Commissioning and testing of the readout system for OSIRIS

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Tim Kuhlbusch

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Abstract

The 20-ton OSIRIS detector is designed to measure radiopurity of the liquid scintillator (LS) for JUNO. This is especially necessary to commission the liquid handling and purification systems. OSIRIS will monitor the scintillator contamination during the filling to ensure that JUNO reaches the design sensitivity. After filling of the central detector OSIRIS can be used for monitoring and detailed studies of scintillator characteristics.

The detector consists of an outer steel tank containing an ultrapure water pool and an inner vessel for the LS. Large Photomultiplier Tubes (PMTs) are positioned around the inner vessel to detect the light emitted from the LS.

OSIRIS will utilize the novel iPMT system where each PMT is equipped with its own “intelligent” readout electronics. All hardware required to digitize the signals is placed in a metal shell at the rear end of the PMTs. This thesis presents and discusses the commissioning and testing of the first prototype of this readout system. A custom ASIC called VULCAN, an analog to digital converter designed for digitizing PMT signals, is used. The digitized signals are buffered and processed by an FPGA and send to the outside via Ethernet by an embedded Linux system.

The VULCAN chip contains three identical 8-bit receivers which can be configured individually. These receivers are used in parallel to digitize the same input signal. VULCAN automatically chooses the receiver with the highest resolution that is not saturated. The receivers are set to different gains to achieve a high dynamic range. After the configuration of each individual receiver a calibration scheme is developed and tested to use all three receivers together. Thus the optimal receiver for the current amplitude can be used.

Due to the low input impedance of the digitizer a matching circuit is needed to get a good transmission of the pulses from the significantly higher impedance PMT. Different wideband matching circuits were designed and tested to reduce reflection and increase signal amplitude.

Once configured the setup is able to reliably digitize PMT pulses resulting from single up to hundreds of photons. Using a threshold trigger a major fraction of detected photons can be selected while keeping the rate of false-positives low.
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1 OSIRIS detector

1.1 JUNO
The 20 kt liquid scintillator Central Detector (CD) of JUNO is currently under construction. Once finished it will be used to study neutrinos and neutrino oscillations. A main focus is the determination of the mass hierarchy using neutrinos from nuclear reactors at a distance of ca. 53 km [4].

But as a large and sensitive neutrino detector it will also be able to measure neutrinos from different sources like geo-neutrinos, solar neutrinos and neutrinos from supernovae. Very low background rates are required to study these low rate neutrino sources.

1.2 OSIRIS
Cleaning contaminated liquid scintillator (LS) once it is filled into the JUNO CD will be difficult and time-consuming. It is therefore necessary to measure the radiopurity before it is filled into the CD. These measurements are done by the Online Scintillator Internal Radioactivity Investigation System (OSIRIS). OSIRIS will also be used to commission the LS purification plant.

![Figure 1: Sketch of the OSIRIS detector. The greyish outer cylinder is the outer steel tank [15]. Inside is the frame holding the iPMTs. The small cylinder in the center is the acrylic inner vessel.](image)

OSIRIS will be a 20 t liquid scintillator detector that measures the radiopurity directly with the scintillator that is being tested. The detector is cylindrical with a height and diameter of 9 m. An inner acrylic vessel separates the outer ultrapure water pool from the scintillator. The inner vessel is cylindrical with a height and diameter of 3 m. 80 Photomultiplier Tubes (PMTs) are mounted to a frame in the water pool. Most of the PMTs face towards the transparent inner vessel to detect...
the scintillation light. Twelve PMTs are used to instrument the waterpool. They act as a water Cherenkov detector to detect radiation entering the detector from the outside.

OSIRIS will be situated in a cavern close to the JUNO central detector to reduce background radiation and to minimize the risk of a radon leak in the pipes between the two detectors.

Heaters at the upper end of the outer OSIRIS vessel are used to establish a temperature gradient inside the detector. This reduces the mixing with already present liquid when new scintillator is filled in. Two modes of operation are planned: One is for measuring one batch of LS that is filled into the detector and remains there; the thermal gradient is then used to swap batches without draining the detector. The other is to have a low constant flow of LS through the detector with a minimized mixing of added and present liquid [7].

Large 20” Photomultiplier Tubes (PMTs) will be used to detect the light emitted from the LS. The classical solution is to connect all PMTs with long coaxial cables to the digitizers situated outside. Instead OSIRIS will be instrumented with the novel intelligent PMTs (iPMTs) where the readout electronics for each PMT channel are directly mounted to the PMTs. One Ethernet connection per PMT is used to send the digitized PMT pulses to the DAQ servers outside the tank. Directly connecting the digitizer to the PMT removes potential issues like cable-losses and crosstalk between cables.

1.3 OSIRIS Physics

The main task for OSIRIS is to measure the $^{238}$U and $^{232}$Th concentrations of the scintillator. Due to the low contamination levels a very good discrimination from background events is required. Both isotopes have a very short lived Polonium isotope in their decay chain. The coincidence with a prior Bismuth decay is used to select events. The monoenergetic $\alpha$ decay of the Polonium isotopes combined with the Bi-Po coincidence can be used to measure the contamination with all isotopes that decay into $^{212}$Bi or $^{214}$Bi [3]. This allows to measure contaminations down to $10^{-9}$ Bq/kg for $^{214}$Bi-$^{214}$Po decays and $10^{-10}$ Bq/kg for $^{212}$Bi-$^{212}$Po [7]. Different detection channels are used to measure the concentrations of e.g. $^{14}$C.

![Figure 2: Bi-Po decays in the $^{238}$U and $^{232}$Th decay chains [12].](image)
Figure 3: Predicted OSIRIS sensitivity. The values shown are the 90% upper confidence limits. [3].
2 iPM T readout system

2.1 Concept

2.1.1 PMT & Potting

Photomultiplier Tubes are vacuum tubes that can detect single photons. The inside of the light detecting surface is coated with a thin metal film called the Photocathode. Photons can release electrons out of the Photocathode via the photoelectric effect. These electrons are then accelerated by electric fields towards electrodes that are called dynodes. When the dynodes are hit by an electron more electrons are released via secondary emission and accelerated towards the next dynode. At the last dynode a current pulse can be measured for every detected photon.

OSIRIS will use R12860HQE 20” PMTs by Hamamatsu. The same model will be used in the central detector. These are large dynode based PMTs with a high photon detection efficiency.

The electronics at the back of the PMT consist of a stack of PCBs that are directly soldered to the PMT pins. To protect the electronics from the surrounding water a stainless steel shell is attached to the PMT. A PMMA piece is glued to the PMT as an adapter. The shell is filled with oil to conduct the heat of the electronics to the surrounding water. Using oil also helps to isolate the electronic components which is especially useful considering the high voltages needed for the PMT.

Figure 4: Sketch of a PMT with iPM T electronics in a steel shell. Around the PMT is the holding structure that is used to attach the PMT to the detector scaffolding (by D. Jahn).
2.1.2 Base & HV

The base is the first in the stack of boards directly connected to the PMT. It contains the HV module and the voltage divider that provides the different voltages for each dynode of the PMT. The used PMTs requires about 2 kV for normal operation. A RS485 connection is used to control and monitor the HV module. This data connection, the power and the analog output signal from the PMT are connected to the above boards.

2.1.3 Power and cables

For simplicity there is only one cable per iPMT that connects them to the outside. These cables are CAT 5 cables. Two of the four twisted pairs in these cables carry the 100 MBit Ethernet link. Power over Ethernet (PoE) is transferred to convey the required power. Another twisted pair is used for the clock. The clock is roughly synchronized via the Network Time Protocol over Ethernet down to second precision. The sub-second synchronization is done with a synchronization signal on the clock wires.

![Figure 5: Overview of the iPMT electronics.](image)

2.2 Surface Board

Due to the non-standard use of the CAT 5 cable an adapter board called the Surface Board is necessary. This board feeds the signal from the wire pairs used for the Ethernet directly to a PoE switch that provides power and the network connection. For the 80 iPMTs of OSIRIS two Surface Boards will be used. Each Surface Board can either provide the reference clock for the iPMTs or act as a fan out for an external clock source. This external clock source can then be another surface board to allow larger iPMT counts. The two Surface boards used in OSIRIS will be synchronized this way.
2.2.1 Readout Board

The central electronic system is the readout board (ROB) where the PMT signals are digitized. These waveforms are then sent via the network connection to DAQ servers on top of the detector. To achieve the necessary timing resolution the digitizer and sections of the FPGA run on the clock received from the surface board.

