

BROADBAND POWER-LINE NOISE MODELING AND COMPENSATION

by

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ABSTRACT

BROADBAND POWER-LINE NOISE MODELING AND COMPENSATION

Broadband power-line communication (BPLC) has become more popular in the Smart Grid context as it allows broadband access, media sharing, automated meter reading etc. However, it still remains as a challenging task to characterize and compensate for the impairments of the power-line networks for high-speed communications.

In the scope of the thesis study, multi-path fading and path loss of the power-line channel are studied and a unit length attenuation profile (ULAP) is derived from measurements as part of the channel characterization. Using ULAP and the other channel parameters, a power-line channel simulator is developed. Correctness of the simulation is verified by measurement. Using the PLC channel simulations, statistics of the channel parameters are also studied.

During the course of the thesis work, both broadband noise and home appliance noise are studied and characterized. Background noise is analyzed in frequency domain and Generalized Gaussian distribution (GGD) is proposed for modeling the noise amplitude distribution of each individual frequency. The proposed model is verified by measurements in the Kullback-Leibler sense of distribution mismatch penalty. Using the proposed noise model, a novel Orthogonal Frequency Division Multiplexing (OFDM) based receiver is proposed. It is shown that the performance of the proposed receiver utilizing the parameters of the proposed noise model reveals promising results.

ÖZET

ELEKTRİK HATTINDA GENİŞBANTLI GÜRÜLTÜ MODELLEME VE İYİLEŞTİRME

Elektrik hattı üzerinde genişbant haberleşme, özellikle Akıllı Şebeke'nin gündeme gelmesi ile birlikte daha fazla önem kazanmıştır. Buna rağmen elektrik hattındaki elektriksel gürültü ve kanal parametreleri genişbant haberleşme için henüz yeterince karakterize edilmemiştir.

Kanalın karakterizasyonu için, elektrik hattındaki çok yollu sönümlenme ve frekansa bağlı zayıflama parametreleri modellenmiş ve deneylerle doğrulanmıştır. Kanal parametrelerinin modellenmesi ile birlikte matris tabanlı bir kanal simulatörü geliştirilmiş ve ölçümlerle doğrulanmıştır. Simulatör kullanılarak kanalın dağılım parametreleri ve istatistikleri üzerinde bir analiz yapılmış ve sonuçlar sunulmuştur.

Tez çalışmasının birinci aşamasında elektrik hattında çalışan ev aletlerinin ürettikleri gürültüler ve arka plan gürültüsü çalışılmıştır. Genişbantlı elektriksel gürültüyü frekans bölgesinde modellemek için genelleştirilmiş Gauss dağılımı (GGD) önerilmiş ve önerilen dağılım ölçümlerle doğrulanmıştır. Önerilen modele uygun yeni bir genişbant alıcı yapısı önerilmiş ve önerilen alıcı yapısının geleneksel alıcı performansına göre dikkat çekici bir iyileştirme sağladığı gösterilmiştir.

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LIST OF SYMBOLS

A	impulse index for Middleton Class A probability density function
$A_k(t)$	amplitude of narrowband noise
$A(f, d_i)$	signal attenuation related to the length and frequency
A_η	amplitude of a square pulse
c_0	speed of light
$c(n)$	coded bits
C	capacitance
CB	matrix that shows the connections among the branching points
CM	matrix that shows the interconnection between nodes
CT	matrix that shows the interconnection between termination points
$d(n)$	information bits
$D(p \parallel q)$	Kullback-Leibler distance between two probability mass functions
$E\{\cdot\}$	statistical expectation operator
f_k	frequency of narrowband noise
f_s	sampling frequency
f_{RX}	receiver bandwidth
$f(Y \Pi_1, H)$	conditional probability density function
G	conductance
$h(n)$	discrete channel impulse response
H_{mod}	transfer function in the z-plane
$H(f)$	frequency response function
$H(p)$	entropy of X with the assigned probability measure $p(x)$
$I(z)$	current on the line
L	inductance
L_c	length of the power-line channel

LB	length matrix of the cables between branching points
LM	length matrix of the cables between nodes
LT	length matrix of the cables between termination points
M	number of signal blocks withing a truncated signal
n_{bg}	background noise
n_{Cbg}	colored background noise
n_{nb}	narrowband noise
N_0	noise power
N_{FFT}	FFT size
$N_{MK}(k, m)$	discrete data samples falling into the k^{th} piece within the m^{th} AC cycle
$p(x)$	probability mass function of random variable X
P	path matrix
P_e	peak excursion
$q(x)$	probability mass function of random variable X
$r(t)$	received signal
R	resistance
R_c	rate of change
$R(\cdot)$	autocorrelation function
$s(n)$	transmitted symbols
$s(t)$	message signal
$S(k)$	discrete fourier transform of the transmitted symbols
$S(t, f)$	power spectral density
v	propagating velocity within the cable
$V(z)$	voltage on the line
$w(n)$	non-white PLC noise
Z_0	characteristic impedance
Z_L	load impedance
Z_{sc}	short-circuit impedance
Z_{oc}	open-circuit impedance
$\mathcal{N}$	normal distribution
$\mathbf{1}$	standard indicator function

α	width of the probability density function peak (variance)
β	shape parameter of Generalized Gaussian Distribution
$\delta(t)$	dirac-delta function
ε_r	dielectric constant
Γ	gamma function
μ	statistical mean
φ_k	phase of narrowband noise
$\Pi_{1,2}$	hypothesis

LIST OF ACRONYMS/ABBREVIATIONS

AC	Alternating Current
AM	Amplitude Modulation
AMI	Automated Metering Infrastructure
AWGN	Additive White Gaussian Noise
AR	Autoregressive
BER	Bit Error Rate
BPSK	Binary Phase Shift Keying
CDF	Cumulative Distribution Function
CFO	Carrier Frequency Offset
CRT	Cathode Ray Tube
DC	Direct Current
DD	Decision-Directed
DFT	Discrete Fourier Transform
DUT	Device Under Test
FEC	Forward Error Correction
FFT	Fast Fourier Transform
FM	Frequency Modulation
GGD	Generalized Gaussian Distribution
HF	High Frequency
IDFT	Inverse Discrete Fourier Transform
KL	Kullback-Leibler
LCD	Liquid Crystal Display
LISN	Line Impedance Stabilization Network
LDPC	Low Density Parity Check
LOS	Line-Of-Sight
LV	Low Voltage
MA	Moving Average
ML	Maximum Likelihood
NRMSE	Normalized Root Mean Square Error

OFDM	Orthogonal Frequency Division Multiplexing
PDF	Probability Density Function
PLC	Power Line Communication
PLN	Power Line Network
PSD	Power Spectral Density
RF	Radio Frequency
RMS	Root Mean Square
RV	Random Variable
SNR	Signal-to-Noise Ratio
TFA	Time-Frequency Analysis
ULAP	Unit Length Attenuation Profile
VNA	Vector Network Analyzer
VSA	Vector Signal Analyzer

1. INTRODUCTION

1.1. Motivation

Similar to conventional communication channels, power line communication (PLC) medium is affected by noise, path loss, and multipath [1]. The level of interference in power-line networks is much larger than any other type of communication medium. The frequency dependent attenuation characteristics of power-lines and multipath stemming from impedance mismatches are the other factors which have to be dealt with in order to establish a reliable high-speed PLC system.

Although narrowband characterization of noise and multipath channel in PLC has been studied to a great extent, broadband noise and channel characterization received much less attention until recently. In the scope of the thesis,

- a reliable channel simulation environment is developed for the following reasons. Multi-path fading and dispersion is not previously studied to a sufficient extent to determine the coherence bandwidth of a broadband power line channel. A reliable power channel simulator is needed to analyze the impact of the physical characteristics of the powerline networks on channel parameters such as RMS delay spread or maximum excess delay.
- frequency dependent attenuation is subject to the type of cables and the quality of the wiring of the power-line networks. Path loss in power-line networks is studied to model the unit length attenuation so that this model could be used in channel simulations of varying length and network topologies.
- broadband PLC noise characterization is studied in order to design a broadband receiver that compensates for the impairments caused by the non-Gaussian power-line noise.

1.2. Contributions

Contributions of the thesis study can be unified in two major themes, one is the broadband channel characterization and the other is the broadband noise characterization and compensation. In this sense, major contributions of the thesis can be listed as the following,

- *Creating a broadband power-line channel simulation environment:* Multi-path and attenuation characteristics of PLC channels are investigated and a matrix based PLC multipath simulation environment is developed. A unique path selection algorithm is proposed. Channel transfer functions of several simple networks are analyzed with relevant discussions. Effect of
 - (i) direct line length between transmitter and receiver,
 - (ii) length of the branches connected to the power line network (PLN),
 - (iii) number of the branches connected to a single node along the direct line,
 - (iv) impedance values seen at termination nodes,
 are investigated. Important relations between the above mentioned physical attributes and the channel transfer functions are drawn based upon the simulation results. The power-line channel simulation results are compared with channel frequency response measurements and it was shown that the simulations are in well accordance with frequency response measurements.
- *Studying the channel statistics:* Besides the investigation of some specific power line networks (PLN), impact of the physical characteristics of PLC network on the channel delay spread parameters, namely root-mean-squared (RMS) delay spread and maximum excess delay, are studied statistically. Effect of
 - (i) number of branching nodes between transmitter and receiver
 - (ii) separation distance between transmitter and receiver
 - (iii) length statistics of branches
 are analyzed. For each attribute examined, statistics regarding the channel delay spread parameters are presented by observing corresponding cumulative distribution function (CDF) curves. Relations between the attributes listed above and the channel delay parameters are revealed.

- *Non-Gaussian noise modeling over the broadband power-line channel:* For the sake of reliability of experimental results, novel and reliable measurement setups are established. The effectiveness of the measurement setups are shown. Based on the broadband noise measurements, a suitable noise model for the amplitude distribution of each individual frequency is proposed. Parameters of this distribution are studied and it is shown how the parameters change across the spectrum. The proposed model is one of the prominent novelties of this thesis study.
- *Transmission over non-Gaussian PLC channels:* Using the proposed noise PDF earlier in the noise characterization study, a novel soft-demodulation based OFDM receiver is proposed. It is seen that the proposed receiver significantly outperforms the conventional receivers performance in terms of BER for fixed a SNR in a multipath fading PLC channel.

1.3. Organization of the Thesis

Chapter 2 focuses on multi-path fading and dispersion analysis of broadband power-line channels. In order to emulate the wiring of power line network, a PLC laboratory is established, a controlled environment is created where the power line is isolated and filtered from the laboratory's power line, and an apartment's power network is wired with all the junctions and outlets. Measurements have been conducted in the laboratory in order to determine a multipath fading and attenuation model.

Attenuation of channel medium has been studied and a simple yet useful attenuation model is proposed. Multipath fading has also been studied to a great extend. Various multipath models proposed in the literature have been studied and combining the best approaches, a reliable simulation environment has been developed. Simulation results showed great correlation with measurement results.

In Chapter 3, as far as the noise characterization is concerned, power-line background noise and each individual noise generated by home appliances are studied. Various measurements have been conducted with home appliances in an anechoic chamber with a proper measurement setup and cyclo-stationary properties of appliance noise are

studied to some extent. Peak excursion and maximum rate of change of instantaneous PSD for each home appliance are investigated. Based on the analysis of broadband power-line noise in frequency domain, a suitable noise amplitude distribution, generalized Gaussian distribution (GGD), model for each individual frequency is proposed.

In Chapter 4, a novel receiver using the broadband noise model developed in Chapter 3 is proposed. An OFDM receiver framework exploiting the proposed noise model's beta shape parameter is proposed and its performance is compared to conventional approaches.

Finally Chapter 5, provides a brief summary of the thesis and the major outcomes of the work done. We also suggest ways to extend and build upon the work presented in this thesis.

2. MULTIPATH FADING AND DISPERSION

In contrast to conventional communication systems, power line networks (PLN) is not designed for communications purposes, instead it was originally designed for power transmission. Hence, instead of following a point-to-point topology, PLN exhibits a random bus topology. This type of a topology introduces severe drawbacks compared to conventional wired communication networks and channel characterization of such a network should therefore be studied very carefully.

On the other hand, PLN is stable and immovable. Analyzing the multipath components, one can analytically compute the frequency response of any point-to-point channel to some degree of correctness. These multipath components are directly affected by the topology and its physical characteristics such as characteristic impedances, number of branches, type of loads connected to the network, length of cables and so on.

In this section, channel characteristics of PLN are analyzed based on the commonly known channel model of Zimmermann and Dostert [2]. As part of the channel model, attenuation in Turkish PLN is also analyzed. Based on this analytical model, a simulation environment is developed and correctness of the simulation is verified by measurement. The simulation environment is used later to test the performance of a novel receiver proposed in the scope of this thesis study.

2.1. Multipath in Power-Line Network

Different from the conventional wired communication systems, many types of different appliances and devices are connected to the PLN to get powered up. These electrical loads cause impedance mismatches at termination points. In addition to these termination points, there are impedance mismatches at branching points as well. All these impedance mismatches cause reflection and transmission of a communication

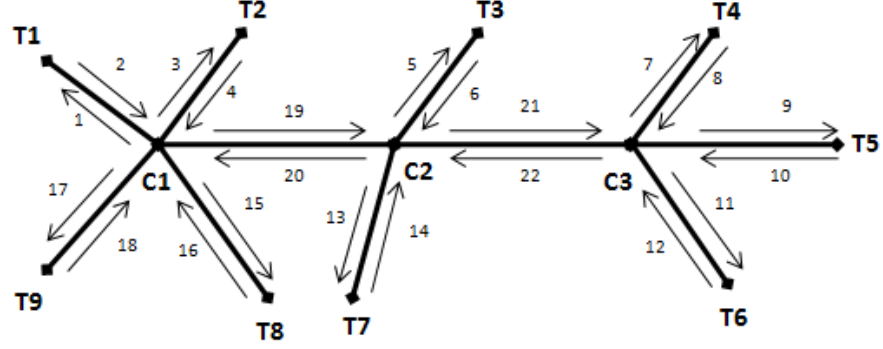


Figure 2.1. An illustration of the occurrence of transmissions and reflections on a PLN.

signal in the power network. Therefore, signal propagation does not only take place through the closest path between a pair of a transmitter and a receiver.

In Figure 2.1, an illustration of the occurrence of transmissions and reflections over a power line network is depicted. As it is seen in the figure, each branching point and termination point on the PLN causes a reflection and a transmission that are in opposite directions. Hence, each multipath segment is numbered because there could be multiple reflections and transmissions on each segment. Hence, the number of possible segments becomes twice as much of the number of lines.

As seen in Figure 2.1, the network consists of 12 nodes, 11 lines, and 22 possible directional segments. For instance, if the transmitter is located at the termination point named as T_1 and the receiver is located at T_7 , various propagation paths can be defined by using numbered directional segments such as

$$2 \rightarrow 19 \rightarrow 13 \quad (\text{which is the closest path})$$

$$2 \rightarrow 19 \rightarrow 20 \rightarrow 19 \rightarrow 13$$

$$2 \rightarrow 1 \rightarrow 2 \rightarrow 19 \rightarrow 13$$

$$2 \rightarrow 17 \rightarrow 18 \rightarrow 19 \rightarrow 13$$

$$\vdots$$

If there is no attenuation or destructive summation of the signals travelling in the network, the propagation would always continue. Anyhow, since there is frequency and distance dependent attenuation and transmissions and reflections cause destructive summation at some nodes, the power of the multipath components will die out as the reflections of the signal travel in the network. Therefore, the number of significant multipath components arriving at the receiver will be limited.

2.2. Signal Propagation over Power-Line Networks

Reflections and transmissions occur at each impedance discontinuity along the propagation path as a signal propagates from one node to another. These discontinuities are due to different characteristic impedance of various cables wired in the same network, electrical loads connected to the power-line network and existence of branching points where there are more than two cables connected to the branching point.

Computing the affect of these reflections and transmissions requires the computation of the reflection and transmission coefficients at each node. Although reflection and transmission coefficients are dependent to the input impedance of the electrical loads and the characteristic impedance of cables, these coefficients are computed as the amplitude and phase ratio of reflected and transmitted energy to the incident signal.

It is commonly known that characteristic impedance of a cable does not depend on the length of the cable. Yet, it depends on the cable's equivalent circuit coefficients and the operating frequency. Hence, Z_0 is calculated as,

$$Z_0 = \sqrt{\frac{R + j\omega L}{G + j\omega C}}. \quad (2.1)$$

where Z_0 represents the characteristic impedance of the line the signal propagates, ω is the angular frequency, and R , L , G , and C represent the per-unit-length resistance,

inductance, conductance, and capacitance values, respectively. As it is shown in [2, 3], equations boil down to $R \ll wL$ and $G \ll wC$ for the frequencies of interest in PLC. The expression for Z_0 can be simplified to

$$Z_0 \cong \sqrt{\frac{L}{C}}. \quad (2.2)$$

In addition, characteristics impedance Z_0 and the impedance of the electrical loads plugged into the termination nodes Z_L are assumed to be real valued and independent from the operating frequency [2].

$$Z_0(f) = Z_0, \quad \text{and} \quad Z_L(f) = Z_L. \quad \text{for} \quad 30 \text{ KHz} \leq f \leq 50 \text{ MHz} \quad (2.3)$$

Reflection and transmission coefficients are derived from the impedance values, frequency-independent and given by the following equations [4]:

$$\Gamma = \frac{Z_L - Z_0}{Z_L + Z_0}, \quad \text{and} \quad T = 1 + \Gamma = \frac{2Z_L}{Z_L + Z_0}. \quad (2.4)$$

where Z_L is the impedance at the discontinuity.

A different way to calculate characteristic impedance of a cable is by measuring the input impedances of the cable with short and open-circuit termination endings [5]. The square-root of the product of these two input impedances gives the characteristic impedance

$$Z_0 = \sqrt{Z_{sc}Z_{oc}}. \quad (2.5)$$

where Z_{sc} and Z_{oc} represent input impedance values of the cable with short and open-circuit endings, respectively.

2.2.1. Reflections and Transmissions in the Channel Model

The amount of loss due to impedance mismatches along the i th path is factored into the channel model as the multiplication of all the reflection and the transmission coefficients the signal sees during its propagation from one node to another. Reflection factor, denoted as $r_i e^{j\theta_i}$, is usually but not necessarily a complex number. Reflection factor of i th path can be given as;

$$r_i e^{j\theta_i} = \prod_{k=1}^K \{\Gamma_{ik}\} \prod_{m=1}^M \{T_{im}\}. \quad (2.6)$$

where K and M are the total reflections and transmissions seen by the transmit signal along a particular propagation path, respectively.

On the other hand, multiple replicas of the transmitted signal arrives to the receiving point. Anyhow, since these replicas travel different paths, denoted as d_i , time delays occur among the replicas and this causes a phase shift at the receiving side. These time delays depend on the velocity of the propagation and the length of the propagation. The time delay of the i th path can be expressed as;

$$\tau_i = \frac{d_i}{v}, \quad \text{where} \quad v = \frac{c_0}{\sqrt{\varepsilon_r}}. \quad (2.7)$$

where v represents the propagating velocity within the cable which depends on the speed of light c_0 and the relative dielectric constant ε_r of the insulating materials of the cable. Note that the speed of the propagation within the power line is assumed to be 60 of the speed of light in this study [6].

2.3. Attenuation Factor in the Channel Model

Although attenuation in PLN is studied extensively in the power line communication society and some rigorous models are introduced in [7, 8], it still suppose to be studied in each country because type of cables used for wiring the power line may differ from country to country. In this sense, attenuation is one of the most challenging issues of the power-line communications since the attenuation increases with distance and frequency. A mathematical model is given in [2], in which the attenuation parameters are calculated individually for each multi-path component.

To model the attenuation of the cable type used in Turkey, several measurements are conducted for different lengths of cables within a particular frequency range. The dependency on length and frequency is extensively investigated. The cross-section of the measured cable is illustrated in Figure 2.2. Measurement campaigns were performed by using Agilent vector network analyzer (VNA). and in order to prevent the reflections, impedance matching circuits are connected to the input and output of the VNA. The measured frequency range was from 30 KHz to 30 MHz and cables with length of 10 m, 15 m, 25 m, 40 m, 50 m, 65 m, 75 m, 90 m, 100 m, and 200 m were examined.

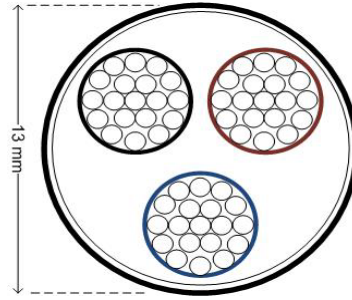


Figure 2.2. Multi-core cable with copper conductor.

To obtain a unit length attenuation profile, different length of the same type of cables are considered. For a fixed frequency, attenuation of different length cables are measured. For each fixed frequency, the attenuation values were fitted via least square fitting. Consequently, a fixed attenuation value for a unit length is obtained for each frequency, namely unit length attenuation profile (ULAP) which basically reveals the attenuation of the cable per meter. Attenuation for i th path having a length of d_i

meters can be calculated as;

$$A(f, d_i) = d_i [20 \log_{10}(\text{ULAP})]. \quad (2.8)$$

where $A(f, d_i)$ is the signal attenuation related to the length and frequency. For instance, the comparison of the measured attenuation of a cable with 100 m and its estimate obtained from Equation 2.8 is plotted in Figure 2.3. As can be seen, the estimate fits well with the measured results.

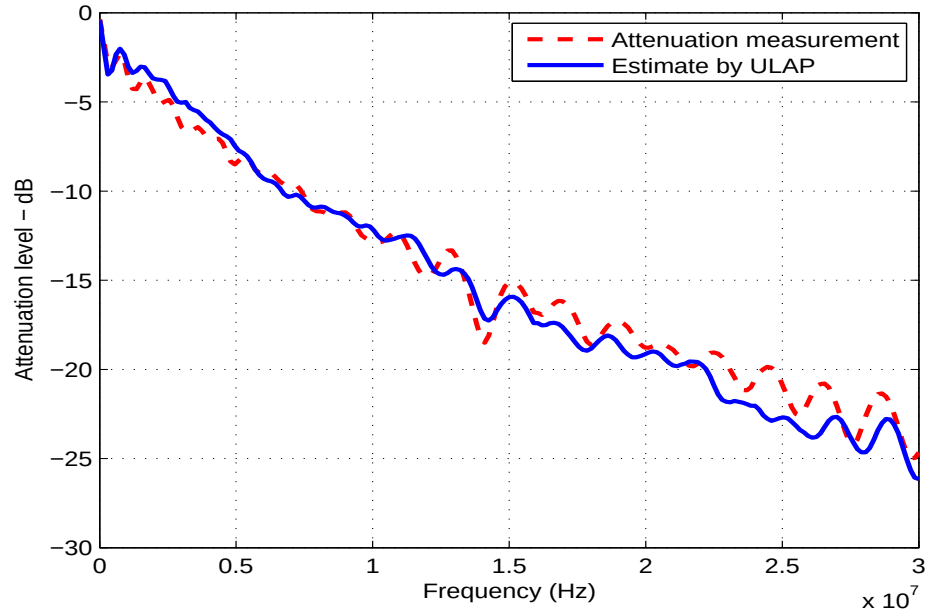


Figure 2.3. Attenuation profile of 100m cable.

As a conclusion, when a signal is transmitted over a PLN, the received signal consists of attenuated, delayed, and phase-shifted version of the transmitted signal. If the total number of received multi-path signals is limited to N , an expression for the frequency response of the channel can be characterized as follows,

$$H(f) = \sum_{i=1}^N \left\{ \left[\prod_{k=1}^K \Gamma_{ik} \prod_{m=1}^M T_{im} \right] A(f, d_i) e^{-j2\pi f \tau_i} \right\}. \quad (2.9)$$

where $A(f, d_i)$ means the frequency and distance dependent attenuation which will be discussed in the next section.

2.4. Multipath Channel Simulation

Evaluation of any proposed transmission method requires the simulation of the channel the signal propagates through. Based on the proposed noise model, a novel receiver structure is part of this thesis study and will be proposed in the next chapter. Characterization of the PLC channel transfer function have been studied and various approaches have been proposed in [8], [2], [9, 10]. In many of the previous studies, channel parameters have to be analyzed individually and some did not even consider the multi-path propagation. In [11], a matrix based algorithm is presented and all of the factors of the Zimmermann's analytical channel model have been taken into account. Using the proposed method in [11], PLNs is successfully simulated. However, a drawback of this method is that the algorithm is not suitable for every type of PLNs, it works for those which have radial topology where no loops exist in the PLN. The algorithm considers the PLN topology as a group of matrices where the connection points, reflection and transmission coefficients, termination impedances, and lengths are registered into matrices. Every connection between the nodes and the physical characteristics of the interconnections are represented as set of matrices. Each reflection and transmission coefficient of the network is calculated by computing the input impedances and applied to each multipath component to estimate their reflection coefficients.

2.5. Registering the Network Topology Information

First step in the algorithm is to register all the information available regarding the network into matrices. All termination points, t_i , branching points, b_i , connections between these points are entered into connection matrix, CM . Entering in this information, some assumptions are made to cope up with the PLN topologies in reality;

- i. termination points are connected to a single point,

- ii. there is not any direct connection between the termination points.

$$CM_{m \times m} = \begin{matrix} & \begin{matrix} t_1 & \dots & t_h & b_1 & \dots & b_k \end{matrix} \\ \begin{matrix} t_1 \\ \vdots \\ t_h \\ b_1 \\ \vdots \\ b_k \end{matrix} & \begin{pmatrix} 0 & \dots & 0 & c_{1,h+1} & \dots & c_{1m} \\ \vdots & \ddots & \vdots & \vdots & \ddots & \vdots \\ 0 & \dots & 0 & c_{h,h+1} & \dots & c_{hm} \\ c_{h+1,1} & \dots & c_{h+1,h} & c_{h+1,h+1} & \dots & c_{h+1,m} \\ \vdots & \ddots & \vdots & \vdots & \ddots & \vdots \\ c_{m1} & \dots & c_{mh} & c_{m,h+1} & \dots & c_{mm} \end{pmatrix} \end{matrix} \quad (2.10)$$

where h represents the total number of termination points, k represents the total number of branching points, and m represents the total number of nodes. Here, m equals to the summation of the total number of branching points and termination points. Every element of the connection matrix $CM[m \times m]$ represents an interconnection between two nodes. Since there is no connection between termination points and every termination point has to be connected to a single branching node, the first $[h \times h]$ elements of CM matrix is set to 0. On the other hand, when there is an interconnection between two nodes, c_{ij} is set to 1. Therefore, the CM matrix exhibits symmetry with respect to its diagonal, which can be expressed as,

$$c_{ij} = c_{ji}. \quad (2.11)$$

If the interconnections between the termination points are represented by the sub-matrix CT and the interconnections between the branching points are represented by the sub-matrix CB , then, CM matrix can also be shown as;

$$CM_{[(h+k) \times (h+k)]} = \begin{pmatrix} 0_{[h \times h]} & CT_{[h \times k]} \\ CT_{[k \times h]}^T & CB_{[k \times k]} \end{pmatrix} \quad (2.12)$$

where $(\cdot)^T$ represents the transpose of the input matrix.

