a}_X$ is determined by

$$J_{a0}(L') = \sum_{l=1}^{L'} \left(\frac{r_{pu}\boldsymbol{y}_{pu}^T(l) \cdot [-1, -1]^T}{\sigma_{pu}^2} - \frac{r_{co}y_{co}(l)}{\sigma_{co}^2} \right)$$
$$J_{aX}(L') = \sum_{l=1}^{L'} \left(\frac{r_{pu}\boldsymbol{y}_{pu}^T(l) \cdot \boldsymbol{x}_{ax,pu}(l)}{\sigma_{pu}^2} - \frac{r_{co}y_{co}(l)x_{ax,co}(l)}{\sigma_{co}^2} \right), \quad (3.6)$$

respectively. If the wrong path $\boldsymbol{a}_X$ is chosen by the MLSE, the relation between the two metrics in Eq.3.6 becomes

$$J_{a0}(L') < J_{aX}(L'). \tag{3.7}$$

In this case, according to the inequality in Eq.3.7, the wrong path becomes the survivor path. By plugging the two metrics in Eq.3.6 into Eq.3.7, we can obtain

$$\sum_{d=1}^{d_{pu}} \frac{r_{pu} y_{pu}(d)}{\sigma_{pu}^2} + \sum_{d=1}^{d_{co}} \frac{r_{co} y_{co}(d)}{\sigma_{co}^2} > 0, \tag{3.8}$$

where d is index of different codewords points between the correct path and the wrong path. The terms dpu and dco are Hamming distances between codewords of $\boldsymbol{a}_0$ and $\boldsymbol{a}_X$ for the NPMCS and the commercial system, respectively. The same codewords between the correct path and the wrong path are eliminated in Eq.3.8. We can understand that the probability of decoding error is affected by the Hamming distances and the level of the noise power.

An error probability is derived from Eq.3.8. First, signal power to noise power ratio (SNR) is derived by isolating the signal power and the noise power from the received signals y_{pu} and y_{co} of Eq.3.9.

$$SNR_{path} = \frac{\left(d_{pu}\frac{r_{pu}^2}{\sigma_{pu}^2} + d_{co}\frac{r_{co}^2}{\sigma_{co}^2}\right)^2}{d_{pu}\left(\frac{\sqrt{E[r_{pu}^2\omega_{pu}^2]}}{\sigma_{pu}^2}\right)^2 + d_{co}\left(\frac{\sqrt{E[r_{co}^2\omega_{co}^2]}}{\sigma_{co}^2}\right)^2} = d_{pu}\frac{r_{pu}^2}{\sigma_{pu}^2} + d_{co}\frac{r_{co}^2}{\sigma_{co}^2}. \tag{3.9}$$

As shown in Eq. 3.9, the signal power (P_s) of SNR is derived by the square of the sum of signals amplitude divided by the noise power because all the signals amplitude is in-phase. Meanwhile, the noise power (P_n) of SNR is derived by the sum of the squares of the noise signals amplitude divided by the noise power because of random phase. As a result of calculation, the SNR becomes sum of two Hamming distances divided by the noise power, d_{pu}/σ_{pu} and d_{co}/σ_{co}. This is similar with the maximum ratio combining (MRC) diversity methods, which is summation of the receive signals multiplied by reciprocal of the noise power if each signal power is constant. In fact, Eq. 3.9 can also be regarded as the MRC because Eq. 3.9 is derived as summation of ratio between the signal power 1 and the noise power d_{pu} and d_{co}. By using the SNR, error probability $P(\boldsymbol{a}_X|\boldsymbol{a}_0)$ is obtained by

$$P(\boldsymbol{a}_X|\boldsymbol{a}_0) = \mathrm{erfc}\left(\sqrt{d_{pu}\frac{r_{pu}^2}{\sigma_{pu}^2} + d_{co}\frac{r_{co}^2}{\sigma_{co}^2}}\right), \tag{3.10}$$

where erfc() is complementary error function.

3.3.3.3 Distance Spectrum

We derive distance spectrum so as to calculate the theoretical BER of the proposed system. After general description of the distance spectrum, we calculate the distance spectrum for the proposed system. Finally, we derive the theoretical BER on the basis of the distance spectrum.

3.3.3.3.1 General Description of Distance Spectrum

As described earlier, the capability of error correcting codes depends on Hamming distances. However, in the case of convolutional codes, Hamming distances can result in various values because the wrong paths can be multiple as shown in Fig. 3.7. For this reason, the capability of the convolutional codes needs to be evaluated using a number of Hamming distances. Herein, we define distance spectrums d_{free}, $af(d_{free}+i)$ and $cf(d_{free}+i)$ for $i = 0, 1, 2, \ldots$. More specifically, d_free denotes free distance which also can be interpreted as the minimum number of the Hamming distance, $af(d_{free}+i)$ denotes the number of the wrong paths for each Hamming distance $d_{free}+i$, and $cf(d_{free}+i)$ denotes the total number of errors on information bit in the wrong paths enumerated in $af(d_{free}+i)$ [36]. Although the distance spectrum can be calculated by the expanding expressions [37], the amount of the calculation increases exponentially with constrains length K. In such cases, the computers utilizing method that tracks the path transitions have been employed [36] [38].

3.3.3.3.2 Distance Spectrum of the Proposed System

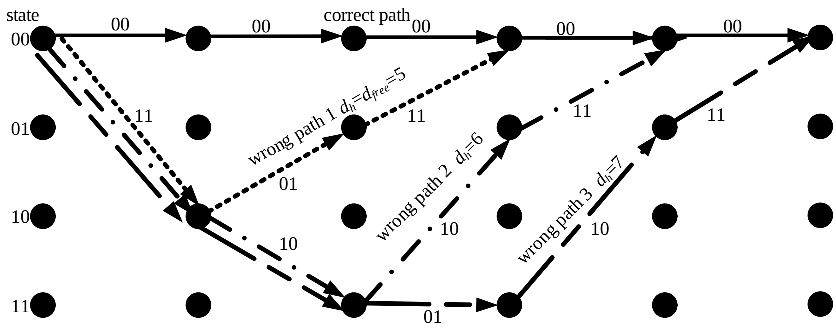

Figure 3.7: Wrong paths of convolutional code.

In this sub-section, We calculate the distance spectrum for the proposed system. First, the Hamming distances of the convolutional codes for the NPMCS and the CMWCS are defined as d_{pu} and d_{co}, respectively. The two Hamming distances have to be computed discretely because they are divided by the different noise power as shown Eq. 3.9. Herein, we consider the two free distances calculated from the codewords of the NPMCS and the CMWCS, which are denoted as $d_{pu,free}$ and $d_{co,free}$, respectively. From the free distances, we can determine $af_{pu,co} \equiv af(d_{pu,free}+i, d_{co,free}+j)$ and $cf_{pu,co} \equiv cf(d_{pu,free}+i, d_{co,free}+j)$ for $i = 0, 1, 2, ...$and $j = 0, 1, 2, ...$ which are the number of the wrong paths for each Hamming distance and the total number of errors on the information bits in the wrong paths enumerated in $af_{pu,co}$, respectively. In this study, I customize the methods proposed in [36] and [38] so that $af_{pu,co}$ and $cf_{pu,co}$ can be calculated. That is, by calculating all the patterns, i.e., patterns $d_{pu,free}+i, d_{co,free}+j, \forall i$and j, the customize method can be performed.

Note that if we employ the removal shown in Fig. 3.2, the distance spectrum varies with the relations between the point that bit error occurs and the point that cordwards are removed. In this case, $af_{pu,co}$ and $c_{fpu,co}$ have to be calculated for all the relations. In this paper, we denote the number of the relations as P. Accordingly, the average total number of the errors on the information bits can be obtained $cf_{pu,co}/P$.

Assuming that the hard decision values of the SI are correct, the distance spectrum can be easily calculated. For example, we can assume a trellis diagram shown in left side of Fig. 3.8. If a third codeword is determined as -1 with the help of the subsidiary information, the only paths whose third codeword is -1 are allowed. It is conflicting if the paths with a third codeword of +1 are allowed as survivor paths. This is because the codeword produced by the above path transitions denies correctness of the subsidiary information. For this reason, path transitions can be limited by the SI of the correct hard decision value (CHD). The right side of Fig. 3.7 shows a case that the path whose third codeword is +1 were limited and eliminated because the third codeword given by the SI was -1.

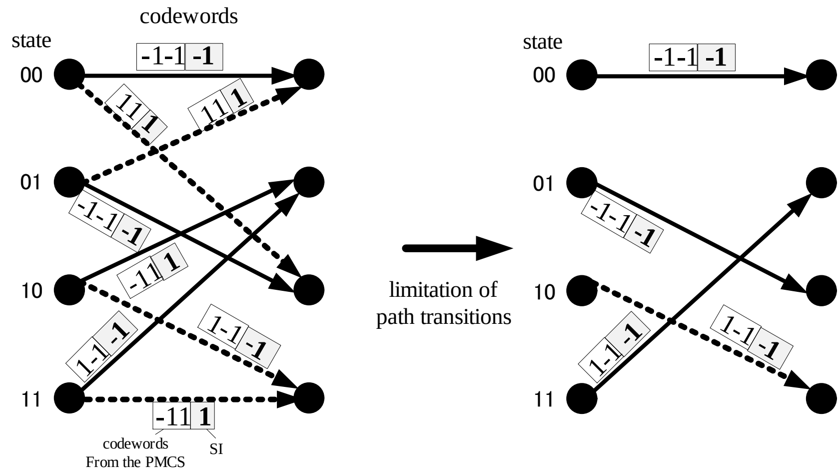

Figure 3.8: The limitation of the path transitions for correct hard decision value (CHD) of the subsidiary information.

In the case of the CHD, calculating the distance spectrum is not difficult, i.e., just by adding the specific path transitioning limitation to the original methods [36][38]. Furthermore, $af_{pu,co}$ and $cf_{pu,co}$ can be simplified. If using the SD values, we need to determine $af_{pu,co} \equiv af(d_{pu,free}+i, d_{co,free}+j)$ and $cf_{pu,co} \equiv cf(d_{pu,free}+i, d_{co,free}+j)$ for all the combinations of i and j. However, in the CHD, $af_{pu,co}$ and $cf_{pu,co}$ become single index, $af(d_{pu,free}+i)$ and $cf(d_{pu,free}+i)$; i is sufficient for calculating the distance spectrum.

3.3.3.3.3 Example of Distance Spectrum Calculation

Table 3.1 shows the parameters of the convolutional code of RCR STD-39[1] that is one of the NPMCS. In this paper, we employ the addition method as shown in Fig. 3.5 and assume three patterns shown in Fig. 3.9. Table 3.2 and

Table 3.1: The convolutional codes of RCR STD-39[1].

constrain length K	6
coding rate R_{pu}	1/2
generator sequency (octal natation)	g(1) =53 g(2)=75

3.3 show the calculated the distance spectrum by the SD and CHD, respectively. As shown Table 3.2 and Table 3.3, the free distances $d_{pu,free}$ and $d_{co,free}$ increase with the amount of the subsidiary information. The free distance of the CHD is larger than that of the SD for the same pattern shown in Fig. 3.9.

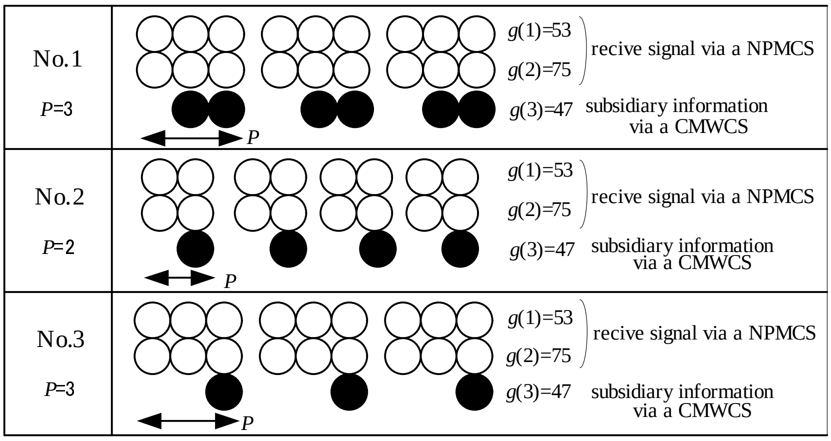

Figure 3.9: Three example of adding subsidiary information (SI).

3.3.3.3.4 Derivation of Theoretical BER

The theoretical BER can be calculated on the basis of the distance spectrum and Eq.3.10. Since Eq.3.10 represents the probability of a wrong path occurrence, we can achieve an upper bound of the theoretical BER by accumulating the values obtained by multiplying the probability Eq..3.11 by $cf(d_{pu,free}+i, d_{co,free}+j)/P$ for each combination of i and j.. Hence, the upper bound of the theoretical BER of the proposed system is determined by

$$P_b < \frac{1}{P} \sum_{d_{pu}=d_{pu,free}}^{\infty} \sum_{d_{co}=d_{co,free}}^{\infty} N(d_{pu}, d_{co}) \mathrm{erfc}\left(\sqrt{d_{pu}\frac{r_{pu}^2}{\sigma_{pu}^2} + d_{co}\frac{r_{co}^2}{\sigma_{co}^2}}\right). \qquad (3.11)$$

Table 3.2: The distance spectrum of the SD subsidiary information(SI).

	d_{pu}	8		9							10							
	d_{co}	2	3	1	2	3	4	5	6	7	0	1	2	3	4	5	6	7
No.1 $g_{max}(3)=41$	$L(d_{pu}, d_{co})$	1	2	2	1	1	6	8	5	1			1	3	3	7	3	4
	$N(d_{pu}, d_{co})$	2	4	2	1	3	22	46	29	5			2	8	14	30	16	26
No.2 $g_{max}(3)=57$	$L(d_{pu}, d_{co})$	2			2	5	4	3	2			2	1	8	1	2		
	$N(d_{pu}, d_{co})$	4			4	19	20	17	12			8	4	34	4	14		
No.3 $g_{max}(3)=61$	$L(d_{pu}, d_{co})$	3		9		13		2			1		16		4			
	$N(d_{pu}, d_{co})$	6		27		69		12			2		76		18			

Table 3.3: The distance spectrum of the CHD subsidiary information(SI).

d_{pu} or d_{ch}		8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24	25
conven-tional	$L(d_{pu})$	1	8	7	12	48	95	281	605										
	$N(d_{pu})$	2	36	32	62	332	701	2342	5503										
No.1 $g_{max}(3)=41$	L_{ch}												1	0	0	0	0	1	1
	N_{ch}												3	0	0	0	0	6	9
No.2 $g_{max}(3)=57$	L_{ch}								1	2	0	0	4	5	5	8	7	9	23
	N_{ch}								5	12	0	0	30	38	49	80	65	104	263
No.3 $g_{max}(3)=61$	L_{ch}			1	0	4	0	15	0	31	0	84	0	269	0	714			
	N_{ch}			2	0	22	0	82	0	256	0	758	0	2806	0	8672			

There are two reasons why Eq. 3.11 is an upper bound on BER. First, as the wrong paths shown in Fig. 3.7 are not independent; a long wrong path may include other short wrong paths. Second, even if the practical convolutional codes are terminated, Eq.3.11 assumes that the length of the convolutional code is infinite[39].

3.3.4 Computer Simulation and its Evaluation

We evaluate the BER by computer simulations so that we can compare it with the BER calculated using the distance spectrum. Moreover, we analyze how much the proposed system can improve the BER performance. The simulation parameter in this paper is shown in Table 3.4.

Figure 3.10 shows the calculated and the simulated BER when using the first adding pattern in Fig. 3.9 (denoted as No.1). The term CP represents the SNR improvement by using the commercial system over the NPMCS when we employ the SD. Figure 3.10 shows that the calculated BER approximates the simulated BER especially for large SNR range of the NPMCS. In small SNR range, the rea-

Table 3.4: The simulation parameter of the computer simulation.

constrain length K	6
coding rate R_{pu}	1/2
generator sequency (octal natation)	g(1)=53 (RCR STD-39) g(2)=75 (RCR STD-39) g(3)=47 (CWMCS)
adding pattern	No. 1,2,and 3(shownin Fig. 3.9)
length of information bits	120 bits
length of codewords	240
tail bit	5bit
environment	AWGN

son why BER performance differs between the simulation and the theory derived from distance spectrum was shown in sub-section 3.3.3.3.4. The long wrong paths which include short wrong paths increase with decreasing SNR of the NPMCS.

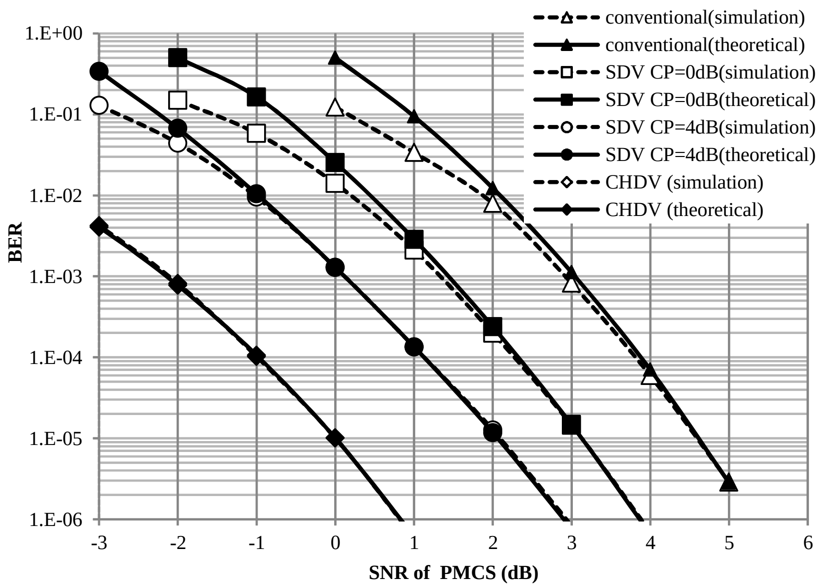

Figure 3.10: Comparison of BER calculated by distance spectrum and computed by computer simulation (No.1).

In addition, we evaluate improvement in BER performance of the proposed system. Figure 3.11 shows the BER curves of the proposed system using the three adding patterns in Fig. 3.9 and the CHD of the subsidiary information. In the figure, we can verify that the proposed system improves the SNR performance by 4dB and higher over the conventional NPMCS when employing the first adding pattern in Fig. 3.9.

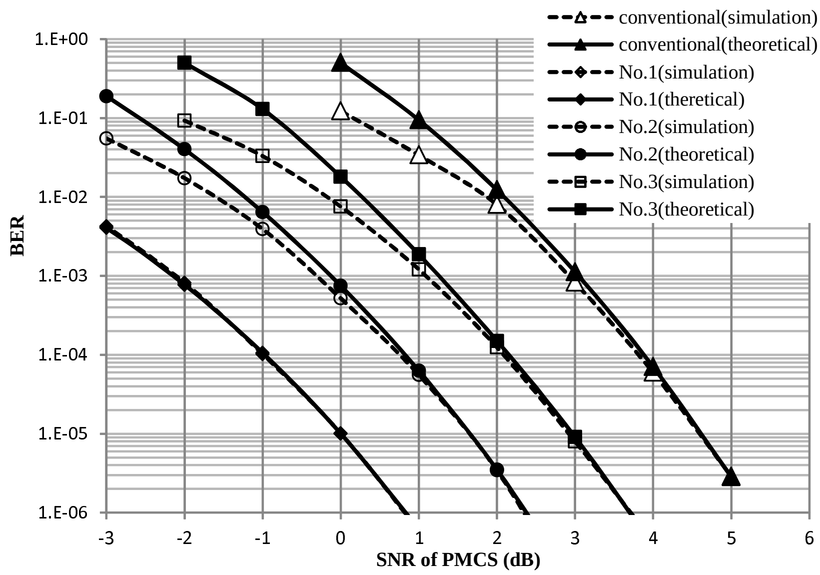

Figure 3.11: BER performance improvement for the proposed system.

