ADDIS ABABA UNIVERSITY

SCHOOL OF GRADUATE STUDIES

Performance Analysis of Digital Subscriber Lines

By

Abel Seife-Selassie

Approved by Board of Examiners

Dr. Bayou Chane	
Chairman, Department Graduate Committee	Signature
Dr. Ing Hailu Ayele	
Advisor	Signature
Prof. P. Roy	
External	Signature
Prof. Wolde-Ghiorgis Wolde-Mariam	
Examiner	Signature
Dr. Nevin Kumar	
Examiner	Signature

Declaration

The thesis is my original work, has not been presented for a degree in any other university and all sources of material used for the thesis have been duly acknowledged.

Abel Seife-Selassie Date

____________________________ ____________________

(Candidate)

Dr. Ing Hailu Ayele Date

____________________________ ____________________

(Thesis Advisor)

Declaration

The thesis is my original work, has not been presented for a degree in any other university and that all sources of material used for the thesis have been duly acknowledged.

Abel Seife-Selassie : Date :

______________________________ ______________________________

(Candidate)

Dr. -Ing. Hailu Ayele : Date :

______________________________ ______________________________

(Thesis Advisor)

Acknowledgement

For this thesis work, a large number of personnel have extended very important presences, which I owe all. Firstly, my gratitude goes to my advisor Dr.-Ing Hailu Ayele. Had it not been for his invaluable comments, suggestions, and guidance, I would not have come up such a directional thesis work.

I also extend my appreciation to my colleagues and classmates in general and Dereje Haile-Mariam in particular for sharing with me his time, ideas and documents. Based on the similarities of our thesis modulation schemes, we had a lot in common to share and to complement. Accordingly the simulation work was geared to more practical presentation through those discussions.

I also would like to thank Ato Anteneh Ayalew for putting his precious time and friendly collaboration throughout the process of the M.Sc. Program, and especially during the period of thesis work. Similarly, Ato Mikiyas Sissay's and Ato Atli Lemma important assistance in the editing of this thesis paper is unforgettable.

Additionally, I am highly grateful to My Parents, B. General Seife-Selassie Alemu, Sister Emnet Alamyehu as well as my brothers, and most importantly my best half Tsege George for their encouragement and tolerance during my moments of solitude through out the period of the M.Sc. Program.

Lastly, I acknowledge W/o Kibrework Alemayehu, Ato Sintayehu Challa and others for their meaningful assistance in the whole period of the M.Sc. Program.

Contents

List of Figures

Chapter 2 The Twisted Pair Telephone Channel

Chapter 3 ADSL Modulations Schemes

Chapter 4 Discrete Multitone Modulation

Chapter 5 Simulations and Results

Abbreviations

ISDN	……	Integrated Services Digital Network
ADSL	……	Asymmetric Digital Subscriber Line
DMT	……	Discrete MultiTone Modulation
QAM	……	Quadrature Amplitude Modulation
MCM	……	Multicarrier Modulation
TEQ	……	Time Domain Equalization
FEQ	……	Frequency Domain Equalization
BER	……	Bit Error Rate
POTS	…….	Plain Old Telephone Systems (also known as PSTN)
PSTN	…….	Public Switched Telephone Network
PSD	…….	Power Spectral Density
CP	…….	Cyclic Prefix
TDM	…….	Time Division Multiplexing
FDM	…….	Frequency Division Multiplexing

Abstract

In recent years, lots of emerging high-speed applications and networking requirements are calling for cost effective access network solutions. These applications are well driven in the high bandwidth transmission media of the backbone network. However, the last mile transmission media that uses the existing modems of maximum speed 64 kb/s, are failing under these high-speed requirements. Firstly, this problem was approached using the very high bandwidth fiber network but it was seen that present time and foreseeable future end user requirements are only within reach to a speed of 2 Mb/s unjustifying the involved high cost. The practical means followed is to use the widely installed copper based access network.

It is the aim of this thesis to use standard techniques to investigate Asymmetric Digital subscriber Line, (ADSL). ADSL is a relatively new modem technology that provides high-speed data rates on the Plain Old Telephone Service (POTS) line. Efficient error control coding schemes and Discrete Multitone (DMT) modulation are the key mechanisms in ADSL. The objective of this work is to develop the software simulation modules for ADSL modem. In this thesis, DMT modulation is realized in an effort to demonstrate the end-to -end transmission of random generated bit streams through the twisted pair channel situated between the transmitter and receiver This simulation in done using a script code with MATLAB® and Bit Error Rate (BER) is used as a measure of performance.

Also largely described in this work are the physical properties of twisted pair channels and their frequency selective distortion. Moreover the high frequency behaviors of copper networks are highly varying among manufacturers. Accordingly twisted pair channel models are used using empirical or measured data.

The results of this work are performance measurement curves obtained from this end-to-end system simulation. These curves demonstrate the decreasing BER values against SNR agreeing with similar works of digital communications system simulation.

Performance Analysis
Of
Digital Subscriber Lines

By

Abel Seife-Selassie

Thesis submitted to the School of Graduate Studies of Addis Ababa University in partial fulfillment of the requirements for the Masters of Science Degree in Electrical Engineering

Advisor DR.-Ing Hailu Ayele

Addis Ababa University
Addis Ababa, Ethiopia

December 2003

Chapter 1

Introduction

1. 1 Present day and near future multimedia requirements

Present day telecommunications facilities and features are very much different from what the industry has been riding on for the last hundred years. The traditional telephony services are being engulfed by the newer digital services of universal data access. It is in the mid eighties that widespread deployment of ISDN basic access began and by now there are already emerging new services that are aimed at a widespread deployment in businesses and residences. Especially the Asymmetric Digital Subscriber Line (ADSL) is quickly achieving a high acceptance, being designed as a transmission platform, allowing several different services, among them being high-speed Internet access and Video on Demand (VoD).

To accommodate the bandwidth requirement of these newly introduced services, different approaches were followed that are largely based on broadband access solutions from an end-to-end connection point of view. The actual requirements, however, called for such largely broadband access only on a level of inter-service center transmission connections. For such realizations, a satellite, a Microwave radio or fiber links are chosen to be used to interconnect service centers and large business units. On the other hand, the service center to customer Small Office – Home Office (SOHO) access solutions could comfortably do with medium broadband access network that usually utilize links with a capacity of up to 8 Mb/s. The question then is “what kind of access link would give such capacities with reasonable cost, utilization of the already existing network and hence minimum installation time?”

As it is pointed out in this paper, one of the preferred alternatives in this regard is the use of the largely existing copper network after boosting the carrying capacity of this network. The other close access solution is the fiber network but it is left out for its relatively expensive cost of introduction into the network.

1. 2 Defining DSL

ADSL is one of the innovative access network technologies generally known as Digital Subscriber Line (DSL). The DSL is a technology, which uses powerful modulation techniques to tame telephone lines making them able to support broadband transmissions [1]. The unique advantage of the modulation techniques used in DSL is their ability to make the untapped frequency spectrum of the already widely existing copper network accessible for signal transmission. Additionally, DSL focus of media is the widely existing copper network; which in effect makes feasible realization of broadband access services at a reasonable cost. The modulation techniques used with DSL also give within-limits Inter-Symbol Interference (ISI) transmission capabilities for this copper network media.

With these and other advantages at hand, over the last fifteen years, xDSL (as it is some times known as) is underway with its different variants. Of the most common types, the first is High speed DSL (HDSL) which uses two or three telephone pairs to carry a two-way data transmission of 1.544 Mb/s (T1) or ~ 2 Mb/s (E1) speed. This is still used today to link small service points like PBXs and Mobile Base Stations (BTS) to the larger central nodes or station controllers. A more recent type of the DSL is the Very high speed DSL (VDSL) that is used to support largely broadband applications (up to 52 MB/s) over short distances.

1. 3 ADSL

ADSL was introduced in the late 1980's and is identified, among other specifications, with its different data carrying capacities for the two data transmission directions. This is called for from the observation of data access flow: usually residential and small office users would download larger data from their service centers than the amount of data they upload. ADSL is then the technology working with this Asymmetric data flow. Generally speaking ADSL provides up to 8 MB/s download speed and up to 640 KB/s upload speed.

Similar to most DSL systems, ADSL is realized with two modems at the two ends of the line. These modems usually take the transmission side copper line and user/service node side LAN, BNC (British Network Connector) or USB (Universal Serial Bus) connections. It is also common to find ADSL systems with lower edge of the transmission spectrum reserved for lower band applications like telephony or Basic rate ISDN. Hence ADSL is also used to transmit more than one service over the telephone line.

The advantages of ADSL stem from its powerful modulation schemes. The choice of a certain modulation scheme depends always on the specific requirements of the service that is to be realized with this transmission system. In the case of ADSL the choice is therefore especially difficult, as there are several different services to be transmitted over one common transmission line. ADSL is realized using two modulation schemes: a Carrierless Amplitude Modulation (CAP) and Discrete Multitone Modulation (DMT). In March 1993 the DMT system was chosen to be the basis of the ANSI standard [2] and this work also uses this modulation scheme.

1.4 Outline

This research is done in an attempt to characterize the ADSL media; the modulation techniques used in ADSL system and the simulation of the data transmission over a typical ADSL media (which is the telephone line). The research, subsequent to this first chapter, is structured as follows.

The Second chapter of this paper is a discussion of the telephone line considered as the media for ADSL system and data transmission environment. This chapter gives a review of the telephone loop plant and describes the characteristics of this media along with its dependence on several parameters. Therein, the imperfections of the copper based network are found to be the major drawback for the transmission of data. By the end of this chapter, a descriptive channel model is presented.

The third chapter covers a number of very basic components of DMT, namely, Quadrature Amplitude Modulation (QAM), Multicarrier Modulation (MCM), Bit Loading and Time Domain Equalization (TEQ). The fourth chapter is a more detailed block-by-block description of the DMT based ADSL system. In this chapter how the modulation techniques discussed in chapter three are combined to form the ADSL system is demonstrated.

Chapter five presents simulation process and results obtained. Accordingly, here clearly shown and discussed are the end-to-end processes of data transmission over the ADSL system using modulation process, effects of the channel, doing away with the effects of the channel and the demodulation processes. In the simulation work, the spectral characteristics of the transmitted signal and the realization of numerical computations using bit loading are presented. Chapter 6 concludes this document with suggestions and recommendations.

Chapter 2

The Twisted Pair Telephone Channel

This chapter is a description of the copper line access network from a switching center to the home or small office. This access network is a channel for the ADSL system and a closer look is made at its behaviors and capacities. Also presented is the effect this network has on data transmitted through it since this effect is extensively used in focusing on the remedies for mitigating the effects of this channel. Finally this chapter concludes with determining the channel model, which is used to reasonably simulate effects of a twisted pair network.

2.1 The Telecommunications Network

A typical telecommunications network is one that consists of the following components :-

A) ***Switching Centers*** **–** These are centers, which, are used to concentrate and route calls to their perspective destinations; usually based on the destination address. For this very reason they are also known as exchanges. Present day switches are usually of types known as a Stored Program Controlled (SPC) exchanges. In addition to their routing functions, Switching centers do also perform the important task of call charging and handling supplementary subscriber data.

B) ***Backbone Transmission Media*** **–** The presence of Switching Centers at more realistic separate locations initiates the interconnection of the centers with a transmission media of a reasonably wide bandwidth. Accordingly, this transmission part of the telecommunications network would do this job where the media can either be a radio media of Satellite, Microwave, HF, VHF or UHF type or cable media of fiber or copper wire type.

C) ***Access Network*** -- This part of the network is responsible for giving the subscribers access to the switches and completes the end to end process of communications. Twisted wire pairs are the dominating cable type in telephone access networks that are built for point-to-point two-way communication [3]. In Addition to this network media to the home, fiber cables and wireless access solutions are also being introduced recently.

The following diagram demonstrates these network components on the basis of two switching centers.

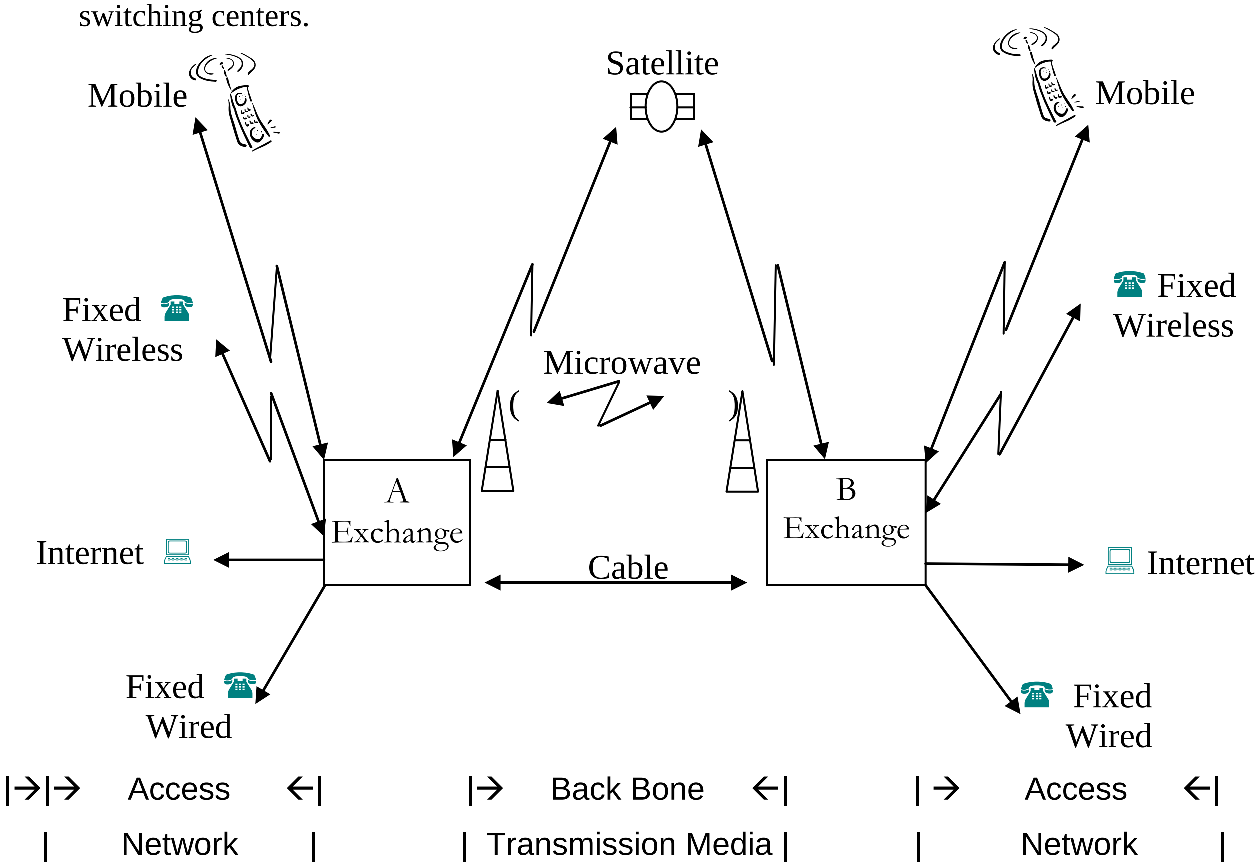


Figure 2.1 A Demonstrative Telecommunications Network

2. 2 Twisted Wire Pairs Characteristics

A telephone line consists of a series of segments of unshielded twisted pair (UTP) cable, with each segment possessing a different (and finite) distributed resistance, capacitance and inductance [5]. Figure 2.2 shows an equivalent circuit model for a transmission line segment.

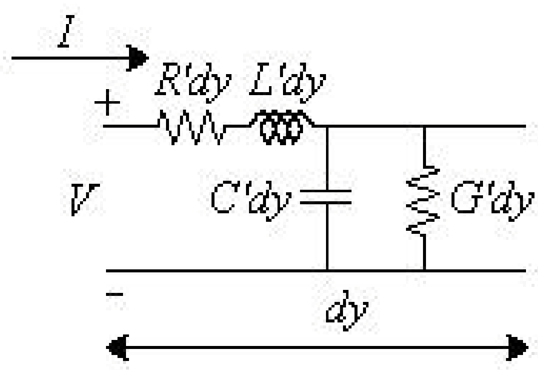


Fig 2.2 Transmission Line Segment Circuit Model

Shown below are physical constants for different cable types; Distribution and Main Cable types are of similar characteristics where as installation cables are of thinner dimensioning. Additionally for DSL applications, the high frequency spectrum is extensively used and the telephone line is considered as a low pass filter.

f (kHz)	R' (Ω/km)	L' (μH/km)	C' (nF/km)	G' (mS/km)
5	179	694,81	55,43	0,003
10	179	694,56	55,40	0,007
50	183	692,06	55,34	0,036
100	193	688,47	55,32	0,073
500	316	661,64	55,27	0,385
1000	438	640,00	55,15	0,789
5000	974	598,93	55,21	4,160
10000	1376	591,53	55,20	8,510
20000	1947	587,94	55,19	17,420
30000	2384	586,83	55,18	26,480

Table 2.1 Primary line constants of 0.5 mm underground cable

f (kHz)	R' (Ω/km)	L' (μH/km)	C' (nF/km)	G' (mS/km)
5	191	724,63	74,72	0,13
10	191	721,82	72,89	0,92
50	195	707,05	69,15	0,99
100	207	694,78	67,75	1,7
500	342	648,79	64,88	6,73
1000	475	624,65	63,81	12,19
5000	1056	578,22	61,62	48,34
10000	1493	565,74	60,79	87,5
20000	2112	557,33	60,03	158,39
30000	2586	553,87	59,61	224,12

Table 2.2 Primary line constants of 0.5 mm aerial cable

f (kHz)	R' (Ω/km)	L' (μH/km)	G' (μS/km)	C' (nF/km)
64	301	949	4	83
256	429	897	13	83
512	646	855	26	83
1000	925	815	47	83
4000	2024	759	156	83
8000	3037	742	298	83
16000	4645	731	586	83

Table 2.3 Primary line constants of category 3 installation cable (0.4 mm)

2. 3 Effects of Transmission in the Twisted Wire Pair

This subsection tries to thoroughly describe effects of the twisted pair. And based on these effects, the twisted pair telephone line is finally put as a communications channel for transmission of ADSL signals.

The makeup of the outside loop plant leads to imperfections of the transmission channel. Imperfection, in this case, means a deviation from an ideal bandlimited channel. An ideal bandlimited channel is characterized by zero noise, a flat transmission spectrum with unit gain inside the channel bandwidth as well as phase linearity with the frequency and having a zero amplitude response outside the band. There are a number of imperfections for data transmission over twisted pair including, but not limited to, amplitude and phase distortions, additive noise, Cross-talk distortions, impulse noise and echo. In this work, the closely looked at imperfections are amplitude and phase distortions, additive noise and the distortion known as intersymbol interference. These imperfections contribute to the general distortion picture and cross talk can be neglected since here the analysis of distortions is considered for ***a single*** twisted pair channel. Hence imperfections of a practical twisted pair wire are discussed and the effects of these imperfections are explained numerically, where possible, specifying some typical computed cases.

2. 3. 1 Amplitude and Phase distortion

The real channel composed of twisted wire pairs has a characteristic of increasing attenuation with respect to frequency and a slightly non-linear phase response. Several effects cause the increasing attenuation. For example the skin effect results in the resistance parameter R' increasing with the square root of the frequency. Also the increasing dielectric loss of the insulation material at higher frequencies, resulting in frequency increasing conductivity parameter G', contributes to the increasing attenuation. As can already be seen from these examples, the dependence of the primary line constants on frequency is non-linear (as can also be seen in tables 2.1, 2.2 and 2.3) and therefore also the resulting channel amplitude and phase responses are non-linear. It is also to be noted here that the higher frequency properties of telephone cables have never been specified and therefore have to be determined by empirical means. Moreover, beyond about 500 kHz these characteristics show a strong variation span from cable to cable due to manufacturing variety.