Replacing the corresponding ones with their cable lengths in CM , we obtain the length matrix, LM as the following,

$$LM_{[(h+k) \times (h+k)]} = \begin{pmatrix} 0_{[h \times h]} & LT_{[h \times k]} \\ LT_{[k \times h]}^T & LB_{[k \times k]} \end{pmatrix} \quad (2.13)$$

where the length matrix of the cables between termination points is denoted as LT and the length matrix of the cables between branching points is denoted as LB .

2.6. Path Selection Algorithm

As part of the channel simulation, one needs to determine the possible paths between a transmitter and a receiver plugged into the PLN. Since the transmitted signal is partially reflected and transmitted at each node (branching and termination point), the signal travels through multiple paths before it is received at the receiving point. A new matrix P then is formed to represent the possible paths. Creation of P depends on the connection matrix CM that describes the interconnections. Considering the dimension of the path matrix and connection matrix, matrix P can be considered as the summary of matrix CM in which interconnections between corresponding nodes are explicitly indicated.

Since it is possible to find infinitely number of different multipath components even for a small PLN, choosing the significant paths can be a complicated task. It may even require to delve into graph theory to optimize the selection in a large network. Since developing such a complicated path selection algorithm is out of the scope of this thesis, a simple random path selector is developed. The random path selector determines the propagating path for each replica of the signal starting from the transmitter until it reaches the receiver. After the random paths are generated, they are

put in order to build a sequence. Each of the sequence from the transmitter to the receives makes up a multipath component that is to be considered in the derivation of Equation 2.9. The length of these node sequences is calculated by considering the interconnection lengths among nodes of interest. Besides, after defining total length of the propagation path, attenuation for each path is computed using Equation 2.8. Significant paths are chosen with respect to a power threshold. Until L multipath components are chosen, the process goes on the same way. The power comparison is made against the first received multipath component which is the most powerful one among all the other replicas. The threshold value is set to 30dB;

$$10\log_{10} \frac{|h_0|^2}{|h_i|^2} \geq 30 \quad (2.14)$$

where h_0 and h_i corresponds to the first and the i th paths, respectively, and $|\cdot|^2$ represents the power magnitude.

In order to find the significant multipath components, one needs to calculate the reflection and transmission coefficients using the Equation 2.4, respectively, noting that the impedance at a termination point Z_{d_i} is equal to the impedance of the load Z_L connected to the termination point and can be computed as it is described in Equation 2.4.

$$\Gamma_i = \frac{Z_{d_i} - Z_0}{Z_{d_i} + Z_0} \text{ and } T_i = \Gamma_i + 1 = \frac{2Z_{d_i}}{Z_{d_i} + Z_0} \quad (2.15)$$

On the other hand, in a homogeneous network, the impedance seen by the incident signal arriving at a branching point can be computed by the following expression:

$$Z_L = \frac{Z_0}{n_i - 1} \quad (2.16)$$

where n_i is the total number of branches extending from a node. Using Equation 2.16, reflection and transmission coefficients for the respective branching nodes can be calculated as:

$$\Gamma_i = \frac{2 - n_i}{n_i} \text{ and } T_i = \Gamma_i + 1 = \frac{2}{n_i} \quad (2.17)$$

Therefore, the matrix R_Γ , representing the reflection coefficients for each node, can be defined as:

$$R = \begin{pmatrix} t_1 & \dots & t_h & b_1 & \dots & b_k \\ \Gamma_1 & \dots & \Gamma_h & \Gamma_{h+1} & \dots & \Gamma_m \end{pmatrix} \quad (2.18)$$

and the transmission coefficient matrix T_T is defined as:

$$T = \begin{pmatrix} t_1 & \dots & t_h & b_1 & \dots & b_k \\ T_1 & \dots & T_h & T_{h+1} & \dots & T_m \end{pmatrix} \quad (2.19)$$

In these calculations, the characteristic and load impedances are assumed to be constant for all frequencies although they are not in reality. Both these impedances depend on the operating frequency in reality. Anyhow, to simplify the calculations they can be considered constant.

Creating all necessary matrices and computing the impedances and multipath components allow us the calculation of the channel transfer function $H(f)$ given by Equation 2.9. By applying the inverse fast Fourier transform (IFFT) operation on channel frequency response, channel impulse response is obtained as follows,

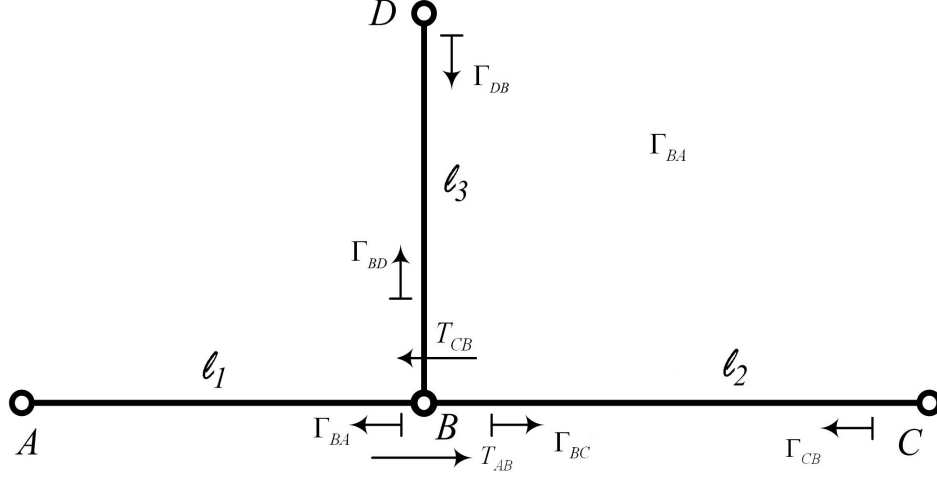


Figure 2.4. An illustration of a simple T network.

$$h(\tau) = \sum_{i=0}^N h_i \delta(t - \tau_i). \quad (2.20)$$

where τ_i denotes the delay of the i th arriving path.

2.7. Verifying the Correctness of the Channel Simulation

To make computations and analysis simpler, we start off the analysis with a simple T network, which is illustrated in Figure 2.4. We assume the transmitter is plugged into the termination A , the receiver is plugged into the termination C and the termination point D is left open. Transmitter and receiver is assumed to be matched to the characteristic impedances of the cables and it is assumed that the network is homogeneous (consisting of one type of cable) and the characteristic impedance of this cable is 100Ω . The load impedance Z_L was represented as $10^8\Omega$ in the simulations. Lengths of the cables l_1 , l_2 , and l_3 are decided to be 15m, 15m, and 10m, respectively. The T network is established and its frequency response is measured using a Vector Network Analyzer to verify the correctness of our simulations. Figure 2.5 and Figure 2.6 show that our simulations are in well accordance with the real measurements.

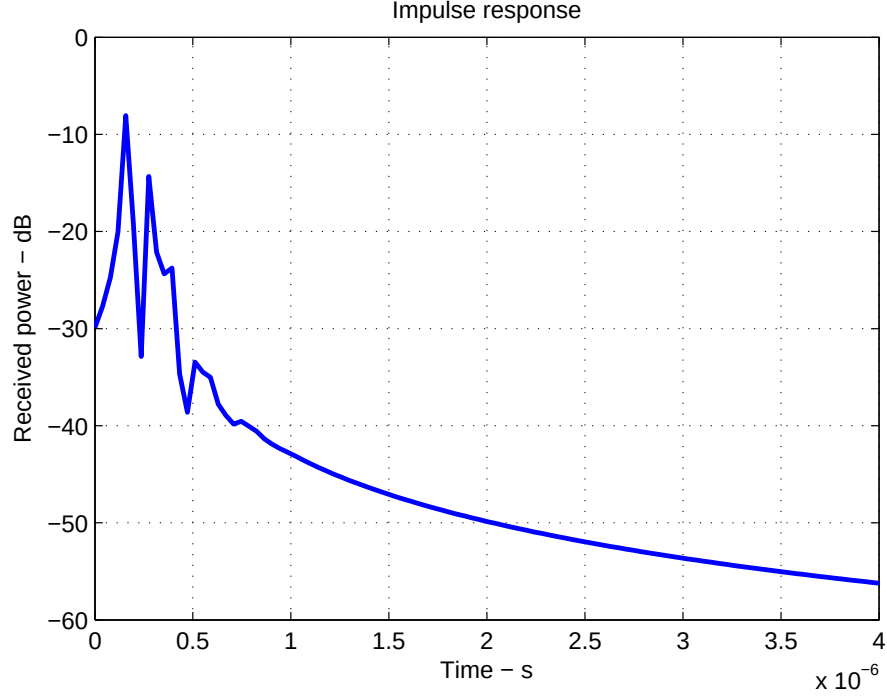


Figure 2.5. Impulse response of the channel between A and C.

2.8. Statistical Analysis of Arbitrary Network Topology

Instead of analyzing specific network topologies for PLC channel characterization as in [2, 3, 11–21], a statistical characterization of arbitrary power network topologies may be more useful. Therefore, without knowing the topology, one might have an idea on how the channel statistics change on average for a specific residential class such as a house, apartment, lab, etc. Impact of the distance between transmitter and receiver, number of branching points and length of each branch on root-mean-squared (RMS) delay spread and maximum excess delay are analyzed in this section.

An arbitrary topology as in Figure 2.7 where nodes denoted by B and T are branching and termination points, respectively, is generated for different sizes to analyze the impact of the physical parameters. The number of branching nodes between transmitter and receiver is denoted by k , number of branches stemming from each i th branching point is denoted by n_i and the length of the cable between node x and y is shown by l_{xy} .

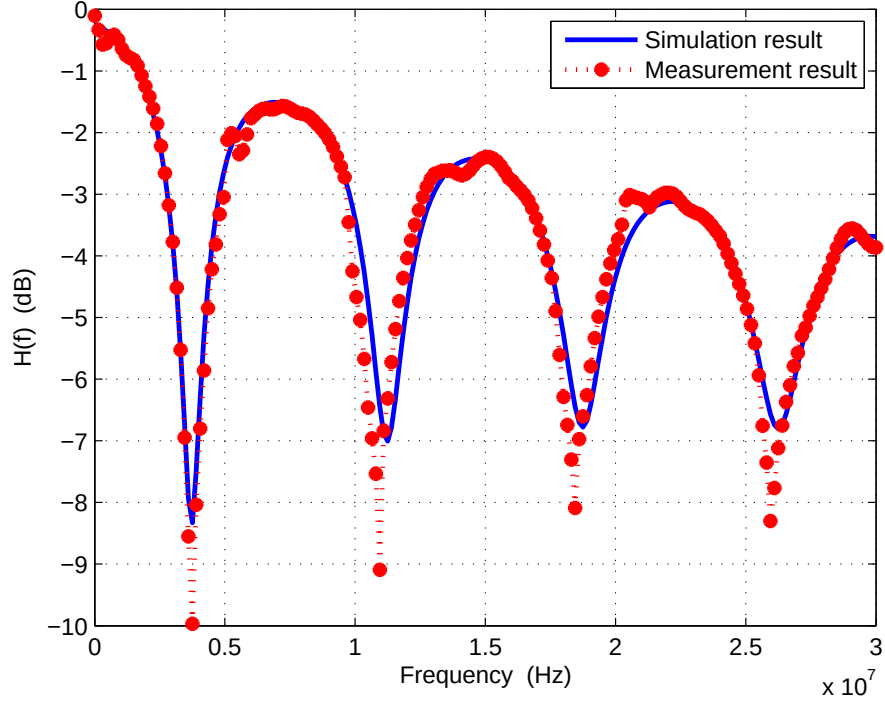


Figure 2.6. Real measurement and simulation of the frequency response of the channel between node A and C.

When the loads at termination points have very low impedance or very high impedance, the delay spread parameters tend towards its highest as analyzed in [12, 13]. To study the worst case scenerios in our analysis, all termination points are assumed to be open circuit [22].

We will be using the channel simulator developed in Section 2.5 with a difference that the simulations are modified to generate random topologies with different physical parameters in a controlled manner. Different topologies are created by manipulating the content of the connection matrix.

$$\mathbf{CM}_{[(h+2+k) \times (h+2+k)]} = \begin{pmatrix} 0_{[(h+2) \times (h+2)]} & CT_{[(h+2) \times k]} \\ CT_{[k \times (h+2)]}^T & CB_{[k \times k]} \end{pmatrix} \quad (2.21)$$

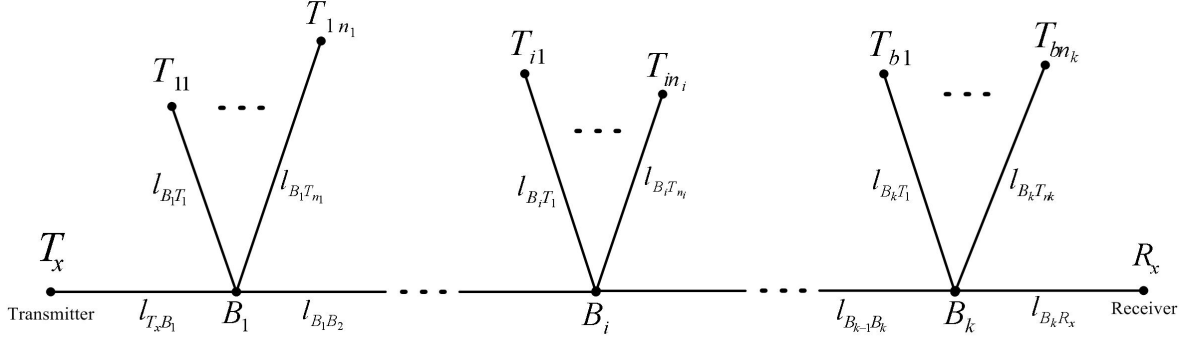


Figure 2.7. Illustration of an arbitrary PLC network topology.

where the interconnections between the termination nodes and branching nodes are denoted by CT , the interconnections among the branching points are denoted by CB and total number of number of branching points and termination points are denoted by k and h , respectively. Number of branches stemming from each branching node is uniformly distributed. Anyhow, this is not the case for the termination points. Size of CT matrix will be changed with the total number of branches h .

$$h = \sum_{i=1}^k n_i. \quad (2.22)$$

As mentioned in the previous subsections, characteristic impedance of the cables is set to 100Ω and transmitter and receiver are also assumed to be matched to the characteristic impedance. Playing with the values of h , k and l_{xy} which correspond to physical parameters of the network, various topologies are generated. Impact of the physical parameters is statistically investigated by generating 20000 realizations of PLC network for each case taken into account.

Please note that, as stated before, each n_i value is uniformly distributed over $[1,4]$. Thus, registration in the CT matrix can be described as

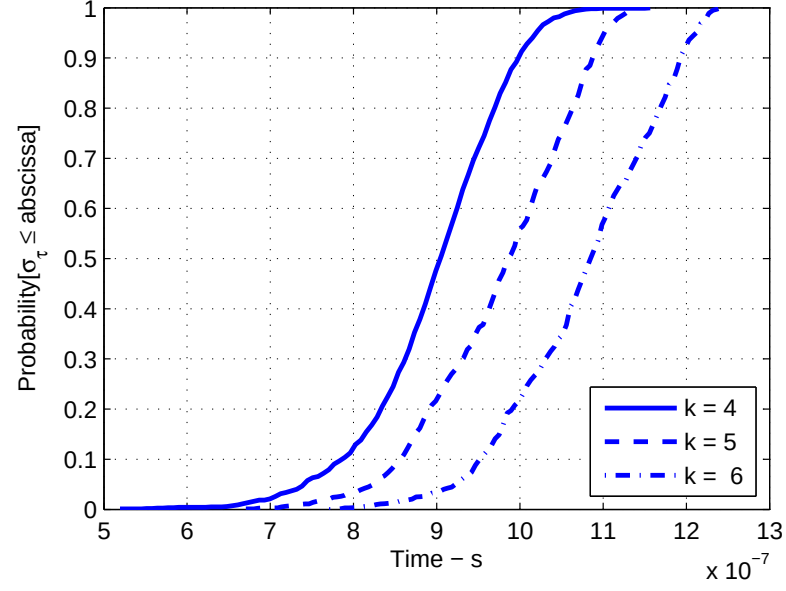
$$CT_{[(h+2) \times k]} = \begin{pmatrix} \boxed{\begin{matrix} 1 & 0 & \dots & 0 \\ 0 & 0 & \dots & 1 \end{matrix}} \\ \boxed{\begin{matrix} 1 & 0 & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 1 & 0 & \dots & 0 \end{matrix}} \\ \boxed{\begin{matrix} 0 & 1 & \dots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 1 & \dots & 0 \end{matrix}} \\ \vdots \\ \boxed{\begin{matrix} 0 & 0 & \dots & 1 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \dots & 1 \end{matrix}} \end{pmatrix} \begin{matrix} \Leftarrow T_x - B_1 \text{ connection} \\ \Leftarrow B_k - R_x \text{ connection} \\ \Leftarrow \text{Branches from } B_1 \\ \Leftarrow \text{Branches from } B_2 \\ \Leftarrow \text{Branches from } B_i \\ \Leftarrow \text{Branches from } B_k \end{matrix} \quad (2.23)$$

Other than the first and the second rows of the matrix CT , all rows of the matrix represent the interconnections between the termination nodes and branching nodes. First and second rows represent the interconnections between the transmitter and branching point B_1 and interconnection between the receiver and the branching point B_k .

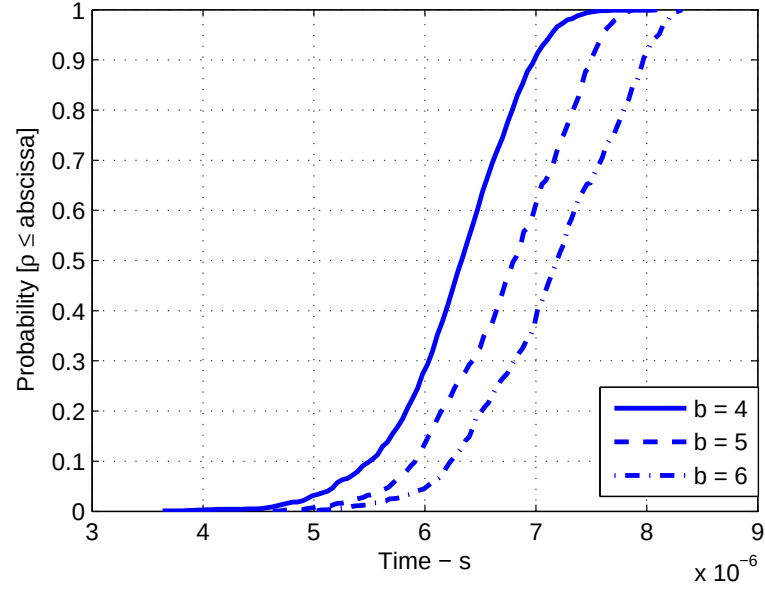
In Figure 2.8a, the impact of the number of nodes between the transmitter and the receiver on the RMS delay spread and maximum excess delay is shown. In the simulation, the distance between the transmitter and the receiver is assumed to be 150m and length statistics of the branches were considered to be uniformly distributed between [10m – 30m]. Looking at the simulation results, it can be seen that an increase in the number of branching points causes an increase in RMS delay spread value. This is due to the fact that as the number of branching points increase, the number of multipath components increase and the propagation time increases. Some components arrive to the receiving point at larger delays as k is increased.

Impact of the distance between transmitter and receiver on RMS delay spread and maximum excess delay is depicted in Figure 2.9a and Figure 2.9b. Note that throughout the simulations, number of branching points are kept constant and considered to be uniformly distributed over $[10\text{m} - 30\text{m}]$. As in the previous case, increasing the distance results in an increase in the delay statistics.

Lastly, in Figure 2.10b, the impact of the branch length statistics are shown. It is seen that out of all the factors analyzed, the impact of the branch length statistic is the most important because it causes the most dramatic change in the delay parameters. It should be noted that throughout the simulations, the number of branching points between transmitter and receiver is set to 4.

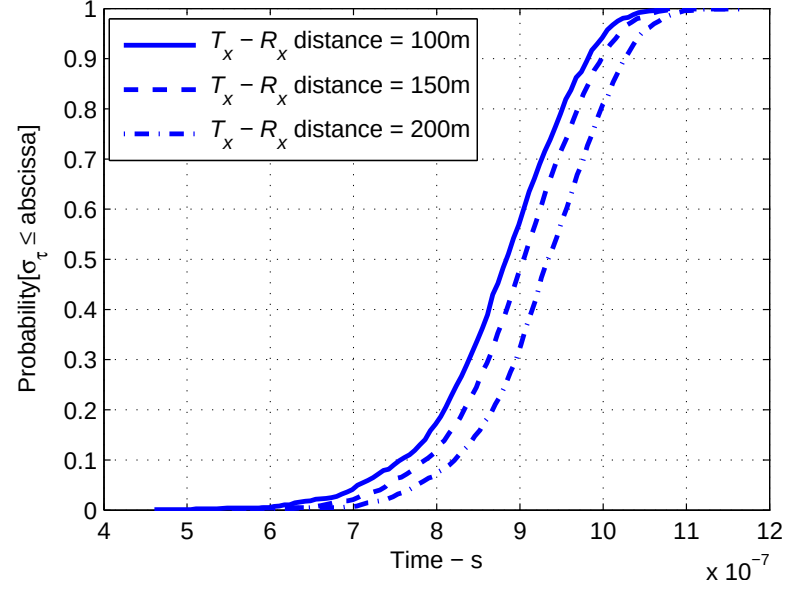


(a) RMS delay spread

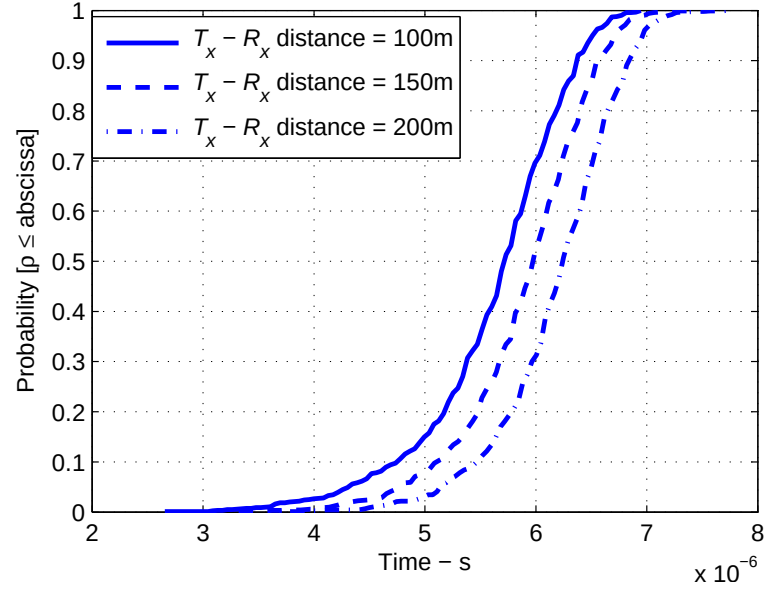


(b) Maximum excess delay

Figure 2.8. Dependency of RMS delay spread (σ_τ) and maximum excess delay (ρ) on the number of branching points k between T_x and R_x .

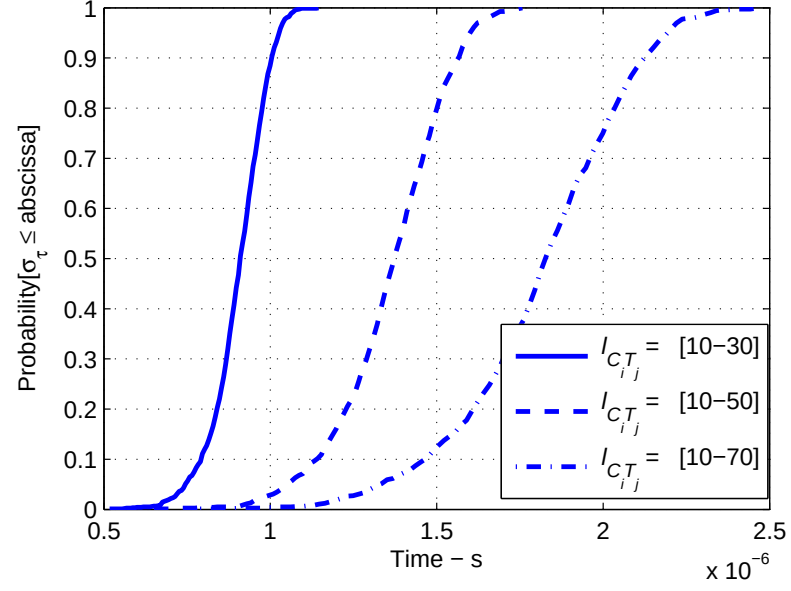


(a) RMS delay spread

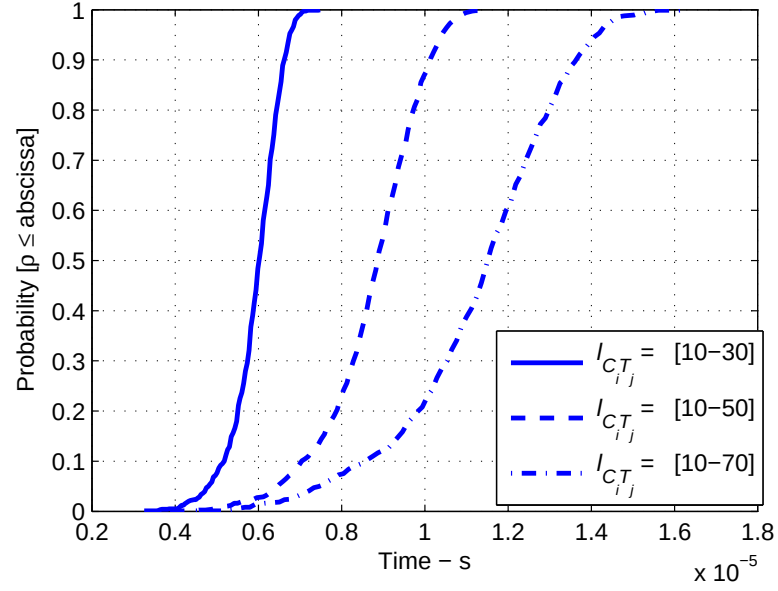


(b) Maximum excess delay

Figure 2.9. Dependency of RMS delay spread (σ_τ) and maximum excess delay (ρ) on the distance between T_x and R_x .