3.4 Chapter Summary

In this Chapter, We proposed the BER improvement method for the HCR combining NPMCSs with CMWCSs for expanding the service coverage of the NPMCSs. The proposed method is designed to improve the decoding performance of the convolutional codes by utilizing the SI which can be transmitted by the commercial systems. I presented a modified soft decision Viterbi algorithm, the distance spectrum and BER performance of the proposed system. Employing the proposed method, the HCR can improve BER performances freely by changing SI's quantity depending on the communication quality.

Since synchronization between the two subsystems, i.e., the NPMCS and the commercial system, is required to realize this method, synchronization methods is considered in the next chapter.

Chapter 4

Self-Synchronization method for Narrowband PMCSs

To realize the HCR combining a NPMCS with a CMWCS, the NPMCS and CMWCS must be synchronized; that is, the HCR must construct the IRC comprising the received signals from the NPMCS and the SI from the CMWCS. In the HCR, the synchronization is a two-step process. First, the NPMCS and CMWCS are self-synchronized. Second, the HCR co-synchronizes the NPMCS and the CMWCS. In cognitive networks, although several co-synchronization methods such as the timing-sync protocol for sensor network (TPSN) [40] and the flooding time synchronization protocol (FTSP) [41] have been proposed, self-synchronization methods have not been researched.

Thus, I mainly focus on the self-synchronization method of the NPMCS. I proposed two self-synchronization methods that are the method employing the GPS(Global Positioning System) signals and the method that can be used without GPS signals.

4.1 Difficulty of Synchronization

First, I mention a difficulty of synchronization. Because the proposed HCR aims to improve the communication quality; the likelihood that the HCR will operate in severe low SNR environments is extremely high. In such environments, NPMCS's timing synchronization included in the HCR may not be achievable using conventional synchronization techniques such as Maximum Amplitude Method (MAM)

[42] and correlation of synchronous words (SW) [43]. Specialized synchronization methods are therefore required.

4.2 Method Employing GPS Signals

In this sub-section, we propose a NPMCS's timing synchronization method derived from the Global Positioning System (GPS). The HCR can acquire the timing synchronization by using time information of the GPS and the SI having the information of transiting time of the NPMCS.

In sub-section 4.2.1, detail of the proposed GPS synchronization method are explained. Then, problems of the proposed method are mentioned and their solutions are proposed in sub-section 4.2.2. In sub-section 4.2.3, BER performance degradation caused by the problem is evaluated by a computation simulation.

4.2.1 Detail of the Synchronization Method Employing GPS

Figure 4.1 shows the HCR system employing the proposed synchronization method. For the timing synchronization, frame transmitting timing information is sent from the base station of NPMCS to HCR by using the SI. The frame transmitting timing information included in the SI is used for both self-synchronization of NPMCS and co-synchronization between NPMCS and CMWCS. The SI consists of the transmitting timing information, association information between the SI and NPMCS's signals and information for BER improvement as shown in Fig. 4.1. The transmitting timing information may be replaced with time-stamp. The association information between the SI and NPMCS's signals is used to construct the IRC.

In the HCR, their location and precise time are measured from GPS signals. By comparing the frame transmitting timing information with the time measured by GPS, the HCR can obtain the self-synchronization as well as the co-synchronization between the NPMCS and CMWCS. GPS signals are used for acquisition of location of HCR and time synchronization. Specially, the HCR synchronizes with UTC (Coordinated Universal Time) of GPS and then synchronizes with time of base station of the NPMCS. After this, the HCR measures own location r and divides distance between r and location of NPMCS's base station r_0 by light speed c to calculate propagation delay τ. By adding τ to transmitting time t_T of the NPMCS's base station, self-synchronization and co-synchronization can be recovered.

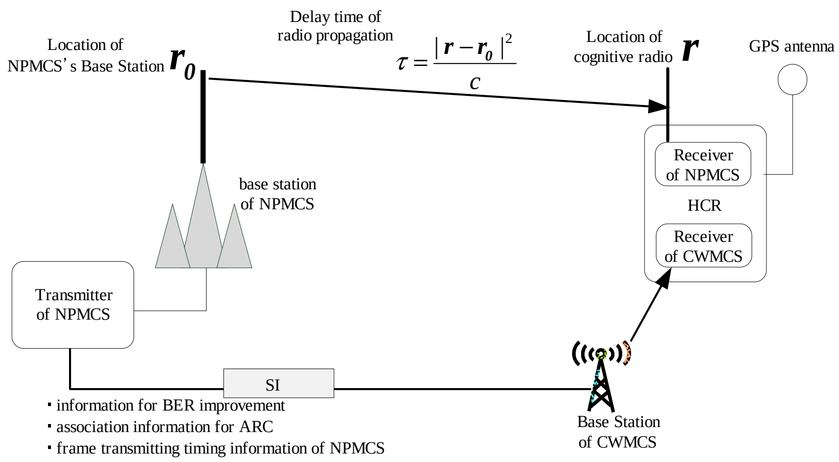

Figure 4.1: HCR employing the proposed GPS synchronization method.

4.2.2 Problems

If accident error of time synchronization calculated from GPS, propagation delay caused by diffraction and delay waves caused by multipath fading occur, calculated synchronization timing is shifted from optimal timing. In this sub-section, detail of three problems are explained.

4.2.2.1 Accuracy of GPS

When GPS are used for timing synchronization, we have to consider whether timing synchronization error derived from GPS is much smaller than symbol duration T. Table 4.1 shows parameters of RCR STD-39 [1] as one candidate of NPMCS. The symbol length of the NPMCS is 62.5μs. On the other hand, There is $\pm 1\mu$s. time error in most of generalized GPS receiver. If dedicated devices of the time synchronization are employed, accuracy for tens ns can be obtained. Sub-section 4.2.4 examines whether error of $\pm$ 1 μs influences BER performance.

Table 4.1: Parameter of RCR STD-39 [1]

Constraint length K	6
Coding rate R_{pu}	1/2
Generator sequency (octal natation)	g(1)=53 (RCR STD-39) g(2)=75 (RCR STD-39) g(3)=47 (CWMCS)
Adding pattern	No. 1,2,and 3(shownin Fig. 3.9)
Length of information bits	120 bits
Length of codewords	240
Tail bit	5bit
Environment	AWGN

4.2.2.2 Delay Caused by Diffraction

In the proposed method, propagation delay is calculated from direct distance between the base station and the HCR as shown Fig. 4.1.However, when diffraction occurs in radio propagation environment as shown Fig. 4.2, actual propagation delay is longer than that calculated from direct distance because of increase of radio propagation. We refer to this increase as $\Delta\ \tau_d$ that caused error of timing synchronization.

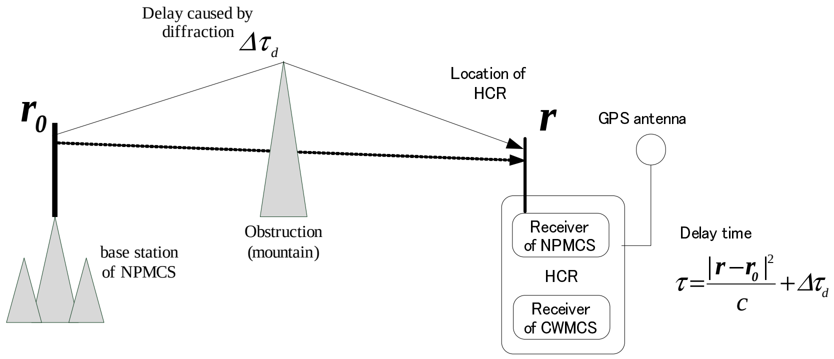

Figure 4.2: Delay caused by diffraction $\Delta\ \tau_d$.

In Fig 4.3, amount of propagation delay is roughly estimated. In this case, direct distance between the base station of the NPMCS and the HCR is 10050m and propagation distance of the diffraction wave becomes 10483m. Thus, difference between the direct distance and the distance of the diffraction wave is estimated as 434m. When this difference distance is divided by light speed c, synchronization timing is shifted for 1.45μs. We therefore must anticipate several us delay to consider worse conditions.

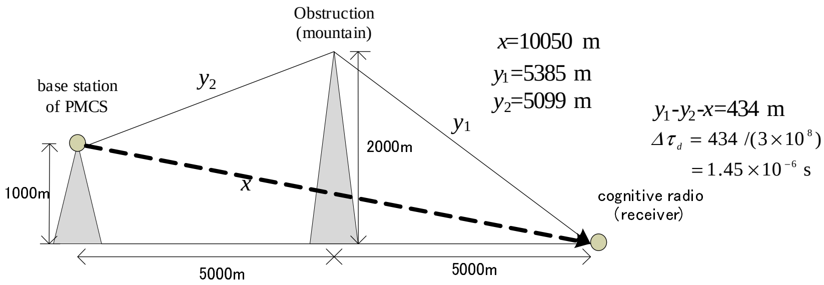

Figure 4.3: Example of diffraction delay.

4.2.2.3 Delay Profile

In actual environments, since many delay waves caused by multipath fading occur, synchronization timing calculated from GPS might not be optimal timing. When we assume that time acquired from GPS is accurate and $\Delta\ \tau_d$ =0, the synchronization timing point calculated from GPS becomes point when preceding wave arrives as shown in Fig 4.4. However, optimal synchronization timing in Fig 4.4 is shifted by $\Delta\ \tau_p$ from the timing calculated from GPS. In this reason, the synchronization timing calculated from GPS probably makes BER performance deteriorate. Optimal synchronization timing in consideration of delay waves and its performance are shown in [44] [45].

4.2.3 Solution of the Problems

In this sub-section, I propose a mitigation method of the timing shift to reduce BER performance deterioration. To mitigate shift of synchronization timing, I proposed that the synchronization timing calculated from GPS, $\tau + t_T$ is added to τ_G as compensated synchronization timing $\tau + t_T + \tau_G$. By adding τ_G to synchronization timing, Delay of $\Delta\ \tau_d$ and $\Delta\ \tau_p$, would be compensated.

However, when $\Delta\ \tau_d = \Delta\ \tau_p$ =0, the synchronization timing is ahead of d optimal synchronization timing. Thus, we propose that firstly $\tau + t_T + \tau_G$ is defined as temporary synchronization timing and then existing synchronization

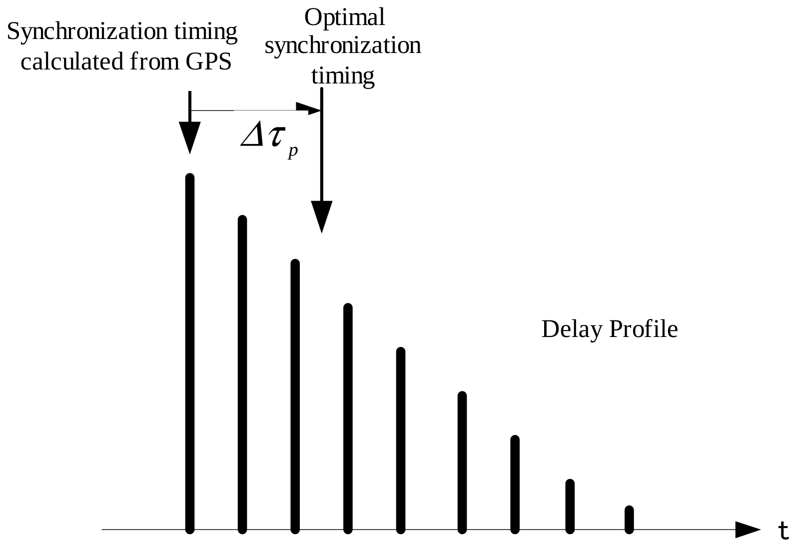

Figure 4.4: Delay profile and synchronization timing.

methods such as MAM(Maximum Amplitude Method), WDM (Waveform Differential Method) and ZCM(Zero Crossing Method)[46] are employed to search an optimal synchronization point. A rough synchronization point is defined by GPS, then the optimal point is searched around the rough synchronization point by the existing methods. Here, MAM is a synchronization method that searches the point where mean-square value outputs of matched filter becomes the largest. WDM is a synchronization method that searches the point where derivative values of outputs of a matched filter become zero. ZCM is a synchronization method that searches sign inversion point of matched filter outputs[46].

4.2.4 Computer Simulation

In this sub-section, the grade of BER deterioration caused by remain synchronization timing shift is calculated by computer simulation. Then, the proposed synchronization method is evaluated by BER performance. Firstly, we explain expected delay profile of NPMCSs. Then we estimate BER deterioration caused by synchronization timing shift. Moreover, we evaluate the proposed BER improvement method combining SI with receive signal of a NPMCS to decode the IRCs shown in sub-section 3.2 when the proposed timing synchronization method is employed. Finally, we consider influence of Doppler frequency in fading environments.

4.2.4.1 Delay Profile

PMCSs employ VHF or low UHF bands (from 100MHz to 400MHz) as carrier frequency so that large-zone systems are realized. In the simulation, we employ two types of delay profile models. One is the radio propagation characteristic that was measured by NICT (National Institute of Information and Communications Technology) in Japan in 2011 [47] as delay profile model for simulation. The other is delay profile models that IEEE802.22 are defined [52].

Although delay profile of VHF band has hardly measured, NICT measured the delay profile of 190MHz in Numazu city to survey radio propagation characteristic of Public Broadband Wireless Communication System (PBWCS). We produce two type of delay profile from the NICT measurement. One is named regular model produced from normal propagation characteristic in NICT measurement. The other is named severe model produced from propagation characteristic that has long delay wave.

The delay profile models that are employed in the simulation are shown in Fig.4.5, Fig.4.6, Fig.4.8, and Fig. 4.9. Moreover, for comparison, I also prepare 1 path Rayleigh fading model and 2 path Rayleigh fading model shown in Fig. 4.7.

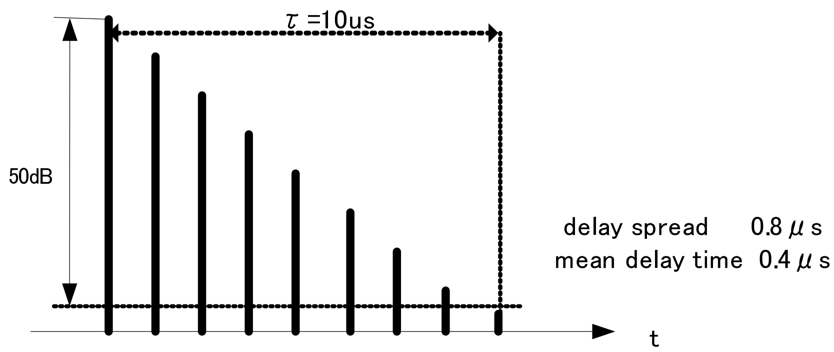

Figure 4.5: Delay Profile 1 (normal model : Exponential decay 9 path Rayleigh fading model) .

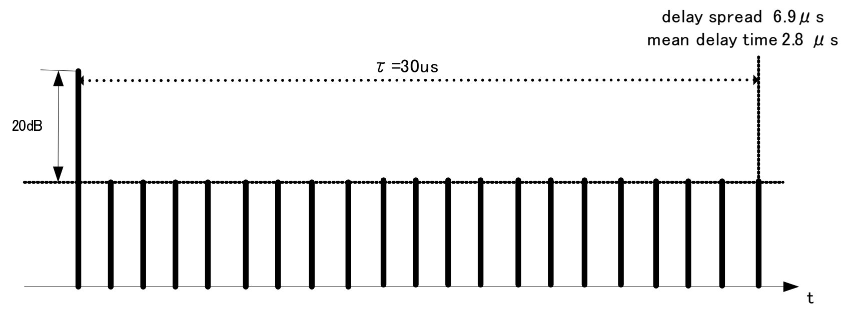

Figure 4.6: Delay Profile 2 (severe model : 24 path Rayleigh fading model) .

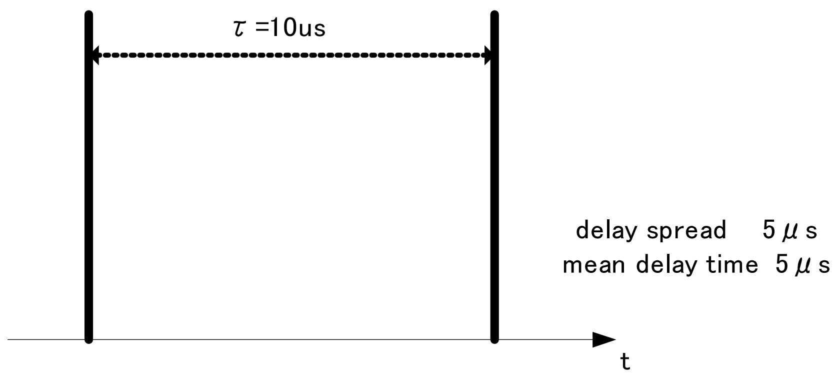

Figure 4.7: Delay Profile 3 (2 path model:equal level 2 path Rayleigh fading model) .

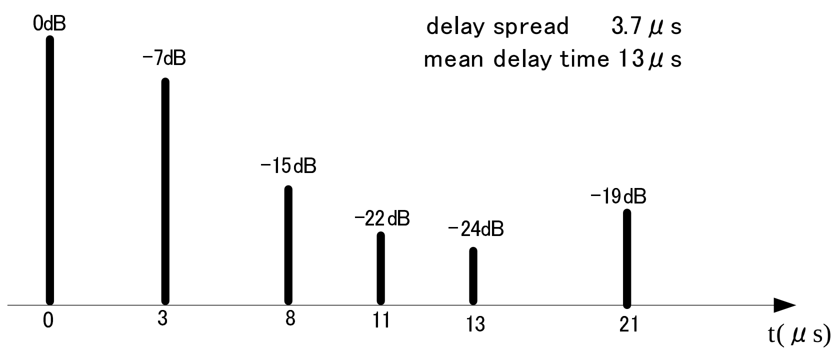

Figure 4.8: Delay Profile 5 (IEEE802.22 WRAN channel model Profile A) .

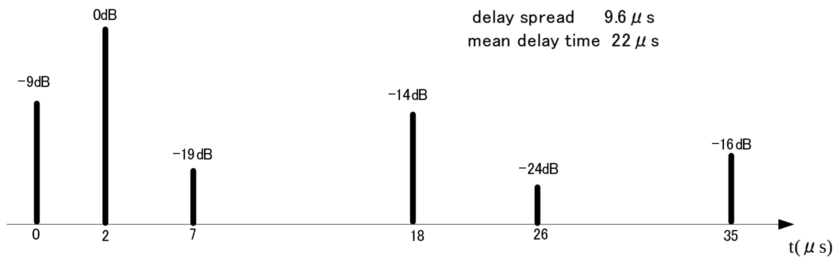

Figure 4.9: Delay Profile5（IEEE802.22 WRAN channel model Profile C）.