**VULCAN** A custom chip called VULCAN is used to convert the analog PMT signal into a digital signal (analog to digital converter, ADC) that can be send to the Data Acquisition (DAQ). This ADC samples the input signal at 500 MHz with three separate 8-bit receivers (RX). The three 8-bit receivers are configured to different gain settings to achieve the necessary dynamic range at a sufficiently high resolution. Thus the resolution of this setup is highest for the receiver with the highest gain (HG). To precisely distinguish between random noise and signals this receiver has a resolution that is smaller than the average signal of one photo-electron. Larger signals require the receiver that is configured for a lower gain (mid-gain, MG). In this range the resolution is still high enough to count photons. For the largest signals the last receiver (low-gain, LG) is used. The spread of the PMT signal is proportional to the square root of the photon count. A higher resolution in the MG and LG receivers would therefore not significantly increase the photon count resolution.

![Figure 6: Sketch demonstrating the range of the three VULCAN receivers [9.]](image)

The VULCAN ADC is configured to continuously output the samples from one of the three receivers. Which receiver is chosen is decided with a threshold. If the signal in one of the receivers is larger than a threshold the internal logic will switch to the next receiver with lower gain. Once the signal falls below another, lower, threshold the output is switched back to the higher gain receiver.

**FPGA & PS** To process the output of the VULCAN chip a Xilinx Zynq hybrid FPGA is used. This chip contains an FPGA (programmable logic, PL) and a dual core ARM processor (programmable system, PS). The FPGA continuously monitors the signal from the ADC to derive a trigger decision. A simple threshold trigger and more complex triggers can be implemented in the programmable logic. To store
the samples a set of ring-buffers is used. A trigger causes a delayed switch to the next ring-buffer. The delay can be configured to adjust the offset of the trigger in the waveform. The values from this ring-buffer are packaged and made available to a program running on the PS. From there the data is sent via TCP/IP over the Ethernet connection.

The FPGA is also powerful enough to analyze the pulses. It is planned to implement an algorithm that can determine the timing and charge of small single PMT pulses. As these small single pulses make up the majority of the dark noise this drastically reduces network bandwidth. It also decreases the workload for the DAQ servers as they don’t have to do the online pulse reconstruction anymore.

To prevent collisions every ROB needs an individual MAC address. An EEPROM with a pre-programmed unique MAC address is read by the first stage bootloader and forwarded to the second stage bootloader. The second bootloader sets the MAC address and starts up the networking. Once it has received an IP address via DHCP it downloads the PS firmware and PL hardware files to load them.

After this is done the Linux on the PS system starts up. The PL system is loaded but large portions of the logic run on clocks driven by VULCAN. So the PL will not run correctly until the VULCAN clock is configured.

### 2.2.2 Slow Control and Configuration Unit

Once the iPMTs are potted the Ethernet cable is the only connection to the outside. The Slow Control and Configuration Unit (SCCU) is a board with an ARM microcontroller designed to alleviate this problem. This microcontroller bridges the hardware interfaces of the iPMT to Ethernet.

One example for this is that the SCCU provides two JTAG interfaces. JTAG is an electrical interface built into chips to simplify hardware and software debugging. A master device can access certain registers in all connected client devices [11]. Depending on the chip this can be used to program or configure devices and to gain valuable information to solve problems during development.

In the iPMT design JTAG is used to configure the VULCAN chip and to flash new firmware and hardware onto the Zynq. The SCCU provides its JTAG interfaces via a Xilinx Virtual Cable (XVC) interface [22]. XVC is a very simple network protocol that is especially useful as it allows using the Xilinx tools to program and debug the Zynq system after potting.

The SCCU also provides I2C and Serial interfaces. A telnet interface is provided to configure the SCCU. This allows for example adjusting the serial port baud rate.

The SCCU also provides a network switch. This is necessary to connect both the ROB and the SCCU via a single cable to the outside. A similar MAC EEPROM to the one used in the ROB ensures that every SCCU has a unique address.

### 2.2.3 Network

To simplify debugging of network related problems every device connected to the readout network will receive a static IP address and a local domain name. Especially
the local domain name will make configuration scripts more readable as e.g. ”rob23”
is much simpler to interpret than ”10.0.1.123”.

This requires some local servers inside the iPMT network. A DNS server that
provides the service of resolving local domain names to IP addresses and a DHCP
server that assigns the IP addresses to the individual devices. To do this every device
in this network needs a unique MAC address. The DHCP server assigns every ROB
and every SCCU a fixed IP address based on their MAC address and distributes the
address of the DNS server.

2.2.4 DAQ

Client The program that runs on the PS, the iPMT-Client, receives complete
network packages from the PL and sends them via TCP/IP to the server. A terminal
interface is provided to configure and monitor the readout system. Commands on
this interface can for example configure the trigger in the PL and monitor the buffer
status in the PS. It is also possible to access waveforms via this interface. The control
interface is also exposed via telnet to allow automated measurements controlled by
software run on a lab computer. Scripts can thus quickly change configuration
settings and collect waveforms for testing.

A simple config file containing a sequence of commands can be provided to this
software to quickly recall predefined configurations.

Development server For testing a simple software is used to collect the wave-
forms sent by the iPMT-Client. This program unpacks the waveforms and saves
them to a root file. Later these files can then be read by analysis scripts. A similar
telnet interface to the one of the client software is built into this server to allow
automated switching of the output file.

Event-building In OSIRIS there will be 80 iPMTs producing dark-noise data at
a rate of ca. 20 kHz per PMT. Assuming 200 bytes of data per waveform this adds
up to roughly 320 MB/s for the whole detector.

<table>
<thead>
<tr>
<th>Data</th>
<th>Size [byte]</th>
<th>Comment</th>
</tr>
</thead>
<tbody>
<tr>
<td>Header</td>
<td>8</td>
<td>Package size</td>
</tr>
<tr>
<td>Subheader</td>
<td>24</td>
<td>Timestamp, trigger source etc.</td>
</tr>
<tr>
<td>Waveform</td>
<td>160</td>
<td>3 samples → 4 byte</td>
</tr>
<tr>
<td>Total</td>
<td>192</td>
<td></td>
</tr>
</tbody>
</table>

Table 1: Overview of the iPMT waveform packet sizes.

Fortunately most of this data can be discarded by a simple coincidence trigger.
To do this the DAQ will need to buffer the incoming data and search for coincidences
once all required data is available. The necessary size of this buffer depends on the
maximum delay between receiving data for the same timestamp. Differences in the
Ethernet cable length, variance of processing time in the iPMT and network delays increase the required size of this temporary storage.

2.3 Prototype

The focus of this thesis are the currently finished prototype boards. As can be seen in Figure 7 this includes the ROB and SCCU. All components are mounted on a board to simplify moving the setup. A fan is used to circulate the air and increase cooling. It is not strictly necessary but reduces stress on the electronic components when operated in a hot environment.

Figure 7: The SCCU and ROB prototypes.

During all measurements described in this thesis the setup was operated inside a container to shield it against electromagnetic interference and light. The light shielding helps when the PMT is used. By reconnecting a cable either the PMT or a signal generator can be used as a signal source.

2.4 Slowcontrol

2.4.1 EPICS

EPICS is a distributed process control system. It will be used to monitor the status of OSIRIS and to control most aspects of the iPMTs.

This simplifies interfacing with e.g. the liquid handling as it is managed with EPICS, too. EPICS provides a stable and well tested control system. A large ecosystem of EPICS compatible software already provides most of the necessary tools. For example there are already logging solutions like the EPICS Archiver Appliance [14]. Simple user interfaces can be created with little work with caQtDM
There are also solutions readily available for other aspects like alarm handling.

Figure 8: Example of temperature data recorded by the EPICS Archiver. \( t_{\text{up}} \) and \( t_{\text{dwn}} \) are readings from temperature sensors on the upper and lower side of the SCCU. The \( t_{\text{rob}} \) values stem from sensors on the ROB.

### 2.4.2 IOCs

EPICS is built around the concept of process variables (PVs). Each variable has a unique string identifier through which it can be accessed by any computer in the same network. Process variables can have different record types (effectively data types) like strings or floating point numbers depending on what they represent.

