

Dissertation

Study of Muon Drift Tube Detectors and Fast Readout Electronics for Very High Counting Rates

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Readout Electronics for Very High Counting Rates

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Abstract

The Large Hadron Collider (LHC) at the European Center for Particle Physics, CERN, collides protons and lead ions with high luminosities at a center-of-mass energy of 13 TeV in the center of four detectors, of which one is the ATLAS experiment. After more than a decade of successful operation with the discovery of the Higgs boson, the instantaneous luminosity will be increased by almost an order of magnitude with the High Luminosity LHC (HL-LHC) in order to perform precision measurements of the Higgs boson and top quark and extend the searches for physics beyond the Standard Model of particle physics. A proposal for a further major increase of the collision energy and luminosities is the Future Circular Collider (FCC). In order to cope with the increased particle fluxes, upgrades of the LHC experiments and the development of new detector technologies are needed.

In this thesis, studies of the potential of small-diameter Muon Drift Tube (sMDT) chambers for precise track reconstruction at high particle fluxes and concepts of new readout electronics are presented. Twelve new sMDT chambers for the ATLAS Muon Spectrometer have already been constructed, tested and installed in ATLAS in order to gain experience and to optimize them for future applications. New read-out electronics concepts with improved signal shaping, which allow for the full exploitation of the sMDT rate capability have been studied. Measurements with generated pulses and on sMDT chambers in a muon beam at CERN under unprecedentedly high γ irradiation show the advantages of the new shaping schemes. The findings are supported by detailed simulations of the high rate effects and allow for prediction of the performance of the sMDT chambers and new readout electronics at the background particle fluxes expected at the FCC.

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Chapter 1

Experiments at Hadron Colliders

Key discoveries in particle physics have been made at high energy colliders. including the discovery of the W and Z bosons, of the top quark and the Higgs boson. With the Higgs boson, the last missing fundamental particle predicted by the Standard Model of particle physics was discovered at the **Large Hadron Collider (LHC)** at CERN near Geneva [1, 2]. Future high energy colliders aim for answering the many questions left open by the Standard Model.

This chapter gives an overview of applications of muon drift tube detectors in experiments at high energy hadron colliders. After a short introduction to the physics goals and to hadron collider concepts, the ATLAS experiment and its high-luminosity upgrade and the detector concept for a future 100 TeV hadron collider and the requirements for muon drift tube detectors at such an experiment are described.

1.1 Physics at High Energy Hadron Colliders

The Standard Model describes the experimental results with the highest precision. Its last missing piece, the Higgs boson, was discovered in 2012 by the two multipurpose experiments ATLAS¹ and CMS² at the LHC. However, there are many questions left open by the Standard Model, which have to be addressed by future high energy experiments with higher collision energies and luminosities. Besides more precise measurements of Standard Model processes, the measurement of the Higgs self coupling is one of the most important goals.

From astrophysical observations there is strong evidence for the existence of Dark Matter and Dark Energy, which are not described by the Standard Model (SM) of particle physics. Supersymmetric extensions of the SM, which postulate partners for all Standard Model particles differing by spin 1/2,

¹A Toroidal LHC ApparatuS

²Compact Muon Solenoid

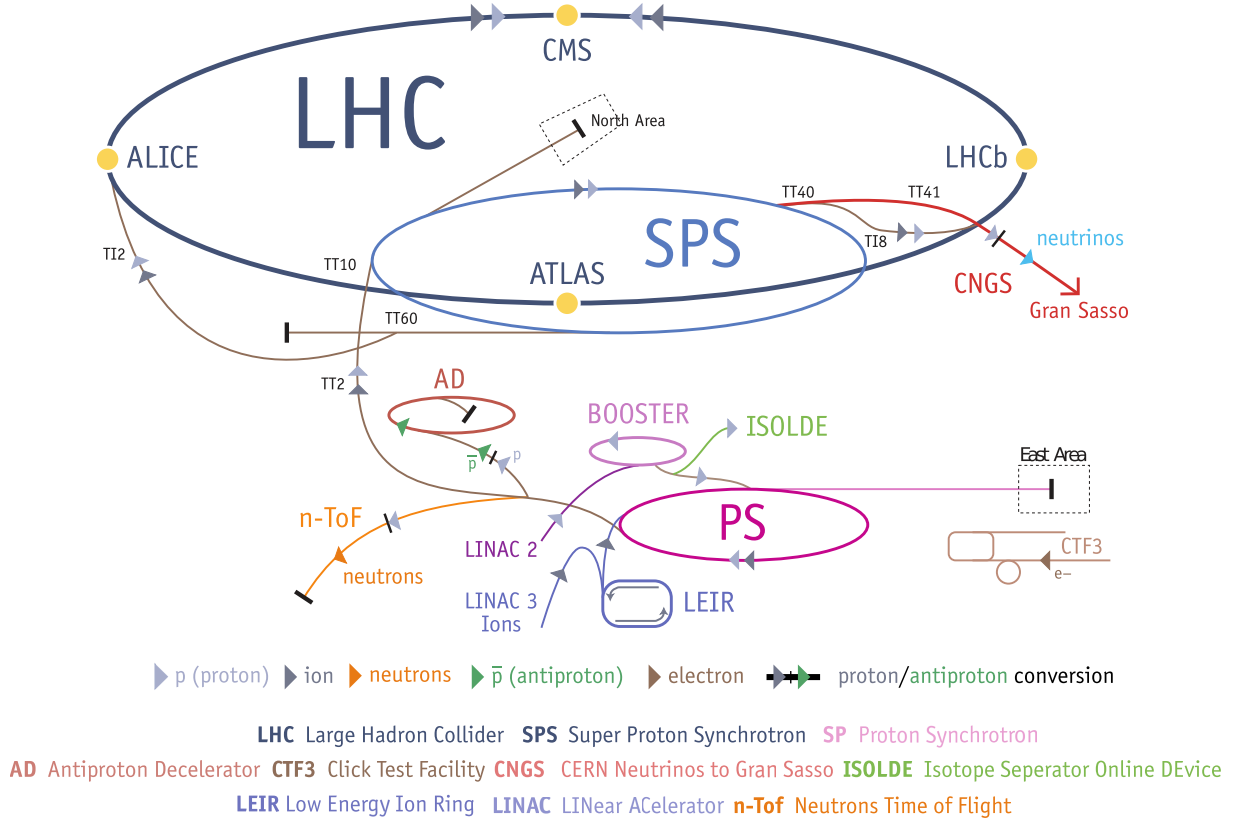


Figure 1.1: Schematics of the CERN accelerator complex with the LHC and its injectors [3].

provide Dark Matter particle candidates, which could be accessible at LHC energies. None of the particles predicted by **Beyond the Standard Model** theories have been found so far.

First hints for physics beyond the standard model, however, have already been found by the LHC experiments in lepton flavor violation. Du to the small branching ratios in the analyzed B decays, more data, i.e. larger collision rates, are needed to achieve statistical significance.

1.2 Accelerators

The **Large Hadron Collider** at CERN with a circumference of 27 km is the largest accelerator ever built. It is located 45 to 170 m under ground in the former LEP tunnel. 1232 superconducting dipole magnets with a maximum field of 8,3 T at 1.9 K bend the beams onto a circular trajectory. The accelerating cavities and the experiments are located along eight straight sections. Two tunnels connect the LHC to the beam injectors of the CERN accelerator complex as shown in Figure 1.1.

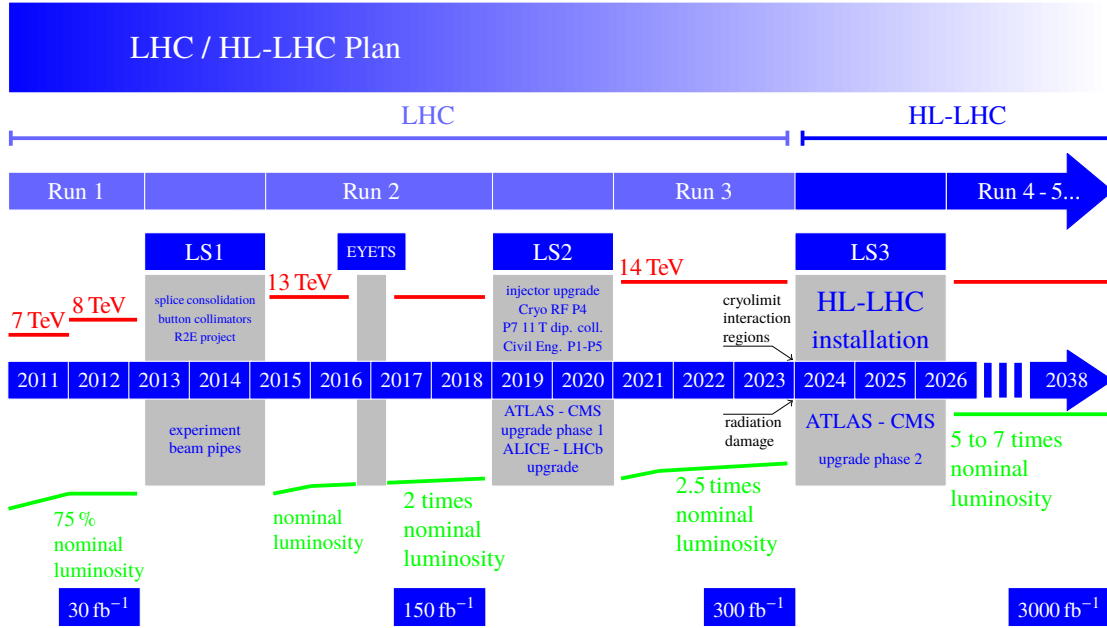


Figure 1.2: LHC baseline schedule for the luminosity upgrades showing the proton-proton collision energy indicated by the red lines and the instantaneous and integrated luminosity indicated by the green lines and the blue boxes, respectively [4].

The dipole magnets generate oppositely oriented magnetic fields of the same strength in two separate vacuum beam pipes, allowing for the circulation of particle beams of the same charge in opposite directions for proton-proton, proton-lead and lead-lead collisions. In 2022, the dipole magnets are expected to be ramped up to their maximal field for 14 TeV pp collisions for the first time. Protons, which are injected with an energy of 450 GeV, will then be accelerated to 7 TeV, and $^{208}_{82}\text{Pb}^{+82}$ ions will be accelerated from 177 GeV to 2.76 TeV per nucleon.

The LHC is operated in the proton mode for about 80 % of the data taking time. Up to 2 808 bunches with 10^{11} protons are successively injected from the SPS into the LHC. The beams are then continuously accelerated during the magnet ramp up and brought to collision at a frequency of 40 MHz at the four interaction points. After the first proton collisions with the injection energy of 450 GeV per beam in 2009, the collision energy was stepwise increased to 7, 8 and 13 TeV.

Due to increased beam focusing in the collision points by the quadrupole magnets, the instantaneous design luminosity of $\mathcal{L} = 10^{34} \text{ cm}^{-2} \text{ s}^{-1}$ was exceeded in 2016 by a factor of two. The LHC delivered an integrated luminosity of $L = \int \mathcal{L} dt = 0.2 \text{ ab}^{-1}$ to ATLAS and CMS within its seven years of operation since its startup in 2009.

The past and future LHC time schedule is sketched in Figure 1.2. Currently, during the second long shutdown (LS2) of the LHC, the linear proton pre-accelerator is replaced to increase the beam intensity in the LHC. After this so-called phase 1 upgrade, an instantaneous luminosity of $\mathcal{L} = 2.5 \cdot 10^{34} \text{ cm}^{-2} \text{ s}^{-1}$ will allow for the collection of an integrated luminosity of $L = 0.3 \text{ ab}^{-1}$ in the years 2021 to 2023.

During the phase 2 upgrade from 2024 to 2026, the CERN accelerator complex will be upgraded to the **H**igh **L**uminosity LHC (HL-LHC) [4]. This requires improved quadrupole magnets, crab cavities and upgrades of the other injectors and will enable five to seven times the LHC design luminosity. Within the following decade, the HL-LHC will then deliver an integrated luminosity of $L = 3 \text{ ab}^{-1}$.

The search for new particles with higher masses, however, requires a more powerful accelerator which delivers much higher collision energies [5]. The largest potential for the discovery of new physics is provided by the **F**uture **C**ircular **C**ollider with a pp collision energy of 100 TeV in a new tunnel of 100 km circumference (cf. Figure 1.3). As a first step, a lepton collider, FCC-ee, could be installed in the tunnel to study the Z, W and Higgs bosons as well as the top quark with high precision in e^+e^- collisions at center of mass energies from 80 to 400 GeV [6].

Due to energy limitations of linear electron-positron colliders and due to energy loss caused by synchrotron radiation of circular electron-positron colliders, the discovery reach for new heavy particles is driven by hadron colliders where the collision energies are limited by the dipole magnetic field strength. With superconducting Nb₃Sn dipole magnets, the second FCC step can be a hadron collider FCC-hh with proton collisions at 100 TeV center-of-mass energy [7].

10^4 bunches with 10^{11} protons each would be injected from the LHC at 3.3 TeV and accelerated to 50 TeV using 16 T dipole magnets. With a maximum instantaneous luminosity of $\mathcal{L} = 3 \cdot 10^{35} \text{ cm}^{-2} \text{ s}^{-1}$, the FCC-hh is designed to deliver an integrated luminosity of $L = 20 \text{ ab}^{-1}$ over ten years to two interaction regions and offers the possibility for two additional interaction points with lower luminosity. This instrument offers a wide reach for searches for physics beyond the Standard Model like supersymmetric or Dark Matter particles.

1.3 Hadron Collider Detectors

Large particle detectors with high solid angle coverage, are installed at the collision points of the colliders to measure the particles emerging from the interaction region and study their production and decay processes. Four big experiments are operated at the LHC. ATLAS and CMS are multi-purpose experiments, designed to study a wide range of hard collision processes including measurements of the properties of the Higgs

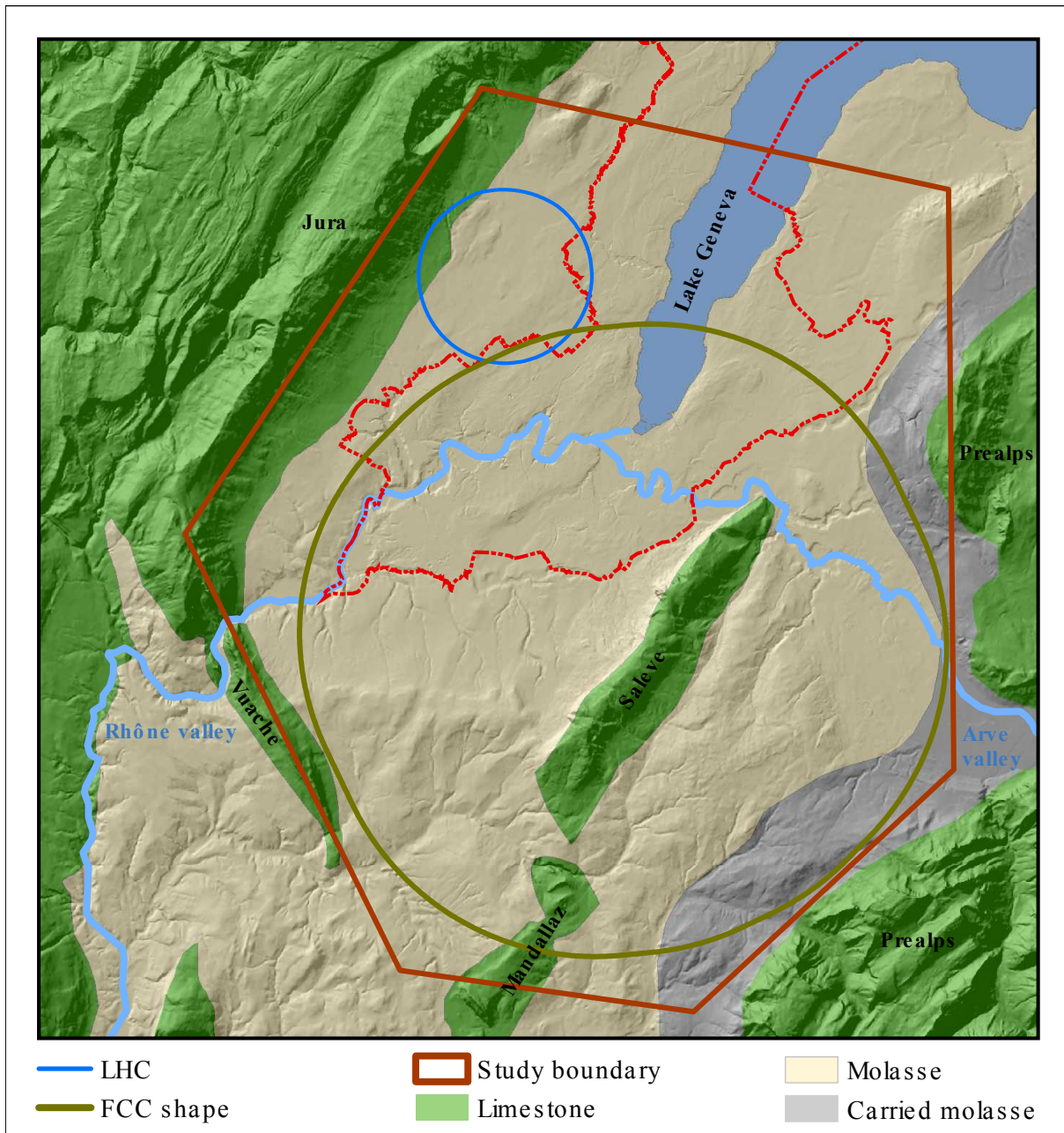


Figure 1.3: Map of the Geneva area with the LHC tunnel and the studied course of the FCC tunnel [7].

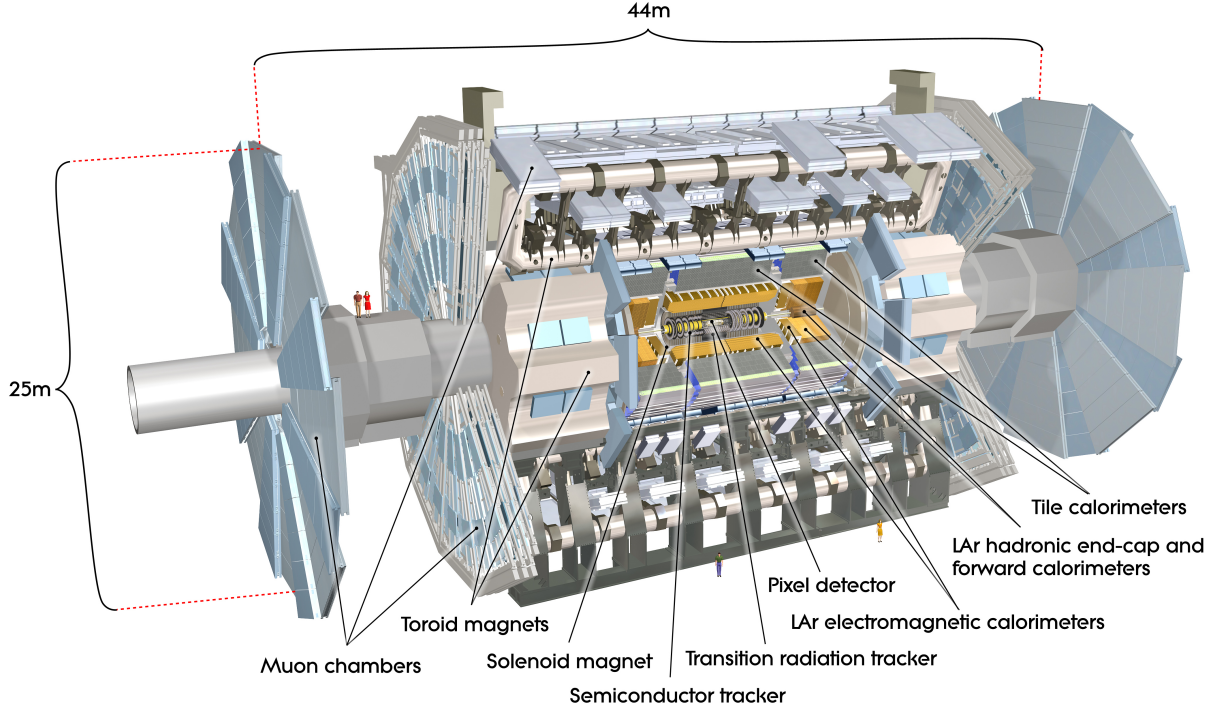


Figure 1.4: Cut-away view of the ATLAS experiment at the LHC [8].

boson and the top quark. ALICE¹ is dedicated to the studies of heavy ion collisions. LHCb² focuses on the measurement of rare decays of B mesons and of CP violation.

The largest LHC detector is ATLAS [8]. It was built in the years 2000 to 2008. Its sub-detector systems, shown in Figure 1.4, are arranged cylindrically around the beam axis in the so-called barrel region and in disks, in the so-called end-caps, in the forward regions. The detector is 25 m in diameter and has a length of 44 m.

The innermost system is the inner tracking detector. It consists of highly granular silicon pixel and strip detectors and a transition radiation tracker in a 2 T solenoidal field generated by a superconducting magnet.

A calorimeter system with electromagnetic and hadronic calorimeters surrounds the inner detector. The electromagnetic calorimeter of at least 22 radiation lengths thickness uses accordion-shaped lead absorbers and liquid argon as active medium. The hadronic calorimeter uses steel as absorber material

¹A Large Ion Collider Experiment

²Large Hadron Collider beauty

and scintillating tiles with photomultiplier readout in the barrel and liquid argon with copper absorbers in the forward regions.

The largest sub-system is the muon spectrometer in the toroidal magnetic field of three superconducting magnet systems, one in the barrel part and each of the two end-caps. Due to the air-core magnet design, the multiple scattering effect is small enough to allow for stand-alone muon momentum measurement. The superconducting magnet coils with eightfold symmetry provide a bending power between 1 and 7.5 Tm depending on the polar angle and resulting in a momentum resolution of about 10 % for 1 TeV muons. **Resistive Plate Chambers (RPC)** and **Thin Gap Chambers (TGC)** are used as fast muon trigger detectors. High precision track reconstruction is performed with **Monitored Drift Tube (MDT)** chambers and **Cathode Strip Chambers (CSC)**.

1.3.1 Upgrade of the ATLAS Muon Spectrometer

After a decade of successful data taking, the ATLAS detector including the muon spectrometer has to be upgraded to cope with the high data and radiation rates at the High Luminosity LHC (HL-LHC) [9]. While the inner detector has to be replaced completely due to radiation damage, the calorimeters and the muon spectrometer need new trigger and readout electronics and only the inner barrel and endcap layers of the muon detector have to be replaced.

The vast majority of hits in the muon detectors is not caused by muons but by the so-called cavern background. The radiation background in the ATLAS cavern consists mainly of γ -ray photons and, to a smaller extent, of neutrons and protons. They are created in interactions of the beam collision products in the beam pipe, the shielding and the detector components and are scattered to fill the whole cavern. The background photons and neutrons have an average energy of about 1 MeV and produce hits in the muon detectors. The background hit rate scales with the instantaneous luminosity.

In order to cope with the radiation of the HL-LHC, the CSC and MDT chambers in the inner endcap layers, the so-called small wheels, are replaced by micropattern gas detectors. The **New Small Wheels (NSW)** are equipped with **Micro-Mesh Gaseous Structure (Micromegas)** detectors for precision tracking and **small-strip Thin Gap Chambers (sTGC)** for triggering [10].

The efficiency loss of the RPC in the ATLAS barrel muon spectrometer due to aging have to be compensated by additional thin-gap RPC trigger chambers in the barrel inner layer of the muon spectrometer. The limited space in the regions of toroid magnet coils in the inner layer requires the replacement of the MDT chambers [9] in these regions by small diameter **Monitored Drift Tube (sMDT)** chambers [11]. sMDT chambers will be discussed in detail in Chapter 3.