4.2.4.2 BER Deterioration Caused by Synchronization Timing Shift

Figures. 4.10 - 4.14 show simulation results of BER deterioration caused by synchronization timing shift. NPMCS signals are decoded by the deferential detection, when Eb/No is 25dB, normalized maximum Doppler frequency is $f_dT = 8.75 \times 10^{-5}$, and FEC (forward error correction) is not used. Synchronization timing points are search in 160 symbols that is equivalent to one slot duration (10ms). In the lateral axis of the graph, 0μs shows timings of preceding wave arrival and synchronization point are the same. In the simulation, we prepare other delay profile models that are expanded toward time domain by using original delay profile models to evaluate deterioration caused by synchronization timing error. By the expansion, mean delay time (MDT) becomes large.

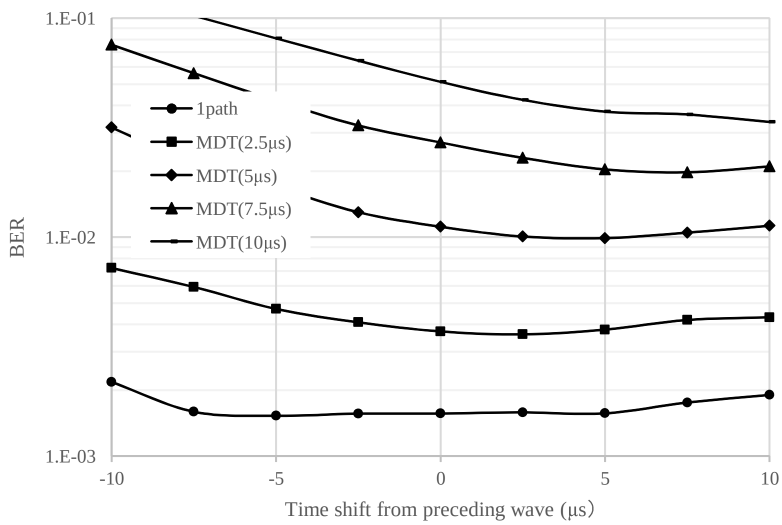

Figure 4.10: BER deterioration caused by synchronization timing shift(1path and 2 path model).

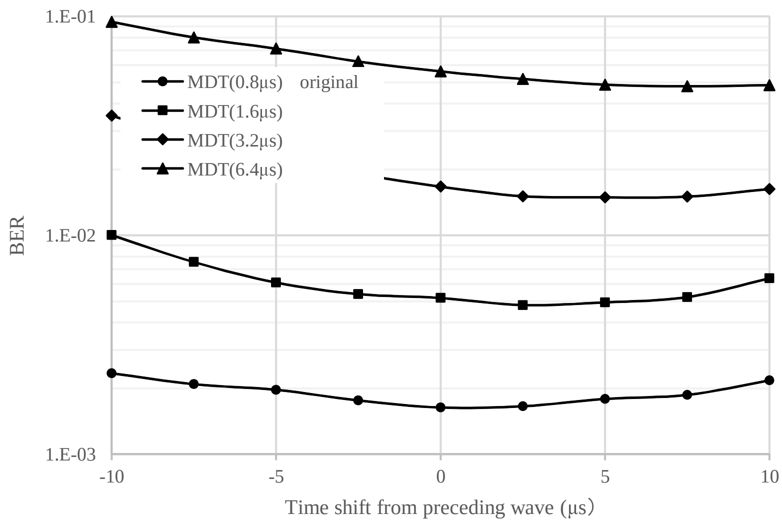

Figure 4.11: BER deterioration caused by synchronization timing shift(normal model:Exponential decay 9 path Rayleigh fading model).

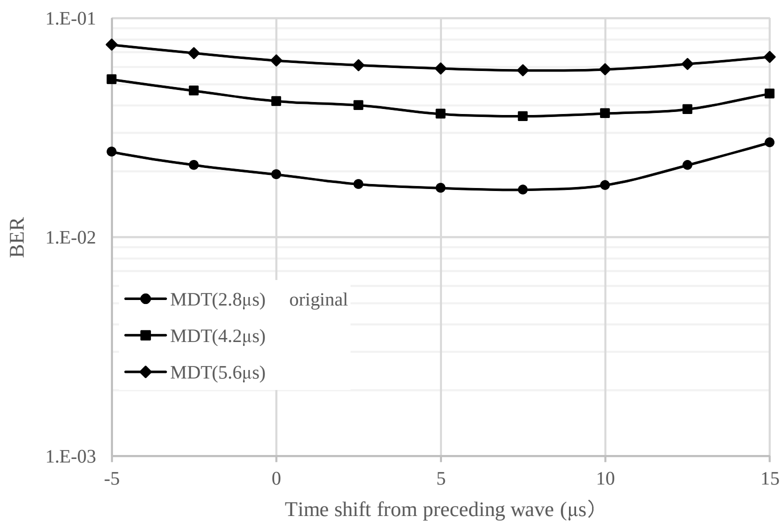

Figure 4.12: BER deterioration caused by synchronization timing shift(severe model:24 path Rayleigh fading model).

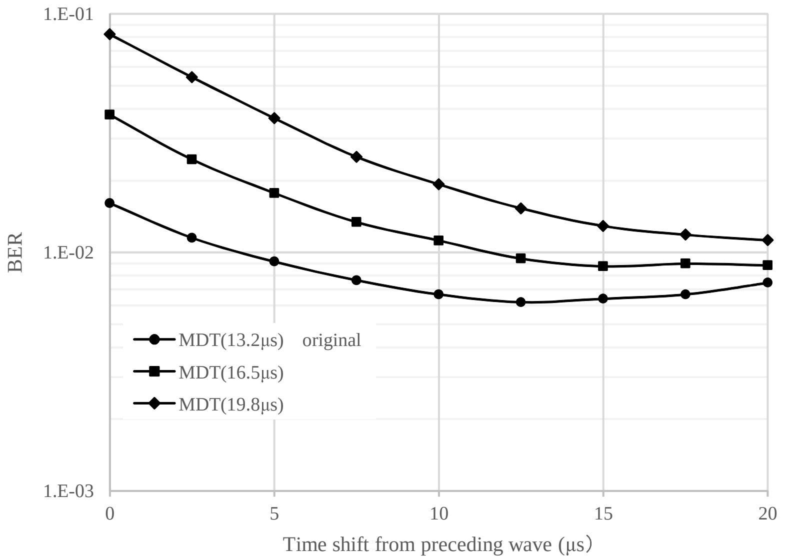

Figure 4.13: BER deterioration caused by synchronization timing shift(IEEE802.22 Profile A).

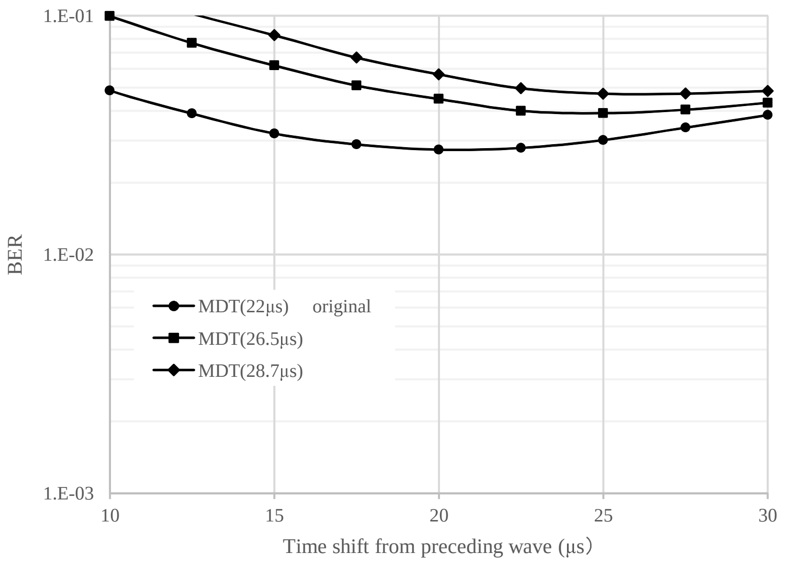

Figure 4.14: BER deterioration caused by synchronization timing shift(IEEE802.22 Profile C).

When evaluating 1path model in Fig. 4.15, BER is hardly deteriorated within $\pm 5\mu$s (5% of symbol duration). In the other models, we find that the optimal synchronization points are MDT [44] [45] except for the 24 path Rayleigh fading model. Moreover, we confirm that BER is on increase in proportion to delay spread [42]. In the case of preceding wave arrival timing synchronization, while the Numazu models do not make BER increase significantly, IEEE802.22 models force systems to deteriorate BER performances. Hence, the synchronization utilizing the GPS requires methods to adjust synchronization timing. One method is adding delay to synchronization timing by forecasting delay waves from ML position obtained from the GPS. We also consider that delay profile can be obtained by employing data base.

4.2.4.3 Comparison of Synchronization Methods

Figures 4.15 - 4.17 show BER performance when MAM, WCM and ZCM to search the optimal synchronization timing after calculating rough synchronization timing by GPS signals. In these figures, GPS 0μs shows the synchronization point is the arrival point of the preceding wave of the delay profile, GPS -5μs shows synchronization point is 5μs earlier than the arrival point of the preceding wave. The other simulation parameters are the same as sub-section 4.2.4.2.

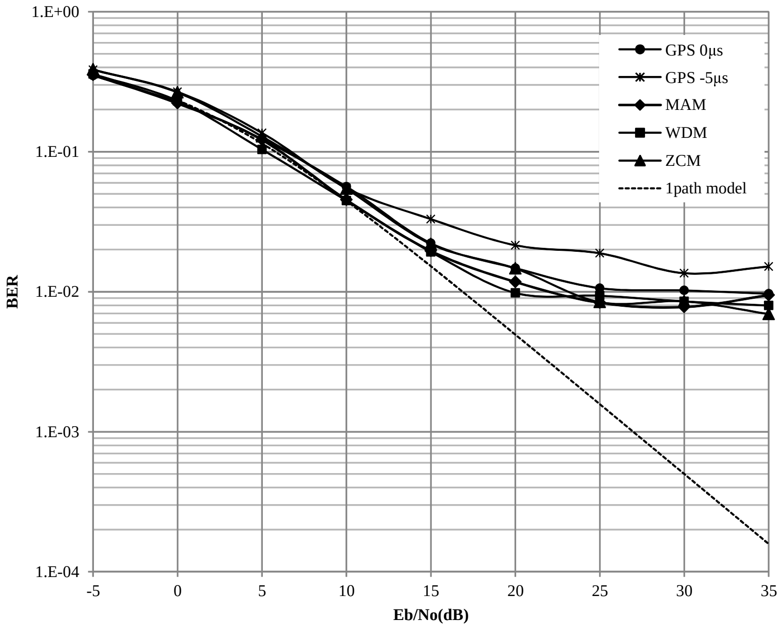

Figure 4.15: Comparison of Synchronization methods（2 path model:equal level 2 path Rayleigh fading model）.

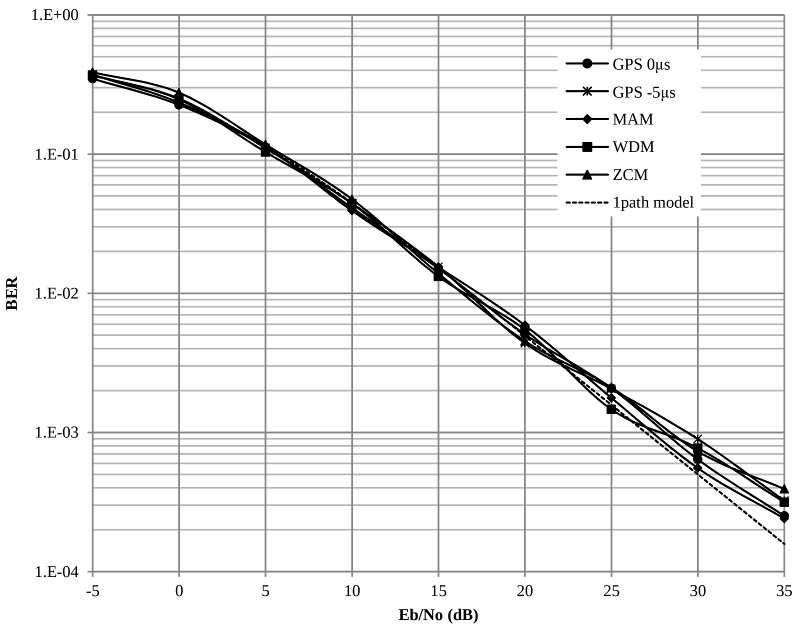

Figure 4.16: Comparison of Synchronization methods (normal model : Exponential decay 9 path Rayleigh fading model）.

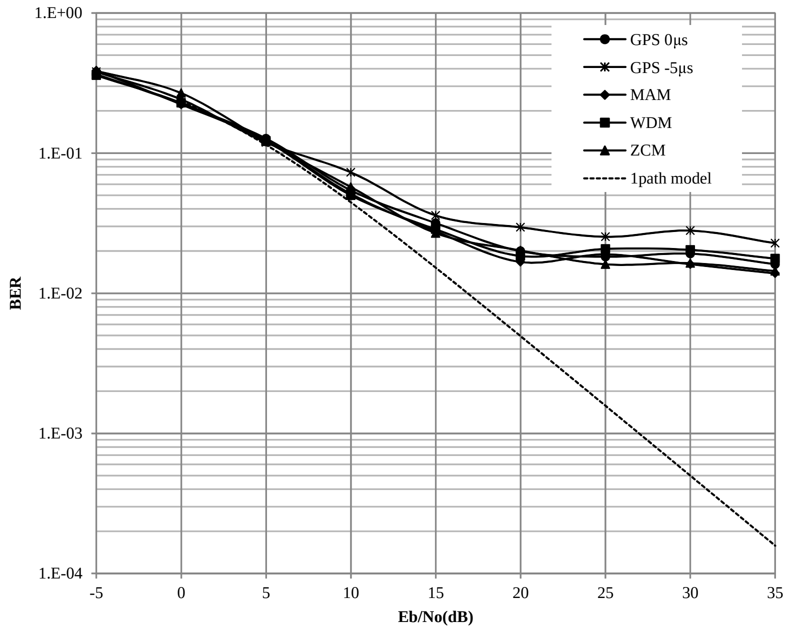

Figure 4.17: Comparison of Synchronization methods (severe model : 24 path Rayleigh fading model) .

By employing MAM, WDM and ZCM, BER performance further improved than that of employing GPS signal alone as shown in Figs. 4.15 - 4.17. The MAM is the best of the three. However, in Figs. 4.15 - 4.17, these BER performance is simulated in $f_d T = 8.75 \times 10^{-5}$. If $f_d T$ increases, BER performance of MAM, WDM and ZCM is worse than that of GPS signal alone. This result is considered in sub-section 4.2.4.5.

In addition, BER performances show floor characteristics. We consider that this reason is not synchronization timing error but ISI (Inter Symbol interference).

4.2.4.4 BER Performance in IRC

BER performance of the proposed synchronization methods shows in Figs. 4.18 and 4.19 when IRCs explained in sub-section 3.2 are decoded. In this sub-section, IRCs consist of two SI bits per three information bits and the IRCs are decoded using MLSE (Maximum Likelihood Sequence Estimation). SI is produced from a generate polynomial that has g3=41 (octal) and removed at a rate of 1bit per 3bits. The other parameters are the same as sub-section 4.2.4.2.

From Fig. 4.18 and 4.19, we confirm that BER performance improvement of around 5dB is expected in the no floor areas. In the floor areas, BER can be reduced to 1/10. In sub-section 3.3.4, while I showed 5dB improvement in AWGN environments, the same results are calculated in the fading environment.

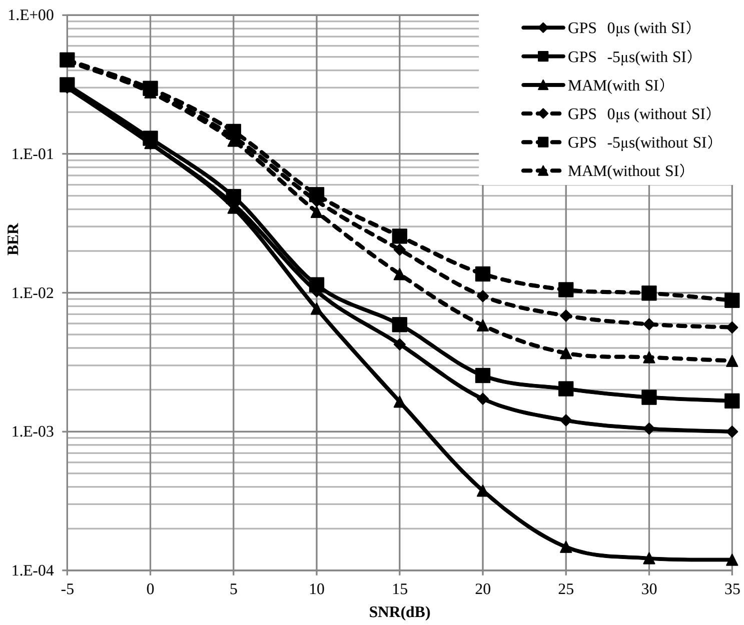

Figure 4.18: BER performance in IRC (2 path model:equal level 2 path Rayleigh fading model).

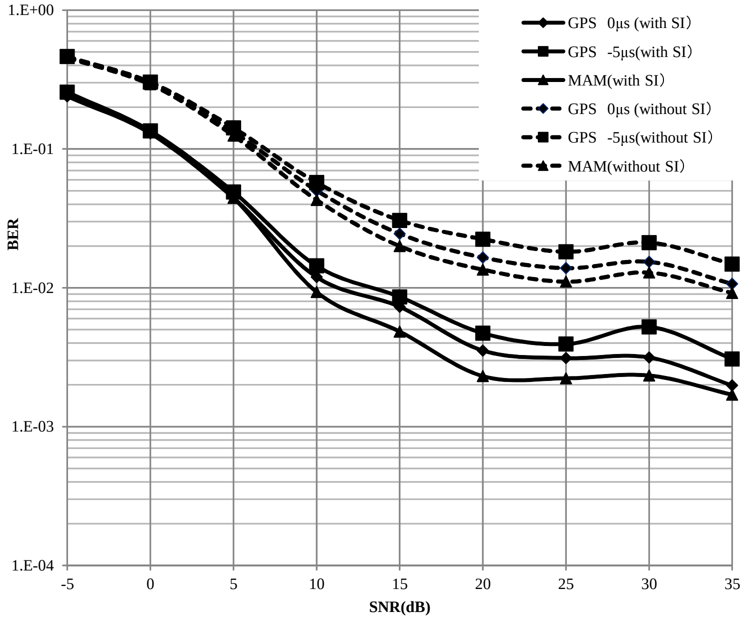

Figure 4.19: BER performance in IRC(severe model : 24 path Rayleigh fading model).

4.2.4.5 Influence of Maximum Doppler Frequency

Figures 4.20 and 4.21 show BER performance when Maximum Doppler Frequency is changed and IRCs are decoded.

From Figs. 4.20 and 4.21 , BER performances of the synchronization method using GPS signals alone are improved as f_dT increases. This is because time diversity and interleave effects increase as Doppler frequency rises.

When $f_dT < 0.008$, BER performances of the method using GPS signals alone are inferior to those of the method adding MAM. However, as f_dT increases, this relationship is reversed. This is because the synchronization point of MAM is defined as the point where Eye[48] opens the most averagely; in rapid propagation channel change environments, the instant most opened Eye are probably far from the average most opened Eye. In the case of high f_dT, the method using GPS signal alone may be superior to the method using MAM.

