MASTER

Powerline carrier communications via low voltage networks

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Powerline carrier communications
via low voltage networks

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Abstract

This graduation report describes the research project powerline carrier (PLC) communications in the Netherlands which is a co-operation between the Radiocommunication Group of the Telecommunication Technology and Electromagnetics department of the Faculty of Electrical Engineering of the Eindhoven University of Technology and NV PNEMfacilitair bedrijf.

The objective of the underlying graduation work is to investigate the possible applications of PLC communications and to calculate the performance of a PLC channel based on measurements from existing literature. There are two areas of interest for PLC communications applications: indoor and outdoor. The difference between these two areas regarding applications and performance is large, so we handle them separate.

In order to calculate the performance of a PLC channel we have to quantify the four most important parameters for the PLC communication channel [16]:

- attenuation
- phase distortion
- noise
- available bandwidth

The maximum allowed transmitting power for PLC communications is, in our case, limited by the general emission standard.

After we characterized the channel parameters and taking the general emission standard as limit we were able to:

- calculate the performance of a basic PLC communication system using a simple PSK scheme,
- compare the performance between indoor and outdoor PLC communication systems
- investigate the performance improvements of the PLC communication systems when using regenerative repeaters, QPSK, spread spectrum modulating techniques and error-correction coding.

We also looked at another solution that could improve the PLC system performance a conditioned network.

Furthermore we considered the problems caused by wideband noise, its origin and anticipated problems.
1 Introduction

1.1 Research goal

Our research goal was to investigate the possibilities for PLC (Power Line Carrier) communications in general and its application in the Netherlands. The research project is a co-operation between the Radiocommunication Group of the Telecommunication Technology and Electromagnetics department of the Faculty of Electrical Engineering of the Eindhoven University of Technology and NV PNEM facilitair bedrijf. It may be expected that the impairments of powerline communication systems are similar to those of mobile radio systems; both operate over a metropolitan area, exhibit deep fades, show large differences in levels of received power and are affected by unpredictable man-made interference.

There are two areas of interest for PLC communications applications: indoor and outdoor. The external powergrid, used for outdoor PLC communications, can be subdivided into three different parts; the high, medium and the low voltage part. We will focus our attention on the low voltage (<10kV) part for the following reasons:

- This part of the network is most widely branched. Hence, it is economically interesting to use this already available net instead of laying new transmission lines.
- The relatively short distances between the transmitter and the receiver on the low voltage part of the powergrid may be able to compensate the large attenuation for high frequencies on a powerline.
- Powerline utilities have already some optical fiberglass / coaxial communication cables in the high voltage grid, therefor PLC communications is redundant in these areas.
- The bandwidth for PLC communications is low compared to coaxial or optical fiberglass transmission lines. So a limited number of users can simultaneously communicate over the powerline. Because of the tree structure of the powergrid, as shown in Figure 1, the number of simultaneous user increases when approaching the top of the tree. Using the low voltage part of the net for PLC communications will give each user the most bandwidth.

Figure 1: Tree structure of powergrid
1.2 Approach

To investigate the possibilities for PLC communications we will look at some of the possible applications and calculated the performance of a PLC communication system. Due to the large amount of literature about PLC measurements we decided to perform a literature research and use the existing measurement results in literature to quantify the most important parameters for the PLC communication channel [16]. These parameter values will be used to calculate the performance of a PLC communication system.

Most European researchers constrain themselves to a small bandwidth because of the existing CENELEC standard (EN 50065-1) for PLC systems which cover the range from 3 kHz – 148.5 kHz. Two problems are encountered when using this frequency band:

- Research has shown that there is a lot of noise in this frequency band [3,5,8].
- The available bandwidth in the separate sub-bands for the EN 50065-1 standard is small (see Section 2.2.2); there is 30 kHz bandwidth (sub-band B) for user PLC applications and 86 kHz bandwidth (sub-band A) for utilities PLC applications. Therefore, it is difficult to overcome the noise with smart coding schemes like spread spectrum and error-correction codes.

Therefore, our approach is to go beyond this 148.5 kHz barrier. For our calculations we will consider the frequency range between 1 MHz – 10 MHz. There are three main reasons for this:

- In this frequency range there are enough measurement results available to quantify the most important parameters for the PLC communication channel. During our literature research there was a lack of available measurements for frequency above 10 MHz, especially in the area of outdoor attenuation.
- Early calculations using the bandwidthB and processing gain PG equations (3.16) and (3.17), showed that a bandwidth of 9 MHz should be enough to use spread spectrum modulation.
- Measurements showed the attenuation in power-lines increases with higher frequencies [3], so frequencies above 10 MHz may cause more problems regarding attenuation.

For the frequency range between 1 MHz – 10 MHz we have to limit ourselves to the general emission standard (EN 50081-1) that covers a range from 0 – 30 MHz. The generic emission standard is much stricter when compared to the CENELEC (EN 50065-1) standard regarding the allowed spectral power of the emission but it offers a much larger bandwidth. This gives us the following advantages:

- The much larger bandwidth, compared to the CENELEC band (EN 50065-1), enables the use of spread spectrum techniques and may have a sufficient processing gain to overcome the narrow bandwidth noise on the powergrid while complying with the generic emission standard.
- The larger bandwidth offers more applications for PLC systems.
- Measurements show that the indoor noise power drops to –90 dBm in the range from 1 MHz, while the worst-case indoor attenuation was 2dB/m in the frequency-range below 20 MHz [3].
- Outdoor noise measurements in the range from 1 MHz to 10 MHz were not found during our literature research, but several studies indicate that the outdoor noise power is equal or less than the measured indoor noise power in this frequency range [3,8,12,13]. Outdoor measurements show attenuation levels of 66dB/m in the frequency-range from 1 MHz to 10 MHz [9].
1.3 Applications

The need for more and more communication media is growing every day. The continuing progress in microelectronics technology leads to a plenitude of discrete and distributed microsystems in a modern household. New applications like home automation using PLC communications become possible [19]: similar to networked computers, the efficiency and functionality of domestic appliances can be increased considerably, if they have the ability to exchange information.

Another growing need for communications is fed by the rising percentage of personal computers that have communications capability. A clear example of this is the enormous growth of the worldwide communication network, the Internet.

Internet communication is based on a packet switched network that is active 24 hours a day. PLC systems are also 24 hours a day connected and have a relative small bandwidth compared to traditional modem communication systems. These properties make PLC systems a good alternative for low bandwidth intensive Internet services such as Email and remote monitoring.

Powerline communications have two areas of application: indoor and outdoor. The difference between these two areas regarding applications and performance is large, so we handle them separate.

1.3.1 Indoor applications

Indoor powerline communication applications can be used to create a smart home. A smart home can be defined as a building equipped with numerous sensors and actuators, e.g. where heating, air-conditioning or illumination is automatically, remotely controlled and supervised.

Each smart device can communicate over a bus (using the existing indoor powerlines) to a central controller. This central controller can read the status of each device and can send some control information to operate each device.

At this moment several home-automation systems that use PLC are under development and some are already on the market.

1.3.2 Outdoor applications

Some outdoor applications for PLC systems could be:

- **Remote meter reading of gas, water, electricity usage by the electricity company**
  The power company can monitor the use of gas, water and electric for billing purposes and can even feedback this information to the customer via an information page on a special TV channel.

- **Load management**
  Because most devices that use power from the powergrid are continuously connected to this grid, load management can be used to make a more efficient use of the available power and to reduce peak use.

- **Remote disconnection of defaulter**
  The power company can do a remote disconnection of customers who didn't pay their bill.

- **Fire and burglary alarm**
  The alarm system can be connected 24 hours a day with the fire department and the police.

- **Alarm system for elderly**
An alarm system that can warn relatives or friends via a PLC system in case of distress.

- **Remote status checking of maintenance sensitive devices.**
  With this system you can do just in time maintenance can be performed, instead of checking every device regularly.

- **E-mail system**
  A system that is connected 24 hours to the Internet for receiving and sending E-mail.

- **Remote monitoring**
  Remote (internet) servers can be monitored 24 hours a day

- **Remote controlled street lights**
  Street lights can be controlled and monitored remotely
2 PLC channel characteristics

2.1 Literature research

To characterize to PLC channel we need at least to quantify the four most important parameters for the powerline communication channel [16]:

• attenuation
• phase shift
• noise
• available bandwidth.

If these parameters are known, then we can compute the powerline channel capacity [16], which represents the maximum rate at which nearly errorless data transmission is theoretically possible.