The rate of low-momentum muon accepted by the RPC barrel first-level muon trigger increases vastly with the instantaneous luminosity. In order not to lose too many muon tracks from W and Z decays, the

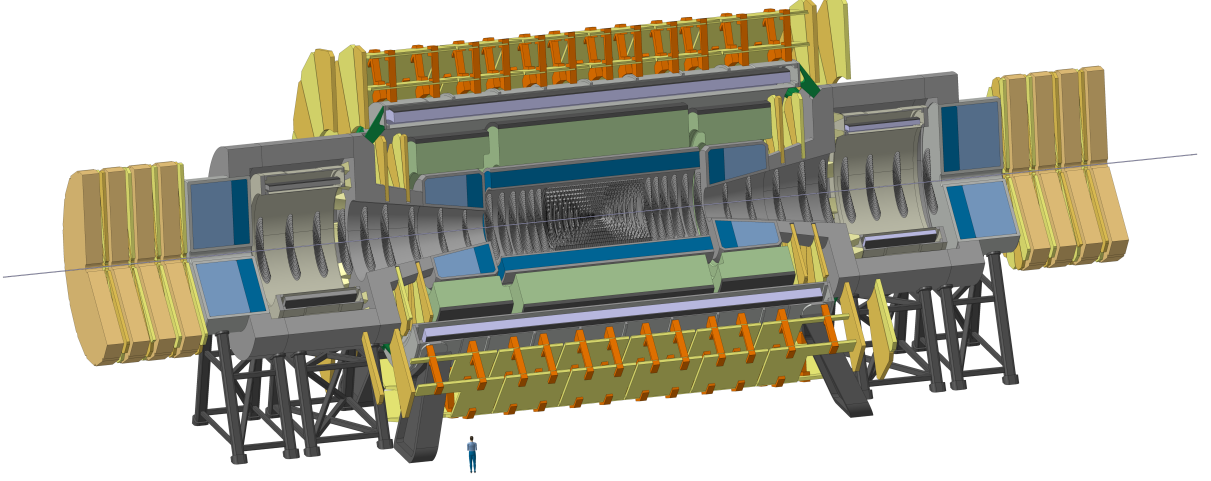


Figure 1.5: Cut-away view of the FCC-hh baseline detector with a central and two forward air-core solenoid magnet systems for precise muon momentum measurement [7].

nominal trigger threshold cannot be increased. Hence, an improved trigger selectivity is needed at the HL-LHC upgrade, to keep the muon trigger rate at an acceptable level. The required increase of the muon momentum resolution of the first trigger level by an order of magnitude requires to use the (s)MDT chamber measurements in the first level trigger. Hence, a new triggerless readout of the MDT chambers has to be implemented, which requires the replacement of the on-chamber electronics.

1.3.2 The FCC Reference Detector

Due to the maximum instantaneous luminosity of $\mathcal{L} = 3 \cdot 10^{35} \text{ cm}^{-2} \text{ s}^{-1}$ in proton-proton collisions at the FCC-hh collider, unprecedentedly high background fluxes are expected in the muon system of the associated detector. Because of their high spatial resolution up to very high background rates over large areas, sMDT chambers are a natural candidate for the instrumentation of the muon spectrometer [12]. γ conversion fluxes of $f = 500 \text{ s}^{-1} \text{ cm}^{-2}$ in the barrel region and up to $10 \text{ ms}^{-1} \text{ cm}^{-2}$ in the inner endcap muon systems are expected.

Figure 1.5 shows the design of the FCC-hh baseline detector. The innermost tracking detector and the electromagnetic and hadronic calorimeters are operated in a magnetic field of 4 T generated by a huge superconducting solenoid magnet surrounding the calorimeters. Two additional superconducting solenoid magnets bend the tracks in the forward regions up to very small polar angles. The total length of the detector is 50 m and the diameter is 15 m.

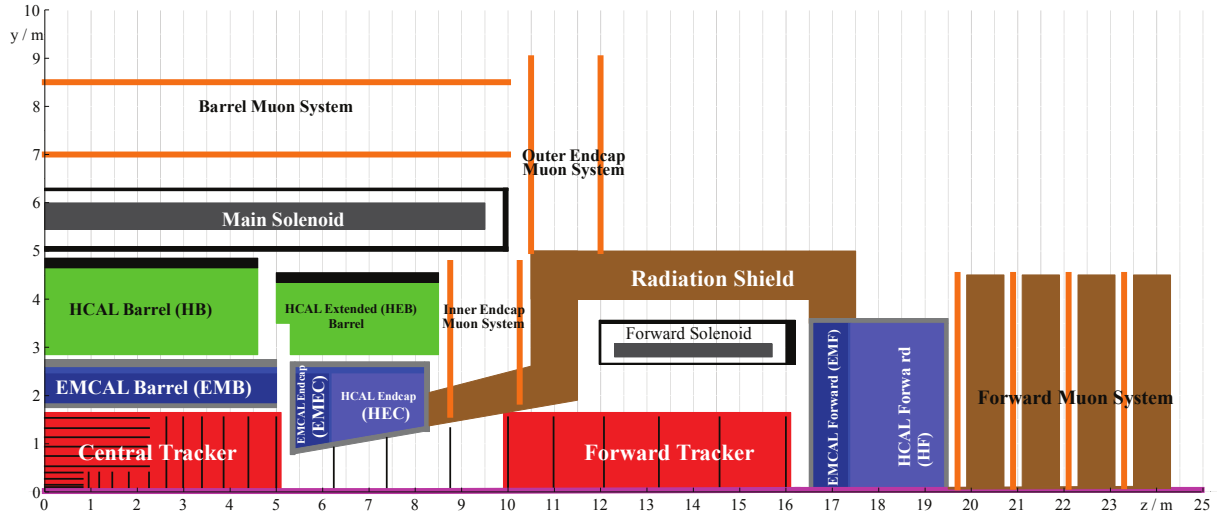


Figure 1.6: Cross section of one quadrant the FCC-hh reference detector with its sub-systems [7].

The arrangement of the detectors sub-systems is shown in Figure 1.6. The barrel muon system consists only of one detector layer for precision muon momentum measurement together with the inner tracker as stand-alone high-resolution first level muon trigger.

Chapter 2

Performance of Drift Tube Detectors

In the ATLAS muon spectrometer, the majority of tracking detectors are MDT¹ chambers [8], gaseous detectors consisting of drift tubes with anode wire read-out. Due to their performance parameters, drift tube detectors are widely used in high energy, nuclear and heavy ion physics experiments for track reconstruction. This chapter is about the operation of drift tube detectors using the example of ATLAS MDT chambers, their limitations at high rates and their successor, the small diameter **Monitored Drift Tube** chambers [13].

In drift detectors, the charge carriers generated by ionizing energy deposition of high energetic particles, e.g. muons, in the active detector volume are separated in an electric field between the cathode and anode wire and drift to the opposite electrodes. The duration of the charge drift is used to determine the distance between the energy deposition of the primary particle and the electrode, i.e. the drift distance, in order to reconstruct the particle track.

Detectors with gaseous active medium are called chambers. The charge carriers are generated by ionization of the gas molecules. The electric field increases rapidly near the anode wires leading to an acceleration of the drifting electrons and an ionization avalanche in the immediate vicinity of the anode wires, which amplifies the ionization signal, which is then read out by electronics connected to the wires. Knowing the time of the particle passage through the chamber from a trigger or the accelerator clock, the drift time is used for precise track reconstruction.

2.1 Resolution and Efficiency of Drift Tubes

In high energy physics experiments, where the chambers have to cope with high particle fluxes and counting rates, the detectors often consist of drift tubes. The anode wires are surrounded by electricly

¹**Monitored Drift Tube**

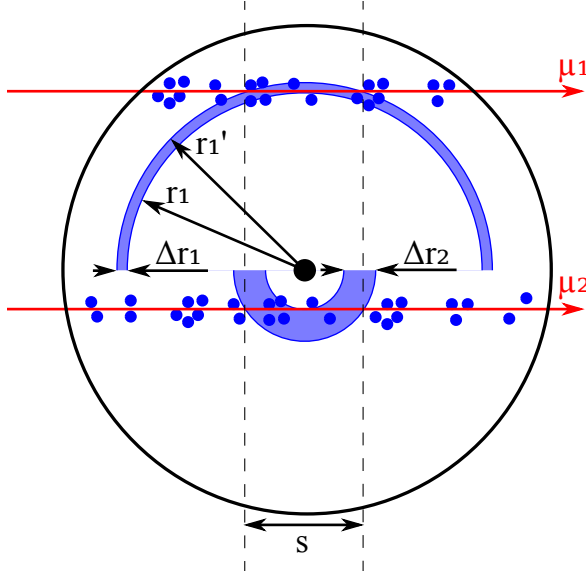


Figure 2.1: Graphical explanation of the improvement of the resolution with increasing drift radius.

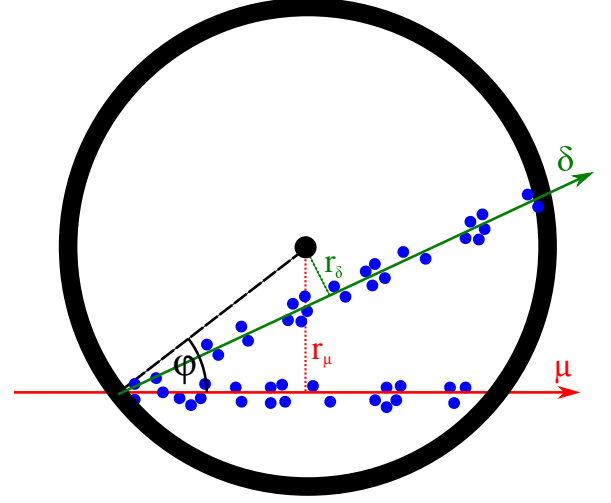


Figure 2.2: Illustration of the masking of muon hits by δ -electrons generated in the tube wall.

conducting tubes over the whole length, which acts as the cathode. In this geometry, the charge carriers, generated in the Townsend avalanche, do not affect the electric field of surrounding wires.

In the ATLAS MDT chambers, the drift tubes have a circular cross-section with a $t = 0.4$ mm thick aluminum wall, a tungsten-rhenium (97/3) wire of $r_{\min} = 25 \mu\text{m}$ radius with a gold coating in the center and are filled with Ar/CO_2 (93/7). Applying a voltage U between the wire and the tube wall, the cylindrical geometry of the tube propagates to the radial electric field

$$E(r) = \frac{U}{r \cdot \ln \frac{r_{\max}}{r_{\min}}} \quad (2.1)$$

in the active detector volume, where $r_{\max}$ is the inner tube radius.

High energetic particles ionize gas molecules and the created electrons are accelerated towards the anode wire. Due to the high gas density, the electrons are frequently scattered and lose energy. Consequently, the electrons drift with a velocity in the order of $30 \frac{\mu\text{m}}{\text{ns}}$ in ATLAS drift tubes.

Within about 0.2 mm around the wire, the electric field is strong enough, that the electrons gain enough kinetic energy between two collisions with gas molecules, that they generate secondary electrons by ionization, resulting in a Townsend avalanche. The desired multiplication factor of $2 \cdot 10^4$, which is called the gas gain, is achieved with a high voltage $U \approx 3$ kV. The resulting charge separation induces an electric signal on the wire, which is large enough to be read out by the electronics. As each tube acts as

a stand-alone detector with two cylindrical electrodes, the signal current on the anode wire by a charge carrier can be calculated analytically using the weighting field method based on Ramo's theorem [14]. The behavior of the intrinsic resolution of drift tubes can be explained in a simple picture, illustrated in Figure 2.1, which sums up the effects of electron diffusion and noise. Let s be the length of the ionization path generating sufficient primary ionization charge for a detectable signal. The radial distance difference between the closest and the most distant ionization to the wire is

$$\Delta r = r' - r = \sqrt{r^2 + \frac{s^2}{4}} - r.$$

Radius straggling increases close to the wire deteriorating the spatial resolution. Hence, the achievable resolution of drift detectors improves with the average drift distance favoring large tube diameters. For large drift radii, the statistical fluctuations of the primary ionization and deteriorate the effective value of s and hence the resolution. MDT chambers with 30 mm outer tube diameter achieve single tube resolutions of better than 200 μm close to the wire and 60 μm at large drift radii with time slewing corrections of the read-out electronics (cf. Figure 2.3 explained in Section 6.3.4).

The ATLAS drift tube detection efficiency is 100% for small and medium drift radii. Within the outermost 1 mm, the particle track through the detector gas is too short to ensure enough primary ionizations for a detectable signal. The 3σ -efficiency $\epsilon_{3\sigma}$ is a stricter measure, additionally requiring the measured radius to be within three times the tube resolution around the muon track and hence considers the information loss due to masking muon hits e.g. by δ -electrons, which are secondary particles produced by muons in the tube wall. With about 20% probability, a muon knocks out a relativistic electron from the tube wall. Due to the smaller density, the contribution of δ -electron from the gas is negligible. If the electron's trajectory is closer to the wire than the muon, i.e. $r_\delta < r_\mu$ in Figure 2.2, the Townsend avalanche of the δ -electron arrives at the wire before the muon signal.

The probability for a scattering angle of the δ -electron which leads to a masking of the muon is

$$p_\varphi = \frac{\varphi(r_\mu)}{\pi} = \frac{1}{\pi} \arcsin\left(\frac{r_\mu}{r_{\max}}\right).$$

according to [15]. The probability for the generation of a δ -electron is proportional to the muon track length a in the tube wall with thickness $t = 0.4$ mm, which depends on the drift radius via

$$a(r_\mu) = \sqrt{(r_{\max} + t)^2 - r_\mu^2} - \sqrt{r_{\max}^2 - r_\mu^2}.$$

The dependence of the 3σ -efficiency on the drift radius caused by δ -electrons is hence

$$\epsilon_{3\sigma} = 1 - \frac{a(r_\mu)}{\lambda} p_\varphi = 1 - \frac{1}{\pi\lambda} \left(\sqrt{(r_{\max} + t)^2 - r_\mu^2} - \sqrt{r_{\max}^2 - r_\mu^2} \right) \arcsin \left(\frac{r_\mu}{r_{\max}} \right), \quad (2.2)$$

where $\lambda \approx 3$ mm gives the average muon propagation length to produce a δ -electron in aluminum. The 3σ -efficiency is 100 % close to the wire and drops to about 90 % in 1 mm distance to the tube wall, leading to an overall of about $\epsilon_{3\sigma} \approx 95$ %.

2.2 High Rate Effects in Drift Tubes

Ionizing particles produce electron-ion pairs along their path. While the electrons arrive at the anode wire within less than 700 ns maximum drift time in a 30 mm diameter MDT, the maximum drift time of the ions from near the wire to the tube wall is about 4 ms (see Equation (2.5)). This means, that in a MDT a homogeneous ion space charge is present for background radiation fluxes $f \gtrsim 150 \text{ s}^{-1} \text{ cm}^{-2}$ which depends on the background flux and modifies the electric field in the tube, leading to gain drop and stochastic drift velocity variations due to fluctuations of the background irradiation of the tubes and, therefore, of the space charge density.

The gas gain G , i.e. the number of ionization electrons produced from a primary electron in the Townsend avalanche, in the electric field $E(r)$ near the anode wire with radius $r_{\min}$ can be calculated using Diethorn's formula [16]

$$G = \left(\frac{E(r_{\min})}{E_{\min}} \right)^{\frac{r_{\min} E(r_{\min}) \ln 2}{\Delta V}}, \quad (2.3)$$

where E_0 and ΔV are Diethorn parameters which have been measured for possible MDT gas candidates [17]. ΔV is the ionization energy of the gas. $E_{\min}$ can be interpreted as the electric field threshold, at which the avalanche begins. It strongly depends on the mean free path of the electrons and its nominal value E_0 has therefore to be corrected for changes of the gas density $D = D(p, T)$ due to pressure and temperature compared to the density $D_0 = D(p_0, T_0)$ at nominal conditions:

$$E_{\min} = \frac{D \cdot E_0}{D_0}$$

The measured values published in [17] are $E_0 = 24 \cdot 10^5 \frac{\text{V}}{\text{m}}$ and $\Delta V = 34 \text{ V}$ with estimated errors of 10 % for Ar/CO₂ (93/7) at $p_0 = 100 \text{ kPa}$ and without specifying the reference temperature T_0 . Hence, calculations in this thesis correct $E_{\min}$ just for the pressure, but not for temperature.

The generated ions, which as first approximation assumed to be homogeneously distributed in the tube, reduce the electric field strength near the anode wire reducing the gas gain. The discussion follows reference [18]. A hit rate ν of ionizing background particles that generate on average N_p primary electron-ion pairs in a drift tube with wire radius $r_{\min}$, inner tube radius $r_{\max}$ and length l , leads to an ion charge density of

$$\rho_{\text{ion}} = \frac{\nu G N_p e t_{\text{ion}}}{\pi l (r_{\max}^2 - r_{\min}^2)}, \quad (2.4)$$

where t_{ion} is the maximum drift time of the ions with drift velocity $v_{\text{ion}}(r)$ from the surface of the wire to the inner tube wall:

$$t_{\text{ion}} = \int_{r_{\min}}^{r_{\max}} \frac{1}{v_{\text{ion}}(r)} dr.$$

In the low field region outside of the avalanche, the ion drift velocity is $v_{\text{ion}}(r) = \mu_{\text{ion}} E(r)$. As the electric field and, therefore, the ion velocity are inverse proportional to the drift radius (cf. Equation (2.1)) and the volume filled with drifting ions increases linearly with the drift radius, the assumption of a homogeneous ion density is underpinned. In Ar/CO₂ (93/7) the Ar⁺ mobility μ_{ion} , is constant $\mu_0 = 1.535 \cdot 10^4 \frac{\text{m}^2}{\text{Vs}}$ and has only to be corrected linearly for density D variations, from the nominal value D_0 at standard conditions. In the case of negligible space charge, the ion drift time is determined using Equation (2.1):

$$t_{\text{ion}} = \int_{r_{\min}}^{r_{\max}} \frac{1}{\mu_0 \frac{D_0}{D} \frac{U}{r \cdot \ln \frac{r_{\max}}{r_{\min}}}} dr = \frac{D \ln \frac{r_{\max}}{r_{\min}}}{2 \mu_0 D_0 U} (r_{\max}^2 - r_{\min}^2). \quad (2.5)$$

Using Gauss's law for a charge Q in a cylindrical volume V of length l and radius r with the anode wire in its center gives a radial electric field $E(r)$ according to

$$\int_0^{2\pi} \int_{-\frac{l}{2}}^{\frac{l}{2}} \vec{E}(r) \cdot \hat{e}_r r d\phi dz = 2\pi l r E(r) = \frac{Q}{\epsilon_0}, \quad (2.6)$$

where the charge is given by

$$Q = \int_0^r \int_0^{2\pi} \int_{-\frac{l}{2}}^{\frac{l}{2}} \rho_{\text{wire}} \delta^2(r') + \rho_{\text{ion}} r' dr' d\phi dz = 2\pi l \rho_{\text{wire}} + \pi r^2 l \rho_{\text{ion}},$$

with the line charge density ρ_{wire} on the anode wire and the radiation induced ion space charge density ρ_{ion} added.

The effective line charge density ρ_{wire} on the anode wire can be determined from the boundary condition $U = \int_{r_{\min}}^{r_{\max}} E(r)dr$, which leads to

$$E(r) = \frac{U - \frac{\rho_{\text{ion}}}{4\epsilon_0}(r_{\max}^2 - r_{\min}^2)}{\ln \frac{r_{\max}}{r_{\min}}} \cdot \frac{1}{r} + \frac{\rho_{\text{ion}}}{2\epsilon_0} \cdot r. \quad (2.7)$$

Comparing the electric fields with (Equation (2.7)) and without (Equation (2.1)) space charge, one can see that at high rates the high voltage is effectively reduced by the amount $\frac{\rho_{\text{ion}}}{4\epsilon_0}(r_{\max}^2 - r_{\min}^2)$. The resulting gas gain drop reduces the signal amplitude which deteriorates the tube drift time and spatial resolution and efficiency.

The additional term, which is linear in r , modifies the field mainly at large drift radii. The drift gas in the ATLAS MDT chambers is nonlinear, meaning that the space-to-drift-time relation is not linear due to the dependence of the electron drift velocity on the electric field [19]. Hence, the field modifications affect the electron drift time and track radius is measurements. As these modifications are stochastic, they can just be corrected for on average and, therefore, deteriorate the resolution.

The field variations at high hit rates can be reduced by reducing the tube diameter, which is, however, in conflict with maximizing the drift tube resolution at low counting rates, for which large drift radii are favorable (cf. Section 2.1).

A more detailed examination takes into account ion divergence at larger drift radii and the ion drift velocity dependence on the electric field. The electrical ion current $I = \nu GN_p e$, with the tube hit rate ν and the number of primary ionizations N_p generated by the Townsend avalanches flows with the average ion velocity $v(r) = \mu E(r)$ through a cylindrical surface $2\pi lr$ of length l concentric with the tube at radius r . With the ion charge density $\rho_{\text{ion}}(r)$, the relation

$$\nu GN_p e = 2\pi l r v(r) \rho_{\text{ion}}(r)$$

is obtained.

Instead of assuming constant ion density from Equation (2.4), Gauss's law gives more precisely

$$2\pi r E(r) = \frac{1}{\epsilon_0} \rho_{\text{wire}} + \frac{1}{\epsilon_0} \int_0^r \frac{\nu GN_p e}{\mu_{\text{ion}} E(r')} dr'.$$

Differentiating with respect to r results in

$$(E(r))^2 + r E(r) \frac{dE}{dr}(r) = \frac{\nu N_p G e}{2\pi \mu \epsilon_0}.$$

This differential equation is solved by

$$E(r) = \sqrt{\frac{\nu N_p G e}{2\pi\mu\epsilon_0}} \frac{k}{r} \sqrt{1 + \frac{r^2}{k^2}} \quad (2.8)$$

where the parameter $k > 0$. It has the dimension of a length can be interpreted as a measure for the range of the space charge field modifications. The high voltage leads to the following constraint in k :

$$k \ln \left(\frac{r_{\max}}{r_{\min}} \frac{k + \sqrt{k^2 + r_{\min}^2}}{k + \sqrt{k^2 + r_{\max}^2}} \right) + \sqrt{k^2 + r_{\max}^2} - \sqrt{k^2 + r_{\min}^2} = U \frac{2\pi\mu\epsilon_0}{\nu N_p G e}. \quad (2.9)$$

The Equation (2.9) has to be solved numerically to obtain the electric field $E(r)$ in Equation (2.8). For given $r_{\min}$ and $r_{\max}$, applied high voltage U and ion mobility μ , k depends on the hit rate ν , the number of primary ionizations N_p and the resulting gas gain G . As G depends on the electric field (cf. Equation (2.3)), the calibration has to be iterated.

2.3 Performance of MDT Detectors under γ -Irradiation

The effect of ion space charge on the spatial resolution 30 mm diameter MDT from measurements [20] in the CERN Gamma Irradiation Facility GIF is shown in Figure 2.3. The space charge is created in Townsend avalanches near the sense wire originating from the ionization electrons drifting to the wire which have been generated by the 662 keV photons from the γ source mostly in the tube walls but also in the gas. At HL-LHC luminosities, ion space charge is non-negligible in most MDT chambers [15]. The resulting drop of the gas amplification deteriorates the resolution especially at small drift radii [20]. At large drift radii $r \gtrsim 6$ mm the resolution suffers from drift time modifications caused by ion space charge fluctuations due to the stochastic nature of the γ interactions.

In order to overcome the resolution deterioration at high rates, small-diameter **M**onitored **D**rift **T**ubes with reduced outer diameter of 15 mm have been developed [23] (cf. Table 2.1 and Figure 2.4).

Ar/CO₂ (93/7) gas has a nonlinear radius-to-drift time relationship due to the dependence of the electron drift velocity on the electric field [19] (see Figure 2.6). While the maximum drift time in 30 mm diameter MDT is about 700 ns, it is reduced to about 190 ns in sMDT (cf. Figure 2.5), where the radius-to-drift time relation becomes more linear and the drift velocity largely independent of the space charge.

