System and receiver design for two-dimensional optical storage

Citation for published version (APA):

DOI:
10.6100/IR597907

Document status and date:
Published: 01/01/2005

Document Version:
Publisher’s PDF, also known as Version of Record (includes final page, issue and volume numbers)

Please check the document version of this publication:
• A submitted manuscript is the version of the article upon submission and before peer-review. There can be important differences between the submitted version and the official published version of record. People interested in the research are advised to contact the author for the final version of the publication, or visit the DOI to the publisher’s website.
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System and Receiver Design for Two-Dimensional Optical Storage

PROEFSCHRIFT

ter verkrijging van de graad van doctor aan de Technische Universiteit Eindhoven, op gezag van de Rector Magnificus, prof.dr. R.A. van Santen, voor een commissie aangewezen door het College voor Promoties in het openbaar te verdedigen op dinsdag 15 november 2005 om 16.00 uur

door

André H.J. Immink

geboren te Lemele, Nederland
The work described in this thesis has been carried out at the Philips Research Laboratories Eindhoven, The Netherlands, as part of the Philips Research Programme.
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Chapter 1

Introduction

Since the introduction of optical storage in the form of the Video Long Play (VLP) and later the Compact Disc (CD) there has been a continuous effort to increase the storage capacity of the discs and the data rate of the read-out and write systems. This thesis presents a new technology referred to as two-dimensional optical storage (TwoDOS) in which an optical storage medium is scanned by multiple laser beams. Clearly this results in a higher data rate of the system. Moreover, by combining the simultaneously obtained signals of adjacent tracks in a digital processing system, it is possible to increase the storage capacity of the disc, without a further modification of the main optical parameters of the system. In Section 1.1 we will first give a brief review of the history of optical storage and explain why it has become such a great market success. Then a basic introduction is given on the optical system in Section 1.2. The most common optical formats are presented in Section 1.2.1. From this format overview trends can be derived for the data rate of optical systems and for the storage capacity of the medium. Based on these trends it is possible to set requirements for the future (4th) generation of optical storage as is done in Section 1.3. To live up to these expectations, research is done on numerous candidate technologies. A disadvantage of most of these technologies is that data rate does not keep pace with storage capacity and that the time to write or read out a complete disc is increasing rapidly. This is described in Section 1.4. Multi-beam read-out as described in Section 1.4.1 appears to be a natural solution to this problem. The TwoDOS system, which additionally increases the storage density, is introduced briefly in Section 1.5. The remainder of the thesis is devoted to a detailed description of the TwoDOS technology. An outline of the thesis is given in Section 1.6.

1.1 Optical Storage: History and Trends

In the 1960’s the enormous popularity of the gramophone record and the rapid growth of television made it an obvious idea to look at the possibility of storing video signals on a disc. The use of a disc, as an information carrier, has the advantage that one has immediate access to any part of the programme, in contrast to the slow accessibility
Introduction

of tape-based storage. Of even greater importance is the low price of disc storage due to the use of production methods similar to that of the gramophone disc [1] i.e. mechanical impressing the information in the disc by using a master stamper. It was also recognized early on that optical read-out of information from the disc has a distinct advantage over mechanical read-out [2]. Any wear is eliminated because there is no contact between the medium and the read-out device. Furthermore, the contact-less read-out makes the medium easily removable. These properties made the optical disc the removable storage medium of choice and they were the basis for its unabated advance in the consumer market.

In 1972 the Video Long Play (VLP) system was introduced with the goal of playing back video content on a television set [3, 4]. The standard was established by Philips, Thomson, Music Corporation of America (MCA) and later Pioneer. The system uses discs of a transparent polymer material with standardized diameters of 20 and 30 cm and a thickness of 2.6 mm. The information on these discs is stored in tracks spiraling outward with a track-to-track distance of $1.6 \mu\text{m}$. The discs are manufactured by mechanical impressing of information in the disc using a master stamper to allow a cheap and fast replication process. The master stamper is made by illuminating a $100 \mu\text{m}$ thin photo-sensitive layer on a glass substrate and developing this photo resist to remove it at positions where it was illuminated. This development process is highly non-linear which limits the choice of modulation techniques to two-level signal recording. For this reason, the information is present in so-called pits and non-pits. The VLP system uses a frequency modulation to store the audio and video information on the disc. Two rotational velocities of 25 and 30 Hz are used depending on the TV standard (PAL/SECAM, NTSC) in such a way that two interlaced TV frames (at a refresh rate of 50 Hz or 60 Hz respectively) can be stored on one revolution of the disc when operating in a Constant Angular Velocity (CAV) mode. This allows, by synchronously jumping back and forward, various trick modes such as still-picture, slow motion and reverse play without the use of a frame memory which wasn’t available at that time. The (high) price for these nice options is a short playing time of only 36 minutes and an areal density on the outer radii that is much lower than technically achievable. In the Constant Linear Velocity (CLV) mode where rotation frequency drops from 25 Hz on the inner diameter to 9.5 Hz on the outer diameter the still-picture option is absent but the areal density is constant across the disc and playing times up to 1 hour can be realized.

It was soon recognized that the small size of the pits (width of $0.4 \mu\text{m}$; average length of $0.6 \mu\text{m}$) requires some special protection of the information layer. Small dust particles and scratches on the medium can easily damage the imprinted information layer and lead to signal deterioration and drop-outs. The idea to use a transparent, protective layer on top of the information layer and more important the idea to use the disc substrate itself as this protective layer [5] has proven to be one of the key ideas


<table>
<thead>
<tr>
<th>Property</th>
<th>Advantage</th>
</tr>
</thead>
<tbody>
<tr>
<td>• mechanical impression of information using a master stamper.</td>
<td>• cheap replication of discs.</td>
</tr>
<tr>
<td>• no mechanical contact between medium and read-out device.</td>
<td>• no mechanical wear during read-out and easily removable storage medium.</td>
</tr>
<tr>
<td>• protective cover-layer in the form of the disc substrate.</td>
<td>• robust against dust and scratches.</td>
</tr>
</tbody>
</table>

Table 1.1: Key differentiators of optical storage systems.

that made the optical storage system the robust information carrier as we know it today. Table 1.1 shows an overview of these key differentiators that make the optical storage system the system of choice for many of today’s applications.

In spite of these advantages the VLP system never became a big market success due to the limited playing time and the lack of a recording option which made competition with the video cassette recorder rather difficult [6]. In the meantime research was done to replace the old gramophone disc (also called Long Play disc or just LP) by an optical system to distribute audio content. The large increase in areal capacity when going from the mechanical ‘needle’ to the optical scanning spot was used in two ways. Firstly, the optical disc was reduced considerably in size compared to the LP. Secondly, the audio signal was digitized causing a large increase of the required bandwidth of the signal. The major benefit of the latter was that digital error correction (ECC) codes could be applied. This made the system even more robust against dust and scratches compared to the VLP but above all it brought the ultimate goal of ‘perfect’ audio replay within reach.

1.2 Read out of Single-Spiral Optical Discs

Before continuing the discussion with an overview of the various optical formats, first the read-out mechanism is briefly discussed. As already mentioned, the data is written in a track spiraling outwards from an inner radius (R1) towards an outer radius (R2), see Fig. 1.1. The data is read out with a focused laser beam generated by the optical part of the system. A drawing of the optical light path is shown in Fig. 1.2. A divergent light beam is generated by a semiconductor laser diode. The light is pointed towards a beam-splitting cube and then directed towards the objective lens via a collimating lens that makes a parallel laser bundle. The objective lens focuses the parallel bundle onto the rotating storage medium. By actuating the objective lens towards and from the disc, ideal focus can be maintained even when the disc is not ideally flat.
Introduction

Figure 1.1: Schematic drawing of the outward spiraling track on an optical disc. In the inset the pits on the disc are shown in detail.

Additionally, by actuating it in the radial direction (the direction perpendicular to the along-track direction) the spiraling track can be followed accurately. The focused light beam is reflected and diffracted by the storage medium, after which the light is collected again by the same objective lens. Via the same optical path and the beam splitter it is now focused onto a photo detector that transfers the optical signal into an equivalent electrical signal. This electrical signal contains information on the pit sequence on the disc from which we can derive the original bit sequence by suitable processing of the signal. Several other optical components can be found in the light path of Fig. 1.2. These ‘auxiliary’ components are not further discussed here.

The electronic part of the system is shown schematically in Fig. 1.3. It is commonly referred to as ‘Figure 1’ because its basic form is presented as the first figure in many books on digital storage, see e.g. [7]. The figure includes references to the chapters in this thesis that are devoted to each of the building blocks of the system. The diagram can be read starting from the source, which generates a stream of (binary) data. Redundant information is added to this stream to be able to correct any errors that might occur when ‘transmitting’ the data across the channel. Then a modulation code is applied. It is the task of modulation coding to transform the input stream of user data into an output stream of channel data that has some desirable properties (such as the ability to transfer it reliably across the channel). The data is now ready to be written on the disc. The data modulates the laser light beam according to a so-called write strategy. This write strategy prescribes a pulse sequence that
1.2 Read out of Single-Spiral Optical Discs

Figure 1.2: The optical light path.

Figure 1.3: 'Figure 1' of optical storage.

Each block in the writing path of the system has its counterpart in the read-out branch. The electrical signal at the output of the photo detectors is processed to recover the original bit stream as reliable as possible. A small fraction of the bits, however, is detected erroneously. A demodulation step reverses the modulation coding step. In this process the erroneously detected user bits might spread over a larger part of the output channel bit stream. The redundant information that was added at the transmitting side of the system is now used to correct those errors. In case the number of errors in the channel bit stream is not too high the final output user bit stream is
Introduction

perfectly equal to the input data. This ‘transparency’ property of the channel offers e.g. perfect audio and video reproduction, and reliable archiving possibilities.

1.2.1 Digital Optical Formats

The digital audio long play disc that originated from the VLP system was renamed compact disc (CD; after the compact cassette). It was officially introduced in 1982. Besides the digitization of the data and a change in laser wavelength to 780 nm, the basic principle was kept the same. The storage density of 680 MByte on a single layer disc with a diameter of 12 cm was reached using a track pitch of 1.6 µm and a channel bit length of 277 nm. This storage density is directly dependent on the size of the optical spot. In the ideal case the spot is diffraction limited and its size depends only on the wavelength and the numerical aperture (NA). The NA is defined as the sine of the opening angle of the light cone that is focused on the storage medium. For CD the NA=0.45. The resulting Airy intensity profile [8] has a full-width at half maximum (FWHM) just over 1 µm. The thickness of the transparent disc (that serves as the protecting cover layer for the data) is 1.2 mm. Fig. 1.4 shows an overview of existing optical storage formats together with the main parameters. By reducing

<table>
<thead>
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<th>CD</th>
<th>DVD</th>
<th>BD</th>
</tr>
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<td><img src="image1.png" alt="CD Diagram" /></td>
<td><img src="image2.png" alt="DVD Diagram" /></td>
<td><img src="image3.png" alt="BD Diagram" /></td>
</tr>
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</table>

*Figure 1.4: Overview of existing optical storage formats.*

the wavelength of the used laser light and by increasing the numerical aperture the storage capacity of the disc has been increased in a few steps. This is the so-called ‘physical roadmap’ of optical storage. The ‘digital versatile disc’ (DVD) uses a laser with a wavelength of 650 nm and the NA is increased to 0.6. By reducing the margins slightly (made possible by more advanced signal processing and manufacturing
1.3 4th Generation Optical Storage

Table 1.2: Key parameters of the various formats. The user bit length is calculated based on the channel bit length, the overhead for error correction and the rate of the modulation code. The d,k-constraints determine the minimum and maximum number respectively of consecutive ones and zeros in the channel bit stream.

<table>
<thead>
<tr>
<th>Property</th>
<th>CD</th>
<th>DVD</th>
<th>BD</th>
<th>NF</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \lambda ) [nm]</td>
<td>780</td>
<td>650</td>
<td>405</td>
<td>405</td>
</tr>
<tr>
<td>NA</td>
<td>0.45</td>
<td>0.6</td>
<td>0.85</td>
<td>1.9</td>
</tr>
<tr>
<td>FWHM [( \mu m )]</td>
<td>1.040</td>
<td>0.650</td>
<td>0.286</td>
<td>0.128</td>
</tr>
<tr>
<td>d,k-constraint</td>
<td>2,10</td>
<td>2,10</td>
<td>1,7</td>
<td>1,7</td>
</tr>
<tr>
<td>Channel bit length [nm]</td>
<td>280</td>
<td>133</td>
<td>74.5</td>
<td>33.3</td>
</tr>
<tr>
<td>User bit length [nm]</td>
<td>700</td>
<td>313</td>
<td>137</td>
<td>61</td>
</tr>
<tr>
<td>ECC rate</td>
<td>0.85</td>
<td>0.85</td>
<td>0.8170</td>
<td>0.8170</td>
</tr>
<tr>
<td>Track pitch [( \mu m )]</td>
<td>1.6</td>
<td>0.74</td>
<td>0.32</td>
<td>0.1432</td>
</tr>
<tr>
<td>Inner radius (R1) [mm]</td>
<td>24</td>
<td>24</td>
<td>24</td>
<td>24</td>
</tr>
<tr>
<td>Outer radius (R2) [mm]</td>
<td>58</td>
<td>58</td>
<td>58</td>
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<tr>
<td>Surface [cm(^2)]</td>
<td>87.59</td>
<td>87.59</td>
<td>87.59</td>
<td>87.59</td>
</tr>
<tr>
<td>Surface [inch(^2)]</td>
<td>13.58</td>
<td>13.58</td>
<td>13.58</td>
<td>13.58</td>
</tr>
<tr>
<td>User Capacity [GB]</td>
<td>0.68</td>
<td>4.7</td>
<td>25.0</td>
<td>125.0</td>
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<tr>
<td>Density [Gb/inch(^2)]</td>
<td>0.40</td>
<td>2.78</td>
<td>14.74</td>
<td>73.65</td>
</tr>
</tbody>
</table>

methods) a storage capacity of 4.7 GB on a single layer is achieved. This has been realized by using a track pitch of 0.74 \( \mu m \) and a bit length of 133 nm (see Table 1.2 for an overview of these parameters). Recently, the Blu-ray Disc (BD) standard was introduced. It offers a capacity of 25 GB and uses a blue-violet laser diode with a wavelength of 405 nm. The NA is 0.85. The tolerance to disc tilt goes with the third power of the numerical aperture. This means that for increasing numerical aperture disc-tilt becomes a serious issue. This is counteracted partially by choosing a thinner protective layer (0.6 mm for DVD and 0.1 mm for BD). This, however, decreases the robustness against dust and scratches.

1.3 4th Generation Optical Storage

From Table 1.2 it can be derived that each new generation of optical storage offers about a factor of 5 to 6 more storage capacity and a factor of 2 to 3 higher data rate. Extrapolation of this trend results in an expected capacity of about 150 GB per layer and a data rate of about 100 Mb/s for the 4th generation of optical storage. The exact capacity and data rate has always been defined by a particular target application for each new format. The CD has been introduced to store 74 minutes of audio. Later it
was used for software distribution and for archiving of data. With DVD (and MPEG2 encoding) it became possible to store a full-length movie (about two hours) with high quality on a single disc. Again the format was also used for software distribution. In particular PC-games with a large video or high-resolution-graphics content needed the increase in storage capacity and the higher data rate. The targeted application of Blu-ray Disc is the distribution of high-definition video and the recording of high-definition television broadcasts. It is expected that archiving computer and multimedia content (photo’s, home-video’s, etc.) will be a second application of the BD format.

For the 4th generation format the application in consumer electronics is less clear. There is a trend to record digital audio and video on hard-disks. The device containing the hard-disk is becoming the multi-media server of the home and also stores personal data such as photo’s, home-video’s, etc. The multi-media servers generally contain an optical drive to archive data and to exchange content with friends and family. The optical disc should have a reasonable storage capacity and at the same time be removable, but above all the system should be able to record a full disc within a short time (preferably a few minutes). Other applications in the consumer domain are 3-dimensional video, multi-angle video, interactive video and gaming with full-resolution video content. For all these applications a high storage capacity combined with a high data rate is required.

1.4 Increasing the Data Rate

The need for a higher data rate is not a contemporary one. This is reflected e.g. in the increased read out speed for the Compact Disc (CD) starting from 1x CD at 4.32 Mb/s channel data rate in the beginning of the nineties to 48x CD at 207 Mb/s (at the outer radius of the disc) at the start of the 21st century, see Fig. 1.5. In this way the time to write or read a complete CD has dropped from 74 minutes to less than 2.5 minutes (note that the rewritable format lags in data rate and does not yield the same improvement). A similar data-rate trend can be observed for DVD and is also expected for BD. Unfortunately, in optical storage, the maximum data rate does not keep pace with the growth in storage capacity. The maximum data rate is limited by the maximum rotation velocity of the disc, limited by the maximum centrifugal forces the polycarbonate disc can endure without breaking. The ultimate linear velocity at the outer radius of a standard 12 cm disc is approximately 56 m/s, and is independent of the optical format. Storage capacity on optical discs has increased mainly by scaling the physical parameters of wavelength, (\(\lambda\)) and Numerical Aperture (NA). The size of a user bit cell scales proportional to \((\lambda/NA)^2\). The maximum data-rate, however, depends on the tangential bit size only and scales linearly with \((\lambda/NA)\). This means that the maximum data rate to capacity ratio drops with each next optical
1.4 Increasing the Data Rate

Figure 1.5: Data rate of storage media as function of time.

storage generation. Fig. 1.6 shows a column chart that underpins this observation. About 5 minutes are needed to record a full Digital Versatile Disc (DVD) disc at the highest possible rotation speed. This increases with a single-layer Blu-ray Disc (BD) to almost 12 minutes and twice this value for a dual layer (DL) version. It will even be more than 25 minutes for a Near Field (NF) format with an NA of 1.9 [9].

1.4.1 Multi-Beam Recording

A rather obvious way out of this problem is parallel access, where a number of tracks are read out and recorded simultaneously. This was already proposed by Zen-Kenwood around 1997 and was known under the ‘TrueX’ trademark. Information can be found in numerous patents [10–12]. A schematic representation of the system as provided by Zen-Kenwood is given in Fig. 1.7. The ‘TrueX’ trademark found its origin in the fact that most commercial drives are CAV drives that advertise the highest speed, which is only realized on the outer radius of the disc. The Zen-Kenwood technology, in contrast, is a CLV one which achieves the advertised speed over the complete disc area. Commercial products in the form of high speed CD-ROM drives were released but didn’t become a market success, possibly due to the fact that at the same time DVD was introduced, with an intrinsically higher data rate. Moreover,
recordable and rewritable systems became popular. At that moment no multi-beam lasers were available that are needed to realize a recordable multi-beam system.

Another major problem was the strong market position of the standardized CD and DVD format. Although the Zen-Kenwood system made use of an array of 7 spots the data rate was not increased by this factor. The reason for this is that the data on the CD and DVD disc is organized in blocks on a single spiral (Fig. 1.1) and one needs to jump (over 7 tracks) after each revolution in order not to read out the same data multiple times. This jumping is time consuming (at best about 5 to 10% of a revolution) and eliminates part of the data-rate gain. Moreover, valid data extraction is only possible after the start of the next logic data block. A related issue is that data is not received in the correct order anymore. The received data blocks need to be stored in memory and rearranged to be able to provide the host with a consistent stream of data. This increases drive cost and might influence access time adversely. It means that parallel access does not improve access time, which depends only on factors such as block-size, rotational velocity of the disc, jump speed (and number of retries before the correct track is found) and the propagation delay of the digital signal processing.

**Figure 1.6:** Ratio of maximum data rate and disc capacity for the various optical formats. Dual layered versions are indicated with DL.
1.5 Two Dimensional Optical Storage (TwoDOS)

It is clear from the previous section that for a high data rate system with a multiple-spot read-out it is very advantageous to adapt the format of the disc accordingly. Instead of a single spiral as shown in Fig. 1.1 we propose a so-called broad spiral containing multiple bit-rows as shown in Fig. 1.8. There is no need for jumps anymore and the data rate of the system increases with a factor equal to the number of bit-rows within the spiral. Apart from a higher data rate, however, also an increase in storage density can be realized by a further change in the data format. This can be explained as follows: For state-of-art densities the intensity profile of the spot extends over multiple bits thereby causing inter symbol interference (ISI). The ISI along the single spiral track can be taken into account by suitable detector algorithms, such as the Viterbi algorithm. For ISI due to bits in adjacent tracks, denoted as radial ISI, this is more difficult. These tracks are not scanned with a laser spot and the bits in these tracks are not taken into account in the detection process. The radial ISI is generally considered to be a disturbance. However, for the broad spiral system as shown in Fig. 1.8 multiple tracks are scanned simultaneously. By ordering the data on a fixed lattice a stationary 2D bit configuration is present under the spot at each detection instant. Two-dimensional (2D) signal processing makes it possible to take the complete 2D ISI into account in a 2D detection process, thereby offering a potential improve-
Figure 1.8: Schematic drawing of a broad-spiral format. The broad spiral consist of multiple bit-rows, which are read-out simultaneously using an array of laser spots. The inset shows an example of how this would look like on the medium.

A further advantage of 2D optical storage (next to higher storage density and data rate) is that it can be applied on top of improvements in numerical aperture of the system and/or a change in wavelength. The signal processing is ‘orthogonal’ to the evolutionary road-map of optical storage and the gain in storage density obtained by 2D signal processing is multiplicative to the gain derived from the physics. This is schematically shown in Fig. 1.9.

This thesis gives an overview of the optical storage technology on which the 2D system is based and to provide insight into the way we can exploit the knowledge about the characteristics of the optical channel by applying suitable 2D signal processing algorithms and by doing 2D bit-detection.
Figure 1.9: Roadmap of optical storage showing the orthogonality between signal processing and the traditional ‘evolutionary physics’ roadmap of optical storage. Near Field and Holographic storage are discussed in Appendix A.

1.6 Contents of this Thesis

In the following chapter the interested reader can find an overview of possible technologies other than the 2D technology to further increase the storage capacity (and data rate) beyond the BD format. This chapter may be skipped for readers merely interested in the 2D optical storage (TwoDOS) technology, which will be dealt with in the remainder of this thesis. The TwoDOS concept will be explained in Chapter 2 without going into too much technical detail. The main purpose of the chapter is to explain the basic techniques that are needed for the realization of a two-dimensional storage system. The chapter will also discuss the basic differences between a conventional 1D system and the novel 2D system. Chapter 3 includes a more detailed analysis of the 2D channel. It starts with a brief overview of 2D lattice characteristics followed by a scalar diffraction model of the readout of an optical disc that is specifically suited for signal processing purposes. The model is extended with a noise characterization by discussing the physical origins of noise in the system. Chapter 4 addresses the short-comings of the optical channel and ways to reduce the effect of these short-comings by proper 2D modulation coding of the user data. Receiver design is discussed in Chapter 5. It focuses primarily on the adaptation loops in the presence of a detector with a large detection delay. In Chapter 6 the 2D bit-detector is described in detail. Here the cost of silicon implementation will be an important
boundary condition for the solution that is finally chosen. The results of simulation and experiments can finally be found in Chapter 7. The knowledge that is developed in preceding chapters is used here to interpret the outcome of experimental results and to be able to identify those issues that hamper a further increase in density or data rate. This leads to final conclusions and recommendations for further research in Chapter 8.
Chapter 2

The TwoDOS Concept

This chapter discusses the basic concept of two-dimensional (2D) optical storage. We will define what we mean by two-dimensional optical storage in Section 2.1. In Section 2.2 it is explained how two-dimensionally organized data can be read out with a linear array of spots. Section 2.3 describes the benefits of 2D optical storage compared to its 1D counterpart. To fully understand the difference between 2D and 1D storage a short introduction on cross-talk cancellation is included. In Section 2.4 several mastering techniques such as laser beam recording (LBR), electron-beam recording (EBR) and liquid immersion mastering (LIM) are discussed in relation to issues that are specific for the mastering of 2D formats. Finally, a brief overview of rewritable and recordable optical storage technology is given in Section 2.6 with a link to the applicability for the TwoDOS concept.

2.1 Storing Data on a 2D Lattice

At a first glance considering two-dimensional optical storage does not seem to be very original at all. Isn’t all disc-based storage two-dimensional, simply because the data is present on a 2D surface? The answer is: No! Conventionally, the data is considered to be one-dimensional in the sense that data is stored as a single 1D bit sequence that is arranged along a spiral track on the disc. Successive revolutions of the spiral are simply referred to as ‘tracks’ along a spiral that evolves along the tangential direction of the disc (see Fig. 1.1). The track separation is chosen such that adjacent tracks do not interfere significantly during read-out. For example in CD the track pitch was chosen to be 1.6 \( \mu m \), which is well beyond the first zero of the intensity profile of the optical spot, which is at about 1 \( \mu m \). Moreover, there is no logical or physical relationship between the bits in adjacent tracks. They are recorded asynchronously and are separately encoded by the modulation encoder during channel coding. The bits in successive tracks even belong to different error correction blocks, because the size of these blocks is intentionally made smaller than the number of bits on one circumference of the disc.

In a 2D system this is different: A number of adjacent tracks are aligned in phase
and the data belongs to one logical entity. A so-called ‘broad spiral’ or ‘meta-spiral’ is defined in which the tracks (in the 2D case called ‘rows’) do have a logical and physical relation to each other. A schematic representation of the proposed TwoDOS format is given in Fig. 2.1 [13].

![Diagram of the TwoDOS format](image)

*Figure 2.1: The TwoDOS format [13].*

The broad spiral consists of a number of bit-rows stacked upon each other with a fixed phase relation between the rows. A 2D closed-packed hexagonal ordering of the bits is chosen because it has a 15% higher packing fraction than the square lattice as proposed in [8, 14]. On the other hand 6 nearest neighbours will cause the bulk of the inter symbol interference (ISI) in the hexagonal case, while on the square lattice there are only 4 of these nearest neighbours. The number of bit-rows in one spiral is typically chosen to be around 10 (for our experimental systems: $N_r = 7$ or $N_r = 11$). The broad spirals are separated by a so-called guard band with a maximum width of one empty bit row (Fig. 2.1). This guard band serves three main purposes:

1. It is used for radial tracking.\(^1\) For 1D systems, push-pull is the most widely used method for generating a tracking error signal. The tracking method is based on the diffraction of the laser beam at the periodic track (or row) structure on the disc. However, for spatial track frequencies beyond the cut-off of the optical channel (see Chapter 3) no push-pull signal is available anymore.

\(^1\)Radial tracking is needed to keep the focussed spot(s) exactly on the track(s). A practical disc has finite tolerance on the position of the center hole. This causes so-called ‘radial run-out’ which can be in the order of tens of microns. Without tracking the spot would sweep over many tracks during one circumference of the disc. Active mechanical tracking of the objective lens using a feedback control system compensates for this problem. Push-pull is a method to generate a track error signal that is used as input for the control loop. See Chapter 2 in [8] for a detailed explanation.
2.2 Read out of a TwoDOS Disc

This situation arises for the high track density in the TwoDOS format. A simple method to do tracking in the 2D case is to measure the ‘low-pass averaged’ amplitudes $A_1$ and $A_2$ of the data signal of the two boundary rows within a broad spiral. These amplitudes should be equal if the corresponding data bitstreams have the same low-pass content (see Chapter 4 for codes that guarantee the latter property). In case a radial offset is present one boundary spot ($S_1$) will partly scan the guard band leading to a larger amplitude (because reflection at the mirror surface yields the maximum signal amplitude). The other boundary spot ($S_2$) will partly scan one of the inner rows of the broad spiral giving an unchanged signal amplitude. The inset of Fig. 2.2 shows the principle of tracking error generation.

2. The guard band can serve as a starting point for 2D bit-detection. A full-fledged 2D bit-detector that covers the complete broad-spiral will appear to have an impractical implementation complexity. A ‘per-stripe’ bit-detector is proposed where a limited sub-set of bit-rows is decoded simultaneously making use of side information from adjacent per-stripe bit-detectors (see Chapter 6). Initially, when no side information is present, per-stripe bit-detection can take place at the boundaries of the broad spiral by making use of the prior knowledge of zeros on the guard band.

3. A third function of the guard band is to introduce a discontinuity in the phase relation between two adjacent broad spirals to enable a constant areal density across the disc. From Fig. 2.2 it is clear that an exactly constant lattice parameter cannot be maintained throughout a contiguous concentric lattice, that is, without the introduction of ‘lattice defects’ to accommodate for the tiny difference in radius.

2.2 Read out of a TwoDOS Disc

Read out of a TwoDOS disc can be accomplished by two means:

1. Generating multiple spots from a single spot laser diode by using a diffraction grating (See Fig. 2.3). Such a phase grating consists of a periodic pattern (as shown in Fig. 2.4). An advantage of this method is that only one laser is needed. On the other hand, the spots on the medium must be separated by some distance to avoid interference between the coherent beams. Furthermore, multi-spot writing is not possible.

2. A multi-spot laser diode. In this device multiple laser sources are integrated on a single substrate. Each laser can be modulated independently, which allows a recordable or rewritable 2D format. The lasers typically have a rather
The TwoDOS Concept

Figure 2.2: Schematic illustration of the role of the guard band in maintaining a constant areal density across the disc and in the generation of a radial tracking signal (as shown in the inset. Here $A_1$ and $A_2$ are the amplitudes of the signals generated by scanning the disc with spots $S_1$ and $S_2$ respectively).

large separation on the substrate (in the order of 100 $\mu$m) to avoid thermal and electrical interference between the lasers.

In our experimental player we have used the first option. An experimental intensity scan of the generated array of laser spots using a CCD camera is shown in Fig. 2.5. In the inset of this figure a photograph of the spot array is shown. The use of a grating is only suitable for the read-out part of the system.

The experimental optical set-up is shown in Fig. 2.6. A photograph of the total set-up is shown in Fig. 2.7. A more detailed picture is shown in Fig. 2.8. After passing the laser beam through the diffraction grating it is collimated to a parallel bundle. A beam-shaper compensates for the elliptic far field intensity profile of the laser. For a small spot a high rim intensity is required. In that case the beam shaper avoids a large loss in light power when passing the beam through the circular aperture of the objective lens. A telescope is inserted to obtain the correct magnification i.e. to set the correct distance between the spots at the information layer on the disc. Furthermore, it avoids vignetting of the outer spots by imaging the pupil of the collimator on the pupil of the objective lens. Via a polarizing beam splitter (PBS) the light is directed towards the disc. A quarter-wave-plate (denoted by $\lambda/4$) is used to rotate the linearly polarized light by 90°. This ensures that the light that is reflected from the disc is transmitted by the PBS towards two normal beam splitters (NBS). These direct part
2.2 Read out of a TwoDOS Disc

Figure 2.3: Simplified representation of the read-out principle of a 2D disc. The 2D spiral is located in a single information layer of the optical disc (the disc is shown in top-view).

Figure 2.4: Design of the periodic phase grating structure. A periodic structure with a depth of 356 nm is etched in a glass substrate material with refractive index $n=1.4696$.

of the light to the segmented detectors on the photo diode IC (PDIC) and part of the light to a Foucault focusing branch [8]. The remaining light is used to make an image
Figure 2.5: Intensity profile of the generated spot array measured with a CCD camera. One can observe a reasonable homogenous spot intensity. Furthermore, the diffraction efficiency of the grating is as high as 85%. Note that the intensity of the central spot is the highest, and that some spurious spots with low but non-zero intensity exist outside of the array of \( N_r = 11 \) intended spots.

of the focal plane at the position of the disc on a CCD camera.

Because of the coherent nature of the laser light and the fact that the spots in the array originate from the same source, interference will occur at areas where adjacent spots overlap, giving rise to typical interference fringes. Each spot is diffraction limited and has the Airy intensity profile (according to Eq. 3.9, see Section 3.3.1). To avoid interference beyond a tolerable level, the minimal distance between two consecutive laser spots on the disc has been chosen such that the local maximum of the so called second Airy rings of adjacent spots are aligned as is shown in Fig. 2.9. This leads to a minimal distance between the spots equal to:

\[
r_s = 2.682 \frac{\lambda}{\text{NA}},
\]

which for \( \lambda = 405 \text{ nm} \) and \( \text{NA} = 0.85 \) is equal to 1.28 \( \mu \text{m} \). The distance between the bit-rows in case of a hexagonal bit-ordering with lattice constant \( a_H \) is \( \sqrt{3} a_H \) (with \( a_H \) typically equal to about 130 to 170 nm). This is a factor of 10 smaller than the distance between the spots for the storage densities of interest. To align the spots on
each of the rows the grating is rotated with respect to the tangential direction (see Fig. 2.1 for a definition) over an angle of:

$$\alpha = \sin \left( \frac{\sqrt{3} a_H}{2 \times 6.82 \frac{\lambda}{NA}} \right)$$  \hspace{1cm} (2.2)$$

resulting in a configuration as illustrated in Fig 2.9. For $a_H = 165$ nm the angle is equal to 6.4°. For $a_H = 138$ nm it is 5.4°. The consequence of this rotation is that the signal of row $n$ is lagging with respect to the signal of row $n + 1$ (for the adopted definition of row numbers). The time delay $D$, expressed in channel bit periods $a_H$, is equal to:

$$D = \sqrt{\left( \frac{2.682 \lambda}{a_H NA} \right)^2 - \frac{3}{4}}.$$  \hspace{1cm} (2.3)$$

For $a_H=165$ nm this delay is equal to 7.7 bit periods while for $a_H=138$ nm it increases to 9.2 bit periods. A dedicated signal processing block has been designed in the receiver to compensate for this delay (see Section 5.3).

**Signal Folding**  In optical recording, the optical wavefront of the laser spot is diffracted by the sub-wavelength structures on the disc. Due to this effect, part of the incoming light is diffracted outside the pupil of the objective lens and is lost for detection,
Figure 2.7: Photograph of the experimental setup. The large cabinet in the lower part of the photograph contains 19-inch racks with electronics for servo and laser control.
causing the modulation in the signal. This modulation enables to discriminate between digital ones and digital zeros. The digital ones are represented by pits, which are impressed in the substrate with a predefined depth to induce a phase difference \( \Phi \), between the wavefront of the light reflected from a land area and the wavefront of the light reflected from a pit area. Therefore, both land and pit areas are needed under the spatial extent of the optical spot to cause diffraction to occur. In 1D optical storage the spot diameter is normally larger than the radial width of a pit on the disc. This means that when the spot scans a pit the reflected light beam loses intensity by diffraction outside the central aperture (CA) of the lens, even in the case of successive pit-bits that are written as one elongated mark in the direction of the track. In the latter case there is only radial diffraction in case the spot scans the central part of such an elongated mark.

In the 2D-case the most simple assumption is that the whole (hexagonal) bit-cell is of the pit-type when the corresponding bit equals a one. Then the problem of ‘signal folding’ arises [13]. No diffraction occurs at large contiguous pit areas, consisting of a number of neighbouring bits that are all of the pit-type. Both areas of one and zero bits, i.e. both a large pit area and a large non-pit (or land) area will act as a mirror and show identical read-out signals (see Fig. 2.10). As a result the channel
becomes highly non-linear. For this reason each pit-bit is recorded as a separate (preferably circular) pit-hole with a size smaller than the size of the hexagonal bit-cell. This is schematically indicated in the left part of Fig. 2.10. In this way, large contiguous areas of pit-marks are avoided so that the non-linearity of the channel is greatly reduced. To estimate the remaining non-linearities, and to evaluate the signal modulation but also to steer the mastering process, a channel model is developed based on scalar diffraction theory. The numerical results of this model, which clearly show the non-linearities, are discussed in Section 3.3.2.

### 2.3 Benefits of 2D Optical Storage

A first reason to pursue 2D optical storage is the increased data rate compared to a conventional 1D system. For a system with $N_r$ rows in the broad spiral that is read out with an array of $N_r$ spots the data rate simply increases by this factor $N_r$ at
2.3 Benefits of 2D Optical Storage

Figure 2.10: Mastering of pit-holes smaller than the hexagonal bit-cell to avoid non-linearity due to signal folding. As a reference the full-width half-maximum (FWHM) of the optical spot is indicated.

the same rotational speed of the disc. A second, somewhat intuitive reason, is the higher robustness of the system. For example, it is known that the data frequency in each row is the same. The timing recovery can make use of this knowledge by combining the timing error information for all rows. When a disturbance affects only a couple of bit-rows (e.g. a small dust particle) the timing recovery still receives timing information. Hence, it is potentially more robust than the 1D system. The third reason is to increase the recording capacity on the disc. This deserves some further explanation.

The intensity profile of the read-out spot has a finite spatial extent determined by the wavelength $\lambda$ and the numerical aperture NA (see Fig. 2.9). Therefore, when reading out a bit, not only information from this bit is present in the read-out signal, but also contributions from adjacent bits appear. This is called intersymbol interference (ISI) and it is present isotropically due to the circular symmetry of the spot intensity profile. Conventionally, the radial component of this ISI is treated as a ‘noise contribution’ to the signal. Accordingly, the distance between tracks, is chosen such that at nominal read-out conditions the disturbance of adjacent tracks is negligible, i.e. radial ISI is much smaller than tangential ISI. As the radial ISI has its origin in bit information of adjacent rows it could, however, be used as information in the detection process (instead of treating it as a noise contribution). This will be elaborated in further sections.

2.3.1 Cross-Talk Cancellation

With the arrival of DVD there was a pressure to increase disc capacity beyond the nominal $\lambda/NA$-scaling, and the track pitch was reduced considerably at the cost of somewhat more interference from bits in the adjacent tracks. To facilitate the read out of a central track in the presence of inter symbol interference (ISI) from adjacent tracks a so-called cross-talk canceller (XTC) was applied [15] (although it appears
that DVD and also BD can be read out without the use of an XTC). In this scheme three tracks are read out simultaneously using three optical spots (generated by a diffraction grating from a single laser source). Filtered versions of the signals of the outer tracks (denoted with subscripts $n - 1$ and $n + 1$) are subtracted from the center signal (with subscript $n$) in an attempt to cancel out that part of the center signal that is due to the bits on the adjacent tracks. Fig. 2.11 shows a block diagram of a cross-talk cancellation scheme. The finite impulse response (FIR) filters are adapted using a least mean square (LMS) algorithm to minimize a certain criterion. A more detailed discussion can be found in Chapter 6, where the XTC method is used in conjunction with a 2D Viterbi bit-detector.

![Figure 2.11: Block diagram of a LMS based cross-talk cancellation scheme. A possible criterion is the minimization of correlation between $\tilde{y}_{m,n}$, $y_{m,n+1}$, and $y_{m,n-1}$.](image_url)

### 2.3.2 2D Inter Symbol Interference

One of the basic ideas behind TwoDOS stems from the fact that the optical channel has a circularly symmetric impulse response, and that the data should be ordered more symmetrically to benefit from this channel property. The tightest ordering of circular pits on the disc is the hexagonal closed-packed ordering. In the case these pits are packed at a density equivalent to two times the BD density the impulse response sampled at the 2D bit positions can be approximated by the one given in Fig. 2.12. The central tap-value $c_0$ equals 2, and the 6 nearest-neighbor taps $c_1$ equal 1. These
2.3 Benefits of 2D Optical Storage

nearest-neighbor taps will be referred to as the 1st shell taps. Along the center track

![Diagram of 2D Optical Storage](image)

*Figure 2.12: Linear approximation of the response sampled on a hexagonal lattice at an equivalent density of two times the BD density for the same physical read-out conditions (NA=0.85; λ=405 nm).*

we have the tap-values \( \{1,2,1\} \) and along each of the two adjacent tracks the tap-values are \( \{1,1\} \). The total energy of this 2D impulse response equals 10, with an energy of 6 along the center track and an energy of 2 along each of the adjacent tracks. From these energy considerations, one can see that simply considering the signal contributions (in the form of ISI) from adjacent tracks as a disturbance would be a waste of valuable information. In fact the use of cross-talk cancellation as described in the previous section would yield a 40% loss of energy-per-bit (i.e. a loss of 2.2 dB). The main advantage of ‘joint 2D bit-detection’ is that all signal energy associated with a single bit is used for detection of this bit. Note that this observation is not limited to the case of a hexagonal bit ordering. Also in (1D) coded cases with a small track pitch the above energy considerations hold. Note further, that the reduction in track pitch (TP) is likely to require a compromise on the tangential channel bit length (CBL). Compared to the conventional 1D-case, the optimum is expected to lie at a point with smaller track pitch and a slight increase in channel bit length. For the coded case however the ratio between track pitch and channel bit length is likely to deviate from 1. This is visualized in Fig. 2.13 where iso-density lines are plotted in a graph with the normalized radial and tangential density on the respective \( x \) and \( y \) axes. The normalization is with respect to the 25 GB BD format. In the tangential direction the user bit length (UBL) is used, which incorporates the rate of the modulation code.

Another way of looking at the principal difference between ‘joint 2D bit-detection’ + 2D equalization and conventional 1D bit-detection + XTC is that the XTC procedure in the latter approach is equivalent to an attempt to do full response equalization in the radial direction (using a limited number of equalizer taps in the radial direction). A possible result is noise enhancement. The fact that XTC works particularly well in optical recording is that media noise, with a power spectral density (PSD) corresponding to the channel transfer characteristic, is prevalent above other noise sources. In the 2D approach a partial response could be targeted in both the tangential and radial directions, which offers more freedom for engineering the noise PSD...
Figure 2.13: Iso-density lines plotted as function of radial and tangential density. The plot is normalized to BD parameters at 25 GB density (CBL=74.5 nm; UBL=111.75 nm; TP=320 nm) indicated with the point at (1,1). For the TwoDOS hexagonal case a lattice constant $a_H=138$ nm is assumed. The code rate is taken 0.98 (see Chapter 4). In case of a TwoDOS format the overhead for the guard band of 1 additional row for each 11 rows is taken into account in the effective radial density. The TwoDOS-RLL parameters are TP=200 nm and CBL=66.7 nm for a d=1 RLL code.

in radial direction. Especially, in cases with a very low track pitch beyond which full response equalization isn’t feasible anymore (TP < $\frac{\lambda}{2\text{NA}}$).

### 2.4 Manufacturing of TwoDOS Read-Only Discs

For read-only discs the problem of signal folding arises as described in Section 2.2. A possible solution is to make pit-holes smaller than the bit-cell. The pit-holes, with a typical diameter in the order of 100 nm, are embossed into the disc by means of a so-called stamper in a mass production process. Several techniques such as Laser Beam Recording (LBR), Liquid Immersion Mastering (LIM) and Electron-Beam Recording (EBR) are used to manufacture these stampers.
2.4 Manufacturing of TwoDOS Read-Only Discs

Stamper fabrication  The procedure to make a stamper using LBR is depicted in the left part of Fig. 2.14. A photo-sensitive resist layer, which is spin-coated on top of a glass substrate is illuminated using a focused laser beam (or a focused electron beam). The photo resist is etched away completely at the positions where it was illuminated. The glass substrate is used as an etching-stop. Note that the thickness of the photo resist layer will determine the pit-depth of the final disc and therefore it should have a good thickness uniformity. After the etching step a thin layer of nickel or nickel-alloy is deposited on top of the structured photo resist layer in a sputtering process. The metallized glass master is electroplated to form a thick nickel so-called father stamper which can be removed from the glass substrate. This father stamper is used in our case to press the final discs. For large-scale production two additional electroplating steps are applied to create a mother stamper (which is the ‘negative’ of the father stamper) in a family process from which more than one stamper can be created (see Fig. 2.14). In the latter case the stampers are oxidized prior to electroplating to allow proper separation of the stampers.

![Figure 2.14: Making the stampers in a family-process.](image)

Replication  When the stampers are available several possible replication methods can be used to produce many identical replicas from a single stamper. One is injection moulding. Here the nickel stamper is mounted on a press to form one half of the mould. The other half is a flat plate to form the optically flat surface at the read-out side of the disc (at least for CD and DVD. A BD disc is read out via a 0.1 mm cover layer that is bonded or spin-coated on top of the information layer). Molten polycarbonate of optical quality is injected into the cavity formed between the nickel stamper and the flat surface. After cooling down the newly formed disc is released
from the mould. The whole replication procedure takes only around 3 seconds which makes mass-manufacturing of discs very cheap. After the replication an aluminum layer is sputter-coated onto the transparent disc to form the mirror. Another method, which is used in our case, is the so-called glass-2P (Photo-Polymerization) process. Here a liquid lacquer is deposited on a glass substrate, after which the stamper is pressed against the substrate (see Fig. 2.15). The lacquer is cured by ultra-violet (UV) light through the transparante substrate. After curing the newly formed disc can be released from the stamper and an aluminum and additional protective layer can be deposited on top of the lacquer.

**Figure 2.15**: Replication using a glass-2P process.

**Liquid Immersion Mastering**  Generally, the spot in a mastering machine is smaller than the spot that is used for read out of a disc. In the 1D case this is necessary to form elongated marks with a radial width smaller than the diameter of the read-out spot (see Section 2.2). The smaller spot is realized by using a higher NA objective lens and a smaller wavelength in the mastering set-up (for example a DVD mastering machine may use an NA of 0.9 and a wavelength equal to 458 nm, whereas a
DVD reader uses NA=0.6 and a wavelength of 650 nm). For BD and also for TwoDOS the requirements on mark size are such that conventional far-field optics are not sufficient to produce high quality masters. A possible solution is to apply near-field mastering [16] (similar to the near-field readout as described in Section A.1). A somewhat less revolutionary approach is to use liquid immersion mastering (LIM) [17,18]. Here a far-field objective lens is used (NA=0.9) and an immersion liquid is applied between the lens and the rotating disc. The immersion liquid, which has a refractive index considerably higher than 1, allows light waves at angles above the critical angle to pass through the liquid film without the problem of total internal reflection. This gives a diffraction limited spot corresponding to an NA of approximately 1.2. A schematic drawing and some photographs of the experimental set-up are given in Fig. 2.16. Many of the experimental results that will be presented in this thesis are obtained from LIM-mastered discs.

**Figure 2.16:** Schematic drawings of the LIM-principle (left) and some photographs of the experimental set-up (right).

**Electron Beam Recording** When even smaller features must be mastered also LIM may not be sufficient anymore. A disruptive technology to master even smaller features is by using an electron beam recorder (EBR) [19]. By accelerating electrons with voltages up to $U=15$ kV a very small wavelength can be obtained as (including the relativistic correction):

\[
\lambda = \frac{hc}{\sqrt{2qUm_0c^2 + (qU)^2}}
\]
with $h$ equal to Planck’s constant, $c$ the velocity of light, $q$ the electron charge and $m_0$ its rest-mass. The effective spot diameter is not directly determined by this parameter. It is rather limited by spherical and chromatic aberrations of the objective lens in the electron-optical setup, and due to the finite energy spread of the high-energy electrons. Within the resist the effective spot size is further increased by forward and backward scattering of the electrons in the resist. Also the repelling action of the electrons in the focal region and in the cross-over points in the electron-optics lens (the so-called Boersch effect \cite{ref20, ref21}) increases the spot-size. A schematic drawing of a recording set-up is shown in Fig. 2.17. The typical beam diameter is 50 nm at beam currents of approximately 20 nA. A beam blanker is used to deflect the beam at high frequencies across an aperture, hereby effectively switching it on and off at the location in the photo-resist layer. Stability of this deflection is very important to minimize pit-position variation, see e.g. 3.6.3. In addition a beam deflector is present that allows the mastering of wobble patterns. Because the accelerated electrons have a very short propagation distance in normal air a high-vacuum chamber is usually necessary. This would be very cumbersome for the mastering set-up because every time a new disc is inserted either a load-lock chamber should be used or the complete system must be pumped vacuum again. The problem is solved by using a differential pumping head assembly as is also used in commercial TEM and SEM instruments (e.g. for fast removal of photographic plates). A combination of vacuum pumps and the flow of a high pressure gas allows a localized high-vacuum region while the complete assembly can still maintain a small distance with respect to the disc by floating on an ‘air’-suspension.

As already mentioned the resolution of electron beam recording is limited amongst others by scattering of electrons in the resist and the substrate. Therefore, in regions with a large number of pits there is a considerable amount of background ‘illumination’ and the pits tend to be somewhat larger. This effect is know as the ‘proximity’ effect. It can be considered as a kind of non-linear inter symbol interference in the write channel, that causes a pattern dependent variation in the signal level (see Section 3.5.1).

**Multiple-pass Mastering** The mastering methods discussed above are all developed for 1D recording. This means that only a single beam is available to master a single track per revolution. The TwoDOS format, however, contains a broad spiral where track-to-track data synchronization within this spiral is extremely essential. To master such a format with LBR or LIM we would require an array of lasers spots similar to the one used for reading, but with the important difference that now we must be able to modulate each of the spots independently. Such an array of independent lasers is not available (yet). For this reason an alternative method, which we will name multi-pass mastering, was developed \cite{ref23}. The main issue with multiple-pass
mastering is the track-to-track synchronization of the data. The mastering clock must have such an absolute accuracy that at each pass it can be synchronized with constant phase offset with respect to the previous pass. This means that in CAV mode the frequency must increase slowly for increasing disc radius otherwise the channel bit length will change slowly as function of the radius. In a practical set-up an arbitrary waveform generator (AWG) was used to generate the data and reference signal. The reference signal is input to a first phase locked loop (PLL), which generates one pulse per revolution, and a second PLL which generates an input reference signal for the controllers of the translation stage and the spindle motor. The latter PLL is adjusted slowly as function of the radius to obtain a nearly CLV or quasi CAV (QCAV) mode of operation. This allows a constant data frequency for increasing radius on the disc. A schematic view of the set-up is shown in Fig. 2.18.