The attenuation and phase characteristics can be calculated based on these primary line constants. By equation 2.1 the transmission exponent is calculated from the primary line constants. Then equation 2.2 leads from this exponent to the frequency transfer function. Figure 2.3 shows the frequency dependence of the attenuation constant for a range of frequencies up to 1104 kHz, which is the actually defined range for ADSL transmission. Attenuation, in general, is undesirable as it lowers the energy of the received signal. Therefore, to keep the received signal level as high as possible, the low frequency portion of the loop is used first and the higher frequency portion is put to use only as a wider bandwidth is required. It can also be seen that especially in the frequency region below about 20 kHz there is extremely strong amplitude distortion [4]. Figure 2.4 reveals that for the three cables, the phase delay exponent is slightly non-linear; resulting in a dispersive effect as it results in different propagation speeds for different frequency contents, as can be seen from equation 2.3 and figure 2.5. The strong decay of propagation speeds at low frequencies, seen in figure 2.5, is due to delay characteristics, formed by the capacity of the wire pair together with the longitudinal resistance. For low frequencies this composition produces effects very similar to the ones produced by an R-C-circuit. The dispersive phase effects together with the non-flat amplitude response are the causes for ISI (intersymbol interference) that will be treated more in detail in 3. 3. 1. 2.

$$\gamma = \alpha + i * \beta = \sqrt{(R' + i * 2 * \pi * f * L') * (G' + i * 2 * \pi * f * C')} \qquad \ldots\ldots\ldots\ldots \quad 2.1$$

where γ Transmission exponent

α Attenuation exponent in Np/km (1 Np = 8,684 dB)

β Phase delay exponent in rad/km

R' Primary resistance constant in Ω/km

L' Primary inductivity constant in H/km

C' Primary capacity constant in F/km

G' Primary conductivity constant in S/km

i Imaginary operator (square root of -1)

$$H(f) = \frac{U_2(f)}{U_1(f)} = e^{-\gamma * l} \qquad \text{............ .... 2 . 2}$$

where H(f) Frequency transfer function

$U_1(f)$ Input voltage

$U_2(f)$ Output voltage

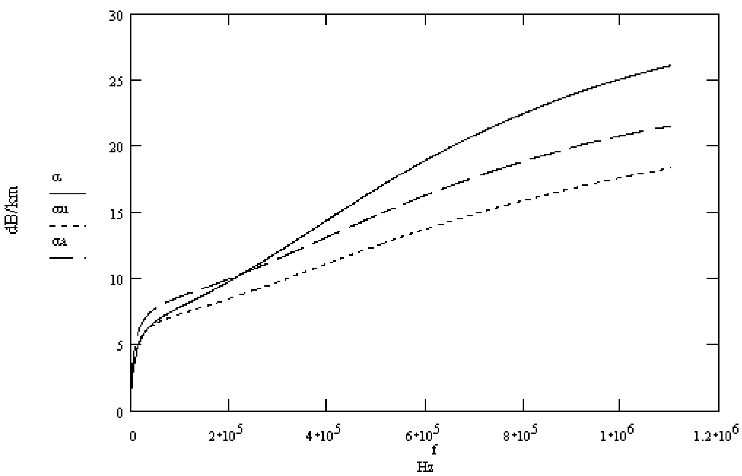
γ Transmission exponent

l Line length in km

$$c = \frac{2 * \pi * f}{\beta(f)} \qquad \text{............ .... 2 . 3}$$

where c Propagation speed

β(f) Phase delay exponent in rad/km



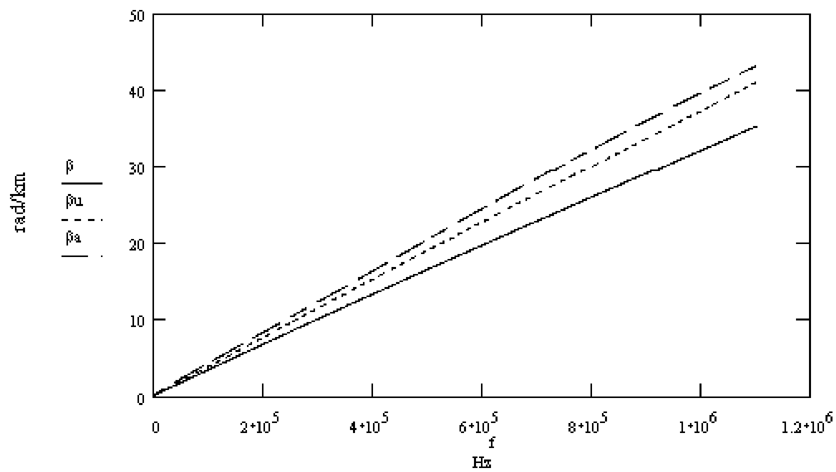
Key

α Category 3 installation cable (0,4 mm)

αu Underground distribution cable (0,5 mm)

αa Aerial distribution cable (0,5 mm)

Figure 2.3 Attenuation exponent as a function of the frequency



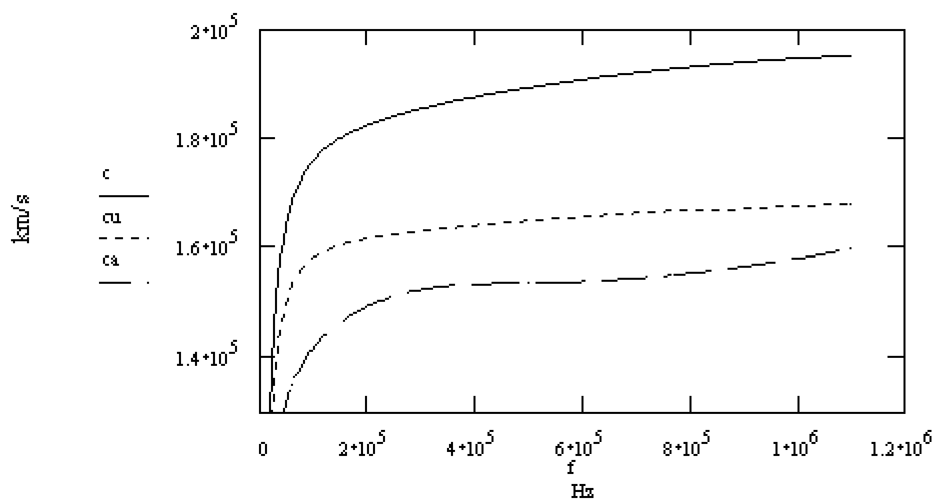
Key

β Category 3 installation cable (0,4 mm)

βu Underground distribution cable (0,5 mm)

βa Aerial distribution cable (0,5 mm)

Figure 2.4 : Phase delay exponent as a function of the frequency



Key

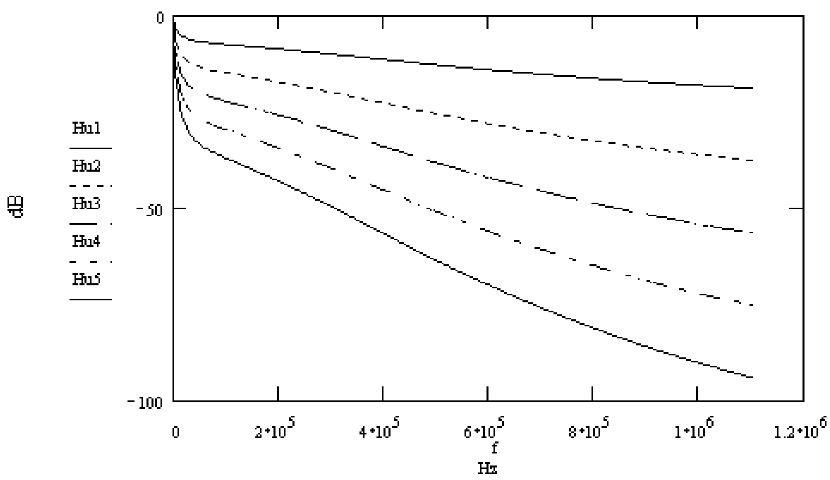
c Category 3 installation cable (0,4 mm)

cu Underground distribution cable (0,5 mm)

ca Aerial distribution cable (0,5 mm)

Figure 2.5 : Dependence of the propagation speed on the frequency for different cable types

Based on the above figures, the attenuation, phase and propagation speeds of common telephone cables over frequency ranges of over 1 MHz, the frequency behaviours of these channels can be arrived at with equation 2.2 using the physical line constant data given in tables 2.1 to 2.3. The following plots were prepared in such a manner.



Hu1 Power transfer function $|H(f)|^2$ for 1 km cable

Hu2 Power transfer function $|H(f)|^2$ for 2 km cable

Hu3 Power transfer function $|H(f)|^2$ for 3 km cable

Hu4 Power transfer function $|H(f)|^2$ for 4 km cable

Hu5 Power transfer function $|H(f)|^2$ for 5 km cable

Figure 2.6 : Power transfer functions for underground cable (0,5 mm)

Figure 2.6 demonstrates the power transfer function $|H|^2$ for underground distribution cables for lengths between 1 km and 5 km, respectively. Here it can be seen, the higher frequency portion is highly attenuated. This indicates that, in system design, a certain margin has to be left and this additional margin may be included in the general performance margin, defined when a DSL system is designed (see also 3.3 and Chapter 4).

Moreover, in the high frequency portion, the high total attenuation reaching up to values well beyond 80 dB [10] is in particular observed. These values are equal to the maximum dynamic range that an ADSL transceiver will have to cope with. On the other hand, loops with a very short length, therefore having hardly any attenuation, have to be considered also. This will be extremely severe for ADSL, as it is not restricted to shorter distance ranges, as for example VDSL, which is used over short distances. It can also clearly be seen in figure 2.6 that the wire pair channel has a low pass characteristic with its 3 dB frequency decreasing with increasing length. However, the non-flat amplitude spectrum of the twisted pair channel has not only effects in the frequency domain, but as well in the time domain as will be shown in the next subsection.

2. 3. 2 Intersymbol Interference (ISI)

As is mentioned earlier, frequency dependent propagation speed leads to a dispersive behaviour of the channel, resulting in a wider received pulse. This causes a time sample to spread also into the neighbouring time slots, or in other words a time sample to contain not only the contribution of the corresponding sent sample, but also of neighbouring samples.

But there is also a second effect to be seen. There exists a duality between the time and the frequency domain. Bandlimited signals in the frequency domain have theoretically infinitely extended pulses in the time domain and vice versa. This also means that the steeper the slope of a signal in the frequency domain, the wider it spreads out in the time domain. Therefore an exclusion of the low frequency portion with its extremely strong attenuation-distortion can also yield an improvement in ISI-behavior.

Equation 2.4 gives a closed expression for the ISI phenomenon [6]. The first term is the contribution of the corresponding sent sample, while the next two terms represent the contributions of the neighbouring samples; so-called postcursors and precursors. If the two ISI terms are equal to zero, the transmission is called ISI free. The last term represents contribution from noise. The delay time t_0 is included in order to use the same origin for the time axis for the received and the sent signal. It represents the transmission delay on the loop and can be seen in the figure 2.3 as the time (in sample periods) between the 0-point and the maximum of the impulse response (cursor). The condition for ISI free transmission can also be transferred to the frequency domain, as shown in equation 2.5. This condition is called Nyquist's first criterion [6].

$$r(t_0 + k*T) = x_k * h(t_0) + \sum_{j<k} x_j * h(t_0 + (k-j)*T) + \sum_{j>k} x_j * h(t_0 + (k-j)*T) + n(t_0 + k*T)$$

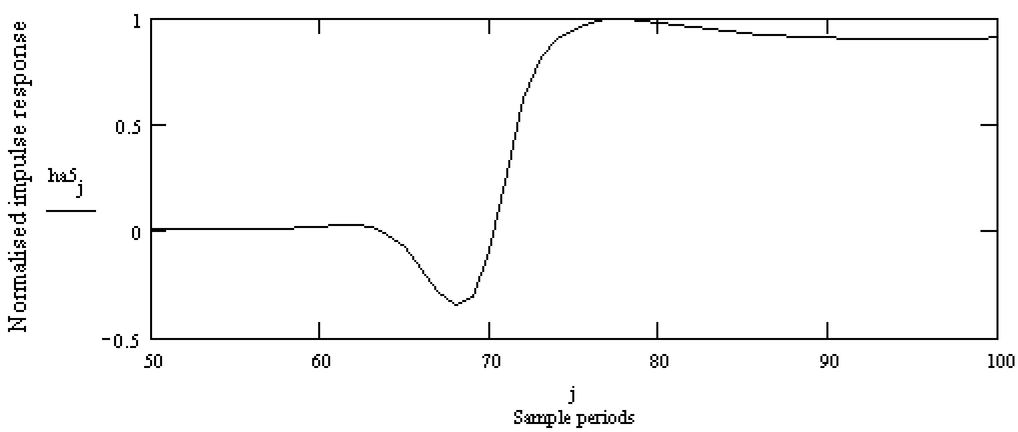
……….……………… 2.4

where
r Received signal sample
x Sent signal sample
t_0 Delay time
h Channel impulse response
n Noise sample

$$H_s(f) = H(f) + \sum_{n \neq 0} H(f - n*f_s) = const.$$

for $|f| \leq \frac{f_s}{2}$ …………..………….. 2.5

where
Hs Sampled frequency transfer function
H Frequency transfer function
f_s Sample frequency



ha5 5 km aerial distribution cable

Figure 2.7: Zoomed impulse response of 5 km aerial cable

The separation into pre- and postcursors is straightforward as there are different schemes used to deal with these two effects [7]:

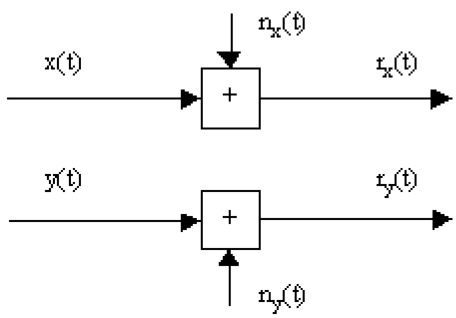
- Linear feedforward equalisation (precursors) [6]

- Feedback equalization (postcursors) [6]
- Gaps (guard time)
- Long symbol periods
- Viterbi maximum likelihood decoding with ISI states (postcursors) [8]

Of these methods several may be combined or they can be applied separately. The process to deal with ISI is called equalisation. Dealing with ISI becomes more complicated since the channel characteristics are normally not known beforehand, as a result adaptive solutions have to be used for time varying channels. More about the different equalization schemes that can handle ISI is to be found in section 3.4.

2. 3. 3 Additive White Gaussian Noise (AWGN)

Another impairment of data transmission (not only in the loop plant) is noise. One special kind of noise is Additive White Gaussian Noise (AWGN). Normally it is characterised by a double sided power spectral density in the transmission band. Figure 2.8 shows a frequently used model to represent the effects of AWGN [9].



$x(t), y(t)$	Transmitted signal in the in-phase and quadrature branch, respectively
$r_x(t), r_y(t)$	Received signal in the in-phase and quadrature branch, respectively
$n_x(t), n_y(t)$	Contribution of AWGN to the in-phase and quadrature branch, respectively

Figure 2.8: AWGN-model

Many research papers seem to neglect the effects of AWGN in a NEXT (Near end crosstalk) dominated environment but nevertheless AWGN influences the possible data throughputs. Especially for low frequencies, where the crosstalk coupling still is low, it creates a performance floor [11]. However, as frequencies increase, crosstalk noise becomes dominant and AWGN becomes negligible. The modeling of AWGN becomes especially important in environments where there is no crosstalk interference, because there AWGN may become the limiting impairment, especially for higher frequencies where the receive power level becomes low.

2. 4 Twisted Pair Channel Transmission Limitations

Unlike the rapid progress of data transmission speeds realized by advances in Very Large Scale Integration (VLSI) technology, there exist some theoretical limits beyond which reliable data transmission is not possible over a given channel. For every kind of data transmission over a specified channel, there may be considered two different theoretical limits that will be presented in detail in the following subsections.

2. 4. 1 Channel capacity

Shannon's capacity theorem of 1949 states the existence of a limit for reliable, i.e. error-free data transmission on every bandlimited channel, impaired with some kind of noise [12]. Error-free in this context means that there exists a coding and modulation scheme for every arbitrarily low constraint on the bit error rate (BER).

The channel capacity gives an important guideline in order to evaluate if a certain data transmission is theoretically possible and a means of comparison of different approaches of implementation. These comparisons are usually stated in terms of power difference in dB that is needed to actually transmit the signal at the channel capacity. With the most sophisticated modulation schemes, known at the moment (like DMT discussed in chapter 4), the difference for uncoded modulation is about 9.8 dB and can be reduced to about 3 dB by means of coding (trellis or block codes). At sufficiently high signal to noise ratios (and no self-crosstalk) an additional bit per symbol requires an approximately 3 dB higher signal to noise ratio, meaning in case of constant noise, 3 dB more transmit power. Shown below in equation 2.6 is the basic formula for the channel capacity.

$$C = \int_0^W \log_2\left(1 + \frac{S(f)}{N(f)}\right) df \qquad \text{.........................} \quad 2.6$$

where
C Channel Capacity in bit/s
W Upper band-edge of the transmit signal
S(f) Signal power spectrum at the receiver
N(f) Noise power spectrum at the receiver

This basic channel capacity formulation can be modified as follows to represent a capacity computation over frequency subdivided system where each frequency division subchannel is of its own noise level as in a typical frequency selective distortion channel is.

$$C = \sum_{i=1}^{N} \frac{W}{N} * \log_2\left(1 + \frac{P_T(i) * |H(i)|^2}{N(i)}\right) \qquad \text{.........................} \quad 2.7$$

where
C Channel capacity (approximated)
N Number of subchannels
W Upper band-edge of the transmit-signal
i Subchannel index
$P_T(i)$ Transmit power in subchannel i
$|H(i)|^2$ Channel power transfer function for subchannel i
N(i) Noise power in subchannel i

2. 4. 2 Cutoff-rate

It has already been discussed that, the channel capacity gives the limit for infinite complexity of implementation. In reality every system is constrained by certain complexity (cost) requirements. In fact, due to the rapid advances in the field of VLSI-technology even more sophisticated and complex schemes become feasible at a tolerable cost. However, it is of interest to have a means of estimating the limits for realisable data transmission over a channel. Table 2.4 shows the cutoff-rates for the different models for a 0.5 mm wire. The data here is obtained using equation 2.8 with data from practical measurements [11], [9].

	Twisted Pair		Quad	
	POTS Mbits/s	ISDN Mbits/s	POTS Mbits/s	ISDN Mbits/s
Real World Model, 2 km	12,70	11,34	14,45	12,85
Real World Model, 4 km	10,19	9,35	11,52	10,44
Worst Case Model, 2 km	11,66	10,30	11,69	9,95
Worst Case Model, 4 km	3,76	3,85	4,23	2,92
Absence of Crosstalk, 2 km	24,89	21,60	24,89	21,60
Absence of Crosstalk, 4 km	15,47	12,71	15,47	12,71
Real World Model, 2 km, upstream	1,51	1,60	1,74	1,80
Real World Model, 4 km, upstream	0,99	0,82	1,22	1,02
Worst Case Model, 2 km, upstream	1,51	1,62	1,87	2,03
Worst Case Model, 4 km, upstream	0,99	0,84	1,35	1,25
Absence of Crosstalk, 2 km, upstream	2,10	2,41	2,10	2,41
Absence of Crosstalk, 4 km, upstream	1,57	1,62	1,57	1,62

Table 2.4 Channel cutoff rates for different models in 0.5 mm underground distribution cable

For all the presented models and distances the cutoff-rates are higher than the E1-rate, being considered the lowest rate for ADSL-transmission in Europe.

$$R_0 = \int_0^W \log_2 \left(1 + \frac{S(f)}{2 * N(f)}\right) df \qquad \text{..........................} \quad 2.8$$

where

R_0 Channel Capacity in bit/s

W Upper band-edge of the transmit-signal

$S(f)$ Signal-power-spectrum at the receiver

$N(f)$ Noise-power-spectrum at the receiver

2. 4. 3 Realizable transmission rates considering a safety margin

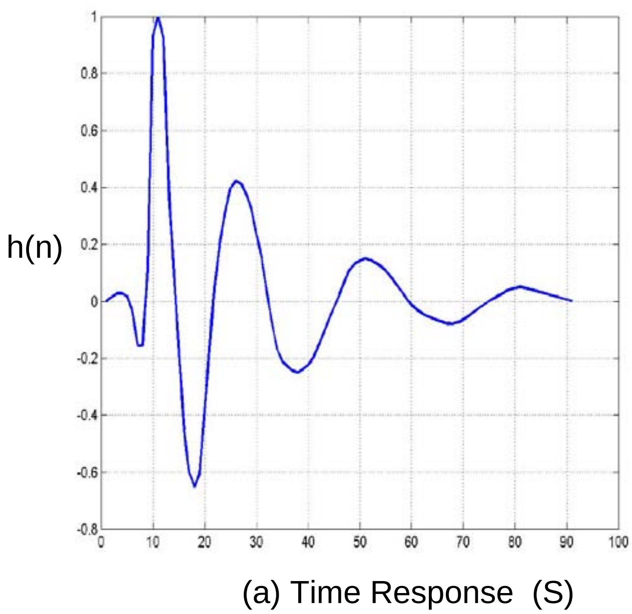
The channel capacity states the absolute theoretical limit for data transmission where as the cutoff-rate is supposed to give the absolute limit for realizable data transmission. However, all the analytic models, as they have been used in many studies, do not contemplate certain impairments, such as impulse noise, radio-interference or additional loss due to discontinuities; since these impairments are difficult to predict or estimate. Therefore, in order to design a reliable system, an additional safety margin has to be incorporated to be able to cope with these other impairments. Special considerations on the BER have not been done in such works as it is assumed that an extra safety margin is to be considered. Hence a 6-dB safety margin used in relation with the cutoff-rate [12], [11] as it is usually done for analytic approaches [11]. Moreover, consideration of the BER would not have been possible in such a general form, as more precise knowledge about the modulation scheme and the symbol rate is required. This is due to the fact that the BER depends on the minimal distance between the signal points in the transmit constellation. Originally 12 dB had been used in these cases, but because the actual transmission systems are more and more nearing the theoretical limits and the analytic models are more and more refined, this requirement has been relaxed to 6 dB.