2.1.1 Attenuation

There are two types of attenuation:

• Coupling loss, due to the impedance mismatch between the transmitter/receiver and the channel.
• Line attenuation

2.1.1.1 Coupling loss

The coupling loss is caused by a difference between the line impedance and the impedance of the receiver/transmitter. In an ideal situation the impedance of the transmitter, powerline and the receiver is the same.

![Line impedance stabilisation network schematic](image)

Figure 2: Line impedance stabilisation network schematic

Early measurements done by Nicholson [2] show that a standard 5-μH LISN (Line Impedance Stabilization Network), when used in its specified 150 kHz to 25 MHz range, is a good representation of the mean commercial power line impedance. Consequently, it is a good choice for standardization among the various laboratories performing conducted interference measurements. The schematic of a LISN is shown in Figure 2. The port on the right, the power source, represents the input section of a commercial power line. Measurements are done at the test-sample port. During the measurements a 50Ω resistor is placed between the monitor-jack and
the ground to represent a matched commercial power line load. The measurements also show that
the power line absolute impedance $|Z|$ increases from $30\Omega - 50\Omega$ in the frequency range between
the 1 MHz - 30 MHz. Using the port between phase-zero is also called differential mode [22]. All
further measurements and calculations are based on driving the PLC network in differential
mode.

More recent measurements performed by Tanaka [3], give also a linear function for the
impedance in the frequency range from 1 MHz – 30 MHz. The absolute impedance $|Z|$ increases
from $25\Omega - 150\Omega$. These exact numbers are confirmed by the measurements from Kwasniok [4].
Although no reports where found with similar measurements in the Netherlands, one can assume
that the results also apply to the powergrid in the Netherlands because the type of powerlines
considered in this reference is very similar to the one in the Netherlands.

Further measurements from Tanaka [3] also indicate that the distance between the transmitter and
the receiver influences the impedance significantly. As soon as the line length becomes
comparable with the wavelength, the line length and the load impedance influence the input
impedance. This means that the attenuation varies greatly with the place of the
transmitter/receiver. These results are confirmed by measurements from Suh [7]. They find signal
maxima and minima occurring at quarter-wavelengths caused by standing wave patterns in high
voltage transmission lines. Low voltage lines are not as long as high voltage lines but it indicates
that, when using high frequencies, we also have to deal with standing wave patterns.

Measurements done by Hooijen [1] indicate that the coupling loss is strongly dependent on the
frequency. The average measured coupling losses range between 15 and 30 dB with a worst-case
loss of 53 dB within the CENELEC EN 50065 frequency range (3 kHz - 148.5 kHz). Although
these measurements are below the frequency range we would like to investigate (1 MHz – 10
MHz), the results combined with the measurements by Tanaka [3] indicate that the coupling loss
will also happen in the frequency range 1 MHz – 30 MHz.

The measurements (from Nicholson, Hooijen and Tanaka) show that the absolute impedance $|Z|$ is
dependent from the frequency and therefore causing coupling losses between the transmitter,
channel and receiver.

The variations of impedance in time were measured by Arzberger [5]. These measurements reveal
that the variance is small and nearly independent from the actual energy load of the network.
These measurements where performed in the CENELEC EN 50065 frequency range. This is only
ture, as indicated by Tanaka [3], in case the line length is NOT comparable with the wavelength.
Higher frequencies, as with the frequency range under investigation (1 MHz – 10 MHz), is likely
to cause more problems combined with changing impedance of the load.

2.1.1.2 Line attenuation
Outdoor measurements done by Hooijen [1], show an average attenuation of 80 dB/km in the low
voltage powergrid, with minimum and maximum values of respectively 60 dB/km and 100
dB/km in urban and rural locations. The length of the measured lines ranges from 200m – 700m.
Signal attenuation levels have been measured of 20 dB over a 725m long line on an industrial
location. The line attenuation was constant in time and frequency. All the measurements took
place in the CENELEC EN 50065 frequency range.

Outdoor measurements by Arzberger [5] in almost the same frequency range (20 – 140 kHz)
show that the attenuation varies from 10 to 30 dB with a maximum at about 90 kHz. An
attenuation increase of 20 dB was measured when connecting a load to the cable. The attenuation is mainly caused by the load or termination of the network and not by the cable itself. Long time measurements revealed a significant impact of changing network load on the communication properties of the low-voltage grid.

Research by Dickinson [9] shows that during practical testing on UK low voltage electricity distribution networks, that for the frequency range from 1 MHz to 10 MHz, the attenuation can be as in the range of 30-90 dB for a network length of 250 meters. This variation in attenuation is caused by reflections created by impedance mismatches between transmitter, cable and receiver. The measured power cables have a pseudo coaxial structure. The results from these measurements do not directly apply to power cables used in the Netherlands because they don't have a coaxial structure. This indicates, however, that attenuation levels of cables used in the Netherlands in the frequency range between 1 MHz to 10 MHz will probably be higher but no measurements are reported yet.

Indoor measurements in the low voltage powergrid, done by Tanaka [3], show that number of connections between two measurement points strongly influence the line attenuation. The measured frequency range was beyond the CENELEC EN 50065 norm, up to 20 MHz. The lengths of the measured lines were 10, 20 and 30 m. The measurements, in the frequency range from 10 kHz to 20 MHz, show that the attenuation is proportional with the frequency. The numerous numbers of graphs show a typical attenuation of 1 dB/m and a worst-case attenuation of 2 dB/m for powerlines below 30 m line length.

2.1.2 Phase shift
To get good results with phase modulating techniques over PLC networks, it is very important that the channel has a linear phase characteristic. Up to now, very little research is done to determine phase response and group delay in PLC networks. Until now only Hooijen [1] has done some measurements to determine if the phase characteristics are linear. A standard telephone line was used as reference. No time varying phase shift was found in the powerline.

2.1.3 Noise

2.1.3.1 Types of noise
An extensive study from Vines [8] shows that interference and noise on residential power distribution circuits can be placed into four categories.

1. Interference synchronous with the power system frequency
2. Noise with a smooth frequency spectrum (white noise)
3. Single-event impulse noise (transients)
4. Non-synchronous periodic interference

2.1.3.1.1 Interference synchronous with the Power Frequency
Switching devices, such as silicon-controlled rectifiers (SCR) and certain power supplies cause this category of interference. An SCR switches when the power voltage crosses a certain value. Since the voltage is cyclic, the SCR switches at 50 Hz or a multiple of 50 Hz and thereby causes interference at 50 Hz and multiples thereof. SCR's are often found in light dimmers, which employ triacs to "switch off" the appliance connected for a fraction of the 50 Hz sinewave. The amplitude of the spectral noise lines depends on the load; the heavier the load the higher is the amplitude of the spectral noise line.
2.1.3.1.2 Noise with a smooth spectrum

Loads on the network often create this noise, which do operate synchronously with the powerline frequency. A typical example is a universal motor. Universal motors are small motors with serial windings and can use a dc or an ac voltage to function. They can be found in numerous household appliances like electrical drills, vacuum cleaners, mixers, blenders etc. Such a motor has brushes that cause current switching at intervals that depend on the speed of the motor. For most practical purposes, this noise can be thought as having a smooth spectrum without stationary spectral lines.

2.1.3.1.3 Single-event impuls noise

Lightning, thermostats and other switching phenomenon cause such impuls noise. Capacitor banks being switched in and out for power-factor correction create also impuls noise. This type of noise disturbs the entire frequency band and can cause problems for wide-band spread spectrum communication systems but only occurs for a small fraction of the time.

2.1.3.1.4 Non-synchronous periodic noise

This type of noise is characterized by line spectra uncorrelated with the 50 Hz power frequency. The mayor source of this type of noise is television sets, but in the last few years, also computer monitors have become an important source. They produce noise spikes at multiples of the horizontal scanning line frequency. 15 kHz for television sets and higher frequencies for computer monitors (80 kHz).

2.1.3.2 Noise power spectrum

The total noise is a superposition of the four types of noise. Research by Arzberger [5] show that the total noise power distribution is non-gaussian in the CENELEC EN 50065 frequency range. Chi-square tests exhibited a notable mismatch. An approximation by a superposition of two variance differing gaussian distributions was found suitable.

2.1.3.3 Measurements

Extensive measurements for determining the noise power spectrum were done by Tanaka [3]. The measurements were done in a large residential apartment and in a research building. The result show that the intensity of the noise power spectrum decreases rapidly with increasing frequencies.

Measurements in the research building at the frequency range from 10 kHz to 100 kHz show large levels in the range from -20 dBm to -40 dBm with a bandwidth of 10kHz. For frequencies between 100 kHz and 1 MHz the noise power has a decay of -40dB/decade. The noise power falls under -90dBm for frequencies above 1 MHz.