Due to the smaller maximum drift radius of $r_{\max} = 7.1$ mm of the sMDT, the effect of space charge fluctuations is essentially eliminated [25] (cf. Figure 2.3). The space charge density and, therefore, the

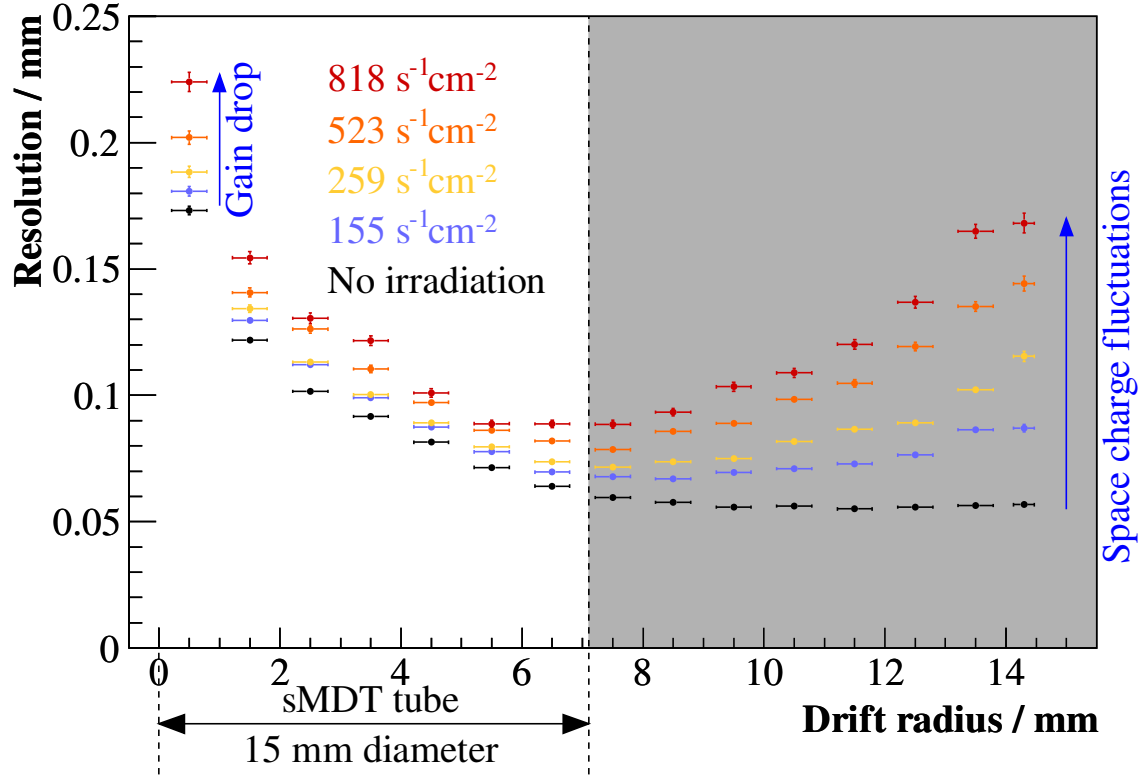


Figure 2.3: Radial dependence of the resolution of MDT with 30 mm diameter for different γ conversion fluxes [20, 21]. With increasing flux, the ion space charge density increases modifying the electric field. The resulting field reduction at the wire leads to gas gain drop, reducing the pulse amplitude and leads to resolution degradation especially near the wire. Fluctuations of the space charge modify the electron drift velocity leading to deterioration of the resolution increasing with the drift radius.

Table 2.1: Comparison of properties of ATLAS drift tube detectors with 30 (MDT) and 15 mm (sMDT) outer tube diameter. Both chamber types have 0.4 mm thick aluminum tube walls and gold-plated W/Re (97/3) wire with a diameter of $50 \mu\text{m}$ and are operated with Ar/CO₂ (93/7) gas at 300 kPa absolute pressure. The nominal high voltage is chosen to result in a gas gain of $G = 2 \cdot 10^4$ according to Equation (2.1).

	MDT	sMDT
Max. drift radius r_{max}	14.6 mm	7.1 mm
Wire pitch	30.1 mm	15.1 mm
Geometric acceptance	97 %	94 %
Maximum electron drift time	700 ns	190 ns
Nominal operating voltage U	3 080 V	2 730 V



Figure 2.4: MDT (left) and a sMDT (right) with their endplugs signal caps [22].

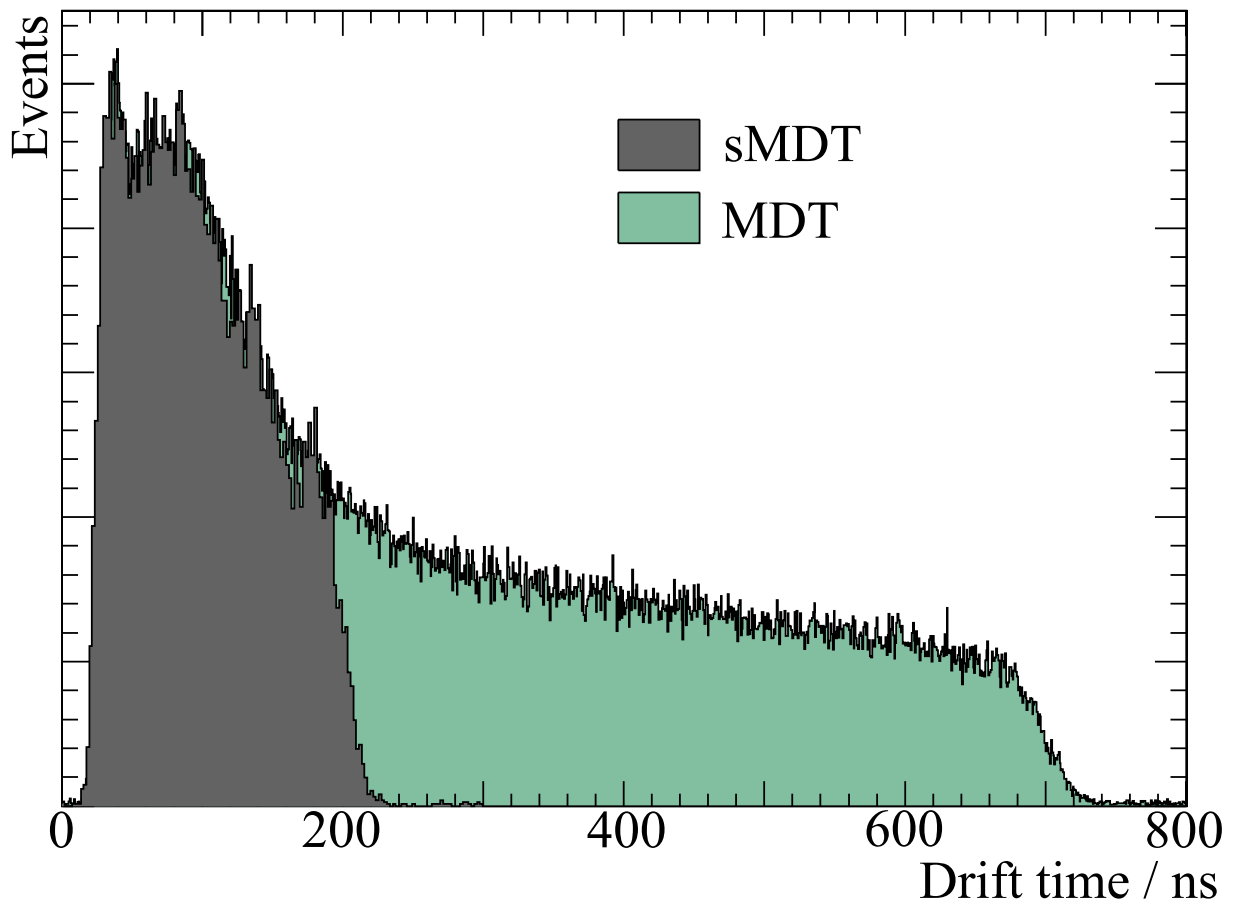


Figure 2.5: Comparison of the drift time spectra of drift tubes with 30mm (MDT) and 15mm diameter (sMDT) [20, 24].

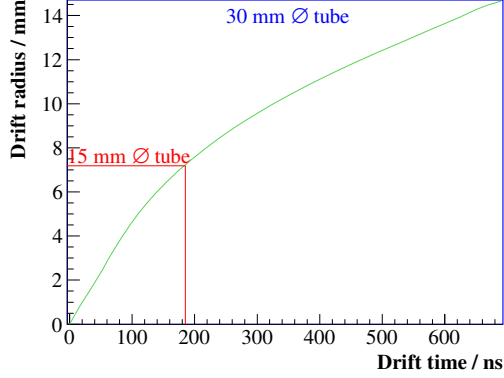


Figure 2.6: Radius-to-drift time relationship for ATLAS drift tubes. The maximum electron drift time in 30 mm diameter tubes is 700 ns and approximately 190 ns in 15 mm diameter tubes (cf. Chapter 7).

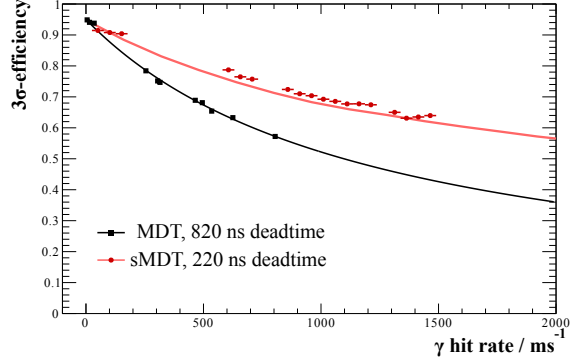


Figure 2.7: 3σ muon detection efficiency of drift tubes with 820 ns (MDT) and 220 ns (sMDT) minimum readout electronics deadtime [8, 15].

gain drop are also strongly reduced due to the reduction of the maximum ion drift time from 4 ms in MDTs to 1 ms in sMDTs (cf. Equation (2.5)). According to Equation (2.7), the gain drop for a given tube hit rate ν is proportional to $(r_{\max}^2 - r_{\min}^2)$. For given γ flux, the conversion rate in the tube walls ν is proportional to the tube diameter causing the gain drop to be proportional to $r_{\max}^3$. The reduction of $r_{\max}$ from 7.1 mm to 14.6 mm consequently reduces the gain drop effects by almost an order of magnitude, due to the reduction of $r_{\max}$ from 7.1 mm to 14.6 mm.

The effect of the tube background occupancy on the muon detection efficiency depends on the electronics deadtime. Frequently, but not necessarily, for larger maximum drift time a larger electronics deadtime of similar magnitude is chosen to suppress afterpulses of the signals and save data transfer bandwidth. Two examples of the effect of electronics deadtime on the muon efficiency, defined as the so-called 3σ efficiency that the measured tube hit is within three times the tube resolution of the reconstructed muon track, depending on the background counting rate are shown in Figure 2.7.

Due to the reduced tube cross section of sMDT compared to MDT with 30 mm diameter, the tube diameter reduction reduces the efficiency loss caused by the tube deadtime by one order of magnitude for a constant flux of primary particles.

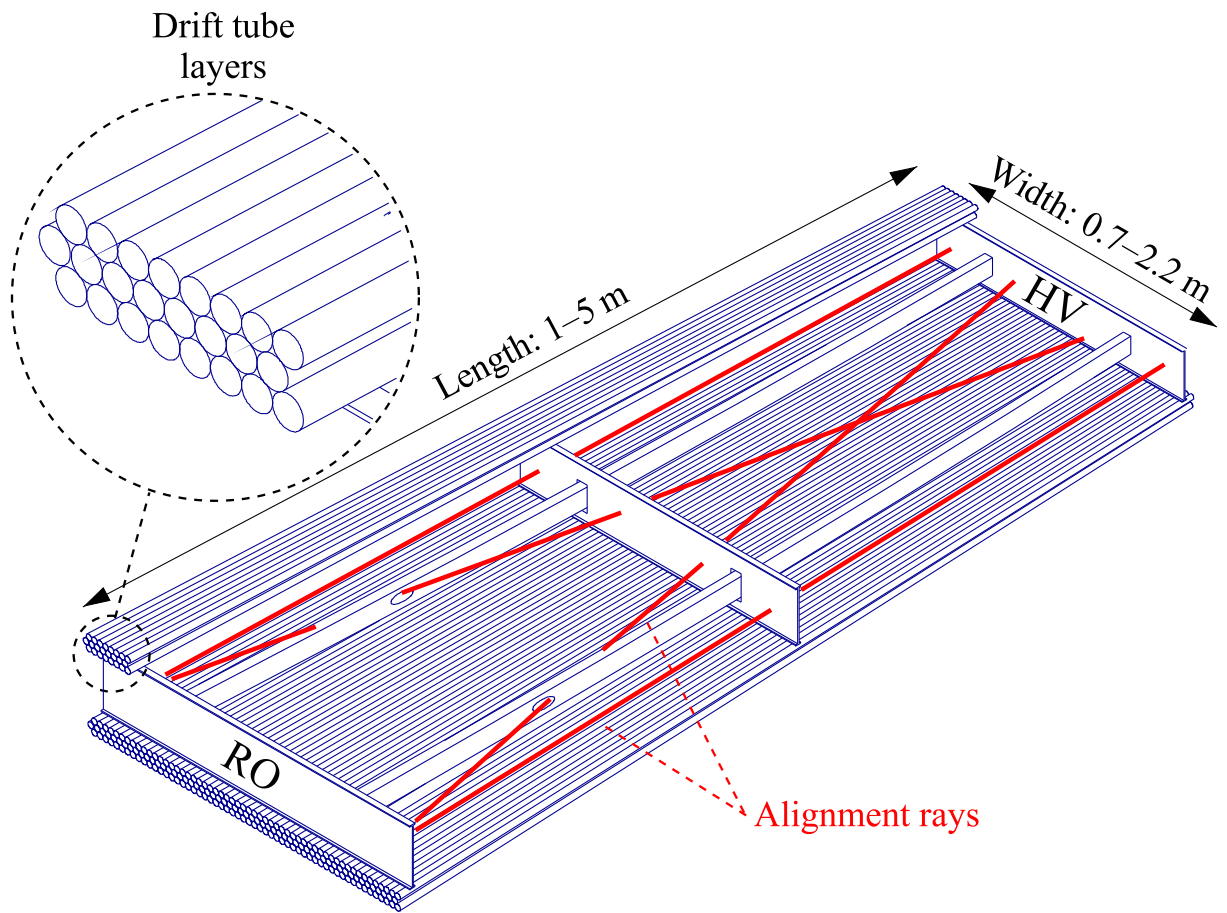


Figure 2.8: Schematic layout of the ATLAS MDT chambers in the barrel region of the Muon Spectrometer [8].

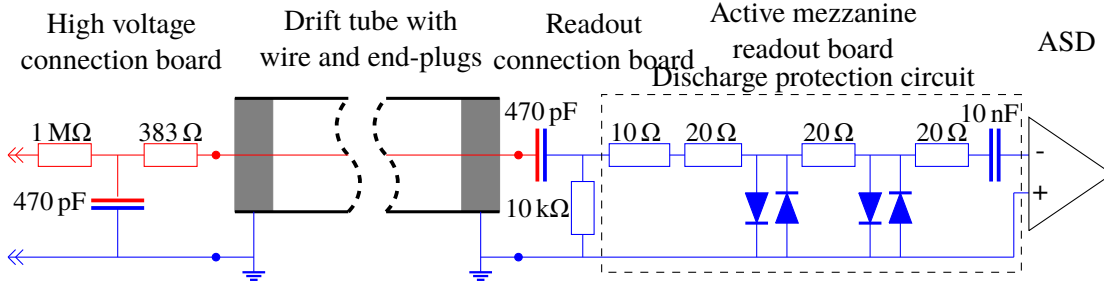


Figure 2.9: Schematics of the front-end electronics of an ATLAS MDT [8]. The dots symbolize the pin connectors on both ends which are gold-plated to ensure permanent reliable low-impedance contact.

2.4 Design of the ATLAS Drift Tube Chambers

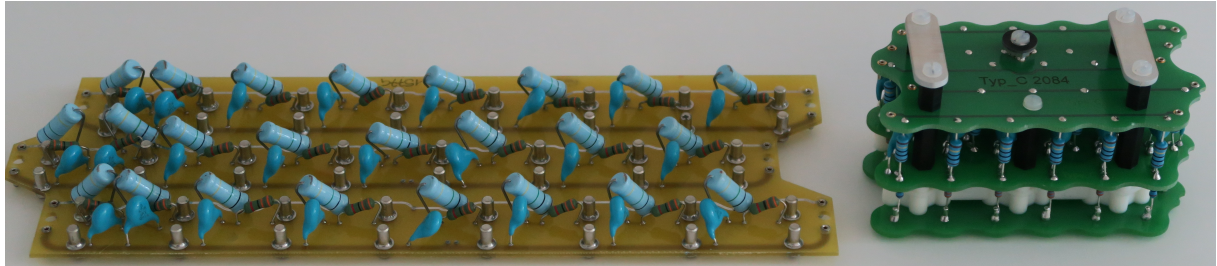
The ATLAS MDT chambers are composed of two multilayers with three or four layers of tubes each which are separated by a support structure (cf. Figure 2.8). The support structure consists of two long beams parallel to the drift tubes and three perpendicular cross plates. The optical in-plane planarity monitoring system with RasNik¹ sensors embedded into the supporting structure measures chamber deformations as a consequence of mechanical stress, temperature changes and gravity. Optical sensors are also mounted on the multilayers to measure the relative chamber positions within the ATLAS muon detector.

The MDT chambers in the barrel region of the muon spectrometer have a rectangular shape. The two muon system end-caps are equipped with MDT chambers of trapezoidal shape. Although the tube length increases with the distance from the beam axis, the innermost muon chamber layer has the highest occupancy. In order to improve the track reconstruction efficiency, the inner barrel chambers consist of four tube layers per multilayer, while the middle and outer chambers consist of three tube layers per multilayer. Particle hits in the aluminum tube walls cannot be detected. The resulting geometrical acceptance is 97 % for 30 mm and 94 % for 15 mm tube diameter. The efficiency loss per layer is compensated by at least three tube layers per multilayer.

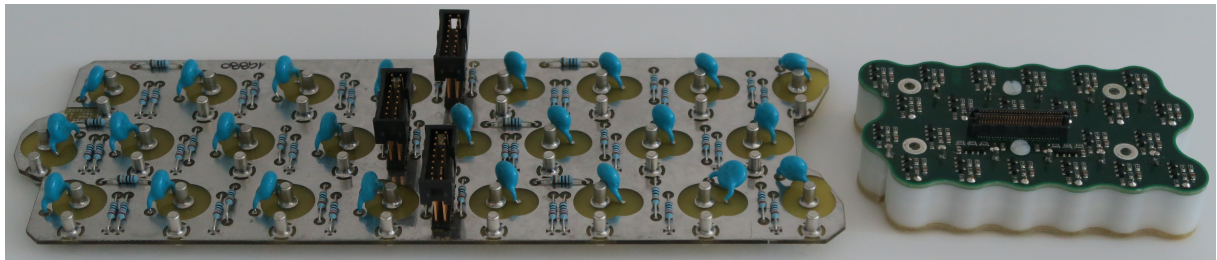
During operation, the tubes are continuously flushed with the Ar/CO₂ (93/7) gas mixture with one volume exchange per day to keep the impurities low. In order to optimize the efficiency and resolution, the gas pressure is kept at $p = 300$ kPa, as a higher gas density increases the ionization rate, reduces ionization fluctuations and electron diffusion [27]. The drift tubes are sealed by endplugs on both ends (cf. Section 3.1.1).

Figure 2.9 shows a schematic view of the electronic components connected to a MDT. The high voltage supply and signal readout boards are connected on opposite tube ends, defining the **ReadOut** and **High**

¹Red alignment system Nikhef [26]



(a) High voltage supply boards.



(b) High voltage readout boards.

Figure 2.10: Photographs of the high voltage boards for ATLAS drift tube chambers, left for the multilayers of 2x3-layer MDT chambers with 30 mm diameter tubes (3x8 channels) and right for the multilayers of 2x4 layer sMDT chambers with 15 mm diameter tubes (4x6 channels). The high voltage coupling capacitors for the sMDT readout are enclosed in plastic containers to ensure high-voltage stability.

Voltage side of the chamber (cf. Figure 2.8). The active front-end electronic and the HV distribution boards group 24 tubes to simplify the readout.

The MDT and sMDT chambers are supplied with high voltage independently for each layer (cf. Figure 2.10a) with high voltage supply boards for groups of 24 tubes. A $383\,\Omega$ termination resistor matching the tube impedance prevents signal reflections, and a low-pass filter consisting of a $0.47\,\text{nF}$ capacitor and a $1\,\text{M}\Omega$ resistor reduces noise and limits the high voltage current in case of a short circuit, e.g. as a consequence of a broken wire. Additional noise filters for each tube layer, composed of a $25\,\text{k}\Omega$ resistor each in the high voltage and the ground supply line and two $2.2\,\text{nF}$ capacitors, are installed on the chamber.

On the readout side, the high voltage readout boards connect to the same group of 24 tubes as the high voltage supply boards, requiring a mirrored geometry of the boards (cf. Figure 2.10b). The wire potential is decoupled with an CR-circuit consisting of a $0.47\,\text{nF}$ capacitor and a $10\,\text{k}\Omega$ resistor per channel. A protection circuit prevents damage of the ASD due to voltage spikes due to discharges. It consists of two stages of diode pairs, which short-circuit large pulses of both polarities to ground. The $10\,\text{nF}$ capacitor at the end decouples the DC offset of the ASD input stage.

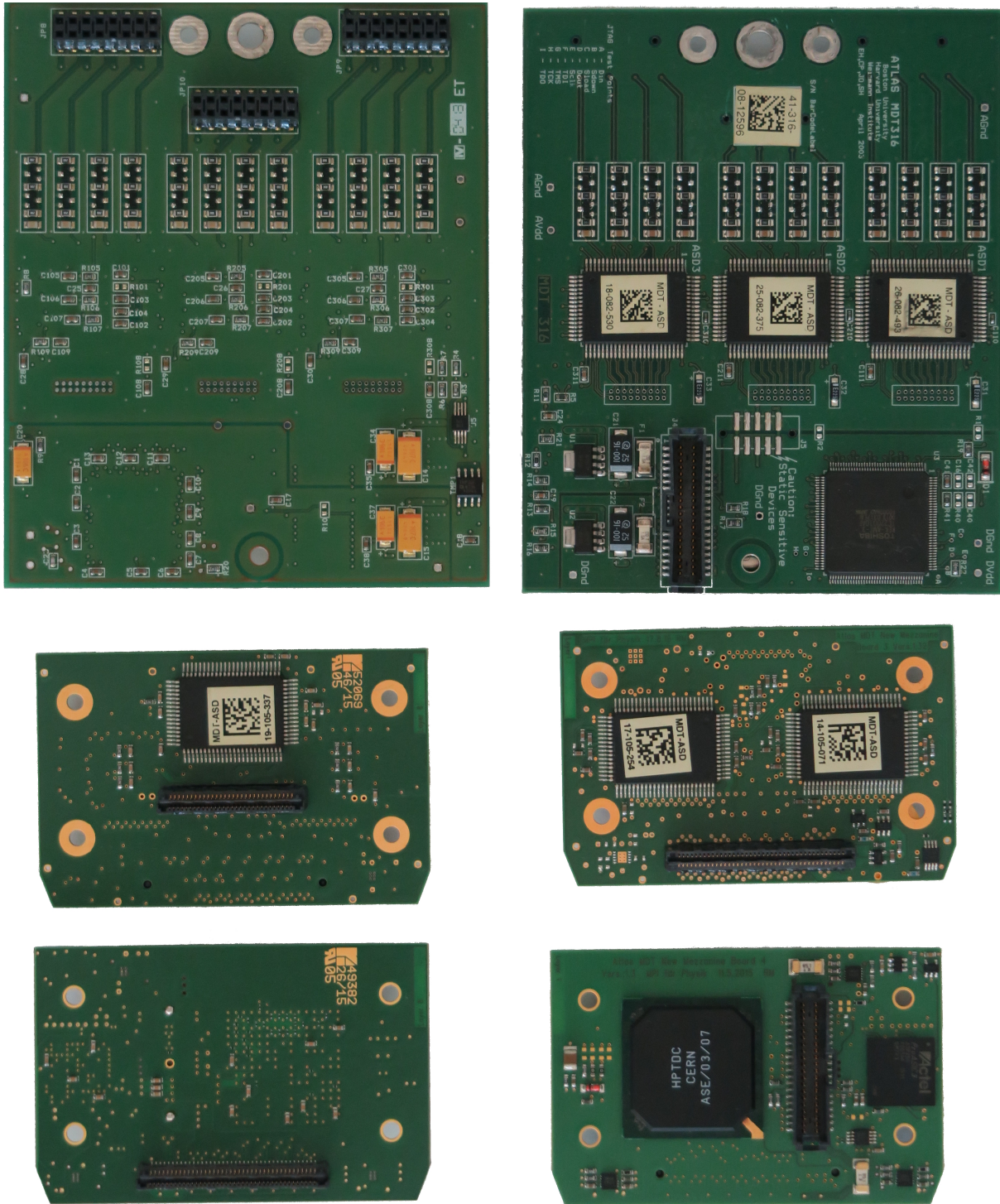


Figure 2.11: Photograph of the back (left) and front (right) sides of the MDT mezzanine cards (top row), and the ASD (middle) and TDC (bottom) sMDT mezzanine cards. Each ASD chip is labeled with a matrix bar code and serial number. For the new sMDT chamber readout, the MDT AMT chip has been replaced by the HPTDC [28] with a FPGA for configuration.