	Twisted pair		Quad	
	POTS	ISDN	POTS	ISDN
	Mbit/s	Mbit/s	Mbit/s	Mbit/s
Real world model, 2 km	9,47	8,45	12,21	9,96
Real world model, 4 km	7,00	4,49	8,30	7,56
Worst case model, 2 km	8,45	7,42	8,49	7,08
Worst case model, 4 km	1,99	1,40	2,58	1,59
Absence of crosstalk, 2 km	21,66	18,71	21,66	18,71
Absence of crosstalk, 4 km	12,24	9,82	12,24	9,82
Real world model, 2 km, upstream	1,12	1,18	1,40	1,39
Real world model, 4 km, upstream	0,65	0,43	0,88	0,62
Worst case model, 2 km, upstream	1,17	1,21	1,53	1,61
Worst case model, 4 km, upstream	0,65	0,45	1,02	0,84
Absence of crosstalk, 2 km, upstream	1,76	2,00	1,76	2,00
Absence of crosstalk, 4 km, upstream	1,23	1,20	1,23	1,20

Table 2.5: Transmission rates including 6 dB safety margin in 0,5 mm wire

2. 5 Twisted Pair (TP) channel model

Consistent with properties of the twisted pair transmission medium as observed in sections 2.3.1.1 and 2.3.2.2, it gives a ground that such mediums can usually be modeled taking a form of finite length impulse response models. In addition, as described in section 2.3.1.1, the coefficients of this finite impulse model are usually obtained from empirical or measured data encompassing the highly varying and not clearly specified high frequency behavior of the TP channel. A typical empirical data is plotted in figure 2.9 [10].



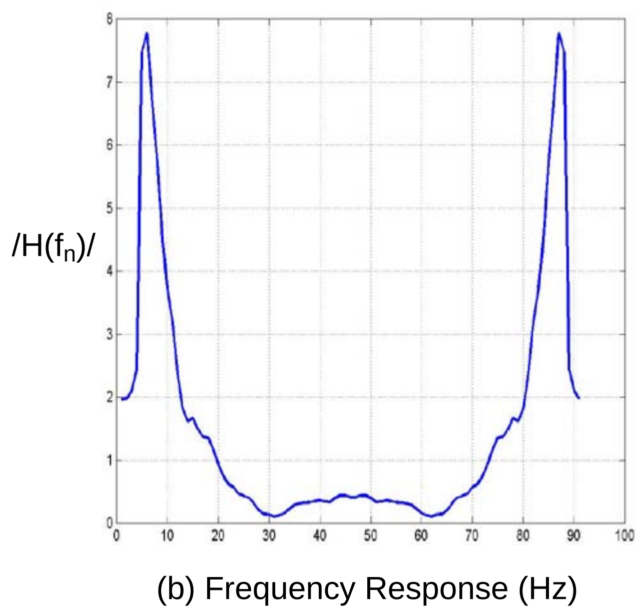


Figure 2.9 Empirical channel model

Chapter 3

ADSL Modulation Schemes

In digital communication systems, modulation is a process of converting the input bits into waveforms to be sent over the channel. While the modulated waveforms pass through the channel, noise corrupts it such that it is the corrupted waveforms that are available at the receiver. Therefore the demodulation is designed so as to recover the bits from the corrupted waveforms. Figure 3.1 illustrates a generic block diagram of a transmitter and a receiver in a communication system.

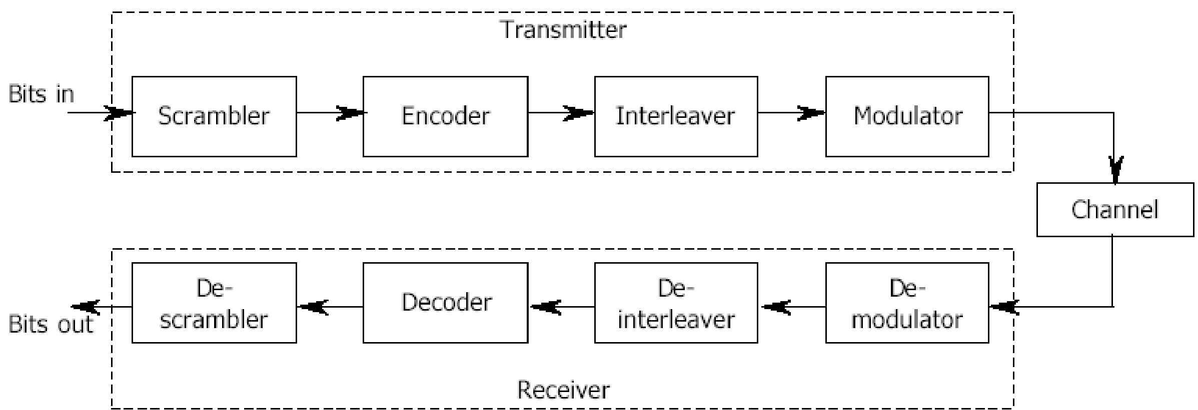


Fig. 3.1 Transmitter and receiver model

The decision for a certain modulation scheme depends always on the specific requirements of the service that is to be realized with this transmission system. In case of ADSL the decision is therefore especially difficult, as there are several different services to be transmitted over one common transmission platform. ADSL uses two modulation schemes: a Carrierless Amplitude Modulation (CAP) and Discrete Multitone Modulation (DMT).

DMT modulation is a common form of multicarrier (as in 3.2) modulation [1]. It was introduced by Peled and Ruiz of IBM in 1980 to take advantage of digital signal processors and Fast Fourier transform (FFT). The use of DMT for ADSL was first proposed by John M. Cioffi in 1991. In March 1993 the DMT system was chosen to be the basis of the ANSI standard [2] and this work also discusses this modulation scheme. A more detailed description of DMT is given in the next chapter. However, a number of building blocks from DMT end-to-end system are of a great importance and these are discussed more closely in this chapter. These blocks are also used in modulation systems other than DMT.

3. 1 Quadrature Amplitude Modulation

This is a modulation technique where a number of bits are mapped into two orthogonal carriers. For f(n) and q(n) be two sequences of carriers, the discrete form of orthogonality definition can be put as given below:

$$\sum f(n)\, q(n) = 0 \quad \text{for all values of index n}$$

Taking these carriers as real and imaginary planes, QAM can be realized with a mapping of data samples onto the complex plane to obtain symbols of variable amplitude and phase; as illustrated in the following example. Let an "m" bit data be QAM modulated. Then a look-up table is prepared to span all the 2^m possible symbols realized by the “m” bits. The mapping of bit streams to symbols and the QAM demodulating process of mapping of symbols to bit streams can then be taken as a straight forward table look-up mechanism.

The look-up table can be put in a more presentable manner using a 2-Dimensional plane of Real (Inphase) components and Imaginary (Quadrature) components. Here, "m" bit (2^m point) possible variations in these 2-Dimensional complex planes are known as constellations. A typical 4 bit (16 point) constellations are given as follows :

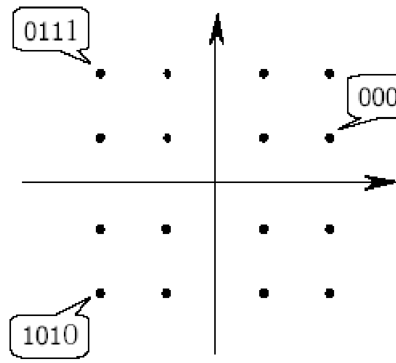


Figure 3.2 An example of 4-bit QAM

Errors on QAM based communications systems are observed only when a symbol value is corrupted; from effects of the channel or the noise. Such corrupted symbols are then seen to fall somewhere in the constellation plane – other than on the points of the transmitted symbol. A decision algorithm would determine the closest constellation point on the constellations plane and does the decoding Using table lookup) based on this point as a received symbol. The other focus of QAM modulations is that "how" usually, the constellation points are prepared. ADSL uses the standard constellation map specified by ANSI [9]. Fig. 3.3 shows the QAM constellation map of the size of 2^b (from 2^2 to 2^5). The label in the map equals the decimal value of binary bits to be encoded. The next even (odd) size constellation can be obtained from the previous even (odd) constellation by replacing each label n by 2X2 block of labels as follows:

$$\begin{matrix} 4n+1 & 4n+3 \\ 4n & 4n+2 \end{matrix} \qquad (3.1)$$

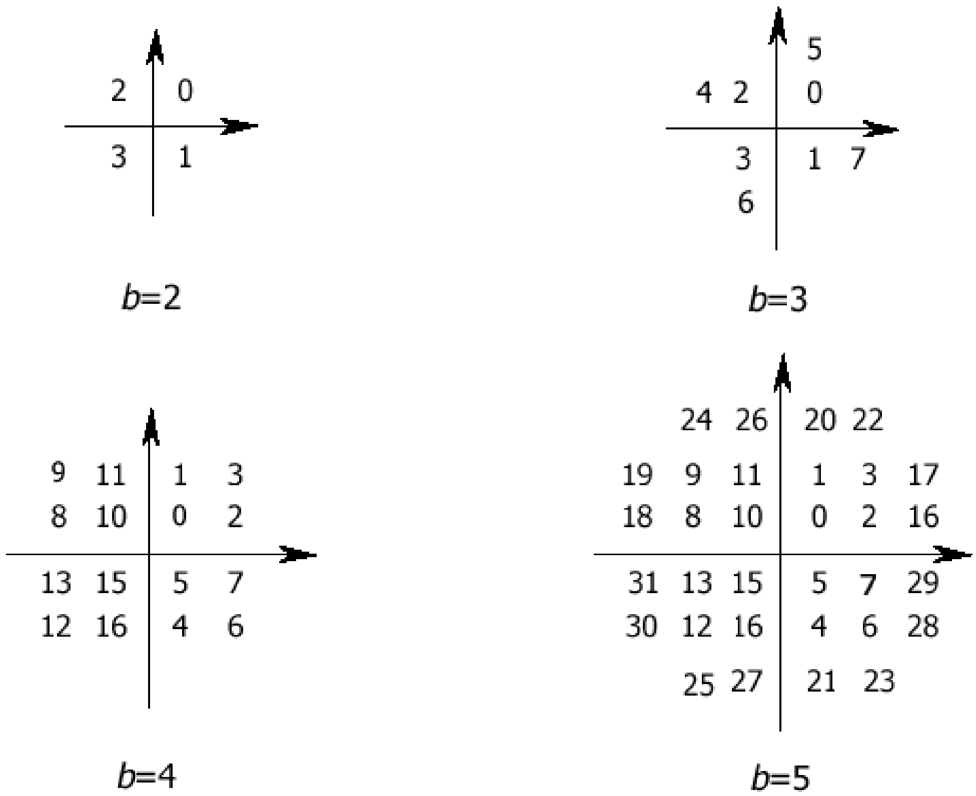


Figure 3.3 QAM Constellation Diagrams for m = 2 to 5

The constellation map described above is centered at the origin and has zero mean value. If we assume that all points in the map are equally likely, the average energy of such a QAM symbol is [16]

$$\varepsilon = \frac{M-1}{6} d^2 \qquad \text{... (3.2)}$$

Where $M = 2^b$ is the number of constellation points and d is the minimum Euclidean distance between two points.

3 . 2 Multicarrier Modulations (MCM)

MCM is also a modulation technique, which has come under utilization in high speed communication requirements where the communications channels are of limited bandwidth. Basically MCM is a frequency division multiplexing technique hence the whole communications channel at hand is divided into segments known as subchannels or tones. Here, a separate orthogonal carrier modulates data streams in each frequency subchannel at the same time. The data streams can either be from different data source hence representing multi-user usage or partitions from one longer data stream hence realizing larger symbol duration. The overall larger symbol duration so achieved (due within-channel flat spectral make-up and hence less ISI) is highly appreciated for relaxing the complexities of circuits designed to mitigate the effects of the communications channel, which can be realized, with the more ideal characteristics of the narrower subchannels. MCM is also chosen for its highly flexible spectrum usage giving a chance for its use of possibly different constellation size per each subchannel.

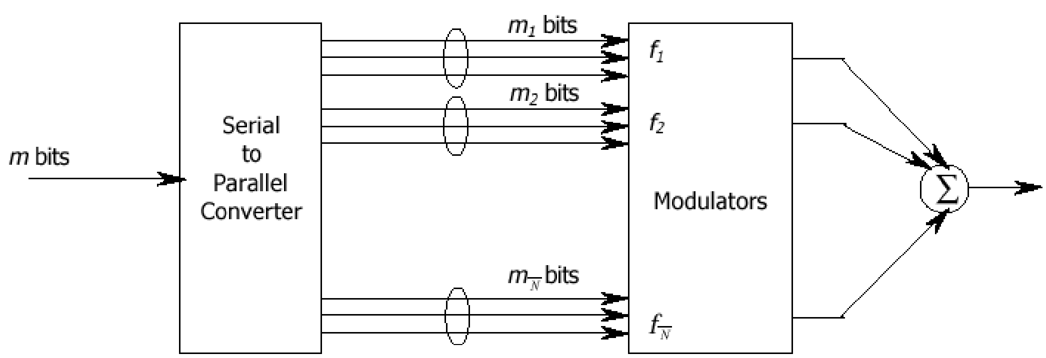


Figure 3.4 Multi-Carrier Modulation for m bits long data streams partitioned into N parts

MCM can be realized with different methods but the most common ones are mentioned below

1) Filter bank Methods (an old method and replaced by methods (2) and (3) given below).
2) Fast Fourier Transform (FFT) Based Methods
3) Wavelet Based Methods.

In all the above three realizations of MCM realization, the focus is on the orthogonality of carriers and care has to be taken to see if this is kept up with even after distortions caused by communications channels (see 4.1.2). Method (1) above has been replaced with method (2) after the discovery of orthogonality of carriers using the FFT methods. Method (3) is a recent approach for the realization of MCM.

For ADSL, DMT modulation uses the FFT realization and also it is used in this work for performance analysis simulation. Additionally in DMT, input data at each carrier of MCM are outputs of a QAM modulation and hence they are of complex nature. Moreover, the length of bits for each QAM at each carrier is variable making the constellation plane of somewhat variable size. This makes the system undoubtedly a little more complex but the rationale for the choice of variable bit lengths per carrier is significant; as it is presented in the following subsection.

3 . 3 Bit Loading

Due to the orthogonality of carriers in MCM based systems, each subchannel will act, as a channel having it is own behaviors such as the channel transmission capacity. As explained in section 2.4.1, the channel capacity depends on the spectral signal power, the spectral channel magnitude and the spectral noise per subchannel [13].

$$C = \sum_{i=1}^{N} \frac{W}{N} * \log_2 (1 + \frac{P_T(i) * |H(i)|^2}{N(i)}) \qquad \text{…………..………….. 3 . 3}$$

where C Channel capacity (approximated)

N Number of subchannels

W Upper band-edge of the transmit-signal i Subchannel index

$P_T(i)$ Transmit power in subchannel i

$|H(i)|^2$ Channel power transfer function for subchannel i

$N(i)$ Noise power in subchannel i

Hence one needs to consider the subchannel capacity so as to define how many bits are to be taken at a time for a QAM modulation. This would, on one hand, very much help in achieving transmission rates very near to the channel capacities. On the other hand, it helps one to dynamically avoid spectral points where the SNR is highly destroyed. This happens, perhaps with frequency selective noises like radio frequency interference as in the cases of amateur radio transmissions. The SNR is based on Spectral Signal Power which depends on the Maximum Amplitude of the QAM Symbol and hence constellation size. With typical cases of SNR values in TP cables within the ADSL frequency range (up to 1104 kHz) it is seen that the capacity of transmission would be range between 0-20 bits/Symbol/Hz. This variation, however, would create a high clipping noise and the industry uses a bit loading values of 0, 2-15 bits/Symbol/Hz as a standard [15].

3 . 4 Equalization

Another important idea in studies of digital communications, especially in the loop plant, is equalization; as ISI is one of the major impairments for reliable data transmission in such media. Multicarrier system is robust against ISI since the longer duration of multicarrier symbols achieved due to the less number of bits used per carrier provides higher immunity against delay spread and ISI. Moreover, even though, an equalizer is unavoidable in case of higher data rate and channels with extensive time dispersion, the complexity of the equalizers for multicarrier systems is different from that of single carrier systems as shown on the next paragraphs.

Basically the equalizers used in multicarrier systems such as DSL are the Time domain (TEQ) and Frequency domain Equalizers. The purpose of Time domain equalization in multicarrier systems is not complete removal but restriction of inter-symbol interference to that within the duration of the symbol duration. Figure 2.7 shows a part of the impulse response of 5 km aerial distribution cable. In the depicted case the cursor itself is hardly distinguishable from the neighboring ISI-contributions, for there is very little difference in signal strength between them. The resulting problems for the detection of received signals can easily be imagined.

In figure 3.5, the impulse responses for an underground distribution cable type are depicted for several lengths and over duration of 1000 sample periods ([17] and [18]) for base-band use without a lower band-limit. The calculations show that in case of no lower band-limit one sample overlaps long into of its neighbors and so contributes considerably to the received signal at these other sample instants.

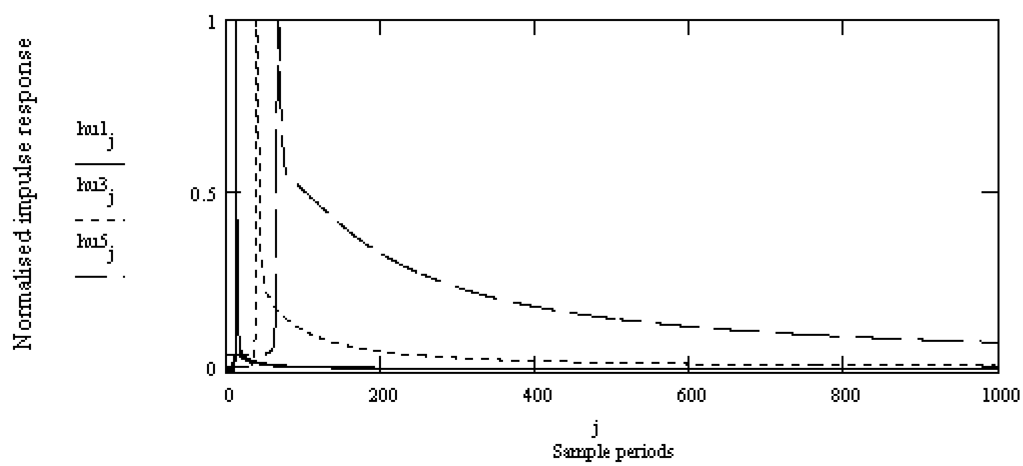


Figure 3.5 : Impulse responses of underground distribution cable for several lengths

where $hu1_j$ 1 km underground distribution cable

$hu3_j$ 3 km underground distribution cable

$hu5_j$ 5 km underground distribution cable

A general problem is that a good part of the received energy is contained in the ISI contributions, as can also be verified by the above diagram. Therefore, for optimal equalization the task can not only be to cancel out the ISI-contributions, but also to integrate them into the demodulation and decision process. In figure 3.5 it can also be seen how much the cable length and type influence the duration of ISI. This can be explained by the duration-bandwidth duality. If the bandwidth with growing length gets more and more restricted, then the signal expands in the time domain.

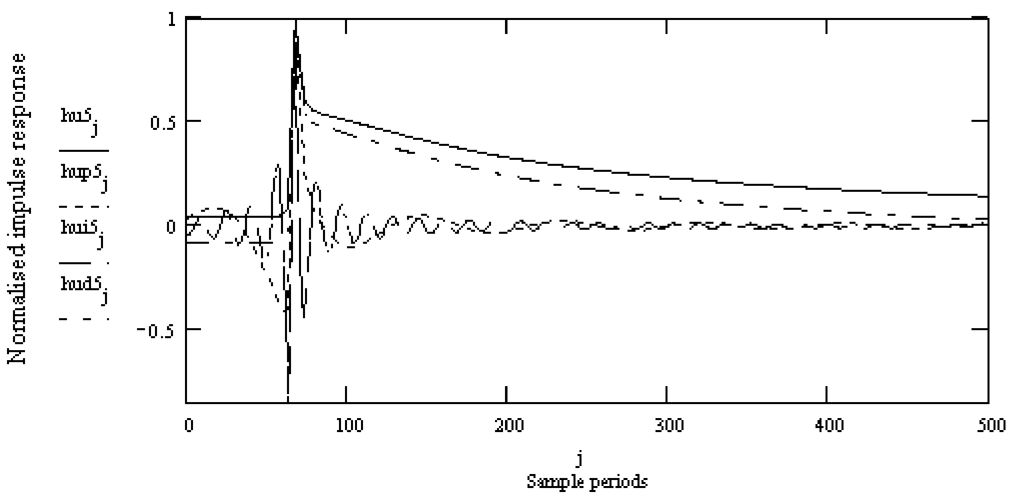


Figure 3.6 : Impulse responses of 5 km underground distribution cable for different lower band-edges

Where $hu5_j$ Impulse response without lower band-edge

hup_j Impulse response with lower band-edge at 25 kHz (POTS overlay)

hui_j Impulse response with lower band-edge at 138 kHz (ISDN overlay)

hud_j Impulse response, excluding DC (band-edge at 1 kHz)

It can also be seen in figure 3.6 that already an exclusion of the lowest 1 kHz yields considerable improvement compared to the non-limited case. The lower band-edges defined for ADSL in overlay with POTS and ISDN lead to even much more performance-improvement, but it can also be seen that the further restriction from the POTS-overlay to the ISDN-overlay does not yield considerable improvement in terms of ISI (see 4.3.1). The effect of ISI could be mitigated somehow, using a non-flat transmit spectrum with a power distribution following the water pouring theorem [14] and some excess bandwidth with a smooth roll off characteristic. The consideration of this approach however is beyond the scope of this work.