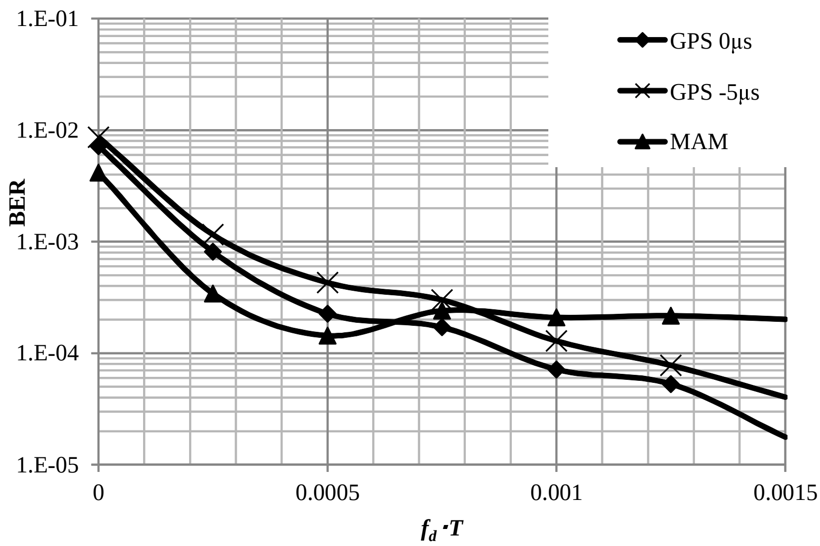

Figure 4.20: Influence of Maximum Doppler Frequency（2 path model:equal level 2 path Rayleigh fading model）.

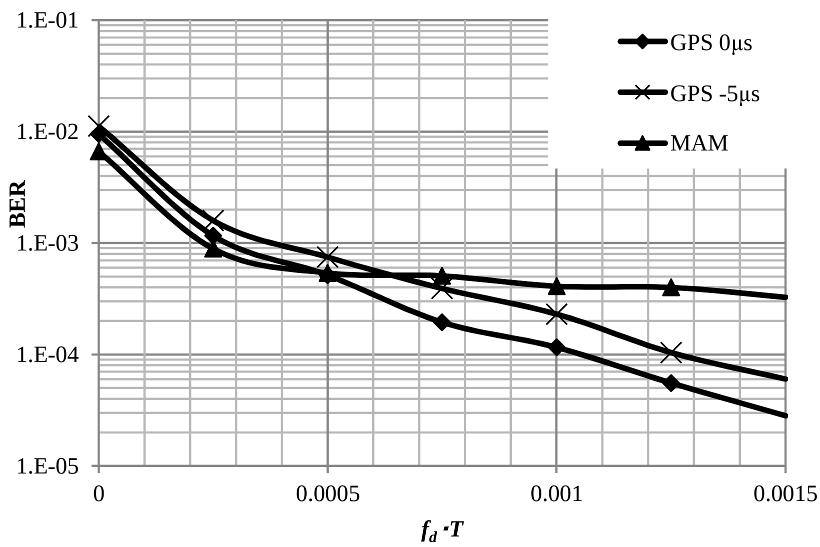

Figure 4.21: Influence of Maximum Doppler Frequency(severe model : 24 path Rayleigh fading model）.

4.3 Method Employing the SI

In this sub-section, We propose SI-based synchronization methods. Although we consider that the GPS method is useful for cognitive network, it has a problem; it cannot be used indoors. To solve this problem, we consider SI-based synchronization methods. Moreover, the incorporation of separate SI for synchronization and BER improvement is inefficient, since the CMWCS must then convey much extra information. We therefore devise a technique in which synchronization and BER improvement can be realized from the same SI. For simultaneous timing synchronization and BER improvement, the SI is transmitted as phase rotations of modulated signals. First, we show that a simple timing synchronization, in which the SI comprises the send signals of the NPMCS, is insufficient. Hence, we propose a novel synchronization technique employing the SI. The proposed technique can accomplish both of the self-synchronization of the NPMCS and the co-synchronization. In this sub-section, we focus on the self-synchronization of the NPMCS especially because we assume that HCR is in low SNR environments where acquisition of the self-synchronization of the NPMCS is difficult.

4.3.1 Simple Method and its Problems

This section explains the problems of simple technique. Figure. 4.22 illustrates the simple technique. The SI is defined as modulated signals, in-phase (I) and quadrature-phase (Q) signals, (Fig. 4.22, signals (A)) that are removed periodically to reduce resource consumption of the CWMCS (Fig. 4.22, modules (B)). In the HRC, the timing synchronization is recovered by correlating the filtered received signal of the NPMCS with the SI from the CWMCS (Fig. 4.22, modules (C)). It appears that the simple technique can easily acquire synchronization by correlating the received signals of the NPMCS with the SI. However, there are two problems in the simple technique. One is as follows. If carrier frequency offset occurs in the down-converter of the NPMCS receiver, the simple timing synchronization is insufficient because the offset gradually rotates the phases of received I and Q signals. These phase rotations reduce the correlation values between the SI and received I and Q signals. The other problem is BER improvement. If the sent data have been coded by a differential encoder (Fig. 4.22, modules (D)) and then I and Q signals are removed (Fig .4.22, modules (B)), the HCR receives insufficient information from the received SI for BER improvement. This is because the decoding of the SI, differential coding signals, requires two continuous (successive) symbols. Since the removal scheme changes the continuous symbols to discontinuous symbols (Fig. 4.22, modules (B)), some symbols cannot be decoded in the differential decoder Fig. 4.22, modules (E)). The removal deprives the continuous SI of the information required for BER improvement.

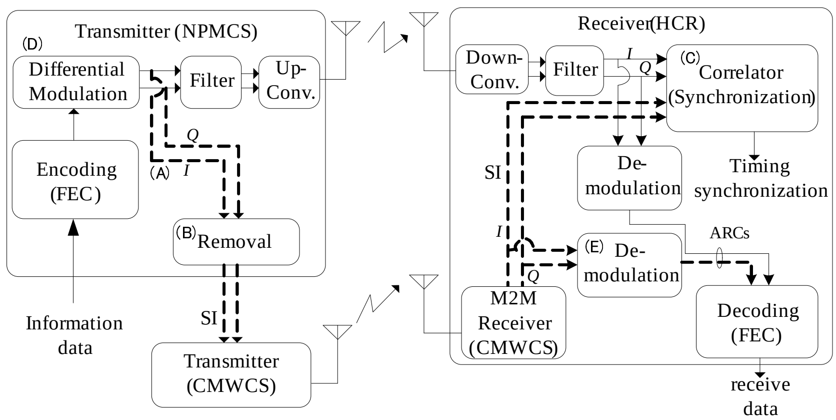

Figure 4.22: The HCR employing simple timing synchronization technique.

4.3.2 Method Employing Envelopes of $\pi/4$ shift QPSK

To prevent reduction of the correlation values and BER improvement information, we propose that the SI contains information of the phase rotations attributed to the differentially encoded $\pi/4$ shift QPSK.

Figure. 4.23 shows the modified HCR that can accomplish the self-synchronization of NPMCSs, the co-synchronization, and the BER improvement for differentially encoded π /4 shift QPSK systems that NPMCSs employs. To produce the SI and compress the sending information, the differentially modulated signals are converted into binary $\pm$ 1 values (see Fig. 4.23, module (A)). If the received phase rotation is $\pm\pi/4$, the output signal is +1. In the same way, a phase rotation of $\pm 3\pi/4$ is converted into -1. The converted SI is removed in module (B) of Fig. 4.23 and transmitted by the CMWCS. The timing synchronization including the

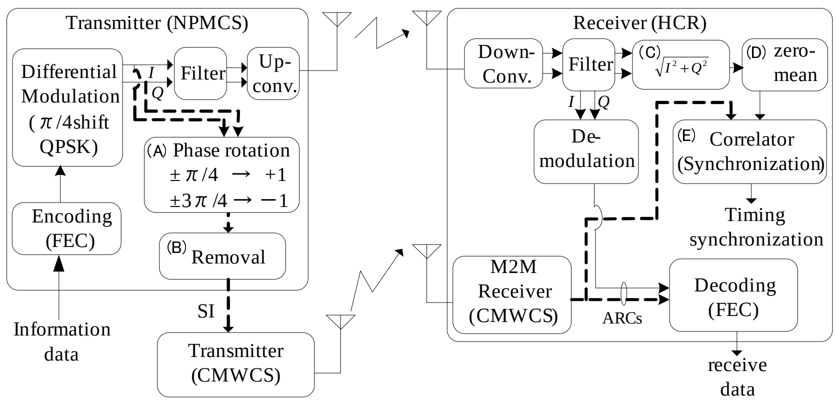

Figure 4.23: The modified HCR employing the proposed timing synchronization technique.

self-synchronization and the co-synchronization, and the BER improvement are shown from next sub-sections.

We explain disquisition of the proposed technique. Since the employed SI consists of the phase rotational information, it cannot be directly correlated with the received signal of the NPMCS. Hence, Timing synchronization is achieved by correlating the SI with the envelopes of the received signals of the NPMCS. Moreover, by introducing the envelopes, the correlation is not affected by carrier frequency offset. The correlation employing envelopes eliminates deterioration caused by the carrier frequency offset.

4.3.2.1 Characteristics of $\pi/4$ shift QPSK

Before explaining the correlation method, we clarify the characteristic of the envelopes of the $\pi/4$ shift QPSK. Figures 4.24 and 4.25 show a constellation and an envelope of $\pi/4$ shift QPSK, respectively. Figure 4.24 shows that if a phase rotates by $\pm\pi/4$, the amplitude at the midpoint of the transition from one symbol to the adjacent symbol is usually smaller than the average amplitude. If the phase rotates by $\pm\pi/4$, the amplitude is generally increased. From the above characteristics, we can forecast the envelope of the $\pm\pi/4$ shift QPSK.

Here, the amplitude at the midpoint is derived using mathematical expressions. Since variations of the envelopes are attributed to band-limits, we first show the impulse response of the raised-cosine filter often employed for QPSK

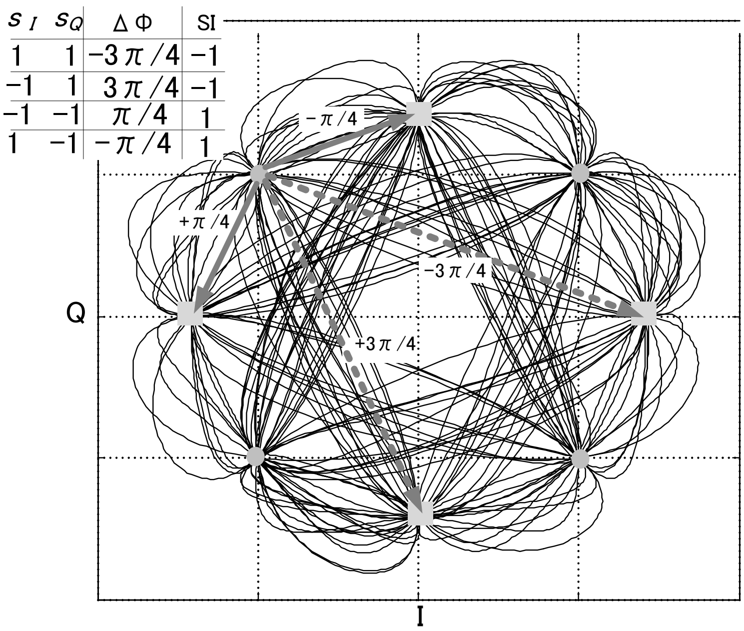

Figure 4.24: The constellation of $\pi/4$ shift QPSK($\alpha=0.5$).

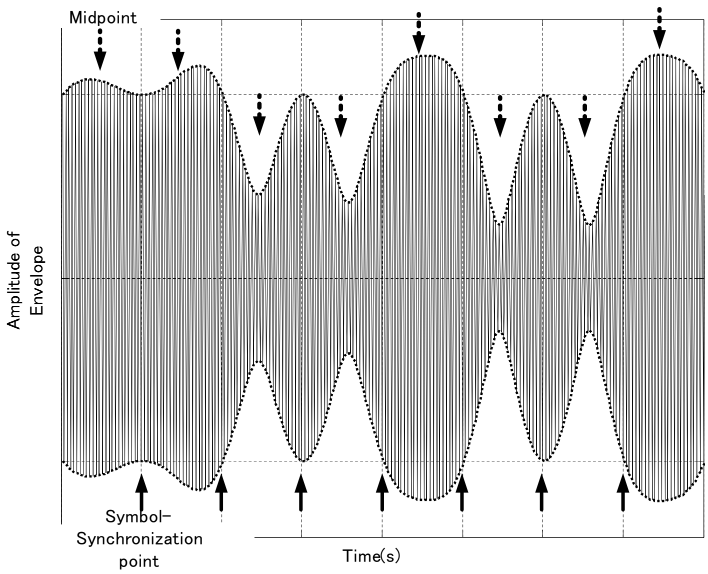

Figure 4.25: An envelope of $\pi/4$ shift QPSK($\alpha=0.5$).

modulation as follows [49]:

$$h(t) = \frac{\sin(\pi t/T)}{\pi t/T} \cdot \frac{\cos(\alpha \pi t/T)}{1-(2\alpha t/T)^2}, \tag{4.1}$$

where t and T indicate represent time and symbol duration, respectively. α is roll-off factor. Figure. 4.26 shows an example of amplitude of in-phase (I) and quadrature-phase (Q) when the bandwidth is limited by $h(t)$.

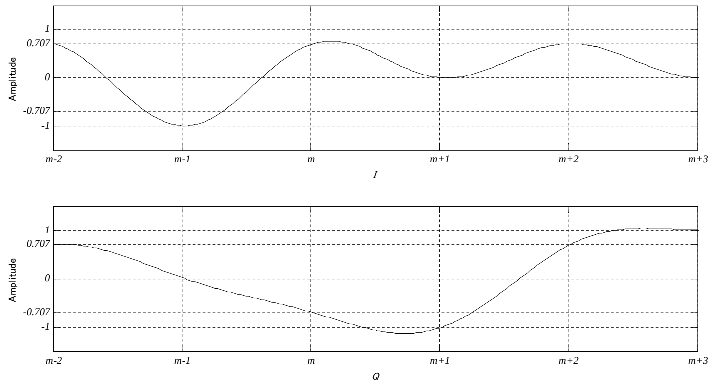

Figure 4.26: In-phase and quadrature-phase signals of $\pi/4$ shift QPSK, when bandwidth is limited by raised-cosine filter (α=0.5).

In the Fig 4.26, we force on the midpoint between m and m+1. The amplitude of I and Q at this midpoint, I_m and Q_m, is as follows:

$$I_m = \sqrt{P} \sum_{n=-\infty}^{\infty} \cos\left(\sum_{k=-\infty}^{n-1} \Delta\theta_k + \theta_0\right) h\{T/2-(n-m)T\}$$
$$Q_m = \sqrt{P} \sum_{n=-\infty}^{\infty} \sin\left(\sum_{k=-\infty}^{n-1} \Delta\theta_k + \theta_0\right) h\{T/2-(n-m)T\}, \tag{4.2}$$

where P represents power of the envelope at symbol decision points. n and k indicate index number of symbol. θ_0 and $\Delta\theta_k$ are initial phase and phase rotation from k th to $k+1$ th symbol, respectively. The amplitude of the envelope at the midpoint is

$$A_m = \sqrt{I_m^2 + Q_m^2}. \tag{4.3}$$

To simplify Eq.4.2 and Eq.4.3, when $h(t)$ is restricted from $T/2$ to $-T/2$, the approximate amplitude A'_m of Eq.4.3 is as follows:

$$\begin{aligned} A'_m &= \frac{2\sqrt{P}\cos(\alpha\pi/2)}{\pi(1-\alpha^2)} \times \\ &\quad \sqrt{\{\cos(\theta_m)+\cos(\theta_m+\Delta\theta_m)\}^2+\{\sin(\theta_m)+\cos(\theta_m+\Delta\theta_m)\}^2} \\ &= \frac{2\sqrt{2P}\cos(\alpha\pi/2)}{\pi(1-\alpha^2)}\sqrt{\cos(\Delta\theta_m)+1}, \end{aligned} \tag{4.4}$$

where $\theta_m = \sum_{k=-\infty}^{m-1} \Delta\theta_k + \theta_0$ standing for the phase of the m th symbol. From Eq.4.4, when $\Delta\theta_m$ is $\pm\pi/4$, $\cos(\Delta\theta_m)$ become positive and A'_m become large. (when $\alpha = 0.5$ and Although A'_m is not accurate amplitude A_m, A'_m shows the characteristic of the amplitude strongly in the midpoint of $\pi/4$ shift QPSK symbol transition. Because pulse responses of two immediate symbols located around focused midpoint, $h(-T/2)$ and $h(T/2)$, affect A_m the most; That is $|h(T/2+n)|$ decreases with symbol number increasing $|n|$.(n is integer number.) To calculate more accurate amplitude A'_m, we utilize computer calculations. Figure. 4.27 shows a histogram when A'_m is calculated 100,000 times by selecting $\Delta\theta_k$, $\pm\pi/4, \pm3\pi/4$, at random. In the calculations, $P = 1, \alpha = 0.5$ and side-lobe of impulse response $h(t)$ is cut off in 21 symbols. From Fig. 4.27, we confirm that if a phase rotates by $\pm3\pi/4$, the amplitude of the midpoint is smaller than 0.8. If a phase rotates by $\pm\pi/4$, the amplitude is larger than 0.8.

4.3.2.2 Correlation with Envelopes

To utilize the characteristic for the timing synchronization, the signals received by the NPMCS are modified as follows: first the received signals are changed to the envelope, then, the envelope is shifted such that its mean is zero without changing the waveform (Fig. 4.23, module (C) and (D)). By shifting the envelope, the amplitude of the envelope at the midpoint of the symbol transition is almost negative (positive) if the phase rotates by $\pm3\pi/4(\pm\pi/4)$. Since the SI has been defined as $\pm3\pi/4 \to -1$ and $\pm\pi/4 \to +1$, we can determine the correlation between the SI and the zero-mean envelope. This is achieved in module (E) of Fig. 4.23. Figure 4.28 shows the zero-mean envelopes and the SI sequence, where

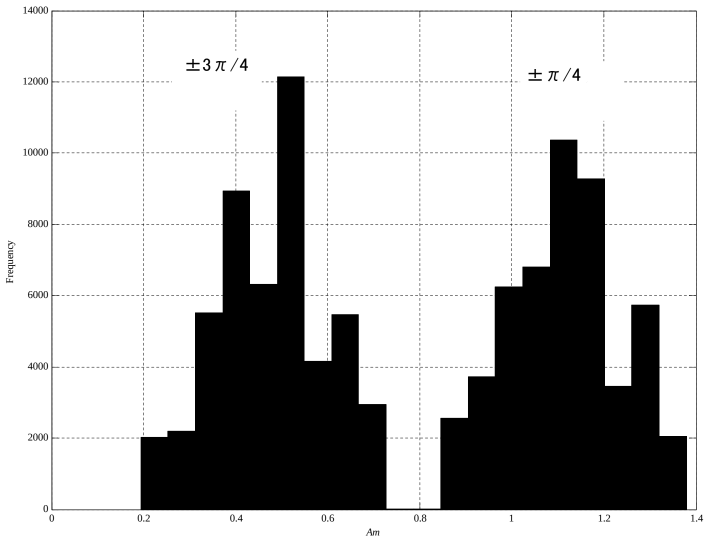

Figure 4.27: A histogram of the midpoint amplitude of $\pi/4$ shift QPSK (roll off factor α =0.5).

some of the SI time-coincides with the symbols. The correlation between the zero-mean envelope and the SI sequence is computed, while the SI sequence is shifted laterally. When the SI and the zero-mean envelope have the same sign, both values are multiplied to a positive result, and the correlation increases. By searching where the correlation is highest, the HCR can achieve timing synchronization. This timing synchronization can obtain not only self-synchronization of the NPMCS but also the co-synchronization.

