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Design and Evaluation of a Modular fNIRS Probe for Employment in Neuroimaging Applications

Semester Thesis

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Abstract

**Background** For functional near-infrared spectroscopy (fNIRS), light of multiple discrete wavelengths is guided into the scalp and the internally scattered portion that exits the head’s surface is measured at some centimeters distance. For sufficiently large source-detector separations (SDS), the light penetrates cortical tissue, whose optical properties depend on brain activity. Increasing the SDS is favorable in terms of penetration depth, but makes the selection of detectors challenging, as the reflected light intensity drastically decreases with SDS.

**Motivation** To date, photomultiplier tubes (PMTs) or avalanche photodiodes (APDs) are considered the gold standard for detectors in fNIRS. However, PMTs are sensitive to overexposure, operate at very high voltages, and are bulky. APDs benefit from being solid state devices, but have reduced sensitivity and also need rather high operating voltages. Silicon photomultipliers (SiPMs) are alternative detectors which have matured in the last few years and overcome some of these limitations. They essentially consist of an array of parallel connected tiny APDs operating in Geiger mode, have a high sensitivity and gain, and operate at lower voltages, while performing comparable to PMTs.

**Methods** In this project we propose the employment of SiPMs and report on the development of an fNIRS modular device. The devices consist of two LEDs with wavelengths of 680 nm and 850 nm respectively, a SiPM detector, the analog signal conditioning circuit and a microcontroller for signal processing and communication between the modular boards. The circuit was realized on two stacked boards with 26 mm x 26 mm footprint size. This device was tested on a phantom with optical coefficients similar to a human head to determine the signal-to-noise ratio (SNR) for different SDS. Measurement data were furthermore obtained from the forehead during rest and from the forearm during arterial occlusion.

**Results** The experiments on a phantom showed an SNR of 60 dB up to an SDS of 38 mm. The measurements on a human forehead with an SDS of 50 mm clearly showed the pulsation and the corresponding change in oxy- and deoxy-hemoglobin. Furthermore the oxy- and deoxy-hemoglobin change on the forearm during occlusion showed the expected change in signal.

**Discussion & Outlook** The measurements showed that SiPM bear big potential for fNIRS instrumentation. The resulting device is small, compact and very sensitive. Its performance is limited by a drift assumed to be caused by self-heating of the SiPM but also excess heat of the residual electronic circuit. To improve the device’s performance, the temperature drift must be addressed first. Furthermore, investigating in more sophisticated signal modulation and processing techniques to further increase the SNR-performance, and thus increase the maximal possible SDS, could be of interest.
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# Nomenclature

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<tr>
<td>ADC</td>
<td>Analog-to-Digital Converter</td>
</tr>
<tr>
<td>APD</td>
<td>Avalanche Photodiode</td>
</tr>
<tr>
<td>ASF</td>
<td>AVR Software Framework</td>
</tr>
<tr>
<td>BORL</td>
<td>Biomedical Optics Research Laboratory</td>
</tr>
<tr>
<td>DAC</td>
<td>Digital-to-Analog Converter</td>
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<tr>
<td>DMA</td>
<td>Direct Memory Access</td>
</tr>
<tr>
<td>ETH</td>
<td>Eidgenössische Technische Hochschule</td>
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<tr>
<td>GPU</td>
<td>Graphics Processing Unit</td>
</tr>
<tr>
<td>HbO</td>
<td>Oxy-Hemoglobin</td>
</tr>
<tr>
<td>HbR</td>
<td>Deoxy-Hemoglobin</td>
</tr>
<tr>
<td>HPC</td>
<td>High-Power Configuration</td>
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<tr>
<td>IDE</td>
<td>Integrated Development Environment</td>
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<td>ISR</td>
<td>Interrupt Service Routine</td>
</tr>
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<td>LDO</td>
<td>Low-Dropout (Voltage Regulator)</td>
</tr>
<tr>
<td>LPC</td>
<td>Low-Power Configuration</td>
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<tr>
<td>MBLL</td>
<td>Modified Beer-Lambert Law</td>
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<tr>
<td>NIR</td>
<td>Near-Infrared</td>
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<tr>
<td>fNIRS</td>
<td>Functional Near-Infrared Spectroscopy</td>
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<tr>
<td>OD</td>
<td>Optical Density</td>
</tr>
<tr>
<td>PCB</td>
<td>Printed Circuit Board</td>
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<tr>
<td>PDI</td>
<td>Program and Debug Interface</td>
</tr>
<tr>
<td>PMT</td>
<td>Photomultiplier Tube</td>
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<td>SiPM</td>
<td>Silicon Photomultiplier</td>
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<tr>
<td>SNR</td>
<td>Signal-to-Noise Ratio</td>
</tr>
<tr>
<td>SPI</td>
<td>Serial Peripheral Interface Bus</td>
</tr>
<tr>
<td>TIA</td>
<td>Transimpedance Amplifier</td>
</tr>
<tr>
<td>USART</td>
<td>Universal Synchronous and Asynchronous Serial Receiver and Transmitter</td>
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Chapter 1

Introduction

Functional near infrared spectroscopy (fNIRS) is based on the absorption and scattering properties of near-infrared light in tissue which provides information about brain activity. Due to high scattering in biological tissue it was long thought to be impossible to recover information from anything else than the most superficial layers of the head. However, it could be shown that it is possible to obtain good enough information such that functional near infrared spectroscopy can be used as a technique for human brain monitoring. Diffuse optical imaging techniques - based on fNIRS, can provide excellent temporal sensitivity as well as reasonable spatial sensitivity of brain activity [1].

The primary advantage - compared to other methods (fMRI, PET, SPECT, MEG, EEG), lies in particular the instrumentation. fNIRS devices can be made low-power, low-cost, small and portable. Furthermore near infrared light is non-ionizing and does therefore not limit the number of scans one can undergo.

An fNIRS device is dealing with very low light intensities requiring very sensitive detectors. To the best of our knowledge, we are the first to use a silicon photomultiplier (SiPM) as a detector for fNIRS instrumentation. Having a high sensitivity but also a small footprint, the SiPM reveals the possibility for the development of a very accurate but also small device. These small devices then allow the setup of a modular network.

1.1 Organization of this Report

Chapter 1 of this report describes the general methodology of fNIRS devices and explains why SiPMs detectors bear a big potential to be used in fNIRS devices and could replace PMTs. In Chapter 2 the methods for designing a modular fNIRS device including a SiPM are described in detail. The evaluation of the relevant components and the design of the hardware and software is documented. In Chapter 3 the measurement results, achieved with the produced fNIRS device, are presented. Finally, the results are discussed in Chapter 4 and improvements are suggested in Chapters 5 and 6.
1.2 fNIRS

The concept of fNIRS is basically as follows: Light with two or more different wavelengths is injected at the head’s surface and detected as it exits the head a few centimeters away. Different light-absorbing molecules, i.e. chromophores, exhibit specific wavelength-dependent absorption. Thus changes in their concentration can be calculated from the detected light variations. Figure 1.1 shows the absorption spectra of the two main chromophores in the NIR range. Wavelengths between 650 and 950 nm show low absorption coefficients for the main chromophores and therefore provide the so-called optical window where fNIRS operates.

![Figure 1.1: Absorption spectra for the two primary chromophores: HbO (oxy-hemoglobin) and HbR (deoxy-hemoglobin) having low absorption factors in the optical window (650-950 nm). Reprinted from [1] with permission from Elsevier.]

1.2.1 MBLL

The most common way to determine the chromophore concentration is by using the modified Beer-Lambert law (MBLL) from Equation 1.1, which defines the optical density ($OD$) as the natural logarithm of the detector and source light intensity ratio.

$$OD = -\log_e \frac{I_{\text{Out}}}{I_{\text{In}}} = \epsilon \cdot C \cdot L \cdot B + G \quad \text{(Source: [2])}$$  \hspace{1cm} (1.1)

- $OD$: optical density
- $I_{\text{Out}}/I_{\text{In}}$: detected/incident light intensity
- $\epsilon$: extinction coefficient of the chromophore
- $C$: concentration of the chromophore
- $L$: length of the photon path between source and detector (without scattering)
- $B$: pathlength factor (increase in photon pathlength due to scattering)
- $G$: offset factor accounting for measurement geometry

To determine the change in chromophore concentration, one has to perform two measurements $I_{\text{Initial}}/I_{\text{Final}}$ before/after the concentration change respectively. This leads to an optical density change as shown in Equation 1.2. When one does this at two different wavelengths $\lambda$ the change in oxy- and deoxy-hemoglobin concentration ($\Delta[HbR]$ & $\Delta[HbO]$) can be calculated by using
Equation 1.3 (see [2] for more details).

\[ \Delta OD = -\log_e \frac{I_{\text{Final}}}{I_{\text{Initial}}} = \epsilon \cdot \Delta C \cdot L \cdot B \]  
(Source: [2]) (1.2)

\[ \Delta OD^\lambda = (\epsilon_{\text{HbO}}^\lambda \Delta[HbO] + \epsilon_{\text{HbR}}^\lambda \Delta[HbR])B^\lambda L \]  
(Source: [2]) (1.3)

\[ B \approx \frac{1}{2} \sqrt{\frac{3\mu'_s}{\mu_a}} \left( 1 - \frac{1}{1 + L\sqrt{3\mu'_s\mu_a}} \right) \]  
(Source: [2]) (1.4)

The pathlength factor \( B \) from Equation 1.4 depends on the reduced scattering \( \mu'_s \) and absorption coefficient \( \mu_a \). It is therefore wavelength dependent.

In [1] simultaneous fMRI and fNIRS measurements for a simple motor task show good correspondence between the two recording modalities for the chromophore concentration changes. Nevertheless caution must be taken when directly interpreting the results obtained via the MBLL of a single source-detector pair as clearly stated in [2]. This is further discussed in the next section.

1.2.2 Sensitivity and Accuracy

Spatial- & Depth Resolution

The further apart the source and detector are placed the more scattering will occur and therefore worsen the spatial resolution of the diffuse light measured. But a small spacing on the other hand leads to a lower depth of maximum brain sensitivity. As nicely shown in Figure 1.2 the brain sensitivity of a given source-detector pair is banana-shaped and has the maximum sensitivity at a depth of half the source-detector distance (This rule of thumb e.g. leads to a sensitivity depth of 2.5 cm for a source-detector spacing of 5 cm). Thus, the choice of the SDS is a trade-off between the depth of maximum sensitivity and the minimal acceptable signal strength at the detector. Nowadays instruments reach a maximal source-detector separation of 5-6 cm [1].

![Figure 1.2: Banana-shaped sensitivities for two source-detector configurations, S-D1 & S-D2. Reprinted from [3] with permission from SPIE.](image-url)
some cases can even fail to return the correct relative sign for changes in [HbR] and [HbO] (not even mentioning absolute or relative magnitudes). One way to overcome this problem is to use diffuse optical tomography (DOT), a technique that uses image reconstruction algorithms using the information of multiple source-detector pairs (without using the MBLL) to more accurately determine the chromophore concentrations. For more details we refer to [2].

1.2.3 Current Device’s Limitations/Drawbacks

To date, photomultiplier tubes (PMTs) or avalanche photodiodes (APDs) are considered the gold standard for detectors in fNIRS [1]. However, PMTs are sensitive to overexposure, operate at very high voltages and are bulky. APDs benefit from being solid state devices but have reduced sensitivity and also need rather high operating voltages.

Silicon photomultipliers (SiPMs) are alternative detectors which matured in the last years [4]. Having a high gain (\(\sim 10^6\)) and comparable photon detection efficiency, the SiPM are fast approaching the performance of vacuum PMTs. Concerning their low bias voltage (\(\sim 24\) V), small size, better robustness and magnetic field insensitivity, they even top the PMTs performance. Compared to APDs their performance is superior due to the higher gain and lower noise as it is shown with several examples in [5]. A more detailed comparison is given in [5] in Table 1.
1.3 Silicon Photomultiplier

A silicon photomultiplier (SiPM) consists of a 2-dimensional array of numerous (∼ 10³ mm⁻²) microcells. Each microcell is composed of a Geiger mode photodiode with an integrated quenching resistor in series. With a bias voltage a bit above the breakdown voltage, the diodes are working in Geiger mode. Photons entering the diode’s active area generate a charge carrier which leads to the generation of an avalanche current. This current would be self-sustaining and is therefore quenched by the resistor. This gives the possibility to count single photons with a microcell. Those microcells are all connected in parallel such that the output voltage results from the sum of each microcell’s current going through the output resistor as it is shown in Figure 1.3. Hence, the total output of the SiPM will be an analog signal proportional to the number of photons reaching all microcells of a SiPM.