These values are provided by programs that are called Input Output Controllers (IOCs). An IOC traditionally runs on a computer that connects to a piece of equipment. In the case of the iPMTs this connection is the network connecting all iPMTs. Any EPICS software can access variables provided by the different IOCs and act upon them [13].

The distributed nature of this concept means that an EPICS system is really expandable. New variables can be added by just running another IOC on any computer on the same network allowing a huge scalability. This also makes the control system more resilient. If one IOC becomes unavailable all other systems still continue to work.

The IOCs could potentially be run on the Linux system of the ROB. This does however increase use resources on the ROB and complicates the update procedure for these IOCs as they need to be cross-compiled for the ARM platform. It is also simpler to monitor and manage IOCs running on a dedicated server than distributed across all iPMTs.

**iPMT Client** The software running on the PS provides a telnet interface. An IOC running on a server outside the detector can connect to this interface and monitor
the status of this software. This provides for example access to the status of the
buffers in the PS and PL.

**I2C**  
I2C is a very common serial bus system. The SCCU acts as an I2C master
to e.g. read data from I2C sensors.

One example for this are TMP100 I2C temperature sensors that are placed on
the electronics boards [21]. These allow monitoring the temperature in the housing
at different places which is important to predict aging of the electronics and to
identify problems. Another example are voltage and current readings provided by
INA226 used to monitor the different voltage rails inside the iPMTs [20].

**Serial**  
The serial ports of the SCCU can be interfaced via a simple telnet. They
are used to access the Linux console of the PS and to configure the HV module. An
IOC allows adjusting and monitoring the HV over EPICS.

**OW**  
A One-Wire (OW) interface is provided via a DS2480B Serial to OW con-
verter chip [16]. OW temperature sensors can be deployed very easily with relatively
long cables. This allows simple monitoring of the electronics outside the detector
with an extra SCCU.
3 The VULCAN ASIC

3.1 Analog frontend

A PMT pulse is produced by the charge that is deposited on the last dynode. To estimate the photon count of overlapping pulses the charge is determined as the integral over the measured current. The actual digitizing circuitry in VULCAN however requires a voltage signal. Thus the first amplifier stage in VULCAN is used to convert the current signal to a voltage signal. This amplifier is a Trans Impedance Amplifier (TIA). It has a low impedance input acting as a current sink and a low impedance output acting as a voltage source.

The next amplifier in the chain is used to switch between two different amplifications with a factor of 10 difference in the gain. The combination of this amplifier and the TIA can be interpreted as a TIA with an effective trans-impedance of 72 Ω or 720 Ω.

![Figure 9: Overview of the internal structure of VULCAN.](image)

The third amplifier is the Programmable Gain Amplifier (PGA). This amplifier has an array of switchable feedback resistors allowing a finer control of the gain.

The bias voltage for the amplifiers can be controlled. This voltage has to be chosen in such a way that the range of the digitizer is within the working range of the amplifier circuitry. Otherwise an amplifier would reach saturation before the full range of the ADC can be utilized.

To achieve a lower amplification in the LG receiver the first two amplifiers are bypassed. As the input signal is directly fed into the PGA this is a high impedance input that can measure the voltage across the other lower impedance receivers.
3.2 Digitization

4 Banks of 64 comparators are used to digitize the signal. Each comparator outputs whether the input is above its threshold. The threshold is increased in equidistant steps to create a linear scale. Digitizers constructed in this manner are referred to as flash ADCs. Their raw kind of output format is called thermometer code.

To achieve the best performance the 8-bit ADCs of VULCAN are split up into the four 6-bit ADCs. Limiting amplifiers with separately controlled input biasing are used to create the offset input signals to each 6-bit ADC.

3.3 Digital logic

The digitized signals of all three receivers are fed into the Control Unit (CU). For the iPMTs the pass-through mode of the CU is used. This mode bypasses the trigger and buffering capabilities of the chip. Instead the FPGA provides these functions making the whole system more flexible. It also simplifies the development process as the FPGA is much simpler to understand and debug than the internals of VULCAN.

The input stage of the CU allows to configure the input signal. All bits can be inverted and the bit-order reversed. In all following measurements this section is configured so that the digitized PMT pulses are positive.

The input to the CU are the states of all comparators. One of two ways to convert these values to a number can be chosen: Counting the ones or counting how many continuous ones are there from the bottom up. The second option is called greycoding (Figure 10). As the count-ones method is more resistant against 'bubbles' in the thermometer code, it is the better choice. Using the greycoding is useful during development and testing because it does not hide errors and gives more insight into what is actually happening.

![Figure 10: Illustration of the thermometer coding and how the greycoding and count-ones decoders work.](image)

Depending on the input signal the CU chooses the optimal ADC. If the input signal reaches a configurable threshold in the current receiver it will switch to the next. When the signal falls below a second lower threshold in a higher gain receiver the CU will switch back to it.

The CU also contains a block that can regulate the baseline of each receiver to a target value. A biasing voltage of the circuit that splits off the signal to each ADC
is controlled by this baseline regulation (BLR) [8]. This reduces the impact of low frequency noise.
4 Single receiver configuration

4.1 Impedance matching & connections

The MG and HG receiver have a low input impedance of about 10 Ω. All three receivers are connected in parallel, so the high impedance LG receiver measures the voltage across the MG and HG receivers. The effective impedance is thus about 5 Ω. Connecting a 50 Ω signal source directly to the low impedance VULCAN input would result in most of the signal being reflected.

Reflections are bad in two ways: On the one hand the signal reaching VULCAN is reduced. On the other hand the reflected signal can be reflected back by the PMT leading to distorted pulses or delayed second pulses. The time at which the reflection reaches the digitizer depends on the delay introduced by the signal line in between. In the final iPMTs this delay will be lower than 5 ns as the digitizer is placed very close to the PMT. During development however a more bulky setup is used that is not mounted directly to the PMT. Thus at least 0.5 m long cables are required resulting in a reflection delay larger than 5 ns.

The reflection coefficient \( \Gamma \) for an impedance mismatch depends on the output impedance \( Z_0 \) of the signal source and on the input impedance \( Z_{load} \) of the receiver [10, p. 34].

\[
\Gamma = \frac{U_r}{U_i} = \frac{Z_{load} - Z_0}{Z_{load} + Z_0}
\]

(1)

During measurements either a signal generator (SG) or a PMT are used as signal sources. Both have \( Z_0 = 50 \Omega \) outputs. Considering the VULCAN impedance \( Z_{load} = Z_{ROB} \approx 5 \Omega \) yields a reflection coefficient \( \Gamma \approx -0.8 \) if no matching is used.

Measuring the impedance matching  The impedances are frequency dependent. It is therefore necessary to measure the quality of the matching at the target frequencies. The achieved impedance of a certain matching circuit can be measured with a network analyzer (VNA).

Due to the sample rate of 500 MHz the maximum usable frequency given by the Nyquist limit is 250 MHz. As the PMT pulses are usually shorter than 50 ns frequencies below 20 MHz will not significantly affect the pulse shape. Blocking them can however lead to an undershoot after the pulse.

The results from the VNA can be difficult to interpret as it is not totally obvious what a high reflection coefficient at certain frequencies means for the relatively wideband PMT pulses.

A more direct approach of measuring a reflected pulse is therefore a good option. Pulses with different spectral components are reflected differently as the impedance might not be totally flat. Therefore when using a signal generator similar pulses to those of the PMT must be used to get meaningful results. An estimation of the impedance mismatch can be derived from the amplitude ratio \( \Gamma \) of an initial and reflected pulse using (2).

\[
\frac{Z_{load}}{Z_0} = \frac{1 + \Gamma}{1 - \Gamma}
\]

(2)
To create as realistic pulses as possible the shortest possible setting of the SG was used; the resulting pulses have a rise and fall time of about 20 ns. The reflected pulse must be delayed so that it can be distinguished from the incident signal. A cable long enough to delay the pulse by more than the full pulse duration was used. Both pulses are recorded by a high impedance oscilloscope connected at the same end of the long cable as the signal source. The result of such a measurement is shown in the left half of Figure 11.

