T.C. DOGUS UNIVERSITY

ENGINEERING FACULTY

DEPARTMENT: ELECTRONICS AND COMMUNICATOIN

CHANNEL CODING AND MODELING

IN WIRELESS SYSTEMS

GRADUATION PROJECT
492

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200134017

INSTRUCTOR
Prof. Dr. LEVENT SEVGİ

ISTANBUL, 05, 2006

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1. Applications and Requirements of Wireless Services

1.1 Introduction

Wireless communications is one of the big engineering success stories of the last 20 years – not only from a scientific point of view, where the progress has been phenomenal, but also in terms of market size and impact on society. Companies that were completely unknown 20 years ago are now household names all over the world, due to their wireless products, and in several countries the wireless industry is dominating the whole economy. Working habits, and even more generally the ways we all communicate, have been changed by the possibility of talking ‘‘anywhere, anytime’’.

For a long time, wireless communications has been associated with cellular telephony, as theses the biggest market segment, and has had the highest impact on everyday lives. In recent times, wireless computer networks have also led to a significant change in working habits and mobility of workers – answering emails in a coffee shop has become an everyday occurrence. But besides these widely publicized cases, also a large number of less obvious applications have been developed, and are starting to change our lives. Wireless sensor networks monitor factories; wireless links replace the cables between computers and keyboards; and wireless positioning systems monitor the location of trucks that have goods identified by wireless RF (Radio Frequency) tags. This variety of new applications causes the technical challenges for the wireless engineers to become bigger with every day. This book aims to give an overview of the solution methods for current, as well as future challenges.

Quite generally, there are two paths to developing new technical solutions: engineering-driven, and market-driven. In the first case, the engineers come up with a brilliant scientific idea –without having an immediate application in mind. As time progresses, the market finds applications enabled by this idea.1 In the other approach, the market demands a specific product, and the engineers try to develop a technical solution that fulfills this demand. In this first chapter, we will describe these market demands. We start out with a brief history of wireless communications, in order to convey a feeling of how the science, as well as the market, has developed in the past 100 years, then follows a description of the types of services that constitute the majority of the wireless market today. Each of these services makes specific demands in terms of data rate, range, number of users, energy consumption mobility and so on .We will discuss all these aspects in Section 1.3.We wrap up this section with a description of the interaction between the engineering of wireless devices and the behavioral changes in society induced by them.

1.1 History

When looking at the history of communications, we find that wireless communications is actually the oldest form – shouts and jungle drums did not require any wires or cables to function .Even the oldest ‘‘electromagnetic’’ (optical) communications are wireless: smoke signals are based on propagation of optical signals along a line-of-sight connection However , wireless communications as we know it started only with the work of Maxwell and Hertz, who laid the basis for our understanding of the transmission of electromagnetic waves .It was not long after their groundbreaking work that Tesla demonstrated the transmission of information via these waves – in essence, the first wireless communications system. In 1898, Marconi made his well-publicized demonstration of wireless communications from a boat to the Isle of Wight in the English Channel .It is noteworthy that while Tesla was the first to succeed in this important endeavour , Marconi had the better public relations, and is widely cited as the inventor of wireless communications, receiving a Nobel prize in 1909.

In the subsequent years, radio (and later television) became widespread throughout the world. While in the ‘‘normal’’ language we usually do not think of radio or TV as ‘‘wireless communications’’, they certainly are in a scientific sense – information transmission from one place to the other by means of electromagnetic waves. They can even constitute ‘‘mobile communications’’, as evidenced by car radios. A lot of basic research – especially concerning wireless propagation channels – was done for entertainment broadcasting .By the late 1930 s, a wide network of wireless information transmission – though unidirectional – was in place.

1.3 The First Systems

At the same time, the need for bi-directional mobile communications emerged. Police departments and the military had obvious applications for such two-way communications and were the first to use wireless systems with closed user groups. Military applications drove a lot of the research during, and shortly after, the Second World War. This was also the time when much of the theoretical foundations for communications in general were laid. Claude Shannon’s [1948] groundbreaking work ‘‘A mathematical theory of communications’’ appeared during that time, and established the possibility of error-free transmission under restrictions for the data rate and the signal-to-noise ratio. Some of the suggestions in that work, like the use of optimum power assignment in frequency-selective channels, are only now being introduced into wireless systems. [2]

The 1940s and 1950s saw several important developments: CB (Citizens’ Band) radios became widespread, establishing a new way of communicating between cars on the road. Communicating with these systems was useful for transferring vital traffic information and related aspects within the closed community of the drivers owning such devices, but it lacked an interface to the public telephone system, and the range was limited to some 100 kilometers, depending on the power of the (mobile) transmitters. In 1946, the first mobile telephone system was installed in the U.S.A. (St. Louis). This system did have an interface to the PSTN (Public Switched Telephone Network), the landline phone system, though this interface was not automated, but rather consisted of human telephone operators. However, with a total of six speech channels for the whole city, the system soon met its limits. This motivated investigations of how the number of users could be increased, even though the allocated spectrum would remain limited. Researchers at AT&T's Bell Labs found the answer: the cellular principle, where the geographical area is divided into cells; different cells might use the same frequencies.

To this day, this principle forms the basis for the majority of wireless communications. Despite the theoretical breakthrough, cellular telephony did not experience significant growth during the 1960s.However, there were exciting developments on a different front: in 1957, the Soviet Union had launched the first satellite (Sputnik), and the U.S.A. soon followed. This development fostered research in the new area of satellite communications. Many basic questions had to be solved, including the effects of propagation through the atmosphere, the impact of solar storms, the design of solar panels and other long-lasting energy sources for the satellites, and so on. To this day, satellite communications is an important area of wireless communications (though not one that we will address specifically in this book).The most widespread application lies in satellite TV transmission.

1.4 Analog Cellular Systems

The 1970s saw a revived interest in cellular communications. In scientific research, these years saw the formulation of models for path loss, Doppler spectra, fading statistics, and other quantities that determine the performance of analog telephone systems. A highlight of that work was Jakes' book Microwave Mobile Radio, that summed up the state of the art in this area [Jakes 1974].The 1960s and 1970s also saw a lot of basic research that was originally intended for landline communications, but later also proved to be instrumental for wireless communications. For example, the basics of adaptive equalizers, as well as multi carrier communications, were developed during that time. [2]

For the practical use of wireless telephony, the progress in device miniaturization made the vision of ''portable'' devices more realistic. Companies like Motorola and AT&T vied for the leadership in this area, and made vital contributions. NTT (Nippon Telephone and Telegraph) established a commercial cellphone system in Tokyo in 1979.However, it was a Swedish company that built up the first system with large coverage and automated switching: up to that point, Ericsson AB had been mostly known for telephone switches, while radio communications was of limited interest to them. However, it was just that expertise in switching technology, and the (for that time daring) decision to use digital switching technology that allowed them to combine the different cells in a large area into a single network, and establish the Nordic Mobile Telephone (NMT) system [Meurling and Jeans 1994].Note that while the switching technology was digital, the radio transmission technology was still analog, and the systems became therefore known as analog systems.

Subsequently, other countries developed their own analog phone standards. The system in the U.S.A., for example, was called AMPS (Advanced Mobile Phone System).

An investigation of NMT also established an interesting method for estimating market size: business consultants equated the possible number of mobile phone users with the number of Mercedes 600 (the top-of-the-line luxury car at that time) in Sweden. Obviously, mobile telephony could never become a mass market, could it? Similar thoughts must have occurred to the management of the inventor of cellular telephony, AT&T. Upon advice from a consulting company, they decided that mobile telephony could never attract a significant number of participants, and stopped business activities in cellular communications.

The analog systems paved the way for the wireless revolution. During the 1980s, they grew at a frenetic pace, and reached market penetrations of up to 10% in Europe, though their impact was somewhat less in the U.S.A. In the beginning of the 1980s, the phones were ‘‘portable’’, but definitely not handheld. In most languages, they were just called ‘‘carphones ’’, because the battery and transmitter were stored in the trunk of the car, and were too heavy to be carried around. But at the end of the 1980s, handheld phones with good speech quality and quite acceptable battery lifetime abounded. The quality had become so good that in some markets, digital phones had difficulties to establish themselves – there just did not seem to be a need for further improvements.

1.5 GSM and the Worldwide Cellular Revolution

Even though the public did not see a need for changing from analog to digital, the network operators knew better. Analog phones have a bad spectral efficiency why, and due to the rapid growth of the cellular market, operators had a high interest in making room for more customers. Also, research in communications had started its inexorable turn to digital communications, and that included digital wireless communications as well. In the late 1970s and the 1980s, research into spectrally efficient modulation formats, the impact of channel distortion and temporal variations on digital signals, as well as multiple access schemes and much more was explored in research labs throughout the world. It thus became clear to the cognoscenti that the real-world systems would soon follow the research.

Again, it was Europe that led the way. The European Telecommunications Standards Institute (ETSI) group started the development of a digital cellular standard that would become mandatory throughout Europe, and was later adopted in most parts of the world: GSM (Global System for Mobile communications).The system was developed throughout the 1980s; deployment started in the early 1990s, and user acceptance was swift. Due to additional features, better speech quality, and the possibility for secure communications, GSM-based services overtook analog services typically within 2 years of their introduction. In the U.S.A., the change to digital systems was somewhat slower, but by the end of the 1990s, also this country was overwhelming digital.

Digital phones turned cellular communications, which was already on the road to success, into a blockbuster .In the year 2004, market penetration in Western Europe exceeded 80%, with some Scandinavian countries approaching the 100% mark (many people have two or three cellphones). Also the U.S.A. is exceeding the 50% mark, and Japan has about 70%. In absolute numbers, China has become the single biggest market, with some 300 million subscribers in 2004.

The development of wireless systems also made clear the necessity of standards. Devices can only communicate if they are compatible, and each receiver can ‘‘understand’’ each transmitter – i.e., if they follow the same standard. But how should these standards be set? Different countries developed different approaches. The approach in the U.S.A. is ‘‘hands-off’’: allow a wide variety of standards, and let the market establish the winner (or several winners).When frequencies for digital cellular communications were auctioned off in the 1990s, the buyers of the spectrum licenses could choose the system standard they would use. For this reason, three different standards are now being used in the U.S.A. A similar approach was used by Japan, where two different systems fought for the market of Second Generation (2G) cellular systems. In both Japan and the U.S.A., the networks based on different standards work in the same geographical regions, allowing consumers to choose between different technical standards.

The situation was different in Europe. When digital communications were introduced, usually only one operator per country (typically, the incumbent public telephone operators) existed. If each of these operators would adopt a different standard, the result would be high market fragmentation (i.e., a small market for each standard), without the benefit of competition between the operators. Furthermore, roaming from country to country, which for obvious geographical regions is much more frequent in Europe than in the U.S.A. or Japan, would be impossible .It was thus logical to establish a single common standard for all of Europe. This decision proved to be beneficial for wireless communications in general, as it provided the economy of scales that decreased cost, and thus increased the popularity, of the new services.

1.6 New Wireless Systems and the Burst of the Bubble

Though cellular communications defined the picture of wireless in the general population, a whole range of new services was introduced in the 1990s.Cordless telephones started to replace the ‘‘normal’’ telephones in many homes. The first versions of these phones used analog technology; however, also for this application, digital technology proved to be superior. Among other aspects, the possibility of listening in to analog conversations, and the possibility for neighbors to ‘‘highjack’’ an analog cordless base station and make calls at other people’s expense, led to a shift to digital communications. While cordless phones never achieved the spectacular market size of cellphones , they constitute a solid market.

Another market that seemed to have great promise in the 1990s was fixed wireless access and Wireless Local Loop (WLL) – in other words, replacing the copper lines to the homes of the users by wireless links, but without the specific benefit of mobility. A number of technical solutions were developed, but all of them ultimately failed. The reasons were as much economical and political as they were technical .The original motivation for WLL was to give access to customers for alternative providers of phone services, bypassing the copper lines that belonged to the incumbents. However, regulators throughout the world ruled in the mid-1990s that the incumbents have to lease their lines to the alternative providers, often at favorable prices. This eliminated much of the economic basis for WLL. Similarly, fixed wireless access was touted as the scheme to provide broadband data access at competitive prices. However, the price war between DSL (Digital Subscriber Line) technology and cable TV has greatly dimmed the economic attractiveness of this approach. The biggest treasure thus seemed to lie in a further development of cellular systems, establishing the ‘‘Third Generation (3G).2G systems were essentially pure voice transmission systems (though some simple data services, like the Short Message Service – SMS – were included as well).The new systems were to provide data transmission at rates comparable with the ill-fated ISDN (Integrated Services Digital Network) (144 kbit/s), and even up to 2 Mbit/s, at speeds of up to 500 km/h. After long deliberations, two standards were established: 3GPP (Third Generation Partnership Project) (supported by Europe, Japan, and some American companies), and 3GPP2 (supported by another faction of American companies).The new standards also required a new spectrum allocation in most of the world, and the selling of this spectrum became one of the big bonanzas for the treasuries of several countries. The development of 3GPP, and the earlier introduction of the IS-95 CDMA (Code Division Multiple Access) system in the U.S.A., sparked a lot of research into CDMA and other spread spectrum techniques for wireless communications; by the end of the decade, multicarrier techniques had also gained a strong footing in the research community. Multiuser detection – i.e., the fact that the effect of interference can be greatly mitigated by exploiting its structure – was another area that many researchers concentrated on, particularly in the early 1990s.Finally, the field of multi antenna systems saw an enormous growth since 1995, and for some time accounted for almost half of all published research in the area of the physical layer design of wireless communications.

The spectrum sales for 3G cellular systems, and the IPOs (Initial Public Offerings) of some wireless startup companies represented the peak of the ‘‘telecom bubble’’ of the 1990s.In 2000/ 2001, the market collapsed with a spectacular crash. Developments on many of the new wireless systems (like fixed wireless) were stopped as their proponents went bankrupt, while the deployment of other systems, including 3G cellular systems, was slowed down considerably. Most worrisome, many companies slowed or completely stopped research, and the general economic malaise led to decreased funding of academic research as well.

1.7 Wireless Revival

Since 2003, several developments have led to a renewed interest in wireless communications .The first one is a continued growth of cellular communications, stimulated by new markets (China, India), new applications (camera phones), and the slow but steady takeoff of 3G systems (where Japan is leading the way).Secondly, wireless computer networks (wireless Local Area Networks– LANs) have become an unexpected success. Devices following the IEEE (Institute of Electrical and Electronics Engineers) 802.11 standards have enabled computers to be used in a way that is almost as versatile and mobile as cellphones. The standardization process had already started in the mid-1990s, but it took several versions and the impact of intense competition from manufacturers, to turn this into a mass product. Thirdly, wireless sensor networks offer new possibilities of monitoring and controlling factories and even homes from remote sites. Of course, the renewed interest in military and security applications is also creating interest in new wireless products.

In general, the ‘‘wireless revival’’ is based on three tendencies: a much broader range of products data transmission with a higher rate for already existing products, and higher user densities. These trends determine the directions of research in the field, and provide a motivation for many of the more recent scientific developments.

2. Requirements for the Services [2]

A key to understanding wireless design is to realize that different applications have different requirements in terms of data rate, range, mobility, energy consumption, and so on. It is not necessary to design a system that can sustain Gigabit/s data rates over a 100-km range when the user is moving at 500 km/h. We stress this fact because there is a tendency among engineers to design a system that ‘‘does everything but wash the dishes’’; while appealing from a scientific point of view, such systems tend to have a high price, and low spectral efficiency .In the following, we list the range of requirements encountered in system design, and we enumerate which requirements occur in which applications.

2.1 Data Rate

Data rates for wireless services span the gamut from a few bits per second to several Gigabit/s, depending on the application: Sensor networks usually require data rates from a few bits per second to about 1 kbit/s.

Typically, a sensor measures some critical parameter, like temperature, speed, etc., and transmits the current value (which corresponds to just a few bits) at intervals that can range from milliseconds to several hours .Higher data rates are often required for the central nodes of sensor networks that collect the information from a large number o sensors, and forward it for further processing. In that case, data rates of up to 10 Mbit/s can be required. These ‘‘central nodes’’ show more similarity to WLANs or fixed wireless access.

Speech communications usually require between 5kbit/s and 64kbit/s depending on the required quality and the amount of compression. For cellular systems, which require higher spectral efficiency, source data rates about 10kbit/s are standard. For cordless systems, less elaborate compression, and therefore higher data rates (32kbit/s) are used.

Elementary data services require between 10 and 100 kbit/s. One category of these services uses the display of the cellphone to provide Internet-like information. Since the displays are smaller, the required data rates are often smaller than for conventional Internet applications. Another type of data service provides a wireless mobile connection to laptop computers. In this case, speeds that are at least comparable with dial-up (around 50kbit/s) are demanded by most users, though elementary services with 10kbit/s (exploiting the same type of communications channels foreseen for speech) are sometimes used as well.
.Communications between computer peripherals and similar devices: for the replacement of cables that link computer peripherals, like mouse and keyboard, to the computer (or similarly for cellphones), wireless links with data rates around 1 Mbit/s are used. The functionality of these links is similar to the previously popular infrared links, but usually provides higher reliability.

High-speed data services: WLANs and 3G cellular systems are used to provide fast Internet access, with speeds that range from 0.5 Mbit/s to 100 Mbit/s (currently under development). Personal Area Networks (PANs) is a newly coined term that refers mostly to the range of a wireless network (up to 10 m), but often also has the connotation of high data rates (over 100 Mbit/s), mostly for linking the components of consumer entertainment systems streaming video from computer or DVD player to a TV) or high-speed computer connections (wireless USB).

2.2 Mobility

Wireless systems also differ in the amount of mobility that they have to allow for the users. The ability to move around while communicating is one of the main charms of wireless communication for the user. Still, within that requirement of mobility, different grades exist:Fixed devices are placed only once, and after that time communicate with their BS, or each other, always from the same location. The main motivation for using wireless transmission techniques for such devices lies in avoiding the laying of cables. Even though the devices are not mobile, the propagation channel they transmit over can change with time: both due to people walking by, and due to changes in the environment (rearranging of machinery, furniture, etc.). Fixed wireless access is a typical case in point. Note also that all wired communications (e.g., the PSTN) fall into this category.
Nomadic devices: nomadic devices are placed at a certain location for a limited duration of time (minutes to hours) and then moved to a different location. This means that during one ‘‘drop’’ (placing of the device), the device is similar to a fixed device. However, from one drop to the next, the environment can change radically. Laptops are typical examples: people do not operate their laptops while walking around, but place them on a desk to work with them. Minutes or hours later, they might bring them to a different location, and operate them there.

Low mobility: many communications devices are operated at pedestrian speeds. Cordless phones, as well as cellphones operated by walking human users are typical examples. The effect of the low mobility is a channel that changes rather slowly, and – in a system with multiple BSs – handover from one cell to another is a rare event.

High mobility usually describes speed ranges from about 30 km/h to 150 km/h. Cellphones operated by people in moving cars are one typical example:Extremely high mobility is represented by high-speed trains and planes, which covers speeds between 300 km/h and 1,000 km/h. These speeds pose unique challenges both for the design of the physical layer and for the handover between cells.

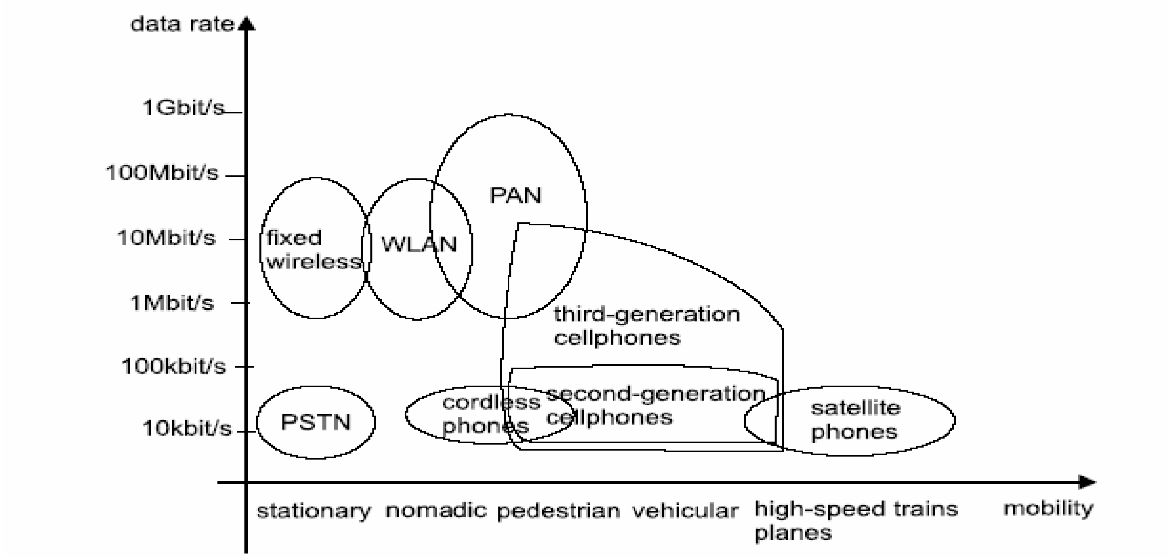

Figure 2.1 Data rate versus range for various applications [2]

2.3 Energy Consumption

Energy consumption is a critical aspect for wireless devices. Most wireless devices use batteries, as they should be free of any wires – both the ones used for communication, and the ones providing the power supply.

Rechargeable batteries: nomadic and mobile devices, like laptops, cellphones, and cordless phones, are usually operated with rechargeable batteries. Standby times as well as operating times are one of the determining factors for customer satisfaction. Energy consumption is determined on one hand by the distance over which the data have to be transmitted (remember that a minimum Signal-to-Noise Ratio – SNR – has to be maintained), and on the other hand by the amount of data that are to be transmitted (the SNR is proportional to the energy per bit).The energy density of batteries has increased slowly over the past 100 years, so that the main improvements in terms of operating and standby time stem from reduced energy consumption of the devices. For cellphones , talk times of more than 2 hours, and standby times of more than 48 hours are minimum requirements. For laptops, power consumption is not mainly determined by the wireless transmitter, but rather by other factors like hard drive usage and processor speed.

One-way batteries: sensor network nodes often use one-way batteries, which offer higher energy density at lower prices. Furthermore, changing the battery is often not an option; rather the sensor including the battery and the wireless transceiver is often discarded after the battery has run out. It is obvious that in this case energy-efficient operation is even more important than for devices with rechargeable batteries.