![Figure 1.3: Schematic Diagram of a SiPM](image)

Originally, SiPMs were developed in the mid 1980s in Russia. Thanks to new integration concepts and modern semiconductor fabrication techniques the SiPM technology has matured in the last years.

1.3.1 Characteristics

A SiPM is characterized by various parameters from which the most important ones are listed below. Timing and some other characteristics, with lower importance for our application, are not discussed here. The interested reader is referred to [4] and [5], which describe all SiPM properties in more detail.

- **Breakdown Voltage, Over-Bias Voltage**: The breakdown voltage is determined by the reverse IV-characteristics of the diode. To work in Geiger mode the applied bias voltage in operation is chosen few volts higher than the breakdown voltage: \( V_{\text{overbias}} = V_{\text{bias}} - V_{\text{breakdown}} \).

- **Gain**: Is defined as the charge generated by a Geiger avalanche event divided by the charge of a single photoelectron. It is a function of bias voltage and temperature. The gain increases linearly with increasing bias voltage (Figure 1.4 left) and is in the order of 10⁶. The gain variation vs. overvoltage or temperature are in the range of \( \frac{dG}{G} \approx 7 \cdot \frac{dV_{\text{bias}}}{V_{\text{bias}}} \) and \( \frac{dG}{G} \approx 1.3 \cdot \frac{dT}{T(\text{°K})} \) respectively [5].

- **Photon Detection Efficiency (PDE)**: Defines the probability with which an incident photon really generates an avalanche event. It depends on the bias voltage as well as the wavelength and is given by Equation 1.5. Even though the quantum efficiency \( \eta(\lambda) \) or the geometrical efficiency (\( F = \frac{\text{OpticalActiveArea}}{\text{TotalChipArea}} \)) cannot be adjusted for a
given SiPM device, we can influence the PDE by adjusting the silicon quantum efficiency \( \varepsilon(V_{\text{bias}}) \) via the bias voltage.

\[
PDE(V, \lambda) = \eta(\lambda) \cdot \varepsilon(V_{\text{bias}}) \cdot F \quad \text{(Source: [4])} \tag{1.5}
\]

- **Dark Rate**: Is the result from avalanche events which are triggered by thermally generated carriers. It is one of the main factors limiting a SiPMs performance and discussed in more detail in the following paragraph about noise.

- **Recovery Time**: Duration which is needed by a single microcell to recover from an avalanche event. It is defined by the RC-combination of the quench resistor and the capacitance of a microcell.

- **Afterpulsing**: Carriers which got trapped in bandgap states at previous avalanche events can generate false counts.

- **Crosstalk**: Photons which get emitted during an avalanche event in one microcell trigger an avalanche in another microcell. This can be reduced significantly by optical isolation.

### Linearity, Dynamic Range

Due to the finite number of total pixels (microcells) \( M_{\text{pixels}} \), the number of simultaneously detectable photons is limited. This results in a saturation of the SiPM signal for high light intensities. The number of fired microcells \( N_{\text{Fired}} \) for a given number of incident photons is given by Equation 1.6. This is also visualized in Figure 1.4 (Right).

\[
N_{\text{Fired}} = M_{\text{pixels}} \left(1 - \exp\left(-\frac{PDE(V, \lambda) \cdot N_{\text{Photons}}}{M_{\text{pixels}}}\right)\right) \quad \text{(Source: [4])} \tag{1.6}
\]

As can be seen the dynamic range is limited to the linear range of this curve which limits the number of incident photons to \( N_{\text{photons}} \leq 0.6 \cdot M_{\text{pixels}} / PDE(V, \lambda) \) [4].

### Noise

The noise is dominated by the dark rate and is in the range of some MHz \( \cdot \) mm\(^{-2}\). It is a function of the over-bias voltage (linearly increasing noise with some MHz/V) and strongly temperature dependent. For sensing small light intensities the sensor needs to be operated in a cold environment. But according to [5], the performance is not affected by the noise at room temperature for sensing larger light signals.

![Figure 1.4: SiPM microcell gain as a function of over-bias voltage (Left) and SiPM response as function of the number of instantaneous photons for two different PDE values (Right). Both © 2008 IEEE [4].](image)
1.4 Project Description

The advantages of SiPMs compared to PMTs, mainly the small size, the small operating voltage and the insensitivity to overexposure, seem to bear great potential for the use in fNIRS instrumentation. The small size of the detectors opens the door for modular probes that each consist of sources and detectors, hence allowing for more flexibility in their placement. The small operating voltage facilitates the use directly on the human body and also future certification. The insensitivity to overexposure simplifies the handling of the probes and reduces the maintenance cost. This project shall provide the experimental evidence that SiPMs are a promising employment in fNIRS instrumentation. The test setup should consist of SiPMs, sources (LEDs or laser diodes) as well as a readout and signal conditioning circuitry integrated on a PCB. Furthermore, a simple interface (e.g. LabVIEW or Simulink) should control the modules.

1.4.1 Goals

The goals stated below are an excerpt of the task list out of the project definition.

- Theoretical determination of the attenuation (i.e. $I_{\text{Out}}/I_{\text{In}}$) in tissue (using Monte Carlo simulations or analytical methods) in dependence of the source-detector separation and identification of losses (coupling efficiencies, etc.).
- Definition of specifications for source (LED/laser diode, wavelengths, intensity, etc.) and detector based on attenuation.
- Definition of requirements for signal processing and communication (BUS) components (microprocessor, source driver, operational amplifiers, transimpedance amplifiers, etc.).
- Selection/ordering of components based on the specifications found above.
- Design of the prototype circuitry (scheme).
- PCB Layout.
- Design and implementation of a simple interface (Simulink, LabVIEW, ...) to control data acquisition.
- Evaluation of the prototypes’ performance:
  - SNR for different source detector separations,
  - Attenuation for different source detector separations,
  - Occlusion measurements.

Design Requirements

The minimal change in light intensity which should be detectable by the device was estimated to be 0.7%. This estimation was done by the supervisor. In order to reliably detect this small light intensity change, we aimed at a resolution of 0.1%, which corresponds to an SNR of 60 dB.
Chapter 2

Methods

2.1 Simulations

The physical phenomena of energy transfer in tissue in form of electromagnetic radiation can be described with the radiative transfer equation. The propagation of radiation through tissue is affected by absorption and scattering processes [6]. Analytical solutions to the radiative transfer equation exist only for simple cases. For complex situations such as a human head, numerical methods like Monte Carlo simulations are required. To obtain analytical solutions for a semi-infinite medium the diffusion equation is often used. It is an approximation of the more exact transport equation.

In the following two sections the Diffusion Equation and the Monte Carlo Simulation are described in more detail. The results of these calculations are the basis for the evaluation of the source and detector pair.

2.1.1 Diffusion Equation

Solutions of the diffusion equation have frequently been applied to derive the optical parameters from experimental data. In [6] the influence of different boundary conditions to the diffusion equation for a semi-infinite turbid medium are investigated, mainly: the partial-current boundary condition (PCBC), the zero-boundary condition (ZBC) and the extrapolated boundary condition (EBC). All the equations, derivations and results in this section were taken from [6].

The results for the different boundary conditions were compared to Monte Carlo simulations. The EBC boundary condition assumes that the fluence rate goes to zero some distance beyond the actual surface. It could significantly improve the solution of the diffusion equation for steady-state spatially resolved applications. With the EBC boundary condition the steady-state diffusion equation leads to the following fluence rate.

$$
\Phi(\rho, z) = \frac{1}{4\pi D} \left( \frac{\exp(-\mu_{eff}((z - z_0)^2 + \rho^2)^{1/2})}{((z - z_0)^2 + \rho^2)^{1/2}} - \frac{\exp(-\mu_{eff}((z + z_0 + 2z_b)^2 + \rho^2)^{1/2})}{((z + z_0 + 2z_b)^2 + \rho^2)^{1/2}} \right) 
$$

(2.1)

Where $D = \frac{1}{3(\mu_a + \mu_s')}$ is the diffusion constant, $\rho$ is the radial distance from the source, $z$ is the distance normal to the boundary and $z_0 = (\mu_a + \mu_s')$ is the position of the point source and $\mu_{eff} = [3\mu_a(\mu_a + \mu_s)]^{1/2}$ is the effective transport coefficient, $z_b$ is the distance between the surface and the location where the fluence rate goes to zero, for EBC this is:

$$
z_b = \frac{1 + R_{eff}}{1 - R_{eff}} 2D
$$

(2.2)

For a better understanding of the geometrical setup of these calculations it is referred to [7]. $R_{eff}$ represents the fraction of photons that are internally diffusely reflected at the boundary. According to [8], this coefficient is calculated for a refractive index of $n = 1.4$ to be $R_{eff} = 0.493$. 
The reflectance in the steady-state case with the EBC boundary condition is according to Equation 2.3:

\[ R_s(\rho) = 0.188 \cdot \Phi_s(\rho, z = 0) + 0.306 \cdot R_s^f(\rho) \]  

The diffuse reflectance \( R_s^f(\rho) \) from the medium is:

\[ R_s^f(\rho) = \frac{1}{4 \cdot \pi} \left[ z_0 \left( \mu_{\text{eff}} \cdot \frac{1}{r_1} \right) \exp\left( -\mu_{\text{eff}} \cdot r_1 \right) + (z_0 + 2z_b) \left( \mu_{\text{eff}} + \frac{1}{r_2} \right) \exp\left( -\mu_{\text{eff}} \cdot r_2 \right) \right] \]  

with \( r_1^2 = z_0^2 + \rho^2 \) and \( r_2^2 = (z_0 + 2z_b)^2 + \rho^2 \).

The results of the Monte Carlo simulation and the diffusion equation under different boundary conditions are shown in Figure 2.1. The spatial resolution of the steady-state Monte Carlo simulations was 0.1 mm, and the distances up to 25 mm were scored. The optical coefficients are \( \mu_s = 1 \text{ mm}^{-1} \) and \( \mu_a = 0.01 \text{ mm}^{-1} \).

![Figure 2.1: Comparison of steady-state spatially resolved reflectance, calculated by means of diffusion theory with different boundary conditions and Monte Carlo simulations. Reprinted from [6] with permission from OSA.](image)

As stated above, the diffusion equation under the EBC boundary condition shows the best correspondence to the Monte Carlo simulation for distances longer than 1.5 mm. Only for distances smaller than 1.5 mm the differences are greater than 5%.

### 2.1.2 Monte Carlo Simulations

In this project, Monte Carlo simulations were used to simulate light transport in semi-infinite tissue. The simulated photons are moved a distance \( \Delta s \) where they may be absorbed, scattered or transmitted out of the tissue. With simulating a sufficiently large number of photons, the light paths are recorded and the reflectance between the source and the detector can be calculated. The Monte Carlo simulations were used to validate the above mentioned diffusion equation.

**GPU Accelerated Implementations**

Monte Carlo simulations are very computational intensive and can sometimes take hours if not even days to complete. Here the graphics processing unit (GPU) of a commodity graphic card with a nVidia CUDA chipset comes into play. With hundreds of processors on one chip, they are able to compute such tasks highly in parallel and provide an interesting way of accelerating algorithms without too much additional cost intensive hardware effort or even a supercomputer.
For two commonly used CPU-based implementations namely mcml [9] and tMCimg [10] there exist the GPU-based counterparts named CUDAMCML [11] and mcxlab [12] respectively. The CUDAMCML\(^1\) (Version: 2009-07-06) implementation was used and successfully compared to the mcml\(^2\) (Version: 1.2.2, 2000) CPU implementation. It simply simulates a pencil beam incident to a semi-infinite medium with one or more layers. A simulation (for the calculations in Section 2.1.3) launching 100 million of photons took 9.35 minutes on the GPU. The same setup but with only 10 million photons took 178 minutes on the CPU. This is a 19-fold speedup. When performing the simulation with the same number of photons a speedup of around 200 is expected. This clearly shows the power of GPU-based Monte Carlo simulations.