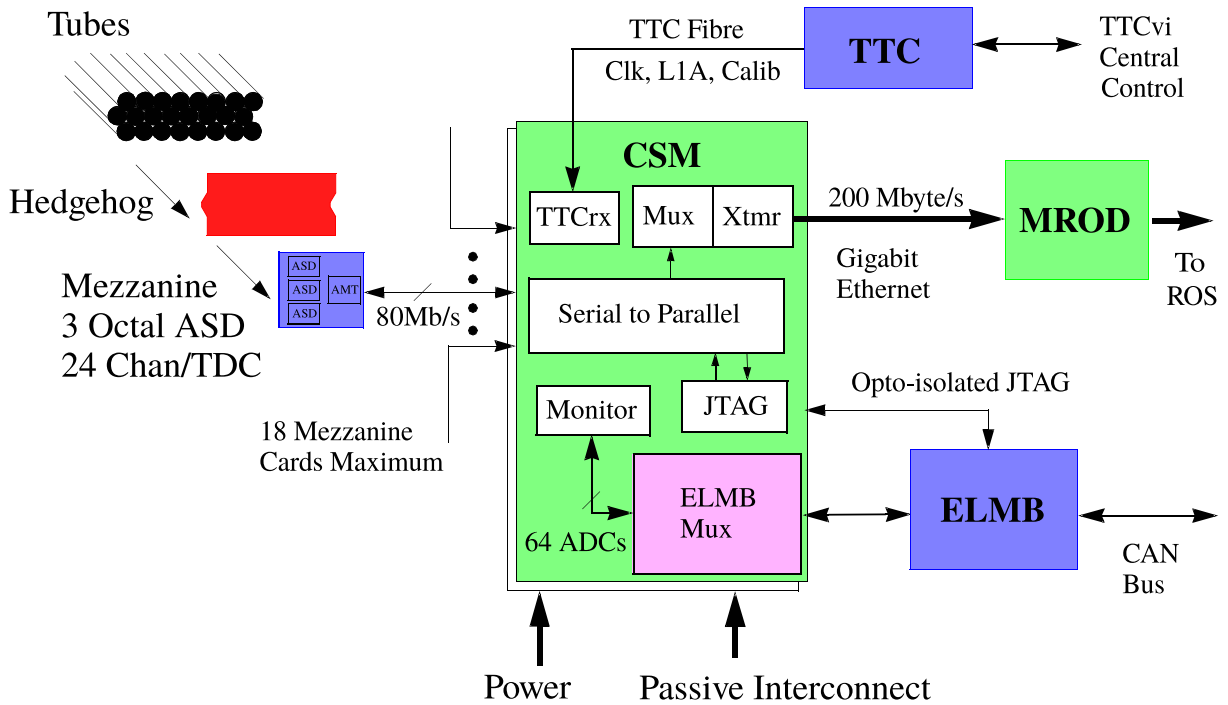


Figure 2.12: Block diagram of the MDT readout chain [29]. The Chamber Service Module (CSM), controls and reads out up to 18 mezzanine boards and supplies them with low voltage and sends the multiplexed data to the ATLAS readout system.

Each high voltage readout board carries a mezzanine board (see Figure 2.11) with three octal Amplifier/Shaper/Discriminator chips and one common Time-to-Digital Converter chip, which are discussed in detail in Chapter 4. Due to the four times higher channel density of sMDT chambers compared to MDT chambers, the mezzanine board is split into an ASD and a TDC board which are stacked. The protection circuits for the sMDT readout electronics are placed on the high voltage readout cards, while it was located on the mezzanine boards for the 30 mm tube diameter chambers except for the loaded $10\ \Omega$ resistor.

Up to 18 mezzanine boards are supplied with low voltage and read out by one Chamber Service Module (cf. Figure 2.12). It multiplexes the 18 TDC output channels and sends the data words via an optical fiber to the readout system. The Timing, Trigger and Control (TTC) system generates the clock and trigger signals, which are distributed to the TDC chips on the optical fiber.

Initialization, control, and temperature sensor readout for each CSM and the upstream electronics is performed by the associated Embedded Local Monitor Board (ELMB). This MDT DCS Module (MDM)

is connected to the central ATLAS Detector Control System (DCS) via CANbus link. The JTAG protocol is used to control the ASD and TDC chips via the CSM and optical filter.

The maximum number of 432 channels per CSM is sufficient for all MDT chambers with 30 mm tube diameter. Two sMDT chambers with eight layers of 78 tubes, installed in 2014 to cover acceptance holes in the middle barrel layer [30], have to be read out by two CSMs. The 16 sMDT chambers under construction for the ends of the barrel inner layer with 600 to 700 tubes each are also read out by two CSMs each (see Chapter 3).

Chapter 3

New sMDT Chambers for the ATLAS Muon Spectrometer

As sMDT chambers with 15 mm diameter tubes are have smaller total height for the same spacer height than the standard ATLAS MDT chambers with 30 mm diameter tubes, it became possible to equip the feet regions of the middle barrel layer of the Muon Spectrometer with twelve new so-called BMG sMDT chambers. They add a third detector layer with precise track measurement, which improves the momentum resolution in the feet regions [31]. In this chapter the construction, test and installation of the BMG chambers and the conclusions from the test for future sMDT chamber design is described.

3.1 Chamber Construction

3.1.1 Tube Production

The basic components of the ATLAS MDT chambers are cylindrical aluminum tubes with $t = 0.4$ mm wall thickness and gold-plated tungsten-rhenium anode wire with $r = 25$ μm radius tensioned to 350 g [32]. The sMDT chambers, like the MDT chambers, are operated with Ar/CO₂ (93/7) gas mixture at 300 kPa absolute pressure. Contaminations of the inner tube wall and the wire can increase the currents between wire and tube wall under high voltage and even lead to discharges. Hence, all chamber components, which are in contact with the gas are carefully cleaned before usage and all production steps until the final sealing of the complete gas system are performed in a temperature stabilized clean room with flow boxes. In order to ensure reliable electric contacts with low impedance to the outside of the aluminum tube walls and to clean the inner tube walls, tube surfaces are etched and chromatized.

A crucial component of the drift tubes are the endplugs, which combine several tasks (see Figure 3.1):

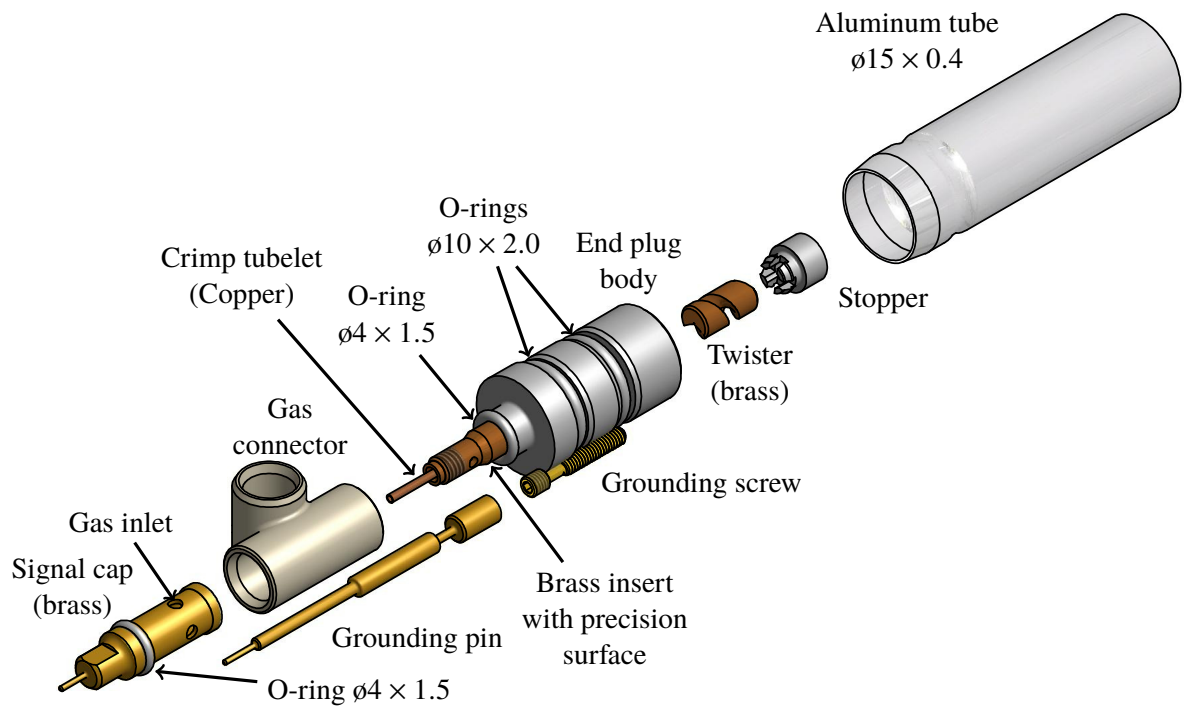


Figure 3.1: Exploded view of a sMDT. All dimensions are in mm. [15, 23]

- Sealing of the gas volume at 300 kPa pressure ensuring very low leakage rate of less then $2 \frac{\text{mPa}\cdot\text{l}}{\text{s}}$ [33] and, therefore, low back diffusion rates of oxygen and nitrogen into the drift gas.
- Gas tight connection to the gas distribution system.
- Positioning of the wire with few micrometer precision.
- Fixating the wire ends under tension.
- Insulation of the wire from the tube wall.
- Electric connections for high voltage supply and signal readout.

Grounding screws with ground pins connect, from the outside, the tube walls of adjacent tubes to the ground plane of the high voltage boards.

The plastics of the insulating injection molded body of the endplugs and of the gas connectors have been carefully chosen to prevent any outgassing which could lead to wire aging of the tubes under the high background irradiation of the LHC and HL-LHC [34]. The body consists of fiber-glass headed Crastin¹, which includes the brass insert with the wire. The brass insert has two radial holes for the gas supply, an axial feedthrough for the wire, a screw thread for the electric connection and gas seal via the signal cap and a cylindrical precisely machined surface connected with the wire on the outside of the tube. On the inside of the tube, the brass insert has a concentric hole into which the helical brass wire locator (twister) is inserted, which has an inner bore of $d = 50 \mu\text{m}$ diameter matching the wire diameter defining the wire position with respect to the outer precision surface with few micrometer precision. The latter is used for precise positioning of the wire in and allows for precise mechanical measurement of the relative wire position in the chamber.

The Pocan² stopper holds the twister in place. O-rings are placed in grooves for gas sealing. The aluminum tubes are mechanically swaged onto them to fix and seal the endplugs. The brass insert also holds on the outside the copper tubelet for crimping of the wire.

Two semi-automated assembly setups are used for the tube production. First (see Figure 3.2), the wire is threaded through the aluminum tube and one prepared endplug by clean airflow, through a vacuum pump connected to the end plug and slight overpressure injected into the tube on its open side. The tube is then mechanically swaged onto the O-rings of the endplug by rotating clamps to generate a lasting mechanical and gas tight connection. Afterwards, the wire is mechanically threaded through the endplug on the other tube end and the copper tubelet is mechanically crimped onto the wire to fix it. The aluminum tube is swaged onto the second endplug to complete the sealing of the tube.

¹A polybutylene terephthalate by DuPont.

²A polybutylene terephthalate by Lanxess.

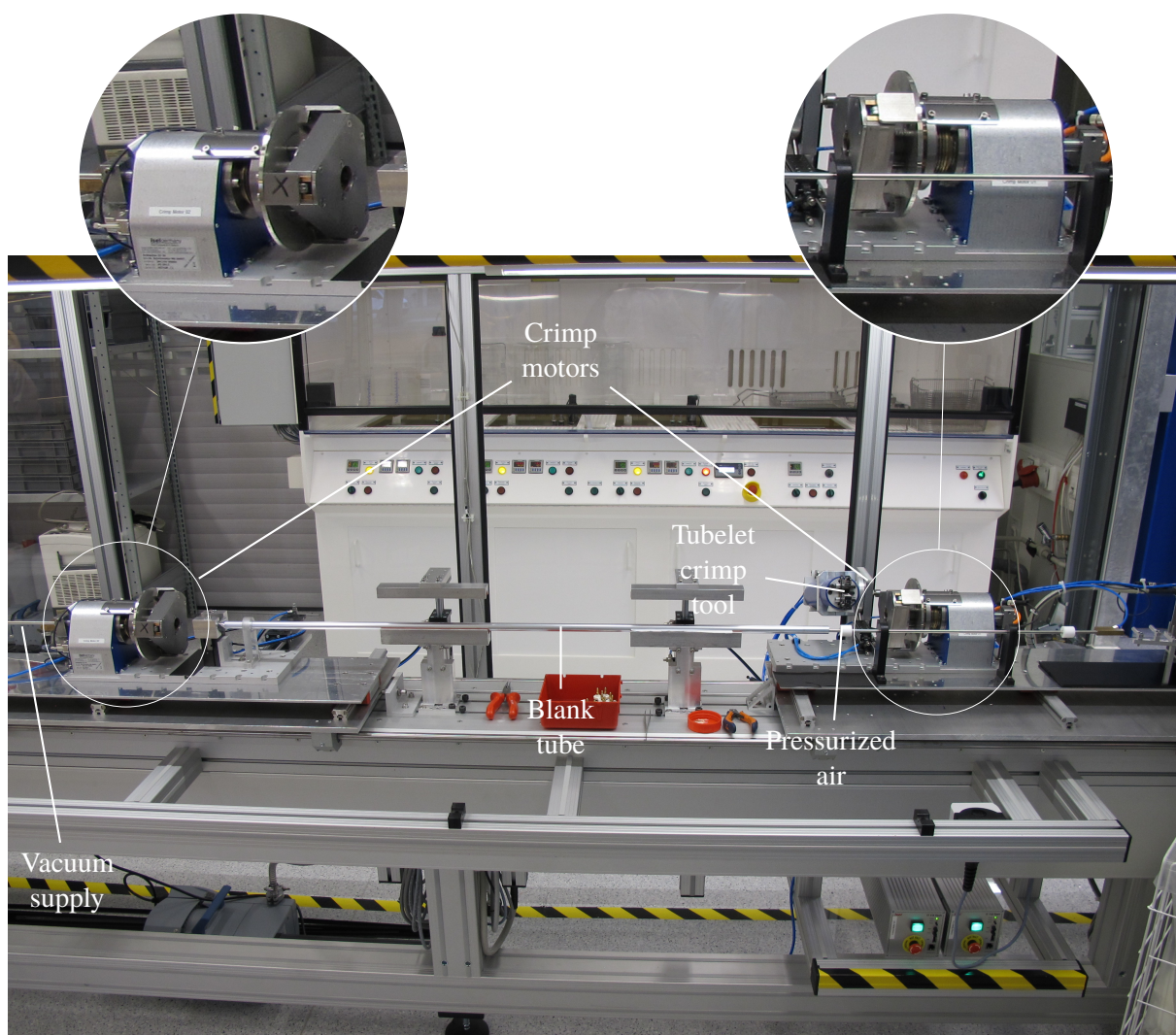


Figure 3.2: Photograph of the wire threading and endplug insertion and swaging machine for sMDT production.

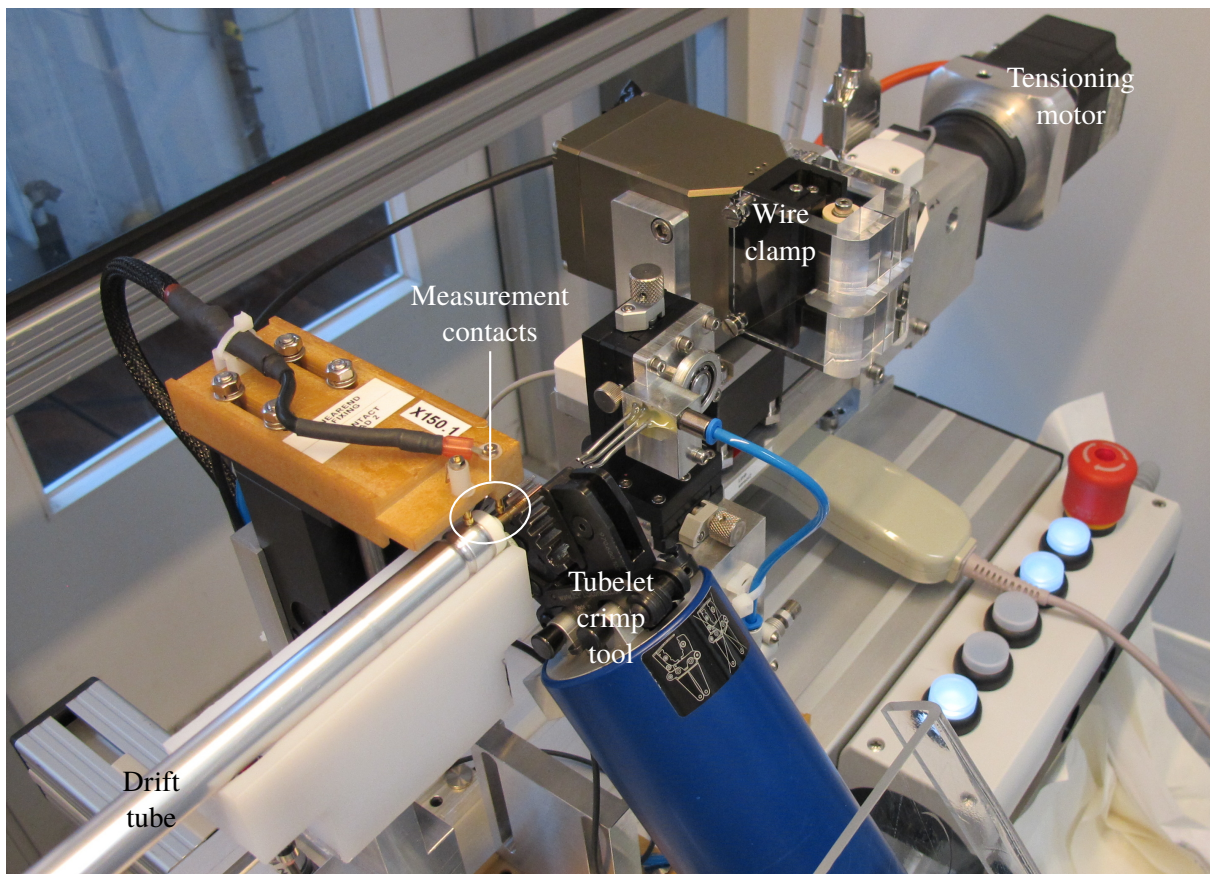


Figure 3.3: Photograph of the wire tensioning and crimping machine for sMDT production. The wire tension is measured with an electric stress sensor.

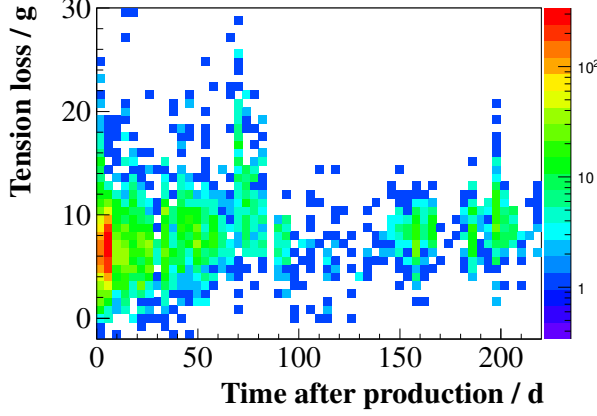


Figure 3.4: Wire tension loss in BMG drift tubes as a function of time.

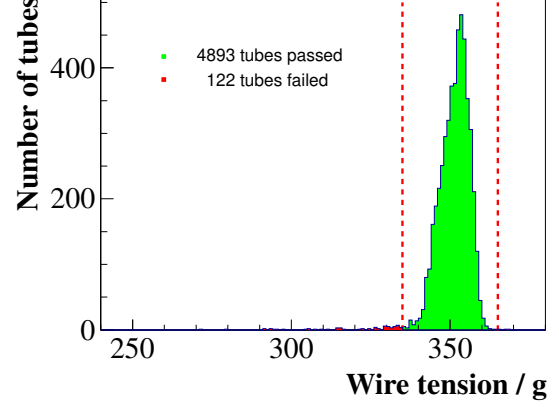


Figure 3.5: Distribution of the wire tension of the BMG drift tubes after the relaxation measured after the first two days after production. The acceptance range is indicated.

The final steps are performed on the tensioning setup (see Figure 3.3). The loose end of the wire is tensioned with a stepping motor controlled by a stress sensor to 350 g. To reduce relaxation of the wire tension over time, overtension to 450 g is applied for 10 s. The wire typically breaks at a tension of 500 g. At 350 g, the wire is crimped in the second endplug.

After the production of the tube, the wire tension T is measured by measuring its fundamental oscillation frequency f according to the relationship

$$T = \frac{\pi \rho d^2 L^2}{g} f^2,$$

where g is gravitational acceleration ρ , d and L are the density, diameter and free length of the wire [35].

The oscillations are excited by an electric current pulse through the wire in the magnetic field of a horseshoe magnet. The wire oscillations in the magnetic field lead to an alternating current of frequency f , which is measured.

Relaxation of the wire leads to tension loss of about 8 g within the first day to the nominal value of 350 g. The wire tension remains stable afterwards (cf. Figure 3.4). The wire tension is measured immediately after the production and again after at least two days. All tubes with a tension loss $\Delta T > 18$ g are rejected, because such a large loss in wire tension could indicate faulty wire crimping and result in a complete loss of tension. The wire sags under gravitation, which changes the distance to the tube wall and hence the electric field in the tube. All tubes with final wire tension $T < 335$ g therefore are rejected (cf.

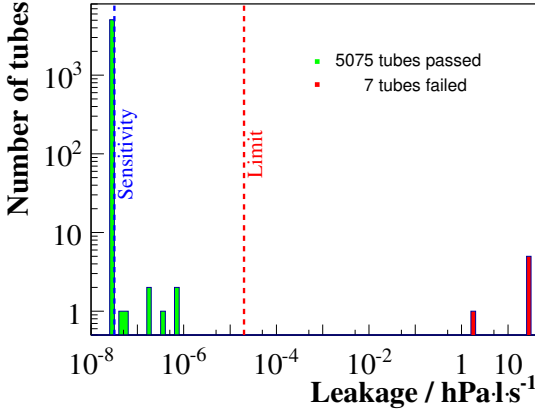


Figure 3.6: Distribution of the Ar gas leak rate of the BMG drift tubes at 300 kPa pressure.

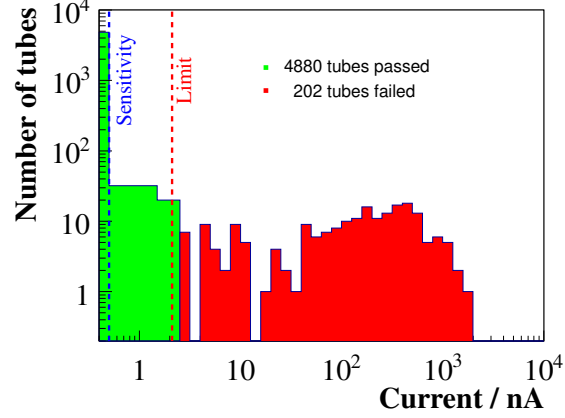


Figure 3.7: Distribution of the dark current of the BMG drift tubes at nominal operating conditions.

Figure 3.5). Tubes with wire tension $T > 365$ g are rejected to reduce the risk of breaking of the wire. Due to frequent readjustments of the tensioning machine, the wire tension distribution is very narrow around the nominal value of $T \approx 350$ g.

The gas tightness of the tubes is measured with a helium leak detector. The tube is sealed gas tight at one end and then inserted into a test tube which is sealed with a cap with feedthrough for the gas supply to the drift tube. The drift tube is then evacuated through its other end plug and filled with a Ar/He (95/5) gas mixture at 200 kPa absolute pressure. The test tube is evacuated within a few seconds with the leak detector, which detects helium leaking from the drift tube. The pressure difference between the drift tube and the surrounding vacuum equals the operating pressure. The Helium leak measurement only has to be corrected for the difference in the diffusion constants of helium and argon. The sensitivity of the setup is $3.2 \cdot 10^{-8} \frac{\text{hPa} \cdot \text{l}}{\text{s}}$, which is well below the maximum allowed leak rate of $2 \cdot 10^{-5} \frac{\text{hPa} \cdot \text{l}}{\text{s}}$ per tube. The leak rate of the vast majority of the tubes is below the sensitivity (cf. Figure 3.6). The leak rates of the few drift tubes, which do not meet the quality requirement, are several orders of magnitude above the limit and easily detectable.