LIM multiple-pass mastering is used for a run length limited (RLL) coded data format (CBL=66.7 nm, TP=200 nm). From the single tone carrier at the beginning of every data-frame, as shown in Figure 2.19, it can be clearly seen that the synchronization between the rows within the broad spiral is very good. Furthermore, the broad spirals are all the same. This indicates an excellent bit positioning accuracy of 4-7 nm.
Figure 2.18: Multiple-pass mastering set-up, where the rotation and translation stage are controlled such that the data frequency can be kept constant for increasing radius on the disc.

Figure 2.19: SEM image of a LIM disc (LIM 513, see Table 2.1). Within the meta-spiral excellent alignment between the bit rows is achieved, which can be seen from the single-tone carriers at the right hand side of the image. The two shown meta-tracks are exactly the same, which indicates very good synchronization between the rotation of the disc and the data-recording during mastering.
For the hexagonal format even smaller pits are required and thus EBR is needed. The EBR recording is performed at a velocity of 0.83 m/s using a beam current of 12 nA and an acceleration voltage of 15 kV. Circular pits with a diameter as small as 70 nm were realized on the stamper and the replicated discs. Figure 2.20 shows three identical meta-spirals, containing the hexagonal code. Note the track pitch (TP=119.5 nm), the small size of the pits (70 nm) and the small separation between the pits (138 nm). Again, the 2D alignment is excellent.

![SEM image of an EBR master used for disc production (disc E490, see Table 2.1). Within the meta-spiral excellent alignment between the bit rows is achieved, which can be seen from the single-tone carriers at the right hand side of the image. The three shown meta-tracks are exactly the same, which indicates very good synchronization between the rotation of the disc and the data-recording during mastering.](image)

Besides multiple-pass mastering the EBR technique offers an additional method to generate the TwoDOS format. This method is based on the fact that the electron beam can be deflected very fast in both tangential and radial directions. When the rotational velocity of the disc is limited it is possible to master pits over the full width of the broad spiral by deflecting the electron beam along a radial path and by blanking the beam at the required positions of the pit-marks. This results in a single-pass method. The scanning method is illustrated in Fig. 2.21.

**Lattice Distortion**  In practice the track-to-track synchronization of data during the multi-pass mastering procedure is not always perfect. This results in time-varying phase offsets between the data in adjacent tracks compared to their nominal phase relation. These phase offsets appear as lattice distortions of the 2D lattice. A first-order approximation of the read-out signal of such a distorted lattice can be obtained by delaying the signals by the appropriate amount. The phase-part of the timing recovery
loop is independently done for each row within the broad spiral, and can therefore
be used to compensate for these time-varying delays as long as the frequencies of
the variations are within the bandwidth of the timing recovery loop. A second-order
effect is that in particular radial ISI changes as function of time. This problem is
solved by the adaptive equalizer, again provided that the frequencies of the variations
fall within the bandwidth of the equalizer loop (which is generally a lot slower com-
pared to the timing recovery loop). Another source of lattice distortion is ‘row-pitch’
variation. During the mastering process no reference is present for the track position.
Instead the sledge makes a linear translation to form a spiral or a circular format. In
case the linear translation velocity is not precisely constant this results in ‘row-pitch’
variation.

2.5 Experimental Discs

In this section a brief overview is given of the discs that were manufactured using
the various technologies. A list is shown in Table 2.1. The numbering in this table is
used throughout this thesis. Note that only the most important discs are mentioned.
The initial discs, for example LBR 1 to 9 that were used for optimization of the
manufacturing process, are not listed in the table.

LBR Discs For LBR mastering it is not possible to achieve the very small feature
size that is needed to for a 35 or 50 GB disc capacity. For this reason, experiments
are performed with a scaled NA equal to 0.4 or 0.57. For the hexagonal lattice with a
lattice period \( a_H \) equal to 248 nm the disc capacity is only 15.5 GB. However, when
the disc is read out at \( NA=0.57 \) this is equivalent to a capacity of 35 GB that is read
out at \( NA=0.85 \). For an \( NA=0.4 \) of the read-out system the equivalent capacity is
even 70 GB.
2.5 Experimental Discs

Table 2.1: Overview of discs replicated from stampers made with different mastering technologies.

<table>
<thead>
<tr>
<th>Techn.</th>
<th>Disc</th>
<th>format</th>
<th>BL (a_H) [nm]</th>
<th>TP [nm]</th>
<th>Disc Cap. [GB]</th>
<th>remarks</th>
</tr>
</thead>
<tbody>
<tr>
<td>LBR</td>
<td>10</td>
<td>HEX</td>
<td>248</td>
<td>214.8</td>
<td>eq. 35</td>
<td>@ NA=0.57</td>
</tr>
<tr>
<td></td>
<td>12</td>
<td>HEX</td>
<td>248</td>
<td>214.8</td>
<td>eq. 35</td>
<td>@ NA=0.57</td>
</tr>
<tr>
<td>LIM</td>
<td>513</td>
<td>17PP</td>
<td>66.7</td>
<td>220</td>
<td>37.5</td>
<td>no WS</td>
</tr>
<tr>
<td></td>
<td>522</td>
<td>17PP</td>
<td>66.7, 50</td>
<td>200, 200</td>
<td>41.25, 55</td>
<td>WS</td>
</tr>
<tr>
<td>EBR</td>
<td>E266</td>
<td>HEX</td>
<td>165</td>
<td>142.9</td>
<td>35</td>
<td>WS: 6T/10</td>
</tr>
<tr>
<td></td>
<td>E267</td>
<td>HEX</td>
<td>165</td>
<td>142.9</td>
<td>35</td>
<td>WS: 4T/10</td>
</tr>
<tr>
<td></td>
<td>E268</td>
<td>HEX</td>
<td>138</td>
<td>119.5</td>
<td>50</td>
<td>WS: 6T/10</td>
</tr>
<tr>
<td></td>
<td>E377</td>
<td>HEX</td>
<td>165</td>
<td>142.9</td>
<td>35</td>
<td>EBR#2, WS: 6T/10</td>
</tr>
<tr>
<td></td>
<td>E378</td>
<td>HEX</td>
<td>138</td>
<td>119.5</td>
<td>50</td>
<td>EBR#2, WS: 6T/10</td>
</tr>
<tr>
<td></td>
<td>E427</td>
<td>HEX</td>
<td>138</td>
<td>119.5</td>
<td>50</td>
<td>RIE Si master</td>
</tr>
<tr>
<td></td>
<td>E490</td>
<td>HEX</td>
<td>138</td>
<td>119.5</td>
<td>50</td>
<td></td>
</tr>
<tr>
<td></td>
<td>E491</td>
<td>HEX</td>
<td>138</td>
<td>119.5</td>
<td>50</td>
<td></td>
</tr>
</tbody>
</table>

LIM Discs The two most important discs made with liquid immersion mastering are disc 513 and disc 522. Here, a run length limited (RLL) modulation code is used similar to the 17PP (parity preserving) code that is used in the Blu-ray Disc format. The shortest mark on the disc has a length of 2 bits, while the longest mark is equal to 8 channel bits. For disc 513 no write strategy is applied, i.e. the laser is simply switched on and off according to the ones and zeros in the data stream. Especially, the longer pits appear to increase in width. For this reason in disc 522 a write strategy (WS) is applied. More information can be found in Section 7.1.1. On each disc various bit lengths (BL) are recorded (50, 55, 60 and 66.7 nm) and various track pitches (TP) are used (200 and 220 nm). For these values disc capacity ranges from 37.5 (TP=220 nm; BL=66.7 nm) to 55 GB (TP=200 nm; BL=50 nm). A few examples with different bit lengths and track pitches are given in Table 2.1.

EBR Discs With the electron beam recorder two disc types were made. The first one has a hexagonal lattice parameter a_H equal to 165 nm (35 GB). The nominal pit-diameter is 120 nm. The second one has a_H = 138 nm (50 GB). Here, the nominal pit-diameter is 102 nm. A write strategy is used where each bit period is divided into 10 equal time intervals. This allows a variation in pulse-width to tune the shape of the pits in the disc. The write strategy with the longest pulse (6T/10) leads to elongated pits, while the strategy with a pulse length equal to 4T/10 shows more circular pits.
Masters E377 and E378 were made using another EBR set-up. Two different mastering set-ups are used denoted with EBR#1 and EBR#2. The advantage of EBR#2 is that it has less rotation jitter (i.e. a better row-to-row synchronization during the multi-pass mastering process). A disadvantage is the larger pit-size variation for masters that are manufactured with EBR#2.

Discs E268 and E378 showed considerable pit-moge (see Section 7.1.3). For this reason master E427, E490 and master E491 were manufactured. Master E490 is a nickel stamper. This stamper is used for the experiments with a release layer (see Section 7.1.3). Master E491 is a silicon master that was manufactured using a reactive ion etching (RIE) step to modify the wall-angle of the pits (to improve the release of disc and master during replication).

2.6 Recordable and Rewriteable TwoDOS Technology

Although the experimental TwoDOS work has focused on a ROM system, for a complete optical disc family one also needs at least a recordable format. We will briefly discuss recordable and rewriteable technologies in relation to issues that are specific for the 2D system. The availability of a light source which allows an independent modulation of each of the spot intensities in the array is preferable, but not strictly needed. Also here a multi-pass writing method could be applied.

Phase Change Recording  For a rewritable system, phase-change technology is generally used. A good introduction to phase-change recording can be found in [24]. Phase-change recording is based on the fact that some alloys can exist in two different states. The poly-crystalline state, where atoms are ordered in small grains of a few tens of nm according to a periodic 3D lattice and an amorphous state, where the atoms do not have such a long-range order. The crystalline state is the equilibrium state and when the alloy is cooled slowly from the melted phase a crystalline structure will be obtained. Writing on such a (poly)crystalline disc is possible by heating the material again using a focused laser beam to beyond the melting point, followed by a rapid cooling or quenching (in the order of nanoseconds). If the cooling period is shorter than the so-called crystallization temperature, the atoms do not have enough time to return to their crystalline positions and they are ‘frozen’ in a random position (the amorphous state; see Fig. 2.22). Direct overwriteability is achieved by the fact that heating the material to just below the melting point, but above the crystallization temperature will give the atoms the opportunity to return to their crystalline position. A sequence of laser pulses to write a pattern of crystalline and amorphous marks is called a ‘write-strategy’. A typical example together with a SEM image of the resulting disc surface is shown in Fig. 2.23. Read-out is based on the considerably lower
reflection of the amorphous marks compared to the crystalline background, i.e. signal modulation is an amplitude effect instead of a phase effect as was the case for a read-only disc. This also means that the problem of signal folding is only partly present for the rewritable system. Of course diffraction still occurs at amplitude transitions, while for a large continuous 2D area with the same amplitude this effect disappears. This means that signal folding does not necessarily disappear for a recordable or rewritable disc.

**Thermal Cross-Erase** When a bit-row is written using the phase change principle, the adjacent row is also influenced due to an increase in temperature in this row. The temperature increase is not only caused by the finite size of the optical intensity profile, but also by heat diffusion in the phase-change stack. One of the effects is the (partial) recrystallization of amorphous marks in the adjacent row during the write process, which is caused by exposure of the marks to a temperature above the crystallization temperature for a certain period of time. This phenomenon is called thermal cross-erase. In a 2D system the data is ordered more symmetrically compared to the 1D case to benefit from the circularly symmetric channel. For the write channel the resulting small row-to-row distances are expected to lead to a large thermal cross-erase. The ultimate radial density is thought to be limited by this effect and not by the ISI in the read channel. However, several physical counter measures seem to be possible. First of all a proper stack design\(^2\) is very important [25]. By manipulation of, for example, the thickness and the thermal conductivity of the different layers in

---

\(^2\)With ‘stack’ we refer to the multi-layer recording material. One layer is the actual phase-change material. Additionally, the stack generally includes a metal layer (reflective layer and cooling layer) and dielectric layers.
the phase-change stacks one could make trade-offs that favor a low thermal cross-erase. Another option is to introduce an additional layer in the recording stacks as is proposed in [26, 27]. This layer, denoted as anti-crystallization (AC) layer, prevents the amorphous marks from crystallization. This means that cross-erase is reduced considerably. A track pitch of 0.24 µm is reported in a land-groove format with 1D RLL encoded tracks, thereby achieving 50 GB capacity for BD optics. A 3-beam cross-talk canceller is used for read-out. A small drawback is that due to the anti-crystallization layer also the overwrite of amorphous marks is difficult and the disc is of the write-once type. Nevertheless, this technique can be used for a 2D recordable system.

From a signal-processing point of view one can also think of some measures that prevent or compensate thermal cross-erase. An example of prevention is the use of a 2D write strategy, which means that the laser power sequence depends not only on the row that needs to be written, but also on adjacent rows. Ultimately, this leads to a write strategy matrix, where a number of input data signals determine a number of parameters from multiple laser pulse sequences. Such a write-strategy matrix is proposed in [28] (see also Fig. 2.24, for a 1D example). A practical implementation of such a matrix in the form of a laser driver integrated circuit can be found in [29, 30]. A triplet of run lengths \((x_{-1}, x_0, x_{+1})\) can be chosen for each row as input to the matrix.
2.7 Conclusions

Figure 2.24: Write strategy matrix [28]. This example is suitable for 1D, but the matrix can easily be extended to 2D. A set of parameters \( p_{x-1,x_0,x+1} \) can be varied depending on a triplet of input run lengths \( x_{-1}, x_0, x_{+1} \).

to be able to also vary the write strategy of run length \( x_0 \) depending on the leading and trailing run length \( (x_{-1}, x_{+1}) \), such that thermal memory of the disc can be handled. Note that the separation of light spots at the information layer can be chosen smaller in case of a laser array. The laser sources are independent and do not coherently interfere with each other. The distance depends on the minimum separation of the laser devices on the single substrate (in view of thermal and electrical cross-talk between the lasers) in combination with the magnification of the optical system, and it depends on the thermal memory of the heated area on the disc. The write-strategy to prevent thermal cross-erase will depend on the distance between the spots as long as this distance is smaller than the thermal memory length on the disc. In case the memory length is longer than the distance between the spots the write-strategy becomes independent of this length. It is then not strictly necessary to have a laser array with independently controllable lasers. Again a multi-pass technique could be used. However, the thermal independency of the laser spots may limit the possible benefits of the application of 2D write strategy.

Yet another method originates from the principles of 2D optical recording. For compensation of the effects of thermal cross-erase the ‘joint 2D-detection’ that is proposed in this work is ideally suited. Cross-erase will cause partial recrystallization of the amorphous marks depending on the data that is written in the adjacent bit-rows. Therefore, the read-out signal will show a non-linear behavior that extends over more than one bit-row. In a joint detector a 2D target response can be adapted such that it resembles the radially extended non-linear behavior. This is not possible in the 1D case and as such the 2D system offers a potential advantage.

2.7 Conclusions

In conventional optical recording systems the data is considered to be one dimensional in the sense that adjacent parts of the spiral track do not have any logical or
physical relation to each other. They are recorded asynchronously, are separately encoded by the modulation encoder and belong to different error correction blocks. Also during read-out the inter-symbol interference due to adjacent tracks is treated as a noise contribution. Accordingly, the distance between tracks is chosen larger in order to minimize this noise contribution.

In contrast, in the two-dimensional optical system a disc with a ‘broad spiral’ or ‘meta-spiral’ is defined containing multiple bit-rows. Adjacent parts of the broad spiral are separated by a guard band that serves three purposes:

- It enables a constant areal density across the disc by introducing a discontinuity in the phase relation between adjacent parts of the spiral.
- It serves as a starting point for 2D bit-detection. The fact that no pits are present in the guard band is used as prior information in the detection process.
- It is used for radial tracking by measuring the low-pass averaged amplitudes of the data signal of the boundary rows.

Read out of such a disc containing a broad can be done by two means:

- Generating multiple spots from a single laser by using a diffraction grating.
- A multi-spot laser diode, where each of the lasers can be modulated independently.

We have used the first option, which limits us to the read-out of discs. Discs are manufactured by mastering. Stamper fabrication is done by liquid immersion mastering (LIM) or electron beam mastering (EBR) in a multiple-pass mastering process. In this process the broad spirals are formed by successive revolutions in a single beam setup. The timing accuracy that is achieved allows a fixed phase relation between adjacent rows in the broad spiral, which makes it possible to form 2D lattices. Practical examples show excellent alignment of adjacent rows within a broad spiral both for LIM and EBR mastering.

The bits are ordered on a hexagonal lattice because it offers a 15% higher packing density compared to the square lattice. A digital ‘1’ is represented by a pit. This pit is chosen considerably smaller than the size of the hexagonal unit-cell. This is needed to avoid so-called ‘signal folding’. With signal folding we mean the occurrence of a non-linearity in the read-out due to the fact that a large contiguous pit-area would cause no diffraction. Both a large non-pit area (land) and a large pit-area will act as a perfect mirror and show identical read-out signals.

By placing the bits on a lattice the complete 2D inter symbol interference can be taken into account in the detection process. The interference from adjacent tracks is not a noise contribution anymore, but valuable information that can be used to improve the detection performance. In contrast to the cross-talk cancellation applied in the 1D case, the 2D signal processing allows more freedom in engineering the
2.7 Conclusions

noise power spectral density in radial direction by transforming the optical channel to a complete 2D partial response channel. The improved detection performance is used to increase the storage density, especially in the radial direction.
Chapter 3

Characterization of the 2D Channel

3.1 Introduction

In this chapter we will make a more detailed analysis of the 2D channel by modelling the system and fitting these models to experimental results. We consider the complete 2D system. By this we mean the disc or medium containing the data, the optical read-out system and the front-end electronics up to and including the signal amplifiers. In Section 3.2 we discuss the 2D lattice on which the bits are arranged on the medium, thereby introducing the notion of a reciprocal lattice. The next section is devoted to the modelling of the optical read-out system. A scalar diffraction model is discussed which is suitable for signal-processing purposes, in the sense that it requires low calculation complexity in order to obtain the synthetic replay signals for long bit sequences. Its correspondence with the experimentally measured signal waveforms is very good. This heuristically shows that there is no need for our 2D channel to consider full-fledged vector diffraction computations. These are several orders of magnitude more computationally intensive, and therefore not of any practical use for our purpose. The model is used in various practical examples to provide insight in the behavior of the optical read-out mechanism in nominal cases as well as cases that deviate from nominal, such as a tilted objective lens with respect to the disc. Combining the lattice analysis results and the results of the scalar diffraction model demonstrates the feasibility of an increase in density by a factor of 2 compared to the BD format for the same parameters of the optical system. This is discussed in Section 3.4. The information in the optical domain is then converted to the electronic domain by a segmented photo detector. In Section 3.5 this detector is discussed in conjunction with its noise sources. For completeness also an overview of other noise sources is given in the same section. Finally, the results of the analysis in this chapter are compared and fitted to practical measurements.
3.2 2D Lattice Characteristics

In this section we will analyse the lattice on which the bits are organized. From this lattice the corresponding lattice in the 2D spatial frequency space will be derived. Results of this analysis will be used to assess the spectral content of the data in relation to the frequency transfer characteristics of the optical channel.

**Hexagonal Lattice**  As already discussed in Chapter 2, the bits are stored in a meta-spiral consisting of a number of bit-rows stacked upon each other with a fixed phase relation in the radial direction (which is normal to the meta-spiral). In this way the bits form a 2D lattice. A 2D close packed hexagonal ordering was chosen because it has a 15% higher packing density than the square lattice as used in [14] [31] [32] [33]. The 2D hexagonal lattice has a six-fold symmetry. In terms of an $xy$ cartesian coordinate system that is defined with respect to the tangential (x) and the radial (y) direction of the meta-spiral, the lattice can be defined by two base vectors:

\[
\begin{align*}
  i : & \quad u_0 = a_H \begin{pmatrix} 1 \\ 0 \end{pmatrix} \\
  j : & \quad u_1 = \frac{a_H}{2} \left( -1, \sqrt{3} \right)
\end{align*}
\]  

(3.1)

with $a_H$ the hexagonal lattice constant. An index system $\Lambda = (i, j)$ is used to define points on the lattice. In crystallography, the above lattice is known as the ‘real space lattice’. A convenient way to denote such a lattice is by means of its ‘generating matrix’ [34], given by:

\[
L = \begin{bmatrix} u_{0x} & u_{1x} \\ u_{0y} & u_{1y} \end{bmatrix} = \frac{a_H}{2} \begin{bmatrix} 2 & -1 \\ 0 & \sqrt{3} \end{bmatrix}
\]  

(3.2)

The real space lattice is drawn in Fig. 3.1. Note that the base vectors and their corresponding generating matrix are not unique for the hexagonal lattice. For instance, the angle between the two base vectors $u_0$ and $u_1$ in Fig. 3.1 is chosen to be $120^\circ$. An alternative choice would be to set this angle to $60^\circ$.

In Fig. 3.1 two types of lattice planes are shown (dashed lines). Diffraction at any set of 2D lattice planes occurs in the direction normal to these planes. The vector normal to the planes, with a length equal to the inverse of the repeat period of the planes, is commonly referred to as the diffraction vector of these planes. A set of diffraction vectors constitute another 2D lattice, which is called the ‘reciprocal (space) lattice’, which is laid out in the 2D spatial frequency plane. The diffraction vectors are unique for a particular set of two real space lattice vectors.

---

1. The ‘generating matrix’ of this real space lattice has the property that it transforms the lattice coordinates in the hexagonal coordinate system back to the coordinates in the Cartesian coordinate system.
Reciprocal Space Lattice  The basic vectors of the reciprocal lattice of the 2D hexagonal bit-lattice with lattice vectors \((u_0, u_1)\) are denoted as \((v_0, v_1)\). They have the property that:

\[
u_k \cdot v_l = \delta_{kl}
\]  

(3.3)

where the dot indicates the inner-product of the two vectors and \(\delta_{kl}\) is the Kronecker delta-function. This indicates that the real space lattice and its reciprocal space counterpart are ortho-normal. In matrix notation we can write this property as:

\[
\begin{bmatrix}
(u_0 \cdot v_0) & (u_0 \cdot v_1) \\
(u_1 \cdot v_0) & (u_1 \cdot v_1)
\end{bmatrix} = \begin{bmatrix} u_{0x} & u_{0y} \\ u_{1x} & u_{1y} \end{bmatrix} \begin{bmatrix} v_{0x} & v_{1x} \\ v_{0y} & v_{1y} \end{bmatrix} = LL^* = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}.
\]

(3.4)

This means that the generating matrix \(L^*\) of the reciprocal space lattice is equal to the inverse transposed of the generating matrix of the real space lattice, that is

\[
L^* = (L^{-1})^T,
\]

(3.5)

which for our hexagonal lattice results in:

\[
L^* = \frac{2}{a_H} \frac{1}{2\sqrt{3}} \begin{bmatrix} \sqrt{3} & 0 \\ 1 & 2 \end{bmatrix}.
\]

(3.6)

\(^2\)The ‘generating matrix’ of the reciprocal lattice has the property that it transforms the lattice coordinates in the Cartesian coordinate system back to the coordinates in the hexagonal coordinate system.
The reciprocal lattice is also a hexagonal lattice with reciprocal lattice constant $a_H^* = \frac{2}{\sqrt{3}a_H}$. The two base vectors $v_0$ and $v_1$ of the reciprocal lattice are shown in Fig. 3.2. One can define a fundamental domain (2D Nyquist cell, Brillouin zone or Wigner-Seitz cell [35]) as an area that has the property that adjacent fundamental domains do not overlap and must fill the complete 2D space (see Fig. 3.2). The size of the fundamental domain of the reciprocal lattice is the inverse of the size of the fundamental domain of the real space lattice. The 2D Shannon sampling theorem now states that a function in the spatial domain can be retrieved from its sampled version if the Fourier Transform is contained within a fundamental domain of the reciprocal lattice [34]. The excess bandwidth is defined similarly as in the 1D case, as the fraction of the bandwidth of the channel that extends beyond the fundamental domain. Inter symbol interference (ISI) increases rapidly as the excess bandwidth decreases. We will see later in this chapter that we are even in the situation of negative excess bandwidth for the final goal of TwoDOS (50 GB of capacity for BD optical parameters). This will have some consequences in e.g. equalization where a full-response target cannot be approached anymore.

**Figure 3.2:** Reciprocal lattice with two base vectors derived from the base vectors in real space, as defined in Fig. 3.1. Note that, in order to satisfy Eq. (3.3) both lattices are rotated over $30^\circ$ with respect to each other, and that the real space base vectors make an angle of $120^\circ$, whereas the reciprocal space base vectors make an angle of $60^\circ$. 
3.3 Modelling Read Out of TwoDOS Information

The idea behind 2D optical storage is to increase the areal density by applying more sophisticated signal processing, while using the same optical system as in 1D. In this way the density increase that we can achieve by applying 2D signal processing is orthogonal to the density increase that could be reached by scaling of the physical parameters $\lambda$ and NA. This means that, from a physical point of view, the 2D optical channel is not different from the 1D optical channel. However, for 1D optical storage it is common practice to model the intrinsically two-dimensional optical channel with a 1D model. In this section we will describe a 2D model for the read-out of TwoDOS media that is suitable for signal processing. The latter requirement means that the signal waveform can be written down as a direct closed-form expression, which is a function of the channel bits. This allows the simulation of signal waveforms for long 2D sequences of channel bits within a limited amount of time. As a basis we start from a standard scalar diffraction model, which due to its computational requirements is not suitable for signal processing purposes. This model is rewritten in a formalism with linear and bi-linear kernels, and is applied to some examples that are of practical interest for TwoDOS. An even further simplification of the bi-linear model is possible by imposing circular symmetry on the channel and limiting the interaction between the bits to the set of nearest neighbour bits. This will result in a so-called 4-parameter model. Another simplification is to neglect the generally smaller bi-linear contributions, resulting in a very simple linear model as shown in Fig. 3.3.

Figure 3.3: Overview of the different optical channel models.
### 3.3.1 Scalar Diffraction Model

From Fourier Optics [36] it is known that the spot profile can be calculated by applying a two-dimensional Fourier Transform on the complex-valued optical wavefront in the plane of the entrance pupil of the objective lens. With ‘wavefront’ we mean the phase and amplitude profile of the light that is focused by the objective lens, \( U(\nu_x, \nu_y) = E(\nu_x, \nu_y)e^{j\phi(\nu_x, \nu_y)} \), with \((\nu_x, \nu_y)\) the spatial frequency in \(x\) and \(y\) direction respectively. Each spatial frequency corresponds to a well-defined position in the pupil. For the moment we will assume the ideal case that the circular aperture of the objective lens is homogeneously filled with light of uniform amplitude and constant phase. This is shown in two plots in Fig. 3.4. We consider a position \( \mathbf{r}_p \) on the medium (real space) where the light beam is focused. Using the Fourier transform, denoted with operator \( \mathcal{F} \), one can write the (in general complex-valued) spot profile (i.e. the probe function) at this position as:

\[
P(\mathbf{r} - \mathbf{r}_p) = \mathcal{F}_{\mathbf{r} \rightarrow \nu} [U(\nu)]
\]

where the aperture function is written as \( U(\nu) \) with \( \nu \) the 2D spatial frequency vector \((\nu_x, \nu_y)\), i.e. the position in the pupil of the objective lens. The intensity profile of the spot centered at \( \mathbf{r}_p \) in real space is simply derived by squaring the absolute value of the probe function:

\[
I(\mathbf{r} - \mathbf{r}_p) = |P(\mathbf{r} - \mathbf{r}_p)|^2.
\]

**Figure 3.4:** Ideal wavefront in the plane of the entrance pupil of the objective lens. The use of a circular aperture is obvious.
### 3.3 Modelling Read Out of TwoDOS Information

Analytical calculation of the intensity profile for the nominal pupil function leads to the well-know Airy pattern:

$$I(r - r_p) = \left[ \frac{2J_1 \left( \frac{2\pi \lambda}{\lambda} \right)}{2J_1 \left( \frac{2\pi \lambda}{\lambda} \right) \left| r - r_p \right|} \right]^2 I(0) \tag{3.9}$$

with $J_1$ the first order Bessel function. Fig. 3.5 shows the 2D and 1D representation of the analytically calculated, normalized Airy profile. In the 1D plot also the cross section through the center of the spot profile is shown. The spot is incident on the information layer where it is diffracted by the pit and land structures on the disc. The information layer can be represented by a complex-valued reflection function which we denote $R(r)$. Normally, the central aperture (CA) detection method is used where the power of the reflected optical wavefront is integrated within the detection aperture of the objective lens. The back-propagation from the disc towards the lens can again be modelled as a 2D Fourier transform. The replay signal $y(r_p)$, obtained while scanning the disc with the optical spot centered at position $r_p$, can therefore be written as:

$$y(r_p) = \int_{CA} \left| \mathcal{F}_{r \rightarrow \nu} \left[ P(r - r_p)R(r) \right] \right|^2 d\nu_s. \tag{3.10}$$

This model is practically not very suitable to calculate waveforms over large areas of the disc. At the input of the receiver one needs a synchronous waveform that is sampled at each instant $kT$ with $T$ the bit spacing in the time domain and $k$ the time index. This means that the 2D Fourier transform, needed for the back-propagation...
Characterization of the 2D Channel

of the wavefront of the light, must be calculated many times, at least equal to the number of possible bit-configurations on the synchronous sampling grid within the ISI-span of the optical spot (that is taken into account by the model).

**Linear Model** To overcome the large calculation complexity (and corresponding long processing times) one can simplify the inherently non-linear model of Eq. 3.10 to a very simple linear model that only considers linear interferences in the optical channel. In this model it is assumed that the linear components are much more important than the non-linear ones. In this linear approximation the synchronous replay signal can be modelled as the convolution of the unipolar channel bits \( b_k \in \{0, 1\} \) on the disc with the impulse response function (IRF) of the channel, denoted as \( h_k \):

\[
y(k) = \sum_i h_i b_{k-i}. \tag{3.11}
\]

or equivalently for the two-dimensional channel:

\[
y(k, n) = \sum_{i,j} h_{2D}^{ij} b_{k-i,n-j}. \tag{3.12}
\]

The second index \( n \) denotes the row number. The impulse response function can be derived from the Fourier transform of the modulation transfer function (MTF), denoted as \( H(\nu_s) \). For the nominal case as shown in Fig. 3.4 the MTF has the well-known low-pass behavior with a hard cut-off at \( \nu_c = \frac{2NA}{\lambda} \). The normalized function as derived in the Braat-Hopkins formalism \[8\] can be written as:

\[
H(\nu_s, \theta) = \begin{cases} 
\frac{2}{\pi} \left[ \arccos\left(\frac{\nu_s}{\nu_c}\right) - \frac{\nu_s}{\nu_c} \sqrt{1 - \left(\frac{\nu_s}{\nu_c}\right)^2} \right] & |\nu_s| \leq \nu_c, \\
0 & |\nu_s| > \nu_c.
\end{cases} \tag{3.13}
\]

In general the MTF depends on the azimuth \( \theta \); in the nominal case however as described above it is circularly symmetric as is shown in Fig. 3.6.

A geometrical derivation of this equation that provides quite some insight in the mechanism of the optical read-out starts from the consideration of periodic pit patterns in the disc as grating structures (actually, the pits are considered to be of infinite extent in the radial direction, such that only one-dimensional diffraction occurs). Furthermore, the focused optical beam is considered such that each point within the aperture of the objective lens emits a spherical wavefront, which is at the medium well described by a plane wave travelling at an unique angle towards the medium \[36\]. When such a plane wave hits the grating at a certain angle \( \alpha_i \), it is diffracted under a different angle \( \alpha_o \) according to the well known grating equation:
3.3 Modelling Read Out of TwoDOS Information

Figure 3.6: Nominal Modulation Transfer Function (MTF).

\[
\sin(\alpha_i) - \sin(\alpha_o) = \frac{m\lambda}{p}, \tag{3.14}
\]

where \(m\) is the diffraction order and \(p\) is the period of the pit pattern on the disc. The configuration is schematically represented in Fig. 3.7, where A and B serve as scattering centers for the incoming plane wave. The equation can be easily derived by noticing that the distance \((AC - BD)\) between the wavefronts \(BC\) and \(AD\) must be equal to an integer number of wavelengths of the light in order to have constructive interference of all diffracted wavefronts.

Figure 3.7: Diffraction from a grating.

The diffraction of the incident light cone is shown in Fig. 3.8. In the model only
two diffracted orders are generated because we consider here a so-called ‘single-
frequency’ grating with a 1D periodicity in the form of a sine function. This al-
lows us to treat the MTF as a frequency transfer function. The two diffracted orders
will interfere with the reflected $0^{th}$ order causing the actual modulation of the signal
waveform that is generated upon detection of the integrated power within the ‘central
aperture’ of the objective lens. Mathematically this is represented by:

$$H(\nu_x, \nu_y) = \int \int U^*(\nu_x0 - \nu_x, \nu_y0 - \nu_y)U(\nu_x0, \nu_y0) d\nu_x0 d\nu_y0,$$

which means that the MTF can be calculated by the autocorrelation of the complex-valued wavefront. Examples of this calculation are shown in Fig. 3.9. For the nominal case the value of the MTF for a given frequency is simply computed as the area of the overlap region between the $0^{th}$ order light cone and the diffracted light cone. This is indicated as the dark gray area in Fig. 3.8.

**Deviations from the nominal spot**  Until now an ideal filling of the aperture of the objective lens has been assumed. This means that the light is homogeneous across the complete aperture and that the phase is constant. In practice the intensity distribution across the aperture is better represented by a truncated Gaussian function:

$$E(\rho, \theta) = e^{-\frac{\rho^2}{2\sigma^2}}$$

where coordinates ($\nu_x, \nu_y$) are transformed to their polar equivalents ($\rho, \theta$). The edge of the ‘central aperture’ (CA) corresponds to $\rho=1$. Generally the Gaussian roll-off with increasing radius in the aperture is characterized by a single parameter, denoted as ‘rim-intensity’, which is equal to:
3.3 Modelling Read Out of TwoDOS Information

\[ \text{rim intensity} = e^{-\frac{1}{2\sigma^2}}. \]  

(3.17)

A rim-intensity lower than 100% will broaden the center part of the optical spot and will reduce the height of the side lobes. Additionally, optical aberrations that might be present, are modelled as phase variations \( \varphi(\rho, \theta) \) within the aperture of the objective lens. Normally these aberrations are split into primary optical aberrations like defocus, coma, and astigmatism. The primary aberrations can be described by a set of polynomials known as Zernike polynomials that are orthogonal to each other over the circular aperture of the lens. Due to this orthogonality an aberrated wavefront can be modelled as the sum of different polynomials by fitting the Zernike coefficients \( Z_{m,n} \):

\[ \varphi(\rho, \theta) = \sum_{m,n} Z_{m,n} \cos(m\theta)\rho^n \]  

(3.18)

An overview of the optical aberrations with their Zernike polynomials is given in [37]. One of the most prominent aberrations is coma, represented by \( Z_{3,1} \), which is e.g. caused by a tilt of the objective lens with respect to the optical disc. The corresponding wavefront and spot-profile are shown in Fig. 3.10. The spot intensity profile is normalized to the maximum intensity value for a nominal spot in the aberration-free case.

**Bi-linear Model** The linear model is very convenient because of its low complexity and it is a quite good approximation when applied to 1D optical storage. Here, the
spot diameter is larger than the radial width of a pit on the disc. This means that when the spot scans a pit the reflected light beam loses intensity by diffraction outside the central aperture (CA) of the lens, even in the case of successive pit-bits that are written as one elongated mark in the direction of the track. In the 2D-case however, the problem of signal-folding occurs as described in Section 2.2. In the TwoDOS format these non-linear effects contribute significantly to the read-out signal and therefore a signal processing model for the scalar diffraction needs to be devised that combines the ability to model non-linear components in the optical channel with a low computational complexity. As already mentioned, the computational complexity originates from the fact that all bit-configurations within the ISI-span must be considered. For example for an ISI-span of 3 hexagonal shells containing 19 bit positions the number of possible bit-patterns is $2^{19}$. However, we will show that the required amount of computations can be decreased drastically [38]. For a linear write-channel it is satisfactory to add only bi-linear contributions to the linear system model. This means that only combinations of two pit-positions need to be considered. To do this we first write the detected signal resulting from an optical spot centered at position $r_p$ on the disc as

$$y(r_p) = \langle \psi | \psi \rangle,$$  \hspace{1cm} (3.19)
3.3 Modelling Read Out of TwoDOS Information

\[< \phi | \varphi > = \int_{\text{CA}} \phi^*(v_s) \varphi(v_s) dv_s, \quad (3.20)\]

From inspection of Eq. 3.19 Eq. 3.10, it is obvious that the complex-valued wavefront \(|\psi >\) has to be written as:

\[|\psi > = \mathcal{F}_{\nu \rightarrow r} [P(r - r_p) R(r)]. \quad (3.21)\]

Note that the bracket notation is used as introduced by Dirac [39] in the field of quantum mechanics. It allows a compact notation in the sequel of this section. The disc reflection function in Eq. 3.21 can be conveniently written as:

\[R(r) = 1 + \sum_{i} a_i W(r - r_i), \quad (3.22)\]

which is a linear function of the unipolar channel bits \(b_i\). The window function \(W(r - r_i)\) defines the area of the pit on the disc which is centered at position \(r_i\). It is equal to 1 inside the pit area and 0 elsewhere. Furthermore, \(a_i\) represents the bit value at position \(r_i\). When no pit is present the reflection must be 1 and \(a_i\) can simply be equal to 0. In case a pit is present the reflection function must be equal to \(e^{j\Phi_p}\), with \(\Phi_p\) the double-pass phase depth of the pit (i.e. from the top of the pit to its bottom and back). This means that in that case \(a_i\) must be equal to \(e^{j\Phi_p} - 1\). This leads to the following convenient definition:

\[a_i = b_i (e^{j\Phi_p} - 1). \quad (3.23)\]

The bits, \(b_i \in \{0, 1\}\) are the unipolar channel bits, representing the pits that are imprinted into the substrate with a predefined depth to induce the phase difference \(\Phi_p\), between light reflected from a land area and light reflected from a pit area. Combining Eq. 3.22 and Eq. 3.21 results in:

\[|\psi > = \mathcal{F}_{\nu \rightarrow r} \left[ P(r - r_p) \left(1 + \sum_{i} a_i W(r - r_i)\right) \right]. \quad (3.24)\]

Now the Fourier transform that models the back propagation of the diffracted orders towards the central aperture of the objective lens can be interchanged with the summation over the wavefront contributions originating from the different bits [38]. This is allowed because the system is linear in the electric fields. Non-linearity in the channel is introduced because the light intensity is detected (and not the field). Using the bracket notation, \(|\psi >\) can be conveniently decomposed into a contribution from the ‘all-land’ case and a sum of contributions from all pit-bits that are within the area of the laser spot:

\[|\psi > = |\psi_L > + \sum_{i} a_i |\psi_i > \quad (3.25)\]
with
\[ |\psi_L> = \mathcal{F}_{r\rightarrow \nu_s}[P(r-r_P)], \]
which is simply the all-land reflection from a perfect mirror without any pits, and
\[ |\psi_i> = \mathcal{F}_{r\rightarrow \nu_s}[P(r-r_P)W(r-r_i)], \]
which is the wavefront contribution from a single pit at position \( r_i \) with a pit-area defined by \( W(r-r_i) \).

Using (3.25) and (3.19) leads to:
\[ y(r_P) = \langle \psi_L|\psi_L > + \sum_i a_i < \psi_L|\psi_i > + \sum_i a_i^* < \psi_i|\psi_L > + \sum_{i,j} a_i^* a_j < \psi_j|\psi_i > \]
(3.28)

A first observation is that the optical read-out system is intrinsically non-linear (or more accurate bi-linear), due to the fact that the power of the light is detected. Normalization with respect to the constant term \( < \psi_L|\psi_L > \) and reworking results in the final form of the bi-linear model:
\[ y(r_P) = 1 - \sum_i c_i b_i + \sum_{i\neq j} d_{i,j} b_i b_j, \]
(3.29)

where the coefficients \( c_i \) for the linear inter-symbol interference (ISI) and \( d_{i,j} \) for the non-linear ISI are given by:
\[ c_i = 2(1 - \cos(\Phi))(< \psi_L|\psi_i > - < \psi_i|\psi_i >) \]
(3.30)
\[ d_{i,j} = 2(1 - \cos(\Phi)) \text{Re}\langle \psi_i|\psi_j >. \]

For the linear ISI coefficients one recognizes the interference term of the pit with the surrounding land (the linear interference term) and the interference of the pit with itself (the so-called self-interference which is in fact a special case of non-linear interference).

**Reduction of Computational Complexity**  
By interchanging the summation over the waveform contributions with the Fourier transform for back propagation the calculation complexity can be reduced drastically. The number of FFTs needed to perform the back-propagation for all possible bit-patterns within a ISI span of 3-shells is normally equal to \( 2^{n_P} \) with \( n_P = 37 \) (the number of pit positions). After interchanging the summation and the Fourier transform this reduces to \( \frac{n_P(n_P+1)}{2} = 703 \) bi-linear components plus 37 linear components (note that for the bi-linear components \( d_{i,j} = d_{j,i} \)). This is still a considerable amount of FFT-calculations. Nevertheless, it results in a workable situation for calculating the coefficients of the bi-linear model.
3.3 Modelling Read Out of TwoDOS Information

3.3.2 Results of Scalar Diffraction Model

As a practical example the model is applied to the hexagonal format with both circular pit holes and the RLL format with elongated pit marks. For the hexagonal format, lattice constants $a_H = 138$ nm and $a_H = 165$ nm are chosen. Signal levels for all possible hexagonal clusters are calculated. A hexagonal cluster consists of a central bit at the lattice center, and of six nearest neighbour bits at the neighbouring lattice sites. This results in $2^7 = 128$ possible bit configurations. Plotting all these levels in a single plot would make interpretation of the main effects difficult. Fortunately, because the (nominal) spot is circularly symmetric it only matters to identify the central bit and the number of pit (or non-pit) bits among the nearest neighbour bits (neglecting non-linear effects of adjacent nearest neighbour bits). This results in a total of 14 unique signal levels, where each of the signal levels is $N_n$-fold degenerate with $N_n = \binom{6}{n}$ and $n$ the number of nearest-neighbour pit-bits. The cluster types are numbered as $7b_0 + \sum_{i=1}^{6} b_i$. An example of such a plot with a hexagonal lattice parameter $a_H=165$ nm and a pit-hole diameter $b=120$ nm is shown in Fig. 3.11. The levels are normalized with respect to the ‘all-land’ reflection level, which is equal to 1.

![Figure 3.11: Signal levels for different hexagonal clusters ($a_H=165$ nm, $b=120$ nm, $\Phi_p=180^\circ$).](image)

In this plot, level-identification is only based on the six nearest neighbouring
lattice cells (the so-called first shell). The scalar diffraction model, however, also takes the other shells into account (up to the 3rd one in this particular case). Signal-level variation due to the influence of these shells is not shown. Instead an average is taken across all possible bit-patterns in these shells. This explains the fact that the signal level of the all land cluster (with cluster-index equal to 0) is not equal to 1 as we would expect from a normalized model.

**Choice of pit-hole diameter**  A clear roll-off as function of the number of nearest neighbour pit-bits is observed. This roll-off is not linear as was already explained in Section 2.2, and which is also expected from the non-zero bi-linear ISI coefficients as derived in the previous section. Simulations using a range of pit-hole diameters show the dependence of the amount of non-linearity on the pit-hole diameter. The results are shown in Fig. 3.12. As already expected from Section 2.2 the amount of non-linearity increases when larger pits are present on the disc. From this point of view it would be beneficial to master pits with a diameter as small as possible to achieve a linearity that is as good as possible. However, for very small pits the total signal modulation drops and a trade-off must be found between linearity and signal modulation. An optimum seems to be obtained with a pit-hole diameter such that the pit-hole covers about half the area of the available hexagonal cell. A qualitative explanation of this optimum is that one needs 50% land and 50% pit area to have complete destructive interference between the reflected light from both areas (neglecting amplitude variations over the spot-profile). The optimum pit-hole diameter \( b_{50\%} \) equals:

\[
b_{50\%} = \sqrt{\frac{\sqrt{3}}{\pi}} a_H
\]  

(3.31)

To verify this choice, the influence of non-linearity on the performance of the bit-detector has been investigated by varying the pit-diameter in a simulation using a stripe-wise Viterbi detector (see Chapter ??). No equalizer or electronic non-linearity compensation has been applied at this stage. The signal-to-noise level (SNR) is defined with respect to half the peak-amplitude of the signal. Additive White Gaussian Noise (AWGN) is assumed for simplicity. It is clear that the best value is close to \( b_{50\%} \) (100 nm for \( a_H=138 \) nm, and 120 nm for \( a_H=165 \) nm). Note that in a final system a 2D equalizer must be incorporated to equalize to a compact target response which is needed to achieve a limited state complexity in the Viterbi detector. However, the optimum pit-diameter is not expected to deviate significantly from the optimum one obtained in this simulation.

**Signal Level Overlap**  The scalar diffraction simulation results for the optimum pit-hole diameter are plotted in Fig. 3.14 for a lattice constant \( a_H=138 \) nm and for
3.3 Modelling Read Out of TwoDOS Information

Figure 3.12: Signal levels for various pit-hole diameters $b$ ($a_H=165$ nm, $\Phi_p=180^\circ$).

It is clear that the number of overlapping signal levels between the branch with a center bit equal to ‘1’ and the branch with a center bit equal to ‘0’ increases with increasing densities. This makes detection more difficult.

Validity of 14-Level Plot The signal level plot with 14 levels was adopted for ease of interpretation. However, it obscures the actual signal levels partly due to the averaging over all possible clusters that belong to a certain cluster type. Fig. 3.15 shows two plots where all 128 signal levels are plotted, but still the classification in 14 cluster types is carried out. Those points that have the highest degeneracy also show
Figure 3.13: Simulations results of the stripe-wise Viterbi Detector on a 7-row broad spiral at densities of 1.4x and 2x that of BD. The pit-diameter, \( b \) is varied. Dotted lines: \( a_H = 165 \text{ nm} \); Solid lines: \( a_H = 138 \text{ nm} \).

Figure 3.14: Signal levels for different densities, \( \Phi_p = 180^\circ \), optimum pit-hole diameter \( b = b_{50\%} \).

the largest variations in cluster levels. It is also clear when comparing (a) and (b) in Fig. 3.15 that the higher the non-linearity the larger the variation in the levels.
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3.3.3 4-Parameter Model

For some applications, like the definition of a target impulse response to be used in Viterbi processing (see Section ??), it may be convenient to further simplify the bilinear model. This simplification can be realized by explicitly imposing rotational symmetry, neglecting all the interferences beyond the nearest neighbours and by neglecting the cross-interference between nearest neighbour bits that are separated by a distance larger than \( a_H \). Furthermore, the phase depth of the pits is taken \( \Phi = \pi \). This model is presented in [38] and can be written as:

\[
y(k) = 1 - 4b_0(l_0 - s_{0,0}) - 4n_1(l_1 - s_{1,1}) + 8b_0n_1x_{0,1} + 8p_{1,1}[1]x_{1,1}[1] \quad (3.32)
\]

The parameters represent the following:

- \( b_0 \): bit-value of the central bit (‘1’ for pit, ‘0’ for land);
- \( n_1 \): number of nearest-neighbors (of the central bit, in shell ‘1’) being of the pit-type;
- \( l_0 \): tap-value of linear interference for central pit-bit;
- \( l_1 \): tap-value of linear interference of (nearest) neighbour pit-bit;
- \( s_{0,0} \): value for self-interference of central pit-bit;
- \( s_{1,1} \): value for self-interference of (nearest) neighbour pit-bit;
- \( x_{0,1} \): value of cross-interference between central pit-bit and (nearest) neighbour pit-bit;
- \( x_{1,1}[1] \): value of cross-interference between two (nearest) neighbour pit-bits, which are also nearest neighbours of each other;
• $p_{1,1}$: number of pit-pairs that are nearest neighbours of each other.

For a hexagonal lattice parameter, $a_H=165$ nm and a pit diameter, $b=120$ nm the values can be derived based on the full-fledged bi-linear diffraction model (note that the 4-parameter model is only a reasonable approximation for limited densities up to $a_H=165$ nm):

\[
\begin{align*}
l_0 & = <\psi_L | \psi_0 > = 0.142855 \\
s_{0,0} & = <\psi_0 | \psi_0 > = 0.021613 \\
l_1 & = <\psi_L | \psi_1 > = 0.041490 \\
s_{1,1} & = <\psi_1 | \psi_1 > = 0.0056859 \\
x_{0,1} & = <\psi_0 | \psi_1 > = 0.0062178 \\
x_{1,1} & = <\psi_j | \psi_{j+1} > = 0.0031908.
\end{align*}
\]

Here $|\psi_L>$ is the complex valued wavefront at the pupil plane for the all-land situation. The wavefront for a pit at the center or neighbour position is given by $|\psi_0>$ and $|\psi_1>$ respectively. Wavefronts of two adjacent nearest neighbour bits are indicated by $|\psi_j>$ and $|\psi_{j+1}>$.

Fig. 3.16 shows the validity of this approximation for $a_H=165$ nm (top) and $a_H=138$ nm (bottom). The scalar diffraction model results are indicated with solid lines, while the results of the 4-parameter model are shown with dashed lines. It is clear from these graphs that the all-zero cluster for the 4-parameter model now gives a value equal to 1 because no pits beyond the nearest neighbours are taken into account. Furthermore, the 4-parameter model shows a steeper roll-off with increasing number of nearest neighbour pits. Further analysis, which is not repeated here, reveals that the steeper roll-off can be attributed to the omitted cross interference of neighbour bits at a distance larger than $a_H$.

The 4-parameter model is suitable for the use as a target response. It has the advantage that it has only few parameters, which can be optimized rather easily by optimizing the bit error rate as function of these parameters. This is explained in more detail in Chapter ???. Furthermore, it could be used as an adaptive target response. The advantage is that only 4 parameters have to be adapted and therefore adaptation is fast. In this thesis a full 128-level adaptive target will be used (Chapter ??).