### 2.1.3 Reflectance

The reflectance specifies the fraction of the incident optical light power which is seen at the detector. It is normed with the area of the detector. The reflectance of a human head’s tissue was calculated using Equation 2.3 and compared to three Monte Carlo simulations. The first two approximated the head as a simple 1-layered semi-infinite medium whereas the third was extended to simulate a 4-layered model. Figure 2.2 shows the results of the aforementioned calculations and simulations. The diffusion equation based reflectance coincides pretty well with the two Monte Carlo results starting from an SDS of about 2 mm. The CPU-based 1 layer simulation is less accurate for larger SDS than its GPU counterpart. This is due to the fact that 10 times less photons were simulated which leads to a less accurate result. The 4 layer model shows a higher reflectance for an SDS of 15 mm and higher. Since the phantom used in the measurements represents a 1-layered homogeneous medium, the reflectance based on the 1-layered model was chosen for further calculations. This way, the diffusion equation (Equation 2.3) could be used to quickly estimate the light power available at the detector for a given input power and SDS:

\[
\text{LightPower}_{\text{Detector}}(\text{SDS}) = \text{LightPower}_{\text{Source}} \cdot R^*(\rho = \text{SDS}) \cdot \text{Area}_{\text{Detector}}
\]

(2.5)

### 2.1.4 Signal-to-Noise Ratio (SNR)

Most of the current literature covers the usage of SiPMs for single photon counting. The - for this project more interesting - topic of using SiPMs for analog mode measurements is scarcely discussed. Therefore the dark noise is mostly the main noise source mentioned. But when continuously measuring light with very low intensities one has to consider the shot noise resulting from statistical fluctuations of the photons. \(^3\) addresses this topic and provides a useful formula for calculating the SNR given by:

\[
\text{SNR}_{\text{lin}} = \frac{I_S}{\sqrt{2q_eFBG(I_S + 2(I_D + I_B))}} = \frac{I_S}{i_S + 2(i_B + i_D)}
\]

(2.6)

where \(I_S\), \(I_B\) and \(I_D\) are the average signal, backlight and dark current at the SiPM output. \(F\) is the excess noise factor, \(B\) the electronic bandwidth (in Hz), \(G\) the SiPM gain and \(q_e\) the

---

3. Please note that the SensL White Paper \(^4\) unfortunately contains some typos in the equations. Furthermore the SiPM SNR contains copy & paste errors originating from the PMT SNR section. Nevertheless the equations for PMT’s and SiPM’s SNR seem to be correct when having compared them to other literature \(^5\) (page 14, ‘e’ SNR’) or \(^6\) (page 2, Equation 1). On top of that the equation of the SiPM was derived from the one of the PMT and validated. One simply has to be very careful with differentiating between capital and small letters (respectively between anode and cathode currents) in the indices when reading the SensL White Paper’s SNR equations!
Figure 2.2: The reflectance of a human head’s tissue calculated by using Equation 2.3 (blue), simulated with a semi-infinite 1 layer model using two mcml-based Monte Carlo simulations running on the GPU [11] (red) or CPU [9] (green) launching $100 \cdot 10^6$ or $10 \cdot 10^6$ photons respectively. All three having the same optical parameters ($\mu_a = 0.01 \text{ mm}^{-1}$, $\mu_s = 1 \text{ mm}^{-1}$, $g = 0.9$, $n = 1.4$, $R_{\text{eff}} = 0.493$). Additionally, a four-layered model GPU based simulation is shown (black). Its optical parameters were taken from Table 1 in [13] with $n = 1.3$. 
electron charge. The individual currents contributing to the shot noise: \( i_S, i_B, i_D \), are given by the following three equations:

**Signal Shot Noise Current:**
\[
i_S = \sqrt{2q_exec \cdot G \cdot FB \cdot G} = \sqrt{2qexec I_S FBG} \tag{2.7}
\]

**Backlight Shot Noise Current:**
\[
i_B = \sqrt{2qexec \cdot G \cdot FB \cdot G} = \sqrt{2qexec I_B FBG} \tag{2.8}
\]

**Dark Shot Noise Current:**
\[
i_D = \sqrt{2qexec \cdot G \cdot FB \cdot G} = \sqrt{2qexec I_D FBG} \tag{2.9}
\]

![Figure 2.3: Signal-to-Noise Ratio (SNR) in function of SDS at two different bandwidths. The solid lines show the SNR from Equation 2.6 whereas for the dashed lines dark and backlight noise were set to zero. The dotted red line shows the SNR limit to detect 0.1% of signal change. The reflected power results from Equation 2.3 with the following optical parameters: \( \mu_a = 0.01 \text{ mm}^{-1} \), \( \mu_s = 0.97 \text{ mm}^{-1} \), \( R_{eff} = 0.493 \), for \( \lambda = 850 \text{ nm} \). The detector is parameterized as follows: \( G = 2 \cdot 10^6 \), \( F = 1.1 \), \( I_D = 135 \text{ nA} \), \( I_B = 392 \mu \text{A} \). The source’s optical power is: \( P_{in} = 2.6 \text{ mW} \).](image-url)

The maximal physically possible SNR limited by the shot, backlight and dark noise is depicted in Figure 2.3. The parameters for source and detector are based on the final hardware (see Section 2.2). The optical parameters are the same as for the reflectance calculations (see Section 2.1.3), modeling a human head.

The dark noise is the smallest of all three noise contributors. Reducing the backlight might improve the actual SNR (solid line) but only to the maximal possible limit defined purely by the shot noise (dashed line).

Rather should one consider to reduce the bandwidth \( B \) or increase the light power \( P_{in} \) when wanting to improve the SNR.
2.2 Optical Components

2.2.1 Source

As source for the fNIRS setup, principally filtered white light, lasers or LEDs are possible [1]. For
the sake of simplicity and safety only LEDs were evaluated as light source. The criteria for the
light sources are listed below.

- appropriate wavelength
- high emitted power
- small view angle
- small footprint
- good availability

Table 2.1 shows the most important properties of the evaluated LEDs.

The wavelengths of the light sources must lie in the optical window which is between 650 nm and
950 nm. Furthermore the wavelengths were chosen such that the difference between the extinction
coefficients of HbR and HbO is maximal. Best results are made with the combination of 680 nm
- 700 nm paired with 830 nm or 750 nm - 760 nm paired with 830 nm [17]. The hemoglobin
absorption spectra does not change much from 830 nm - 900 nm, but the detection efficiency of
SiPMs decreases rapidly in this region. Therefore the second light source should be chosen close
to 830 nm.

The footprint of the light source is preferred to be small in size to meet the requirements of a small
sensor module and to place both sources as close together as possible. Therefore high power LEDs
are not a convenient choice as they are often bulky in size and have a rather high heat dissipation.
The companies VISHAY and Panasonic could not offer products of the same type with the desired
wavelengths. The SMT735/850 and SMT735/780/810 of Epitex would have been an interesting
solution as both/all three LED-dies are combined in the same housing, but these products are
currently not available. The SMT690 and SMT830N of Epitex have a rather high view angle and
are only sold in units of 50 pieces.

In the end the LEDs from Osa Optolight were chosen as they meet best the specifications and
are available. This type of LED has a tiny lens which makes the small view angle possible. The
weak point of this LED is the rather low continuous forward current of 30 mA which corresponds
approximately to an optical power of 1.3 mW for 680 nm and 2.65 mW for 850 nm.

<table>
<thead>
<tr>
<th>Manufacturer</th>
<th>Type</th>
<th>Max Forward Current $I_F$ [mA]</th>
<th>$1/2$ View Angle</th>
<th>Wavelength [nm]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Osa optolight</td>
<td>330-1206</td>
<td>30</td>
<td>20°</td>
<td>680 ± 8 850 ± 8</td>
</tr>
<tr>
<td>VISHAY</td>
<td>VSMG2720</td>
<td>100</td>
<td>60°</td>
<td>830</td>
</tr>
<tr>
<td>Panasonic</td>
<td>LN1261CTR</td>
<td>20</td>
<td></td>
<td>700</td>
</tr>
<tr>
<td>Epitex</td>
<td>SMT690</td>
<td>50</td>
<td>55°</td>
<td>690</td>
</tr>
<tr>
<td>Epitex</td>
<td>SMT830N</td>
<td>100</td>
<td>63°</td>
<td>830</td>
</tr>
<tr>
<td>Epitex</td>
<td>SMT735/850</td>
<td>75</td>
<td>55°</td>
<td>735/850</td>
</tr>
<tr>
<td>Epitex</td>
<td>SMT735/780/810</td>
<td>75</td>
<td>55°</td>
<td>735/780/810</td>
</tr>
</tbody>
</table>

Table 2.1: Overview of evaluated light sources. The second last LED combines two LED-dies in
one housing and the last one even three. The chosen LED is shaded in gray.
2.2.2 Detector

Detectors of several companies producing SiPMs were evaluated. The detector’s requirements for the current application are listed below:

- high detector SNR
- low dark current
- low recovery time
- large active area
- high dynamic range
- low supply voltage

Table 2.2 shows how the requirements above are influenced by the physical property of the SiPM device.

<table>
<thead>
<tr>
<th>Requirements</th>
<th>Physical Property</th>
</tr>
</thead>
<tbody>
<tr>
<td>high detector SNR</td>
<td>high PDE</td>
</tr>
<tr>
<td></td>
<td>low dark count rate</td>
</tr>
<tr>
<td></td>
<td>large active area</td>
</tr>
<tr>
<td></td>
<td>large size of microcells (high gain)</td>
</tr>
<tr>
<td>low dark count rate</td>
<td>small active area</td>
</tr>
<tr>
<td>low recovery time</td>
<td>small size of microcells</td>
</tr>
<tr>
<td>large active area</td>
<td>large size of microcells</td>
</tr>
<tr>
<td></td>
<td>or high number of microcells</td>
</tr>
<tr>
<td>high dynamic range</td>
<td>high number of microcells</td>
</tr>
</tbody>
</table>

Table 2.2: The qualitative influence of the physical properties to the detector’s requirements is shown. The requirements are mainly adjusted and limited by the size and number of the microcells.

As one can see in Table 2.2, the desired specifications can not all be optimized as for example a high detector SNR requires a large active area. On the other hand a low dark count rate requires a small active area. Since not all the requirements can be optimized at the same time, the focus was mainly set on a high detector SNR. The detector SNR is mainly affected by the dark current (noise) and the photon flux on the detector (signal). The signal can be amplified by increasing the PDE or the active area of the SiPM.

- The PDE is a characteristic of the sensor and can be increased by adjusting the bias-voltage. But an increase of the bias-voltage also amplifies the dark current and therefore does not change the SNR significantly.

- A large size of the active area can be achieved by a large size of the microcells or a high number of microcells. But a large active area affects the dark count rate in a negative way.

As it can be seen out of the above description, the active area of the detector has always an influence on the signal as well as on the dark current. These dependencies were observed only qualitatively. Therefore the detector SNR of different SiPMs was calculated in order to investigate the quantitative influence of these properties.

