Study on the transportation of management signals in a WDM network

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1 INTRODUCTION

A communication system transmits information from one place to another, whether separated by a few
kilometres or by transoceanic distances. Information is often carried by an electromagnetic carrier
wave whose frequency can vary from a few megahertz to several hundred terahertz.

Fiber-optic communication systems are lightwave systems that employ optical fibers for information
transmission. Such systems have been employed worldwide since 1980 and have revolutionised the
technology behind telecommunications. The lightwave technology, together with microelectronics, is
believed to be a major factor in the advent of the "information age".

Nowadays the performance of the lightwave systems is increasing rapidly. Not only the capacity or bit
rate increases, but the distance increases as well. A commonly used figure of merit for communication
systems is the bit rate-distance product, BL, where B is the bit rate and L is the repeater spacing. In
Figure 1-1 we can see the development of the lightwave systems over the last years. Section 1.1 gives a
historical perspective on the development of optical communication systems.

Figure 1-1 Progress in lightwave communication technology over the period 1974-1996. Differnt
curves show increase in the bit rate-distance product BL for five generations of fiber-optic
communication systems.

With the development of lightwave systems the control and management systems develop as well. The
lack of standardisation leads to a large diversity of solutions and discussions on how to implement
these tasks in the total network. In section 1.2 we discuss several options.

1.1 Five generations of lightwave systems

The commercial deployment of lightwave systems followed the research closely. After many field
trials the first generation lightwave systems operating near the 0.8μm became available commercially
in 1980. They operated at bit rates of 45 Mb/s and allowed a repeater spacing of about 10 km. The
larger repeater distance was an important motivation for system designers, as it decreased the
installation and maintenance costs associated with each repeater.

It was clear during the 1970s that operating the lightwave system in the wavelength near 1.3 μm could
increase the repeater spacing, where fiber loss is below 1 dB/km. Furthermore, optical fibers exhibit
minimum dispersion in this wavelength region. The second generation of fiber-optic communication
systems became available in the early 1980s, but the bit rate of early systems was limited below 100
Mb/s because of dispersion in multi-mode fibers. This limitation was overcome by the use of single-
mode fibers. By 1987 the second-generation 1.3 µm lightwave systems, operating at bit rates up to 1.7 Gb/s with a repeater spacing of about 50 km, were commercially available. The repeater spacing of the second-generation lightwave systems was limited by the fiber loss at the operating wavelength of 1.3 µm (typically 0.5 dB/km). The loss of silica fibers is minimum near 1.55 µm. However the introduction of third generation lightwave systems operating near 1.55 µm was considerably delayed by large fiber dispersion near 1.55 µm. The dispersion can be overcome either by using dispersion-shifted fibers designed to have minimum dispersion near 1.55 µm or by limiting the laser spectrum to a single longitudinal mode. Both approaches were followed during the 1980s. Third-generation 1.55 µm systems operating at 2.5 Gb/s became available commercially in 1990. The fourth generation of lightwave systems makes use of optical amplification for increasing the repeater spacing and of wavelength-division multiplexing (WDM) for increasing the capacity. In such systems, fiber loss is compensated periodically by using erbium-doped fiber amplifiers (EDFA) spaced 60-100 km apart. Such amplifiers were developed during 1980s and became available commercially by 1990. This system appeared to be feasible for intercontinental communication. By 1996, not only transmission over 11,300 km at a bit rate of 5 Gb/s had been demonstrated by using actual submarine cables, but a commercial transpacific cable (TPC-5) also became operational. Clearly the fourth-generation systems have revolutionised the state of art of the lightwave systems.

The current emphasis of fourth-generation lightwave systems is on increasing the system capacity by transmitting multiple channels through the WDM technique. Optical amplifiers are ideal for multi-channel optical transmission since all channels can be amplified simultaneously without requiring demultiplexing of individual channels. Despite the use of dispersion-compensation schemes, dispersive effects limited the total transmission distance to about 600 km. Commercial WDM operating at a bit rate up to 40 Gb/s were available by the end of 1996.

The fifth generation of fiber-optic communication systems is concerned with finding a solution to the fiber-dispersion problem. Optical amplifiers solve the loss problem but, at the same time, make the dispersion problem worse since the dispersive effects accumulate over multiple amplification stages. An ultimate solution is based on the concept of optical solitons, optical pulses that preserve their shape during propagation by counteracting the effect of dispersion through the fiber non-linearity. In 1996 soliton transmission over 9400 km was demonstrated at a bit rate of 70 Gb/s by multiplexing seven 10 Gb/s channels. Even though the fiber-optic communication technology is barely two decades old, it has progressed tremendously.

### 1.2 Management and control

Important parts of a communication network are the management and control functions. When we look at the different functions we have to make a distinction between two different types of control:

- management and control of the data (information);
- management and control of the network.

For example, when we look at a telephone call, the telecommunication network has to detect the request for a telephone call. Further the call has to be routed to the receiver and eventually ended. Besides this tasks the management and control functions have different administration tasks for billing and statistics.

Another task of the management system is to monitor and maintain the state of the network. Laser temperatures have to be monitored and eventually adjusted, wavelengths have to be reallocated (see section 2.2.1).

There are many different ways to transport the management information. When we look at ATM-cells, part of its capacity is reserved for management signal. In ISDN a special channel is reserved for signalling called the D-channel.

When it is not possible to use part of the capacity of the data signal for management communication, we have to look for new options. In Figure 1-2 we see three different options. A simple solution of course is to use a different network for the optical management. A disadvantage of this option is that it is very expensive.

So it is very interesting to know if there are possibilities of using the same network for management communication. As we have seen in section 1.1 we can use a different wavelength to arrange the management communication.
Another option and this option is discussed in this paper, is that we try to electrically multiplex two signals using the same wavelength.

Figure 1-2 Solutions for the transportation of management. a) Transportation of management using a special network. b) Transportation of management using a different wavelength or a multiplexing technique in the electrical domain.

1.3 TOBASCO project

In Europe several companies are trying to find solutions to upgrade a CATV network to an interactive high capacity network. They combined their research activities in as project sponsored by the European Commission, called TOBASCO.

In Chapter 2, the TOBASCO project and the way in which the new network offers high capacity interactive services will be explained.

As explained in the former two paragraphs the development in capacity and services of optical networks demands better management and control protocols. As we will see the wavelength selection can take place in this project. The management signals for the switching of the wavelength and temperature control of the lasers is exchanged using the RS232 protocol.

This report discusses a way to transport these management signals over the existing network. In Chapters 3 to 5 a model is presented in which the penalty of combining the management and data signals is calculated.

In Chapter 6 the theory is implemented in a transceiver for the management signal. The design of this transceiver is discussed and the basic principles are explained.

In a measurement set-up the performance of the data and management signals are measured. The results are discussed in Chapter 7.

Finally we will give a short review, in which the whole model is evaluated.
2 TOBASCO PROJECT

The TOBASCO project is a European project sponsored by the European Commission [1]. The objective is to upgrade existing CATV networks to a high capacity interactive network. This chapter describes the TOBASCO project objectives and gives an overview of the network architecture and components.

2.1 TOBASCO objectives

Today's CATV headend stations are feeding distributive services to large numbers of subscribers (>1000) in a network with abundant splitting. The introduction of interactive services in such a network is hampered by the limited bandwidth and addressing space of conventional TDMA/SCMA techniques. Until now, only narrowband interactive services can be carried; larger bandwidths per subscriber would require broadband switching and concentration functions to be installed closer to the subscribers. This would affect the network topology and complicate maintenance and operational aspects, resulting in high costs.

The scope of the TOBASCO (TOwards Broadband Access Systems for CATV Optical networks) project is to extend the interactive services to larger bandwidths (≥2 Mbit/s) by applying High Density Wavelength Division Multiplexing (HDWDM). With HDWDM, a large subscriber group can be partitioned into smaller groups, each characterized by a specific wavelength; TDMA/CDMA techniques offering peak rates of more than 2 Mbit/s per subscriber are used for addressing within these groups. Using these techniques, the switching functions for these interactive services are concentrated in the headend station, and the CATV network topology remains unchanged, both entailing significant techno-economic advantages.

The objectives of the TOBASCO project are:

1. to develop a strategy for upgrading partly optical CATV networks for provisioning of interactive broadband services to subscribers' homes;
2. to develop HDWDM system techniques which, in combination with Time Division or Sub-Carrier Division Multiple Access techniques, offer the transport capacity for 2 Mbit/s or more (ATM/Ethernet-oriented) connection capability via a partly fiber, partly coaxial cable based CATV network;
3. to develop strategies for flexible allocation of wavelengths, yielding an optimum Quality of Service;
4. to develop cost-effective optical components for multi-wavelength signal generation and routing;
5. to demonstrate in a laboratory the technical feasibility of the physical layer of the system, in combination with a limited network management and wavelength allocation control system;
6. to demonstrate the viability of the system in a field trial of a commercial cable operator, in order to show compatibility with ATM/Ethernet-based broadband communication services and with CATV services, and to perform user evaluation/acceptance studies with the commercial operator;
7. to contribute to standardization processes for (partly-) optical local access networks, in particular for CATV networks.

The impact of the project will represent a significant advancement in the introduction of interactive services in existing CATV networks. As an example of a promising area of application, the growing need of teleworkers to communicate with their industrial homebase may be considered. In an "Ethernet-to-the-Home" situation, a broadband interface from the teleworker's PC to the office can be created via the vast CATV local network infrastructure.

The main goal of the TOBASCO project is to identify and demonstrate in a practical environment an approach to upgrade CATV networks for the transport of interactive broadband services to and from the subscribers' homes. A similar approach will be valid to upgrade other non-CATV networks with high network-splitting factors.
2.2 Network Architecture

As a starting point for identifying strategies for interactive broadband services upgrades of CATV networks, existing CATV networks using optical fiber feeder links will be investigated. A typical example of such a system is shown in Figure 2-1; this type of system is being installed in the OPAL project in eastern Germany. Typical splitting factors are N=8 in the LSC (Local Splitting Center, which is installed in the field) and P=16 in the last optical part of the network. The number of houses connected to the optical network unit (ONU) via the coaxial access network varies considerably; a realistic number for modern CATV Fiber-to-the-Curb networks is 12-40 homes (e.g., quoted by Bellcore, OFC'95). Thus some 1536-5120 homes are served with CATV broadcast signals via a single fiber from the headend (HE).

The basic approach in the TOBASCO project is to apply multiple wavelengths to partition the large group of subscribers in a CATV network; each wavelength is used to carry the bi-directional broadband traffic of a part of the subscriber group. For instance, when N=8 wavelengths are used, each wavelength can handle P=16 ONUs corresponding to 192-640 subscribers.

The traffic for the subscribers may differ in its characteristics; this project will pay special attention to the packet-wise ATM characteristics. With a peak bit rate of 2 Mbit/s per subscriber, an Erlang-distributed traffic density of 60 mE, and a blocking probability of less than 0.6%, 40 subscribers per ONU would allow a concentration factor of 5.7 in the ONU and thus an average bit rate of 225 Mbit/s per wavelength; 12 subscribers allow a concentration factor of 3 only, implying 107 Mbit/s per wavelength.