### 3.4 Density Calculations

From Section 3.3.1 it became clear that the spectral transfer characteristic of the channel has a low-pass behavior with a hard cut-off at $\frac{2\text{NA}}{\lambda}$. Preferably the optical
channel should pass all spectral content of the data, which means that a circle with radius $\frac{2\lambda}{A}$ includes the complete 2D Nyquist cell in the 2D spatial frequency space. On the other hand the TwoDOS concept targets a density that is at least 2 times the density of the 25 GB BD-format. In the latter format the user bit takes up an equivalent area on the disc equal to:

**Figure 3.16:** Signal-level plots for the 4-parameter model (indicated with dashed lines) compared to the full-fledged bi-linear scalar diffraction model (solid). The occurrence of even negative signal levels indicates that the model is non-physical.
Characterization of the 2D Channel

\[ S_{BD} = \frac{p_t l}{R} = \frac{320 \cdot 75 \text{[nm]}^2}{2/3} = 0.6343 \left( \frac{\lambda}{2 \text{NA}} \right)^2 \]  

(3.34)

where \( p_t \) is the track pitch, \( l \) is the channel bit length and \( R \) is the rate of the 17PP (d=1) modulation code (we omit an additional factor for the overhead of error correction coding, DC-control and frame synchronization because we assume it to be equal for 1D and 2D). It appears that a factor of 2 in density is reached when the cut-off frequency of the 2D MTF forms exactly an inscribed circle in the fundamental domain of the 2D frequency space (see Fig. 3.17). Note that the spectrum (and also the circle) repeats around the reciprocal lattice points in the same way as the repeating spectrum around the Nyquist frequency in the 1D case. The corresponding hexagonal lattice parameter is equal to:

\[ a_H = \frac{1}{\sqrt{3}} \left( \frac{\lambda}{2 \text{NA}} \right) = 137.5 \text{ nm}. \]  

(3.35)

\[ S_{\text{sub}} = \frac{\sqrt{3}}{2} a_H^2 = \frac{1}{2 \sqrt{3}} \left( \frac{\lambda}{2 \text{NA}} \right)^2. \]  

(3.36)

Figure 3.17: Reciprocal lattice with the circles formed by the cut-off frequency of the 2D MTF, which are exactly inscribed circles of the hexagonal fundamental domain.

In that case the size of the hexagonal user bit-cell is:

Note that the rate of the 2D modulation code as designed in Section 4.3 is nearly equal to 1. Comparing (3.34) and (3.36) and accounting for the factor \( \frac{11}{12} \) due to the
3.4 Density Calculations

presence of a guard band of one empty bit-row indeed reveals that in this case the achievable density for TwoDOS is a factor 2 higher than that of BD. For this density the level-plot was already calculated in the previous section (see Fig. 3.14). It is clear from this plot that there is considerable overlap between cluster levels for clusters that have a central pit and a central non-pit (or land). This means that the ‘eye-pattern’, commonly used in 1D optical storage, is completely closed (see Fig. 3.18) and that straightforward detection using a slicer with suitable threshold is not possible anymore. For 1D optical storage Viterbi detection was already proposed to be able to read out higher tangential densities, giving non-open eye-patterns [40]. This type of detection will be extended to the 2D case (Section ??). Also clock recovery, which is normally based on detection of the zero-crossings, has to be done in a different way (Section ??).

![Figure 3.18: ‘Eyepattern’ for aH=137.5 nm and b=51 nm, see also [41].](image)

Square Lattice For the square lattice with lattice constant $a_S$ the situation is shown in Fig. 3.19 in case of a density of 2x BD (again taking the overhead of the guard band into account). For this density the size of the square bit-cell must be equal to:

$$S_{S_{ub}}^S = a_S^2 = \frac{11 S_{BD}^{ub}}{12},$$

leading to a value of $a_S = 0.539 \frac{\lambda}{2NA}=128$ nm. The inverse of this (the reciprocal lattice constant) is slightly smaller than $2 \cdot \frac{2NA}{\lambda}$, meaning that the circles formed by the cut-off of the optical channel would slightly overlap in frequency space. The
frequency of the diagonal lattice planes lies beyond the cut-off of the optical channel. This means that longer alternating bit-sequences in the direction perpendicular to these lattice planes are prone to be erroneously detected as a shifted sequence.

![Diagram of a square lattice showing the overlap of circles formed by the cut-off of the optical channel.](image)

**Figure 3.19:** Frequency space for the square lattice showing the overlap of the circles formed by the cut-off of the optical channel.

3.5 Noise Characterization

This section will start with a brief analysis of the conversion of the signal from the optical to the electronic domain using photo-detectors and current amplifiers. The most important noise sources will be described and a simple model will lead to an estimate of the power spectral density of the electronics noise. To complete the discussion on noise sources an overview will be given of the most important sources of noise in the total system (see also [42]). These are:

- Shot noise.
- Electronics noise.
- Quantization noise.
- Media noise.
3.5 Noise Characterization

The first four will be discussed briefly in subsequent paragraphs. The latter one is discussed in more detail in Section 3.5.1. Although it should be clear from the equations, in some cases dimensions are given to avoid any ambiguity on the definition of the noise parameters.

**Relative Intensity Noise** The intensity of semiconductor lasers fluctuates due to several reasons, but the most important is spontaneous emission at low laser power levels [41]. Due to the fluctuations, the current generated in the photo detector fluctuates as well, thus causing noise. The parameter ‘Relative Intensity Noise’ (RIN in \([\text{dB}\sqrt{\text{Hz}}]\)) is used to describe these intensity (or power) fluctuations and is defined as:

\[
\frac{\sigma_I^2}{I_{pd}^2} = \int_0^{BW} \text{RIN}^2(f)df
\] (3.38)

With \(\sigma_I^2\) the variance of the photo-detector current that can be attributed to variation in laser power and \(I_{pd}\) the average value of the photo-detector current. Frequency is denoted with \(f\) and \(BW\) is the bandwidth of the system (for simplicity reasons we assume an ideal cut-off and do not introduce the notion of effective noise bandwidth). RIN can be regarded constant within the bandwidth of interest for optical recording. Another source of noise may be fluctuations in the current that drives the laser. In general, it can be safely assumed that this noise is low enough to be neglected.

**Shot Noise** Shot Noise arises due to the quantum nature of light. Photons arrive at the photo-detector according to a Poisson distribution. Therefore, even at a constant average incident power on the detector the number of photons arriving within a certain limited amount of time is fluctuating leading to a non-zero variance of the detected photo-current. The corresponding power spectral density \(S\) of this noise is constant and equal to:

\[
S_{\text{shot}} = 2qI_{pd} \left[\frac{A^2}{\text{Hz}}\right]
\] (3.39)

where \(I_{pd}\) represents the current generated in the photo-detector, and \(q\) the electron-charge. In an electronic-circuit-based noise model this noise can be inserted as a current source with rms value:

\[
I_{\text{shot}} = \sqrt{2qI_{pd}BW} \quad [\text{A}]
\] (3.40)

with \(BW\) the bandwidth of the system. Another convenient representation which allows scaling of the bandwidth of the equivalent circuit is obtained by normalizing the rms value to a 1 Hz bandwidth, resulting in current spectral densities \((i_n)\) or voltage spectral densities \((v_n)\). In that case:
\[ i_{\text{shot}} = \sqrt{2qI_{pd}}. \quad \left[ \frac{A}{\sqrt{\text{Hz}}} \right] \quad (3.41) \]

**Electronics Noise**  Electronics Noise is a accumulation of all kinds of noise sources that are present in the integrated circuit of the photo-detector and current amplifier. We will divide these noise sources into two single noise sources, a voltage one and a current one that are calculated at the output of the current amplifier. To obtain a convenient model however, the noise sources are introduced at the input of the amplifier according to Fig. 3.20. The photo-detector is reversely biased such that the depletion layer is wide and the capacitance of the photo-diode denoted with $C_{pd}$ is low. This is important as we will see further on. Another capacitor is present at the input of the amplifier, which represents its total input capacitance.

![Figure 3.20: Equivalent circuit used to model the noise sources in the analogue front-end of an optical storage system.](image)

The voltage noise source can be converted to an equivalent current input noise source such that we have only current sources in our final model:

\[ i_{a,e} = 2\pi f (C_{pd} + C_{in}) v_a. \quad \left[ \frac{A}{\sqrt{\text{Hz}}} \right] \quad (3.42) \]

Note that the equivalent noise current increases linearly with total capacitance $C_t = C_{pd} + C_{in}$. This is the reason why the total capacitance at the input of the detector must be kept as low as possible. Moreover, the noise current increases with frequency. Thus this noise component becomes especially important when the read-out speed increases. A parallel read-out of the disc, as applied in our 2D storage system, allows a lower linear read-out speed compared to a 1D system for an equal data rate of both systems, thereby solving problems caused by the current noise source (assuming that an even higher data rate is not required by the application). Because the noise sources are not correlated, their powers add up to the total noise power. The total equivalent noise current at the output of the amplifier is given by:
3.5 Noise Characterization

\[ i_{n,tot} = G \sqrt{i_n^2 + i_{\text{shot}}^2 + (2\pi f C_i v_a)^2} = G \sqrt{i_n^2 + (2\pi f C_i)^2 v_a^2} \] (3.43)

with \( G \) the gain of the trans-impedance amplifier. For modelling reasons the noise current sources considered so far are taken together in one parameter \( i_n \). Now the total current noise variance originating from voltage noise can be calculated as:

\[ \sigma_{i,v}^2 = \int_0^{\text{BW}} (2\pi f C_i)^2 v_a^2 df = \frac{(2\pi C_i)^2 v_a^2}{3} G^2 \text{BW}^3 \] [A^2] (3.44)

and the total current noise variance originating from all current sources together is:

\[ \sigma_{i,I}^2 = G i_n^2 \text{BW} \] (3.45)

**Quantization Noise** In the ADC, the analog signal is mapped onto a limited set of amplitude levels. This mapping causes an error which results in quantization noise. For a sinusoidal input signal that covers the complete input range of the ADC it can be derived that [43]:

\[ \text{SNR}_Q = 6.02 \cdot \text{ENOB} + 1.76 \] [dB] (3.46)

where ENOB stands for the effective number of bits in the ADC (note that by using ENOB instead of the nominal number of bits also other noise sources in the ADC are incorporated). In practice the number of bits is chosen to be large enough for quantization noise to be negligible. In our case an 8-bit ADC is used.

**Signal to Noise Ratio** Conventionally the total SNR is defined as the integrated noise from zero to the Nyquist-frequency compared to the rms signal value of the I2 runlength (which, for low densities is equal to half of the eye-height). This definition is not suitable anymore for systems where the eye is closed and the highest frequency is beyond the cut-off of the optical channel. Therefore, the SNR for the 2D system is defined as:

\[ \text{SNR} = 20 \log \frac{\sigma_{\text{data}}}{\sigma_{\text{noise}}} = 20 \log \frac{\sqrt{\int_{0}^{f_s/2} i_{\text{data}}^2 df}}{\sqrt{\int_{0}^{f_s/2} i_{n,tot}^2 df}} = 20 \log \frac{K_{MTF} \int_{0}^{f_s/2} H^2(f) df}{\sqrt{\int_{0}^{f_s/2} i_{n,tot}^2 df}} \] (3.47)

where \( f_s \) is the sampling frequency of the ADC, and \( H(f) \) the MTF of the optical channel. Note that the data spectrum as measured by a spectrum analyzer can be fitted by the constant \( K_{MTF} \) and that the noise can be fitted with the various parameters
discussed above. Analytical derivation of the integral of $H^2(f)$ is possible and for a system bandwidth larger than the cut-off frequency $f_c$ of the channel leads to:

$$\int_{0}^{f_c} H^2(f) \, df = f_c \times \frac{4\pi}{\pi^2} \left[ \frac{4\pi}{6} - \frac{64}{45} \right]. \quad (3.48)$$

3.5.1 Media Noise

Media noise originates from small deviations in the storage medium from its ideal form. For example roughness of the mirror-like surface may cause media noise. This type of noise can be easily separated from the previously discussed noise source by first performing a noise measurement on a system where the focused laser beam is reflecting from a static disc followed by a second measurement on a rotating mirror-disc. This type of noise does not contribute significantly to the overall noise and is therefore not discussed in more detail. Another source of media-noise is inaccuracy in the pit-shape. One possible inaccuracy is that the pit-size varies from one pit to the other. This effect is clearly observed in some worst case EBR-mastered discs. One reason for this is that only a limited number of electrons is available to master a single pit, which results in shot-noise. A photograph of a clear, worst-case example is shown in Fig. 3.21. The same source of noise is present in LBR-mastered discs due to the fluctuations in laser power from one pulse to the next during the mastering process. Another source of pit-size variations in EBR masterd stampers is the proximity effect, where the pit-size depends on the number of pits in a wide neighbourhood. The effects of pit-size variation will be discussed in one of the next paragraphs. Finally, pit-position variation is discussed as a last source of media-noise. Here, the ‘center of gravity’ of the pit deviates from its nominal position on the 2D hexagonal lattice.

In general, media noise can be considered as a random (uncorrelated) variation at the level of the medium having a constant power spectral density (PSD). During the read-out process however, it is filtered by the MTF. As a consequence, the replay signal has a PSD corresponding to the transfer of the optical channel. Due to media noise the variances of the HF-signal levels will depend on the type of cluster (with the cluster defined in the hexagonal lattice as a center bit together with its six nearest neighbour bits, see Section 3.3.2). The reason for the cluster dependence of the pit-noise is due to the non-linearity of the read-out channel. It implies that pit-noise will be data-dependent. This knowledge can be used to derive some qualitative estimates on the amount of media noise during experiments. For a fair comparison, no variance due to the degeneration of the levels is allowed. Therefore, the set of clusters that give unique levels in a circular symmetric channel model has 26 entries. Any arbitrary cluster (1 of 128) can be derived from these 26 basic types by rotation or point-inversion with respect to the center bit. The 26 cluster are listed in Fig. 3.22 and in Table 3.1 where the notation ‘c.nnnnnn’ denotes the center bit followed by the nearest
3.5 Noise Characterization

Figure 3.21: Scanning electron microscope image of a replica of stamper E378 showing clear pit-size variation. Note that this is a worst-case example to make the phenomenon more clear and that the discs used in the ‘final’ evaluation show a much better uniformity.

neighbour bits. Note that the relative ordering of the neighbour bits do matter, but not its cyclic permutations.

Figure 3.22: Overview of 7-bit cluster types.

Media noise simulations To simulate media noise we will start with the disc reflection function of Eq. (3.22). and rewrite it slightly as:

\[ R(r) = 1 + \sum a_i W_i(r - r_i - \delta r_i) \] (3.49)
Table 3.1: Overview of 7-bit cluster types.

<table>
<thead>
<tr>
<th>cluster nr.</th>
<th>c.nnnnnn</th>
<th>cluster nr.</th>
<th>c.nnnnnn</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0.000000</td>
<td>13</td>
<td>1.000000</td>
</tr>
<tr>
<td>1</td>
<td>0.100000</td>
<td>14</td>
<td>1.100000</td>
</tr>
<tr>
<td>2</td>
<td>0.110000</td>
<td>15</td>
<td>1.110000</td>
</tr>
<tr>
<td>3</td>
<td>0.101000</td>
<td>16</td>
<td>1.101000</td>
</tr>
<tr>
<td>4</td>
<td>0.100100</td>
<td>17</td>
<td>1.100100</td>
</tr>
<tr>
<td>5</td>
<td>0.111000</td>
<td>18</td>
<td>1.111000</td>
</tr>
<tr>
<td>6</td>
<td>0.110100</td>
<td>19</td>
<td>1.110100</td>
</tr>
<tr>
<td>7</td>
<td>0.101010</td>
<td>20</td>
<td>1.101010</td>
</tr>
<tr>
<td>8</td>
<td>0.111100</td>
<td>21</td>
<td>1.111100</td>
</tr>
<tr>
<td>9</td>
<td>0.111010</td>
<td>22</td>
<td>1.111010</td>
</tr>
<tr>
<td>10</td>
<td>0.110110</td>
<td>23</td>
<td>1.110110</td>
</tr>
<tr>
<td>11</td>
<td>0.111110</td>
<td>24</td>
<td>1.111110</td>
</tr>
<tr>
<td>12</td>
<td>0.111111</td>
<td>25</td>
<td>1.111111</td>
</tr>
</tbody>
</table>

Pit-position noise is modelled as a random displacement vector \( r_i \) and pit-size noise is considered with respect to the nominal pit-size window \( W \) as:

\[
W_i = W + \delta W_i.
\] (3.50)

**Pit-size noise** The pit-size variations are practically accounted for by calculating tables of pit-wave functions for a large number of different pit-sizes. Pit-size noise is quantized through a parameter \( \sigma_{\delta W} \) with dimension of an area (m\(^2\)). However, to be able to compare simulations and measurements at different areal densities the pit-size noise is denoted relative to the area of a hexagonal fundamental domain:

\[
q_{\delta W} = \frac{2\sigma_{\delta W}}{\sqrt{3}a_H^2}.
\] (3.51)

To get some insight into the amount of (tolerable) pit-size noise, simulations are performed based on the scalar diffraction model. In each simulation a set of wavefront contributions \( |\psi| \) (see Eq. 3.19) is calculated for a single pit with various pit-sizes that is located at different lattice positions (no position variation is included yet). Then the replay waveform is simulated by picking wavefronts from the table under the assumption that the pit-size is distributed normally with standard deviation \( \sigma_{\delta W} \). Average signal levels are derived and also the standard deviations for each of the clusters are calculated. The result for \( q_{\delta W} = 6.3\% \) is shown in Fig. 3.23 for various nominal pit-sizes. Further simulations show that \( \sigma_{\delta W} \) is linear with \( q_{\delta W} \) for not too large values of \( q_{\delta W} \).
3.5 Noise Characterization

Figure 3.23: Cluster dependent standard deviation of the detected signal level (scalar diffraction model) in case of media noise due to pit-size variation: $q_{\delta W} = 6.3\%$; $a_H = 138$ nm.

Pit-position noise  Pit-position noise is taken into account by approximating the pit-wave function of the displaced pit as follows:

$$
|\psi_i(r_i + \delta r_i, W + \delta W_i) > = e^{i2\pi \nu_s \cdot \delta r_i} P(r_i - r_p + \delta r_i) \frac{P(r_i - r_p)}{P(r_i - r_p)} |\psi_i(r_i, W + \delta W_i) >
$$

A displacement at the media plane is represented by a phase factor $2\pi \nu_s \cdot \delta r_i$ in the pupil plane due to the Fourier transform that models the back propagation. The 2D frequency vector in the plane of the pupil is denoted by $\nu_s$. A second factor $\frac{P(r_i - r_p + \delta r_i)}{P(r_i - r_p)}$ is introduced that accounts for the ratio between the amplitude of the spot at the nominal pit-position and at the displaced pit-position. Note that for the non-displaced pit-wave function one of the table entries is chosen depending on the pit size. In this way a combined simulation of pit-size and pit-position variation is possible. The whole scheme is visualized in Fig. 3.24. Pit-position noise is quantized through the parameter $\sigma_{\delta r}$ which has the dimension of length (m). Again a relative measure is adopted which is normalized to the hexagonal lattice constant:

$$
q_{\delta r} = \frac{\sigma_{\delta r}}{a_H}.
$$
Characterization of the 2D Channel

**Figure 3.24**: Scheme for a combined simulation of pit-size noise (via a table) and pit-position noise (via an analytical correction factor).

Using the scalar diffraction model the replay waveform is simulated by randomly choosing pit-position variations according to a normal distribution. The resulting standard deviations per cluster for various pit-sizes are shown in Fig. 3.25. Also here it appears that this plot is almost linear with $q_{SR}$ for not too large values of this parameter.

**Proximity Effects** The so-called proximity effect reveals itself as a pit-size variation depending on the number (and position) of the neighbouring pits. The origin of the so-called proximity effect in case of EBR mastering is the back (and forward) scattering of electrons in the resist during mastering as briefly discussed in Section 2.4. These scattered electrons generate a background illumination which will increase the pit-size compared to the nominal situation without the background illumination. For LBR mastering the Airy rings of the optical spot, although quite different in nature, may cause a similar effect. From a signal processing point of view the proximity effect causes data-dependent noise already at the level of the write-channel. The influence of the effect on signal levels can in first order be simulated by enlarging the pit-hole diameter in the simulation depending on the number of nearest neighbour pits. The amount of enlargement is varied in order to be able to fit the results to the experimental observations. An example of such a fit is shown in Fig. 3.26. It can be observed that in this case a 2 to 3% enlargement of the pit-hole diameter for each nearest neighbour pit seems to result in a good fit of the experimental data. In general the roll-off in signal level as function of cluster number will be more steep, especially in the second branch where a center pit is present. The fact however, that
3.6 Fitting Experimental Results

In this section experimental results are compared with the theoretical characteristics of the 2D channel as derived in this chapter. First the actually detected signal values are compared with the signal level plots derived from the scalar diffraction model. This is followed by fitting of measured noise spectra and the comparison of measured signal level variances to values derived from the pit-size and pit-position noise models.

3.6.1 Diffraction at the 2D Lattice

In Fig. 3.8 the diffraction of light at the disc surface was shown for a simple 1D evolution of the data. In practice always 2D diffraction occurs both in tangential and radial direction. For the hexagonal lattice diffraction with 6-fold symmetry takes place. An image of the pupil taken with a CCD camera clearly shows this pattern with...
6-fold symmetry, see Fig. 3.27(a). In Fig. 3.27(b) a photograph of a disc is shown. The reflection of the disc also clearly reveals the 6-fold diffraction.

The exact diffraction geometry in the pupil provides us with information about the exact pit-geometry on the disc. This might be used in the detection process by segmenting the photo-detector in 6 equal segments [44]. For example, when integrating the signal over the total pupil one cannot distinguish patterns that transform into each other by rotation only based on a single sample. By segmenting the photo-detector the information about rotational equivalent patterns becomes available. An initial study revealed that about 2 dB could be gained (in case of AWGN noise). A further study would be needed (e.g. to what extend partitioning is allowed in view of the reduced signal amplitude per segment) to really assess the benefits of this approach. This lies outside the scope of this thesis, however.

### 3.6.2 Fitting the Channel Model

The scalar diffraction model can be used to fit to experimental data to obtain some qualitative insights into the quality of the mastering process. This is done for two LBR-mastered discs (known as disc-10 and disc-12) and for two EBR-mastered discs
3.6 Fitting Experimental Results

(a) Pupil Image for the case of a disc with all bits equal to ‘1’, that is, each bit-cell has a pit. Density is very low leading to a large overlap of the orders. Also clear interference fringes due to defocus can be observed.

(b) Disc Reflection

Figure 3.27: Image of the pupil and photograph of a TwoDOS disc. Both clearly show the 6-fold diffraction at the hexagonal lattice.

(known as E266 and E268) with densities of 1.4x and 2.0x BD respectively. The result for disc-10 is shown in Fig. 3.28. The left part shows an AFM picture of the replicated disc showing that the pits are much too large. This is clearly reflected in the measured average levels as shown in the right part of the same figure. The levels were fitted with the scalar diffraction model. To achieve a good fit the phase-depth of the pits required quite some adjustment to $\Phi = 120^\circ$. This is thought to be caused by the pit-walls which do not have infinite steepness resulting in a lower effective depth of the pit. A similar procedure was followed for disc-12 (See Fig. 3.29). This disc was mastered using a special technique where, before exposure in the LBR, the resist layer was pre-developed for a fixed amount of time. This results in smaller pits, which can be seen in the AFM-scan and the level-plot that is much more linear.

Subsequently two EBR discs were measured. The fitted level plots are shown in Fig. 3.30. Again an adjustment of the phase to $\Phi = 160^\circ$ is necessary in both cases. Apparently the pit-walls are much steeper for the EBR discs than for the LBR discs. Even after this adjustment the data of EBR disc E268 does not fit too well, especially in the second branch. This is attributed to the proximity effect as discussed previously in this chapter. Based on these experimental results we can state that the scalar diffraction model gives us very good qualitative insight in the read-out process.
Characterization of the 2D Channel

(a) AFM image of LBR ‘disc-10’. It can be clearly observed that the pits are much too large.

(b) Experimental and calculated level-plot for LBR ‘disc-10’: NA=0.57; \(\lambda=405\) nm; \(\Phi=120^\circ\); \(a_H=248\) nm; \(b=248\) nm.

Figure 3.28: Experimental results for LBR-mastered ‘disc-10’.

Fitting of the level-plots gives us a basic tool to steer the mastering process.

(a) AFM image of LBR ‘disc-12’. The pits are clearly smaller due to the pre-development step.

(b) Experimental and calculated level-plot for LBR ‘disc-12’: NA=0.42; \(\lambda=405\) nm; \(\Phi=180^\circ\); \(a_H=248\) nm; \(b=100\) nm.

Figure 3.29: Experimental results for LBR-mastered ‘disc-12’.
3.6 Fitting Experimental Results

(a) Experimental (dashed, squares) and calculated (solid, circles) level-plot for EBR disc E266: NA=0.85; λ=405 nm; Φ = 160°; \( a_H = 165 \) nm; \( b = 120 \) nm.

(b) Experimental (dashed, squares) and calculated (solid, circles) level-plot for EBR disc E268: NA=0.85; λ=405 nm; Φ = 180°; \( a_H = 138 \) nm; \( b = 100 \) nm.

Figure 3.30: Experimental results for EBR-mastered discs E266 and E268. The discrepancy in the second branch of the right graph is attributed to the proximity effect (see Section 3.5).

3.6.3 Noise Measurements

Spectral Measurements The above presented noise model is used to fit measurements that are carried out on the experimental 2D optical player using a spectrum analyzer (Agilent 4395A) with a resolution bandwidth (RBW) of 30 kHz and a video bandwidth (VBW) of 1 kHz. First the noise spectra of the photo-detector IC (PDIC) were measured for various levels of illumination with a bright, shot noise limited, white-light source (to eliminate RIN from the measurement). The results are shown in Fig. 3.31. All measured voltages \( U_m \) are scaled in dBm, which is defined as:

\[
1\text{dBm} \equiv 20\log_{10} \left( \frac{U_m}{223.6\text{mV}} \right) \tag{3.54}
\]

i.e. 223.6 mV corresponds to 1 mW dissipation in a resistor of 50 Ω.

When the illumination level is increased an increase in current spectral density is observed over the complete frequency range. The high frequency noise level is determined by voltage noise, which is generally independent of illumination level. This is not observed in our case due to the design of the PDIC, where the signal current is also used as the bias current for the circuit (see Section 5.2). For even higher levels of illumination shot-noise starts to play a role and its ‘white’ contribution increases the total noise level especially in the low frequency range. The plotted curves can be
fitted using Eq. 3.43 (note that due to the design of the PDIC it is difficult with this measurement to distinguish all the independent noise sources and that the lumped model with $I_n$ and $V_n$ is used as a basis).

Subsequently, measurements were done on a replicated disc (E266), at a disc rotation speed of 5 Hz at a radius of 42.4 mm. The RBW is again 30 kHz. The results are shown in Fig. 3.32. The rotating-mirror-spectrum shows an increase in noise especially in the low frequency range. This is due to small defects on the mirror surface. These are picked-up by the scanning spot and filtered by the frequency transfer function of the optical channel. A very good fit is obtained for the following parameters: $G=330.000 \text{ V/A}$; $V_{DC}=0.438 \text{ V}$; $C_t=6 \text{ pF}$; $\nu_a=4.2 \text{ nV/} \sqrt{\text{Hz}}$; $i_a=1.6 \text{ pA/} \sqrt{\text{Hz}}$; $RIN=-120 \text{ dB/} \sqrt{\text{Hz}}$; MediaNoise=-22 dB (with respect to the data; on the alignment pattern).

Finally, some measurements were done on a silicon direct disc (E491) in exactly the same setup. With 'direct disc' we mean that the master itself is provided with a transparent cover layer and read out directly. No replication step is needed. This is done to avoid problems with ‘pit-moge’ as discussed in Section 7.1.3. The fitted noise spectra are shown in Fig. 3.33. The corresponding parameters are: $G=330.000 \text{ V/A}$; $V_{DC}=0.214 \text{ V}$; $C_t=6 \text{ pF}$; $\nu_a=4.2 \text{ nV/} \sqrt{\text{Hz}}$; $i_a=1.6 \text{ pA/} \sqrt{\text{Hz}}$; $RIN=-120 \text{ dB/} \sqrt{\text{Hz}}$; MediaNoise=-20 dB (with respect to the data; on the alignment pattern as discussed in Section 2.4). The fitted parameters can be used to derive SNR values according
3.6 Fitting Experimental Results

Figure 3.32: Noise spectra at the output of photo-detector when measuring a replica disc from stamper E266.

to Eq. 3.47. The results are listed in Table 3.2 for two different bandwidths (BW). One can observe that for disc E266 the limiting factor is media noise in both cases. For disc E491 the electronics noise is the dominant noise source. This is due to the fact that the silicon disc has a much lower reflection coefficient than disc E266 (with an aluminum mirror) and, more importantly, a much smaller modulation. Another observation is that for increasing bandwidth of the system (i.e. higher read-out speeds) the medianoise remains constant, while SNR due to voltage noise drops with 9 dB/octave and SNR due to all other noise sources drops with 3dB/octave.

Signal Level Measurements For disc 10, 12, E266 and E268 the standard deviations of the signal levels were determined for each of the 26 cluster types of Table 3.1. Some results for discs 12 and E266 are shown in Fig. 3.34. Note that in the outer rows some clusters do not occur. For these clusters the standard deviation is set to zero. Note also that only a 1-shell cluster is taken into account to characterize the signal levels. Therefore, higher-shell ISI may have a large contribution to the standard deviation per cluster and may obscure media noise due to e.g. variations in the illumination level during the multiple-pass mastering (per-shot power variation in
Figure 3.33: Noise spectra at the output of photo-detector when measuring a silicon direct disc (E491).

Table 3.2: Fitting results of the measurements on replica disc E266 and silicon direct disc E491 for different bandwidths (8 and 16 MHz).

<table>
<thead>
<tr>
<th>disc/parameter</th>
<th>E491 [dB] (8 MHz)</th>
<th>E491 [dB] (16 MHz)</th>
<th>E266 [dB] (8 MHz)</th>
<th>E266 [dB] (16 MHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>SNR (Voltage noise)</td>
<td>31.5</td>
<td>22.5</td>
<td>46.7</td>
<td>37.7</td>
</tr>
<tr>
<td>SNR (Current noise)</td>
<td>20.0</td>
<td>17.0</td>
<td>35.2</td>
<td>32.2</td>
</tr>
<tr>
<td>SNR (Shot noise)</td>
<td>30.9</td>
<td>27.9</td>
<td>43.0</td>
<td>40.0</td>
</tr>
<tr>
<td>SNR (RIN)</td>
<td>27.8</td>
<td>24.8</td>
<td>36.8</td>
<td>33.8</td>
</tr>
<tr>
<td>SNR (Media noise)</td>
<td>20.0</td>
<td>20.0</td>
<td>22.0</td>
<td>22.0</td>
</tr>
<tr>
<td>SNR (Total)</td>
<td>16.3</td>
<td>13.9</td>
<td>21.6</td>
<td>21.2</td>
</tr>
</tbody>
</table>

LBR or variations in per-shot electron-count for the EBR). For this reason a second set of standard deviation values is calculated by collecting signal values from exactly the same position within a frame over a large number of frames. These are shown in Fig. 3.35. Here the variations due to ISI are eliminated and standard deviations are much smaller. However, also the signal level variations that are caused by the
proximity effect are eliminated. Nevertheless, it is clear that the standard deviation as function of cluster number for disc E266 shows a different trend than the one for disc 12. Similar trends are observed for other EBR and LBR measurements. When comparing these plots with Fig. 3.25 and Fig. 3.23 one can conclude that generally the LBR discs are more hampered by media noise due to pit-size variations and that the EBR discs seem to suffer more from pit-position variations. The pit-position variation in EBR discs is attributed to the stability of deflection during the beam blanking procedure, see Section 2.4. Recall indeed that Fig. 3.21 is a worst-case situation. A more typical SEM photograph is shown in Fig. 3.36.

![Figures](a) Disc 12 (b) Disc E266

**Figure 3.34:** Standard deviation the Signal Level as function of the cluster number for all 7 rows in the broad spiral (cluster numbering according to Table 3.1). Standard deviation is taken over all clusters with same cluster number in one row.

### 3.7 Conclusions

In this chapter the characterization of the 2D channel is discussed. A hexagonal lattice is proposed for the ordering of the bits on the medium. This lattice offers a 15% higher packing than the square lattice. The reciprocal version of this hexagonal lattice is again a hexagonal lattice. At a density of two times the density of Blu-ray Disc, which is the goal of the TwoDOS project, the cut-off of the modulation transfer function forms an inscribed circle in the hexagonal fundamental domain. This means that the channel has a small negative excess bandwidth.

A scalar diffraction model is conventionally employed to simulate the read-out of an optical disc. Such a model is not suitable for signal processing purposes. The
Figure 3.35: Standard deviation of the Signal Level as function of the cluster number for all 7 rows in the broad spiral (cluster numbering according to Table 3.1). Standard deviation is taken over all samples at same position in a frame.

Figure 3.36: SEM photograph of typical EBR master (E268).

number of fast Fourier transforms (FFTs) needed to calculate all possible signal levels within an area with 37 bit positions (a central bit and 3 ‘shells’ of neighbours) is equal to $2^{37}$. A substantial reduction in number of FFTs is achieved when the Fourier transform to model the back propagation of the light towards the objective lens is
interchanged with the summation of the wavefront contributions originating from the different bits. It results in a workable situation, where 703 bi-linear components and 37 linear components must be calculated only once (using an FFT) for the same 37 bit positions. The results of this scalar diffraction model show a very good fit with experimental results. Signal level calculations for a particular bit show a gradual roll-off of the signal level as function of the number of nearest-neighbour pit-bits. The signal levels of pit-bits overlap with those of bit with value zero, i.e. no pit. This means that the ‘eye-pattern’ is closed, and that straightforward detection with a slicer with suitable threshold is not possible anymore. The results also show considerable non-linearity (or more accurately bi-linearity) in the optical read-out channel, especially for large pit-sizes (so-called ‘signal-folding’). The use of smaller pits makes the channel more linear but reduces modulation. An optimum is found for a pit-diameter such that the pit-hole covers about half the area of the available hexagonal cell.

Noise characterization of the system is done by reviewing the following noise sources:

- Shot noise.
- Electronics noise.
- Quantization noise.
- Media noise.

A combination of these noise sources is fitted successfully to the experimental results. This is done by measuring the noise spectrum and by fitting a model to these measured spectra. The media noise component is assessed using the scalar diffraction model and distinguishing three main sources:

- Pit-size noise.
- Pit-position noise.
- Proximity effect.

By comparing signal level plots it can be concluded that generally the LBR discs are more hampered by media noise due to pit-size variations, while the EBR discs seem to suffer more from pit-position variations.
Chapter 4

2D Modulation Coding and Test-Format

The characterization of the 2D optical channel in Chapter 3 made clear that the channel starts to exhibit several undesired properties when one wants to increase the storage density on the medium [45]. For example, the optical channel has a low-pass characteristic with a hard cut-off beyond which it is not able to transmit frequency components in the spectrum of the data. Therefore, especially data that has a lot of high frequency components at the input of the storage channel may be prone to mis-detection at the receiver side of the system. In particular it is the task of channel modulation coding to transform the input data sequence into a sequence of so-called channel bits that have certain desirable properties. A typical desirable property is to achieve high reliability of transmission across the optical storage channel while simultaneously making efficient use of the channel’s capacity. Another reason for modulation coding can be some specific requirements of the receiver or read-out system. Clock recovery is a good example. The receiver needs to know at which instants it needs to sample the signal in order to make a decision whether the stored information was a zero or one. Generally, in optical recording no clock information is stored explicitly on the disc for example in the form of timing marks. Instead, using a suitably constrained code, guaranteed clock information can be provided in the data stream itself such that the receiver is able to regenerate a synchronous clock at the baud-rate at the receiver side of the system. This property of the channel bitstream is generally referred to as the ‘self-clocking’ property. Additionally, conventional clock recovery systems are based on zero-crossing detection that require an ‘open eye-diagram’ of the read-out signal. The bit clock is adjusted at each occurrence of a transition in the signal waveform. This chapter will start with possible coding schemes to achieve these goals (see Section 4.1). It will proceed with a way of coding to eliminate those binary sequences from the channel input data stream that have a high probability of detection error. This analysis can be done on a theoretical basis by for example looking to worst case sequences in terms of the matched filter bound (MFB; see Section 4.2) or on a practical basis by ranking the error patterns in dependency of their occurrence during detection of data in a typical experimental system. A practical example of a channel code that removes certain error patterns (including those that
have a high loss with respect to the MFB) is given in Section 4.3. Finally, the chapter concludes with the definition of a suitable test format (Section 4.4) containing the coded data.

4.1 Low-Pass Coding

In 1D coding the eye pattern was an important parameter during the design of a modulation code [7]. The eye height can be considered as a measure of the robustness of the signal against noise, while the eye width can be used as an indication of the operating margins with respect to phase or transition jitter. The opening of the eye depends on the amount of inter-symbol interference (ISI). For this reason 2D-constraints will be proposed that relate to the number of nearest-neighbouring bits of the same polarity as the center bit in a 1-shell cluster. The constraints have the objective to open the closed eye-pattern (See Fig. 3.18), i.e. to reduce the influence of ISI. Another reason for low-pass coding is at the side of the write channel and will be discussed in Section 4.1.2.

4.1.1 Prior Art

**Checkerboard Codes** Prior art on coding for 2D systems can be found in [14]. Here the capacity of checkerboard codes, which are defined on a square lattice, is studied. Various 2D constraints were considered as shown in Fig. 4.1. These constraints are defined as ‘a two-dimensional arrangement of zeros that must surround every one in a two-dimensional binary code’.

The capacity of such a 2D code is defined similar to the 1D case:

$$C = \lim_{m,n \to \infty} \frac{\log_2 N_c(n,m)}{nm}$$

(4.1)

with $N_c(n,m)$ the number of 2D bit-patterns with size $n \times m$ that satisfy some constraint $c$. Because it appears to be difficult to calculate $N_c(n,m)$ (e.g. based on a recursive relation), it is proposed to find a recursive relationship for a connection matrix $A_n$. This connection matrix defines how the $n \times m$ pattern can be extended to an $n \times (m+1)$ pattern by adding a column vector of size $n \times 1$. The entries of the connection matrix $a_{i,j}$ equal one if such a column of size $n \times 1$ denoted as $j$ can be placed right next to a column with same size denoted as $i$. The largest eigenvalue of the connection matrix then provides a way to calculate the capacity $C_n$ of the code.

---

1 Another way of dealing with the worst case error patterns is scrambling. In [46] it is shown that it is possible to do guaranteed scrambling in the sense that the predicted bit error rate of a particular input sequence is not larger than that of random data. To achieve this, a scrambling code word is selected that is summed modulo-2 with the input sequence such that the resulting sequence does not have a more than average amount of worst case patterns. Scrambling will not be further discussed in this thesis.
4.1 Low-Pass Coding

with columns of height \( n \). Final capacity is reached by calculating the limit of \( C_n \) for \( n \) approaching infinity. It appears that convenient recursive relations for the connection matrix can be found for all constraints given in Fig. 4.1.

**Nearest Neighbour Constraints** Another way of low-pass coding is to define constraints on the nearest neighbours. This is done in [47] where several constraints are defined on a 2D pattern of limited size. The constraints are defined on an orthogonal grid. It is proposed for example that the four nearest neighbours to a zero cannot all be ones and similarly that the four nearest neighbours of a one cannot all be zeros. The codes are applied in strips where the total two-dimensional array is seen as a collection of several long horizontal strips, all of the same height. Special attention is needed when stacking these strips in order not to violate the 2D constraints on the border of two strips. The latter constraints can be guaranteed by forbidding additional patterns at the border of strips that are a subset of the original forbidden patterns. The codes can be implemented by a simple finite state machine and decoded by a sliding block decoder.

4.1.2 Definition of Constraints on the 2D Hexagonal Bit-Lattice

For the design of 2D low-pass codes in this thesis an approach similar to [47] is followed, which means that constraints are defined on the nearest neighbours depending on the value of the central bit. However, we apply this type of constraints on the 2D
hexagonal bit lattice where each lattice site has 6 nearest neighbours (instead of 4 for the square lattice).

**Bulk Constraints**  On a hexagonal lattice, the constraints can be described by two parameters: The first one is the minimum number of nearest neighbours \((N_{nn})\) of the same type as the bit located on the center lattice site:

\[
|6x_0 + \sum_{i=1}^6 x_i| \geq 2N_{nn}
\]  

(4.2)

with \(x_i \in \{-1, 1\}\) the bipolar channel bits at the central site \((i=0)\) and the 6 neighbouring sites \((i=1, \ldots, 6)\). The second type of constraint specifies the minimum number of azimuthically-contiguous nearest neighbours \((N_{ac}\), with \(1 \leq N_{ac} \leq N_{nn}\)):

\[
\exists J \in \{0, 1, \ldots, 5\} : |x_0 + \sum_{i=1}^{N_{ac}} x_{i+J}| = N_{ac} + 1.
\]  

(4.3)

The parameter \(N_{nn}\) reduces the effect of ISI. This can be derived from a simple linear model as shown in Fig. 4.2 where the 2D impulse response function (IRF) has values \(c_0\) at the central site, and \(c_1\) at the nearest-neighbour sites. The largest deviation of the waveform from its central response value \(c_0\) is when there are just \(N_{nn}\) nearest neighbours of the same type. In that case the signal level is equal to:

\[
c_0 - (6 - 2N_{nn})c_1.
\]  

(4.4)

The constraint in terms of azimuthically contiguous nearest neighbours could offer an additional advantage by realizing a minimum mark size that relieves the burden at the side of the write channel. In that case, however, three pits must be replaced by one (larger) pit in the center of the triangle formed by the three pits. This central pit must give the same signal contribution to all surrounding samples compared to the three pit case. To achieve this \(N_{ac}\) must be at least 2. Additional coding constraints might be necessary in order to avoid excessive signal folding due to the formation of large contiguous pit-areas.

![Figure 4.2: Simple linear channel model for bits on a 2D hexagonal lattice.](image-url)
4.1 Low-Pass Coding

Figure 4.3: Clusters of Bit-Sites on the Hexagonal Lattice. (a): Bulk Cluster; (b): Bottom Boundary Cluster; (c): Top Boundary Cluster.

Boundary Constraints When the code is applied on one-dimensional evolving strips instead of on an infinite 2D lattice one must also take into account the boundary constraints (a further explanation of strip-based coding is given in Section 4.3.3). The simplest way to guarantee the coding constraints is to satisfy the constraints within the strip irrespective of the bits that are present in a neighbouring strip. This enables stacking of the strips on top of each other without violation of the 2D constraints. An incomplete cluster at the boundary consists of 5 bits, compared to 7 bits for the bulk cluster i.e. two nearest-neighbour sites are then omitted from the constraint (see Fig. 4.3). The first constraint is then formulated as:

\[ |6x_0 + \sum_{i=1}^{4} x_i| \geq 2(N_{nn} + 1) \]  

(4.5)

while the second one reads:

\[ \exists J \in \{0, 1, ..., 4 - N_{ac}\} : |x_0 + \sum_{i=1}^{N_{ac}} x_{i + J}| = N_{ac} + 1. \]  

(4.6)

These constraints can be used to open the eye diagram of the 2D replay signal. For example if \( N_{nn}=1 \) then the two middle HF signal levels disappear from the level-plot as introduced in Fig. 3.11. The rate-loss induced by a code that implements this constraint must be accounted for by scaling the lattice parameter \( a_H \) to make a fair comparison of the eye-opening. An efficient coding scheme will approach the capacity and therefore we will estimate the capacity of the 2D code. This is done by deriving the underlying finite state machine (FSM) that drives the generation of the 2D sequences. Since all constraints that we currently present relate to nearest neighbours only, it is sufficient to consider states based upon two successive columns on the hexagonal grid, and covering all \( N_r \) rows of the strip. Note that the bits are ordered in a zig-zag pattern, so we actually consider zig-zag columns. The number of such states is then simply \( 2^{2N_r} \). By transition from a given state (the departure state) towards the next state (the arrival state), a complete column of channel bits is
emitted. By definition, the last column of the first state is identical to the first column of the successor state. For the case $N_r = 7$, Fig. 4.4 shows a typical state (one out of 8192), and one of its possible successor states.

**Figure 4.4:** Possible Transition from State $i$ to State $j$.

Crucial in the derivation of the capacity and the design of 2D channel codes is the connection matrix $D$ [7]. This is a square matrix with a maximum size of $2^{2N_r} \times 2^{2N_r}$ (note that it may be smaller if some departure states already violate the constraint). It has ”1” as matrix elements $D_{ij}$ in case one state $i$ can have state $j$ as its successor. All other matrix elements are zero. Transitions from state $i$ to state $j$ are possible under a number of conditions:

- Condition 1: the last column of state $i$ is identical to the first column of state $j$;
- Condition 2: state-transitions may not cause a constraint violation for the bulk clusters (bulk constraint). This constraint is the only one to be considered for the derivation of an upper bound of the capacity;
- Condition 3: concatenation of strips may not cause a constraint violation at the boundaries of the strips. Therefore, we apply the boundary constraint. This constraint is needed for computation of the lower capacity bound.

The capacity values are calculated when applying the bulk constraint only (high capacity bound) and when applying both the bulk and the boundary constraint (low capacity bound). This is done by calculating the largest eigenvalue of the connection matrix $D$. The capacity values serve as an upper and lower boundary respectively for the actual capacity for an infinite 2D area. It appears that for increasing number of rows the upper and lower boundary converge to this capacity for an infinite 2D area. Two examples are shown in Fig. 4.6 for $N_{nn} = 1$ (left) and for $N_{nn} = 2$ (right).
4.2 Patterns with a High Loss with Respect to the Matched Filter Bound

Note that for $N_{nn} = 1$ the capacity in case of a single row is equal to the 1D, $d = 1$ RLL constraint. Similar calculations can be done for the constraints related to azimuthically-contiguous bits.

![Graph](image)

(a) Capacity for $N_{nn} = 1$

(b) Capacity for $N_{nn} = 2$

*Figure 4.5: Upper and lower bound capacities as function of the number of rows in a broad spiral for different values for the nearest-neighbour constraint.*

In Fig. 4.6 the procedure is followed for a fixed number of rows ($N_r = 7$) and for increasing values of $N_{nn}$. For values of $N_{nn}$ larger than 3 it is not possible anymore to tile the complete 2D area with clusters that satisfy the bulk constraint. Only in the boundary rows some coding entropy is left. For the calculation of the HF signal levels the lower bound on the capacity is chosen, i.e. boundary conditions included. It can be observed that for small values of $N_{nn}$ the benefit in terms of improvement in the eye-opening is significant, but this diminishes when the penalty of code rate loss becomes higher. Moreover, for high densities the scaled pits become very small leading to smaller modulation and moreover to pit-moge problems (see Section 7.1.3). Therefore, for 2D-storage this technique only seems viable for low to modest densities and for mild coding constraints.

4.2 Patterns with a High Loss with Respect to the Matched Filter Bound

*Loss With Respect to Matched Filter Bound* Another criterion to remove patterns from the data stream is to remove those patterns that support error sequences which have a high loss with respect to the matched filter bound (MFB). When a single
bit (or in the case of the linear system equivalently a single bit error) is transmitted there is no ISI and one can expect the optimum situation for detection. In maximum-likelihood detection where the total euclidian distance between waveforms is calculated the detection distance for such a single bit error is equal to $d_{\text{min}}^2 = \sum g_k^2$ (in case of a linear channel model $g_k$; A more detailed analysis can be found in the paragraph on Sequenced Amplitude Margin in Section 6.5). This lower limit is called the MFB for single bit-errors. For error patterns that consist of more than one bit, ISI must be taken into account and a smaller value for $d_{\text{min}}$ is possible due to partial cancellation of the impulse responses of the neighbouring bits in the error pattern. The smaller value for $d_{\text{min}}$ may lead to higher error rates. For this reason the ratio between the $d_{\text{min}}^2$ values for single and multiple bit errors is called the ‘loss with respect to the matched filter bound’ if it is smaller than 1, that is, the MFB is then simply too optimistic for the performance of the bit-detector. With the use of a proper code one can eliminate either those bit-patterns that allow this error to occur (i.e. the erroneous sequence must also be a valid data sequence for the modulation code that is used) and/or those bit-patterns that may result from the particular error sequence.

**Worst Case Pattern for the linear 2D Channel** Let us assume a channel with linear coefficients as shown in Fig. 4.7. These coefficients correspond with a linear approximation of the channel with hexagonal lattice parameter $a_H=138$ nm. The sum of the squared coefficients is 2.027. Now a pattern according to Fig. 4.8a is transmitted across the channel resulting in the response of Fig. 4.8b. The sum of the squared coefficients of this response results in 1.147, which means a loss with respect
4.2 Patterns with a High Loss with Respect to the Matched Filter Bound

to the matched filter bound of:

\[ L_{\text{MFB}} = 10 \log \left( \frac{1.147}{2.027} \right) = -2.47 \text{dB} \]  

(4.7)

Figure 4.7: Linear coefficients corresponding to \( a_H=138 \text{ nm} \).