4.3.2.3 BER Improvement

Moreover, the SI can be used for BER improvement as well as the synchronization. This is because the SI consisting of the phase rotation represents sending information itself in the differential coding. From the definition of the SI —the phase rotations $\pm 3\pi/4$ and $\pm\pi/4$ are assigned to -1 and +1, respectively—the HCR can obtain $-s_Q$ information shown in Fig.4.24 for BER improvement. The integrated received codewords are produced by combining the received signals of the NPMCS with the SI_A shown in Fig. 4.29. By the decoding the integrated received codewords using the Viterbi algorithm, which is employed for most of

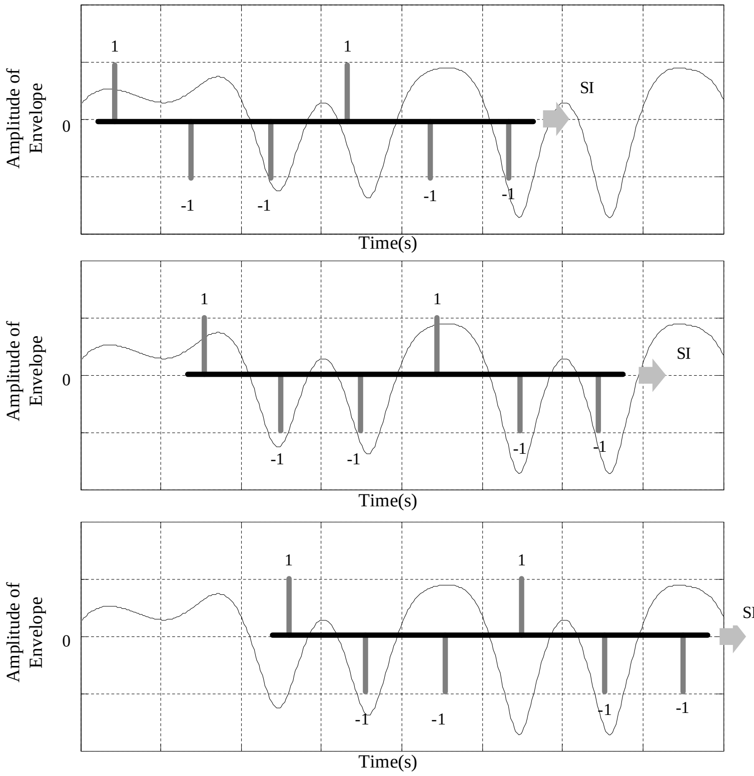

Figure 4.28: The correlation between the zero-mean envelopes of the NPMCS and the SI sequences in the timing synchronization.

the NPMCS, BER of the NPMCS is improved. When coding-rate R is 1/2 in the NPMCS, the integrated receivedcodewords are decoded as the punctured code of coding rate 1/3, where zero is inserted in the removed SI (codewords).

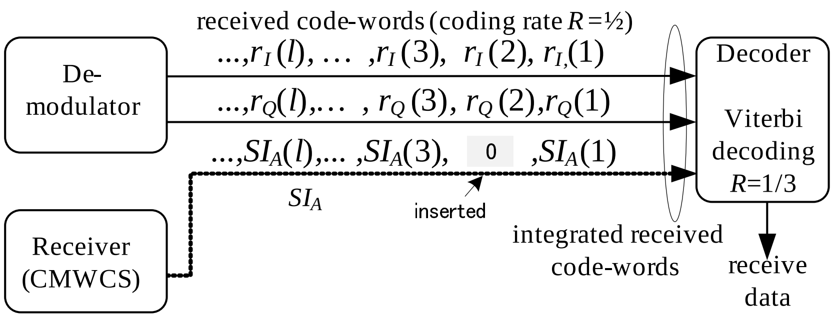

Figure 4.29: Decoding of IRCs (integrated received codewords).

In terms of the BER improvement, optimal SI is not SI_A. We therefore introduce SI_B that makes the free distance [50] of thecodewords maximize. In the Viterbi decoding, the BER decreases with increasing the free distance. By employing the SI_B, further BER improvement is expected. We summarize feature of the SI_A and SI_B. The SI_A is used for both the timing recovery and the BER improvement. However, the capability of the BER improvement is inferior to the SI_B. The employment of SI_B improves BER effectively by maximizing the free distance of the integrated received codewords, although not being used for the timing recovery.

Table 4.2 shows the free distance of the integrated received codewords. SI_A and SI_B are compared under the same removal rate. In the case of SI_B, the generator polynomial, which shows the relationship between the codewords and the past inputs, must be changed by the removal rate so that the free distance can be the largest. In the table 4.2, the parameters of the convolutional code of the NPMCS are as follows: constrain length $K = 6$, coding rate R=1/2, generator polynomial g1 =47 and g2 =75 (octal). Additionally, we assume that SI_A and SI_B have no-error and infinite reliability.

Table 4.2: Free distance of IRC when the removal rate is the same

Removal Rate	Free distancegenerator polynomial of SI (octal)	
	SI_A	SI_B
1/3	15[75]	19[41]
1/2	11[75]	15[57]
2/3	8[75]	10[61]
1/1	8	

4.3.2.4 Adaptive Coding

In order to reduce quantity of the transmitted SI, we consider the method that controls quantity of sending both SI_A and SI_B. The method is referred to as the adaptive coding in the thesis. Figure 4.30 depicts the HCR that deals with both SI_A and SI_B for the adaptive coding. The integrated received codewords of Fig. 4.30 are shown in Fig. 4.31. The generator polynomial of SI_B is switched when the removal rate is changed. In the case of the adaptive coding, the free distance of the integrated received codewords becomes intermediate value between SI_A and SI_B alone.

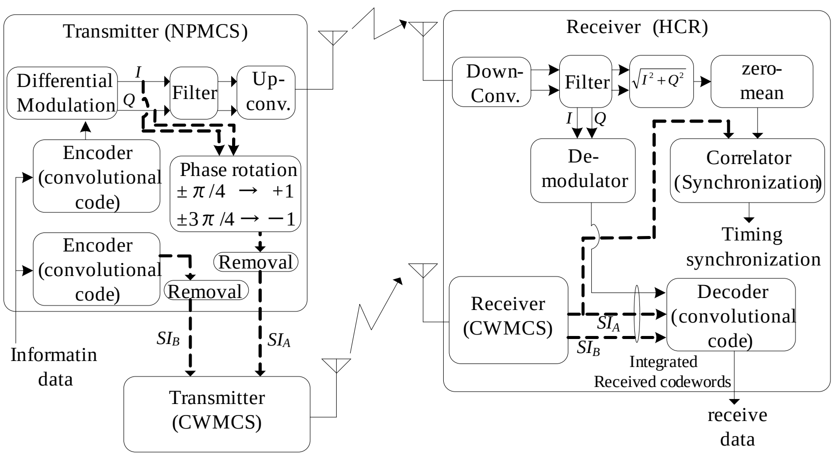

Figure 4.30: The proposed system adopting the adaptive coding.

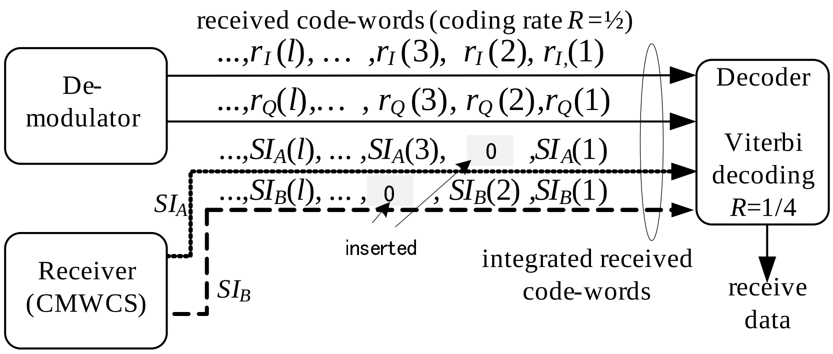

Figure 4.31: Decoding of IRCs in the case of the adaptive coding.

The concrete procedure of the adaptive coding is as follows. First, the HCR evaluates if the timing synchronization is stable. When the timing synchronization is not stable, quantity of sending SI_A will be increased by reducing the removal rate to stabilize the timing synchronization. After stabilizing the timing synchronization, sending SI_B will be increase until the BER reaches specified value, for example, BER$< 10^{-4}$, which is required in most speech communication systems. In the case of difficulty of timing synchronization, quantity of the SI_A

Table 4.3: Parameter of RCR STD-39 [1]

Access method (Down-link/Up-link)	TDMA(4multiplexing)/TDMA
Modulation	Differentially encoded π shift QPSK
Symbol duration T	62.5 μs
Slot length	10ms (160 symbols)
Frame length	40ms
Transmission rate	16kbaund(32kbps)
FEC	convolutional code
Coding rate R_{pu}	1/2
Constrain length K	6
Generator sequency (octal natation)	g(1)=53 g(2)=75 (octal)
Tail bit	5bit
Interleave	Interleave in a slot 2 frames interleave
Synchronous word (SW)	20bit

is on increase. The SI_B increasing cases are that BER is deterioration even if the timing synchronization is stable.

4.3.2.5 Computer Simulation

In the computer simulation, we employ the RCR STD-39[1] as the NPMCS. Table 4.3 shows parameters of RCR STD-39.

In the simulation, the SI_A and SI_B received by the HCR have no-error. This is because we assume that a ready-made M2M (Machine to Machine) module is used as the CMWCS receiver included in the HCR as mentioned in sub-section 3.1.4. The received SI_A and SI_B via the M2M module become correct hard decision values owing to CRC (Cyclic Redundancy Check) and ARQ (Automatic repeat-request). We simulate the performance in AWGN (Additive White Gaussian Noise), 1 path Rayleigh fading, and IEEE802.22(WRAN) multipath model Profile C [52]. We consider that the Profile C can be representative of PMCS's real multipath fading when long-delay multipath waves occur. The Maximum Doppler frequency fd is 10Hz and 40Hz in the simulation.The fading parameters are shown in table 4.4.

Table 4.4: Fading parameters

Model	(1) AWGN (2) 1 path Rayleigh fading (3) IEEE802.22(WRAN) multipath model Profile C (mean excess delay:21 μs, RMS delay spread:8.4 μs)
Maximum Doppler frequency	(1) $fd = 10$ Hz, $fd \cdot T = 62.5 \times 10^{-5}$ (Speed:28 km/h Carrier frequency:400 MHz) (2) $fd = 40$ Hz, $fd \cdot T = 250 \times 10^{-5}$ (Speed:108 km/h Carrier frequency:400 MHz)

Figures 4.32 and 4.33 show the synchronization probabilities when the removal rate of SI_A is changed. Here, the synchronization probability is referred to as the probability that the simulated timing falls within ± 0.5 symbols of the optimal synchronization point. In the case of the Profile C, we define the mean excess delay as the optimal synchronization point [53][54] in this thesis. The synchronization timing is searched from the range of one frame. The quantity of the SI removal (removal rate) is changed from 1/10 to 2/3. For comparison, Figs.4.32 and 4.33 include the performance of the conventional synchronization method [43] that correlates the received signal of the PMSC with the 20bits synchronous words (SW) shown in Table 4.3.

As shown in Figs.4.32 and 4.33, we can confirm that the synchronization probability increases with decreasing removal rate. In the 1 path Rayleigh fading, a higher SNR is required to reach high synchronization probabilities compared with the AWGN environment. This is because the received signals of the PMCS sometimes almost disappear in slow and non-selective fading environments. In the Profile C, although the signals rarely disappear owing to frequency selective fading (the time diversity effect), inter-Symbol-interference (ISI) makes the synchronization probabilities deteriorate. To mitigate the signal disappearance and ISI, Space diversity techniques [55] may be required.

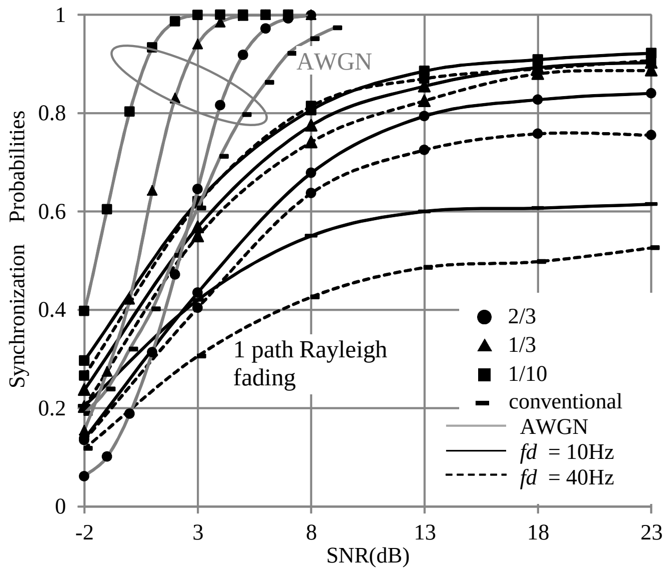

Figure 4.32: Synchronization probabilities (AWGN and 1path Rayleigh fading).

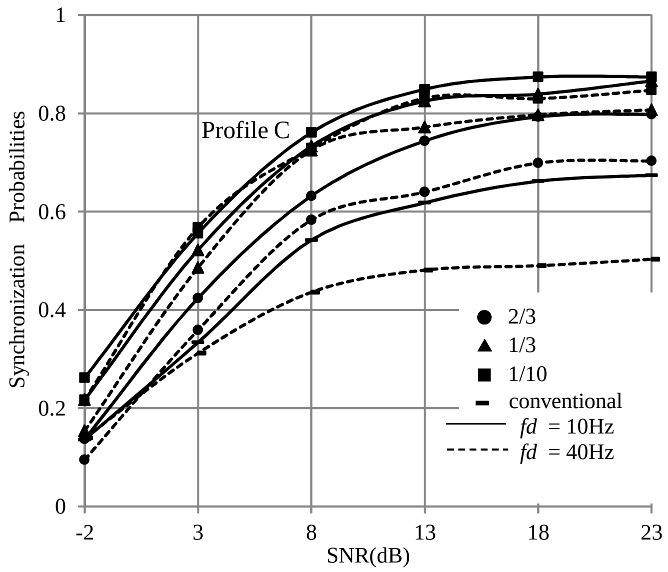

Figure 4.33: Synchronization probabilities (IEEE 802.22 multipath model Profile C).

Aside from space diversity techniques, we consider another synchronization improvement method combining the proposed method with the conventional method. Figures 4.34 and 4.35 show the performance of this combined method that achieves synchronization by adding the correlation values between the received signals and the SW (the conventional method [43]) to the correlation values between the envelopes and SI_A (the proposed method). By combining, a further improvement in the synchronization probability is achieved for low SNRs.

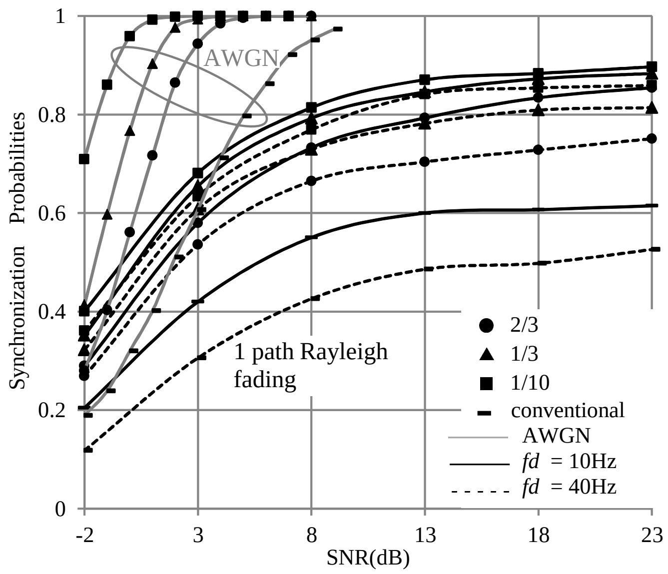

Figure 4.34: Synchronization Probabilities by combining the proposed and the conventional method (AWGN and 1path Rayleigh fading).

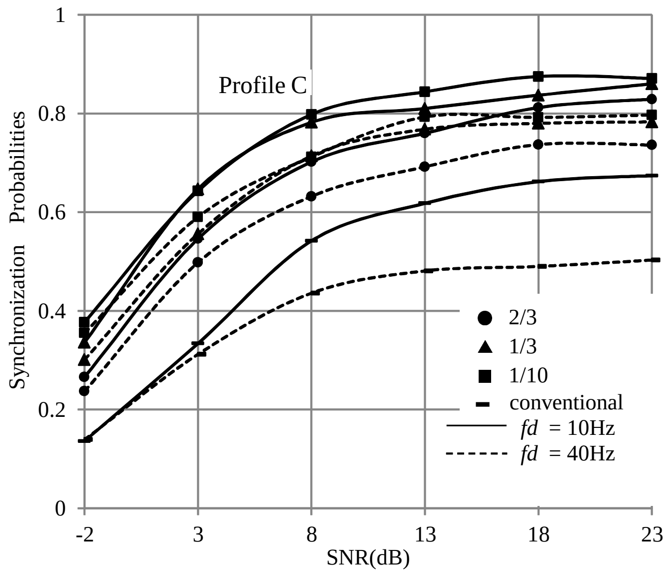

Figure 4.35: Synchronization Probabilities by combining the proposed and the conventional method (IEEE 802.22 multipath model Profile C).

Figure 4.36 shows the root mean sequence (RMS) jitter [46] characteristics of this combined method in the case of fd=40Hz. In the case of AWGN and 1 path Rayleigh fading, the jitter decreases as the removal rate decreases. In the case of the Profile C, the jitter can hardly decrease with increasing SNR. It appears that the jitter is large. However, since we define the mean excess delay as the optimal synchronization point, we cannot correctly assess the jitter shown in Fig.4.36 shows suitable values. In the frequency selective fading channels, the instant optimal synchronization point usually differs from the mean excess delay. The influence of the jitter is evaluated in Section 4.3.4.

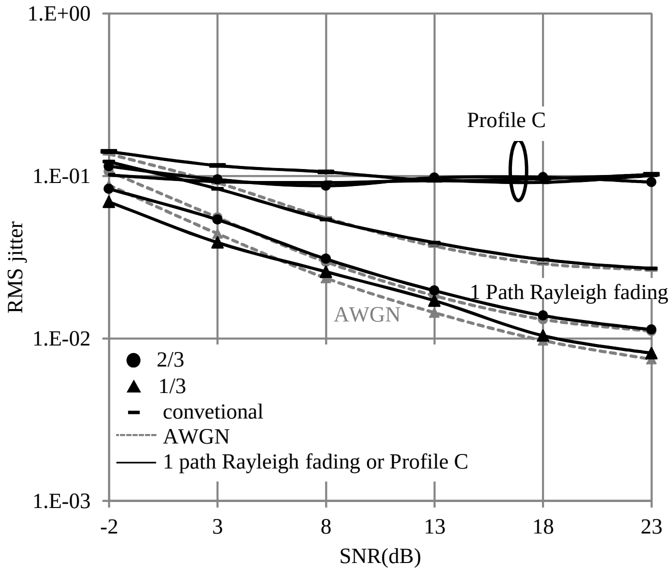

Figure 4.36: RMS jitter characteristics(fd=40Hz).