The advantages of this approach are:

- Very-high-speed TDMA protocols with associated complicated ranging problems are avoided.
- The maintenance of the network is facilitated; e.g., by doing fault location on a specific wavelength, the associated part of the network can be scanned with more detail than if only a single wavelength for the total network would be available.
- The IS (Interactive Services) signals will be carried together with the CATV signal in the optical network.

2.2.1 Wavelength selection

Figure 2-2 shows that wavelength selection takes place at the ONUs. The splitter in the LSC distributes, besides the CATV signal at $\lambda_0$, all interactive wavelengths $\lambda_1$ to $\lambda_N$ to all ONUs. Each ONU is preferably equipped with a wavelength-selective receiver and transmitter, where selection is achieved by switching to the appropriate wavelength(s). As the receiver of each ONU, as well as the transmitter, may be switched to process one or more wavelengths, a very flexible wavelength assignment to the ONUs can be obtained, allowing dynamic capacity assignment (bandwidth on demand). Multiple wavelengths may be received and/or transmitted in an ONU by using a wavelength-selective receiver array and a multi-wavelength laser diode array. Wavelength switching is achieved via the network control circuit, managed by the headend. Thus an optimum allocation of bandwidth capacity to the ONUs can be realised. When upgrading from the CATV network in Figure 2-1, the only items to be replaced in the field are the optical amplifiers (OAs) in the LSC, which have to be replaced by bi-directional ones.
An evolutionary path towards the full wavelength-flexible IS ONU is envisaged, following a number of options for the ONU as illustrated by Figure 2-3. These options may coexist in the same network without compatibility problems. Initially, when introducing the multi-wavelength interactive signals in the network, in those ONUs still needing only CATV broadcast signals the distributive ONU can remain in place, and only a wavelength-blocking filter to suppress the interactive wavelengths is needed (Option 0). Those ONUs wishing broadband interactive services can be upgraded by simply adding a HDWDM splitter/combiner (to decouple interactive and distributive services), and the multi-wavelength IS ONU part. ONUs that do not need a re-arrangeable downstream or upstream capacity, can be equipped with a fixed-wavelength laser transmitter and fixed-wavelength receiver (Option 1), respectively. ONUs needing the full wavelength flexibility can employ the wavelength-switchable laser array and receiver array (Option 2). In this way, a graceful evolution towards the full wavelength-flexible broadband IS system is possible.
For the overall system the wavelength allocation has been specified; four wavelengths for the interactive upstream transmission of the interactive services, and four wavelengths carrying the interactive downstream traffic. An additional wavelength has been allocated for the distribution of the CATV signal. The wavelength allocation is shown in Figure 2-4.

![Figure 2-4 Wavelength allocation for the TOBASCO system.](image)

The four wavelengths carrying interactive services to the subscriber are specified as 1535.04, 1536.61, 1538.19, 1539.77 nm. The wavelengths for the interactive services from the subscriber to the Headend are 1535.82, 1537.40, 1538.98 and 1540.56 nm. Note that we are conforming to the international consensus of wavelength spacings of multiples of 100 GHz relative to the Kr-line at 1547.824 nm, which is in line with current standardization proposals (IEC TC86). The up- and downstream wavelengths are interleaved to minimize the amount of spectrum consumed. The distributive wavelength $\lambda_0$ for CATV distribution is to be located between 1550 and 1560 nm, to obtain maximum output power of the Erbium-doped optical amplifiers.

### 2.3 Optical Network Unit

The Optical Network Unit (ONU) has the following functions: to split off the CATV signal using a CWDM and feed this to an analogue CATV receiver already developed. The (remaining) IS signals are fed to a wavelength-switchable receiver. The upstream IS signal is generated at the selected wavelength by the wavelength-switchable transmitter. A network control circuit will take care of selection of the proper wavelength for both the transmitter and the receiver.

The wavelength switching speed is specified to be faster than 1 ms between the four wavelengths. The optical transmitter has to be a low-chirp, high-output-power device with internal optical isolation. Wavelength tuning is accomplished by laser temperature tuning. Because the ONU may have to operate in uncontrolled environmental conditions, it should be able to operate at an ambient temperature between 0 °C and 60 °C. The wavelength receiver should be able to receive at least one of the downstream wavelengths. Its adjacent cross-talk should be better than 25 dB, and its sensitivity better than -32 dB at 622 Mbit/s. The CATV receiver, like the transmitter, will be procured from outside the project.

On the coaxial medium standard protocols such as ATM forum 25.6 Mbit/s ATM or Ethernet will be used as far as possible. The upstream frequency range reserved for interactive services on the coax is between 20 and 30 MHz. State-of-the-art Medium Access protocols like CDMA or S-CDMA will be used by the cable modem. The downstream information is QAM-64 modulated and transported between 450 and 470 MHz. The capacity per subscriber for the interactive services should be at least 2 Mbit/s peak rate. CSO, CTB and CNR levels of the analog CATV channels at the subscriber premises, will have to be better than -60 dBc, -60 dBc and 45 dB, respectively.

As mentioned before, for provision of the interactive services, TOBASCO will build on equipment developed in an earlier RACE project, i.e., R2024 BAF. As the BAF equipment uses the 1.3 $\mu$m window for transmission, the optical part of the BAF-ONU has been replaced by a new 1.5 $\mu$m
switchable multi-wavelength optical board. Figure 2-5 shows a more detailed view of the TOBASCO ONU without the CWDM.

The non-shaded blocks in this figure indicate the electronic functions that have been designed and realised:

- 4 downstream receivers at 622 Mbit/s for each of the 4 photodiodes of the monolithic multi-wavelength photodetector. These receivers use transimpedance amplifiers and clock and data recovery chips by MAXIM.
- 4 optical burst-mode transmitters at 622 Mbit/s operating at the 4 upstream wavelengths. These transmitters use a 622 Mbit/s laser driver chip.
- A high-speed electrical switch to select one of the received wavelengths.
- A high-speed electrical switch to select one of the optical transmitters operating at the right wavelength for the upstream burst-mode data.
- A single-chip 8-bit micro controller system (Philips 89CE558) for temperature control of the monolithic multi-wavelength receiver, the 4 burst-mode upstream transmitters and the HDWDM which multiplexes all 4 transmitters, for switching the up- and downstream wavelengths, for adjustment and storage of laser bias and modulation current settings using EEPROMs and to interface with the NM&C system via a RS-232 interface. The firmware for this micro controller will be written in C using an 8051 cross-compiler from Franklin.

2.4 Optical Network Devices

2.4.1 The Coarse WDM (CWDM)

High-performance optical filters (CWDMs) are being used to separate the CATV from the IS signals. The role of this device is to separate the IS wavelengths, e.g. eight wavelengths between 1535 and 1541 nm, from the CATV signal located between 1550 and 1560 nm. The two groups of wavelengths are fixed, no tuning being needed.

The CWDM function is required:
1. at the Headend to combine the downstream interactive signals (4 wavelengths) and the CATV signal; it also directs the interactive upstream signals towards the multi-wavelength receiver;
2. at the ONU where it separates the downstream interactive signals from the CATV and sends the upstream interactive signals into the network.
3. in the bi-directional amplifiers to have separate IS and CATV amplifications.

The coarse WDM is a 3-port device with different specifications for the CATV and IS signals, because of different signal formats and levels and no wavelength tuning.
2.4.2 ONU Multi wavelength receiver:
The technical approach of the multi-wavelength receiver is based on the opto-electronic integrated
PHASAR based demultiplexer in InP with PIN photodetectors. PHASAR devices are based on an array
arrangement of bent waveguides, the radius of the bent controlling the length of each waveguide and
therefore the phase distribution at the output aperture of the array. This array type is dispersive and any
change in wavelength will move the focal spot at the output of the array. Based on this principle, a
HDWDM device can be made using a topology like [input waveguide - fan-out waveguide region -
waveguide array - fan-in waveguide region - output waveguides].

In TOBASCO, a 4-channel photodetector array has been monolithically integrated on a PHASAR
based demultiplexer chip. The required wavelength for the receiver is determined by the wavelengths
set by the transmitter lasers in Headend; the channel spacing is 1.6 nm. An important requirement is
polarisation independent operation of the demultiplexer. Figure 2-6 shows the layout of a PHASAR
demultiplexer: the overall device size with no detectors included yet is 1.2×3.3 mm².

Figure 2-6 Layout of eight-channel PHASAR demultiplexer. Size is 1.2 x 3.3 mm².

2.4.3 The TOBASCO lasers:
CCM-SIPBH DFB lasers with a λ/4 phase shift for high single mode operation and high yield of
specific wavelengths have been fabricated. By realising a wide wavelength range in the grating periods
of the laser structure using EBPG techniques, the required wavelengths can be obtained. Figure 2-7
shows this new laser structure:

Figure 2-7 Left: layer structure of CCM-DFB SIPBH laser. Right: Sketch CCM principle with
S/4 phase shift in the grating written by EBPG technique. The grating height is varied in the
longitudinal direction to reduce spatial hole burning at higher output powers.
3 MODELLING THE PHYSICAL LAYER

In this chapter a model is presented of the lightwave system of the TOBASCO project, which includes the transceiver of the management signal based on the RS232 interface.

In the first paragraph the different components and their physical characteristics are discussed. The second paragraph gives a small introduction in the way the data and management signal are multiplexed by using the same laser and fiber.

3.1 The Lightwave system of the TOBASCO project

In Figure 3-1 we can see, that the model consists of a laser, fiber, amplifier, optical filter and two receivers. At this time we do not take into account that the TOBASCO project is a WDM system. So at this stage we assume that we work with one data signal and one management signal operating at the same wavelength.

The different components and their parameters are discussed shortly.

![Figure 3-1 The Lightwave system of the TOBASCO project.](image)

- **The laser**
  The output-power of the laser depends on the output power of the management-signal $P_m$ and the power of the data-signal $P_d$. Furthermore the laser generates intensity noise, which is represented by the RIN.

- **Fiber**
  In our model the influence of dispersion is neglected and the fiber loss can be modelled by an attenuation factor $L$. (The losses due to imperfect filters or optical splitters are also taken into account.)

- **The amplifier**
  Besides amplifying the input signal with a factor $G$, the amplifier degrades the Signal-to-Noise Ratio (SNR), because of spontaneous emission. The SNR degradation is given by the amplifier-noise figure $F_{ap}$. The spontaneous-emission noise power $P_{sp}$ can be calculated out of these parameters.

- **Optical filter**
  In the TOBASCO project High Density Wavelength Division Multiplexers (HDWDM) are used to separate and combine different wavelengths. The optical filter used in our model has the same bandwidth as a channel from the HDWDM.

- **Receivers**
  The signal received by the photodiode has three components, the received power from the management-signal $P_{m}$, the power from the data-signal $P_{d}$, and the power from the spontaneous emission $L_{2}P_{sp}$ (with $P_{sp}$ in Watts). The current, caused by the received power, through the photodiode can be calculated by introducing the responsivity $R$. Furthermore there is some dark-current $I_{d}$ leaking through the diode, which causes extra shot-noise. The transimpedance preamplifier has a load resistor $R_{L}$, which causes thermal noise.
The other parameters which influence the noise sources are the operating temperature $T$ of the receiver, the equivalent noise bandwidth $\Delta f$, which is matched to the bitrate and the noise figure of the amplifier in the receiver $F_{nr}$.