(a) RMS delay spread



(b) Maximum excess delay

Figure 2.10. Dependency of RMS delay spread (σ_τ) and maximum excess delay (ρ) on the length statistics of branches.

3. POWER-LINE NOISE MODELING

In contrast to conventional communication systems, instead of stationary additive white Gaussian, noise is modeled as cyclo-stationary additive non-white non-Gaussian in power-line communication. Noise characteristics of power line networks in various buildings are investigated in [23–26] by performing measurements at different power outlets. Measurement campaigns in these studies have shown that power-line communication channel is impaired mostly by impulsive noise of bursty nature. Therefore, noise in PLC systems is classified into three main categories that are colored background noise, narrow band noise, and impulsive noise as depicted in Figure 3.1.

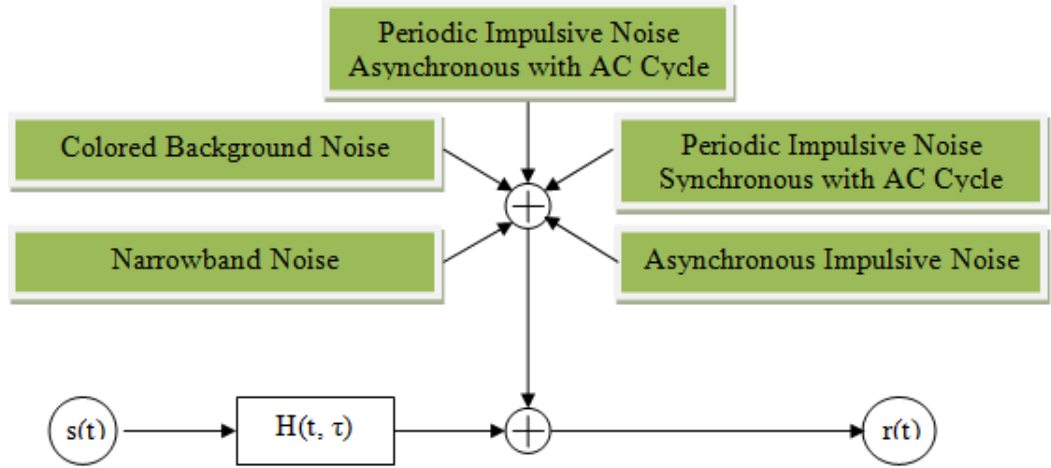


Figure 3.1. Noise types observed in PLC systems.

- *Colored background noise*: the summation of different noise sources of low power present in the network is the cause of background noise, which is usually characterized with a power spectral density (PSD) decreasing with the frequency.
- *Impulsive noise*: electrical appliances generate noise of impulsive nature and this type of noise is the most significant among the above-mentioned noise types. Studying impulsive noise proposes that it can be further classified as:

- (i) *Periodic impulsive noise synchronous with the AC cycle* is caused by the non-linear components such as diodes used in power supplies which operates synchronously with the mains cycle.

- (ii) *Periodic impulsive noise asynchronous with the AC cycle* is generated by pulse width modulators commonly found in switched-mode power supplies and AC/DC power converters.
- (iii) *Aperiodic impulsive noise* is caused by on/off switching transients, electrical motors such as washing machine and drills present in the network.
- *Narrow-band noise*: the cause of this noise is the RF coupling of radio broadcasts in long, middle, and short wave ranges.

Impulsive noise needs special attention because average duration of impulses are in microseconds level which make them highly time varying and they cause most of the errors at data transmission [27–29].

It is worth mentioning that understanding the impulsive noise characteristics of electrical devices individually is also essential from the communication aspect. Recently, as a first step towards decoder design for PLCs where noise memory is decidedly non-negligible [30] considered a two-state Markov process to model the memory of the associated noise. In a power-line context, however, partitioned Markov chains (PMCs), have been shown to provide a better statistical characterization of the noise process [1]. PMCs were initially used by Fritchman [31] to describe the statistical dependence of errors in binary bursty channels. By comparison with experimental measurements, [1] showed that PMCs are excellent means to realistically model the additive noise in PLC as it allows one to account for the bursty nature of impulses. PMCs have been successfully used to characterize memory in communication channels in [31–33]. However, performance analysis of transmission schemes for PLC based on a PMC model has received limited attention. Efforts on these lines were made in [34, 35].

In addition to studying noise in these categories, holistic approaches were also taken to analyze noise as the summation of all these categories. Time-varying colored background noise has been modeled by Katayama *et al.* in [36] and its time-varying and colored features are expressed with only a small number of parameters. Katayama *et al.* also showed that sampling at random timing without synchronization to the AC voltage results in random samples that follow the probability density function of Middleton

Class A [36], which expresses the noise PDF as a sum of Gaussian distributions with different variances in time.

Middleton proposed a canonical model of impulsive noise categorizing the noise distributions as Class A, B and C [37]. The classification is based on the temporal coherence of the noise process whereby for Class A noise the duration of the impulsive noise ($T_{impulse}$) is greater than the inverse of the receiver bandwidth (f_{RX}), i.e.,

$$T_{impulsive} \ll \frac{1}{f_{RX}} \quad (3.1)$$

The opposite is true for Class B noise, and Class C is a mixture of the Class A and Class B noise. Effectively, hence, Class A noise is referred to as *narrowband* interference and Class B as *broadband* interference. Mathematically, the noise distribution is an infinite sum of weighted Gaussians depicted as follows

$$f(x) = \sum_{m=0}^{\infty} \frac{\exp(-A) A^m m!}{e} \exp\left(\frac{-x^2}{2\sigma_m^2}\right) / \sqrt{2\pi\sigma_m^2} \quad (3.2)$$

The parameter A is referred to as the impulsive index and is the parameter that governs the frequency of occurrence of the impulses. A lower value of A ($\rightarrow 0$) indicates structured interference while a high value of A ($\rightarrow \infty$) makes the noise more Gaussian.

The variances σ_m^2 are related to the physical parameters of the noise and are given by

$$\sigma_m^2 = \frac{(m/A) + \gamma}{1 + \gamma} \quad (3.3)$$

where γ is the ratio of power of the Gaussian component to the impulsive component and thus is a measure of the strength of the impulsive noise. The appeal of this model derives from its good fit to various noise and interference measurements in PLC. Interestingly, it has been observed in several studies, foremost of them being [38], that usually the first $M(\leq 3)$ coefficients are sufficient to provide excellent approximations to the actual noise processes and thus the infinite sum in Equation 3.2 can be conveniently truncated to a finite value without loss in estimation quality.

Katayama *et al.* and Meng *et al.* studied the background noise also in frequency domain and showed that the amplitude distribution of each individual frequency is different and that Gaussian noise model is not a good approximation for the whole spectrum in PLC [36, 39]. The power-line noise was also modeled by spectrum fitting in [40, 41]. Anyhow, spectrum fitting only captures the average noise spectrum and does not provide any information regarding the noise amplitude distribution of each individual frequency. The previously proposed probability density functions (PDF) for modeling the noise amplitude distribution for each individual frequency in the literature include Nakagami- m distributions [39], "sum of two Rayleigh" [42], log-normal [43], and Gaussian distributions. As part of the thesis work, the noise amplitude distribution of each frequency is studied and a suitable distribution is proposed. The proposed model is discussed in the following sections.

For appliance noise characterization and modeling noise amplitude in frequency domain, two particular measurement set-ups are established. First set-up is established in order to investigate the individual noise characteristics of various electrical loads. The measurement set-up is designed in a way that it significantly reduces the noise power of other undesired noise sources such as narrow band noise and the noise generated by the other electrical loads within the power line network. The other set-up is established to receive noise as a summation of all noise types to analyze the noise

in a more holistic way. It will be described in more detail how the outcomes of these measurements are used in the course of this thesis.

3.1. Noise Measurement Set-up

3.1.1. Appliance Noise Measurement

Since the focus of this measurement campaign is only to investigate the noise characteristics of the electrical loads, an experimental set-up which is designed to suppress all other types of noise, namely colored background noise and narrow band noise, as much as possible was established as depicted in Figure 3.2.

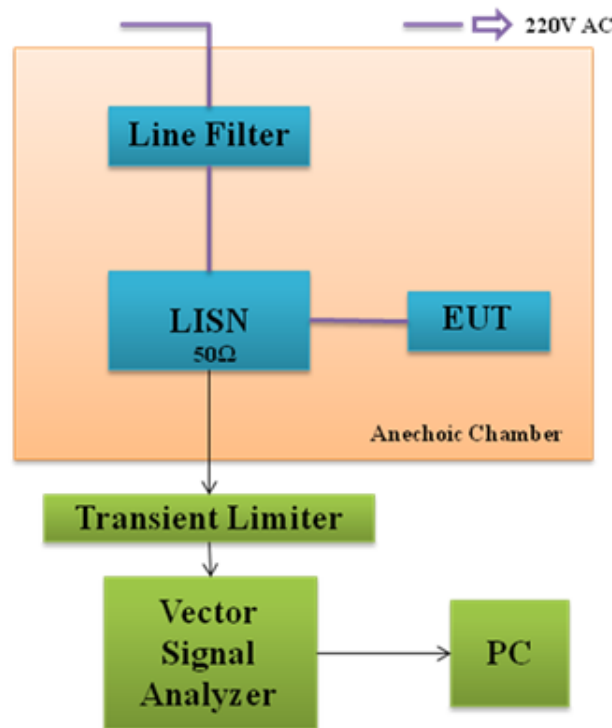


Figure 3.2. Pictorial description of the measurement set-up.

List of the equipment that was used in the experiments is as follows:

1. Anechoic chamber
2. Power-line filter (TGH-16)
3. Line Impedance Stabilization Network (LISN) (963350TS50N)

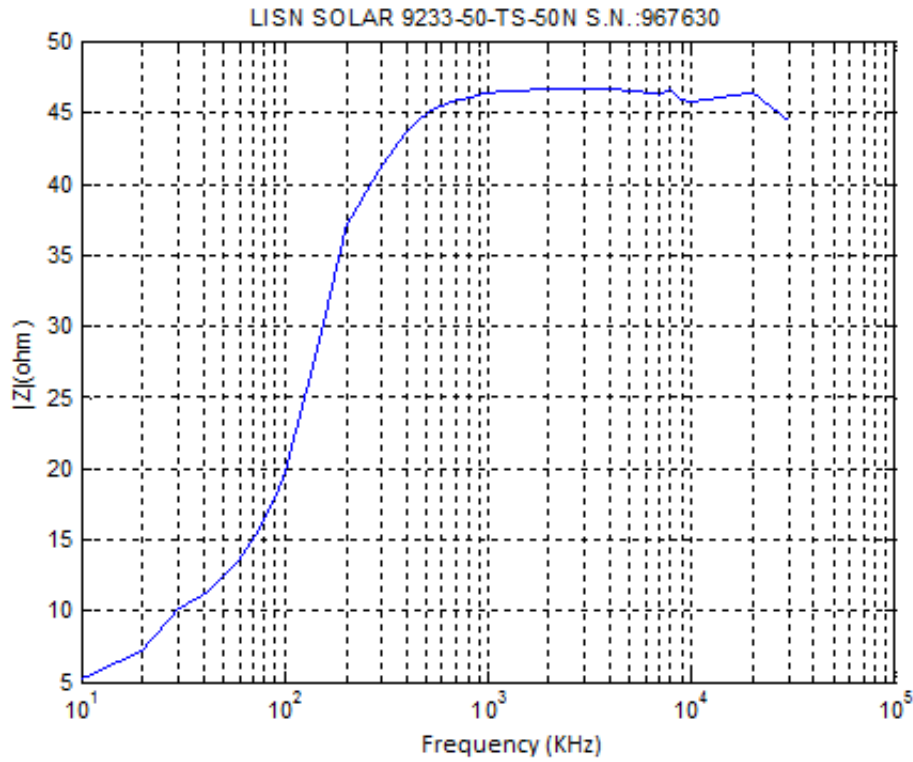


Figure 3.3. Frequency response of the LISN used in the measurements.

4. Line coupler
5. Transient limiter (HP 11947A)
6. Agilent E4440A PSA series spectrum analyzer

Anechoic chamber is a shielded room fitted with Radio Frequency (RF) absorbers in order to form an experimental zone that is completely isolated from the electromagnetic radiation in the outside world. In our experimental set-up, anechoic chamber is used for rejecting narrow band and the colored background noise.

Power line filter suppresses the undesired noise, stemming from electrical loads, which may leak into the measurement set-up from the mains line. The operating frequency of the power line filter ranges from 30 KHz to 1 GHz.

Another component of our measurement set-up is the Line Impedance Stabilization Network (LISN). Since the impedance of the power-line network is varying with mains voltage and this variation causes an impedance mismatch between measurement

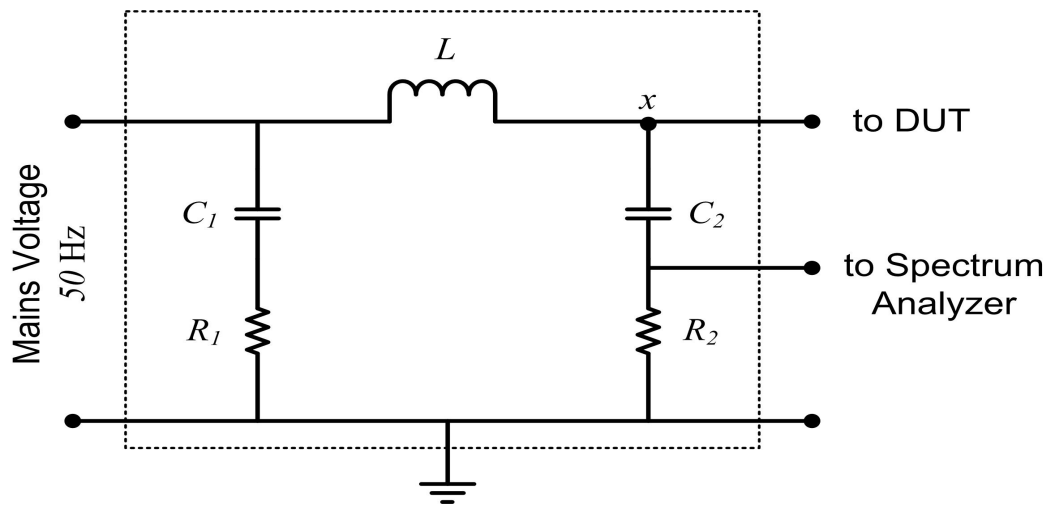


Figure 3.4. Block diagram of LISN.

equipment and the line, this mismatch must be taken care of in order to enhance the accuracy of the measurement results. An LISN is used to compensate this inconvenience.

A simple circuit equivalent of a LISN is shown in Figure 3.4. It is a “Pi-type” filter and will transmit the higher frequency components of the device under test (DUT) while blocking the components belong to low frequencies such as the mains voltage. As a result, only high frequency components of the noise, generated by the DUT, will be captured by the spectrum analyzer with the help of LISN.

Measurement equipment is isolated from the mains voltage via a line coupler circuit. The line coupler is designed as a high pass filter whose frequency response is flat within the frequency range of 30 KHz to 47 MHz. Transient limiter is integrated into the measurement set-up to reject transients which may possibly damage the measurement system hardware. Agilent E4440A PSA series spectrum analyzer was used in order to capture the noise characteristics of DUT. In every measurement, data with a duration of 200ms was recorded. Span of the spectrum analyzer is adjusted in a way that the frequency content of captured signal extends from 30 KHz to 30 MHz.

Throughout the measurement campaigns, all these measurement equipment itemized above as well as the DUTs except for the spectrum analyzer are placed inside the

anechoic chamber. In addition, noise data is obtained by plugging each DUT individually into an outlet within the anechoic chamber not allowing any other device to share the same outlet at the time of the measurement.

In order to investigate clearly the impulsive noise characteristics of each electrical load, suppression of other types of noise is crucial. Hence, in order to evaluate the effectiveness of the measurement set-up before starting the experiments, noise level of two outlets, one inside and the other outside the anechoic chamber, are compared. As can be seen from Figure 3.5, a remarkable change is observed. For most of the frequencies, more than 20 dB suppression of noise is achieved.

In order to show the effectiveness of the measurement set-up, noise seen from two different outlets, one is a regular outlet outside the set-up which is open to narrow band noise and the noise generated by the electrical loads connected to the network, and the other one is the outlet located inside the experimental environment, are considered in Figure 3.5. As can be seen obviously, a remarkable change is observed. For most of the frequencies, more than 20 dB suppression of noise is achieved.

3.1.2. Power-line Noise Measurement

The second measurement set-up is fairly simple compared to the previous set-up. Since the purpose here is to measure the summation of all the noise types mentioned in previous sections. Isolation or narrowband noise (HF, AM, FM) rejection is not considered in this set-up. The line coupler consists of a wideband PLC transformer with an almost flat frequency response from 4 KHz up to 50 MHz. National Instruments 100 MSPS digitizer, NI-PXIE-5122, is used to sample the noise and a high-speed PCIe interface, NI-8262 PCIe x, is used to store the sampled data into a high capacity (terabyte) hard-disk drive, NI-HDD8264, as seen in Figure 3.6.

Noise records are taken in a laboratory and residential house environment with the measurement set-up shown in Figure 3.6. Measurements of noise spectrum up to

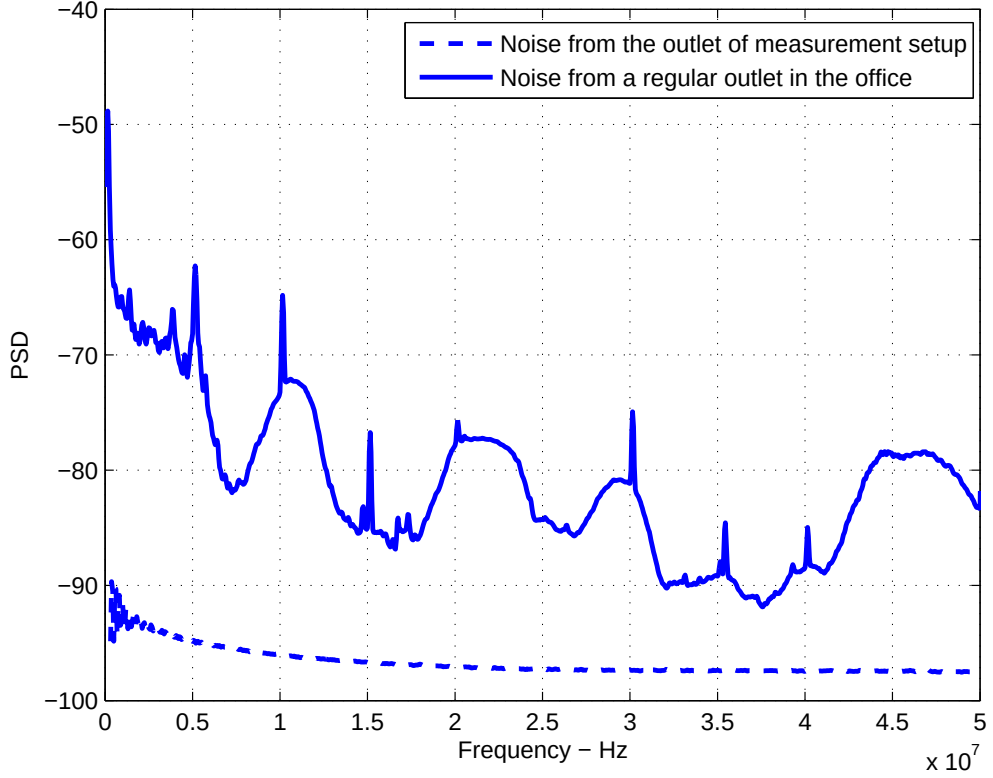


Figure 3.5. Noise levels of two outlets one inside and the other outside the anechoic chamber. It is clearly seen that the narrow band noise and the noise due to the electrical loads connected to the power line network are rejected successfully.

50 MHz are conducted for 60 minutes in blocks of 10 minutes at six different times of a day from the morning to the evening at two sites.

3.2. Background Noise

Commonly, background noise is assumed to be consisting of the colored background and the narrowband noise and it can be expressed as the summation of the two in the frequency domain as stated in the following,

$$n_{bg}(f) = n_{Cbg}(f) + n_{nb}(f) \quad (3.4)$$

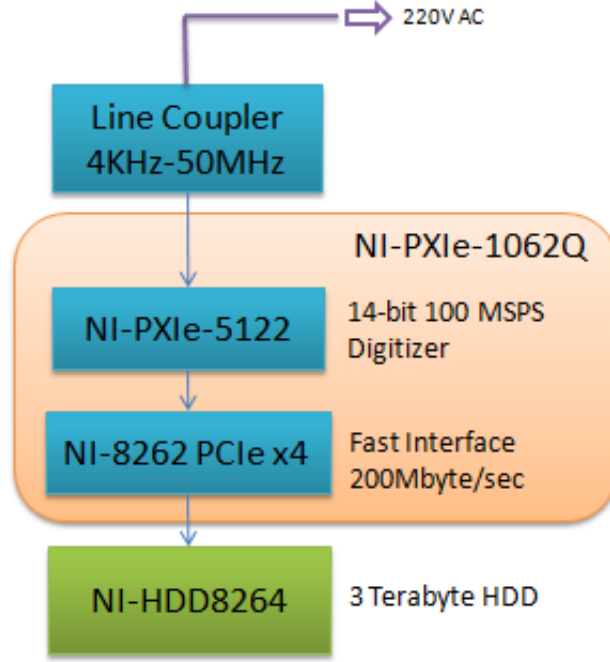


Figure 3.6. Power-line noise measurement set-up.

where $n_{bg}(f)$ represents the total background noise, $n_{Cbg}(f)$, represent the the colored background noise and $n_{nb}(f)$ represents the narrowband noise. The total background noise $n_{bg}(f)$ is assumed to be stationary for several minutes or even hours [1].

The colored background noise signal $n_{Cbg}(t)$ can be easily synthesized by filtering of a white noise source as depicted in Figure 3.7. The noise shaping filter is described by its transfer function $H_{mod}(z)$ in the z-plane:

$$H_{mod}(z) = \frac{B(z)}{A(z)} = \frac{1 + \sum_{i=1}^m b_i z^{-i}}{1 + \sum_{i=1}^m a_i z^{-i}} \quad (3.5)$$

The transfer function consists of an moving average (MA) portion in the numerator $B(z)$ and an autoregressive (AR) portion in the denominator $A(z)$. The parameters of the model are the variance σ^2 of the noise source and the coefficients of the filter. By use of an AR-process model, which means $B(z) = l$, the parameters can be determined from a measured noise signal with an AR-spectral estimator. Due to the fact that the

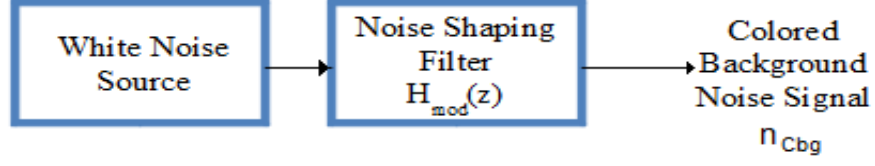


Figure 3.7. Generating the colored background noise.

PSD of the background noise changes only slowly over time, the model parameters have only to be changed for the simulation of a new state of the noise scenario.

On the other hand, narrowband noise sources are the RF transmissions in the LF, HF and UHF frequencies and consist of sinusoidal signals with modulated amplitudes [23]. Narrowband noise can be described as

$$n_{nb}(t) = \sum_{k=1}^N A_k(t) \sin(2\pi f_k t + \varphi_k) \quad (3.6)$$

where N represents the total number of narrowband interferers, and $A_k(t)$, f_k , and φ_k describe the amplitude, central frequency and phase of the received narrowband noise, respectively. Number of the interferers depend on the where the measurements are taken. One may assume that there will be more narrowband interferers in an urban area than a suburban area. Regardless of where the measurements are taken, phase φ_k of each narrowband source can be assumed uniformly distributed between $[0, 2\pi]$.

3.3. Impulsive Noise

It is commonly known that the most significant noise sources that affect the communications quality in PLC channels are the impulsive noise sources and studied extensively in [1, 28, 44, 45]. Most of these impulses are located int low frequencies and mathematically, they can be shown as the summation of damped sinusoidals with different periods as in the following

$$n_{imp}(t) = \sum_{i=1}^{N_s} A_i \cdot \sin(2\pi f_i(t - t_p) + \varphi_i) \cdot e^{(-t-t_p)/\tau_i} \cdot \eta\left(\frac{t - t_p}{t_\eta}\right). \quad (3.7)$$

where N_s is the number of damped sinusoids that form the impulse, A_i represents the amplitude of the i th sinusoid, f_i denotes the pseudofrequency of the damped sinusoid, t_p is the arrival time of the impulses, φ_i represents the phase of the i th sinusoid, τ_i is the damping factor, and $\eta(t)$ denotes a square pulse with a duration of t_η . The amplitude value A_η of square pulse $\eta(t)$ is

$$A_\eta = \begin{cases} 1, & \text{if } 0 < t < t_\eta \\ 0, & \text{elsewhere} \end{cases} \quad (3.8)$$

The amplitude of each damped sinusoid A_i is selected to be $\sim \mathcal{N}(0, G_k \sigma_n^2)$ where G_i denotes the increment of the impulse over the background noise with a variance of σ_n^2 . Values of G_i can change between 20 – 30 dB.

3.3.1. Periodic Impulsive Noise

Studies in the literature and our measurements show that most of the impulses encountered in PLN are periodic in nature and synchronous to the mains frequency [23]. They usually have a repetition rate of 50 – 100 Hz. Hence, arrival time of these impulses are also somewhere between 10ms or 20ms. Each impulse has an average duration of 50 μ s which is set to be the duration of the square pulse t_η . It was stated in [28, 44], that the number of damped sinusoids N_s is determined as 3 and pseudofrequencies of the damped sinusoids are set as 300 KHz, 2 MHz, and 3.5 MHz.

3.3.2. Aperiodic Impulsive Noise

On the other hands, another impulsive noise type is the aperiodic impulsive noise. Aperiodic impulses do not have any periodic arrival time and their arrival time can be modeled as a random variable. It was shown that interarrival time between two successive aperiodic impulses has an exponential distribution with a mean of 100ms [1] and can be formulated as

$$t_{int} = t_p(n) - t_p(n - 1) \quad (3.9)$$

where $t_p(n)$ represents the arrival time of n th aperiodic impulse. Amplitude distribution of the aperiodic impulses are assumed to be Gaussian as in the case of periodic impulses.