The Frequency domain equalization is concerned with doing away of the effects left over from the equalized channel. Once it is assured that orthogonality of the sub-carriers is maintained, there will be no interaction among them. However, after demodulation the sub-carriers will be subjected to different losses and phase shifts while passing through the channel & the TEQ. Therefore, FEQ consists of adjustments of sub-carrier gain and phase. This is performed by multiplying symbols in each sub-channel with separate complex constants determined during initialization.

Chapter 4

Discrete Multitone Modulation

In the previous chapters it was noted that the ANSI standardizing body selected DMT as the modulation scheme for ADSL. DMT is a special form of implementation of multicarrier modulation, based on the discrete Fourier transform that can conveniently be implemented in a fully digital way. In comparison with other modulation schemes, the main advantages of DMT are basically the same as the ones that have already been pointed out in general, as the advantages of multicarrier modulation in section 3.2. The advantage of DMT, compared with other multicarrier modulation (MCM) approaches, is the possibility of fully digital implementation and relatively low complexity, as it will be pointed out in the following sections.

In 4.1 and 4.2, the DMT-system will be presented in a step by step fashion, emphasizing on the elements that have not already been sufficiently explained in chapter 3. After that some specific topics, as for example echo cancellation (4.3.3) will be presented.

Figure 4.1 shows the block-diagram of the DMT-system as envisaged in a basic make-up. Additional elements in order to realize equalizer adaptation and echo cancellation have been omitted for simplicity of presentation. In the following subsections each block will be explained with its special features of DMT that distinguish this system from other multicarrier schemes (see 4.2.2).

4.1 Blocks of DMT Modulation – The Transmitter

4.1.1 Serial to parallel converter and block encoder

In this block the incoming serial bit-stream is converted into parallel data and grouped into blocks. Due to the flexible structure of DMT the number of bits carried with each block is not fixed, but determined during the start-up procedure (subsection 4.5). The bits in a block again are sub-grouped into the bits that will be carried by each subchannel. This number is not fixed either, but is also determined during initialization as bit loading [7], [19], [20], [21], [17], [3] and [12].

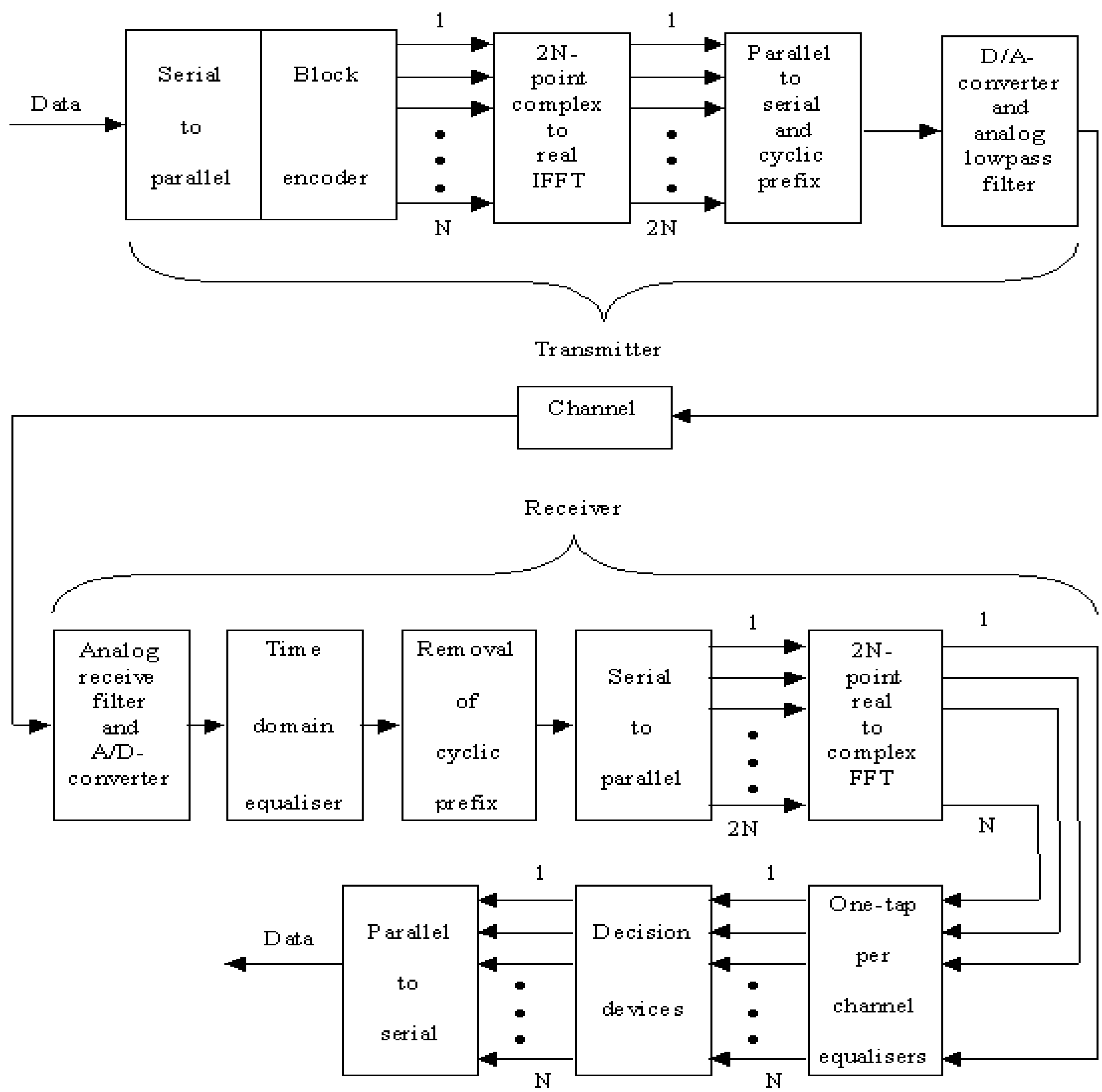


Figure 4.1 Structure and make-up of the DMT-system

Bit loading, as described in 3.3, is based on the idea of assigning the optimal number of bits per subchannel and the optimal amount of power to each subchannel. Depending on the environmental conditions and other design constraints, bit loading can be optimized towards different parameters. These can be the Bit Error Rate (BER), transmit power or total bit rate. Usually two of them are fixed and then the bit power allocation is optimized for the third. In the specific case of ADSL, optimizations are possible with the BER (fixed to 10^{-7}) as a design constraint. In these allocation schemes a common point of departure is the assumption that the bit power allocation can only be optimal, if the BER in all the subchannels are about equal [7], [21]. With this work, however, the total achievable bit rate is not so much of interest, but rather the number of bits assigned to each subchannel (hence the overall transmission rate) and the BER are followed up as performance measurement criterion. This leads to equation 4.1.

$$b_n = \log_2\left(1 + \frac{3}{\left(Q^{-1}\left(\frac{BER}{4}\right)\right)^2} * SNR_n\right) \qquad \text{……………………… 4.1}$$

where

$$Q(x) = \frac{1}{\sqrt{2 * \pi}} * \int_x^{\infty} e^{\frac{-y^2}{2}} dy$$

b_n	Number of bits assigned to subchannel n
BER	Bit error rate
SNR_n	Averaged signal to noise ratio (SNR) of subchannel n

A simplified number of bits assignment to subchannels is shown in equation 4.2 where there is only one variable, influencing the bit allocation, and it can be manipulated directly by the designer of the system. This is the PSD-function (power spectral density) of the transmitted signal. In all the previous computations (especially in chapter 3) always white transmit spectra have been assumed and will therefore be assumed here as well for a first set of computations so as to ease the complexity of the system.

$$b_n = \log_2 (1 + \frac{PSD_{signal}(n) * |H(n)|^2}{PSD_{noise}(n) * \Gamma} \qquad \text{………….………….. 4 . 2}$$

with

$$\Gamma = 10^{9,8-\gamma+10*\log(\frac{2*N}{2*N-CP})}$$

where b_n Number of bits assigned to subchannel n

PSDsignal(n) PSD of the transmitted signal in subchannel n

$|H(n)|^2$ Averaged channel power transfer function in subchannel n

PSDnoise(n) PSD of noise in subchannel n

Γ SNR-gap toward capacity, including cyclic prefix and coding gain

γ, N, CP Coding gain in dB, N is the number of subchannels and CP is Cyclic prefix

A certain problem arises by the fact that in uncoded DMT-systems, only integer numbers of bits can be transmitted in each subchannel. Without any further processing, this granularity would lead to a noticeable loss in the achievable data rate, as the next lower value would always have to be chosen in order to maintain the projected BER. The sub-optimality of data rates from using integral number of bits can be dealt by either transmission of fraction of bits [19] or more complex near optimal methods of total power allocation over the subchannels [22]. The great advantage of such "near optimization" approaches is the fact that the PSD of the transmitted signal is always near to the value of a flat spectrum. That means that it produces approximately the same crosstalk spectrum as would be produced by a spectrally white (but unrealizable because of the granularity of the usable numbers of bits) signals.

In most practical cases it is common to assume a system having a block length 2*N=512 and a cyclic prefix of length CP=40 [21]. It can be seen that for the long distances (3 and 4 km) in the worst case model with ISDN overlay the presently required 2 Mbit/s can not be achieved. For the case without ISDN and the worst case model in the 4 km twisted pair cable (0.5 mm) the system also fails to transport the minimum bit rate. It has to be noted that all these loops are inside the specified range for ADSL, while some loops in the real world model exceed the specified ranges, but achieve to transport the minimum bit rate. This shows clearly that not only the length and the wire diameter, but also the loop population and therefore the crosstalk environment will decide whether an ADSL system works properly or not.

Capacities of transmission calculated with Γ are slightly lower than those obtained with the theoretical ones. This is due to the fact that the gap between capacity and DMT is 9.8 dB (instead of 9 dB assumed for the safety margin) and, moreover, an additional loss of about 0.35 dB had to be added to take the cyclic prefix (subsection 4.1.3) into account [22].

$$PSD_{signal}(n) = \frac{\Gamma * PSD_{noise}(n) * (2^{b_n} - 1)}{\left|H(n)\right|^2} \qquad \text{……………..………….. 4 . 3}$$

$$P_{total} = f_{symb} * \sum_{1}^{N} PSD_{signal}(n) \qquad \text{……………..………….. 4 . 4}$$

Where P_{total} Total average power of the ADSL signal

f_{symbol} Symbol rate in Hz

N, PSDsignal(n) See equation 4.2

			Without ISDN			With ISDN		
			2km	3km	4km	2km	3km	4km
			Mbits/s	Mbits/s	Mbits/s	Mbits/s	Mbits/s	Mbits/s
0,4mm	Twisted pair	Real world	8,28		3,38	7,40		2,97
		Worst case	5,01	2,15		4,11	1,43	
	No crosstalk		15,00		6,25	13,17		4,12
	Quad	Real world	9,87		3,97	8,80		3,34
		Worst case	5,22	2,72		3,96	1,64	
0,5mm	Twisted pair	Real world	8,41		6,13	7,51		5,71
		Worst case	7,45		1,66	6,57		1,14
	No crosstalk		15,92		10,74	14,10		8,77
	Quad	Real world	10,02		7,34	8,94		6,70
		Worst case	7,52		2,20	6,25		1,32

Table 4.1: Achievable data rates for the different channel models, using the flat spectrum optimization

The sub-optimality of the scheme, shown above, inspires further thought in order to find a scheme to make better use of the available transmit power. From theory it is clear that the optimal distribution of transmit power resembles the water pouring solution of Gallager [14], [7]. However, a disadvantage of every other scheme, but the flat spectrum one, is the higher complexity, where as the bit power allocation following the flat spectrum optimization is in fact calculated in a straight-forward manner.

In Figure 4.2 bit allocation for an ADSL system with POTS overlay in a typical real world twisted pair cables is depicted as a function of the subchannel index. In these diagrams, a line corresponding to the PSD of an ADSL signal (POTS overlay) without any optimization (flat spectrum) is also shown as a point of reference. In general, for the short loops there is no big difference between the PSD curve of the "small optimization" to that of the matrix optimization or other optimization schemes, only the curves for the longer loops (3 km for worst case in 0,4 mm wires and 4 km for all the others) are shown.

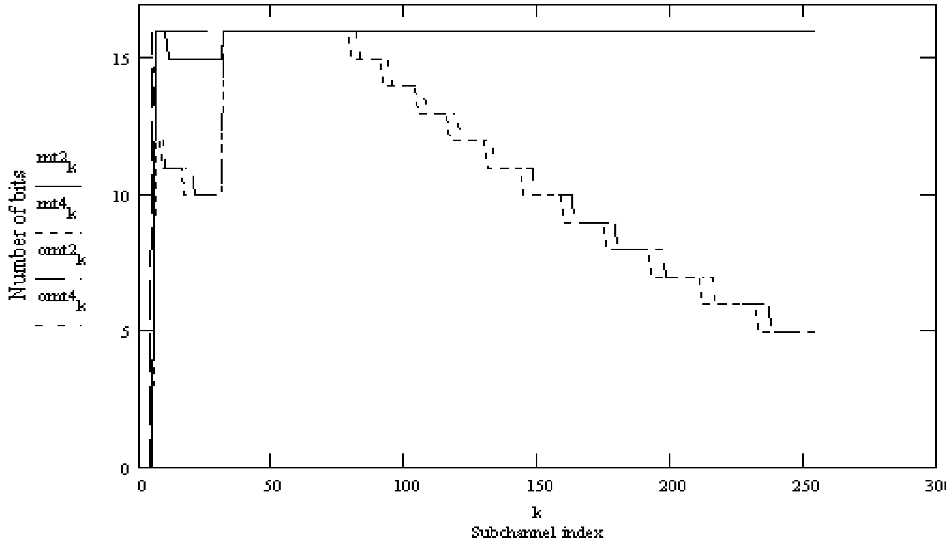


Figure 4.2: Bit allocation for an ADSL system (POTS overlay) without crosstalk noise in twisted pair cable

Where mt2 2 km loop (0,5 mm) with flat spectrum optimization

mt4 4 km loop (0,5 mm) with flat spectrum optimization

omt2 2 km loop (0,5 mm) with matrix optimization

omt4 4 km loop (0,5 mm) with matrix optimization

In this subsection only the bit power allocation itself had been considered, assuming that the channel transfer function and the spectral noise distribution are known. In practice this is normally not the case. In real world applications this information has to be obtained by measurement during the start-up procedure [13].

4.1.2 Inverse Fast Fourier Transform (IFFT)

In the DMT-system, MCM modulation is realized by a 2N-point IFFT. Here, the inverse transformation in the transmitter is due to the fact that the data is assigned to sub-symbols, represented by constellation points in parallel independent subchannels in the frequency domain, where as the DMT transmit signal is created in the time domain. The set up of DMT with its transmit signal in time domain does away with filter banks and banks of modulators otherwise necessary in the classical multicarrier systems (3.2).

The inputs to the 2N-point transform are the constellation points of the N subchannels. In order to achieve a real valued output (in time domain), Hermitian symmetry has to be maintained at the input of the IFFT. The symmetry rule for the IFFT in order to receive a real valued time signal is shown as equation 4.5.

$$X_{2xN-k} = (X_k)^* \qquad \text{for} \quad k < N \qquad \text{……………………….. 4.5}$$

Where X_k Signal point in the constellation of subchannel k

$(X_k)^*$ The star denotes complex conjugate

2 x N Block length of the transformation

In a DMT-system the Fourier transform and its inverse is usually performed by the well-known fast Fourier transform (FFT) algorithm [13]. This means the computational complexity can be reduced considerably, while the normal discrete Fourier transform (DFT) needs about M^2 operations (with M being the total length of the transform), the FFT only needs about 1.5*M*log(M) for lengths beyond 128, as they are usual for DMT systems [20].

Here, there exists a small, but for DMT not very important, disadvantage of the application of the FFT-algorithm. It only performs optimally for transformation lengths that are integer powers of 2. As the block length (equal to the transformation length) for a fixed sample rate (and fixed length of the cyclic prefix, (4.1.3)) defines the bandwidth of the subchannels, it has an impact on performance. Since the basic idea is to create subchannels that are virtually distortionless and only impaired by white noise, the bandwidth of each subchannel would have to tend toward zero. In this limit the subchannels would be free of distortion and no equalisation would be needed [7].

In reality, however, the number of subchannels cannot be chosen arbitrarily high. It was shown that an increase in block length beyond 512 does not yield significant improvement in performance, but it does increase complexity and throughput delay [17], [20]. A longer delay, although undesirable in order to maintain compatibility with a maximum number of services that might be transported over the ADSL platform, eases the requirements on equalisation and computational speed. In practical systems a combination is made between a small amount of equalisation (subsection 4.2.1) and a large amount of multicarrier modulation. The relationship between sample rate, block length, symbol rate (= subchannel bandwidth), symbol period, subchannel distance and length of the cyclic prefix is shown in equation 4.6. There it can also be seen that the subchannel bandwidth is smaller than the spacing between the subchannels, which means a certain loss in bandwidth efficiency. This guard band is introduced by the cyclic prefix as in 4.1.3

The modulation by the IFFT corresponds to modulation of the respective carrier frequencies with a rect-pulse. The use of a rect-pulse results in a sinc-spectrum in the frequency domain, causing one subchannel to overlap with many of its neighbours. As long as the channel is non-dispersive, orthogonality is maintained and neither interchannel (ICI) nor interblock (IBI) interference does occur. On a dispersive channel, however, additional measures, such as synchronization and timing recovery mechanisms [19], have to be taken in order to maintain orthogonality.

In recent years there was a growing interest in applying other pulse shapes in order to increase the spectral containment of the subchannels in their own frequency range [19], [22]. With the rect-pulse the first side lobe is only about 13.6 dB lower than the main lobe and the strength of the further side lobes decreases as f^{-2} [19]. One of the great advantages of the present DMT-system, using the IFFT for modulation, is its relatively low complexity and its ease of implementation. The whole system can be built in a single chip fashion. A great improvement could be made in future by designing an FFT peripheral in order not to have to perform all these operations in the main signal processor [17], [7], [20].

$$f_{symbol} = \frac{1}{T} = \frac{f_{sample}}{2 * N + CP} \qquad \text{………….………….. 4 . 6}$$

and

$$\Delta f = \frac{f_{sample}}{2 * N}$$

Δf	Subchannel spacing
f_{symbol}	Symbol rate
T	Symbol period
f_{sample}	Sample rate
N	Number of subchannels (=1/2 block length)
CP	Length of the cyclic prefix (in samples)

4.1.3 Parallel to serial conversion and addition of the cyclic prefix

In this block the parallel output (2N values) of the IFFT is converted into a serial bit-stream and the so-called cyclic prefix is added. The principle of the cyclic prefix is shown in figure 4.4. The last CP samples are taken from the end of the block and copied to the beginning of it. This makes the transmitted signal look somehow like a periodic signal [3], [19], [20], [23].

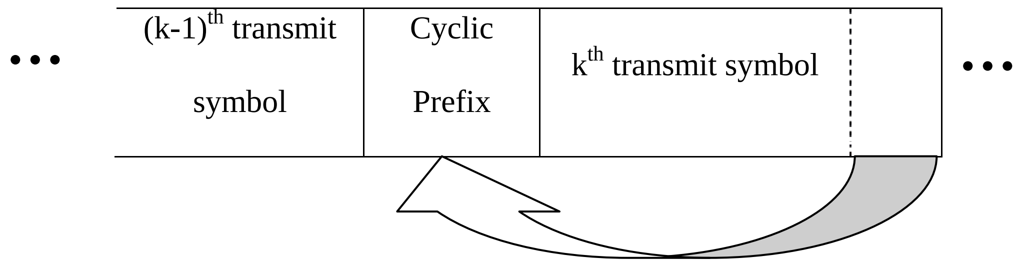


Figure 4.3: Make-up of the cyclic prefix

The effect of the cyclic prefix is of twofold. The most obvious one is to create a guard space between neighbouring transmit symbols in the time domain. Instead of zeros this guard space is filled with the cyclic extension as shown in figure 4.4, but it has basically the same effect and therefore combats ISI efficiently in the time domain by concentrating it in the cyclic prefix that is later discarded in the receiver.