Here, we consider the influences of the roll-off factor. The synchronization probabilities are shown in Fig.4.37 when the roll-off factor α is changed to 0.2, 0.5, and 0.8. The other parameters are the same as in Tables 4.3 and 4.4. The removal rate is 1/3 and fd=40Hz. Figure 4.37 shows that the roll-off factor does not significantly affect the synchronization probabilities.

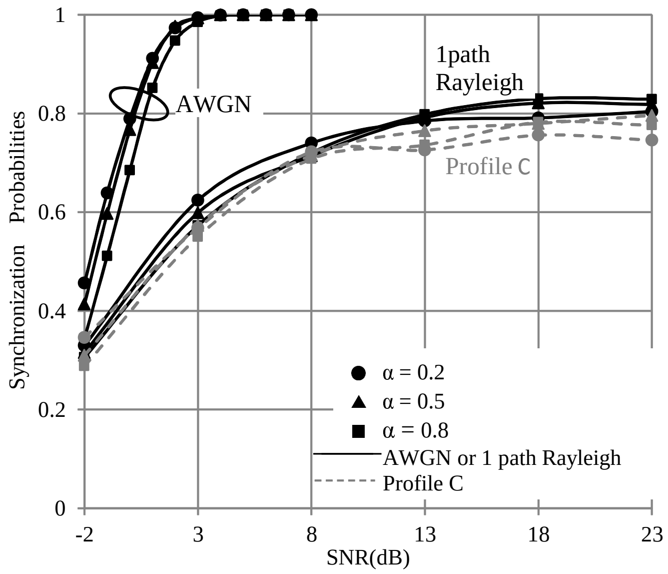

Figure 4.37: Influences of the roll-off factor(fd=40Hz).

Figures 4.38 and 4.39 show the BER performance using the SI_A shown in Fig.4.23 and Fig.4.29, when the synchronization falls within ±0.5 symbols of the optimal synchronization point, and there is no frequency error in the down-converter of the NPMCS. In the case of AWGN and 1 path Rayleigh fading, although the simulated BER performance includes the effect of the jitter shown in Fig.4.36, the BER performance is hardly degraded. The BER performance is improved as the removal rate decreases; the quantity of SI_A increases. In the case of the Profile C, the BER performance of the proposed method (with jitter) can also be improved by adding SI. However, the performance is inferior to that of the optimal synchronization point (without jitter), which is the mean excess delay of delay profile. From this result, we think that there is large jitter when employing the proposed method in frequency selective fading environments as shown in Fig. 4.36. However, since the proposed method has high synchronization probability, we consider that the deterioration caused by the jitter can be compensated by increasing amount of SI. For reference, the reason why BER decreases with increasing fd is attributed to enhancement of interleave effect. The benefit of the interleave is to provide time diversity (when used along with FEC.) [56]

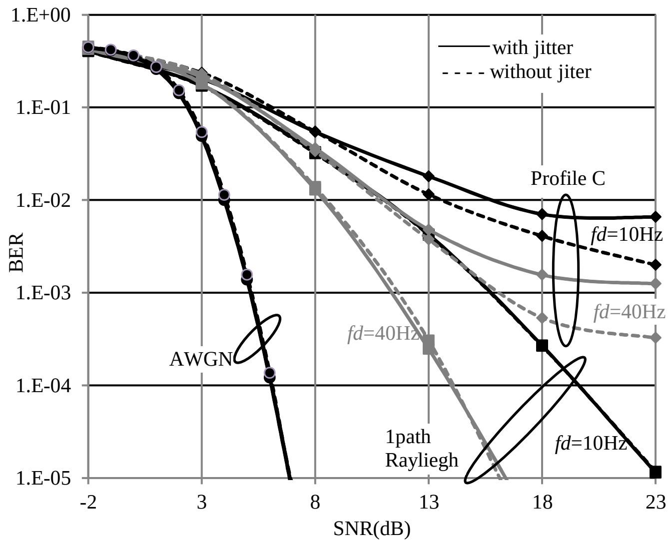

Figure 4.38: The BER improvement when changing removal rate of SI_A (Removal Rate 2/3).

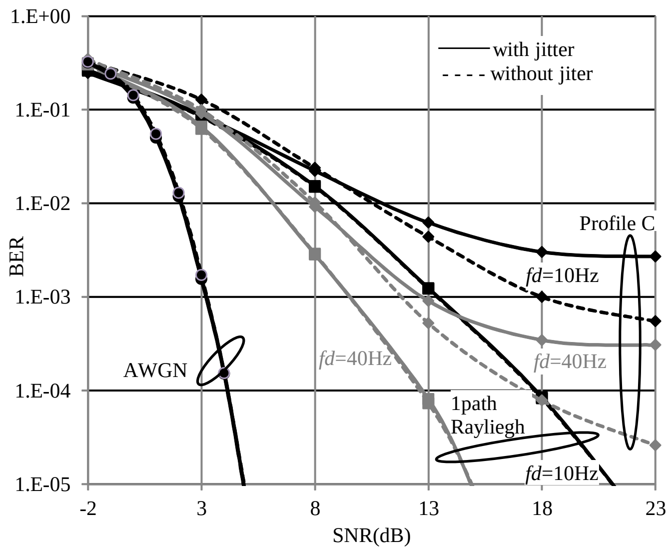

Figure 4.39: The BER improvement when changing removal rate of SI_A (Removal Rate 1/3).

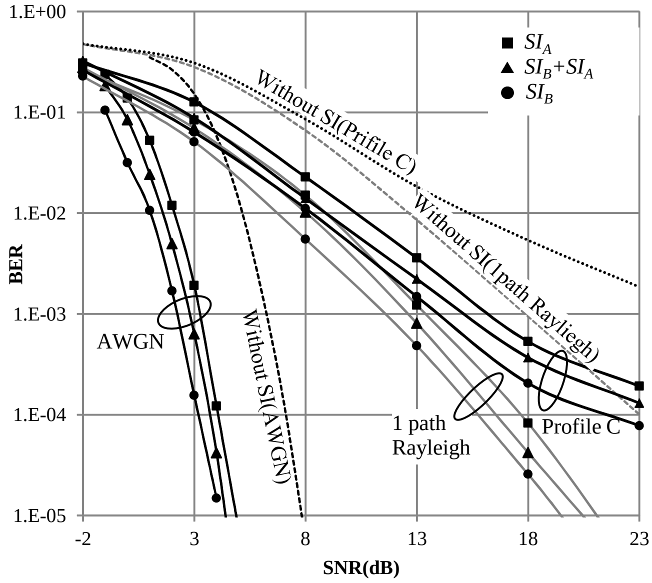

Figure 4.40: The BER performance when using SI_A, SI_B, and SI_A + SI_B (fd=10Hz).

Figure 4.40 shows the performance of the BER improvement method employing SI_A, SI_B, and SI_A+SI_B shown in Figs. 4.30 and 4.31, when we assume that the synchronization has already been obtained, fd=10Hz, and the jitter is set to zero. In this figure, the total removal rate is fixed at 1/3; i.e., the total quantity of SI is the same. In the case of SI_A+SI_B, the removal rates of SI_A and SI_B are 2/3. In terms of the BER improvement, we confirm that SI_B is superior to SI_A if the system can synchronize properly.

4.4 Chapter Summary

In this chapter, we described difficulty of synchronization when NPMCSs are in low SNR environments. In the sub-section 4.2, we proposed method employing GPS signals. The method can obtain self-synchronization of NPMCSs and co-synchronization. In the simulation, we found that BER performances were deteriorated by synchronization timing error when error from the optimal synchronization point is within 4% symbol duration. Additionally, to come close the optimal synchronization point, we proposed the method employing both GPS and

MAM. As a result, the method was effective in low fd environments. Moreover, when considering indoor environments where GPS signals cannot be available, we proposed that the synchronization and BER improvement method employing the SI transmitted by the CMWCS. The timing synchronization was obtained by forecasting envelopes from the SI and then correlating them with the envelopes of the NPMCS's signals in differentially-encoded $\pi/4$ shift QPSK. The method can also be used for BER improvement by decoding IRCs constructed from the SI and the received signals of the NPMCS. Moreover, to realize both the synchronization and the BER improvement simultaneously using minimum SI, the adaptive coding was considered.

Chapter 5

Heterogeneous Cognitive Radio (HCR) combining PMCSs

5.1 Combination of Narrowband PMCSs

In this subsection, we consider advantages and problems when narrow band PMCSs are integrated by SDR.

5.1.1 Advantages of Integration in Narrowband PMCSs

First, we explain five advantages when SDR is employed to NPMCSs.

5.1.1.1 Integration of NPMCSs

SDR techniques can integrate several PMCSs to deal with one MS as shown Fig. 5.1. So far, there are many PMCS's standards such as RCR STD-39, T-79. However, as shown in Fig.5.1(a), we must bring a couple of MSs when using several systems because of incompatibility.

Application of SDR lets us release to bring several MSs. Moreover, cellular systems, radio, TV, and GPS can be realized on one MS. We therefore expect splendid MSs that have a lot of functions.

5.1.1.2 Intercommunication

SDR can operate several systems simultaneously and exchange data each other freely. (Hereinafter, we call this function a link function.) For example, in the present circumstances, when two systems are connected, we must prepare two

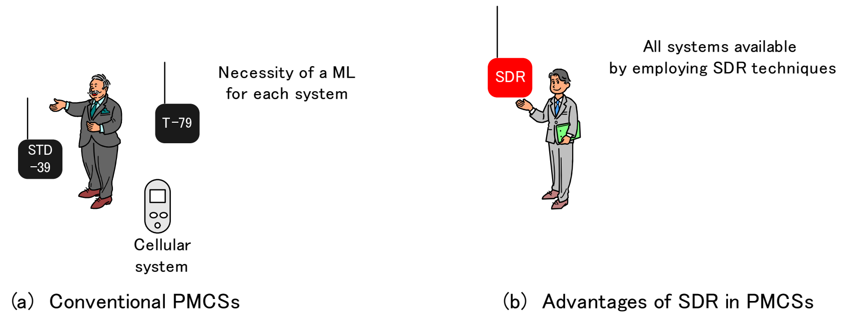

Figure 5.1: Integration of several NPMCSs using SDR techniques.

MSs and connect the MSs using analog cables as shown Fig.5.2(a). In SDR, only one MS can realize the link function as shown Fig.5.2 (b).

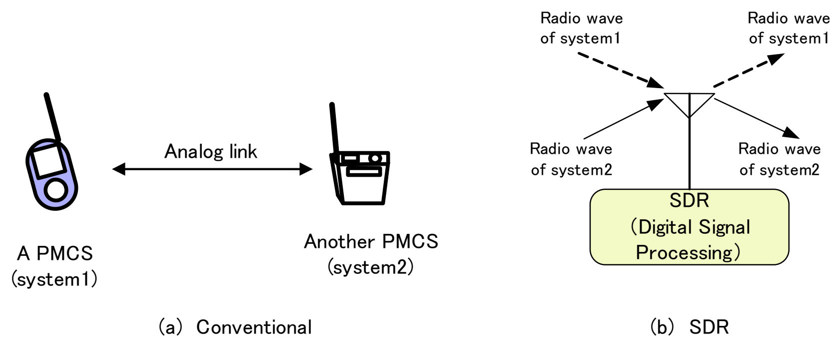

Figure 5.2: Example of the link function.

5.1.1.3 Expansion of Services Areas

To link PMCSs is also useful for expansion of service areas in a manner similar to explanation of chapter 2. By application of SDR, no-service areas can be reduced since each PMCS has various service areas. Moreover, handover between systems are realized smoothly. If a MS belonging to a certain system moves in no-service areas, the MS can realize handover to other systems so that users can prevent isolation. Figure 5.3 shows an example of integrated BSs by employing the SDR technique. In this case, the BSs can handle data that comes from MSs of several systems simultaneously. That means several systems can be linked at BSs. Moreover, by being able to receive all PMCS's signals at one BS, service areas are expanded because the number of BSs increases from a MS's point of view. However, in transmission of BSs, we must consider avoidance of interference when BSs transmit several system signals.

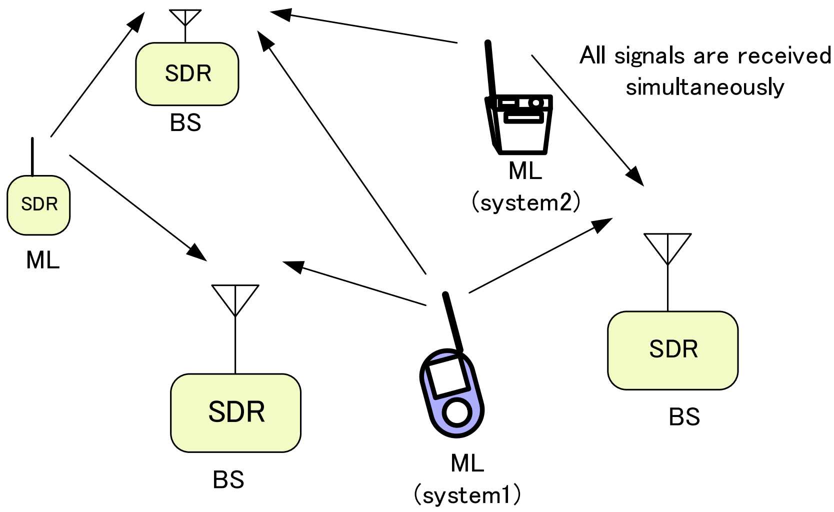

Figure 5.3: BSs employing SDR techniques..

The contents described above relate to the heterogeneous cognitive techniques [10][11]. The HCR employing SDR techniques can employ existing systems effectively and enhance reliability of communication.

5.1.1.4 Reduction in Development and Manufacture Costs

By SDR techniques, hardware can be communalized even if different PMCSs are used. That means we do not need to develop and manufacture new hardware. In case of introducing new system, since we develop only software, the cost and time of development can be reduced. However, we do not have knowhow enough for realization of SDR so far, and the development of SDR that can be used in real environments requires considerable cost. For the future, progress of SDR techniques is further required for introducing SDR that can reduce the cost.

5.1.1.5 Easy Version up and Bug Fix

Version up and bug fix of SDR are easily accomplished by downloading update file through recording medium, wired network, and wireless network. In the case of conventional radio, hardware must be directly upgraded after recalling all radios.

On the other hand, SDR requires only distributing software for version up and bug fix instead of recalling all radios. That means we expect reduction of amount of repair and maintenance works.

5.1.2 Features of Introduced SDR and Introducing Timing

In sub-section 5.1.1, we explain advantages when SDR is adapted to PMCSs. In sub-section 5.1.2, we explain suitable SDR that should be introduced in PMCSs while considering costs of development and manufacturing. In sub-section 5.1.2.1, we separate SDR into two types, "SDR with full functions" and "SDR with limited functions". Here, "limited functions" means modulation schemes, carrier frequency, and bandwidths are restricted. Next, in sub-section 5.1.2.2, we consider introduction timing when above two types of SDR are employed in PMCSs. This is because SDR will be further developed and then manufacturing costs will be reduced. In sub-section 5.1.2.3, we summarize features of introduced SDR and introduction timing.

5.1.2.1 Functions of SDR

- SDR with full functions

  Ideally, SDR needs to be able to change any frequency and modulation. We call this ideal SDR "white radio" because white color can be changed to any colors (functions). In the case of PMCSs, the realization of the white radio is also expected. This white radio can realize not only link function and handover between systems but also realizable system update. New devices are not perhaps required for renewed systems. However, since the white radio (SDR with limited functions) must have several problems such as development of high performance A/D, it will take a long time to realize the white radio. Moreover, we forecast that high development costs are needed to develop the white radio.

- SDR with limitecd functions

  We consider a SDR with limited functions to reduce development costs and duration. We restrict functions of the SDR to easily realize the SDR.

Table 5.1: Advantages and disadvantages attributed to limiting functions

SDR with full funtions	1)New system will be installed in the future. 2) available for systems other than PMCSs	1)Huge amounts of costs for development 2)required long time for accomplishment
SDR with restricted functions	1)available at present techniques 2)costs can be suppressed.	1)limited systems can be used. 2)future systems cannot be used. 3)costs are high compared to conventional NPMCS's systems.

The SDR cannot freely install software to change. So far, realized SDR have this type. Considering that conventional NPMCSs are integrated to SDR, we can construct SDR sooner rather than later. This is because carrier frequency of NPMCSs is lower than one GHz and their bandwidth is narrow compared to cellular systems that use GHz carrier frequency and several MHz bandwidths. The means technical difficulty of A/D and D/A can be mitigated when SDR techniques are employed only for conventional NPMCSs. Moreover, we consider that RF signals can be sampled directly for signal processing in a SDR. In the case of using superheterodyne method and IF sampling method, the SDR can avoid becoming big size because of limited carrier frequency and bandwidth. From these reasons, SDR only for NPMCSs can be realized easily. However, the SDR with limited functions only for PMCSs are not adapted to next generation PMCSs. New devices of SDR must be purchased in this case. We will face with a dilemma because the feature of SDR, which can change any systems, cannot be available. Table 5.1 shows advantages and disadvantages when SDR with full functions and SDR with restricted functions are employed, respectively.

5.1.2.2 Introducing Timing of SDR for PMCSs

In 5.1.2.1, we classified SDR into SDR with full functions and SDR with restricted functions. In this sub-section, we summarize introduction timing of classified SDR, considering development and manufacture costs. We consider two timing when SDR are employed as soon as possible or SDR are not employed until popularization.

- SDR with full functions and swiftly introduction

  So far, SDR with full functions have not been realized because technical problems cannot be solved. If the SDR are forcibly realized using present techniques, the size of the SDR would be very big and cost would also be high. Hence, immediate realization is impossible.

- SDR with restricted functions and swiftly introduction

  If we accept limitation of functions, SDR can be realized using present techniques. However, costs of development and manufacturing will be high. Moreover, performances and power consumption of the SDR would be inferior to those of ordinary PMCSs.

- SDR with restricted functions and introduction when SDR techniques prevail

  When waiting for introduction of SDR until SDR techniques prevail, we can realize the SDR at a moderate price. In this case, link of several PMCSs and handover to other PMCSs can be realized. Hence, this timing is one candidate when SDR are employed to PMCSs. However, since the SDR might not be employed to the next generation PMCSs, obsolescence of the SDR is ineluctable.

- SDR with full functions and introduction when SDR techniques prevails

  Ideally, the SDR with full functions should be employed to PMCSs. In this case, when new systems are introduced, we do not need to prepare new devices for SDR, that is, costs will reduced. However, the ideal SDR require a long time before it can be used, and it takes more time for becoming popular.

5.1.3 Problems for Realization

In sub-section 5.1.3, we show problems that must be solved for realization of SDR.

5.1.3.1 Technical Problems

As explained in chapter 2, we need to solve technical problems, broadband antennas, high performance transmitters and receivers, high speed computing digital signal processers, etc. In the SDR with full function, to solve the problems are especially difficult. On the other hand, although the SDR with restricted function can be realized using present techniques, manufacturing costs are expensive. For prevalence of SDR, it is necessary to enhance reliability as well as to solve the problems.