Power mains: BSs and other fixed devices can be connected to the power mains. Therefore, energy efficiency is not a major concern for them.It is thus desirable – if possible – to shift as much functionality (and thus energy consumption) from the MS to the BS.

User requirements concerning batteries are also important sales issues, especially in the market for cellular handsets. The weight of an MS is determined mostly (70–80%) by the battery. Weight and size of a handset are critical sales issues .It was in the mid-1980s that cellphones were commonly called ‘‘carphones’’, because the MS could only be transported in the trunk of a car, and was powered by the car battery. By the end of the 1980s, the weight and dimensions of the batteries had decreased to about 2 kg, so that it could be carried by the user in a backpack. By the year 2000, the battery weight had decreased to about 200 gr. Part of this improvement stems from more efficient battery technology, but to a large part, it is caused by the decrease of the power consumption of the handsets. Nomadic Figure 2.1 Data rate versus mobility for various applications. 20 Applications and requirements of wireless services

2.4 Direction of Transmission

Not all wireless services need to convey information in both directions: Simplex systems send the information only on one direction – e.g., broadcast systems and pagers. Semi-duplex systems can transmit information in both directions. However, only one direction is allowed at any time .Walkie-talkies, which require the user to push a button in order to talk, are a typical example. Note that one user must signify (e.g., by using the word ‘‘over’’) that (s) he has finished his/her transmission; then the other user knows that now he can transmit.
Full duplex systems allow simultaneous transmission in both directions – e.g., cellphones and cordless phones.

Asymmetric duplex systems: for data transmission, we often find that the required data rate in one direction (usually the downlink) is higher than in the other direction. However, even in this case, full duplex capability is maintained.

2.5 Service Quality

The requirements for service quality also differ vastly for different wireless services. A first main indicator for service quality is speech quality for speech services and file transfer speed for data services. Speech quality is usually measured by the Mean Opinion Score (MOS).It represents the average of a large number of (subjective) human judgments about the quality of received speech (see also Chapter 15).The speed of data transmission is simply measured in bit/s – obviously, a higher speed is better.

An even more important factor is the availability of a service. For cellphones and other speech services, the service quality is often computed as the complement of ‘‘fraction of blocked calls7 plus 10 times the fraction of dropped calls’’. This formula takes into account that the dropping of an active call is more annoying to the user than the inability to make a call at all. For cellular systems in Europe and Japan, this service quality measure usually exceeds 95%; in the U.S.A., the rate is closer to 50%.

For emergency services and military applications, service quality is better measured as the complement of ‘‘fraction of blocked calls plus fraction of dropped calls’’. In emergency situations, the inability to make a call is as annoying as the situation of having a call interrupted. Also, the systems must be planned in a much more robust way, as service qualities better than 99% are required.

A related aspect is also the admissible delay (latency) of the communication. For voice communications, the delay between the time when one person speaks, and the other hears the message, must not be larger than about 100 ms. For streaming video and music, delays can be larger, as buffering of the streams (up to several tens of seconds) is deemed acceptable by most users. In both voice and streaming video communications, it is important that the data transmitted first are also the ones made available to the receiving user first. For data files, the acceptable delays can be usually larger, and the sequence with which the data arrive at the receiver is not critical (for example, when downloading email from a server, it is not important whether the first or the seventh of the emails is the first to arrive).However, there are some data applications where small latency is vital – e.g., for control applications, security and safety monitoring, etc.

3. The Digital Modulation Techniques

3.1 FSK

Frequency-shift keying (FSK) is a form of frequency modulation in which the modulating signal shifts the output frequency between predetermined values. Usually, the instantaneous frequency is shifted between two discrete values termed the mark frequency and the space frequency. FSK is the simple and low performed modulation technique. This is the type of constant phase angle modulation like a classical frequency modulation. The difference between this types, modulated signal has the continuous wave form, but in FSK modulated signal has the two levels changed by wanted voltages. [3]

FSK Transmitter is the signal output which modulating the signal by the voltage levels. During the design of the FSK transmitter, the band with is the most important requirement. Because, the band with must be narrow not to include unwanted interference. Actually, FSK is a kind of frequency modulation and this modulation refers the bits sinusoidal waves include different frequencies. In addition, FSK modulator is a kind of FM transmitter. It includes voltage controlled oscillator. And this modulated according to zero and ones.

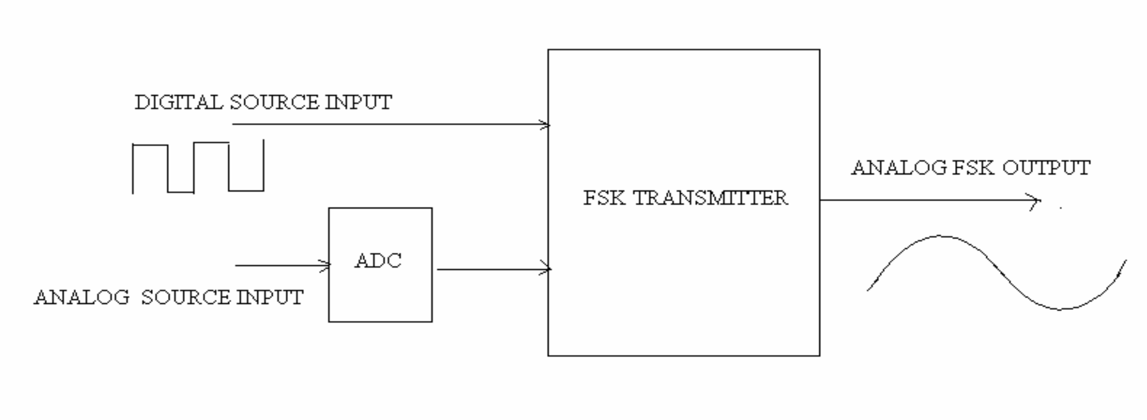

Figure 3.1 FSK transmitters [1]

FSK Receiver is the demodulator circuit signal received form the transmitter. The most popular receiver circuit is PLL – FSK demodulator. When the input of the PLL shift the wave and interval frequencies, at the output of the Dc error voltage follows the DC error voltage. There are two output frequencies. That’s why, one frequency refers the 1, and the other frequency refers ‘0’ digital bits at the output. That’s why, output is sported by the two binary bits.(1,0).

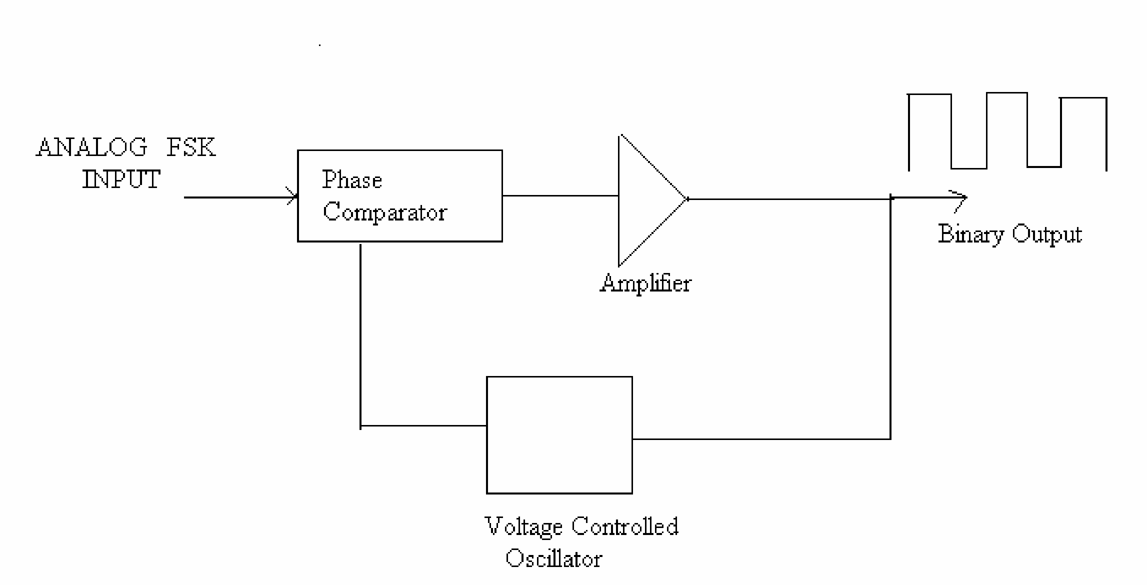

Figure 3.2 FSK reciever [3]

3.2 PSK

Phase-shift keying (PSK) is a method of digital communication including the phase variation of a transmitted signal. There are several types of PSK modulation. The simplest PSK technique is called binary phase-shift keying (BPSK). It uses two opposite signal phases (0 and 180 degrees). The digital signal is broken up time wise into individual bits (binary digits). The state of each bit is determined by the state of the sending bit. If the phase of the wave does not change, the bit (1,0) does not change according to phase differences . And the output stay as a same bit 1 or 0.If the phase of the wave changes by 180 degrees or '0' degrees according to bit changing , bits changes from 0 to 1, or from 1 to 0. This called BPSK modulation.[3]

More sophisticated forms of PSK exist. In m-array or multiple phase-shift keying (MPSK), there are more than two phases, usually four (0, +90, -90, and 180 degrees) or eight (0, +45, -45, +90, -90, +135, -135, and 180 degrees). If there are four phases (m = 4), the MPSK mode is called quadrate phase-shift keying or quaternary phase-shift keying (QPSK), and each phase shift represents two signal elements. If there are eight phases (m = 8), the MPSK mode is known as octal phase-shift keying (OPSK), and each phase shift represents three signal elements. In MPSK, data can be transmitted at a faster rate, relative to the number of phase changes per unit time, than is the case in BPSK.

Amateur-radio operators use a special form of BPSK or QPSK known as PSK31. In this mode, the data transmission rate is 31.25 baud (state changes per second), and the signal bandwidth is approximately 31 Hz. The main advantage of PSK31 is its excellent signal-to-noise ratio (S/N or SNR), which allows communication under adverse conditions such as severe fading, noise, or interference where other communications modes fail.

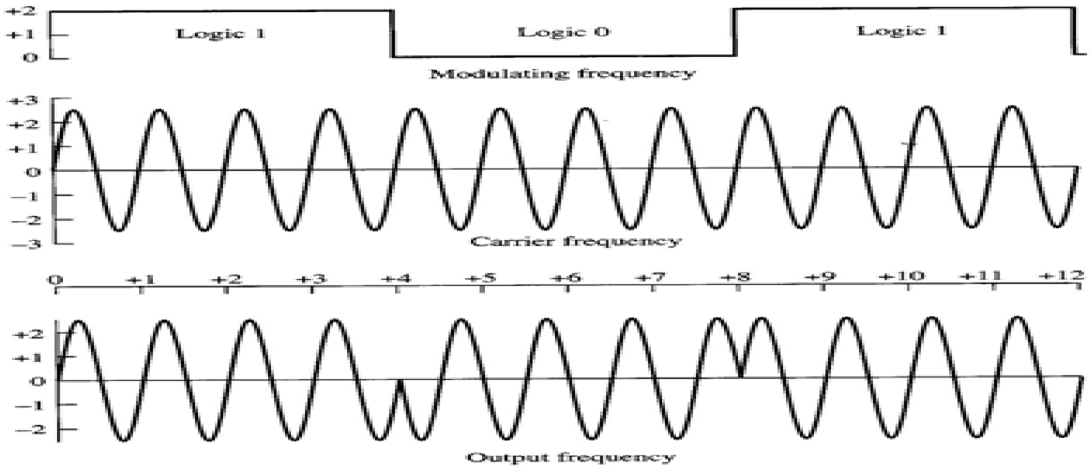

Figure 3.3 PSK Modulation

3.3 FDMA

Frequency Division Multiple Access (FDMA) is a kind of analog multiplexing technique. In this system, data is the analog during the transmission. For example, every station has the 0 – 5 kHz bandwith. If the every data of the station is transmitted by the one frequency interval, the detection of the data will be the impossible. And data on the every station is modulated by the carriers include the different frequencies. These frequencies are different with 10 kHz band with. That's why, data can be transmitted successfully. With this process, we can take the signal successfully and communicate the channels.

This modulation technique is used to for FM broadcast, TV broadcast and high voice telecommunications systems. For all commercial broadcast, the broadcast of the stations are independent. That's why, all station broadcast is independent from others and multiplexing doesn't effect the communication of the stations badly.

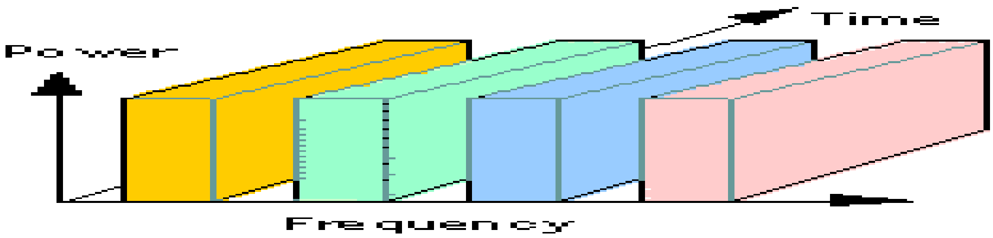

Figure 3.4 FDMA [1]

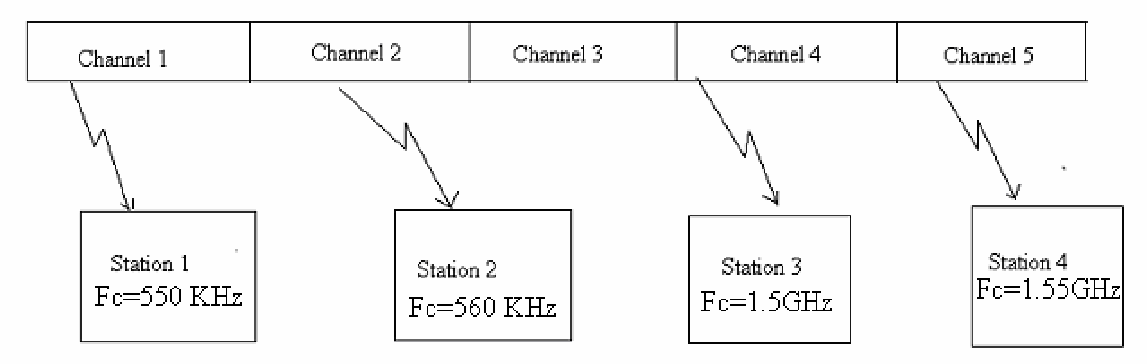

Figure 3.5 Channel locations in FDMA Systems

3.4 TDMA

TDMA is the multiplexing of the data came from a lot of sources. And this multiplexing is realized at the different time intervals and received data is returned at the same time domain. The most wide speared modulation type is PCM –TDM. At this system , a lot of voice band is multiplexing and sampling. They are returned PCM codes and multiplexed.

As you see figure (2.a), there is block diagram of the two channel PCM-TDM carrier system. Every channel is sampled by the time interval and returned to PCM code. Firstly, PCM-CODE of the first channel is transmitted , at the same time, second channel PCM-CODE is sampled. And PCM-CODE of the second channel is transmitted, at the same time, first channel PCM-CODE is sampled. This process continuous and samples are taken from the channels. PCM-CODE is constructed by this process .The detection and demodulation is realized by the PCM-CODES.

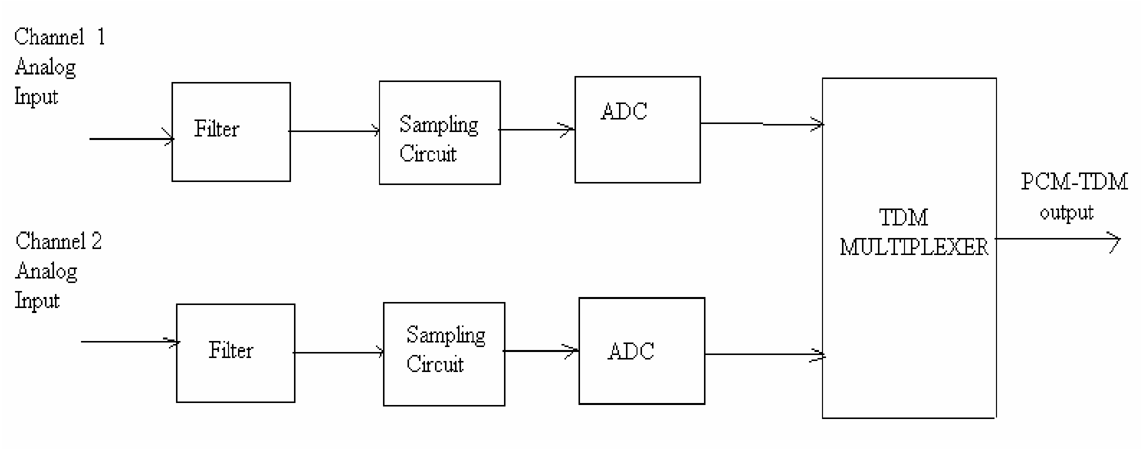

Figure 3.6 TDMA system constructions [3]

4. Channel Coding

4.1 Channel Models for Channel Coding [5]

Figure 4.1 presents a few basic versions of the channel models useful for analysis of the channel coiling process. The simplest model version is known as a *binary symmetric memoryless channel* model (Figure 4.1). The channel inputs and outputs are binary. The transmitted and received blocks are observed at the input and output of the channel model on the bit-after-bit basis. Every bit of the encoded sequence appears unchanged at the channel output with the probability. With the probability p the transmitted bits are negated, which is equivalent to the bit errors.

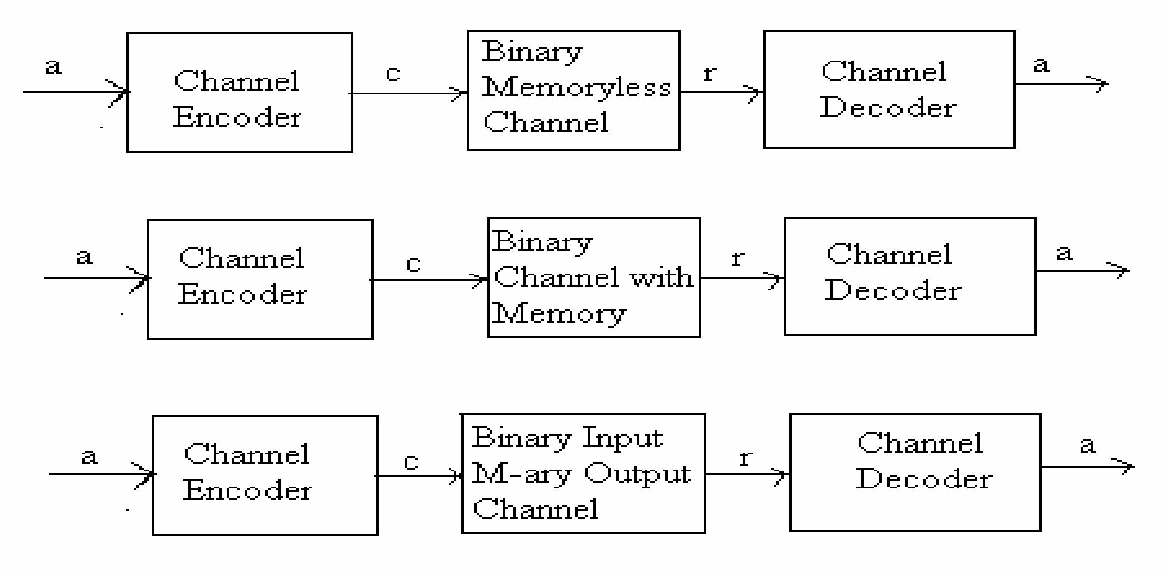

Figure 4.1 Channel models from the point of view of channel coding [5]

The decoder makes a decision about the transmitted coded sequence c on the basis of the received binary sequence r. In the decision process it can apply only the algebraic interdependences among particular bits of the transmitted sequence which have been implied by the coding rule. Due to the memoryless nature of the considered model, the occurring errors are mutually statistically independent, i.e. the occurrence of errors at previous moments does not have any influence on the error probability at the current moment. In reality only some transmission channels can be considered as memoryless. In most channels the errors occur in bursts. On the other hand there are many decoding algorithms which are designed for the correction of random errors, i.e. for memoryless channels. In order to ensure the error correction with the sufficient quality, additional means are undertaken in order to spread channel error bursts in the receiver. A widely applied method of destroying the error bursts is *interleaving.* It will be explained further in this chapter**.**

The second channel model reflects the bursty nature of errors occurring in the transmission channel. In this case the occurrence of a single error at one moment increases the probability of errors at the following moment. In this sense the channel has a memory of its previous states. There are special codes and decoding algorithms fitted to such a situation.