Finally, the high voltage stability and dark current is tested. The tubes are evacuated and filled with the operating gas mixture. 3.01 kV are applied between wire and tube wall, i.e. the nominal voltage of 2.73 kV plus a safety margin of 10 %. The vast majority of tubes has a dark current below the sensitivity of the high voltage power supply of 0.5 nA (cf. Figure 3.7). Unwanted dust particles inside the tube volume can lead to dark currents of up to about 2 μA . Dark currents of up to 2 nA per meter is accepted, which

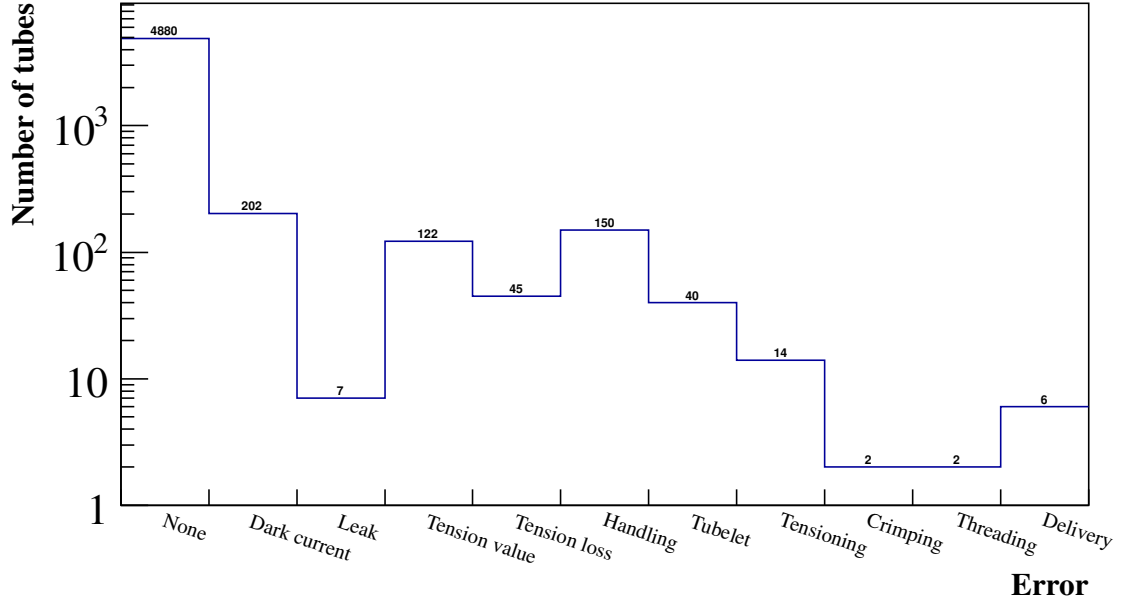


Figure 3.8: Number of the different failures during drift tube production.

corresponds to a hit rate of $3 \text{ s}^{-1} \text{ cm}^{-2}$ which is in the order of the cosmic muon hit rate. The BMG drift tube length is 1.1 m.

4800 drift tubes were needed for the BMG chambers including 3 % spares [36]. An overview of the tube production errors is shown in Figure 3.8. 202 tubes were rejected due to too high dark current, which is the most frequent cause of failure. Improved cleaning of the components, mainly cleaning the inside tube surfaces with ethanol soaked lint free paper towels before production, reduced the losses. 150 tubes were damaged during handling and assembly, which occurred especially during the beginning of the production. This loss was reduced by optimization of the production processes to well below 3 %. The number of cases of wrong wire tension was kept at 122 by monitoring the settings of the tensioning machine. Defects of crimp tubelets were detected by optical inspection of all tubes after production. Gas leaks, threading, crimping and tensioning errors and tube delivery defects were all on the per-mille level, demonstrating the reliability of the tube assembly machines. In total 4880 out of 5470 tubes were accepted for chamber construction, corresponding to an overall failure rate of 10 %, which was significantly reduced during the tube production, especially the failure rate due to handling, wrong wire tension and crimp tubelet defects, as well as dark current.

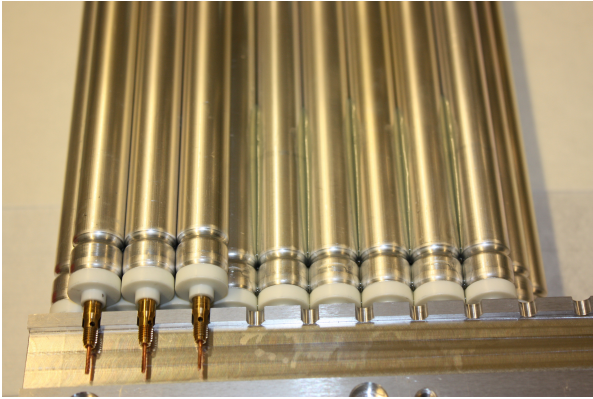


Figure 3.9: Photograph of endplugs placed in the assembly combs. The uppermost tube layer is fixated by the next comb layer.

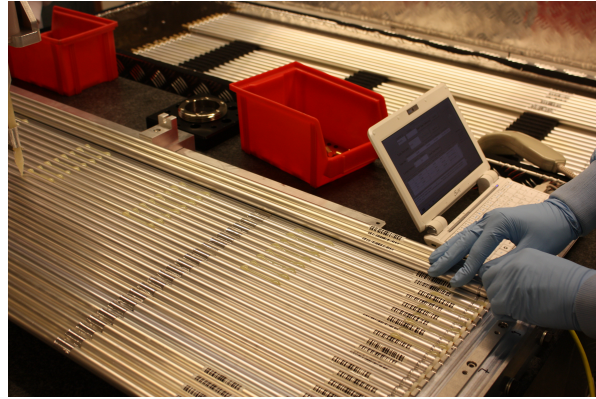


Figure 3.10: Chamber gluing. The glue is dispensed in short strips between two adjacent tubes of a layer by an automated glue dispenser. Then the tubes of the next layer are placed on the glue. The combs are removed horizontally after the curing of the glue.

3.1.2 Chamber Assembly

The chamber assembly is performed tube layer by tube layer. The tubes are placed in aluminum combs at both ends with bores which hold the cylindrical precision surfaces of the endplugs. The combs are stacked for each tube layer. Glue is dispensed between every two adjacent tubes and the tubes of the next layer are placed on it shifted by half a tube diameter (see Figure 3.10). Immediately after the completion of the layer, all its endplugs are fixed by the next comb layer to ensure precise positioning of the wires during curing of the glue. Grounding screws are screwed into the holes between adjacent endplugs of layers 1 and 2 as well as 3 and 4 of each multilayer for low impedance ground connections. For the BMG chambers there are no grounding screws between layers 2 and 3. After the completion of the first multilayer, the spacer structure is glued onto the fourth tube layer and the chamber gluing is continued with multilayer 2. The combs are removed after the curing of the glue of the last tube layer after 24 hours. A wire positioning accuracy of $5\ \mu\text{m}$ RMS has been achieved in the BMG chambers.

The chamber positions in the ATLAS Muon spectrometer require cutouts in all chambers (see Figure 3.11) to allow for the passage of the light rays of the optical alignment monitoring system. Due to different positions and angles of the alignment rays of the twelve BMG chambers, there are eight different types [37]. The BMG chamber have no in-plane alignment system, because they are rigid enough.

After installation of the gas distribution system, the active chamber volume is sealed so that the chamber can be taken out of the clean room. After the temperature sensors and the Faraday cage have been installed, a protective cover is glued onto the top multilayer. The high voltage distribution box, the

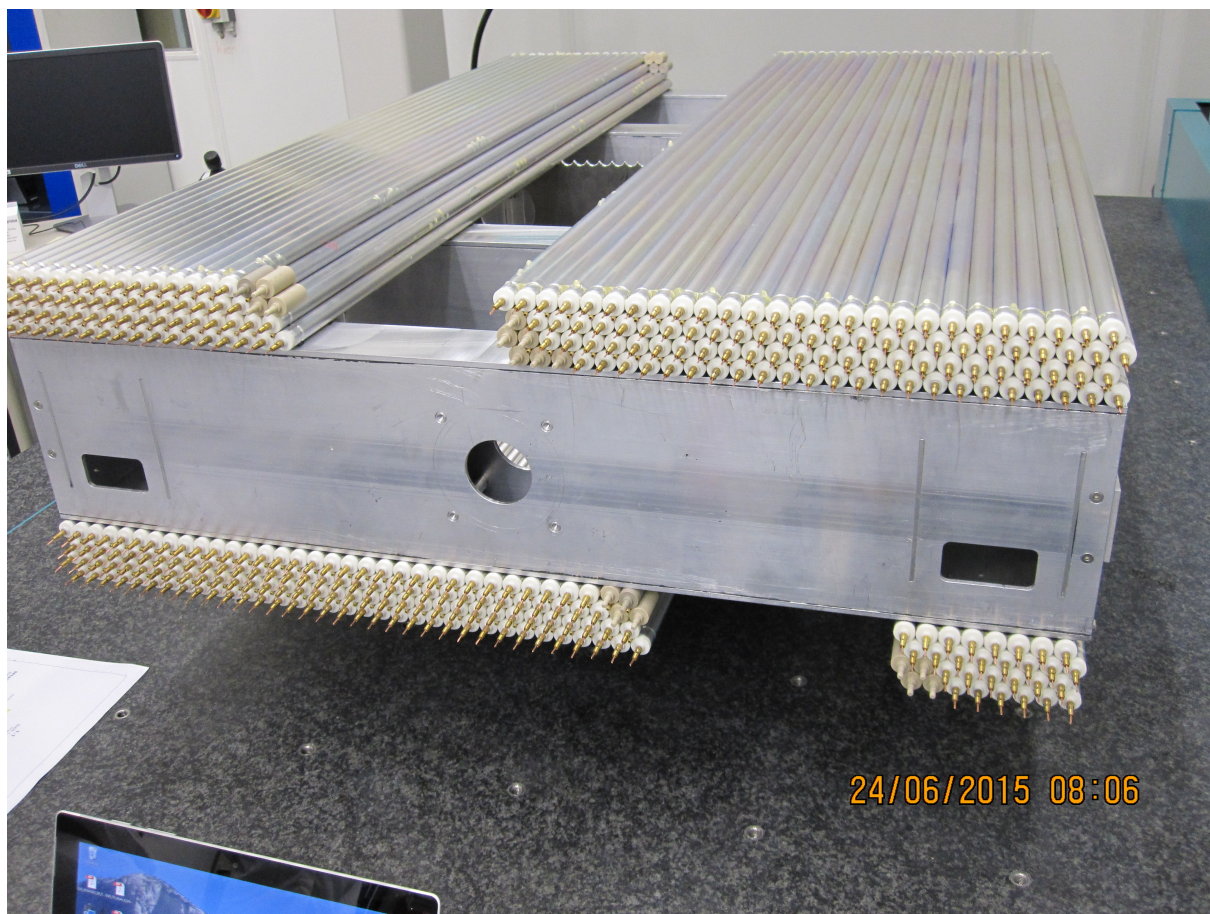


Figure 3.11: Photograph of the BMG-6C12 sMDT chamber after gluing. The cutouts allow for passage of the light rays of the projective alignment system of the muon spectrometer.

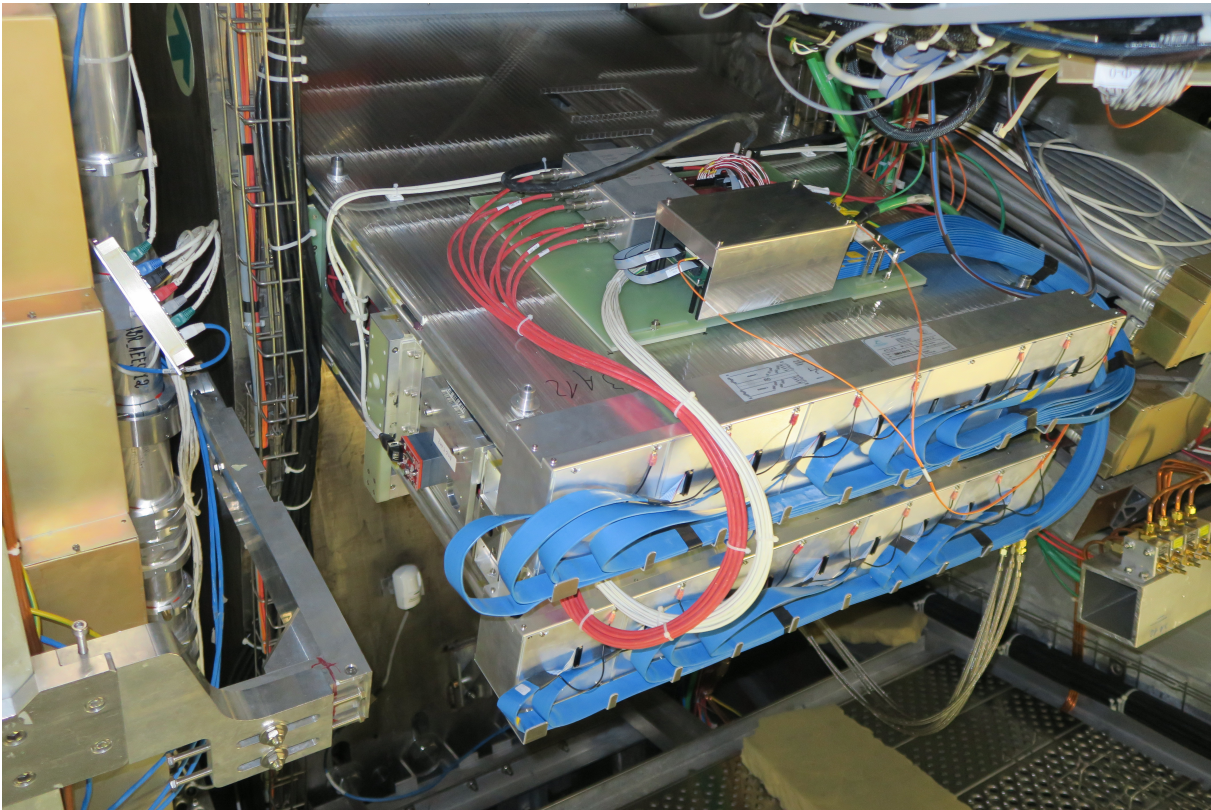


Figure 3.12: Photograph of the BMG-6A12 chamber installed in ATLAS. Each tube layer is supplied with high voltage by one of the red cables. The white cables connect the temperature sensors to the DCS. The blue signal cables between the mezzanine boards and the CSM (under a protective aluminum cover) are for low-voltage power supply and data readout.

Table 3.1: Summary of the test results values of the BMG chambers. The tube hit efficiency does not include the geometric acceptance of 94 %. The drift tube resolution is not corrected for time-slewing of the drift time measurement and for multiple scattering which contributes on average $110 \mu\text{m}$.

Chamber BMG	Pressure loss [$\frac{\text{mPa}}{\text{s}}$]		Dark current [nA]		Avg. noise rate [ms^{-1}]	Drift tube resolution [μm]	Tube hit efficiency [%]
	ML1	ML2	ML1	ML2			
2A12	1	8	7	25	2.0 ± 0.6	192 ± 5	98.8 ± 0.6
2A14	184	77	16	19	2.9 ± 1.1	196 ± 6	98.7 ± 0.4
2C12	167	4	20	20	4.0 ± 1.5	221 ± 3	98.1 ± 0.6
2C14	46	83	10	10	2.3 ± 0.5	202 ± 7	98.7 ± 0.6
4A12	58	42	10	10	9 ± 5	203 ± 8	98.5 ± 0.4
4A14	14	2	40	25	2.7 ± 0.7	260 ± 22	98.3 ± 1.6
4C12	38	28	22	20	1.8 ± 0.3	273 ± 27	95.8 ± 2.3
4C14	67	76	25	10	2.2 ± 0.4	171 ± 2	98.8 ± 1.2
6A12	135	176	10	7	2.7 ± 0.5	191 ± 5	98.7 ± 0.8
6A14	123	16	17	16	3.4 ± 1.1	205 ± 8	98.3 ± 0.5
6C12	78	138	10	10	2.0 ± 0.6	196 ± 6	98.3 ± 0.6
6C14	36	36	14	23	13 ± 11	197 ± 6	98.9 ± 0.7

CSM and the DCS box are mounted on it. High voltage supply and signal readout boards are mounted on the signal caps before the Faraday cage is closed and the high voltage and signal cables are installed. A complete chamber is shown in Figure 3.12.

3.2 Chamber Testing

Before the installation of the twelve BMG chambers in ATLAS, each chamber was tested, to correct operation of all components. Gas tightness, HV dark current, electronics noise and drift tube resolution and efficiency are determined for each chamber. In this section, the test procedures are described. The test results are summarized in Table 3.1.

3.2.1 Gas Tightness

To determine the gas leakage, each chamber was filled with the nominal Ar/CO₂ (93/7) gas mixture to 300 kPa and the temperature sensors were read out. The gas valves were then closed to determine the pressure loss of the chamber over a time period of at least seven hours. The gas connection to the two multilayers were opened again separately to measure the pressure drops in each. Simultaneously the



(a) An opened BMG Faraday cage. The gas distribution system is hidden behind the shielding plates between the high voltage signal readout board and the active ASD board.

(b) Close-up of the edge between two neighboring high voltage boards.

Figure 3.13: Photograph of the BMG read-out electronics.

temperature was read out again to be able to correct for temperature variations. The results are summarized in Table 3.1.

To ensure high purity of the drift gas at a volume exchange per day, a leak rate below $1 \frac{\text{mPa}\cdot\text{l}}{\text{s}}$ per end plug is the specification for MDT chambers in ATLAS. For the BMG chambers, the resulting maximum pressure loss is $0.11 \frac{\text{Pa}}{\text{s}}$. This limit was easily met for each chamber during the first pressure loss measurement, which was performed just after the installation of gas distribution system. As the temperature sensors were not yet mounted at this time, the measurement could not be temperature corrected. Temperature fluctuations of $\Delta T < 0.5 \text{ K}$ in the temperature stabilized clean room for gas system installation limit the measurement precision to $30 \frac{\text{mPa}}{\text{s}}$ after 10 h of measurement, which is precise enough to ensure pressure loss below the ATLAS limit. In the final measurement of the completed chambers equipped with electronics, some of the multilayers had leak rates, which exceeded the measurements in the clean room and the ATLAS specifications by approximately a factor of two. Even though the leaks could not be detected directly due to inaccessibility of the gas connectors behind the electronics boards, mechanical stress due to conflicts between adjacent high voltage and readout boards on the gas connectors to the tubes (see Figure 3.13) was identified as most likely cause. As the BMG chamber gas volumes are relatively

small, the excessive leak rates can be compensated by somewhat higher gas flow. For the future, the mechanical tolerances of the electronics boards have to be more tightly controlled.

3.2.2 High Voltage Test

To disconnect a broken wire from the high voltage supply, the gas system has to be opened to replace the signal cap with an insulator. Hence, the integrity of the wires has been tested before the full installation of the electronics boards at a voltage of only 100 V. A broken wire touching the tube wall leads to a short circuit. If the wire did not break close to the HV side end plug, this is easily detectable by high current, limited only by the 1 M Ω protection resistor on the HV supply board. Otherwise a broken wire is just discovered during the cosmic ray test described in Section 3.2.4 from missing tube hits.

Afterwards the chamber is tested for high voltage stability. The chamber is filled with the nominal gas mixture and the nominal voltage of 2730 V is applied. All BMG multilayers stayed well below the maximum dark current limit of 0.2 μ A per chamber, corresponding to 0.5 nA per tube (cf. Table 3.1).

3.2.3 Accidental Hit Rates

Accidental hit rates (noise hits) increase the readout bandwidth and can reduce drift tube efficiency and resolution, if the discriminator threshold has to be raised. Although the readout electronics has been certified separately, the actual accidental hit rates can only be determined under realistic conditions on the chamber with the complete readout chain, the detailed grounding conditions and with high voltage connected. 10^5 events, recorded with a random trigger provide sufficient statistics for noise rate determination.

Accidental hits occur, when the discriminator input voltage crosses the threshold level due to stochastic voltage fluctuations. Hence, their rate depends on the threshold settings. Figure 3.14 shows the noise hit rate for all BMG chambers at two different threshold settings. The noise rates differ between channels because of actual threshold variations from the chip production, which are rather large for the ATLAS MDT ASD chips produced 15 years ago. As the effective threshold affects the tube performance, all ASD chips were analyzed after production to determine the threshold offset of each channel. These data were not available anymore for the left-over ASD chips used for the BMG chambers, such that the threshold offsets had to be determined by threshold scans. The dependence of the accidental hit rate on the threshold settings was measured for each channel at 2.5 mV hysteresis (cf. Figure 3.15). The threshold voltage is generated by a DAC with a resolution of 2 mV. In order to determine the threshold offset more precisely, a fit (cf. Section 6.2.1) to the five measurement points of the threshold scan was used to interpolate between them and determine the threshold value at which the maximum reference noise rate of 1 ms⁻¹ is reached.

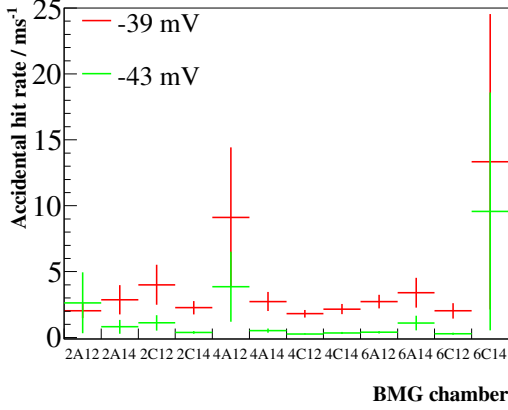


Figure 3.14: Average noise rate per channel for each of the twelve BMG chambers at the nominal threshold of -39 mV and an increased threshold of -43 mV for 2.5 mV hysteresis setting.

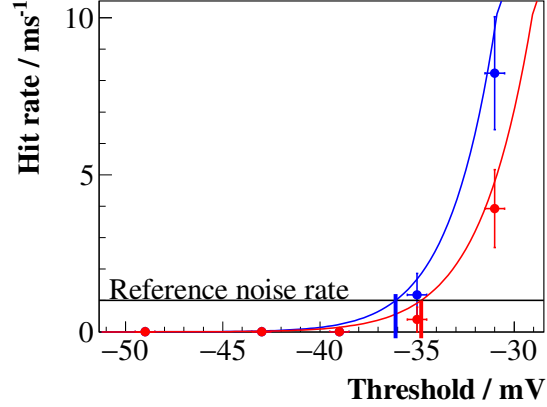


Figure 3.15: Dependence of the noise rate of two channels (red and blue) of the BMG-2A12 chamber on the nominal ASD threshold. A Gaussian fit through the five measurement points was used to determine the threshold at which the acceptable noise rate of 1 ms^{-1} is reached. (The sign of the ASD signal voltage pulse is negative.)

A noise rate of 1 ms^{-1} still leads to negligible occupancy of 0.02 % for a readout time window of the size of the maximum sMDT drift time.

The resulting threshold distribution of all tubes of the chamber BMG-2A12 is shown in Figure 3.16. For the hysteresis setting of 2.5 mV, the mean is at -37.8 mV, which is below the current nominal ATLAS MDT threshold of -39 mV. For the nominal hysteresis setting of 8.75 mV used in ATLAS, the noise rates are expected to still be lower. The reference threshold spread for the BMG-2A12 chamber is $2.3 \pm 0.1 \text{ mV}$, which comes down to the uniformity specifications, as the chips used are supposed to be from the best channel uniformity class, which is defined by a maximum threshold spread between the channels of 2 mV. Figure 3.17 shows that the reference threshold distribution mean and spread are very uniform for all BMG chambers.

3.2.4 Cosmic Ray Test

All 12 BMG chamber were tested with cosmic ray muons. The setup allowed for the measurement of the spatial resolution, efficiency and drift time spectrum of all drift tubes. Missing high voltage connections, faulty mezzanine boards and wrongly plugged readout cables could be detected and fixed before installation of the chambers.

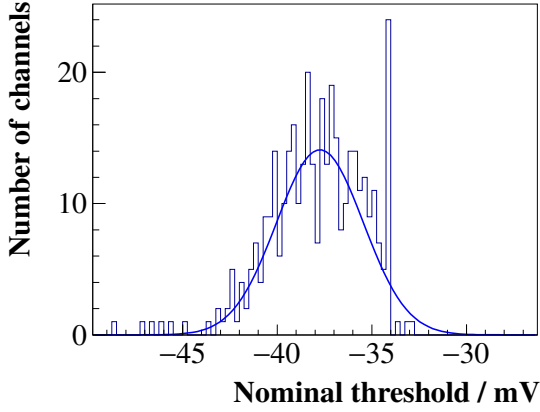


Figure 3.16: Distribution of the reference threshold of the readout channels of the chamber BMG-2A12 with mean at -37.8 mV and spread of 2.3 mV from a Gaussian fit.

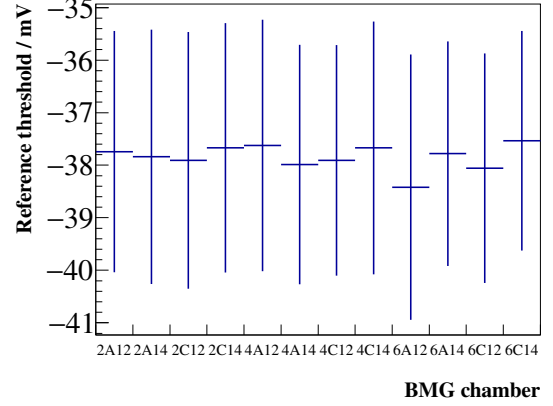


Figure 3.17: Mean value and spread of the reference threshold distributions of the BMG chambers corresponding to the reference noise rate of 1 ms^{-1} and a hysteresis setting of 2.5 mV. The error bars show the spread (standard deviation) of the Gaussian fit to the threshold distribution.