According to (1) a short ($Z_{\text{load}} = 0$) will result in total negative reflection while open ($Z_{\text{load}} = \infty$) results in total positive reflection.

$$\lim_{Z_{\text{load}} \to 0} \Gamma = -1 \quad \text{and} \quad \lim_{Z_{\text{load}} \to \infty} \Gamma = 1$$

(3)

Reflections on an open and short termination at the end of the long cable are used to compensate for the cable attenuation. As the cable attenuation is always the same we can simply compare the amplitude of the reflection at the ROB with the amplitude of the total reflection. This should also compensate for any change in the pulse shape due to frequency dependent cable attenuation. Doing this results in the estimated impedance of ca. 4 Ω, which matches will with the estimate from the prior section.

The right half of Figure 11 shows a PMT pulse being reflected at the unmatched ROB input and the PMT. A long cable is used to delay the reflection. This method uses real PMT pulses and thus gives the most representative results but is highly dependent on the reflection at the PMT. Without knowing the reflection factor of the PMT it is not possible to estimate the impedance mismatch. But anyway the most important result of this measurement is the reflected amplitude. We can see...
that using no matching results in a negative reflection of effectively 40% in the final digitized signal.

**Required signal level**  A central requirement for the system is to reliably discriminate between random noise and PMT pulses. This requires a certain minimal signal level compared to the noise level.

**Using no matching**  The most simple option is to use no matching. However as shown in Figure 11 this results in more than 80% of the incoming signal being reflected. When a short connection is used the already small signal is further reduced by the first reflection with an effective amplitude of ca. −40% in the ADC waveform. Every second reflection will be positive and thus increase the signal amplitude, but every reflection is also just 40% of the previous one.

Using a short 50 Ω cable and no matching did not yield a high enough signal amplitude to reliably trigger upon single PE events. This might however change when the electronics are directly mounted to the PMT.

**Resistive matching**  The most simple matching circuit is a series resistor that increases the total input impedance.

\[
Z_{tot} = Z_{ROB} + R \Rightarrow R_{opt} = Z_{tot} - Z_{ROB}
\]

In this case the total impedance is simply the sum of both impedances. The disadvantage of this method is that the pulse amplitude is reduced due to resistive losses. But on the positive side this results in a very good wideband impedance match as the resistor has a flat impedance curve in the required frequency range and is the larger fraction of the total impedance.

The output coupling capacitor in the PMT base needs a bleed resistor to discharge it once the HV is switched on or off. Unfortunately the VULCAN input biasing cannot provide enough current for this resistor. To solve this problem an additional DC blocking capacitor is used as can be seen in the upper half of Figure 12. The capacitor must be chosen large enough to pass all necessary low frequencies.
4.2 Analog frontend

There are a lot of configuration options for the analog frontends of the VULCAN chip. Some of them like the multiplexers that are used to route the signals have relatively obvious correct values. Other values like the input biasing for the amplifier stages are not that simple and need to be tuned to achieve the maximum performance.

4.2.1 Biasing voltages

The input biasing voltages to all used amplifiers have to be set within their working range. If one amplifier reaches saturation only a part of the range of the ADC will be usable. Due to production variation these settings can vary from chip to chip. The input biasing of each amplifier changes the output offset. Thus changing any input biasing value moves the baseline of the ADC.

To find the optimal settings the biasing voltages of the first amplifier and the PGA are scanned. With no signal connected the baseline dependence on the biasing voltages can be seen (Figure 13). \texttt{tia\_ctrl} (tial1) controls the input biasing of the PGA. \texttt{tia\_ctrl2} (tia2) does the same for the initial amplifier. The linear section in the middle indicates the usable range while the horizontal sections are where the second amplifier reached saturation.

The biasing settings also have a backward effect. Changing both values also influences the receiver input biasing. If all receivers are directly connected then changing the settings of one will influence the others. This can be seen in Figure 14.
Figure 13: Plot showing the change of the HG receiver baseline when adjusting the two main biasing voltages.

where the baseline dependency on other receivers biasing values is depicted. If the biasing voltages are kept constant this is not an issue.

Putting a capacitor in front of each receiver could block the biasing voltages so that the values for each receiver can be chosen independently. This would simplify choosing the biasing voltages but can lead to issues with the matching of the different ranges.

As can be seen in Figure 14 changing the MG biasing voltages has a significant effect on the other receivers. During alignment it therefore makes sense to start with the configuration of the MG receiver before that of the HG.

During most measurements a baseline between 10 and 15 was chosen. This is a compromise to have enough distance to the lower end of the receivers range for baseline fluctuations and not to clip undershoots. Increasing the baseline reduces the positive range of the ADC that is used for the actual signals.

The biasing voltages change the operating point of the amplifiers. t1a1 values between 50 and 150 are usable because no ADC is in saturation yet. As can be seen in Figure 15 the differences are below 20% within the usable range.

SPE charge spectra for different biasing voltages are shown in Figure 51. There is overall a wide range of working bias settings. As long as no amplifier reaches saturation within the ADC range a decent PTV is visible. Small variations are visible which is consistent to the measurements with the signal generator.
Figure 14: Influence of other receivers biasing voltage settings.

Figure 15: Amplitude of the same square pulse for different biasing voltage.
4.2.2 PGA setting

Eight parallel switchable resistors make up the feedback branch for the PGA. They can be used to adjust the total gain of the amplifier chain. This allows adjusting the relative gain of the ranges.

\[ g \propto \sum_i \frac{1}{b_i r_i} \]  
\[ (4) \]

The effective resistor values can be found by comparing the gain when the bit of one resistor is turned on \( (g^*) \) or off \( (g) \).

\[ g = \frac{1}{r_{\text{other}}} = r_{\text{other}} \quad \text{and} \quad g^* = \frac{1}{r_{\text{other}}} + \frac{1}{r_{\text{toggled}}} \Rightarrow r_{\text{toggled}} = \frac{1}{g^* - \frac{1}{g}} \]  
\[ (5) \]

A histogram of all estimates calculated this way is shown in Figure 49. From this \( r_i = 120 + 25 \cdot i \) was guessed and used to generate the leftmost plots in Figures 17 and 18.

The middle figure shows the linear relation of the baseline and the gain. In the rightmost figure the standard deviation of the baselines for different gain settings is plotted. The strong correlation suggests that changing the PGA gain does not significantly increase the signal to noise ratio for the HG receiver. As the PGA is not the first amplifier in the chain it will amplify the noise from all prior amplifiers. For the HG receiver the relative noise level from the frontend is higher than for the later
sections. So increasing the PGA gain will not significantly increase the separation between random noise and signals.

![Figure 17: HG PGA behaviour for all non-oscillating gain settings. The orange cross indicates the gain setting used in all other measurements (0x01). One data point with the highest gain saturated the receiver and thus has a wrong gain.](image)

Figure 17: HG PGA behaviour for all non-oscillating gain settings. The orange cross indicates the gain setting used in all other measurements (0x01). One data point with the highest gain saturated the receiver and thus has a wrong gain.

![Figure 18: MG PGA behaviour for all non-oscillating gain settings. The orange cross indicates the gain setting used in all other measurements (0x01).](image)

Figure 18: MG PGA behaviour for all non-oscillating gain settings. The orange cross indicates the gain setting used in all other measurements (0x01).

The PGA can be used to adjust the range of the MG and LG receiver. Adjusting the PGA gain can increase the gain by 2.2 or reduce it by 0.22 compared to the gain used in all other measurements.

### 4.3 Digitization

The digitizer of each receiver consists of four identical 6-bit ADCs. Each 6-bit ADC has an additional amplifier with configurable offset. This effectively allows to individually adjust the offset of the range of each 6-bit ADC. These offsets (`vref_buffer`) have to be chosen in such a way that the next ADC range starts at the end of the previous one.

The first step to align them is to find reasonable starting values by changing the settings manually and observing the waveforms. It is important that all values...
are far enough from the upper and lower limits to prevent saturation. A saw-tooth pulse is used as any misalignment is clearly visible on the rising edge. As shown in Figure 19 using count-ones coding hides misalignment, especially if the overlap of the ranges is small. A measurement of the effect shown in the left image can be found in Figure 20 in the appendix. The jitter and noise of the signal generator combined with count-ones makes it even more difficult to measure the misalignment. Therefore greycoding is used for these measurements.