Figure 2.4 shows the detector SNR of different SiPM detectors. The detector SNR is defined as the detected signal photons per second divided by the dark noise rate. The specification of the detectors shown in Table 2.3 were used for this calculation. Acceptable SNRs can be achieved for an SDS in the range of 15 mm up to 60 mm. The lower bound of the SDS (indicated with a circle) is limited by the maximal linear range of the SiPM. It can easily be extended by reducing the LED power for small SDS. The upper bound of the distance (indicated with a square) limits the region where only one photon per recovery time generates a Geiger avalanche. This limit can be extended by integrating the signal over a multiple of the recovery time.
### Chapter 2. Methods

<table>
<thead>
<tr>
<th>Manufacturer</th>
<th>Type</th>
<th>Number of Cells</th>
<th>Cell Size [µm]</th>
<th>Dark Count Rate [kHz]</th>
<th>Recovery Time [ns]</th>
<th>PDE @850 nm</th>
<th>V Bias [V]</th>
</tr>
</thead>
<tbody>
<tr>
<td>SensL</td>
<td>10020</td>
<td>1296</td>
<td>20</td>
<td>600</td>
<td>30</td>
<td>5%</td>
<td>29.5</td>
</tr>
<tr>
<td>SensL</td>
<td>30035</td>
<td>4774</td>
<td>35</td>
<td>5550 ¹</td>
<td>130</td>
<td>5%</td>
<td>29.5</td>
</tr>
<tr>
<td>Hamamatsu</td>
<td>S10931−025P</td>
<td>14400</td>
<td>25</td>
<td>8000</td>
<td>63 ²</td>
<td>4%</td>
<td>70.0</td>
</tr>
<tr>
<td>Hamamatsu</td>
<td>S10931−050P</td>
<td>3600</td>
<td>50</td>
<td>10000</td>
<td>320 ³</td>
<td>8.5%</td>
<td>70.0</td>
</tr>
<tr>
<td>Photonique</td>
<td>SPPM 0905V13MM</td>
<td>810</td>
<td>40</td>
<td>300</td>
<td>16.7</td>
<td>13%</td>
<td>29.0</td>
</tr>
<tr>
<td>KETEK</td>
<td>PM5350</td>
<td>3600</td>
<td>50</td>
<td>4500</td>
<td>150</td>
<td>5%</td>
<td>32.0</td>
</tr>
<tr>
<td>KETEK</td>
<td>PM60A1</td>
<td>3600</td>
<td>50</td>
<td>2250</td>
<td>150</td>
<td>5%</td>
<td>28.5</td>
</tr>
</tbody>
</table>

¹ Not given in the data sheet, calculated out of the dark current.
² Communicated by Hamamatsu Japan based on E-Mail request
³ Photonique is bankrupt, products are sold by Advatech-UK

Table 2.3: Overview of the evaluated detectors. The PDE is often only graphically given and not specified up to the wavelength of 850 nm. The values are therefore extrapolated and must be taken with caution.

![Detector SNR in function of SDS for different detectors](image)

Figure 2.4: Detector SNR in function of SDS for different detectors. The optical power of the LED is set to 5 mW and corresponds to a wavelength of 850 nm. According to the diffusion equation (Equation 2.3) the incident power on the detector was calculated. As noise source only the dark noise of the detectors was taken into account. The working range of the detectors is indicated with a circle on the lower bound off the distance and with an square on the upper bound.
Remark: At the time of the detector evaluation the sensor KETEK PM60A1 was not available on the market and was therefore not respected. The SNR of the other detectors vary only by half a decade with the exception of the Photonique detector. The Photonique and Ketek PM3350 sensor were tested in the lab, where the Photonique detector showed a much worse behavior than the Ketek sensor. Mainly the gain of the Photonique sensor was much lower and was therefore not considered.

Furthermore the Hamamatsu sensors need a supply voltage of approximately 70 V whereas the SensL and Ketek sensors need only around 30 V. The high SNR and the low bias voltage tipped the scales for the Ketek PM3350 sensor.

During the progress of the project the Ketek PM3350 showed a very strange behavior when exposed to high light exposure. There were inexplicable spikes on the signal. For that reason the sensor was replaced to the newest sensor edition by the manufacturer KETEK. The new sensor KETEK PM60A1 has an even better SNR behavior as Figure 2.4 shows and was used for all measurements presented in Chapter 3.

2.3 Hardware

Both PCB and electronic schematics were designed using Altium Designer (Version: 9.3.1.19182). Hardware simulations were taken out using TINA-TI (Version: 9.3.50.40 SF-TI). The low-pass filter was designed with the help of TI’s FilterPro Application (Version: 3.1.0.23446).

2.3.1 Test Setup

To familiarize with the SiPM detectors and evaluate the basic electronic circuits two test boards were manufactured. Their schematics can be found in Appendix B.1.

- The PreAmp-board holds the possibility to connect various detectors from different manufacturers. Besides the usual transimpedance amplifier stage two additional filter stages are present. Their op-amps are surrounded by numerous pads such that different filter configurations can be tested.

- Two voltage-to-current converter stages are present on the LEDriver-board to properly control the two LED’s currents and thus their light intensity.

2.3.2 Final Setup

The final device consists of two boards stacked on top of each other as it can be seen in Figure 2.5 or on the title page of this report. The two boards basically contain the following circuits:
• **AnalogBoard**: SiPM detector with adjustable bias voltage regulator, transimpedance and low-pass amplifier stages, data converters (ADC, DAC), NIR LED driver

• **ControllerBoard**: AVR xMega microcontroller, power supplies, BUS (providing the possibility for a modular network)

A third board containing a UART-to-USB bridge and the microcontroller’s debug interface was designed. This **USB-UART-Debug Board** can be attached on top of a module during debugging. The schematics of the three boards can be found in Appendix B.2 (Please note that all the part numbers referenced in the text correspond to revision A).

The whole hardware interaction is depicted in Figure 2.6 by a block diagram.

PCBs

The stacking of two boards permitted a small footprint of 26x26 mm. Thanks to this, an easy separation of the analog and digital electronics was achieved which is necessary for a low noise level. To reach the goal of a small footprint and low noise electronics the PCB was designed using 4 layers, whereas the two inner layers simply contain a plane with ground or main power supply respectively.

SiPM

To generate a positive output voltage at the transimpedance amplifier stage - such that it can be directly measured by the ADC - a negative **SiPM bias voltage** was chosen. It is regulated by the low-dropout (LDO) **voltage regulator** (U6) driven with a maximum of $-35$ V at the input. With the resistor divider configuration controlled by the DAC, an output voltage in the range of
−24 to −33 V can be set. This allows the adjustment of the SiPM gain in real-time and therefore increases the dynamic range of the system such that it can be used for very short but also for large SDS.

The output voltage is given by the simple Equation 2.10 whereas the reference \((V_{ref})\) and output \((V_{out})\) voltages have to be negative - as they are in reality. The regulator’s quiescent current \(I_q\) was neglected in our calculations. A more important step was to take the reference voltage inaccuracy into account.

As validated with a simple simulation (Figure 2.7 (Left)) the chosen resistor configuration enables the adjustment of the output in the full range for all possible reference voltage variations.

\[
V_{out} = V_{ref} + \frac{R_7 \cdot I_q}{R_{11}} + \frac{R_7(V_{ref} - V_{DAC})}{R_{10}}
\]  

\(2.10\)

Figure 2.7: (Left) Simulation of the SiPM voltage regulator output in function of the DAC adjustment voltage. Three reference voltages were chosen to cover the full reference voltage variation. (Right) Simulated transfer function of the two amplifier stages (\(V_{TI}\): transimpedance amplifier, \(V_{LP}\): low-pass filter including the preceding transimpedance amplifier). (Note that for the simulation the OPA656 was used as op-amp since OPA(2)356 was not available as model.)

The first amplifier stage consists of a transimpedance amplifier with 1st-order low-pass characteristics with a gain of \(R_{Gain} = 510\) and a corner frequency of \(f_c = 1\ kHz = 1/(2\pi R_{Gain} C_{Filter})\).

It is fed to the first channel of the ADC and can also be measured accurately with the oscilloscope using connector X1.

The second amplifier stage consists of an inverting 2nd-order low-pass filter \((f_c = 1\ kHz)\) with adjustable offset voltage. Its gain is set to 2. Figure 2.7 (Right) shows the simulated transfer function of the two amplifier stages. The whole setup is a 3rd-order low-pass filter with cut-off frequency at 1 kHz and a gain of 1020.

The 2nd stage’s output is also fed to a connector for oscilloscope measurements (X2) and connected to ADC’s second channel. The offset voltage should be adjusted such that the output of the filter is always positive and in the measurable range of the ADC. If the 2nd amplifier stage does not saturate due to too high input signal it is recommended to always measure from the low-pass output due to the lower noise level at this node.

Even though the amplifier U5 (OPA2356) has rail-to-rail outputs their lower limits are at around 100mV which made it still necessary to introduce a negative supply voltage exclusively for the op-amp.

Data Converters

To digitize the analog signals a 16-bit ADC (U1, ADS8330) connected to an accurate 3.0 V voltage reference (U2, REF3230) is used. Furthermore a 12-bit DAC (U3, DAC124S85) is available to precisely set the NIR LED’s current, bias and offset voltage. The ADC and DAC are both communicating via the same SPI bus with the microcontroller.
LED Driver

The voltage-to-current converter used to drive the two NIR LEDs is based on Analog Devices' Circuit Note CN-0125 [18]. It is designed to drive a maximum current of 30 mA for each LED. This is the case when the input voltage is set to 495 mV.

In the initial circuit on the test setup there was always a little bit of current flowing through the LED even when the input voltage was set to zero. This is due to a minimal input offset voltage which is amplified by the huge open loop gain of the op-amp. Because of this leakage problem, so-called force-off lines connected to the negative op-amp input were added to the suggested setup. When setting those lines to logic high by the microcontroller, the LED is completely switched off. In normal operation the force-off lines must be set to input (high-impedance) on the microcontroller’s side.

Microcontroller & Bus

The ATxMega32A4U 8-bit AVR microcontroller represents the control unit of each module. It is the interface between the on-board logic and the other devices connected to the bus. It can be programmed using Atmel's PDI interface via the connector P4 using the USB-UART-Debug board.

The controller is clocked with the internal 32 MHz oscillator. This is tuned to a more precise value with the external 32.768 kHz clock crystal (Y1) as a reference.

To interface the modules with the PC, the I2C bus lines are available on the main connector (P1). All the modules are then connected to a host device (e.g. Atmel's AVR Xplain evaluation kit) which acts as the I2C bus master and translates the commands to a virtual serial interface via USB. Unfortunately this could not be implemented anymore, due to lack of time. But all the hardware is available such that is just a matter of time and firmware programming effort.

Power Supply

To generate the +3.3 V supply voltage used by the microcontroller and the analog circuitry, a LDO voltage regulator (U4, TPS79933) is used.

To overcome the problem of 100 mV minimum op-amp output voltage - as mentioned above in the SiPM section - a negative supply voltage is necessary. To not exceed the op-amp’s maximum operating voltage of 5.5 V an asymmetric negative voltage of \(-1.2 \text{ V}\) was chosen as output of the negative LDO regulator (U3, TPS7A300).

The power consumption of two modules were measured and are listed in Section 3.1.1.

Current Monitor

To protect the SiPM from overcurrent a current monitor circuit (U2, LMP8646) is implemented. It is connected to the microcontrollers analog comparator input such that a unwanted overcurrent event can be detected rapidly and the SiPM’s bias voltage can be disabled immediately.

USB-UART-Debug

The USB-UART-Debug board can be plugged onto the controller board via connector P4 (on ControllerBoard) during debugging. It provides a program and debug interface (PDI) connector (P3) compatible to the one on Atmel’s AVR Dragon programming device.

Furthermore a FTDI USB-to-UART bridge (U1, FT230X) is available to provide a possibility to easily communicate between PC and microcontroller via a virtual COM port.
2.3.3 Circuit Noise

In order to be able to resolve the desired signal change of about 0.1% - which corresponds to an SNR of 60 dB - precise analog components are used. This way the electronic circuit’s noise is kept negligibly low compared to that introduced by the SiPM.

A rough noise calculation has been carried out which is available in Appendix A. It can be shown that the SiPM’s dark noise is the main noise contributor compared to all the residual electronic noise.

The low circuit noise has been practically measured using a 9.1 kΩ test resistor connected in place of the SiPM. The bias voltage was set to $-28.5$ V. This leads to a current of 3.1 mA through $R_{\text{gain}}$ and thus to a voltage of 1.5973 V which corresponds to a normalized ADC value of 0.5324.

The results can be found in Section 3.1.2.

2.4 Microcontroller Firmware

To develop the microcontroller (Atmel, AVR xMega ATxmega32A4U) firmware Atmel’s proprietary IDE AVR Studio (Version: 5.1.208) was used. Thanks to the supplied software framework ASF (AVR Software Framework, Version: 2.11.1), fast development was possible. The ASF provides powerful libraries for interfacing the microcontroller’s hardware modules such as the DMA controller, USART module, etc.

The structure of the current firmware is shown by the block diagram in Figure 2.8.

Figure 2.8: Block diagram of the microcontroller Firmware. Self-written modules are framed in blue and those provided by ATMEL in green.

The source code was documented using Doxygen (Version: 1.8.1.1) and the corresponding documentation is available in HTML format on the DVD.