ISI, however, is not the only internal interference problem that can affect a DMT-system. The other problem to be coped with cyclic prefix is the interchannel interference (ICI). By the periodicity of the transmitted signal (due to the cyclic prefix) a cyclic convolution, instead of a linear convolution, between the channel impulse response and the transmitted signal is efficiently simulated. This means that with using the CP, the effect of the channel is reduced to an element by element multiplication between the Fourier transforms of the channel impulse response and that of the transmitted signal, thus introducing only different gains and delays on each subchannel [19]. These different gains and phases can be coped with by one-tap per channel equalizers (see 4.2.3), figure 4.6a, figure 4.6b and no ICI is produced.

Equation 4.6 demonstrates that the cyclic prefix but introduces a difference between the subchannel bandwidth and the spacing of the subchannels and can therefore be understood as a guard band in the frequency domain. Moreover, the cyclic prefix enhances the spectral containment of the main lobe of the subchannels' transmit spectra, although only slightly. The additional difference between the main lobe and the first side lobe of the subchannels' spectra is calculated using equation 4.7 [22].

However, the inclusion of a cyclic prefix also gives rise to some disadvantages. Firstly, it brings a certain loss in SNR due to the fact that energy is used in order to transmit redundant contents. Depending on the length of the cyclic prefix, in typical ADSL systems the loss due to cyclic prefixing is about 0.35 dB, therefore a relatively low price to pay for the relaxed requirements on equalisation (4.1.5). The cyclic prefix, however, only works in the described manner, if its length (CP) is at least as long as or longer than the channel impulse response. As the impulse response of typical ADSL channels may have lengths of up-to and more than several hundred samples (2.3.1.1}, figure 2.12 and 2.13), the block length would have to be about 13 times longer than the longest impulse response of a typical ADSL channel; as this could only maintain the previously mentioned additional loss of about 0.35 dB [3]. Block lengths of several thousand samples, however, are not realisable in practice, firstly, because of the increased complexity and, secondly, because of the large additional delay that would be introduced.

Therefore in practical systems a combination of cyclic prefixing and a small amount of time domain equalizing is employed (4.1.5) in order to restrict the length of the equalized impulse response to the length of the cyclic prefix [3]. A shortening of the channel impulse response, however, is much easier to achieve than zero ISI, as it is necessary in baseband and single carrier modulation schemes. For ADSL systems lengths of cyclic prefixes of 8 [17] and more recently 32 [24] and 40 [21] have been proposed. All computations in this work assume a cyclic prefix of less than 40 samples therefore somehow representing a worst case, as it is the one with the highest power loss. Although the inclusion of a cyclic prefix is, after all a clearly sub-optimal solution, the additional power loss is a low price to pay for maintaining the orthogonality between the subchannels and avoiding ISI between subsequent symbols. Moreover, the cyclic prefix facilitates synchronization and timing recovery as will be shown in 4.2 [19].

$$\Delta L = 20 * \log\left(\frac{2*N + CP}{2*N}\right) \qquad \text{.........................} \quad 4.7$$

ΔL Additional suppression of the side lobes [22]

N Number of subchannels

CP Length of the cyclic prefix

4.1.4 D/A-, A/D-converters and transmit and receive filters

The A/D- and D/A-converters in a DMT-system have to be more sophisticated than those in single carrier or baseband systems. This is mainly due to the many possible values in the transmitted signal, due to the addition of many (in case of ADSL 256) components, each having complex Gaussian amplitude probability [3]. The converters also have to offer a large dynamic range to cope with the high peak to average ratio of multicarrier systems (see 4.2.2). The analog filter in the transmitter is not included for pulse shaping as it may be done in single carrier and baseband systems, but only to remove the higher frequency copies of the transmit signal that are created due to the fully digital make-up of the DMT-system. At the input of the receiver there is to be found the same kind of filter, here used in order to exclude the noise of the higher frequency portion. It is also needed for clear band limitation or, in other words, as an anti-aliasing filters previous to sampling. As this point, it should be noted that the

POTS/ISDN-splitter (as in 4.3.1) was assumed to be part of the channel and has therefore neither been depicted, nor been described in this section. For further information about this device refer to [25].

4.2 Blocks of DMT Modulation – The Receiver

4.2.1 Time domain equalizer (TEQ) and removal of the cyclic prefix

As it had already been mentioned in previous subsections, the DMT-system requires some sort of equalisation although it is much less than in comparable single carrier or passband systems. Here, the task of the time domain equalizer is to shorten the impulse response of the channel to a length, shorter or equal to the one of the cyclic prefix. This task is much easier to achieve than to create zero ISI. The equalizer itself is typically a short tapped delay line of between 10 [20] and 64 [21] taps. To find the tap-settings there exist different possibilities. The most straightforward one is to measure the channel impulse response and to derive a pole zero model of the form shown in equation 4.8.

$$h(D) = \frac{a(D)}{b(D)} \qquad \text{.................} \qquad 4.8$$

Where *h(D)* Channel impulse response, modeled by the pole-zero model

a(D) Zero-polynomial of the pole-zero model

b(D) Pole-polynomial of the pole-zero model

If a(D) has as its maximal length the one of the cyclic prefix and b(D) has maximum the length of the TEQ, then b(D) is the optimal setting for the equalizer taps. If in the derived model b(D) is shorter than the TEQ, the rest of the taps can be filled with zeros. With the equalizer taps set equal to b(D), a(D) becomes the impulse response of the equalized channel. In the typical ADSL environment there are slightly more poles than zeroes, which means in other words that b(D) will be slightly longer than a(D) [21]. Even on bad loops and up to data rates of 6 Mbit/s the channel response can be shortened to 10-30 samples [3], requiring a TEQ of maximum about 32 taps of length. This form of calculation of the tap-settings, however, requires relatively complex mathematical operations after channel estimation. This causes additional complexity for the start-up procedure (4.4).

Other schemes work in an iterative manner, applying some form of least mean square (LMS) algorithm in order to converge the tap-settings. This does not require any complex mathematical operations and therefore does not increase computational complexity unduly, but has the disadvantage of long convergence times [6]. For common DMT-equalisers it takes several millions of iterations to converge the equalizer, resulting in considerable delays in the start-up procedure. An approach to shorten the necessary time-span without increasing prohibitively computational complexity has been made by Lee et al. [23]. They observed that a DMT-system with cyclic prefix and TEQ could be understood as a DFE (Decision feedback Equalizer) where the TEQ corresponds to the feedforward filter and the cyclic prefix to the feedback filter.

Their method is therefore based on the fast adaptation algorithm for the DFE. Channel estimation is used in order to measure the frequency transfer function of the channel and the spectral noise distribution using MMSE (Minimum Mean Square Error) or similar algorithms. Depending on the optimization scheme that is being used in the system, a part or all of this information is also needed for the calculation of the bit power allocation and therefore this measurement does not mean much additional complexity. The frequency transfer function of the channel is then transferred into the time domain by an inverse FFT. Now the output of the channel with TEQ is compared to a desired impulse response (as in figure 4.5) and the equalizer is adapted in order to match both responses. The desired response is calculated based on the measured channel data and optimizing towards highest possible data rate. In this calculation a simplification is used in order to avoid matrix inversion and to replace it by FFT operations.

As the FFT has to be implemented anyway in the DMT-transceiver, this function can be shared and does not yield additional complexity. Due to the simplification a slight error is being made, resulting in a slightly sub-optimal performance. The performance gap is about 0.2 to 0.3 dB for usual ADSL loops and TEQ lengths, but approaches zero with increasing filter lengths (up to 64) or decreasing loop lengths. Even including the channel estimation, this algorithm still needs much less iterations than a normal LMS scheme as only a few thousand iterations are needed for the channel estimation and another few thousand for the optimization of the equalizer tap-settings (see also 4.4). After initial convergence with the fast algorithm, a LMS optimization may be used for continuous tracking.

In this case the performance gap for the fast algorithm diminishes, causing little cost in additional complexity. It has to be checked more profoundly, however, if the about two tenths of a dB, achievable by concatenation with the LMS algorithm, are worth the additional complexity. After equalisation the cyclic prefix is removed and simply discarded, but before its removal it can be used for synchronization and timing recovery [19].

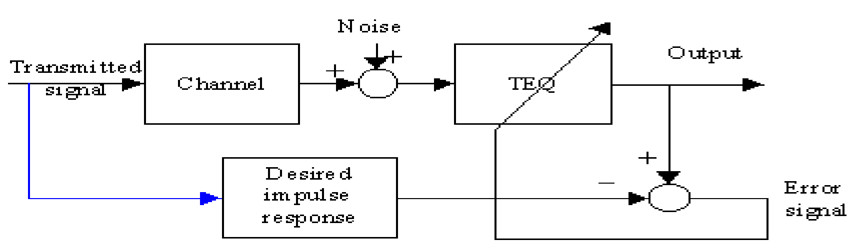


Figure 4.4: TEQ for ADSL application, using an optimized desired impulse response

4.2.2 Serial to parallel conversion and fast Fourier transform (FFT)

After equalization and removal of the cyclic prefix the incoming serial stream of samples is converted into blocks of parallel data with 2N parallel values. These are fed into a 2N-point real to complex FFT, therefore transferring the time domain signal again into the frequency domain.

The transfer into the frequency domain also means the separation of the N parallel independent subchannels whose contents can now be further processed on a per subchannel base. Of the 2N outputs of the FFT, only N are used for further processing due to the Hermitian symmetry that had already been discussed in 4.1.2.

4.2.3 Frequency domain equalizer (FEQ)

The FEQ is a set of complex one-tap per subchannel equalisers that are included in order to cope with the different gains and phase delays that the equalized channel introduces on the different subchannels. By multiplication with a complex number, the amplitude levels and the phase positions are readjusted as shown in figure 4.6a and figure 4.6b. The initial tap-values for the FEQ are learned during the start-up procedure (4.4). First, the channel frequency transfer function of the equalized channel is estimated and then it is inverted in order to receive the tap values for the FEQ. If during the start-up procedure the impulse response of the equalized channel is determined, then the initial tap-settings are the inverse FFT of it [21]. During the data mode, the equalizer taps are continuously updated in a decision directed manner where the decision element is included to make decisions at the output of the FEQ. The output of this decision device is then compared with its input for each subchannel and the FEQ-taps are then adjusted accordingly. As the decisions, made without a Viterbi-receiver in systems with trellis coded modulation, are not very reliable, the step size for the tap-adaptation has to be chosen small in order to avoid problems due to error propagation, leading to adjustments into the wrong direction due to wrong decisions [23].

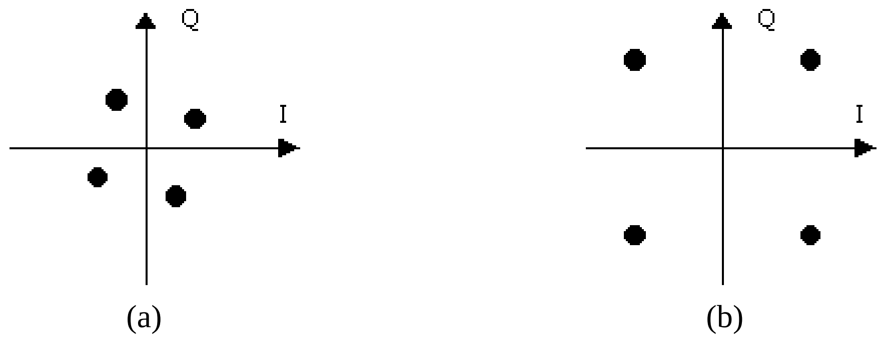


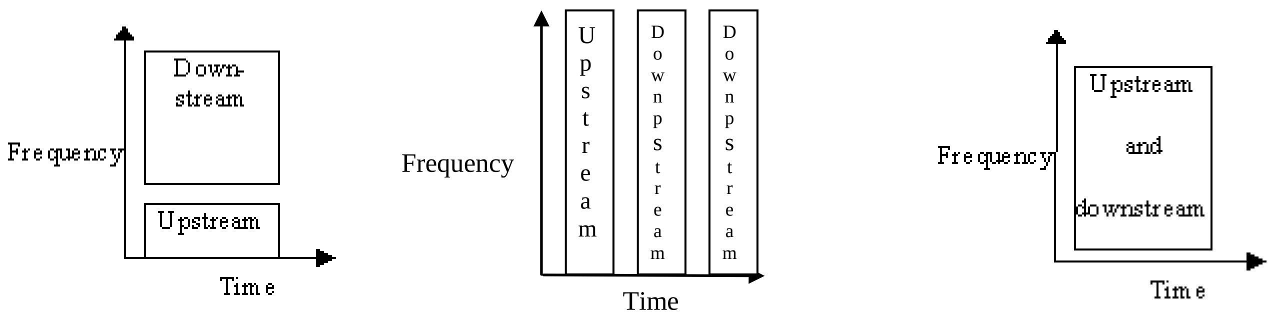
Figure 4.5 : Received subchannel signal (a) Before one-tap per channel equalisation and (b) After one-tap per channel equalisation

4.2.4 Decision device and parallel to serial conversion

In the decision device the symbols of the subchannels are detected, using the knowledge about the bit power allocation size and as nearest point in the constellation, and therefore the sub-blocks of bits that had been assigned to the subchannels are recovered. Once the sub-blocks are detected, the data is re-converted into a serial bit stream in a block-by-block fashion.

4.3 Duplex schemes

A duplex scheme is always needed, if transmission in both directions at the same time is desired. "At the same time" in this context does not necessarily mean exactly the same instant, but at least virtually at the same time, meaning that the differences are minimal, as will be discussed in this section. For duplex transmission basically there exist three different possibilities as is shown also in the figures 4.6a, 4.7b and 4.7c. The signals in both directions can be separated in time, in frequency or simply by their direction of propagation. In the following subsections the three duplex schemes are briefly presented and their applicability for ADSL as well as their advantages and disadvantages for this application are discussed.



(a) (b) (c)

Figure 4.6a: Frequency division Figure multiplexing (FDM)

Figure 4.6b: Time division multiplexing (TDM)

Figure 4.6c: Transmission with echo cancellation

4.3.1 Frequency division multiplexing (FDM)

The most obvious choice for an asymmetric service, such as ADSL, would be FDM (see figure 4.7a), as the narrow upstream signal does not occupy a wide frequency band and still allows a lot of usable bandwidth for the downstream in the higher frequency region. Another point in favor of FDM is its ease of implementation, as only splitting filters are needed in order to divide the frequency band into an upstream and a downstream section (as in 1.3). FDM totally avoids the problem of self-NEXT, as there is no transmission in the frequency band where the receiver is receiving and vice versa. In fact, ADSL already uses FDM in order to realize the overlay facility for POTS and ISDN. In the first standard for ADSL, FDM was defined as the duplex scheme and it is still up to the decision of the manufacturer to implement his equipment with this scheme. The big disadvantage, however, is that the low frequency portion, where in case of FDM only the upstream signal is transmitted, is not fully used (in both directions) and therefore valuable transport capacity for the wideband downstream is lost. Anyway, even if echo cancellation has to be used, DMT has the flexibility to work as an FDM-system. This can be done by the bit power allocation, not allocating bits for the downstream where already bits for the upstream are allocated and vice versa. Finally it may be said that FDM is a cost-attractive, but definitely sub-optimal choice for the ADSL system.

4.3.2 Time division multiplexing (TDM)

TDM has never been considered as a strong candidate for ADSL, mainly because the line bit rate has to be increased considerably in order to transmit the data in the two directions in different time slots (see figure 4.7b. In this case the line bit rate not only has to take the value of the sum of the upstream and the downstream rate, but there also has to be considered a certain guard time. This is needed for the impulse response and the echoes of the transmission in one direction to linger off before transmission in the other direction is started. The overhead due to the guard space depends on the maximum permitted line length and also on the length of the bursts in each direction.

4.3.3 Echo cancellation (EC)

EC is the second possibility that the present standard for ADSL offers to the choice of the manufacturer (together with FDM, see 4.3.1). In this scheme the signals are only separated by their direction of propagation inside the cable.

It is to be expected that on the long term all transceiver systems, offered for ADSL, will use echo cancellation in order to make full use of the available channel. This will be especially interesting for rate-adaptive ADSL as in this case the positive effect of the echo canceller in terms of achievable data rate will be fully available to the customer.

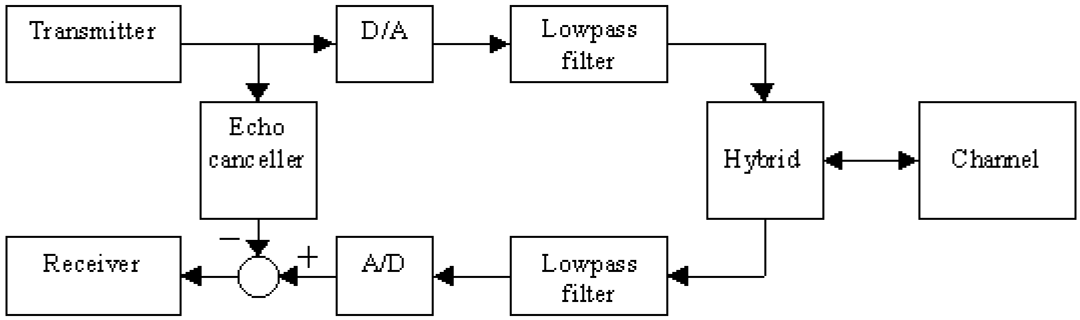


Figure 4.7: Block-diagram of a transceiver with signal driven echo cancellation

4.4 Start-up procedure

Several preparations have to be done in receiver and transmitter parts before reliable data transmission in an ADSL system can be started [23], [17], [20], [3]. The initial settings for the equalisers and echo cancellers have to be computed and the bit power allocations have to be performed (all the tasks as well in the central office as in the remote transceiver). In order to do these steps, several measurements have to be done. The initialization routine begins with a transmit request and several test patterns in order to make sure that the channel is in operational conditions. Afterwards some test patterns are exchanged and timing and synchronization is established. Then predefined periodical pseudorandom sequences are transmitted in order to estimate the channel transfer function and the spectral noise distribution. This periodical sequence is at least of the length of the channel impulse response, producing a line spectrum in the frequency domain (comb function) [23]. As the receiver knows the test sequence and its spectrum, the channel transfer function can be computed, using equation 4.9. The result has to be averaged over several periods of the test sequence in order to mitigate the impact of noise on the result.

$$H_n = \frac{1}{L} * \sum_{i=1}^{L} \frac{Y_{i,n}}{X_n} \qquad \ldots\ldots\ldots\ldots\ldots\ldots \qquad 4.9$$

H_n Channel transfer function for frequency index n

L Number of repetitions of the test sequence for channel estimation

$Y_{i,n}$ Channel output for test sequence repetition number i and frequency index n

X_n Test sequence content for frequency index n

Equation 4.9: Channel transfer function estimation [21]

With an observing time of 40 sequence periods the excess mean square error can be reduced to about 0,1 dB as can be calculated with equation 4.10.

$$EMSE = 10 * \log(1 + \frac{1}{L}) \qquad \ldots\ldots\ldots\ldots\ldots\ldots \qquad 4.10$$

Where *EMSE* Excess mean square error [23]

L Number of repetitions of the test sequence for channel estimation

By transmitting the test pattern a little longer, the noise spectrum can be estimated by comparing the received signal with the element by element product between the (by then) known channel transfer function and the transmit signal [24]. This whole operation has to be done for the upstream and for the downstream path. Then the bit power allocations are computed and transmitted in a safe mode to the respective transmitter at the opposite end of the line. Also the tap settings for the time domain and the frequency domain equalisers in both receivers have to be computed. Then the test patterns to train the echo cancellers can begin. During this procedure there is only one transmitter at a time sending its test sequence in order to estimate the transfer function of the echo path in the frequency domain, much in the same way as previously the channel transfer function had been estimated. With the results the initial settings of the echo cancellers are computed. The restriction to only one transmitter at a time is made in order to avoid the influence of the far end signal on the settings of the echo cancellers. In order to achieve reliable transmission, the estimates have to be performed with high accuracy. Especially the measurements that are the base for the computation of the bit power allocation have to be accurate, because there is no later update of the bit power allocation during transmission. Echo cancellers and equalisers, on the other hand, are adapted continuously and therefore have the chance to correct errors in the initial settings. After all these steps have been performed, frame synchronization is established and data transmission may begin. The feedback channel between both transceivers that had been used during the initialization procedure is also maintained active during normal data mode in order to transmit information, used for system control and monitoring [17].

Chapter 5

Simulations and Results

To simulate this end-to-end communications system, a MATLAB® code was written and this code is prepared into two parts; of initialization procedure and a realization stage.