A brute-force search shows that the 'closed-ring' of alternating '+1' and '-1' symbols (referred to as 'Nyquist-ring') is the worst case error pattern for the channel of Fig. 4.7. The spectrum of such a ring is shown in Fig. 4.9. It also makes clear that in general Nyquist rings with different sizes and shapes have a high loss with respect to MFB. Also Nyquist rings at the boundary of the broad spiral as used in the TwoDOS format (see Fig. 4.10) have a high loss due to energy leakage to the guard band (assuming that no energy outside the broad-spiral is picked up, which could, of course, be realized via additional read-out spots on the guard bands). To eliminate these patterns from the input data sequence it is sufficient to simply eliminate the patterns resulting from inverting the error patterns. For the worst case one this appears to be automatically achieved for the high-rate modulation code that will be described in Section 4.3.

Worst Case Patterns for the non-linear 2D Channel  In most cases the 2D channel exhibits non-negligible non-linearities. Therefore, it is not allowed to simply
transmit error patterns through the channel and select the one with the lowest energy. Instead, again a brute-force search is carried out, where all bit-patterns within a predefined 2D-area are transmitted through the channel, and where Euclidian distances are calculated between the resulting replay signals. The lattice constant is again 138 nm and pit-diameter is 102 nm. A minimum Euclidian distance is found

Figure 4.9: Spectrum of the first-shell Nyquist-ring.

Figure 4.10: Nyquist-ring patterns.
between the signal sequences of the ‘all-1’ cluster and the cluster with a center zero surrounded by all ‘1’s. The ‘all-1’ cluster (in an ‘all-1’ environment) results in signal values 0.107 (using the 4-parameter model with $b_0 = 1; n_1 = 6; p_{1,1} = 6$). Then a single zero is transmitted surrounded by all ones. For the central sample this means $b_0 = 0; n_1 = 6; p_{1,1} = 6$ resulting in the value 0.294 and for the next nearest neighbour samples $b_0 = 1; n_1 = 5; p_{1,1} = 4$ giving the value 0.150. The distance between these two data patterns is 0.214. This value is compared to a bound, which is a bit similar to the matched filter bound and is calculated as follows: First the ‘all-0’ sequence is transmitted leading to signal value 1. Then a single ‘1’ is transmitted. The Euclidian distance appears to be equal to: 0.599. The equivalent ‘loss’ of the above pattern with respect to this ‘matched filter bound’ is 8.94 dB. Although this value is very large compared to the worst case pattern derived from the linear channel, it is difficult to compare the two, because the definition of MFB is quite different in both cases. What matters in practice is which error patterns are occurring when increasing density. In Section 7.2.2 it will become clear that Nyquist patterns account for most of the errors in a 50 GB disc.

4.3 High Rate Coding for Elimination of Critical Patterns

In this section, which is based on [48], critical patterns are considered of which the spectral content is highly attenuated by the channel or does not pass the channel at all. Several patterns can be found that have spectral components beyond the channel’s cut-off frequency, but within the hexagonal fundamental domain (Section 4.3.2). This is possible because, at the targeted capacity of 50 GB on a 12 cm disc, the channel cut-off forms an inscribed circle within the fundamental domain as discussed in Section 3.4. In Section 4.3.3 strip-based 2D modulation coding is proposed, where strips coded with a first modulation code are stacked one upon the other in the radial direction and are glued together with merging bit-rows that are 1D coded using a second modulation code. In this way the total bit-stream satisfies the 2D-constraints. The 2D code for the strips is presented in Section 4.3.4. The 1D code is discussed in Section 4.3.5. Finally, the relation with the error correction code (ECC) and the basic choices for a possible format are discussed in Section 4.3.6.

4.3.1 1D-Case with the MTR-constraint

As already discussed in the introduction of this chapter, it is well known that in traditional 1D storage, the Nyquist sequence of alternating bits $\cdots + - + - + - \cdots$ is badly transmitted by the channel (because of its low-pass characteristics). Based on this knowledge, a class of so-called maximum transition run (MTR) runlength-limited (RLL) codes has been derived in [49]. The MTR-constraint specifies the
maximum number of consecutive ”1”-bits in the \((d, k)\) bitstream of ‘differential’ bits (where a ”1” indicates the start of a new run). Note that the channel bits that are written to the medium (which we refer to as RLL bits or polar bits) are obtained from the \((d, k)\) bitstream after feeding it to a so-called 1T-precoder, which is essentially a modulo-2 integrator. Thus, in the RLL bitstream, the MTR constraint limits the number of consecutive 1T runs. Originally, in [49], the maximum number of consecutive ”1”-bits in the \((d, k)\) bitstream was limited to 2, which implies that only isolated 1T runs are present in the RLL bitstream. A pure Nyquist sequence of alternating bits obviously corresponds with the highest spatial frequency for a \(d=0\) encoded system. Alternating minimum run lengths with a period equal to \(2(d+1)\) correspond to the highest spatial frequency in the case of a general \(d\)-constraint. Similar concepts apply to the two-dimensional case.

### 4.3.2 2D-Case on Hexagonal Bit-Lattice

Recall that the boundary point of the 1D Nyquist interval in frequency space (i.e. the Nyquist frequency) represents the most critical frequency in 1D signal processing. Similarly, the corner points of the 2D hexagonal fundamental domain represent the most critical two-dimensional periodic structures (also called super-structures here, in analogy to the usual terminology in crystallography) on the 2D hexagonal bit-lattice that are subject to the worst transfer by the MTF of the channel. This is due to its low-pass characteristics as shown in Fig. 3.6. The super-structure can be described in terms of its generating matrix in direct space given by (e.g. for the diamond-shaped pattern at the top of Fig. 4.11a):

\[
L_{\text{diam.}} = \frac{a_H}{2} \begin{bmatrix} 3 & 3 \\ -\sqrt{3} & \sqrt{3} \end{bmatrix}.
\]  
(4.8)

The corresponding generating matrix (for the super-structure) in reciprocal space then becomes:

\[
L_{\text{diam.}}^* = \frac{1}{a_H \sqrt{3}} \begin{bmatrix} \sqrt{3} & \sqrt{3} \\ \frac{\sqrt{3}}{3} & 1 \end{bmatrix},
\]  
(4.9)

which can be seen in Fig. 3.2 to constitute the corner points of the fundamental domain in reciprocal space. It can be easily analyzed in Fig. 4.11a that this super-structure consists of a concatenation of two types of clusters: one cluster having a ‘+’-bit at its center, and surrounded by six ‘−’-bits as its nearest neighbors, the other cluster having a ‘−’-bit at its center with a Nyquist ring consisting of the six alternating bits ‘+−−−+−’ as its nearest neighbors. These bits can also have the opposite polarity. It is further obvious from Fig. 4.11a, that the periodic super-structure can
also be described in terms of two lattice planes (indicated with the dashed lines in the figure) making an angle of $60^\circ$.

Figure 4.11: Worst-case patterns of channel bits on the 2D hexagonal bit-lattice. In (a) three equivalent orientation variants are shown. The two types of bit-clusters that generate the 2D pattern are encircled: one has a ‘+’-bit at its center, the other a ‘−’-bit (of which the nearest neighbours could also have opposite polarity). At the bottom, the repetitive pattern of fish-bone-shaped bit-triplets is indicated with a repeat-period of three fish-bones. (b) shows a pattern of alternating lattice planes of ‘+’ and ‘−’. Intuitively, this could be mistakenly considered as the pattern that generates the 2D-equivalent of the Nyquist sequence. That is, pattern (a) is the genuine Nyquist equivalent on a 2D hexagonal bit-lattice.

Finding this pattern as the most critical pattern in terms of frequency transfer maybe somewhat counter-intuitive at first sight, because alternating lattice planes along one of the main hexagonal directions may seem to generate the 2D-equivalent
of the 1D Nyquist sequence. The pattern of alternating lattice planes along the \( u_1 \) direction (defined in Eq. 3.1) is shown in Fig. 4.11b. It can be described by the generating matrix:

\[
L_{\text{diag. planes}} = \frac{a_H}{2} \begin{bmatrix} 4 & -1 \\ 0 & \sqrt{3} \end{bmatrix}
\]  

(4.10)

with reciprocal counterpart:

\[
L^*_{\text{diag. planes}} = \frac{a_{H}^*}{4} \begin{bmatrix} \sqrt{3} & 0 \\ 1 & 4 \end{bmatrix}.
\]  

(4.11)

Figure 4.12: Reciprocal lattice vectors of the alternating planes along the main axes of the hexagonal lattice.

One can see that the diffraction vector of the alternating planes \( \frac{a_{H}^*}{4} \begin{bmatrix} \sqrt{3} \\ 1 \end{bmatrix} \), lies at the point of intersection of the boundary of the 2D fundamental domain and the inscribed circle at the cut-off of the MTF (see Fig. 4.12). In conclusion this means that although the alternating ‘+’ and ‘−’ planes along the main hexagonal directions as shown in Fig. 4.11b seem the 2D equivalent of the 1D Nyquist sequence, it is the 2D superstructure described in the beginning of this section that represents the real 2D equivalent of the 1D Nyquist sequence and the most critical pattern to transfer over the 2D optical channel. It is shown in Fig. 4.11a.

4.3.3 2D Modulation Coding for Meta-Spiral: Two Building Blocks

Similar to the case of the MTR-constraint [49] in modulation coding for 1D storage, we advocate the use of a 2D modulation code that forbids the occurrence of the
super-structure patterns on the 2D hexagonal bit-lattice that give rise to frequency components at the set of 2D Nyquist frequencies as defined above. In order to realize this 2D constraint, two types of building blocks are considered to construct a multi-track meta-spiral. A first type of building block consists of strips as used in strip-based 2D coding proposed in [14, 47]: The 2D area is divided into consecutive strips. A strip is aligned "horizontally", that is, along the tangential direction of the meta-spiral, and consists of a small number of bit-rows, for example three. Coding is carried out along this horizontal direction, and becomes essentially one-dimensional with M-ary symbols representing the bits along the columns. Strips are concatenated one upon the other in the radial direction. In the formalism of [14, 47], codewords do not cross the boundaries of a strip. In addition, a second type of building block is considered for the 2D modulation code: These are single bit-rows that are 1D encoded and are inserted between consecutive strips such that these rows serve as merging bit-rows effectively gluing consecutive strips together. The code constraints for both types of building blocks are described by a state-transition diagram (STD) where each branch between a given departure state and the corresponding arrival state has one of the possible channel symbols as its label. The STD defines the direction of evolution of the code. Note that in the meta-spiral, there is one single direction of evolution for both types of modulation codes (of 2D strips and 1D merging bit rows).

4.3.4 Modulation Coding for Building Block Type I: 2D-Encoded 3-Row Wide Strips

As a practical case strips consisting of 3 bit-rows are taken. Note however that the basic idea is generic for any number of rows larger than one.

3-row NRZ and NRZI Channel Symbols  For the case of 3-row strips, an 8-ary channel symbol comprises three NRZI bits (the bipolar bits as they are written on the disc), one bit per bit-row in the strip. The 3 NRZI bits in the symbols have a staggered stacking, so that these symbols may be reminiscent to fish-bones (see Fig. 4.11a at the bottom). The 2D code is designed such that it generates NRZ symbols. Just as in 1D modulation encoding, a basic ingredient of the 2D modulation encoding is a 1T-precoder, carried out in a row-wise manner along the tangential direction of the meta-spiral: this 1T-precoder transforms the NRZ bits (differential bits, or \((d,k)\)-bits of the runlength-limited (RLL) code in the traditional 1D RLL terminology) into NRZI bits. Since the 2D constraint does not depend on the overall polarity, 4 STD-states can be defined for a 3-row strip as listed in Table 4.1. Each of the transitions in the STD from one STD-state to another is labelled by an 8-ary NRZ symbol. Each of the NRZ bits of the bit-triplet corresponds to a given bit-row in the 3-row strip. An NRZ-bit equal to ‘1’ indicates a transition in the NRZI
Table 4.1: STD States for a 3-row strip \((b \in \{-1, 1\}; \bar{b} \equiv -b)\).

<table>
<thead>
<tr>
<th>STD-State (\sigma_1)</th>
<th>STD-State (\sigma_2)</th>
<th>STD-State (\sigma_3)</th>
<th>STD-State (\sigma_4)</th>
</tr>
</thead>
<tbody>
<tr>
<td>(b)</td>
<td>(b)</td>
<td>(b)</td>
<td>(b)</td>
</tr>
<tr>
<td>/</td>
<td>/</td>
<td>/</td>
<td>/</td>
</tr>
<tr>
<td>(b)</td>
<td>(\bar{b})</td>
<td>/</td>
<td>/</td>
</tr>
<tr>
<td>(\bar{b})</td>
<td>(b)</td>
<td>(\bar{b})</td>
<td>(b)</td>
</tr>
</tbody>
</table>

Table 4.2: 8-ary NRZ and NRZI symbols for 3-row strip

<table>
<thead>
<tr>
<th>(k)-th NRZI Symbol</th>
<th>(k)-th NRZ Symbol</th>
<th>((k + 1))-th NRZI Symbol</th>
</tr>
</thead>
<tbody>
<tr>
<td>(b_k^0)</td>
<td>(\rightarrow i \rightarrow)</td>
<td>(b_{k+1}^0 = b_k^0(-1)^i)</td>
</tr>
<tr>
<td>(b_k^1)</td>
<td>(\rightarrow j \rightarrow)</td>
<td>(b_{k+1}^1 = b_k^1(-1)^j)</td>
</tr>
<tr>
<td>(b_k^2)</td>
<td>(\rightarrow l \rightarrow)</td>
<td>(b_{k+1}^2 = b_k^2(-1)^l)</td>
</tr>
</tbody>
</table>

bitstream along the considered bit-row, a ‘0’-bit indicates the absence of such a transition. For a 3-row strip, this is illustrated in Table 4.2 (the superscript index denotes the bit-row in the strip, the subscript index denotes the symbol number along the strip, in the direction of evolution of the code).

Generation of Worst-Case Super-Structure Patterns The periodic super-structure patterns (in a 3-row strip) that need to be forbidden, can be described as a succession of a number of bit-triplets (or fish-bones), given by:

\[
\ldots \ b \ \bar{b} \ b \ b \ \bar{b} \ \ldots \\
\ldots \ / \ / \ / \ / \ / \ / \ / \ / \ / \\
\ldots \ \bar{b} \ b \ b \ \bar{b} \ b \ \bar{b} \ \ldots \\
\ldots \ \bar{b} \ b \ b \ \bar{b} \ b \ \bar{b} \ \ldots 
\]

The repeat period of this pattern comprises exactly three fish-bones. Note that the first state in the above pattern equals \(\sigma_3\). The corresponding NRZ-symbols are derived as
(with the NRZ symbols denoted $S$, with $S = i + 2j + 4l$ where $i$, $j$ and $l$ represent the NRZ-bits along each of the three bit-rows):

\[
\begin{pmatrix}
i \\
j \\
l
\end{pmatrix} = \begin{pmatrix}1 \\
1 \\
1
\end{pmatrix} \quad \rightarrow \sigma_3
\]

\[
\begin{pmatrix}
i \\
j \\
l
\end{pmatrix} = \begin{pmatrix}1 \\
0 \\
1
\end{pmatrix} \quad \rightarrow \sigma_3
\]  

\[
\begin{pmatrix}
i \\
j \\
l
\end{pmatrix} = \begin{pmatrix}0 \\
1 \\
0
\end{pmatrix} \quad \rightarrow \sigma_3
\]  

\[
(S = 7) \quad (S = 5) \quad (S = 2)
\]

The last state is again equal to the first state ($\sigma_3$), and the sequence can be repeated to form a longer worst-case pattern.

**Forbidden Sequences** The worst-case patterns can be eliminated by forbidding particular sequences of NRZ-symbols that would leave from certain STD-states. As an example, it is proposed to truncate the sequence as soon as two consecutive NRZI bit-triplets result that form part of the worst-case pattern. This implies that, for instance, for STD-state $\sigma_3$, after emission of NRZ-symbol $S = 7$, it is forbidden to emit as subsequent symbol $S = 5$. There are two additional cases that have to be eliminated: for STD-state $\sigma_3$, after emission of NRZ-symbol $S = 5$, one may not emit as subsequent symbol $S = 2$, and for STD-state $\sigma_1$, after emission of NRZ-symbol $S = 2$, it must be forbidden to emit as subsequent symbol $S = 7$.

**Connection Matrix and Capacity** The adaptation of the state transition diagram (STD) in order to consider the forbidden sequences is done as follows. Three additional states are generated. The first one, denoted $\sigma_5$, can only be reached when emitting symbol $S = 7$ from state $\sigma_3$; further, the fan-out of $\sigma_5$ is identical to that of $\sigma_3$ (the state that is reached in the original STD when emitting $S = 7$ from state $\sigma_3$), with the single exception that emission of symbol $S = 5$ is not allowed:

\[
\rightarrow \sigma_3 \rightarrow \sigma_5 \rightarrow 7 \quad \neq \quad 5
\]

The second additional STD-state, denoted $\sigma_6$, can only be reached when emitting symbol $S = 5$ from state $\sigma_3$; further, the fan-out of $\sigma_6$ is identical to that of $\sigma_1$ (the state that is reached in the original STD when emitting $S = 5$ from state $\sigma_3$). Now, the single exception is that emission of symbol $S = 2$ is not allowed from $\sigma_6$:

\[
\rightarrow \sigma_3 \rightarrow \sigma_6 \rightarrow 5 \neq \quad 2
\]

Finally, the third additional STD-state, denoted $\sigma_7$, can only be reached when emitting symbol $S = 2$ from state $\sigma_1$; further, the fan-out of $\sigma_7$ is identical to that of $\sigma_3$.
Table 4.3: State-Transition Diagram (STD) for 3-row strip.

<table>
<thead>
<tr>
<th>NRZ-Symbol</th>
<th>STD-state</th>
<th>(0)</th>
<th>(1)</th>
<th>(2)</th>
<th>(3)</th>
<th>(4)</th>
<th>(5)</th>
<th>(6)</th>
<th>(7)</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \sigma_1 )</td>
<td>( \sigma_1 )</td>
<td>( \sigma_2 )</td>
<td>( \sigma_7 )</td>
<td>( \sigma_4 )</td>
<td>( \sigma_4 )</td>
<td>( \sigma_3 )</td>
<td>( \sigma_2 )</td>
<td>( \sigma_1 )</td>
<td></td>
</tr>
<tr>
<td>( \sigma_2 )</td>
<td>( \sigma_2 )</td>
<td>( \sigma_1 )</td>
<td>( \sigma_4 )</td>
<td>( \sigma_3 )</td>
<td>( \sigma_3 )</td>
<td>( \sigma_4 )</td>
<td>( \sigma_1 )</td>
<td>( \sigma_2 )</td>
<td></td>
</tr>
<tr>
<td>( \sigma_3 )</td>
<td>( \sigma_3 )</td>
<td>( \sigma_4 )</td>
<td>( \sigma_1 )</td>
<td>( \sigma_2 )</td>
<td>( \sigma_2 )</td>
<td>( \sigma_6 )</td>
<td>( \sigma_4 )</td>
<td>( \sigma_5 )</td>
<td></td>
</tr>
<tr>
<td>( \sigma_4 )</td>
<td>( \sigma_4 )</td>
<td>( \sigma_3 )</td>
<td>( \sigma_2 )</td>
<td>( \sigma_1 )</td>
<td>( \sigma_2 )</td>
<td>( \sigma_3 )</td>
<td>( \sigma_4 )</td>
<td>( \sigma_4 )</td>
<td></td>
</tr>
<tr>
<td>( \sigma_5 )</td>
<td>( \sigma_3 )</td>
<td>( \sigma_4 )</td>
<td>( \sigma_1 )</td>
<td>( \sigma_2 )</td>
<td>( \sigma_2 )</td>
<td>( \sigma_4 )</td>
<td>( \sigma_4 )</td>
<td>( \sigma_5 )</td>
<td></td>
</tr>
<tr>
<td>( \sigma_6 )</td>
<td>( \sigma_1 )</td>
<td>( \sigma_2 )</td>
<td>( \sigma_4 )</td>
<td>( \sigma_4 )</td>
<td>( \sigma_4 )</td>
<td>( \sigma_3 )</td>
<td>( \sigma_2 )</td>
<td>( \sigma_1 )</td>
<td></td>
</tr>
<tr>
<td>( \sigma_7 )</td>
<td>( \sigma_3 )</td>
<td>( \sigma_4 )</td>
<td>( \sigma_1 )</td>
<td>( \sigma_2 )</td>
<td>( \sigma_2 )</td>
<td>( \sigma_6 )</td>
<td>( \sigma_4 )</td>
<td>( \sigma_4 )</td>
<td></td>
</tr>
</tbody>
</table>

(the state that is reached in the original STD when emitting \( S = 2 \) from state \( \sigma_1 \)). Again, the single exception is that emission of symbol \( S = 7 \) is not allowed from \( \sigma_7 \):

\[ \rightarrow \sigma_1 \rightarrow \sigma_7 \rightarrow 2 \quad 7 \]

The state-transition diagram (STD) for the 2D-constraint in the 3-row strip is given in Table 4.3 (with the additional states underlined, and with the forbidden symbols indicated by "X"). This 7-state STD has the following connection matrix (the elements of which indicate how many branches exist that connect one state to another):

\[
D = \begin{bmatrix}
2 & 2 & 1 & 2 & 0 & 0 & 1 \\
2 & 2 & 2 & 0 & 0 & 0 & 0 \\
1 & 2 & 1 & 2 & 1 & 1 & 0 \\
2 & 2 & 2 & 0 & 0 & 0 & 0 \\
1 & 2 & 1 & 2 & 1 & 0 & 0 \\
2 & 2 & 1 & 2 & 0 & 0 & 0 \\
1 & 2 & 1 & 2 & 0 & 1 & 0 \\
\end{bmatrix} \quad (4.12)
\]

The largest eigenvalue of \( D \), denoted \( \lambda_{\text{max}} \), is equal to 7.91697. The capacity corresponding with the 2D constraint for a 3-row strip is given by:

\[
C = \frac{1}{3} \log_2(\lambda_{\text{max}}) = 0.994983. \quad (4.13)
\]

The factor 1/3 originates from the fact that each symbol comprises three channel bits.
2D $k$-Constraint  In traditional 1D storage, very long runs lead to inaccuracies in the timing recovery, which is dealt with by a device called a phase-locked loop (PLL). The PLL regenerates the internal “clock” that is matched to the length of the bits on the medium: the bit clock is adjusted at each variation in the detected signal waveform. Areas on the medium with too few transitions may cause “clock-drift”. Similarly, in a 2D-encoded strip, it is necessary that the 2D channel bitstream shows some variation which can be used as timing information or as information for the adaptive control loops. Therefore, the number of successive $S = 0$ symbols is limited to a small number, denoted $k$, similarly to the $k$-constraint in 1D RLL channel coding. Such a so-called ‘joint’ $k$-constraint over a number of contiguous bit-rows is well known in the 2D coding literature, and was originally proposed in [50, 51].

To combine the 2D-constraint with such a joint $k$-constraint, the following procedure is followed. First, all branches in the STD of Table 4.3 are identified that correspond with the $S = 0$ symbol; with these branches, a $7 \times 7$ sub-matrix $Y$ is constructed. Further, the sub-matrix $X$ is defined as the original connection matrix of Eq. (4.12) minus $Y$. $X$ is thus that part of the original connection matrix that corresponds to the emission of symbols other than the 0-symbol. With these two sub-matrices, the number of times a 0-symbol is emitted is easily controlled via a new connection matrix (including the $k$-constraint) that is denoted $D[k]$. It is obtained as a $7(k+1) \times 7(k+1)$ matrix, which can be written as a $(k+1) \times (k+1)$ block matrix:

\[
D[k] = \begin{pmatrix}
X & Y & 0 & \cdots & 0 & 0 \\
X & 0 & Y & \cdots & 0 & 0 \\
& \cdots & \cdots & \cdots & \cdots & \cdots \\
X & 0 & 0 & \cdots & 0 & Y \\
X & 0 & 0 & \cdots & 0 & 0
\end{pmatrix}
\] (4.14)

where each “0” matrix element denotes a $7 \times 7$ all-zero matrix, and where the sub-matrices $X$ and $Y$ are given by:

\[
X = \begin{pmatrix}
1 & 2 & 1 & 2 & 0 & 0 & 1 \\
2 & 1 & 2 & 2 & 0 & 0 & 0 \\
1 & 2 & 0 & 2 & 1 & 1 & 0 \\
2 & 2 & 2 & 1 & 0 & 0 & 0 \\
1 & 2 & 0 & 2 & 1 & 0 & 0 \\
1 & 2 & 1 & 2 & 0 & 0 & 0 \\
1 & 2 & 0 & 2 & 0 & 1 & 0
\end{pmatrix}
\] (4.15)

and
It is interesting to note that the connection matrix in Eq. (4.14) has a structure that is reminiscent of the connection matrix for a 1D RLL constrained sequence with \( d = 0 \) and a finite \( k \)-constraint (just replace \( X \) and \( Y \) in Eq. (4.14) by unity).

For \( k = 2 \), the capacity amounts to \( C[k] = 0.994123 \); hence, a rate \( R = 152/153 = 0.99346 \) code is possible for the 2D-encoded strip (with 3 bit-rows). This code-mapping is obtained as follows. The number 152 is the largest number \( n \) of user bits that satisfies three conditions: (i) it is a multiple of 8 (since we use byte-oriented error-correction coding), (ii) \( n + 1 \) is a multiple of 3 (the number of bits in a fish-bone symbol), and (iii) \( R = n/(n+1) \leq C[k = 2] \).

**Enumerative Coding Scheme** A convenient way to implement the high-rate 152-to-153 2D modulation code is by means of enumerative coding (see [7, 52], where it is applied to 1D runlength limited (RLL) coding). Since the channel symbols in the 3-row wide 2D-encoded strips comprise three bits \( (M = 8) \), the \( M \)-ary version for the lexicographic ordering as given by Cover [52] must be used, which yields for the rank \( i_S \) of a channel word \( x \):

\[
i_S(x) = \sum_{j=1}^{n} \sum_{m=0}^{x_j-1} n_S(x_1, x_2, \ldots, x_{j-1}, m).
\]

In this equation, a 2D channel word \( x \) comprises \( n \) channel symbols (bit-triplets or fish-bones) denoted \( x_1, x_2, \ldots, x_{n-1}, x_n \). Further, \( n_S(x_1, x_2, \ldots, x_{j-1}, x_j) \) represents the number of channel words of length \( n \) for which the first \( j \) channel symbols are equal to \( x_1, x_2, \ldots, x_{j-1}, x_j \). Note that in the binary case with 1D \( d \)-constraint RLL coding, the set of numbers \( n_S(x_1, x_2, \ldots, x_{j-1}, x_j) \) is simplified to \( x_j N_d(n-j) \) (for a \( d \)-constrained sequence, with \( x_j \) the binary bit value of the differential bitstream, with \( n \) the channel-word length, and with \( N_d(p) \) the number of \( d \)-constrained sequences of length \( p \)). A similar reduction cannot be made for the current case. Instead, the following procedure is used. The current state after emission of symbol with index \( j \) is indicated with \( \Sigma_j \), where \( \Sigma_j \) is one of the basic STD-states, that is, \( \Sigma_j \in \{ \sigma_1, \sigma_2, \ldots, \sigma_7 \} \). The encoding of a channel word is always started from the same STD-state in order to avoid channel errors to propagate from one word into the next one. This initial state is denoted as \( \Sigma_0 \), which can be chosen as \( \Sigma_0 \equiv \sigma_2 \) (the reason for this particular
For the index sets $n_S$ that are to be used in Eq. (4.17), it is required to know what the resulting state will be when symbol $m$ is emitted to reach symbol position $j$, hereby considering the history of the previously emitted channel symbols $x_1, x_2, \ldots, x_{j-1}$ (which represent the coding path through the state-transition diagram). This is shown schematically below, with the resulting state denoted $\Sigma^m_j$, where the superscript $m$ is used to indicate that for symbol number $j$, we have $0 \leq m < x_j$:

$$
\Sigma_0 \xrightarrow{x_1} \Sigma_1 \xrightarrow{x_2} \Sigma_2 \xrightarrow{x_3} \Sigma_3 \ldots \xrightarrow{x_{j-1}} \Sigma_{j-1} \xrightarrow{m} \Sigma^m_j.
$$

This means that

$$n_S(x_1, x_2, \ldots, x_{j-1}, m) = n'_S(\Sigma^m_j), \quad (4.18)$$

where $n'_S(\Sigma^m_j)$ indicates the total fan-out from state $\Sigma^m_j$ up to any possible arrival state $q$ after completion of the channel word (with emission of the $n - j$ extra symbols so that a total of $n$ channel symbols in the channel word is realized). Therefore, it can be derived that

$$n'_S(\Sigma^m_j) = \Sigma_q D^{n-j}_{\Sigma^m_j, q} \quad (4.19)$$

where $D$ is the connection matrix of the state-transition diagram, the required power is $n - j$ (the remaining number of symbols that are still to be emitted), the required matrix element is $(\Sigma^m_j, q)$ and the sum is over all possible arrival states $q$.

In the practical case of the example it is chosen to have $n = 51$. Concatenation of consecutive channel words is realized by means of an extra merging symbol, that redirects the coding path to the initial coding state that was agreed upon (e.g. $\Sigma_0 \equiv \sigma_2$, see above). At this stage, a sufficient condition for a proper choice of the initial state $\Sigma_0$ is that it can be reached from any other state of the STD by emission of a single symbol: therefore, the three states that were added to realize the 2D constraint, $\sigma_5$, $\sigma_6$ and $\sigma_7$, are no potential candidates for $\Sigma_0$. In case additional DC-control must be realized along one of the bit-rows of the 2D 3-row strip, as will be discussed in the next paragraph, then an extra condition will further restrict the candidates for $\Sigma_0$.

**DC-control**  In a 2D optical drive, servo systems are used for tracking of the position of the laser spot array along the radial direction of the meta-spiral, and for focusing of the laser spot on the information layer. These servo systems typically require a DC-free data signal. The signals used for the servo control are derived from one of the laser spots that are scanning the 2D meta-spiral, and thus result from a
single bit-row within the meta-spiral. If the spectral power density of a considered bit-row reveals a "notch" (or zero) at DC, then the signal is considered to be "DC-free".

The sequence of the considered bit-row with bipolar bits $b_1, b_2, \ldots$ is called DC-free if its running digital sum (RDS)

$$RDS_i = \Sigma_{j=-\infty}^{i} b_j$$

takes on only finitely many different values. In that case, the power spectral density function vanishes at DC. A well known procedure to restrict the value of the RDS is by insertion of an extra NRZ channel symbol that offers the freedom to flip the polarity of the subsequent channel bitstream (and thus the sign of its contribution to the RDS, not its magnitude). Note that a ‘1’ instead of a ‘0’ in the NRZ bitstream of a single bit-row changes the polarity of the subsequent RLL (or NRZI) bitstream due to the use of the 1T-precoder. For the current case with the 2D-encoded 3-row wide strips, DC-control needs to be realized for only one out of the three bit-rows (in view of servo-control in the 2D optical drive that uses only few rows of the total meta-spiral, see Chapter 2). For this purpose, the same NRZ-symbol that was required for the merging of two consecutive channel words can be used so that the coding path could be redirected towards the same initial state $\Sigma_0$ in the STD.

![Figure 4.13](image)

**Figure 4.13:** DC-control for one bit-row of the 2D-encoded 3-row strip via the extra symbol that is also used to redirect the coding path of the enumerative coding to STD-state $\sigma_2$ as the initial state $\Sigma_0$. The situation is shown where a choice has to be made between NRZ-symbols $S_0 = 2$ and $S_1 = 5$.

Assume $S_0$ and $S_1$ represent two complementary NRZ symbols, that is, $S_0 + S_1 = 7$. By selection of either $S_0$ or $S_1$ as merging NRZ symbol, it is possible to flip the
4.3 High Rate Coding for Elimination of Critical Patterns

polarity of all three bit-rows at once. Note that it is only needed to control the RDS of only a single bit-row; therefore, any of the three bit-rows can be selected to be made DC-free. Note that, upon emission of either $S_0$ or $S_1$, it is required to arrive in the same initial state $\Sigma_0$: this extra condition eliminates the STD-states $\sigma_1$ and $\sigma_3$ from the list of potential candidates for $\Sigma_0$. Now choose $\Sigma_0 \equiv \sigma_2$. Then the following 4 cases can be distinguished: (i) if the last state of the preceding $n$-symbol channel word equals $\sigma_1$ or $\sigma_6$, then the two symbols $S_0$ and $S_1$ equal 1 and 6; (ii) if said last state equals $\sigma_2$, then the two symbols $S_0$ and $S_1$ equal 0 and 7; (iii) if said last state equals $\sigma_3$, $\sigma_5$ or $\sigma_7$, then the two symbols $S_0$ and $S_1$ equal 3 and 4; and (iv) if said last state equals $\sigma_4$, then the two symbols $S_0$ and $S_1$ equal 2 and 5: the latter situation is depicted in Fig. 4.13. A similar reasoning would apply if the alternative initial state would have been chosen, that is, $\Sigma_0 \equiv \sigma_4$.

4.3.5 Modulation Coding for Building Block Type II: 1D-Encoded Single Bit-Rows

In strip-based coding as proposed in [14, 47], strips are constructed so that concatenation of strips in the radial direction does not lead to violations of the constraints across the strip boundaries: for this purpose, the clusters at the boundary of a strip have to satisfy a special boundary constraint. In the approach discussed in the previous paragraph, the 2D encoded strips cannot be freely concatenated in the radial direction. Therefore, when strips are stacked upon each other, it is possible to identify at the boundary area of the two strips two other 3-row wide areas that overlap with the two considered strips. This is illustrated in Figure 4.14. It is obvious that, al-

![Figure 4.14](http://example.com/figure414.png)

**Figure 4.14:** Stacking of two 3-row strips.

though the 2D constraint applies to all strips individually, it may be violated for each of these two 3-row wide areas that cross a boundary between strips. The purpose of the building blocks of the second type, which are the 1D encoded bit-rows positioned in between two consecutive strips as shown in Fig. 4.15, is to glue these consecutive strips together such that the 2D constraint also applies in the boundary area of the
strips.

![Figure 4.15: Stacking of two 3-row strips with a merging bit-row in between.](image)

**Forbidden Sequences in 1D-Encoded Single Bit-Rows** In order to satisfy the 2D constraint also in the boundary area of consecutive strips, it is sufficient to forbid the following two repetitive sequences in the NRZI bitstream of the 1D encoded single bit-rows: \( \cdots 00100100100100 \cdots \) and \( \cdots 11011011011011 \cdots \). These two NRZI sequences correspond with the single NRZ sequence given by: \( \cdots 0110110110110 \cdots \). By means of this constraint on the single bit-rows, the actual phase of the modulation codes of the two building blocks has become irrelevant. Therefore, the 1D modulation code is designed such that the NRZ pattern 11011 is prohibited. The corresponding state-transition diagram (for a \( k = 5 \) constraint) is shown in Fig. 4.16.

![Figure 4.16: State-transition diagram for the 1D modulation code of the merging bit-rows.](image)

**Modulation Code** A sliding block code for the above 1D constraint is constructed using the well known state-splitting algorithm or ACH-algorithm [53]. The code maps user words of 12 user bits onto channel words of 13 channel bits (code rate \( R = 12/13 \)). Its \( k \)-constraint is chosen to be \( k = 5 \). The approximate eigenvector equals \( \{2,2,2,1,1,2,1,1\} \). The code is not designed to be DC-free (DC-control is reserved for one of the bit-rows of the 3-row strips).
4.3 High Rate Coding for Elimination of Critical Patterns

4.3.6 Bliss-like Scheme for the Combination of Modulation Coding and ECC Coding

The proposed scheme uses two modulation codes. The first modulation code that needs to be applied for the 3-row strips has a high coding rate, and is implemented via a coding scheme that uses very long codewords (with a mapping of 152 user bits onto 153 channel bits). It is used for the largest fraction of the source data. For such a high-rate code, the traditional ‘Figure 1’ (see Chapter 1) is not very well suited. Its very long codewords that comprise (at the user side) a number of consecutive ECC symbols (or bytes) are subject to excessive error-propagation: a single channel error during read-out will affect one complete channel codeword, but will lead to a large number of erroneous bytes prior to ECC decoding. For traditional high-rate 1D RLL codes to be used in conjunction with error-correction coding, Bliss [54] has proposed a variant on ‘Figure 1’ that does not suffer from the above drawbacks.

First the Bliss-scheme is briefly explained for traditional 1D modulation encoding. In the encoder (shown in Fig. 4.17), there are two modulation codes instead of a single one. The first modulation code (step (1)) is positioned before the ECC encoder, unlike the traditional ‘Figure 1’. The first modulation code has a (very) high code rate, which necessitates to use long codewords, but leads to severe error propagation. Subsequently, the ECC encoder (step (2)) operates on the constrained sequence that is produced by the first modulation encoder. Because of the high rate of the latter, the ECC is only slightly less effective than in the case the ECC would be applied on the original source data (since the first modulation encoder with its high rate adds only limited correlation). The parities generated by the ECC encoder are then encoded by the second modulation encoder (step (3)), which has not that very high rate (of the first modulation encoder), and which does not suffer from severe error-propagation. The constrained sequence for the parity part, produced by the second modulation encoder, is then cascaded in step (4) with the constrained sequence for the data part, produced by the first modulation encoder: the cascading process might require some merging procedure. In the decoder of the Bliss-scheme, the part of the channel bitstream corresponding to the parities is decoded first by the second modulation decoder (which yields only little error-propagation). Then, the ECC decoder operates with as input the (decoded) parities and the part of the channel bitstream corresponding to the data: the ECC decoder produces at its output the error-free channel bitstream. In this way, channel errors that occur in the channel bitstream corresponding to the data part do not lead to error propagation. Finally, the decoder of the first channel code produces the corresponding source data for the user.

Now let’s go back to the 2D situation. The first modulation encoder was applied for encoding the 3-row strips at a high rate. The second modulation code that needs to be applied for the single merging bit-rows has a lower coding efficiency (with a 12-to-13 mapping), but (unlike the first code) does not suffer from excessive error-
propagation. Recall that an additional purpose of this second modulation code is to glue 3-row wide strips together while maintaining the 2D constraint also at the boundary area. The second modulation code is used for the ECC parities, and for a certain fraction of the source data. In the next paragraphs the encoder and decoder for the 2D modulation scheme in conjunction with a Bliss-like scheme for the error-correction coding is described.

### 4.3.7 Encoder

The different steps at the side of the encoder are shown in Fig. 4.18. In a first step (step-1), the source data is partitioned into two parts, denoted Part-1 and Part-2. This subdivision is convenient since, for an 11-row meta-spiral, the two single bit-rows (second type of building block, Section 7) represent a data capacity that exceeds the overhead that is typically needed for the ECC parity symbols (for Blu-ray Disc: 15%). Then, the source data Part-1 is encoded with (the first) modulation encoder-1 (in step-2), producing the channel bitstream in the distinct building blocks of the first type. The latter channel bitstream together with the source data Part-2 are then the inputs (denoted as (3.a)) for the ECC encoder (step-3), which produces ECC parities (3.b) at its output. Subsequently, the source data Part-2 together with the ECC parities are the input of (the second) modulation encoder-2 (in step-4), producing the channel bitstream in the distinct building blocks of the second type (in this case, the single merging bit-rows). In step-5, the different building blocks are assembled (or multiplexed) to generate the overall channel bitstream of the full strip. The latter overall bitstream is then ready to be transferred over the channel (step-6).
4.3 High Rate Coding for Elimination of Critical Patterns

4.3.8 Decoder

The different steps at the side of the decoder in our 2D coding scheme are shown in Fig. 4.19. In a first step (step-1), the as-detected overall channel bitstream is demultiplexed into the respective parts corresponding to the 3-row strips and merging bit-rows. In step-2, modulation decoder-2 decodes the as-detected channel bitstream of the building blocks of the second type into ECC parities and source data Part-2. Modulation decoder-2 has only little error-propagation (it is based on 12-bit input words, which are 1.5 bytes long). ECC decoding is performed in step-3: at its input (3.a), the ECC decoder uses the as-detected channel bitstream of the building blocks of the first type, the ECC parities and the source data Part-2: all these may contain channel errors produced by the bit-detection module that uses the read-out signals. At its output (3.b), the ECC decoder (of step-3) produces the error-free (corrected) channel bitstream of the building blocks of the first type, and the corrected source data Part-2. Next, the error-free channel bitstream of the building blocks of the first type are decoded by modulation decoder-1 (step-4), hereby generating the source data Part-1. In step-5, the two parts of source data are reassembled to generate the overall source data, that are to be sent to the user in step-6.

4.3.9 Data-Allocation in the Meta-Spiral

Now consider the data allocation for the practical case of an 11-row meta-spiral consisting of three 3-row stripes and two merging bit-rows. DC-control is applied in the outer two bit-rows of the meta-spiral for the purpose of the radial tracking servo, and in the center bit-row for the focusing servo: these bit-rows are the top and bottom
outer bit-rows of the corresponding two outer 3-row strips, and the center bit-row for the central 3-row strip. No DC-control is applied on the merging bit-rows. Consequently, for each block of 52 channel symbols (including one merging symbol) along the tangential direction of the meta-spiral, we have 19 (152/8) byte-symbols for each 3-row strip, and 6 (1/8 × 12/13 × 52) byte-symbols for each merging bit-row. Thus, for each 52 symbols along an 11-row meta-spiral, a total of 69 bytes can be encoded.

### 4.4 Test Format

For evaluation purposes, it is convenient to use the receiver (as discussed in the next chapter) in a data aided mode, which means that the a-priori known data which is recorded on the medium is used to generate the error signals for the feedback control loops. The final receiver however, will be of the decision directed type, meaning that feedback control loops make use of the detected data to generate error signals that drive the loop filters. In the latter case most control loops will suffer from a ‘Baron-von-Münchhausen’ syndrome in the sense that they cannot ‘pull themselves out of the swamp’. A badly converged control loop will cause erroneous detection. The erroneous detected data will be used for calculating error signals to drive the control

---

2Karl Friedrich Hieronymus, Baron von Münchhausen (May 11, 1720 - February 22, 1797) was a German nobleman who served in the Russian army against the Turks, about which he was used to tell very tall stories. These stories were collected by Rudolf Erich Raspe and published in 1785 with the title: 'The Surprising Adventures Of Baron Munchhausen'. According to these stories the baron could do amazing things like travelling on a canon-ball and pulling himself out of the swamp from his own boots (from which the word ‘bootstrapping’ is derived).
loops, thereby further deteriorating the pre-detection processing conditions. To overcome this problem the data is organized in frames, where each frame is preceded by a sequence of a-priori known data called the preamble. The data in the preamble is chosen such that it contains a maximum of information for the control loops that need to be converged before the actual pay-load of the frame starts. A periodic pattern of two ‘1’s and two ‘0’s (2T/2T repeat pattern) is chosen for the control loops timing, DC and gain in the evaluation format. This pattern is repeated 64 times. At the end of the preamble a sync-pattern is inserted to mark the start of the actual data. This pattern is chosen such that it has a large detection distance with respect to the pattern in the preamble [55]. The total length of the frame is 2560 bits. Fig. 4.20 shows a schematic representation of the test format. Note that in our test-format 10% of the data is ‘wasted’ with this preamble-length. This was done to ensure that in this test format it was possible to reach convergence of the control loops on the preamble. In a final design a trade-off must be made between overhead for the control loops and robustness of the system. For example, one can choose long frames such that overhead due to the preamble is low. However, if the control loops loose lock (due to a scratch or other disturbance) they cannot converge anymore until the next preamble. Another way is to decrease the length of the preamble. However, this reduction in length may be limited by the convergence speed of the control loops and the allowed residual error of a particular controlled parameter at the start of the data.

![Image](image.png)

Figure 4.20: Schematic representation of the frame-based evaluation format.

### 4.5 Conclusions

It is the task of modulation coding to transform the input data into a sequence of channel bits that have desirable properties, such as the ability of being transferred with high-reliable across the optical channel. Two different types of modulation codes are discussed:

- **Low-pass Codes**: These codes impose constraints on the channel bit sequence in terms of the nearest neighbours. They lead to a ‘more open eye-diagram’.
We defined constraints on the hexagonal lattice in terms of (i) the minimum number of nearest neighbour bits of the same type as the bit located on the center lattice site or (ii) the minimum number number of azimuthically contiguous nearest neighbours. It appears that it is possible to open the eye-diagram for a density equal to two times BD, however, at a large code rate loss. To compensate for this loss smaller pits are required leading to problems during the manufacturing of the discs.

- Elimination of worst case patterns: This is done by removing those data patterns that support error sequences that are prone to errors during transfer across the optical channel. Examples are patterns with a high loss with respect to the matched filter bound. A brute-force search using a linear channel shows that 2D sequences of alternating ‘+1’ and ‘-1’ symbols are especially vulnerable. The ‘close-ring’ of 6 alternating bits appears to be the worst case one. Other examples are 2D periodic structures that have most spectral content beyond the cut-off of the optical channel.

A 2D modulation code has been designed that forbids the occurrence of 2D periodic structures in the data stream that have a lot of high-frequency spectral content. As a practical implementation strip-based 2D coding is applied. The strips consisting of 3 rows can be concatenated one upon the other in the radial direction. The periodic structures are eliminated by modifying the fan-out of certain states in the state transition diagram of the code in such a way that emission of symbols that lead to a further build up of the worst case pattern are not allowed.

This leads to 2D modulation codes with a high rate (152 to 153), which can be conveniently implemented by means of enumerative coding. For the 2D implementation of 3-row strips the $M$-ary ($M=8$) version of the lexicographic ordering must be used to be able to find the index that corresponds to the received 2D channel word. The length of the channel word is chosen to be 51. The stripes cannot be freely concatenated in radial direction and therefore, a single ‘glue’ row is inserted between each pair of strips. This row forbids a few repetitive sequences in a 12-to-13 low-rate modulation scheme in order to satisfy the constraints in the boundary area. DC control is done only on those rows that are used for servo by inserting merging symbols.

The problem of the very high rate code with that is efficiently implemented with the long codewords is that it has excessive error-propagation. For this reason a ‘Bliss-scheme’ had been applied that exchanges the order of the modulation and error correction coding.

A test format has been designed that consists of frames with a length of 2560 bits that start with a preamble followed by the payload containing the modulation coded data.
Chapter 5

The 2D Receiver

This chapter will discuss the signal processing path from Photo Detector IC (PDIC) up until the bit-detector (the bit-detector itself is the subject of Chapter 6). In Section 5.1 some key design considerations are given on which the block diagram of the receiver is based. The blocks are then discussed more elaborately in the order of appearance in the signal processing chain of the receiver block diagram, starting with the analogue front-end (Section 5.2). After digitization of the signals the samples need to be delayed by a row-dependent delay in order to compensate for the delays due to the slanted orientation of the grating with respect to the tangential direction of the broad spiral (see Section 2.2). Then, a number of processing steps are performed (Section 5.4) to transform the raw signal into a signal at the input of the bit-detector that matches the ideal target response ‘as well as possible’. Generally, the processing is adaptive to account for variations in the actual channel response. The processed signal is input to the 2D bit-detector, which, in its full-fledged version, is considerably more complex than a 1D bit-detector. A modified configuration makes practical implementation possible but shows a significant detection delay (see Chapter 6). This delay is part of the adaptation loops and therefore influences the maximum bandwidth and the stability of these loops. A novel strategy to partly circumvent the problems of detection delay in the 2D system is discussed in Section 5.5. Final conclusions are given in Section 5.6.

5.1 Block Diagram and Key Design Considerations

A block diagram of the receiver is shown in Fig. 5.1. The optical signals are transduced into a set of analogue electrical signals by a segmented PDIC [56]. As a drawing convention we will use wide arrows for multiple signals and double lines for blocks that are instantiated multiple times. At the input of the receiver one analog-to-digital converter (ADC) per row is used to digitize the analogue replay signals. The ADCs run in parallel at a fixed sampling clock with frequency $f_s = \frac{1}{T}$. On the one hand it is convenient for practical reasons to choose this sampling frequency as low as possible. On the other hand it must be high enough to avoid aliasing of data
components, i.e. it should obey $f_s > 2f_c$. Here $f_c$ is the cut-off frequency of the optical channel. When $f_s = 2f_c$, an ideal analogue anti-aliasing filter would be needed to avoid aliasing of the high frequency noise components. In order not to lose or distort data components a flat pas-band is needed with an amplitude equal to 1 for $f \in [0, f_c]$, and for ideal suppression of the noise an infinitely steep transition band at $f_c$ and an infinite attenuation at $f > f_c$ is needed. In practice such a filter can not be realized. For this reason the analog signal is over-sampled at $2.5\ T$ and more relaxed filter requirements are taken as a basis of the design (see Section 5.2). After digitization a more steep digital noise filter is applied to the signals $y_n$ (in combination with delay compensation). Now the sampling frequency can be further reduced (a factor of 2 by simply throwing away each second sample) to reduce the calculation effort needed for further digital processing of the signal. The delay compensation block implements an adaptive delay which compensates the relative signal delay between the rows due to the slanted orientation of the grating with respect to the tangential direction of the broad spiral. Adaptivity is needed because the delay is not perfectly known beforehand, and may be time-varying, e.g. due to track pitch variation. It should therefore be estimated ‘on-the-fly’. The loop is based on the knowledge that during the preamble the signals are equal in each row and should be aligned perfectly. During this time the control loop is closed and an update of the relative row delay is generated. To indicate the presence of this preamble a preamble detection circuit is included [57].