To prevent an overlap with the range that is currently being aligned all unaligned \texttt{vref\_buffer} values are adjusted to create large gaps. An overlap with the ADC that is currently being aligned would reduce the fit range and thus worsen the result.

The 6-bit ADCs must be aligned one after the other. Either the highest or lowest ADC is kept at the initial value from the previous manual alignment. Then the next ADC is adjusted so that it is aligned to the prior ADC.

Waveforms are recorded for offset values above and below the initial guess taken from manual alignment. Average waveforms from this data are shown in Figure 21. The horizontal sections result from a gap between the ADCs and a vertical jump is
created if they overlap.

A very simple way to test the matching would be to fit a line to the whole waveform, spanning both ADC ranges. The $\chi^2$ of the fit could then be used as an indicator. But this will only give a good result if the slope is the same in both ADC ranges. To be more resistant against such problems both ADC ranges can be evaluated separately. Then the intercept point of both lines is used. Figure 22 shows the calculated intercept point for different $v_{\text{ref\_buffer}}$ offset values. The target value is subtracted from the y-axis values so the best value is at the x-axis intercept point.

This procedure is repeated to match all four ADCs of all three receivers.

![Figure 21: Average waveforms for different $v_{\text{ref\_buffer}}$ values.](image)
5 Range Matching

5.1 The three ranges

The raw data, because it is recorded from the VULCAN ASIC, cannot directly be used for physics analysis. Range switches appear as jumps in the raw data as can be seen in Figure 23. A calibration scheme is required to make these waveforms continuous and usable for further analysis.

To prevent unnecessary switching the threshold for using the higher gain has to be higher than the threshold for switching back to the lower gain. But increasing the hysteresis means that the lower gain setting is used longer resulting in a lower resolution.

The hysteresis of the range switching was 10 LSB during the following measurements. As this value is large compared to the standard deviation $\sigma_{HG} = 2$ LSB of the digitized values in the HG receiver it reliably prevents unnecessary range switching.

A mapping of the ADC values from all three ranges to current can be used to reconstruct the input signal. To get this data a controllable current source is necessary. One way to do this is to use a voltage source with a series resistor. Voltage sources have a very low internal resistance. If the internal resistance of a source is large compared with the load it behaves as a current source.

A 500 $\Omega$ resistor was used in the following to use a signal generator as a current source. It was chosen as a compromise as it is significantly larger than the ROB input impedance and low enough to achieve a decent current range with the signal generator. As can be seen in Figure 24 a 50 $\Omega$ resistor is used to terminate the input signal and prevent reflections.

Figure 25 shows the correlation between pulse amplitudes and charges. The charge is calculated from the raw ADC values without any corrections. This leads...
to the non-linear behaviour in the lower part of the MG and LG receiver. In these regions significant parts of the pulse are within the lower range leading to a different charge value. The gaps between the ranges are a result of the non-zero baselines and the fact that the higher ranges are only used when the signal amplitude is high enough to saturate all lower ranges. Due to the different gains of the receivers the slope is different in each receiver.
5.2 Determination of the relative gain

It is necessary to get the output from all three ranges for the same input signals. There are different ways to achieve this. One is to read out the VULCAN status register via JTAG. The updating of the status register can be stopped. Then values from all three ADCs at the same point in time can be read out. A disadvantage of this method is that it is relatively slow as the data rates of the JTAG interface are not very fast. If a few status register entries are read the readout rates are in the order of 10 Hz. This method works best with a constant current source at the input.

A faster approach is to record saw-tooth pulses that are large enough to cover more than one range. Then the relative gain can be deduced by finding the relative gain at which waveforms of two ranges fit together best.

This works, but a more simple solution that provides more information is to use square pulses and compare their amplitude when recorded by the different receivers. The setup for this is similar to that used in Section 4.2.2 but the amplitude of the pulses is changed. By comparing the amplitude of similar pulses in the different receivers the relative gain can be deduced.

**Gain matching with square pulses** All receivers are connected to the signal generator that is set to deliver square pulses. The timing is chosen so that the first half of each waveform is just a baseline and the second half contains the pulse. A script switches the receiver ordering so that the pulse can be digitized with each receiver. 1000 waveforms are recorded each. The pulse height over the baseline is

---

Figure 25: Histogram showing the correlation between pulse amplitude and charge for raw data of PMT pulses selected with a threshold trigger. Data from the HG receiver is below 256 LSB. The LG receiver produces values above 512 LSB and the MG receiver data is in between. As a result of the non-zero baselines and the range-switching gaps appear between the receiver ranges.
determined for every waveform. When the amplitude is adjusted the measurements are repeated.

Results from this analysis are shown in Figure 26. Each data point represents one measurement at a different amplitude. The error on the input signal is given by the datasheet of the signal generator [19].

The HG receiver is not entirely linear, but a second order correction term works really well. A parabola is not bijective which could be a problem as we need a unique mapping from ADC values to the actual input current. This is not a problem as the resulting parabola is invertible on the full ADC range.

This non-linearity is not a problem as it can be compensated for. It might be possible to reduce or eliminate this nonlinearity by adjusting the VULCAN configuration.

A better linearity is shown by the MG receiver. Thus a simple linear model was used. The fit range for the calibration starts at 10 ADC counts as there is a slight difference compared to the rest of the data. Values in this range are not used during normal operation because are within the HG range. So it has no effect during normal operation that the calibration is bad for these points.

For the low-gain a similar procedure is used. Unfortunately the used signal generator is limited to amplitudes of 5 V. This is not sufficient to saturate the MG range when the 500Ω adapter is used. Thus the calibration could not be done for the whole range of the MG and LG receivers.

A triangular pulse reconstructed using this calibration is depicted in Figure 27. The black reference waveform was recorded by an oscilloscope. To measure the baseline of all three ranges ADC samples were recorded via JTAG during the measurement. All values that are more then 4 sigma above the median were not included in the mean BL calculation to prevent the few samples that are recorded during a pulse from changing the baseline estimate.
Figure 26: Relations of input pulse height to average pulse height measured by the three receivers. The pulse height over baseline is determined as average of the plateau values. Each point represents the average of 1000 measurements.

Data from all three receivers is first adjusted based on the average baselines calculated from the JTAG data. Then the inverted calibration functions are applied to calculate the currents.

\[ y = mx + b \Rightarrow x = \frac{y - b}{m} \]
\[ y = mx + sx^2 \Rightarrow x = \sqrt{\frac{y}{s} + \frac{m^2}{4s^2} - \frac{m}{2s}} \]

The 500 Ω adapter is slightly mismatched and creates slight reflections. When comparing both waveforms one has to take into account that the reference curve is a superposition of the input signal and the delayed reflected signal. Due to the limited input amplitude a lower threshold for switching between the MG and LG range was chosen for this measurement. Overall both curves align well and there are no visible jumps when the range is switched.

It is necessary to know the baselines values to use this calibration. This means either logging baseline samples via JTAG or regulating all three baselines to known values.

Figure 27: Left: Reconstructed signal of a triangular pulse. The reference is the same pulse recorded with an oscilloscope. The timing is adjusted by hand to roughly align. Right: A large reconstructed PMT pulse.

Figure 28 showcases the range matching for different input signal slopes at the switching point between the HG an MG receiver. A line is fitted to the linear section of the rising edge of the input signal. For steep curves the count of usable data points is very low due to the fixed sample rate. Considering the limited sample rate and linearity of the signal generator there is no significant mismatch at the switching point.
5.3 Absolute current calibration

These measurements can also be used to determine the current per LSB. As the $R = (504 \pm 5) \, \Omega$ resistor is much larger than the VULCAN input impedance $Z_{\text{ROB}} \approx 5 \, \Omega$ it is very simple to estimate the current:

$$i(t) = \frac{u(t)}{Z_{\text{tot}}} = \frac{u(t)}{R + Z_{\text{ROB}}}$$

Using this the slopes $m$ determined in the previous measurements can be used to calculate the current factor $m_i$.

$$m = \frac{a_{\text{LSB}}}{u} = \frac{a_{\text{LSB}}}{iZ_{\text{tot}}} \Rightarrow m_i = m \cdot Z_{\text{tot}} = \frac{a_{\text{LSB}}}{i}$$

The thus determined values are listed in Table 2. Considering a baseline of 10-20 LSB and switching to the next range at 240-250 LSB the full usable range is assumed to be 230 LSB. Single PE pulses have an average amplitude of 6 mV at 50 Ω with the used PMT. This is equivalent to $120 \, \mu A$.