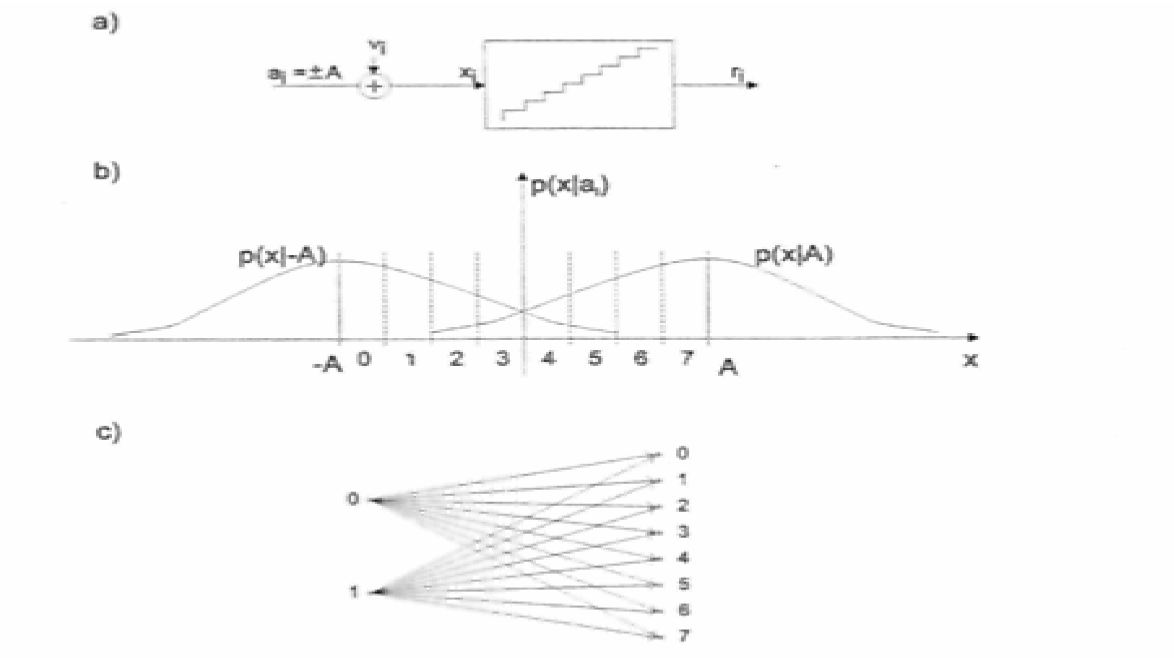

Figure 4.2 Example of use of the additional knowledge from the channel output for soft-decision. [5]

The third model, similar to the first one is also memoryless; however, it illustrates the case when more than binary information only is retrieved from the channel output. It means that the decoder uses not only algebraic interdependencies among particular bits in the coded sequence but also additional knowledge received from the channel, allowing for the improvement of the decoding process. Figure **4.2** illustrates a simple example of this case. Binary symbols are represented by bipolar pulses having values equal to $\pm A$. They are distorted by additive statistically independent Gaussian noise samples. Let the binary pulse -A represent the binary symbol "0", whereas the pulse $+A$ represent binary "1". The sample *x,* being the sum of the pulse and the noise sample, has the probability density function conditioned on the transmitted symbol +A or –A. In the receiver the sample *x* is quantized by an W-level quantization , giving the output symbol r. Assigning a digit in the range from 0 to M - 1 to each possible quantization level ,we obtain a channel model with binary input and **M-array** output. In the case of a **binary** quantization this channel model is reduced to the binary symmetric memoryless channel model. As we see. in our channel model the channel output is measured much more precisely as compared with the binary channel model. This allows us to use that additional knowledge on the received symbols to improve the decoding quality, i.e. to decrease the probability of a false decision upon the received coded sequence. In **Figure 4.2b** the dashed lines indicate the subsequent quantization levels. The type of decoding in which additional channel knowledge is used is called *soft-decision decoding,* as opposed to *hard-decision decoding* when only binary symbols are used by the decoder. Most of the decoding algorithms applied in modern digital cellular telephony use soft decisions. The kind of knowledge used in the soft-decision decoding resulting from the **M-level** quantization is not the only one used to improve the decoding quality. There are other methods of measuring the hit reliability applied by soft-decision decoding. The power level of the signal carrying the information bit is one of them.

4.2 The Essence of the Redundant Coding

As already mentioned, the channel coding relies on appending the information sequence by additional bits which constitute information redundancy. Let the subject of coding be a fc bit information sequence a. Assume that the information source can generate any combination of bits in the fc-bit block .Thus, 2^k different information sequences are possible. As a result of supplementing fc-bit information blocks by *n - k* additional bits we receive n-bit sequences. There are 2" different binary sequences of length *n.* however, only 2* sequences are selected from them. Each of them represents one of the possible information sequences a. Let us call them *code words.* The n-bit sequences are selected in such a way that the sequences should differ from one another as much as possible. Thus, despite the erroneous reception of some bits, the decoder can assign with a high probability that coded sequence to the received sequence, which has been sent by the transmitter. Difference among the coded sequences can be measured by the number of positions on which the bits of any pair of two coded sequences are different. This number is called the *Hamming distance* between two sequences. One can show that if binary errors occur statistically independently of each other (which means that we represent the channel by the binary, symmetric memoryless channel model], then 2 code words of length *n* should be selected in such a way that the minimum Hamming distance occurring between some pairs of them should be maximized. The optimum *maximum likelihood* decoder finds that sequence among 2* of code words which is the closest to the received n-bit sequence in the sense of the Hamming distance. If the minimum Hamming distance d_{min} between the coding sequences is minimized, the coding sequence can be erroneous in no more than t = [(d_{min} — 1) /2] positions- and the maximum likelihood decoder is still able to make a correct decision upon the received sequence.

Let the decoder have the unquantized channel output samples *XI (I* = l….n) at its disposal. Assume that the additive noise samples are Gaussian and statistically independent. One can prove that in case of the channel model shown in **Figures 4.2a** , the optimum decoder finding the maximum likelihood code word should select the coding sequence c = (ci,c_2....,cn,) which is the closest in the sense of the Euclidean distance to the received sequence x = *(xi, x*2,....,x_n*).* It means that the decoder selects that code word which fulfills the criterion [5]

$$\min c \sum_{i=1}^{n} (x_i - c_i)^2 \qquad (4.1)$$

In practice, the decoder does not handle the ideal values of the samples, but their quantized versions. Moreover, from the implementation point of view, it is much easier to calculate the distance between the received sequence and the code word in a sub-optimum way in form of the sum of modules of the differences between the elements of both sequences, i.e. the decoder has to search for such a coding sequence c, for which is fulfilled.

$$\min \sum_{i=1}^{n} | r_i - c_i | \tag{4.2}$$

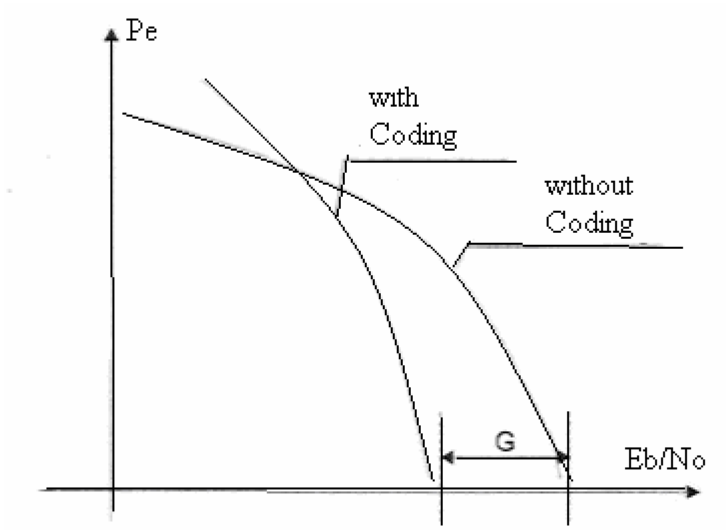

Figure 4.3 Illustration of the coding gain on the plot of block error probability versus Eb/No with and without channel coding [5]

The term coding gain is also strictly associated with the essence of coding. If we wish to compare the communication system with channel coding with the system without it, we have to assume that the time periods used for the transmission of sequences representing the same k-bit information block are equal in both cases. If the signal energy per bit in the system without coding is equal to E_b then the energy of a single bit in the system with coding has to be smaller due to the fact that redundant bits must be additionally transmitted. Thus, for the n-bit sequences representing the fc-bit information block the energy per bit is equal to $\frac{k}{n} E_b$. In general, the probability of erroneous decoding of a code word is a function of the ratio of the energy per bit to the noise power density N_o. Despite the fact that the energy per bit in the system with coding is lower than that for the system without coding, the performance of the system with coding is higher if the energy per bit to the noise power density ratio exceeds a certain threshold value. **Figure 4.3** shows that feature. For the raising value of *Eb/No* the plots of the probability of the erroneous block decoding for coded and un coded systems are becoming more and more parallel and asymptotically they are shifted one with respect to the other by *G* dB along the *Eb/No* axis. The value *G* is called an **asymptotic coding gain**.

4.3 Classification of Codes

There are several criteria of classification of channel codes. The first one is the function they perform. From this point of view we divide channel codes into ***error correction*** and ***error detection*** **codes**. The differences between these two code categories have already been explained.

The second criterion is the way the codes are created. Let us assume, similar to the previous section, that the binary information stream is divided into fc-bit blocks aj, where *j* is the block number. If the code word cj is a function exclusively of the current information block a., for each *j,* the code is called a *block code.* If the code word is a function of the current information block *aj* and a few previous information blocks (*aj-1, aj-2, aj-,aj-i)* it is called a **convolutional code**. From the point of view of the logical circuit theory a block code encoder can be implemented using only the combinatorial circuitry (logical gates), whereas the convolutional code encoder is an automaton and requires some memory cells. The term "convolutional code" originates from the observation that the binary sequence at the output of the encoder can be considered as a discrete convolution of the binary input stream with the **encoder impulse response**. The encoder impulse response is understood as the response of the encoder to a single "one" followed by a stream of zeros.

The basis of the next classification criterion is the number of different symbols of which code words are built. The symbols are mostly binary. A code in which binary symbols are used to compose code words is called a *binary code.* All operations performed on the elements of code words are realized in an algebraic field which consists of two elements: zero and one. In consequence, an additive operation is the addition[3] modulo-2, whereas a multiplication operation is a logical conjunction. In some special applications *nonbinary codes* are used. The number of different symbols used for representation of the code words is a primary number or is its power.

An example of a practical application of a nonbinary code is correction of a binary stream distorted by errors concentrated in bursts. A code word of the nonbinary code, in which the symbols selected from the set consisting of digits {0….. (2^m - 1)}are used, is created in such a way that subsequent symbols of the code word are represented by m-bit blocks. The additive operation is then addition modulo-2^m, whereas the multiplication operation is multiplication modulo-2^m. If the error burst does not exceed *m-bit* subsequent bits, it distorts at most two subsequent nonbinary code symbols. Then, in order to correct all m-bit long error bursts, it is sufficient to apply a nonbinary code which is able to correct at least two erroneous symbols.

According to another criterion of code classification, we divide codes into **systematic and nonsystematic**. In systematic codes the information blocks appear in the code words in the direct form and they are followed by parity bits. On the other hand, the symbols of the code word in a nonsystematic code are the sum of information symbols calculated in conformity with a selected coding rule and the information symbols do not appear in a direct form.

5. Channel Modeling

5.1 Introduction

Specification of the transmission channel characteristics has a crucial meaning for the design of any communication system. Channel propagation properties, introduced distortions and disturbances as well as the allowable transmitted signal bandwidth determine achievable transmission rate and its quality. Thus, before the specification of any communication system, the designer has to know the properties of the channel applied in the designed system. In this respect a mobile communication channel is not an exception. Therefore, before considerations of functioning of mobile communication systems we present transmission channel properties which are specific for mobile communication systems. Characteristic features of the transmission channel strongly depend on the type of the system. Channel properties are quite different for various systems, e.g. for indoor communications, cellular or satellite communication systems. In this chapter we will concentrate on some common features of communication channels currently used in mobile communications.

We will start with a basic set of terms associated with antennae. Then we will present basic expressions dealing with signal propagation in free space and calculation of the link power budget. Subsequently we will investigate the influence of the multipath effect on the transmission properties. In turn, we will present transmission channel models allowing for evaluation of communication system quality. We will also give examples of standardized channel models characterizing signal propagation in typical types of terrain, which are used in the GSM cellular system design. We will also survey the most important channel propagation models useful for evaluation of the cellular base station coverage. At the end of this chapter we will consider the phenomena of fiat and selective fading and the diversity reception as a method combating these phenomena.

5.2 Free-Space Signal Propagation [5]

Signal transmission in a radio system is based on converting the electrical signals generated by the transmitter into electromagnetic waves, propagation of waves in space and conversion back into electrical signals at the receiver side of the system. The properties of a mobile communication channel depend on many factors, in particular on the characteristics of the applied antennae, the properties of the physical medium in which the radio waves propagate, the features of the electronic circuits participating in signal transmission and reception, and the velocities of mobile stations. In order to give a simple presentation of mobile communication channel properties, we will first consider an ideal case - the signal propagation in free space. We will precede these considerations by explaining a few basic terms referring to antennae.

Let us consider a theoretical case of an antenna which emits the signal equally in all directions at the power level of *Pt* Watts. Such an antenna is called an *isotropic antenna.* It is an ideal device which cannot be built in practice; however, it serves as a reference for other antenna types. If we draw a sphere of radius r around the isotropic antenna, at each point of its surface the electromagnetic field induced by this antenna is identical. Real antennae focus radiates energy at the specific directions; therefore we usually describe normalized radiation antenna characteristics with the expression:

$$F(\theta,\varphi)=\frac{E(\theta,\varphi)}{E_{\max}} \tag{5.1}$$

where $E(\theta,\varphi)$ is the field at point *P* of the sphere whose coordinates are determined by the angles θ and φ , whereas E_{max} is the maximum field on the surface of the sphere. Figure 3.1 presents the coordinate system in which the angles θ and φ are denoted. The isotropic antenna is located in the origin of the coordinate system. It is easy to see that the normalized characteristic does not depend on the sphere radius r. The term *radiation density* is closely related to the normalized characteristic. It is the power radiated in the determined direction within a unit solid angle. Both antenna characteristics appear in the expression.

$$U(\theta,\varphi)=U_{\max}\ |\ F(\theta,\varphi\ |^{2} \tag{5.2}$$

where ($U_{\max}$ is the maximum radiation density. The total power P_T radiated by the antenna is the integral of radiation density with respect to the solid angle.

$$p_T=\int_{4\pi}U(\theta,\varphi)d\Omega=4\pi U_{mean} \tag{5.3}$$

s

where $d\Omega = \sin\theta\, d\theta\, d\varphi$ Let us note that the radiated power can be expressed as the product of the mean radiation intensity U_{mean} and the value of the full solid angle which is equal to *ix.* Mean radiation intensity can be interpreted as the radiation density of the isotropic antenna radiating the same power P_T as the given antenna. The ratio of the radiation density $U(\theta,\varphi)$ to the mean radiation density is called *directivity gain* of the antenna. Its maximum value is called the *antenna directivity D* and is described by the expression

$$D=\frac{U_{\max}}{U_{mean}} \tag{5.4}$$

The term *directivity* means that the radiation intensity in the direction of the maximum radiation is *D* times larger than the radiation intensity of the isotropic antenna radiating the same power as the given antenna.

In the case of a real antenna the radiated power is only a part of the power given to its connector. Part of the input power is dissipated and converts into heat. Thus the antenna is characterized by the *power efficiency* $\eta = P_T / Pi_{nput}$.
The antenna gain is often used in the determination of the *Effective Isotropic Radiated Power* (EIRP) which is described by the product Pi_{nput} with G. EIRP is the power which would have to be supplied to the isotropic antenna in order to achieve the same field at the reception point as that which is received at that point due to the antenna with the gain *G* and supplied by power Pi_{nput}.

Another type of the reference antenna is the *half-wave dipole.* If we compare the power of the antenna with the gain *G* with the power of the half-wave dipole, we define the so-called *Effective Radiated Power* (ERP). The half-wave dipole has a gain equal to 1.64 or 2.15 dB as compared with the isotropic antenna. The ERP radiated by a given antenna is 2.15 dB lower than the EIRP power. Depending on the reference antenna, the units of the antenna power gain are denoted as dBi for the isotropic antenna or as dBd for the half-wave dipole.
Let us consider a transmit antenna which radiates the power of *Pr* Watts into free space. Let the antenna gain in a given direction be G_T. The density of power sent in this direction and measured at the distance *d* is $P_T G_T / 4\pi d^2$ W/m^2. The receive antenna directed at the transmit antenna and located at a specified distance from the latter "collects" only a part of the radiated power. This received power depends on the *effective area of the antenna* A_R and is given by the formula (5.5),

$$P_R(d) = \frac{P_t G_t A_R}{4\pi d^2} \tag{5.5}$$

shows the dependence of the received power on the distance between the transmit and receive antennae.

The electromagnetic field theory shows that the effective area of the receiving antenna can be described by the expression where G_R is the receiving antenna gain, A =c/f is the wavelength of the transmitted signal, *c* is the light velocity and f is the frequency of the transmitted signal. Substituting in (5.6), we obtain the formula for the power of the received signal

$$A_R = \frac{G_R \lambda^2}{4\pi}[m^2] \qquad\qquad P_R(d) = \frac{P_T G_T G_R}{(4\pi d / \lambda)^2} \tag{5.6}$$

which allows us to write the received power, taking into account additional loss, for example the atmospheric loss L_a, by the formula

$$P_R(d) = P_T G_T G_R L_s L \tag{5.7}$$

5.3 Influence of the Multipath Effect on the Signal Propagation

Let us come back to the signal propagation. The power received at distance *d* from the transmit antenna can be expressed with respect to the power measured at a certain standard distance *do,* i.e. the reference **power** $p_R(d_0)$. Thus, on the basis of formula (5.6) *the* power received at the distance *d* can be expressed as:

$$P_R = P(d_0)(\frac{d_0}{d})^{\lambda} \quad \text{d>=}\ d_0 \qquad \lambda = 2 \tag{5.8}$$

From (5.8) we see that the received power of the signal propagating in free space is inversely proportional to the square of the distance from the transmit antenna. The reference distance d_0 has to be appropriately large in order to consider the receiver power at distance *d* in the antenna far-field determined by the so-called *Fraunhofer distance df* given by the expression

$$d_f = \frac{2L^2}{\lambda} \tag{5.9}$$

where *L* is the maximum physical linear size of the antenna and A is the wavelength. In practice, in the frequency range of 1 to 2 GHz the reference distance do is assumed to be 1 m for antennae used in the indoor environment (for example in the wireless telephony systems) or to be 100 m or 1 km in the outdoor environment .

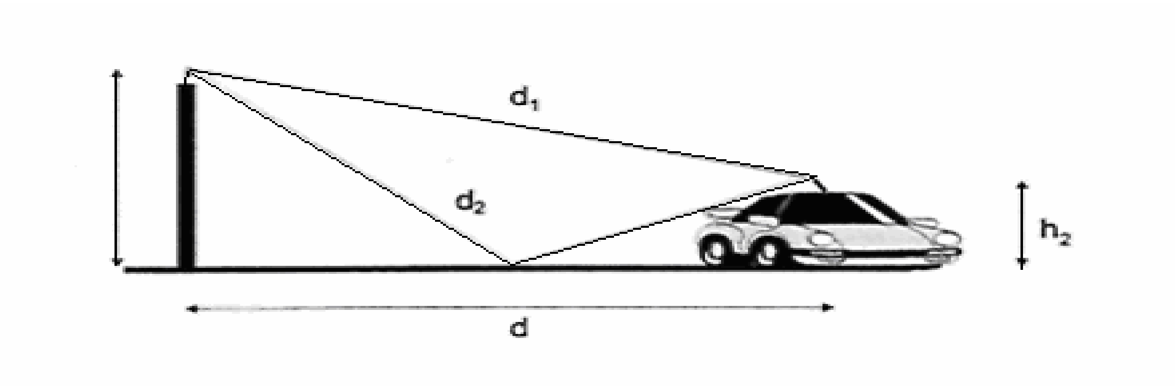

Figure 5.1 Illustration of two-path effect [5]

Let us consider a simplified model of signal propagation observed in mobile communications. Figure 3.3 presents the model of two-path propagation, which is a simplification of the real situation. Let the transmit antenna and receive antenna be placed /ii and h_2 meters above the ground level, respectively. The distance on the ground between both antennae is equal to d meters and is much longer than the heights of both antennae. Let us assume that the signal reaches the receiver along two paths: the direct one (line-of-sight) and one with reflection from the ground, as shown in Figure 3.3. Let the reflection coefficient $a_1= -1$, which means that the ground behaves as a lossless reflecting surface. The power of the signal arriving at the receiver is then

$$P_R = p(d_0)d_0^{\,2}\left|\frac{1}{d_1}\exp(j\varphi_1) - \frac{1}{d_2}\exp(j\varphi_2)\right|^2 \tag{5.10}$$

Simple geometrical dependencies allow deriving the lengths of both paths. They are

$$d_1 = \sqrt{(h_1 - h_2)^2 + d^2} \qquad d_2 = \sqrt{(h_1 + h_2)^2 + d^2} \tag{5.11}$$

Figure 3.1 presents the normalized power *Pn (d)* with respect to *P {do)* as a function of distance d for a few carrier signal frequencies in case of two-path propagation. Let us note that the level of the received power depends not only on the distance but also on the signal frequency. Besides a visible tendency of the power level to decrease one can also observe fast power level fluctuations as a function of the distance. This is in fact an illustration of the fading phenomenon. For certain values of the distance between antennae the signals traveling along both paths combine at the receive antenna with the opposite phases which causes the decrease of the signal power. For some other values of d the incoming signals add to each other constructively, which results in the increase of the signal level. Of course there are many intermediate cases as well. The model of a two-path channel is a simplification of the reality: however, it shows the essence of the multipath influence on the signal reception.

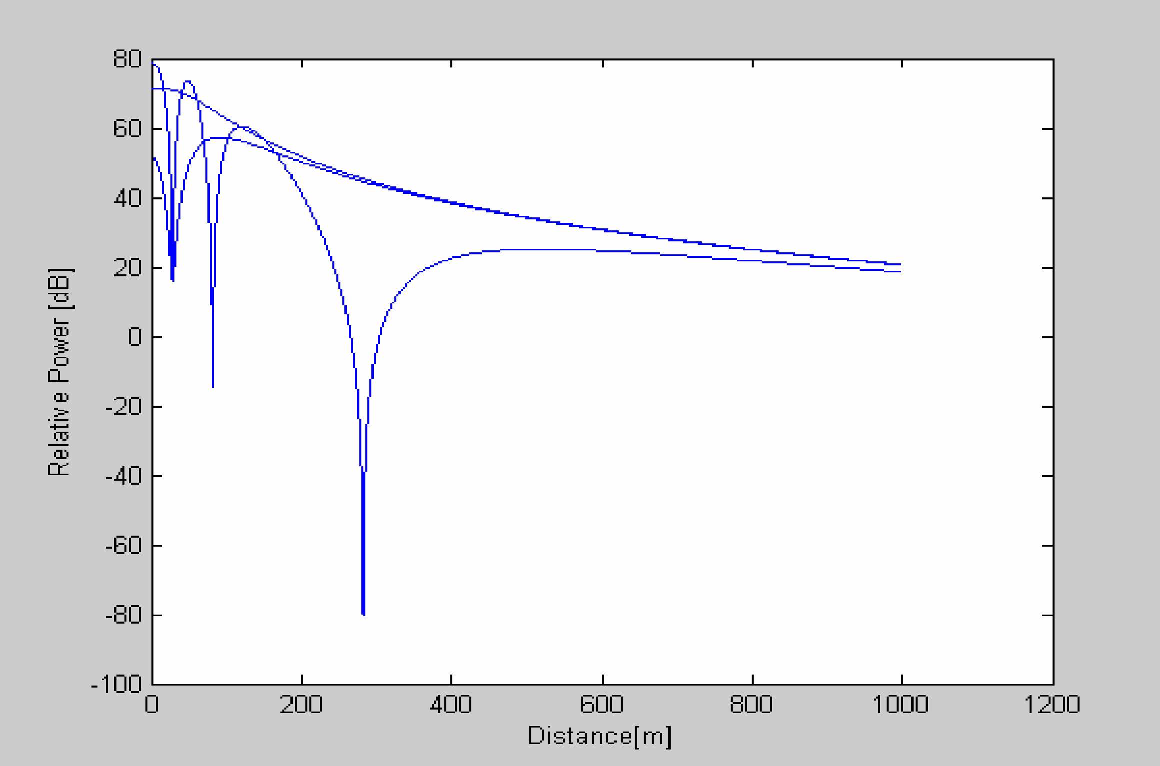

Figure 5.2 relative power level versus distance [m] ($h_1 = 50m$, $h_2 = 3m$ for $f_c = 100MHz$, $f_c = 500MHz$ $f_c = 1GHz$ and distance(d)=1000m.