Measurements in the residential apartment show that the noise level is lower than that in research buildings. For frequencies above 300 kHz the level drops to -95 dBm with a bandwidth of 1kHz.

High noise power levels are observed in summer in whole frequency range, particularly the levels in summer is 30 – 50 dB higher than in winter at the range 200kHz – 600kHz. Its is considered that usage of the rotating machines such as electric fan and air conditioners increase noise power. With frequencies above 1 MHz the noise power level drops to -90dBm with a bandwidth of 1 kHz in the summer.

Outdoor noise measurements in the range from 1 MHz to 10 MHz were not found during our literature research, but several studies indicate that the outdoor noise power is equal or less than
the measured indoor noise power in this frequency range [3,8,12,13]. For further calculations we will assume the same noise power for outdoor and indoor powerlines.

2.2 Standards

2.2.1 general
The PLC link is nowadays defined to some extend by international or national standards. The most important signaling parameter, which is specified by these standards, is the maximum transmitted power in a specified frequency range.

Recently, some European countries, especially Germany, had their own standards for PLC communications. But now there is a European standard for PLC communications on the low voltage powergrid using frequencies from 3 kHz to 148.5 kHz: The CENELEC EN 50065-1 standard. For higher frequencies one has to comply with the generic emission standard EN 50081-1.

2.2.2 CENELEC EN 50065-1 standard
CENELEC EN 50065-1 defines four communication bands (AB, C and D) and a signalling band as described in Table 1.

<table>
<thead>
<tr>
<th>Band</th>
<th>Frequency range (kHz)</th>
<th>Power output (μV)</th>
<th>User dedication</th>
<th>Protocol</th>
</tr>
</thead>
<tbody>
<tr>
<td>Signaling</td>
<td>3 – 9</td>
<td>Not defined</td>
<td>Utilities</td>
<td>Not defined</td>
</tr>
<tr>
<td>A</td>
<td>9 – 95</td>
<td>134 dB</td>
<td>Utilities</td>
<td>Not defined</td>
</tr>
<tr>
<td>B</td>
<td>95 – 125</td>
<td>116 dB</td>
<td>Private</td>
<td>No access protocol</td>
</tr>
<tr>
<td>C</td>
<td>125 – 140</td>
<td>116 dB</td>
<td>Private</td>
<td>Access protocol</td>
</tr>
<tr>
<td>D</td>
<td>140 - 148.5</td>
<td>116 dB</td>
<td>Private</td>
<td>No access protocol</td>
</tr>
</tbody>
</table>

2.2.3 GENERIC EMISSION STANDARD 50081-1
When using higher frequencies one has to comply with the generic emission standard EN 50081-1. The limits in this standard are much stricter than those in the EN 50065-1 standard regarding power emissions on the AC mains. The power emission limits are shown in Table 2.

<table>
<thead>
<tr>
<th>Frequency range (MHz)</th>
<th>Limits (μV)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.15 – 0.5</td>
<td>66 - 56 dB quasi peak*</td>
</tr>
<tr>
<td></td>
<td>56 - 46 dB average</td>
</tr>
<tr>
<td>0.5 – 5</td>
<td>56 dB quasi peak*</td>
</tr>
<tr>
<td></td>
<td>46 dB average</td>
</tr>
<tr>
<td>5 – 30</td>
<td>60 dB quasi peak*</td>
</tr>
<tr>
<td></td>
<td>50 dB average</td>
</tr>
</tbody>
</table>

* Measured in a bandwidth of 9kHz

Average detectors are intended for, and best used for, continuous signals and are not generally considered suitable for the measurement of impulsive components. Impulsive components can be measured better with a peak detector which follows the signal envelope.
The quasi peak detector is simply a peak detector with an inherently long time constant. This will de-emphasise a low repetition rate peak response indicating a level, which is below the actual peak signal level and dependant upon the duty cycle.

The frequency range we investigate, 1 MHz to 10 MHz, can be seen as large series of impulsive components. Hence, we will use the quasi peak power limit for our calculations.
3 Analysis

3.1 Powerline communication model

3.1.1 Channel data

It is not well possible to create an accurate channel model on the basis of the figures found in Section 2.1; Most parameters are not constant in time and can vary enormously due to a lot of different influences. Also some parameters are not exactly known because we lack some measurements in certain areas. We have to extrapolate to approximate these needed figures. Together with the figures found we can calculate the performance of a PLC communication system.

As stated in Section 1.2 we investigate PLC communications for frequencies above 1 MHz. Furthermore we would like to use spread-spectrum techniques in order to overcome the noise problems and power limiting standards. Hence, we need a large bandwidth to get a proper processing gain. For the development of our channel model, we will consider the frequency range between 1 MHz - 10 MHz. In this frequency range there are enough measurement results available to quantify the most important parameters for the PLC communication channel.

3.1.2 Noise power

As shown in Section 2.1.3.3; for frequencies above 1 MHz only background noise exists. Indoor measurements show that the worst-case noise power was around −85 dBm and on average it was −90 dBm, both calculated with a bandwidth \(B_{\text{measured}}\) of 1 kHz. As explained in Section 2.1.3.3 we will use these values for indoor and outdoor calculations. Research shows that the noise, in frequencies above 1 MHz, now only exists of white noise \[3,8\]. Hence, one can assume that this background noise has a more Gaussian character because it is not a superposition of multiple types of noise anymore, as is the case with lower frequencies. Considering a frequency range from 1 MHz to 10 MHz will give a bandwidth \(B_{\text{used}} = 10 \text{ MHz} - 1 \text{ MHz} = 9 \text{ MHz}\).

The noise power \(N\) can be written as \(N = \eta \cdot B\), where \(\eta\) denotes the single sided power spectral density, whereas \(B\) represents the bandwidth. Using the numbers above, the worst - and average-case noise power can be calculated. Research shows that noise power values vary between worst - and average-case with time intervals of hours \[3,8\].

**Worst case**

\[B_{\text{used}} = 9 \text{ MHz},\]
\[B_{\text{measured}} = 1 \text{ kHz},\]
\[N_{\text{worst case}} = -85 \text{ dBm} + 10 \log (9 \cdot 10^6) - 10 \log (1 \cdot 10^3) = -75.5 \text{ dBW}.\]

**Average case**

\[B_{\text{used}} = 9 \text{ MHz},\]
\[B_{\text{measured}} = 1 \text{ kHz},\]
\[N_{\text{average}} = -90 \text{ dBm} + 10 \log (9 \cdot 10^6) - 10 \log (1 \cdot 10^3) = -80.5 \text{ dBW}.\]
3.1.3 Signal power and optimum signal-to-noise ratio

The maximum signal power is limited by the generic emission standard EN 50081-1 for frequencies above the CENELEC EN 50065-1 band. To use these values we have to convert these values to watt instead of volt. Based on research [2,3,4], we use an absolute impedance \( |Z| \) of 50 Ohm to represent a matched network load in the equations below. The maximum allowed power \( P_{\text{max}} \) can be written as

\[
P_{\text{max}} = \frac{U_{\text{max}}^2}{|Z|},
\]

where \( U_{\text{max}} \) represents the voltage limit. Considering a frequency range from 1 MHz to 10 MHz, the quasi peak voltage limit from the generic emission standard EN 50081-1 is 56 dB\( \mu \)V for the frequency range 1 MHz - 5 MHz and 60 dB\( \mu \)V for the frequency range 5 MHz - 10 MHz. Both quasi peak voltage limits apply to a standard bandwidth of 9 kHz.

To prevent that the complexity of the PLC system will become too high, due to changing signal power levels, a single power level limit for the frequency range from 1 MHz to 10 MHz is used. So for all further calculations, the lower emission bound (56 dB\( \mu \)V) is used. With this given value, the maximum allowed power, in the frequency range can be calculated.

\[
U_{\text{max d-B=9kHz}} = 56 \text{dB } \mu \text{V}
\]

\[
P_{\text{max d-B=9kHz}} = \frac{(6.31 \times 10^4)^2}{50} = 7.96 \times 10^{-9} = -81 \text{dBW},
\]

(3.2)

The voltage limit values of the generic emission standard apply to a standard bandwidth of 9 kHz. Assuming a uniform spread of power we can recalculate the maximum allowed power levels for the bandwidth needed.

\[
P_{\text{max d-B=9kHz}} = \frac{B_{\text{std}}}{{B_{\text{std}}} \times 10^6} \times P_{\text{max d-B=9kHz}} = \frac{10 \log(9 \times 10^6) - 10 \log(9 \times 10^3) - 81 \text{dBW}}{9 \text{kHz}} = -51 \text{dBW},
\]

(3.4)

Without signal attenuation we can get a maximum signal to noise ratio of

\[
\left( \frac{S}{N} \right)_{\text{max worst case}} = -51 + 75.5 = 24.5\text{dB},
\]

\[
\left( \frac{S}{N} \right)_{\text{max average}} = -51 + 80.5 = 29.5\text{dB},
\]

3.1.4 Attenuation

Indoor attenuation is worst-case 2 dB/m and typically 1 dB/m for frequencies below 20 MHz and 30 m cable length (see Section 2.1.1.2).