The setup consisted of a gas system with a vacuum pump, pressure controller and a capacitance manometer. The chambers were evacuated before being filled with Ar/CO₂ to 300 kPa. The capacitance manometer allowed for precise pressure control. The temperature sensors on the chambers were tested by readout via the DCS box, which was controlled by a CAN Bus to USB interface and a slow control computer, with which also the readout electronics was configured. The chambers were stored and transported on wheels in stacks of two (see Figure 3.18). A scintillator long enough to cover the whole chamber width, was slid underneath the chambers perpendicular to the tubes and read out by a photomultiplier tube at each end. The photomultiplier signals were discretized to logic NIM signals. The coincidence of the two photomultiplier signals was used as trigger signal. A fixed deadtime of 10 ms was added to the trigger signal. For noise studies, an accidental trigger signal injected after the end of the deadtime, was available. In both cases, the signal was sent to a TTC unit module in a VME crate consisting of a TTCVI [38] and a TTCvx [39]. The TTC unit multiplexes the clock and trigger signal and sends them to the CSM via an optical fiber. The CSM was read out directly by a DAQ computer via an optical fiber. A picture of the setup at CERN is shown in Figure 3.18.

In ATLAS, the time measurements of all events are referenced to the LHC bunch crossing time, which is provided with very high precision by the LHC clock with a frequency of 40 MHz. Hence, it is sufficient to record the bunch crossing number, to determine the start of the electron drift with high precision.

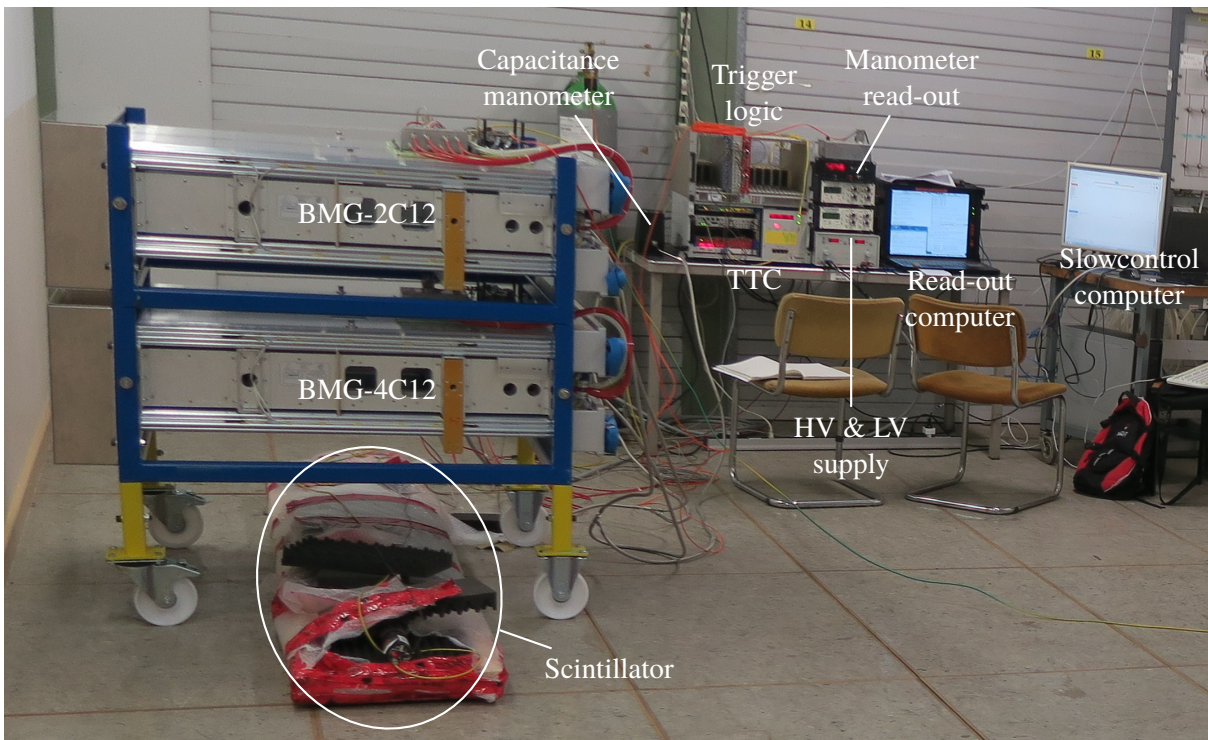


Figure 3.18: Photograph of the BMG test setup at CERN with a stack of two BMG chamber on transport wheels.

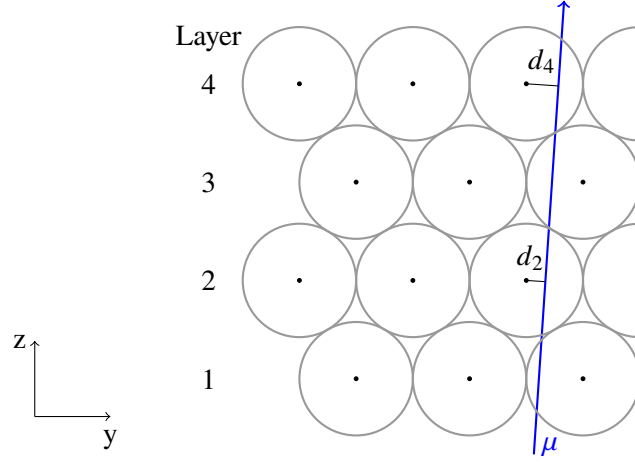


Figure 3.19: Illustration of the determination of the spatial resolution of the drift tubes in the BMG test setup. Muon tracks are fitted as straight lines to the hits in the drift tubes. The resolution is determined from the difference of the drift radii in two tubes with the same y position in each multilayer of the chamber in the chamber coordinate system (see text).

Cosmic muons arrive continuously at the chamber. The start time of the ionization electron drift has no constant offset to the LHC clock, to which the TDCs are synchronized. The HPTDCs of the BMG chambers digitize the threshold crossing times, corresponding to the end of the drift time, with a precision of 0.2 ns. The test setup only allowed for the recording of the scintillator trigger times, as alternative to the determination of the drift start time with a precision of 25 ns. Therefore, the drift time measurement in the test stand could only be performed with a resolution of 25 ns.

3.2.5 Drift Tube Spatial Resolution

With this 25 ns time jitter, track points can be reconstructed with a precision of about 0.3 mm. This is not sufficient to perform the autocalibration method of the space-to-drift time relation (see Section 6.2.4) and determine the spatial resolution from the track residuals to the measured hits [40]. Consequently, a fixed reference space-to-drift-time relation was used. The resolution is determined with a method, in which the 25 ns jitter of the trigger time cancels out [41]. A schematic drawing of a muon track in a BMG chamber is shown in Figure 3.19. Let $y(z) = m \cdot z + b$ be the projection of a muon trajectory into the y - z plane perpendicular to the tubes. In each crossed tube k with wire position y_k and z_k , the expected drift distance is given by the closest distance of the track to the wire

$$d_k = \frac{|m \cdot z_k + b - y_k|}{\sqrt{1 + m^2}}. \quad (3.1)$$

If the measured drift radius in the tube is r_k , the spatial resolution σ_{tube} is determined from the track residual $\delta_k = r_k - d_k$

$$\sigma_{\text{tube}} = \sqrt{\text{Var}(\delta_k)} = \sqrt{\frac{1}{N} \sum_{k=1}^N \delta_k^2 - \frac{1}{N^2} \left(\sum_{k=1}^N \delta_k \right)^2} = \sqrt{\text{Var}(r_k - d_k)} = \sqrt{\text{Var}(d_k - r_k)}. \quad (3.2)$$

In order to eliminate the intercept b of the muon trajectory, the measurements of pairs of tubes, which are two layers apart, e.g. layer 2 and 4 or 1 and 3, are combined. Equation (3.2) then becomes

$$\sigma_{\text{tube}} = \sqrt{\text{Var} \left(\frac{1}{2} (r_2 - d_2 + d_4 - r_4) \right)}$$

and

$$\sigma_{\text{tube}} = \sqrt{\text{Var} \left(\frac{1}{2} (r_1 - d_1 + d_3 - r_3) \right)},$$

respectively. Rejecting all tracks where d_2 and d_4 or d_1 and d_3 have opposite signs, i.e. muons with large $|m|$, inserting Equation (3.1) gives

$$\delta = \begin{cases} \sqrt{2} \left(r_2 - r_4 + \frac{m \cdot (z_2 - z_4)}{\sqrt{1+m^2}} \right) & \text{if } d_2, d_4 > 0. \\ \sqrt{2} \left(r_2 - r_4 + \frac{m \cdot (z_4 - z_2)}{\sqrt{1+m^2}} \right) & \text{if } d_2, d_4 < 0. \end{cases} \quad (3.3)$$

The slope of the muon trajectory m can be determined with high precision because of the large distance of 340 mm between the outermost tube layers of a chamber. It is limited to $|m| \leq 0.01$ in order to fulfill $\sigma(r_2) \approx \sigma(r_4)$ such that $\sigma_{\text{tube}} = \frac{1}{\sqrt{2}} (\sigma(r_2) + \sigma(r_4))$ and to keep the error due to the linear approximation of the space-to-drift time relation,

$$r_2 - r_4 = r(t_2) - r(t_4) \approx \frac{dr}{dt}(t_2) \cdot (t_2 - t_4),$$

small. This method slightly overestimates the track point measurement error when the space-to-drift-time relation has a larger curvature, namely at large drift radii (cf. Figure 6.14).

The residual distribution is fitted by a modified Gaussian function plus a constant offset A , which accounts for noise and fake hits, in the range $-1 \text{ mm} \leq \delta \leq 1 \text{ mm}$.

$$f(\delta) = C \cdot e^{-\frac{1}{2} \left(\frac{\delta - \mu}{\sigma} \right)^2} + A. \quad (3.4)$$

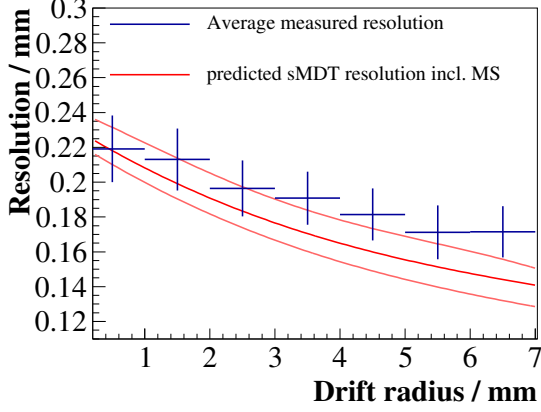


Figure 3.20: Dependence of the drift tube spatial resolution of the BMG chambers on the drift radius. The measurement is shown in blue. The expected sMDT resolution from muon beam measurements and simulation of multiple scattering (MS) of cosmic muons is shown in red with uncertainty band for comparison.

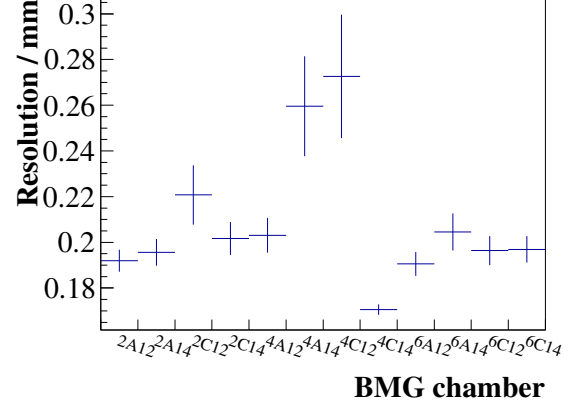


Figure 3.21: Mean drift tube spatial resolution in the twelve BMG chambers.

The width of the residual distribution is used as a measure for the average spatial resolution of the tubes. Reducing the fit range to $-0.6 \text{ mm} \leq \delta \leq 0.6 \text{ mm}$ allows for the determination of the systematic uncertainty of the mean spatial resolution. Without background irradiation, the systematic uncertainty is about $3 \mu\text{m}$.

The scintillator in the cosmic ray test stand used for triggering of the readout was 20 cm wide covering about 20 % of the tube length. This results in a small geometric acceptance of the trigger system with respect to the active chamber area which limits the statistics and requires several hours of data taking per chamber to perform the performance tests.

Combining the cosmic ray recorded data of all twelve BMG chambers, the dependence of the spatial resolution of the drift tubes on the drift radius could be determined (cf. Figure 3.20). Due to the strong energy dependence of the cosmic ray muon flux $j \sim E^{-2}$, most of the cosmic ray muons are of low momentum $p_\mu < 1 \frac{\text{GeV}}{c}$ and affected by multiple scattering in the chamber. Multiple scattering therefore contributes significantly to the spatial resolution σ_{meas} determined using cosmic muon tracks:

$$\sigma_{\text{meas}} = \sqrt{\sigma_{\text{tube}}^2 + \sigma_{\text{MS}}^2}, \quad (3.5)$$

where σ_{MS} is the contribution of multiple scattering. According to simulations of the chamber based on Geant4¹ [42], $\sigma_{\text{MS}} = (0.11 \pm 0.01)$ mm for the cosmic muon momentum spectrum in the test stand [41]. σ_{tube} is taken from high energy muon test beam measurements (cf. Section 6.4.1). Figure 3.20 shows the resulting prediction for the sMDT spatial resolution including multiple scattering in comparison with the measurement. The uncertainty on the determination of the multiple scattering contribution is indicated as a band around the mean value. For small drift radii, the agreement between measurement and expectation is very good. For large drift radii, the expected resolution falls systematically below the predicted resolution. This is a consequence of the increasing deviation at large drift radii of the space-to-drift-time relation from the linear approximation.

Figure 3.21 shows the mean tube resolutions in each BMG chamber including the contribution from multiple scattering. The differences between the chambers are a consequence of differences in gas pressure and temperature. Due to the design of the gas system, the pressure varied up to 10 kPa during and between measurements. The temperature variations were on the order of 5 K because the room was not climatized. The resulting differences in gas density linearly affect the gas gain and the actual space-to-drift time relation and hence the spatial resolution.

3.2.6 Drift Tube Efficiency

The drift tube efficiency was also determined using the cosmic ray data. Reconstructed tracks are extrapolated to each tube layer. The tube hit efficiency is measured as the probability to detect a hit in the tube that the muon passed through. As only one scintillator underneath the chamber was used for triggering, muons with a small inclination angle θ to the chamber can produce a reconstructable track in the chamber without propagating through the tube layers of the upper multilayer (see Figure 3.22), as only a minimum of four hits per track is required in the reconstruction. Hence, only tracks that have at least one hit on the track in one of the two uppermost tube layers (i.e. layer 1 or 2 of multilayer 1) are used for the efficiency determination.

Some tubes initially showed major inefficiency due to bad electric connections which could be recovered by replugging the electronics boards. One tube in layer 4 of multilayer 2 of chamber BMG-4C14 has a broken wire and was electricly disconnected.

The final measured average tube hit efficiency of all BMG chambers is shown in Figure 3.23 as a function of the drift radius. The drift tubes are fully efficient at small and intermediate drift radii. At the largest drift radii, the muon track length in the gas volume becomes increasingly smaller and therefore the ionization charge and the tube hit efficiency.

¹Geometry and tracking

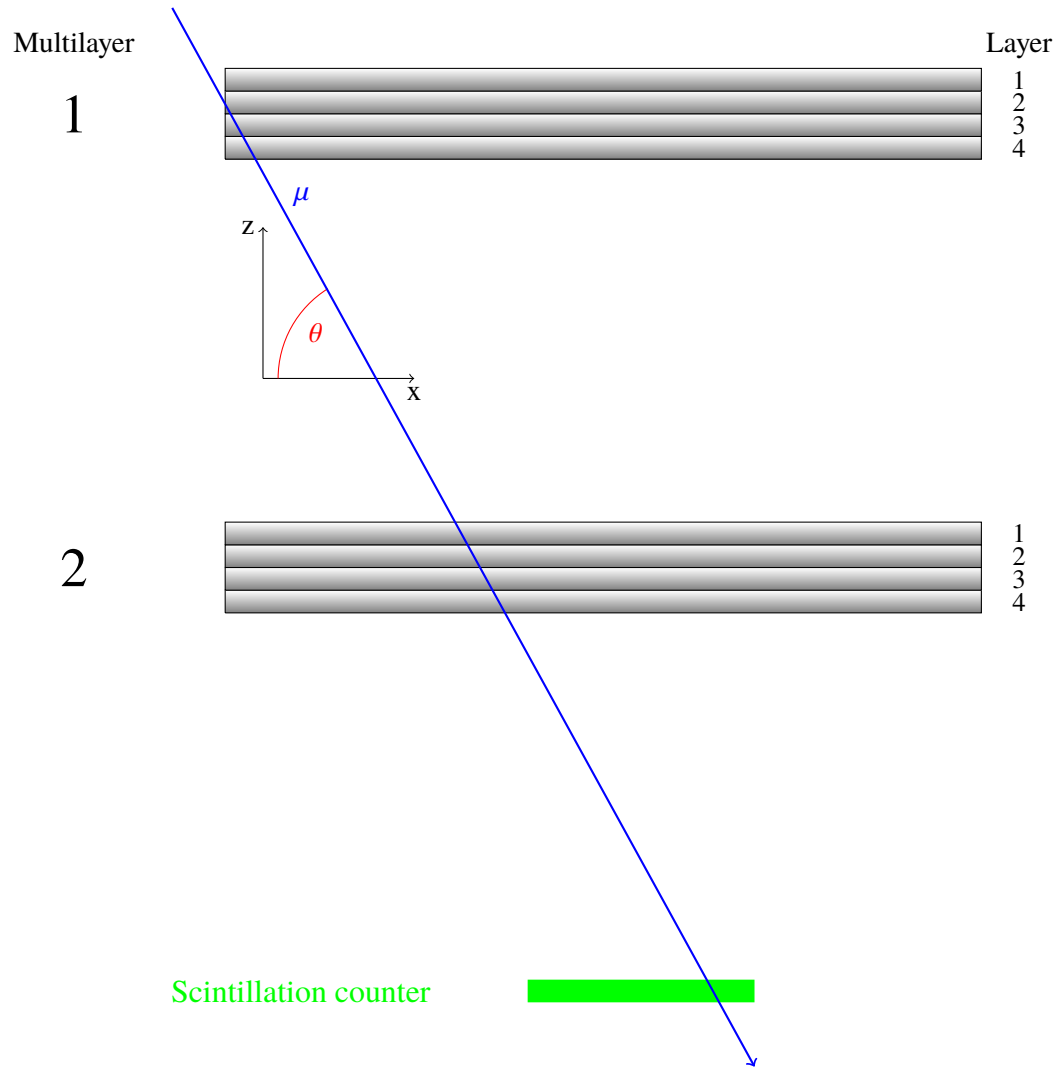


Figure 3.22: Illustration of the drift tube layer orientation and definition of the inclination angle θ in BMG cosmic ray test setup.

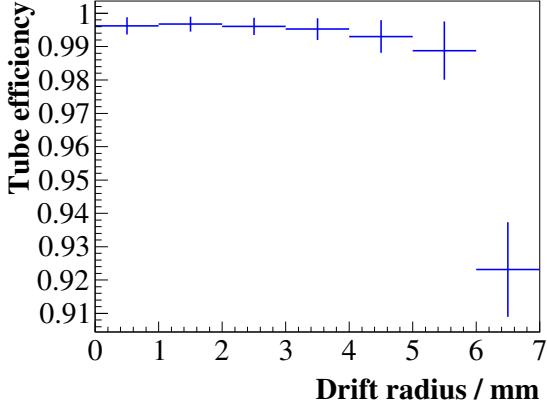


Figure 3.23: Dependence of the tube hit efficiency of the BMG chambers on the drift radius.

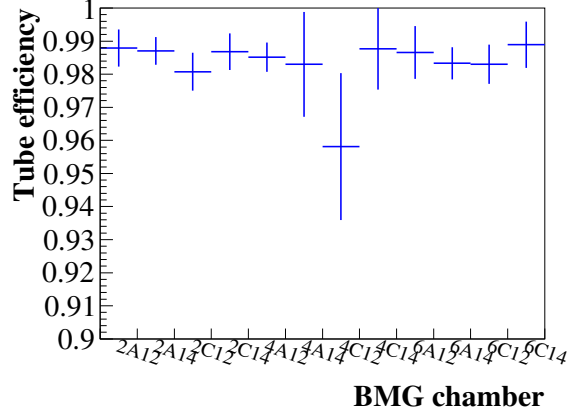


Figure 3.24: Average drift tube muon efficiency of the twelve BMG chambers. The average over all chambers is $98.3 \pm 0.8 \%$.

The average tube efficiencies of the different BMG chamber are shown in Figure 3.24. The average efficiency of all chambers is $98.3 \pm 0.8\%$.

3.3 Sensitivity to Pick-up Noise

After the installation of the BMG chambers in the ATLAS muon spectrometer, the noise hit rate increased compared to the commissioning on the surface in Munich and at CERN, where linear Iseg SHQ 224M HV power supplies were used. Figure 3.25 shows a comparison of the accidental hit rates of the tubes in an outer BMG sMDT chamber layer before and after the installation. On the surface, only very few tubes had hit rates larger than 10 ms^{-1} . These tubes are located at the cutouts and at the edges of the chambers, where the readout circuits are more exposed to external pick-up and where the ground connections of the tube walls have less redundancy. In the experiment, the BMG chambers are exposed to more noise sources than during commissioning in particular due to the switching HV power supplies used. A large number of tubes now show dark counting rates between 10 and about 100 ms^{-1} . The locations of the noisy tubes in the chamber and their noise hit rates depend on the environment in the experiment and are not stable in time and from chamber to chamber.

The increased noise hit rates mainly have the effect of consuming more readout bandwidth. The solution to the problem is discussed in the following.

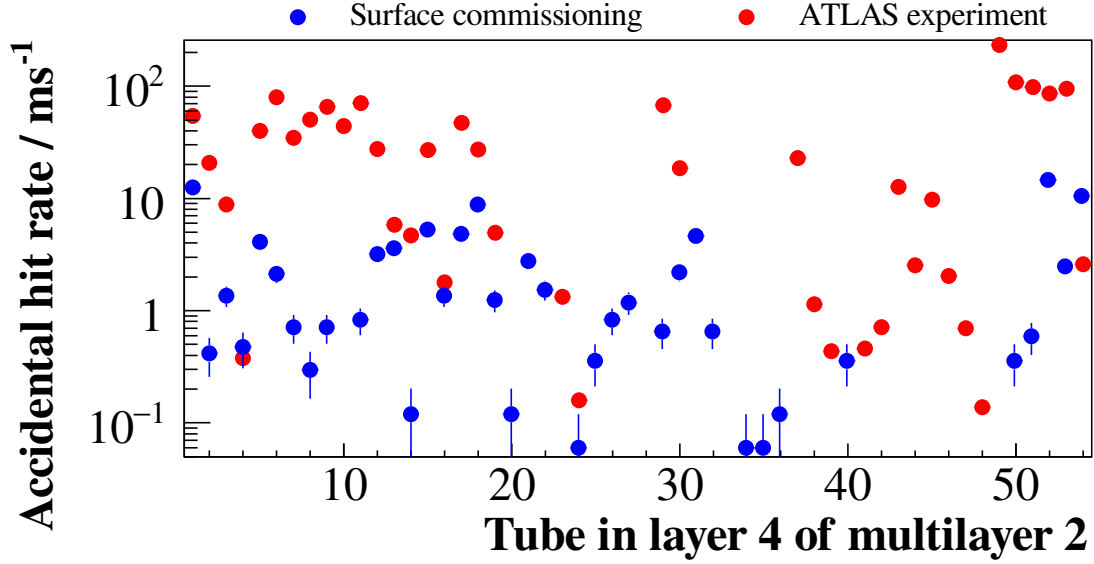
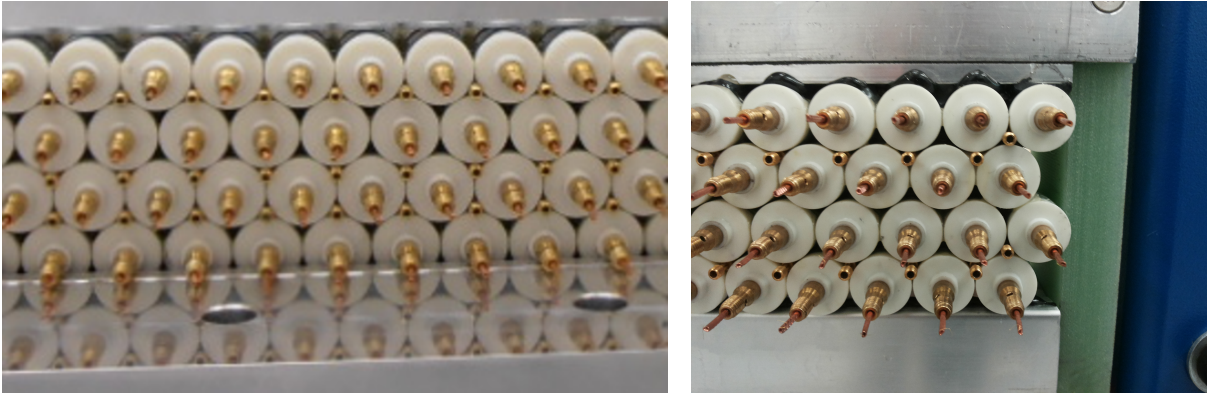


Figure 3.25: Comparison of the noise rates of layer 4 of multilayer 2 of chamber BMG-6C14 on the surface and in the pit.