In the following paragraphs only the most important facts concerning the measurement procedure and the bus protocol are mentioned. For a more detailed documentation of the firmware please refer to the generated HTML documentation mentioned above.
2.4.1 NIR LED Switching

The devices can be connected to a modular network where one device is acting as light source and the others are acting as detectors such that various light paths can be measured easily. When a device should act as the optical light source the LEDs are switched in on-mode. To synchronize the measurement timing of the other devices (detecting the LEDs’ light signal) with the light emitting device, a synchronization line was introduced. The falling edge of this further called LED-SyncPort signalizes the beginning of each measurement period. In on-mode, the LEDs are switched consecutively as follows:

- \( t = 0 \text{ ms} \): Both LEDs are switched off (Backlight can be measured). LED-SyncPort is pulled low to signalize start of measurement period.
- \( t = 3.3 \text{ ms} \): LED A (680 nm) is switched on, LED B (850 nm) remains off. LED-SyncPort is pulled high.
- \( t = 6.6 \text{ ms} \): LED A (680 nm) is switched off, LED B (850 nm) is switched on.
- \( t = 9.9 \text{ ms} \): The cycle is restarted, which leads to a total measurement period of 9.9 ms.

This leads to a measurement rate of approximately 100 Hz. For a better understanding see also Figure 2.9.

2.4.2 SiPM Readout - Measurement

During normal operation, the device is switched in measurement mode which is described in the following subsection. The subsequent subsections concerning single-shot measurement and ADC DMA readout provide a deeper insight into the firmware but are not crucial for the understanding of the results.

Measurement Mode

To measure both LED as well as the backlight signals the microcontroller is running a timer which is started on every falling edge of the LED-SyncPort. For each of the three measurement sources (LED A, LED B, Backlight) 50 ADC samples are taken with a sampling frequency of 50 kHz, which leads to a measurement period of 1 ms per LED and backlight. Each of the three measurements is triggered by the measurement timer’s interrupts. The 50 samples are then averaged such that finally three 16-bit values represent one single sample point. This sample is then combined with a counter value (which is simply incremented for each measurement period) and sent to the PC via the bus. The exact sampling timing is shown in Figure 2.9.

Single-Shot Measurement

To facilitate the adjustment of the SiPM signals on the transimpedance- and low-pass outputs, the single-shot mode was introduced. It provides a kind of digital oscilloscope in combination with the PC software.

During 10 ms, 500 samples are taken from one ADC channel (see Figure 2.10). These values are directly sent to the PC without averaging or other modifications. This way, the pure ADC values are accessible on the PC side such that the SiPM bias voltage and the low-pass offset voltage can easily be adjusted to the ADC’s measurement range without the need for an oscilloscope. The result of a single shot in the GUI is seen in Figure 2.13.

When a device’s LEDs are on, the single-shot measurement is started with the LED-SyncPort going low. Otherwise a single-shot measurement is started in a random fashion.
Figure 2.9: Oscilloscope Screenshot of one Measurement Period (CH1: Transimpedance Amplifier Output, CH2: ADC_CS - ADC Chip-Select, CH3: LED-SyncPort). The backlight/LEDA/LEDB sampling is started at 1.8/5.1/8.4 ms and stopped at 2.8/6.1/9.4 ms respectively as it can be seen from the ADC_CS activity (active when sampling).

Figure 2.10: Oscilloscope Screenshot of a Single-Shot Measurement (CH1: Transimpedance Amplifier Output, CH2: ADC_CS - ADC Chip-Select, CH3: LED-SyncPort).
ADC DMA Readout

As shown in Figure 2.11, the ADC-DMA interaction incorporates two DMA channels and one sampling timer besides the ADC itself. One DMA channel (Request Data) constantly sends two bytes (Read Data Command) to the ADC as soon as one sample is available (ADC_INT pulled low). The second DMA channel (Read Data) stores the samples received from the ADC into a buffer and generates an interrupt if sufficient samples are read. To get \( N \) samples from the ADC the procedure is as follows:

- The two DMA channels and the corresponding buffers are initialized. The ADC is switched to the desired channel. The sampling timer is initialized.
- The Request Data DMA channel remains disabled but is initialized. The Read Data DMA channel is enabled and configured to generate an interrupt after \( N \) samples.
- Starting the sampling timer starts the measurement.
- Each sample is read one by one with the following sequence:
  1. The falling edge of the ADC_CONVST (generated by the PWM output of the sampling timer) starts ADC sampling.
  2. When ADC conversion has completed ADC_INT is pulled low by the ADC. This generates an interrupt on the microcontroller.
  3. The ISR activates the ADC (ADC_CS low) and triggers the Request Data DMA channel which sends the Read Data Command to the ADC.
  4. The SPI data read during the two byte transfer is automatically read by the Read Data DMA channel.
  5. When the two bytes have been transferred the Request Data DMA channel ISR is executed. It deactivates the ADC (ADC_CS high) and deactivates the DMA channel itself.
- As soon as \( N \) samples have been read by the Read Data DMA channel the sampling timer as well as the DMAs get disabled and the measurement can be further processed.

This then looks like on the oscilloscope screenshot in Figure 2.12.

![Figure 2.11: Schematic Overview of the ADC-DMA Interaction.](image-url)
Figure 2.12: Oscilloscope screen-shot of a single ADC sample showing the ADC-DMA interaction (CH1: ADC_CONVST, CH2: ADC_INT, CH3: ADC_CS, CH4: SCLK). The following events are represented by the red numbers: Start Conversion (1), Conversion Finished (2), Read ADC Sample (3), Deselect ADC and Disable DMA (5).
2.4.3 Bus

Due to lack of time, the I2C-Bus was not implemented. Instead, the debug UART interface was used to interface two devices on a pseudo bus. The communication protocol was programmed such that it can easily be extended to multiple device mode. Each device holds its unique bus address which is stored in the non-volatile EEPROM memory.

The basic communication block is a packet which consists of the following bytes:

\[
\text{[START]}|\text{DEVICE-ID}|\text{COMMAND}|\text{DATA}_1|\text{DATA}_2|\ldots|\text{DATA}_N|\text{END}\]

The data bytes are variable in length and can also be omitted completely. A command with device ID of zero (0x00) addresses all the devices and will therefore be executed by all of them.

Table 2.4 shows a brief summary of all the commands used. The measurement response is generated every time a measurement was triggered by the LED-SyncPort as long as the measurement mode is on for one of the two channels. This means that the measurement mode command must be executed only once at the beginning of the measurement and from then on the bus gets flushed with data as soon as a new measurement has been performed.

<table>
<thead>
<tr>
<th>Command Name</th>
<th>Command</th>
<th>N_{DATA}</th>
<th>Data Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>Set LED Mode 'L'</td>
<td>'L'</td>
<td>1 Byte</td>
<td>1 (on) or 0 (off)</td>
</tr>
<tr>
<td>Set LED Current 'I'</td>
<td>'I'</td>
<td>2x2 Bytes</td>
<td>16-bit DAC Values: [LEDA][LEDB]</td>
</tr>
<tr>
<td>Set Bias Voltage 'V'</td>
<td>'V'</td>
<td>2 Bytes</td>
<td>16-bit DAC Value</td>
</tr>
<tr>
<td>Set Offset Voltage 'O'</td>
<td>'O'</td>
<td>2 Bytes</td>
<td>16-bit DAC Value</td>
</tr>
<tr>
<td>Set Measurement Mode 'M'</td>
<td>'M'</td>
<td>1 Byte</td>
<td>0 (off), 1 (TI channel), 2 (LP ch.)</td>
</tr>
<tr>
<td>Measurement Response 'M'</td>
<td>'M'</td>
<td>4x2 Bytes</td>
<td>[Counter][LEDA][LEDB][Backlight]</td>
</tr>
<tr>
<td>Get Single-Shot 's'</td>
<td>'s'</td>
<td>1 Byte</td>
<td>1 (TI channel), 2 (LP channel)</td>
</tr>
<tr>
<td>Single-Shot Response 's'</td>
<td>'s'</td>
<td>501x2 Bytes</td>
<td>[Channel][Sample1][Sample500]</td>
</tr>
</tbody>
</table>

Table 2.4: Bus Protocol Commands: All [COMMAND] bytes are represented as ASCII characters. The data bytes are raw bytes. Note that all 16-bit values are transferred little-endian (LSB first).

2.5 Matlab GUI

The Matlab GUI is the interface between the User and the devices and allows to set the properties on the devices as well as to plot the data of the detectors. The software code is divided into the class SerialDriverSiPM and the GUI SiPM\_fNIRS. The class SerialDriverSiPM controls the data stream from the device over the USB-Link to Matlab. All received packages are checked for their integrity and converted according to their data type. The bus protocol commands are stated in Table 2.4. In the current version of the GUI SiPM\_fNIRS, it is possible to control one source device as well as one detector device. Figure 2.13 shows a Print Screen of the GUI. The possible settings in the GUI are explained in the following Sections.

Port

In the panel Port\(^4\) the port number of the virtual COM port is selected. With a click on the button Connect, the Matlab GUI is connected to the device, the feedback Connected is prompted in the Matlab Command Window and the other buttons are enabled.

\(^4\)The first time, the FTDI driver must be installed. To activate the COM port set the checkbox VCP laden/load VCP in Device Manager / USB Serial Controller / Properties / Extended-Menu.
Figure 2.13: Print Screen of the Matlab GUI, showing the result of a Single Shot. The red curve corresponds to the TIA channel and the blue one to the low-pass channel.

Source Device

The source and target addresses are unique numbers and depend on the devices used. The three modules are named with A, B and C and the corresponding addresses are 2, 3 and 1. 0 is the broadcast address. The LED current can be adjusted from 0 mA to 30 mA and the LEDs can be switched on and off.

Detector Device

The SiPM Bias Voltage can be adjusted in the range of \(-24\) V to \(-33\) V with the slider. The nominal value recommended by KETEK is between \(-28\) V and \(-29\) V. It might be necessary to adjust the SiPM Bias Voltage above or below this range in order to get a detector signal which lies in the range of the amplifiers.

The LP Offset Voltage is only used in connection with the Lowpass Channel and allows to shift the signal level. SiPM Bias Voltage and LP Offset Voltage are best adjusted by checking the result with the Get Single Shot measurement.

Plot Options

In the panel Plot Options the desired variables for plotting can be chosen before the measurement is started.

Measurement

The Get Single Shot measurement is very useful to check the settings and the range of the signals. Using this measurement one period of backlight, LED A and LED B is measured and plotted. With the button Start Measurement a continuous measurement is started, which stops again when the time in the field Time Left is expired or if it is manually stopped with the button Stop Measurement. It is recommended to use the Lowpass channel for the measurement, because the
data measured on this channel is filtered with an additional second order low-pass filter. The specification of the SDS is necessary for the calculation of the chromophore concentration.

**Post Processing**

The raw data can be filtered with a Butterworth filter of adjustable Order and Cut Off Frequency according to the settings in the panel Post Processing. The chromophore concentration is also recalculated when filtering. The SNR of the signal can be calculated but does only make sense when steady state signals, for example on a phantom, are measured.

**Save**

All the raw data and the settings are saved in a *.mat file in the specified folder. The settings are also saved in a *.txt file together with the Comment.

**HbO/HbR calculation**

The concentration of HbO and HbR is calculated employing the MBLL according to Equation \[1.3\]. The extinction coefficient are taken from [19] and are as follows: 
\[\epsilon_{\text{HbO}}^{680} = 0.974 \frac{1}{\text{cm mM}}, \epsilon_{\text{HbO}}^{850} = 2.526 \frac{1}{\text{cm mM}}, \epsilon_{\text{HbR}}^{680} = 6.011 \frac{1}{\text{cm mM}}, \epsilon_{\text{HbR}}^{850} = 1.798 \frac{1}{\text{cm mM}}.\]

The pathlength factor was calculated using Equation \[1.4\] with the absorption coefficients given in Section 2.6.2.

### 2.6 Measurement Setup

**2.6.1 Power Configuration**

For the powering of the devices three supply voltages are necessary, labeled in the schematics with: -35V, -5V0, 5V0. In the usual high power configuration (HPC) all the negative LDO-regulators were powered with -35 V. This configuration was used for the SNR, pulse, occlusion and HPC drift measurement.

For the low power configuration (LPC) the supply voltages were reduced according to Table 3.1. With this configuration one additional supply voltage for the -5V0 is necessary but the power consumption is drastically reduced. The second drift measurement was performed with the LPC.