### 3.2 Multiplexing two signals in a fiber

A laser starts to generate stimulated emission, as soon as the current flowing through the laser exceeds the threshold-level. We can write:

$$I_{\text{laser}} = I_{\text{threshold}} + I_{\text{modulation}}$$ \hspace{1cm} \text{Equation 3-1}

In a Direct-Detection Lightwave system, as in the TOBASCO-project, the bits are represented by optical pulses generated by the laser. The output power of the laser is proportional to the modulated current through the laser:

$$P_{\text{laser}} \sim I_{\text{modulation}}$$ \hspace{1cm} \text{Equation 3-2}

This modulated signal is a Non-Return-to-Zero pulse with a different positive value for a “0” and “1” and operates at 622 Mbit/s.

When we add another modulated positive current to the total current, the output-power of the laser will be proportional to the sum of the two modulated currents. This second signal is a signal which represents a management-signal and operates at low bitrates (10 kbit/s).

In the future we will address these different signals with the suffix “d” for the data-signal and “m” for the management-signal. In this case the output power of the laser exists out of two components:

$$P_{\text{laser}} = P_d + P_m \sim I_d + I_m$$ \hspace{1cm} \text{Equation 3-3}

Now we have multiplexed two different signals, which we have to separate at the receiver (see Figure 3-2).

The management-signal is retrieved by using a filter, so that the influence of the data-signal is minimised. The interference from the management-signal with the data-signal is kept to a minimum by using small current variations and therefore small power variations, in the modulated management current $I_m$.

![Figure 3-2 Two modulated currents affecting the output of a laser.](image)

As we will see in Chapter 5, the sensitivity of these two signals can be expressed by the Bit-Error-Rate (BER) of both signals. To model the BER for both data and management signal, we have to look at all noise sources in the system.
4 NOISE SOURCES

A measure of sensitivity for the signalling channel is the Bit-Error-Rate (BER). Before we can calculate the BER, we have to know all different noise sources and their influence on the system. The model we are working with is based on a Direct-Detection Lightwave system and uses a transimpedance preamplifier.

The different noise sources we have to cope with in this system are

- Receiver shot noise;
- Receiver thermal noise;
- Laser Relative Intensity Noise (RIN);
- Optical amplifier noise;
  - Spontaneous-Spontaneous Beat noise;
  - Signal-Spontaneous Beat noise;
  - Shot-Spontaneous Beat noise;
- Interference from other signals.

Each of the above discussed will be discussed separately in the following paragraphs.

4.1 Receiver shot noise

Shot noise [1] is a manifestation of the fact that the electric current consists of a stream of electrons that are generated at random times. The photodiode current generated in response to a constant optical signal can be written as

\[ I(t) = I_p + i_s(t) \]  

Equation 4-1

where \( I_p = R(P_d + P_m) \) is the average current and \( i_s(t) \) is a current fluctuation related to shot noise. Mathematically, \( i_s(t) \) is a stationary random process with Poisson statistics which in practice can be approximated by the Gaussian statistics. The autocorrelation function of \( i_s(t) \) is related to the spectral density \( S_s(f) \) by the Wiener-Khinchin theorem [3].

\[
< i_s(t) i_s(t + \tau) > = \int_{-\infty}^{\infty} S_s(f) \exp(2\pi if\tau) df
\]

Equation 4-2

The spectral density of shot noise is constant and is given by \( S_s(f) = qI_p \). So we can see shot noise as a white noise source, which has a two-sided spectral density.

The noise variance is obtained by setting \( \tau = 0 \) and is given by

\[
\sigma_s^2 = < i_s^2(t) > = \int_{-\infty}^{\infty} S_s(f) df = 2qI_p \Delta f
\]

Equation 4-3

where \( \Delta f \) is the effective noise bandwidth of the receiver.

All photodiodes generate some current even in the absence of an optical signal, because of stray light or thermal generation of electron-hole pairs. This residual current is referred to as the dark current.

Besides the dark current, the photodiodes generate current, because of the appearance of other signals. Their contribution is included in Equation 4-3 by replacing \( I_p \) by \( I_p + I_{dark} + I_m \). The total shot noise is then given by

\[
\sigma_s^2 = 2q[R(P_{d-r} + P_{m-r}) + I_{dark}] \Delta f
\]
4.2 Receiver thermal noise

At a finite temperature, electrons in any conductor move randomly. Random thermal motion of electrons in a resistor manifests as a fluctuating current even in the absence of an applied voltage. The load-resistor in the front-end of an optical receiver adds such fluctuations to the current generated by the photodiode. This additional noise component is referred to as thermal noise, Johnson noise or Nyquist noise [1 and 4].

Mathematically $i(t)$ (the thermal noise current) is modelled as a stationary Gaussian random process with a spectral density that is frequency independent up to ~ 1 THz (nearly white noise) and is given by

$$S_T(f) = \frac{2k_B T}{R_L}$$

Equation 4-4

where $k_B$ is the Boltzmann constant, $T$ is the absolute temperature and $R_L$ is the load resistor.

The noise variance is obtained by

$$\sigma_T^2 = \langle i^2_T(t) \rangle = \int_{-\infty}^{\infty} S_T(f) df = \left(\frac{4k_B T}{R_L}\right) \Delta f$$

Equation 4-5

where $\Delta f$ is the effective noise bandwidth. The same bandwidth appears in the case of both shot noise and thermal noise.

Equation 4-5 includes thermal noise generated in the load resistor. An actual receiver consists of many other electrical components, some of which may add additional noise. A simple approach accounts for the amplifier noise by introducing a quantity $F_n$ referred to as the amplifier noise figure and modifying Equation 4-6 as

$$\sigma_T^2 = \left(\frac{4k_B T}{R_L}\right) F_n \Delta f$$

Equation 4-6

Physically, $F_n$ represents the factor by which thermal noise is enhanced by the preamplifier.

4.3 Laser Relative Intensity Noise (RIN)

The output of a semiconductor laser exhibits fluctuations in its intensity even when the laser is biased at a constant current with negligible current fluctuation. The two fundamental noise mechanisms are spontaneous emission and electron-hole recombination. Noise in semiconductor lasers is dominated by spontaneous emission.

The intensity-autocorrelation function is defined as

$$C_{pp}(\tau) = \langle \delta P(t) \delta P(t + \tau) \rangle / \bar{P}^2$$

Equation 4-7

where $P$ is the average value and $\delta P$ represents a fluctuation. The Fourier transform of $C_{pp}(\tau)$ is known as the Relative Intensity Noise (RIN) [1] and is given by

$$RIN(\omega) = \int_{-\infty}^{\infty} C_{pp}(\tau) \exp(\imath \omega \tau) d\tau$$

Equation 4-8
Figure 4-1 shows the calculated RIN spectra at several power levels for a typical 1.55 μm InGaAsP laser. In our model the influence of the RIN will be modelled as a white noise source with an electrical-noise-bandwidth (Equation 4-9).

\[ \sigma_{\text{RIN}}^2 = (P_{d} + P_{m})^2 \cdot R^2 \cdot \Delta f \cdot \text{RIN} \]  

Equation 4-9

![Figure 4-1 Calculated RIN spectra at several power levels for a typical 1.55 mm semiconductor laser.](image)

**4.4 Amplifier noise**

The amplified signal received by the photo-detector can be written as

\[ P_{\text{dode}} = L_1 L_2 G (P_d + P_m) + L_2 P_{sp} \]  

Equation 4-10

where \( L \) is the attenuation caused by the two fibers, \( G \) is the amplifier gain, \( P_d \) and \( P_m \) is the power of optical signals at the output of the laser and \( P_{sp} \) is the spontaneous-emission noise power added to the signal such that

\[ P_{sp} = S_{sp} \Delta \nu_{sp} \]  

Equation 4-11

\( \Delta \nu_{sp} \) is the effective bandwidth of spontaneous emission and can be approximated by the amplifier bandwidth in practice. The spectral density \( S_{sp} \) is given by

\[ S_{sp}(\nu) = (G-1)n_{sp}h\nu \]  

Equation 4-12

where \( \nu \) is the optical frequency and \( n_{sp} \) is called the spontaneous-emission factor or population-inversion factor. We can also find a relationship between the population-inversion factor and the amplifier-noise figure, which is

\[ F_{na} = 2n_{sp}(G-1) / G \approx 2n_{sp} \]  

for \( G >> 1 \)  

Equation 4-13
The contribution of the spontaneous-spontaneous beat noise, signal-spontaneous beat noise and shot-spontaneous Beat noise is given by [1 and Olsson model]

\[
\sigma_{sp-sp}^2 = 4L_2^2 R^2 S_{sp}^2 \Delta v_{opt} \Delta f \tag{4-14}
\]

\[
\sigma_{sig-sp}^2 = 4L_2^2 R^2 G L_1 (P_d + P_m) S_{sp} \Delta f \tag{4-15}
\]

\[
\sigma_{shot-sig}^2 = 4qL_2^2 R S_{sp} \Delta v_{opt} \Delta f \tag{4-16}
\]

where \( R = \eta q/\hbar v \) is the photoconductor responsitivity and \( \Delta f \) is the receiver bandwidth. The three contributions \( \sigma_{sp-sp}^2, \sigma_{sig-sp}^2 \) and \( \sigma_{shot-sig}^2 \) originate from beating of spontaneous emission against itself, signal and shot-noise respectively.

### 4.5 Interference from other signals

Since we're using the fiber for more than one communication channel, we also have to deal with interference from other signals (besides the influence in shot noise, already mentioned). Here we need a different approach for two cases:

- the interference from the data-signals on the management-signal;
- the interference from the management-signal on the data-signal.

These different approaches are made because the two signals have a large difference in the frequency domain. As we will see in the following sections, in the first case, we can model the interference from the data-signals as a white noise source. The latter, we have looked at several signal states to get the model.

#### 4.5.1 Interference from the data-signals with the management signal

In the former paragraph is said, that we, mathematically, can model the interference of the data-signals on the management signal as a white noise source. If we look at the power spectral densities, we can see why.

The power spectral density of a random signal is the Fourier-transform of the autocorrelation [5]. In Figure 4-2 and Figure 4-3 the autocorrelation and power spectral density of an optical signal of a Non-Return-to-Zero random signal is given. The symbols used are explained in Table 4-1.

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>B-A</td>
<td>“Zero”-level of binary signal in Watts</td>
</tr>
<tr>
<td>B+A</td>
<td>“One”-level of binary signal in Watts</td>
</tr>
<tr>
<td>T</td>
<td>Bit-time</td>
</tr>
</tbody>
</table>
Figure 4-2 Autocorrelation of a random NRZ signal.

Figure 4-3 Power Spectral Density of a random NRZ signal without DC component.

The optical power spectral density is given by

\[
S_{dd}(\omega) = 2\pi B^2 \delta(\omega) + A^2 \frac{4 \sin^2 \left(\frac{1}{2} T \omega \right)}{\omega^2}
\]

Equation 4-17

As we can see, the plot of the power spectral density function is getting wider when shorter bit-times are used. If the difference of the bit-time between the management-signal and data-signal is large enough and a low-pass filter is used, the influence of the data-signal on the management-signal can be seen as a white noise source with an equivalent noise bandwidth, because the optical spectral density is nearly a constant between the limits of the filter.

If we translate the power transmitted by the laser into the current flowing through the photodiode Table 4-1 changes into Table 4-2.
Table 4-2 Binary representation of the current of through the photodiode.