3.3.1 Elimination of High Impedance Ground Loops

The BME chambers [30], two sMDT chambers which have been installed in the ATLAS muon spectrometer in 2013, do not show increased noise rates. The difference in the construction is that, in contrast to the BME chambers, the tube walls of layers 2 and 3 of a BMG multilayer were by mistake not directly connected with each other by grounding screws (see Figure 3.26) but rather via the the common ground plane of the high voltage signal readout boards which introduces ground loops which are sensitive to pick-up noise especially at the edges of the chambers. The ground screws between BMG tube layers 2 and 3 could not be inserted during the chamber assembly (see Section 3.1.2) because the necessary holes in the assembly combs had been omitted by accident and could not be screwed into the holes between the tubes belated, as they are filled with cured glue. Future sMDT chambers will be equipped with grounding screws between the layers 2 and 3 and, like the other screws, drilled into the wet glue between the tubes during production, to ensure a low impedance ground connection between all tube layers.

In order to nevertheless connect the tube walls of layers 2 and 3 directly with low impedance, the concept of a so-called ground foil as back plane of the Faraday cage which is electricly connected to all grounding screws was introduced again. This concept had already been used for the ATLAS MDT chambers to connect the tube grounds as in the MDT chambers the ground pins are connected to the



(a) Grounding screws on BME chambers [43].

(b) Grounding screws on BMG chambers.

Figure 3.26: Photograph of the positioning of grounding screws on the sMDT chambers installed in ATLAS. The grounding screws, which were not connected to the boards, were not installed on the BMG chambers. On each BMG chamber side four grounding screws were added belated to establish a ground connection to the outermost tubes, which were not connected electricly before.

aluminum rings of the individual endplugs. For the BMG chambers installed in ATLAS, this solution came to late, but it was successfully tested on the prototype BMG chamber not installed in ATLAS.

Figure 3.27 shows the ground foil installed on the BMG chamber prototype before mounting of the partially installed gas distribution system.

The effect of the ground foil on the noise rates is shown in Figure 3.28. To allow for comparison under the same external noise conditions, the foil was only installed on one multilayer on HV and RO side. Without direct low impedance connections between all tube layers, the noise hit rates exceed 10 ms^{-1} in several tubes, with a maximum of 1000 ms^{-1} . In multilayer 2, the accidental hit rate is reduced by at least two orders of magnitude due to the ground foil. Only two channels show a noise rate above 1 and below 2 ms^{-1} , the others are in the few hundred s^{-1} range as expected from the MDT and the BME sMDT chambers, which are equipped with ground screws between tube layers 2 and 3.

3.3.2 Validation of the Grounding

For future sMDT chambers for ATLAS, the grounding screws between tube layers 2 and 3 will of course be used. For additional safety, also the grounding foil will be used. In order to validate the grounding network of a chamber and to exclude unwanted high impedance ground loops, a new testing procedure of the frontend electronics on the chamber has been developed.

Due to production tolerances during the MDT ASD chip production, the discriminator thresholds of the ASD chips and channels vary quite significantly, and as a consequence also the noise rates per channel.

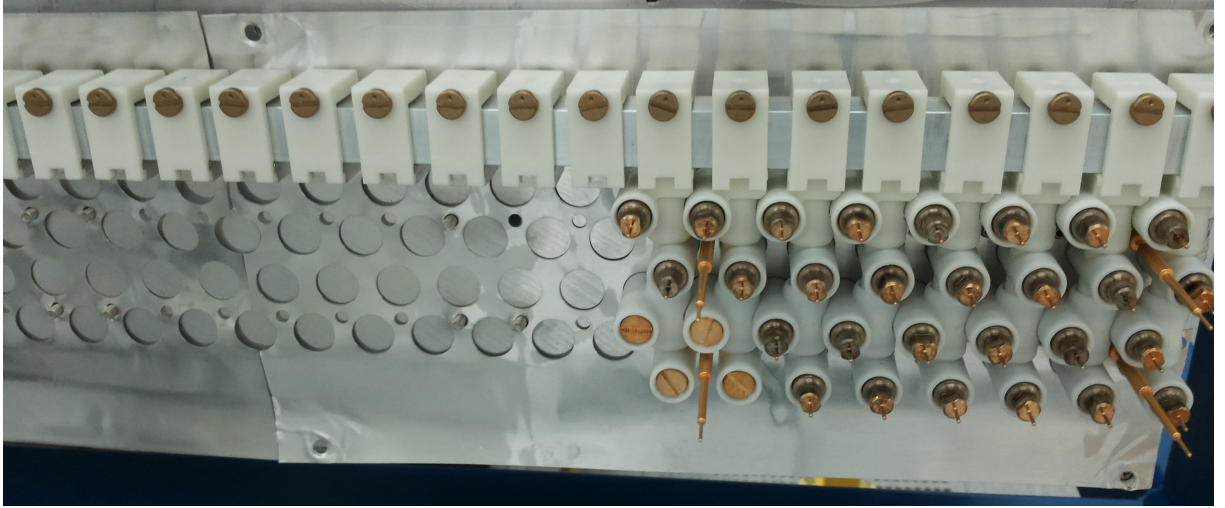


Figure 3.27: Photograph of the ground foil installed on one of the BMG multilayers underneath the gas distribution system.

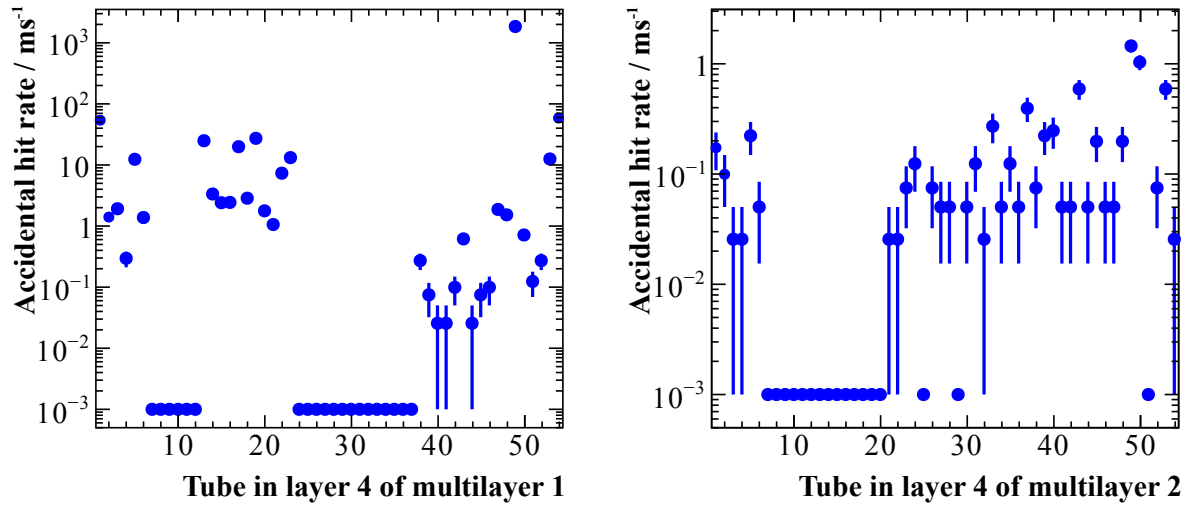
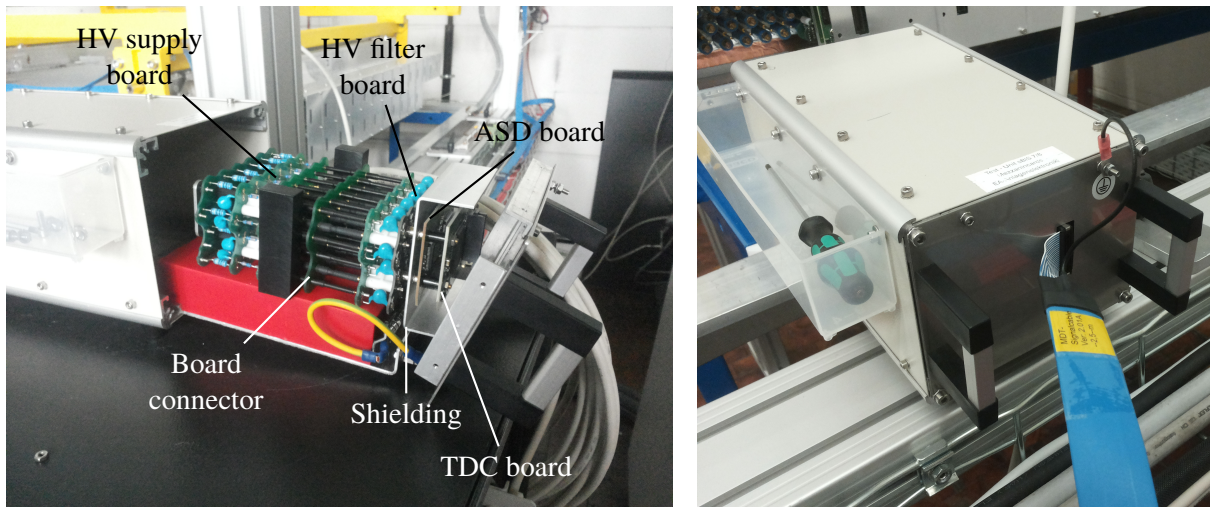


Figure 3.28: BMG noise hit rates in a tube layer without (multilayer 1, left) and with (multilayer 2, right) direct low impedance connections between all tube layers.



(a) Opened electronics test setup.

(b) Closed Faraday cage.

Figure 3.29: Photographs of the front-end board noise test setup.

To obtain reference noise measurements under ideal conditions for the different ASD channels, the whole electronics chain for a set of 24 tubes from HV distribution with the termination resistors to mezzanine cards is tested in a Faraday cage which is constructed like the ones on the chambers (see Figure 3.29). The drift tubes are replaced by pin connectors in the setup.

The noise rates are measured with the electronics installed on the chamber with high voltage applied. If the difference in noise rate between ideal test setup and chamber is larger than a factor of 2, the ground connection and shielding are checked and improved. Figure 3.30 shows an example for the accidental hit rates of three mezzanine cards on a new chamber with all grounding screws and ground foils. The on-chamber rates follow the expectation of the off-chamber test, and are only slightly higher. This shows that optimum grounding and shielding and minimum achievable noise rates for the given ASD chips have been reached.

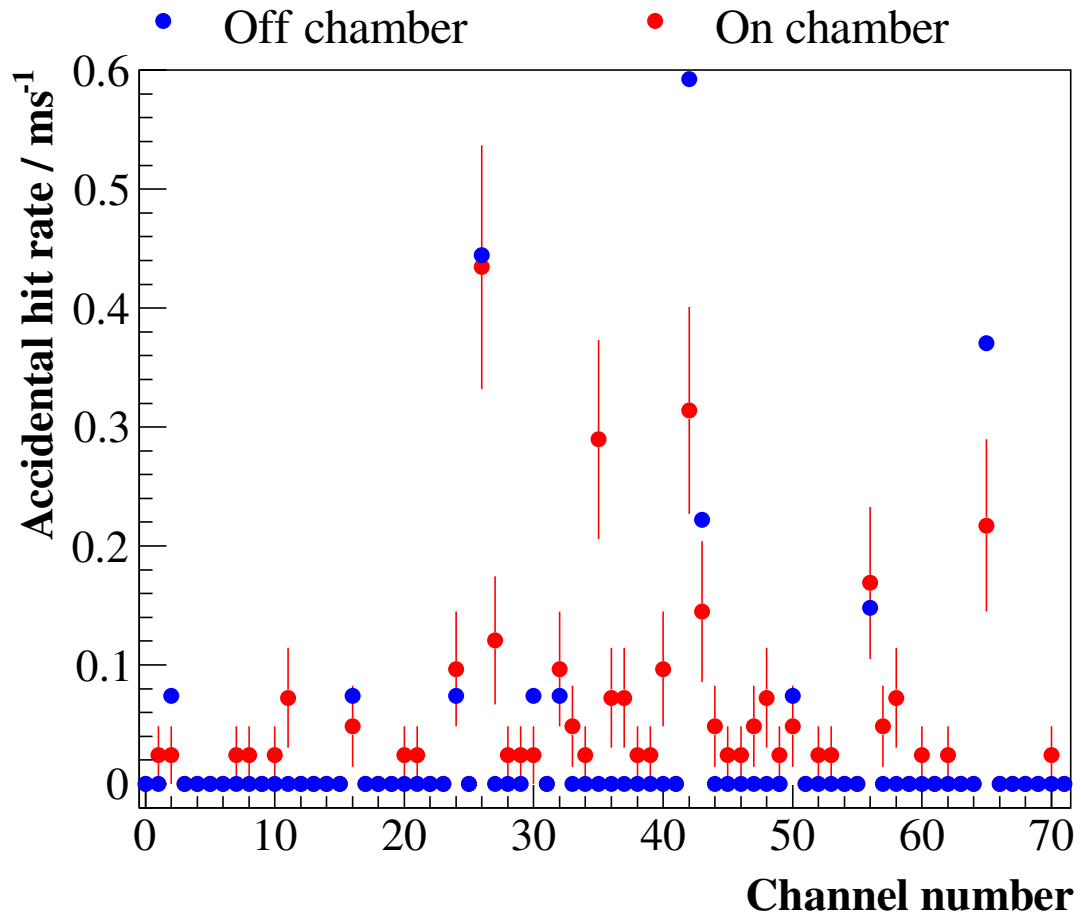


Figure 3.30: Comparison of off- and on-chamber noise rates for the hardware channels of three mezzanine boards with the nominal ATLAS settings.

Chapter 4

Readout Electronics Concepts

In this chapter the electronics used for the readout of the ATLAS MDT chambers, its limitations and concepts for new readout electronics for high counting rates are discussed.

4.1 The ASD Chip

The first active component of the front-end electronics for the readout of the ATLAS MDT chamber is the **A**mplifier/**S**haper/**D**iscriminator chip [44]. It is an octal ASIC with a size of 3 mm times 4 mm, realized in 250 nm CMOS technology. Each channel consists of five subcircuits shown in section 4.1.

1. The pre-amplifier consist of the “active” amplifier and a “dummy”, each with a gain of 40 dB. The input of the first one is connected to the sense wire and amplifies the actual signals. The input of the latter is connected to ground. This quasi-differential circuit compensates the DC levels and produces a differential signal for the subsequent stages.
2. The shaper consists of three stages of differential amplifiers (DA). The first stage further amplifies the signal and ensures that the output is fully differential. The second and third stage use R-C combinations as input impedances and as pole-zero in the feedback loop in order to obtain frequency dependent amplification with high gain in the relevant range around 5 MHz. DA 3 differentiates the signal for bipolar shaping (see Section 4.3) and is nonlinear, in contrast to the other stages, suppress large background γ pulses and maximize the dynamic range.
3. The shaped signal is fed into the timing discriminator. Its threshold and hysteresis are controlled by a digital-to-analog converter (DAC).

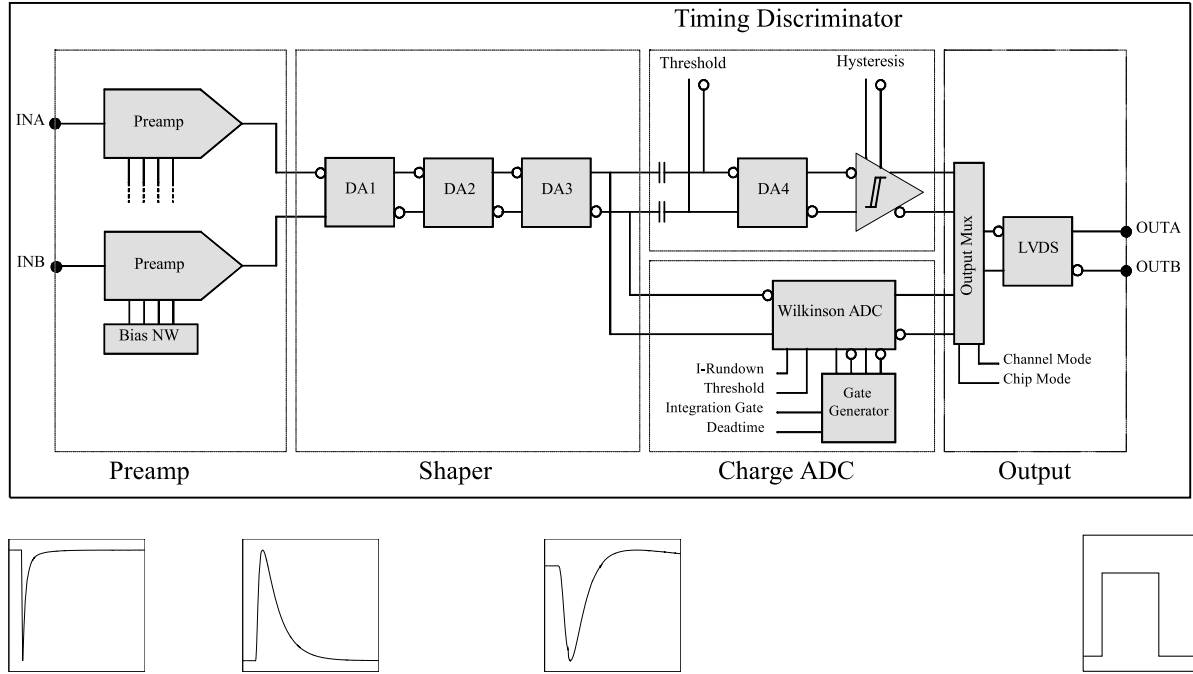


Figure 4.1: Block diagram of the ATLAS MDT ASD [44] with systematic pulse shapes between the blocks.

Table 4.1: Programmable parameters of the MDT ASD chip with the minimum, maximum and default settings used in ATLAS [15, 44]. The last column gives the integer value that has to be set in the configuration file to obtain the default setting.

Parameter	Min.	Max.	Nominal	Unit	Set value
Main threshold	-255	255	-39	mV	108
Main hysteresis	0	18.75	8.75	mV	7
Wilkinson threshold	126	14	94	mV	2
Wilkinson gate	11	33.5	18.5	ns	5
Wilkinson rundown	0.7	7.3	4.5	μ A	4
Programmable deadtime	17	535	535	ns	7
Calibration select	1	8	1	-	0
Calibration mask	0	255	0	-	0
Channel mode	HI/LOW/ON		ON	-	0
Chip mode	TOT/ADC		ADC	-	1

4. The shaper output is also fed into a charge measurement stage, which is often called Wilkinson ADC [44]. It collects the pulse charge within an adjustable time window of 11 to 34 ns after the discriminator threshold crossing in a hold capacitor. After the integration window, the capacitor is discharged with adjustable constant current. The discharge time is a measure of the rising edge charge of the signal and coded into the output signal length of discrete amplitude. The ADC charge measurement can be used offline for time slewing corrections, which are described in Section 6.3.4.
5. The outputs of the discriminator and the charge measurement stage are multiplexed (Mux) in the output stage. A LVDS¹ driver produces the differential output signal, which is transmitted to the time-to-digital converter (TDC).

All chip settings are programmed in a register with the JTAG protocol [45]. The analog control parameters, like threshold and hysteresis values, are generated by DACs, which are controlled via the same JTAG interface. Each of the eight channel outputs can be activated or deactivated separately, while all other settings are common for all channels. There are two different operating modes of the ASD. In the time-over-threshold (TOT) mode, the ASD codes the input pulse width into the discrete output signal, which goes high and low again when the signal surpasses the discriminator threshold and when it falls below it, respectively. In the ADC mode, in contrast, the rising edge charge measurement is coded into the output width. In the ADC mode, a programmable deadtime can be set in addition to the output pulse width, in order to avoid multiple threshold crossings due to successive ionization clusters and to reduce the needed readout bandwidth. If the programmable deadtime is set to its maximum, the deadtime is, due to the output pulse width, 820 ns, slightly larger than the maximum drift time of 30 mm diameter ATLAS drift tubes. The minimum deadtime in the ADC mode is 200 ns to 300 ns, depending on the chip. The deadtime is determined by the minimum distance between two successive hits (cf. Section 6.2.2). An overview over all programmable ASD parameters and their nominal settings used in ATLAS is given in Table 4.1.

4.2 Readout Electronics Effects at High Counting Rates

4.2.1 Deadtime Effects

The deadtime $\tilde{t}_{\text{dead}}$ of a readout channel of the ASD chip after a threshold crossing is programmable between about 200 ns and 800 ns in the ADC mode (see Section 4.1). In the TOT mode there is an effective deadtime of about the negative signal width belonging to the threshold crossing.

¹Low-voltage differential signaling

All hits arriving during the deadtime are lost. This reduces the detection efficiency, the efficiency to detect the hit belonging to the muon track. For background radiation hits arriving randomly in time, the reduction depends on the deadtime and the counting rate. Two different cases of deadtime can be distinguished [46]. The extendable, also called paralyzable, deadtime, where the deadtime is extended by a further hit arriving during the primary deadtime, and the non-extendable, also called non-paralyzable case, in which all hits arriving during the deadtime have no effect. The electronics deadtime of the ASD chip is non-extendable.

Let ν be the tube hit rate and $k = n \cdot T$ the number of recorded hits in the time interval T . The actual number of tube hits is then

$$\nu \cdot T = k + \nu \cdot k \cdot t_{\text{dead}},$$

because k registered hits lead to a total deadtime of $k \cdot t_{\text{dead}}$, in which hits were not recorded with rate ν . If ϵ_0 is the detection efficiency without background irradiation, t_{dead} the deadtime and n the recorded counting rate, the efficiency under irradiation is given by

$$\epsilon(n) = \frac{\epsilon_0(\nu T - \nu k t_{\text{dead}})}{\nu T} = \epsilon_0 \cdot (1 - n t_{\text{dead}}). \quad (4.1)$$

or

$$\epsilon(\nu) = \epsilon_0 \frac{nT}{nT + \nu n T t_{\text{dead}}} = \frac{\epsilon_0}{1 + \nu \cdot t_{\text{dead}}}. \quad (4.2)$$

Because of masking of particle conversions in the tube wall by δ electrons created by the particles in the tube walls with ionization clusters arriving earlier at the wire, $\epsilon_0 \approx 93\% < 100\%$. Taking into account the possibility of multiple threshold crossings caused by several primary ionization clusters per hit, this is modified to

$$\epsilon(\nu) = \frac{\epsilon_0}{1 + \nu \cdot m \cdot t_{\text{dead}}}, \quad (4.3)$$

where m is the hit multiplicity, i.e. the average threshold crossings per tube hit. With the nominal ATLAS settings, the MDT hit multiplicity is very close to unity [15]. In order to achieve a high detection efficiency at high counting rates, the deadtime and threshold counting rates must be kept low.