**2.6.2 SNR Measurement Setup**

All SNR measurements were done using an ISS phantom (Phantom 2 of the BORL lab) which was placed into a black opaque plastic box. The phantom was covered with a thin (approximately 0.5 mm) piece of black imitation leather to prevent from direct optical source detector coupling. Ten holes were cut in the imitation leather with a spacing of approximately 10 mm for the placement of the source and detector which were fixed with tape on the phantom. The cables for the devices were inserted through a tiny hole into the box and the hole was sealed with black tape. The absorption coefficients are taken from the data sheet of Phantom 2: \[\mu_{\text{a690}} = 11.0 \frac{1}{\text{cm}}, \mu_{\text{a830}} = 0.104 \frac{1}{\text{cm}}, \mu_{\text{a690}} = 9.7 \frac{1}{\text{cm}}, \mu_{\text{a830}} = 0.100 \frac{1}{\text{cm}}.\]

The SNR was measured twice. First, the analog signals were measured directly on the AnalogBoard with the Oscilloscope (Model: TPS MSO7104B, Agilent Tech., Santa Clara, CA, USA) via the connector X2. The vertical resolution of the Oscilloscope is 8 Bit. Second, the SNR was measured with the ControllerBoard with the 16-Bit ADC via the microcontroller, called the Final Setup. In both setups 50 measurement periods of the signal were sampled at 50 kHz. Samples at 100 Hz were obtained by averaging the data points of the LEDs on-periods and subtracting the backlight level (The exact procedure is described in the next section). The SNR was calculated as the mean of the samples divided by the standard deviation. The SDS was varied from 20 to 60 mm, and the
SNR was determined 10 times each. Please note that SDS has its lower bound at 13 mm given by the mechanical dimensions of the device.

**SNR Measurement Analog Board with Oscilloscope**

The signal was sampled with the Oscilloscope at 200 kHz and then downsampled to 50 kHz to get the same resolution as in the Final Setup. The 100 Hz samples were obtained by averaging 50 data points (corresponds to 1 ms) belonging to 50-90% of the LED’s on-periods and subtracting the backlight level which was calculated analogously. The averaged range of the signal is marked in Figure 2.14. This selection avoids that the transient part of the signal is averaged.

![Figure 2.14: One period of the measurement sequence showing backlight, LedA and LedB. The signal was sampled with 100 kHz using the low-pass output (X2). For a more intuitive understanding the y-axis of the plot was flipped such that an increasing light intensity corresponds to a rising slope.](image)

**SNR Measurement Final Setup**

The selection of data points for the 100 Hz samples is shown in Figure 2.9 and is time based. This selection is comparable to the procedure in Section 2.6.2.

### 2.6.3 Pulse and Occlusion Measurement

The devices were covered on the bottom side of the PCB with a piece of black silicone. This was to prevent any contact of the skin with the electrical circuit. The devices were placed on the skin, fixed and afterwards covered with a changing bag to reduce backlight. Changing bags are opaque and normally used for photo development.
Chapter 3

Results

3.1 Circuit Performance

3.1.1 Power Consumption

Table 3.1 shows the current and power consumption of two connected devices in different modes (Source and Detector) in the LPC. The values were directly read from the power supply’s display (Hameg Programmable Power Supply HMP4030).

During a measurement the overall power consumption of the two modules is around 545 mW. When the modules are in standby (LEDs and SiPM Bias Voltage disabled) they still consume 375 mW.

3.1.2 Noise

The results of the noise measurement described in Section 2.3.3 are shown in Figure 3.1. In the left figure, the quantization steps are clearly visible. Over 70 seconds the ADC reading only varies within a maximum of ±2 quantization steps (≈ 92 µV). Whereas on the right, blowing at the resistor and therefore changing its - and also the surrounding circuit’s - temperature results in a clearly visible change of the ADC reading of 100 quantization steps (≈ 4.5 mV).

The start up drift for this configuration (results not shown in figure) was roughly quantified with 0.0083%/s.

<table>
<thead>
<tr>
<th>Supply</th>
<th>Voltage [V]</th>
<th>Current [mA]</th>
<th>Power [mW]</th>
<th>Module State</th>
</tr>
</thead>
<tbody>
<tr>
<td>-35V</td>
<td>-32.0</td>
<td>-3.2</td>
<td>102.6</td>
<td>Measuring, VBias on</td>
</tr>
<tr>
<td></td>
<td>-32.0</td>
<td>0.0</td>
<td>0.0</td>
<td>VBias off</td>
</tr>
<tr>
<td>5V0</td>
<td>3.5</td>
<td>105</td>
<td>367.5</td>
<td>Measuring, LEDs &amp; VBias on</td>
</tr>
<tr>
<td></td>
<td>3.5</td>
<td>103</td>
<td>360.5</td>
<td>LEDs enabled, VBias off</td>
</tr>
<tr>
<td></td>
<td>3.5</td>
<td>85.4</td>
<td>298.9</td>
<td>VBias &amp; LEDs off</td>
</tr>
<tr>
<td>-5V0</td>
<td>-3.0</td>
<td>-24.8</td>
<td>74.4</td>
<td>Measuring, LEDs &amp; VBias on</td>
</tr>
<tr>
<td></td>
<td>-3.0</td>
<td>-25.3</td>
<td>75.9</td>
<td>VBias &amp; LEDs off</td>
</tr>
</tbody>
</table>

1 VBias Regulator 2 +3.3V Regulator 3 -1.2V Regulator

Table 3.1: Power consumption of two modules: one acting as LED source and the other one as detector equipped with a SiPM. The configuration was as follows: -28.0 V SiPM bias voltage, 3.0V offset voltage and 30mA LED current.
Figure 3.1: (Left) TI amplifier output reading of ADC values resulting from the constant current through a 9.1 kΩ resistor (VBias = −28.5 V) present instead of the SiPM. The device was warmed up during 5 minutes to avoid measuring startup (temperature) drift. (Right) Same configuration as on the left but with twice strongly blowing at the resistor at around 15 s and 35 s to change its resistance by a temperature change.

3.2 SNR Measurements

The SNR measurements were executed as described in Section 2.6.2. The resulting measurements were compared to the SNR limit stated in Equation 2.6 as discussed in Section 2.1.4. This is depicted by the gray shaded area in Figure 3.2 and 3.3.

3.2.1 Analog Board with Oscilloscope

Figure 3.2 shows the SNR in function of SDS. The signals were measured at the low-pass output, except for SDS = 20 mm where the data was picked up directly at the TIA output. For an SDS of 20 to 30 mm, the LED intensities were adjusted to prevent amplifier saturation. The SNR is continuously increasing when the SDS is reduced but shows a kink at an SDS of 40 mm. At an SDS of 30 mm the SNR reaches the maximum at approximately 60 dB for a wavelength of 850 nm.

3.2.2 Final Setup

Figure 3.3 shows the SNR in function of SDS for the Final Setup. The signals were measured at the low-pass output and the LED currents were set to 30 mA for all SDS. These settings became possible (in contrast to the Analog Board with Oscilloscope) as the bias voltage could be changed in a wider range in the Final Setup. The SNR is continuously increasing when the SDS is reduced, but shows a kink at an SDS of 30 mm. At an SDS of 20 mm the SNR reaches the maximum for both wavelengths at approximately 70 dB.
Chapter 3. Results

Figure 3.2: SNR in function of SDS for the analog board measured with the oscilloscope. Error bars indicate standard deviation of ten measurements and estimated 1.5 mm positioning uncertainty, respectively. The measurements were carried out for two LEDs with a wavelength of 680 nm and 850 nm. The theoretical SNR - mainly limited by shot noise - is shown by the gray shaded area. The limitation is calculated for a bandwidth of 100 Hz, a wavelength of 850 nm and an optical LED power of 2.6 mW.

Figure 3.3: SNR in function of SDS for the Final Setup. Error bars indicate standard deviation of ten measurements and estimated 1.5 mm positioning uncertainty, respectively. The measurements were carried out for two LEDs with 680 nm and 850 nm wavelength. The theoretical SNR - mainly limited by shot noise - is shown by the gray shaded area. The limitation is calculated for a bandwidth of 100 Hz, a wavelength of 850 nm and an optical LED power of 2.6 mW.
3.3 Pulse

The two measurements in Figure 3.4 and Figure 3.5 show the pulse of one of the author’s forehead with an SDS of 40 mm and 50 mm, showing a pulse of approximately 64 bpm and 72 bpm, respectively.

3.4 Occlusion

The two occlusion measurements performed at the lower-left arm of each of the two authors are shown in Figure 3.6 (Top) and Figure 3.7 (Top). During all occlusions the expected increase in HbR- and decrease in HbO-concentration is clearly visible for both measurements. All three signals in Figure 3.7 (Bottom) show a constant positive drift (rising light intensities) over the whole measurement period which is of significant magnitude (backlight signal 0.086 %/s). Also for the measurement depicted in Figure 3.6 (Bottom) a drift is visible. It is about 0.063 %/s for the backlight signal. The drift coefficients are normalized to the maximal ADC value (This is the same normalization as in all measurement plots). This drift stagnates for a short time (Figure 3.6) or even changes to positive direction (Figure 3.7) each time the arm is released from occlusion. Please note that the time interval of occlusion, shaded in gray, does not perfectly fit to the change in signal (due to timing inaccuracies of the measurement and on the microcontroller clock). This is best visible in Figure 3.7 (Top) at the times of 500 and 800s.

3.5 Drift

Two measurements were performed to evaluate the drift of all three signals (LED A, LED B, backlight) right after startup of the device during 20 minutes. The usual HPC, powering all the negative LDO-regulators with $-35$ V, was used (Figure 3.8) and compared to the LPC (Figure 3.9). The overall power consumption is 1500 mW or 545 mW for the HPC and the LPC respectively. This means that the residual 955 mW were simply "burned" to heat up the device. The power consumption values are based on the measurements shown in Table 3.1. Table 3.2 summarizes all the drift coefficients corresponding to the Figures 3.8 and 3.9.

<table>
<thead>
<tr>
<th></th>
<th>High Power Configuration (HPC) [%/s]</th>
<th>Low Power Configuration (LPC) [%/s]</th>
</tr>
</thead>
<tbody>
<tr>
<td>Back Light</td>
<td>0.071</td>
<td>0.012</td>
</tr>
<tr>
<td>LED A 0-200 s</td>
<td>-0.11</td>
<td>-0.020</td>
</tr>
<tr>
<td>LED B 0-200 s</td>
<td>-0.043</td>
<td>-0.0095</td>
</tr>
<tr>
<td>LED A - Backlight</td>
<td>-0.052</td>
<td>-0.016</td>
</tr>
<tr>
<td>LED B - Backlight</td>
<td>-0.045</td>
<td>-0.012</td>
</tr>
</tbody>
</table>

Table 3.2: Drift coefficients for different power configurations. A positive coefficient corresponds to a higher light intensity. The drift coefficients are normalized to the maximal ADC value.
Chapter 3. Results

<table>
<thead>
<tr>
<th>Time [s]</th>
<th>LED A − Backlight</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>−0.1</td>
</tr>
<tr>
<td>4</td>
<td>−0.095</td>
</tr>
<tr>
<td>6</td>
<td>−0.09</td>
</tr>
<tr>
<td>8</td>
<td>−0.085</td>
</tr>
<tr>
<td>10</td>
<td>−0.08</td>
</tr>
<tr>
<td>12</td>
<td>−0.075</td>
</tr>
<tr>
<td>14</td>
<td>−0.07</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Time [s]</th>
<th>LED B − Backlight</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>−0.21</td>
</tr>
<tr>
<td>4</td>
<td>−0.2</td>
</tr>
<tr>
<td>6</td>
<td>−0.19</td>
</tr>
<tr>
<td>8</td>
<td>−0.18</td>
</tr>
<tr>
<td>10</td>
<td>−0.17</td>
</tr>
<tr>
<td>12</td>
<td>−0.16</td>
</tr>
<tr>
<td>14</td>
<td>−0.15</td>
</tr>
</tbody>
</table>

**Figure 3.4**: Pulse measurement performed at one of the author’s forehead with an SDS of 40 mm, $I_{\text{LED}} = 30$ mA, $V_{\text{bias}} = -29.7$ V, $V_{\text{offset}} = 0.72$ V. The signals were measured at the low-pass output and filtered off-line with a digital 4 Hz low-pass filter. (Top) Shows the backlight corrected normalized ADC values of both LEDs. (Bottom) Resulting chromophore concentration change.