5.1.3.2 Miniaturization and Lightweight

In PMCSs, small and lightweight MSs are required. So far, SDR's MSs are inferior to conventional NPMCS's MSs in terms of radio performances, power consumption and size. When employing SDR in PMCSs, we must consider not only functions but also small and lightweight MSs.

5.1.3.3 Guarantee of Confidential Communication

- Supervision of software

  We must strictly consider that SDR can protect confidential communication. When software installed in a SDR leaks out, the SDR has potential danger of leaking information. Nowadays, since leaking information frequently happens, the software of SDR must be strictly supervised in order not to leak out. Moreover, when we assume that MSs constructed by a SDR is stolen, the robber may easily be able to unravel the cryptograms of the stolen MS compared to a conventional MS. In the case of the SDR, software must be protected such that programs can never be analyzed by disassembly and other techniques. To protect confidential communications, we must not only supervise software strictly but also encrypt the programs of the software in order not to be analyzed. We worry that confidential communication is not sufficiently considered at present.

- Unique Specification of Hardware and Operating System (OS)

In the future, if commercial SDR and general OS are employed for PMCSs, we can construct SDR for PMCSs easily and the costs are not high. Moreover, software of radio systems such as TV, GPS, and a cellular phone can be easily installed in PMCSs constructed by commercial SDR, that is, PMCSs that have many functions can be realized. However, in this case, we must verify whether eavesdropping is surely prevented. SDR installing commercialized OS have potential danger of computer virus infection. Hence, in terms of security, we hope that hardware and OS of SDR for PMCSs is uniquely developed although the costs are not low. SDR installing commercialized OS cannot completely prevent computer virus infection even if a countermeasure system against virus infection works strongly. This is because computer viruses for commercialized OS are produced one after another, that is, the countermeasure system may look over new type of viruses.

5.1.3.4 Revision of Radio Act (Law)

To realize SDR, revision of the radio act is necessary. In Japan, although ministry of internal affairs and communications (MIC) has discussed the revision about standards and conformity assessment system, when and how to revise are undefined because of premature discussion. Hence, in SDR for PMCSs, while we survey revision of the radio act and then check whether the SDR observe the radio act, we should make specifications of the SDR.

5.1.4 Summary

We considered SDR combining narrowband PMCSs. As a result, application of SDR is effective to realize high performance radios and to reduce no-service areas. Moreover, in the future, costs of manufacturing of SDR will be reduced because radio systems can freely be changed by installing software, that is, new hardware is not required. However, at present, realizable SDR do not have enough radio performances. Additionally, security of SDR must be further strengthened to prevent eavesdropping. Hence, application of SDR combining PMCSs may be waited until SDR are realized by commercial systems. So far, realization of

SDR for PMCS is difficult because of technical problems and high costs. One of the ideal forms of PMCSs is that all PMCSs can be integrated into one small SDR so that PMCSs can freely be changed and no-service areas can be covered by handover of different systems. We expect application of SDR techniques is crucial for PMCSs in the future.

5.2 Combination of Broadband PMCSs and Narrowband PMCSs

5.2.1 Motivation

Recently, several institutes and govements develop broadband PMCSs (BPMCS) such as the public broadband wireless communication system (PBWCS) [6] and the long term evolution (LTE) for public safety [5]. The PBWCS have been researched and developed by NICT. Now, the PBWCS realizes high speed transmission by employing frequency band of 200MHz that are produced by frequency assignment reconstruction. In the national police agency, the PBWCS has already been operated to transmit motion videos of disaster areas to police headquarters. The LTE for public safety is discussed in the 3GPP [5] to realize group communication and direct communication, which provide users with wireless communication without base stations. Now, in the U.S., the LTE for public safety is operated on a trial basis.

Although BPMCSs are expected to expand gradually, we do not think that the conventional NPMCSs are completely replaced with BPMCSs. There are two reasons why the NPMCSs are necessary. One is that all areas that PMCSs must service are not covered by BPMCSs solely. Since the BPMCSs employ small or middle zone scheme to deploy service areas, an enormous budget is required to cover all service areas that have already been covered by the NPMCSs employing large zone scheme. Because PMCSs are required as the crucial systems offering communication for protecting human life, reduction of service areas is not allowed. The other reason is that users require crucially robust PMCSs, which the conventional NPMCSs have already realized. The realized NPMCSs have very high reliability. For example, in the case of Great East Japan earthquake (March 11, 2011), although cellular phone systems were unavailable, PMCSs were able to keep their functions. In emergency cases, users of the PMCSs require more stable communication than broadband communication, even if the PMCSs can only provide users with speech communication. Detecting disconnection in backbone lines, a base station of NPMCSs can autonomously deal with many mobile stations located in their extensive area without center control systems owing to large zone scheme. However, BPMCSs employing small or middle zone scheme can hardly realize it.

From these reasons, we consider that both BPMCSs and NPMCSs must be employed in future. We have therefore researched HCR combining BPMCSs, NPMCSs and CWMCSs so that PMCSs can provide users with the best communication quality. In this sub-section, we research the HCR that can make NPMCSs support BPMCSs to improve BPMCS's communication quality. We do not deal with the cognitive radio that can make BPMCSs support NPMCSs because BPMCSs can be replaced with CWMCSs and this HCR has already been considered in chapter 3.

In the sub-section, to improve communication quality of BPMCSs, we propose HCR that can perform site diversity by receiving uplink BPMCS's signals at base stations of NPMCSs employing large zone scheme. Figure 5.4 shows the proposed systems. The reason why we focus on uplink is that uplink is more important than downlink, because BPMCSs are usually used for transmitting video of scene of accidents to the head office. By employing the site diversity, a mobile station of BPMCSs not only can improve its communication quality but also can reduce transmitting power in order to miniaturize the mobile station. A problem of this site diversity is interference of uplink when several mobile stations transmit radio wave in the service area. In this thesis, we propose site diversity based on the HCR that cognizes status of mobile station's transmitting through NPMCSs so that the interference can be avoided by the adaptive array [61]. Moreover, we consider a method that can send information to the head office through narrowband backbone line. The proposed site diversity needs to convey soft decision values of received signals using backbone lines to perform Maximum Rate Combining (MRC) [60] at the head office. However, the backbone lines of NPMCSs are usually narrowband because of employing microwave links, which are suitable for link from base stations located at top of a mountain to

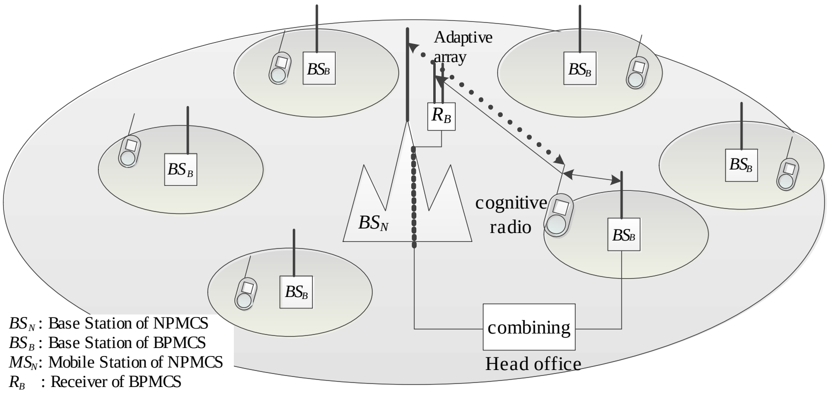

Figure 5.4: HCR that can realize site diversity.

the head office in a city. Hence, since sending of the soft decision values may be difficult through this backbone line, we consider method that compresses the soft decision value without deterioration of the diversity performance. In this thesis, we propose two compression methods. One is employment of logarithmic compression. The other method is to send level of channel state information (CSI) and hard decision values of received signals individually.

This sub-section consists of five sub-sub section as follows. In sub-section 5.2.2, we explain detail of the proposed site diversity based on the HCR. In subsection 5.2.3, the compression methods for the site diversity are proposed. The proposed site diversity system and compression methods are evaluated by computer simulation in sub-section 5.2.4. In sub-section 5.2.5, we summarize the proposed site diversity systems.

5.2.2 Proposed System

We consider improvement of uplink of BPMCSs. This is because uplink, such as transmitting videos of disaster sites to a head office, is more important than downlink in public safety activities. Besides, mobile stations of BPMCSs need to be miniaturized to improve mobility. At present, mobile stations of the PBWCS are large because of long distance transmission (several km in non-line of sight). The proposed HCR plans improvement of uplink communication quality and miniaturization of mobile stations for BPMCSs by employing the site diversity.

5.2.2.1 Realization of Site Diversity

Figure 5.5 shows the structure of the site diversity for improving uplink commutation quality and miniaturizing mobile stations. Uplink signals transmitted by BPMCS's mobile stations (MS_B) are received not only by BPMCS' base station (BS_B) but also by BPMCS' receivers (R_B) located at NPMCS' base stations (BS_N) for the site diversity. The received two signals are sent to the head office using soft decision values and then combined for improving BER performance. Generally, although distance between the HCR and R_B is further than distance between the HCR and BS_B, R_B can receive enough power owing to high antenna height. The detail is described in next sub-sub section.

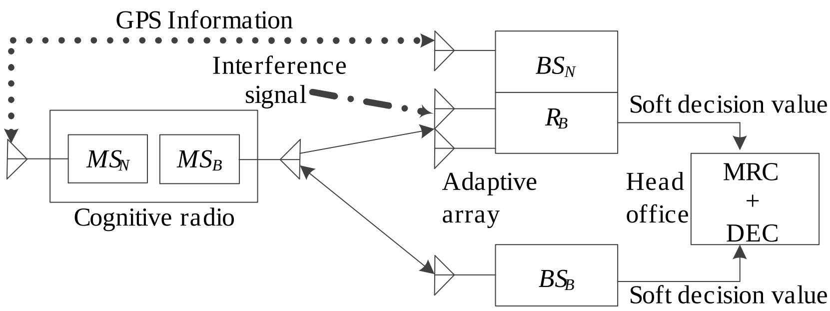

Figure 5.5: System structure when the site diversity is employed.

5.2.2.2 Confirmation of Received Power

Before discussing effect of the site diversity, we estimate average received power of BPMCS's signals received at R_B using Okumura curve [57] [58]. Table 5.2 shows parameters for calculation of the average received power. Since the NPMCS's base station BS_N needs to cover areas whose radius exceeds 30km, the height of NPMCS's antenna is usually higher than 1000 m above sea level. On the other hand, in the case of BPMCSs, antenna height is not high because of small or middle zone systems. At present, since the base stations are temporarily deployed in the PBWCS, most of antenna height is lower than 10m. Figure 5.6 shows average received power calculated by hata model [57] [58], which are mathematical expressions of okumura curve. In the hata model, land type is defined as sub-urban. Transmitted power of BS_B is 37dBm. Moreover, antenna of R_B can get high gain owing to adaptive array described in next sub-sub section.

Table 5.2: Simulation Parameters for calculation of propagation loss

Narrowband PMCS Base Staion R_B	Antenna Hight	500m
	Antenna Gain(adaptive array)	10dBi
Broadband PMCS Base Staion BS_B	Antenna Hight	10m
	Antenna Gain(adaptive array)	7dBi
Broadband PMCS Mabile Staion MS_B	Antenna Hight	1.5m
	Antenna Gain(adaptive array)	3dBi

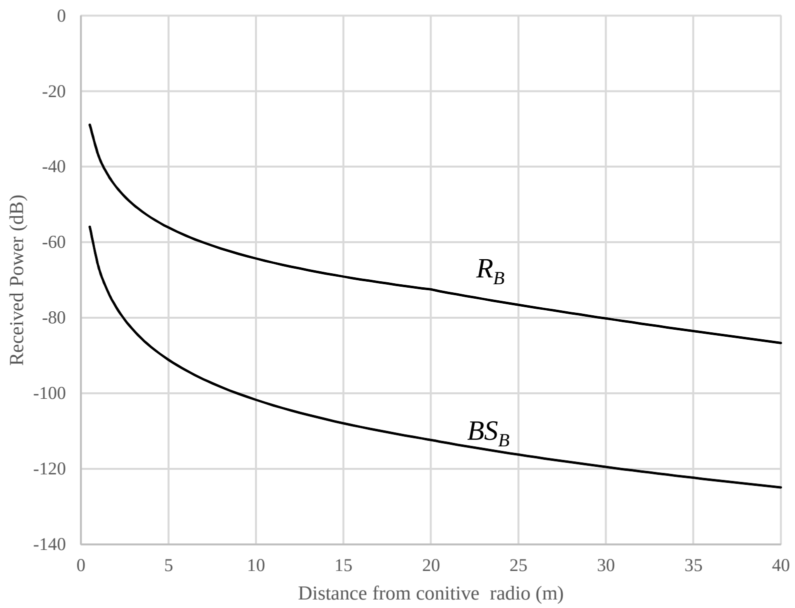

Figure 5.6: Average received power using the hata model.

From Fig. 5.6, since R_B has high antenna gain and employs the adaptive array, R_B can receive enough power despite of long distance. For example, when we define reference average received level as -70 dBm, the distance that communication can be continued within is shorter than 2 km in the case of BS_B. In the case of R_B, this distance can be extended until 15km.

5.2.2.3 Avoidance of Uplink Interference

Realizing the site diversity, we must solve a problem of uplink interference shown in Fig. 5.5, if several BPMCS's mobile station MS_Bs are in the area near the R_B. To avoid uplink interference, we propose the adaptive array in this thesis. The HCR transmits status information of own MS_B transmitting and position calculated by GPS by utilizing own MS_N to the BS_N. The BS_N sends its information to the head office. The head office comprehends all MS_B information of transmitting the status and position. One branch (R_B) of the site diversity acquires this information and then directs beam to desired MS_B and directs null to interference MS_Bs. Moreover, since other controls are required for combining, the required information for these controls is shared by utilizing the NPMCSs. Although adopting the adaptive array in this thesis, we will apply MIMO to our HCR in the future. The cognitive system should separate all uplink signals without interference to improve system performance. In the case of adoption of MIMO, since CSI estimation are required to separate signals, we would like to deal with it as problem in the future.

5.2.2.4 Sending Information through Narrowband Backbone Line

Another problem of the site diversity is to send information through narrowband backbone lines. Since the site diversity of the MRC requires soft decision values before combining, large amount of information is input to backbone lines. However, the backbone lines of BS_N and R_B located at top of mountains are narrowband because of difficulty of laying optical fiber cables. In the case of a base station located at top of a mountain, microwave links, whose transmission rate is around 100 Mbps, are used for backbone lines. Hence, using the microwave links, we must consider whether the soft decision value for the site diversity can be sent or not.

Here, defining transmission rate of a BPMCS as 5 Mbps, we estimate necessary transmission rate. The modulation scheme and coding rate are QPSK and 1/2, respectively. In the case of 8 bit quantization, the transmission rate becomes 80Mbps, which occupies most of bandwidth of the microwave link. Since the microwave links transmits other system's information, to convey the BPMCS's information of 80Mbps is difficult. Hence, to realize the proposed site diversity system, information conveyed through the microwave backbone lines should be compressed. In the next sub-sub section, we consider how to reduce information transmitted through the microwave backbone lines.

5.2.3 Information Compression for Narrowband Backbone Lines

As we mentioned above sub-section, information of soft decision values sent through narrowband backbone lines need to be compressed. First, we consider the compression method that quantizes received signals multiplied by level of CSI to realize MRC. This is because if we send the received signals and the CSI level individually, information input in backbone lines increase. In this quantization method, quantization bits are reduced by compressing the multiplied received signals by employing logarithm. Next, we would like to tackle a compression method that sends received signals and the CSI level individually. The method sends hard decision value of the received signal and compresses the CSI level by employing logarithm and quantization so that amount of information can be reduced.

5.2.3.1 Method Compressing Multiplied Received Signals by Logarithm

First, we consider the method compressing the multiplied received signals by employing logarithm. The method reduces required quantization bits by widening sampling intervals with increasing signal level. We suppose that rough sampling does not make BER performance deteriorate significantly if the signal level is high enough. In the method compressing signals by logarithm, A-Law and μ-Law methods [62] have been proposed for a voice codec. In this thesis, we try to apply the A-Law and μ-Law methods to the CSI multiplied received OFDM signals that are destroyed by frequency selective fading. The compress equations are shown as follows.

$$y_{A-law} = \begin{cases} \dfrac{A|x|/x_{max}}{1+lnA}\text{sgn}(x) & 0 < \dfrac{|x|}{x_{max}} < \dfrac{1}{A} \\ \dfrac{1+lnA|x|/x_{max}}{1+lnA}\text{sgn}(x) & \dfrac{1}{A} < \dfrac{|x|}{x_{max}} < 1. \end{cases} \tag{5.1}$$

$$y_{\mu-law} = \frac{ln(1+\mu|x|/x_{max})}{ln(1+\mu)}\text{sgn}(x). \tag{5.2}$$

In the parameter, x and x_{max} are signals before compression and maximum value of x, respectively. y_{A-Law} and $y_{\mu-Law}$ are compression signals of A-Law and μ-Law methods, respectively. For voice compression, standard values A and μ are 87.56 and 255, respectively.

5.2.3.2 Method Sending Hard Decision Values and CSI Level Individually

In the case of very narrowband backbone lines, further compression may be required. To reduce the sending information drastically, we propose the method whose sending information is hard decision values shown in Fig.5.7. Moreover, to close to MRC performance as much as possible, we propose that the CSI level is send beforehand and multiplied by the hard decision values at the head office. In the case of slow Doppler frequency, CSI does not need to be updated frequently. That means information compression. Moreover, we propose that the CSI level is compressed by A-Law and μ-Law as mentioned in 5.2.3.1.

5.2.4 Computer Simulation

5.2.4.1 Simulation Parameter

Table 5.3 shows simulation parameters for our proposed system. The parameters are set by referring specifications of mobile WiMAX (IEEE802.16e) [63]. In this simulation, we assume that CSI is estimated completely, and Maximum Doppler frequency is almost 0 Hz for using video transmission of an accident area.

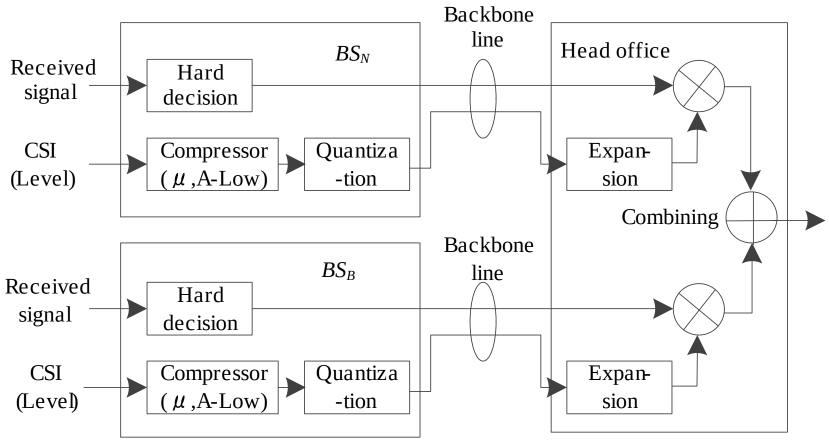

Figure 5.7: The method that sends hard decision values and CSI.