In the first part, the empirical channel model impulse response data, from subsection 2.5, is fed to the program. This data can be used to prepare a bit-loading array, through an SNR array depicting the SNR of each of the subchannels. Here, two spectrally variable SNR tables are used in this paper to simulate the low-pass channel of twisted pair wire; hence comparing higher and lower rates of data transmission. Here the data rates simulated are ranging from around 1 Mb/s to around 7 Mb/s where as an ADSL system is seen to accommodate a rate up to 8 and in good lines up to 10 Mb/s. Using this SNR array, then, a bit loading array is calculated defining the number of bits, b_i, in Number of bits/ Symbol/Hertz.

Moreover, the channel impulse response is used so as to prepare the time domain equalizer. The last process of initialization is the frequency domain equalizer (FEQ) component of the system. However, this is calculated with one full end-to-end transmission realization and hence the first process of end to end transmission is done to determine FEQ.

The second and realization part of the process is the basic simulation process and this is summarized in the following block diagram given in figure 5.1 as follows :-

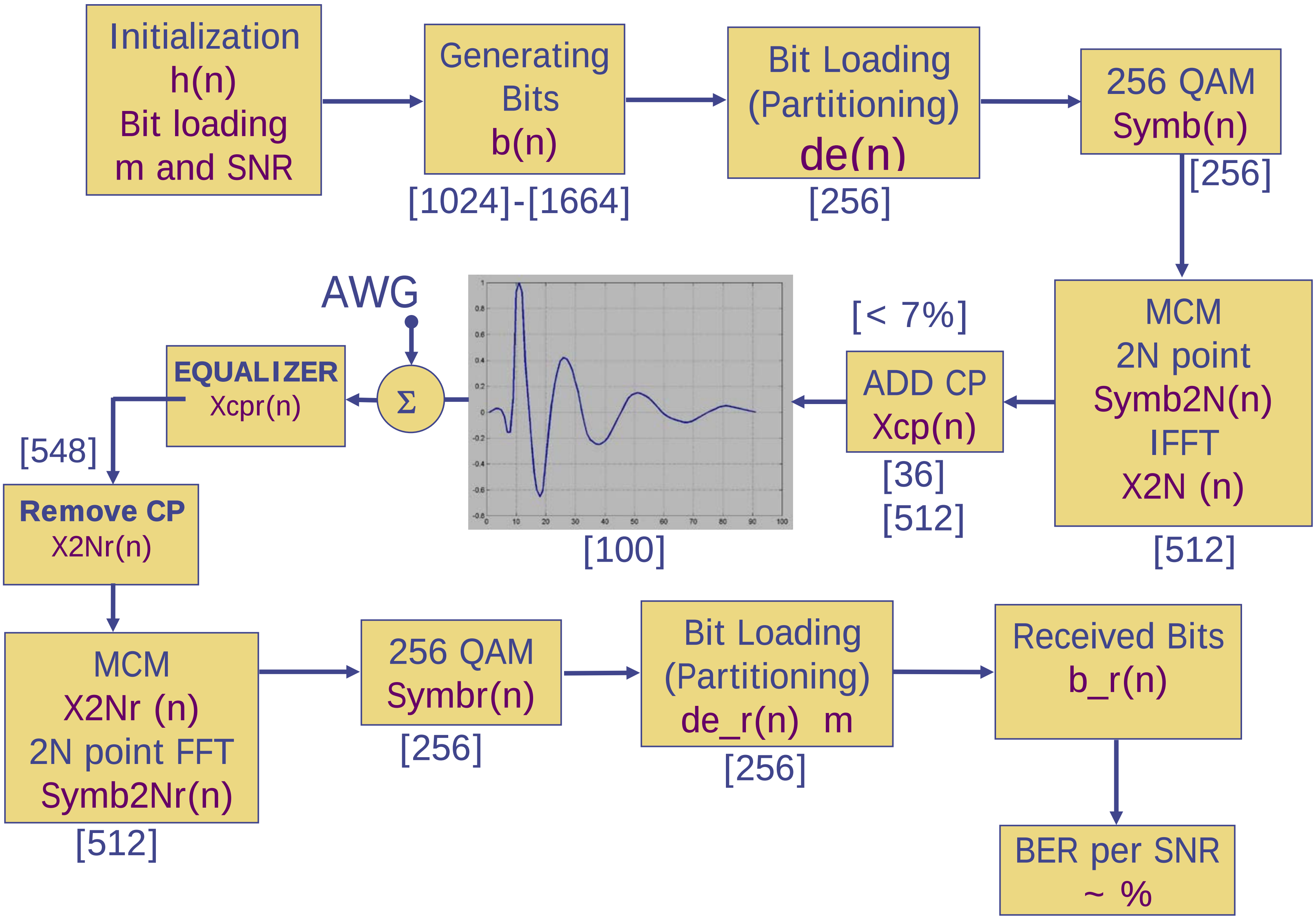


Figure 5.1 A flow chart for the simulation of an end to end ADSL based system.

[5] Liang C. Chu, ADSL System Enhancement with Multiuser Detection, Georgia Institute of Technology, July 2001.

[10] Daniel Franklin et al., An Improved Channel Model for ADSL and VDSL Systems, February 1999.

Jump this page

The MATLAB® code written basically for this work is used for the simulation of the above flow chart and given in the Appendix A.

When this code is run, a number of results and intermediate signals can be obtained. The first is a representation of the two bit-loading arrays used within this project and this is demonstrated in figure 5.2.

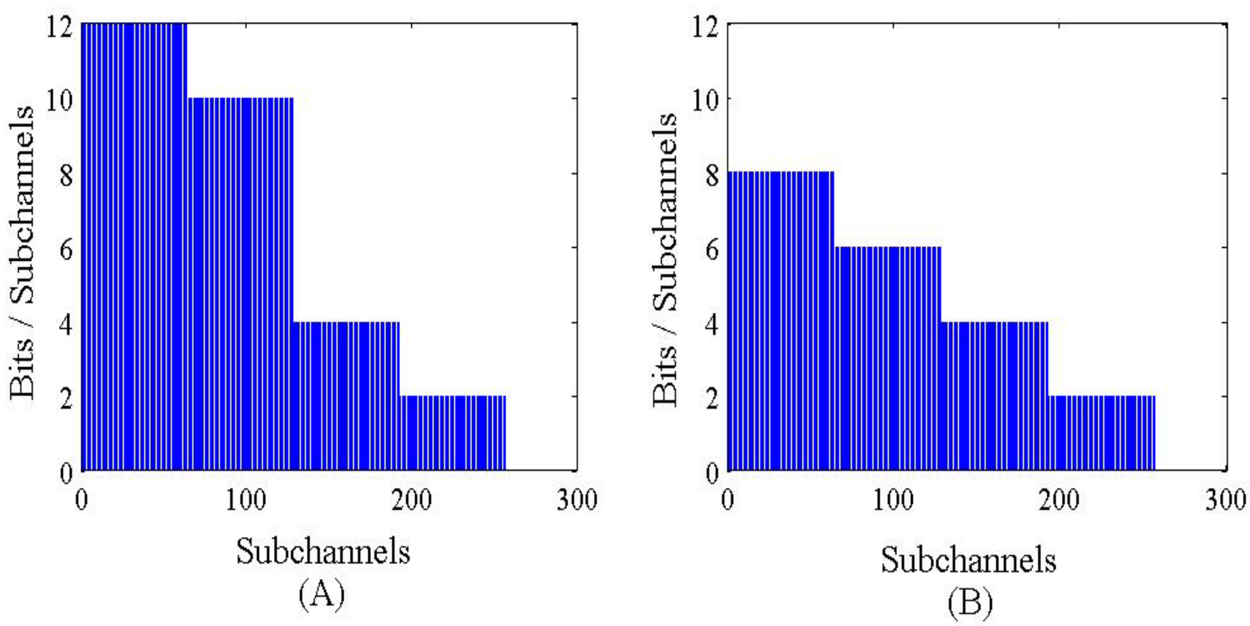


Figure 5.2 Bit Loading array A) Higher data rate & B) Medium data rate simulation

Next shown are the spectral values of the data to be sent. From the diagram of end to end ADSL system in Fig 5.1, it is seen that these spectral values, Symbn(k), are QAM symbols allocated to each subchannel and a typical such symbol array is shown in figure 5.3 a,b.

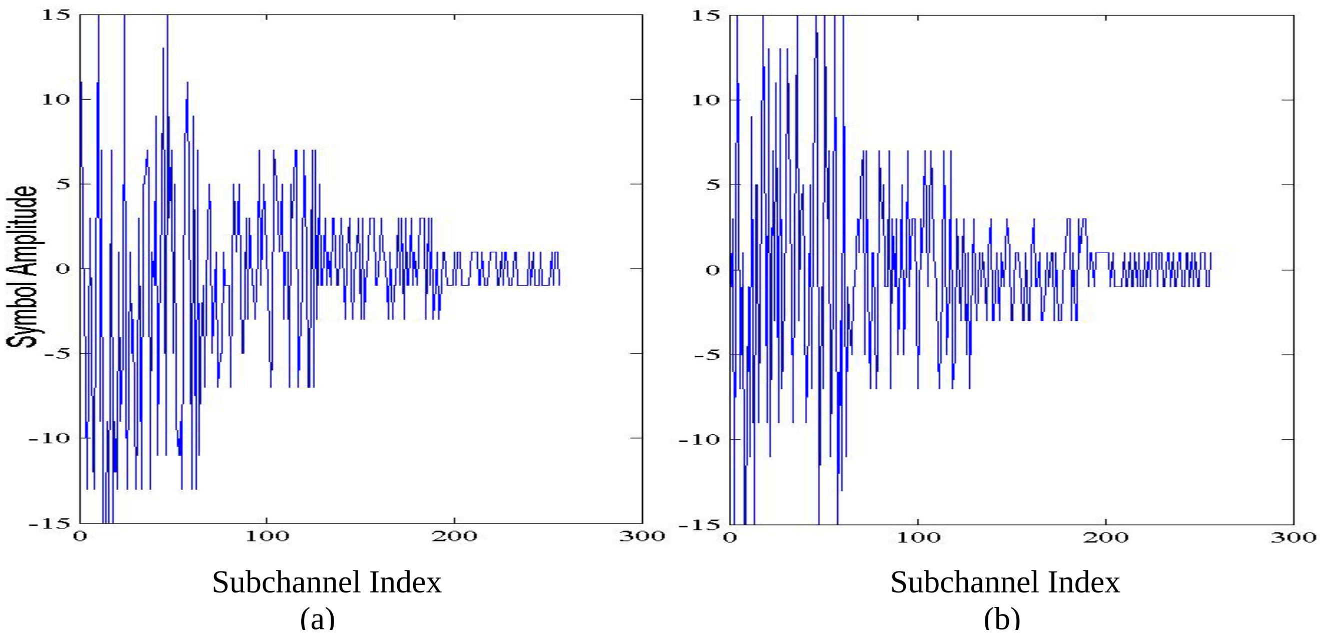


Figure 5.3 A QAM symbols array (a) Real Part and (b) Imaginary Part

A 2N-point even symmetry to this data would be obtained using equation 4.5. The time domain representation of the 2N point Symbol stream array, Symb2N(k) is X2N(n) and adding a $\boldsymbol{v}$=36 length cyclic prefix to X2N(n) produces the Xcp(n) time series and these two signals would look like as shown in fig 5.4 a & b.

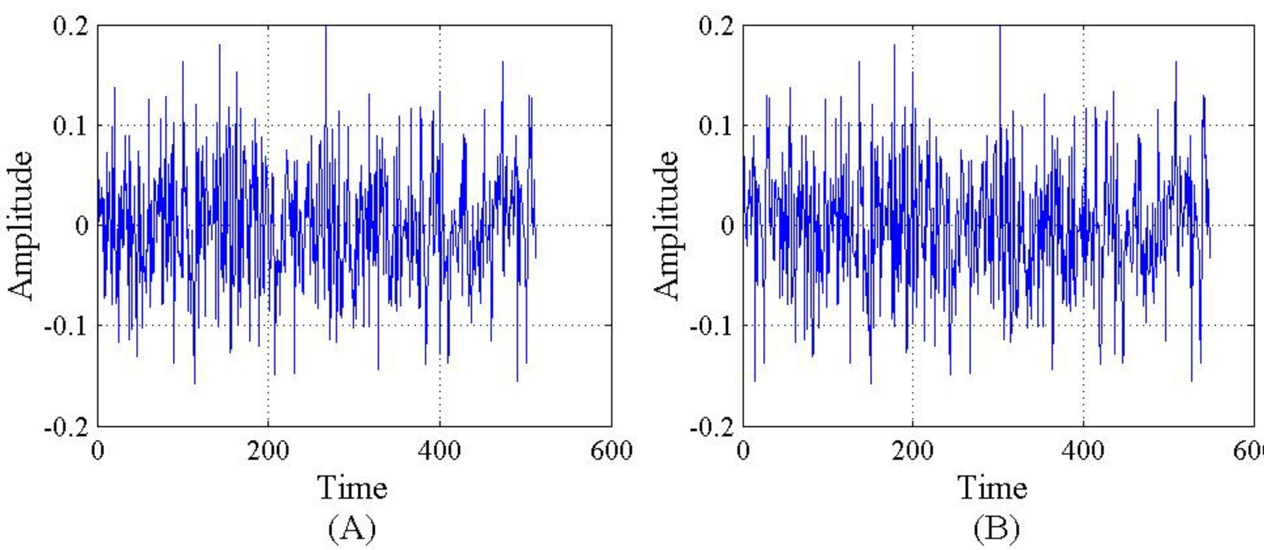
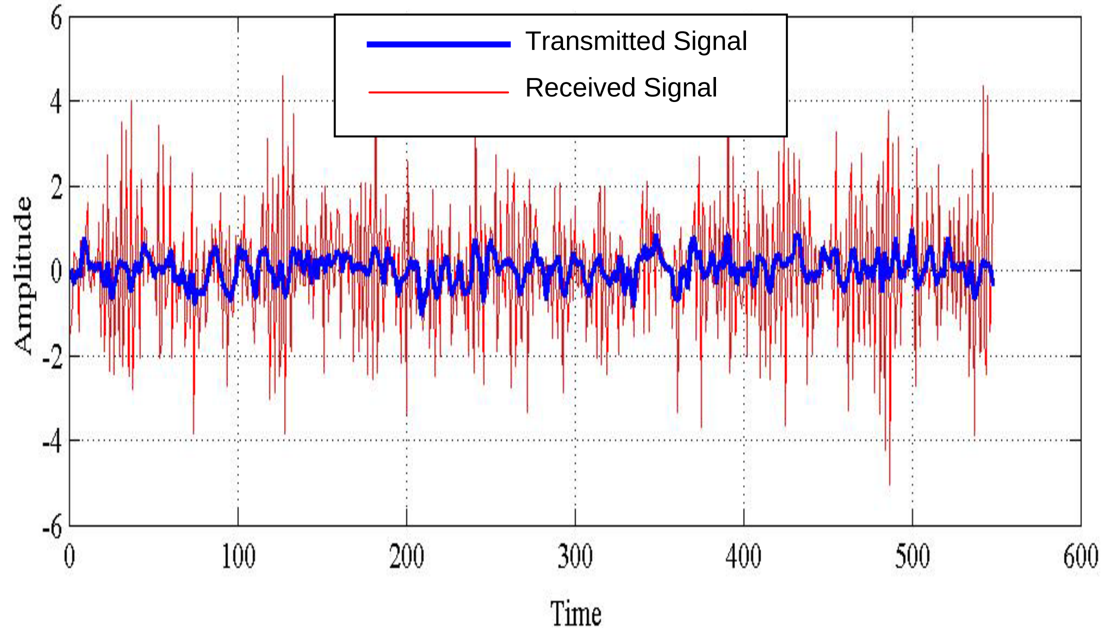
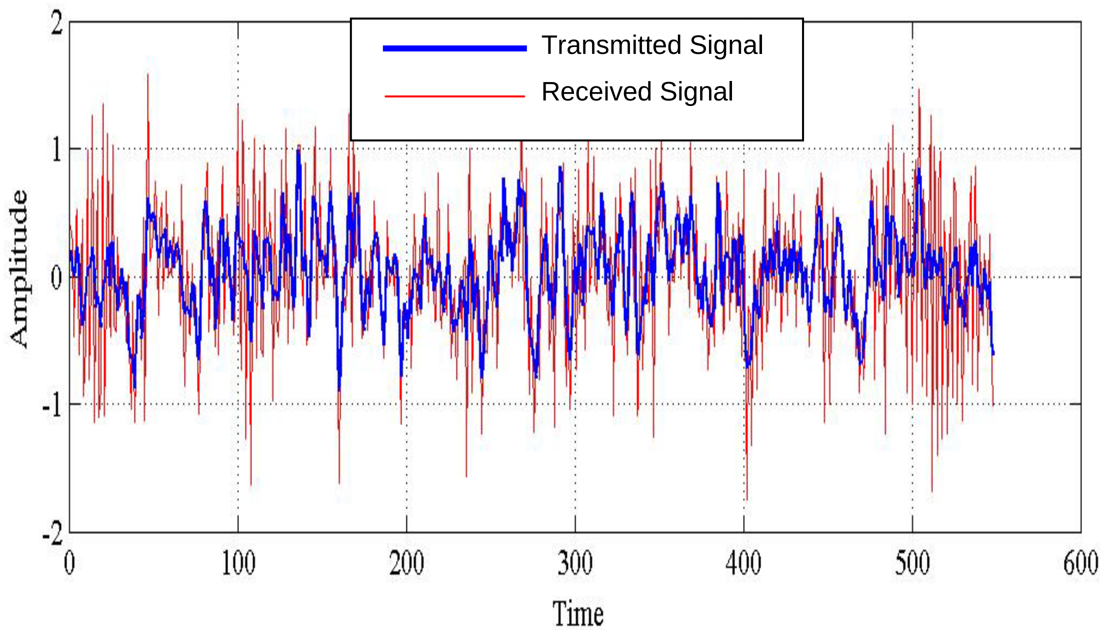


Figure 5.4 Typical time domain representations of (a) data before and (b) after cyclic Here the addition of CP has made the length of Symbols to 2N + $\boldsymbol{v}$

To this data stream, Additive White Gaussian Noise is added and the two figures below show the resulting signal for two SNR values.



(a)



(b)

Figure 5.5 AWGN added signal (Symbr (k) over Symb(k)) with a) SNR = 0 dB and b) SNR = 20 dB

As it is possible to see in figure 5.5, the deviation of the noise added signals is more pronounced when the SNR is of lower values. That is, in figure 5.5.b, where the noise is of less power with respect to the signal compared to that in figure 5.5a, the received signal follows the sent signal more closely.

Next seen are the values of the time domain equalizer, heq(n) and /Heq(k)/ settings. Especially the heq(n) shows a "comb like" curve which is similar to typical equalizer curves in literatures.

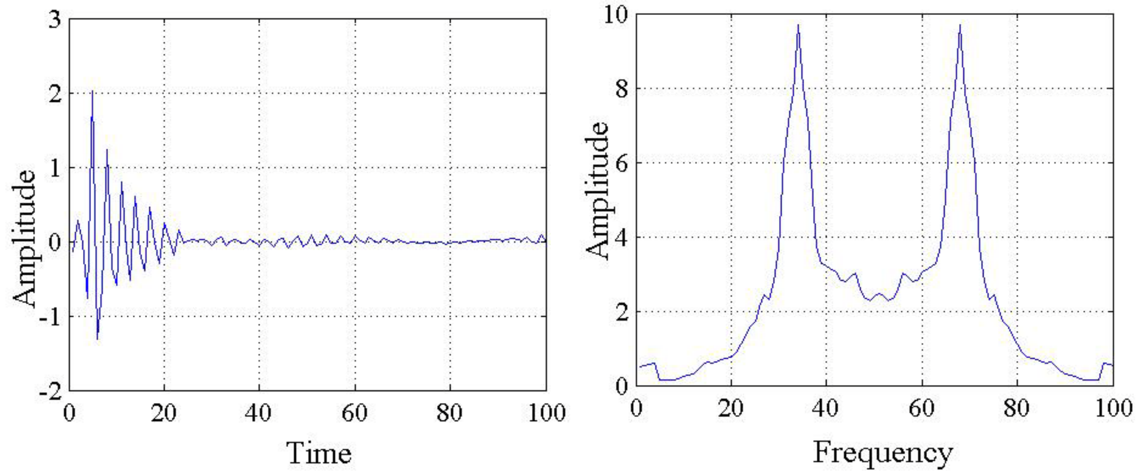


Figure 5.6 Time Domain Equalizer a] Time samples & b) Magnitude of its frequency response.

In Figure 5.7 below, the spectral containment of the data before the channel & after the effects of the channel & noise are shown. Seen here is that the low pass effects of the channel have played a frequency selective distortion over the transmitted data.