The estimated PE values here are just rough estimates as the current per SPE is only estimated based on measurements performed with a commercial ADC. As the current calibration was done with the ROB directly connected via the 500 Ω adapter, the current per LSB changes when matching solutions like the transformers are used.
Table 2: Range matching results. The lower table shows rough estimates based on measurements with a different ADC and does not regard gain or attenuation from impedance matching.
6 Baseline regulation

6.1 Concept

As could be seen in the previous section the positions of the baseline of all three receivers are very important for the correct operation. This can be a problem because for example temperature and aging can influence the baseline. Electromagnetic interference can also influence the baseline. Lower frequency signals can be picked up from external sources or from the iPMT electronics themselves.

Without any compensation for baseline movement the effective trigger level will vary too. Another problem is that baseline movement in the MG and LG receivers cannot be monitored from the output data. Data from the MG and LG receiver will only be recorded in the waveforms for high signal amplitudes.

It is possible to monitor the baseline to reduce these negative effects. For example the trigger logic in the FPGA can monitor the HG baseline position and use a baseline dependent trigger level. The charge could also be corrected based on baseline monitoring data. It could also be corrected on a per waveform basis by recording enough baseline data in front of every pulse.

Data recorded via JTAG can also be used to monitor the otherwise unaccessible baseline of the MG and LG receivers.

But all methods that are based on monitoring the digitized data are limited by the resolution of the ADC. There is no way to know the actual BL value more precise than the ADC resolution if the baseline is very stable and at only one value. This can be mitigated by actively regulating the baseline to the edge between two ADC values.

Actively regulating the baseline of all receivers to a known value also simplifies the matching of the three ADC ranges. Monitoring the baseline values is then, although still a good idea, not strictly necessary anymore.

![Overview of the VULCAN baseline regulation](image)

The baseline regulation built into the VULCAN chip calculates the difference between the actual digitized value and the target value for each sample. This error signal is multiplied by a configurable coefficient before adding it to an accumulation register. A larger scaling factor for the error signal can be used for signals above a configurable threshold to shorten the time it takes the BLR to settle. The higher
bits of the accumulator control the DACs that provide the reference voltage to the
ADC discriminators [17].

Additionally a second regulation loop can be turned on that changes one of the
frontend bias voltages if the previously described loop hits a limit. This can be used
to automatically find the correct input biasing settings at startup and to adjust the
values if the DAC control loop reaches a limit during normal operation [5].

6.2 Testing

The VULCAN baseline regulation has many parameters. A few of them are the
regulation directions of different aspects of the loop. This is necessary as the polarity
of the input signals can be swapped via configuration registers. Others are the
regulation settings that can tune the regulation speed.

Additionally there is a register that completely disables the regulation of the
DACs that would otherwise be the subject of the BLR.

Figure 30 displays the parameters of the baseline regulation loop that are ac-
cessible via the status register. A script regularly stops the updating of the status
register and reads the values via JTAG. Ca. 17s after the logging is started a second
script connected to same JTAG port and enabled the baseline regulation. Due to
the limited readout rate via JTAG it is not possible to watch the BLR as it turns
on. The plot just shows the values without the BLR and then values once the BLR
has already stabilized.

Once the baseline regulation is turned on the error accumulator starts to change
from the default value. Based on the accumulator the DAC values are regulated and
stabilize around the optimal value. It can be seen that the baseline target value of
10 is reached. Even if the difference to the target value is zero a small error value is
added to the accumulator. This allows to potentially achieve the sub-LSB precision
as the baseline values oscillate close to the target value.

The BLR tries to shift the ADC range so that the ADC values are moved towards
the baseline target. So if the BLR is active while a signal is present at the input
it will work against the it and reduce the pulse height. After the pulse the BLR
needs time to recover to the correct values. This creates an undershoot after a signal
pulse.

To prevent this the VULCAN BLR can be automatically paused if the input
signal is above a configurable threshold. So if a pulse arrives it will surpass this
threshold quickly. Due the comparatively slow response of the BLR and the low
amplitude below the threshold the effect of the BLR before surpassing the threshold
is very small. This threshold needs to be lower than the minimum signal ampli-
tude and large enough so that most random baseline fluctuations don’t stop the
regulation. Using a threshold limits the BLRs reduction of high amplitude signals.

If the tia_control loop is not enabled the BLR is limited by the range of the DAC
values. So if the BLR is already close to the lower limit it will not be able to do
much regulation affecting the pulses.

It is important that the baseline mean value never surpasses the threshold. Once
the baseline value is above the threshold the regulation loop is paused until the
baseline randomly sinks below the threshold again which might take a long time. This is mainly a problem when starting the BLR. One way to prevent this is to start the BLR with a very high baseline target and gradually reduce it. An even simpler solution is to just disable the threshold at startup and enable it after a short amount of time.

The same pulse recorded with different BLR threshold settings is shown in Figure 31. Enabling the baseline regulation results in a slightly reduced signal waveform. Without the threshold there is a significant undershoot after the pulse. As can be seen enabling the threshold significantly reduces the effect.

Figure 31: Average waveforms of similar pulses for different settings of the baseline regulation.
The baseline regulation effectively acts as a high pass filter. Low frequency signals would lead to a deviation from the target baseline that is being compensated. The same happens for higher frequency signals, but the dampening effect of the accumulator strongly reduces the effect for high frequencies.

This behavior can be seen in Figure 32. To test the frequency response a sine wave was fed to the ROB input. Resistive matching was used to get the maximum bandwidth. Unfortunately the necessary blocking capacitor acts as a high-pass and attenuates lower frequencies.

To compensate this the amplitude was adjusted manually depending on the input frequency to get a usable amplitude of the disturbing signal. As the BLR only works for low baseline values the possible amplitude of the disturbing signal was limited. A higher signal would reach the lower end of the digitizer range.

To measure the effect of the disturbing signal the standard deviation of 1000 waveforms was calculated. Simply looking at the difference of the maximum and minimum would lead to problems as the dynamic range below the baseline is really limited. The standard deviation provides a useful measure of the amplitude of fluctuations around the baseline. This measurement was performed with the BLR turned on and off to see the effect of the baseline regulation.

![Figure 32: Measurement of the ability to reject disturbing signals of different frequencies. The ratio of baseline standard deviation $\sigma_{BL}$ with and without baseline regulation is plotted for an injected sine wave signal.](image)

The measured standard deviation is the result of both the disturbing signal and the normal noise of the setup. Unfortunately it is not possible to choose the amplitude of the input signal high enough that the normal noise is negligible (see Table 3). Calculating the standard deviation caused by the sine signal alone based on the known noise level cannot be done by gaussian error propagation because the values sampled from the sine wave are not normally distributed. Thus the
Table 3: Table listing the standard deviation of the HG receiver baseline with and without BLR for an input sinewave of varied frequency.

<table>
<thead>
<tr>
<th>f</th>
<th>$\sigma_{BLR, on}$</th>
<th>$\sigma_{BLR, off}$</th>
<th>$\sigma_{BLR, on}/\sigma_{BLR, off}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>–</td>
<td>1.870 ± 0.007</td>
<td>1.819 ± 0.013</td>
<td>1.028 ± 0.008</td>
</tr>
<tr>
<td>1 kHz</td>
<td>2.019 ± 0.018</td>
<td>3.714 ± 0.015</td>
<td>0.544 ± 0.005</td>
</tr>
<tr>
<td>3 kHz</td>
<td>2.25 ± 0.05</td>
<td>4.23 ± 0.05</td>
<td>0.532 ± 0.014</td>
</tr>
<tr>
<td>10 kHz</td>
<td>2.165 ± 0.013</td>
<td>3.593 ± 0.018</td>
<td>0.602 ± 0.005</td>
</tr>
<tr>
<td>30 kHz</td>
<td>2.362 ± 0.016</td>
<td>4.16 ± 0.05</td>
<td>0.567 ± 0.008</td>
</tr>
<tr>
<td>100 kHz</td>
<td>2.447 ± 0.015</td>
<td>3.062 ± 0.021</td>
<td>0.799 ± 0.007</td>
</tr>
<tr>
<td>300 kHz</td>
<td>4.660 ± 0.029</td>
<td>4.83 ± 0.04</td>
<td>0.964 ± 0.011</td>
</tr>
<tr>
<td>1 MHz</td>
<td>5.581 ± 0.025</td>
<td>5.34 ± 0.04</td>
<td>1.045 ± 0.009</td>
</tr>
<tr>
<td>3 MHz</td>
<td>5.420 ± 0.020</td>
<td>5.191 ± 0.020</td>
<td>1.044 ± 0.006</td>
</tr>
<tr>
<td>10 MHz</td>
<td>4.705 ± 0.008</td>
<td>4.694 ± 0.019</td>
<td>1.002 ± 0.004</td>
</tr>
<tr>
<td>25 MHz</td>
<td>3.705 ± 0.007</td>
<td>3.588 ± 0.010</td>
<td>1.033 ± 0.004</td>
</tr>
</tbody>
</table>

absolute values shown in Figure 32 don’t hold much meaning. Comparing them for the different frequencies does however show that the cutoff frequency of the BLR is roughly 100 kHz.