<table>
<thead>
<tr>
<th>Time [s]</th>
<th>$\Delta$[HbO] [µM]</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>−0.62</td>
</tr>
<tr>
<td>4</td>
<td>−0.6</td>
</tr>
<tr>
<td>6</td>
<td>−0.58</td>
</tr>
<tr>
<td>8</td>
<td>−0.53</td>
</tr>
<tr>
<td>10</td>
<td>−0.52</td>
</tr>
<tr>
<td>12</td>
<td>−0.51</td>
</tr>
<tr>
<td>14</td>
<td>−0.5</td>
</tr>
</tbody>
</table>

<table>
<thead>
<tr>
<th>Time [s]</th>
<th>$\Delta$[HbR] [µM]</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>−0.03</td>
</tr>
<tr>
<td>4</td>
<td>−0.02</td>
</tr>
<tr>
<td>6</td>
<td>−0.01</td>
</tr>
<tr>
<td>8</td>
<td>0</td>
</tr>
<tr>
<td>10</td>
<td>0.01</td>
</tr>
<tr>
<td>12</td>
<td>0.02</td>
</tr>
<tr>
<td>14</td>
<td>0.03</td>
</tr>
</tbody>
</table>

**Figure 3.5**: Pulse measurement performed at one of the author’s forehead with an SDS of 50 mm, $I_{\text{LED}} = 30$ mA, $V_{\text{bias}} = -28.5$ V, $V_{\text{offset}} = 0.81$ V. The signals were measured at the low-pass output and filtered off-line with a digital 4 Hz low-pass filter. (Top) Shows the backlight corrected normalized ADC values of both LEDs. (Bottom) Resulting chromophore concentration change.
Figure 3.6: Occlusion measurement performed at one of the author’s (T. Achtnich) lower-left forearm (musculus brachioradialis) with an SDS of 40 mm, $I_{\text{LED}} = 30$ mA, $V_{\text{Bias}} = -28.7$ V, $V_{\text{Offset}} = 0.60$ V. The upper arm was occluded using a pneumatic cuff inflated to a pressure of 280 mmHg during the time intervals of 50-250 s and 450-650 s (shaded in gray). (Top) Shows the resulting chromophore concentration change. (Bottom) Shows the three raw signals measured at the low-pass output with subsequent 0.1 Hz digital filtering. These signals were used as input values for calculating the chromophore concentration. For a more intuitive understanding the y-axis of the plot was flipped such that an increasing light intensity corresponds to a rising slope.
Figure 3.7: Occlusion measurement performed at one of the author’s (F. Braun) lower-left forearm (musculus brachioradialis) with an SDS of 45 mm, $J_{\text{LED}} = 30$ mA, $V_{\text{Bias}} = -29.5$ V, $V_{\text{Offset}} = 0.78$ V. The upper arm was occluded using a pneumatic cuff inflated to a pressure of 280 mmHg during the time intervals of 50-200 s, 350-500 s and 650-800 s (shaded in gray). (Top) Shows the resulting chromophore concentration change. (Bottom) Shows the three raw signals measured at the low-pass output with subsequent 0.1 Hz digital filtering. These signals were used as input values for calculating the chromophore concentration. For a more intuitive understanding the y-axis of the plot was flipped such that an increasing light intensity corresponds to a rising slope.
Figure 3.8: Startup drift measurement performed with the HPC on the phantom with an SDS of 40 mm, \(I_{\text{LED}} = 30\ \text{mA, } V_{\text{Bias}} = -28.0\ \text{V, } V_{\text{Offset}} = 0.94\ \text{V. The } V_{\text{Bias}}-\text{LDO- and the } -1.2\text{V- LDO-regulator were powered with } -35\ \text{V. The signals were measured at the low-pass output and filtered off-line with a digital 1 Hz low-pass filter. For a more intuitive understanding the y-axis of the two upper plots were flipped: such that increasing light intensity corresponds to a rising slope in all three plots.}

Figure 3.9: Startup drift measurement performed with the LPC on the phantom with an SDS of 40 mm, \(I_{\text{LED}} = 30\ \text{mA, } V_{\text{Bias}} = -28.0\ \text{V, } V_{\text{Offset}} = 0.78\ \text{V. The } V_{\text{Bias}}-\text{LDO- and the } -1.2\text{V- LDO-regulator were powered with } -32\ \text{V and } -3.0\ \text{V respectively. The signals were measured at the low-pass output and filtered off-line with a digital 1 Hz low-pass filter. For a more intuitive understanding the y-axis of the two upper plots were flipped: such that increasing light intensity corresponds to a rising slope in all three plots.}
Chapter 4

Discussion

4.1 Circuit Performance

The overall power consumption (see Section 3.1.1) of the devices (545 mW) must be related to the power consumption of the SiPM (approximately 150 mW) and the both LEDs (approximately 66 mW). The power consumption of the SiPM and the LEDs can not be reduced without lowering the signal strength. But there is some room for improvement as the hardware of the devices was not trimmed to extremely low power consumption. It was not a goal of this project to design a very low power device but the experience showed that heat dissipation worsens the temperature drift of the signals (see Section 4.5). Therefore reducing the overall power consumption and thus the excess heat is strongly suggested.

The two measurements to quantify the circuit noise (setup with test resistor instead of the SiPM) in Section 3.1.2 clearly show the high accuracy and sensitivity of the analog circuit. Therefore the circuit noise can be neglected compared to other noise sources such as dark noise and especially shot noise. The start-up drift is influenced by the temperature drift of the test resistor and the drift of the circuit itself, hence the measurement of the start-up drift shows only the order and not a precise value of the circuit drift.

4.2 SNR Measurement

The SNR measured with the oscilloscope (see Figure 3.2) was expected to be higher at SDS smaller than 40 mm. The SNR is reduced in this range for three reasons. First, the LED intensities were reduced to prevent amplifier saturation and therefore the signal was limited. Second, the SNR is limited by the signal-to-quantization-noise ratio (SQNR). The SQNR is stated in Equation 4.1.

\[
\text{SQNR} \approx 20 \cdot \log_{10}(2^N) + 10 \cdot \log(n_{\text{ovs}}) \quad (\text{Source: }[20]) \quad (4.1)
\]

Where \(N\) is the number of bits of the ADC and \(n_{\text{ovs}}\) the oversampling factor. The SQNR for \(N = 8\) and \(n_{\text{ovs}} = 50\) is approximately 65 dB and shows the limitation caused by the quantization. Third, the signal has a temperature drift which is discussed in Section 4.5.

The SNR measurement performed with the Final Setup has an improved SNR. First, the LED intensities stayed the same for all SDS, only the bias voltage was adjusted which improved the SNR. Second, the ADC of the Final Setup has a resolution of 16 Bits and the SQNR is approximately 113 dB. Nevertheless, the SNR in Figure 3.3 has still a kink at SDS = 30 mm. Besides the shot noise, the electronic noise induced by the LED driver and the residual electronic circuit, one assumes that the drift is a crucial source of the reduced SNR. This is discussed in Section 4.5. Furthermore we see in Figure 3.3 that the Final Setup achieves an SNR of 60 dB up to an SDS of 38 mm.
4.3 Pulse

The pulsation in Figure 3.5 (Bottom) is best seen in the change of HbO (red line). This seems to be plausible as with each pulse “fresh” (oxygenated) blood is pumped through the body delivering new HbO. The shape of the HbO change corresponds best with the light of the 850 nm wavelength (compare Figure 3.5 (Top), blue line). This makes sense, since light with a wavelength of 850 nm is more absorbed by HbO, than light with 680 nm wavelength (see Figure 1.1). The same considerations are true for the measurements with SDS = 40 in Figure 3.4.

The unexpected signal change in Figure 3.5 between five and seven seconds is assumed to be caused by motion artifacts.

Interestingly, the signal amplitudes of the LEDs are larger for the SDS of 50 mm than for 40 mm. One must keep in mind that the SDS is not the only source for attenuating the signal. The type of tissue below the source and detector can have an even higher influence. It makes a difference whether the detector or source is placed directly on a blood vessel, muscle or even bone. This could explain the higher LED amplitudes for SDS of 50 mm compared to SDS of 40 mm.

4.4 Occlusion

In Figures 3.6 (Top) and 3.7 (Top) the change of both chromophore’s concentration is nicely visible. The decrease of HbO is due to missing supply of “fresh” (oxygenated) blood and the accumulation of deoxygenated hemoglobin leads to an increase in the HbR concentration. The concentration change of up to 18 $\mu$M is comparable to the results in [21].

These results of the chromophore’s concentration changes provide evidence for the correct qualitative functioning of our setup.

The drift of the two occlusion measurements is discussed in the following section.

4.5 Drift

The drift of the backlight signal is approximately 0.071%/s for the HPC but only about 0.012%/s for the LPC (see Table 3.2). This shows clearly that the 1000 mW of additional power consumption of the LDO regulators - resulting in excess heat - heated up the SiPM detector. Thus the SiPM’s gain changed significantly by this increase in temperature. (Based on the results in Figure 3.1 - showing a low temperature dependency of the amplifier and data converter circuit - it is assumed that the signal change originates mostly from the SiPM.). This effect is referred to as external heating.

In Figure 3.6 (Bottom) it is visible that the backlight signal has an unexpected sharp decrease right after the occlusion is stopped. Less light enters the SiPM at this moment and the SiPM is heated up less by the light of the signals. This effect is referred to as internal heating.

Internal and external heating change the temperature of the SiPM and hence the SiPM’s gain (The temperature coefficient of the SiPM’s gain is specified with $\leq 1%/°C$.).

The drift of the signals LED A and LED B in Figure 3.8 (Middle) shows an unexpected behavior. From zero to approximately 200 s the signals were decreasing and after 200 s they were increasing again. As the drift coefficient is negative for the first 200 s and opposite to the drift coefficient of the backlight, this drift is assumed to result from the LED circuit. In other words, the intensity of the LEDs has decreased. In general, LEDs are known to have a negative nonlinear temperature coefficient which could describe this behavior. It is suggested to further isolate, quantify and fix the source of this problem.

The increase of the signal after the first 200 s is assumed to stem again from the temperature coefficient of the SiPM’s gain as mentioned in the last paragraph.

The Figures 3.8 (Bottom) and 3.9 (Bottom) show the backlight compensated LED signals and it can be seen that the drift is not completely leveled off as one would expect. The reason for that
is supposed to stem from the nonlinearity of the SiPM for higher light intensities, as shown in Figure 1.4. In other words, low level signals (backlight) were amplified linearly while for high level signals (LED A, LED B) the SiPM enters the non-linear region. Especially for high light intensities (short SDS) the SiPM looses its linear behavior and the backlight compensation starts to fail.

Based on the backlight corrected signals in Figure 3.8 and 3.9 the chromophore concentration change was calculated. A clear drift was visible with a maximum amplitude of 1 $\mu$M over the whole measurement period of 20 minutes. This shows that the drift is a severe problem when measuring small concentration changes. It also explains why the drift in Figure 3.7 and 3.6 is barely visible due to the high signal change.

The LPC could reduce the drift of the signals by approximately a factor of three compared to the HPC (see Table 3.2 last two rows). Nevertheless, the drift is a severe problem which must be investigated further on, especially on the side of the LED circuit.
Chapter 5

Conclusion

The main goal of this project was to build a modular fNIRS probe with a SiPM detector and to evaluate the prototype’s performance.

Two PCBs have been designed and stacked on top of each other. They include LEDs and SiPM - as fNIRS source and detector, the corresponding signal processing circuit, a microcontroller and the power management. The result is a very compact (26 mm x 26 mm footprint) device with very low bias voltage and robustness to light overexposure. The performance of this device is shown with the measurements in Chapter 3.

The goals stated in Section 1.4.1 are all fulfilled with the exception of the BUS implementation. The hardware of the devices is ready for BUS communication but the implementation of the software is not yet done, due to lack of time. With the future implementation of the BUS the devices can be interconnected to a modular network.

The device’s signal-to-noise performance was determined on a phantom with an SDS of 20 mm up to 60 mm. The required SNR of 60 dB, to reliably detect cortical activities in the brain, is achieved up to an SDS of 38 mm. When the requirements for the SNR are reduced to 40 dB (corresponds to 1% resolution), the SNR is attained up to an SDS of 56 mm, which is an excellent result.