<table>
<thead>
<tr>
<th>RL₁/L₂G(B-A)</th>
<th>“Zero”-level of binary signal in Amperes</th>
</tr>
</thead>
<tbody>
<tr>
<td>RL₁/L₂G(B+A)</td>
<td>“One”-level of binary signal in Amperes</td>
</tr>
<tr>
<td>T</td>
<td>Bit-time</td>
</tr>
<tr>
<td>R</td>
<td>Responsivity photodiode</td>
</tr>
</tbody>
</table>

The current spectral density is then given by

\[ S_\mu(\omega) = (RL₁L₂G)^2 [2\pi B^2 T^2 + \frac{A^2}{\omega^2}] \]

Equation 4-18

The autocorrelation function of \( i_d(t) \) is related to the spectral density \( S_{dd}(\omega) \) by the Wiener-Khinchin theorem:

\[
< i_d(t) i_d(t + \tau) >= \frac{(RL₁L₂G)^2}{2\pi} \int_{-\infty}^{\infty} S_{dd}(\omega)e^{i\omega\tau} d\omega = (RL₁L₂G)^2 \int_{-\infty}^{\infty} S_{dd}(f) e^{2\pi f\tau} df
\]

Equation 4-19

where angle brackets denote an ensemble average over fluctuations.

Since the bandwidth of the management signal is very small in comparison with the data signal and the delta-pulse of the DC-component is filtered the spectral density can be seen as a constant between the bandwidth and is given by \( S_d(0) \). Since the DC-component is filtered the average current \(<i_d(t)>\) is 0, thus the noise variance is obtained by setting \( \tau=0 \) in Equation 4-19:

\[
\sigma_d^2 = <i_d^2(t)> = (RL₁L₂G)^2 \int_{-\infty}^{\infty} S_{dd}(f) df = 2(RL₁L₂G)^2 S_{dd}(0) \Delta f
\]

Equation 4-20

where \( \Delta f \) is the effective noise bandwidth and

\[ S_{dd}(0) = (RL₁L₂GAT)^2 \]

Equation 4-21

**4.5.2 Interference from the management-signal with the data-signal**

In this case we cannot model the interference of the management-signal on the data-signal as a white noise source, because the management-signal will be constant over a lot of bits from the data-signal, so it is not unpredictable.

Here we assume that the detection-system moves automatically with the centre of the eye-pattern of the receiver and there are four instead off two possibilities for an error to occur.

\[
BER = p(1,1)P(0/1,1) + p(1,0)P(0/1,0) + p(0,1)P(1/0,1) + p(0,0)P(1/0,0)
\]

Equation 4-22

where \( p(a,b) \) is the probability of transmitting “a” for the data-signal and “b” for the management-signal and \( P(c/a,b) \) is the probability of receiving c when a and b are transmitted.

We assume that \( p(1,1)=p(1,0)=p(0,1)=p(0,0)=\frac{1}{2} \) and we neglect the possibility of a sample at a crossing of a level of the data-signal.
4.6 Summary Noise Sources

The photo-current generated at the detector can be written as

\[ I = I_p + i \]  

Equation 4-23

where \( I_p = R(P_d + P_m + P_{sp}) \) is the signal and \( i \) represents current fluctuations which should include the effects of shot noise, thermal noise, spontaneous emission noise, RIN and the interference noise from other signals. The variance \( \sigma^2 = \text{<}i^2\text{>} \) of current fluctuations can be written as

\[ \sigma^2 = \sigma_T^2 + \sigma_{\text{shot}}^2 + \sigma_{sp-sp}^2 + \sigma_{\text{sig-sp}}^2 + \sigma_{\text{shot-sp}}^2 + \sigma_{\text{RIN}}^2 + \sigma_d^2 \]  

Equation 4-24

where \( \sigma_T^2 \) is the thermal noise, \( \sigma_{\text{shot}}^2 \) is the shot noise, \( \sigma_{\text{RIN}}^2 \) is the intensity noise caused by the laser, \( \sigma_{sp-sp} \) and \( \sigma_{\text{sig-sp}} \) is the noise caused by the amplifier and \( \sigma_d^2 \) is the noise caused by the interference from the data-signal on the management (is 0, when the BER from the data-signal is calculated).

\[ \sigma_{\text{shot}}^2 = 2q[R(P_{d-r} + P_{m-r}) + I_{\text{dark}}] \Delta f = 2q[RL_1L_2G(P_d + P_m) + I_{\text{dark}}] \Delta f \]  

Equation 4-25

\[ \sigma_T^2 = (4k_BT / R_f) F_{nr} \Delta f \]  

Equation 4-26

\[ \sigma_{\text{RIN}}^2 = (P_{d-r} + P_{m-r})^2 \cdot R^2 \cdot \Delta f \cdot \text{RIN} = (RL_1L_2G)^2(P_d + P_m)^2 \Delta f \cdot \text{RIN} \]  

Equation 4-27

\[ \sigma_{sp-sp}^2 = 4L_2^2R^2S_{sp}^2 \Delta v_{\text{opt}} \Delta f \]  

Equation 4-28

\[ \sigma_{\text{sig-sp}}^2 = 4L_2^2G L_1(P_d + P_m)S_{sp} \Delta f \]  

Equation 4-29

\[ \sigma_{\text{shot-sp}}^2 = 4qL_2RS_{sp} \Delta v_{\text{opt}} \Delta f \]  

Equation 4-30

\[ \sigma_d^2 = 2(RL_1L_2G)^2S_{dd}(0) \Delta f \]  

Equation 4-31
5 PERFORMANCE

The performance criterion for digital receivers is governed by the bit-error rate (BER), defined as the probability of incorrect identification of a bit by the decision circuit of the receiver. A commonly used criterion for digital optical receivers requires BER ≤ 10⁻⁹. The receiver sensitivity is then defined as the minimum average received power required by the receiver to operate at a BER of 10⁻⁹.

In our model the management signal disturbs the data signal and the data signal disturbs the management signal. As we have seen in chapter 4 the noise sources depend on the power of both data and management signal. Hence the BER-curve depends on the modulation scheme used. The modulation scheme of the data signal is predetermined (direct-detection), but the modulation scheme of the data signal can be chosen.

In the following sections we will discuss several modulation schemes, but we will begin with the explanation of a BER-curve of one signal [1].

5.1 Bit-Error Rate of one signal

Figure 4.16a shows schematically the fluctuating signal received by the decision circuit, which samples it at the decision instant t₀ determined through clock recovery. The sampled value I fluctuates from bit to bit around an average value I₁ or I₀ depending on whether the received bit corresponds to “1” or “0” in the bit stream. The decision circuit compares the sampled value with a threshold value I₀ and calls it bit 1 if I>₁₀ or bit 0 if I<₁₀.

Figure 5-1 (a) Time-dependent fluctuating signal generated in the receiver in response to a digital bit stream. Signal is sampled at the instant t₀ by decision circuit and compared with a threshold level I₀. (b) Gaussian probability distributions centered at the average signal level I₁ and I₀. Dashed region shows the probability of incorrect identification when I₁ falls below I₀ or I₀ exceeds I₁.

An error occurs if I<₁₀ for bit 1 because of noise. An error also occurs if I>₁₀ for bit 0. Both of the sources of errors can be included by defining the error probability, or the BER, as
BER = p(1)P(0/1) + p(0)P(1/0)  \hspace{1cm} \text{Equation 5-1}

where p(1) and p(0) are the probabilities of receiving "1" and "0", respectively, P(0/1) is the probability of deciding 0 when 1 is received and P(1/0) is the probability of deciding 1 when 0 is received. We assume that in the bit stream "1" and "0" are equally likely to occur, so that p(0)=p(1)=\frac{1}{2} . The BER is therefore given by

\[ BER = \frac{1}{2} [P(0/1) + P(1/0)] \hspace{1cm} \text{Equation 5-2} \]

Fig. 4.16b shows how P(0/1) and P(1/0) depend on the probability density function p(I) of the sampled value I. The functional form of p(I) depends on the statistics of noise sources responsible for current fluctuations. Here we assume that every noise source can be described by Gaussian statistics with zero mean and variance \( \sigma_I^2 \). However, both the average and the variance are different for bit 1 and bit 0 depending on the received bit. If \( \sigma_1^2 \) and \( \sigma_0^2 \) are the corresponding variances, the conditional probabilities are given by

\[ P(0/1) = \frac{1}{\sqrt{2\pi}\sigma_1} \int_{-\infty}^{I_0} \exp\left[-\frac{(I-I_0)^2}{2\sigma_1^2}\right]dI = \frac{1}{2} \text{erfc}\left(\frac{I_0-I_1}{\sigma_1\sqrt{2}}\right) \hspace{1cm} \text{Equation 5-3} \]

\[ P(1/0) = \frac{1}{\sqrt{2\pi}\sigma_0} \int_{I_0}^{\infty} \exp\left[-\frac{(I-I_0)^2}{2\sigma_0^2}\right]dI = \frac{1}{2} \text{erfc}\left(\frac{I_0-I_1}{\sigma_0\sqrt{2}}\right) \hspace{1cm} \text{Equation 5-4} \]

where erfc stands for the complimentary error function, defined as

\[ \text{erfc}(x) = \frac{2}{\sqrt{\pi}} \int_{x}^{\infty} \exp(-y^2)dy \hspace{1cm} \text{Equation 5-5} \]

By substituting Equation 5-3 and Equation 5-4 in Equation 5-2, the BER is given by

\[ BER = \frac{1}{4} \left[ \text{erfc}\left(\frac{I_1-I_D}{\sigma_1\sqrt{2}}\right) + \text{erfc}\left(\frac{I_D-I_0}{\sigma_0\sqrt{2}}\right) \right] \hspace{1cm} \text{Equation 5-6} \]

Equation 5-6 shows that the BER depends on the decision threshold \( I_D \). In practice, \( I_D \) is optimized to minimize the BER. The minimum occurs when \( I_D \) is chosen such that

\[ \frac{I_1-I_D}{\sigma_1} = \frac{I_D-I_0}{\sigma_0} = Q \hspace{1cm} \text{Equation 5-7} \]

Explicit expressions for \( I_D \) and \( Q \) are

\[ I_D = \frac{\sigma_0 I_1 + \sigma_1 I_0}{\sigma_0 + \sigma_1} \hspace{1cm} \text{Equation 5-8} \]

\[ Q = \frac{I_1-I_0}{\sigma_1 + \sigma_0} \hspace{1cm} \text{Equation 5-9} \]

The BER with optimum setting of the decision threshold is obtained by using Equation 5-8 and Equation 5-9 and is given by Equation 5-10.
The approximate form of the BER is obtained by using the asymptotic expansion of \( \text{erf}(Q/\sqrt{2}) \) and is reasonably accurate for \( Q > 3 \).

### 5.2 Direct Detection modulation scheme of the management signal

In Figure 5-2 we see how the management and data signal are combined. As we can see there are four different levels of the laser driver current. Each of these levels fits to a possible combination (1 and 2; 3 and 4) of both signals.

![Figure 5-2 Direct Detection modulation scheme of both data and management signal](image)

Since the frequency of both signal differ significantly both signals need a different approach.