Real and imaginary parts of a simulated impulse with a time duration of $50\mu s$ are illustrated in Figure 3.8 and Figure 3.9, respectively. Using these simulations, a more realistic impulsive noise is generated and depicted in Figure 3.10.

3.4. Cyclo-stationarity Analysis of Appliance Noise

The statistics of the noise generated by the electrical loads connected to the PLC network is found to be periodic with the AC cycle. This periodicity in the statistics stems mainly from circuitry of the devices and can be 50 Hz or 100 Hz due to the fact that some of the devices are affected by the absolute value of the mains voltage, whereas some by its polarity. This periodicity in the statistics of the noise favors the idea of treating it as a cyclostationary or periodically stationary process.

Defining the autocorrelation as,

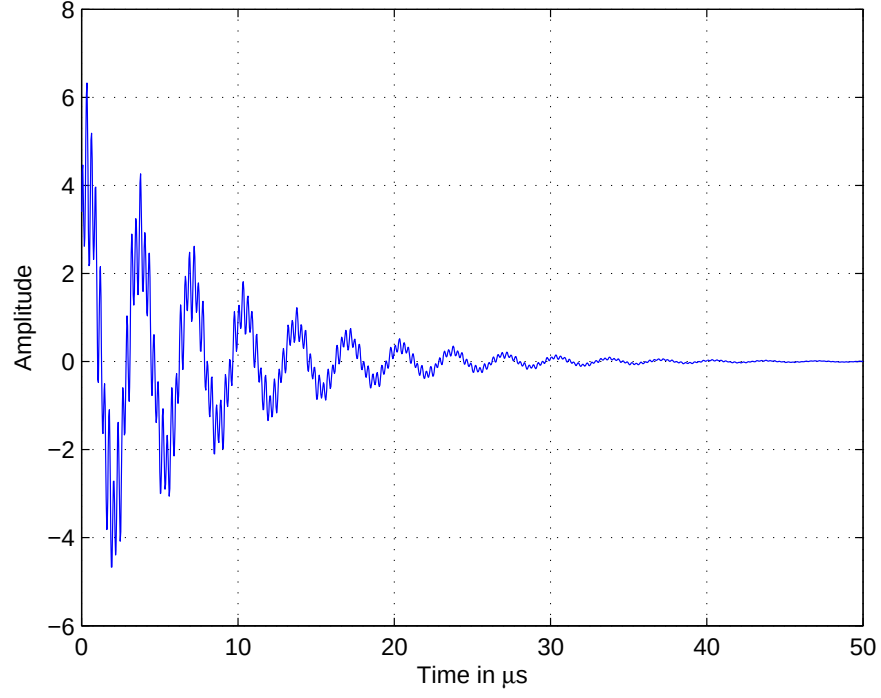


Figure 3.8. Real part of an impulse noise.

$$R(t + \tau, t) = E\{n(t)n^*(t + \tau)\}. \quad (3.10)$$

where $E\{\cdot\}$ is the statistical expectation operator, $(\cdot)^*$ denotes the complex conjugate of its input, τ is the time shift in the correlation operation.

A process, $n(t)$ is cyclostationary if its mean μ and autocorrelation functions denoted as $R(\cdot)$ satisfy the following conditions [46]:

$$\mu(t + mT) = \mu(t) \quad (3.11)$$

$$R(t + \tau + mT, t + mT) = R(t + \tau, t) \quad (3.12)$$

It is implied in Equation 3.12 that the mean and correlation properties of the process do not change at the integer multiples of a period T , which corresponds to

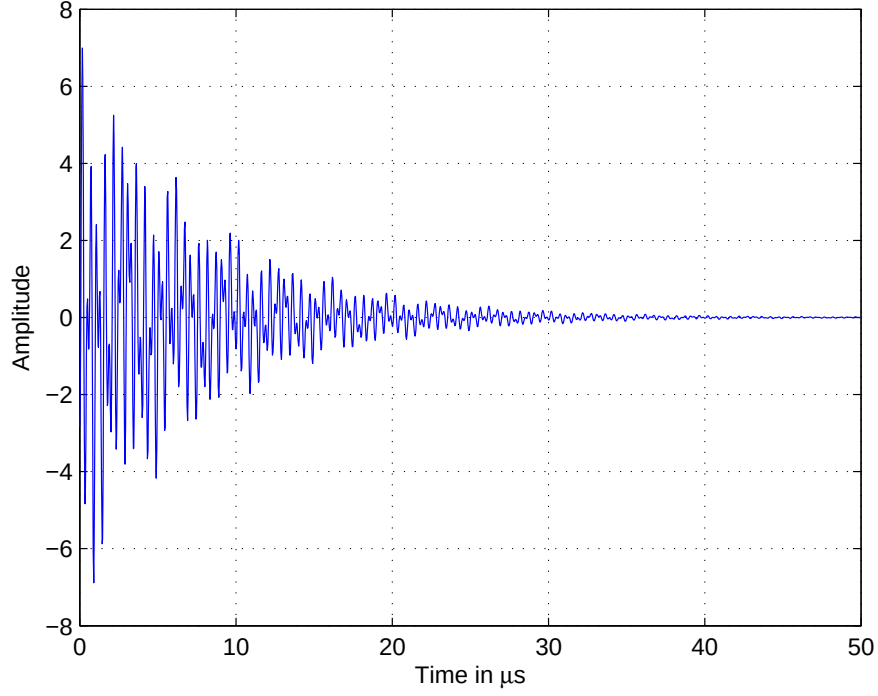


Figure 3.9. Imaginary part of an impulse noise.

half or one alternating current (AC) cycle period. A straightforward cyclostationarity analysis performed on one of the devices verify this in a very clear way. Looking at the autocorrelation of the absolute value of the captured noise waveform over a duration of multiple AC cycles, result of the analysis shown in Figure 3.11. confirms our assumption.

However, frequency content of the noise which is revealed by the observation of power spectral density (PSD) is of equal importance considering the communication systems operating at various frequencies. Instantaneous PSD and the autocorrelation function of a random process are related to each other via forward fast Fourier transform (FFT) operation as follows:

$$S(t, f) = \int_{-\infty}^{\infty} R(t + \tau, t) e^{-j2\pi f\tau} d\tau \quad (3.13)$$

There is a time dependency in Equation 3.13 and we need to eliminate it to look at the average PSD of a cyclostationary process. Hence, an averaging over an AC cycle

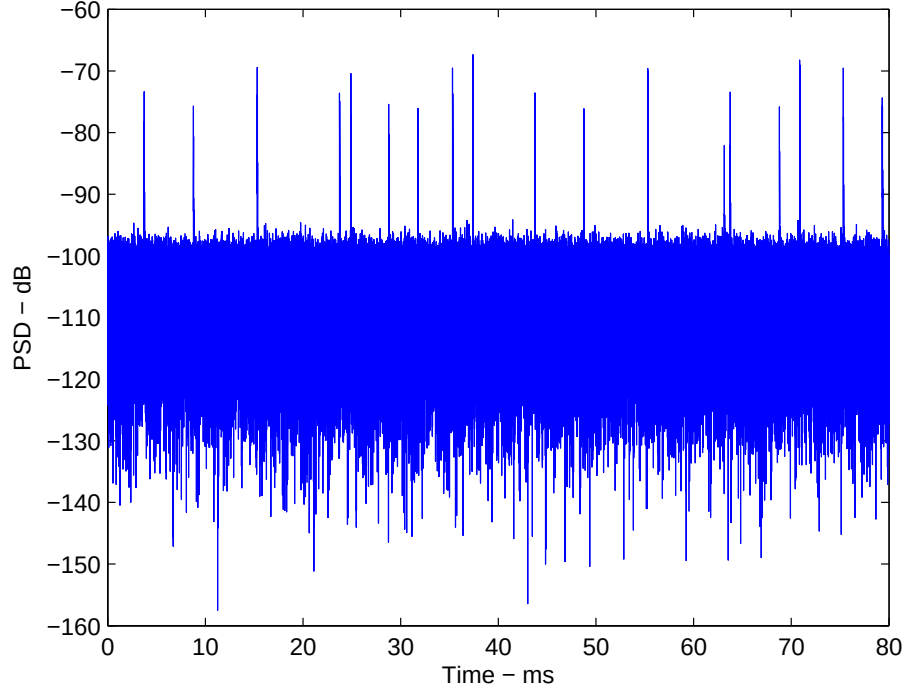


Figure 3.10. Simulation results of the impulsive noise in PLC.

is required. Averaging out the PSD, PSD of a cyclostationary process becomes

$$S(f) = \frac{1}{T_0} \int_{-T_0/2}^{T_0/2} S(t, f) dt \quad (3.14)$$

It is seen in Equation 3.13 that the periodicity in the autocorrelation function appears as a periodicity in the instantaneous PSD of the random process. Therefore, in order to reveal the cyclostationary features of the noise, one might want to look into time–frequency analysis (TFA) analysis of the noise generated by the electrical loads. TFA is commonly used to study the type of signals whose statistics vary over time. As in the case of PSD estimation, one need to average TFA over multiple cycles in order to average the signal out and reveal the features of the noise. On the other hand, to see how the signal evolves in time, the noise data n can be divided into smaller chunk of signals.

In the analysis, it is assumed that the captured signal duration equals to the M multiple of T_0 and each section with the duration of T_0 is divided into K pieces.

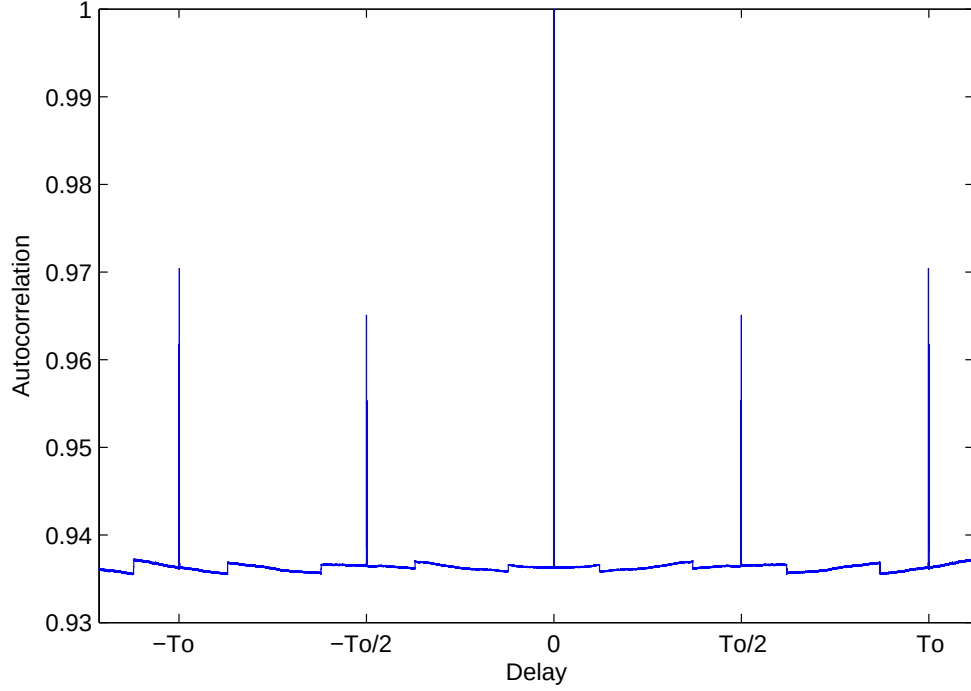


Figure 3.11. Autocorrelation of the light dimmer.

Hence, the total duration of the captured data becomes MT_0 . Let $N_{MK}(k, m)$ denote the discrete data samples falling into the k th piece within the m th AC cycle:

$$N_{KM}(k, m) = n(\frac{kT_0f_s}{K} + mT_0f_s : \frac{(k+1)T_0f_s}{K} + mT_0f_s) \quad (3.15)$$

where k and m are the factors that assume values from 0 to $K-1$ and from 0 to $M-1$, respectively. f_s is the sampling frequency of the measurement equipment that records the noise data, $(x : y)$ implies the inclusion of discrete data samples from start position x till y . If the periodogram of Equation 3.15 is averaged by considering each of the pieces falling into the same phase of the AC cycle, the following expression is obtained:

$$S_{N_{KM}}(k, m) = \frac{\sum_{m=0}^{K-1} \frac{K}{T_0f_s} |FFT(N_{MK}(k, m))|^2}{M} \quad (3.16)$$

Looking at the result of the analysis for power change and the rate of power change given in Table 3.1, some significant inferences can be made for future studies,

- *Computer tower*: A significant increase is observed over the background noise up to 55 dB. It can also be seen that most of the power is distributed up to 4 MHz.
- *Dimmer*: A significant increase is observed over the background noise up to 35 dB. It can also be seen that most of the power is distributed up to 5 MHz.
- *Drill*: A significant increase is observed over the background noise up to 45 dB. It can also be seen that most of the power is distributed between 11 MHz–17.5 MHz.
- *Vacuum cleaner*: A significant increase is observed over the background noise up to 30 dB. It can also be seen that most of the power is distributed over various frequencies.
- *TV set*: A significant increase is observed over the background noise up to 35 dB. It can also be seen that most of the power is distributed between 12.5 MHz–21 MHz.
- *LCD monitor*: A significant increase is observed over the background noise up to 20 dB. It can also be seen that most of the power is distributed over almost all frequencies with some spikes at some particular frequencies.
- *CRT monitor*: A significant increase is observed over the background noise up to 20 dB. It can also be seen that most of the power is distributed between 10 MHz–12.5 MHz with some spikes around 4 MHz.
- *Fluorescent*: A significant increase is observed over the background noise up to 10 dB. It can also be seen that most of the power is distributed around 5 MHz.
- *Laptop charger*: A significant increase is observed over the background noise up to 15 dB. It can also be seen that most of the power is distributed up to 25 MHz.
- *Washing machine*: A significant increase is observed over the background noise up to 20 dB. It can also be seen that most of the power is distributed around 7 MHz.

Power change in time domain is also analyzed and some comments are made on the time domain features of the electrical loads in the following;

- *Computer tower*: Increase up to 45 dB above the noise floor for most of the mains cycle duration
- *Dimmer*: Increase up to 40 dB above the noise for a very short time over the mains cycle duration
- *Drill*: Increase up to 50 dB above the noise floor for a duration of ≈ 8 ms
- *Vacuum cleaner*: Increase up to 20 dB above the noise floor covering half of the mains cycle duration
- *TV set*: Increase up to 30 dB above the noise floor for a duration of 4ms
- *LCD monitor*: Increase up to 13 dB above the noise floor covering most of the mains cycle duration
- *CRT monitor*: Increase up to 10 dB above the noise floor covering most of the mains cycle duration
- *Fluorescent*: Increase up to 6 dB above the noise floor covering half of the mains cycle duration
- *Laptop charger*: Increase up to 13 dB above the noise floor covering half of the mains cycle duration
- *Washing machine*: Increase up to 18 dB above the noise floor covering half of the mains cycle duration

The cyclostationary behavior of the noise generated by electrical devices has been quantified by considering another two parameters, peak excursion and maximum rate of change of the instantaneous PSD as defined in [47]. These parameters are denoted as P_e and R_c , respectively. P_e represents the maximum power change value during one mains cycle among all the frequencies. On the other hand, R_c represents the maximum power change in T_r second along the same frequency axis in one mains cycle. P_e and R_c can be formulated as the following,

$$P_e = \max\{10\log_{10}[S_{N_{KM}}(t, f)]\} - \min\{10\log_{10}[S_{N_{KM}}(t, f)]\}, \quad (3.17)$$

Table 3.1. P_e and R_c of electrical loads.

Measured Device	P_e (dB)	R_c (dB)
Computer Tower	61.26	17.43
Dimmer	45.61	36.83
Drill	59.66	37.03
Vacuum Cleaner	42.49	21.92
TV Set	42.13	4.95
LCD Monitor	30.32	21.68
CRT Monitor	28.75	15.45
Fluorescent	24.30	7.98
Laptop Charger	23.47	21.12
Washing Machine	40.81	26.72

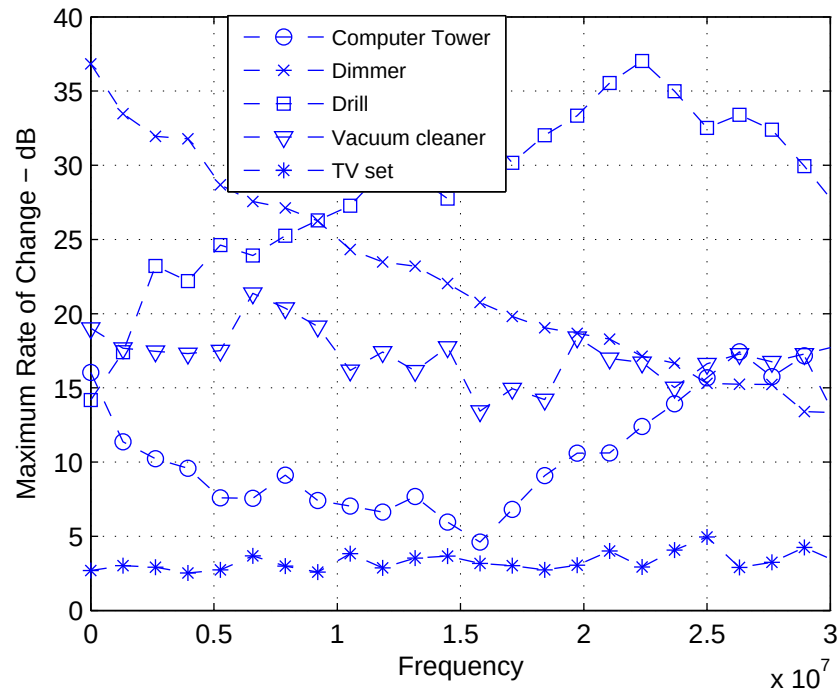
$$R_c = \max\{|10\log_{10}[S_{N_{KM}}(t + T_r, f)] - 10\log_{10}[S_{N_{KM}}(t, f)]|\}. \quad (3.18)$$

Results for the measured electrical loads are given in Table 3.1. It is seen that computer tower and drill are the most significant noise sources in PLC.

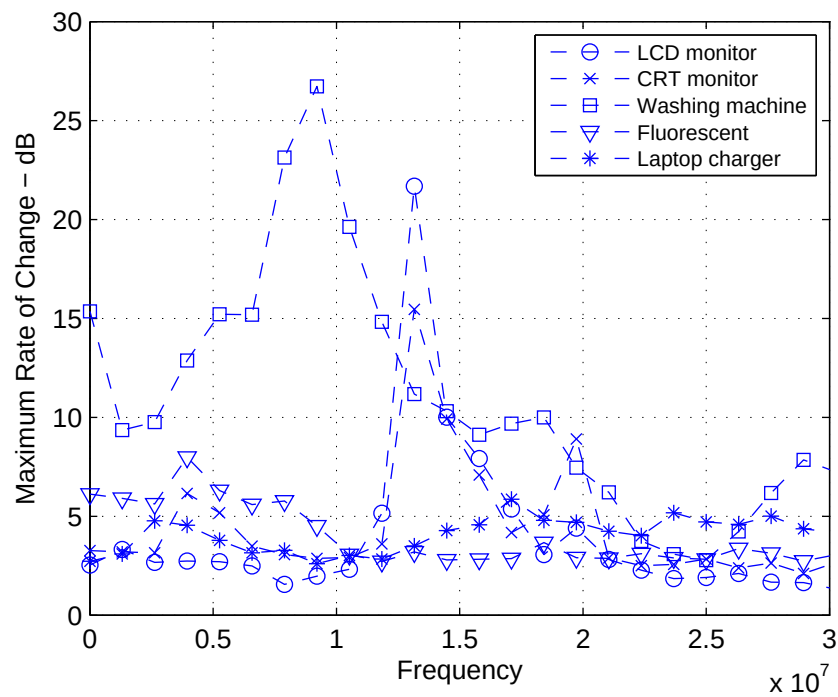
In order to have a better understanding of the significance of electrical loads, the maximum rate of change for each frequency has been depicted in Figure 3.13a and 3.13b.

3.5. Noise Amplitude Distribution Modeling at a Single Frequency

It was shown that Gaussian assumption for amplitude distribution of each individual frequency is not correct [36, 39], rather the noise amplitude at a single frequency follows the Nakagami-m distribution [39].



(a)



(b)

Figure 3.13. Maximum power of change for each frequency over an AC cycle.

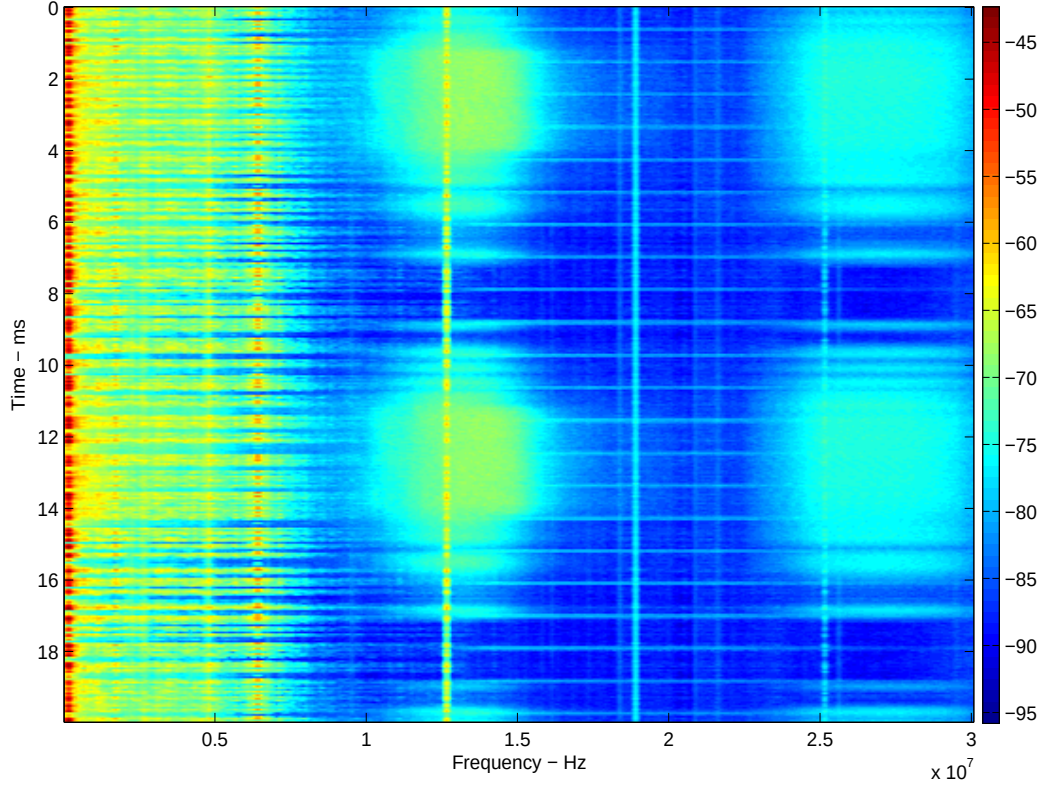


Figure 3.14. TFA of background noise captured from outside of the measurement set-up.

Although, Middleton Class-B noise amplitude distribution in time domain and Nakagami-m noise amplitude distribution in frequency domain seem to capture the behaviour of the broadband power-line noise, these approaches are mathematically not tractable and require numerical integration and approximations [39] to compute the real part of the PDF. Therefore, it can be concluded that these models do not lead to closed form equations to design optimal receivers.

Hence, in this thesis study, new measurements are taken from power-line to look for a more suitable distribution so that one can design an optimal receiver using the closed form of the probability density function (PDF) of the proposed noise model.

Measurements recorded using the set-up depicted in Figure 3.6 is used for analysis. In Figure 3.23, an estimate of power spectral density (PSD) from the measurement data is shown.

3.5.1. Proposed Noise Model

In the literature, Nakagami- m distribution has been proposed to model the noise amplitude distribution in frequency [39]. However, Nakagami- m may not be the best choice for the following reason. The drawback of Nakagami- m distribution is that Nakagami- m does not have a closed form when the shape parameter m has a value smaller than 0.5. Given a random variable X , the Nakagami- m PDF can be written as

$$p(r) = \frac{2}{\Gamma(m)} \left(\frac{m}{\omega}\right)^m r^{2m-1} e^{-\frac{mr^2}{\omega}} \quad (3.19)$$

and its shape parameter estimation is given as;

$$m = \frac{E^2[X^2]}{Var(X^2)} \quad (3.20)$$

Estimation of the shape parameter m based on field data using Equation 3.20 ends up being smaller than 0.5 as given in [39]. On the other hand, the closed form expressions for evaluation of the BER under Nakagami- m optimal receivers have not been obtained and are not very trivial [39]. When the noise is not independent and identically distributed as in the case of PLC, one would need these expressions to design an optimal receiver in the BER sense. Hence, in this study we analyzed an analytically suitable distribution named GGD for modeling power-line noise amplitude distribution of each individual frequency.

Since GGD is a parametric family of symmetric distributions, it has been commonly used to model and detect non-Gaussian and Gaussian signals. As limiting cases, it can also be used to model uniform distributions and Laplace distributions on the bounded intervals of the real line. [48]. The probability density function of GGD is given by

$$p(y; \mu, \alpha, \beta) = \frac{\beta}{2\alpha\Gamma(\frac{1}{\beta})} e^{-(\frac{|y-\mu|}{\alpha})^\beta} \quad (3.21)$$

where $\Gamma(\cdot)$ denotes the Gamma function,

$$\Gamma(z) = \int_0^\infty e^{-t} t^{z-1} dt \quad (3.22)$$

$z > 0$. Parameter α models the width of the PDF peak (variance), while parameter β is inversely proportional with the decreasing rate of the peak [49].

This distribution converges to the normal distribution when $\beta = 2$ with mean μ and variance $\alpha^2/2$ and it converges to Laplace distribution when $\beta = 1$. As $\beta \rightarrow \infty$, the density converges to a uniform density on $(\mu - \alpha, \mu + \alpha)$.

Maximum likelihood estimation and the method of moments have been studied to estimate model parameters [50]. Since the deriving closed form of the estimators is not trivial, estimates are obtained numerically [50].

In order to test the validity of our assumption, normalized KL distance that will be described later is utilized between the estimated p.m.f and discretized version of pdf of our empirical data.

We denote the true probability mass function (PMF) of the sampled (empirical) data by $p(\cdot)$, and the discretized PDF of the GGD approximation by $q(\cdot)$. The empirical PMF $p(\cdot)$ computed from the sampled data set $\mathbb{Y} = \{y_1, y_2, \dots, y_N\}$, where N is the number of samples in the observation set, is given by

$$p_i = \frac{M_i}{N\Delta y}, \quad i = 1, \dots, D, \quad (3.23)$$

where $\Delta y = (\lceil y_{\max} \rceil - \lfloor y_{\min} \rfloor)/D$, D is the number of bins in the histogram, M_i is the number of samples in the i th bin, and $y_{\max}$ and $y_{\min}$ are the maximum and minimum values in the sample set $\mathbb{Y}$.