Outdoor attenuation is worst-case 0.36 dB/m and average-case 0.12 dB/m for frequencies between 1 MHz and 10 MHz (see Section 2.1.1.2). We have to keep in mind that these attenuation levels could be higher for non-coaxial power cables as used in the Netherlands, but no measurements are reported yet.
When using high frequencies, in the range of 1MHz to 10MHz, standing wave patterns may cause serious attenuation, even on short powerlines. This attenuation depends significant on the location of the transmitter and the receiver as stated in Section 2.1.1.1. Research shows that the velocity of propagation on an unloaded distribution line is approximately 94 percent the speed of light in air [7]. In calculations we will use the speed of light for the propagation. Calculation shows that minima and maxima can occur every 15 meters when using a frequency of 5 MHz.

The result is that transmitter and receiver have to be placed at certain optimum distances so that standing wave patterns don't have minima at the receiver's position.

3.2 Improvements

3.2.1 Regenerative repeaters

Extrapolating the attenuation values as shown in Section 2.1.1.2 for longer line lengths will give very high attenuation levels. These high levels make it almost impossible to communicate over a powerline network. To overcome this problem we will first look at a schematic overview of a low voltage network, typical for the Netherlands, as shown in Figure 3.

![Schematic overview of a low voltage network](source: PNEM Facilitair bedrijf)

**Figure 3: Schematic overview of a low voltage network**

Each numbered white-square represents a house, which is attached to one or more (for redundancy) low voltage distributing transformers (M1-M2) via a long low voltage power-cable. The schematic shows that the line distance between the two transformers is approximately 1Km, so distance between two houses is average 1Km / 30 = 33m.
This redundant number of transformers could also be used to create a redundant receiver/transmitter system. Transformer node M1 can then be used as the location for the primary receiver/transmitter. While the backup receiver/transmitter, located at the transformer node M2, takes over when there is a problem with the primary receiver/transmitter. During normal operation there is only one primary receiver/transmitter active together with the receivers/transmitters located in each house.

Using digital regenerative repeaters at each receiver/transmitter can improve the performance enormous compared with a system without repeaters. A block diagram of a repeater system is shown in Figure 4.

![Figure 4: Block diagram of a repeater system](image)

A regenerative repeater consists of a complete receiver and transmitter back-to-back in one package. The receiving part converts incoming message digits and errors occur due to channel noise in each regenerating section. The transmitting part uses these digits and generates a new signal for transmission to the next repeater station. The regenerated signal does not contain any random noise anymore but does contain instead some random digit errors.

### 3.2.1.1 Binary PSK modulation

To analyze the performance for a digital regenerative repeater system we will use a coherent binary PSK modulation scheme.

Let $S$ denote the PSK signal power at the output of the $M$th repeater and $N$ denote the noise power within the received bandwidth $B$ (Hz).

\[ S = \frac{E_b}{T_b} \]

\[ N = N_0 B \]  \hspace{1cm} \text{(3.4)}

where $E_b$ is the energy per bit, $N_0$ is the noise density and $B$ the bandwidth.

Let $\varepsilon$ be the error probability at each repeater, given by

\[ \varepsilon = Q\left(\sqrt{\frac{A^2 T_b}{\eta}}\right) = Q\left(\frac{2E_b}{\sqrt{N_0}}\right) = Q\left(\frac{2T_b B}{N} S\right). \]  \hspace{1cm} \text{(3.5)}

The bit duration $T_b$ - bandwidth $B$ product of an ideal PSK signal is $T_b B = 1$. Hence the signal-to-noise ratio $S/N = \text{ratio of energy per bit to noise density } E_b / N_0$.

If the number of erroneous conversions is even, the error cancels out for binary data. The probability of $n$ errors in $M$ successive conversions is given by the binomial distribution:

\[ P_M(n) = \binom{M}{n} \varepsilon^n (1 - \varepsilon)^{M-n}. \]  \hspace{1cm} \text{(3.6)}
The net error probability in a binary transmission after $M$ repeating sections is then the probability that $n$ is odd:

$$P_e = \sum_{n \text{ odd}} P_m(n) = \binom{M}{1} \varepsilon (1-\varepsilon)^{M-1} + \binom{M}{3} \varepsilon^3 (1-\varepsilon)^{M-3}. \quad (3.7)$$

Provided that the bit error probability at each repeater is $<< 1$.

Hence, the net error probability after $M$ repeating sections, is:

$$P_e = MQ \left\{ \sum_{n \text{ odd}} \frac{2E_b}{N_o} \right\} = MQ \left\{ \sqrt{2T_b \frac{S}{N}} \right\}. \quad (3.8)$$

### 3.2.1.2 Quaternary PSK modulation

With QPSK modulation it is possible to get a higher bitrate while using the same bandwidth. The principle is based on that M-signaling schemes such as QPSK require less bandwidth compared to PSK while using more signaling power.

The average probability of symbol error for a coherent QPSK signal is

$$P_s = 2Q \left( \frac{E_s}{N_o} \right) \quad (3.9)$$

where $E_s/N_o$ = ratio of energy per symbol to noise density

When Gray encoding is employed for symbol mapping and the adjacent phase is selected for the true phase, the symbol error contains only one bit error. Thus for Gray encoding the average probability of coherent QPSK is

$$P_b = \frac{1}{2} P_s = Q \left( \frac{E_s}{N_o} \right) \quad (3.10)$$

Since $E_s = 2E_b$, where $E_b$ is the energy per bit, because there are 2 bits per symbol in QPSK signaling. In term of the input signal-to-noise ratio $S/N$, we have

$$\frac{E_s}{N_o} = \frac{2E_b}{N_o} = T_b \frac{S}{N} \quad (3.11)$$

Thus the average probability of bit error for QPSK is

$$P_b = Q \left( \sqrt{\frac{E_s}{N_o}} \right) = Q \left( \sqrt{\frac{2E_b}{N_o}} \right) = Q \left( \sqrt{T_b \frac{S}{N}} \right) = Q \left( \sqrt{2T_b \frac{S}{N}} \right) \quad (3.12)$$
The bit duration-bandwidth product $T_B$ of an ideal QPSK signal is $T_B = 0.5$. Hence $S/N = 2 E_b/N_0$. Thus, for the same channel bandwidth, twice the bitrate can be achieved with QPSK compared with PSK.

Hence using the QPSK modulation schema in digital repeaters gives the following net error probability after $M$ repeating sections

$$P_e = MQ \left( \sqrt{ \frac{2E_b}{N_0} } \right) = MQ \left( 2T_B \frac{S}{N} \right)$$

(3.13)

where $T_B = 0.5$ for a ideal QPSK signal.

3.2.2 Spread spectrum

The properties of spread spectrum communication technology can be used to overcome narrowband noise interference and to achieve much higher bandwidth efficiency by serving a large group of multiple access users. Both goals can be achieved by spreading the signal over a wide frequency spectrum using a function that is independent of the message and is known to the receiver. A block diagram of a coded Direct Sequence spread spectrum system using coherent PSK as carrier modulation is shown in Figure 5. The resulting PSK carrier at de modulator output is spread by multiplying it by another carrier that has been modulated by a pseudorandom or pseudonoise (PN) sequence represented by the bipolar waveform $p(t)$ with a chip rate $R_p$ that is much larger than the information rate $R_b$. At the receiver, the information is recovered by multiplying the received waveform by a synchronized replica of the PN sequence.

The spreading ratio $R_p/R_b$ is called the Processing Gain $PG$.

$$PG = \frac{R_p}{R_b}$$

(3.14)

A periodic sequence consists of an infinite sequence of plus or minus ones divided up into blocks of length $N$, in which the particular sequence in each block is the same. Such a sequence is said to be pseudorandom if it satisfies the following conditions:

- In every period the number of plus ones differ from the number of minus ones by exactly one. Hence $N$ is an odd number. Thus

$$N_+ - N_- = N$$

$$|N_+ - N_-| = 1$$
• In every period half of the runs of the same sign have length 1, one fourth have length 2, one eighth have length 3, and so forth. Also the number of positive runs equals the number of negative runs.