![Block diagram of the total 2D receiver.](image)

In Section 3.4 an eye-diagram was plotted for a density equal to two times the BD density. The eye-diagram is completely closed due to the little (or even negative) excess bandwidth of the system. This means that e.g. zero-crossing based timing error detection is not feasible due to the very large data-induced zero-crossing jitter (pattern dependent jitter). An error signal, which we denote $e_{\Lambda}$, must be derived in another way. With $\Lambda$ we indicate the two-dimensional index $(k, n)$ where $k$ denotes
5.1 Block Diagram and Key Design Considerations

the time index in the domain synchronous to the bit clock (\( t \)) and \( n \) the row index. We generate \( e_\Lambda \) by subtracting an ideal reference signal \( d_\Lambda \) from the actual detector input \( d_\Lambda \). A fixed bi-linear target response \( g_\Lambda \) is used to calculate the ideal reference signal based on detected bits \( \hat{a}_\Lambda \). Several signal processing blocks are present to transform the replay signal into a sequence that resembles the sequence \((g * \hat{a})_\Lambda\) as well as possible. Here ‘*’ indicates the two-dimensional convolution defined as:

\[
d_\Lambda = (g * \hat{a})_\Lambda = \sum_{(i,j)\in \mathcal{N}} g_{i,j} \hat{a}_{k-i,n-j}
\]  

(5.1)

with \( \mathcal{N} \) some 2D neighbourhood equal to the size of the target response. As a first signal processing block a 2D adaptive linear equalizer is present. The taps of this equalizer are updated via an adaptation loop such that the actual response of the 2D channel is transformed into the response that resembles the target response. Several adaptation criteria can be applied. We use a zero-forcing (ZF) technique where the loop acts to drive a particular error signal on average to zero. Another algorithm that has been used is the least mean square (LMS) algorithm where the loop acts to minimize the power of the error. More details can be found in Section 5.4.1. The equalizer is followed by automatic gain control (AGC; see Section 5.4.3) and DC compensation (see Section 5.4.2), which will try to compensate any variations in the amplitude or DC-content of the signal. The last block is the sampling rate converter (SRC). It converts the sequence of samples taken at the fixed free-running sample clock to a sequence of samples at the decision instants, i.e. at the baud-rate. The first samples are said to be in the asynchronous domain (with respect to the bit-clock), while the latter sequence of samples are in the synchronous domain. To be able to generate a sample at the correct phase the SRC receives a new phase value from the adaptation (or timing-recovery) block at each decision instant. A more detailed discussion will follow in Section 5.4.4.

The order of the processing (AdEq-DC-AGC-SRC) is based on the bandwidth requirements of each of the loops. For this reason, the DC, AGC and adaptive equalizer loops are placed in front of the sampling rate converter (SRC) because the timing-recovery loop must be able to track fast timing variations. If the propagation delays of the AdEq, DC, and AGC processing blocks would be part of the timing-recovery loop it would reduce its stability margins (or equivalently its maximum bandwidth) [58]. Moreover, the adaptive equalizer is inherently slow due to the large number of taps that need to be adapted simultaneously. Time constants of the equalizer loop are in the order of thousands of bits, which makes the adaptive equalizer a suitable candidate to be the first processing block in the chain. The order of the DC and AGC loop is not that important since their processing delay is minimal and will not influence either of the loops. Another, very significant, source of delay in the control loops is the bit-detector. This will be discussed in more detail in Section 5.5.
5.2 Analogue Front-end

Photo Diode IC  The analogue front-end starts with a 11-spots photo detector IC (PDIC). This IC implements a segmented photo-detector according to the configuration shown in Fig. 5.2. The middle and outer detectors are segmented as quadrant detectors to allow the generation of focus error signals and radial tracking error signals. The currents generated in the reverse-biased photo-diodes are amplified by multiple-stage current amplifiers. The gain is switchable between 2048 and 8192. A block diagram of the amplifier stages is shown in Fig. 5.3 (in this example 4 stages are shown with a total amplification $M = M_1 \cdot M_2 \cdot M_3 \cdot M_4$). The photodiode current is

\[ I_{\text{PD}} \]

the collector current of transistor $T_1$. Due to this current the voltage on this collector node then tends to increase, which is sensed by the negative input of the operational amplifier $D_1$. The voltage on the output of this amplifier, which is connected to the base nodes of transistors $T_1$ and $T_2$, decreases. This causes the collector currents of both $T_1$ and $T_2$ to increase in such a way that the collector node of $T_1$ is kept at the reference voltage $V_{\text{ref}}$. Due to the different emitter sizes of transistor $T_1$ and $T_2$ with a ratio $1:M_1$ the collector current of transistor $T_2$ is $M_1$ times as large as the collector current of transistor $T_2$. Hence, amplifying the current signal. One of the advantages of the circuit is that the voltage on the photo-diode is kept constant. This minimizes...
the influences of its parasitic capacitance on the bandwidth of the input stage. It is clear that the photo-diode signal itself is used as a bias current for the amplification circuit. This means that some minimum illumination of the detector is required for the IC to operate properly and to achieve a sufficiently high bandwidth. The IC is designed in a CBiMOS3 process and measures $4.87 \times 5.86 \text{ mm}^2$. A plot of the layout of the IC is shown in Fig. 5.4. The PDIC is mounted on a mechanically adjustable holder that allows a proper alignment of the PDIC with respect to the light path (see Fig. 5.5). The currents are transported via flex-foils to a printed circuit board (PCB) with trans-impedance amplifiers and analogue anti-aliasing filters. A photograph of the PCB is shown in Fig. 5.6.

**Anti-aliasing Filter** As a starting point for the specification of the anti-aliasing filter a user data rate of 35 Mb/s per row was taken, which is equal to the user data rate of 1x BD. Table 5.1 gives an overview of some of the key system parameters under the assumptions that the ECC overhead is 15% and the code-rate is 0.98 for a code that only eliminates some 2D worst case patterns (2D-wcp) equivalent to a RMTR-constraint in the 1D system (see Chapter 4). In the table also an ‘excess bandwidth factor’ is listed, which is defined as the ratio between the cut-off of the MTF and the Nyquist frequency. From the table it becomes clear that the channel cut-off frequency for the required user data rate of 35 Mb/s per row in the broad spiral, is equal to 23.8 MHz. As an anti-aliasing filter a 5th order Chebyshev-II filter has been applied with a cut-off at 37 MHz, and a minimum attenuation of 35 dB in the stop-band. The filter
Figure 5.5: PDIC setup.

(a) PDIC photograph.  (b) Mechanically adjustable holder

Figure 5.6: Printed circuit board containing the trans-impedance amplifiers and the analogue anti-aliasing filters.
5.2 Analogue Front-end

Table 5.1: Parameters for phase-1 (1.4x BD density) and phase-2 (2x BD density).

<table>
<thead>
<tr>
<th>Parameter</th>
<th>BD</th>
<th>TwoDOS $a_H=165$ nm</th>
<th>TwoDOS $a_H=138$ nm</th>
<th>Unit</th>
<th>Note</th>
</tr>
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<tbody>
<tr>
<td>NA</td>
<td>0.85</td>
<td>0.85</td>
<td>0.85</td>
<td>nm</td>
<td></td>
</tr>
<tr>
<td>$\lambda$</td>
<td>405</td>
<td>405</td>
<td>405</td>
<td></td>
<td></td>
</tr>
<tr>
<td>spatial cut-off MTF</td>
<td>4.19</td>
<td>4.19</td>
<td>4.19</td>
<td>$\mu m^{-1}$</td>
<td></td>
</tr>
<tr>
<td>Required User Data Rate</td>
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<td>35</td>
<td>35</td>
<td>Mb/s</td>
<td>per row</td>
</tr>
<tr>
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<td>0.98</td>
<td>0.98</td>
<td>Mb/s</td>
<td></td>
</tr>
<tr>
<td>ECC Overhead</td>
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<td>15</td>
<td>15</td>
<td>%</td>
<td></td>
</tr>
<tr>
<td>Channel Data Rate</td>
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<td>41.07</td>
<td>Mb/s</td>
<td></td>
</tr>
<tr>
<td>Bit-size/lattice parameter</td>
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<td>165</td>
<td>138</td>
<td>nm</td>
<td></td>
</tr>
<tr>
<td>Linear Disc Velocity (CLV)</td>
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<td>6.78</td>
<td>5.67</td>
<td>m/s</td>
<td>r=50 mm</td>
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<tr>
<td>Disc Rotational Speed</td>
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<td>21.57</td>
<td>18.04</td>
<td>Hz</td>
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<tr>
<td>temporal cut-off MTF</td>
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<td>28.45</td>
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</tr>
<tr>
<td>bit-time</td>
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<td>24.35</td>
<td>ns</td>
<td></td>
</tr>
<tr>
<td>d-constraint</td>
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<td>0</td>
<td></td>
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<tr>
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<td>MHz</td>
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</tr>
<tr>
<td>Excess bandwidth factor</td>
<td>1.25</td>
<td>1.39</td>
<td>1.15</td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

circuit, with an impedance of 50Ω is shown in Fig. 5.7. Its transfer characteristic is plotted in the left part of Fig. 5.8, together with the aliased versions (light grey lines) for a sampling frequency of 102.5 MHz (2.5 times oversampled with respect to the bit-frequency). The right part shows the group delay of the filter. This group delay is constant within $\pm 1$ ns within the pass-band of the filter. This is achieved by adding 3 equalization orders to the filter. The adjustable coils to tune the group delay can be clearly observed in the photograph of Fig. 5.6.

Figure 5.7: Fifth order low pass Chebyshev-II filter with three equalization orders to obtain a flat group delay.
5.3 Delay Compensation

The Algorithm  The algorithm makes use of the fact that the data in the preamble is equal in each row (Section 4.4). A block diagram of the delay-compensation function is shown in Fig. 5.9.

A variable delay is present in each row except for one row. This row is used as a reference row. A timing error detector (TED) is used to measure the phase difference between each of the rows and the reference row. The principle of this detector is similar to the one used in the phase locked loop and is well known in literature [59]. Let us assume that the replay signals are equal for each row except for noise, and except for the relative delay between the rows. An error sequence $e(mT_s)$ is generated by subtracting the signal of the reference row $y_{REF}^mT_s(mT_s)$ from each of the signals of
one of the other rows \( y^n(mT_s + \Delta T) \) which are shifted in phase by an unknown amount \( \Delta T \):

\[
e^n(mT_s) = y^n(mT_s + \Delta T) + n(mT_s) - y^{\text{REF}}(mT_s). \tag{5.2}
\]

A sampled noise source \( n(mT_s) \) is assumed with zero average. Using a first order Taylor approximation (which is allowed when the delay between the rows is sufficiently small) we obtain:

\[
e^n_m = y^n_m + \Delta T \left( \frac{\partial y}{\partial t} \right)_m + n_m - y^{\text{REF}}_m = \Delta T \left( \frac{\partial y}{\partial t} \right)_m + n_m. \tag{5.3}
\]

For notational convenience the subscript \( m \) is used to indicate the time instant \( mT_s \). The second equality is only valid when \( y^n_m = y^{\text{REF}}_m \) meaning that the algorithm requires equal data in each of the rows at least for a particular part of the broad spiral. This is guaranteed by the fact that the loop is only activated during the preamble (the so called acquisition window). The error vector is now mapped on the ‘delay space’ by correlating the error sequence with the derivative of the reference signal itself. The derivative is a so-called signature signal for the delay:

\[
\chi_m = e_m \left( \frac{\partial y}{\partial t} \right)_m = \Delta T \left( \frac{\partial y}{\partial t} \right)_m^2 + n_m \left( \frac{\partial y}{\partial t} \right)_m. \tag{5.4}
\]
Upon integrating this phase error in time (using an integrating loop filter) the right term in Eq. (5.4) averages to zero while the left term results in a measure for the relative delay $\Delta T$ times a constant factor equal to the power $P(\frac{\partial y}{\partial t})$ in the derivative of the signal:

$$\chi_m = \Delta T P(\frac{\partial y}{\partial t}).$$  \hspace{1cm} (5.5)

A measure for the derivative of the signal can be obtained by the following difference operator:

$$\left(\frac{\partial y}{\partial t}\right)_{m-1} \propto y_m - y_{m-2},$$  \hspace{1cm} (5.6)

which is denoted as $1 - D^2$ with $D$ a unit delay. The delay of the difference operator is $D$ and for compensation of this delay all error signals are delayed by this value before correlating them with the derivative of the reference signal. The correlation is implemented as a sample-by-sample multiplication. The output is the required measure for relative phase delay. It is input to a first order loop filter consisting of an integrator in series with a programmable gain. The actual delay function is implemented as a polyphase filter structure in combination with linear interpolation in the same way as is done for the SRC (see Section 5.4.4). However, in this case the delay function is combined with a noise - and anti-aliasing function that is required before the reduction in sample rate by a factor of 2.

**Reference Row Selection**  At a first glance it seems efficient to take row 1 as a reference row and delay row 2 to $N$ with respect to this row. However, row 1 is a boundary row and suffers only from ISI from the inner rows since at the other side the guard band is present. On the contrary the inner rows suffer from ISI from both adjacent rows. In adjacent rows the preamble carrier, consisting of a repetition of the '1100' pattern, is shifted by half a bit-interval caused by the hexagonal lattice. Therefore, in a particular inner row we can write the signal as follows:

$$y(t) = A\sin(\omega t) + \alpha A\sin(\omega t + \varphi) + \alpha A\sin(\omega t - \varphi).$$  \hspace{1cm} (5.7)

Here it is assumed that the optical channel only passes the fundamental harmonic of the '1100' pattern resulting in a sinusoidal replay signal with amplitude $A$. Further, $\alpha$ is a measure for the strength of the ISI between adjacent rows and $\varphi$ is the phase shift of the signal due to the half-bit shift caused by the hexagonal lattice. Simplifying this equation leads to:

$$y(t) = A \left[1 + 2\alpha \cos(\varphi)\right] \sin(\omega t)$$  \hspace{1cm} (5.8)
showing that the ISI from adjacent top and bottom rows cancel out in terms of phase. This is not the case in the boundary rows of the broad spiral where a guard band is present on one side of the rows. This becomes more clear from a scalar diffraction simulation for both an inner row (solid) and a boundary row (dashed), see Fig. 5.10. A phase offset as large as 8% of the bit-period makes the boundary row unsuitable to act as a reference row for delay compensation. This consideration has led to the choice of the central row as the reference row.

![Figure 5.10: Bit-synchronous HF-samples at part of the preamble in bit-rows '0' and '1' (lattice parameter 165 nm; pit-hole diameter 120 nm).](image)

**Delay Calculation for Outer Rows**  The boundary rows are not only unsuitable to act as a reference row, but they also cannot be used to measure the relative delay of the data between the reference row and the boundary rows because of the problem discussed above. A solution was found in the fact that the spots, generated by the diffraction grating, are equidistant. This means that we can simply extrapolate the delay of the outer rows from the measured delays for the inner rows. The block diagram of the compensation block now changes to the one shown in Fig. 5.11.

**Lattice Distortion Measurement**  The delay compensation function can be used to estimate lattice distortion that is present due to a time varying relative phase between the rows. Such a phase variation is likely because of the multiple-pass mastering process as discussed in Section 2.4. An estimation of the peak-to-peak phase variations
can be derived by plotting the fractional delay vector derived in the delay compensation block as function of time. This is shown in Fig. 5.12 for replay signals from disc E490. The center row is the reference row and shows no variation. In the ideal case the other rows should have a constant delay that is linearly dependent on the row number. The fact that the delays are not constant is a clear indication of lattice distortion due to relative phase variations between the rows. Peak-to-peak variations of about 0.2 sample periods can be observed, which is equal to about 8% of a channel bit period.

5.4 Adaptation Loops

Adaptation of the receiver is necessary in order to deal with variations of the channel parameters. On the one hand we would like to compensate for these variations as fast
as possible to make sure the variations are adequately tracked. On the other hand, fast adaptation loops will lead to worse steady state performance due to gradient noise and therefore it is preferable to adapt with no more bandwidth than is strictly needed to capture the bulk of the variations. For the same reason it is preferable to adapt not more parameters than strictly needed. The receiver is designed according to this reasoning: a large and slow adaptive equalizer corrects for the overall channel variations, while a few important and fast-varying parameters are identified that are tracked by dedicated, fast control loops.

### 5.4.1 2D Adaptive Equalization

The 2D Inter-Symbol Interference (ISI) in the TwoDOS readout system extends over a large area. Furthermore, it is generally time-varying (for example due to a time-varying tilt). An adaptive partial-response (PR) equalization is used to transform the channel response to a shorter response with a controlled amount of ISI (the so-called target response). The 2D equalizer is implemented as a set of 2D hexagonal filters, one 2D filter per row:

\[
y_{l,n} = \sum_{i=-L}^{L} \sum_{j=-W}^{W} w_{i,j}^n \cdot x_{l-i,n-j},
\]  

(5.9)
where $x$ is the input signal, $y$ is the output signal, and $L$ and $W$ indicate the length and width of the 2D equalizer, respectively. Note that the equalizer is linear. Also non-linear 2D equalizers have been proposed [60]. Generally, these lead to improved performance in terms of bER at the output of the total system, at cost of a higher implementation complexity. Non-linear equalizers are not further discussed in this thesis.

**Size of the Equalizer** To determine the required size of the equalizer the ISI coefficients $C_s$ are calculated as function of the distance from the central bit using the scalar diffraction model of chapter 3 with a lattice parameter $a_l = 138$ nm and a pit radius $b = 51$ nm. The coefficients are indicated with the shell index $s$, where a shell includes all bits having the same distance to the central bit, i.e. obeying the property that $i^2 + j^2 - ij$ is constant. The results for different values of tangential tilt (0 and 0.5°) indicated by the Seidel\(^1\) coefficient $W_{31}$ are shown in Fig. 5.13. The figure shows in particular that the 4\(^{th}\) and 5\(^{th}\) ISI coefficient are not negligible because they account for 6% of $C_0$ at 0.5° tilt. Moreover, the number of bits in subsequent shells is distributed as: \{1, 6, 6, 6, 12, 6, 6, 12\} (for the first 7 shells), and unfortunately the 4\(^{th}\) shell includes more bits than most other shells.

![Figure 5.13: ISI coefficients as function of shell index (along-track projection).](image)

Especially for the boundary rows in a broad spiral a complete 2D equalization is not possible. Part of the equalizer will extend over the guard band to the next broad spiral, which is asynchronous to the broad spiral subject to read-out. In these

\(^1\)Seidel coefficients give a weight to the so-called Seidel aberration polynomials, which describe optical aberrations similarly to the Zernike polynomials mentioned in Chapter 3.
cases the corresponding taps are taken zero and are not adapted. This means that equalization performance decreases.

**Choice of Target Response for Equalization** The ISI after equalization is handled by a 2D maximum likelihood detector in the form of a stripe-wise Viterbi detector (see Chapter 6). The combination of a partial response and a maximum likelihood detector is referred to as a PRML (partial-response maximum-likelihood) system. To limit the complexity of the detector the target response is chosen as short as possible. Furthermore, the short response, having controlled ISI, is chosen such that noise enhancement is limited. At the same time, it is preferably chosen such that the noise power spectral density at the input of the Viterbi detector is white. The actual target response will be a trade-off between these criteria (see also Fig. 6.17). Once the target response is chosen the equalizer is updated based on a suitable criterion in order to transform the actual channel response into the target response as well as possible.

**Position of the Equalizer** The equalizer may be placed completely in the synchronous domain by cascading it with the other control loops as is schematically indicated in Fig. 5.14. A big disadvantage of this topology, however, is that the timing recovery does not profit from the benefits of adaptive equalization. Since in most cases the timing recovery is the most vulnerable link in high-density receiver systems this is not a viable option. Moreover, the timing error detector (TED) must operate in a non-data-aided mode, which becomes difficult for high-density storage systems.

![Figure 5.14](image-url)  
**Figure 5.14:** Receiver topology with a synchronous adaptive equalizer. For simplicity only the SRC is included as one of the other adaptive control loops and also the delay compensation has been omitted.

An alternative is to place the equalizer filter itself in the timing recovery loop (see Fig. 5.15). Now the timing recovery benefits from the adaptive equalization of the replay waveform. However, the latency of the equalizing filter now contributes to the loop-delay of the high-bandwidth timing recovery loop. This will adversely affect the stability and convergence speed of this loop. Therefore, in the chosen configuration
Figure 5.15: Receiver topology where the equalizing filter is part of the timing recovery loop such that this loop benefits from the adaptive equalization of the signal.

the equalizing filter is placed in front of all the other adaptive control loops including the timing recovery loop. It is located in the domain asynchronous to the bit-clock (see Fig. 5.16). This has the advantage that the latency of the equalizing filter does not contribute to the loop delay of the other, generally faster, control loops. The equalizer adaptation itself is rather slow since a large number of filter taps need to be updated simultaneously. Therefore, the total propagation delay of the other signal processing blocks does not lead to a significant limitation in the loop bandwidth of the adaptive equalizer. Another advantage of the asynchronous adaptive equalizer may be the lower speed at which processing takes place, allowing us to choose fewer equalizer taps. Note, however, that this is only true for a system with a considerable amount of negative excess bandwidth as for example the case in the RLL coded CD and DVD data formats. In our case with uncoded data on the hexagonal format the processing rate is higher in the asynchronous domain and this advantage disappears.

Figure 5.16: Receiver topology with an asynchronous adaptive equalizer.
A disadvantage of the asynchronous adaptive equalizer is the increase in complexity of the total loop due the need for an inverse sample rate converter (ISRC) to translate the error from the synchronous to the asynchronous domain. Furthermore, in case the data rate varies (for example in a constant angular velocity system) and the sampling rate of the ADCs is kept fixed the ratio between the clock period in the synchronous and asynchronous domain will vary, causing a data-rate-dependent performance of the equalizing filter. A commonly chosen solution is to insert an additional sample rate converter into the system in front of the equalizing filter (see Fig. 5.17). This SRC is preset to a ratio $R$ that can be derived for example from the wobble of the data track in case of recordable or rewriteable formats. For the ROM format no wobble is present and methods such as measuring the average run-length is possible to derive the SRC preset value. The ratio between $T$ and $RT_s$ is then fixed with a deviation of only a few percent. In the TwoDOS test format no wobble is present and the topology of Fig. 5.16 is used. Therefore, the rotational velocity of the disc is tuned such that the ratio between the sampling period and the bit-period is nearly constant (at a value of 2.5). There are, however, proposals for a TwoDOS wobble format. Some examples are indicated in Fig. 5.18. The left part of the figure shows a wobbled broad spiral with a constant angular frequency (CAF), while the right part of the figure shows a wobble with a constant linear frequency (CLF). The CLF format is preferred because it leads to a constant areal density on the disc. To achieve the same property in case of a CAF format the disc should be organized in zones, where each zone has its own constant angular wobble frequency.

**Equalizer Adaptation** Two adaptation methods are considered for updating the taps of the equalizer. The first one is the Zero Forcing (ZF) adaptation, which forces...
all residual ISI to zero. The second one is using the least mean square (LMS) algorithm to obtain a Minimum Mean Square Error (MMSE). Only the ZF method is discussed briefly. The equalizer update for a 2D system is not principally different compared to conventional 1D systems. Note that the ZF and LMS methods do not directly minimize the bit-error rate (bER). Recently methods using the sequence amplitude modulation (SAM) criterion for Viterbi detectors attempt to directly minimize bER. This allows a further performance improvement compared to the ZF or LMS methods [61, 62]. More information on SAM can be found in Section 6.5.

**Zero Forcing Adaptation** In the ZF adaptation scheme all residual ISI is forced to zero. This means that the input signal of the detector should ideally be a noisy version of the reference signal. The structure of the ZF adaptation loop is shown in Fig. 5.19. It is often chosen for its relative simplicity. The desired (or reference) signal $d_\Lambda$ is generated by passing the channel bits through a target impulse response filter:

$$d_\Lambda = (g * \hat{a})_\Lambda \quad (5.10)$$

with ‘*’ denoting the 2D convolution, and $\Lambda = (k, n)$ with $k$ the time index and $n$ the row number. The result is compared to the actual signal $\tilde{d}_\Lambda$ to derive the error signal $e_\Lambda$ according to:

$$e_\Lambda = \tilde{d}_\Lambda - d_\Lambda. \quad (5.11)$$

The desired signal is also input to a shift registers (SHR). The output of the SHR is a vector of desired signal samples. One vector for each row. Therefore, the SHR output
5.4 Adaptation Loops

Figure 5.19: Zero forcing equalizer adaptation structure. For simplicity the timing recovery loop has been omitted from the drawing.

is indicated with a double arrow. The desired vector is then spatially interpolated (SI). This is needed because the desired signal samples are spaced at the bit-period $T$, while the equalizer taps are spaced at the double sampling period $2T_s$, i.e:

$$w = w_{2iTS,j} \quad i \in \{-L, -L + 1, \ldots, L\}, j \in \{-W, -W + 1, \ldots, W\}. \quad (5.12)$$

The number of equalizer taps is equal to $N = (2L + 1) \cdot (2W + 1)$. The spatial interpolator is needed to calculate the samples at the correct spacing in the $i$-direction. Along the other direction both the equalizer taps and the desired samples are simply spaced according to the row-distance and no spatial interpolation is required. The interpolation is schematically shown in Fig. 5.20. The output of the SI block, which is denoted as the signature vector $s_A$ is correlated with the error signal to extract that part of the error signal that is due to mis-equalization:

$$\Delta_A = e_A s_A. \quad (5.13)$$

The resulting vector is input to a set of loop filters (LF). One for each equalizer tap. Note that the LF outputs are still sampled at the bit-period, while the equalizing filter and its tap-update are performed each sampling period. The required time-base conversion is done by a temporal interpolation (TI) function. Since the taps are varying only slowly the temporal interpolation can be implemented as a simple zero-order hold function. Note that the whole scheme is not principally different from its 1D counterpart. When the loop is in its steady-state mode of operation, i.e. after convergence of the (stable) loop, the tap values of the equalizer will have finite values $w_{iTS,j}$. This means that the integrator inputs must be zero on average and that the loop will force the following expectation to zero:

$$E[\Delta_A] = E[e_A s_A] \quad (5.14)$$

hence the name zero forcing equalization.
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Figure 5.20: Spatial interpolation of samples when going from the synchronous domain sampled at bit-period $T$ to the asynchronous domain sampled at period $2T_s$.

**Tap Leakage** In case ideal integrators are used, the update rule for ZF-adaptation can be written as:

$$w_{k+1} = w_k + \mu e$$  \hspace{1cm} (5.15)

This update rule has the problem that the loop will become unstable in case the channel has spectral nulls, i.e. frequency ranges with zero transfer, while the target is non-zero. In its attempt to force the ISI to zero the frequency transfer function of the equalizer will blow up to large values. A solution that is used in the TwoDOS receiver is the introduction of leakage. The tap update is now modified according to:

$$w_{k+1} = (1 - \alpha)w_k + \mu es$$  \hspace{1cm} (5.16)

with $\alpha$ denoting the leakage factor. When the error is zero the equalizer taps slowly leak to zero. This means that for frequencies where the channel has spectral nulls the equalizer transfer will not blow up to large values. In Fig. 5.21 the equalizer transfer function is plotted in a 2D representation (left) and a 1D representation (right) where $x$ is the along-track direction and $y$ is the radial direction. The plots are normalized with respect to the bit frequency in both directions. In case no leakage is applied the equalizer transfer becomes large at high frequencies where the transfer of the channel is low or zero. This leads to excessive noise enhancement. Application of a small leakage ($\alpha = 0.025$) improves bit-detection performance considerably. The equalizer transfer for the minimum $bER$ is shown in Fig. 5.22. As input a replay signal from disc E490 is taken at nominal situation (no tilt).
5.4 Adaptation Loops

Figure 5.21: Transfer of the ZF adaptive equalizer without any leakage. The frequency scales are normalized with respect to the bit-period.

Figure 5.22: Transfer of the ZF adaptive equalizer using a leakage value of $\alpha = 0.025$. The frequency scales are normalized with respect to the bit-period.

Interaction between Control Loops  Note that there might be interaction with other control loops. For example, if all equalizer taps are chosen twice as large and the gain factor from the automatic gain control (AGC) is chosen half its normal value, then the total signal amplitude after adaptive equalizer and AGC will be the same. This means that both control loops might drift in opposite directions and that this will not be noticed by the error detection algorithms of both loops. To avoid such an interaction between the two loops we simply set the central tap of the equalizer to a fixed non-zero and positive value (equal to 1 in our case).
5.4.2 DC Control

The detector used in the TwoDOS receiver is a 2D Viterbi detector (see Chapter 6). The Viterbi algorithm evaluates a number of possible bit-sequences by comparing the actual received signal with a reference signal. The output of the detector is the bit-sequence that has the best ‘fit’ according to the Euclidian distance criterion (see Eq. (6.6)). The reference signal is calculated by passing the bit-sequence under evaluation through the target response $g_k$. In case the target response is DC-free, i.e.:

$$\sum_k g_k = 0, \quad (5.17)$$

a change in the bit-sequence under evaluation does not change the DC-level of the reference signal. This means that a change in DC-value of the input signal has no influence on the outcome of the detection process. The Viterbi detector using a DC-free target is insensitive to DC offsets. Examples are the $1 - D$ and $1 - D^2$ 1D detectors [59].

In our case, however, the target response is not DC-free. According to the above reasoning the bit-detection becomes sensitive to variations in the DC (an example is the $1 + D$ 1D Viterbi detector). There are several effects that may cause a time-varying DC-offset:

- DC offsets in the analogue or mixed-signal front-end of the receiver. Examples are offsets in the PDIC amplifier stages and the ADCs. These offsets might vary slowly as function of temperature, or due to ageing or supply voltage variations. Part of the offsets can be filtered out by a simple AC-coupling (via a series capacitor) in front of the ADC.

- baseline wander. In case the signal is AC-coupled (to eliminate DC offsets in the analogue front-end) a DC-content of the signal is not passed by the channel. This means that if the data is not DC-free a time-varying DC error (low frequency ISI component) is present at the input of the detector. A solution to this problem is to insert DC-control information in the bit-stream (see Section 4.3.4). By bounding the running digital sum (RDS) the low-frequency content of the signal is suppressed and baseline wander is limited to a large extend.

- variations in the optical path. Reflection variations of the disc (e.g. due to fingerprints, scratches or material inhomogeneities) may cause a large drop in the signal. This generally causes a fast variation in the DC content of the signal. A fast DC restoration loop is needed to compensate for these variations.

It is clear from the above list that many of these effects are common for all rows in the broad spiral. The analogue front-ends for the different rows in the broad spiral
are generally integrated in a single IC. They are subject to a common supply voltage or temperature variation. Additionally, fingerprints and scratches on the disc affect a large 2D area (due to the substrate incident read-out of the disc), which means that all rows in the broad spiral show simultaneous signal drops. This observation influences our adaptation strategy considerably (see Section 5.5) and leads to a configuration of coupled DC compensation loops. In this section we limit ourselves to the derivation of the behavior of independent DC loops. We use the block diagram of Fig. 5.23. No row index $n$ is used. The channel is considered ideal except for some additive noise with zero mean and a (time-varying) DC offset $o_k$. To compensate for the offset in the output signal $r_k = (a \ast g)_k + n_k + o_k$ a correction signal $c_k$ is subtracted. The resulting signal $\tilde{d}_k = (a \ast g)_k + n_k + o_k - c_k$ is compared to the desired signal $d_k = (a \ast g)_k$ resulting in an error $e_k = n_k + o_k - c_k$. Generally, the error signal is correlated with a signature signal $s_k$ which is ideally equal to the error $e_k$ that occurs in presence of the particular parameter that is subject to estimation. In case of a DC offset the signature is simply a DC signal itself and can be taken equal to 1 for reasons of simplicity. In that case the real-time, estimated DC offset is equal to:

$$\eta_k = n_k + o_k - c_k \quad (5.18)$$

The noise with zero mean drops from this equation upon averaging over time. Hence the expectation $E[\eta_k] = o_k - c_k$, will reduce to zero in case $c_k$ will reach its optimum value $c^*_k = o_k$. If this would not be the case the integrator would accumulate the residual signal (multiplied by some proportional gain factor $K_p$) in order to force the residual DC offset to zero (hence the name Zero Forcing (ZF) adaptation).

**Dynamic Loop Behavior** From Eq. 5.18 we can observe that the actual compensation value $c_k$ is compared with its optimum value or setpoint $c^*_k = o_k$. This simplifies the block diagram of Fig. 5.23 to the basic control block diagram of Fig. 5.24. The resulting error signal has a noise component $n_k$ which is injected inside in the control loop. For the discrete-time transfer functions in the $z$-transformed domain this

\[ a_k h_k = g_k \quad + \quad \Sigma \quad + \quad \eta_k \quad z^{-1} \quad \Sigma \quad + \quad g_k \quad \hat{a}_k = a_k \quad \text{to detector} \]
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Figure 5.24: Simplified diagram to analyse the dynamic behavior of the DC control loop.

means:

$$G_e(z) = \frac{E(z)}{C^*(z)} = \frac{z-1}{z-1+K_p};$$

$$C(z) = \frac{E(z)}{C^*(z)} = \frac{K_p}{z-1+K_p};$$

(5.19)

with $G_p(z)$ and $G_e(z)$ the transfer functions from an input excitation $c^*$ to the error $e_k$ and to the compensation value $c_k$. Furthermore, $H(z)$, $E(z)$, $C(z)$ and $C^*(z)$ are the $z$-transform of $\eta$, $e$, $c$, and $c^*$ respectively. In order to assess the dynamic behavior of the loop a unit step function $u_k$ (with $z$-transform $U(z) = \frac{z}{z-1}$ for $|z| < \infty$) is applied to each of the transfer functions $G_e(z)$ and $G_c(z)$. The steady state situation is derived according to:

$$\lim_{k \to \infty} e_k \iff \lim_{z \to 1} E(z) = \lim_{z \to 1} \frac{z}{z-1} G_e(z) = 0$$

$$\lim_{k \to \infty} c_k \iff \lim_{z \to 1} C(z) = \lim_{z \to 1} \frac{z}{z-1} G_c(z) = 1.$$  (5.20)

As expected from this first order loop the compensation value $c_k$ will converge to unity in the steady state situation. The resulting error $e_k$ is zero in that case. The same conclusion can also be derived from the transient output signal after a step response:

$$c_k^{\text{step}} = u_{k-1} - (1-K_p)^k u_{k-1}. $$  (5.21)

For small values of $K_p$ the time constant is equal to $\tau \approx \frac{1}{K_p}$. Note that in Fig. 5.23 it is assumed that the bits $\hat{a}_k$ are present without any delay. In reality the situation is more complex. The 2D Viterbi detector has considerable detection delay, especially for the inner rows of the broad spiral. The consequences of this delay for the dynamic behavior of the loop are discussed in more detail in Section 5.5.
5.4.3 Gain Adaptation

Another time-varying parameter in the system is the total gain of the channel. Several effects may cause a time-varying gain:

- gain variation in the analogue front end. Due to temperature variations, ageing or supply voltage variations the gain of the PDIC amplifier stages or the ADCs may change.

- variations in the optical path due to fingerprints and scratches. In the TwoDOS case also spot-to-spot variations in the diffraction efficiency of the grating (see Fig. 2.5) might contribute to the gain differences between the rows.

A similar approach as in the case of DC adaptation can be followed. Also here the fact that gain variations are expected to occur simultaneously for all rows can be used in the final adaptation scheme. This will be discussed in Section 5.5. Here, a single independent adaptation loop will be discussed. The block diagram of the loop is shown in Fig. 5.25. One can verify that $\eta_k = d_k^2 \lambda_k (c_k - \frac{1}{\lambda_k}) + d_k \lambda_k c_k n_k$ [59]. The corresponding control block diagram is given in Fig. 5.26. Here, the optimum value or setpoint is equal to $c_k^* = \frac{1}{\lambda_k}$. The result is a first order control loop similar to the DC control loop. The main difference is that the loop gain depends on the gain of the channel. A small signal at the receiver input makes the loop slow, while a large input signal leads to a fast loop. It even might lead to instabilities in the loop. This can be solved by a so-called exponential gain control [59]. In the TwoDOS system an exponential gain control is implemented, although the input signal is properly scaled to the input range of the ADC and a linear gain control loop gives sufficiently stable loop parameters.

5.4.4 Timing Recovery

Timing recovery for the 2D system does not seem principally different from timing recovery in a 1D system. One could simply instantiate independent timing recovery...
The reciprocal lattice of the hexagonal lattice is again a hexagonal lattice (see Section 3.2). A simple instantiation of 1D sample rate converters results in a rectangular grid in the real space and in the reciprocal frequency space. This means that by using 1D sample rate converters aliasing can occur that could be avoided by using 2D filtering in the sample rate converter.

The frequency (and ideally also the phase) of all rows in the broad spiral is equal. We can use this property to our advantage by deriving the frequency error based on all rows simultaneously. This makes the system more robust (note that generally timing recovery is the 'Achilles heel' of the system). Furthermore, the loop bandwidth may be chosen larger (relative to the bit frequency) because the total timing error signal is averaged over more rows and causes less gradient noise. The joint derivation of the frequency error also helps in dealing with the large detection delay in the timing-recovery loop due to the 2D stripe-wise Viterbi detector.

At each instant the sample rate converters calculate a new set of output samples that are input to the detector. This is done by interpolation between input samples. These input samples are taken in parallel at each instant of the ADC clock. Because phase might differ slightly between the rows in the broad spiral the SRCs do not skip or add samples simultaneously. Some buffering of samples is needed in the SRC to avoid a relative shift between the data at the output of the receiver.

We will briefly discuss the principle of timing recovery and show the implementation of the sampling rate conversion. In this discussion specific attention is paid to
the above mentioned issues that are specific for 2D storage.

**Principle of Timing Recovery** In optical recording it is customary to transmit clock information as part of the data itself (the so-called self-timing property of the data). Upon reception of the replay waveform it is necessary to extract this clock signal again to determine the correction decision instants. Timing recovery in the TwoDOS system is done in the form of a phase-locked loop (PLL). A block diagram of the loop is shown in Fig. 5.27. In this loop a sample rate converter (SRC) is used to resample the data at the baud-rate with a correct phase. The phase value is determined using a numerically controlled oscillator (NCO) in the form of an integrator. At the input of the NCO the (instantaneous) frequency value is present, which is generated by a loop filter (LF). The output of the SRC is used as input for a bit detector and for a timing error detector (TED). The TED is of the data-aided (DA) type and uses the data sequence $a_k$ when this is available in the form of the preamble (see Section 4.4). When the preamble ends it is assumed that the PLL has converged to a point such that bit detection becomes sufficiently reliable. At that moment the input of the TED is switched to the detected bit sequence $\hat{a}_k$ and the PLL is operating in a decision-directed (DD) mode. In the DD-mode a large detection delay is present in the loop and optimization of loop parameters is needed to guarantee stability.

![Figure 5.27: Block diagram of the phase locked loop as used in the TwoDOS system.](image)

We will start with a brief dynamic behavior analysis of a single second order timing recovery loop without any detection delay. Later we will consider a configuration with coupled second order loops. The single second order loop has a proportional branch with weighing factor $K_p$ that assures an average steady-state phase error equal to zero and an integrating branch with weighing factor $K_i$, which lets the loop converge towards the correct frequency value. The total PLL loop with loop filter, and NCO is shown in Fig. 5.28. From this block diagram we can derive the total transfer function $G_\psi(z)$:

$$G_\psi(z) = \frac{\Psi(z)}{\Phi(z)} = \frac{K_p(z-1) + K_i}{(z-1)^2 + K_p(z-1) + K_i} \quad (5.22)$$
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Figure 5.28: Discrete-time model of the second order PLL in the phase domain.

A second order continuous time PLL is generally written in the form:

\[ G_c(\omega) = \frac{(j\omega)2\zeta\omega_n + \omega_n^2}{(j\omega)^2 + (j\omega)2\zeta\omega_n + \omega_n^2} \]  

Here \( \omega_n \) is the natural frequency and \( \zeta \) is the damping factor. By comparing Eq. 5.23 and Eq. 5.22 and by replacing \( z^{-1} \) by \( j\omega T \) (under the assumption that \( \omega T \ll 1 \)) we can compare the discrete-time PLL transfer function with the continuous-time one. For the natural frequency \( \omega_n \) and damping factor \( \zeta \) one can derive that:

\[ \omega_n T = \sqrt{K_i}, \quad \zeta = \frac{K_p}{2\sqrt{K_i}}. \]  

1D versus 2D SRC filter

A first issue that must be dealt with in case of 2D optical recording is whether or not a full 2D SRC is needed. In principle, we only need to deal with timing variations (and not with radial offset variations e.g. due to track pitch variation. The compensation of these variations is left to the adaptive equalizer). We have chosen to use \( N_r \) 1D SRCs in parallel and to process each row individually. In that case the corresponding 2D reciprocal space can be divided into vertical strips with a width equal to \( \frac{1}{a_H} \), see Fig. 5.29. The fact that the circle denoting the cutoff of the optical channel is crossing the Nyquist-line at \( \frac{1}{2a_H} \) means that, without proper pre-filtering, some aliasing of data components is introduced in the tangential direction. From Fig. 5.29 One can see that this is not the case when applying a full 2D hexagonal SRC filter.

Due to the shape of the MTF the effect of aliasing is not large and it appears in this specific case that applying \( N_r \) 1D SRCs does not deteriorate performance (in terms of bit error rate) compared to using full 2D hexagonal SRC filters.

1D Sampling Rate Conversion

We will now focus on the 1D sample rate converter. It must be able to convert the input signal sampled at a fixed frequency \( \frac{1}{2T_r} \) to...
5.4 Adaptation Loops

a frequency equal to the baud-rate $\frac{1}{T}$. In general the ratio between $\frac{1}{2T_s}$ and $\frac{1}{T}$ is a fractional number. Moreover, this second frequency is subject to small variations caused by several phenomena like eccentricity of the disc, motor-speed variations, variations in mastering/writing speed, etc, which cause this fractional number to vary over a small range. We will first derive an efficient sampling rate converter for rational factors and we will then use it in such a way that fractional sampling rate conversion is achieved. The basic circuit [63] for rational sampling rate conversion is shown in Fig. 5.30. The sampling frequency can be changed by a rational factor $\frac{K}{L}$. Increasing

the sampling rate is done by insertion of $(K - 1)$ zeros in between the input samples and filtering the resulting sequence by an interpolation filter $H(z)$. The low pass filter $H(z)$ will filter out the baseband signal from the repeated spectrum which was caused by insertion of the zeros. The second function of $H(z)$ is to act as an anti-aliasing filter for the subsequent decrease in sampling rate by a factor $L$. The simplest possible filter for $H(z)$ is a linear interpolation filter. However, this leads to large errors as indicated in Fig. 5.31. In the time domain this is clearly visible by the fact that the interpolated sample indicated by the cross is not on the solid line representing the underlying analog signal. From the frequency domain in the right figure one can see that the repeated spectra are not suppressed completely. For this reason a more complex interpolation filter is required. For optical recording the accuracy of the output sample phase must be such that no timing jitter is added to the system. Therefore, the value of $K$ is large resulting in a large number of repeated spectra and a steep and thus extremely complex filter. To reduce this complexity a practical structure is devised
that combines a first interpolation stage with modest complexity followed by simple linear interpolation. Such a structure is shown in Fig. 5.32. The spectra of the signals in this practical structure are shown in Fig. 5.33. The interpolating filter $H_0(z)$ removes the repeated spectra caused by zero insertion in the sample rate increaser. Due to the fact that the TwoDOS system has some excess bandwidth it is unavoidable to introduce aliasing unless we apply a steep filter that not only filters out the repeated spectra but also a small part of the signal content in the channel.

**Implementation of the SRC** The practical design has been split into multiple stages as indicated in Fig. 5.32. Nevertheless, the interpolating filter $H_0(z)$ is still a complex filter with a large length. For the SRC it is very important to have a small latency since it is part of the other control loops such as DC, gain and adaptive equalizer (note that the length of the filter does not necessarily add to the delay in the timing recovery loop itself). Therefore, the latency of the interpolating filter must be as small as possible. A polyphase implementation has been chosen to reach this goal. The polyphase structure also allows an easy implementation of 2D filtering in the SRC. Here, the original finite impulse response (FIR) filter $H_0(z)$ is decomposed
As an example the original filter is decomposed in three components. This decomposition in the so-called first polyphase structure is shown in Fig. 5.34. Note that when a sample rate increaser is put in front of this structure that each filter will run at the increased sampling frequency and will process input samples that are zero for most of the time. This means that although the latency has been minimized the implementation is still not very efficient. A second and much more efficient polyphase structure for our purpose can be obtained by applying the transposition theorem on the block diagram of Fig. 5.34. This means that the signal flow is reversed and all nodes are interchanged with adders [64]. The result is shown in Fig. 5.35 together with the sampling rate increaser in front of the filter. Due to the fact that the filters are now written in terms of $z^3$ the output of the filters will be non-zero only once in 3 samples at the moment that the non-zero output samples of the sample rate increaser coincide with the non-zero filter taps. Therefore, it is allowed to exchange the order

\begin{align}
H(z) &= \sum_{i=0}^{N} b_i z^{-i} = b_0 + b_1 z^{-1} + b_2 z^{-2} + \ldots \\
&= (b_0 + b_3 z^{-3} + b_6 z^{-6} + \ldots) + z^{-1} (b_1 + b_4 z^{-3} + b_7 z^{-6} + \ldots) \\
&\quad + z^{-2} (b_2 + b_5 z^{-3} + b_8 z^{-6} + \ldots) \\
&= H_{3,0}(z^3) + z^{-1} H_{3,1}(z^3) + z^{-2} H_{3,2}(z^3) = \sum_{j=0}^{2} z^{-j} H_{3,j}(z^3)
\end{align}
of the sample rate increase and the filters. This improves the efficiency of the filters considerably since the processing frequency is lowered by a factor equal to interpolation ratio compared to the initial situation. A further reorganization of the delay elements leads to the final structure of Fig. 5.36.

**Polyphase Filter Design** As a basis for the design the sampling frequency is taken to be $f_s \approx \frac{2.5}{T}$. The sampling rate at the input of the SRC is (after downsampling
by a factor of 2 in the delay compensation block) equal to $\frac{1.25}{T}$. The required stop-band attenuation is assumed to be 40 dB and the pass-band ripple cannot exceed 0.1 dB. It appears that 8 polyphase components are required to meet these specifications in combination with the linear interpolation that follows the polyphase components [65]. The situation in the frequency domain is sketched in Fig. 5.37. The cut-off of the optical channel is at 23.8 MHz and the sampling frequency is equal to 102.5 MHz (for a bit time of 24.35 ns, see Table 5.1).
Practical Filter Implementation  When implementing a finite impulse response (FIR) filter one normally needs a number of multipliers equal to the length of the filter (or more accurate the number of non-zero coefficients). A more efficient implementation of the FIR-filters is possible when the coefficients are forced to a so called Canonic Signed Digits (CSD) representation. This representation is particularly efficient because a multiplier can be implemented in hardware as a number of adders and subtractions. A redundant notation is possible that allows us to select the notation with minimum implementation complexity. Each filter coefficient can be written as the sum of a limited number of base-2 contributions wherein each CSD coefficient can have three values \{−1,0,1\} (instead of two in the binary representation):

\[ c_i = \sum_{i=1}^{m} \gamma_i 2^{-i} \quad \text{with } \gamma_i \in \{-1, 0, 1\} \]  

(5.29)

The redundancy in this system is clear from the fact that particular numbers can be represented in more than one way. For example 31 = 32 − 1 or 31 = 1 + 2 + 4 + 8 + 16. In the first case only one subtraction operation is needed while in the second case 4 additions are needed. It was derived by Reitwiesner [66] that the CSD notation offers the minimum number of non-zero coefficients (and thus the minimum number of operations) by requiring that:

\[ \gamma_i \gamma_{i+1} = 0 \quad \forall \ i, \]  

(5.30)

which states that between each two non-zero coefficients there has to be at least one zero coefficient. Using a linear programming (LP) algorithm (DESFIL software) it is found that the total SRC interpolation filter (composed of 8 polyphase components) can be implemented with 144 non-zero coefficients. The total number of coefficients is 80 and the ripple scale constant is 0.97 (meaning that the realized ripple is 0.97 times smaller than the required ripple). To make the design practical the initial filter requirements are relaxed somewhat to a stop-band starting at 35 Mhz (instead of the initial 27.25 Mhz). It appears that complexity increases sharply for frequencies lower than this value. The realized amplitude transfer function of the filter designed with the LP method with CSD constraints is given in Fig. 5.38.

Linear Interpolation

For linear interpolation the number of input samples required is only 2. So at each instant only two subsequent polyphase filter outputs need to be selected to serve as input to the linear interpolation function. This selection can be done with multiplexers as indicated in Fig. 5.39. Which outputs are selected depends on the phase of the sample that is required at the output. This has the advantage that a change in phase instantaneously leads to a change in output sample. No delay is present in the SRC. Note that there is a special case where we have to interpolate between the output
Figure 5.38: Transfer function of the interpolation filter in the SRC for a sampling frequency of 2.5 times the baud rate. Note that the stop-band starts at 35 MHz.

of the last polyphase filter at instant \( l \) and the first polyphase filter at instant \( l + 1 \). For this reason it is needed to store the polyphase filter outputs at least one clock cycle. This is indicated in the figure by the \( z^{-1} \) blocks. The output of the numerically controlled oscillator (NCO) at instant \( k \) denoted \( \Psi_k \), which determines the required phase of the sample at the output of the SRC is split into an integer component \( \Psi^i_k \) and a fractional component \( \Psi^f_k \). The integer component is used to select the input sample from a simple shift register running at the input clock \( \frac{1}{2T_s} \) (as will be explained in more detail in the continuation of this discussion). The fractional component \( \Psi^f_k \) is again split into two values. The first value \( \Phi^\text{select}_k \) indicates the multiplexer selection value (in fact it indicates the number of the first polyphase component that must be selected). It is found by multiplying the value \( \Phi^\text{select}_k \) by the number of polyphase filters \( P \) and taking the integer part of the resulting value. The remaining fractional value \( \Phi^\text{delta}_k \) is used as the linear interpolation coefficient. The whole scheme
can be described as follows:

$$
\begin{align*}
\Psi^i_k &= \lfloor \Psi_k \rfloor; \\
\Psi^f_k &= \Psi_k - \Psi^i_k; \\
\Psi^\text{select}_k &= \lfloor P\Psi^f_k \rfloor; \\
\Psi^\text{delta}_k &= \Psi^f_k - \Psi^\text{select}_k
\end{align*}
$$

(5.31)

with $P$ the number of polyphase filters.