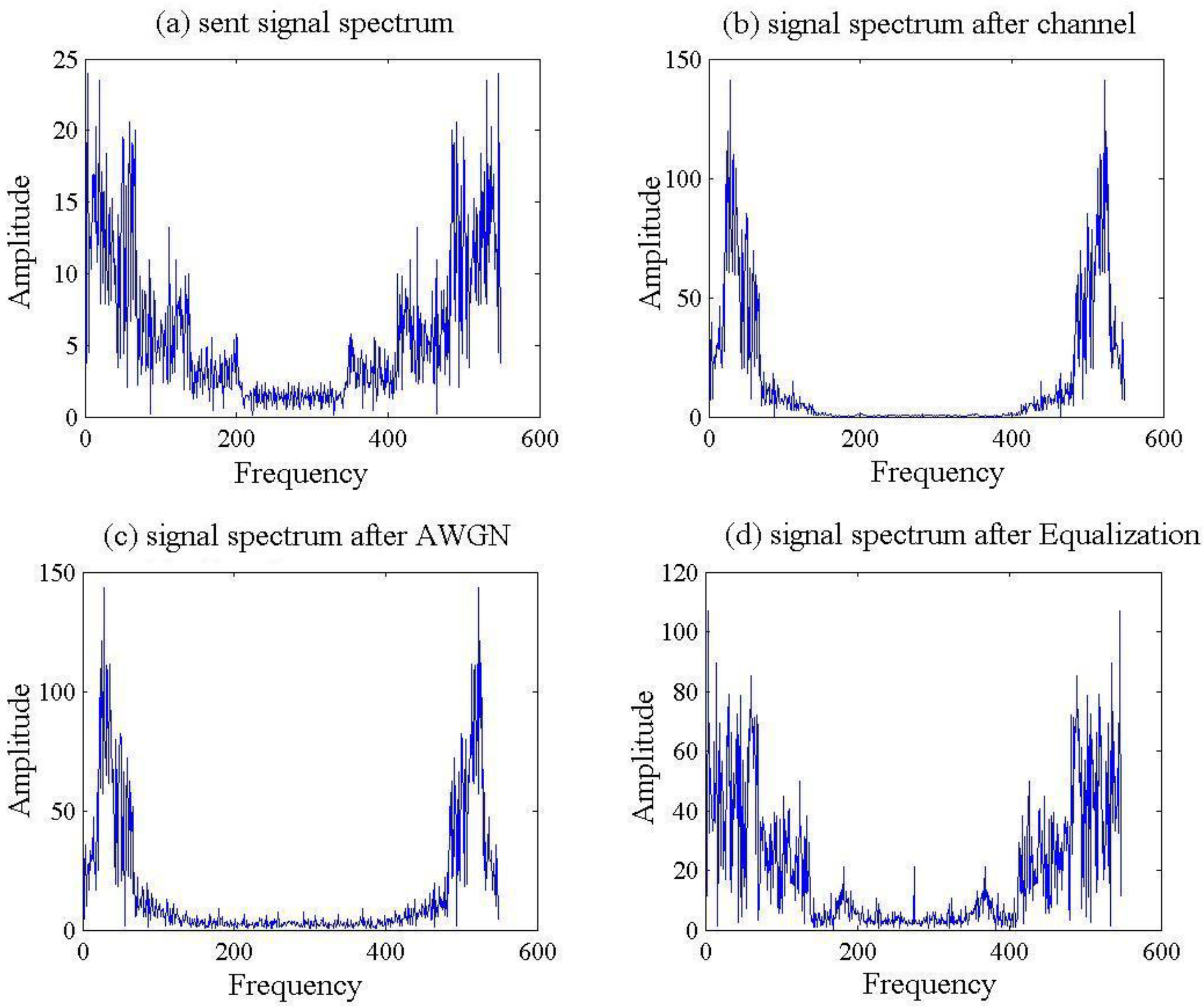


Figure 5.7 Frequency Response Amplitude of different signals followed from transmitted signal up to after equalization stage (Spectral Magnitude per Subchannel Index)

Xcpr(n) is the first signal after equalization and the next signal, X2Nr(n), is received after cyclic prefix removal. An FFT applied to this signal would give Symb2Nr(k) which is spectral representation of X2Nr(n). A removal of even symmetry is done using equation 4.5 and (optionally after multiplication with Frequency domain equalization (FEQ)) the value obtained is the received estimate of the sent symbol stream. A magnitude plot of this symbol estimate is given in figure 5.8 and this is to be compared to the sent symbol given is figure 5.3.

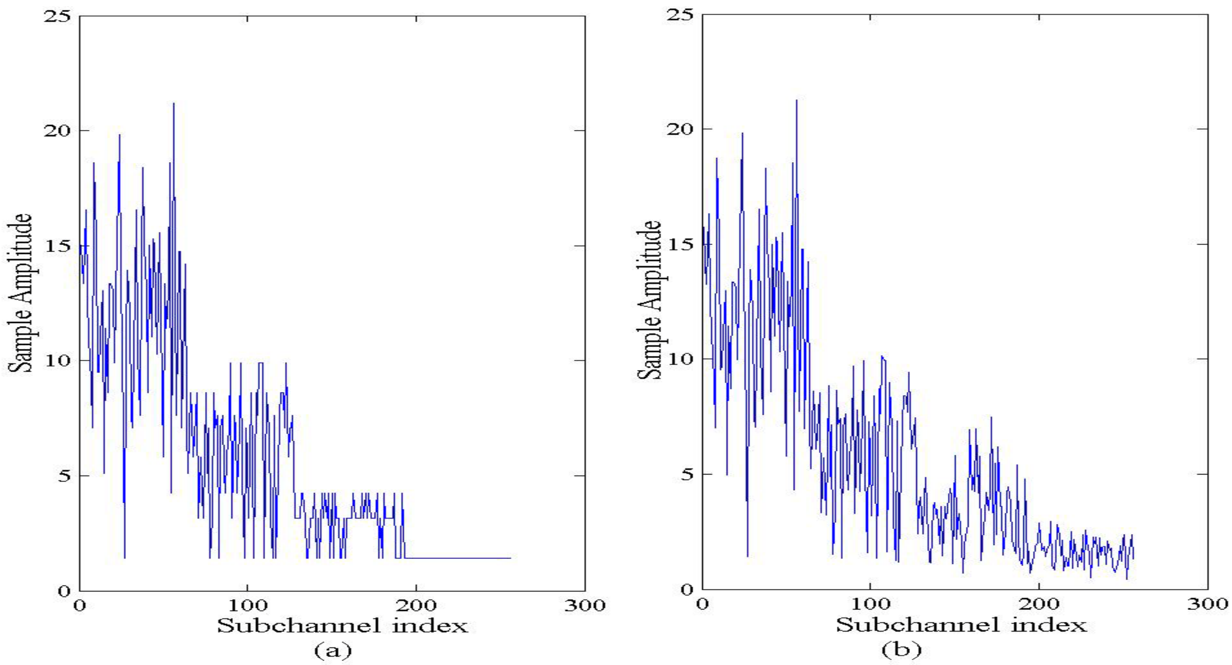


Figure 5.8 A magnitude plot of typical (a) transmitted and (b) estimate symbol stream.

Here, for each subchannel, a decision algorithm tries look for the closest symbol within its corresponding QAM constellation diagram. Upon deciding on so, then, the algorithm decides that this was the symbol sent for that subchannel. Finally, corresponding look up tables are used to decode the symbols into bit groups of subsequent subchannels. A conversion of these bit groups per subchannel into a serialized format over the whole channel would then give the received bit stream.

As it can be seen, then, the BER is obtained as the number of different points between the sent bit stream and the received bit stream. The result would of course be complete if this number is divided by the total number of bits sent or received. This total number of bits sent /received/ will be the sum of elements of the bit loading array. In this work this comes to an average of loading of 1152 bits/channel/Hertz.

The final presentation of performance measurement is the Bit Error Rate analysis effected using different possible analysis for different SNR values. The figures given below show these different analysis approaches and reveal the unbeaten advantage of the equalization resulting in significant reduction of BER.

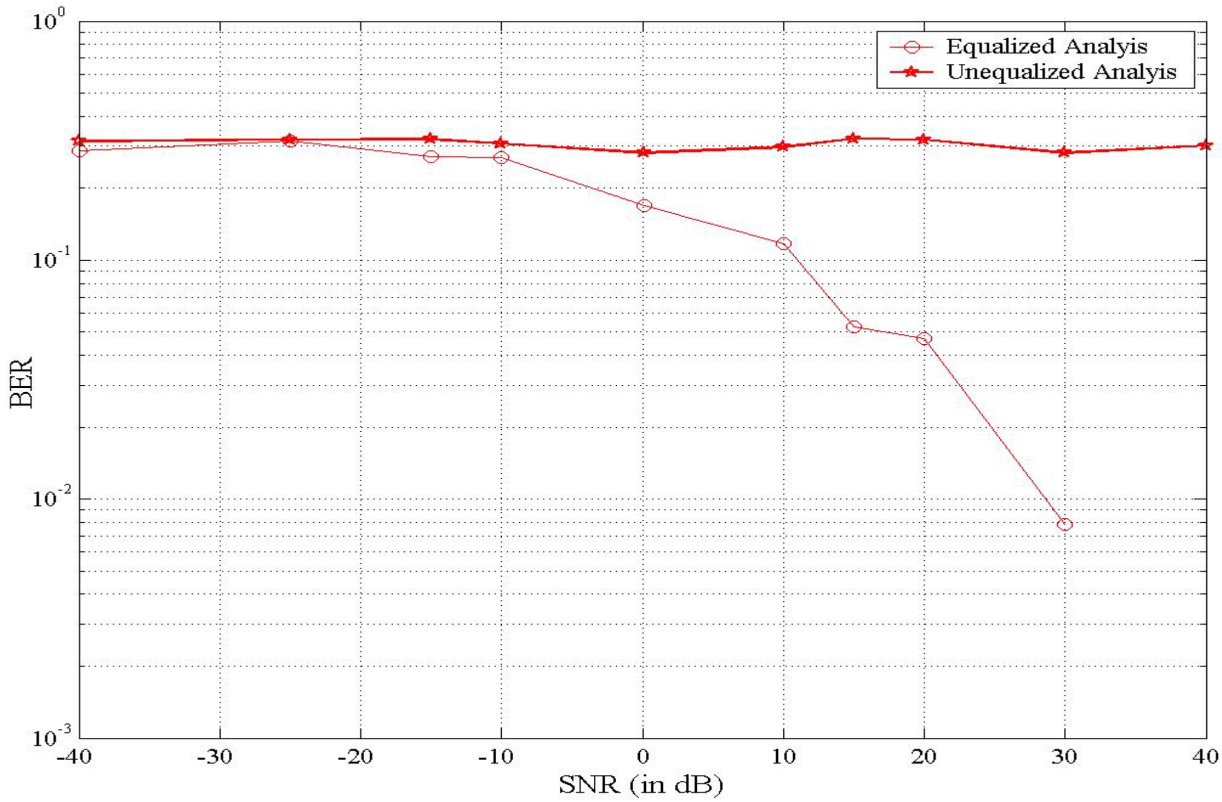


Figure 5.9.a) Performance plot for DMT based end to end realization

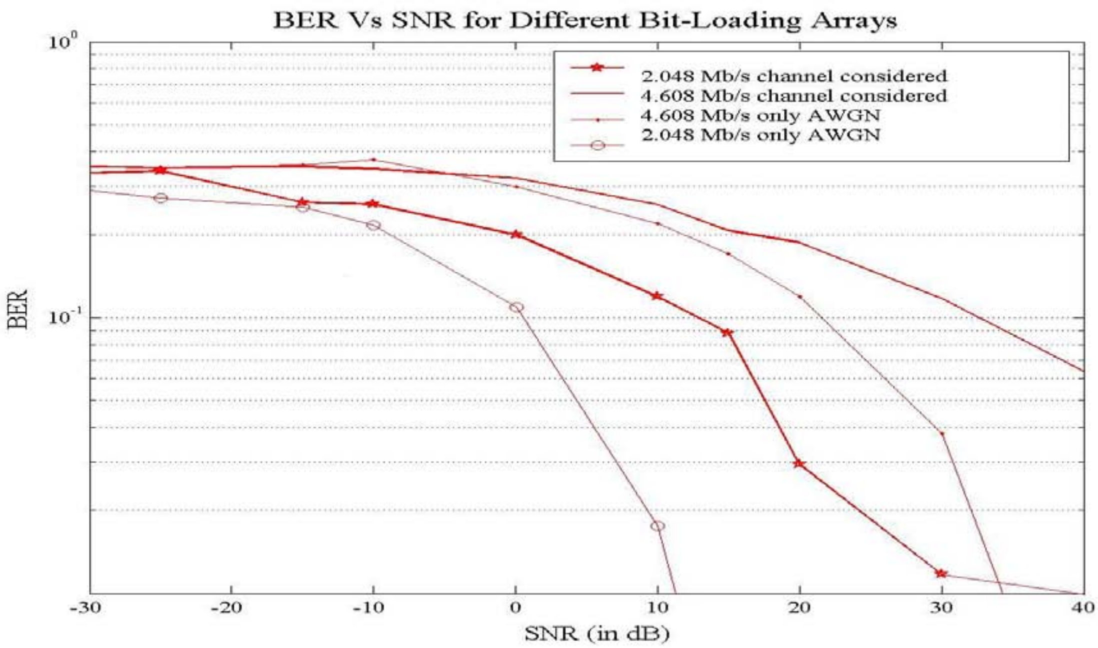


Figure 5.9.b) Performance plot when only AWGN impairments are considered and when effects of the channel are considered.

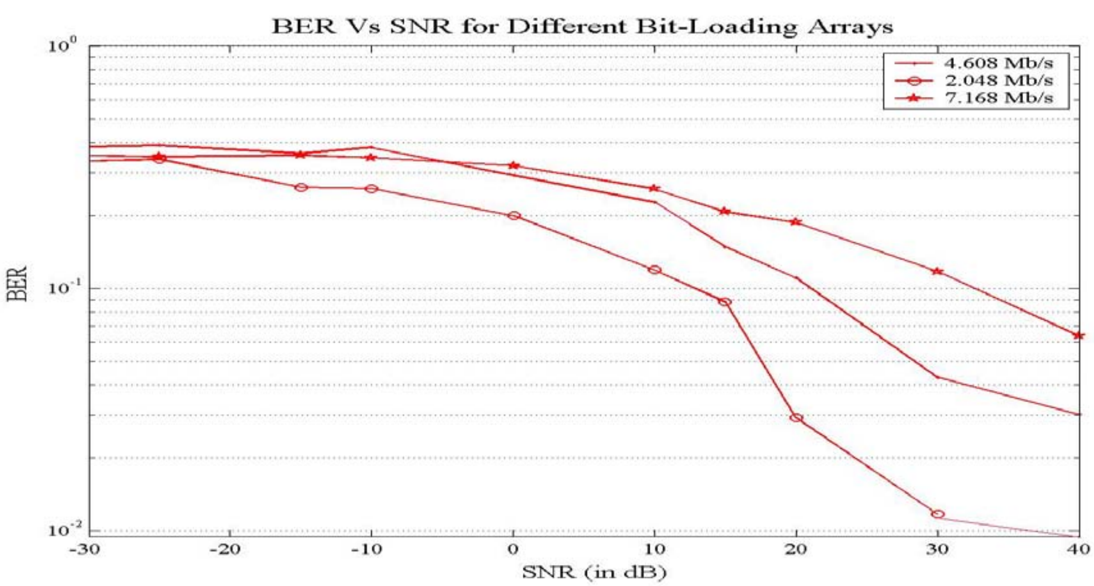


Figure 5.9.c) Performance plot with bit loading arrays of different bits/symbols/hertz -- In this plot the bitloading ranges of 2 bits/symbols/hertz (2.048 Mb/s), 2 - 8 bits/symbols/hertz (4.608 Mb/s) and 2 - 12 bits/symbols/hertz (7.168 Mb/s).

Chapter 6

Conclusions and Recommendations

6.1 Conclusions

6.1.1 Project Work Preliminaries

This paper is a report of a simulation work of a Discrete Multitone (DMT) based Asymmetrical Digital Subscriber Line (ADSL) System. The simulation is performed for a typical telephone line (with length of 2 kM & diameter of 0.4 mm) and it is assumed that this line is carrying a signal, which is obtained after a Quadrature Amplitude Modulation (QAM), and Multicarrier Modulation (MCM) performed over a 1024 to 1664 bits length bit stream.

The telephone (wire) line is one of the most widely used; most studied and well documented telecom media. The studies however, are almost always performed with respect to its behaviors over the 4 kHz and below parts of the frequency spectrum. This was done because basically the telephone line is used for voice and telephony applications. The present day choice of this transmission medium for broad band applications to the home is due to the practicality of this widely existing copper network. Nonetheless, the properties of the copper based telecom network are not well studied for higher parts (greater than 500 kHz) of the frequency spectrum. Additionally, these properties are highly differing between manufacturers.

Therefore, studies of signal transmission applications using the higher frequency spectrum of the telephone line network are usually done using measured or empirical data of the telephone line. In this paper, measured data of the telephone line is used to simulate the effects of this transmission medium over the transmitted signal. Moreover, this data is directly used to produce an inverse filtering equalizer and also indirectly the frequency domain equalizers (FEQ).

The other point of interest stems from the coverage area of the project work under report. Instead of focusing on the parts/ subunits of the ADSL communications system, this project work is work on an end-to-end simulation of an ADSL system. To simplify this end-to-end system simulation, and to show the presentable results within the scope of the work, some of the subunits of the system are purposely omitted. The largest one of these is the error control coding part. When this subunit is considered it is possible to realize error control coding in either of Reed Solomon (RS) type block/Cyclic coding or of the Viterbi type Convolutional coding schemes. A number of papers [7], [10], [13] show that inclusion of error control coding schemes result in a significant reduction of bit error (BER) values.

6.1.2 Summary of the project work results

As it is demonstrated in Section 5, all the components shown in figure 5.1 have been simulated in MATLAB script code. The telephone line is used as a channel & its highly low-pass characteristics have been observed. The results obtained can be summarized as follows:

A) When effects of the channel are neglected and only AWGN is considered, it is seen that values of less than 1 % BER (for 10 dB SNR) were achieved. Also consistently enough, higher SNR was yielding lower BER values (see figure 5.9.b).

B) When the dominant effects of channel impairments are considered in addition to the effects of AWGN, there are two scenario

i. When no equalization is used

Here effects of the channel are seen to significantly distort the signal transmitted. This has put a very destructive performance over the bit transmitted resulting in the BER values of more than 30 % for even large SNR values; as seen in figure 5. 9. a)).

ii. When an equalizer stage is included

With an inclusion of an equalization scheme, the BER values obtained were much lower (in the range of 1–8 % at 40 dB, see figure 5.9.a), hence enhancing performance significantly.

C) As a final comparison stage, different bit loading arrays are used to approach the effects of loading the channel with different bit rates hence testing the channel capacity. In this approach, it was seen (in figure 5.9.c) that higher bit rates indeed are faced with higher BER. Hence approaching the channel capacity (cut-off rate) does in deed only gets away with larger data loss.

6.2 Recommendation

From the results of this project work, two points of interest are straight forwarded for better performance sought:

A) From engineering and academics point of view, project works of an end-to-end communication link study are the best mechanisms for a complete and detailed understanding of communications systems. However, these studies should be based on subunits previously well studied with accessibility of possible parameter variations. Only then can a focused study of the end-to-end system can be envisaged. Also, in this work, all the subunits were prepared from a scratch and the greater part of the study was invested with development of subunits (some of them complicated – requiring of project works of their own – for example inverse filtering in direction of the Time Domain Equalizer /TEQ/)

B) With this work, most of the subunits considered have been well addressed and hence it is hoped that a continual of this work is much better facilitated. This work can be enhanced in two major ways:

a. Perfection of the equalization schemes and other process algorithms – With this, the performance measures (BER and speed of results sought) can be enhanced to much lower values.

b. Inclusion of Error Control Coding schemes and considering pulse shaping filters – Also these subunits are known from [7] and [10], for their enhancement of system performance. Possibly a comparison could be made with systems with out these subunits, with RS error control coding subunit, with Viterbi Convolutional error control coding subunit and the above options with pulse shaping filters.

The other big advantage of MCM for ADSL is the possibility of small step rate-adaptation without great increase of complexity, as it is to be expected that in future rate-adaptive ADSL will take over a dominant position on the market. All the reasons that finally have led to the decision for MCM (DMT (see chapter 6) is a specific form of MCM) will also have to be considered in the selection of the modulation scheme for VDSL. Here again it is to be expected that the ease of rate-adaptation and the flexibility in spectrum usage (especially important to combat radio frequency interference (RFI) in the case of VDSL) will play a main role in the decision.

For the the rate-adaptive ADSL (RADSL), envisioned for the future, the situation is somehow different, as there the total permitted transmit power will be used in order to achieve the maximal bit rate.However, up to now there also appeared other forms of MCM (especially DWMT) that might be serious competitors for the DMT scheme, defined for ADSL.