Let $\mathbb{S} = \{s_1, \dots, s_D\}$ denote the bin centers of the histogram of the empirical noise data. Then the discretized GGD PDF can be written as

$$q_i = f_Y(s_i; 0, \alpha, \beta), \quad i = 1, \dots, D, \quad (3.24)$$

where $\mu = 0$ since our measurements show that the noise distribution is always zero-mean, and the parameters α and β are estimated from noise measurements.

The normalized Kullback-Leibler (KL) distance $D(p||q)/H(p)$ where $H(p)$ is the entropy associated with the PMF $p(\cdot)$, and

$$D(p||q) = \sum_{i=1}^D p_i \log \frac{p_i}{q_i} \quad (3.25)$$

is the Kullback-Leibler (KL) distance, is employed as the measure of "closeness between distributions in this paper (see e.g. [49, 51] for probabilistic distance measure). In order to verify that the PLC noise amplitude distribution for each frequency can be modeled by GGD, the normalized KL distance between the estimated PMF, p_i in Equation 3.23, and the discretized version of the PDF of our empirical data, q_i in Equation 3.24, is evaluated. Figure 3.26 shows that the GGD assumption is valid over the entire 50 MHz bandwidth. Since man-made impulsive noise is encountered mostly in low

frequencies, the GGD PDF deviates from the empirical distribution the most in this range. However, the amount of mismatch in the KL sense is still low in the 0-to-12 MHz range. We observe a constant and negligible degree of mismatch from 12 MHz to 50 MHz. In Section 3.5.2, a method for estimating the GGD parameters is given and in Figure 3.27, the variation of the GGD shape parameter is illustrated.

Computing the normalized KL distance between the discretized PDF of analytical GGD and the discrete empirical PDF, it is seen that GGD assumption remains acceptable over the whole 50 MHz bandwidth. Increase of normalized KL distance values at low frequencies is due to man-made noise since man-made noise drives the distribution out of the noise model.

3.5.2. Moment Estimation for β

Since we have shown that GGD is an acceptable assumption for the noise in frequency domain, now, we would like to investigate how the parameters of GGD changes over the whole bandwidth. This information gives us some insight into how to design optimal receivers against this type of noise in an OFDM set-up.

The first order absolute and second order moments (normalized) of the sample set $X = \{x_i\}_{i=1}^N$ are respectively

$$m_1 = \sum_{i=1}^N \frac{|x_i|}{N} \quad \text{and} \quad m_2 = \sum_{i=1}^N \frac{|x_i^2|}{N} \quad (3.26)$$

while the counterparts in the GGD calculated respectively as [50]

$$M_1 = \int_{-\infty}^{\infty} |x| p(x, \alpha, \beta) dx = 2K \frac{\alpha^2}{\beta} \Gamma(2/\beta) \quad (3.27)$$

$$M_2 = \int_{-\infty}^{\infty} x^2 p(x, \alpha, \beta) dx = 2K \frac{\alpha^3}{\beta} \Gamma(3/\beta) \quad (3.28)$$

where

$$K = \frac{\beta}{2\alpha\Gamma(1/\beta)}.$$

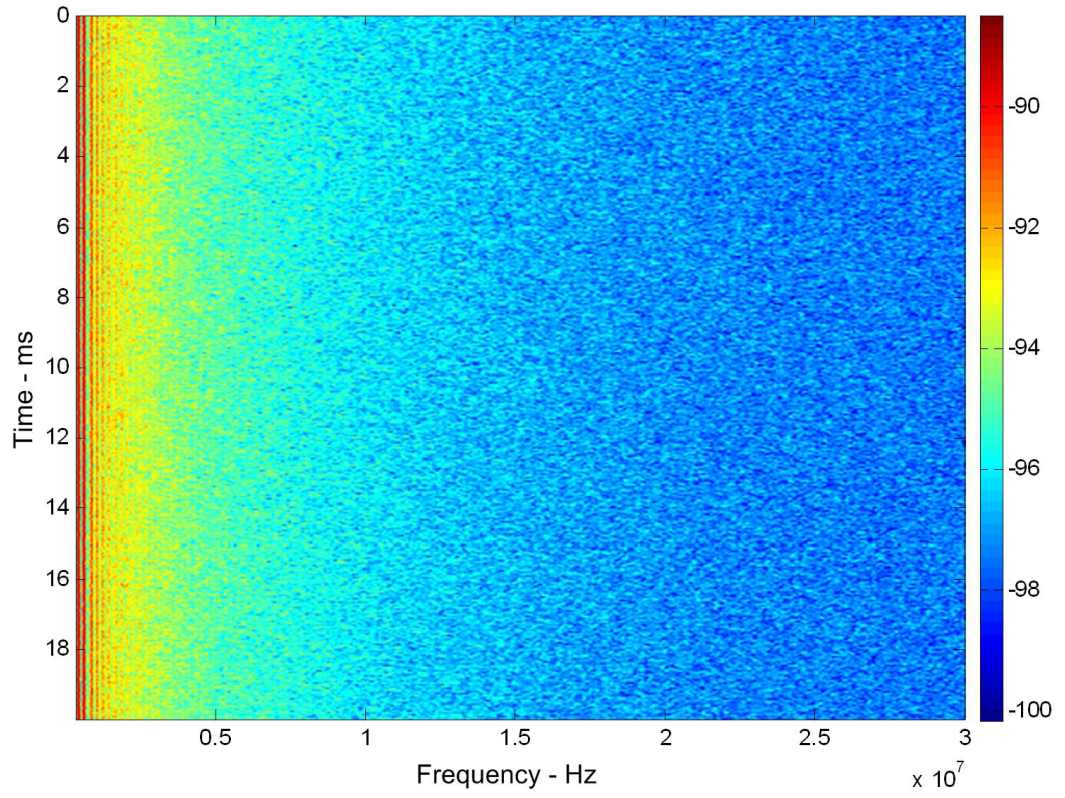
Setting the moments of the sample set equal to the GGD moments, the PDF parameters can be estimated by solving for them in $M_1 = m_1, M_2 = m_2$ so that

$$\hat{\beta} = F^{-1} \left(\frac{m_1^2}{m_2} \right) \quad \text{and} \quad \hat{\alpha} = m_1 \frac{\Gamma(1/\hat{\beta})}{\Gamma(2/\hat{\beta})},$$

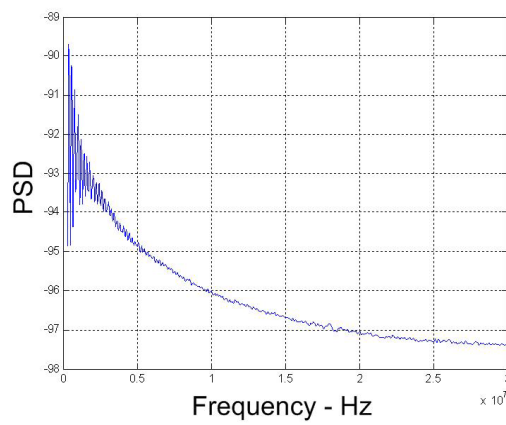
where

$$F(x) = \frac{\Gamma^2(2/x)}{\Gamma(1/x)\Gamma(3/x)}.$$

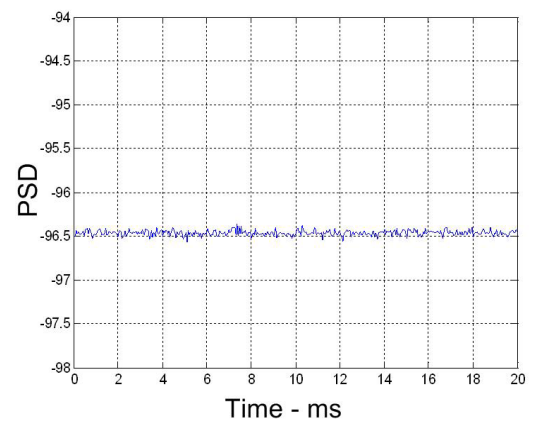
Moment estimates of the β parameter in Figure 3.27 show that the shape of power-line noise amplitude distribution in frequency tends to be non-Gaussian most of the time and the variation of the noise shape is greater in the low frequency region up to 12 MHz.



a.

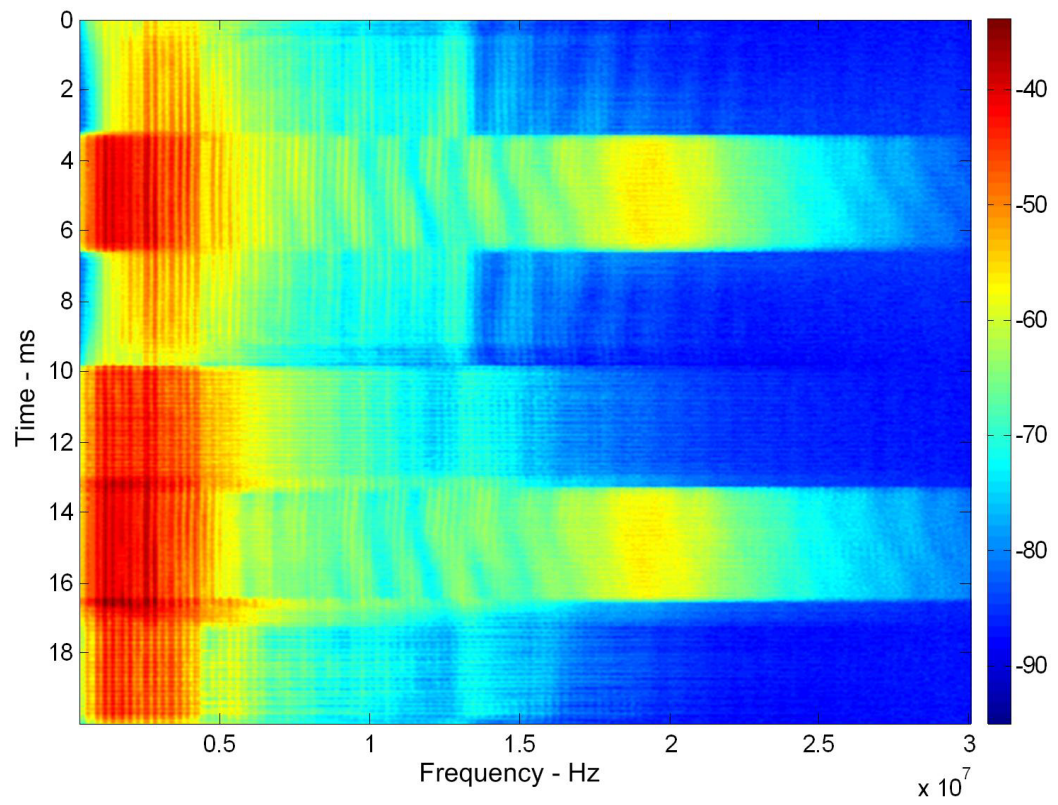


b.

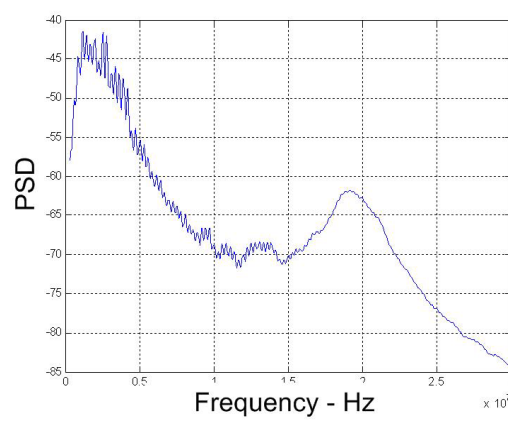


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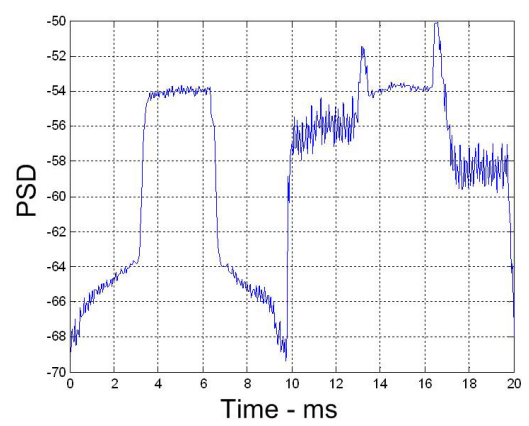
Figure 3.15. TFA of background noise in anechoic chamber.



a.

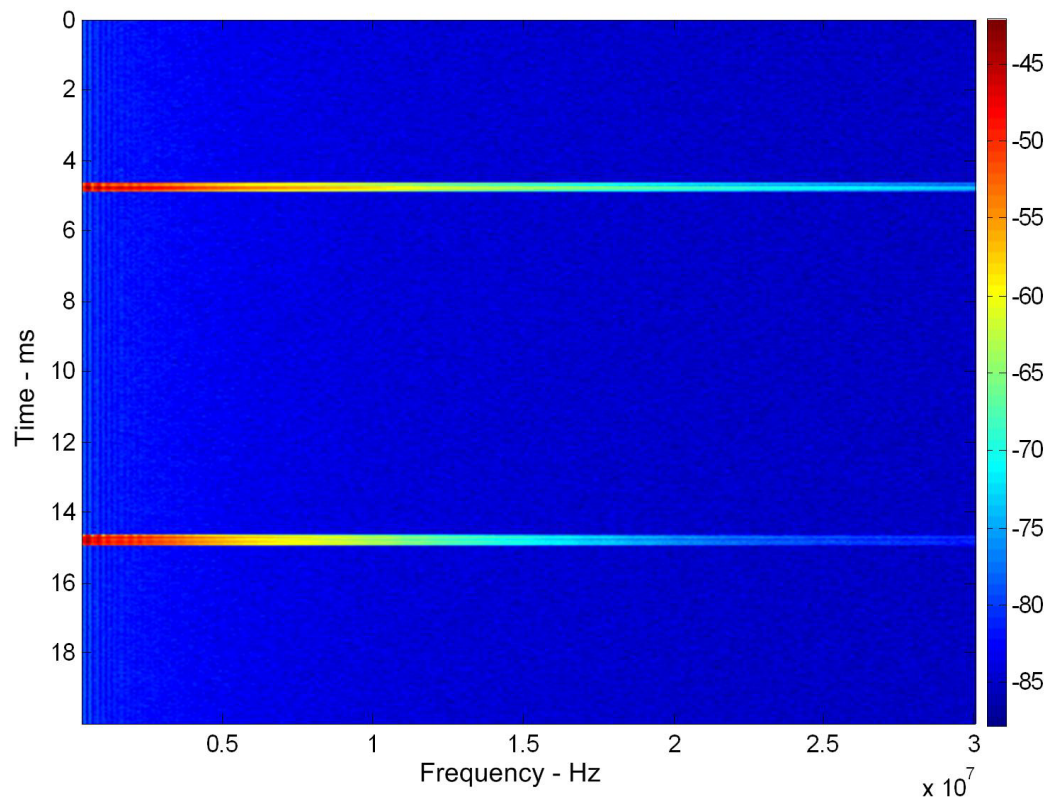


b.

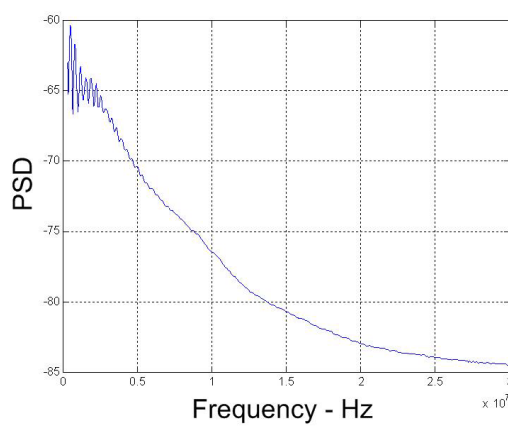


c.

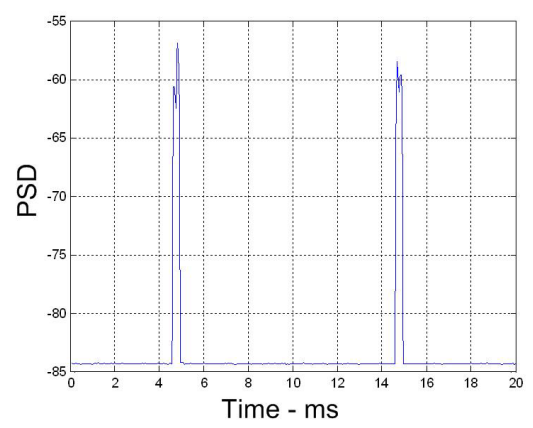
Figure 3.16. TFA of computer tower.



a.

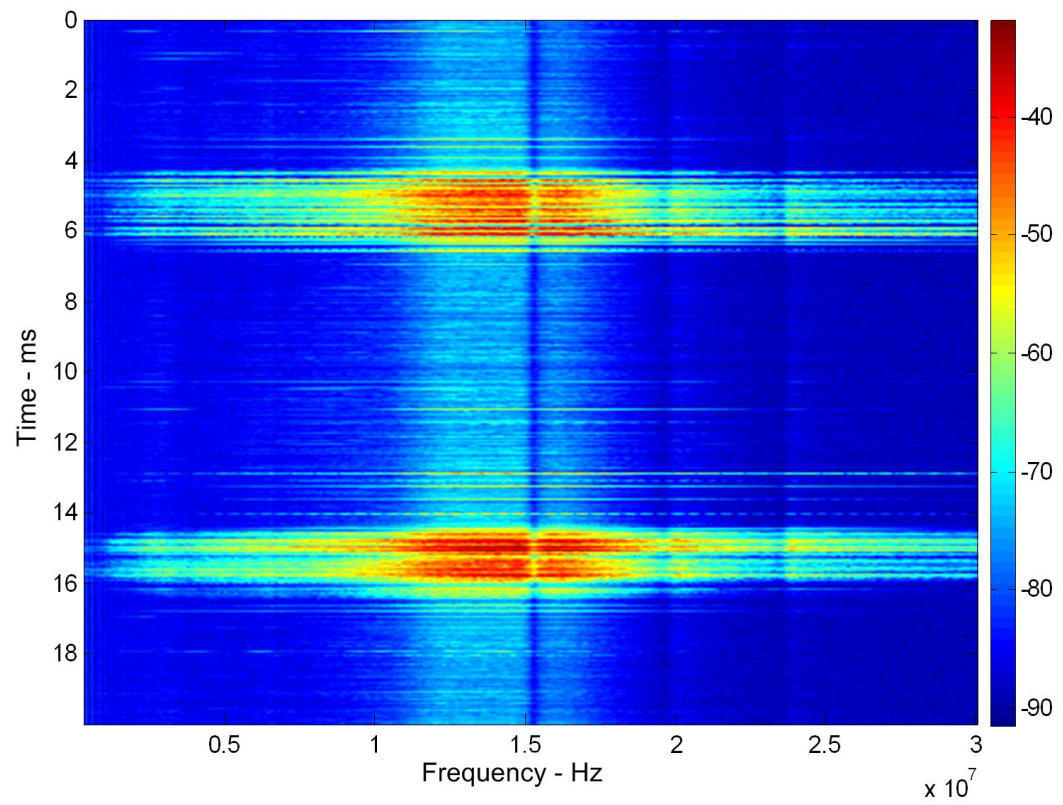


b.

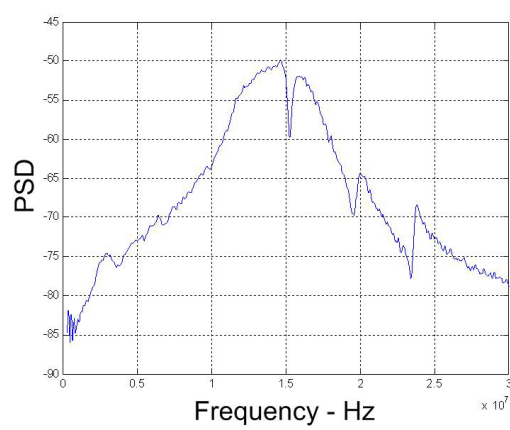


c.

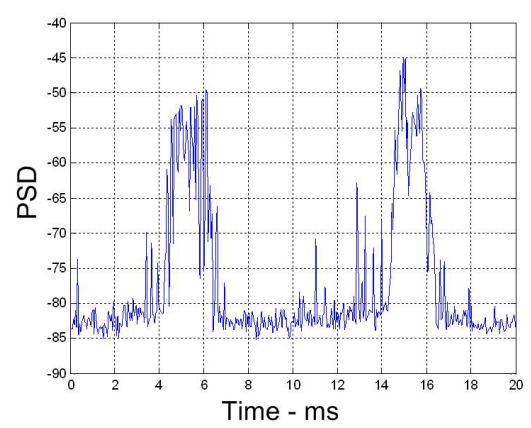
Figure 3.17. TFA of dimmer.



a.

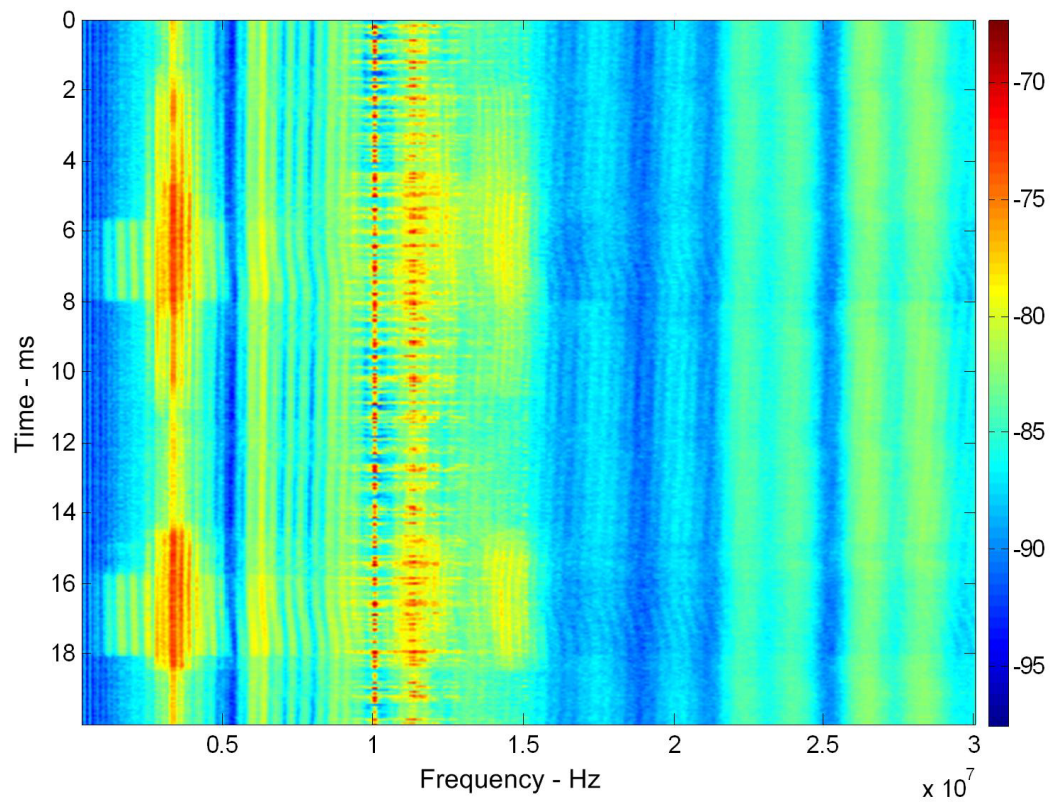


b.

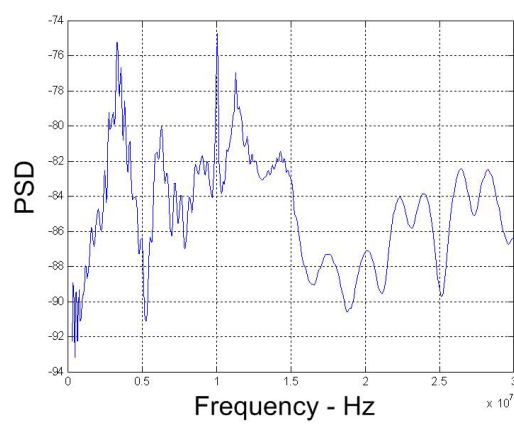


c.

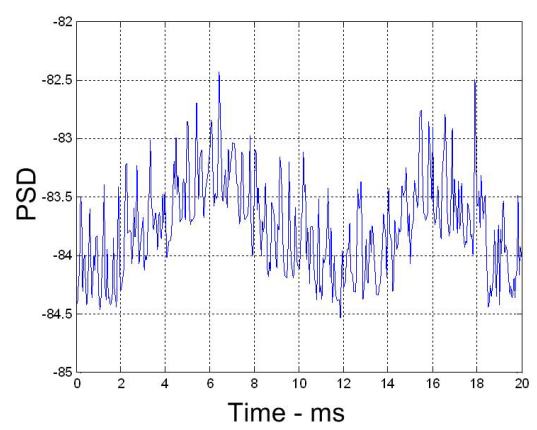
Figure 3.18. TFA of drill.



a.

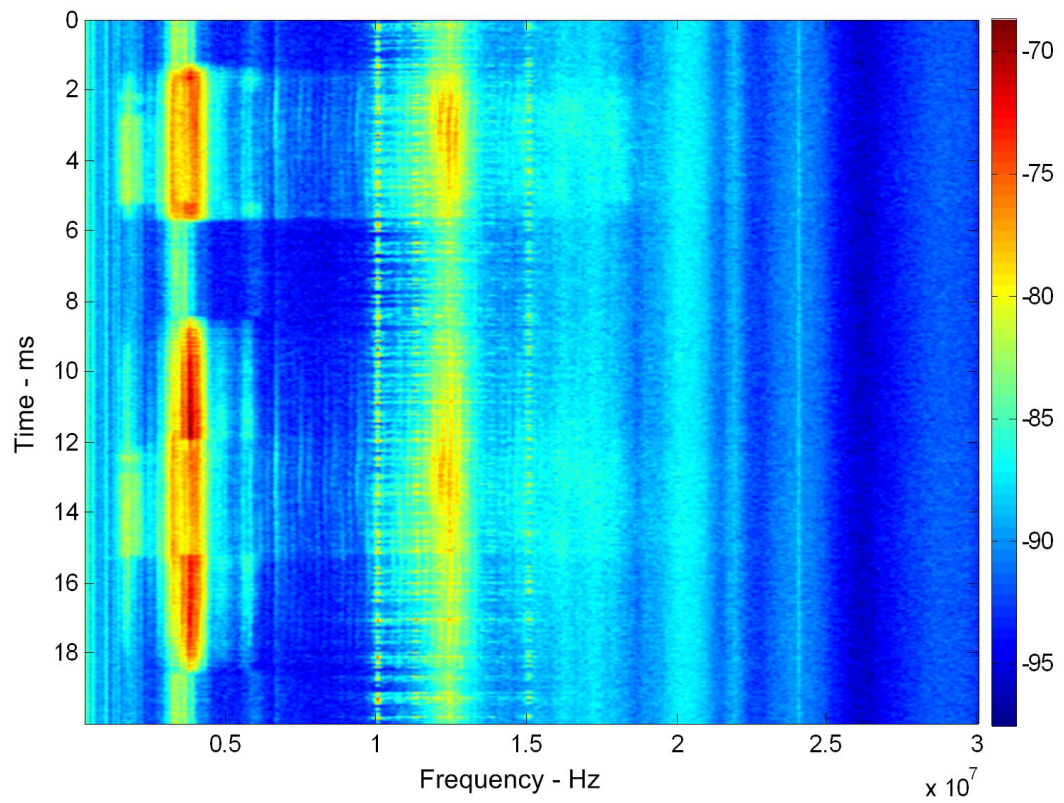


b.