5.4 Transmission Channel in Mobile Communication System

Consider a cellular telephony system as an example of a mobile communication system. The area covered by the system is divided into smaller areas with base stations in their centers (see Chapter 5). Communication with mobile stations is performed through base stations. Typically, base station antennae are omnidirectional or emit the signals into three sectors of the angle width of 120 degrees each. Mobile stations, due to their movement and continuous change of location with respect to the base station, have omnidirectional antennae. These facts imply the properties of the transmission channel. Because the power is emitted in all directions (or at least within a wide angle), the signal, prior to reaching the receiver, is the subject of reflections, diffractions and dispersion caused by several obstacles. The actual environment has a crucial meaning for these propagation phenomena. Figure 3.6 presents typical signal propagation from the base station to the mobile station.

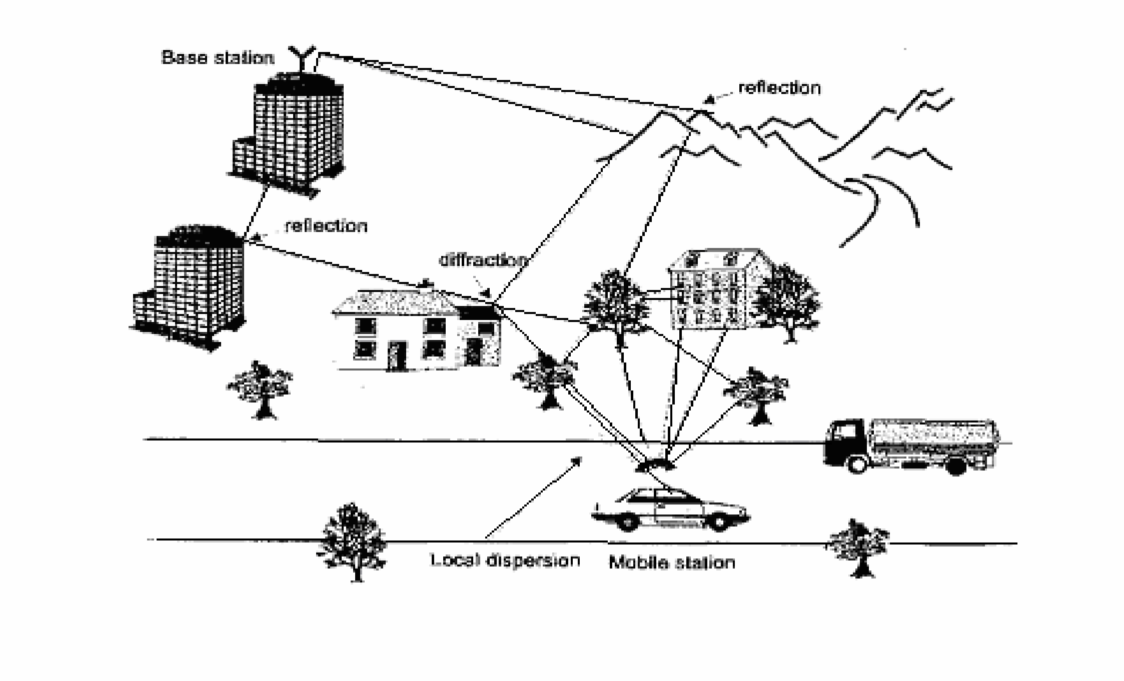

Figure 5.3 Example of signal propagation in the mobile communication systems. [5]

Let us notice that the signal arrives at the receiver along a few distinguishable paths. In the direct vicinity of the mobile station each signal component is additionally dispersed due to terrain obstacles which slightly differentiate its delays and phase shifts. Very often, in particular in hilly and urban areas there is no direct visibility of the base and mobile station antennae and the signal arrives at the receiver exclusively in the form of the reflected and dispersed components. Although the channel model shown in Figure 5.3 is very simple, it reveals the basic phenomena occurring in a short time scale during digital signal transmission at the rate ranging from a few tens to a few hundred kbps in the frequency range between a few hundred MHz and 2 GHz.

Let us denote the signal emitted by the base station as *s(t).* Its analytic[2] form is given by the expression

$$\mathrm{S(t)}=u(t)\exp(j2\pi f_c t) \tag{5.12}$$

where *u(t)* is the complex baseband equivalent signal. For example, in the GSM system the carrier frequency is around 900 MHz. Let us consider a mobile station installed in the vehicle moving at a speed of 250 km/h. Due to fast speed the Doppler affect is becoming significant. The signal components arriving along different paths at the receiver undergo different frequency shifts, which depend on the value of the carrier frequency, the vehicle velocity *v* and the angle between the direction of arrival of the signal component and the direction of the vehicle movement. In general, the Doppler frequency *fu* depends on the above mentioned factors according to the formula:

$$f_D = f_c \frac{v}{c} \cos\varphi \tag{5.13}$$

where *c* is the light velocity. For the channel model shown in Figure 3.6 we can write the following formula describing the received signal in the analytic form

$$r(t) = \sum_{k=1}^{M} \tau_k(t) = \sum_{k=1}^{M} \sum_{\mathrm{k}=1}^{\mathrm{N}} \tau_{ki}(t) \tag{5.14}$$

Let τ_{ki} be the delay after which the signal component $\tau_{ki}(t)$ arrives at the mobile station. Let us note that for the given path index *k* the components $\tau_{ki}(t$ travel along almost the same path and undergo additional dispersion and reflections around the mobile station, so their delays τ_{ki}(İ=I,………, N) are similar. Let T_k be the mean delay of the signals originating from the same *k-th* path. Then the single component $\tau_k(t)$ can be expressed in the form

$$r(t) = u(t - T_k) e^{j2\pi f_c \tau_{ki}} \sum_{i=1}^{N} \alpha_{ki} e^{-2jf_c \tau_{ki}} e^{j2\pi f_{Dki} t} \tag{5.15}$$

where α_{ki} is the attenuation introduced by the path denoted by indices *k* and *i.* Let us note that the simplification made in formula (3.35) is justified because the phase rotation caused by the factor $\exp(j2\pi f_{Dki} \tau_{ki})$ is negligible. For the Doppler frequency equal to 200 Hz and the maximum value of the relative delay equal to 20 μs, the angle $2\pi f_{\max} \tau_{ki}$ is not larger than 1.6 degrees. In general, the signal received by the mobile station can be described by the expression

$$r(t) = \left[\sum_{k=1}^{M} u(t - T_k) c_k(t) \right] e^{j2\pi f_c t} = v(t) e^{j2\pi f_c t} \tag{5.16}$$

$$c_k = \sum_{i=1}^{N} \alpha_{ki} e^{-j2\pi f_c \tau_{ki}} e^{j2\pi f_{Dki} t} \tag{5.17}$$

Equation (3.36) indicates that the baseband equivalent channel model can be considered is a tapped delay line with time-varying tap coefficients $c_k(t),(k=1\dots\dots M)$. Without loss of generality we can assume that $T_i < T_2 \dots\dots < T_M$

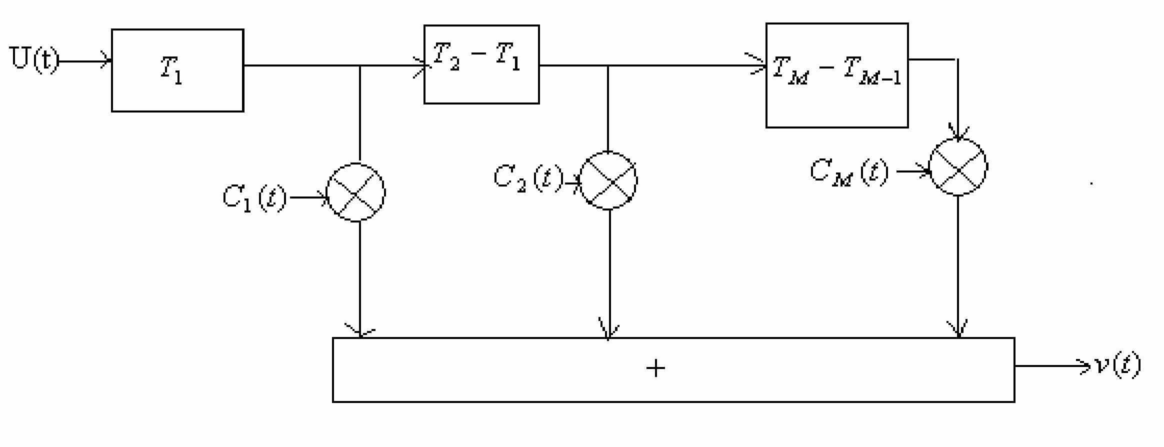

Figure 5.4 Short-term mobile communication channel model

5.5 Modeling the Propagation Loss [5]

The channel model described in the previous section explains the origin of multipath propagation and time variability of the channel characteristics caused by the movement of the mobile station and by signal dispersion in its direct vicinity. The model shown in Figure 5.2 is associated with the fast power level changes around the mean value. As we already know, even small changes in the location of the mobile station can be the reason of substantial changes in the received signal level. From the point of view of mobile system design the following factor has to be taken into account as well. This factor is the mean power level as a function of the distance from the base station. Usually the measurements are averaged in the distance interval of 5λ to 40λ, where A is the carrier wavelength. In the frequency range between 1 and 2 GHz the local mean power is averaged over the distance of 1 to 10 meters. The measurement results are the function of the distance from the transmit station (compare (5.8)) and actual configuration of the main obstacles and distorting and reflecting elements along the paths to the receiver but not in close vicinity of the receiver. This type of information is necessary in the design of cellular systems.
Let us come back to the analysis of formula (5.8). It turns out that this formula shows an agreement with the measurements if they arc averaged over all possible positions of the receive antenna at distance d from the transmit antenna. After calculation of the logarithms of both sides in (5.12) we obtain the expression

$$P(d) = (p(d_0)) - 10\lambda \log(\frac{d}{d_0}) \qquad (5.18)$$

We conclude that the mean power decreases linearly with distance *d* in the decibel scale. The rate of decrease is 10"; dB per decade. As already mentioned, the parameter 7 depends on the propagation environment. Table 3.1 presents typical values of the path loss exponent 7 for several types of environment.

Table 5.1 Values of λ for several environmental types.

Type of environment	Path loss exponent(λ)
Free space	2
Urban area cellular radio	2.7 to 3.5
Shadowed urban cellular radio	3 to 5
In building line-of-sight	1.6 to 1.8
Obstructed in building	4 to 6
Obstructed in factories	2 to 3

Formula (3.47) shows how the mean received power depends on the distance from the transmit antenna. It has been **observed** that the power **levels** measured in **two** different locations equally distant from the transmit antenna can be quite different due to different layout of the reflecting, attenuating and diffracting obstacles. This phenomenon is called *shadowing.* The measurements indicate that the level of the received power is random. Moreover, it has been found that the power level in dB-scale is a Gaussian random variable,

$$P(d) = (p(d_0)) - 10\lambda \log(\frac{d}{d_0}) + X(0,\sigma) \qquad (5.19)$$

where $X(0,\sigma)$ is a Gaussian random variable with zero mean and variance σ. Thus, in the linear scale, the received power has a log-normal distribution. Knowing the distribution in the dB-scale, in particular knowing the variance a^2, it is possible to calculate the probability of the event that the received signal level at the given location point exceeds a selected threshold. Such calculations can be useful in the evaluation of the base station coverage area.

Radio propagation in a terrain featuring several obstacles such as buildings, terrain irregularities and trees and bushes is such a complicated process that the system designers often perform practical electromagnetic field measurements in specified terrain locations in order to determine the real base station coverage. Such measurements are very expensive, so on the basis of the measurement campaigns performed in different representative environments, several propagation models have been established which estimate the mean power loss in function of distance *d* from the base station, the type of environment and transmit and receive antenna heights. In the following sections we will present the most representative examples of the experimental propagation models.

5.5.1 The Lee Model

W.C.Y. Lee proposed a very simple signal propagation model originating from a series of measurements made in the USA at the carrier frequency f_r= 900 MH? According to the Lee model, the mean power measured at distance *d* from the transmit station is determined by the expression

$$p(d) = p_0 (\frac{d}{d_0})^{-\lambda} (\frac{f}{f_0})^{-n} F_0 \qquad (5.20)$$

The symbol P_o is the reference median power measured at distance do = 1 km, whereas F_o is the correction factor selected on the basis of a series of component factors according to the formula

$$F_0 = \coprod_{i=1}^{5} F_i \qquad (5.21)$$

$$F_1 = (\frac{\text{Actual BS antenna height [m]}}{30.5[m]})^2 \qquad (5.22)$$

$$F_2 = (\frac{\text{Actual MS antenna height [m]}}{3[m]})^v \qquad (5.23)$$

The power v = 1 for the mobile station antenna heights lower than 3 m and v = 2 for the heights larger than 10 m. In turn

$$F_3 = (\frac{\text{Actual Power}}{10\ W})^2 \qquad (5.23)$$

$$F_4 = (\frac{\text{BS antenna gain with respect to half - wave dipole}}{30.5[m]})^2 \qquad (5.24)$$

$$F_5 = (\text{BS antenna gain with respect to half - wave dipole})^2 \qquad (5.25)$$

The parameters P_o and 7 are selected experimentally based on the performed measurements. Table 5.2 presents the values of P_o and λ for some characteristic types of environment.

Table 5.2 Values of **Po** and λ for different environments

Environment	P0(dBm)	$\lambda(dB/decade)$
Free space	-41	20
Open (rural area)	-40	43.5
Suburban, small city	-54	38 1
Philadelphia	-62.5	36.8
Newark	-55	43.1
Tokyo	-18	30.5

The mean power loss in function of frequency is modeled by the factor $(\frac{f}{f_0})^{-n}$ and the choice of power n. The value of n is contained in the range between 2 and 3 for the frequencies from 30 MHz to 2 GHz and the distances between mobile and base stations contained in the range between 2 and 30 km. The power *n* also depends on the terrain topography, It is recommended to select n = 2 for suburban and when operating at the frequencies below 450 MHz. and n = 3 for urban environment and carrier frequencies over 450 MHz.

5.5.2 The Okumura Model

The Okumura model is also a result of intensive **measurements.** The measurements in the frequency range between 150 and 1020 MHz were performed in the Tokyo area. The authors proposed the following formula for the median loss $(L_{50})_{dB}$ function of the distance *d* from the transmit antenna of the base station

$$(L_{50})_{dB} = L_s + A(f,d) + G(h_{MS},eff) + G(h_{MS}) \qquad (5.26)$$

where L_s is the free space loss . The symbol *A(f,d)* denotes the median attenuation relative to free space in an urban area over quasi-smooth terrain with the effective base station antenna height $h_{BS},eff = 200$ m and the mobile antenna height h_{MS}=3m. $G(h_{MS}, eff)$ IS the correction term in dB associated with the base station antenna and depending on its effective height if the latter is different from 200 m. $G(h_{MS})$ is the correction factor in dB associated with the mobile station antenna if its height is different from 3 m. The free space loss is calculated (5.26) in the dB-scale. Equation (5.26), together with Figures 5.5 and 5.6, allows evaluation of the signal attenuation in an urban area for the frequency range between 150 and 2000 MHz if the distance between the base and mobile stations is contained between 1 and 100 km and the effective base station antenna height is in the range of 30 to 1000 m.

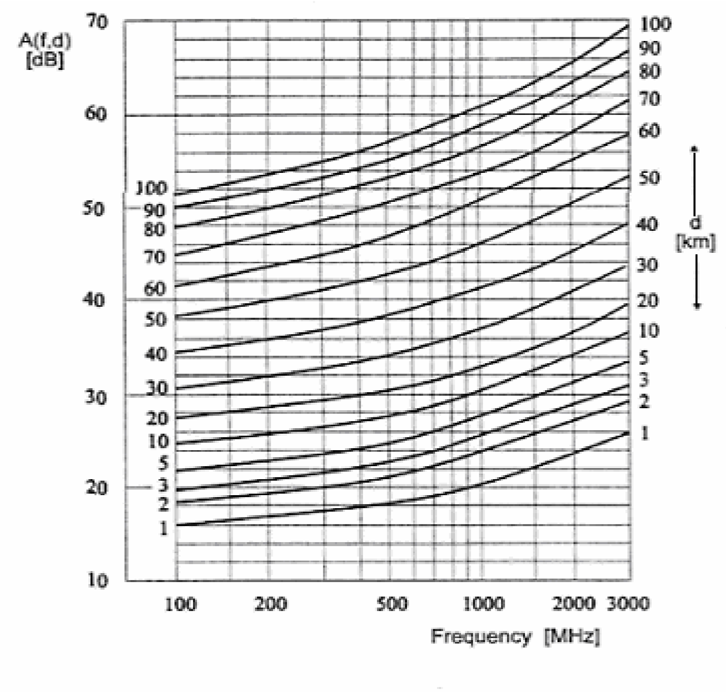

Figure 5.5 Median A (f,d) relative to free space over a quasi-smooth terrain

In the literature one can find **another** equivalent to (5.26) version of the formula describing the Okumura model. It has the form

$$(L_{50})_{dB} = L_s + A(f,d) - G(h_{MS}, eff) - G(h_{MS}) - G_{AREA} \tag{5.27}$$

The parameter $A(f,d)$ is as previously, read from the plots in Figure 5.5, whereas the correction terms $G(h_{MS}, eff)$ and $G(h_{MS})$ are given by expressions:

$$G(h_{MS}, eff) = 20\log(\frac{h_{BS,eff}}{200}) \qquad 10 \text{ m} < h_{BS,eff} < 1000\text{m} \qquad (5.28)$$

$$G(h_{MS}, eff) = 10\log(\frac{h_{BS,eff}}{3}) \qquad h_{MS} <= 3\text{m} \qquad (5.29)$$

$$G(h_{MS}, eff) = 20\log(\frac{h_{BS,eff}}{3}) \qquad 3\text{m} < h_{MS} < 10\text{m} \qquad (5.30)$$

The correction term G ARE A (in dB) depends on the type of terrain and carrier frequency and is read from the plots in Figure 5.6.

The Okumura model is very simple. It is based exclusively on the measurement data collected in the Tokyo area. The characteristic of .Japanese urban areas is slightly different than in Europe or the USA. Nevertheless, the Okumura model is popular and is considered one of the best models used in cellular acid other land mobile systems. The main drawback of the Okumura model is its slow reaction to the change of terrain type. The **Okumura** model is suitable for urban and suburban environment. However, it is not too practical for the rural area.

5.5.3 The Hata Model

The Hata model has been developed as a result of proposing empirical formulae to describe the plots created by Okumura and his collaborators. These formulae well approximate the plots for certain carrier frequency ranges and for the quasi-smooth terrain. Hata proposed the following empirical formulae for estimation of the signal attenuation.
For an urban area in the frequency range from 150 to 1500 MHz and for the effective base station antenna heights 30 <= h_{BS}, eff <= 200m we have:

$$(L_{50})_{dB}\,|_{urban} = 69.55 + 26.16\log f - 13.83\log(h_{BS}, eff) - a(h_{MS}) + (44.9 - 6.55\log(h_{BS,eff}))\log d \quad (5.31)$$

where the correlation term depends on the height of the mobile station antenna and in the range 1<= h_{MS} <=10m is calculated from expression:

$$\text{a}(h_{MS}) = 8.29(\log 1.54 h_{MS})^2 - 1.1[dB] \text{ for } f <= 400 \text{ MHz} \qquad (5.32)$$

$$a(h_{MS})=3.2(\log 11.75h_{MS})^2-4.79[dB] \quad f>=400\text{ MHz} \qquad (5.33)$$

For suburban area the propagation loss can be estimated according to formula

$$(L_{50})dB=(L_{50})dB|_{urban}-2(\log(\frac{f}{28}))^2-5.4 \qquad (5.34)$$

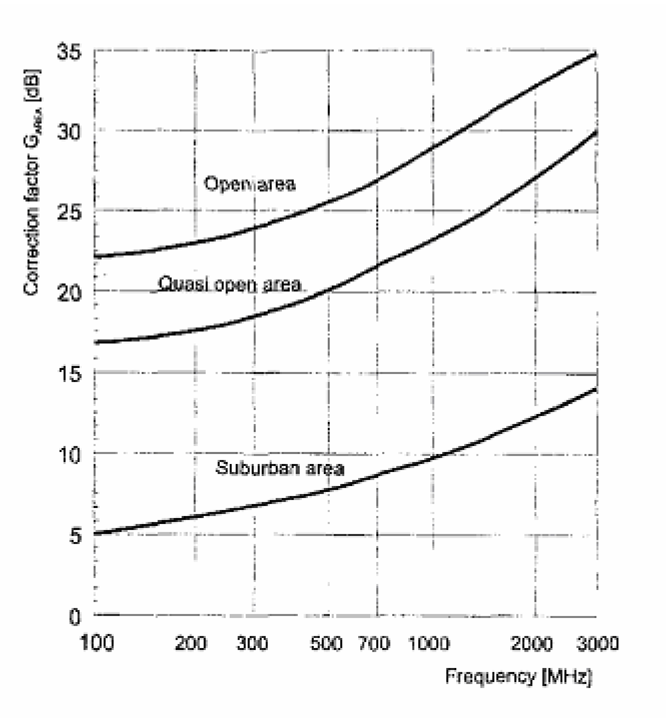

Figure 5.6 Correction term G areas in function of frequency and area type

whereas for the open area:

$$(L_{50})dB=(L_{50})dB|_{urban}-4.78(\log f)^2+18.33\log f-40.94 \qquad (5.34)$$

The propagation models presented so far allow for the estimation of the signal loss L function of the carrier frequency, the base and mobile station antenna heights and type of terrain. They better or worse reflect signal propagation at a large distance from the base station, exceeding 1 km. They are mostly valid for the frequency ranges caching up to 1.5 GHz. However, *Personal Communication Systems* operate in the 8, to 2.0 GHz range. Two examples of such systems are DCS 1800 or PCS 1900 two versions of the GSM system operating in Europe and the USA, respectively, Therefore, a lot of experiments and measurements have been performed in order to instruct the propagation models for the band 1.8 to 2.0 GHz for the environments and at assumptions characteristic for the PCS systems. Due to larger signal attenuation L the 1.8 GHz band as compared with 900 MHz band traditionally used by cellular systems the basic difference between PCS and classical cellular systems is the smaller size of the cells. The research on new propagation models was intensively conducted within the European Union COST#231 projects. There are at least two well-known propagation models reported in literature and developed within COST activity.