• If the pseudo-random sequence is shifted by any nonzero number of elements, the resulting sequence will have an (almost) equal number of agreements and disagreements with the original sequence, i.e. the discrete-time autocorrelation function of the sequence equals the sequence length at zero phase and $-1$ otherwise.

Many different classes of periodic PN sequences are known. Among these, the ones that can be generated in shift registers are of the greatest practical importance. The Maximal-Length Linear Shift Register Sequences ($m$ Sequences) are most commonly used because they are easily generated in shift registers with relatively small number of stages. The $m$ sequences all have a length that is defined by

$$N = 2^m - 1 \quad m = 1, 2, 3.$$  \hspace{1cm} (3.15)

They require only $m$ stages of shift registers to generate a $m$ sequence. The bipolar waveform $p(t)$

\[ \begin{array}{c}
\text{"1"} \\
T_b \\
\text{"1"} \\
\text{"0"}
\end{array} \]

is created via a standard maximal length linear feedback shift register.

An example of a PN sequence is shown in Figure 6.

**Figure 6: Example of PN sequence**

The power spectral density of the $m$-sequence waveform $p(t)$ is plotted in Figure 7.

**Figure 7: Power spectral density of an $m$-sequence waveform**
$T_c$ is the chip duration. When spread spectrum is used for Direct Sequence Code Division Multiple Access (DS-CDMA), where multiple users share the same frequency range, the cross-correlation properties of PN sequences are as important as their autocorrelation properties. The cross-correlation ($|R_{uv,max}|$) should be as low as possible to make it possible to allow multiple users share the same frequency range with each their own PN sequences. Examples of these are the Gold, Kasami and Bent sequences that are obtained by combining two m sequences. Table 3 lists the peak values of the peak cross-correlation for DS-CDMA.

Table 3: Properties of sequence sets with a period of $2^m-1$

| Family          | m        | Set size         | $|R_{uv,max}|$            |
|-----------------|----------|------------------|--------------------------|
| Gold            | Odd      | $2^{m+1}$        | $1 + 2^{m+1} + 2^l$      |
| Gold            | 2 (mod 4)| $2^{m+1}$        | $1 + 2^{m+2} + 2^{l+1}$  |
| Kasami (small set) | Even    | $2^m$            | $1 + 2^{m+1}$            |
| Kasami (large set) | Even    | $2^m(2^m+1)$     | $1 + 2^{m+1} + 2^l$      |
| Bent            | 0 (mod 4)| $2^{m+1}$        | $1 + 2^{m+1}$            |

In a sequence synchronous DS-CDMA system the bit duration $T_b=1/R_b$ is chosen to be the period $N/T$, of the addressed PN sequences and $T_c=1/R_p$ is the chip duration. So, the processing gain $PG$ is equal to the number of chips in a period $N$.

$$PG = N = \frac{T_b}{T_c} = \frac{R_p}{R_b}.$$ (3.16)

When PSK is used, the spread bandwidth $B_s$ is common taken to be the null-to-null bandwidth, hence

$$B_s = 2 \cdot R_p.$$ (3.17)

where $R_p$ is the chip rate.

Introducing $E_b/N_0$, which is the overall ratio of the received energy per bit to noise density. In terms of the processing gain $R_p/R_b$ and the signal-to-noise ratio, $E_b/N_0$ is given by

$$\frac{E_b}{N_0} = \frac{R_p}{R_b} \left( \frac{S}{N} \right).$$ (3.18)

To estimate the system performance we will consider sequence-synchronous DS-CDMA for our calculations [14]. The fixed positions of the transmitters and receivers in a PLC communication system should allow the use synchronous CDMA. When assuming that the background noise can be approximated by Additive White Gaussian Noise (AWGN) as proposed in Section 3.1.2 the bit error probability, for sequence-synchronous DS-CDMA system using PSK modulation [14], is

$$P_e = \Phi \left[ \frac{2}{k_p N_{\max}^2 + \left( E_b / N_0 \right)^{-1}} \right]^{1/2}. $$ (3.19)
\[
\rho_{\text{max}} = \left| \frac{R_{u,v,\text{max}}}{R(0)} \right|,
\]
(3.20)
Where

- \( k \) : number of simultaneous users
- \( R(0) \) : autocorrelation
- \( R_{u,v,\text{max}} \) : maximum cross-correlation between the sequence \( u \) and sequence \( v \)
- \( \rho_{\text{max}} \) : normalized maximum cross-correlation

### 3.2.3 Error-correction coding

The two common features of all error-correction codes are structured redundancy and noise averaging. Structured redundancy is a method of inserting extra redundant symbols into the information message. Noise averaging is obtained by making the redundant symbols depend on a span of several information symbols.

The ratio of information bits \( k \) to the total bits in a code word \( n \) is called the code rate \( R \).

\[
R = \frac{k}{n},
\]
(3.21)

\[
R_c = \frac{n}{k} \cdot R_i,
\]
where \( R_c \) is the coded bit rate and \( R_i \) is the information bit rate at the input of the encoder. Practical values for \( R \) range from 1/4 to 7/8.

The coding gain depends on the type of coding, the code rate and the preferred probability of a bit error. A widespread type of forward error correction (FEC) is convolutional coding. This type of codes trades bandwidth efficiency for increased power efficiency. The resulting bandwidth expansion equals the reciprocal of the code rate \( R \), while the increased power gain equals the coding gain. The most well known convolutional decoding technique is the Viterbi maximum likelihood-decoding algorithm. The achievable coding gain for Viterbi can be seen in Figure 8.
Figure 8: Bit error probability of coded coherent PSK

For our performance calculations we will use the bit error probability upper bounds for the viterbi algorithm determined by Meerberg [21].

The minimum Hamming distance $d_f$, using a code rate $R=1/n$, is determined using the following bound [15].

$$d_f \leq n \min_{k \geq 1} \left[ \frac{(K - 1 + k)2^{k+1}}{2^k - 1} \right]$$

(3.22)

Where $d_f$ is at most equal to the largest integer less than equal to the bound.

Meeberg [21] has found the following upper bound of the bit error probability $P_B$ for the Viterbi algorithm as a function of the channel crossover error probability $p$ and the minimum Hamming distance $d_f$:

$$P_B < \Gamma(d_f) \cdot 320 \cdot \frac{1 + 3.2p}{(1 - 16p)^2}$$

(3.23)

considering that,
\[
\Gamma(d_f) = \left(\frac{d_f - 1}{d_f / 2}\right) \cdot 2^{-d_f} = \frac{(d_f - 1)!}{\left(\frac{d_f}{2} - 1\right)! \cdot \left(\frac{d_f}{2}\right)!} \cdot 2^{-d_f}
\]

(3.24)

3.2.4 Conditioned network

NORWEB's research group developed a directional coupling or conditioning unit (CU) which make it possible to create a conditioned network' [11]. The concept behind the conditioned network' is as follows: the communication (high frequency) signals and the electricity distribution (low frequency signals with high frequency interference/noise) signals are both filtered before combining them on a single line. This prevents each type of signals from interfering the other.

The system is based on a conditioning unit, which exist of two filters (one high and one low pass filter element) as shown in Figure 9.

Figure 9: Conditioning Unit

The network port (NP) interconnects to the electricity utilities cut-out and the low voltage electricity distribution network (LVEDN), the communications distribution port (CDP) interconnects to a HF communications transceiver and the electricity distribution port (EDP) interconnects to the customer electricity meter and hence to the customers in-house electricity distribution network. The idea of this conditioning unit is to create a gateway between the utility outdoor powergrid and the indoor powergrid. A low pass filter will filter all disturbances from electrical appliances while communication signals pass through the high pass filter and both filtered signals are combined in the conditioned network.

A conditioned network allows communications between the conditional units, using PLC communications, with less interference caused by electrical appliances. An example of a conditioned network is shown in Figure 10.
It is also possible to create a conditioned indoor powergrid for indoor PLC communications. All devices connected to this powergrid can be retrofitted with an adapter (which has a low pass filter inside) and new devices in the future can have a built-in filter into their powerplug.

NORWEB's research group [11] did some measurements on the performance gain of a conditioned network in the frequency range from 0 to 20MHz. The signal power-output improvement of the conditional network compared to an unconditional network ranges from 17 dBm for low frequencies to 10 dBm for high frequencies.

3.3 Wideband noise

3.3.1 The problem with wideband noise

Even when the wideband noise is at a level lower than the one specified for narrowband spurious level, the amount of spectrum pollution increases dramatically the likelihood of interference to other systems.