Normally the NCO outputs are between 0 and 1 (i.e. $\Psi^i_k = 0$). In case overflow occurs (i.e. the NCO output value is larger than 1) the NCO value is decremented by 1 and a sample is skipped from the input sequence. In case of an underflow (i.e. the NCO output value is smaller than 0) the NCO value is incremented by 1 and an additional sample is calculated with the same set of input samples. Note that in the 2D case some NCOs might overflow, while other NCOs still have values smaller than 1. This means that for some rows a sample is skipped at a certain instant and for other rows this is not the case. Additional buffering of samples in a shift register is needed to be able to implement this row-selective skipping. In the practical system this is implemented by adding additional memory to the output of the polyphase filters. The NCO value is nominally kept between 1 and 2, but some NCOs may temporarily have values below 1 or above 2. The integer part $\Psi^i_k$ is used for selecting the tap in the
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shift register. A block diagram of this final scheme is shown in Fig. 5.40. Note that the 1D polyphase filters could easily be replaced by 2D polyphase filters followed by a simple 1D linear interpolation to approach a real 2D SRC filter.

![Block diagram of the final SRC implementation in the experimental set-up.](image)

Figure 5.40: Block diagram of the final SRC implementation in the experimental set-up.

Lattice Distortion

Time-varying lattice distortion occurs due to inaccuracies in the multiple-pass mastering process (see Section 2.4). When disabling the delay compensation loop the time-varying lattice distortion will be dealt with by the adaptive equalizer and by the timing recovery loop. The timing recovery loop takes the linear delay into account, while the adaptive equalizer compensates for the time varying 2D inter-symbol interference that results from time-varying lattice distortions. The time-varying lattice distortion can be easily derived by comparing the phase values of the NCOs of each row within the broad spiral. This is done for the LBR disc 12 (with an equivalent density of 50 GB). The result is shown in Fig. 5.41. It shows that the peak-to-peak delay variation between the rows is in the order of 10%, which is about equal to the value observed in the EBR discs (see Fig. 5.12).
5.5 Adaptation Strategy

In all adaptation loops discussed so far it has been implicitly assumed that the error signals for the various parameters are available instantaneously without any delay. In reality the derivation of error signals is done in a data aided mode. The data is needed to calculate the reference signals with which the actual signals at the input of the detector are compared. During the preamble (see Fig. 4.20) the data is known a priori without any detection delay. In the rest of the data frame, however, the data is only available at the output of the stripe-wise 2D Viterbi detector. A so called decision-directed scheme is used. It is a property of the stripe-wise detector that detection delay is low for the outer rows near the guard band, but increases rapidly for the inner rows (see Fig. 6.8). First, we will analyse the behavior of the loops in the presence of detection delay. We will see that large detection delay leads to instability of the loop or a lower achievable bandwidth. Then some alternative adaptation schemes are proposed that improve the performance of the 2D receiver in the presence of detection delay.

As a starting point of the discussion the control block diagram of the DC loop is taken (see Fig. 5.24). It is assumed that the loop incorporates a detection delay $M$. The open loop response is equal to:

$$H_{ol}(z) = \frac{K_{p}z^{-M}}{z-1}. \quad (5.32)$$
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The magnitude and phase of this response can be derived by setting $z = e^{j\omega T}$ [58]:

$$H_{ol} = |H_{ol}| e^{j\phi(\omega T)} = \frac{K_p}{2\sin \left( \omega \frac{T}{2} \right)} e^{-j(\omega T(M+0.5)+\frac{\pi}{2})}. \quad (5.33)$$

The phase margin (i.e. the phase difference from $-\pi$ at the frequency $\omega_o$ where $|H_{ol}| = 1$) can be derived from this equation as:

$$\phi_m(\omega_o T) = \frac{\pi}{2} - \omega_o T(M + 0.5) \quad (5.34)$$

with $\omega_o T = 2\sin \left( \frac{K_p}{2} \right) \cong K_p$ for $K_p \ll 2$. The corresponding closed loop transfer functions $G_e(z)$ and $G_c(z)$ are equal to (see also Eq. (5.19)):

$$G_e(z) = \frac{1}{1+H_{ol}} = \frac{z-1}{z-1+K_pz^{-M}},$$

$$G_c(z) = \frac{H_{ol}}{1+H_{ol}} = \frac{K_pz^{-M}}{z-1+K_pz^{-M}}. \quad (5.35)$$

The magnitude transfer of $G_e(z)$ (also denoted as mis-adjustment transfer function) is plotted in Fig. 5.42 as function of the normalized frequency $\Omega = \frac{\omega T}{2\pi}$. The detection delay $M$ is varied from zero to the value where the phase margin is zero ($M=79$ for $K_p = 0.02$, which is a very fast loop). It is clear that for increasing detection delay

![Figure 5.42: Mis-adjustment transfer function as function of the normalized frequency for $K_p = 0.02$.](image)

the damping in the loop decreases and that ultimately the loop becomes unstable.
Due to the resonance peak the loop will enhance rather than suppress variations in the frequency region of resonance. For a safe phase margin of 60° and a detection delay \( M = 48 \) for the center rows assuming a trace-back depth of 12 (see Fig. 6.14) and 4 subsequent 2-row processors on a 7-row broad spiral (see Fig. 6.8) the maximum value of \( K_p \) is \( \frac{\pi}{6(M+0.5)} \approx 0.011 \). Here, we assumed that decisions are taken from the first iteration and that no additional delay is present in the practical implementation of the detector. Especially, the DC loop appears to require a small time constant (in the order of 50 bit periods) to achieve optimum receiver performance in terms of bit-error rate. This means that for the DC loop in the TwoDOS receiver we already have problems for the situation with \( M = 48 \). Moreover, at high data densities it might be needed to increase the trace-back length or even take decisions from the second iteration. In all these cases the large detection delay becomes a problem and it would be convenient to have an adaptation scheme that does not suffer too much from this delay.

A solution is found by doing fast control on all the rows based on information from the outer rows only. This implicitly assumes that the fast variations are common to all the rows. When looking at the causes of time-varying DC offsets and time-varying gain (see Section 5.4.2) one indeed notices that many fast variations, such as fingerprints and scratches, are likely to be common to all rows. Some slow variations, however, such as baseline wander, are unique for each row in the broad spiral. These slow variations can be handled by correction loops that use delayed information from the inner rows. A block diagram of the configuration with coupled inner and outer loops is shown in Fig. 5.43. It consist of two standard first order loops, where the inner loop receives information from the outer loop. A different detection delay \( M_o \) and \( M_i \) for the outer and inner control loop is inserted in the model.

The response of the inner loop to variations in the DC setpoint can be calculated from the following transfer function:

\[
\frac{C_i(z)}{C^*(z)} = \frac{(K_{op}z^{-M_o} + K_{ip}z^{-M_i}) (z - 1) + K_{op}K_{ip}z^{-(M_o+M_i)}}{(z - 1 + K_{ip}z^{-M_i}) (z - 1 + K_{op}z^{-M_o})} \tag{5.36}
\]

For a first analysis we take a continuous time approximation by setting \((z - 1) = e^{j\omega T} - 1 \approx j\omega T \) for \( \omega T \ll 1 \) and by taking the delays equal to zero \((M_o = M_i = 0)\). The resulting transfer function is equal to [67]:

\[
\frac{C_i(j\omega)}{C^*(j\omega)} = \frac{j\omega(K_{op} + K_{ip}) + K_{op}K_{ip}}{(j\omega + K_{ip})(j\omega + K_{ip})} \tag{5.37}
\]

The corresponding step response (in the continuous domain) is:

\[
c_i(t) = 1 - \frac{K_{ip}}{K_{op} - K_{ip}} e^{-K_{ip}t} + \frac{K_{op}}{K_{op} - K_{ip}} e^{-K_{op}t}, \tag{5.38}
\]
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Figure 5.43: Block diagram of the inner-outer control loop configuration for the case of DC control.

which is plotted in Fig. 5.44 together with the normal first order response for the outer loops. Additionally, a simulation result based on experimental data is plotted that shows a perfect fit to the calculated response. The damping factor for the inner loops is equal to:

$$\zeta_i = \frac{1}{2} \left[ \sqrt{\frac{K_p^i}{K_p^o}} + \sqrt{\frac{K_p^o}{K_p^i}} \right]$$  \hspace{1cm} (5.39)

and reaches a minimum equal to 1 when $K_p^o = K_p^i$. This means that the coupled first order loops show a guaranteed stable behavior under the applied approximations. When the delays are not zero the situation changes. For example for a realistic scenario with $K_p^o = 0.04$, $K_p^i = 0.01$ and $M_o = 12$, $M_i = 50$ one obtains the magnitude and phase responses of Fig. 5.45.

Another way to verify the correct behavior of the inner outer configuration is by plotting the RMS error as function of the time constants of the loops for both the configuration with independent loops and the configuration with coupled loops. This is done for the gain adaption loop in Fig. 5.46. Note that in this configuration the time constants of all loops are kept the same, which means that the damping of the coupled loop reaches its minimum value equal to 1 according to Eq. (5.39). It is clear that in the configuration with coupled loops a much smaller time constant can be allowed than in the case of independent loops. It is also clear from this figure that in
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Figure 5.44: Step response for the inner control loop according to analytical calculation and as a result of simulation with experimental data.

Figure 5.45: Responses for the inner loop when delay values are non-zero

\( K_p^o = 0.04, K_p^i = 0.01, M_o = 12, M_i = 50, T=1 \).

this particular case the gain loop does not need a very high bandwidth. The bit-error rate reaches a minimum for a time constant equal to 4096 bits.

second order loops For the timing recovery loop, which is a second order phase locked loop (PLL), we can apply the same inner-outer structure to solve the problem of detection delay of the stripe-wise Viterbi detector. We start from the scheme in Fig. 5.27 and assume the following:
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![RMS error as function of time constant of the gain adaptation loops.](image)

- The frequency in all rows is equal (the bits are ordered on a hexagonal lattice and no bits slips occur between the rows). There is no frequency offset between the individual rows within the broad spiral. This means that the frequency part of the loop can be based on information from the outer rows only. In this way the frequency error can be derived with relatively low latency. Note also that the errors of the outer loops can be combined, thereby extracting more timing information per unit of tangential distance. This makes the system more robust (for small media defects) and it allows a higher bandwidth of the frequency loop.

- Also the phase is ideally equal for all rows, e.g. rotation speed variations will lead to equal frequency and phase errors in all the rows. However, due to lattice distortions (e.g. caused by inaccuracies in the synchronization during multiple-pass mastering) a time-varying phase between the rows might occur. Fortunately, Fig. 5.41 and Fig. 5.12 show that these phase variations are relatively slow (periods of variation in the order or even larger than 10,000 bits). This means that a low-bandwidth, first-order phase loop for the inner rows is enough to achieve low overall phase errors.

Based on the above assumptions also here a configuration with coupled loops for the inner and outer rows is designed. The outer loops are second order loops that compensate for phase and frequency variations. The information from the outer loops is inserted in the inner loop, which is a low-bandwidth, first-order phase loop. The scheme for a single outer loop and a single inner loop is schematically shown in Fig. 5.47. Here $\Delta \phi^2_k$ is the phase error of outer loop 2 with detection delay $M_o$. The phase error of outer loop 1 (at the other side of the broad spiral) is indicated with
These values are averaged by adding them together and dividing the result by 2. The average is input to the integrating (frequency) branch of the loop filter (LF). The proportional branch of the loop only receives the phase error of the corresponding row. The result of the integrating branch is inserted in the inner loop. Here only a phase correction is added, which is based on the phase error $\Delta_{k}$. The NCOs produce sampling phases $\Psi_{k}^{i}$ and $\Psi_{k}^{o}$ for the inner and outer loop respectively. These are compared to the ideal sampling phases $\phi_{k}^{i}$ and $\phi_{k}^{o}$ respectively to produce the phase error signals.

**Figure 5.47:** Configuration with an inner loop and outer loop in case of a second order timing recovery scheme.

Fig. 5.48 shows the step responses of the timing recovery loop for different rows in the broad spiral. The left figure shows the step responses in case $N_{r}$ independent timing recovery loops are applied. The loop parameters are optimized with a brute-force search to achieve lowest bit-error rate at the output of the stripe-wise Viterbi detector for zero detection delay ($\zeta = 1; \omega_{n} T = 0.03$). The actual detection delay is equal to 10 for each stripe-processor (i.e. the total detection delay is 10 for row 0, 20 for row 1, 30 for row 2, etc.). According to Fig. 6.14 this is realistic value. It is clear from Fig. 5.48 that the damping is lower for the inner rows. The right figure shows the step responses for the configuration of Fig. 5.47. The loop parameters are not changed. Note that the damping of the inner rows is nearly as good as the damping of the outer rows. Furthermore, the outer loops improved in dynamic behavior due to the averaging of the phase errors of the outer rows.
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Figure 5.48: Response on a phase step (0.4 bit period) for the timing recovery configuration with \( N_r \) independent loops and for the proposed scheme with different inner loops and outer loops. Detection delay of a single-stripe processor is assumed to be 10.

The inner loop is a simple first order loop with a relatively large loop delay that has transfer functions according to Eq. (5.35). It is able to handle phase variations. It is not able to handle frequency variations. However, it is assumed that frequency variations in \( \phi_i^k \) are also present in \( \phi_o^1 \) and \( \phi_o^2 \). In order to analyze the dynamic behavior of the coupled loops, we calculate the transfer function from \( \phi_o^1, \phi_o^2 \) and \( \phi_i^k \) to the sampling phase \( \Phi_i \) in case we excite the inputs simultaneously with a common signal. Under the continuous time and ‘delay=0’ approximation one can write for this total transfer function:

\[
\frac{\Phi_i(j\omega)}{\phi(j\omega)} = D \left[ \frac{A}{j\omega + K_p^i} + \frac{j\omega B + C}{(j\omega)^2 + K_p^o j\omega + K_o^i} \right]
\]

with A, B, C and D constants that only depend on \( K_p^i, K_p^o \), and \( K_o^i \). It is clear from the equation that the total system is stable as long as the second order, outer loop is stable. Furthermore, the error can be shown to go to zero in case of a frequency step (phase ramp) by observing that:

\[
\lim_{k \to \infty} \Delta_i^k = \lim_{z \to 1}(z - 1)\Delta'(z) = \lim_{z \to 1}(z - 1)\phi(z) \left( 1 - \frac{\Phi(z)}{\phi(z)} \right) = 0
\]

when for \( \frac{\Phi(z)}{\phi(z)} \) we fill in the discrete-time equivalent of Eq. 5.40 and for \( \phi_i^k \) we apply a phase ramp \( \left( \frac{z}{(z-1)^2} \right) \). This means that stability and convergence of this system are
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guaranteed under the assumed approximations. More details on stability analysis of second order phase locked loops in the presence of loop delay can be found in [68].

A further verification of the scheme with coupled inner and outer loops is done by measuring the root mean square (RMS) error as function of the normalized natural frequency $\omega_n T$ of the timing recovery loops. This is done for both the coupled loops and the independent loops. Replay signals of disc E490 are taken as input to the receiver. The results are shown in Fig. 5.49. The RMS error is shown for row 0 (dashed line), row 2 (dotted line), and row 3 (solid line, center row). The square markers indicate the configuration with independent loops, while the circular markers indicate the scheme with coupled inner and outer loops. The optimum RMS error is -15 dB. Note that for increasing bandwidth at some point the RMS phase error increases dramatically. This occurs first for the center loops, and for even higher loop bandwidths also for the rows that are nearer to the guard band (but at that moment the center loops lost phase and frequency lock already). It is clear from the figure that for

Figure 5.49: Root mean square (RMS) error as function of normalized natural frequency of the timing recovery loop for different rows in the broad spiral, indicated with the dashed line (row 0), dotted line (row 2) and solid line (row 3). The square markers indicate the configuration with independent loops. The circular markers indicate the scheme with coupled inner and outer loops. Experimental data of disc E490 is used for the experiment.
the situation with coupled inner and outer loops the bandwidth of the timing recovery can be increased to significantly higher values without problems with instability than in the case with independent loops. Equivalently the loop delay can increase without problems with stability. This proves the basic concept of our proposal. Note that in the example all loop parameters are taken equal. A row-dependent parameter optimization might lead to further improvements in performance.

5.6 Conclusions

A 2D receiver has been designed starting from the analogue front-end up until the bit-detector. The analogue signals coming from the multi-spot photo diode IC are digitized in parallel using a fixed sampling clock. Some over-sampling is applied in order to relax the requirements for the analog anti-aliasing filter. A steep digital noise filter is applied after digitization, which is combined with the delay compensation block. The delay compensation block is needed to eliminate the relative row-delays caused by the slanted orientation of the linear array of read-out spots compared to the tangential direction of the broad spiral. The delay compensation is done using a feedback control loop that is only enabled during the preamble. Here, the fact that the data (and for the inner rows also the signal) is equal in each row is used as prior knowledge. The signal for the outer rows suffers from inter symbol interference of the guard band and is excluded from delay estimation. Using the delay compensation block lattice distortions in the tangential direction can be measured. For the 50 GB disc (E490) a peak-to-peak distortion of about 8% of a channel bit period can be observed.

After delay compensation several adaptation loops are present. The order in which the adaptation is done is very important. The loops have different bandwidth requirements and therefore tolerate less or more delay in the loop. For this reason the DC, AGC and adaptive equalizer loops are placed in front of the timing recovery loop, which needs a high bandwidth. The DC and AGC are placed after the equalizer loop. The equalizer deals with the remaining, slow channel variations that reside after the fast-varying DC and gain parameters have been identified and tracked. It appears there are several effects that cause a time-varying DC-offset:

- drift in the analogue front-end of the receiver due to temperature variations, ageing, or supply voltage variation. Part of the offsets can be filtered by an AC-coupling of the signal.
- baseline wander. A DC-content of the signal is not passed by the AC-coupling. A solution is to insert DC-control information in the bit-stream.
- variation in the optical path. The resulting fast variations in DC content of the signal require compensation with a fast DC restoration loop.
It appears that many of the above variations (especially the fast ones) are common in all the rows. This observation is crucial in our design of the adaptation loops. One of the major problems in this design is the large loop delay caused by the large detection delay of the stripe-wise 2D Viterbi detector. This causes either instability of the loops or a limitation in maximum achievable bandwidth of these loops.

The observation that many variations are common to all the rows in the broad spiral allows us to compensate variations in the inner loops based on information derived from the outer loops. Analysis of the dynamic behavior of these coupled loops shows that the overall transfer from outer to inner loops shows a second-order behavior but has a guaranteed stability.

For the second order timing recovery loop a similar inner-outer configuration can be applied to solve the large detection delay of the stripe-wise Viterbi detector. The proposed scheme is based on two assumptions:

- the frequency in all rows is equal, which means that the frequency part of the loop can be based on the low-latency information from the outer rows only. Also errors can be combined, thereby extracting more timing information per unit of tangential distance.

- the phase is ideally also equal for all rows. However, due to lattice distortions a time-varying phase between the rows may occur. Fortunately, these phase variations are relatively slow such that a low-bandwidth, first-order phase control loops can be applied.

Based on these assumptions an inner-outer configuration is designed, where the outer loops are second order loops that compensate from phase and frequency variations. The information from the outer loops is inserted in the inner loop, which is a low-bandwidth, first-order phase loop. The performance of the configuration with coupled inner and outer loops is verified by deriving the phase step response for different rows in the broad spiral. A further verification of the scheme is done by measuring the root mean square phase error as function of loop bandwidth. It appears that the coupled loops allow a significantly higher bandwidth than the independent loops. Equivalently, a higher detection delay can be allowed compared to the independent loops. This result validates the proposed adaptation configuration for the 2D receiver.

Another issue in the timing-recovery loop is the design of the SRC filter. It appears that the application of 1D filters causes aliasing in case of a hexagonal lattice. An implementation with polyphase filters in combination with linear interpolation allows relatively straightforward modification of 1D to 2D filters.
Chapter 6

2D Bit Detection

6.1 Introduction

In the TwoDOS system the track pitch is reduced considerably compared to the traditional 1D formats in optical recording (by a factor even larger than 2). Such a system results in considerable 2D inter-symbol interference (ISI): For the hexagonal case the inter-track interference is just as significant as the ‘along-track’ ISI. Also for RLL coded data this isotropic ISI condition may be approached. Chapter 2 already showed that considerable energy of the impulse response is ‘located’ in the neighbouring bit-rows, and that one of the main advantages of a 2D format can be argued to be ‘joint 2D bit-detection’, which means that all energy associated with a single bit is used in the detection process. This contrasts with 1D bit-detection in the conventional 1D format where inter-track interference is considered to be a disturbance and is sometimes tackled via cross-talk cancellation (see also Chapter 2). In the latter case, only the energy ‘along-the-bit-row’ is being used, thus yielding approximately 40% loss of energy per-bit (40% corresponds to the linear 2-1 target response, with 2 as the central tap value and 1 for all the neighbouring tap values). This chapter will discuss a practical algorithm to perform the ‘joint 2D bit-detection’. To define the nomenclature the 1D Viterbi detector is explained very briefly in Section 6.2. A straightforward extension of this detector to the 2D case would result in a complexity that is prohibitive for practical implementation of this detector. As a solution a stripe-wise approach is proposed. A 2D stripe-wise Viterbi detector is explained in Section 6.3, where a stripe is defined as a set of adjacent bit-rows. One stripe passes information to the next stripe to retain the ‘joint 2D detection’ property. At one side of the stripe this information is either unreliable or even not available. In that case cross talk cancellation can be used (see Section 6.4). Section 6.5 describes a new performance measure for signal quality at the input of the detector. Finally, a 2D detector for RLL coded data is discussed (Section 6.6).
6.2 1D Viterbi Detection

A Viterbi detector is used to perform maximum-likelihood detection of the TwoDOS data sequence. Before we explain the 2D detector we will briefly describe the principle of Viterbi detection in the conventional 1D case to define the nomenclature. Viterbi detection is possible because our optical channel has a finite memory length $M$, which means that the equivalent digital channel output $r_k$ only depends on the current bit $b_k$ and $M$ previous bits $b_{k-M}, \ldots, b_{k-1}$. So for a channel with memory length $M=2$ the output of the channel is dependent on a sequence of 3 input bits (for simplicity we use a linear channel according to Eq 3.11):

$$r(s_{k-1}, s_k) = \sum_{i=0}^{M} h_i b_{k-i} + n_k,$$  (6.1)

where $s_k = \{b_{k-M+1}, \ldots, b_k\}$ is the state that reflects the memory in the channel, and $n_k$ is a noise sample (note that a Mealy model is used where the output of the channel depends on the departure state $s_{k-1}$ and the arrival state $s_k$). Now define a neighbourhood $I$ as an ordered index set $I = \{0, 1, 2, \ldots, M\}$ of relative indices with respect to $k$ that define the positions of those bits that contribute to the channel output. The absolute indices can then be calculated according to:

$$k - I = \{k - i | i \in I\}.$$  (6.2)

Furthermore, when we denote $S$ as the set of all possible states $s_k$ then the cardinality of $S$ is the number of possible memory states of the channel. In the current case of binary (uncoded) data this number is equal to:

$$|S| = 2^M.$$  (6.3)

We can represent each of these states as a node in a trellis diagram. For our example-channel with $M=2$ and uncoded input data the trellis diagram is shown in Fig. 6.1. In this trellis, each state $s_k$ at instant $k$ has two possible predecessor states $s_{k-1}$. Note that we assume no modulation coding or run length coding, otherwise some states would have only one predecessor state due to the run length constraints. The number of branches $B$ is then equal to:

$$|B| = 2^{M+1}.$$  (6.4)

Next we introduce a reference channel model $g$ with which we can calculate the preferred (or reference) output of the channel, denoted as $\text{REF}_{\{s_{k-1}, s_k\}} : \{0, 1\}^I \rightarrow \mathbb{R}$, according to:

$$\text{REF}_{\{s_{k-1}, s_k\}} = \sum_{i=0}^{M} g_i b_{k-i} = \sum_{i \in I} g_i b_{k-i} = (g * b)_k.$$  (6.5)
6.2 1D Viterbi Detection

Figure 6.1: Trellis diagram for a channel with memory length 2 and uncoded input data

Note that the reference channel output is determined by the set \( \{ s_{k-1}, s_k \} \). So once we are in a particular state \( s_{k-1} \) and we receive the actual channel output \( r_k \) we can evaluate each of the arrival states \( s_k \) of departure state \( s_{k-1} \) based on some criterion. In general the criterion is to find the combination \( \{ s_{k-1}, s_k \} \) that matches \( r_k \) ‘best’. Equivalently, once we are in a state \( s_k \) we can evaluate each predecessor state \( s_{k-1} \) based on the same criterion. In case we assume that \( n_k \) represents additive white Gaussian noise (AWGN) it appears that maximum likelihood detection is reached when the criterion of ‘best’ fit is the Euclidian distance criterion (see Appendix 3C in [59]):

\[
\beta_{\{s_{k-1}, s_k\}} = (r_k - \text{REF}_{\{s_{k-1}, s_k\}})^2 = (r_k - (g * b)_k)^2. \tag{6.6}
\]

Here \( \beta_{\{s_{k-1}, s_k\}} \) is the metric on the branch going from state \( s_{k-1} \) to state \( s_k \). It is calculated in the so-called Branch-Metric (BM) unit of the Viterbi. The task of the detector is now to find a path (consisting of a set of branches) through a number of subsequent states that has the least total metric sum, i.e. the best fit of received samples \( r_k \) to the calculated reference samples \( \text{REF}_{\{s_{k-1}, s_k\}} \). To carry out this task the total metric sum (also called path cost) is tracked during the build-up of the trellis diagram. The path cost on going from state \( s_{k-1} \) towards state \( s_k \) can be written as:

\[
\gamma_k^{s_{k-1}, s_k} = \gamma_k^{s_{k-1}} + \beta_{\{s_{k-1}, s_k\}}. \tag{6.7}
\]

At an instant \( k \) each predecessor state is evaluated and a selection is made of that branch that leads to the lowest total path cost, i.e:

\[
\gamma_k^{s_k} = \min \{ \gamma_k^{s_{k-1}, s_k} + \beta_{\{s_{k-1}, s_k\}} | s_{k-1} \in S \}, \quad s_k \in S. \tag{6.8}
\]

This is done in so-called Add-Compare-Select (ACS) units. Here \( \gamma_k^{s_k} \) is the best path metric of state \( s_k \) at instant \( k \). To avoid an unbounded increase in the value of the path metric (\( \beta_{\{s_{k-1}, s_k\}} \) is always positive) a modulo operator may be applied to the
metrics [69]. The set of states having the lowest total path cost is called the survivor path. After an initialization period it is possible to continuously trace back along this survivor path. This is done by the Survivor-Memory (SM) unit. Because of the finite memory $M$ of the channel each of the surviving paths will, after some amount of back-tracking, pass through the same node in the trellis diagram. This node contains the detected bit that is sent to the output of the detector. An example is shown in Fig. 6.2, where the survivor branch is indicated with a solid line and where the trace-back paths are shown with a thicker line.

**Figure 6.2:** Trellis diagram for a channel with memory length 2 and uncoded input data. The numbers on the branches indicate the branch metrics.

It is common to start trace back from the state that has the lowest path metric for the current state. Although, one can expect that this leads to the correct output with a minimum back-tracking depth, it is not obligatory to start from this state with minimum path metric. Due to the merging of paths all paths converge to the single best path upon back-tracking.

### 6.3 2D Viterbi Detection

In the TwoDOS system we want to apply a two-dimensional Viterbi detector to fully exploit the two-dimensional character of the data on the disc. Preferably we want to do joint detection of bits over all rows $n = \{0, 1, \ldots, N_B - 1\}$ in the broad spiral. This paragraph describes a practical implementation to perform this joint detection [70]. It describes some refinements of multi-track Viterbi detection of which the underlying principles have been described previously in the M.S. Thesis of Krishnamoorthi [71] and in the Ph.D. Thesis of Weeks [72], where it is referred to as the multi-track Viterbi algorithm (MVA). The refinements include a number of additional features that significantly improve the bit-detection performance: (i) weighing of the separate contributions to the branch metric from each bit-row in a stripe; (ii) inclusion of an additional contribution to the branch metrics from a bit-row adjacent to a stripe; and (iii) processing the consecutive stripes starting from the guard bands towards the
center. In addition, unlike the MVA which uses straightforward decision feedback from one stripe to the next, our approach uses the output of a previously processed stripe to condition the reference levels in the branch metric for the next stripe. In this way, we can handle non-linear ISI. Finally, the computational complexity is reduced by varying the number of bit-rows per stripe during successive iterations of the MVA, and through the use of local sequence feedback [73, 74].

In public literature, 2D storage is studied mainly in the area of optical recording. In [31] for example it is also proposed to use a Viterbi detector for 2D stored data. However, in this publication a set of independent Viterbi detectors is applied, e.g. Viterbi machines that process three consecutive bit-rows and output the bits of the center bit-row. The stripe-wise approach of the MVA-algorithm on the other hand is based on a concatenation scheme of interconnected Viterbi detectors in which the output of previous stripes is passed as side-information for subsequent stripes. Similarly, also magnetic storage can be adapted towards a 2D format, and some papers start to appear in the literature (see e.g. [75]). In [76] a 2D maximum likelihood detection in the form of the so-called M-algorithm is proposed. This algorithm performs a tree search similar to the Viterbi detector. However, it only retains M states in each stage of the trellis. By tuning M a trade-off can be made between complexity and detection performance. Another scheme which has similarities with the MVA scheme is proposed in [77]. It passes on soft-decision information in between iterations (in an iterative detection scheme) and in between stripes. The soft information is obtained by performing a detection algorithm with soft output (i.e. probabilities instead of bit-decisions) on stripes with 2 bit rows. A 3-tap channel is used with impulse response:

\[
h = \begin{bmatrix} 1 & \alpha \\ \alpha & 0 \end{bmatrix}
\]

(6.9) with some parameter that is swept from 0 to 1 to explore the algorithms behavior in the presence of varying amounts of ISI. It appeared from our simulations that passing soft-decision information from one stripe to the next in the stripe-wise Viterbi bit-detection scheme does not improve bit-error rate and therefore a hard-decision information passing was pursued.

To explain our stripe-wise Viterbi algorithm we assume two-dimensional states denoted with \(s_{k,n}\). We avoid the introduction of new symbols for the 2D situation. For the reader it should be no problem to distinguish between the 1D and 2D case. The states have a finite length along the tangential direction of the broad spiral and may not necessarily include all rows in the broad spiral. An example where all rows in the broad spiral are included though is schematically illustrated in Fig. 6.3. The light gray and medium gray positions belong to state \(s_{k-1}\) while the medium gray and dark gray positions belong to state \(s_{k}\). We can drop the subscript \(n\) in this case because the states cover the full width of the broad spiral.
It should be no surprise that the number of states grows exponentially to:

$$|S| = 2^{MN_B}.$$  \hfill (6.10)

For example for a system with (tangential) channel memory length $M=2$ and 7 rows in a broad spiral the number of states would be equal to $2^{14} = 16384$, which is far too much for a practical implementation. For this reason it was proposed to limit the number of rows over which joint detection is done simultaneously to $N_S < N_B$ [78], where $N_B$ is the number of rows in the broad spiral. Such a subset of rows within the broad spiral is called a ‘stripe’. In our configuration all stripes have the same number of rows $N_S$ and each stripe is shifted by one bit-row with respect to the previous stripe leading to a total number of stripes equal to:

$$T_S = N_B - N_S + 1.$$  \hfill (6.11)

The shift of only one bit-row allows us to associate each stripe with one row and to number the stripes with index $n$. The total detector now consists of a number of Viterbi processors each acting on such a stripe. For this reason the complete detector is referred to as ‘stripe-wise Viterbi detector’. Each stripe is passing its detected results to the next stripe to be used as side-information to condition the reference level calculation for this next stripe. Contrary to the impractical full-fledged Viterbi bit-detector as discussed above, a Viterbi processor that operates on a stripe of only 2 or 3 rows still has a reasonable complexity. For the 1-shell channel model as discussed in Section 3.3.2, the tangential span of the 2D impulse response is 3 taps ($M=2$). This means that the number of states is 16 for a 2-row stripe, and 64 for a 3-row stripe. The
number of branches per state is then 4 and 8, respectively. An additional advantage of the stripe-wise approach is that all the Viterbi stripe-processors can operate in parallel using different hardware components. This maximizes the achievable data throughput of the detection scheme in a practical IC implementation. Now let us consider one stripe processor (e.g. $V_{00}$) as indicated in Fig. 6.4. A state can be defined as an ordered set of relative indices:

$$ S = \{ (i, j) | i = -M + 1, \ldots, 0, \quad j = 0, 1, \ldots, N_s - 1 \} \quad (6.12) $$

The index set of a departure state for stripe $n$ located around $(k - D_n - 1, n)$ with $D_n$ the row-dependent delay of each of the stripe processors, is now given by:

$$ S_{k-D_n-1,n} = (k - D_n - 1, n) + S = \{ (i, j) | i = k - D_n - M, \ldots, k - D_n - 1, \quad j = n, n + 1, \ldots, n + N_s - 1 \} \quad (6.13) $$

whereas an arrival state index set is shifted forward one position in the (tangential) i-direction with respect to the departure state,

$$ S_{k-D_n,n} = S_{k-D_n-1,n} + (1, 0) = \{ (i, j) | i = k - D_n - M + 1, \ldots, k - D_n, \quad j = n, n + 1, \ldots, n + N_s - 1 \} \quad (6.14) $$

The delay $D_n$ is present because each next processor uses the bits from the previous detector as side information and must wait until these bits become available in view of the finite detection delay of the stripe-processors. We will omit $D_n$ in cases where it is not absolutely necessary for the explanation. The shift of an index set (or neighbourhood) as used in Eq. 6.13 and Eq. 6.14 is defined similar to Eq. 6.2 as follows:

$$ S + (k, n) = \{ (k + i, n + j) | (i, j) \in S \} \quad (6.15) $$

In general $(k, n)$ will be used for absolute positions and $(i, j)$ as relative indices. Note that the departure state $s_{k-1,n}$ is an assignment of bit values on the index set $S_{k-1,n}$, and an arrival state $s_{k,n}$ is an assignment of bit values on an index set $S_{k,n}$. Due to the limitation of the number of rows per stripe there are only $|S| = 2^{MNS}$ possible arrival states (and the same number of departure states). A departure state $s_{k-1,n}$ and an arrival state $s_{k,n}$ are connected in the trellis if their bit values agree in the common positions $S_{k-1,n} \cap S_{k,n}$. The branch is indicated, similar to the 1D case, as $(s_{k-1,n}, s_{k,n})$.

The decisions made by a Viterbi processor are stored in array $\hat{b}$ of most recent bit-estimates that are available. Per-stripe bit-detectors simply overwrite bit rows in this array. Initially, $\hat{b}$ is filled with e.g. all zeroes, or with best guesses of the channel bit values as obtained from a preliminary bit-detector like threshold detection. At an intermediate stage of the stripe-wise scheme, $\hat{b}$ is filled with the most recent bit-decisions available from the per-stripe bit-detectors that have already been executed;
they can be used as side-information for subsequent per-stripe bit-detectors. After
the stripe-wise scheme has finished, \( \hat{b} \) contains the final bit estimates.

**Branch metric calculation** The total branch metric is calculated as the sum of con-
tributions from each of the rows within the stripe. For each contribution the reference
level is dependent only on a local 2D cluster of bits within the neighbourhood defined
by the ordered set (see the circle in Fig. 6.4):

\[
\mathcal{N}^{2D} = \{(0,0), (1,0), (0,1), (1,1), (-1,0), (0,-1), (-1,-1)\},
\]

which can also be written as:

\[
\mathcal{N}^{2D} = \{(i,j)|i^2 + j^2 - ij \leq 1, i, j \in \mathbb{Z}\}. \quad (6.16)
\]

In this neighbourhood the reference level can be calculated according to (see Eq. 3.12,
again for a linear model):

\[
\text{REF}_{k,n,\mathcal{N}^{2D}} = \sum_{(i,j) \in \mathcal{N}^{2D}} g^{2D}_{i,j} b_{k-i,n-j}.
\] \quad (6.17)

Observe that for the most general (non-linear) case the table:

\[
\text{REF}_{\mathcal{N}^{2D}} : \{0,1\}^{\mathcal{N}^{2D}} \rightarrow \mathbb{R}
\]

has 27 entries for the assumed 1-shell target (with \( \mathbb{R} \) the set of real numbers). In case
a certain symmetry of the optical spot and of the pits on the disk can be assumed, the
size of the table can be reduced (e.g. by calculating the entries according to the 4-
parameter model as discussed in Chapter 3). The absence of full rotational symmetry
can e.g. be due to the presence of tilt between the optical disc and the scanning laser spots. Equalizer filters can be used to enforce a certain symmetry (see Section 5.4). Due to the intrinsic non-linearity of the channel response, such equalizer filters may require non-linearity compensation and may become quite complex [79]. The advantage from a full table is then that the stripe-wise bit-detection scheme does not need such non-linear equalizer filters because it can deal with arbitrary responses \( g^{2D} \). Those arbitrary responses account for the non-linearities of both the write-channel and the read-channel. Optionally it is also possible to keep track of a list of variances:

\[
\sigma^2_{N^{2D}} : \{0, 1\}^{N^{2D}} \rightarrow \mathbb{R},
\]

which can be used later on to further improve the performance of the detector. Observe also that in Eq. 6.17 it was deliberately chosen to use \( N^{2D} \) as a subscript to \( \text{REF} \) instead of \( \{s_{k-1,n}, s_{k,n}\} \). The reason is that in the trellis of a stripe, a branch or state transition \( \{s_{k-1,n}, s_{k,n}\} \) specifies bit values only for points in \( S_{k-1,n} \cup S_{k,n} \). The set of neighbourhoods which are needed for the branch metric extent beyond this index set (as is clear from Fig. 6.4 where bits that are not defined by the branch are indicated with crosses). So knowledge of the bits in the branch does not suffice. We also need certain bits from \( \hat{b} \), the 2D array of current estimates, specifically \( \hat{b}_{k-D_n,n-1} \) and \( \hat{b}_{k-D_n,n-1} \) for the top boundary and \( \hat{b}_{k-D_n-2,n+N_S} \) and \( \hat{b}_{k-D_n-1,n+N_S} \). The total branch metric is now calculated over the medium gray bit-positions as the sum of the branch metric for each row:

\[
\beta_{\{s_{k-1,n}, s_{k,n}\}|\hat{b}} = \sum_{j=0}^{N_S-1} \left( r_{k-D_n-1,n+j} - \text{REF}_{\{s_{k-1,n}, s_{k,n}\}|\hat{b}} \right)^2. \tag{6.18}
\]

**‘High Certainty’ Side and ‘Low Certainty’ Side** Assume we have an 11-row meta-spiral and use 3-row stripes as indicated in Fig. 6.5. The top row of the figure corresponds to row 0. A uni-directional sequence of per-stripe Viterbi bit-detectors named \( V_0, V_1, \ldots, V_8 \) is suggested, one for each stripe. As already mentioned in Eq. 6.11 only \( (N_B - N_S + 1) \) detectors are needed to cover the complete broad spiral of \( N_B \) rows. All Viterbi stripe-processors, except the last one \( V_8 \), have an identical structure, and they produce as output their top bit-row. The last stripe processor \( V_8 \) differs in this respect since it outputs all its bit-rows. The reason that most stripe processors only produce the output bits of a single row is the following: As already mentioned not all bits are defined by the branch in the trellis. Some of them lie either within the bit-row immediately above the stripe or within the bit-row immediately below the stripe. These bits are considered as the side-information that is required for the Viterbi bit-detector of this stripe. For the stripe corresponding to \( V_0 \), the aforementioned border row immediately above the stripe is the guard band \( (n=-1) \). This bit-row is known -i.e. is of ‘high certainty’- as it is modelled as containing all
zero bits. Here one of the functions of the guard-band becomes clear: it can serve as a starting point for bit-detection. Now, for all other stripe-processors the row above the stripe is also of ‘high certainty’ because these bits are provided by the previous stripe and can be taken from the array $\hat{b}$ of most recent bit-estimates. For $V_{00}$ the difficulty lies in the aforementioned border row immediately below the stripe. This row contains unknown data - i.e. it is of ‘low certainty’ - as it contains bits that are yet to be detected, or that have been detected during a previous iteration of the stripe-wise scheme (as we will discuss later). This is also the case for $V_{01}$ to $V_{07}$. Fig. 6.6 shows the bER as function of peak signal-to-noise ratio (PSNR) for each of the 3 rows after a single iteration with a 3-row Viterbi processor ($V_{00}$ in this case). PSNR is defined as:

$$\text{PSNR} \triangleq 20 \log \frac{1}{\sigma}$$

(6.19)

where the signal at the input of the detector is assumed to be ‘1’ in case the detected light is reflected from a perfect mirror and ‘0’ in case no light falls on the detector. It is clear that the bER is limited by the low certainty border. For $V_{08}$ also the row immediately below the stripe is of ‘high certainty’ because it is a guard band filled with zeros. Normally, for each stripe, the bit-row detected with the highest reliability is considered to be the output of that stripe (see also Fig. 6.6), and is stored either for further use by Viterbi processors of subsequent stripes, or for further iterations, or for the final output. For processors $V_{00}$ to $V_{07}$ this is the top row. For $V_{08}$ both sides have a high certainty and therefore, the bit decisions in all rows covered by this stripe are expected to be of equal reliability and are used as output of the processor.

![Configuration of Viterbi processors in a Viterbi detector to cover the complete broad spiral. Note that one horizontal line indicates a single bit-row.](image)

**Figure 6.5:** Configuration of Viterbi processors in a Viterbi detector to cover the complete broad spiral. Note that one horizontal line indicates a single bit-row.

**Iterating the stripe-wise processing**  As already mentioned in the previous paragraph, side-information can be taken from the array of most recent bit-estimates. To do this the delay of $V_{01}$ with respect to $V_{00}$ must at least be the processing delay of $V_{00}$ due to the trace-back processing of the Viterbi algorithm. Thus the output row of the first stripe $V_{00}$ serves as high certainty border row of the second stripe $V_{01}$ and can be used in the computation of the branch metrics of that stripe. This
procedure is continued for all stripes in the meta-spiral: The full procedure from top to bottom of the meta-spiral is considered to be one iteration of the stripe-wise bit-detector. In a similar fashion, the output bit-rows of a given iteration can be used as side information at the low certainty side of all stripes of the next iteration. That is, for the low-certainty border rows, during all subsequent iterations, it is no longer necessary to use arbitrary guesses or initial threshold decisions. Therefore, we can expect that the bit-error rate (bER) after the second iteration has decreased compared to that of the first iteration. Possibly a third iteration can be implemented for a further improvement, etc. Beyond a certain iteration however, the bER will not decrease significantly anymore. At this point the bER is limited by noise or by the occurrence of error-events comprising a number of bit-errors in patterns that typically extend in the radial direction well beyond the number of bit-rows that make up a stripe. This phenomenon demarcates the error floor of the stripe-wise bit-detection scheme.

**<‘\> Shaped Detector Configuration**  
Bit error rate analysis of the output rows of the per-stripe processors reveals -not surprisingly- that the bER increases for bit-rows with increasing distance from the top guard band. As already mentioned the guard band with its 100% reliability constitutes the anchor point for the stripe-wise bit-detection scheme. This becomes clear from Fig. 6.7, where the bER is plotted as function of the row number for various values of PSNR. The configuration of
Fig. 6.5 is taken with 7 rows. Row number 1 is the starting point of the detection. An increasing bER is obtained up to row 4. Row 5, 6 and 7 are detected simultaneously with a 3-row processor and therefore, show decreasing bER towards the guard band. This observation suggests to use the two guard bands of the meta-spiral in a more symmetric fashion. In that case successive stripes will be arranged in a ‘<’-shape as becomes clear from Fig. 6.8 which indicates the configuration for the practical case of an 11-row broad spiral and stripes consisting of three bit-rows. The even-indexed per-stripe processors $V_{00}, V_{02}, \ldots, V_{06}$ have their top bit-row as output row since this row has the highest reliability. The odd-indexed per stripe processors however, have their bottom bit-row as output row, because here this one has the highest reliability output. The two cascades of stripes are terminated in the middle of the broad spiral with a last stripe processor $V_{08}$, which outputs all its bit rows. A same reasoning can be applied to the second or further iteration of the stripe-wise scheme.

An additional advantage of the ‘<’-shaped configuration is that the detection delay of the bottom half of the broad spiral decreases considerably. The reason to pursue a small detection delay is that for a decision directed adaptation scheme of the total receiver the detector will be part of the adaptation loops and hence will
determine the dynamic behavior of these loops. A more detailed discussion is given in Section 5.5.

**Smaller Stripes During Earlier Iterations** It is obvious that a stripe processor for 2-row stripes has an exponentially lower complexity than a stripe processor for 3-row stripes. Stripe processors of different complexity during different iterations can be considered. The best performance in terms of bER is achieved when the last iteration uses the most powerful stripe processor. Using a very powerful (and more complex) bit-detector for the first iteration is not effective, as initially the detection performance is limited anyway by the use of postulated bits for the low certainty border rows. We can take advantage of this phenomenon by the use of lower complexity stripe processors during the first (or earlier) iterations. It has been observed from simulation experiments that satisfactory bER performance can be achieved by the use of 2 iterations, with 3-row stripes for the second iteration and 2-row stripes for the first iteration. This is shown in Fig. 6.9.

**Reducing the Contribution of the Low-Certainty Border** In a previous paragraph it became clear that most stripe processors in the Viterbi detector possess a ‘low certainty border’ and a ‘high certainty border’. The high bit-error rate for the low certainty border causes a frequent use of wrong reference levels REF in the branch...
metric contribution of the signal sample in the bit-row in the stripe closest to this border. Reduction of the contribution to the branch metric of this signal sample can mitigate this effect. Let us denote the weight of the contribution of the \((n+j)\)-th bit-row in the stripe by \(w_j\). The branch metric for a branch \(\{s_{k-1}, s_k\}\) now becomes:

\[
\beta_{\{s_{k-1}, s_k, \hat{b}\}} = \sum_{j=0}^{N_S-1} w_j \left( r_{k-D_n-1,n+j} - \text{REF}_{\{s_{k-1}, s_k, \hat{b}\}} \right)^2.
\]  

(6.20)

In a typical case (assuming for simplicity that \(N_S-1\) is even as it is in the iteration of Fig. 6.8) in the top half of the broad spiral \((0 \leq n < (N_S-1)/2)\),

\[
w_0 = w_1 = w_2 = \ldots w_{N_S-2} > w_{N_S-1},
\]

(6.21)

and in the bottom half of the broad spiral \((N_S/2 < n < N_S)\)

\[
w_0 < w_1 = \ldots = w_{N_S-2} = w_{N_S-1}.
\]

(6.22)

This is shown schematically in Fig. 6.10. Finally, for the center stripe all weights are equal because both borders are of the ‘high-certainty’ type, as the bit estimates \(\hat{b}\) of both borders have been updated in the current iteration by previous Viterbi processors. The weights \(w_j\) can be varied from one iteration to the next because the bits in the low-certainty border row that serve as side information become more reliable at subsequent iterations. Accordingly, the differences between the weights for the various rows can be made smaller. Fig. 6.11 shows the impact of weighing the different branch metric contributions. In case the relative weight of the border row contribution is chosen \(\frac{1}{16}\) for the first 2-row iteration and \(\frac{1}{3}\) for the second 3-row iteration the optimum bER curve as function of PSNR is found. As a reference the unweighed graph is also included. A very clear performance improvement (up to 3 dB at bER=1e-4) is observed.

**Branch Metric Contribution of High Certainty Border** From the discussion in Chapter 2 it became clear that about 20 percent of the signal energy of a channel bit leaks away in adjacent bit-rows. This means that the channel bit in the output bit-row spills 20 percent of its energy in samples in the high certainty border. Inclusion of a contribution to the branch metric of the replay samples in the ‘high certainty’ border improves bER of the bit-detection scheme (as long as the reliability of the bits in the ‘high certainty’ border is good enough). We again assume that the number of stripes is even. For the top half of the broad spiral \((0 \leq n < (N_S - 1)/2)\), the high certainty border had index \(j=-1\) and the branch metric now equals:

\[
\beta_{\{s_{k-1}, s_k, \hat{b}\}} = \sum_{j=-1}^{N_S-1} \omega_j \left( r_{k-D_n-1,n+j} - \text{REF}_{\{s_{k-1}, s_k, \hat{b}\}} \right)^2.
\]  

(6.23)
6.3 2D Viterbi Detection

**Figure 6.10:** Weighing the contributions of the different rows to the branch metric with different coefficients.

**Figure 6.11:** Bit error rate (bER) as a function of peak signal to noise ratio (PSNR). The bottom curve corresponds to the preferred configuration, where weighing is used. The top curve results when all weights are set to 1: a clear bER performance degradation results.