Figure 5.7 Spectral containment of the data before the channel & after the TEQ

The advantage of DMT is the possibility of fully digital implementation and (compared with other MCM schemes) relatively low complexity

The other big advantage of MCM for ADSL is the possibility of small step rate-adaptation without great increase of complexity, as it is to be expected that in future rate-adaptive ADSL will take over a dominant position on the market. All the reasons that finally have led to the decision for MCM (DMT (see chapter 6) is a specific form of MCM) will also have to be considered in the selection of the modulation scheme for VDSL. Here again it is to be expected that the ease of rate-adaptation and the flexibility in spectrum usage (especially important to combat radio frequency interference (RFI) in the case of VDSL) will play a main role in the decision. However, up to now there also appeared other forms of MCM (especially DWMT) that might be serious competitors for the DMT scheme, defined for ADSL.

```
%%********************************************************************* %%
%%                                                                      %%
%%              Implementation of the ADSL Simulation                   %%
%%                                                                      %%
%% ******************************************************************** %%
%%                                                                      %%
%%                                                                      %%
%%     Note                                                             %%
%%                                                                      %%
%%     Input is a random (spectrally white) bit-stream                  %%
%%     Bit Loading done for two prespecified Bit-Loading Array          %%
%%     Channel Model is obtained from [16]                              %%
%%     Equalizer is of Inverse Filtering Model -- With Assumption of    %%
%%     Slowly Time Varying Channel                                      %%
%%                                                                      %%
%%     By   Abel Seife Selassie                                         %%
%%     November 2003                                                    %%
%%                                                                      %%
%%                                                                      %%
%% ******************************************************************** %%

%%%%  *********          Initialization     *************    %%%%

%%%%  *********          Channel
hn1 = [0 0.0157 0.0275 0.0284 0.018 -0.032 -0.156 -0.156 0.114 0.93 1 0.924 0.4 0.101
     -0.18 -0.45 -0.6 -0.652 -0.604 -0.385];
hn2 = [-0.144 0.065 0.214 0.316 0.397 0.423 0.412 0.376 0.324 0.228 0.15 0.03 -0.08
     -0.17 -0.2118 -0.2295 -0.245 -0.248 -0.239 -0.223 ];
hn3 = [-0.198  -0.153 -0.10874 -0.06401 -0.01928 0.023 0.07018 0.11 0.132 0.143 0.15
     0.146 0.135 0.123 0.108 0.085 0.061 0.036 0.009 ];
hn4 = [-0.012 -0.03 -0.042 -0.0526 -0.060 -0.067 -0.073 -0.077 -0.079 -0.073 -0.063
     -0.0516 -0.0382 -0.0234 -0.0107 0.0013 0.012 0.023 ];
hn5 = [0.033 0.041 0.0469 0.05 0.047 0.0422 0.0366 0.0312 0.0264 0.021 0.0164 0.0111
     0.006 0.0005];

hn = [hn1 hn2 hn3 hn4 hn5];
hn_v_sh=[hn(1:91) zeros(9,1)'];
H = fft(hn_v_sh,548);                        % Channel Spectral Values

%%%%  *********          Equalizer
           EQ = 1./(H);                     % Inverse Filtering Equalizer

%%%%  *********          Frequency Equalizer
%%    FEQ = SymbNk./SymbNkr;

%%%%  *********   Analysed Transmission SNR Values

%%  Various Transmission SNR Values are used for performance measurement
%% Prescribed SNR of Transmission in dB
     SNR = [ -40 -25 -15 -10 0.1 10 15 20 30 40];

%% ******************************************************************** %%
```

```
%%%%  *********          Bit-Loading

%%  Subchannels divided into four groups of equal bit carrying capacity
len_1=64; len_2=64; len_3=64; len_4=64;

Bit_Loading_Choice = input ( ' Press H for Higher Bit-Loading or L for
Lower Bit-Loading : - ','s')

  if  Bit_Loading_Choice == 72    %Higher Bit-Loading <-> ascii code of H
      SNR_CH_1 = 46; SNR_CH_2 = 40;  SNR_CH_3 = 22; SNR_CH_4 = 16; %in dB
      %%   m_1 = 12       m_2 = 10        m_3 = 4        m_4 = 2

  elseif Bit_Loading_Choice == 76  %Lower Bit-Loading === ascii code of L
        SNR_CH_1 = 34; SNR_CH_2 = 28;  SNR_CH_3 = 22; SNR_CH_4 = 16;
    %%    SNR_CH_1 = 16; SNR_CH_2 = 16;  SNR_CH_3 = 16; SNR_CH_4 = 16;
        %%   m_1 = 8        m_2 =  6        m_3 = 4        m_4 = 2
    else
      disp(' Ungiven choice -- Lower Bit-Loading Value taken as default')
        SNR_CH_1 = 34; SNR_CH_2 = 28;  SNR_CH_3 = 22; SNR_CH_4 = 16;
    end;

m_1 = round ( log2( 1+10^(0.1*SNR_CH_1 - 0.98) ) );
m_2 = round ( log2( 1+10^(0.1*SNR_CH_2 - 0.98) ) );
m_3 = round ( log2( 1+10^(0.1*SNR_CH_3 - 0.98) ) );
m_4 = round ( log2( 1+10^(0.1*SNR_CH_4 - 0.98) ) );

M_1 =2^m_1; M_2 = 2^m_2;  M_3 =2^m_3; M_4 =2^m_4;


%% ******************************************************************* %%

%%%%  *********          Data Generation & Transmission

for jj=1:10
            % Generate a matrix of length len_i random binary array
            % of m_i bits .
    b1 = randint(len_1 ,m_1);
    b2 = randint(len_2 ,m_2);
    b3 = randint(len_3 ,m_3);
    b4 = randint(len_4 ,m_4);

    %%%%  ********   Converting the bit stream decimals

    for i = 1:len_1
        de1(i) = bi2de(b1(i,:)); de2(i) = bi2de(b2(i,:));
        de3(i) = bi2de(b3(i,:)); de4(i) = bi2de(b4(i,:));
    end;


    %%%%  ********  QAM modulation impementing gray coding
```

```
[x1, y1] = qaskenco(de1,M_1);
[x2, y2] = qaskenco(de2,M_2);
[x3, y3] = qaskenco(de3,M_3);
[x4, y4] = qaskenco(de4,M_4);


%%% **********  Preparing the QAM Symbol streams for MCM
  xrev1 = x1'; xrev2 = x2'; xrev3 = x3'; xrev4 = x4';
  yrev1 = y1'; yrev2 = y2'; yrev3 = y3'; yrev4 = y4';
%% Transposed so as to easily make the SymbNk array as follows !!!

Symb = [xrev1+j*yrev1  xrev2+j*yrev2 xrev3+j*yrev3 xrev4+j*yrev4];

N = len_1 + len_2 + len_3 + len_4;            % Sub_ch_len  = 256


%%% **********  And putting in a 2N point Hermitian Symmetry
  Symb2N(1) = real(Symb(1));
  Symb2N(2:N)= Symb(2:N);
  Symb2N(N +1)= imag(Symb(1));

for k = N +2 : 2*N                                   % 258 -- 512
    Symb2N(k)= conj(Symb(2*N-(k-2)));
end;

%%% ***********  Realization of the MCM modulation

  X2N = ifft(Symb2N);     %  Max of imag(Symb2Nk) ~~ 1.7039e-016
  X2N = real(X2N);

%%%  **********  Adding Cyclic Prefix
          %% Assuming a length of cp= 36  --- i.e, ~ < 7% of 512
CP = 36;
Xcp = [X2N(2*N -(CP-1):2*N) X2N];


%% ******************************************************************* %%

%%%  **********  Channel effects

XCPR_H = (fft(Xcp).*(H));
Xcpr_h = real(ifft(XCPR_H));


%%%  **********  Inclusion of the Additive White Gaussian Noise
  Xcpr_h_noisy = (awgn(Xcpr_h,SNR(jj),'measured'));

%% ******************************************************************* %%

%%%%   *************  Equalization Usage Choice

 Equalizer_Usage_Choice = input ( ' Press E for Equalized or U for
 Unequalized Simulation : - ','s')
```

```
if Equalizer_Usage_Choice == 69        % Using channel equalization
                                       % <-> ascii code of E
    XCPR_NSY = fft(Xcpr_h_noisy);
    XCPR_EQ = XCPR_NSY .* EQ;
    Xcpr = real(ifft(XCPR_EQ));
    X2Nr = Xcpr(CP+1 : 2*N + CP);

elseif Equalizer_Usage_Choice == 85       % Not using channel
                                   % equalization === ascii code of U
    X2Nr = Xcpr_h_noisy(CP+1 : 2*N + CP);
else
  disp(' Ungiven choice pressed  -- Option is lost !!! -- Press enter
       to continue ')
    return;
end;

%%% **********  Generating the N-point Symbol Stream
          %% 2N point to N point Symbol Stream Conversion --
          %% Doing away with the Hermitian Symmetry
Symb2Nr = fft(X2Nr);



SymbNr(1)= Symb2Nr(1) + j*Symb2Nr(N+1);
SymbNr(2:N)= Symb2Nr(2:N);

%%% **********  Frequency Domain Equalization
%% SymbNr = SymbNr .* FEQ;

            %% Preparing for QAM demodulation with real & Imaginary
            %% partitioning of the Symbol Stream

xrev1r(1:len_1) = real(SymbNr(1:len_1)) ;
xrev2r(1:len_1) = real(SymbNr(len_1+1:2*len_1));
xrev3r(1:len_1) = real(SymbNr(2*len_1+1:3*len_1));
xrev4r(1:len_1) = real(SymbNr(3*len_1+1:4*len_1));

yrev1r(1:len_1) = imag(SymbNr(1:len_1));
yrev2r(1:len_1) = imag(SymbNr(len_1+1:2*len_1));
yrev3r(1:len_1) = imag(SymbNr(2*len_1+1:3*len_1));
yrev4r(1:len_1) = imag(SymbNr(3*len_1+1:4*len_1));

x1r = round(xrev1r'); x2r = round(xrev2'); x3r = round(xrev3');
x4r = round(xrev4');
y1r = round(yrev1r') ; y2r = round(yrev2r');
y3r = round(yrev3r'); y4r= round(yrev4r');


%%%  **********   QAM demodulation
           %% Determining the received bit stream
           %% from the received Symbol
```

```
    de1r = (qaskdeco(x1r,y1r,M_1))';
    de2r = (qaskdeco(x2r,y2r,M_2))';
    de3r = (qaskdeco(x3r,y3r,M_3))';
    de4r = (qaskdeco(x4r,y4r,M_4))';

     %%%  ********** Converting received samples into bit streams
                   %% for Performance Measurement  (BER Vs SNR)
    for i = 1:len_1
        b1r(i,:) = de2bi(de1r(i), m_1);
        b2r(i,:) = de2bi(de2r(i), m_2);
        b3r(i,:) = de2bi(de3r(i), m_3);
        b4r(i,:) = de2bi(de4r(i), m_4);
    end;

    %%%  Calculating the Bit Error Rate

    [number1,ratio1] = biterr(b1,b1r);
    [number2,ratio2] = biterr(b2,b2r);
    [number3,ratio3] = biterr(b3,b3r);
    [number4,ratio4] = biterr(b4,b4r);

    %%%  BER per each Transmission SNR
    BER (jj) = (number1 +number2 +number3 +number4) / (len_1*m_1 +
len_2*m_2 +len_3*m_3 +len_4*m_4 );

end; %% jj

plot (SNR,BER,'r')

%% ******************************************************************* %%
```

Appendix

Appendix A. MATLAB ® Script code for Implementation of a DMT based DSL System Performance Analysis

Appendix B. Bibliography

Bibliography

11[6] [1] John A. C. Bingham, “Multicarrier Modulation for Data Transmission: An Idea Whose Time Has Come,” IEEE Communication Magazine, pp. 5-14, May. 1990
11[7] [2] John A. C. Bingham, “ADSL, VDSL, and Multicarrier Modulation,” John Wiley & Sons, Inc. 2000

[13] [3] David W. Lin et al., Video on Phone Lines: Technology and Applications, Proceedings of the IEEE, Vol. 83, NR. 2, February 1995

[5] [4] Joseph W. Lechleider, High Bit Rate Digital Subscriber Lines: A Review of HDSL Progress, IEEE Journal on Selected Areas in Communications, Vol. SAC-9, NR. 6, August 1991

[12] [5] DAVIC (Digital Audio Video Council) , http://media.etri.re.kr/DAVIC/spec/spec10/part8/part8.html

[18] [6] Sahid V. H. Qurechi, Adaptive Equalization, Proceedings of the IEEE, Vol. 73, NR. 9 September 1985, pages 1349-1487

[19] [7] John A. C. Bingham, Multicarrier Modulation for Data Transmission, An Idea whose Time has come, IEEE Communications Magazine, May 1990, pages 5-14

[20] [8] Kamran Sistanizadeh, Kenneth J. Kerpez, A Comparison of Passband and Baseband Transmission Schemes for HDSL, IEEE Journal on Selected Areas in Communication, Vol.9, NR. 5, August 1991, Pages 885-894

[1] [9] G. David Forney, jr, et al., Efficient Modulation for Band-Limited Channels, IEEE Journal on Selected Areas in Communications, Vol. SAC-2, NR. 5, September 1984

[22] [10] Joseph. W. Lechleider, Line Codes for Digital Subscriber Lines, IEEE Communications Magazine, September 1989, pages 25-32

[21] [11] James T. Aslanis et al, Achievable Information Rates on Digital Subscriber Loops: Limiting Information rates with Crosstalk-Noise, IEEE Transactions on Communications, Vol. 40, NR. 2, February 1992, pages 361-172

[32] [12] Irving Kalet, The Multitone Channel, IEEE Transactions on Communications, Vol. 37, NR. 2, February 1989, pages 119-124

[28] [13] Simon Haykin, Communication System, 3rd edition, Wiley, 1994

[31] [14] R.G. Gallager, Information Theory and Reliable Communication, New York: John Wiley and sons, 1968

11[1] [15] ANSI Standard T1.413 issue 2, “Asymmetric Digital Subscriber Line Metallic Interface,” 1998

11[8] [16] John M. Cioffi, “A Multicarrier Primer,” Amati Communications Corporation and Stanford University, 1991

[17] [17] Peter S. Chow et al., Performance Evaluation of a Multichannel Transceiver System for ADSL and VHDSL Services, IEEE Journal on Selected Areas in Communications, Vol. 9, NR.6, August 1991, pages 909-919

[23] [18] Orckit Knowledge Center, http://www.orckit.com/knowledge.html

[45] [19] Ove Edfors et al. An introduction to orthogonal frequency-division multiplexing, Luleå University of Technology, September 1996

[48] [20] Jacky S. Chow, Jerry C. Tu, John M. Cioffi, A Discrete Multitone Transceiver System for HDSL Applications, IEEE Journal on Selected Areas in Communications, Vol. 9, NR. 6, August 1991, pages 895-908

[49] [21] Melbourne Barton, Michael L. Honig, Optimization of Discrete Multitone to Maintain Spectrum Compatibility with other Transmission Systems on Twisted Copper Pairs, IEEE Journal on Selected Areas in communications, Vol. 13, NR. 9, December 1995, pages 1558-1563

[15] [22] Stuart D. Sandberg, Michael A. Tzannes, Overlapped Discrete Multitone Modulation for High Speed Copper Wire Communications, IEEE Journal on selected areas in communications, Vol. 13, NR. 9, December 1995, pages 1571-1585

[50] [23] Inkyu Lee, Jacky D. Chow, John M. Cioffi, Performance Evaluation of a Fast Computation Algorithm for the DMT in High-Speed Subscriber Loop, IEEE Journal on Selected Areas in Communications, Vol. 13, NR. 9, December 1995, pages 1564-1570

[51] [24] David C. Jones, Frequency Domain Echo Cancellation for Discrete Multitone Asymmetric Digital Subscriber Line Transceivers, IEEE Transactions on Communications, Vol. 43, NR. 2/3/4, February/March/April 1995, pages 1663-1672

[36] [25] John Cook, Phil Sheppard, ADSL and VADSL Splitter Design and Telephony Performance, IEEE Journal on Selected Areas in Communications, Vol. 13, NR. 9, December 1995, pages 1634-1642

Intentionally blank

Intentionally blank

Appendix B

Bibliography

[1] John A. C. Bingham, “Multicarrier Modulation for Data Transmission: An Idea Whose Time Has Come,” IEEE Communication Magazine, pp. 5-14, May. 1990

[2] Ahmed R.S. Bahai and et al, “Muti-Carrier Digital Communications,” Kluwer Academic/ Plenum Publishers. 1999

[3] David W. Lin et al., Video on Phone Lines: Technology and Applications, Proceedings of the IEEE, Vol. 83, NR. 2, February 1995

[4] Joseph W. Lechleider, High Bit Rate Digital Subscriber L ines: A Review of HDSL Progress, IEEE Journal on Selected Areas in Communications, Vol. SAC-9, NR. 6, August 1991

[5] Liang C. Chu, ADSL System Enhancement with Multiuser Detection, Georgia Institute of Technology, July 2001.

[6] Sahid V. H. Qurechi, Adaptive Equalization, Proceedings of the IEEE, Vol. 73, NR. 9 September 1985, pages 1349-1487

[7] John G. Proakis Digital Communications, 3rd McGraw-Hill International, Electrical Engineering Series, New York, U.S.A. 1995.

[8] Kamran Sistanizadeh, Kenneth J. Kerpez, A Comparison of Passband and Baseband Transmission Schemes for HDSL, IEEE Journal on Selected Areas in Communication, Vol.9, NR. 5, August 1991, Pages 885-894

[9] G. David Forney, jr, et al., Efficient Modulation for Band-Limited Channels, IEEE Journal on Selected Areas in Communications, Vol. SAC-2, NR. 5, September 1984

[10] Daniel Franklin et al., An Improved Channel Model for ADSL and VDSL Systems, February 1999.

[11] James T. Aslanis et al, Achievable Information Rates on Digital Subscriber Loops: Limiting Information rates with Crosstalk-Noise, IEEE Transactions on Communications, Vol. 40, NR. 2, February 1992, pages 361-172

[12] Irving Kalet, The Multitone Channel, IEEE Transactions on Communications, Vol. 37, NR. 2, February 1989, pages 119-124

[13] Simon Haykin, Communication System, 3rd edition, Wiley, 1994

[14] R.G. Gallager, Information Theory and Reliable Communication, New York: John Wiley and sons, 1968

[15] ANSI Standard T1.413 issue 2, “Asymmetric Digital Subscriber Line Metallic Interface,” 1998

[16] John M. Cioffi, “A Multicarrier Primer,” Amati Communications Corporation and Stanford University, 1991

[17] Peter S. Chow et al., Performance Evaluation of a Multichannel Transceiver System for ADSL and VHDSL Services, IEEE Journal on Selected Areas in Communications, Vol. 9, NR.6, August 1991, pages 909-919

[18] Orckit Knowledge Center, http://www.orckit.com/knowledge.html

[19] Ove Edfors et al. An introduction to orthogonal frequency-division multiplexing, Luleå University of Technology, September 1996

[20] Jacky S. Chow, Jerry C. Tu, John M. Cioffi, A Discrete Multitone Transceiver System for HDSL Applications, IEEE Journal on Selected Areas in Communications, Vol. 9, NR. 6, August 1991, pages 895-908

[21] Melbourne Barton, Michael L. Honig, Optimization of Discrete Multitone to Maintain Spectrum Compatibility with other Transmission Systems on Twisted Copper Pairs, IEEE Journal on Selected Areas in communications, Vol. 13, NR. 9, December 1995, pages 1558-1563

[22] Stuart D. Sandberg, Michael A. Tzannes, Overlapped Discrete Multitone Modulation for High Speed Copper Wire Communications, IEEE Journal on selected areas in communications, Vol. 13, NR. 9, December 1995, pages 1571-1585

[23] Inkyu Lee, Jacky D. Chow, John M. Cioffi, Performance Evaluation of a Fast Computation Algorithm for the DMT in High-Speed Subscriber Loop, IEEE Journal on Selected Areas in Communications, Vol. 13, NR. 9, December 1995, pages 1564-1570

[24] David C. Jones, Frequency Domain Echo Cancellation for Discrete Multitone Asymmetric Digital Subscriber Line Transceivers, IEEE Transactions on Communications, Vol. 43, NR. 2/3/4, February/March/April 1995, pages 1663-1672

[25] John Cook, Phil Sheppard, ADSL and VADSL Splitter Design and Telephony Performance, IEEE Journal on Selected Areas in Communications, Vol. 13, NR. 9, December 1995, pages 1634-1642

[5] Liang C. Chu, ADSL System Enhancement with Multiuser Detection, Georgia Institute of Technology, July 2001.

[10] Daniel Franklin et al., An Improved Channel Model for ADSL and VDSL Systems, February 1999.

OLD ONES

[2] John A. C. Bingham, "ADSL, VDSL, and Multicarrier Modulation," John Wiley & Sons, Inc. 2000

[5] DAVIC (Digital Audio Video Council), http://media.etri.re.kr/DAVIC/spec/spec10/part8/part8.html

[10] Joseph. W. Lechleider, Line Codes for Digital Subscriber Lines, IEEE Communications Magazine, September 1989, pages 25-32