6. The Design Of The Multichannel Propogation with FSK and FDMA

6.1 Theoretical Structures of the Model

In this project we use the deterministic model to define the signal transition from the transmitter to receiver.

$$S_R(t) = \sum_{i=1}^{N} V_i \sin[(w_s \pm w_D)(t-\tau)+\phi_i)] + n_i(t) \qquad (6.1)$$

S_R = Transmitted signal form.
V_i =Amplitude of the modulation signals
w_s (2*pi* f_s) = Angles of the signals used for FSK modulation.
w_D (2*pi* f_D) = Doppler frequency shift.
ϕ =Phase shifting.
n_i =Noise taken from the air.

6.1.1 Transmitter Block

$$Sc_1 = V_1 \sin[(2\pi f_{s1} \pm 2\pi f_{D1} + \phi_1)] + V_2 \sin[(2\pi f_{s2} \pm 2\pi f_{D2} + \phi_2)] + n_1(t) \qquad (6.2)$$

$$Sc_2 = V_3 \sin[(2\pi f_{s3} \pm 2\pi f_{D3} + \phi_3)] + V_4 \sin[(2\pi f_{s4} \pm 2\pi f_{D4} + \phi_4)] + n_2(t) \qquad (6.3)$$

$$Sc_3 = V_5 \sin[(2\pi f_{s5} \pm 2\pi f_{D5} + \phi_5)] + V_6 \sin[(2\pi f_{s6} \pm 2\pi f_{D6} + \phi_6)] + n_3(t) \qquad (6.4)$$

$$Sc_4 = V_7 \sin[(2\pi f_{s7} \pm 2\pi f_{D7} + \phi_7)] + V_8 \sin[(2\pi f_{s8} \pm 2\pi f_{D8} + \phi_8)] + n_4(t) \qquad (6.5)$$

$$\text{St(t)} = Sc_1 + Sc_2 + Sc_3 + Sc_4 + nt(t) \qquad (6.7)$$

In this model, Sc_1(t), Sc_2(t), Sc_3(t), Sc_4(t) are the signals transmitted to receiver on the channel 1, channel 2, channel 3, channel 4.St(t) is the sum of the transmitted signals at the channels. And this signal is sent by the carrier to the receiver. During the signal transitions, noise that comes from the air adds the total signal. The signal will be changed by this factor. But noise is at the all frequencies. And noise is calculated as a dB scale according to SNR value and defined distribution type. But we must define the noise level according to formulas. Firstly, we must define the SNR value and calculate the Pn value by this SNR while the variance (Vm) of the signal is 1. Fluctuate limit is Gmax for the signal. We define the variance (Nk) of the noise according to this fluctuate limit as you see below:

$$SNR = 10^{\left(\frac{SNRdB}{10}\right)} \qquad \text{SNR}=\frac{P_s}{P_n} \tag{6.8}$$

$$s(t) = Vm\sin(Wct) \qquad Pn = \frac{V_m^{\ 2}}{SNR} \tag{6.9}$$

$$G\max = \sqrt{Pn} \qquad \text{Uk= rand (1, N2);} \qquad Nk = 2 * G\max * Uk - G\max \tag{6.10}$$

At the end, transmit signal will be constructed by this block diagram shown at figure 6.1.

6.1.2 Receiver Block

At the receiver part, We take the signals (Sc_1(t), Sc_2(t), Sc_3(t), Sc_4(t)) coming from the Transmitter and we can select the signal according to the modulation signal frequencies. F_{s1}, F_{s2}, F_{s3}, F_{s4}, F_{s5}, F_{s6}, F_{s7}, F_{s8} are the modulation signals frequencies to use the modulate the binary input according to FSK modulation. And we select the original channel signal with the filter according to channel frequencies. F_{s1}, F_{s2} , is the first channel frequency. F_{s3}, F_{s4} is the second channel frequency. F_{s5}, F_{s6} are the third channel frequency. F_{s7}, F_{s8} is the fourth channel frequency.

We design the band pass filter and select the original signals according to Channel frequencies. And we can determine the output of the signal as a 1 or 0. We determine the output bits according to zero crossing numbers. And we define the threshold and take the bits from output of the receiver. Normally, we must take the zero crossing numbers as a average of the zero crossing numbers of the two modulation signals. And then, If the zero crossing number is higher than threshold number , take bit as a 1 or And then, If the zero crossing number is lower than threshold number , take bit as a 0.Threshold zero crossing numbers are different at the different channels. It is defined according to channel frequencies.

At the end of the receiver output, there are bit error rates for every channels .An it is evaluated as percentage by this formula:

$$\text{Bit Error Rate} = \frac{\text{Number of corrupted bits}}{\text{Number of total bits}} * 100 \qquad (6.11)$$

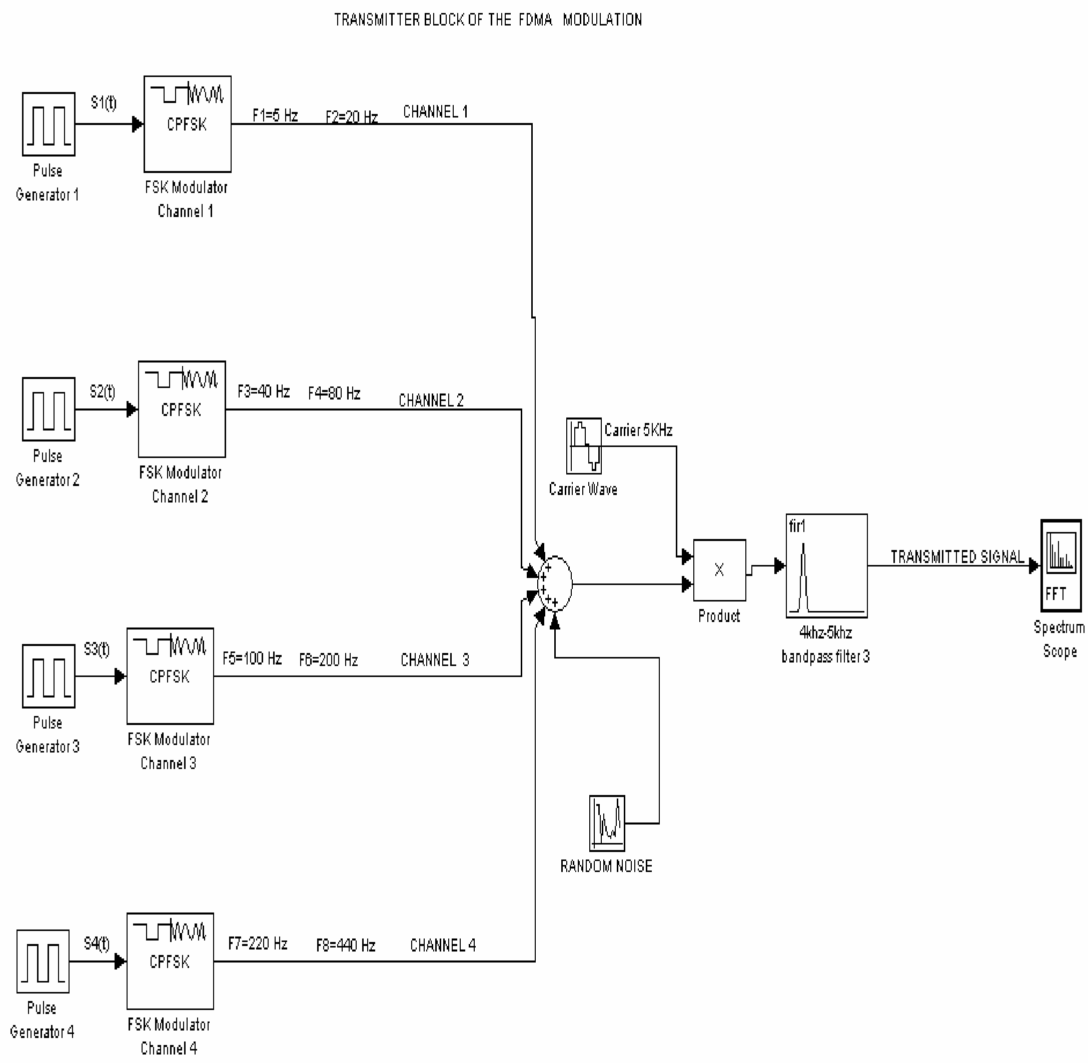

Figure 6.1 Transmitter block diagram

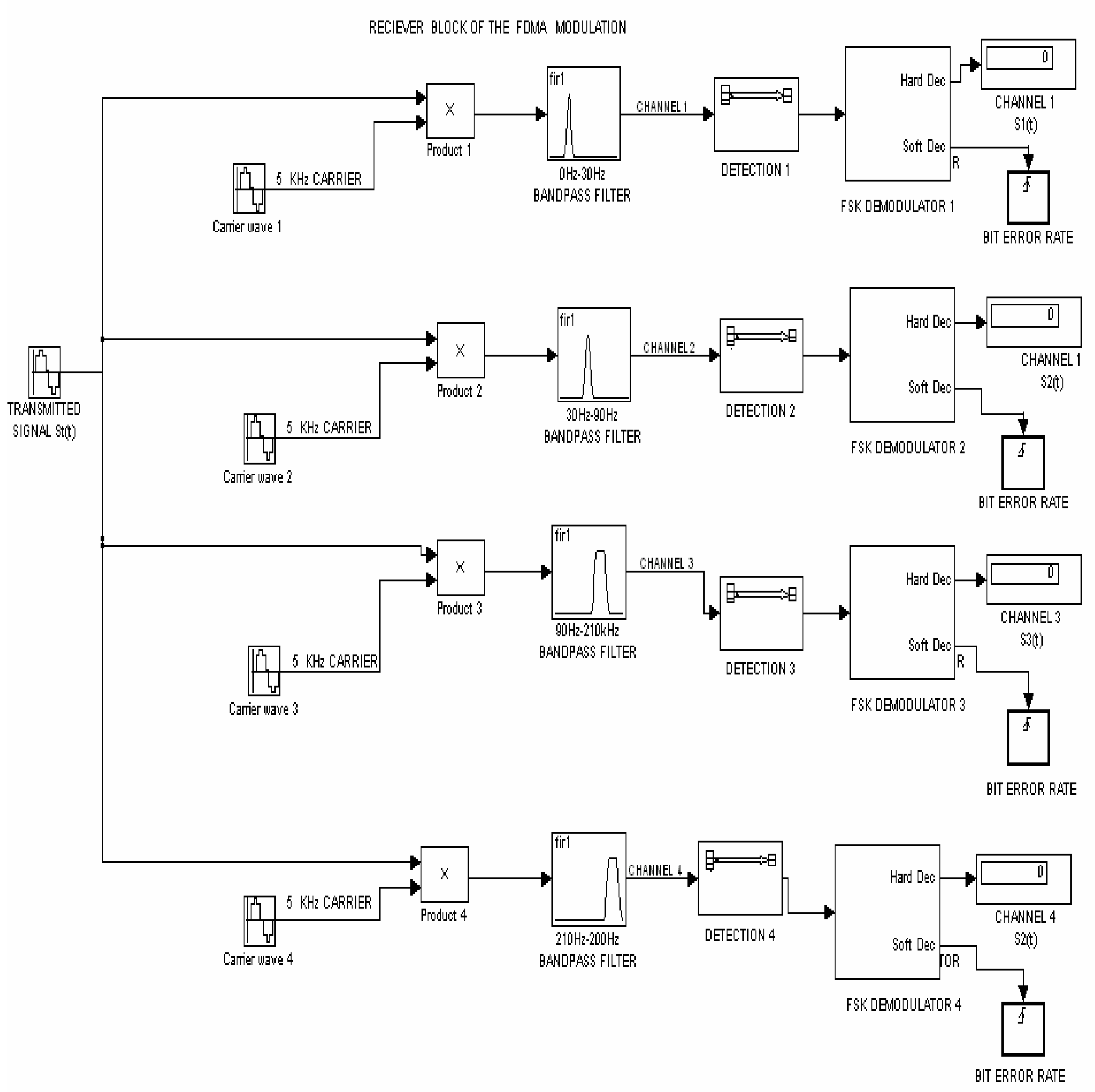

Figure 6.2 Receiver block diagram

6.2 Simulation of the Model at MATLAB

In this simulation program, firstly we modulate the bit streams with the FSK modulation and we send modulated signals at the fourth channels. But in this process, bit streams are random and its length is taken from the used. In addition, we send a signal having the sum of the all.

Transmitted channel .Then we must send this signal (St (t)) on the carrier. At the transmission signal, noise is added to transmission signal. And we must detect the signal from the noise.

At the receiver part, we take the total signal, sum of the signals sending with transmission channels. And then the total signal is filtered by the band pass filter at the known channel frequencies (f1-f2,f3-f4,f5-f6,f7-f8).As a result, all transmitted signals are reached at the receiver part at the wanted channels. After this process, we must detect and demodulate of the taken FSK signals at the wanted channels. At the end of the program, we detect the signals with counting the zero crossing number according to given threshold zero crossing number.

At the end of the program, we generally reach the bits sent at the transmitter part. But some time the bit error rates occur for each channel according to SNR value. Because , SNR value defines the corruption of the signal. If the corruption is high level, it is very difficult to detected output bit steam correctly. If the corruption is low level, it is easier than high level to detected output bit steam correctly. This program is simulated and designed by the given specifications. The design and construction of the program is explained below.

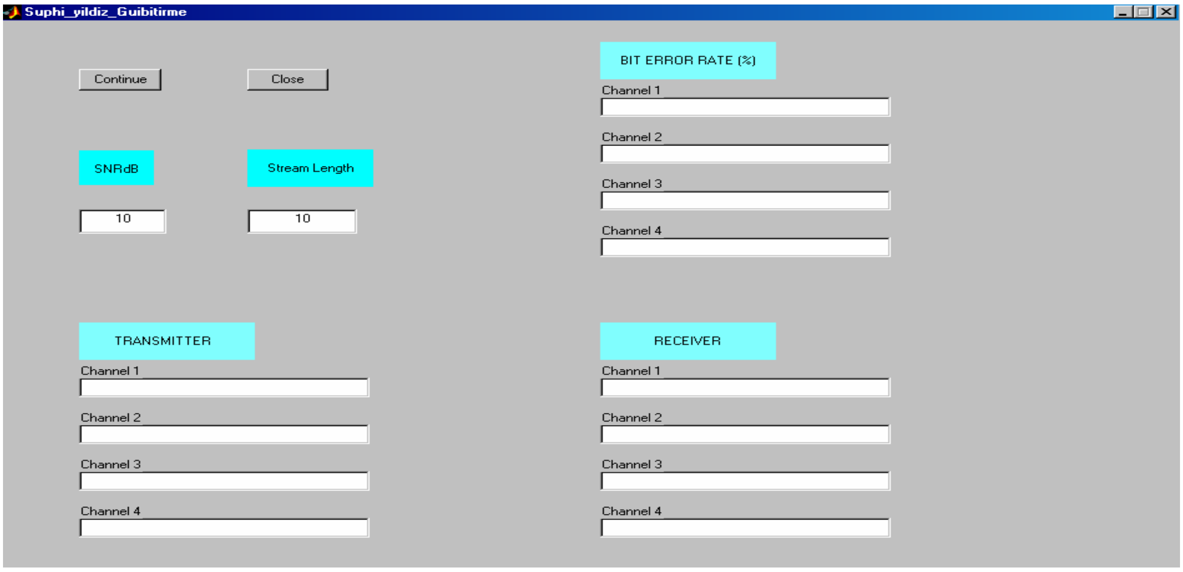

Figure 6.3 User interference of the program

6.2.1 Program Inputs

As you see above, this program takes the inputs from the users. The inputs are Signal to noise Ratio (SNR) and the Length of Streams (Stream Length) taken from the user.

6.2.2 Frequency and Time Parameters

According to these inputs we define the frequency and time domain parameters.

$deltat$: Time Changing interval.
$T\max$: Time interval which is responsible for sending a bit.
$f\max$: Maximum frequency value.
$deltaf$: Frequency Changing Interval.

$$deltat = \frac{1}{2f\max} \qquad T\max = deltat * N \qquad deltaf = \frac{1}{T\max} \qquad (6.11)$$

Mathlab code of the parameters

```
dt =0.001             .
T1=0.25
T2=T1*Q
N2=T2/dt
N1=T1/dt
fmax=(1/(2*dt))
df=1/T2
```

In this program, dt is the time Changing interval. T1 is time interval which is responsible for sending a bit. Q is length of the bit streams.T2 is total time which is responsible for sending bit stream or the modulated signal. N2 is number of iteration of bit stream or modulation signal. N1 is number of iteration for a bit. Fmax is maximum operation frequency. Df is frequency changing interval.

6.2.3 Channel Frequencies

In this program we use the four channels and define the frequencies as at least two times of first frequency:
Channel 1: 0-30 Hz
Channel 2: 30-90 Hz
Channel 3: 90-210 Hz
Channel 4: 210-500 Hz

6.2.4 Noise Adding

We calculate the noise according to SNR value taken from the user. The signal variance (Vm) is 4 , because there are four channel in the system , as you see formulas (6.8, 6.9,6.10).

Mathlab code of the noise adding

```
SNR=10^(SNRdB/10)
Vm=1
Pn=Vm^2/SNR
gmax=sqrt(Pn)
Uk=rand(1,N2)
Nk=2*gmax*Uk-gmax
```

In this part, **SNR** is signal to noise ratio in linear. **Vm** is the variance of the FSK signal. **Pn** is noise power. **Gmax** is fluctuation limits of the uniformly distributed noise Nm.**Uk** is random number generator between 1 and 0.**Nk** is shifting type of gamx between 1 and -1.

6.2.5 Transmitter Part of the Program

Bit streams

In this simulation, we take the data having the random bit streams. And we modulate and send these bit streams at the four channels.

Mathlab code of the random bit streams

```
s1=randint(1,Q)    Take the Random Bit Stream 1
s2=randint(1,Q)    Take the Random Bit Stream 2
s3=randint(1,Q)    Take the Random Bit Stream 3
s4=randint(1,Q)    Take the Random Bit Stream 4
```

s1 is random bit stream1.**s2** is random bit stream 2 .**s3** is random bit stream 3 .**s4** is random bit stream 4.

6.2.6 FSK Modulation

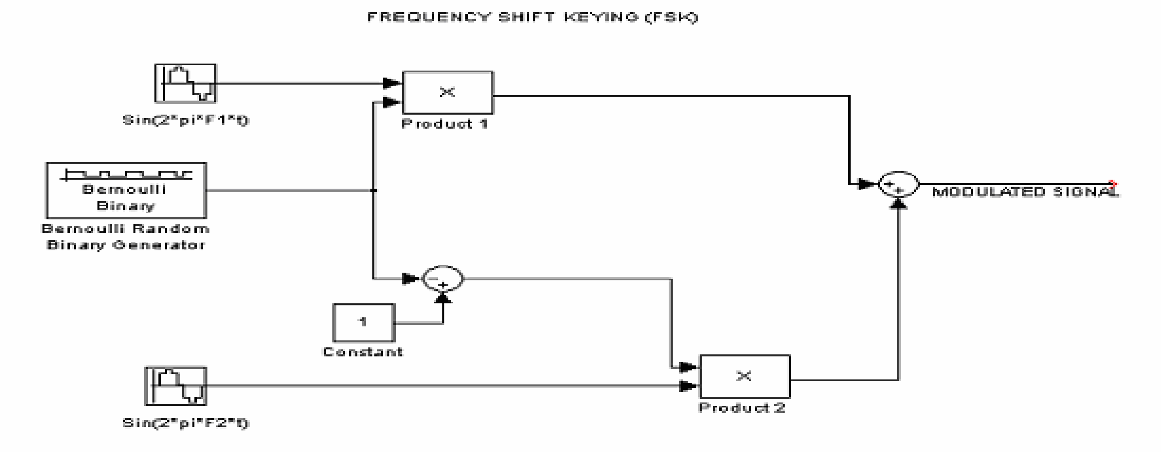

Figure 6.4 FSK modulator

Firstly, we modulated Random bit streams with the FSK modulation according to modulation frequencies. In this modulation, if the modulation bit is zero, signal shows this bit at the F1 frequency. If the modulation bit is one, signal shows this bit at the F2 frequency. And this process is the same for the modulated signals at the all channels .The FSK modulation is shown below:

For first channel, modulation frequencies are f1 (5Hz), f2 (20Hz).
For second channel, modulation frequencies are f3 (40Hz), f4 (80Hz).
For third channel, modulation frequencies are f5 (100Hz), f6 (200Hz).
For fourth channel, modulation frequencies are f7 (220Hz), f8 (400Hz).