Although the use of spread spectrum can reject some of these interferences (see Section 4.3.2.3), wideband noise can cause a significant drop in performance of PLC communication systems.

3.3.2 The origin of wideband noise

There are numerous possible origins of wideband noise [20].

1. It is well know that some very simple modulation schemes like 4-PSK might have very wide modulation sidebands if no counter-measures, such as appropriate baseband filtering, are taken. Officially, appropriate baseband filtering is mandatory, so generally this should not be a problem.

2. The non-linearity of power amplifiers also tend to widen the signal. It generates $3^\text{rd}$, $5^\text{th}$, or $7^\text{th}$ order intermodulation products expanding the signal up to 3, 5, or 7 times the occupied bandwidth.
3. The phase noise of local oscillators is also generating low-level, very wide band noise. It mainly depends on the quality of the local oscillator but sometimes worsened by the fact that the wanted LO frequency is obtained by the multiplication of a lower frequency source, which also increases the relative level of the phase noise.

4. In some cases, thermal noise might also be an issue. A high power amplifier with a very bad noise figure can generate this thermal noise.

3.3.3 Anticipated problems

As computer and microcontroller systems become faster and faster due to an increasing clock frequency, their frequencies are now falling within the frequency span of the recent European Community (EC) electromagnetic compatibility (EMC) standards for radiated interference (30 MHz to 1 GHz). As more spectral energy is held within the lower order harmonics of a square wave signal, screening the equipment to prevent radiated emissions becomes increasingly more difficult. So some proposals have been made to use Spread Spectrum Modulation of the system clock [10]. When this is implemented in future computer systems it could cause a mayor wideband noise pollution.
4 Performance Evaluation

4.1 Channel capacity

Following the Shannon-Hartley theorem the capacity $C$ of a channel with bandwidth $B$ and additive Gaussian bandlimited white noise is

$$C = B \log_2(1 + S/N) \text{ bps},$$

(4.1)

where $S$ and $N$ are the average signal power and noise power, respectively, at the output of the channel. This theoretical channel capacity gives the upper limit that can be reached in the way of reliable data transmission rate over Gaussian channels in an ideal system. In practice this ideal channel capacity can not be reached, but a system designer always tries to optimize his system to have a data rate as close to $C$, given in (4.1), as possible with an acceptable error rate.

Although the PLC channel has not only Gaussian bandlimited white noise (see Section 3.1.2), it has been shown that the result obtained for the Gaussian channel provides a bound on the performance of a system operating over a non-Gaussian channel [16].

4.1.1 Indoor

The signal-to-noise ratio $S/N$ for indoor communications can be determined using the values from Section 3.1.3 and the attenuation levels found in Section 2.1.1. The high attenuation levels for indoor cables, in the frequency range from 1-10MHz, make it difficult to cover more than 10 meters without repeating the signal [3]. So we choose a cable length of 10m for our calculations.

$$\left( \frac{S}{N} \right)_{\text{worst case indoor 10m}} = \left( \frac{S}{N} \right)_{\text{max worst case indoor 10m}} - L_{dB} = 24.5 \text{ dB} - 20 \text{ dB} = 4.5 \text{ dB}$$

$$\left( \frac{S}{N} \right)_{\text{average indoor 10m}} = \left( \frac{S}{N} \right)_{\text{max average indoor 10m}} - L_{dB} = 29.5 \text{ dB} - 10 \text{ dB} = 19.5 \text{ dB}$$

Using (4.1) the channel capacity for indoor PLC communications, with frequencies between 1-10 MHz and a cable length of 10m, is

$$C_{\text{worst case indoor 10m}} = 9 \cdot 10^4 \log_2\left(1 + 10^{0.45}\right) = 17.4 \text{ Mbps},$$

$$C_{\text{average indoor 10m}} = 9 \cdot 10^4 \log_2\left(1 + 10^{1.95}\right) = 58.4 \text{ Mbps}.$$
4.1.2 Outdoor

The signal-to-noise ratio $S/N$ for outdoor communications can be determined using the values from Section 3.1.3 and the attenuation levels found in Section 2.1.1. A typical outdoor cable length in a low voltage network is 33m (see Section 3.2.1).

$$\left( \frac{S}{N} \right)_{\text{worst case outdoor 33m}} = \left( \frac{S}{N} \right)_{\text{max worst case outdoor 33m}} - L_{\text{dB worst case outdoor 33m}} = 24.5 \text{ dB} - 11.9 \text{ dB} = 12.6 \text{ dB}$$

$$\left( \frac{S}{N} \right)_{\text{average outdoor 33m}} = \left( \frac{S}{N} \right)_{\text{max average outdoor 33m}} - L_{\text{dB average outdoor 33m}} = 29.5 \text{ dB} - 4.96 \text{ dB} = 24.5 \text{ dB}$$

The channel capacity for outdoor cables with a length of 33m and using the same frequency range is

$$C_{\text{worst case outdoor 33m}} = 9 \cdot 10^6 \log_2 \left( 1 + 10^{1.26} \right) = 38.3 \text{ Mbps}$$

$$C_{\text{average outdoor 33m}} = 9 \cdot 10^6 \log_2 \left( 1 + 10^{2.45} \right) = 73.3 \text{ Mbps}$$

Although the above Shannon-Hartley calculations give only the theoretical feasible channel capacity, it shows that the outdoor PLC channel capacity is higher when compared with the indoor PLC channel even though the cable length used is more than 3 times as long.

4.2 Indoor communications

We examine the use of PSK as modulation scheme to calculate the bit error probability using Eq. (3.5). Results are shown in Figure 11.

A higher bitrate can be achieved by using an ideal QPSK modulation scheme. QPSK needs more transmitting power but requires the same bandwidth with a bitrate two times the bitrate of a similar system using PSK. In the case of PLC communications we are power limited by the general CENELEC standard. To calculate the bit error probability Eq. (3.12) is used. Results are shown in Figure 11.

4.2.1 Error-correction coding

As mentioned in Section 3.2.3 convolutional coding improves the bit error probability considerably. One of the most used decoding schemes is the Viterbi soft decision algorithm with a constraint length $K=7$ and a code rate $R=1/2$. Using Eq. (3.22), (3.23) and (3.24) we can calculate the bit error probability using Viterbi with $K=7$ and $R=1/2$.

$$d_f = 10,$$

$$\Gamma(10) = 0.123047$$

Therefore,
Using Eq. (3.12) for QPSK we can calculate the performance improvement using QPSK modulation together with Viterbi error correcting coding. Results are shown in Figure 11.

\[ P_B < \Gamma(10) \cdot 320 p^3 \cdot \frac{1 + 3.2p}{(1 - 16p)^2} \]  

(4.3)

**Figure 11:** Bit error probability using PSK, QPSK and QPSK with Viterbi coding

The error probability calculations for Viterbi coding are based on the upper bound equations of van de Meeberg [21]. The upper bound equations are more accurate for a low crossover error probability \((p<10^{-3})\). This inaccuracy for lower \(S/N\) ratios is causing the QPSK Viterbi trace to cross the PSK trace, in Figure 11, when \(S/N\) gets lower than 6dB.

Calculations reveal that an indoor PLC network with a cable length of 10m and average \(S/N\) (19.5dB) has a small \((P_e<10^{-7})\) bit error probability. The worst-case \(S/N\) (4.5dB), for this same network, has a bit error probability which is too large \((P_e>10^{-3})\) for reliable digital communications.

This indicates that indoor PLC communications are not suitable for real-time applications like indoor-telephony, real-time controlling or monitoring because in the worst-case there is no communication possible over the PLC network. However the average-case has a bit error probability \((P_e<10^{-3})\) and a bitrate which is ideal for non real-time applications like Email, server replication, etc.

To communicate over larger distances than 10 meter, digital regenerative repeaters can be used. Calculations with these repeaters can be found in Section 4.3.1.
4.2.2 Bitrate
The bit duration-bandwidth product $T_e B$ of an ideal QPSK signal with coding is $T_e B = 0.5$. The Viterbi code rate $R = 1/2$. So in case with a bandwidth $B = 9$ MHz, the available information bitrate $R_b$ for a QPSK signal with Viterbi soft decision $(K=7, R=1/2)$ can be calculated using Eq. (3.21),

$$T_e B = 0.5 \Rightarrow R_b = 2BR = 9 \text{Mbps} \quad (4.4)$$

So the available information bitrate $R_b$ for QPSK with Viterbi soft decision $(K=7, R=1/2)$ is the same compared to standard PSK while the bit error probability is better for a given S/N ratio.