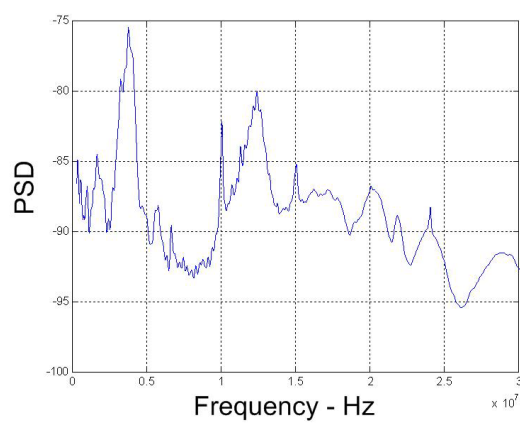


c.

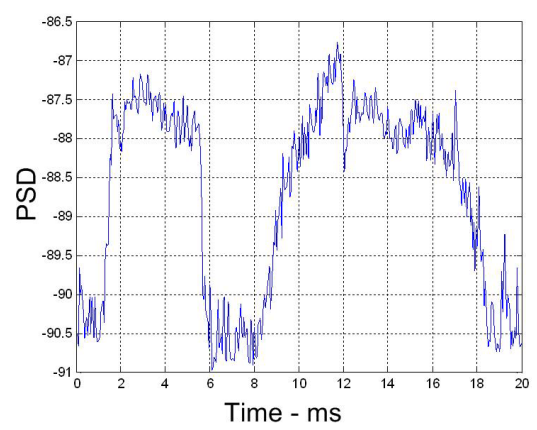
Figure 3.19. TFA of LCD monitor.



a.

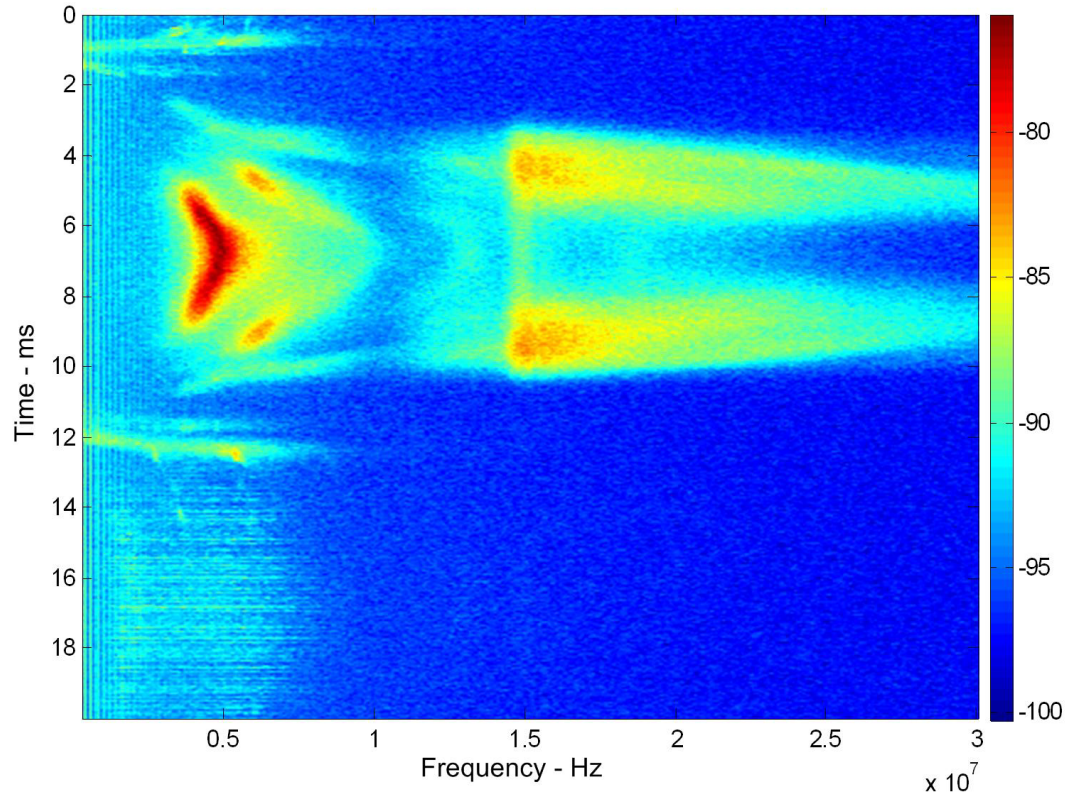


b.

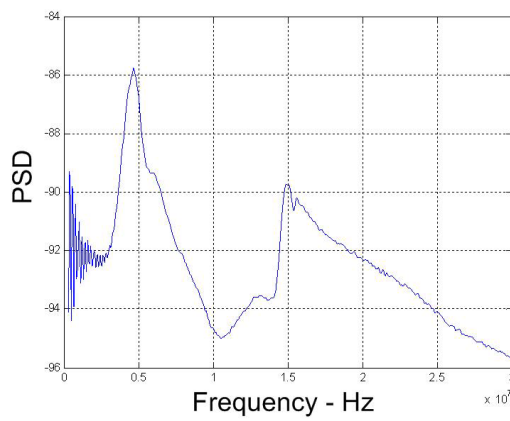


c.

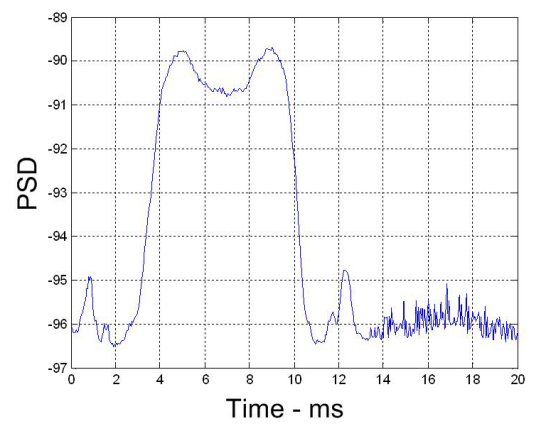
Figure 3.20. TFA of CRT monitor.



a.

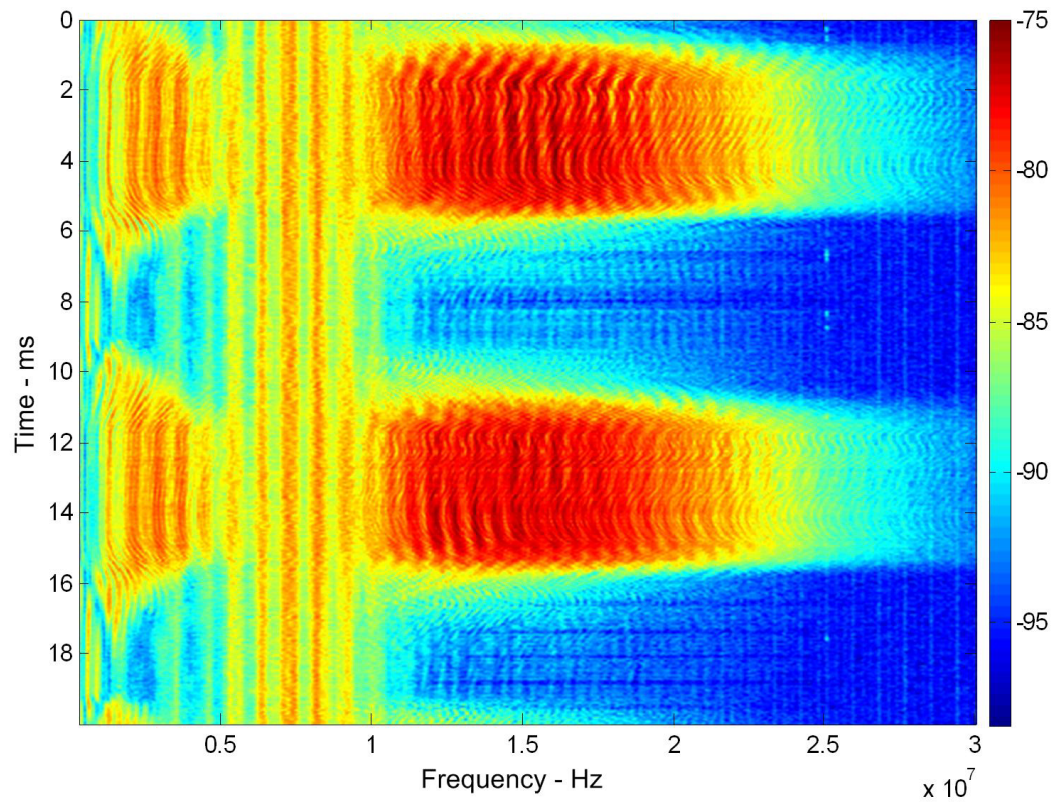


b.

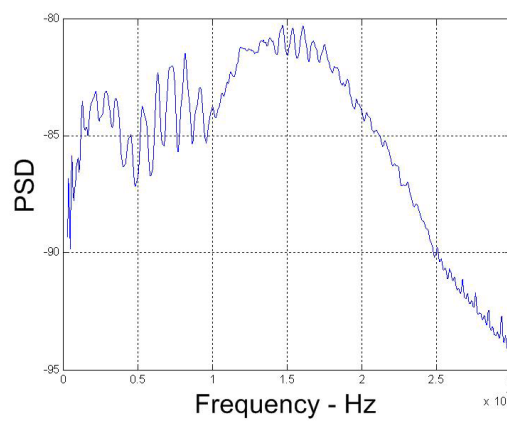


c.

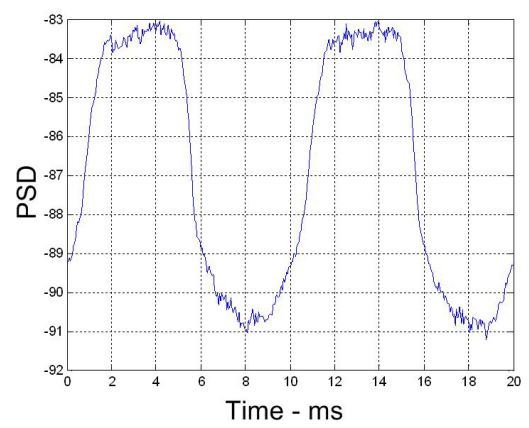
Figure 3.21. TFA of fluorescent.



a.



b.



c.

Figure 3.22. TFA of a laptop charger.

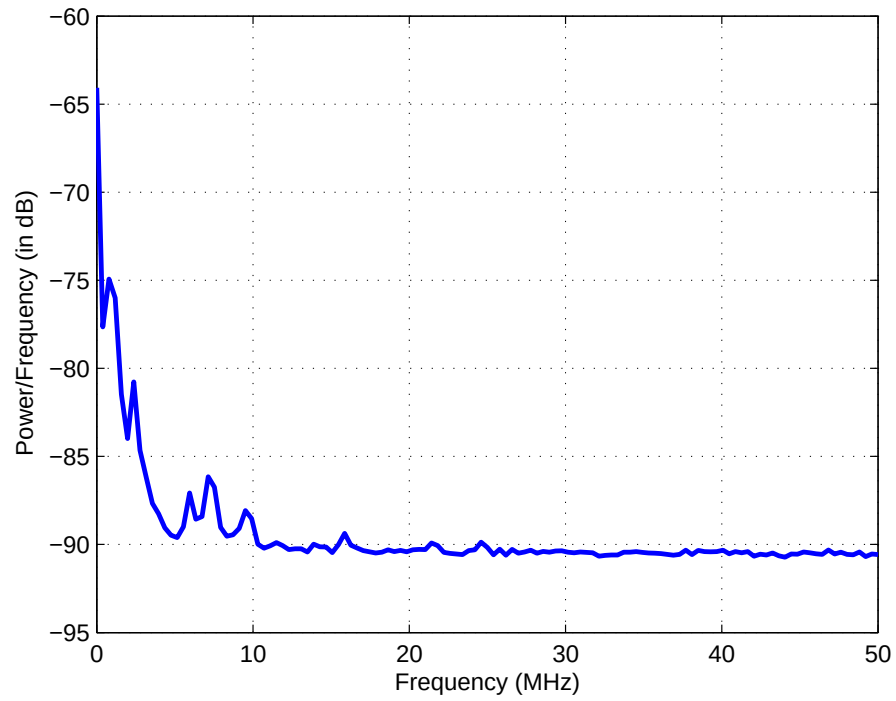


Figure 3.23. Power spectral density of the background noise recorded in a residential house.

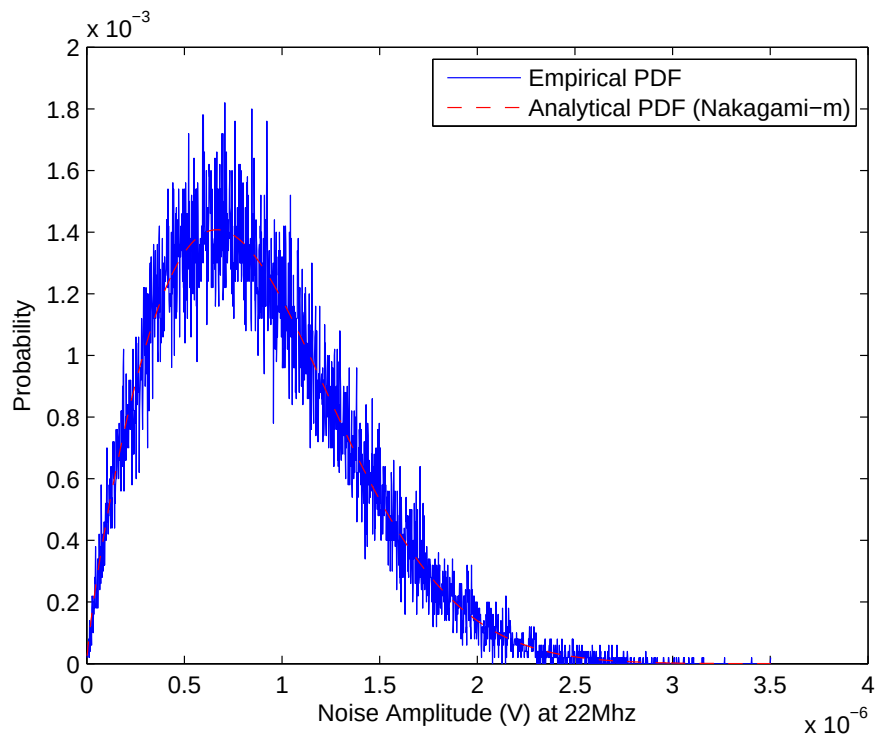


Figure 3.24. Empirical PDF versus Analytical Nakagami-m PDF.

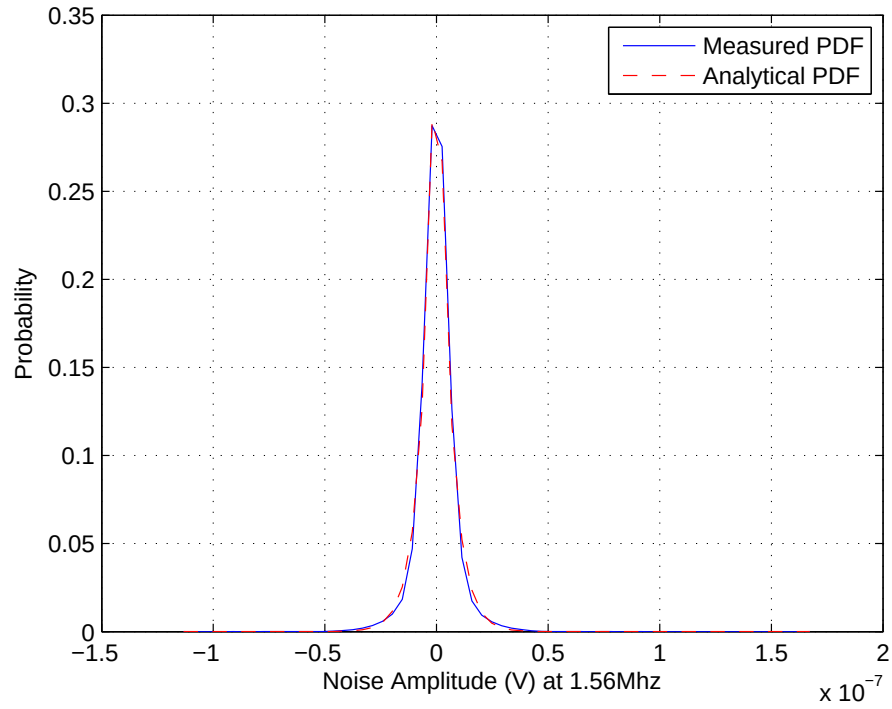


Figure 3.25. Empirical PDF versus Analytical GGD PDF whose MLE parameters are estimated from measured data.

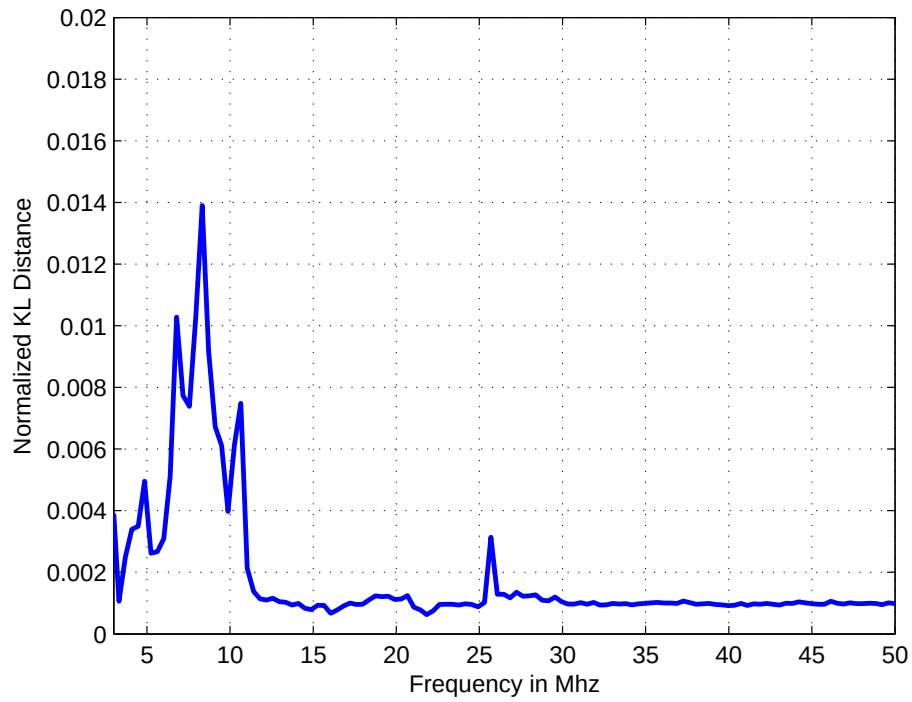


Figure 3.26. Normalized KL distance for each frequency of broadband power-line noise.

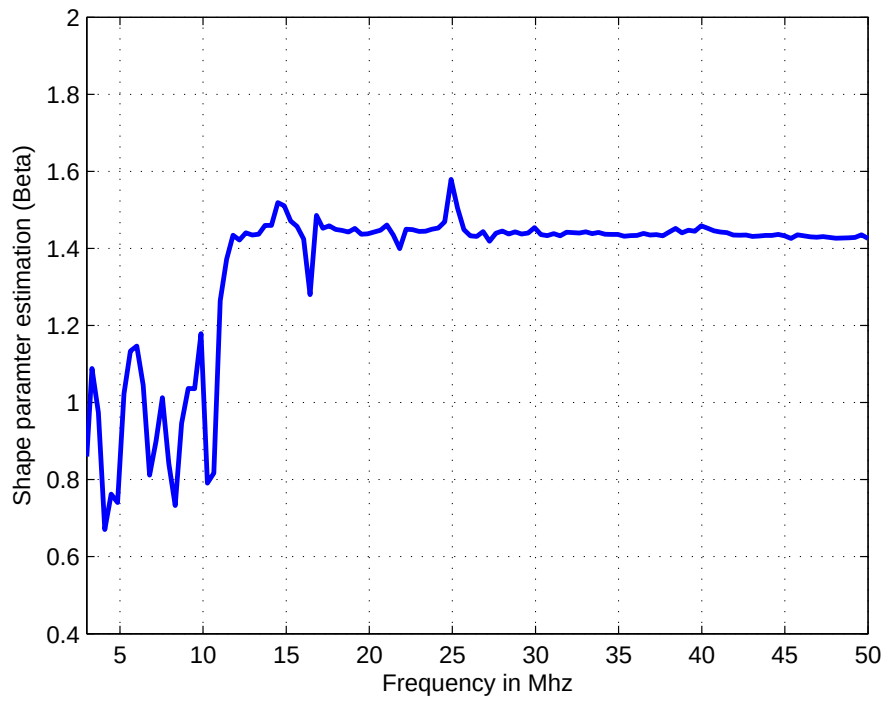


Figure 3.27. Change of shape parameter β over the 50 MHz bandwidth. Distribution is considered "normal distribution" for $\beta = 2$ and Laplacian for $\beta = 1$.

4. TRANSMISSION OVER BROADBAND PLC CHANNELS

Conventional communication systems make the assumption that the noise is additive Gaussian in nature. As we have shown in Chapter 3, shape of the noise amplitude distribution in PLC is impulsive below 12 MHz, and it is closer to Gaussian above in the 12 – 50 MHz band. The performance loss due to this non-Gaussianity can be compensated by utilizing the noise PDF in the design of the receiver. Herein, the noise PDF is used in the orthogonal frequency division multiplexing (OFDM) framework to model the noise amplitude distribution of each sub-carrier using our proposed model, GGD.

4.1. GGD Based Viterbi Decoder

4.1.1. System Model

Consider the binary communication system in the presence of a frequency selective channel determined by Equation 2.9 and additive GGD noise shown in Figure 4.1.

4.1.2. Transmitter

The information bits $d(n) \in \{0, 1\}$ are emitted by a source with equal probability and encoded by a rate-1/3 convolutional encoder to produce the coded bits $c(n)$. The coded bits $c(n)$ are then mapped to equiprobable ± 1 binary phase shift-keyed (BPSK) symbols. Inverse DFT (IDFT) of the BPSK symbols are taken and a cyclic-prefix of length N_{CP} is added to construct the transmitted symbols, $s(n)$, which pass through the power-line channel $h(n)$. Hence, $s(n)$ is defined as

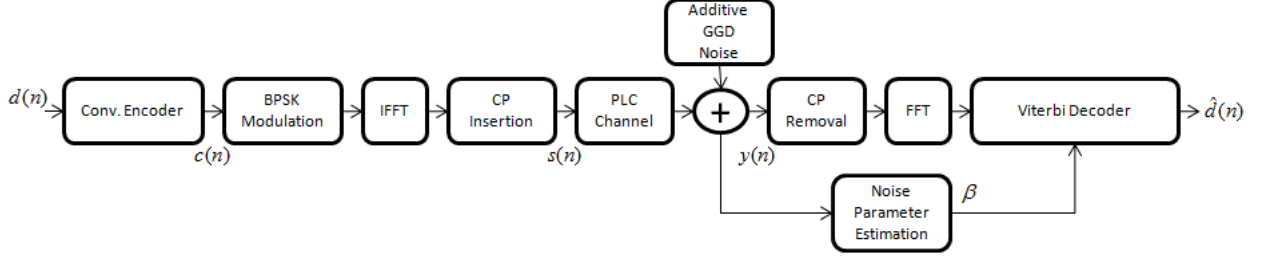


Figure 4.1. Proposed OFDM based receiver structure.

$$s(n) = \frac{1}{\sqrt{N_{FFT}}} \sum_{k=0}^{N_{FFT}-1} S(k) e^{\frac{j2\pi}{N_{FFT}} kn} \quad (4.1)$$

where $-N_{cp} < n < N_{FFT} - 1$, k is the frequency index, N_{FFT} is the FFT size used for each OFDM symbol and $N_{CP} = N_{FFT}/4$. This signal is transmitted through the PLC channel as in Figure 4.1.

4.1.3. Channel

The channel used in these simulation is generated by the power-line channel simulator developed in the previous chapter. 50, out of which 20 taps are non-zero, channel taps are generated using the simulator. The RMS delay spread of the channel corresponds to the delay spread of a 4-floor building.

4.1.4. PLC non-Gaussian Additive Noise Model

An additive non-white PLC noise $w(n)$ is also added to the degraded signal. The noise generated here is of generalized Gaussian distributed nature. In accordance with the measurements, the shape parameter, β , of the noise is varied from 0.8 to 2, which corresponds to Gaussian distribution. The noise shape parameter is started from 0.8 because our measurements show that within the frequency region of interest, the shape parameters varies around 0.8 on average.

4.1.5. Proposed Decoder

The received signal $y(n)$ degraded by the channel $h(n)$ and the additive white power-line noise $w(n)$ is given

$$y(n) = s(n) * h(n) + w(n). \quad (4.2)$$

After removing the cyclic-prefix and taking the DFT at the receiver, we have

$$Y(k) = \frac{1}{\sqrt{N_{FFT}}} \sum_{n=0}^{N_{FFT}-1} y(n) e^{\frac{-j2\pi}{N_{FFT}} kn} \quad (4.3)$$

where $-L_c + 1 < n < N_{FFT} - 1$, where L_c is the length of the PLC channel. Here, assuming the signal $S(k)$ is passed through the fading and noisy channel $H(k)$, $Y(k)$ is received as

$$Y(k) = S(k)H(k) + W(k). \quad (4.4)$$

After taking the DFT of the cyclic-prefix removed symbols to obtain $Y(k)$, the symbols are passed on to the Viterbi decoder in which demodulation and decoding of the transmitted symbols are performed. At the output of the Viterbi decoder we obtain $\hat{d}(n)$.