#### 5.2.1 BER management signal

As we can see in Equation 5-11 and Equation 5-12 the decision current \( I_D \) and \( Q \) is determined by the average current received by a "0" and "1" and the two different standard deviations.

\[
I_D = \frac{\sigma_0 I_1 + \sigma_1 I_0}{\sigma_0 + \sigma_1}
\]  

Equation 5-11

\[
Q = \frac{I_1 - I_0}{\sigma_1 + \sigma_0}
\]  

Equation 5-12

The average currents of the management signal are given by

\[
I_0 = [L_1 G(P_{d-\text{average}} + P_{m-0}) + P_{sp}] L_2 R
\]  

Equation 5-13

\[
I_1 = [L_1 G(P_{d-\text{average}} + P_{m-1}) + P_{sp}] L_2 R
\]  

Equation 5-14
In some cases the variance of the current fluctuation depends on the transmitted power of the laser. So there are two different standard variances:

\[ \sigma_0 = \sqrt{\sigma_T^2 + \sigma_{sp}^2 + \sigma_{shot}^2 + \sigma_{s-d0-m0}^2 + \sigma_{RIN-d0-m0}^2 + \sigma_{sp-d0-m0}^2} \]  

Equation 5-15

\[ \sigma_1 = \sqrt{\sigma_T^2 + \sigma_{sp}^2 + \sigma_{shot}^2 + \sigma_{s-d1-m0}^2 + \sigma_{RIN-d1-m0}^2 + \sigma_{sp-d1-m0}^2} \]  

Equation 5-16

The different terms for the shot noise, RIN and signal-spontaneous beat noise are given by Equation 4-25, Equation 4-27 and Equation 4-29, where \( P_d \) is \( \frac{1}{2}(P_{d-1} + P_{d0}) \) and \( P_m \) is the power transmitted by a "0" or "1" of the management signal.

### 5.2.2 BER data signal

As said, we assume, that the detection-system moves automatically with the center of the eye-pattern of the receiver and so there are four instead off two possibilities for an error to occur.

\[ BER = p(1,1)P(0/1,1) + p(1,0)P(0/1,0) + p(0,1)P(1/0,1) + p(0,0)P(1/0,0) \]  

Equation 5-17

where \( p(a,b) \) is the probability of receiving "a" for the data-signal and "b" for the management-signal and \( P(a/b) \) is the probability of deciding c for the data signal when a and b are received.

We assume that \( p(1,1)=p(1,0)=p(0,1)=p(0,0)=\frac{1}{4} \) and we neglect the possibility of a sample at a crossing of a level of the data-signal.

We assume, that the possibilities are equal to occur, so \( p(1,1)=p(1,0)=p(0,1)=p(0,0)=\frac{1}{4} \). If we compare this situation to the former, we can conclude, that we can determine the BER by taken the average of the BER of two different situations, equally to occur.

\[ BER_{total} = \frac{BER_{management=0} + BER_{management=1}}{2} \]  

Equation 5-18

#### 5.2.2.1 BER\(_{management=0}\)

In this case the two average currents are

\[ I_0 = [I_0 G(P_{d-0} + P_{m-0}) + P_{sp}]L_2 R \]  

Equation 5-19

\[ I_1 = [I_1 G(P_{d-1} + P_{m-0}) + P_{sp}]L_2 R \]  

Equation 5-20

and the two standard deviations are

\[ \sigma_0 = \sqrt{\sigma_T^2 + \sigma_{sp}^2 + \sigma_{shot}^2 + \sigma_{s-d0-m0}^2 + \sigma_{RIN-d0-m0}^2 + \sigma_{sp-d0-m0}^2} \]  

Equation 5-21

\[ \sigma_1 = \sqrt{\sigma_T^2 + \sigma_{sp}^2 + \sigma_{shot}^2 + \sigma_{s-d1-m0}^2 + \sigma_{RIN-d1-m0}^2 + \sigma_{sp-d1-m0}^2} \]  

Equation 5-22

where \( P_m \) is given by the power transmitted for a "0" from the management-signal, and \( P_d \) depends on the transmitted bit.
5.2.2.2 BER\text{management}=1

Now the average current is determined by the transmitted power from a "1" of the management-signal, so:

\[ I_0 = [L_1 G(P_{d-0} + P_{m-1}) + P_{sp}]L_2 R \]  

\[ I_1 = [L_1 G(P_{d-1} + P_{m-1}) + P_{sp}]L_2 R \]

and the two standard deviations are

\[ \sigma_0 = \sqrt{\sigma_T^2 + \sigma_{sp-sp}^2 + \sigma_{shot-sp}^2 + \sigma_{s-d0-m1}^2 + \sigma_{RUN-d0-m1}^2 + \sigma_{sig-sp-d0-m1}^2} \]

\[ \sigma_1 = \sqrt{\sigma_T^2 + \sigma_{sp-sp}^2 + \sigma_{shot-sp}^2 + \sigma_{s-d1-m1}^2 + \sigma_{RUN-d1-m1}^2 + \sigma_{sig-sp-d1-m1}^2} \]

where \( P_m \) is given by the power transmitted for a "1" from the management-signal, and \( P_d \) depends on the transmitted bit.

5.3 ASK modulation scheme of the management signal

Another modulation method we can use is ASK [6]. The transmitted signal is plotted in Figure 5-3.

![Figure 5-3 ASK modulation of the management signal.](image)

Instead of having two different combinations, we have three combinations; 1 and 2, 3 and 4, 5 and 6. By the use of an ASK modulation scheme we introduce a four-sided spectral density instead of two sided spectral density (see Figure 5-4). The extra noise is taken into account by increasing the effective noise bandwidth.

Again we follow the same approach as in paragraph 5.2.
5.3.1 BER management signal

Before we can calculate the BER-curve, we should take a closer look at the design of the receiver (see Figure 5-5). After the photodiode the signal passes through a band-pass filter to separate the management signal from the data signal. The lower frequencies like the DC-component of the signal are filtered, so that the signal is centered around the central axis (f=0). The higher frequencies are filtered and as a result of this the bloc-shaped pulses are translated into a sinus-shaped-pulse. After the band-pass filter the signal passes a double-sided rectifier. After the double-sided rectifier we recover the original 10kbit/s management signal with the use of a Low Pass filter. Since the DC-level of a bloc-shaped pulse is higher than the DC value of a rectified sinus pulse (the surface below the rectified pulse is smaller than the surface of an rectified bloc-shaped pulse) we have to take into account some losses in signal power.

In Figure 5-6 the transformation process of the signal through the receiver is plotted. The constant \( c \) in Figure 5-6 d is the penalty mentioned.

Figure 5-5 Design of an ASK receiver.
Figure 5-6 Transformation process of management receiver. a) Signal after photodiode. b) Signal after bandpass filter. c) Signal after double-sided rectifier. d) Signal after Low Pass Filter.

Now we understand the transformation process of the receiver, we can formulate the relations between the transmitted powers and the Bit-Error rate.

Let us assume that the front-end of the receiver has filtered the incoming signal, what leaves us with the following three different situations at the input of the band-pass filter.

\[
I_0 = [L_1 G (P_{d-\text{average}} + P_{m-0}) + P_{sp}] L_2 R
\]

Equation 5-27

\[
I_{1L} = [L_1 G (P_{d-\text{average}} + P_{m-1L}) + P_{sp}] L_2 R
\]

Equation 5-28

\[
I_{1H} = [L_1 G (P_{d-\text{average}} + P_{m-1H}) + P_{sp}] L_2 R
\]

Equation 5-29

and their variations

\[
\sigma_0 = \sqrt{\frac{\sigma_T^2 + \sigma_{sp-sp}^2 + \sigma_{\text{shot-sp}}^2 + \sigma_m^2 + \sigma_{z-d-\text{average-m}0}^2 + \sigma_{\text{RIN-d-\text{average-m}0}}^2 + \sigma_{\text{sig-sp-d-\text{average-m}0}}^2}{\sigma_{\text{RIN-d-\text{average-m}0}}^2 + \sigma_{\text{sig-sp-d-\text{average-m}0}}^2}}
\]

Equation 5-30

\[
\sigma_{1L} = \sqrt{\frac{\sigma_T^2 + \sigma_{sp-sp}^2 + \sigma_{\text{shot-sp}}^2 + \sigma_m^2 + \sigma_{z-d-\text{average-m}1L}^2 + \sigma_{\text{RIN-d-\text{average-m}1L}}^2 + \sigma_{\text{sig-sp-d-\text{average-m}1L}}^2}{\sigma_{\text{RIN-d-\text{average-m}1L}}^2 + \sigma_{\text{sig-sp-d-\text{average-m}1L}}^2}}
\]

Equation 5-31

\[
\sigma_{1H} = \sqrt{\frac{\sigma_T^2 + \sigma_{sp-sp}^2 + \sigma_{\text{shot-sp}}^2 + \sigma_m^2 + \sigma_{z-d-\text{average-m}1H}^2 + \sigma_{\text{RIN-d-\text{average-m}1H}}^2 + \sigma_{\text{sig-sp-d-\text{average-m}1H}}^2}{\sigma_{\text{RIN-d-\text{average-m}1H}}^2 + \sigma_{\text{sig-sp-d-\text{average-m}1H}}^2}}
\]

Equation 5-32

where \( P_{d-\text{average}} = \frac{1}{2} (P_{d-1} + P_{d-0}) \).

with the following chances of the management signal:
we get a probability function as given in Figure 5-7a. When this signal passes the double sided rectifier and the filters, negative values become positive.

\[ p(0) = \frac{1}{2} \]
\[ p(1\text{L}) = \frac{1}{4} \]
\[ p(1\text{H}) = \frac{1}{4} \]

Figure 5-7 Gaussian probability densities of the received signal. a) Densities before rectifier. b) Densities after rectifier and LPF.

The new Gaussian probability function is given by

\[ I_a = 0 \]  
\[ I_b = c(I_{1\text{H}} - I_0) \]
\[ \sigma_a = c\sigma_0 \]  
\[ \sigma_b = c\sigma_{1\text{H}} \]

Since \( I_{1\text{H}} - I_0 = I_0 - I_{1\text{L}} \) Equation 5-35 is straightforward, when the DC-component is filtered and the signal is rectified.

Both standard deviations when the management signal carries an “1” differ. Since \( \sigma_{1\text{H}} \) is higher than \( \sigma_{1\text{L}} \) we will chose for the worst case scenario and equal \( \sigma_b \) to \( c\sigma_{1\text{H}} \). Here we see that the standard deviations decrease with the same factor \( c \) as the average received current.

Here we see that the BER is equal to:

\[ BER = \frac{1}{4} \left[ 2 \cdot \text{erfc} \left( \frac{I_{1\text{H}} - I_0}{\sigma_{1\text{H}} \sqrt{2}} \right) + \text{erfc} \left( \frac{I_0 - I_0}{\sigma_0 \sqrt{2}} \right) \right] \]  
\[ \text{Equation 5-37} \]

where \( I_0 \) is numerically solved so that the BER is set to a minimum.
So when we define the extinction ratio of the management signal (ASK modulated) as the difference between the "1H" and "1L" (see Figure 5-8) the extinction ratio is not directly matched to the distance of the average current received when receiving a "0" or an "1".

5.3.2 Simulation results

In Figure 5-9 we see the results of several simulations. Since we want to focus upon the penalty induced by the combining of both data and management signal, the optical amplifier is excluded in our simulations.