Mathlab code of the FSK modulation at wanted frequencies

For first bit stream at frequencies (f1, f2)

```
nn=0; n=0; m=s1(1)
for tt=dt:dt:((T2))
if mod(n,(N1))==0
m=s1(nn+1)
nn=nn+1
end
Freq(n+1)=-fmax +(n-1)*df
time(n+1)=tt
if m==1
x3(n+1)=cos(2*pi*f2*dt*(n+1))
elseif m==0
x3(n+1)=cos(2*pi*f1*dt*(n+1))
end
n=n+1; end
```

For second bit stream at frequencies (f3,f4)

```
nn=0; n=0; m=s2(1)
for tt=dt:dt:((T2))
if mod(n,N1)==0
m=s2(nn+1)
nn=nn+1
end
if m==1
x4(n+1)=cos(2*pi*f4*dt*(n+1))
elseif m==0
x4(n+1)=cos(2*pi*f3*dt*(n+1))
end
n=n+1;
end
```

For third bit stream at frequencies (f4, f6)

```
nn=0; n=0; m=s3(1)
for tt=dt:dt:((T2))
if mod(n,N1)==0
m=s2(nn+1)
nn=nn+1
end
if m==1
x4(n+1)=cos(2*pi*f6*dt*(n+1))
elseif m==0
x4(n+1)=cos(2*pi*f5*dt*(n+1))
end
n=n+1;
end
```

For fourth bit stream at frequencies (f7, f8)

```
nn=0; n=0; m=s4(1)
for tt=dt:dt:((T2))
if mod(n,N1)==0
m=s2(nn+1)
nn=nn+1
end
if m==1
x4(n+1)=cos(2*pi*f8*dt*(n+1))
elseif m==0
x4(n+1)=cos(2*pi*f7*dt*(n+1))
end
n=n+1;
end
```

In this program we built the FSK modulation array by this code. In this code, cosine wave at the f2 refers 1 bits and cosine wave at the f1 refers 0 bits .We repeat this code for all bit streams (s1,s2,s3,s4) coming from four channels at the different frequencies (f1,f2;f3,f4;f5,f6;f7,f8). This figure is shown FSK modulation graph for first bit stream at f1 and f2 modulation frequencies:

S1= 0001100001

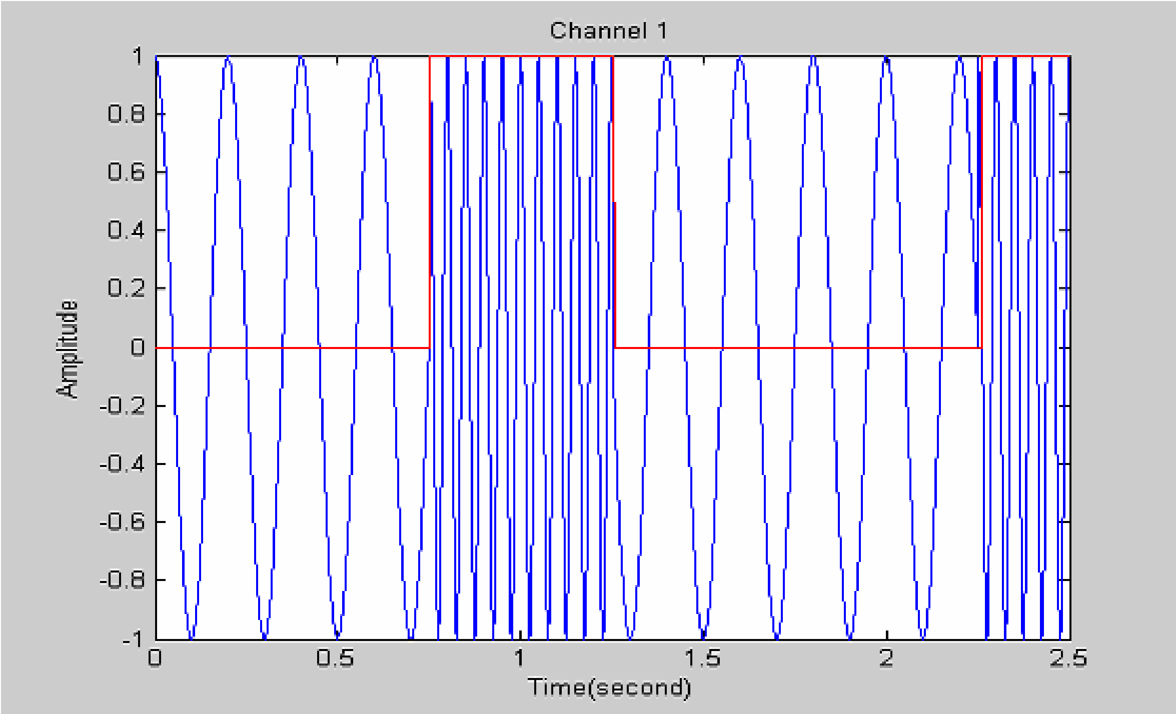

Figure 6.5 Modulated signal at the transmitter side for channel 1 for f1=5 Hz and f2=20Hz

6.2.7 Transmitted Signal

FSK modulation is constructed for all bit streams coming from the four channels .At the end of the FSK modulation we reach the modulated signals (x3,x4,x6,x8). And then , we add the modulated signals (x3,x4,x6,x8) and we reach the Transmission signal St(t). S(t) is the sum of the x3,x4,x6,x8 signals. In addition, we add noise to transmission signal and send this signal with the carrier wave. We will detect the transmitted signal at the receiver part .Because noise is added on the four channel and we use this formula at the program .Because variance of the modulation signal change between 1 and 0 for every channel. Transmitter block diagram is explained at **figure 6.2.**

$$\text{Transmitted Signal} = x3+x4+x6+x8+4*\text{Noise (Nk)} \qquad (6.12)$$

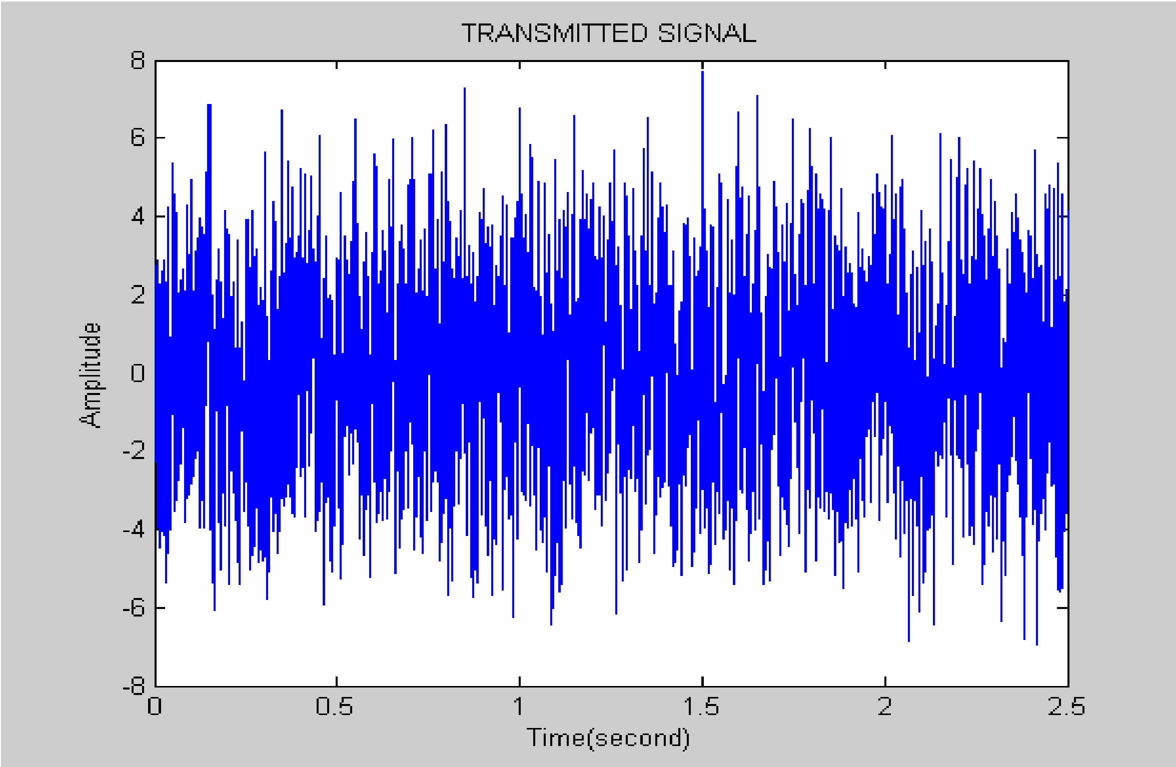

Figure 6.6 Modulated signal sent form transmitter to receiver for SNR=0 and Stream Length =10 bits.

6.2.8 Receiver Part of the Program

Frequency Spectrum of the Transmitted Signal

In this program, we find the furrier spectrum of the transmitted signal and select the signals sent on the four channels according to channel frequency band with. As you see, the furrier spectrum of the **signal in figure (6.3).**

Mathlab code of taking furrier spectrum of the signal

```
ffttsignal=fft(tsignal)
ffttsignal1=abs(fftshift(ffttsignal))
```

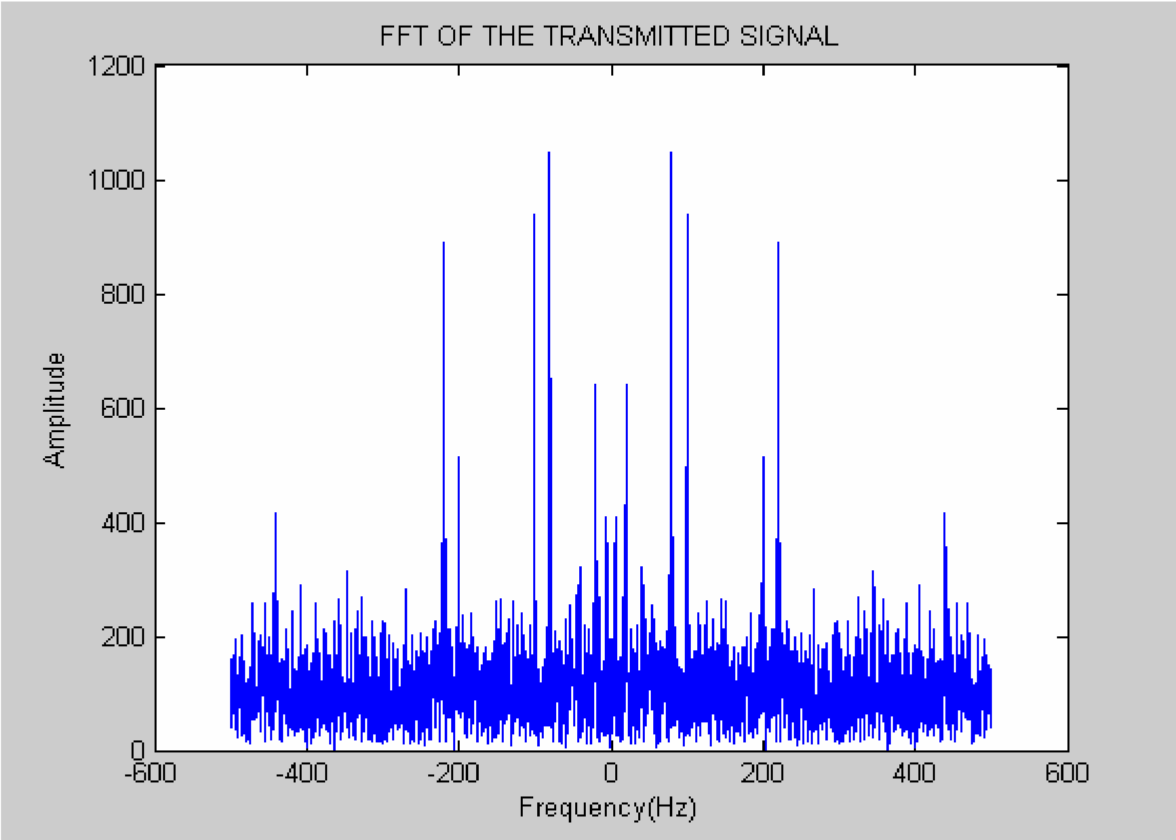

Figure 6.7 Frequency spectrum of the signal taken from the receiver.

Filtering

Channel 1 = 0-30 Hz
Channel 2 = 30-90 Hz
Channel 3 = 90-210 Hz
Channel 4=210-500 Hz

And then, we use the filters to take the modulated signals from the receiver. If we filtered transmitted signal with the band pass filter having bandwidth between 0-30 Hz, We will take the modulated signal (S1 (t)) at channel 1. If we filtered transmitted signal with the band pass filter having bandwidth between 30-60 Hz, We will take the modulated signal (S2 (t)). If we filtered transmitted signal with the band pass filter having bandwidth between 90-210 Hz, We will take the modulated signal (S3 (t)). If we filtered transmitted signal with the band pass filter having bandwidth between 210-500 Hz, we will take the modulated signal (S4 (t)).

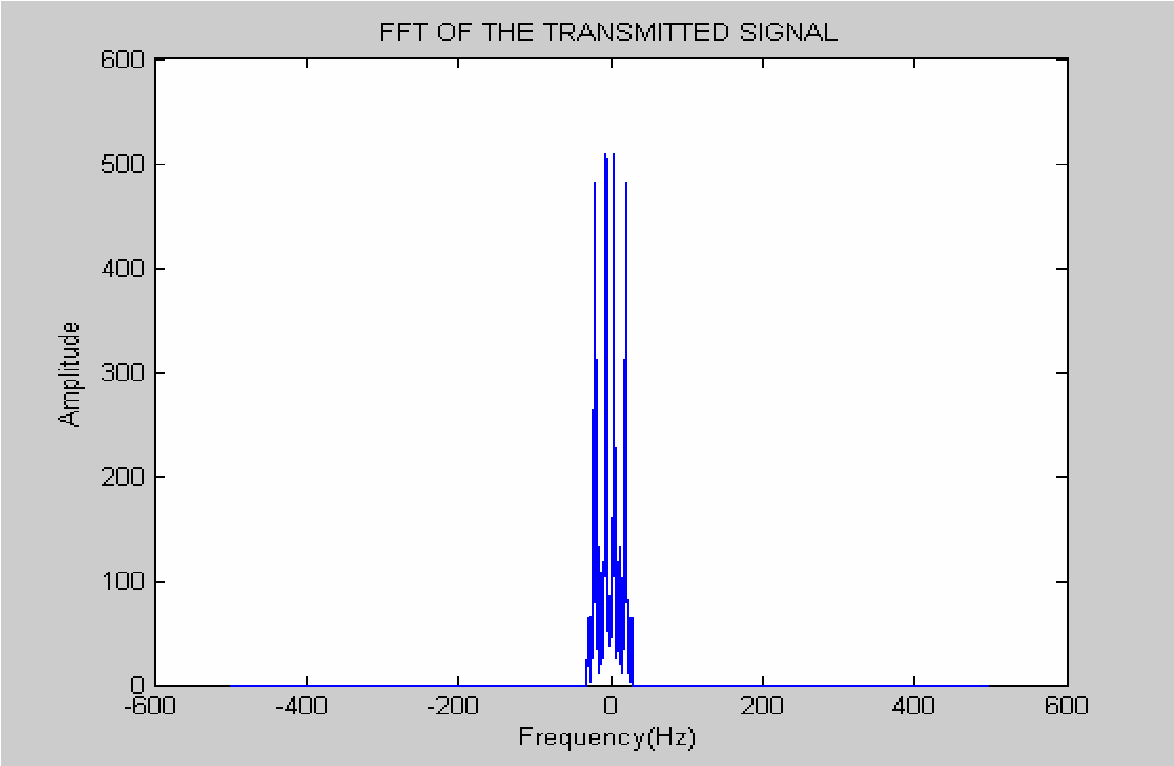

Figure 6.8 Frequency spectrum of the received signal (S1 (t)) having frequencies (5Hz and 20 Hz) .

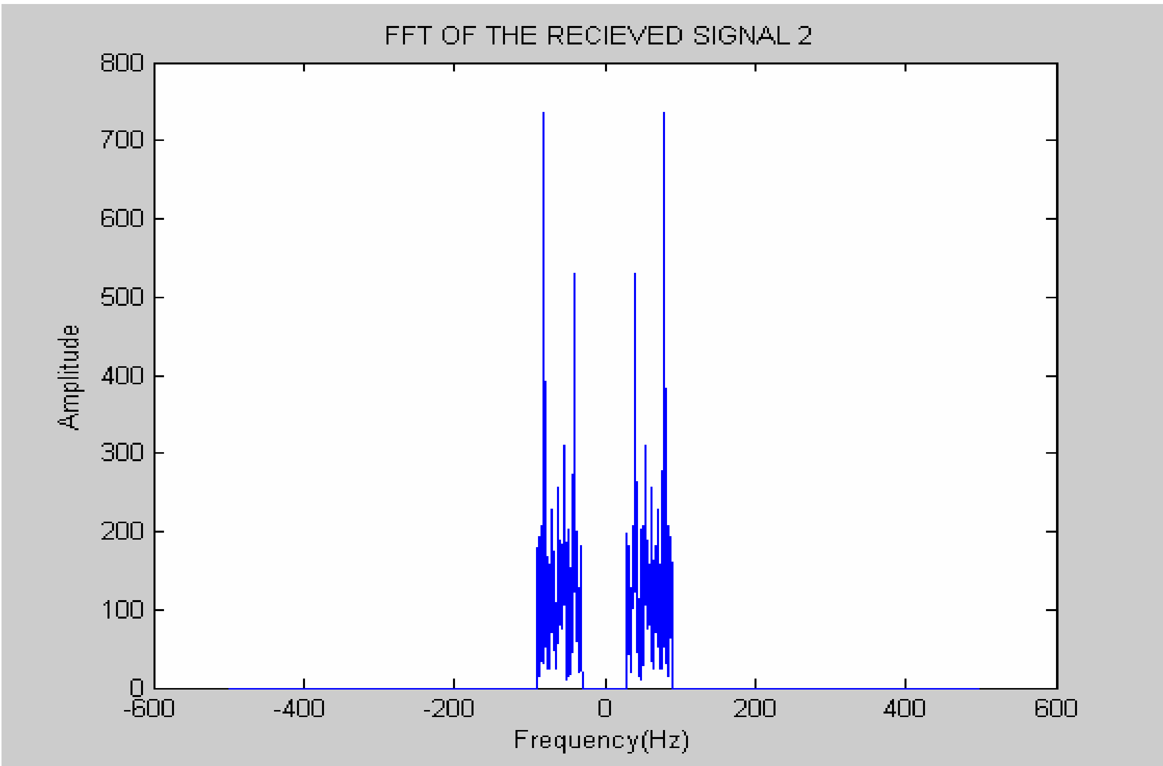

Figure 6.9 Frequency spectrum of the received signal (S2 (t)) having frequencies (40Hz and 80 Hz).

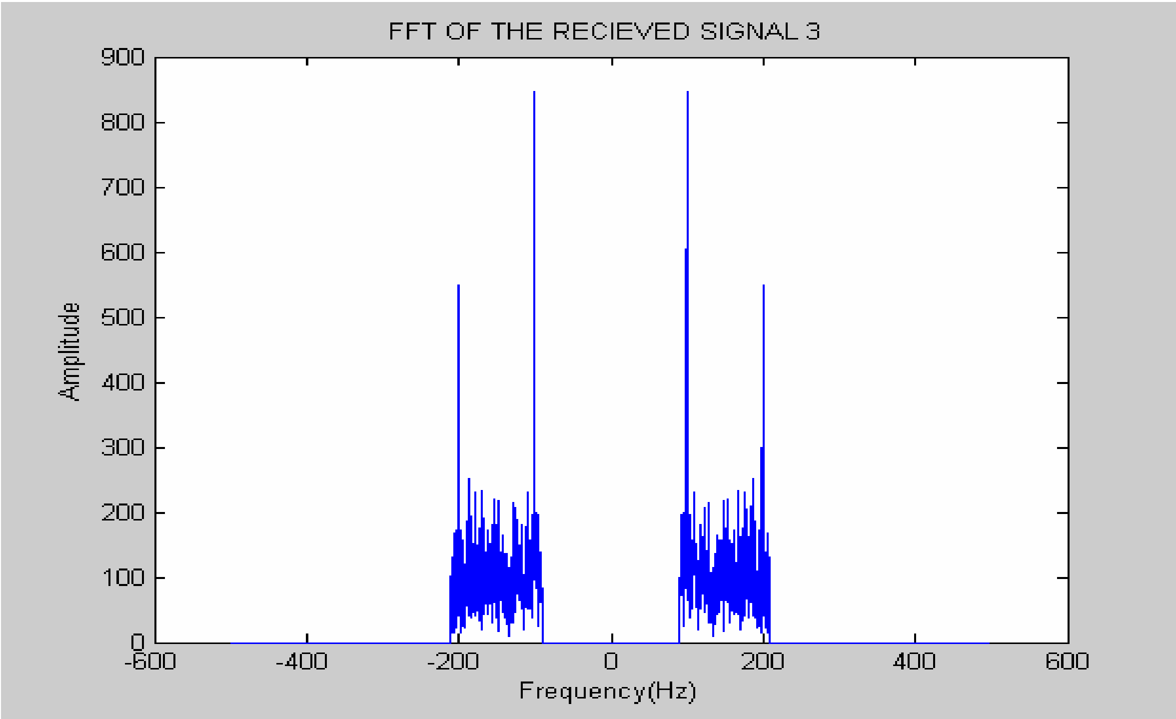

Figure 6.10 Frequency spectrum of the received signal (S3 (t)) having frequencies (100Hz and 200Hz)

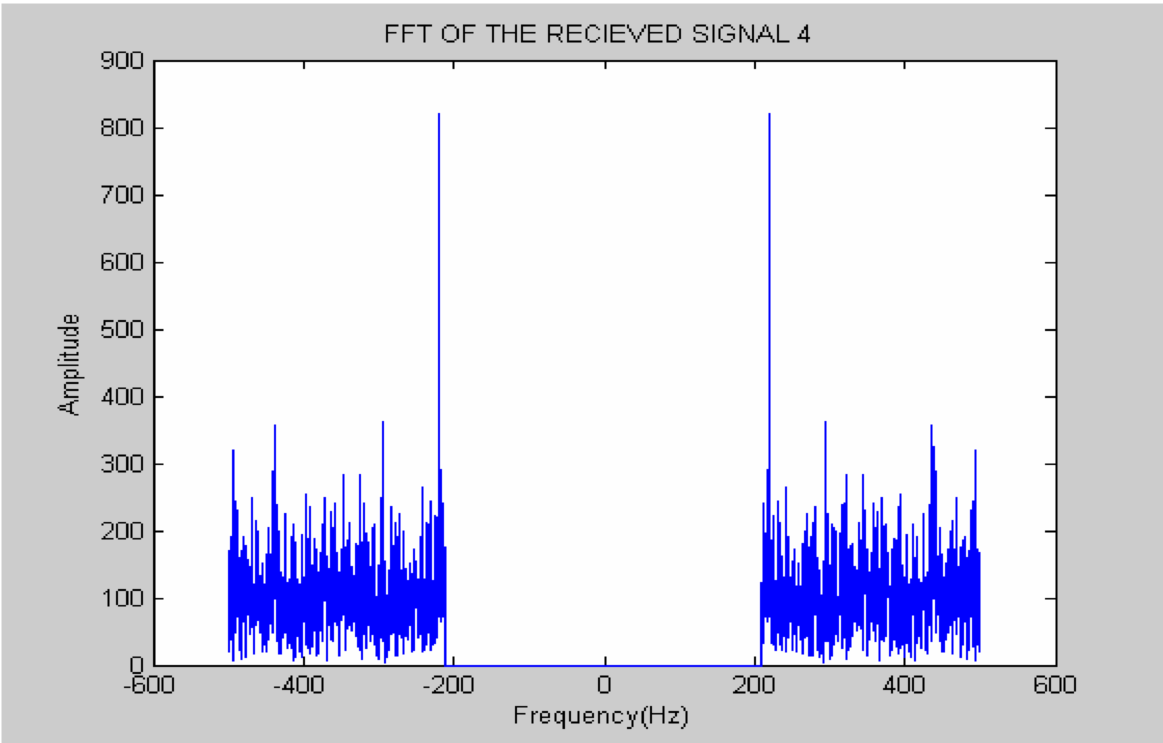

Figure 6.11 Frequency spectrum of the received signal (S4 (t)) having frequencies (220Hz and 440 Hz)

Mathlab code of the filtering

For Channel 1

```
for j=1:N2;
if  (j<=(f3+f2)*T2/2)||(j>=(N2-(f3+f2)*T2/2))
FFT1(j)=ffttsignal(j)
else
FFT1(j)=0
end
end
```

For Channel 2

```
for j=1:N2
if (j>=(f3+f2)*T2/2 & j<=(f4+f5)*T2/2)||(j<=N2-(f3+f2)*T2/2 & j>=N2-(f4+f5)*T2/2)
FFT2(j)=ffttsignal(j)
else
FFT2(j)=0
end
end
```

For Channel 3

```
for j=1:N2
if (j>=(f4+f5)*T2/2 &j<=(f6+f7)*T2/2)||(j<=N2-(f4+f5)*T2/2 & j>=N2-(f6+f7)*T2/2)
FFT3(j)=ffttsignal(j)
Else
FFT3(j)=0
end
end
```

For Channel 4

```
for j=1:N2;
if (j>=(f6+f7)*T2/2) & (j<=N2-(f6+f7)*T2/2)
FFT4(j)=ffttsignal(j)
Else
FFT4(j)=
end
end
```

In this process, we define the filter frequencies according to channel frequencies as you see above. After filtering frequency spectrum of the total signal according to channel frequencies, we take the received signal on wanted channels. Then we take the inverse FFT of the FFT of the received signal at the frequency domain. With this process, we reach the transmitted signal at the receiver part for all channels again.