It should be noted that in the Viterbi decoder an Euclidean distance metric can not directly be used for decoding since the noise amplitude distribution at a single frequency is not Gaussian in PLC. Therefore, the GGD based distance metric is derived and employed in the Viterbi decoder. The BPSK symbol $S(k)$ takes the values $+1$ and

-1 with equal probabilities and the noise is assumed to be complex, independent and identically Generalized Gaussian Distributed random variable with zero-mean and variance of N_0 defined as $W(k) = W = W_r + jW_i$. Additionally, $H(k)$ is carried on as $H(k) = H = H_r + jH_i$. Hence, the received signal PDF can be written as

$$f(Y | H) = \frac{\beta}{2\alpha\Gamma(1/\beta)} e^{-(\frac{|Y-SH|}{\alpha})^\beta}, \quad (4.5)$$

where from the definition of the variance for GGD,

$$\frac{N_o}{2} = \frac{\alpha^2\Gamma(2/\beta)}{\Gamma(1/\beta)}. \quad (4.6)$$

Solving Equation 4.6 for α , we have

$$\alpha = \sqrt{\frac{N_o\Gamma(1/\beta)}{2\Gamma(2/\beta)}}. \quad (4.7)$$

For the information symbol $\hat{S}(k)$ at time k given the received sample $Y(k)$, the distance metric becomes,

$$\lambda(\hat{S}(k) | Y(k)) = - | Y(k) - \hat{S}(k)H(k) |^\beta, \quad (4.8)$$

where $\lambda(. | .)$ is the distance metric that is used for computing the branch and path metrics. Note that, since α in Equation 4.5 is a scaling factor, it is omitted in the metric computation.

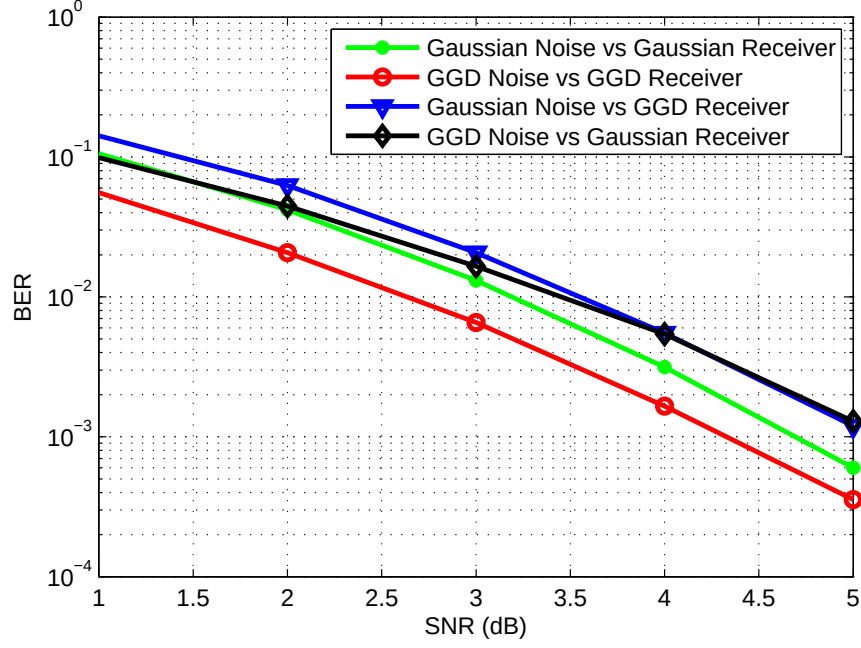


Figure 4.2. Coded OFDM with PLC fading channel, convolutional coding with rate 1/3 and constraint length of 7, number of data bits=85, GGD shape parameter $\beta=1$.

It will be shown that using the GGD based path metric defined in Equation 4.8 in the Viterbi decoder, one can achieve better performance compared to Gaussian receiver against GGD noise.

4.1.6. Numerical Results and Discussion

In this OFDM framework designed for broadband PLC channel, carrier spacing is chosen to be 31.25 KHz based on the coherence bandwidth results in [52]. An FFT size of 256 is used for simulation purposes. The bandwidth in this simulation is 8 MHz.

As seen in Figure 4.2, when each individual frequency is contaminated by Generalized Gaussian distributed noise with a shape parameter $\beta = 1$, which corresponds to Laplace noise, in a multipath fading PLC channel, the GGD receiver performs slightly more than 1 dB better at BER of 1×10^{-2} than the Gaussian receiver.

In Figure 4.3, it is seen that the enhancement on receiver performance is much better as the noise shape parameters becomes smaller. As the shape of the noise am-

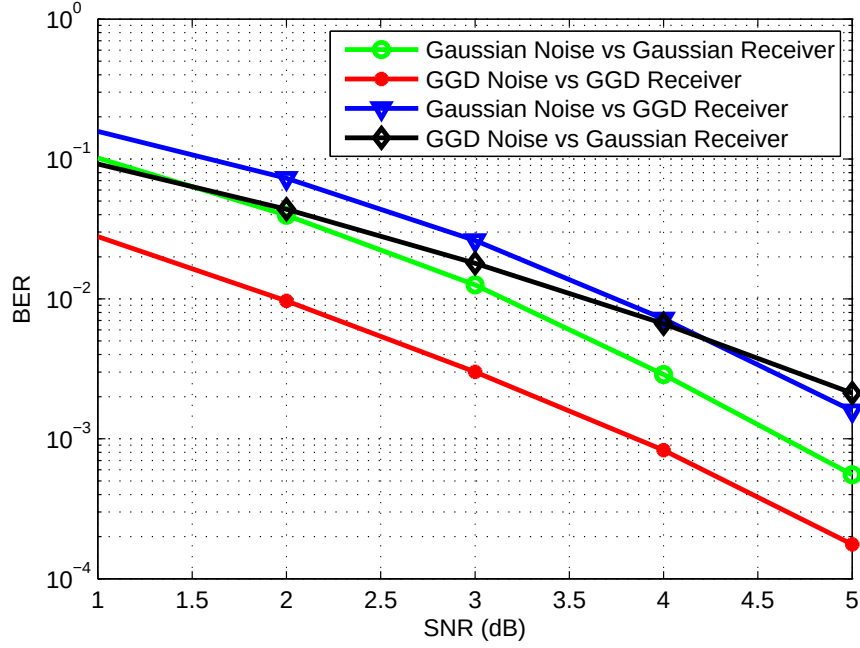


Figure 4.3. Coded OFDM with PLC fading channel, convolutional coding with rate $1/3$ and constraint length of 7, number of data bits=85, GGD shape parameter $\beta=0.8$.

plitude distribution becomes more impulsive, the gap between the performance of a Gaussian receiver and Generalized Gaussian receiver versus PLC noise widens. Parameter estimations in Figure 3.27 show that the shape parameter swings between 0.8 and 1.44 and after a certain frequency it does not significantly change. The shape of the noise amplitude distribution at $\beta = 1.44$ resembles Gaussian distribution which is in coherence with the previous studies. However, performance of the GGD optimal receiver when $\beta = 1.44$ is fairly close to Gaussian optimal receiver's performance as oppose to the results for lower values of the shape parameter β as seen in Figure 4.2 and 4.3. For $\beta = 0.8$, it is seen that the BER difference for a fixed SNR can go up to 2.2 dB. Hence, it makes sense to use this receiver in the low frequency sub-carriers of an OFDM system even if the estimation of noise parameters are not readily available to the receiver when needed.

Therefore, assuming AWGN and using Gaussian optimal receiver over the whole bandwidth may cause a significant loss in BER for a fixed SNR. It is seen that GGD receiver significantly outperforms the Gaussian receiver's performance in terms of BER

for fixed SNR when β shape parameter has low values ($\beta < 1.44$) in a multipath fading PLC channel. However, The shape parameter is only lower than 1.44 up to about 10 Mhz. Beyond this frequency, it is almost fixed to 1.44 and our receiver's BER performance is fairly close to the Gaussian receiver's performance for those frequencies.

Hence, it was shown that the use of β shape parameter information could provide significant enhancement on receiver performance.

4.2. GGD Based Viterbi Decoder with Correlated Noise

The correlated noise W_c can be easily synthesized by filtering a white noise source as depicted in Figure 3.7. The noise shaping filter is described by its transfer function $H_c(z)$ in the z-plane:

$$H_c(z) = \frac{B(z)}{A(z)} = \frac{1 + \sum_{i=1}^m b_i z^{-i}}{1 + \sum_{i=1}^m a_i z^{-i}} \quad (4.9)$$

Following filter coefficients are used in our simulations and the frequency response of this filter is shown in Figure 4.4.

$$b = \{0.049922035, \quad -0.095993537, \quad 0.050612699, \quad -0.004408786\}$$

$$a = \{1, \quad -2.494956002, \quad 2.017265875, \quad -0.522189400\}$$

Welch estimate of the power spectral density of white noise and colored noise generated by the filter coefficients given above is shown in Figure 4.6 and 4.5.

Performance of the Viterbi decoder along with an interleaver is depicted in the Figure 4.7, 4.8. It is seen that the as the lenght of the data block or depth of interleaver is increased, the interleaver brings in a better performance against correlated noise.

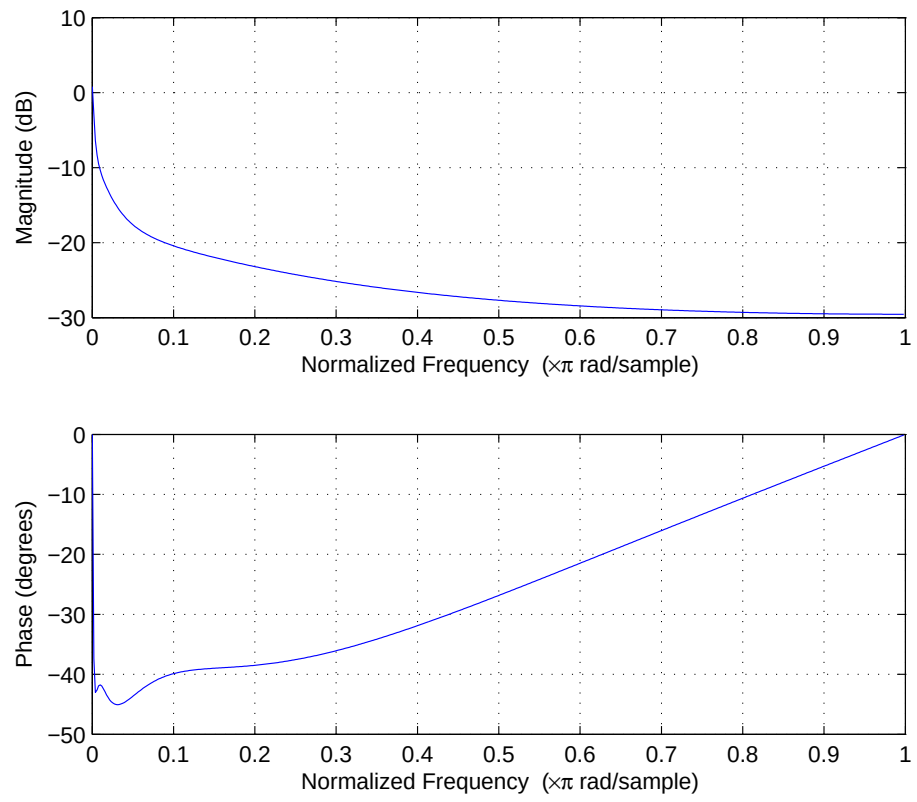


Figure 4.4. Frequency response of the Colored Noise filter, H_c .

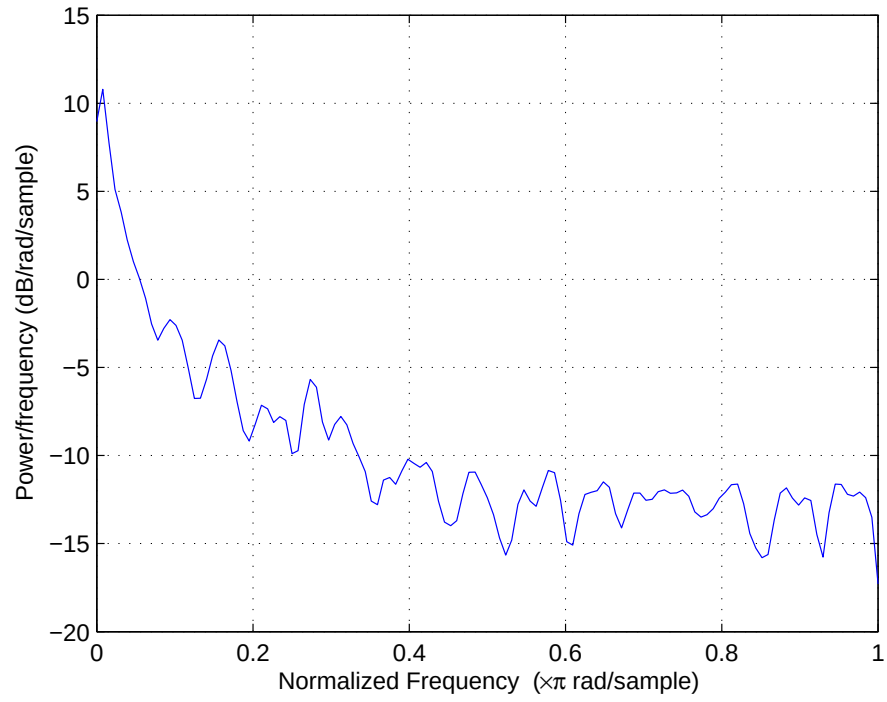


Figure 4.5. Welch estimate of the PSD of noise used to generate the colored Gaussian noise for interleaver simulations.

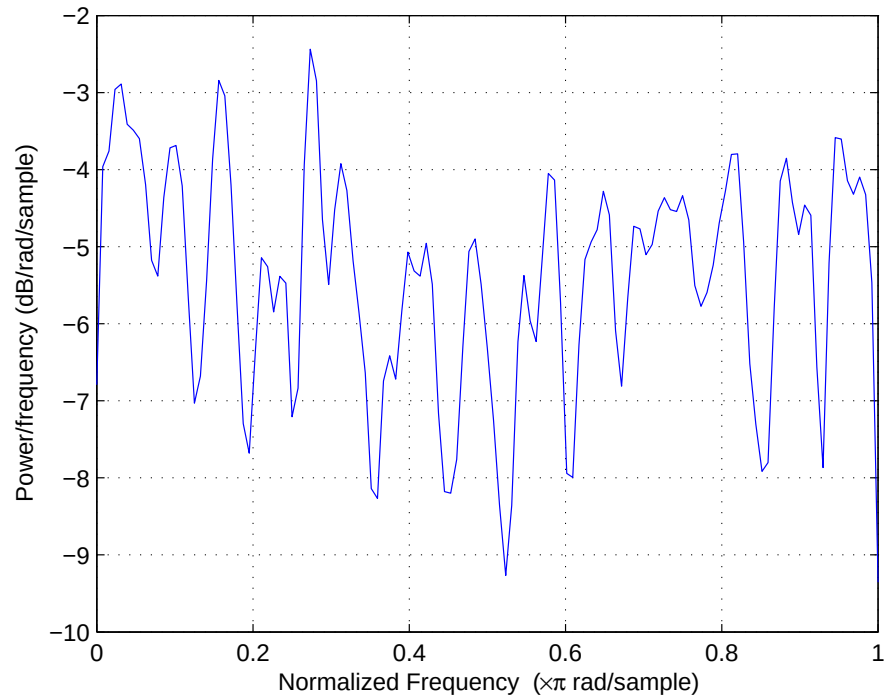


Figure 4.6. Welch estimate of the PSD of noise used to generate the white Gaussian noise for interleaver simulations.

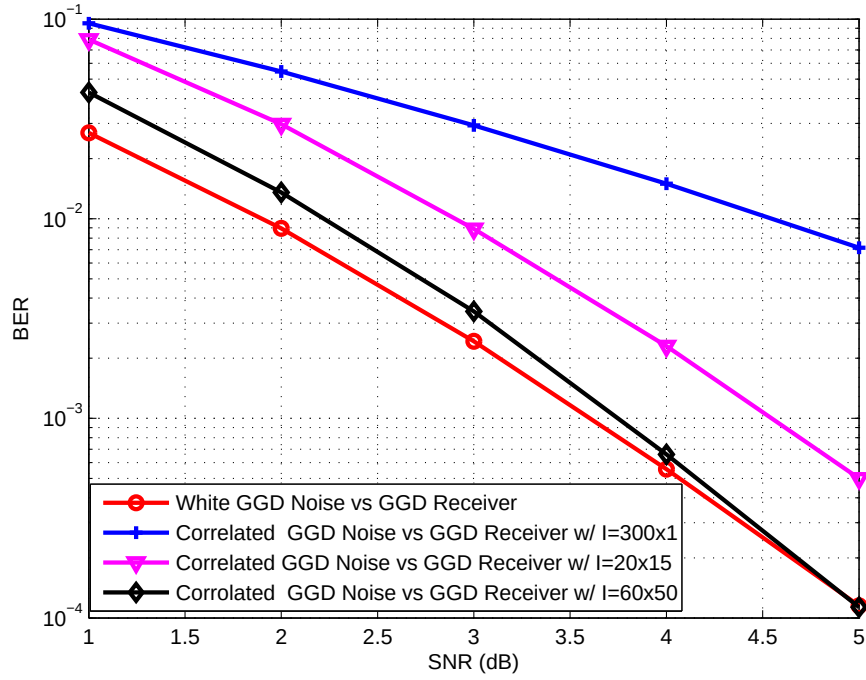


Figure 4.7. Coded OFDM with PLC fading channel, convolutional coding with rate $1/3$ and constraint length of 7, GGD shape parameter $\beta=0.8$.

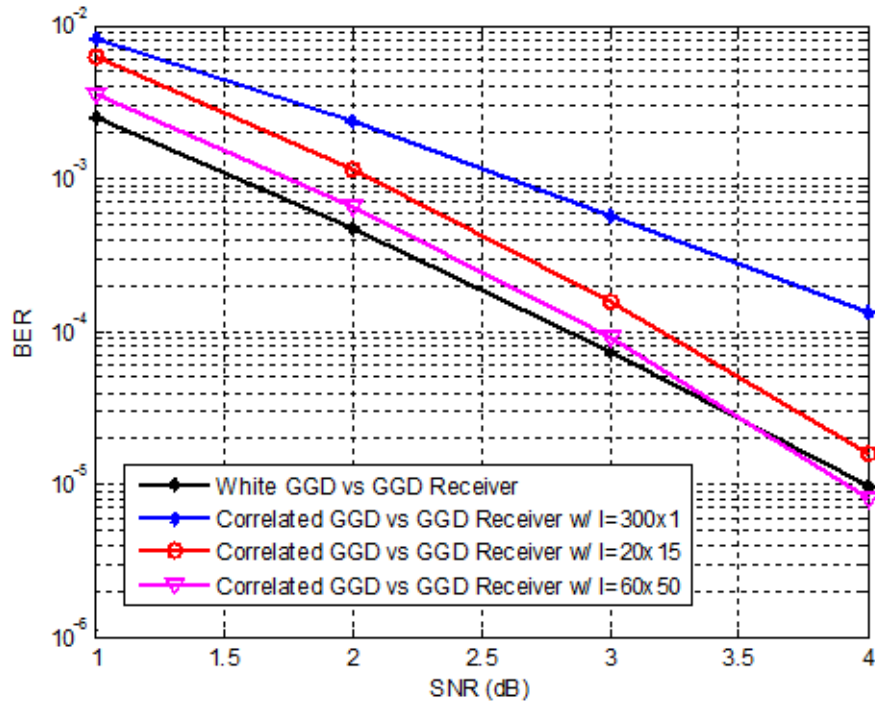


Figure 4.8. Coded OFDM with PLC non-fading channel, convolutional coding with rate $1/3$ and constraint length of 7, GGD shape parameter $\beta=0.8$.

4.3. GGD Based LDPC Decoder

Recently, there has been an interest in using low density parity check (LDPC) codes in power-line communication. It is because the LDPC codes are very robust against large noise bursts. A structured LDPC code can use the memory in the channel to give the best erasure correction performances. Therefore, an interleaver may not be needed if LDPC codes are used for channel coding which is contrary to the case of convolutional coding schemes [53]. Although interleavers are not computationally intensive processing blocks, lack of an interleaver in a system is desired to overcome the problems arising from the delays introduced by interleavers. Therefore, since PLC noise is known to be non-white, it is worth looking into how LDPC codes would perform with our proposed noise model under consideration.

Although LDPC codes were invented by Gallager [54] in 1963, it was not spreadly used until the recent years. This is due to the fact that the processing required for LDPC decoding was way beyond the existing systems' performance at the time. Nowadays, computers can handle such processing heavy tasks in a very timely fashion.

The naming "low density" comes from the definition of the parity check matrix which is always binary and sparse. Most of the entries of parity check matrix H are zero and only a small percentage are 1's. Although there may be some weak constraints to construct the parity check matrix, in general, the parity check matrix is constructed at random subject. LDPC codes are linear codes and can be obtain from sparse bipartite graphs as seen in Figure 4.9. The parity check matrix that corresponds to the bipartite graph represented in Figure 4.9 is illustrated in Figure 4.10.

Let $\mathcal{G}$ is a graph with n left nodes and r right nodes. Left nodes are called the message nodes represented by v and right nodes are called the check nodes represented by c . Such a graph represents a linear code of block length n and its dimension would be at least $n - r$. As seen in Figure 4.9, codewords are the vectors $(c_1, \dots, c_n)$ for which sum of the all neighboring lines among the message nodes is zero for all check nodes.

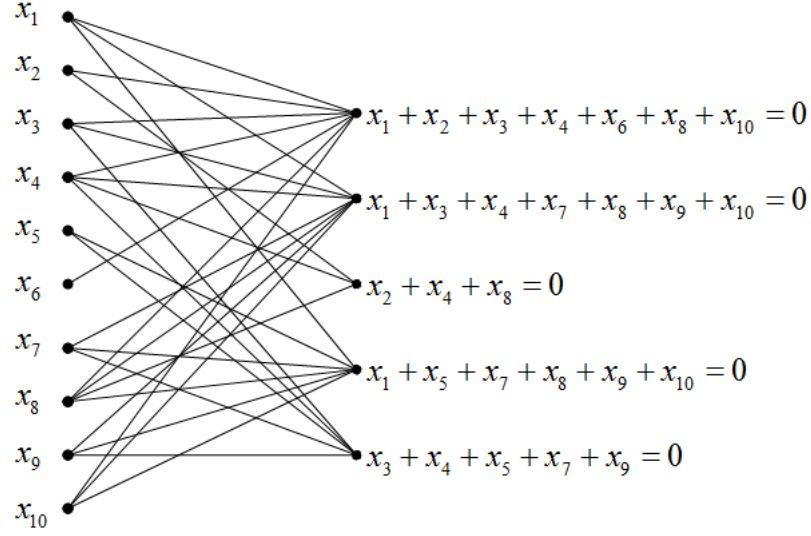


Figure 4.9. An LDPC Code.

$$H = \begin{bmatrix} 1 & 1 & 1 & 1 & 0 & 1 & 0 & 1 & 0 & 1 \\ 1 & 0 & 1 & 1 & 0 & 0 & 1 & 1 & 1 & 0 \\ 0 & 1 & 0 & 1 & 0 & 0 & 0 & 1 & 0 & 0 \\ 1 & 0 & 0 & 0 & 1 & 0 & 1 & 1 & 1 & 1 \\ 0 & 0 & 1 & 1 & 1 & 0 & 1 & 0 & 1 & 0 \end{bmatrix}$$

Figure 4.10. The parity check matrix for the bipartite graph shown in Figure 4.9.

The graph representation can also be expressed in a matrix form by looking at the adjacency matrix of the graph. Suppose H is an $r \times n$ binary matrix in which the entries (i, j) are 1 if and only if the i th check node is connected to the j th message node in the graph. Therefore, the LDPC code that is defined by the graph becomes the set of vectors $(c_1, \dots, c_n)$ such that $Hc^T = 0$. The matrix H , hence, is called a parity check matrix for the code. In terms of linear algebra, the null space of matrix H is the code associated to this graph and any binary $r \times n$ matrix could give a rise to a bipartite graph between the messages and check nodes. However, not every binary linear code has a representation by a sparse bipartite graph. If it does, then the code is called a low density parity check code [55]. The sparsity of the graph structure determines the algorithmic efficiency of LDPC codes.

In this study, we consider a binary regular (w_r, w_c) code is used to construct the parity check matrix H , where w_r and w_c are the row and column weight of H ,

respectively and the commonly known method of probability propagation or sum-product algorithm is used for decoding. Note that a LDPC code is regular if the number of 1's is constant for every column and row. Only difference of the decoding algorithm in our case compared to conventional decoders is the initialization of the Log domain decoding. It should be noted that since Gaussian assumption does not hold here anymore, the GGD probability density function is taken into account while computing the soft values before decoding bits using the LDPC decoder.

The same setup described in Section 4.1.1 is used. Only this time the LDPC encoder and decoder is used for forward error correction instead of convolutional encoder and Viterbi decoder as depicted in Figure 4.11. Since the symbol $S(k)$ takes the values $+1$ and -1 with equal probabilities, the problem can be modeled as the following hypothesis testing problem,

$$\begin{aligned}\Pi_1 : Y &= -H + W, \\ \Pi_0 : Y &= H + W.\end{aligned}\tag{4.10}$$

Therefore, the log-likelihood ratio (LLR) based on Equation 4.5 can be written as

$$\begin{aligned}\gamma &= \ln \frac{f(Y | \Pi_1, H)}{f(Y | \Pi_0, H)} = \frac{|Y_r + H_r|^\beta - |Y_r - H_r|^\beta}{\alpha^\beta} \\ &\quad + \frac{|Y_i + H_i|^\beta - |Y_i - H_i|^\beta}{\alpha^\beta}.\end{aligned}\tag{4.11}$$

From the definition of α in Equation 4.7, the LLR becomes,

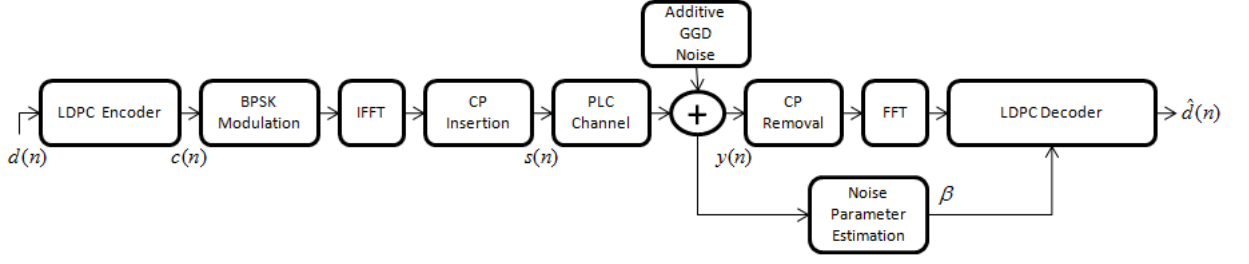


Figure 4.11. Proposed OFDM based receiver structure with LDPC coding.

$$\gamma = \frac{|Y_r + H_r|^\beta - |Y_r - H_r|^\beta}{\sqrt{\frac{N_o}{2} \frac{\Gamma(1/\beta)}{\Gamma(2/\beta)}}} + \frac{|Y_i + H_i|^\beta - |Y_i - H_i|^\beta}{\sqrt{\frac{N_o}{2} \frac{\Gamma(1/\beta)}{\Gamma(2/\beta)}}} \quad (4.12)$$

The LLR, γ , is used in the LDPC decoder to compute the soft values and the message passing algorithm is used for decoding in the LDPC decoder. At the output of the LDPC decoder we obtain the estimate $\hat{d}(n)$ of the transmitted information.