4.2.3 Longer indoor distances
Longer indoor distances between the transmitter and the receiver are more convenient but create also more attenuation. Using the values from Section 3.1.3 and the attenuation levels found in Section 2.1.1 we can calculate the signal-to-noise ratios $S/N$ for an indoor PLC system using cables with a length $L$ of resp. 20m and 30m.

$$\left(\frac{S}{N}\right)_{\text{worst case indoor 20m}} = \left(\frac{S}{N}\right)_{\text{max worstcase indoor 20m}} - L_{\text{dB worstcase indoor 20m}} = 24.5 \text{dB} - 40 \text{dB} = -15.5 \text{dB}$$

$$\left(\frac{S}{N}\right)_{\text{average indoor 20m}} = \left(\frac{S}{N}\right)_{\text{max average indoor 20m}} - L_{\text{dB average indoor 20m}} = 29.5 \text{dB} - 20 \text{dB} = 9.5 \text{dB}$$

$$\left(\frac{S}{N}\right)_{\text{worst case indoor 30m}} = \left(\frac{S}{N}\right)_{\text{max worstcase indoor 30m}} - L_{\text{dB worstcase indoor 30m}} = 24.5 \text{dB} - 60 \text{dB} = -35.5 \text{dB}$$

$$\left(\frac{S}{N}\right)_{\text{average indoor 30m}} = \left(\frac{S}{N}\right)_{\text{max average indoor 30m}} - L_{\text{dB average indoor 30m}} = 29.5 \text{dB} - 30 \text{dB} = -0.5 \text{dB}$$

Using these $S/N$ ratios and the calculations in Figure 11 it can be shown that the worst-case $S/N$ in both cases ($L=20$ or $L=30$) has a $P_e > 10^{-3}$, which is not acceptable for digital communications.

Further the $S/N$ ratio of the indoor PLC network with $L=20$ has a $P_e = 10^{-7}$, which is enough for reliable communications. The worst-case and average-case $S/N$ for $L=30$ however has also a $P_e > 10^{-3}$, which is not acceptable for digital communications.

So using a indoor cable length of 20m, between the transmitter and receiver, is possible but this prevents the indoor PLC system to be used for real-time applications.
4.3 Outdoor communications

4.3.1 Digital regenerative repeaters

Digital regenerative repeaters can be used to improve the performance of an outdoor PLC network with relative long cable lengths compared to indoor PLC networks. The following calculations used the measured line attenuation described in Section 2.1.1.2, the schematic overview of a typical Dutch outdoor low voltage network and the equations from Section 3.2.1.

The repeaters can be installed in every house, to enhance the signal after each hop in the PLC network. The distance between the houses is approximately 33 meters.

The signal-to-noise ratio after one hop is

\[
\left( \frac{S}{N} \right)_i = \left( \frac{S}{N} \right)_{\text{max}} - L_{db}
\]  

(4.5)

The worst-case $S/N=12.6$ dB and the average $S/N=24.5$ dB (see Section 4.1.2). The number of repeaters $M$ in an average PLC network is 30 (see Section 3.2.1). Using (3.8) we can now calculate the bit error probability when using these regenerative repeaters for PLC communications. The result are shown in the Figure 12.

![Figure 12: Bit error probability of a PLC network using regenerative repeaters with PSK modulation](image)

Figure 12: Bit error probability of a PLC network using regenerative repeaters with PSK modulation
When using PSK modulation it can be seen from Figure 12 that the signal-to-noise ratio \( S/N \) needs to be at least 9 dB, with 30 repeaters used, to get the generally maximum accepted bit error probability \( P_e = 1 \cdot 10^{-3} \) for modern digital communication systems.

The calculations of the bit error probabilities for given \( S/N \) in an outdoor PLC network, indicate that reliable real-time PLC communications are possible using digital regenerative repeaters with ideal PSK modulation.

### 4.3.1.1 Error-correction coding

As mentioned in Section 3.2.3 convolutional coding improves the bit error probability considerably. One of the most used decoding schemes is the Viterbi soft decision algorithm with a constraint length \( K = 7 \) and coding rate \( R = 1/2 \). Using Eq. (3.22), (3.23), (3.24), (4.2) and (4.3) we can calculate the bit error probability for Viterbi with \( K = 7 \) and \( R = 1/2 \).

Using the equations for the digital regenerative repeaters we can calculate the performance improvement using PSK modulation together with Viterbi error correcting coding with digital regenerative repeaters. For all further calculations we set the number of repeaters used \( M = 30 \). The result can be seen in Figure 13.

![Figure 13: Performance improvement using Viterbi error correcting coding](image)

When using PSK modulation with Viterbi coding it can be seen from Figure 13 that the signal-to-noise ratio \( S/N \) needs to be at least 4.3 dB, with 30 repeaters used, to get the generally maximum accepted bit error probability \( P_e = 1 \cdot 10^{-3} \) for modern digital communication systems.

The code rate \( R \) used here is \( 1/2 \) so the information bitrate of PSK with Viterbi is now half compared to standard PSK.
Other modulation schemes such as QPSK can improve the bandwidth/bitrate ratio, but need more power. In this PLC network we are power limited by the Cenelec general emission standard. We already saw that Viterbi error correcting coding could decrease the bit error probability with the same signal-to-noise ratio. So using ideal QPSK ($\eta_B=0.5$) together with Viterbi error correcting code could be a nice solution to overcome the extra carrier power requirements for QPSK while doubling the bitrate and using the same bandwidth compared with ideal PSK ($\eta_B=1$) with Viterbi coding. The bit error probability is calculated for PSK, QPSK both with and without using the Viterbi error correcting code to get a good comparison. The results are found in Figure 14.

![Bit error probability of a PLC network using regenerative repeaters with PSK, QPSK modulation schemes both with and without Viterbi coding](image)

**Figure 14:** Bit error probability of a PLC network using regenerative repeaters with PSK, QPSK modulation schemes both with and without Viterbi coding.

When using QPSK modulation with Viterbi coding it can be seen from Figure 14 that the signal-to-noise ratio ($S/N$) needs to be at least 7.3 dB, with 30 repeaters used, to get the generally maximum accepted bit error probability ($P_e=1\cdot10^{-3}$) for modern digital communication systems.

The large margin between the given $S/N$ (worst-case 12.6 dB and average-case 24.5 dB) and the needed $S/N$ (7.3 dB) indicate that the cable length can be made longer without compromising the reliability of the outdoor PLC network.

However the attenuation measurements are based on outdoor power cables with a coaxial structure, which have probably less attenuation than the cables used in the Netherlands with a non-coaxial structure (see Section 2.1.1). So this large margin could be needed when using low
voltage power cables in the Netherlands for PLC communications. Future measurements are needed to confirm this.

4.3.1.2 Bitrate

The bit duration-bandwidth product \( T_e B \) of an ideal QPSK signal with coding is \( T_e B = 0.5 \). The Viterbi code rate \( R = 1/2 \). So in case with a bandwidth \( B = 9 \) MHz, the available information bitrate \( R_b \) for a QPSK signal with Viterbi soft decision (\( K = 7, R = 1/2 \)) can be calculated using Eq. (3.21).

\[
T_e B = 0.5 \implies R_b = 2BR = 9 \text{Mbps}
\]

So the available information bitrate \( R_b \) for QPSK with Viterbi soft decision (\( K = 7, R = 1/2 \)) is the same compared to standard PSK while the bit error probability is better for a given S/N ratio.

The bitrate per user in a multiple access PLC network using a standard Time Division Multiple Access (TDMA) protocol can easily be calculated as follows [14]:

\[
R_{\text{user}} = R_b/n \text{ where } n \text{ is the number of simultaneous users on the PLC network.}
\]

To get the real information bitrate per user the frame efficiency have to be taken into account for the particular TDMA protocol used.

A PLC system using TDMA with a frame efficiency \( \eta \) of 80% and 120 users has a the following information bitrate per user:

\[
R_{\text{user}} = \frac{R_b \eta}{120} = 60 \text{Kbps}
\]

4.3.2 Spread spectrum

Spread spectrum can be used to overcome powerline interference and to create an efficient PLC network for multiple access.