Matlab code of Inverse FFT

IFFT1=real (ifft(FFT1))
IFFT2=real (ifft(FFT2))
IFFT3=real(ifft(FFT3))
IFFT4=real (ifft(FFT4))

IFFT1 shows received signal located at the channel 1. IFFT2 shows received signal located at the channel 2. IFFT3 shows received signal located at the channel 3. IFFT4 shows received signal located at the channel 4.In this process, there are a lot of corruptions at the received signal reaching from the transmitter at all channels. That's why sometimes we can not detect some bits reach from the transmitter because of noise and Inverse OF The Furrier Transform. We can see the all received signals for SNR (dB) and stream length (10 bits) on all channels below:

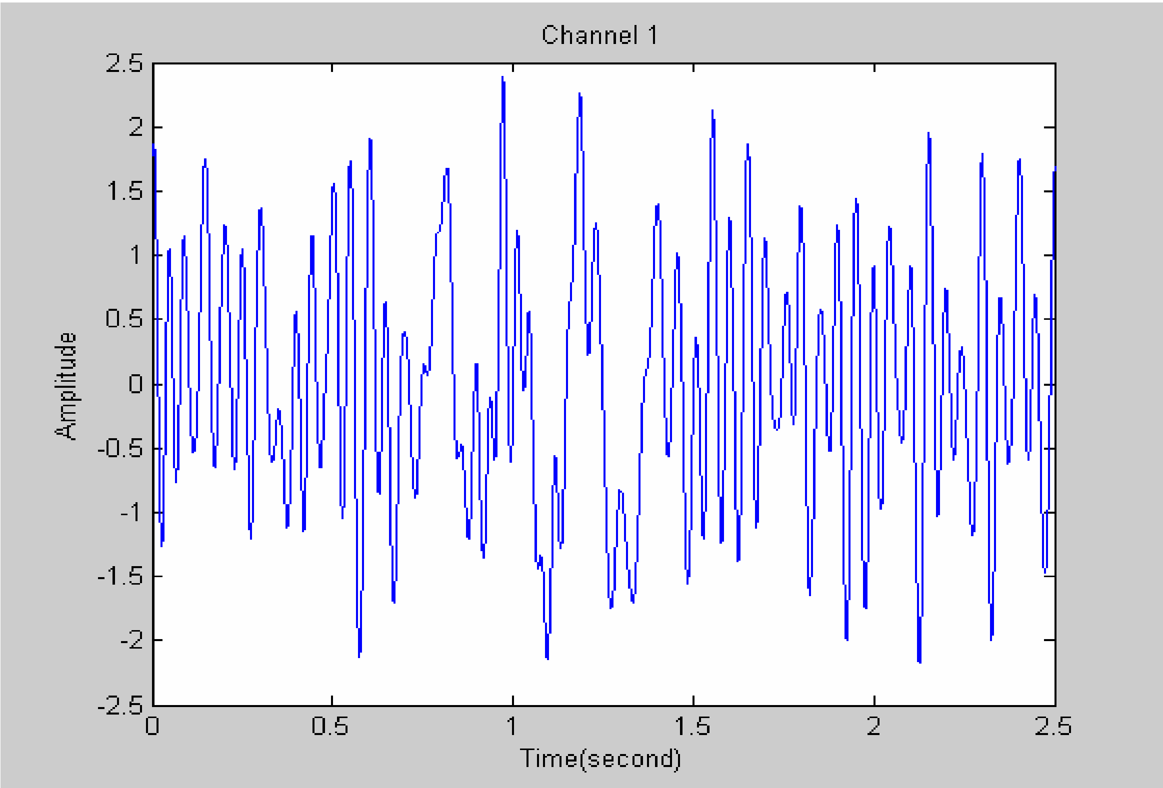

Figure 6.12 Received signal 1 for SNR =0 dB and Stream Length=10 bits

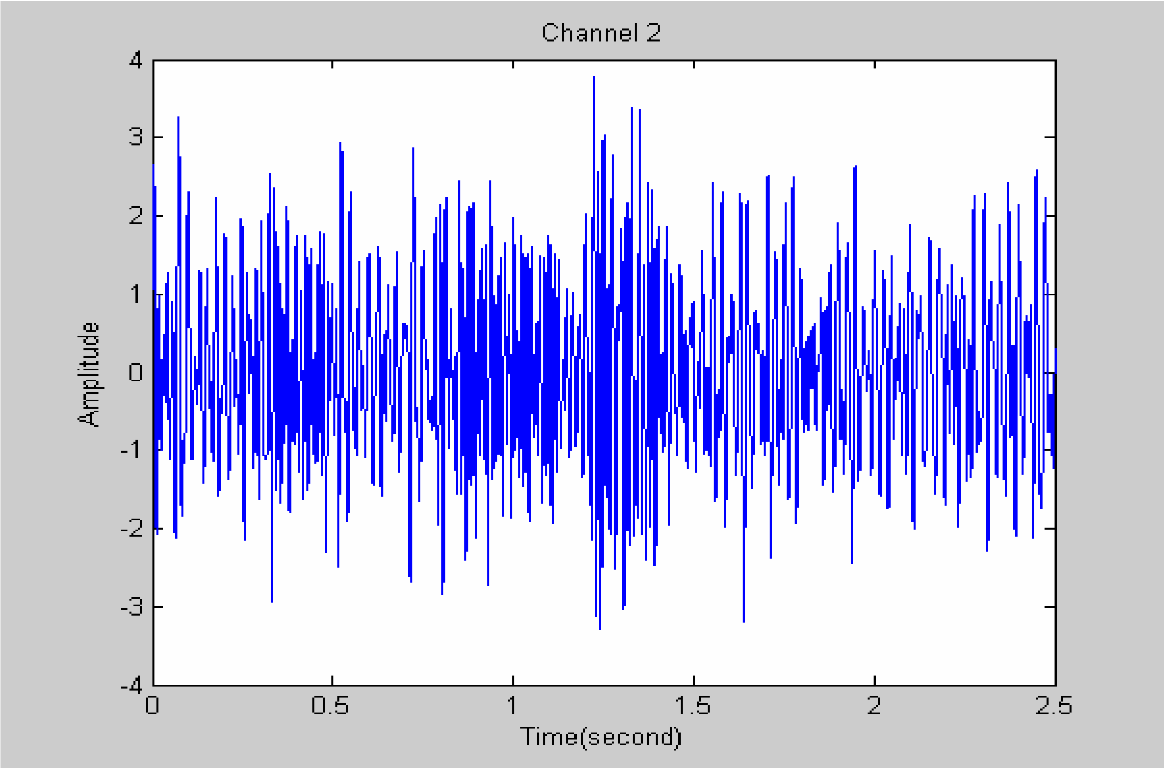

Figure 6.13 Received signal 2 for SNR=0 dB and Stream Length=10 bits

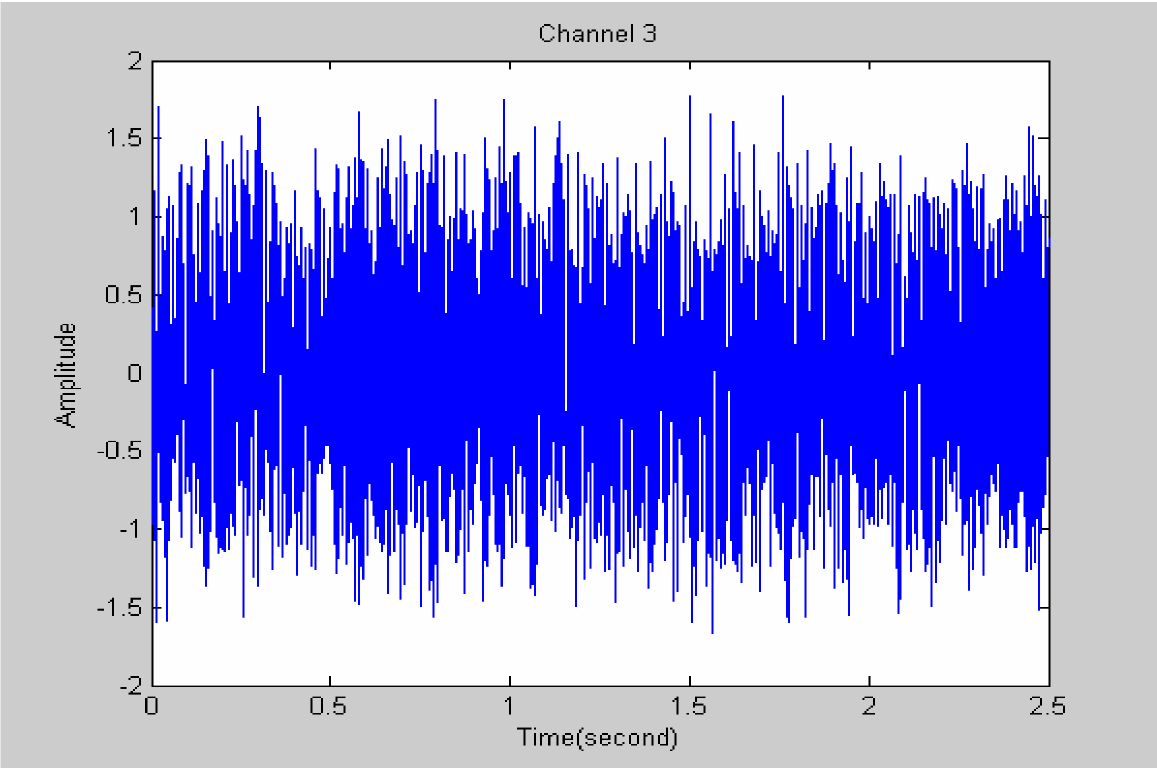

Figure 6.14 Received signal 3 for SNR=0 dB and Stream Length=10 bits

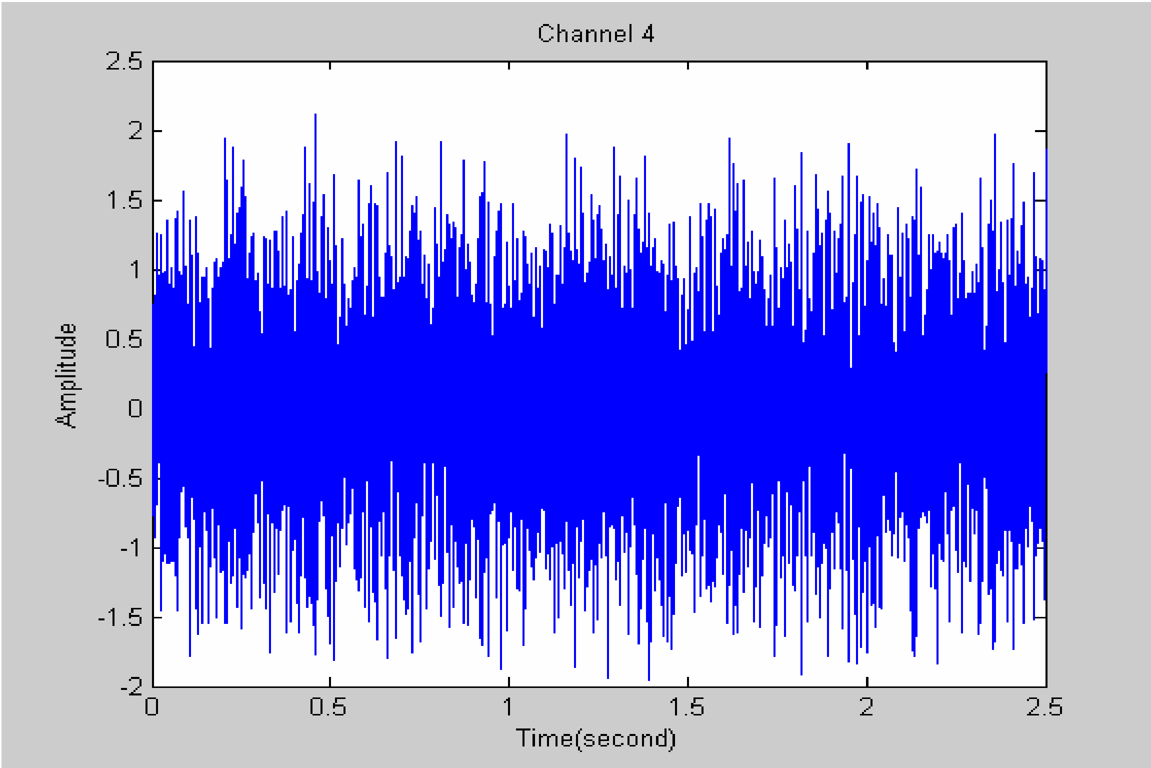

Figure 6.15 Received signal 4 for SNR=0 dB and Stream Length=10 bits

6.2.9 Demodulation

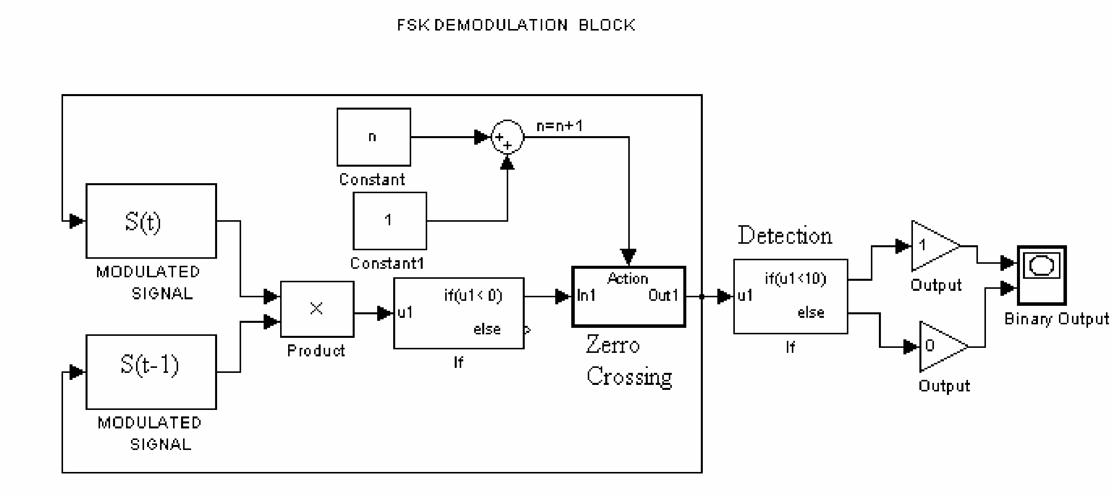

Figure 6.16 Demodulation block diagram

After we take the received signals on every channels, we must demodulate and detect the signals to determine the binary outputs (0 or 1).As you see block diagram above, we use the zero crossing numbers to determine the 1 or 0 outputs. Firstly, we define a threshold zero crossing number and we count the zero crossings numbers of the received signals. And if this number is higher than threshold zero crossing number , output bit is 1 or if this number is lower than threshold zero crossing number , output bit is 0 as you see in figure(6.9,6.10,6.11,6.12).

In this figure, firstly, we multiply the elements of the received signals with shifting one array number. if S(t)*S(t-1) higher than zero , don't increase the zero number or if S(t)*S(t-1) higher than zero , increase the zero number (n=n+1). With this method , we calculate the number of zero numbers for a bit interval. And then we must detect the bit as a 1 or 0.

At the Detection Part, we detect zero crossing numbers with the defined threshold zero crossing numbers. And we take the zero crossing number as a average of the zero crossing numbers of modulated signals. We define the binary output according to threshold zero crossing numbers as you see matlab code. Threshold zero Cross num = Zero Cross num of signal 1(F1) + Zero Cross num of signal 2(F2).
The demodulation and detection is processed by this matlab code. And in this mathlab code we take the threshold zero crossing numbers as T1*(f1+f2).This number is average of all zero numbers of a signal showing zero, and a signal showing ones.

Mathlab code of the binary demodulation and detection

For channel 1

```
dd=0; nn=0; n=0;
for k=1:N2-1;
if (sign(IFFT1(k+1))*sign(IFFT1(k)))<0;
n=n+1;
end
if mod((k),(N1-1))==0 & n>=ceil(T1*(f1+f2));
Recieve1 (dd+1) =1;
G (dd+1) =n;
dd=dd+1;
n=0;
elseif mod((k),(N1-1))==0 & n<ceil(T1*(f1+f2));
Recieve1 (dd+1) = 0;
G (dd+1) = n;
dd= dd+1;
n=0;
end
nn=nn+1;
end
```

For channel 2

```
dd=0; nn=0; n=0;
for k=1:N2-1;
if (sign(IFFT2(k+1))*sign(IFFT2(k)))<0;
n=n+1;
end
if mod((k),(N1-1))==0 & n>=ceil(T1*(f3+f4));
Recieve1 (dd+1) =1;
G (dd+1) =n;
dd=dd+1;
n=0;
elseif mod((k),(N1-1))==0 & n<ceil(T1*(f3+f4));
Recieve1 (dd+1) = 0;
G (dd+1) = n;
dd= dd+1;
n=0;
end
nn=nn+1;
end
```

For channel 3

```
dd=0; nn=0; n=0;
for k=1:N2-1;
if (sign(IFFT3(k+1))*sign(IFFT3(k)))<0;
n=n+1;
end
if mod((k),(N1-1))==0 & n>=ceil(T1*(f5+f6));
Recieve1 (dd+1) =1;
G (dd+1) =n;
dd=dd+1;
n=0;
elseif mod((k),(N1-1))==0 & n<ceil(T1*(f5+f6));
Recieve1 (dd+1) = 0;
G (dd+1) = n;
dd= dd+1;
n=0;
end
nn=nn+1;
end
```

For channel 4

```
dd=0; nn=0; n=0;
for k=1:N2-1;
if (sign(IFFT4(k+1))*sign(IFFT4(k)))<0;
n=n+1;
end
if mod((k),(N1-1))==0 & n>=ceil(T1*(f7+f8));
Recieve1 (dd+1) =1;
G (dd+1) =n;
dd=dd+1;
n=0;
elseif mod((k),(N1-1))==0 & n<ceil(T1*(f7+f8));
Recieve1 (dd+1) = 0;
G (dd+1) = n;
dd= dd+1;
n=0;
end
nn=nn+1;
end
```

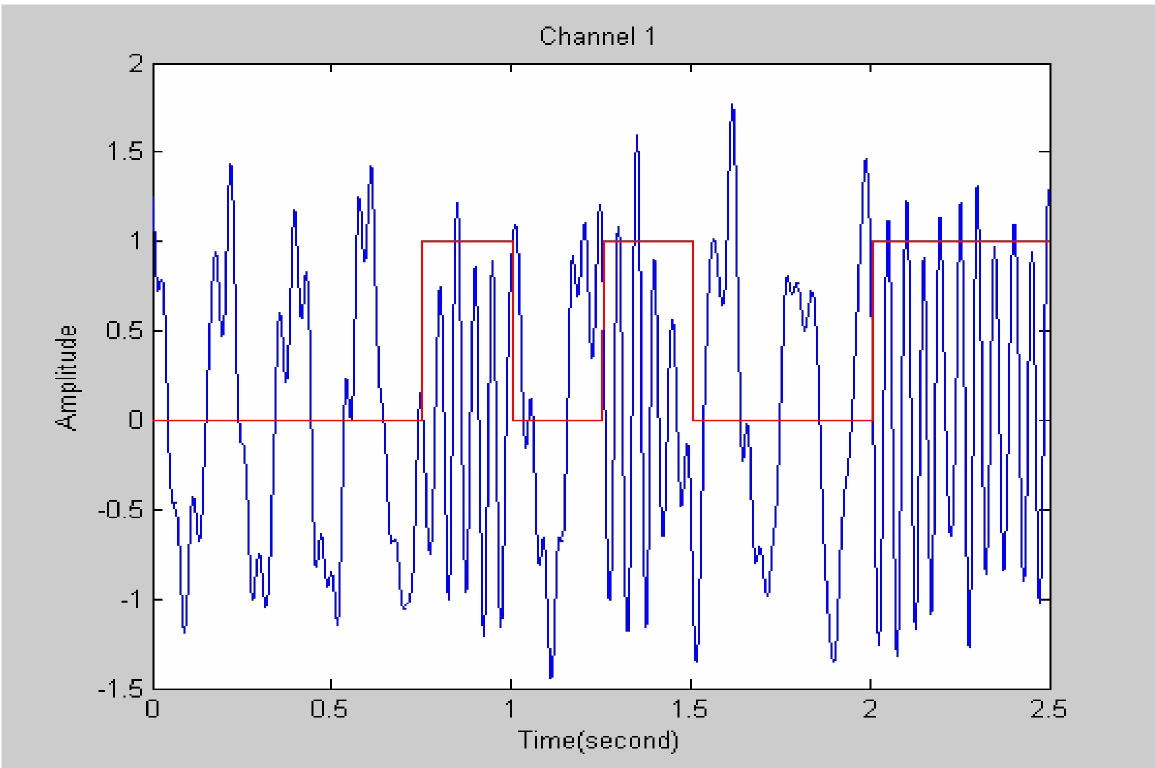

Figure 6.17 Demodulation and detection for channel 1 SNR=0 dB and stream Length=10 bits.