We see the BER as a function of the average power received of the management signal. The BER curve is plotted at several different coupling ratios with the data signal.
Figure 5-9 Theoretical BER of the management signal.

In the simulation we used the following values:

<table>
<thead>
<tr>
<th>Characteristic values of the simulation of Figure 5-9.</th>
</tr>
</thead>
<tbody>
<tr>
<td>RIN&lt;sub&gt;liner&lt;/sub&gt;</td>
</tr>
<tr>
<td>F&lt;sub&gt;ref&lt;/sub&gt;</td>
</tr>
<tr>
<td>Δf</td>
</tr>
<tr>
<td>R</td>
</tr>
<tr>
<td>I&lt;sub&gt;dark&lt;/sub&gt;</td>
</tr>
<tr>
<td>R&lt;sub&gt;l&lt;/sub&gt;</td>
</tr>
<tr>
<td>Extinction ratio&lt;sub&gt;data&lt;/sub&gt;</td>
</tr>
<tr>
<td>Extinction ratio&lt;sub&gt;management&lt;/sub&gt;</td>
</tr>
<tr>
<td>T</td>
</tr>
<tr>
<td>Bitrate data signal</td>
</tr>
<tr>
<td>Bitrate management signal</td>
</tr>
</tbody>
</table>

5.3.3 BER data signal

Instead of having two different combinations, which we had in the former paragraph, we now have three. When we assume having three different combinations, we should have three different eye patterns. Since it is hardly recognisable since the power of the management signal is far below of the power of the data signal, we raised the power of the management signal. In Figure 5-10 we see three different eye-patterns, of which each belongs to a level of the management signal.
Figure 5-10 Three different eye-patterns can be found, when both signal powers are equal

Once more, we assume that the detection-system moves automatically with the center of the eye-pattern of the receiver and so there are four instead of two possibilities for an error to occur.

\[
\begin{align*}
\text{BER} &= p(1,1H)P(0/1,1H) + p(1,1L)P(0/1,1L) + p(1,0)P(0/0,1H) + \\
&+ p(0,1H)P(1/0,1H) + p(0,1L)P(1/0,1L) + p(0,0)P(1/0,0) \\
&= \frac{1}{2} (\text{BER}_{\text{management}=1H} + \text{BER}_{\text{management}=1L}) \\
\end{align*}
\]

where \( p(a,b) \) is the probability of receiving "a" for the data-signal and "b" for the management-signal and \( P(c/a,b) \) is the probability of deciding c for the data signal when a and b are received.

We assume that \( p(1,1H)=p(1,1L)=p(0,1H)=p(0,1L)=1/8 \) and \( p(0,0)=p(1,0)=1/4 \) and we neglect the possibility of a sample at a crossing of a level of the data-signal.

If we compare this situation to the former, we can conclude, that we can determine the total BER by taken the BER of three different situations.

\[
\text{BER}_{\text{total}} = \frac{\text{BER}_{\text{management}=0} + \frac{1}{2} (\text{BER}_{\text{management}=1H} + \text{BER}_{\text{management}=1L})}{2}
\]

5.3.3.1 \( \text{BER}_{\text{management}=0} \)

In this case the two average currents are

\[
\begin{align*}
I_0 &= [L_1G(P_{d-0} + P_{m-0}) + P_{sp}]L_2 R \\
I_1 &= [L_1G(P_{d-1} + P_{m-0}) + P_{sp}]L_2 R
\end{align*}
\]

and the two standard deviations are

\[
\sigma_0 = \sqrt{\sigma_I^2 + \sigma_{sp}^2 + \sigma_{skew}^2 + \sigma_{d0-m0}^2 + \sigma_{BER-d0-m0}^2 + \sigma_{bias-sp-d0-m0}^2}
\]
where \( P_m \) is given by the power transmitted for a “0” from the management-signal, and \( P_d \) depends on the transmitted bit.

5.3.3.2 BER_{management=1L}

Now the average current is determined by the transmitted power from a “1L” of the management-signal, so:

\[
I_0 = [L_d G(P_{d-0} + P_{m-1L}) + P_{sp}] L_2 R \quad \text{Equation 5-44}
\]

\[
I_1 = [L_d G(P_{d-1} + P_{m-1L}) + P_{sp}] L_2 R \quad \text{Equation 5-45}
\]

and the two standard deviations are

\[
\sigma_0 = \sqrt{\sigma_1^2 + \sigma_{sp-0}^2 + \sigma_{shot-0}^2 + \sigma_{1-d0-0}^2 + \sigma_{RIN-d0-0}^2 + \sigma_{sig-sp-d0-0}^2} \quad \text{Equation 5-46}
\]

\[
\sigma_1 = \sqrt{\sigma_1^2 + \sigma_{sp-0}^2 + \sigma_{shot-0}^2 + \sigma_{1-d1-0}^2 + \sigma_{RIN-d1-0}^2 + \sigma_{sig-sp-d1-0}^2} \quad \text{Equation 5-47}
\]

where \( P_m \) is given by the power transmitted for a “1” from the management-signal, and \( P_d \) depends on the transmitted bit.

5.3.3.3 BER_{management=1H}

Now the average current is determined by the transmitted power from a “1H” of the management-signal, so:

\[
I_0 = [L_d G(P_{d-0} + P_{m-1H}) + P_{sp}] L_2 R \quad \text{Equation 5-48}
\]

\[
I_1 = [L_d G(P_{d-1} + P_{m-1H}) + P_{sp}] L_2 R \quad \text{Equation 5-49}
\]

and the two standard deviations are

\[
\sigma_0 = \sqrt{\sigma_1^2 + \sigma_{sp-0}^2 + \sigma_{shot-0}^2 + \sigma_{1-d0-0}^2 + \sigma_{RIN-d0-0}^2 + \sigma_{sig-sp-d0-0}^2} \quad \text{Equation 5-50}
\]

\[
\sigma_1 = \sqrt{\sigma_1^2 + \sigma_{sp-0}^2 + \sigma_{shot-0}^2 + \sigma_{1-d1-H}^2 + \sigma_{RIN-d1-H}^2 + \sigma_{sig-sp-d1-H}^2} \quad \text{Equation 5-51}
\]

where \( P_m \) is given by the power transmitted for a “1” from the management-signal, and \( P_d \) depends on the transmitted bit.

5.3.4 Simulation results

In Figure 5-11 we see the results of several simulations. We see the BER as a function of the average power received of the data signal. The BER curve is plotted at several different coupling ratio with the management signal.
5.4 The performance of management communication in a WDM network

Although this report main emphasis is the disturbance of one data signal on one management signal and of one management signal on one data signal, we will take a preview of the disturbance of several different data and management signals in a WDM network, like TOBASCO. The statements mad in this chapter can be used as guidelines for further study.

In a WDM network different wavelengths are multiplexed in one fiber. The TOBASCO project uses four different wavelengths for upstream and four wavelengths for downstream communication. If we superpose our management signal on all channels we can achieve a higher sensitivity (see Figure 5-12).
Figure 5-12 The summation of currents results in a higher sensitivity.

Since the noise sources are independent the variations may be summarised just like the new current. Hence the new current and standard deviation are:

\[ I_t = \sum_{i=1}^{n} I_i \]  
\[ \sigma_t = \sqrt{\sum_{i=1}^{n} \sigma_i^2} \]  

Equation 5-52  
Equation 5-53

Since the standard deviation increases with the square root of the summation of the variations and the total current increases with the summation of the currents, we are able to achieve higher receiver sensitivities when we choose this set-up.

As we have seen in Chapter 4, the variance depends on different noise sources. In Equation 5-54 (replica of Equation 4-24) we have separated these noise sources.

\[ \sigma^2 = \sigma_T^2 + \sigma_{\text{shot}}^2 + \sigma_{\text{sp}}^2 + \sigma_{\text{sp-sp}}^2 + \sigma_{\text{sig-sp}}^2 + \sigma_{\text{shot-sp}}^2 + \sigma_{\text{RIN}}^2 + \sigma_d^2 \]  

Equation 5-54
When we take a closer look at the set-up described (see Figure 5-13a), we see that we de-multiplex the different channels in the HDWDM (High Density Wavelength Division Multiplexer) and multiplex the signals by adding the currents through the different photodiodes. By placing the photodiode for the HDWDM we can avoid the losses and more important we can reduce the total noise. This measure improves the sensitivity of the management signal, but introduces extra losses, because we have to use an optical splitter to divide the data and management signal. This option may be interesting, when a large number of wavelengths are multiplexed and the losses of the optical splitter are replaced by a smaller penalty of the sensitivity of the data signal, due to smaller power variations in the optical signal of the management signal (1L and 1H).

The output-current of the photodiode depends linearly on the optical power, which falls on the photodiode. Since the responsitivity R of the photodiode is nearly constant in the 1500 nm region, the photodiode in the set-up of Figure 5-13b summates the different channels.

When we only use one photodiode, we only have one thermal noise source and one shot-noise source due to the dark current, instead of having four times these noise sources (see Figure 5-13b). Hence the sensitivity of the network improves. Another advantage is of course the reduction of total costs.

Another advantage of the transmission of the management signals using all wavelengths is, that the management signal still is received, when one ore more wavelengths have fallen out. So the network is redundant.

When we use an ASK modulation scheme, we do not take any advantage of the changing levels between 1L and 1H. When we make use of CDMA (Code Division Multiple Acces) and we increase the sample rate we can divide the bits in chips and achieve a higher sensitivity. This technique is beyond the framework of this report and therefore the reader is referred to other articles and literature.

Figure 5-13 Placement of management receiver. a) After HDWDM. b) Before HDWDM.
6 TRANSCEIVER DESIGN

The challenge of the transceiver design is to find a suitable modulation scheme for the management signal, so that the influence of the management at the data signal is reduced to a minimum and that there is an "error-free" transmission path for the management signal.

If we look at the transceiver design, we have to take the following properties of the TOBASCO project into account:

- the transmission of the data signal is based on a direct detection lightwave system, working at a speed of 622 Mbit/s;
- the transmission of the data signal is sent in bursts (upstream);
- the management signal is based on a RS232 interface, which transmits data in burst mode at a rate of 9600 bit/s (is approximated by 10 kbit/s).

Due to the fact that the DC-component of the data signal changes, we have to choose a modulation scheme for the management signal, that is able to adapt to these changes of circumstances. The option chosen is to use a modulation scheme, which doesn't have a DC-component. This means, that we can't use the baseband direct detection method. So we have to use ASK or FSK modulation schemes [6].

If we use FSK, we can achieve a higher sensitivity, but the design will be more complex. Because we still are working in an experimental stage, we have chosen for ASK modulation, to keep the design less complex. (The complete schematic is given in Appendix A.)

6.1 Optical transmitter

When we look in Figure 6-2 we see the basic principle of an optical transmitter. In our design the frequency of the clock is a major issue.

![Figure 6-2 Scheme of an Optical ASK transmitter](image)

When a ASK modulation scheme is used, the original signal is recovered by recovering the envelope of the signal. Most of the time this is done by a Low Pass Filter at the receiver. To be able to filter the envelope there has to be a certain distance between the cut-off frequency (3 dB point) of the LPF and
the clock. On the other hand the spectrum of the management and the data signal have to be apart so that they can be separated. The best clock frequency in our system is therefore 200 kHz (see paragraph 6.1.1).