### 6.3 BLR & 6-bit ADC alignment

The baseline regulation changes the offset of the 6-Bit ADCs. Unfortunately the change is not the same for all ADCs as can be seen in Figure 33. The left half of the figures shows the rising edge of a sawtooth signal. Both waveforms were recorded with the same setup apart from the BLR that was turned on in one and off in the other. The data recorded without BLR looks fine because the 6-bit ADC alignment was done without BLR. Once the BLR is turned on there are suddenly significant gaps between the 6-bit ADCs.
Figure 33: Left: The same SG pulse digitized with BLR on and off. The MG receiver is used due to the more linear response. Right: The same signal recorded with BLR target 10 and 20 (bl). Both measurements were done once with the offset values tuned for the baseline of 10 and 20 (cal). The blue and green graph are recorded with the correct calibration. The other two were captured with the wrong calibration values.

This is not a problem if the alignment changes just a because the BLR is turned on. But the right half of Figure 33 shows that changing the baseline target also affects the alignment. When a different baseline target is chosen than the one used for the calibration the waveform effectively looks tilted as a result of the misaligned 6-bit ADCs.

As a consequence of this the 6-bit ADC alignment has to be done after the configuration of the BLR if the BLR is used. To use the BLR the temperature range where changes of the 6-bit ADC alignment do not affect the system performance should be determined. This measurement requires a temperature controlled room.
7 VULCAN evaluation

7.1 Trigger

A central requirement for the system is a working trigger. The rate of random mistriggers should be considerably lower than the dark rate of the PMTs. At the same time the ratio of missed real signals should be as low as possible.

The most simple way is a plain threshold trigger. In Figure 34 a histogram of the amplitude of waveforms is shown demonstrating this. The peak at 16 LSB is the result of empty waveforms. Waveforms containing SPE pulses are distributed around 54 LSB. For this exact setup reasonable trigger values are between 25 and 40. Below random fluctuations of the baseline would make up the majority of recorded waveforms and above too much of the actual signal would be lost.

Another way to look at this is demonstrated in Figure 35 where the trigger rate of the readout board is plotted against the trigger threshold. Fitting an error function to the baseline and SPE peak allows to estimate the mean SPE response. The SPE fit also includes an exponential function.

\[
R_{\text{SPE}}(a_{\text{thr}}) = \frac{b}{2} \cdot \left(1 + \text{erf}\left(\frac{a_{\text{thr}} - \mu}{\sqrt{2}\sigma}\right)\right) + c \cdot \exp(-a_{\text{thr}}/\tau)
\]  

(6)

The amplitude spectrum estimates a SPE amplitude of 38 LSB. Looking at the trigger rates yields 46 LSB. The difference is probably a result of the non-linearity that can be seen in the amplitude histogram.

![Figure 34: Frequency of amplitudes in waveforms. Pulses induced by light from a LED are recorded.](image)

A test of this can be seen in Figure 36 where different charge spectra are shown. The slight misalignment of the baseline peak is probably an artifact of the external trigger. Using a threshold trigger also selects high amplitude dark noise resulting in the increased count of above SPE pulses compared to the externally triggered data.
Figure 35: Trigger rate for different thresholds for PMT pulses. The recorded spectrum is the result of the ADC baseline and dark noise pulses.

The external trigger will just rarely catch a few large pulses if they happen to fall into the trigger window. And even then their timing might not be perfect and part of the pulse is outside the fixed charge integration window.

Slight coupling of the trigger signal to the input leads to a small difference in the average baseline. A problem with this measurement is that the used LED does not homogeneously illuminate the photocathode. This results in a different gain of the dark noise as the electrons released are differently distributed across the photocathode. The fact that the PMT was operated without magnetic shielding increases this effect. Nonetheless a threshold of 35 results in a good recreation of the SPE peak. Adjusting the threshold in 1 LSB steps gives a fine control of the trigger.
Figure 36: Comparison of the charge spectrum with externally triggered with an LED and with a threshold trigger.

7.2 Follow-up trigger

Once a trigger condition is met a fixed window is read out. This is a problem if a pulse is larger than this window or more pulses arrive within the window that would be cut off at the end of the waveform. The follow-up trigger will start recording a new waveform directly after the prior if the trigger condition stays active until the end of the prior waveform or a new trigger condition arises. For this to work the delay settings after which the switch to the next buffer happens has to set correctly.

Figure 37 shows the reconstruction of a long saw-tooth pulse from individual waveforms. The joining of the waveform works well and no samples are missed.
7.3 Resolution

The resolution is different depending on the signal amplitude as the lower gain receivers have a lower absolute resolution. Based on the range matching results (Section 5.1) and the SPE amplitude taken from Figure 34 the maximum PE count for each receiver can be calculated. These values are listed in Table 4. A low maximum PE value in the HG receiver compared to Table 2 due to the higher current from the transformer is necessary to reliably trigger upon SPE pulses. Using transformers for the impedance matching results in this increased gain. If necessary the PGA can be used to adjust the MG and HG receiver gain. Adjusting the PGA gain can increase the gain by 1.5 or reduce it by 0.4 compared to the default, resulting in the maximum and minimum values listed in the table.

If needed it might be possible to additionally reduce the MG receiver gain with an external circuit. For example a voltage divider could be used. It might also be possible to further attenuate the LG receiver through configuration. Adjusting the ADC ranges is an option that might work.

<table>
<thead>
<tr>
<th>Range</th>
<th>LSB / PE</th>
<th>PGA default (0x01)</th>
<th>PGA max (0x80)</th>
<th>PGA min (0xFF)</th>
</tr>
</thead>
<tbody>
<tr>
<td>HG</td>
<td>38.0</td>
<td>6.1</td>
<td>1.3</td>
<td>13.3</td>
</tr>
<tr>
<td>MG</td>
<td>3.7</td>
<td>61.3</td>
<td>13.5</td>
<td>134.9</td>
</tr>
<tr>
<td>LG</td>
<td>0.5</td>
<td>442.1</td>
<td>97.3</td>
<td>972.5</td>
</tr>
</tbody>
</table>

Table 4: Estimated resolution for photoelectron counting and maximum PE per range.
The PE/full range values listed in the table are the maximum PE count that can be digitized at the same time. In reality the PE pulses from high energy events will not all arrive at the exact same time but will be distributed. That means that events with a higher PE/PMT count can still be recorded without saturating the ROB input.

7.4 Clock leakage

As can be seen in Figure 38 the even and odd samples of the waveforms have different average values. This is most likely because the internal clock is slightly coupled into the signal lines somewhere. The amplitude changes depending on the configuration which is most likely because the configuration adjusts signal paths that are differently coupled to clocked signal lines.

This does not affect the general performance much as this effect is much smaller than the normal noise levels of the receivers. For example the standard deviation of the baseline of the HG ADC of approximately 2LSB is more than a magnitude larger than this coupled clock noise of less than 0.1 LSB.

![Average waveform showcasing the minor difference between even and odd samples in the HG ADC.](image)

7.5 Non-linearity

A histogram of all samples in a set of recorded waveforms is shown in Figure 39. There are significant differences in the frequency of neighbored bins. This non-linearity is probably due to the thresholds in the ADCs not being perfectly regularly spaced. In principle these thresholds could be optimized but this was not attempted yet due to time constraints.