All SiPM signals are affected by a significant drift which is assumed to be caused by temperature changes in the SiPM and the LEDs itself and their surrounding electronic circuit. This drift must be further investigated and reduced with methods suggested in Section 6.1.

The above mentioned SNR performance made it possible to measure the pulse on the forehead up to an SDS of 50 mm. It seems to be possible to even increase the SDS of the pulse measurement by a clever placement of the source and detector.

Furthermore, the operational reliability was evidenced with measurements showing the chromophore concentration change when the arm of a human is occluded by using a pneumatic cuff. This occlusion measurements were performed with an SDS of 40 and 45 mm and led to satisfying results.

To the end, the measurements can be comfortably performed, evaluated and stored with the GUI implemented in Matlab.
Chapter 6

Outlook

Based on our experiences during this project we propose the following steps for continuing this work.

6.1 Drift

The main problem of the current device is the drift which is assumed to originate mostly from temperature changes. This problem cannot easily be overcome as the SiPM heats up itself when exposed to light. A decrease of the light intensities would diminish the SNR and therefore also the resolution of the device.

As a first step, the temperature dependency of the LED circuit must be investigated in detail and probably a redesign of the circuit is necessary.

As a second step, the external warming caused by the surrounding electronic components must be reduced to a minimum. The LDO voltage regulators should be replaced by switching mode power supplies, paying attention not to deteriorate the supply voltage quality.

As a third step, the SiPM should be actively cooled and temperature controlled. This is already done in the MiniSL modules of the company SensL which are cooled using a Peltier thermoelectric cooler. This points out the necessity of cooling the SiPM.

6.2 Hardware

PCB The size of the PCBs can be further reduced by using BGA footprints for the active components and 0402 (or even smaller) footprints for the passive components where possible.

SPI ADC/DAC The device’s on board SPI BUS, connecting the ADC and DAC with the microcontroller, was operated at 16 MHz. This is a challenging clock rate and the signal quality on the BUS lines was highly perturbed by reflections. Therefore a matching of the transmission lines using serial resistors and terminating R/C "snubber" networks is necessary. The matching of the transmission lines could not be implemented retroactively as their is no place on this very compact device. But it is warmly recommended to do so for future versions of the PCB.

Current Monitor A current monitor is provided on the PCB but it was never used because the maximal current was adjusted with the power supply. If in future a power supply without adjustable current limiter is used, it is recommended to implement the current monitor by software.
6.3 GUI

The measurement data, sent by the detector device over the serial link, is sequentially processed and plotted in the Matlab GUI. When Matlab is too busy with processing the data, the serial port is processed to irregularly and the serial link breaks which leads to a complete crash of Matlab. This problem has also been reported by other Matlab users in forums, but a solution was not presented. The workaround which is currently implemented is to adapt the refresh rate of the plot according to the processing demand. Nevertheless, it can happen, that the link gets stuck after more than 20 min of a measurement procedure. Of course, this problem must be improved in future versions.

6.4 Further Improvements

Besides the improvements tackling the drift as mentioned in Section 6.1 one should try further methods to improve the SNR.

If the focus lies on increasing the SNR for SDS longer than 38 mm, the limitation caused by the shot noise must be addressed. The shot noise can be reduced by lowering the bandwidth. Reducing the bandwidth by a factor of $n$ will increase the linear SNR by a factor of $\sqrt{n}$. For the current circuit setup this will also reduce the sampling rate which should be avoided. If a high sampling rate is required together with a small bandwidth, more sophisticated signal modulation and processing should be used, e.g. a lock-in amplifier could solve the problem. This brings up the question whether implementing this rather complex type of circuit is worth the effort and whether it can be miniaturized to the small footprint available.
References


Appendix A

Noise Estimation

A.1 Identified Noise Sources

- SiPM Dark Noise
  - Dark Rate: \( DR_{SiPM} = 500 \frac{kHz}{mm^2} \) [Source: Ketek Datasheet]
  - Detector Area: \( A_{SiPM} = 3 \times 3 mm^2 = 9 mm^2 \)
  - Microcell Charge: \( Q_{MC} = 60 fC \) [Source: T. Ganka, Ketek]
  - Dark Current: \( I_{Dark} = Q_{MC} \cdot A_{SiPM} \cdot DR_{SiPM} = 270 nA \)
  - Resulting Voltage Noise at Transimpedance Amplifier Output:
    \( V_{Dark} = I_{Dark} \cdot R_{Gain} = 270 nA \cdot 510 \Omega = 138 \mu V \)

- LED Noise (due to DAC Noise)
  - DAC Noise: \( V_{n-DAC} < 80 \mu V \) [Source: TI Forum]
  - LED Voltage: \( V_{LED}(@I_{LED} = 30 mA) = 500 mV \)

- SiPM Gain Noise (due to VBias & DAC Noise)
  - DAC Noise: \( V_{n-DAC} < 80 \mu V \) [Source: TI Forum], results in triple the noise due to voltage setting resistor configuration: \( V_{nBias-DAC} = 240 \mu V \)
  - LDO Noise: \( V_{nBias-LDO} < 20 \mu V \)
  - Total Noise at \( V_{Bias} \): \( V_{nBias} = V_{nBias-DAC} + V_{nBias-LDO} < 260 \mu V \)
  - Gain Sensitivity: \( \frac{dG}{G} \approx 7 \cdot \frac{dV_{Bias}}{V_{Bias}} = 6.1 \cdot 10^{-5}(@V_{Bias}=30 V, dV_{Bias} = V_{nBias} = 260 \mu V) = 0.006\% \) [Source: Buzhan et al. [5]]

- ADC Noise
  - Quantization Step: \( \Delta = \frac{V_{Ref}}{2^{n \text{bits}}} = \frac{3.0 V}{2^{16}} = 45.8 \mu V \)
  - Total ADC Noise: \( V_{n-ADC} < 100 \mu V \)

- Lowpass Offset Noise (due to DAC Noise)
  - DAC Noise: \( V_{n-DAC} < 80 \mu V \) [Source: TI Forum]
  - Gain of Lowpass Stage: \( G = 2 \)
  - Noise at Lowpass Output: \( V_{n-LP-Offset} = G \cdot V_{n-DAC} = 160 \mu V \)

- OpAmp Noise (Transimpedance & Lowpass Stage)
  - According to Tina-TI Simulation: \( \ll 100 \mu V \)
A.2 Noise at Transimpedance Output

- Assume Minimal Signal Amplitude of $V_s = 100\text{mV}$
- Noise Contributions:
  - Dark Noise: $V_{Dark} = 138\mu V$
  - LED (via DAC) Noise: $V_{n-LED} = V_s \cdot \frac{V_{n-DAC}}{V_{LED}} = 100\text{mV} \cdot \frac{80\mu V}{500\text{mV}} = 16\mu V$
  - SiPM Gain Noise: $V_{n-Gain} = \frac{dG}{G} \cdot V_s = 6.1 \cdot 10^{-5} \cdot 100\text{mV} = 6.1\mu V$
  - ADC Noise: $V_{n-ADC} < 100\mu V$
  - OpAmp Noise with OPA2356 very small and therefore negligible
- Total Noise: $V_n = 259.8\mu V$
- $\text{SNR}_{lin} = \frac{100\text{mV}}{259.8\mu V} = 385; \quad \text{SNR}_{dB} = 20\log_{10}(\text{SNR}_{lin}) = 51.7\text{dB}$

A.3 Noise at Lowpass Output

- Assume Minimal Signal Amplitude of $V_s = 200\text{mV}$ (2x gain of previous example)
- Noise Contributors:
  - Same as for TI output, but amplified by factor 2: $V_{n-LP-TI} = 519.5\mu V$
  - Lowpass Offset Noise: $V_{n-LP-Offset} = 160\mu V$
- Total Noise: $V_n = 679.5\mu V$
- $\text{SNR}_{lin} = \frac{200\text{mV}}{679.5\mu V} = 294; \quad \text{SNR}_{dB} = 20\log_{10}(\text{SNR}_{lin}) = 49.4\text{dB}$
Appendix B

Hardware Schematics

B.1 Test Setup

- **PreAmp**: transimpedance amplifier with two freely configurable (low-pass) filters
- **LEDriver**: voltage-to-current converters to drive the NIR LEDs
B.2 Final Setup - PCB Version (Revision A)

- **AnalogBoard** (4 pages)
  1. Main Schematic
  2. SiPM Bias and Amplifier Stages
  3. Data Converters
  4. LED Driver

- **ControllerBoard** (3 pages)
  1. Main Schematic
  2. Power Supply
  3. Microcontroller
V_Adjust = 0...3V --> V_SiPM = -27...-33V
REM: OFF_LED_A, OFF_LED_B put high (3.3V) by microcontroller to ensure LED is switched OFF (if offset voltage of OpAMP might cause problems). In working mode this pin must be set to high impedance by microcontroller.
Power Supply: -35V, +3.3V, -3.3V (SiPM fNIRS Project)

V_{max} = 100mV @ 5mA

-35V_SENSED

100n
C10
GND

R10
360k

R6
R_sense = 200E

R9
22k

C11
470p

R12
1M

R13
10n

R14
10\,\mu\text{F}, 6.3V

R15
10n

R16
10\,\mu\text{F}, 10V

R17
10\,\mu\text{F}, 10V

C12
100n

C13
180k

C14
10n

C15
10\,\mu\text{F}, 50V

C16
10\,\mu\text{F}

C17
[470p]

C18
10\,\mu\text{F}, 10V

C19
10\,\mu\text{F}, 6.3V

B_{\text{bias current sense}}

470p
C11
GND

V_{out} = 1V @ 5mA (relative to -35V)

\rightarrow -34V (rel. to GND) @ 5mA

\rightarrow V_{\text{ADC}} = 1.09V (@ 0mA) vs. 1.15V (@ 5mA)

\begin{align*}
    \text{V}_{\text{out}} &= 1V @ 5mA \\
    \text{V}_{\text{ADC}} &= 1.09V (@ 0mA) \text{ vs. } 1.15V (@ 5mA)
\end{align*}
B.3 Final Setup - Corrected Version (Revision B)

- **AnalogBoard** (4 pages)
  1. Main Schematic
  2. SiPM Bias and Amplifier Stages
  3. Data Converters
  4. LED Driver

- **ControllerBoard** (3 pages)
  1. Main Schematic
  2. Power Supply
  3. Microcontroller
Transimpedance Amplifier
* Gain = R_Gain = 510
* Cutoff-Frequency = 1/(2*pi*R_Gain*C_Filter) = 1 kHz

2nd Order Lowpass with adjustable Offset
* Gain = 2
* Cutoff-Frequency = 1kHz

Zeichnungs Titel:
SiPM Bias, TI-Amplifier and Lowpass (SiPM fNIRS Project)
File:
Datum: 02.07.2012 17:33:44
Zeichnungsnummer: 2

ETH
Eidgenössische Technische Hochschule Zürich
Swiss Federal Institute of Technology Zurich

SiPM Bias, TI-Amplifier and Lowpass (SiPM fNIRS Project)

Zeichnungsnummer: 2
Rev: B
Datum: 02.07.2012 17:33:44

File: C:\sipm\Hardware\V2_RevB\SiPM_Amp.SchDoc
REMARK:
OFF_LED_A, OFF_LED_B: put high (3.3V) by microcontroller to ensure LED is switched OFF (if offset voltage of OpAMP might cause problems). In working mode this pin must be set to high impedance by microcontroller.
Power Supply: -35V, +3.3V, -3.3V (SiPM fNIRS Project)

V_max = 100mV @ 5mA
V_{out} = 1V @ 5mA (relative to -35V)
\rightarrow -34V (rel. to GND) @ 5mA
\rightarrow V_{ADC} = 1.09V (@ 0mA) vs. 1.15V (@ 5mA)
Actual Pseudo "Bias" (UART) Config.:

* BUS_SDA connected to UART_RXD for all devices
* BUS_SCL connected to UART_TXD only for 1 master (with SiPM, address 0x01)

Need to swap SCLK/SDI to be compatible with USARTC1 module in SPI mode. For now USARTD1 is used in SPI mode and the pins are routed as follows: SDI<->PD7; SDO<->PD6; SCLK<->PD5

Furthermore all the high frequency lines (SCLK, SDO, SDI, CS, ... ) must be matched to avoid ringing/overshoots. (Use serial resistance in all lines and R&C (in series, snubber network) as terminators on the input side.)