The neighbourhood needed for the extra contribution is indicated in Fig. 6.12 with a circle. To calculate the reference level at \( j = -1 \) for the high certainty border row one requires the bits from row \( n - 2 \). Consequently, this extension is not used for the stripe adjacent to the guard band, as the bit row at \( n - 2 \) refers to another revolution of the broad spiral. Similarly, for the bottom half of the broad spiral \( (N_S/2 < n < N_S) \),
the high certainty border has index \(j = N_S\) and the branch metric now equals:

\[
\beta_{(s_{k-1,n},s_{k,n}|\hat{b})} = \sum_{j=0}^{N_S} \omega_j \left( r_{k-D_n-1,n+j} - \text{REF}_{(s_{k-1,n},s_{k,n}|\hat{b})} \right)^2 .
\]  

(6.24)

For the middle most stripe, both borders actually are of the ‘high-certainty’ type, as the bit estimates in \(\hat{b}\) have been updated in the current iteration of the stripe-wise detector. In that case, additional contributions from both borders of the stripe can be included into the branch metric (i.e. corresponding to both \(j=-1\) and \(j = N_S\)).

**Figure 6.12:** Neighbourhood needed for the extra contribution to the branch metric to take into account the leaked energy at the high certainty border.

**Signal-Dependent Noise Variance** Another technique to improve bER is to use the signal-dependent noise variances. In Section 3.6.3 it became clear that the noise variances are considerably data-dependent. The validity, however, of using the squared error in the branch metric formula derives from the assumption that the noise is additive, white and, especially, Gaussian (AWGN). In practice however, laser intensity noise, for example, can be modelled as multiplicative noise. Also media noise may be cluster dependent, since the non-perfect sidewall of pits contributes to the media noise (while such a contribution is of course absent for non-pit areas). In that case, the squared errors can be normalized by means of these signal-dependent variances. Thus the metric for a branch \((s_{k-1,n},s_{k,n})\) generalizes to:

\[
\beta_{(s_{k-1,n},s_{k,n}|\hat{b})} = \sum_{j=0}^{N_S} \omega_j \left( \frac{r_{k-D_n-1,n+j} - \text{REF}_{(s_{k-1,n},s_{k,n}|\hat{b})}}{\sigma^2_{(s_{k-1,n},s_{k,n}|\hat{b})}} \right)^2 .
\]  

(6.25)

**Survivor management** The problem of detection delay in a decision directed receiver configuration was already indicated above. Note that the practical delay of the detector consist of two main contributions. The first contribution is the time
needed to build up the trellis which is equal to the finite truncation depth that is chosen for the surviving paths. This delay is unavoidable due to the recursive nature of the Viterbi detector. The second delay is of a more practical nature and depends on the survivor management method. Generally, we can distinguish two survivor memory management methods. The first one is the already mentioned trace-back method. Here in each stage the address of the predecessor is stored in a random access memory (RAM). During trace-back the survivor memory at instant $k$ is read out. The content of this memory will point to the address of the predecessor in the path at instant $k - 1$. In its turn it will point to its predecessor, and so forth. For each memory access in a practical implementation one normally needs at least one clock cycle. If the detector logic runs at the baud rate, this means that the additional delay of the trace-back procedure is at least equal to the finite truncation depth and the total detection delay is equal to at least twice the finite truncation depth. A second way of survivor management is called the register-exchange method. Here for all states the instantaneous back-tracking path is kept in memory. It reduces the latency at the cost of higher complexity (i.e. larger silicon area). The register-exchange method is implemented by a set of cross-coupled shift registers as shown in Fig. 6.13. The structure clearly reflects the structure of the trellis. Depending on the decision, a whole backtracking path is selected and the last ACS-result is added to this path, which is shifted into a new set of registers. In the TwoDOS receiver the register-exchange method is implemented because detector delay is critical in the fast control loops (see Section 5.5). Register-exchange allows immediate backtracking by accessing the proper register. In fact, after some trace-back depth, all registers will have the same content.

**Preliminary decisions** Another way to reduce detection delay is to use preliminary decisions for the purpose of error calculation. The actual decisions are derived at a larger delay, but are only used as output of the total receiver. Of course the preliminary decisions taken with a smaller truncation depth $D_p$ are less reliable (see Appendix 7B in [59]. The $|S|$ paths originating from different states at instant $k$ may not have merged at $k - D_p$). In that case the preliminary decisions are taken from that path that has the smallest path metric. Fig. 6.14 shows the bER as function of detection delay for experimental data from a 50 GB disc (E490). A trace-back depth of at least 12 seems to be necessary for a saturation of the bER.

**Comparison with Full-fledged Viterbi Detector** Before further optimizing the detector configuration it is worthwhile to compare the low-complexity stripe-wise approach with the full-fledged two-dimensional Viterbi detector. Fig. 6.15 shows the achieved bER as function of PSNR for both detectors in the case of 7 rows in the broad spiral. Simulations are done without the use of local sequence feedback (as will be discussed in the continuation of this chapter). A performance degradation
2D Bit Detection

of approximately 1 dB is observed. In view of the enormous complexity difference between the two configurations this is considered acceptable.

Local Sequence Feedback  The large number of states (especially in case of a 3-row Viterbi detector) becomes unpractical. A method to reduce the effective number of states, which is used in the TwoDOS detector is sequence feedback. In this scheme preliminary decisions (determined by taking the best predecessor at each state) are used to replace some of the bits within the current state combination \( \{ \sigma_{k-1}, \sigma_k \} \) to effectively reduce the number of bit-permutations over which the Viterbi detector must iterate to find the best predecessor. Each additional sequence feedback bit reduces the state complexity by a factor of 2. In fact the trailing part of the target response \( g \) is subtracted from the incoming impulse response by taking one of the state bits fixed.

**Figure 6.13:** Block diagram of the register-exchange method in case of 8 states with two-input ACS units [80].
in the calculation of the $\text{REF}_{k-1,n,cl}$-signal. The performance of the detector will slightly decrease because of two main reasons. The first one is error propagation. If the preliminary decision is wrong a trailing response of opposite sign is erroneously subtracted and influences all decisions. Local sequence feedback (LSFB: In contrast to global sequence feedback, where the state with lowest path metric could be used as starting point for the trace back) reduces this effect by taking the current state as starting point for trace-back, i.e. states might have different preliminary decisions. A second reason for performance degradation is the loss of data power which is not exploited in the detection process (similar to the data power present in adjacent tracks that is not used when doing 1D detection). Simulations were done for uncoded data on a hexagonal lattice using a 1-shell target response using 1, 2 or 3 bits derived from local sequence feedback. Fig. 6.16 shows the bER as function of SNR in case AWGN is added for all these cases and the reference case of no sequence feedback. It can be observed from this graph that the loss due to LSFB is limited to about 1 dB in case of 3 LSFB bits. The corresponding reduction in complexity of the Viterbi detector is a factor of 8.

**Target Response adaptation** One of the main criteria to optimize the target response (i.e. the set of reference levels of the receiver) is to ‘whiten’ the noise at the input of the detector. The Viterbi algorithm using the squared error in the branch
Figure 6.15: Bit error rate as function of peak signal to noise ratio for the full-fledged 2D Viterbi and the stripe-wise approach using 2 iterations with 2 and 3 rows per stripe respectively.

Figure 6.16: Bit error rate (bER) as a function of peak signal to noise ratio (PSNR) after the second iteration of the preferred stripe-wise bit-detection scheme, with 3, 2, 1, and 0 local sequence feedback bits, respectively.
metric calculation is based on the assumption of AWGN and performs best when the equalized noise of the actual channel is approaching this ideal characteristic (see Appendix 3C in [59]). This means that we are not looking for the target that is a best fit to the unequalized channel, or a target that provides the minimum mean square error (MSE) between the equalized channel response and the target. To find the optimum target for practical channels an exhaustive search over the parameter space is done. For the linear, 1-shell circular symmetric case there is only one parameter $c_1$ as a weight for the nearest neighbour bits which can be varied. The result of the search is shown in Fig. 6.17. From this figure one can see that the minimum MSE does not coincide with minimum bER. The minimum in MSE allows a good fit with respect to the equalized replay signals, but it does not offer a white noise power spectral density. In case we suffer from media noise only (which has a power spectral density that resembles the psd of the uncoded data when passed through the optical channel) we can perform full-response equalization and ideally $c_1 = 0$. In practice however some other noise sources are present which become important at the highest frequencies passed by the channel. Therefore, an optimum bER is found at $c_1 = 0.26$, while minimum MSE is reached at $c_1 = 0.42$. For the optimum value of $c_1$ in terms of bER the fit of the replay signal to the target may be rather poor e.g. due to non-linearities. The fit can be improved by adapting the full table of reference levels (128 values)

![Graph showing bER and MSE as function of parameter $c_1$ for replay signal from disc E266.](image-url)

Figure 6.17: bER and MSE as function of parameter $c_1$ for replay signal from disc E266.
according to the following update rule:

\[ \text{REF} = (1 - \alpha)\text{REF} + \alpha r_{k,n} \]

(6.26)

A reference level (i.e. an entry in the table) is only updated when the particular cluster occurs in the detected data. When we apply this method two related problems occur:

(i) The target that was found initially using an exhaustive search is the optimum circularly symmetric one. By updating the reference levels one might slowly drift from the optimum condition of a white noise power spectral density.

(ii) The equalizer is using the target for generation of its error signal in such a way that its output signal matches the target optimally according to the chosen criterion. If the target changes the equalizer will try to follow the changed target, which leads to interaction between the two adaptation loops. For example, a trivial solution would be the all-zero solution where the reference levels and the equalizer taps are zero. This situation can be simply avoided by constraining the center equalizer tap to 1 for example. A more comprehensive approach to solve these problems is to use two targets. This is schematically shown in Fig. 6.18. A first target is a constant one with parameters derived from the exhaustive search. It is generally a simple 2D filter and is used to generate error signals for all control loops. A second target is initialized using the first target during start-up of the receiver (so that both targets are equal at start-up), but then it is adapted according to the described method. The reference levels from this second target, which are generally stored in a table, are used in the Viterbi detector. In this way any interaction between the adaptation loops is avoided. The

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**Figure 6.18:** Scheme for target adaptation.
6.4 Single-Sided Cross-Talk Cancellation

merit of this method are clear from Fig. 6.19, where the bER is plotted as function of the adaptation constant $\alpha$ for replay signals from disc E490 (50 GB) in the nominal case (no tilt) and for disc E266. A data-aided method was applied where a delay equal to the detection delay was incorporated. A value $\alpha = 0$ means no adaptation. A clear gain in bER can be observed for $\alpha$ values larger than zero. Higher values of $\alpha$ (to decrease the adaptation time constant) appear to be not beneficial for bER, due to the introduction of gradient noise. Finally, at $\alpha=1$ the reference level is just equal to the sample that was received previously for the particular cluster of bits. Note in this respect that not all reference levels are updated with the same speed, simply because they occur with different frequencies. Therefore, the adaptation time constant is level dependent.

![Figure 6.19: bER as function of target adaptation constant $\alpha$ for signals from disc E266 and E490. The oscillation in the curve for E490 is due to the limited number of detected errors.](image)

6.4 Single-Sided Cross-Talk Cancellation

In the 2D detection process side information is taken from the array of most recent bit decisions. However, in the first iteration no bit decisions are available yet. This means that side information is either not present or has a very poor quality for example when it is generated by simple threshold detection. This is reflected in the
fact that bER in the rows next to the boundary with unreliable side information is high (see Fig. 6.7). The results of these bit rows are not used. However, the unreliable side information also influences the bER of the output row adversely. A possible alternative that enables us to avoid using unreliable side information is cross talk cancellation (XTC). The principle of XTC has already been briefly discussed in Section 2.3.1. Here, we will start the discussion with a single-sided version of this scheme as shown in Fig. B.1, i.e. only using one side-row for compensation of cross talk.

![Diagram of cross-talk cancellation scheme](image)

**Figure 6.20:** Single-sided version of the cross-talk cancellation scheme.

The filtering is performed using a Finite Impulse Response (FIR) filter. The taps of this filter are denoted \( f_{p,n+1} \), and are adapted using a least mean square (LMS) algorithm based on some suitable criterion. We can write the XTC scheme as:

\[
\tilde{y}_{k,n} = y_{k,n} - \sum_{p} f_{p,n+1} y_{k-p,n+1} \tag{6.27}
\]

with \( y_{k,n} \) the read-out signal of row \( n \) at time instant \( k \). Signal \( \tilde{y}_{k,n} \) is the compensated signal.

Several criteria are possible to update the filter taps. One that is commonly used is a minimization of a measure \( \mathcal{M} \) for jitter given by:

\[
\mathcal{M}^2 = \begin{cases} 
\frac{(\tilde{y}_{k,n} + \tilde{y}_{k+1,n})}{2}^2, & \text{sign}(\tilde{y}_{k,n}) \neq \text{sign}(\tilde{y}_{k+1,n}) \\
0, & \text{otherwise}
\end{cases} \tag{6.28}
\]

As input to the LMS algorithm we take \( J(f) = \mathcal{M}^2 \) leading to a tap update equal to:

\[
f_{p,n+1}^{k+1} = f_{p,n+1}^k - \mu \frac{\partial J(f)}{\partial f_{p,n+1}} \tag{6.29}
\]

with \( \mu \) a suitably chosen adaptation constant and \( f \) the set of all filter taps. The final update rule, which is only carried out at a signal transition, can be derived from the
above equation as:

\[ f_{p,n+1}^{k+1} = f_{p,n+1}^{k} + \mu (\tilde{y}_{k,n} + \tilde{y}_{k+1,n}) (y_{k-p,n+1} + y_{k+1-p,n+1}). \]  (6.30)

The jitter criterion is suitable for moderate densities as in DVD and BD, but may not be optimal for higher densities (like extended versions of BD beyond 25 GB) where jitter is not a good performance measure. Furthermore, it must be applied synchronous to the bit clock (or at least the error signal must be derived in this synchronous domain). An alternative method could be to seek minimum correlation between the compensated, central signal and each of the signals from the adjacent tracks. This criterion can be used asynchronously. The input to the LMS algorithm can then be written as:

\[ J(f) = (\tilde{y}_{k,n} y_{k,n+1})^2. \]  (6.31)

The corresponding update rule is derived as:

\[ f_{p,n+1}^{k+1} = f_{p,n+1}^{k} + 2\mu \tilde{y}_{k,n} y_{k-p,n+1} y_{k,n+1}^2. \]  (6.32)

The term \( \tilde{y}_{k,n} y_{k-p,n+1} \) in the update can be recognized as the correlation between the compensated signal of the central track and a signal from the adjacent track. This means that the update will stop when the correlation becomes zero as we would expect. The term \( y_{k,n+1}^2 \) is less clear. It seems a weighing of the update with the signal power in the satellite signal. It appears that removing this weighing factor even improves the results (in terms of jitter). The removal of the factor is equivalent to a criterion equal to:

\[ J(f) = (\tilde{y}_{k,n})^2, \]  (6.33)

i.e. a minimization of the power in the signal after compensation. This criterion can be understood from the fact that the track signals are independent sources. An addition of independent sources (which happens in case of cross-talk) always increases the resulting power. Therefore, a power minimization automatically leads to minimum addition of the satellite signal to the central signal, i.e. minimum cross-talk.

For a practical implementation we have chosen to minimize the mean square error, where the error is taken as the difference between the actual result of cross-talk cancellation and the signal that we would expect after ideal cross-talk cancellation based on the target response. In Appendix B it is derived that a data aided LMS algorithm to minimize this squared error can be replaced by a non-data aided zero forcing (ZF) algorithm by scaling the resulting filter coefficients and adding a DC-term to the output signal. The big advantage of this scheme is that it does not need any preliminary decisions. The ZF based algorithm is applied in the first iteration of the stripe-wise detection at the low-certainty boundary according to the diagram in Fig. 6.21.
Figure 6.21: Scheme showing the first iteration of the stripe-wise detection including cross-talk cancellation. The thick black arrows indicate where the compensated signals are used. Note that the compensated signals are only used at the boundaries with low certainty.

As input to the receiver we use replay signals from disc E490 under nominal situations (no tilt). Before XTC the correlation between two adjacent rows is calculated. The result is plotted in Fig. 6.22(a). As expected the cross correlation reflects the target response that is used. Then the XTC is switched on according to the scheme of Fig. B.3 (see Appendix B for more details). The length of the FIR filter is taken equal to 8 (although a length of 4 suffices to obtain the bulk of the improvement). The result in Fig. 6.22(b) shows that indeed the algorithm forces 8 correlation coefficients to zero. The resulting impulse response of the filter is shown in Fig. 6.23.

Figure 6.22: Correlation between two adjacent rows (row 1 and row 2) for disc E490, after equalization to a target with $c_1=0.25$.

The bER improvement appears to be more than 1 order of magnitude. Fig. 6.24 shows the bER as function of row number. Only for row 1 and row 2 cross-talk cancellation is applied. Note that to benefit from cross-talk cancellation the weight
6.5 Performance Measurement

of the row closest to the uncertain boundary must be increased. Without cross-talk
cancellation the optimum weight \( w_j \) (see Eq. 6.20) is equal to \( \frac{1}{16} \). After applying
cross-talk cancellation the optimum weight increases to \( \frac{1}{4} \). Note also that, in most
cases, the rows that do not have cross-talk cancellation show an increasing bER for
larger weights. For row 3, 4 and 5 this is not true for a small increase of the weight.
This is due to the information passing mechanism from the very reliably detected top
rows, which mitigates the effect of the larger weight. The bER improvement can
be advantageously used in cases where ISI cancellation is done based on detected
bits from the first iteration to improve the performance of the second iteration.
Furthermore, the error calculation for the adaptation loops based on the bits from
the first iteration becomes much more reliable. This is very important when fast varying
channel parameters need to be tracked and decisions with a small detection delay
need to be available (see Section 5.5).

6.5 Performance Measurement

Conventionally, jitter is used as a good performance criterion of an optical storage
system, e.g. for optimization of media. However, jitter does not provide a direct
measure on the bit-detection performance of the system, i.e. it is not a direct indication
of bER. Furthermore, for high density systems jitter is not a good criterion.
In this section sequenced amplitude margin (SAM) is described as a performance
criterion that is a direct indication of bER and is suitable for high-density systems.
Figure 6.24: Bit error rate as function of row number. For row 1 and row 2 cross-talk cancellation was switched on. It is clear that in case of cross-talk cancellation the weight of the row at the uncertain boundary can (and should) be increased (in this plot the weights are taken equal to the value in the legend for all rows).

**SAM Analysis**  The principle of Viterbi detection is to find the path with minimum cost through the trellis. This is done by selecting at each time instant and for each state the predecessor that yields the minimum path metric. The error probability of the Viterbi detector is related to the probability that an erroneous path is selected. We can analyse this probability using the so-called Sequenced Amplitude Margin (SAM), which was first introduced in [81]. To understand this criterion let us assume an ideal 1D, linear channel response i.e. a response that perfectly matches the desired channel response. The theory for 2D is a straightforward extension of the 1D case although the use of a stripe-wise implementation has some consequences for the SAM analysis as will become clear later in this paragraph. The signal is only corrupted by sampled AWGN, denoted as $n_k$. The total Euclidian distance between the received signal and the desired signal can be written as:

$$\varepsilon = \sum_{-\infty}^{\infty} [r_k - (b \ast g)_k]^2.$$  \hspace{1cm} (6.34)

Note that in the Viterbi detector this ideal ‘maximum-likelihood receiver’-criterion is approached by making decisions with a certain decision delay such that most en-
6.5 Performance Measurement

The energy of the impulse response is included in the path metric. This path metric now resembles the Euclidian distance criterion. Bit detection is performed by selecting the data sequence \( \hat{b} = b \) that results in the smallest distance \( \varepsilon \). In case we select the correct data sequence, i.e. the transmitted sequence, the Euclidian distance is only determined by the noise samples:

\[
\varepsilon = \sum_{-\infty}^{\infty} [(b \ast g)_k + n_k - (b \ast g)_k]^2 = \sum_{-\infty}^{\infty} n_k^2. \tag{6.35}
\]

Now let us assume that one or more errors are made according to the sequence \( e = \frac{1}{2}(\hat{b} - b) \), with \( \hat{b} \) an erroneously detected bit-sequence. The Euclidian distance then becomes:

\[
\varepsilon_e = \sum_{-\infty}^{\infty} [(b \ast g)_k + n_k - ((b + 2e) \ast g)_k]^2 = \sum_{-\infty}^{\infty} [n_k - 2(e \ast g)_k]^2. \tag{6.36}
\]

A decision error will occur when the desired signal for the erroneously detected sequence gives a better match to the actually received signal than the desired signal for the transmitted data, i.e:

\[
\varepsilon_e \leq \varepsilon \iff \sum_{-\infty}^{\infty} [n_k - 2(e \ast g)_k]^2 \leq \sum_{-\infty}^{\infty} n_k^2, \tag{6.37}
\]

which can be rewritten as:

\[
\sum_{-\infty}^{\infty} (e \ast g)_k^2 + \sum_{-\infty}^{\infty} n_k (e \ast g)_k \leq 0. \tag{6.38}
\]

As a first term we recognize the signal energy resulting from the transmission of an error sequence \( e \) through the channel. It is called the squared Euclidian distance \( d^2 \) of the particular error sequence. The second term is linearly dependent on \( n_k \) which means that it has zero mean and variance \( \sigma_n \), related to the variance of \( n_k \). The value on the left part of inequality 6.38 is generally denoted SAM value:

\[
\text{SAM} = \sum_{-\infty}^{\infty} (e \ast g)_k^2 + \sum_{-\infty}^{\infty} n_k (e \ast g)_k \tag{6.39}
\]

Graphically the distribution of SAM values can be represented as in Fig. 6.25. The part of the normal distribution that is left of zero leads to an erroneous decision in the maximum-likelihood receiver. The relative area under the curve left from zero compared to the total area under the curve indicates the probability that an error occurs. This probability is denoted \( Pr(\text{error}) \). It is clear that if \( \sum (e \ast g)_k^2 \) has a small value, a large tail of the curve will be left from zero resulting in a high bER. Therefore, the
worst-case pattern that will govern the bER is the pattern that has the minimum value of $d_{\text{min}}^2 = \sum (e * g)^2$. For this pattern the error probability $Pr(e)$ can be calculated as:

$$Pr(e) = Q\left(\frac{d_{\text{min}}}{\sigma_n}\right)$$

(6.40)

where $Q$ is the function that defines the integral under the normal distribution as indicated with the gray area in Fig. 6.38. Now why is this analysis of the SAM distribution of importance to receiver design? There are four main reasons:

- as already mentioned the traditional criterion for optimization of recording media, jitter, is not valid anymore at high densities. The eye pattern is closed and zero-crossing information is not a reliable performance measure anymore. Therefore, it is often proposed to use SAM as an alternative [82, 83],

- the analysis of the SAM distribution offers the possibility to estimate the bER based on a relatively short sequence of processed data. Normally, about 100 errors are needed to obtain a statistically reliable bER estimate. For bER values in the order of $10^{-5}$ this means that approximately $10^7$ bits must be processed in order to achieve these reliable bER estimates. Especially in the case of Two-DOS where a rather complex detector is used, this leads to long simulation times. The SAM distribution offers a reliable bER estimate for much shorter, processed data sequences by fitting a normal distribution to the measured distribution and simply integrating the area under the distribution left of zero (note that it is implicitly assumed that the noise is well modelled with AWGN noise),

- by analyzing SAM distributions we are able to separate the intrinsic disc quality in terms of SNR or random errors from the occurrence of disc defects (such as dust and scratches). This is intuitively clear from the observation that disc defects generally cause a large signal waveform deviation compared to the desired waveform leading to large Euclidian distance values. By setting a suitable threshold such that SAM values below this threshold contribute to the fit of the normal distribution and SAM values above this threshold do not contribute,
it is possible to estimate bER that could be achieved in case the disc had no large defects. This procedure is shown schematically in Fig. 6.26. A related quality measure is the SAM-based signal-to-noise ratio (SAMSNR). It can be calculated from the estimated bER (SAMbER) according to [40]:

\[
\text{SAMSNR} = 20 \log_{10} \left( \sqrt{\frac{\sum (e + g)^2}{\sigma_n^2}} \right) \tag{6.41}
\]

\[
= 20 \log_{10} \left( \sqrt{2 \text{erfinv}(1 - 2 \text{SAMbER})} \right) \tag{6.42}
\]

with erfinv the inverse error function and the error function defined as:

\[
\text{erf}(x) = \frac{2}{\sqrt{\pi}} \int_0^x e^{-t^2} dt \tag{6.43}
\]

\[\text{Figure 6.26: Threshold method to distinguish random errors from disc defects.}\]

For TwoDOS this feature is very important in view of the problems with pit-moge as discussed in Section 7.1.3. Using SAM we are in a position to predict bER values which could be obtained in the case that pit-moge would be absent, although the situation for pit-moge is a bit more complicated than for large defects such as scratches. As it happens for pit-moge the defect is only a single missing pit. The detector simply detects the (partial) absence of this pit by choosing a trellis path that has a zero in this position instead of a one. This might still lead to small SAM-values that contribute to the error estimate. It is expected that a data aided SAM analysis solves this problem.

- The SAM values can be used in feedback control loops. The SAM values offer a direct criterion to estimate bER and they can be derived from the Viterbi detector by simple means. It is e.g. possible to directly minimize bER, by adapting, with a suitable algorithm, the equalizer taps based on the SAM values derived from the Viterbi detector [61, 62]. In this case it is important that the SAM values are derived with minimum delay.
To determine the SAM-values we determine at each time instant and for each state the difference between the optimum path metric (i.e. the one that is actually chosen) and the second best path metric leading to the same state. This can be done by adding registers that store for each state in the Viterbi the SAM value given by the difference in optimum path metric and the second best path metric during the ACS operation. In the backtracking procedure only one state is finally selected. From this state the output bits of the Viterbi detector are derived. In a very similar way the SAM value can be derived from this same state. Off-line a histogram of these SAM-values can be made. One thing that is observed immediately is that the SAM-values are all positive. This is no surprise because the Viterbi algorithm is designed to choose the minimum path metric. This means that this optimum path metric is always lower than the second best path metric and thus that their difference is always positive. What happens in fact is that the tail of the distribution is mirrored (and added) to the distribution right from zero. This is schematically depicted in Fig. 6.27. The problem of fitting the resulting curve is solved by simultaneously fitting two Normal distributions with the constraint that they are mirrored with respect to the line SAM=0. Another problem related to SAM, which is more specific for TwoDOS is the fact that we use a stripe-wise Viterbi detector where branch metrics are calculated over more than one row, while the final bit-decisions are derived from the most reliable row only. This means that the unreliable bits contribute to the SAM value, but do not influence the bER in a direct way. It is therefore not straightforward to do bER estimation based on SAM values from the stripe-wise Viterbi detector. For this reason it is proposed to execute one separate iteration of a simple 1D Viterbi detector that is working in a data aided mode (this also solves the fitting problem and the pit-moge issue as discussed above). In a non-data-aided mode one extra iteration of the Viterbi detector could be executed using the very reliable output of the previous iteration. In practical cases the error level is at the saturation level for this SAM-iteration, i.e. further iterations of the detector does not improve bER.

**Experimental SAM measurements** Experimental SAM measurements are performed by executing a third iteration of the 3-row stripe-wise Viterbi detector. The
distribution of SAM values for a 50 GB EBR-mastered disc (E490) are given in Fig. 6.28. Fitting the SAM values below a certain threshold leads to a SAMbER of 8.6E-6 and a corresponding SAMSNR equal to 12.67 dB. For the used target response and for the case of single bit-errors the mean value of the distribution is expected at 0.44. From the plot a higher value can be observed. This is due to the fact that the single bit-error is not the most frequently occurring error pattern. Error patterns in the form of Nyquist sequences as discussed in Section 4.2 appear to contribute significantly to the total number of errors and also to the SAM distribution. For example a double bit-error ‘+1-1’ is found at a nearly equal Euclidian distance of 0.48, a triple bit error has an expected mean value of the distribution at 0.56, etc. The total distribution is the weighed sum of all these individual distributions. It is expected that this also causes the slight underestimation of the bER, which in reality is approximately equal to 2E-5.

Fig. 6.29 shows the time-dependent behavior of the SAM values during acquisition of the receiver. The adaptive equalizer and reference level adaptation are switched on after 4 data frames (10240 bit periods). One can clearly see the increase in mean value. The variance does not change much in this case. A special equalization scheme that focuses both on the mean and the variance of the SAM distribution can improve detection performance considerably [61].
Figure 6.29: Transient behavior of SAM values during acquisition of the receiver (disc E490). The adaptive equalizer and reference level adaptation are switched on after 4 data frames.

6.6 1D-RLL Coded Data

Up till now Viterbi detection is applied to uncoded data. This means that all possible bit permutations within the states are allowed. In case we apply a run length limited (RLL) code to the data the number of permutations is limited due to the run length constraints $d$ and $k$. RLL sequences can be derived from $(dk)$ sequences. In a $(dk)$ binary sequence two logical ones are separated by at least $d$ and at most $k$ zeroes. The number of $(d)$ sequences with length $l$ can be found using the following recursive relations [7]:

\[
N_d(l) = l + 1, \quad 1 \leq l \leq d + 1,
\]
\[
N_d(l) = N_d(l - 1) + N_d(l - d - 1), \quad l > d + 1.
\] (6.44)

A run-length-limited (RLL) sequence is derived from such a $(dk)$ sequence by applying a modulo-2 integration to this sequence. i.e. the ‘one’s in the $(dk)$ sequence

---

$^1$The Viterbi detector is applied to data where the worst-case patterns are eliminated (see Section 4.3). The trellis of the Viterbi detector, however, is not modified to take advantage of this worst-case pattern coding.
denote the position of a transition in the RLL sequence. From a \((dk)\) sequence of length \(l\) we can derive an RLL sequence of length \(l + 1\) that can either start with a ‘one’ or a ‘zero’ (note that we can neglect the \(k\)-constraint as long as \(k\) is larger than the number of bits in a state of the Viterbi bit-detector). It is easy to verify that the number of RLL sequences of length \(l\) (and thus the number of states with \(l\) bits) can be written as:

\[ N_{RLL}(l) = 2N_d(l - 1). \]  

For example for a \(d=1\) code and a channel with memory length 4 (i.e. a 5-tap Viterbi detector, because the reference channel output can be calculated from 5 consecutive input bits) we find for the number of states \(|S| = 2N_1(3) = 10\). The corresponding state machine can be described with the graph of Fig. 6.30.

For the Viterbi processor with \(N_{Rstr}\) rows this means we have a total of \(10^{N_{Rstr}}\) possible states. In the TwoDOS format the reference signal is calculated over an area covered by 13 bits as indicated in Fig. 6.31.

\[ \text{Figure 6.30: State diagram according to the Mealy model corresponding to a 5-tap Viterbi detector for a } d=1 \text{ code (for unipolar channel bits).} \]

\[ \text{Figure 6.31: 2D bit configuration for calculating the reference signal in a 5-taps Viterbi detector.} \]
The index set associated with this target is:

\[ \mathcal{N}^{ Stap} = \{ (-1, 1), (0, 1), (1, 1), (2, 1), \\
( -2, 0), (-1, 0), (0, 0), (1, 0), (2, 0), \\
( -2, -1), (-1, -1), (0, -1), (1, -1) \} \]  

The procedure as described for the uncoded case can now be followed for \( M = 4 \). The trellis can be derived from the 1D state machine by going over all possible 2D states (with number \( \Sigma_{k,T} \)) and determine the sub-states (with number \( \Sigma_{k,1} \) and \( \Sigma_{k,2} \)) in each of the bit-rows. For each of these sub-states the predecessor state \( \Sigma_{k-1,1} \) and \( \Sigma_{k-1,2} \) can be determined, which together form the 2D predecessor state \( \Sigma_{k-1,T} \). For \( M = 4 \) the number of states per row is 10. Therefore, we can write (for a 2-row stripe):

\[
\begin{align*}
\Sigma_{k,T} &= \Sigma_{k,1} + 10 \Sigma_{k,2}; \\
\Sigma_{k,2} &= \left\lfloor \frac{\Sigma_{k,T}}{10} \right\rfloor; \\
\Sigma_{k,1} &= \Sigma_{k,T} - 10 \left\lfloor \frac{\Sigma_{k,T}}{10} \right\rfloor.
\end{align*}
\]  

(6.47)  

(6.48)  

(6.49)

For each of the states we now find the predecessor based on the state machine of Fig. 6.30 and we calculate the total predecessor state according to:

\[
\begin{align*}
\Sigma_{k-1,1} &= \text{Pred} (\Sigma_{k,1}) \\
\Sigma_{k-1,1} &= \text{Pred} (\Sigma_{k,1}) \\
\Sigma_{k-1,T} &= \Sigma_{k-1,1} + 10 \Sigma_{k-1,2}
\end{align*}
\]

(6.50)  

(6.51)  

(6.52)

**Local sequence feedback** The large number of states (\( 10^3 = 1000 \) in case of a 3-row Viterbi detector) again becomes unpractical. The local sequence feedback method as discussed before can again be applied in order to reduce the effective number of states.

**SAM Analysis for Coded Data** For coded data the histogram of SAM values has a somewhat different shape. Single bit-errors are still present in the form of transition shifts of the run-length limited marks. However, Nyquist sequences are not allowed anymore by the \((d,k)=(1,7)\) modulation code, simply because a data sequence that supports these kind of error patterns violates the coding constraints. A common error pattern for the coded sequence is a shifted-I2, where an ‘I2’ is a mark with a run-length of 2 channel-bit periods. A SAM histogram of a LIM-mastered disc (LIM 513) is shown in Fig. 6.32. For a manually optimized 13-tap 2D target response
the Euclidian weight of both a single bit-error and a shifted-I2 error pattern is 0.66. Fitting a Gaussian to the tail leads to an estimated bER of 1.6E-4, which is equivalent to a SAMSNR of 11.14 dB. The obtained bER at the output of the detector is equal to 8.2E-5. The long tail of the SAM distribution is probably caused by shifted versions of the longer run lengths, e.g. a shifted-I3 has Euclidian distance 1.08. These kinds of errors do not contribute significantly to the overall bER.

![Experimentally derived distribution of SAM values](image)

**Figure 6.32:** Experimentally derived distribution of SAM values (disc LIM513; BL=66.7 nm; TP=220 nm; no tilt).

### 6.7 Conclusions

A 2D bit-detector has been designed that is suitable for the TwoDOS format and allows practical implementation due to its manageable complexity. A full-fledged Viterbi detector covering the complete broad spiral would be preferable from a performance point of view, but would be far too complex to implement in a practical system. Instead a modified Viterbi detector is proposed that is composed of a number of Viterbi processors, each processing a subset of rows with the broad spiral. To approach the performance of the full-fledged 2D Viterbi detector the detected information is passed from one stripe processor to the next to condition the reference level calculation for this next stripe. It is shown that this stripe-wise Viterbi detector suffers a performance loss of only 1 dB with respect to the full-fledged solution.
Considering the very large complexity advantage this is thought to be acceptable. The stripe-wise solution is equally well applicable to 1D-RLL coded data, although the number of states tends to grow rapidly in this case (this is merely due to the longer channel memory, because of the shorter tangential bit-length). A solution can be found in local sequence feedback. A disadvantage of the information passing is the increase in detection delay for each next processor in the total chain of Viterbi detectors. This hampers the compensation for rapidly varying parameters in the total recording channel, especially for the inner rows of the broad spiral. This has been discussed elaborately in Chapter 5. Another problem in the stripe-wise approach is the unreliability of side information at one side of the stripe, especially in the first iteration. This problem can be alleviated by application of single-sided cross-talk cancellation, which improves bER after the first iteration considerably. This might be advantageous in case ISI cancellation is performed based on the output of the first iteration to improve the detection performance of the second iteration.

As an alternative to jitter it is proposed to use the sequenced amplitude margin as a good performance measure of the receiver and the Viterbi detector. This measure has some distinct advantages over the jitter criterion:

- It allows us to estimate bER based on short sequences of processed data.
- By analyzing SAM distributions it is possible to separate the intrinsic disc quality in terms of SNR or random errors from the occurrence of disc defects (such as dust and scratches).
- The SAM values can be used as a suitable criterion in adaptation loops since they provide a direct indication of bER at the output of the detector.

Some experimental examples of SAM distributions are shown for data on the hexagonal format and RLL-coded data.
Chapter 7

Experimental Results

Many experimental results have already been presented in previous chapters as ‘proof of principle’ for the various topics that have been discussed. In this chapter, however, some of the most important experimental results are presented. Overall system experiments will be described that prove the feasibility of the original project targets. Section 7.1 discusses some results related to disc technology. Especially the problems encountered when replicating small structures are addressed. Problems during disc replication occur for disc with a capacity of 50 GB (with the hexagonal format). It is shown that the replication problems can be solved elegantly by making use of a special ‘release’ layer during the replication process. Results of signal processing are presented in Section 7.2. These results demonstrate, the read-out of a 50 GB ROM disc with more than sufficient margins. This means that the initial goal of the work described in this thesis (a doubling of the capacity with respect to the 25 GB BD-ROM disc) has been achieved.

7.1 Disc Technology

In this section some results related to the manufacturing and replication of discs are presented.

7.1.1 Write Strategies for Mastering of RLL-Coded Data

In Section 2.4 it was explained how LBR and LIM can be used during the mastering process of stampers, which are used to emboss the pits in ROM discs. When the laser is modulated in a straightforward manner, i.e. by just switching it on or off (as shown in the top part of Fig. 7.1) it appears that the longer pits in case of RLL coded data become too wide in the radial direction. It is intuitively clear that a pit-width that is approximately equal to half the track pitch is optimal (due to a similar reasoning as for the hexagonal case, leading to a pit-hole size that covers 50% of the total bit-cell, see Section 3.3.2). Simply reducing the laser power to solve this problem will lead to pits that are too short, resulting in asymmetry (non-linearity) in the read-out sig-
A good solution is the application of write strategies. Because an electro-optical modulator is used in the mastering set-up, the laser light can only be switched completely on, or completely off at the position of the disc. Modulation of laser power in between these values is not possible. Therefore, power modulation is performed in the time domain. A write strategy according to the bottom part of Fig. 7.1 was chosen. A similar approach can be found in [22]. It consist of a first pulse, a repeated version of the so-called multi-pulse and a last pulse, which is preferably equal to the first pulse. These pulses are generated by increasing the resolution of the write-clock by a factor of 4 and by placing the transitions on this grid with this 4-fold increased resolution. The first and last pulse have a width of $3T/4$ (with $T$ the bit-interval); the multi-pulse have a width of $T/2$.

![Figure 7.1: Schematic representation of the laser pulses using no write strategy (top) and with the new write strategy applied (bottom).](image)

A striking experimental result of the application of a write strategy is shown in Fig. 7.2. The upper image shows a SEM photograph of disc 513. It is replicated from a master for which no write strategy was used. Although the pits have the correct length, it is clear that especially the longer pits are too wide. In between longer pits almost no land area is left. This leads to severe inter-track dependence of the diffraction from this type of structures and to non-linearity in the channel, which manifests itself by significant signal folding. The lower image shows disc 522 where the proposed write strategy was applied during the mastering of the stamper. One can observe a clear improvement. The pits have a homogeneous width and their lengths are still equal to the nominal values.
7.1 Disc Technology

7.1.2 Write Strategy for EBR mastering of the Hexagonal Format

Similar to write strategies for LIM mastering of RLL-coded data, it is also possible to increase the resolution of the write-clock in case of EBR mastering. With this feature it is possible to control the exact shape of the ideally circular pits. A resolution increase of a factor of 10 has been used. Initially, a duty-cycle of 6T/10 has been used for disc E266. This leads to elongated pits along the tangential direction (see the left part of Fig. 7.3), where the length of the pits is approximately 126 nm, while the width is around 82 nm. It appears that even a symmetrical pulse shape gives an elongated pit due to the linear movement of the disc during mastering. A duty-cycle of 4T/10 result in pits shown in the right part of Fig. 7.3 for disc E267. Here the pits are nearly circular (pit width ≈ 76 nm; pit length ≈ 98 nm). However, in this particular case they are also quite small and modulation of discs of this stamper is too low for proper read-out (although a power increase of 17% has been applied).

7.1.3 Pit Moge for 2D Hexagonal Discs

When very small structures are replicated, as in the case of TwoDOS, the pillars on the stamper (which emboss the pits in the disc) become very prone to mechanical forces. The strength of these pillars is proportional to the cross-sectional area of the pillars, which decreases quadratically when reducing the diameter of the pits. During the separation of the nickel stamper from the glass-substrate (see Fig. 2.14), forces act

![Figure 7.2: SEM image of a RLL-encoded 2D disc, replicated from a master that was made without write strategy (top) and a SEM image of a disc, for which a write strategy was applied during the recording of the master (bottom). In both cases tp=200 nm, and CBL=66 nm, resulting in 45 GB nett disc capacity (nett capacity means that the guard space is taken into account).]
Experimental Results

(a) Disc E266. Duty-cycle of mastering pulse is 6T/10. Pit length $\approx 126$ nm; pit width $\approx 82$ nm.

(b) Disc E267. Duty-cycle of mastering pulse is 4T/10. Pit Length $\approx 98$ nm; pit width $\approx 76$ nm.

Figure 7.3: SEM images of disc replicated from EBR-mastered stampers using write strategies with different duty-cycles.

on the pillars that are proportional to the surface area of the side-wall. As these forces decrease only linearly when reducing the diameter of the pits, at some moment the pillars show an increased probability to break off or wrench off during the stamper-substrate separation. A similar problem can occur when separating the disc and the stamper during the replication process, see Fig. 2.15. Indeed it appears that the problem of ‘wrench off’ occurs during the replication process for the EBR-mastered discs with a density of 50 GB ($a_H = 138$ nm; $b = 100$ nm). All EBR discs that are used in this work have been mastered by M. Furuki of Sony Corporation (Tokio, Japan). In Japanese this phenomenon is called ‘pit-moge’ (see Fig. 7.4). It appears primarily for pits with steep walls. Here, the separation forces are apparently larger than the strength of the pillars and the probability of pit-moge increases. Furthermore, pits with a high aspect ratio are more susceptible to pit-moge. This is no surprise because for deep pits the force on the surface during separation is larger. The SEM image

![Image](image_url)

Figure 7.4: Pit-moge: Translation from a Japanese-English dictionary.
shown in Fig. 7.5 shows the typical defects caused by pit-moge. The image shows the stamper E427 after a first replication. As the broad spirals should be identical, it is clear that some pillars are missing. It is not a-priori clear whether this damage occurred during the stamper fabrication process or that it occurred during the first replication. If the latter would be the case we would expect some nickel to be present on the first replicated disc, but local ‘electron probe micro-analysis’ (EPMA) could not reveal any traces of nickel.

![Image of damage caused by pit-moge on stamper E427](image)

**Figure 7.5:** Damage by pit-moge on stamper E427 after one replication step. *At the center-position in the white circles no pillar can be observed although there should be one present.*

Another example of pit-moge is shown in Fig. 7.6. This atomic force microscope (AFM) image of a replica of stamper E268 clearly shows bumps instead of pits. This can be explained by the fact that during break-off of the pillars from the stamper some extra material of the stamper surface was torn off. The newly formed irregular hole in the stamper causes a bump in the replica.

Not only the pillars on the stamper may break off, but it is also possible that, during the separation of the stamper and the replicated disc, the UV-curable polymer that is present between the pillars of the stamper may stick to the stamper and breaks from the disc. This causes large holes in the disc and it especially occurs at places on
Experimental Results

Figure 7.6: AFM image of a replica of stamper E268.

the disc with a large density of pits. There the structures between pits are extremely narrow and therefore most fragile. In Fig. 7.7 a SEM image of a replica of stamper E427 shows this type of defect. The proposed mechanism that causes the damage is schematically shown in the right part of Fig. 7.8.

Yet another artefact that can be observed from the photograph of Fig. 7.7 is that a small amount of debris is present always at the same side of most pits. The origin of this debris is thought to be the scraping of the pillars at the sides of the just-formed-pits during the separation of the disc from the stamper. This is shown schematically in the left part of Fig. 7.8. The debris indicates that the pillars have very steep walls with wall-angle $\alpha_w$ close to 90° (note that the drawing even shows an angle larger than 90°).

A more direct proof of the ‘steep wall hypothesis’ was found by cross-sectional SEM analysis of the first replica of stamper E427. Images are shown in Fig. 7.9. It is clear that wall-angle of the pits is larger than 90°. Yet, separation is possible due to the small shrinkage of the photo-polymer during the polymerization reaction. The pit-moge problem appears to be the most important hurdle to achieve densities larger than 35 GB on ROM discs with the hexagonal format.

Solution to Pit Moge The problems discussed above relate mainly to the relative increase in frictional forces between stamper and disc when separating them during the replication process. Therefore, a process is proposed to modify the surface of the stamper with a suitable ‘release’ layer. This layer should have a very good (chemical) bond with the nickel stamper, whereas it should have a very low adhesion to
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Figure 7.7: SEM image of a replica from stamper E427 showing the damage between the pits and the debris present at the side of the pits.

Figure 7.8: Schematic drawing of the mechanism that causes the debris at the side of each pit (left) and the mechanism that damages the land area between pits (right).

the UV-cured polymer layer of the disc. Preferably, this interface layer should be as thin as possible, in the order of a few nanometers, to prevent a change in pit diameter and to avoid an increase in media noise due to inhomogeneous thickness of this interface layer. In case of nickel surfaces, self-assembling monolayers of alkanethiols have been investigated in great detail in literature [84–86]. The used layer has a hydrophilic nature and it was expected that it would reduce the adhesion between the nickel stamper and the photo-polymer on the replicated disc. A modified version of
Figure 7.9: Cross-sectional SEM images of the first replica of stamper E427. The enlargements in the inset clearly show the large wall-angles.
7.2 Signal Processing

The layer has been applied to improve the stability of the layer. The method works very well. Fig. 7.10 shows SEM images of the stamper surface (E490; with a density of 50 GB) before and after replication. No pit-moge can be observed, even after multiple replications. The bER improves by two orders of magnitude from the 1E-3 range to the 1E-5 range.

7.2 Signal Processing

This section starts with the results of the measurement of tilt margins for various densities. Subsequently, we show an analysis of error patterns in detected data sequences from a 50 GB disc. The distribution of the errors appears to be asymmetrical. This can be very well explained with the results of the channel characterization.

7.2.1 Tilt Margin

Frequently used figures of merit for optical recording are the radial and tangential tilt margins. Tilt is defined in this thesis as the angle between the optical axis and the normal of the disc surface, see Fig. 7.11 (note that sometimes the tilt angle is defined as the angle between the incident and reflected beam. This results in values twice as large as the ones reported here).

In Fig. 7.12 the radial and tangential tilt margin of a 35 GB EBR-mastered disc (E266) is given. Tangential tilt margin is larger than \( \pm 0.6^\circ \), whereas the radial tilt margin is even larger than \( \pm 1.0^\circ \). This is much larger than the required tilt margins according to the Blu-ray Disc specification [87] which are \( \pm 0.15^\circ \) and \( \pm 0.3^\circ \) for radial and tangential tilt respectively (these are tilt margins allowed for the disc. The margins of the total drive are typically twice the mentioned values). For 50 GB the margins reduce slightly as shown in Fig. 7.13 but they are still large ( \( \pm 0.35^\circ \) for tangential tilt and \( \pm 0.63^\circ \) for radial tilt). This suggests that even higher densities should be possible. Moreover, the results are achieved using 2 and 3 local sequence feedback bits in the 2-row and 3-row Viterbi processors in the first and second iteration respectively. Simulations show that reducing the number of local sequence feedback bits further improve the bER and tilt margins.

7.2.2 Error Patterns

In Section 4.2 a simple theoretical analysis has been performed to search for worst case error patterns. A linear channel approach reveals that Nyquist-ring patterns are expected to be the most prominent error patterns (see Fig. 4.10). Generally, all 2D sequences of alternating ‘+1’ and ‘-1’ symbols have a large loss with respect to the matched filter bound. A further extension with a non-linear channel model showed a large asymmetry in the error patterns. For example a 1-shell cluster of all ones is more
Figure 7.10: SEM images of master E490 before and after replication using a self-assembling monolayer as a release layer. No ‘pit-moge’ damage can be observed after the replication.

likely (in terms of Euclidian distance) to have an erroneously detected central bit than a cluster with all zeroes. To check the validity of this analysis an error analysis has been performed on experimentally detected bit sequences from a 50 GB disc (E490). The stripe-wise Viterbi detector was used in combination with the receiver
as described in this thesis. It appears that the predicted Nyquist sequences actually are responsible for a large part of the errors on disc E490. Note that some Nyquist sequences, such as the closed-ring of 6 alternating symbols, do not occur, because they are eliminated by the modulation code as described in Chapter 4. Fig. 7.14 shows a selected set of errors that actually occur for data detected at a radial tilt angle of 0.3°. The number of erroneously detected ones does not deviate a lot from the number of erroneously detected zeroes, i.e. no asymmetry has been observed in this sense. Apparently, the channel is more linear than expected due to the fact that the pits are smaller than nominal (with a diameter of ∼80 nm instead of 102 nm).

It also appears that at nominal conditions (no tilt) most errors are single bit errors. For increasing values of tilt multiple-bit errors start to appear in the form of Nyquist sequences. The total number of ‘+1’ and ‘-1’ symbols in such a Nyquist sequence is denoted the length of such a sequence. The sequence length tends to become larger for increasing tilt angles. Table 7.1 reflects this observation. The large contribution of sequence length 10 under nominal conditions is a statistical anomaly that is caused by the low number of errors.

### 7.2.3 Error Distribution

The distribution of errors over the rows within the broad spiral can be very asymmetric. For the ideal case one would expect a low bER for the boundary rows and an increasing bER going inwards to the central row. This is due to the fact that in the stripe-wise bit-detection scheme detected bits from one stripe processor of the stripe-wise Viterbi detector, are passed as side information to a next processor. The detected bits may contain errors thereby decreasing the reliability of the output of the next stripe. However, readout of a 50 GB disc with the experimental player shows

---

**Figure 7.11: Definition of tilt angle.**
Table 7.1: Distribution of error patterns for various situations of disc-tilt. In the table the contribution to the total bER is given in percentages for error sequences of varying length. The sequence length is defined as the total number of ‘+1’ and ‘-1’ symbols in the Nyquist sequence.