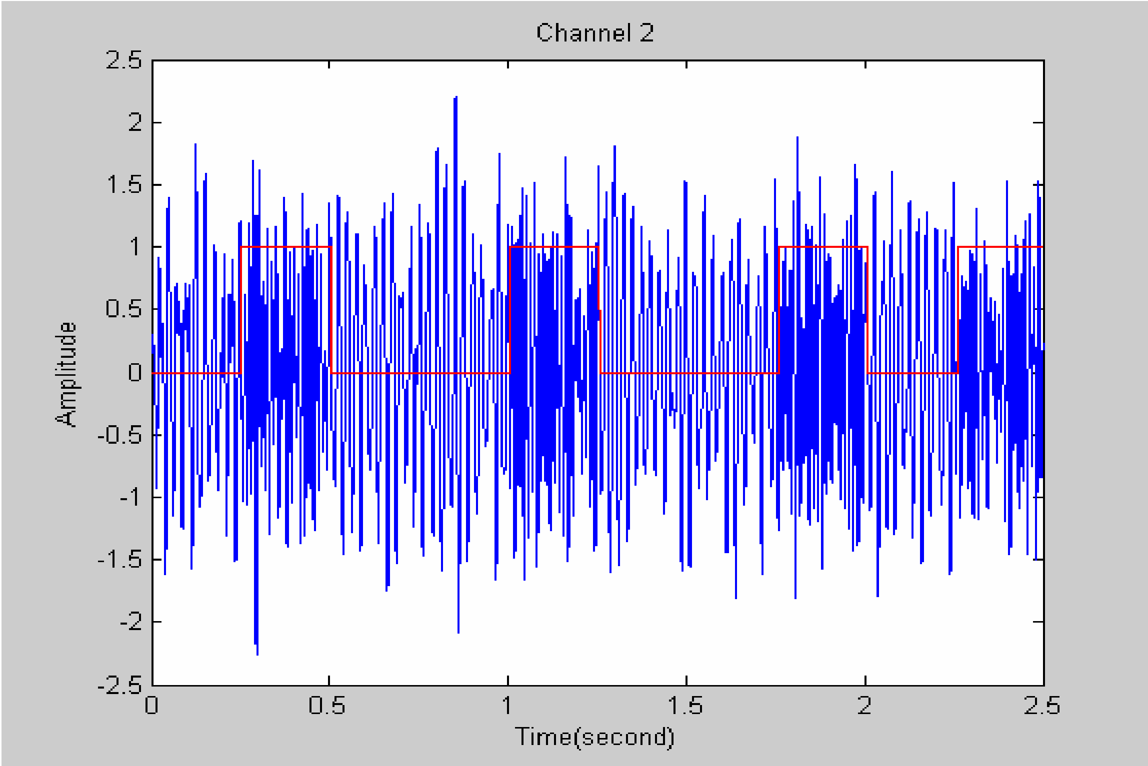

Figure 6.18 Demodulation and detection for Channel 2 SNR=0 dB and Stream Length=10 bits.

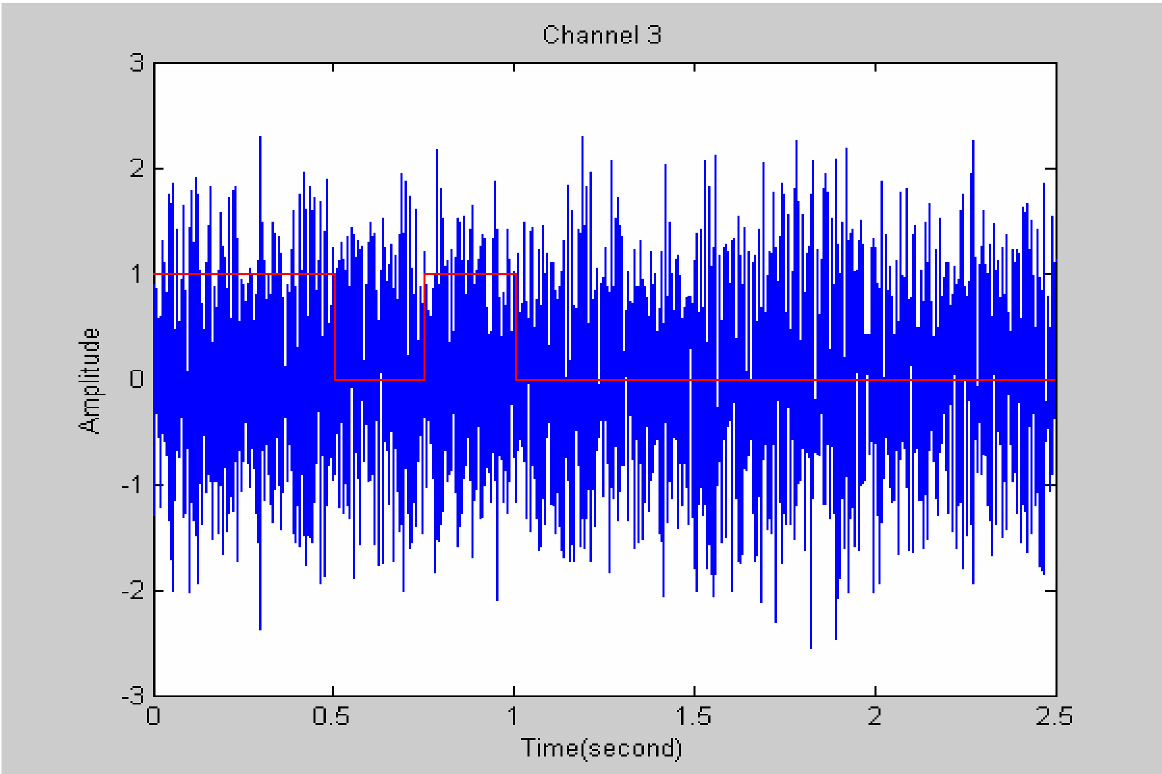

Figure 6.19 Demodulation and detection for channel 3 SNR=0 dB and Stream Length=10 bits.

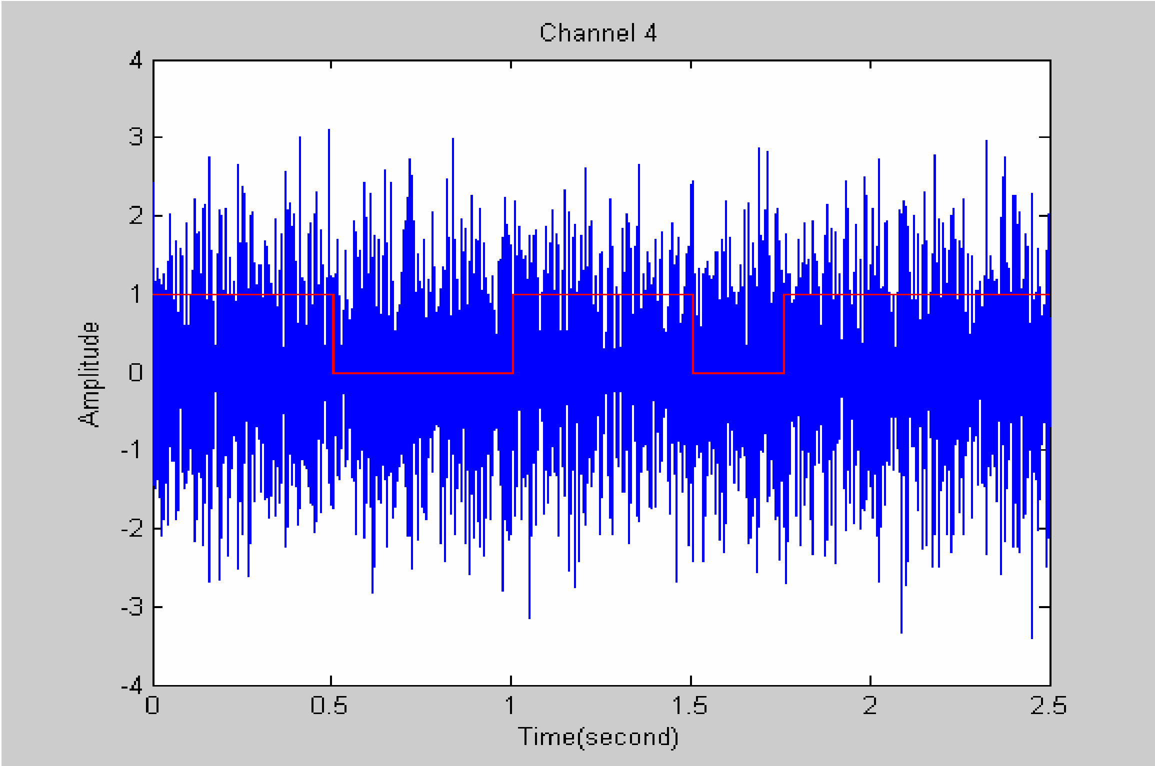

Figure 6.20 Demodulation and detection for channel 4 SNR =0 dB and Stream Length=10 bits.

6.2.10 Bit Error Rate

Bit error rate is calculated with ratio by number of corrupted bits to numbers of total bits on the bit stream. And we repeat this calculation for every channel. We find four error rate (E1, E2,E3, E4) in this system.

$$\text{BER}=\frac{\text{number of corrupted bits}}{\text{numbers of total bits}} \tag{6.13}$$

Error rate calculation for channel 1

```
nn=0; L=0;
for k=1:(N2);
if  mod((nn),(N1+1))==0;
L=L+1;
if s1(L)~=Recieve1(L);
E1=E1+1;
end
end
receive1(nn+1)= Recieve1(L);
nn=nn+1; end;
```

Error rate calculation for channel 2

```
nn=0; L=0;
for k=1:(N2);
if  mod((nn),(N1+1))==0;
L=L+1;
if s1(L)~=Recieve2(L);
E2=E3+1;
end
end
receive1(nn+1)= Recieve2(L);
nn=nn+1; end;
```

Error rate calculation for channel 3

```
nn=0; L=0;
for k=1:(N2);
if mod((nn),(N1+1))==0;
L=L+1;
if s1(L)~=Recieve3(L);
E3=E3+1;
end
end
receive1(nn+1)= Recieve4(L);
nn=nn+1; end;
```

Error rate calculation for channel 4

```
nn=0; L=0;
for k=1:(N2);
if  mod((nn),(N1+1))==0;
L=L+1;
if s1(L)~=Recieve3(L);
E4=E4+1;
end
end
receive1(nn+1)= Recieve4(L);
nn=nn+1; end;
```

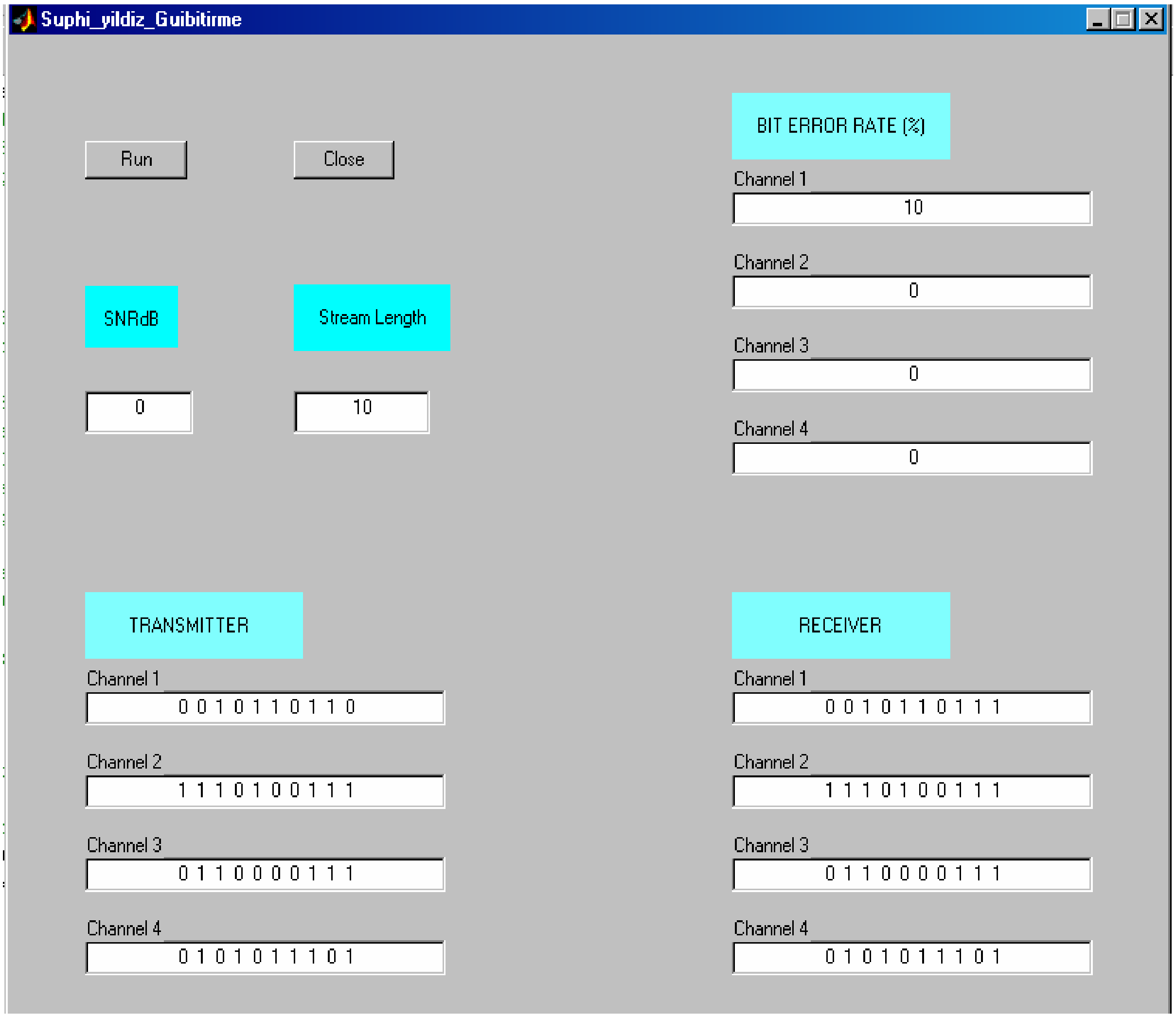

Figure 6.21 Bit error rate calculation for SNR=0dB and Stream Length=10

You can reach the program 'Channel Coding and Modeling with FSK and FDMA simulation' from disk locating at the end of the thesis.

7. Conclusions and Suggestions

In this project, there are two main parts, the theoretical structure and simulation structure parts. Firstly, we must construct the theoretical structure of the project before simulation. Because, building theoretical structure is the fundamental process to build simulation and to realize it in the real life. In this theoretical structure, firstly, we consider the signal propagation model. As you know, there are two main models to determine the propagation of the signal. One of them is deterministic and other is statistical model. In deterministic model we use the random noise and random Doppler frequency shifting and all variable about the noise depends on random values .Because , we can not know that where fading and frequency shifting are occurred. So, we use the deterministic model in this project. And we add the noise according to deterministic model and we build the parameter according to SNR (signal to noise ratio). As you see below:

$$SNR = 10^{\left(\frac{SNRdB}{10}\right)} \qquad \text{SNR} = \frac{P_s}{P_n}$$

$$s(t) = Vm\sin(Wct) \qquad Pn = \frac{V_m^{\ 2}}{SNR}$$

$$G\max = \sqrt{Pn} \qquad \text{Uk= rand (1, N2);} \qquad n_i(t) = 2 * G\max * Uk - G\max$$

Nk is the editable noise according to SNR value and variance of the original signal. At the same time, doppler frequency shifting is editable according to velocity of the receiver. But Doppler is not a problem for systems generally, because, this effect can be deleted by the filters. On the other hand, noise is a big problem for the wireless systems. Because, noise locates at all frequencies and it destructs the frequency construction of the wireless systems. We can decrease the affect of the noise to use some ways. We use the zero crossing number detection method in this project. And we see that, it is so suitable way to detect the sending bits and delete noise. In this project we use the deterministic model to define the signal transition from the transmitter to receiver.

$$S_R(t) = \sum_{i=1}^{N} V_i \sin[(w_s \pm w_D)(t - \tau) + \phi_i)] + n_i(t)$$

S_R = Transmitted signal form.
V_i =Amplitude of the modulation signals
w_s (2*pi* f_s) = Angles of the signals used for FSK modulation.
w_D (2*pi* f_D) = Doppler frequency shift.
ϕ =Phase shifting.
n_i =Noise taken from the air.

In conclusion, we try to take the modulated bit stream at the receiver part. After we take the modulated signal from the receiver part we detected and demodulated received signal. And we use the zero crossing numbers to detect the bits as '1' or '0'. At the end of the project we calculated received bit error rate. Bit error rate is generally %10 and %20 for SNR=0 dB and Stream Length is 10 bits for all channels. But if the FSK modulation frequencies of the channel are near as a value, bit error rate will be increase as a percentage.

In addition, we must calculate the maximum bit rate at for all channels. We can calculate the bit rate as you see below:

For first Channel having a bandwith (0-30) Hz:

Maximum Modulation frequency is 30 Hz. Bit Time $= T_s = \dfrac{1}{f_s}$

We calculate $T_s = \dfrac{1}{30Hz}$ We can transmit bits minimum at $5T_s$ time interval.

Maximum Bit Rate= $\dfrac{1}{5T_s}$

We can send maximum 6 bits for a second in this channel. Maximum bit rate is 6 bit/second.

For second channel having a bandwith (30-90) Hz:

Maximum Modulation frequency is 90 Hz. Bit Time $= T_s = \dfrac{1}{f_s}$

We calculate $T_s = \dfrac{1}{90Hz}$ We can transmit bits minimum at $5T_s$ time interval.

Maximum Bit Rate= $\dfrac{1}{5T_s}$

We can send maximum 15 bits for a second in this channel. Maximum bit rate is 15 bit/second.

For second channel having a bandwith (90-210) Hz:

Maximum Modulation frequency is 90 Hz. Bit Time $= T_s = \dfrac{1}{f_s}$

We calculate $T_s = \dfrac{1}{210Hz}$ We can transmit bits minimum at $5T_s$ time interval.

Maximum Bit Rate= $\dfrac{1}{5T_s}$

We can send maximum 42 bits for a second in this channel. Maximum bit rate is 42 bit/second.

For second channel having a bandwith (210-500) Hz:

Maximum Modulation frequency is 90 Hz. Bit Time $= T_s = \dfrac{1}{f_s}$

We calculate $T_s = \dfrac{1}{500Hz}$ We can transmit bits minimum at $5T_s$ time interval.

Maximum Bit Rate= $\dfrac{1}{5T_s}$

We can send maximum 100 bits for a second in this channel. Maximum bit rate is 100 bit/second.

In my opinion, this project is more wondered and more practical. This can be used in a lot of wireless system as channel multiplexing and modulating. And with this method, the signals coming from transmitter can be received to receiver with less destruction. These kinds of systems are used in a lot of communication systems.

At the next of the project, we can use the PSK to modulate the bit streams and TDMA to send the receiver at the multiple channels. Otherwise, we can use CDMA to send the signal to receiver at wanted numbers of channels.

8. References

1. Http;// www.radio-electronics.com ,(20.05.2006).

2. Molisch, F.M., (2001), *Wireless Communications,* West Sussex PO1983Q, England: John Willey & Sons .Ltd, page (1-24)

3. Tomasi, W., (2004), *Electronic Communication Systems Fundamental Through Advanced*, New Jersey Columbus, Ohio : Pearson Education Ltd, page 345- 503.

4. Uzunoğlu, M., (2005), *Kolay Anlatımlı İleri Düzeyde Matlab 6.0-6.05.* , Beyazıt, Istanbul : Türkmen Kırtasiye Tic. San.Ltd.Sti.

5. Welsolowski, K., (2001), *Mobile Communication Systems*, West Sussex PO1983Q, England: John Willey & Sons .Ltd, page (12-113).

10. Autobiography

I was born in 06.01.1983 in Sinop. I have a sister . I have lived in Istanbul since 1987. I finished to first school in Istanbul in Bagcılar. Then, I started to high school in 2001. I finished high school with 4.38 averages. And then I continued to graduate at DOGUS University. Firstly, I worked in a company, designing the electronic and communication devices at 2002 - 2005 in summers. I made my first summer practice in this company. I made my second summer practice in other company, solving the communication problem and processing the code to communicate terminals with Delphi in 2004. I made my last summer practice a company, servicing the telecommunication and telephone wide of the Turkey in 2005. It is the biggest company of the Turkey in this area.In addition, I worked in Software Company, writing code for graduation sets, commercial products and English learning sets, in 2005-2006.