4.3.2.1 Bit error probability

Given the equations (3.16), (3.17), (3.18), (3.19) and (3.20), the properties of spread spectrum sequence sets (see Section 3.2.2) and the bandwidth defined in Section 3.1.1, the bit error probability vs simultaneous users is calculated with Gold code PN sequences. For this calculation the following channel properties are used:

\[
\left( \frac{S}{N} \right)_{\text{max}} = 29.5 \text{dB.}
\]

An outdoor cable length of 33 meters with an attenuation of \( L_{\text{db}} = 4.96 \text{dB} \), where

\[
E_b \quad N_0 = \frac{R_p}{R_b} \cdot \left( \frac{S}{N} \right) = PG \cdot \left( \frac{S}{N} \right) = \left( 2^m - 1 \right) \cdot \left( \frac{S}{N} \right)
\]

where \( 2^m - 1 \) is the sequence length that can be created by an \( m \)-stage shift register

In Section 3.2.2 is shown that the bit error probability for sequence-synchronous DS-CDMA system using PSK is

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Calculations show that when the processing gain \( PG \) gets large (>200) Eq. (4.8) can be reduced to

\[
P_b = Q \left[ \frac{2}{k \rho_{\text{max}}^2 + (E_b / N_0)^{-1}} \right]^{1/2}
\]

(4.9)

So the bit error probability depends now only on the number of simultaneous users \( k \) and the factor \( m \) which determines the set size, the processing gain and the cross-correlation between the gold sequences. Four different PN sequences \((m = 8, 9, 10, 11)\) are used with the calculations. The results from these calculations are shown in the Figure 15.

Figure 15: Bit error probability of a PLC DS-CDMA system
Figure 16 shows a 3D plot of the bit error probability of the same PLC DS-CDMA system with different setsizes versus simultaneous users.

Figure 16: The bit error probability of a PLC DS-CDMA system with different setsizes versus simultaneous users.

Note: The setsize in Figure 16 means the m factor, which determines the setsize (see Table 3).
4.3.2.2 Bit error probability with Viterbi coding

Using Eq. (3.22), (3.23), (3.24) and (4.9) the bit error probability can be calculated for a PLC DS-CDMA system with Viterbi coding. Two different PN sequences (n = 8 and 9) are used with the calculations. The results are shown in Figure 17.

![Figure 17: Bit error probability of a PLC DS-CDMA system using Viterbi coding](image)

Results show that Viterbi coding can increase the number of users in a PLC system while performing with the same bit error probability. A PLC system with DS-CDMA (n=9) and Viterbi (K=7, R=1/2) can perform with an acceptable bit error probability (P_e<10^{-3}) for up to 120 simultaneous users.

4.3.2.3 Interference rejection

To investigate the interference capability of DS spread spectrum systems we will look at the effect of jamming.

The performance of a DS spread spectrum system in the presence of jamming may be estimated from the average probability of bit error.

The ratio of the energy per information bit to jamming density after despreading is given by

\[
\frac{E_b}{J_0} = \frac{R_p}{R_o} \left( \frac{S}{J} \right)
\]  

(4.10)
Hence the jammer-to-signal ratio, which defines the jamming margin of the receiver, is

$$\frac{J}{S} = \frac{R_p}{R_b} \frac{E_b}{J_0}. \quad (4.11)$$

Taking into account the AWGN, the required $E_b/N_0$ after despreading is

$$\frac{E_b}{N_0} = \frac{E_b}{N_0 + J_0} = \left( \frac{E_b}{N_0} \right)^{-1} + \frac{R_p J}{R_p S} \left( \frac{E_b}{N_0} \right)^{-1}. \quad (4.12)$$

Eq. (4.12) indicates an interference rejection capability in a DS spread spectrum system a large processing gain ($R_p / R_b$) will reduce the influence of the jamming power ($J$). In the Figure 18 is calculated how much the bit error rate (ber) is effected by jamming.

![Figure 18: Interference rejection capability of a PLC DS-CDMA system](image)

These calculations indicated that for a given value of $E_b/N_0 = 24.5$ dB the value of the jamming margin $J/S$ heavily depends on the processing gain ($2^m - 1$) and the jamming can be very severe for larger setsizes without causing the bit error probability to drop under $10^{-3}$. 

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4.3.2.4 Interference rejection with Viterbi coding

Using Eq. (3.22), (3.23), (3.24) and (4.12) the jamming margin \( J/S \) can be calculated. The results are shown in Figure 19.

![Diagram showing ber performance with uniform jamming](image)

**Figure 19:** Interference rejection capability of a PLC DS-CDMA system using Viterbi coding

Clearly can be seen that using Viterbi coding can increase the interference rejection capability of a PLC DS-CDMA system with a uniform jamming. The jamming margin \( J/S \) has increased with almost 5 dB using Viterbi coding compared to the same system without Viterbi coding.
4.3.2.5 Bitrate

The bitrate of a PLC DS-CDMA system can be calculated using Eq. (3.15), (3.16) and (3.17).

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**Figure 20:** Information bitrate $R_b$ of a PLC DS-CDMA system using Viterbi coding

Note: In Figure 20 the information bitrate $R_b$ of $m=9$ coincide with $m=8$, Viterbi ($R=1/2$)
5 Conclusions

5.1 Technical

5.1.1 Indoor
Calculations and measurements revealed that indoor PLC communications with a distance of 10m, between the transmitter and receiver, exhibit an acceptable bit error probability in the average -and worst-case when using PSK, QPSK with or without Viterbi decoding. Indoor PLC systems with QPSK and Viterbi decoding can have a bitrate of 9 Mbps. So indoor PLC systems with these distances can be used for real-time applications like telephony, fax, (computer) data networks, real-time controlling or monitoring.

Calculations and measurements revealed that indoor PLC systems with distances of 20m, between the transmitter and receiver, can not be used for real-time applications because the worst-case attenuation causes a high bit error probability \( P_e \gg 10^{-3} \), which disables any reliable communication. However the average-case has an acceptable bit error probability \( P_e < 10^{-3} \) and a bitrate 9 Mbps which is ideal for non real-time applications like email transmissions.

Calculations and measurements revealed that indoor distances of 30m or longer, between the transmitter and receiver, can not be used for indoor PLC communication using standard PSK, QPSK with or without Viterbi coding. The bit error probability is too high \( P_e \gg 10^{-3} \). A way to overcome these distances is to work with digital regenerative repeaters.

5.1.2 Outdoor
Calculations and measurements revealed that distances of tens of meters could be covered with moderate transmission power that complies to the generic emission standard using advanced techniques like digital regenerative repeaters, DS-CDMA and error correcting code.

Calculations and measurements indicate that the frequency range up to 10 MHz is usable for PLC communications. However to create a more accurate model more (accurate) attenuation measurements, on outdoor powerlines in the Netherlands, are needed.

Performance calculations of outdoor communication in a PLC network over distances of 33m in the low voltage grid with digital regenerative repeaters yield the following results:

- The calculations of the bit error probabilities for a given \( S/N \) in an outdoor PLC network, indicate that reliable real-time PLC communication is possible using 30 digital regenerative repeaters with ideal PSK modulation
- When using QPSK modulation with Viterbi coding there is large margin between the given \( S/N \) (worst-case 12.6dB and average-case 24.5dB) and the \( S/N \) needed (7.3 dB) to get the generally maximum acceptable bit error probability \( P_e=1\cdot10^{-3} \) for modern digital communication systems.

This large margin could be needed to compensate for the probably higher attenuation levels when using the non-coaxial low voltage power cables in the Netherlands for PLC
communications instead of the coaxial cables used in the outdoor attenuation measurements. Future measurements are needed to confirm this.

A PLC system with DS-CDMA (m=9) and Viterbi (K=7,R=1/2) can perform with an acceptable bit error probability \( P_e < 10^{-3} \) for up to 120 simultaneous users.

Outdoor PLC systems using digital regenerative repeaters perform better than when DS-CDMA is used. The information bitrate per user in PLC network (with 120 users) using digital regenerative repeaters (QPSK, Viterbi K=7, R=1/2) and TDMA (\( \eta=80\% \)) is higher (60kbps) compared to a PLC DS-CDMA (m=9, Viterbi K=7, R=1/2) system (4.4Kbps) while in both systems the bit error probability is sufficient \( P_e < 10^{-3} \) for reliable communications.

Future research work should be done on: phase response and group delay, analysis of possibilities for equalization.

### 5.2 Commercial

It is wise to compare outdoor PLC with other possible future communication systems such as cable TV networks. They also have the advantage (as PLC networks do) that they reach to the end user but cable TV networks can possibly achieve a much higher performance because of the coaxial structure of the cables. Another important factor is that all the wild disturbances that exist on the powerline are less likely to exist on the cable TV network.

### 5.3 Other developments

Prototypes of high-speed PLC networks are already available in Germany using Orthogonal Frequency Division Multiplexing (OFDM) together with digital repeaters. This system is developed by Siemens and Ascom and can handle a maximum of 200 houses connected to a single low voltage transformer. However, it is unclear if this system can be applied in the Netherlands using the limits of the Cenelec general emission standard.
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