<table>
<thead>
<tr>
<th>Sequence length</th>
<th>nominal</th>
<th>( \alpha_t = \alpha_r = 0 )</th>
<th>( \alpha_t = 0.3^\circ )</th>
<th>( \alpha_t = 0.5^\circ )</th>
<th>( \alpha_t = 0.5^\circ )</th>
<th>( \alpha_r = 1.0^\circ )</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>29.5</td>
<td>7.4</td>
<td>3.7</td>
<td>7.4</td>
<td>3.6</td>
<td></td>
</tr>
<tr>
<td>2</td>
<td>9.1</td>
<td>9.1</td>
<td>10.8</td>
<td>3.7</td>
<td>9.6</td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>20.4</td>
<td>12.0</td>
<td>16.3</td>
<td>23.9</td>
<td>15.0</td>
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<td>4</td>
<td>18.2</td>
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<td>14.7</td>
<td>10.6</td>
<td></td>
</tr>
<tr>
<td>5</td>
<td>0.0</td>
<td>12.8</td>
<td>13.0</td>
<td>15.3</td>
<td>10.1</td>
<td></td>
</tr>
<tr>
<td>6</td>
<td>0.0</td>
<td>5.1</td>
<td>10.7</td>
<td>3.7</td>
<td>13.0</td>
<td></td>
</tr>
<tr>
<td>7</td>
<td>0.0</td>
<td>8.0</td>
<td>11.5</td>
<td>21.5</td>
<td>8.5</td>
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</tr>
<tr>
<td>8</td>
<td>0.0</td>
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<td>4.6</td>
<td>9.8</td>
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<tr>
<td>9</td>
<td>0.0</td>
<td>2.6</td>
<td>3.5</td>
<td>0.0</td>
<td>5.5</td>
<td></td>
</tr>
<tr>
<td>10</td>
<td>22.7</td>
<td>5.7</td>
<td>3.2</td>
<td>0.0</td>
<td>5.4</td>
<td></td>
</tr>
<tr>
<td>11</td>
<td>0.0</td>
<td>3.1</td>
<td>3.1</td>
<td>0.0</td>
<td>3.7</td>
<td></td>
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<tr>
<td>12</td>
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<td>3.4</td>
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<td>0.0</td>
<td>4.0</td>
<td></td>
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<tr>
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<td>0.0</td>
<td>7.4</td>
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<td>0.0</td>
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<td></td>
</tr>
<tr>
<td>14</td>
<td>0.0</td>
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<td>0.0</td>
<td>0.0</td>
<td>0.9</td>
<td></td>
</tr>
<tr>
<td>&gt;14</td>
<td>0.0</td>
<td>0.0</td>
<td>2.2</td>
<td>0.0</td>
<td>1.0</td>
<td></td>
</tr>
</tbody>
</table>
asymmetric error distributions as shown in Fig. 7.16 for various tilt conditions. The cause of this asymmetry becomes clear when performing channel estimation. Channel estimation is done based on a mean square error (MSE) criterion. A block diagram of the used configuration is shown in Fig. 7.15. Estimation is carried out in the synchronous domain. The output of the bi-linear model $y_{k,n}$ (according to Eq. 3.29) is compared to the synchronized replay signal $r_{k,n}$ with $k, n$ the indices for tangential (column) and radial (row) position, respectively. The error signal is weighed with a constant $\mu$ that determines the convergence speed of the loop. The weighed error signal is correlated with the known binary data and the result of this correlation is integrated with an ideal integrator and added to the linear and bi-linear coefficients of the model. Results of this model reveal two main reasons for the asymmetric error distribution. Firstly, there seems to be a small misalignment of the grating angle. This causes a linear variation in track offset when going over the rows in the broad spiral. Fig. 7.17 shows the estimated linear part of the channel response for rows 2, 4, and 6 in case of a tangential tilt angle of $-0.5^\circ$. When connecting the centers of the channel profiles a deviation in grating angle of about $0.5^\circ$ becomes clear.

A second, more serious, cause of the asymmetry is radial tilt. As already became

\textbf{Figure 7.12:} Radial and tangential tilt margins of a 35 GB EBR-mastered disc (E266).
clear in Chapter 3 (see Fig. 3.10), tilt of the disc causes the spot to widen slightly on one side, and causes some serious side lobes in the airy intensity profile at the other side. It appears that in case of positive tilt the main side lobe is still on the broad spiral for spots on row 1 to 5, see Fig. 7.19. This side lobe causes severe intersymbol
Figure 7.15: Configuration for channel estimation based on a mean square error criterium.

Figure 7.16: Error distribution over the rows for actual error sequences measured from a 50 GB disc (E490), for various tilt angles.

interference. A corresponding high bER is observed. For spot 6 and 7 the main side lobe is not on the broad spiral anymore and bER drops by more than 2 orders of magnitude. This large drop can only occur in the test format where individual broad spirals are placed at a very large separation i.e. the width of the guard band is much larger than the row-distance. For the actual format with a guard band in the order of one bit row the large side lobe will cause cross talk between adjacent broad spirals. This will affect bER adversely and cross-talk cancellation will most certainly
Experimental Results

Figure 7.17: Results of linear channel estimation showing a clear deviation of the grating angle. Estimation is based on the measurement at a tangential tilt of $-0.5^\circ$.

be necessary to achieve reliable read-out. In that case additional spots on the guard band are needed to generate the signals as input to the cancellation scheme.

For negative values of radial tilt a mirrored version of the asymmetry profile is expected. Although the drop in bER is less clear this is indeed the case as shown in Fig. 7.18.

7.3 Analysis of Residual ISI

7.4 Conclusions

In this chapter some important experimental results have been presented that were not discussed in previous chapters. A first result relates to the manufacturing of ROM discs using LIM. A write strategy is adopted for the mastering process in case of a TwoDOS format with run length limited coding along the bit-rows. This improves
Figure 7.18: Error distribution over the rows for actual error sequences measured from a 50 GB disc (E490) in case of positive and negative radial tilt.

the linearity of the channel considerably. For a conventional mastering strategy the marks are not uniform in width and the long marks become too wide. By applying a pulsed strategy the width of the marks can be made independent of the run length, leading to improved linearity of the channel.

A second important result is the improvement of the replication process of discs from a master in the case of the hexagonal TwoDOS format. The process to replicate ROM discs becomes more and more difficult for higher densities. The pillars on the stamper that are needed to impress the pits in the medium decrease in diameter and as a consequence become more vulnerable to wrench off during the separation of disc and stamper. This is especially true in the case of the hexagonal format, where pits are significantly smaller compared to the RLL format. Various types of disc defects were observed, such as:

- missing pits due to wrenched off pillars on the stamper,
- debris at the side of the pits, attributed to the scraping of the pillars at the side walls of the newly formed pits with steep walls,
- missing land areas between pits caused by sticking of the UV-curable polymer to the stamper. This sticking is also thought to be caused by the high wall angle of pillars and pits.

All these defects are called ‘pit-moge’. The issue of pit-moge severely hampered an
Experimental Results

Figure 7.19: Results of a linear channel estimation for different rows showing that for row 6 and 7 the large side lobe falls outside the broad spiral causing a considerable improvement in bER.

increase in capacity above 35 GB. The problem can be solved by the application of a monolayer thick ‘release’ layer that has a low-friction property and reduces adhesion at the surface. Replication using this layer is very successful and read-out of a 50 GB disc becomes possible at excellent tilt margins: $\pm 0.35^\circ$ for tangential tilt and $\pm 0.63^\circ$ for radial tilt.

The error patterns in the case of 50 GB appear to consist of 2D sequences of ‘+1’ and ‘-1’ symbols. This was theoretically predicted based on matched filter bound calculations. The somewhat asymmetric error distribution as function of row number, especially at large radial tilt angles, could be attributed to the specific format of the 2D disc. The large side lobe that arises in the airy intensity profile of the read-out spot in case of tilt is not on the broad spiral for some of the rows within the broad spiral. Instead it is located on the large guard space that was present in the test format. In
the final format, where the guard space has a width approximately equal to one bit row, cross-talk is expected to occur between broad spirals. This will affect the bER adversely. In order to maintain similar radial tilt-margins as in the case of the wide guard band of the test format, cross talk cancellation techniques can be applied for reliable read-out.
Chapter 8

Discussion and Prospects

In this chapter the state of art of optical storage is sketched (Section 8.1). For the next generation of optical storage, TwoDOS is considered to be a technology option that can be applied orthogonally to improvements in the physics of the storage system. In Section 8.2 some hints are given on possible application of TwoDOS as part of the candidate 4th generation storage systems. In the discussion also counter-arguments against the use of 2D technology are taken into account. Section 8.3, finally, discusses the lack of a clear market application for the huge amount of storage capacity that will become available. It also gives an example for alternative product approaches that deal with this issue.

8.1 Battle for the Third Generation Format

At the time of writing of this thesis the 3rd generation optical recording standard has been set, although there are still two competing formats. The first one is Blu-ray Disc (BD). It was formulated by a consortium of nine manufacturers called the ‘Blu-ray Disc Founders’. These companies include Philips, Sony, Matsushita, Hitachi, LG Electronics, Pioneer, Samsung, Sharp, and Thomson. The standard is based on a blue-violet laser diode with a wavelength of 405 nm. The NA of the optical system is 0.85. Three capacities are defined on a single-sided, single-layered disc: 23.3 GB, 25 GB, and 27 GB. The format supports a data rate of up to 36 Mb/s, which is enough for recording high-definition video in MPEG-2 format, which consumes about 25 Mb/s.

Sony was the first to commercialize the BD standard by introducing a video-recorder product (BDZ-S77) in 2004. It is capable of recording high-definition TV broadcasts using its built-in satellite tuner. Recording is done on discs in a cartridge or caddy using phase-change technology (see Chapter 2). The caddy is needed to protect the disc (with only 0.1 mm protective cover layer) from dust and scratches (see Fig. 8.1). Many other companies, including Philips, have working prototypes and will introduce first products soon. Improvements in cover-layer technology such as protective hard-coats have made the use of a caddy for the disc superfluous for most applications.
The second, competing format is called advanced optical disc (AOD). It is created by Toshiba in cooperation with NEC. The AOD format is also based on a 12cm-disc and uses the same blue-violet laser. The main difference with BD is the NA of the system. The AOD founders propose an NA equal to the one of DVD and a substrate of 0.6 mm instead of 0.1 mm with the arguments that a caddy could be omitted, discs could be manufactured cheaper, and the system will be more robust. The lower NA has direct consequences for the storage capacity. This will be 15 GB and 30 GB for a single and dual-layer disc respectively.

8.2 The Use of TwoDOS Technology

In research the focus shifts to a next generation of optical storage [88]. History has shown that each new generation of optical storage offers about a factor of 5 to 6 more storage capacity and a factor of 2 to 3 higher data rate (see Chapter 1). Extrapolation of this trend results in an expected capacity of about 150 GB per layer and a data rate of about 100 Mb/s for the 4th generation of optical storage. These high density and data rate requirements impose quite some technological challenges in the research work on 4th generation of optical storage. Until now the main improvements in density and data rate have been achieved by decreasing the wavelength and increasing the numerical aperture (NA). Continuing along this path is not self-evident for the following reasons:
8.2 The Use of TwoDOS Technology

- The wavelength of the laser diode depends on the bandgap of the material that is used. Today’s blue-ultraviolet lasers are based on gallium nitride (GaN) [89]. It is not expected that a current-driven semiconductor laser diode with a wavelength considerably shorter than 400 nm, a sufficient operating lifetime, and enough light output will become commercially available in the coming years [90]. Moreover, if this will happen, it will certainly take many years before the price has dropped to levels that become interesting for integration in optical recording systems.

- The NA for BD is equal to 0.85. For a far field optical recording system, where focusing is done in air, the maximum attainable value is NA=1. This means that only a small improvement is theoretically possible.

The far-field NA-limitation may be circumvented by going to near-field (NF) optical storage where focusing is done inside a material with high refractive index (see Section A.1.1). A storage capacity of approximately 150 GB per disc is then possible. One of the main issues for industrialization of this technology will be robustness of the system. The evanescent coupling of the light from the objective lens towards the disc at an NA larger than 1 requires working distances in the order of tens of nanometers. A protective caddy for the disc seems unavoidable. Another major issue is the read-out time of a complete disc, which amounts to 25 minutes (see Section 1.4). Application of TwoDOS technology, even with only 3-5 rows per broad spiral, can solve this issue.

An alternative approach for the 4th generation is to apply advanced signal processing to achieve 35 GB per storage layer (for 1D detection) or over 50 GB per storage layer (for 2D detection using TwoDOS technology) and to develop a multi-layer system using BD optics (see Section A.1.2). Such a multi-layer system would approach (in case of 4 layers) or even exceed the disc capacity achieved by NF-storage. A strong competition between both approaches can be expected.

An outsider in the next-generation technology competition is the ever promising holographic storage (see Section A.1.6). The intrinsic 2D nature of the technology requires a 2D signal processing approach. The techniques developed for TwoDOS can be applied in that case.

Besides the obvious advantages of increased data rate and increased density and the orthogonal applicability (see Section 1.5), there are, however, also counter-arguments that point at some disadvantages of the 2D system. These include:

- increased complexity of the digital signal processing, especially in the Viterbi detection. The total silicon area required for implementation of the 2D signal processing increases approximately linearly with the number of rows in the broad spiral. For the detector the complexity depends mainly on the number of branches which need to be evaluated at each decision instant. A rough estimate
is derived from the calculation shown in Table 8.1. Here a true 2D system taking the complete 2D ISI into account is compared to multi-track 1D system with the same number of rows ($N_r$). For the 1D system a 6-tap Viterbi detector is chosen, which is capable of detecting data from a 35 GB disc. The 2D system uses 3-row stripe-wise processors and a 3-bit local sequence feedback (LSFB) (see Chapter 6), which is capable of detecting data from a 50 GB disc. For an 11-row 2D system an increase in complexity of the detector of a factor of 6 is estimated compared to a 1D multi-track system. Here, it is taken into account that for the 2D system with 3-row stripe-wise processors the branch metric calculation is the sum of 3 contributions (this is indicated in Table 8.1 in the row ‘relative complexity per branch’). The increase in complexity is deemed acceptable because Moore’s law predicts that the number of transistors and thereby the signal processing power doubles about every 18 months [91]. At the moment a 4th generation product will be introduced in the market the advance in IC process technology has compensated for the complexity increase due to 2D processing. Moreover, we anticipate that there is much room for improvement left (e.g via 2D phase equalization, ISI cancellation and cross-talk cancellation). Ultimately the difference is likely to be modest.

- power dissipation of the system’s main printed circuit board (PCB) increases due to the more complex digital processing. Again one can argue that Moore’s law will solve this problem to a large extent. Unfortunately, a large increase in power consumption is also unavoidable on the optical pick-up unit (OPU). In state-of-art systems the OPU contains the analogue pre-processing in the form of a photo-diode IC (PDIC) and a laser driver IC (LADIC). Both ICs consume a considerable amount of power due to the required high bandwidth in the read-channel and the high operating current of the high-power lasers needed for writing. For a 2D system both the PDIC and the LADIC must have a parallel

<table>
<thead>
<tr>
<th></th>
<th>1D Viterbi detector (35 GB)</th>
<th>2D Viterbi detector (50 GB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of branches</td>
<td>26 (6-taps)</td>
<td>64 (3-row, 3b LSFB)</td>
</tr>
<tr>
<td>Number of processors</td>
<td>$N_r$</td>
<td>$N_r \cdot 2$</td>
</tr>
<tr>
<td>Some Examples</td>
<td>branches</td>
<td>branches</td>
</tr>
<tr>
<td>N=4</td>
<td>104</td>
<td>128</td>
</tr>
<tr>
<td>N=5</td>
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<td>192</td>
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<tr>
<td>N=11</td>
<td>286</td>
<td>576</td>
</tr>
<tr>
<td>relative complexity per branch</td>
<td>1</td>
<td>3</td>
</tr>
</tbody>
</table>
8.3 Future Applications

The application areas for CD (audio) and DVD (video) have been very clear. Also for BD (high definition video) there seems to be no discussion on what to do with the huge amount of storage capacity. The next generation optical storage, however, suffers from a lack of such an obvious application area in the consumer electronics market. No consumer application requires such large amounts of storage (of course application in the PC-domain as a medium to archive data is always possible). This has paved the road for other product approaches. One is to decrease the disc size and the drive size to allow portable applications in mobile phones and portable gaming consoles.

One of the first to propose a small disc was DataPlay [92–94]. Their disc, with a diameter of 32 mm, offers a storage capacity of 350 MB per side, which is real-
ized with a wavelength of 650 nm and a NA of 0.65. Also Philips has proposed a small form factor optical disc (SFFO) as a storage medium (see Fig. 8.2). A small drive based on miniaturized BD optics (NA=0.85; \( \lambda = 405 \) nm) is able to read out a medium with a diameter of about 30 mm and a capacity of 1 GB [95, 96]. A similar device is already present in Sony’s portable play station (PSP). The UMD (universal media disc) offers a capacity of 1.8 GB on a diameter of 60 mm. The disc is read out at an NA of 0.6 with a 660 nm red laser. It is a read-only system that is used for distribution of games and movies. Compared to the conventionally used masked ROM technology the small form factor optical disc technology offers a higher capacity, short turnaround time for manufacturing and low media costs. Application of TwoDOS in these portable storage systems allows a higher storage capacity (or equivalently an even smaller system) and a higher data rate at the cost of a higher complexity and higher power dissipation. An additional advantage for the distribution of games would be that the track-pitch of the 2D system can be engineered such that recording of such a disc will become very difficult.

In general the research work on TwoDOS has initiated a large number of innovations. A number of these innovations will certainly be part of future storage systems.

Figure 8.2: Photograph of a small form factor optical disc system able to read out discs with a storage capacity of 1 GByte.
Appendix A

Optical Storage Technologies and Trends

This Appendix gives an overview of optical storage technologies that potentially offer an increase in capacity and data rate over the state of the art in optical recording systems that are on the market. First the trends in storage density are discussed. To be able to continue the trend in the future several technologies are discussed, such as near field recording (Section A.1.1), multi-layer media (Section A.1.2), multi-level modulation (Section A.1.3), and various super-resolution methods (Section A.1.4 and A.1.5). A last method is the application of advanced signal processing (Section A.1.7).

A.1 Technologies to Increase Density

Figure A.1 shows the trends in density of various storage media. At the time digital optical storage started with the CD format its storage capacity and density seemed to be enormous. However, around 1996 the density (and capacity) of hard disk drives (HDD) came up to the level of CD. The HDD line is a continuous one because the storage medium is not bounded by standardized formats and every technological improvement can be directly applied. In the same period DVD was introduced together with its dual layer version (DVD DL). The most recently standardized format is BD dual layer. A complete overview of the trends in storage density is shown in Fig. A.1. The figure shows that Magneto-optical (MO) storage (see Section A.1.4) lags slightly with respect to the normal optical storage road map. It is mainly used for professional archiving. Furthermore, flash becomes a strong competitor for optical storage. Although the storage density is smaller, its robustness and small size, make it the favored storage medium for portable applications.

Further improvements can be expected from various technologies, which are discussed in more detail in this section. A short technical explanation is followed by some pro’s and con’s of the technology.
A.1.1 Near-Field Recording

Optical storage density has increased considerably over the past 10 years in going from CD, via DVD to BD. In every step an evolutionary approach was taken by decreasing the wavelength and by increasing the numerical aperture of the objective lens. These systems use a so-called cover-layer-incident configuration in which the data storage layer is protected by a cover layer with refractive index $n$ (see Fig. A.2b). Due to refraction of the light at the air-cover layer interface the angular extent of the focused light beam is increased such that the internal numerical aperture $NA' = NA/n$. However, simultaneously the wavelength in the substrate also decreases by the same factor $\lambda' = \lambda/n$ such that the focussed spot size does not change compared to the air-incident configuration as shown in Fig. A.2a. This means that for an optical system where the objective lens is located a certain distance from the medium (far-field optical recording) the optical resolution is limited by the maximum angular extent of the focussed light cone i.e. $\theta_{\text{max}} = 90^\circ$.

From the reasoning above one can infer that if we would be able to focus the light spot within a material with refractive index $n$, which is higher than the refractive index of air, without a change in angular extent of the focussed light cone we would be able to increase optical resolution. Indeed this is possible by the use of a solid immersion lens (SIL). Such a SIL consist of a half sphere centered at the data storage
A.1 Technologies to Increase Density

Figure A.2: Configurations for far-field optical recording.

layer. All light enters at right angles at the SIL’s surface and passes without refraction. Hence, the SIL’s operation principle is to reduce the effective wavelength by a value \( \lambda' = \lambda/n \), while leaving the numerical aperture unaltered (\( NA' = NA \)). This means that total optical resolution improves by a factor \( n \) equal to the refractive index of the SIL. Again an air-incident configuration can be used (see Fig. A.3a) or a cover layer protected disc can be used. In the latter case the SIL reduces to a cut section of the sphere.

Figure A.3: Configurations using a solid immersion lens (SIL).

Until now it was assumed that the SIL is in close contact with the data storage layer or the cover layer. To allow a free rotation of the disc however, a small air gap is necessary between the SIL and the medium as indicated in Fig. A.3. Now, at the
bottom surface of the SIL Snell’s law will hold which means that:

\[ n \sin(\theta) = n_{\text{air}} \sin(\theta). \]  (A.1)

In practice the left-hand of the equation will become larger than 1 (for example \( \sin(\theta) = 0.85 \) and \( n = 2 \)). At the right-hand side of the equation the refractive index of air is equal to 1, which means that the sine of the angle should increase to values larger than 1, which is physically impossible. What will happen is that rays with \( n \sin(\theta) > 1 \) show total internal reflection (TIR) within the SIL. These rays will not reach the storage layer and still the optical resolution is not increased. However, when solving the Maxwell equations at the boundary for the case of total internal reflection one finds that the electric field is protruding from the bottom of the SIL: the so-called evanescent field. This field drops off exponentially with increasing distance from the SIL’s surface. By bringing the substrate or storage layer of the disc close to the bottom of the SIL the light will propagate by means of this evanescent field across the air gap. For a propagation with high efficiency (say in the order of 70-80%) the air gap should be smaller than \( \approx \frac{\lambda}{10} = 40 \text{ nm} \).

A further increase in resolution is possible by using a super SIL, which in contrast to a normal SIL shows refraction at the entrance surface of the light. By choosing a hemispherical surface such that the SIL’s total height is equal to \( r + \frac{r}{n} \) with \( r \) the radius of the SIL and \( n \) its refractive index one can achieve another factor of \( n \) in optical resolution making the total increase equal to \( n^2 \). Without the super SIL the focal point would have been at a distance \( r + nr \) below the SIL’s top surface. The configuration with a super SIL is shown in Fig. A.4.

![Figure A.4: Configuration with a super SIL.](image-url)

An obvious disadvantage of the system seems to be the very small free working distance of only tens of nanometers. Initially, hard-disk technology in the form of air-bearing sliders was applied for this purpose \([?]\). This appeared to be rather
cumbersome in view of robustness of the system and tolerance to dust and scratches. Furthermore, the air gap determined by the flying-height of the slider is dependent on the linear velocity of the flying head with respect to the disc [97].

A breakthrough has taken place by switching to an actuated head containing the SIL [98]. Here, conventional optical actuators (like the ones commercially used for DVD) are applied to become a remarkable reliable and robust system (operation times of months in a non-clean room environment without crashes between the optical head and the disc are reported []). The light that is reflected at the bottom of the SIL is used to generate a gap error signal (GES) by selecting the proper polarization of the light. This signal is high when no disc is present, i.e. light with \( n \sin(\theta) > 1 \) is reflected. When approaching the disc more light is coupled into the disc's substrate causing the GES to drop. The GES is used in a feedback control loop that reduces the residual error to values as low as \( \pm 2 \text{ nm} \). Special repetitive control procedures can be applied to handle the large axial run-out of the disc, which is in the order of \( 20 \mu\text{m} \).

Other issues are that the substrate material needs at least a comparable refractive index as the SIL material. Currently, a search is done for very high-index cover layer materials. Similarly, in case of multi-layer near field applications very high-index spacer materials need to be applied that at the same time are suitable for future mass production of the multi-layer discs [99] (for example a spin-coating process might be used to apply these materials).

In spite of the above described issues, near field recording is a promising candidate as a successor to BD.

### A.1.2 Multi-Layer or Double-Sided Discs

During the standardization of DVD also double-layer and double-sided standards were included (see Fig. A.5). This allowed storage of 8.5 Gb for a double-layer DVD disc and up to 17 GB for a double-layer, double-sided DVD disc. Also recordable double-layer systems have been realized [100, 101] and are now commercially available. A big issue for this development is the backwards compatibility with the existing double-layer DVD ROM in view of the high absorption of the recording materials involved. Other issues are coherent inter-layer cross-talk and the compensation of spherical aberration which is induced by focusing on layers at different depths in the medium.

Also in the case of BD, the use of more layers is a possible way to increase the storage capacity of a disc. Research is done to 4 or even 8 layers of data [102]. Generally, signal-to-noise ratio is a big issue due to the small reflectivity of each of the layers. Therefore, the combination of many layers and a high data rate seems to be cumbersome.
A.1.3 Multi-Level Recording

Virtually all practical (optical) recording systems use a binary alphabet, which means that at most 1 bit of information can be conveyed per symbol interval. This does not mean that no research has been done to systems with larger data alphabets. In case of high SNR, larger alphabets can increase channel capacity to come close to the so-called Hartley-Shannon limit [59]. In optical recording, Calimetrics has proposed systems with more than two levels [103]. A rewritable system is realized by recording very small marks using phase change technology (as will be discussed in Chapter 2). The multi-level signal is obtained due to the low-pass behavior of the optical channel [104], i.e. due to the fact that the marks are much smaller than the size of the optical spot. For ROM systems the reduction in mark size is difficult because the pits have to be mastered and replicated. To overcome this problem pit-edge modulation has been proposed. With this technique, denoted SCIPER [105, 106], a multi-level signal is generated by shifting the rising and falling edge of the binary modulated signal in discrete steps during mastering of the disc. Another option is to modulate the pits in radial direction as discussed in [107]. Also pit-depth modulation has been proposed [108], but this is not very practical, because pit-depth is normally determined by the thickness of the resist during mastering, which is uniform over the master. Another problem with the multi-level approach is that it is not compatible with the existing formats. An approach to circumvent this problem is presented in [109], where a limited multi-level (LML) format is proposed. Here, only the longest run lengths of a run length limited (RLL) format are modified to give multiple signal amplitudes. This is done by a short reduction in illumination level during the mastering of the long run lengths. The additional storage capacity (24% for the CD format if run lengths larger than 5 are used to store an LML bit) can be considered as an add-on channel, that does not influence the performance of the main data channel.

Multi-level recording never became a market success due to the lack of cost-
effective modulation methods that work in a combined ROM/rewritable system, and probably due to the fact that a multi-level modulation technique requires a high-SNR recording channel. If the latter is the case, the conventional path of squeezing bit-length and/or track pitch is mostly followed until the SNR reached a level where a larger data alphabet can not further increase capacity significantly.

A.1.4 Magnetic Super Resolution: MSR

MO Recording Magnetic super resolution is possible in magneto-optic (MO) recording. MO recording is based on a thermo-magnetic process. For a comprehensive overview on MO recording one is referred to [110]. For the storage medium a rare earth-transition metal (RE-TM) alloy is commonly used. The RE-TM alloys are ferrimagnetic with the RE magnetization anti-parallel to that of the TM magnetization. The net magnetic moment is the vector sum of the magnetic moment of each of the components. Both the RE and TM magnetic moments decrease with temperature caused by the increasing thermal disorder, which competes with the exchange coupling that tends to keep the atomic dipole moments aligned. This relation is shown in Fig. A.6a. At a temperature $T=0$ K the magnetization of the RE material is larger than the magnetization of the TM materia and the net magnetization is parallel to the RE direction. However, at increasing temperatures the magnetization of the RE material drops faster than that of the TM material. At the so-called compensation temperature $T_{\text{comp}}$, both materials show the same absolute magnetization and the net value $M_s$ approaches zero. For even higher temperatures, the TM magnetization is larger than the RE magnetization. The saturation magnetization can be calculated by taking the absolute value of the difference between RE and TM magnetization (the absolute value stems from the fact that the sub-lattice with highest magnetization is being aligned with an external field to minimize the total energy of the system):

$$M_s = |M_{\text{RE}} - M_{\text{TM}}|.$$  \hfill (A.2)

Normally, for MO-recording the material shows positive uni-axial anisotropy (i.e. the preferred magnetization direction is perpendicular to the recording layer). This anisotropy is expressed in the anisotropy energy $K_u$, needed to rotate the magnetization from the easy-axis (preferred magnetization direction) to the hard axis (non-preferred magnetization direction). $K_u$ appears to decrease approximately linear with temperature to zero at the so-called Curie temperature $T_c$. At the compensation temperature $M_s$ is zero, while $K_u$ still has a finite value. Therefore, the coercivity (i.e. the threshold for the external magnetic field to switch the magnetization direction) $H_c$ approaches infinity (see Fig. A.6b):

$$H_c = \frac{2K_u}{\mu_0 M_s}.$$  \hfill (A.3)
with $\mu_0$ the permeability of free space. It is this property that allows the formation of very small and thermally stable magnetic marks. Normally, in magnetic recording magnetization tends to decrease as a function of time with a certain time constant $\tau$ according to:

$$M(t) = M(0)e^{-\frac{t}{\tau}}$$  \hspace{1cm} (A.4)

with the time constant $\tau$ equal to:

$$\tau = 10^{-9} e^{\frac{E_{\text{switch}}}{kT}}.$$  \hspace{1cm} (A.5)

This means that grain size cannot be made smaller than a certain critical limit well known as the super-paramagnetic limit. On the other hand a minimum number of grains per bit is needed to achieve a sufficiently good signal to noise ratio (SNR).

**Figure A.6:** Magnetic properties of an MO medium as function of temperature.

In an MO system data is written on the medium by applying an external magnetic field generated by a coil that is relatively large compared the the magnetic domain that needs to be written. At room temperature this field is below the coercive field and domain formation on the disc is not possible. A focused laser beam is used simultaneously to locally heat the material to values near the Curie temperature (see Fig. A.7 [111]). The coercivity drops and at the center of the temperature profile a domain is written, which remains stable after cooling down the medium. Several recording strategies are known:

- Laser intensity modulation (LIM): In this method the recording layer is oriented in a predefined direction prior to recording. During the recording process the external field has the opposite direction as the magnetization in the
A.1 Technologies to Increase Density

recording layer, but it does not switch the layer since the coercive force of the layer is too high at room temperature. By intensity modulation of the laser the recording layer is heated locally, and the resulting drop in $H_c$ makes a switch in magnetization direction possible. LIM results in round or ellipse-shaped domains on the disc. A few large disadvantages of this technique can be mentioned [112]. Firstly, an initialization procedure is necessary prior to recording, which decreases data transfer rate considerably. Secondly, it is difficult to write small marks. The thermal profile must be controlled accurately resulting in small write power margin. Above all the domains are prone to collapse when they are below a critical size.

- Magnetic field modulation (MFM): During MFM the laser is on continuously, bring the recording layer at the temperature higher than the Curie temperature $T_c$. The marks are defined by switching the magnetic field direction allowing adjacent recorded marks to overlap partially. The resulting domain has a crescent shape and very small tangential bit lengths are possible. Disadvantages of this method according to [112] are that the data rate is determined by the maximum switching speed of the generally bulky coil. Furthermore, cooling down of the disc is rather slow (determined by the dwell time of the spot) and unnecessary transition regions will be present between magnetic domains with opposite magnetization, thereby limiting the density. Finally, the power burden of the disc is relatively high due to the continuous laser irradiation. Cross-erase problems will occur limiting the achievable track pitch.

Figure A.7: Illustration of the recording configuration in a MO system [111].
• Laser pulsed magnetic field modulation (LP-MFM): This combination of LIM and MFM was first published in [113]. It has the advantage that clear-edge marks can be formed by synchronization of the magnetic field transition and the laser pulse transition.

Fig. A.8 gives an overview of the recording methods [112]. It shows that by partly overwriting adjacent domains it is possible to form magnetic marks that are considerably smaller than the FWHM size of the optical spot, at least in the tangential direction. These marks are thermally stable because of the coercivity that approaches infinity at room temperature. The big problem is to read out these very small marks using the optical spot (readout of MO is based on the Kerr rotation of the polarization direction upon reflection from a perpendicular magnetized medium). This has been a subject of many years of research.

![Figure A.8: Recording methods for a magneto-optical system (from left to right: LIM, MFM and LP-MFM) [112].](image)

**Super Resolution** Magnetic super resolution by so-called rear aperture detection (RAD) was first reported by Aratani [114, 115]. Here an exchange-coupled double layered medium is used. The bottom layer is the recording layer. On top of this layer a read-out layer is present. Initially this read-out layer is initialized by an external field $H_i$ in the downward direction as shown in Fig. A.9. The bits in the recording layer are not affected by this field because it is still below the switching field $H_{s2}$ of this layer. When the read-out layer is heated above a certain threshold temperature its magnetization will be reversed to the upward direction only when the magnetization in the recording layer is assisting the read-out field. There, the signal can only be read out from the aperture at the rear side of the light spot. The bits in the recording layer on the front side are masked by the initialized read-out layer.

A similar technique is called front aperture detection (FAD) also reported by Aratani [114, 115]. Now a triple-layer disc is used. Again the recording layer contains the bits. These bits are copied to the read-out layer via a so-called switching
layer by exchange coupling. This occurs at normal ambient temperatures when the switching layer is below its Curie temperature $T_c$. When the temperature is increased to above this value the intermediate layer loses its magnetization and the read-out layer is not coupled anymore to the recording layer, which allows the magnetization in the read-out layer to align to an small external read-out field $H_r$. The heated rear aperture region does not contribute to the signal anymore and acts as a mask. The disc configuration is shown in Fig. A.10.

One of the most important disadvantages of RAD and FAD is the strong dependence of the read-out performance on linear disc speed. The reason for this is that the thermal profile depends on the linear velocity of the medium. At lower speeds a circular thermal profile is formed that largely overlaps with the optical spot. A higher disc speed results in a more elongated thermal profile with smaller overlap with the optical spot. A possible solution is to use a nanosecond laser pulse to create a quasi-circular aperture [116, 117]. This method of detection is called central aperture detection (CAD).
Domain Expansion Techniques Although all of these methods improve read-out resolution with a significant factor they have one common problem and that is the decrease in SNR. A smaller aperture means a decrease in ISI for small bit lengths, but also a decrease in area generating the replay signal. This means that there is always a trade-off between SNR and density increase. Awano et al [118] proposed a solution to this problem by amplifying the small, copied domain. This technique is called magnetically amplifying magneto-optic system (MAMMOS). It is based on the RAD technique and consist of three basic steps (see Fig. A.11). First, a bit is copied to the readout layer in the area where the stray field of the bit $H_s(T)$ is assisting the external field $H_{ext}$ such that the sum is larger than the switching threshold. This area is called the copy window. Simultaneous with the copy process an external magnetic field is applied in the direction of the copied mark that causes expansion of the domain to the full size of the optical spot. This results in a large, saturated readout signal. Subsequently, the external magnetic field is reversed in order to collapse the expanded domain completely and to leave the readout layer behind in its initial state. This procedure can be repeated for each channel bit on the disc. Although

Figure A.11: Principle of the magnetically amplifying magneto-optic system (MAMMOS).

bit-detection for the MAMMOS system is very simple (a simple slicer is satisfactory for the saturated signal) it has some large other problems [119]. The first one is the control of the size of the copy window. A too large window will destroy the resolution of the system, while a too small window easily disappears due to laser power fluctuations [120]. The second problem is clock recovery from the readout signal. The data is present in the recording layer and becomes visible only when an external expansion field is applied. Therefore, incomplete clock information is present. Later zero-field MAMMOS was introduced [121] where the timing recovery problem is eliminated.

A second technique to increase SNR is domain wall displacement detection (DWDD). It was first reported by Shiratori et al. [122] and is based on FAD in the sense that it also uses a triple layer medium where the bit is copied from the recording layer to
the read-out layer by exchange coupling in the front part of the optical spot. In the rear part no coupling is present due to an elevation in temperature above the Curie temperature. The copied domain now expands due to a force caused by a gradient in the domain wall energy as soon as the front line of the temperature profile hits the domain. The gradient in domain wall energy is caused by the temperature gradient. The principle is shown in Fig. A.12 (see e.g. [123]). A more complete explanation of the process can be found in [124, 125]. Major problems of this technique are domain wall pinning, the necessity of non-magnetic guard spaces between tracks and the appearance of ghosts in the readout signal due to domain wall movement along the other slope of the temperature gradient [126].

![Figure A.12: Principle of domain wall displacement detection (DWDD)](image)

**MO versus Phase-change Recording** Despite the possibility of super resolution and the near-to infinite number of overwrite cycles, MO recording never became a success in the high-volume consumer market [24]. The key issues that are unfavorable for MO recording are the need for a magnetic coil, which needs to be close to the disc and shows increasing power dissipation at higher data-rates, and the complicated disc structure with many magnetically coupled layers. Moreover, signal detection is rather complicated compared to phase-change recording (see Section 2.6). Last, but not least, discs with multiple recording layers are impossible or difficult to realize in (super resolution) MO systems.
A.1.5 Other Super-Resolution Methods

Various other super-resolution methods have been proposed, for example optical super resolution where the transmittance of an extra layer changes as function of illumination (or induced temperature changes). The transmitting region created in the center of the illuminated area acts as an optical mask for the underlaying data layer. Examples can be found in [127, 128]. Yet another method is Super-RENS as introduced in [129]. The frequency response of this system indeed extends slightly beyond the cut-off of the optical channel. The effect is thought to be caused by non-linearities in the optical path due to changes in refractive index of the medium upon illumination.

A.1.6 Holography

Holography has been a promise for many years in optical storage. A good overview on holographic data storage can be found in [130]. Data is stored in pages. One page is a 2D array of bits. The 2D array modulates the phase or amplitude profile of a light beam by using a spatial light modulator (SLM). The modulated beam is directed towards the recording beam where it interferes with another beam, called the reference beam (see Fig. A.13). The resulting interference pattern is recording in the material e.g. as a change in refractive index. The pattern may be recorded in three dimensions in a ‘thick’ hologram. When illuminating the recorded interference pattern again with the same reference beam the original wavefront of the modulated beam is reconstructed. The most appealing characteristic of holography is that by choosing a suitable set of different reference beams it is possible to store multiple data pages in the same location in the storage medium. This is called multiplexing. For example the reference beam can change in angle (angle-multiplexing). Other
methods are wavelength-multiplexing, shift-multiplexing and phase-multiplexing or combinations of these. The reconstruction is also done page-by-page. This means that, in theory, a very high data rate could be achieved.

The reason that holographic data storage has never been a success in the market is the lack of suitable recording materials. Because the principle is based on diffraction from interference or fringe patterns with sub-nanometer features, the materials need to be very stable. This means e.g. no shrinkage depending on temperature, humidity, or further illumination steps to record other pages of data. Furthermore, the diffraction efficiency must be as high as possible to receive sufficient light on the detector. Because the holograms are recorded in the same recording volume, each hologram can only use part of the modulation range (e.g. refractive index range) for storage of the fringe pattern. A higher diffraction efficiency means that the modulation range can be kept limited and that more holograms can be multiplexed in the same recording volume.

Conventionally, photorefractive crystals such as lithium niobate (LiNbO$_3$) have been used. Recently, a breakthrough in material design has been claimed [131] in the form of photo-polymers and new research initiatives are developing in this field of data storage [132, 133].

A.1.7 Signal Processing and Coding

The first CD players incorporated a simple slicer to do the bit-detection. A signal above the threshold was detected as a one and a signal below the threshold as a zero. The increase in digital signal processing power allowed by Moore’s law offers many possibilities to increase storage density. The TwoDOS system as proposed in this thesis incorporates many of the advances in digital signal processing that have been proposed up till now.
Appendix B

Single Sided Cross Talk Cancellation

This Appendix discusses the derivation of the optimum filter for cross-talk cancellation (XTC). A very straightforward solution to derive the optimum filter is to use a zero forcing (ZF) scheme. A direct application of the zero forcing method, however, does not converge to a correct solution. This is attributed to the fact that the filtered signal that is subtracted from the row that is subject to cancellation is partly dependent on the same bits in that row. For this reason a more detailed analysis is made of the XTC method. An error is derived by comparing the actual signal with a target signal resulting from an ideal XTC scheme. By comparing the solution that leads to the minimum mean square error (MMSE) with the ZF solution it appears that with a slight modification of the ZF method it can be used to achieve the MMSE target.

B.1 Analysis of the XTC Method

A single-sided XTC scheme is considered according to the block diagram of Fig. B.1.

![Block Diagram](image)

*Figure B.1: Single-sided version of the cross-talk cancellation scheme.*

To analyze the behavior of this scheme the 2D impulse response is separated in a set of 1D impulse responses according to Fig. B.2. The one for the central row is denoted $h_c$ and the two for the adjacent (or side) rows are denoted $h_{s1}$ and $h_{s2}$ (note that these responses can be taken equal in case of the circularly symmetric target.)
Single Sided Cross Talk Cancellation

Figure B.2: Model for analyzing the behavior of the XTC scheme.

i.e. $h_{s1} = h_{s2} = h_s$. According to Eq. (3.29) the replay signals from two adjacent bit streams $b_n$ and $b_{n+1}$ can now be written:

$$y_{k,n} = 1 - [(b_{n-1} * h_s)_k + (b_n * h_c)_k + (b_{n+1} * h_s)_k];$$
$$y_{k,n+1} = 1 - [(b_n * h_s)_k + (b_{n+1} * h_c)_k + (b_{n+2} * h_s)_k]$$

(B.1)

with ‘∗’ denoting a 1D convolution. Note that $b \in \{0, 1\}$, similarly to the definition in the scalar diffraction model in Chapter 3. In the cross-talk cancellation scheme a filtered version of $y_{k,n+1}$ is subtracted from signal $y_{k,n}$:

$$\tilde{y}_{k,n} = y_{k,n} - (f * y_{n+1})_k$$
$$= (1 - \sum_p f_p) - (b_{n-1} * h_s)_k - (b_n * h_c)_k - b_{n+1} * (h_s - (f * h_c))$$
$$+ (b_n * h_s * f)_k + (b_{n+2} * h_s * f)_k$$

(B.2)

with $\sum_p f_p$ the sum over all filter coefficients. Ideally, after cross-talk cancellation, the signal $\tilde{y}_{n,k}$ is only given by:

$$\hat{\tilde{y}}_{n,k} = 1 - [(b_{n-1} * h_s)_k + (b_n * h_c)_k].$$

(B.3)

By comparing Eq. (B.2) and Eq. (B.3) the error $\tilde{y}_{k,n} - \hat{\tilde{y}}_{k,n}$ can be derived as:

$$e_k = -\sum_p f_p - b_{n+1} * (h_s - (f * h_c)) + (b_n * h_s * f)_k + (b_{n+2} * h_s * f)_k.$$  

(B.4)

The tap update following the steepest descent can be calculating as the partial derivative of the error with respect to the equalizer taps:

$$\frac{\partial e^2}{\partial f_p} = 2e \frac{\partial e}{\partial f_p} = -2e [1 - (b_{n+1} * h_c)_{k-p} - (b_n * h_s)_{k-p} - (b_{n+2} * h_s)_{k-p}].$$

(B.5)
B.2 Zero Forcing Solution

The steady state solution of the update algorithm using this error can be derived by rewriting this derivative in the frequency domain and setting it to zero:

\[ -2 \left[ \frac{1}{4} H^* H_c - \frac{1}{4} F |H_c|^2 - \frac{1}{2} F |H_s|^2 \right] 
+ \frac{\delta(0)}{2} \left( -F + 2FH_s + FH_c - \frac{1}{2}H_s + \frac{1}{4}H_c^* H_s 
- \frac{1}{4} F |H_c|^2 - F |H_s|^2 + \frac{1}{2} |H_s|^2 - H_c^* H_s F \right) = 0. \]  

(B.6)

Here, it is assumed that the bits \( b \in \{0, 1\} \) are uncorrelated, which means that a correlation between signals from unequal bit streams only results in a DC-term with a proportionality factor equal to \( \frac{1}{4} \), i.e. \( (b_1 * f)(b_2 * g) \) leads to a frequency domain contribution equal to \( \frac{1}{4} \delta(0)FG \), with \( \delta(0) \) the kronecker delta in frequency domain. Furthermore, \( H_s, H_c \) and \( F \) are the Fourier transforms of \( h_s, h_c \) and \( f \), respectively, and \( H^* \) indicates the complex conjugate of \( H \). It can be easily shown that the solution is equal to:

\[ F(\Omega) = \frac{H_s^* H_c}{|H_c|^2 + 2|H_s|^2} \quad \text{for} \quad \Omega \neq 0. \]  

(B.7)

For \( \Omega = 0 \) a special situation occurs, which we will discuss later in this appendix. Although the above method leads to the minimum mean square error (MMSE), it is not very convenient, since the bits are needed to calculate the error. A data-aided algorithm is needed where decisions are taken from the first or second iteration of the stripe-wise Viterbi detector. A large delay in the loop is unavoidable. Nevertheless, it is a viable solution in view of the minor (and relatively slow) variations in the impulse response of the equalized signal.

B.2 Zero Forcing Solution

As described in the previous section it is possible to achieve a minimum mean square error solution in a data-aided configuration. However, another solution is possible that is non-data-aided. This alternative solution is based on a zero-forcing of the average correlated signal. The correlated signal can be written as:

\[ f_{k,n} \otimes y_{k,n+1} = \left[ -\sum_p f_p - (b_{n-1} * h_s)_k - (b_n * h_c)_k - b_{n+1} * (h_s - (f * h_c)) 
+ (b_n * h_s * f)_k + (b_{n+2} * h_s * f)_k \right] 
\otimes \left[ 1 - (b_n * h_s)_k - (b_{n+1} * h_c)_k - (b_{n+2} * h_s)_k \right] \]  

(B.8)
with $\otimes$ denoting correlation. A zero forcing of this signal leads to a steady state solution in the frequency domain equal to:

$$F' = \frac{2H_c^*H_c}{|H_c|^2 + 2|H_s|^2}$$

for $\Omega \neq 0$. \hfill (B.9)

Note that this solution is equal to Eq. B.7 except for a factor of 2. This means that it is possible to converge to the MMSE solution using a simple zero forcing algorithm. In order to achieve this the resulting taps of the compensating filter obtained with the ZF algorithm need to be divided by 2.

Now let us consider the DC situation. It appears that a DC-term needs to be added after compensation with the cross-talk cancellation filter. The final XTC algorithm then works according to Fig. B.3. One way to deal with the additional DC-term is to adapt the reference levels in the Viterbi detector accordingly. No additional processing is needed to achieve this, since the automatic reference level update is already implemented in the detector. However, we would need two sets of reference levels. One for the signals with XTC and one for the signals without XTC. A more elegant way is to determine the DC level prior to detection. The calculation of the DC-term is possible by starting from the zero-forcing assumption:

$$E \left[ (y_{k,n} - (f'_k * y_{n+1})_k) \otimes y_{k+1,n} \right] = 0$$

with $E [\cdot]$ the expectation operator. In frequency domain we can write:

$$\int (Y_n - F'_n Y_{n+1})Y_{n+1}^* d\Omega = 0.$$ \hfill (B.11)

When both $y_{k,n}$ and $y_{k,n+1}$ are equalized to the same target response they also have the same DC value, indicated by $\beta$. When $F'$ is long enough to satisfy Eq. (B.11) it will...
B.2 Zero Forcing Solution

have the property that $F'(0)=1$ (this is confirmed from simulations with experimental data). The DC value amounts to:

$$\beta = 1 - \frac{1}{2} (2H_s(0) + H_e(0)) = 1 - 3c_1 - \frac{1}{2}c_0. \quad (B.12)$$

This means that the resulting ideal signal after cross-talk cancellation (Eq. (B.3)), has DC-value:

$$\text{DC}(\tilde{y}_{k,n,\text{target}}) = \beta - \frac{1}{2} H_s(0). \quad (B.13)$$

In reality the XTC scheme will produce a signal according to Eq. (B.2) with a DC value equal to:

$$\text{DC}(\tilde{y}_{k,n,\text{actual}}) = \beta - \frac{1}{2} F'(0). \quad (B.14)$$

Here, the factor 2 between the MMSE and ZF solution is taken into account. By comparing Eq. B.14 and Eq. B.13 one can derive that:

$$\Delta \text{DC} = \frac{1}{2} H_s(0) - \frac{1}{2} \beta F'(0). \quad (B.15)$$

Filling in $\beta$ and $F'(0) = 1$ in Eq. (B.15) leads to a final DC difference of:

$$\Delta \text{DC} = -\frac{1}{2} + \frac{5}{2}c_1 + \frac{1}{4}c_0 \quad (B.16)$$

Indeed, adding this $\Delta \text{DC}$ value to the output signal according to the scheme of Fig. B.3 leads to the correct DC level of the output signal and the minimum bER at the output of the detector. As an example, when using values of $c_0 = 0.8$ and $c_1 = 0.2$ (see Fig. B.2) we get $H_c(0) = -0.2$, $H_s(0) = 0.6$ and $\Delta \text{DC} = 0.2$ respectively. The $\Delta \text{DC}$ value can be calculated prior to detection based on the target response that is used.
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