One important subclass of message passing algorithms is the belief propagation algorithm. This algorithm is present in Gallager's work [54]. The messages passed along the edges in this algorithm are probabilities, or beliefs. More precisely, the message passed from a message node v to a check node c is the probability that v has a certain value given the observed value of that message node, and all the values communicated to v in the prior round from check nodes incident to v other than c . On the other hand, the message passed from c to v is the probability that v has a certain value given all the messages passed to c in the previous round from message nodes other than v .

if d is a equiprobable random variable, then $L(d|y) = L(y|d)$ by Bayes's rule. Therefore, if $y_1, y_2, \dots, y_N$ are independent random variables, then we have

$$\ln L(d|y_1, \dots, y_N) = \sum_{i=1}^N \ln L(d|y_i). \quad (4.13)$$

Now suppose that $d_1, \dots, d_M$ are binary random variables, $y_1, \dots, y_M$ are random variables and denote addition over F_2 by $\oplus$. $\ln L(d_1 \oplus \dots \oplus d_M|y_1, \dots, y_M)$ can be computed as the following

$$\ln L(d_1 \oplus \dots \oplus d_M|y_1, \dots, y_M) = \ln \frac{1 + (\prod_{i=1}^M \tanh(M_i/2))}{1 - (\prod_{i=1}^M \tanh(M_i/2))}, \quad (4.14)$$

where $M_i = \ln L(d_i|y_i)$. The belief propagation algorithm is derived from Equation 4.13 and 4.14. In round 0, the check nodes send along all the outgoing edges their log-likelihoods conditioned on their observed value. For example, if the channel used is the binary symmetric channel (BSC) with error probability p , then the first message sent to all the check nodes adjacent to a message node is $(\ln(1-p) - \ln p)$ if the node's value is zero, and it is the negative of this value if the node's value is one. In all the subsequent rounds of the algorithm a check node c sends to an adjacent message node v a likelihood according to Equation 4.14. A message node v sends to the check node c its log-likelihood conditioned on its observed value and on the incoming log-likelihoods from adjacent check nodes other than c using the relation Equation 4.13.

4.3.1. Numerical Results and Discussion

For simulation purposes, a $(3, 6)$ regular parity check matrix H of size 540×1080 is chosen. Firstly, the impact of using our proposed noise model with an LDPC decoder is investigated. In Figure 4.12 and 4.13, it is seen that using the GGD assumption during the initialization of the decoder makes a significant difference when the noise is actually GGD and performance of the Gaussian receiver degrades against the GGD

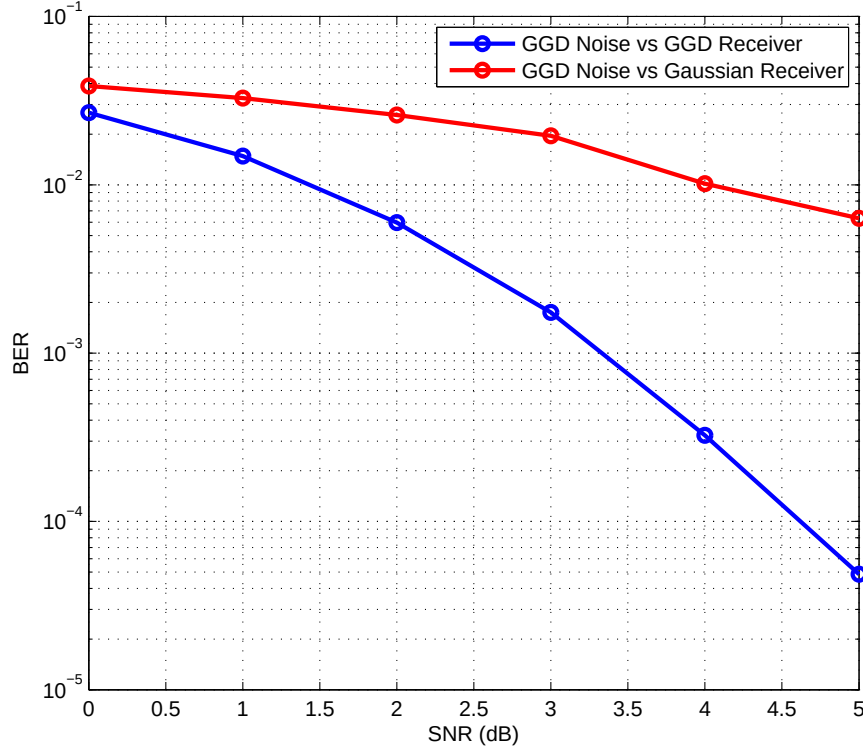


Figure 4.12. BER using the proposed LDPC decoder in a non-fading channel with GGD noise shape parameter $\beta=0.8$, where size of the parity check matrix H is 540×480 , maximum iteration number is 20, maximum block error number is 40.

receiver. Like in the case of the Viterbi decoder, the receiver performance gets better as the beta shape parameter gets smaller, meaning the shape of the noise distribution gets more closer to Laplacian.

We investigated the performance of the LDPC decoder with the correlated noise as well. As in the case of Viterbi decoding, using GGD noise model in LDPC, a better performance on decoding is obtained. As seen in Figure 4.12, when each individual frequency is contaminated by Generalized Gaussian distributed noise with a shape parameter $\beta = 0.8$, in a non-fading channel, the GGD receiver performs almost 2.5 dB better at BER of 1×10^{-2} than the Gaussian receiver. This gap in performance decreases in the fading channels. The GGD receiver outperforms the Gaussian receiver only up to 1.25 dB at BER of 1×10^{-2} in the case of a PLC multipath fading channel as seen in Figure 4.13.

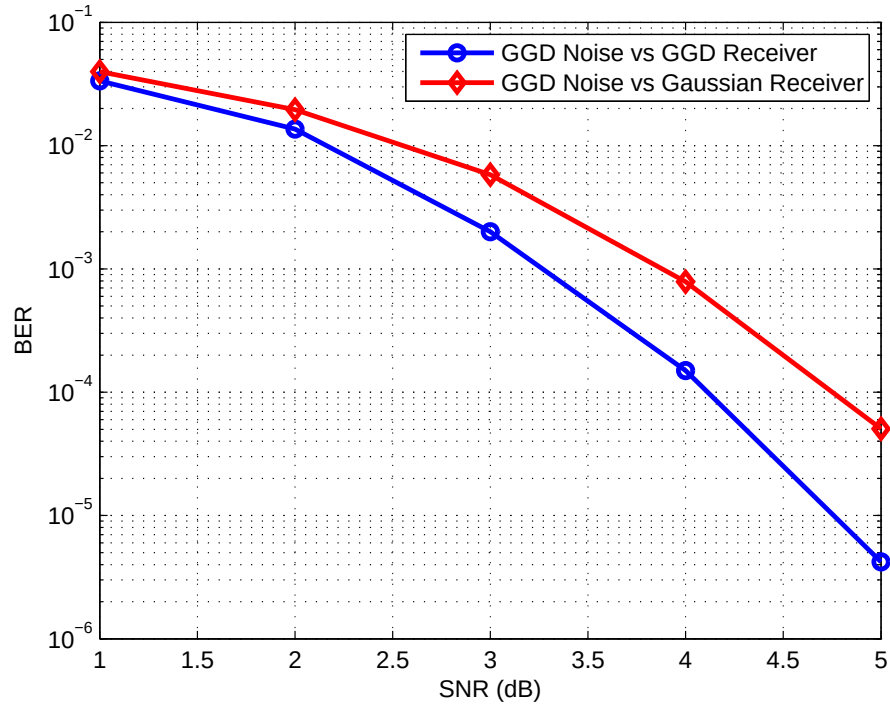


Figure 4.13. BER using the proposed LDPC decoder in a PLC-fading channel with GGD noise shape parameter $\beta=0.8$, where size of the parity check matrix H is 540×1080 , maximum iteration number is 20, maximum block error number is 40.

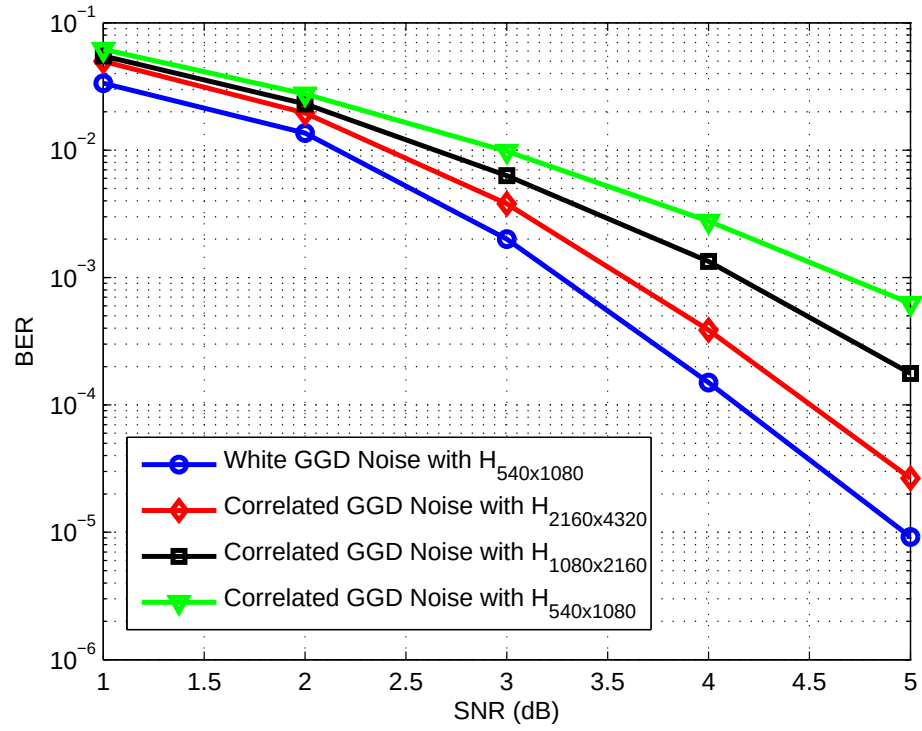


Figure 4.14. Impact of increasing the size of parity check matrix is shown using the proposed LDPC decoder with correlated GGD noise, where the maximum iteration number and the maximum block error number is set to 20 and 40, respectively.

Although LDPC decoder does not have a significant performance gain over convolutional coding, LDPC decoding has an advantage over convolutional coding in the sense that LDPC decoder does not require an interleaver in the system for very long data blocks. It is seen in Figure 4.14 that as the number of bits in a packet grows the parity check matrix also grows and that results in a better decoding performance against correlated noise. This is due to the fact that the sparse parity check matrix gives rise to an uncorrelated bit sequence after encoding. Therefore, it is commonly known that LDPC coded OFDM is better than that of the Turbo coded OFDM on a frequency-selective fading channel [56] and LDPC codes perform as close as possible to the Shannon limit of turbo codes [57].

Finally, to see how the proposed receiver will perform on the recorded noise for a broader bandwidth that is defined in current PLC standards such as HomePlug AV [58], we use the recorded noise as the noise source and the HomePlug AV specification is used to determine the OFDM symbol duration, subcarrier spacing and FFT size. Performance of the Viterbi decoder and LDPC decoder can be seen in Figure 4.15, 4.17 and 4.16, respectively. Here, the utilized bandwidth reaches to 37.5 MHz and the number of subcarriers are chosen to be 1536 for a subcarrier spacing of 24.414 KHz.

It is seen that there is about 1 *dB* gain at 1×10^{-3} and 1.5 *dB* gain at 1×10^{-4} in using LDPC decoder versus Viterbi decoder with the GGD decoding metric derived as part of this thesis study.

Hence, it was shown that the use of β shape parameter information in LDPC decoding yields a significant enhancement on receiver performance and an interleaver may not be required for systems that have long data block length. The simulations ran with recorded noise shows that LDPC receiver slightly outperforms the Viterbi decoder's performance.

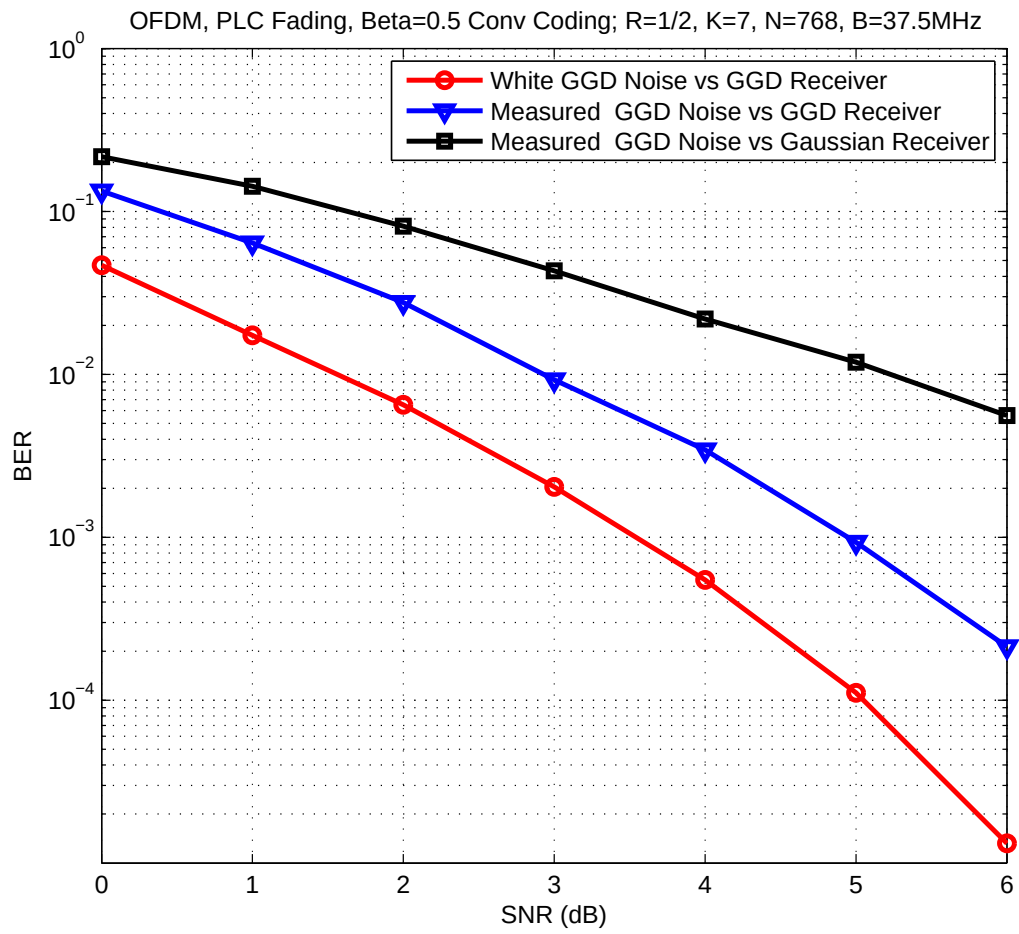


Figure 4.15. GGD based Viterbi decoder simulation is run with Homeplug AV system description and real measurements are used while adding noise to the received signal.

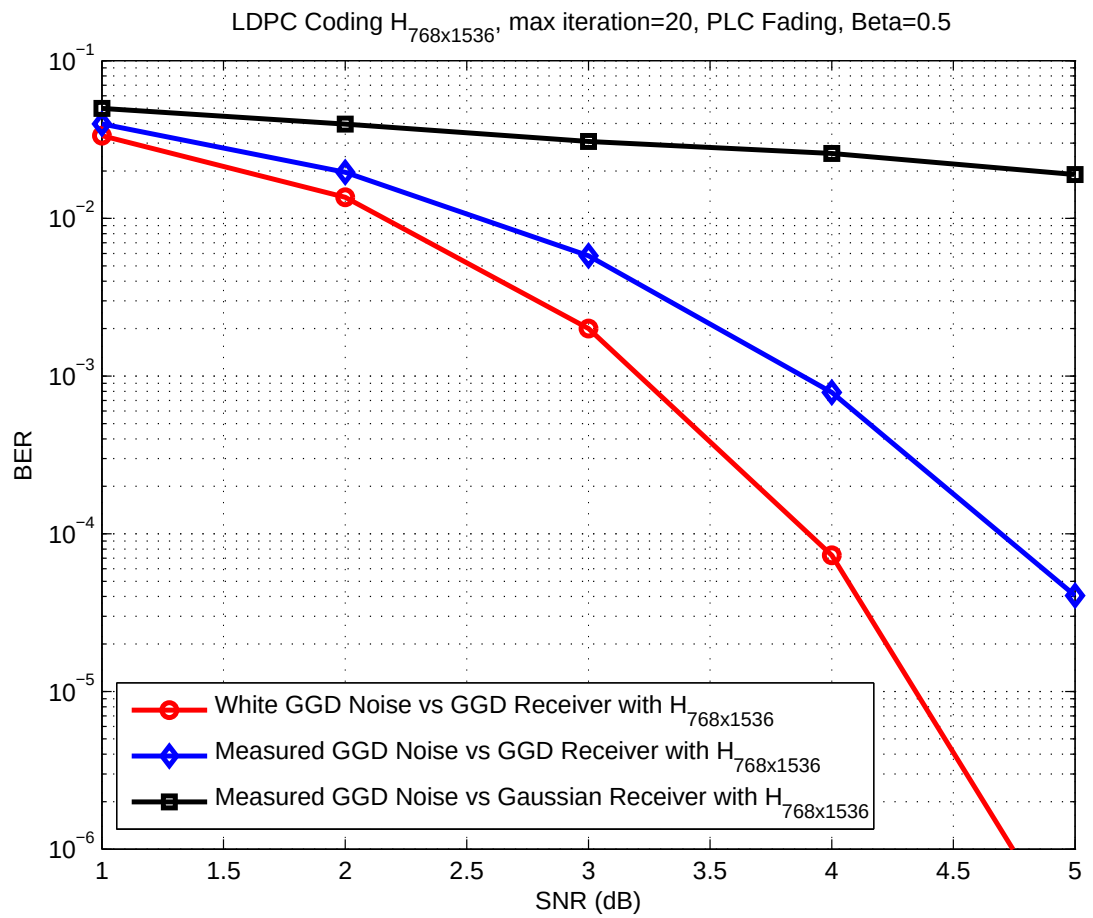


Figure 4.16. GGD based LDPC decoder simulation is run with Homeplug AV system description and real measurements are used while adding noise to the received signal.

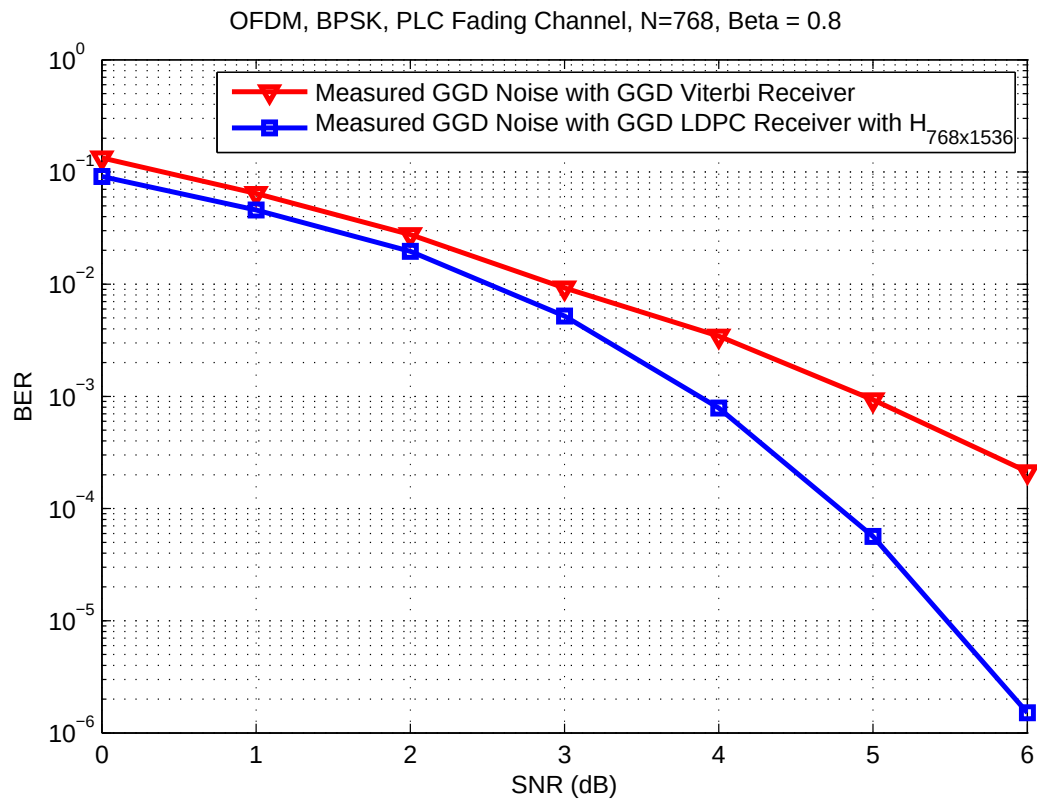


Figure 4.17. Performance of the GGD based LDPC and Viterbi decoder with measured noise. Homeplug Plug AV system parameters are used during simulations.

5. CONCLUSION

Broadband power-line communication is received a great amount of attention lately as the Smart Grid concepts have become popular. However, high speed PLC systems require a very good understanding of noise, attenuation and multi-path characteristics of the power-line channel. In this thesis study, all of the aspects of power-line communication are studied.

In Chapter 2, Multi-path and attenuation characteristics of PLC channels were investigated. Reflection and transmission coefficients were derived. Multipath analysis for a T network is given. In order to have a better understanding on attenuation, a reliable measurement setup is established and the attenuation profile, namely ULAP, has been developed. The measurements have shown that the proposed ULAP model is in well accordance with the measurements.

A matrix based PLC multipath simulation environment was introduced. A unique path selection algorithm was proposed. Channel transfer functions of several simple networks are analyzed with relevant discussions. Effect of

- (i) direct line length between transmitter and receiver,
- (ii) length of the branches connected to the PLN,
- (iii) number of the branches connected to a single node along the direct line,
- (iv) impedance values seen at termination nodes,

were investigated. Important relations between the above mentioned physical attributes and the channel transfer functions are drawn based upon the simulation results. The power-line channel simulation results are compared with channel frequency response measurements and again, it was shown that the simulations are in well accordance with frequency response measurements.

Besides the investigation of some specific PLNs, impact of the physical characteristics of PLC network on the channel delay spread parameters, namely RMS delay spread and maximum excess delay, are studied statistically. Effect of

- (i) number of branching nodes between transmitter and receiver
- (ii) separation distance between transmitter and receiver
- (iii) length statistics of branches

are analyzed. For each attribute examined, statistics regarding the channel delay spread parameters are presented by observing corresponding cumulative distribution function (CDF) curves. Relations between the attributes listed above and the channel delay parameters are revealed.

In Chapter 3, in order to have a better understanding of noise generated by electrical devices, two novel and reliable measurement setups are established. The effectiveness of the measurement setups are shown. Several electrical home appliances are analyzed and the most significant noise sources have been identified. Their features both in time and frequency are extracted. Peak excursion and maximum power change of each device has been identified.

More prominently, based on the noise measurements, a suitable distribution, namely Generalized Gaussian Distribution, for the amplitude distribution of each individual frequency is proposed to model the noise in frequency domain in PLC. Parameters of this distribution are studied to see how they change across the spectrum and it is concluded that above a certain frequency depending on the location of the measurement, the noise amplitude distribution of each frequency tends towards the Gaussian distribution. It is also seen that below this frequency band, the distribution shape converges to Laplacian.

In Chapter 4, using the proposed noise PDF in Chapter 3, a novel OFDM receiver with a GGD decoding metric is proposed. It is seen that the Generalized Gaussian Distribution based receiver significantly outperforms the Gaussian receivers performance

in terms of BER for fixed a SNR in a multipath fading PLC channel. The LDPC based decoder is compared with the Viterbi based decoder with both utilizing the GGD decoding metric and it is seen that GGD based LDPC decoder outperforms the performance of the Viterbi decoder.

This thesis can be considered as a basis for future studies in this field. Future works are planned as

- (i) investigation of noise, multipath, and attenuation characteristics of power line channels for higher frequency ranges up to 100 MHz,
- (ii) improvement of multipath simulation environment by introducing the AC cycle and frequency dependent impedance values,
- (iii) improving the design of the optimal OFDM based PLC receiver by introducing a method to estimate the β shape parameter given very short amount of noise and computing confidence intervals and bounds on the parameters of the proposed noise PDF.

APPENDIX A: PUBLICATIONS RELATED TO THESIS

The following papers have been published/submitted for publication from the work done in this thesis.

Chapter 2

- Guzelgoz, S. and H.B. Celebi and T. Guzel and H. Arslan and M. K. Mihcak, “Time Frequency Analysis of Noise Generated by Electrical Loads in PLC,” *IEEE International Conference on Telecommunications*, vol. 17, pp. 864-871, April 2010.

Chapter 3

- Celebi, H.B. and S. Guzelgoz and T. Guzel and H. Arslan and M. K. Mihcak, “Noise and Channel Statistics of Indoor Power Line Networks,” *IEEE International Conference on Telecommunications*, vol. 18, pp. 523-527, March 2011.

Chapter 5

- Guzel, T. and E. Ustunel and H. B. Celebi and H. Delic and M. K. Mihcak, “Noise Modeling and OFDM Receiver Design in Power-Line Communication,” *IEEE Transactions on Power Delivery*, (Accepted for publication and currently in print process.), July 2011.

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