Table 5.3: Simulation Parameters of Mobile MiMAX

Number of FFT Points	512
Symbol Length	102.9 μs
Modulation Method	QPSK
Guard Interval Lenght	11.4 μs
Delay Profile	Typical Urban (6 path) [59]
Maximum Doppler Frequency	≒ 0
Forward Error Correction	Convolutional Code
	Constraint Length K=7
Interleave	Random Interleave (frequency domain)

5.2.4.2 MRC Peformance

Figure 5.8 shows BER performance when site diversity is employed. In Fig. 5.8, selective combining(SC), equal gain combining(EG), maximum radio combining (MRC) are employed. We confirm that diversity can reduce BER greatly and MRC is the best performance of the three.

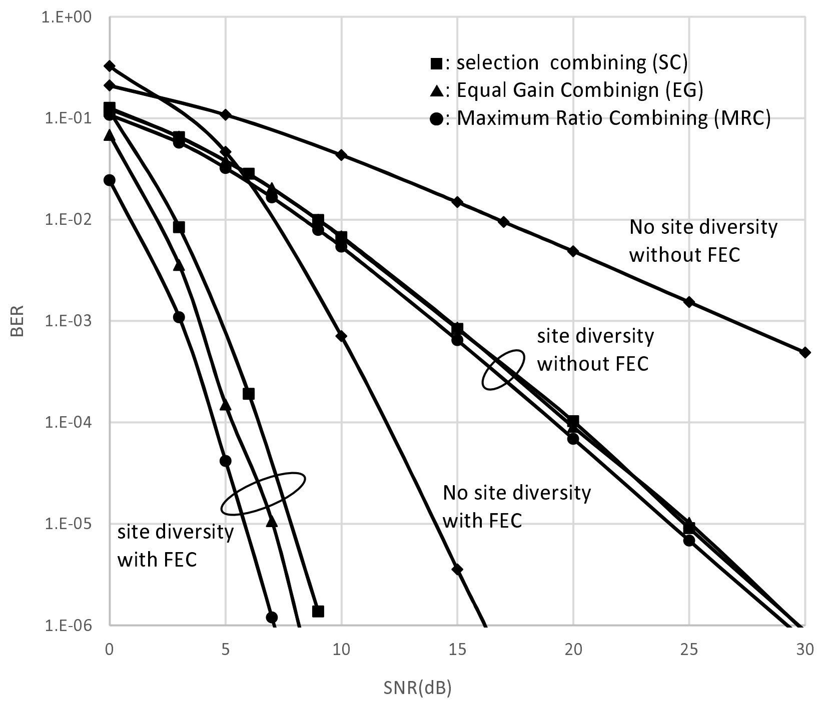

Figure 5.8: MRC performance.

5.2.4.3 Method Compressing Multiplied Received Signals by Logarithm

To reduce amount of data and to realize MRC , demodulated each OFDM subcarrier multiplied by the CSI level is compressed by logarithm. When signals are compressed by the A-Law and μ-Law, maximum values are send with the compressed signals because of normalization. In the head office, the received signals are expanded and then multiplied by the maximum values. Figures 5.9 and 5.10 show BER performances of not using logarithmic compression and of using the A-Law method, respectively. In the case of A-Law, parameter A is 10.

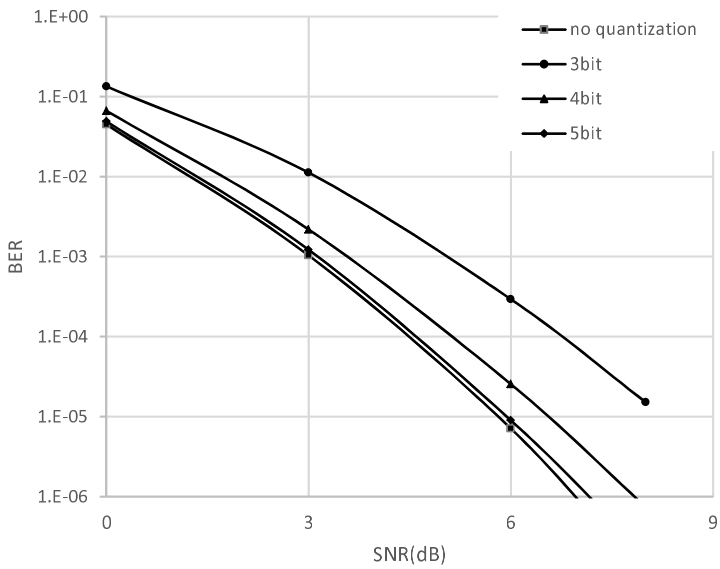

Figure 5.9: Diversity BER performance in quantization without logarithmic compression.

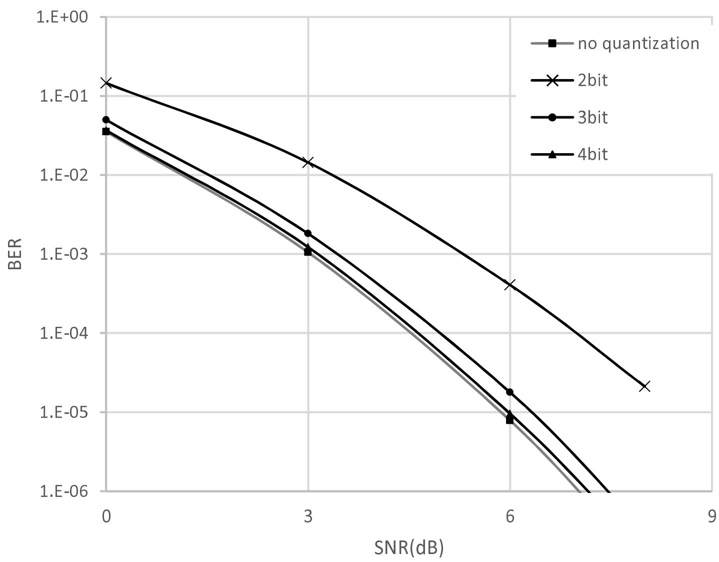

Figure 5.10: Diversity BER performance in quantization with logarithmic compression (A-Law).

When comparing Fig.5.9 with Fig.5.10 by employing A-Law logarithmic compression, quantization bit can be reduced from 5 bit to 4bit. In the simulation, although we simulated BER performance using μ-Law method, difference of A-law and μ-law can hardly be observed. Moreover, we found that BER performances do not depend on fading models. In the parameters A and μ, the best values are around 10 and 100, respectively. However, difference between these values and default values, A =87.56 and μ=255, is lesnn than 0.1dB. Figures 5.11 and 5.12 show BER performances when average received powers of one branch are 3dB and 6dB lower than these of the other, respectively. In this simulation, the signals are also compressed by A-Law(A=10). From Fig. 5.10 to Fig .5.11, we confirm that

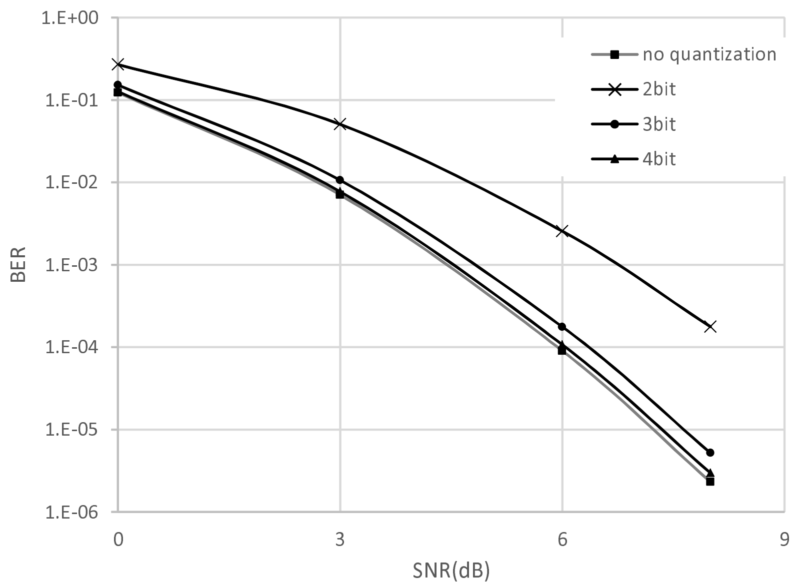

Figure 5.11: Diversity BER performance in the case of 3dB reduction in one branch received Power.

more than 4 bit quantization does not make BER performance deteriorate compared with MRC performance. By normalization and sending maximum values, BER performances are not be deteriorated even if two branches have different received power as shown in Figs. 5.11 and 5.12.

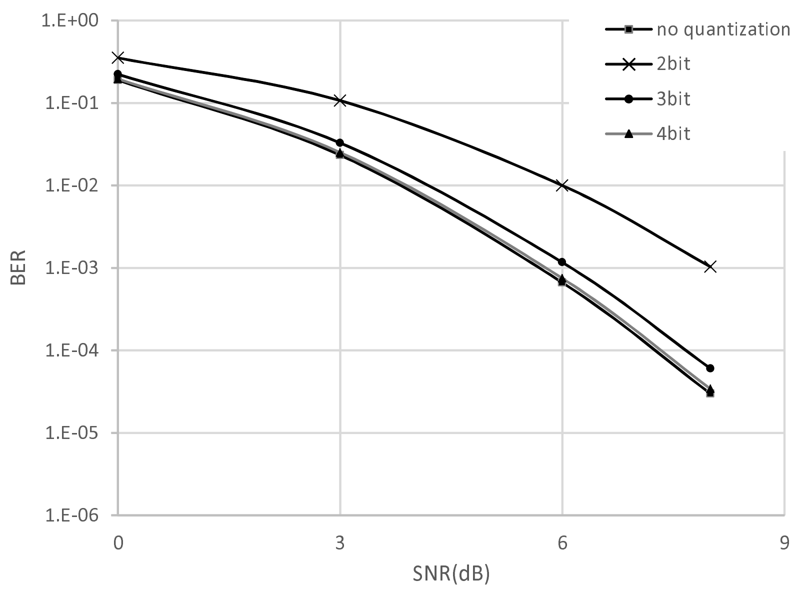

Figure 5.12: Diversity BER performance in the case of 6dB reduction in one branch received Power.

5.2.4.4 Method Sending Hard Decision Values and CSI Level Individually

Next, Fig. 5.13 shows diversity BER performances when hard decision values and CSI level are individually sent. The CSI level is compressed with 3bit and average received power of two branches are same. As references, Fig. 5.13 also shows diversity BER performances when the received signals without multiplied the CSI are quantized with 2 ∼ 4 bits. Deterioration caused by hard decision is around 2 dB. Hence, sending hard decision values is available to prevent large deterioration compared with the MRC performance. Next, Fig. 5.14 shows BER performance when error of CSI estimation happens. Here, SN indicates noise power to CSI power radio.

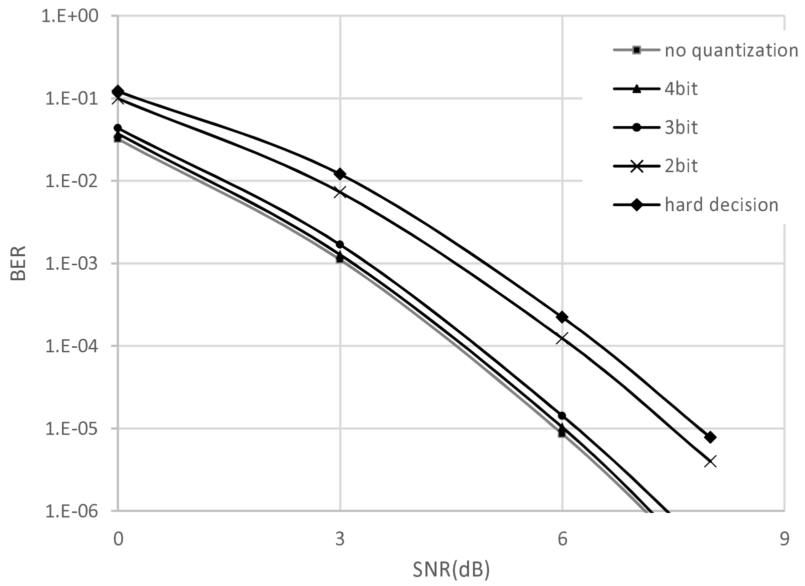

Figure 5.13: Method sending hard decision values and CSI level individually.

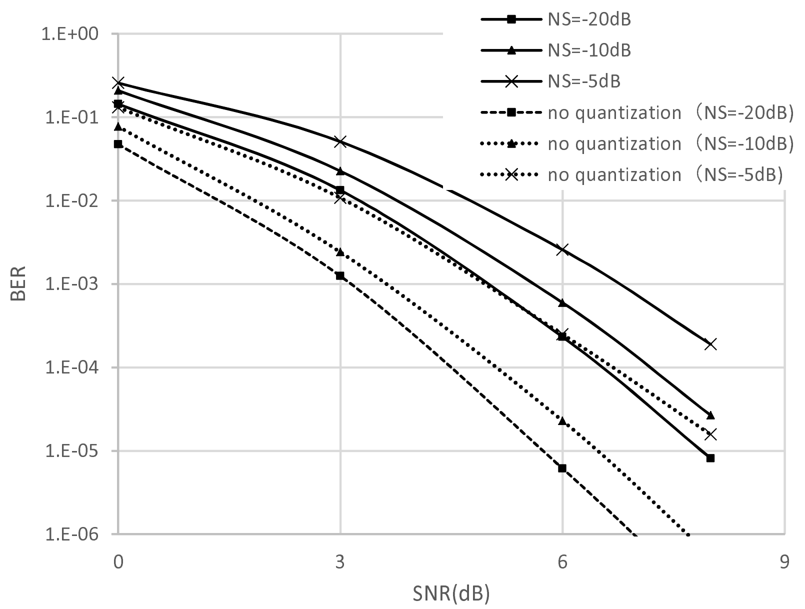

Figure 5.14: Method sending hard decision values and CSI level individually with CSI estimation error.

In the Fig. 5.14, "without quantization" indicates combining received signals with CSI level having noise without compression and quantization. In the other words, Fig. 5.14 shows MRC BER performance when multiplied the CSI level is contaminated by noise. In the noise contamination, difference between the MRC performance and the proposed method is also around 2 dB. Hence, our proposed method can be performed in real systems.

5.2.5 Summary

In this sub-section, we considered combination of BPMCS and NPMCS to improve BPMCS communication quality. To realize uplink improvement of BPMCS, we proposed site diversity based on HCR. Since the problem of the proposed site diversity is uplink interference and narrowband backbone lines, we considered applying the adaptive array and the information compression method, respectively. In the information compression method, we proposed the logarithmic compression by employing A-Law method and μ-Law methods. Moreover, the method that sends hard decision values of the received signal and the CSI level individually. After this, we are going to study theoretical concept of logarithmic compression and updating cycle of the CSI in the moving environment.

Chapter 6

Conclusion

This chapter concludes our research work based on the study of software defined Radio (SDR) and heterogeneous cognitive radio (HCR) combining public safety mobile wireless communication systems (PMCSs) with commercial wireless mobile communication systems (CWMCSs). First, we describe the proposed SDR, HCR, their advantages, and contributions. Secondly, we discussed about the potential future research direction.

6.1 Contribution and Advantages of the Proposed Systems

SDR and HCR are a new trend in wireless communication promising greater flexibility, reliability, and performance over conventional wireless systems. On the other hand, PMCSs are crucial to protect safety and security of communities. We therefore considered introduction of SDR and HCR to PMCSs. By employing SDR and HCR, we expect that PMCSs will be much more reliable. Moreover, owing to handover to multiple systems, wide service areas will be realized. Additionally, in utilization of CWMCSs, we can expect not only wide service areas but also high speed transmission.

In this thesis, firstly, we proposed the HCR for expanding service areas of Narrow-band PMCSs(NPMCSs). The proposed system can improve communication quality (or BER performance) of NPMCSs by obtaining subsidiary information (SI) from CWMCSs when communication quality of NPMCSs becomes poor. We evaluated BER improvement analytically when employing the proposed system. We confirmed that the results are correct by computer simulation.

Next, we researched synchronization of the HCR combining NPMCSs with CWMCSs. In particular, since NPMCSs are utilized in extremely low SNR environments, acquisition of self-synchronization of NPMCSs is probably difficult. Hence, we considered self-synchronization of NPMCSs and proposed two self-synchronization methods. In thesis, we proposed a self-synchronization method employing GPS signals at first. Then, we proposed another self-synchronization method utilizing SI so that the HCR can be used in areas where GPS is unavailable.

We considered SDR that can integrate NPMCSs. By integration of NPMCSs, upgrade will be ease and then higher quality communication will be provided. Moreover, since handover to other systems is available, service areas will be expanded. Although costs of SDR realization are high so far, it will be significantly reduced in the future and then high quality SDR of NPMCSs will be realized.

Finally, since Broad-band PMCSs(BPMCSs) are getting popular, we studied a HCR combining NPMCSs with BPMCSs to enhance usability of the BPMCSs. By employing HCR techniques, we expect that communication quality improvement and miniaturization of BPMCSs are realized. We believe that our proposed SDR and HCR are useful for development of future PMCSs.

6.2 Future Research Work

In this thesis, we studied SDR and HCR for PMCSs. Our proposed systems were researched mainly by computer simulation. In the future research work, we must employ real radio to confirm if the proposed systems works properly. Moreover, although we used convolutional code as forward error correction (FEC), application of turbo code must be considered in future work. Additionally, since the public broadband wireless communication system (PBWCS) and the long term evolution (LTE) for public safety will be developed and improved, we continuously need to research these next generation PMCSs employing SDR and HCR.

In conclusion, CWMCSs will be developed increasingly. To keep pace with CWMCSs, we must research and develop remarkable PMCSs for keeping stable and firm public safety. We surely believe that our work in the thesis is useful for future PMCS's development.

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Publications

List of Publications Directly Related to The Dissertation

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1. Masafumi Moriyama, Takeo Fujii, **"Theoritical Analyses of Viterbi Decoding Employing Heterogeneous Cognitive Radio for Digital Public Private Mobile Radio Systems",** IEICE Transaction On Communication (Japanese), Vol.J97-B no.2, Feb. 2014.

2. Masafumi Moriyama, Takeo Fujii,**"Novel synchronization and BER improvement method for public safety mobile communication systems employing heterogeneous cognitive radio",** IEICE Transaction On Communication. Vol.E98-B no.4, Apr. 2015.

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1. Masafumi Moriyama, Takeo Fujii, **"Viterbi Decoding Employing Heterogeneous Cognitive Radio for Digital Public PMR Systems"**, in Proceedings of SDR'14-WinnComm, Schaumburg Illinois U.S.A, Mar. 2014.

2. Masafumi Moriyama, Takeo Fujii, **"Novel Timing Synchronization Technique for Public Safety Communication Systems Employing Heterogeneous Cognitive Radio "**, in Proceedings of *IEEE International Conference on Computing, Networking and Communications 2015 (ICNC2015)*, Anaheim, California, U.S.A, Feb. 2015.

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1. Masafumi Moriyama, Hiroshi Harada, Seiichi Sampei, Ryuhei Funada, **"A Novel Method of Estimating the Signal-to-Interference Ratio for One-Cell-Frequency-Reuse OF/TDMA Systems",** IEICE Transaction On Communication, Vol.E97-B no.1.Jun. 2008.

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1. Masafumi Moriyama,Hiroshi Harada, Seiichi Sampei,**"A novel desired signal to undesired signal power ratio estimation method in one-sell reuse DPC-OF/TDMA,"** IEICE Technical conference on Communication Syetems (IEICE CS 2004), Okinawa, Japan, IEICE Tech. Rep., vol. 104, no. 595, Jun. 2004.

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