Since we are in an experimental stage it was not practical to add the management signal to the same laser, which send the data. Although this is the most efficient option in economic and technical sense, we have to make a transmitter, which is able to send the ASK signal. The option chosen is to use a 3-State Buffer Device, which sets us to be able to send the same DC-component, when transmitting a “0” or an “1” (see Figure 6-3).

![Figure 6-3 3-State Buffer Device, which delivers a symmetrical output at V*.

Using the function table of the 3-State Buffer Device in a right way, we can achieve the goal of sending 3 different levels, which are symmetrical to the centre level (see Figure 6-4).

![Figure 6-4 3-state buffer device used to achieve symmetrical levels.

Since the lowest level (IL) of the management signal is far above the threshold of the laser, we add an extra DC-component to the received signal. The results will therefore deteriorate a little bit, but testing should be possible.
6.1.1 Clock frequency

As mentioned the clock frequency is centred between the spectrum of the management signal and the spectrum of the data signal.

The data signal has a bit rate of 622 Mbit/s. The coding technique used is a code in which the maximum number of ones or zeros in succession is 100. As a result the minimum frequency of switching between a one and a zero of the data signal is 6.22 Mbit/s. Using Nyquist we know that the minimum carrier frequency of the data signal is half the bit-rate. Although the data signal is transmitted by using a direct detection modulation scheme, we assume that the spectrum of the data signal begins at half the minimum bit-rate.

When we take a look at the management signal, we see that the spectrum is largely concentrated in the area between $-1/f$ and $1/f$ (see Figure 4-3). The middle of these two frequencies is approximately (see Figure 6-5) 200 kHz.

![Figure 6-5 The clock frequency in the middle of the data and management spectrum.](image)

6.2 Optical receiver

The design of an optical receiver depends largely on the modulation format used by the transmitter. In particular, it depends on whether the signal is transmitted in an analog or digital format. Since we work with a digital format, we take a closer look at the block diagram of such a receiver (see Figure 6-6) [1].

![Figure 6-6 Diagram of a digital optical receiver showing various components. Vertical dashed lines group receiver components into three sections.](image)

Its components can be arranged into three groups: the front end, the linear channel and the data-recovery section. In this chapter we discuss each group separately.

6.2.1 Front end

The Front End of a receiver consists of a photodiode followed by a preamplifier. In the configuration a transimpedance front end is used, because of the high sensitivity. During this stage we receive both data and management signal. If do not make any changes to the transimpedance amplifier [1 and 7] the maximum gain is limited by the sum of both received powers. As we can see in Figure 6-7, in our design there are two changes to the standard transimpedance front end.
The first change is the capacitor, that is placed parallel to the load resistor. The effect of this capacitor is a band limitation. It works as a Low Pass Filter with the result that the noise decreases and the spectrum of the management signal is separated from the data signal.

The second change is the addition of a High Pass Filter between the photodiode and the transimpedance amplifier. As a result of this the gain can be increased, because we cancel out the DC-component.

Another side-effect is that the thermal noise increases as a result of adding the resistor in front of the transimpedance amplifier. The extra amount of noise is taken into account by increasing the noise figure.

6.2.2 Linear Channel

Normally the linear channel in optical receivers consists of a high-gain amplifier and a low-pass filter, but because we are working with a ASK modulation our Linear Channel consists out of three components (see Figure 6-8).

In the first section the ASK modulated signal is filtered. As a result of this LPF (and of the HPF in the front end) we have limited the influence of the data signal and noise. After this section we use a double-sided rectifier and a LPF to recover the 10 kbit/s signal.
Equation 6-1 Transformation process of management receiver. a) Signal after photodiode. b) Signal after bandpass filter. c) Signal after double-sided rectifier. d) Signal after Low Pass Filter.

Because we work with a burst mode receiver, we cannot work with a gain, which is automatically controlled to limit the average output voltage to a fixed level irrespective of the incident average optical power at the receiver. Therefore the gain of each filter is fixed to operate between certain limits.

6.2.3 Data Recovery

The data recovery circuit is drawn in Figure 6-9.

![Figure 6-9 Schematic of the Data Recovery circuit.](image)

We can compare the basic principle of this circuit with a feedback control system with the exception, that the feedback consists out of a peak-detection. The following equations will clarify this statement.

For the first part of the circuit we can derive Equation 6-2 (We eliminate the effect of the capacitor, because we work with frequencies far below the cut-off frequency of the filter.).

$$V_{\text{input}} - \frac{V^*}{2} = \frac{V^*}{2} - V_1$$

Equation 6-2
Simplifying Equation 6-2 we get:

\[ V_1 = V^* - V_{\text{input}} \]  

Equation 6-3

For the peak detector and the amplifier the following equations apply:

\[ V_2 = -V_1 \]  

Equation 6-4

\[ V^* = \hat{V}_2 \]  

Equation 6-5

Substituting Equation 6-4 and Equation 6-5 in Equation 6-3 we get

\[ V_1 = \frac{V_{\text{input}}}{2} - V_{\text{input}} \]  

Equation 6-6

Another option is to couple input the peak-detection (V2) to the input of the Data Recovery Unit and use a circuit which is able to subtract half of the peak of the input of the Data Recovery Unit. This option will result in Equation 6-6 also.
7 Measurements

7.1 Bit-Error Rate measurement of the management signal

In the following paragraphs the measurements results are given of the BER-measurements of the management signal. The properties of the transmitters and receivers are given in Table 7-1. Before we discuss the BER of the management signal, we take a closer look at the cell structure of signal, that is used by the RS232 protocol. Then we discuss the measurement set-up, followed by the results and conclusions.

Table 7-1 Properties of the data and management signals

<table>
<thead>
<tr>
<th>Property</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Δf_data</td>
<td>500 MHz</td>
</tr>
<tr>
<td>Δf_management</td>
<td>250 kHz</td>
</tr>
<tr>
<td>R</td>
<td>0.9 A/W</td>
</tr>
<tr>
<td>I_dark</td>
<td>10^-9 A</td>
</tr>
<tr>
<td>R_s-data</td>
<td>6 kΩ</td>
</tr>
<tr>
<td>R_s-management</td>
<td>105 kΩ</td>
</tr>
<tr>
<td>Extinction ratio_data</td>
<td>7.1 dB</td>
</tr>
<tr>
<td>Extinction ratio_management</td>
<td>8.1 dB</td>
</tr>
<tr>
<td>T</td>
<td>293 K</td>
</tr>
<tr>
<td>Bitrate data signal</td>
<td>622 Mbit/s</td>
</tr>
<tr>
<td>PRBS_s-data</td>
<td>2^23-1</td>
</tr>
</tbody>
</table>

7.1.1 Bit-Error-Rate coupled to the cell structure

Since the RS232 protocol describes a burst mode signal, we have to find a way to correspond the Bit Error Rate to the cells of the burst mode signal. To comprehend this connection, we take a closer look at the cell structure of a signal, which uses the RS232 protocol. In Figure 7-1 we see that the cell consist out of 10 different bits and that the first and last bit are respectively the start and stop bit.

![Figure 7-1 Structure of a RS-232 cell.](image)

The first way we can achieve the BER is by using the following formula:

$$BER = 1 - \left( \frac{NCR}{NCR + NER} \right)^{1/8}$$

Equation 7-1

where

- BER = Bit Error Rate
- NCR = Number of Correct cells Received
- NER = Number of Errored cells Received
This formula is based upon the fact that a received cell is correct when every data-bit is received correctly; so the ratio of correct cells is equal to the chance that eight bit in a row are received correctly (see Equation 7-3):

\[
\frac{NCR}{NCR + NER} = (1 - BER)^8
\]

Equation 7-2

The other way is to check the number of bit-faults in the received cells.

\[
BER = \frac{\# \text{ faults}}{8 \cdot (NCR + NER)}
\]

Equation 7-3

where

\# faults = Number of bit-faults in the data

The latter is more precise than the first one, so we apply Equation 7-3.

### 7.1.2 Measurement set-up

Since the transmission capacity is low (10 kbits/s) and the transmission occurs in cells, we cannot use a normal BER-set for the error measurements. Therefore a special program is written to calculate the BER using Equation 7-3.

The measurement set-up is given in Figure 7-2. Here we can see, that we use a 3dB coupler to combine the management signal and the data signal. The data signal is generated by a normal BER-set. By using the two attenuators before the coupler so that we can arrange the power-coupling ratio of both signals. With the use of the attenuator after the coupler we can change the power received by the photodiode without changing the power coupling-ratio of the combined signal.

![Figure 7-2 BER measurement set-up of the management signal](image)
7.1.3 Measurements results

Because the transmission capacity of the management signal is very low (10 kbit/s), we need to measure for a very long time to be able to measure an error rate of $10^{-9}$, which corresponds to the sensitivity of the receiver. For this reason we only measured considerably higher error rates.

In Figure 7-3 we see the results. The BER of the management signal is plotted as a function of the average power of the management signal. The BER-curve is plotted for several different coupling ratios.

![BER curves of the management signal](image)

**Figure 7-3 BER of the management signal at different coupling ratios.**

In Figure 7-3 we see the degradation of the BER when the coupling ratio is getting higher (the signal power is getting higher). If we extrapolate the BER-curve we see that the receiver sensitivity lies around -45 dBm.

Furthermore we see that if the signal power is 15 dB above the average received power of the management signal the transmission reliability decreases rapidly.

A possible cause is that in our design, we only take care of amplitude recovery instead of amplitude and clock recovery. When the distortion is increasing pulse narrowing can occur (see Figure 7-4). Since the RS232 receiver (UART) has prearranged sample moments after receiving a high flank of the start-bit, the result can be that the pulse width is smaller than the distance needed for an error free transmission.

If we introduce a data clock recovery we may be able to cope with these errors.
Another possible explanation is an increase of the RIN of the laser of the management due to reflections or low speed switching of the laser. In Figure 4-1 we see that the RIN increases as the power decreases. Since we are working at very low speed the RIN may rise, when the level of the laser is set to an "0" or "1L". Reflections in the laser and internal reflections (Rayleigh scattering) can increase the RIN up to −70 dB/Hz.

When we adjust the RIN of the laser of the management signal and the noise figure of the receiver we get results as given in Figure 7-5.

In the simulation we used the following values:
Table 7-2 Characteristic values of the experiment of Figure 7-5.

<table>
<thead>
<tr>
<th>Characteristic</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>RIN_{laser-data}</td>
<td>-140 dB/Hz</td>
</tr>
<tr>
<td>RIN_{laser-management}</td>
<td>-70 dB/Hz</td>
</tr>
<tr>
<td>F_{se}</td>
<td>42 dB</td>
</tr>
<tr>
<td>Δf</td>
<td>250 kHz</td>
</tr>
<tr>
<td>R</td>
<td>0.9 A/W</td>
</tr>
<tr>
<td>I_{dark}</td>
<td>10^{-9} A</td>
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</tr>
<tr>
<td>Bitrate data signal</td>
<td>622 Mbit/s</td>
</tr>
</tbody>
</table>

A third explanation is given by the random bit pattern of the data generator of the data signal. The power spectral density (see Figure 4-3) of this pattern (PRBS=2^{27}-1) is more concentrated in the lower frequencies. Therefore S_{dd}(0) and the effective noise bandwidth increase. As a result the sensitivity is becoming worse.