This effect is not visible in the trigger rate scan shown in Figure 35. The trigger rate scan shows effectively the integral over the amplitude spectrum. In this integration all values above the threshold are added. This effectively averages out the differences between the samples.
Figure 39: A histogram of all ADC samples in $10^6$ waveforms. The data was recorded with the PMT connected.

7.6 Undershoots

Undershoots as shown in Figure 40 can arise for different reasons. The BLR creates undershoots if it is turned on during a pulse. But even without undershoots can be the result of bandwidth limiting. This is demonstrated in Figure 41 where a log-normal pulse is plotted after different filters. The effect of the lowpass filter is relatively low. It mainly delays the input signal. The highpass filter on the other hand creates an undershoot because the lower frequency components keep the baseline stable.

The VULCAN ASIC has an internal circuit that can actively compensate undershoots. This Overshoot Compensation (OSC) has not been tested yet.
Figure 40: Waveform of a PMT pulses with overshoot recorded with transformers in the signal path. The baseline regulation was turned off during the measurement.

Figure 41: Effects of a limited bandwidth on a simulated PMT pulse.
8 System evaluation

8.1 Hardware

To test the stability of the whole setup regular measurements were taken over a few days. Figure 42 effectively shows the baseline values recorded during these measurements. There are significant fluctuations that are most likely an effect of the changes in ambient temperature as these tests were not done in a temperature controlled room. This is not a significant problem as it can be either corrected by the BLR or compensated for afterwards based on data recorded via JTAG.

Figure 43 shows the correlation between the baseline and the temperature measured by an I2C temperature sensor on the ROB. This shows that the temperature has an influence on the baseline. A line fit estimates a temperature dependent baseline shift of $-2 \text{ LSB/}^\circ\text{C}$.

![Figure 42: Baseline over longer times.](image)

The gain calculated from the same measurements is shown in Figure 44. Overall the gain is really stable. It just changes after the gaps. The gaps are pauses when other measurements were executed. For these other measurements the HV had to be turned off and the cabling between the PMT and the ROB was changed. Whenever the HV module is turned on again there is a slight change in the HV value that will also change the PMT gain. The many SMA connectors used in the test setup to connect the PMT to the ROB are not always tightened by exactly the same amount leading to a different attenuation of the signal. Overall the gain is stable. Looking at the correlation between gain and temperature would not provide much insight as the gain of the PMT also changes with temperature. To get meaningful results from such a correlation the temperature dependent gain of the PMT would have to be known.

Although the power scheme will be changed in the next prototype, the rough power consumption of the system will not change. The ROB consumes a total of...
8 W from a 24 V supply when fully configured. Figure 46 shows measurements from the on-board power sensing. The 24 V are first stepped down to 5 V. Two separate 3.3 V rails are generated for the VULCAN ASIC and the ZYNQ system. These are the main power sink totaling more than 5 W. Due to the conversion losses more than 1 W of the input power is dissipated in the DC-DC converters.

Compared to this the power needed for the SCCU is small at about 1.2 W. In the final design there will also be a HV module that requires about 2 W. The total power draw for the final iPMT design will probably be 12 W.

All energy consumed by the iPMT electronics will be dissipated as heat. For the prototype a fan was used to reduce stress on the electronics (Figure 45).

The iPMTs will be operated in the temperature controlled water pool of OSIRIS. If the heat produced by each iPMT is constant in time the system will stabilize at a constant temperature. This temperature should be as low as possible because the failure rate increases with the temperature [9].

The final iPMTs electronics will be encased in a metal can at the end of the PMT. Oil will be filled into this enclosure to achieve a good thermal coupling to the water pool. The oil also helps to isolate the electronics which is especially useful for the high voltages required to operate the PMTs.
Figure 44: Gain over longer times.

Figure 45: Recorded temperature and power data. At around 13:20 the cooling fan moved resulting in rising temperatures. Later the fan moved completely away at 15:30. At 20:50 the cooling fan was put back into the correct position.
Figure 46: Screenshot of the power consumption monitoring of the ROB board prototype. The values are recorded by INA226 I2C power and current monitors [20]. An EPICS IOC monitors the values via the SCCU. A caQtDM GUI is used to display data that was accessed via EPICS.
8.2 Software

Software stability A fixed version of the SCCU firmware was used during all measurements that performed without any problems. The JTAG and I2C interfaces were used continuously during the measurements and showed no problems.

The ROB firmware was continuously being updated to add new features and improve performance. Overall the system works and was operating stable for weeks.

Readout rate and network bandwidth Figure 47 shows the used network bandwidth for different trigger rates. It can be seen that up to 60 kHz the used bandwidth increases. For higher trigger rates the maximum transfer capacity using TCP/IP over the 100 MBit/s \( \approx 12 \text{ MB/s} \) link is reached and waveforms are lost.

A readout rate of 60 kHz is sufficient for the use in OSIRIS. The total expected event rate resulting from decays in the scintillator, water pool, detector structure and surrounding rock is below 10 s\(^{-1}\) [3, 7.2 Radiopurity Assumptions]. Considering the overburden of more than 700 m [6] of rock the muon background will be negligible compared to the other rates. Even if these result in more than one recorded waveform per iPMT the rate is negligible compared to the dark count. The dark count rates of the used PMTs are typically 10-30 kHz but definitely below 50 kHz [1]. Considering all this the event rate will be below 60 kHz which the ROB can handle.

![Network Bandwidth](image)

Figure 47: Used network bandwidth for different trigger rates.

Network setup The test setup runs in a local network organized by a lab computer. A dnsmasq DNS server and an ISC-DHCP server run on this lab computer to manage the network. Both were chosen because they very available via the package manager on the lab computer. Replacing them would be straightforward.

The ROB is connected via the Ethernet switch on the SCCU as it will be in the final design. A short cable connects the prototype to the lab computer. In the
detector the network cables will be longer leading to a slightly larger delay. This should not significantly affect the network throughput.

This setup should work without problems for the 80 iPMTs in OSIRIS.

**Scalability** Increasing the size of the iPMT readout system is really simple concerning the hardware. Putting more iPMTs into a detector system requires more Surface Boards and more POE switches. There would still be one Surface Board that provides the clock for all others.

Using the 10.0.0.0/8 private address range provides more than 16 million addresses. So even without using IPv6 there is enough address space for large detectors.

A larger device count is not a problem for the control system due to the distributed design of EPICS.

The more complicated aspect of increasing the readout system channel count is the DAQ as it needs to be able to process data coming from all computers. The DAQ servers need the timing information of trigger events from all iPMTs to determine coincidences. This is probably the major factor that would limit increasing the channel count for this design.
9 Conclusion and outlook

The prototype of the iPM T readout system was successfully commissioned. With calibration and adjustments to the configuration of the VULCAN ASIC an acceptable performance could be achieved.

Each individual receiver of the VULCAN ASIC was tuned to achieve the necessary performance. To do this an impedance matching circuit was tested to achieve the necessary resolution in the high gain receiver.

A concept was developed and tested to reconstruct the actual waveform from the data taken from the three ranges. This allows for the necessary dynamic range to detect signals resulting from single up to hundreds of photons. Due to a limited amplitude of the available signal generator the range matching could not be calibrated on the full range. The concept does however work for real PMT waveforms of large amplitudes.

Although the baseline regulation was tested and performed as expected, side effects render it unusable for now. The baseline regulation affects the alignment of the 4 6-bit ADCs that each receiver uses leading to an effective change in the ADC linearity.

In it’s current state the setup is capable to trigger on SPE events with a simple threshold trigger. It can also continuously record which is necessary if many photons are detected close to each other. A large buffer size and a fast enough continuous readout rate are available to record all dark noise and signals expected from the PMTs in the OSIRIS detector.

Overall the prototype was stable over extended periods of time and meets the tested requirements. Fluctuations of the baseline can be compensated afterwards based on samples of all three receivers recorded via the JTAG status register.
Appendix

Figure 48: Schematic of one receiver of VULCAN.
[Schematic of one receiver of VULCAN. (by J. Steinmann)]

Figure 49: Estimates for the effective resistor values in the PGA feedback network.
Figure 50: Linear fits to the data from the lower 6-bit ADC to determine the optimal offset. The upper ADC data is plotted in black.

Figure 51: PTV measured with different TIA settings.
References


REFERENCES


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