### 7.1.4 Conclusions

When we compare the experimental data of Figure 5-9 with the measurement results, we see that we lost approximately 20 dB.

Possible explanations for this loss are the limitations in our design. The use of many different building blocks increases the noise figure. Implementation in an integrated circuit (IC) should reduce the noise figure. Furthermore we should introduce a clock recovery, so we can cope with the errors due to pulse narrowing. Furthermore we should reduce the optical reflections by the use of optical isolators to decrease the RIN.

Since the measurement data reveals only a small area it is hard to adapt our model to the measurement. For this reason the values of the set-up cannot be fitted with great accuracy.

### 7.2 Bit-Error Rate measurement of the data signal

In the following paragraphs the measurements results are given of the BER-measurements of the data signal. First the measurement set-up is given, followed by the results and conclusions.

#### 7.2.1 Measurement set-up

The set-up for measuring the BER of the data signals differs slightly from the one mentioned in section 7.1.2. Again we see three attenuators to set the power ratio of the data and management signal and to change the total power received by the photodiode. Instead of using a computer program to determine the BER, we now can use a normal BER-set.

The management signal now exists of a continued bit-stream with a pattern, which consists out of switching bits (101010..). Since we want to determine the influence of the management signal on the data signal, we need a continuous stream of management data and not a burst mode stream. Another reason for choosing this pattern, is that we want to see if the switching of levels in the bit-stream of the management influences the results.

The measurement set-up is given in Figure 7-6.
The receiver of the 622 Mbit/s has a DC-cancellation loop. Since the receiver is not specially designed for receiving a coupled signal, we try to change the speed of adapting to a new DC-level to the new level of the management signal by changing a capacitor outside the cavity of the photodiode.

7.2.2 Measurement results

Since the receiver of the data-signal has a DC-cancellation loop, we did several experiments. Following the data-sheets the capacitor of the DC-cancellation loop should be 390 pF. We tried two other capacitors to influence the speed of adapting to a new DC-level. The following capacitors were used with the following assumptions:

- 0 F (short circuit) | Fast adjustment to new DC-level
- 390 pF (data-sheets) | Medium adjustment to new DC-level
- 10nF | Slow adjustment to new DC-level

The results of the different experiments are given in Figure 7-7 to Figure 7-9.
Figure 7-7 BER of the data signal, without capacitor (fast)

In the above figure we see the BER as a function of the average power of the data-signal. The measurement is done for several different coupling ratios of the data and management signal. If we compare the three different figures, we see the best result, when we use a capacitor of 390nF. It seems impossible to adapt the laser to fast adjustments to new DC-levels.

Figure 7-8 BER of the data signal, with a capacitor of 390pF (following the datasheets)
A way of verifying the fact that this receiver is not suitable for adapting to changes in the DC-level is to look at the electrical eye-patterns of the data signal (see Figure 7-10 and Figure 7-11). The eye-pattern of a receiver after its decision circuit should be distortionless. However, here we can see that the distortion of the signal mainly exists due to the fluctuations of the DC-level when transmitting a "1" of the management signal. (This can be seen when we change the bit-rate of the management signal from 10 kbit/s to 1 bit/s)

Our conclusion is indeed, that this receiver is not suitable for adapting to DC-levels, which change with the clock-frequency of 200kHz.

Figure 7-10 Eye-pattern of the data-signal when sending a "0" of the management signal after the decision circuit of the receiver.
Figure 7-11 Eye-pattern of the data-signal when sending an "1" of the management signal after the decision circuit of the receiver.

7.2.3 Conclusions

If we compare the measurement results with the theory of section 5.3 we see that the measurements follow the numerical results closely. If we tolerate a total penalty of 1.5 dB we will be able to transfer a management signal, of which the average power is 10 dB down.

Another thing we see is that our data receiver is not able to adjust to changes in the DC-level, which exists due to the transmission of management signals. The conclusion that we can draw is that the results may be worse than the theory due to fact, that the decision threshold is not moving as fast as the received eye. So we have to look for other receivers, which adapt better to these circumstances.

Furthermore we see that the difference between the experiment of Figure 5-11 and the measurement results is increasing when the coupling ratio decreases (management signal power is increasing in comparison with the data signal power). In section 7.1 we concluded that the RIN of the management laser was a possible cause of the degradation in the receiver sensitivity. When the average power of the management signal increases and the gap between the average power of the management and data signal is decreasing, the influence of the RIN of the management laser is becoming an important noise source. As a result the error rate increases.
8 CONCLUSIONS & RECOMMENDATIONS

In this report we discussed a model in which management and data signals are multiplexed in one fiber. As a result of the coupling of these signals the Relative Intensity Noise (RIN) and the shot noise increase, which leads to a lower sensitivity of the receiver. Furthermore both signals interfere. Since the two signals have a large difference in the frequency domain, we used two different approaches to calculate the disturbance. To determine the interference of the data signal on the management signal we used a bandpass filter to select a small bandwidth. Between this bandwidth the power spectral density is constant, so we treated the disturbance as a white noise source.

The interference of the management signal on the data signal was calculated by dividing the combination of both signals in separate signal states.

We discussed two different modulation schemes of the management signal. Since we had to filter the DC-component of the data-signal, the direct detection modulation scheme was replaced by an ASK modulation scheme. This approach needed a few adjustments in our theory. The use of an double-sided changed the probability densities of the received signal, which resulted in a translation of the threshold current. Furthermore some recommendations were given to implement this principle in a WDM network like the TOBASCO project. The summation of currents and the possibility of the placement of a photodiode before the HDWDM and the use of another modulation technique (CDMA) are interesting options, which enables us to reduce the extinction ratio of the management signal and minimise the penalty of combining the data and management signals.

Since this project is in an experimental phase the laser of the data signal did not transmit the management signal. A transmitter was designed to modulate the RS232 signal into an ASK modulated signal. In the receiver design a High Pass Filter was add to a standard transimpedance amplifier and filters and a rectifier were used to obtain the RS232 format.

The transceiver designed was used in several experiments. To measure the Bit Error Rate (BER) of the management signal as special program was made to calculate, which was able to send and transmit signals based upon the RS232 protocol (burst mode). With the use of this program, we did several experiments, in which we saw that the sensitivity of the receiver was 20 dB worse than an ideal receiver (in theory). Possible explanations for this decrease in sensitivity are:

- the increase of the Relative Intensity Noise (RIN) due to reflections and due to the low power output of the management laser during parts of the signal;
- the introduction of errors due to pulse narrowing. The receiver designed recovers the amplitude, but does not recover the clock frequency and the pulse shape;
- the power spectral density of the data pattern (PRBS=2^{23}-1) consists of high values in the lower frequencies.

In the BER-measurements of the data signal, we saw that the receiver we used was unable to adapt to the changing circumstances, caused by the different power levels of the management signal. We saw noise on the electrical output of the receiver after the decision circuit, which is extraordinary. This perception shows us that the receiver was not able to coop with these changes in power.

The theory discussed seems to be a theory which can be used to calculate the penalty in the receiver sensitivity. But the possible explanations have to be checked to be sure this theory is beyond any discussion. Furthermore the transmitter of the management signal has to be implemented in the laser of the data signal. And the receiver has to be improved to become a higher sensitivity. Implementation of the circuit in an Integrated Circuit (IC) should reduce the thermal noise and by the use of optical isolators we should be able to reduce the RIN. The receiver of the data signal should be able to coop with the different power levels of the management signal. Since it is very important to reduce the penalty, study to the speed in which receivers adapt to a moving a patterns is an interesting subject, which can play a major role in future receiver design.
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APPENDIX A) LAYOUT OF THE OPTICAL RS232 TRANSMITTER
**APPENDIX B) LIST OF SYMBOLS AND ABBREVIATIONS**

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>ASK</td>
<td>Amplitude Shift Keying</td>
</tr>
<tr>
<td>BER</td>
<td>Bit Error Rate</td>
</tr>
<tr>
<td>CDMA</td>
<td>Code Division Multiple Access</td>
</tr>
<tr>
<td>EDFA</td>
<td>Erbium Doped Fiber Amplifier</td>
</tr>
<tr>
<td>$F_n$</td>
<td>Noise figure (EDFA or receivers)</td>
</tr>
<tr>
<td>G</td>
<td>Optical gain EDFA</td>
</tr>
<tr>
<td>$h$</td>
<td>Planck’s constant</td>
</tr>
<tr>
<td>HDWDM</td>
<td>High Density Wavelength Division Multiplexing</td>
</tr>
<tr>
<td>$I_{\text{avg}}$</td>
<td>Average current</td>
</tr>
<tr>
<td>$I_d$</td>
<td>Dark current</td>
</tr>
<tr>
<td>$I_{\text{th}}$</td>
<td>Threshold current</td>
</tr>
<tr>
<td>$k_B$</td>
<td>Boltzmann’s constant</td>
</tr>
<tr>
<td>$L$</td>
<td>Attenuation (fiber and optical splitter losses)</td>
</tr>
<tr>
<td>$n_{sp}$</td>
<td>Spontaneous emission factor (EDFA)</td>
</tr>
<tr>
<td>ONU</td>
<td>Optical Network Unit</td>
</tr>
<tr>
<td>$P_{\text{opt}}$</td>
<td>Optical power data signal</td>
</tr>
<tr>
<td>$P_{\text{m}}$</td>
<td>Optical power management signal</td>
</tr>
<tr>
<td>$P_{sp}$</td>
<td>Spontaneous emission noise power</td>
</tr>
<tr>
<td>$q$</td>
<td>Electron charge</td>
</tr>
<tr>
<td>RIN</td>
<td>Relative Intensity Noise</td>
</tr>
<tr>
<td>$R_L$</td>
<td>Load Resistor (receiver)</td>
</tr>
<tr>
<td>$S_{dd}$</td>
<td>Spectral density data signal</td>
</tr>
<tr>
<td>$T$</td>
<td>Temperature or bit time (see context)</td>
</tr>
<tr>
<td>$\Delta f$</td>
<td>Effective noise bandwidth</td>
</tr>
<tr>
<td>$\Delta \nu_{\text{opt}}$</td>
<td>Optical filter bandwidth</td>
</tr>
<tr>
<td>$\nu$</td>
<td>Frequency of the light</td>
</tr>
<tr>
<td>$\sigma_{\text{std}}$</td>
<td>Standard deviation of total noise</td>
</tr>
<tr>
<td>$\sigma^2_d$</td>
<td>Variation in the current due multiplexing of signals</td>
</tr>
<tr>
<td>$\sigma^2_{\text{RIN}}$</td>
<td>Variation in the current due to Relative Intensity Noise</td>
</tr>
<tr>
<td>$\sigma^2_{\text{shot}}$</td>
<td>Variation in the current due to shot noise</td>
</tr>
<tr>
<td>$\sigma^2_{\text{shot-sp}}$</td>
<td>Variation in the current due to shot-spontaneous beat noise</td>
</tr>
<tr>
<td>$\sigma^2_{\text{sp-sp}}$</td>
<td>Variation in the current due to signal-spontaneous beat noise</td>
</tr>
<tr>
<td>$\sigma^2_{\text{sp-sp}}$</td>
<td>Variation in the current due to spontaneous-spontaneous beat noise</td>
</tr>
<tr>
<td>$\sigma^2_T$</td>
<td>Variation in the current due to thermal noise</td>
</tr>
</tbody>
</table>