4.2.2 Pileup Effects

With increasing rate, the time distance between two subsequent hits decreases. If the time between two hits is smaller than the time it takes for the shaper output to return to the baseline after the first hit, the second hit superimposes the first signal. This effect is called pileup. Due to the superimposition, the second hit bases on an effectively shifted threshold and its amplitude is changed. The effects of pileup strongly depend on the pulse shaping. The two commonly used shaping concepts are unipolar and bipolar shaping

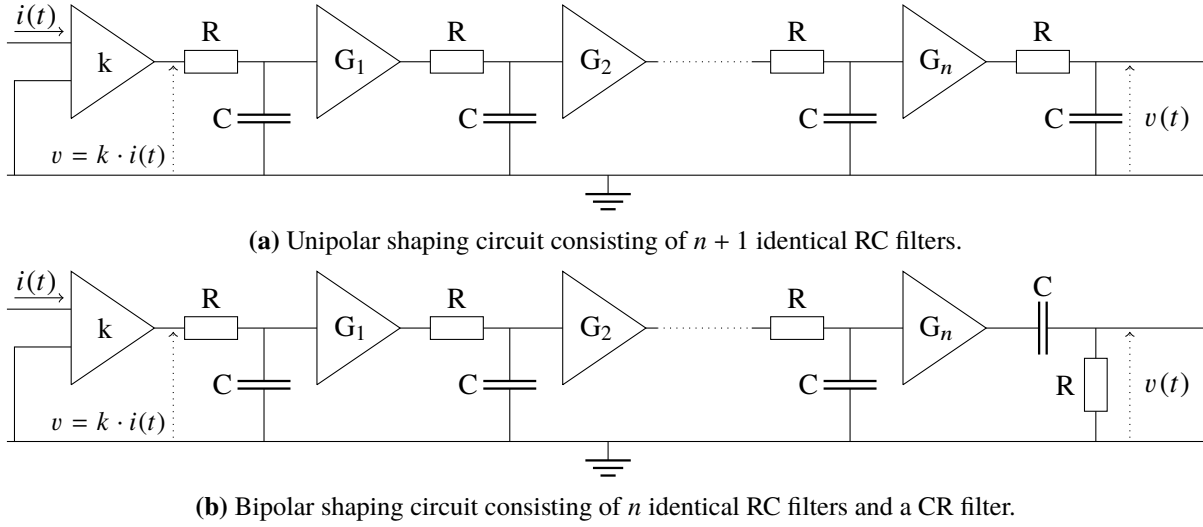


Figure 4.2: Generic schematics of shaping circuits consisting of $n + 1$ active stages with operational amplifiers [47].

(cf. Figure 4.2). Unipolar shaping is achieved using just RC filters. Adding a CR filter leads to bipolar pulse shaping. The resulting pulse shapes and pileup effects of both shaping concepts are described in Section 4.3. The ASD chip uses bipolar shaping, where a signal is followed by an undershoot of the same area. For the ATLAS ASD chip, this differentiating behavior is achieved by a CR filter in the last shaping stage. Due to finite rising times of the signal and limited amplitudes, the spatial resolution and the detection efficiency is deteriorated in case of pileup. This will be explained in detail in Section 4.3.4.

4.2.3 Readout Bandwidth and Buffer Limitations

The ATLAS Muon TDC (AMT) has been developed to record the ASD output signals and digitize the drift time measurement between bunch crossing and threshold crossing time [48]. It has 24 LVDS input channels to read out three ASD chips. The times of the rising and falling edges of each input signal is converted into a data packet (word) consisting of 32 bits with 0.78125 ns time resolution. Each channel has a FIFO buffer with a capacity of four measurements. If a hit arrives while the FIFO is full, it is rejected and a data word with the corresponding error code is generated. The hits in the channel buffer are read out and written into the central circular (level 1) buffer with 40 MHz frequency, which has a capacity of 256 data words. Data arriving while the buffer is full are again rejected and an error word is generated. Data in the central buffer are either transferred to the readout FIFO if there was a trigger or rejected if there

was no trigger within the trigger latency. Also the trigger signals are stored in a FIFO, before they are processed. The data in the readout FIFO is read out with a bandwidth of 80 Mbps.

The buffer size and the readout bandwidth limit the efficiency of the detector readout. The channel buffer will overflow only at extreme high rates, because it only has to store the hits from one tube for a short time. The central buffer, however, has to store the information of all 24 tubes read out by the TDC during the time of the trigger latency, which is on the order of $2\ \mu\text{s}$. Above an average hit rate of $400\ \text{ms}^{-1}$ per channel, data start to get lost. The data carrying error codes from buffer overflows enhance this problem. Already a small excess of the buffer capacity generates error words, which, in addition to the hit information, consume buffer space and bandwidth.

The maximum tolerable trigger rate depends on the data transmission bandwidth and the buffer sizes. The AMT can handle trigger rates of up to $200\ \text{ms}^{-1}$ for low hit rates. At higher background hit rates, more data words per triggered event have to be transmitted, which decreases the maximum acceptable trigger rate. Trigger signals, which arrive while the trigger FIFO is full, are rejected and the whole event is lost. Overflowing level 1 or channel buffers lead to a loss of individual hits. As long as the data flow is not interrupted due to too many error words, e.g. after a noise hit burst, this has the same effect as increased channel deadtime. In order to achieve high efficiency, it is crucial to have sufficient buffer size and data bandwidth between all the memories.

The MDT chamber in ATLAS are read out with the AMT chip. Only the 12 BMG sMDT chambers in the feet region of the middle barrel layer (see Chapter 3) use the HPTDC [28] chip instead. It has a similar architecture like the AMT, but can achieve a time resolution of up to 24 ps. As a time resolution of 800 ps is sufficient for optimum MDT performance, the HPTDC is operated with a resolution of 195 ps to save data bandwidth. A significant advantage of the HPTDC over the AMT for the MDT readout is increased memory size. The HPTDC has a level 1 buffer size of 256 words for each block of eight channels. This prevents data loss up to hit rates of 1 MHz per channel. This chip is also used on all prototype electronics boards discussed in Section 4.3 and used for the studies in this thesis.

4.3 Signal Shaping

The ATLAS MDT readout electronics has been optimized for drift tubes with 30 mm diameter and background rates up to $500\ \text{ms}^{-1}\text{cm}^{-2}$. In this section, the limitations of the MDT ASD chip are explained, before the new analog readout concepts for much higher counting rates are described.

Table 4.2: Comparison of properties of the readout electronics operated with their nominal settings (Table 6.1). The minimum deadtime is the minimum time between two discriminator threshold crossings recorded by the TDC. It was determined from the muon beam data (cf. Section 6.2.2). The deadtime spread of the MDT ASD chip is a result of production tolerances and caused by the charge measurement. It is 0 ns if the chip is operated in the TOT mode.

Parameter	MDT ASD chip	Discrete unipolar shaper	Discrete bipolar shaper
Signal peaking time	15 ns	5 ns	5 ns
Minimum deadtime	200...300 ns	0 ns	0 ns

4.3.1 Limitations of the MDT ASD Chip

Bipolar pulse shaping was chosen for the ATLAS MDT readout to avoid the necessity of active baseline restoration [49]. In addition, fine tuning of the signal shaping time constants is not necessary, which is difficult due to chip production variations [47]. The ASD chip with bipolar shaping guarantees baseline stability for MDT hit rates of up to 400 ms^{-1} without significant performance loss.

Figure 4.3 shows a comparison of the dependence of the MDT and the sMDT spatial resolution on the background hit rate. Besides the clear superiority of the 15 mm diameter tubes over the 30 mm diameter tubes at high γ conversion flux, the improvement of the resolution with increased deadtime is shown. For the sMDT data, taken with minimum adjustable ASD deadtime, leading to 220 ns minimum deadtime, for optimum efficiency, artificial deadtime is introduced by requiring a minimum time interval of 600 ns between successive hits in order to suppress signal and resolution deterioration due to signal pile-up effects.

Avoiding resolution deterioration due to signal pileup effects is especially important for the 15 mm diameter drift tubes because they are supposed to operate at much higher rates than the standard MDT chambers.

To avoid signal pileup, the detector deadtime would have to be significantly increased, which reduces the detection efficiency. Optimized readout electronics with fast baseline restoration (BLR), which returns the shaper output to the baseline as quickly as possible after the end of the input pulse is a solution to fully exploit the rate capability of the sMDT chambers [50].

First tests have already been performed with the ASDQ¹ chip, which has a BLR stage for the readout of the CDF² Central Outer Tracker [52] which has to cope with very high rates. It turned out that this chip is not suited for the readout of sMDT chambers due to its high threshold and fast pulse shaping and peaking time [53]. Two different BLR concepts have, therefore, been developed and tested specifically for

¹Amplifier/Shaper/Discriminator with charge(Q) measurement [51]

²Collider Detector at Fermilab

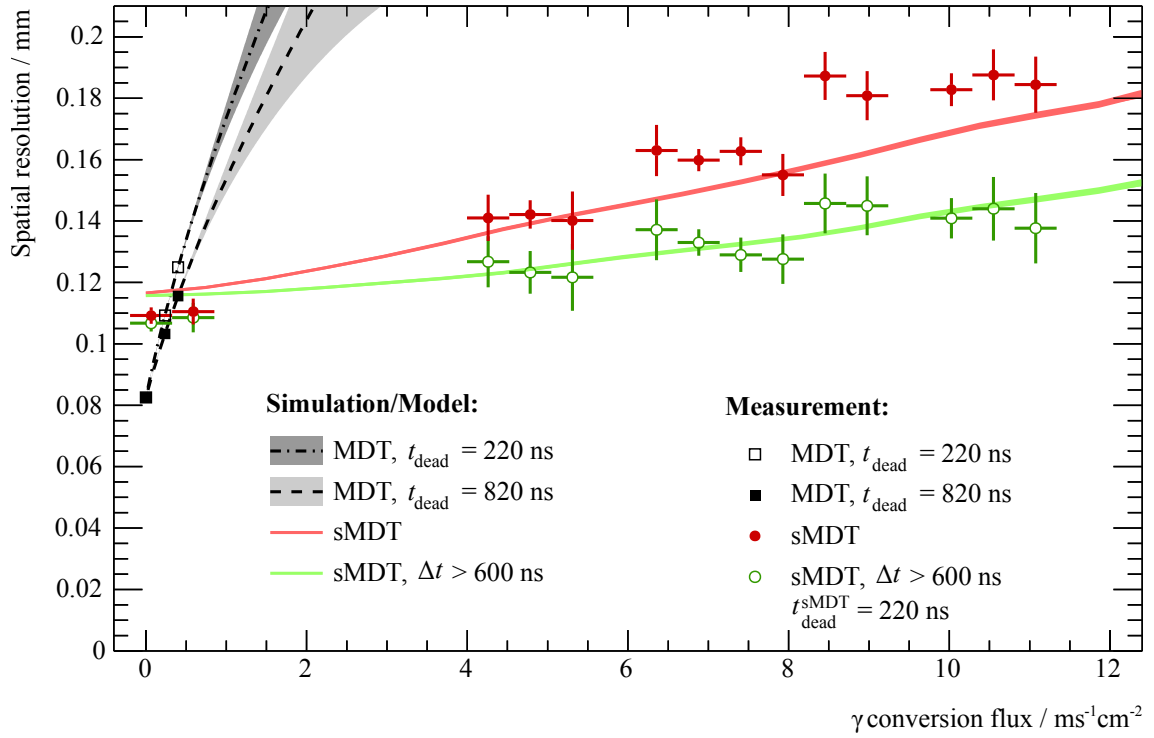


Figure 4.3: Spatial resolution of sMDT and MDT as a function of the γ conversion flux in the Gamma Irradiation Facility GIF at CERN, measured with cosmic ray muons [15] for different electronics deadtime settings. The points show the measurement results. 30 mm diameter MDT with 820 ns maximum drift time were tested for the minimum and maximum deadtime settings of the ASD. For 15 mm diameter sMDT with 190 ns maximum drift time only the minimum electronics deadtime was used. The offline requirement $\Delta t > 600$ ns on the time interval between two successive hits corresponds to an electronics deadtime of 600 ns. The curves show the result of the simulation model [15].

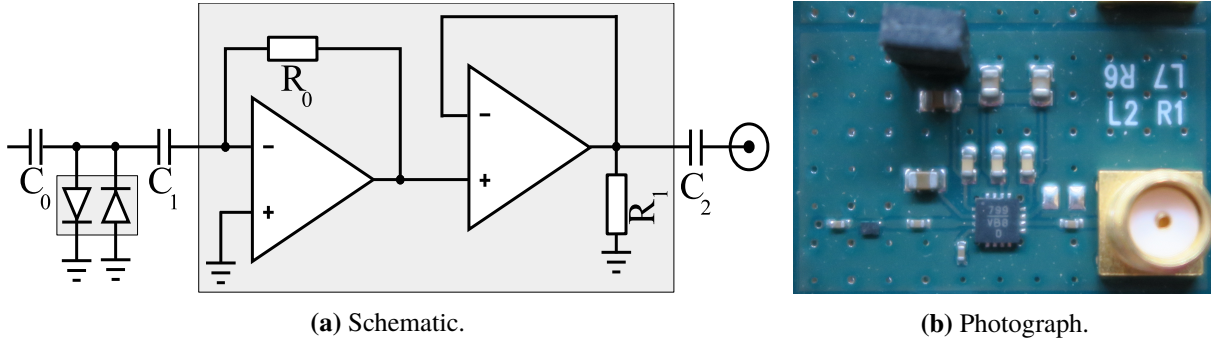


Figure 4.4: Schematic (a) and photograph (b) of one preamplifier channel. The filter capacitors $C_0 = C_1 = C_2 = 100\text{ nF}$ eliminate DC offsets. A protection diode pair (Infineon ESD0P4RFL [54]) prevents damage of the amplifier due to voltage spikes from discharges in the tube or the HV decoupling capacitor. The transimpedance amplifier with $R_0 = 10\text{ k}\Omega$ and the output driver stage with an output impedance of $R_1 = 50\text{ }\Omega$ are integrated in a commercial chip [55]. 24 channels are combined on one board to amplify the signals from all tubes connected to a signal hedgehog board.

the readout of sMDT chambers. They are described in the following and compared to the MDT ASD chip in Table 4.2.

4.3.2 Preamplifier Stage

The first stage of the new readout electronics is a 700 MHz bandwidth inverting preamplifier¹ with 10 k Ω transimpedance, shown in Figure 4.4. In order to read out a complete chamber, 24 discrete channels are densely arranged on one board [56] with a connector to a standard sMDT high voltage signal board mounted on the drift tubes. The LV supply is realized with a common connector per board. By removing a jumper, each channel can be disconnected from the power supply to turn it off. Each amplifier output is connected to a SMA connector for a cable transmitting the signal output to the off-chamber shaper board, of which there are two types, unipolar, i.e. with negligible pulse differentiation, or bipolar shaping, as will be discussed in the two following subsections.

4.3.3 Unipolar Shaping

As BLR functionality is needed anyway for sMDT operation at very high rates, bipolar pulse shaping loses its advantages over unipolar shaping, which is affected by baseline shift already at intermediate rates. Figure 4.5 shows the pileup of two subsequent generic shaped pulses for purely unipolar shaping. The output voltage slowly returns to the baseline, such that the second pulse piles up onto the tail of the

¹Analog Devices, HMC799LP3E [55].

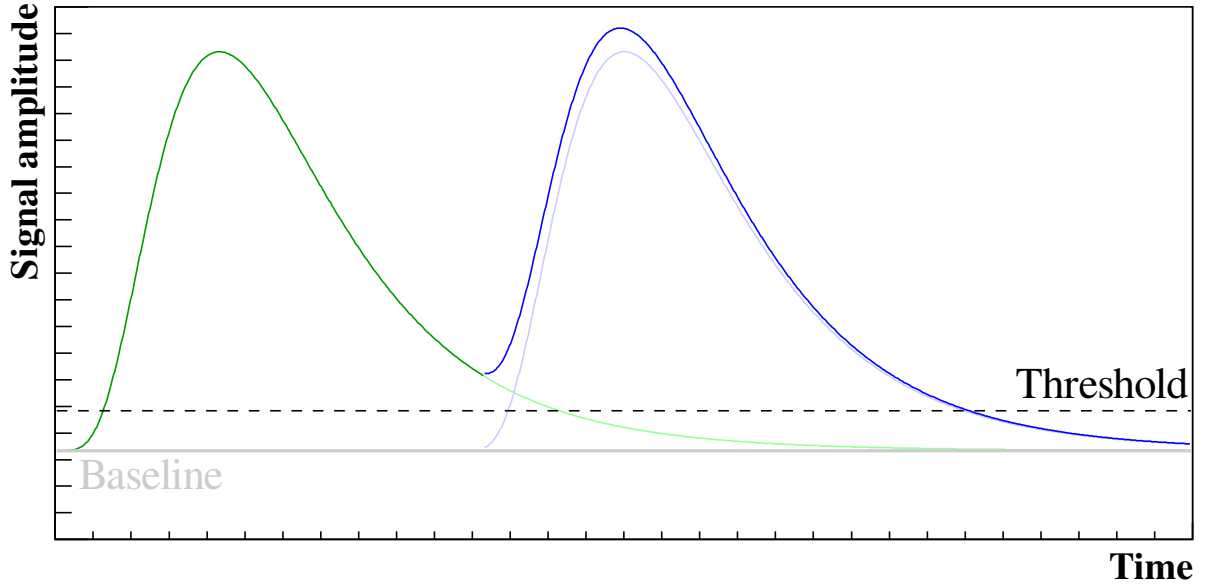


Figure 4.5: Illustration of signal pileup for unipolar shaping without capacitive coupling. The second signal (blue) adds upon the tail of the previous. As the signal amplitude does not drop below the discriminator threshold in between the pulses, the second hit is not detected by the read-out electronics.

previous one, before the voltage finally falls below the discriminator threshold allowing for the detection of a new hit. The time constant of the shaper has to match the detector signal shape and noise level and can hence not be arbitrarily adjusted to accelerate the return to the baseline. As the pulses are purely positive, continuous pileup leads to a shift of the baseline to higher voltages. Due to the long signal tails of more than $2\ \mu\text{s}$ (cf. Figure 5.3), already moderate rates of $r \geq 500\ \text{ms}^{-1}$ lead to a baseline shift which fluctuates, because the tube hits occur randomly. An increased baseline voltage reduces the dynamic range and increases the probability for saturation of the circuit.

Purely unipolar shaping is only achieved, if the whole analog readout chain contains no capacitor in the signal path or parallel to it. This is not possible for any detector, where the HV is applied to the anode, like in the ATLAS MDT chambers, because the HV decoupling capacitor (cf. Figure 2.9) differentiates the signal. Hence, every signal pulse will be followed by an undershoot of the same area (effectively bipolar shaping). For large capacitors and discharge resistors, this undershoot is long with small amplitude, so that the signal resembles unipolar shaping. An example of pulses with this shaping is shown in Figure 4.6. In contrast to the purely unipolar shaping, a successive hit can already be detected long before the end of the negative tail of the previous signal. However, the threshold crossing time of the second hit is delayed depending on the undershoot amplitude leading to additional time slewing and the amplitude of the second hit above the threshold is reduced. The long duration of the undershoot of each pulse leads to a

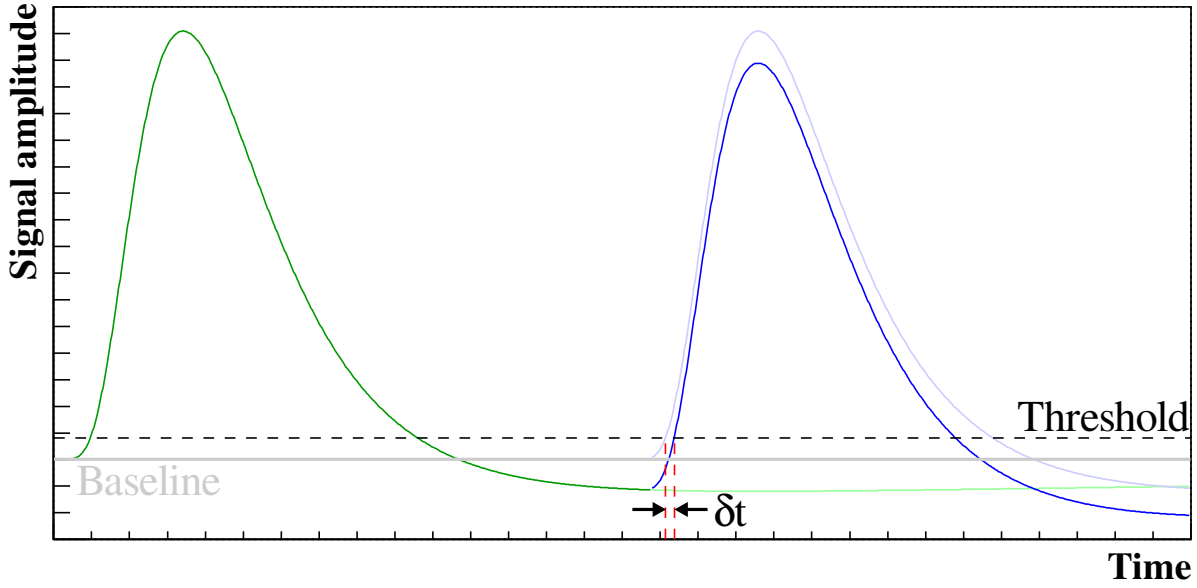


Figure 4.6: Illustration of signal pileup for unipolar shaping with capacitive coupling. The second signal (blue) adds upon the long undershoot of the previous pulse. Due to this signal pileup, the output amplitude of the second hit is reduced and the threshold crossing time is delayed by δt .

negative baseline shift increasing with hit rate, as the undershoots of successive pulses add up already at moderate rates. Each capacitive coupling through a CR filter in the signal path differentiates the signal. Long undershoots are, due to their small amplitude slope, effectively eliminated by the next CR filter with smaller time constant $\tau = CR$. The length and amplitude of the resulting undershoot is determined by the time constant τ of the CR filter, leading to baseline shift in case of signal pileup. This baseline shift is fluctuating, as the tube hits arrive randomly. To avoid jitter of the threshold crossing times, the baseline must be restored directly before the discriminator.

Figure 4.7 shows the circuit diagram of the unipolar shaper board with BLR stage, which was developed for the readout of the 15 mm diameter drift tubes at high rates. The output signal of the preamplifier board is injected to the shaper via a SMA connector. The input is terminated with $R_0 = 50\ \Omega$ and capacitively decoupled with $C_0 = 10\ \text{nF}$ from the input operational amplifier¹ to cancel DC offsets. The first active stage amplifies the signal and cancels the long tails of the drift tube pulses. The second stage further amplifies the signal and inverts it, restoring positive signal amplitude. Using two active stages allows for staging of the gain and, thereby, widening of the bandwidth. Because of the last active shaping stage (cf. shaper stage response in Section 5.1.2.1), this circuit is referred to as unipolar shaper in the following.

¹Texas Instruments, THS4304 [57]

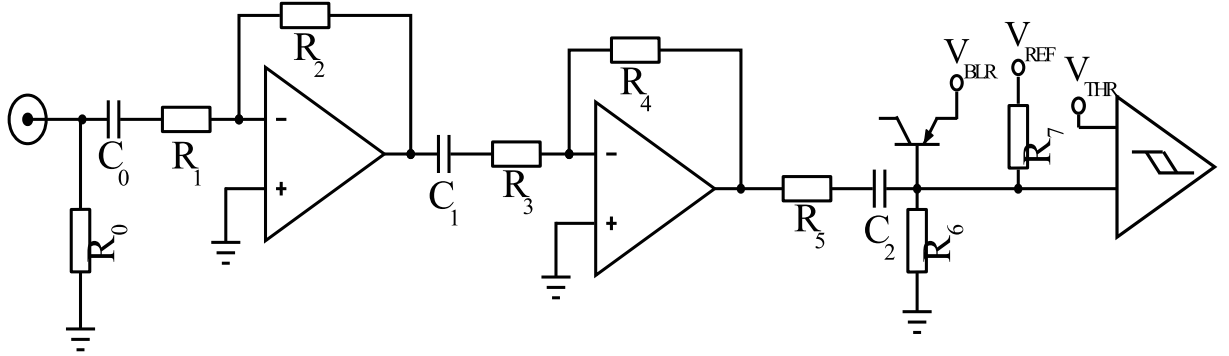


Figure 4.7: Simplified schematic of the discrete unipolar shaper circuit. The detailed circuit diagram with values of the components and additional buffering capacitors is shown in Figure A.1.

The BLR stage is an enhanced version of the simple diode restorer [58]. It begins with the capacitor $C_2 = 30 \text{ pF}$. Due to its time constant $\tau = CR = \tau = C_2 \cdot (R_6 || R_7) \approx 200 \text{ ns}$, it generates a significant undershoot and differentiates away the flat undershoot from the previous stages. The remaining undershoot is short-circuited to ground by the diode via the pull down resistor $R_6 = 10 \text{ k}\Omega$ depending on V_{BLR} . The diode can be in three different operational states.

- **No signal input:** The diode is slightly conducting at a working point adjustable via the BLR voltage V_{BLR} .
- **Positive input:** The voltage drop at the diode is reduced shifting the diode into the non-conducting region. Only a small amount of charge is drained to ground through R_6 reducing the signal amplitude.
- **Negative input:** During the undershoot, the voltage drop at the diode is increased leading to an exponential increase of the current through the diode canceling the undershoot.

As the transitions between these operational stages are smooth, the functionality is enhanced for larger voltage amplitudes.

An analogous circuit for a differential signal path is realized in the ASDBLR¹ chip, which has been designed for the readout of the transition radiation tracker (TRT) in the ATLAS inner tracking detector. As sMDT pulses are approximately four times longer than the TRT straw tube signals [60], the value of the input capacitor had to be adapted for the BLR circuit. A transistor² is used as diode to make parasitic capacities small, as they would distort the high frequency behavior. The absolute value of the baseline depends on the voltage drop at R_6 and, hence, on the BLR voltage. A reference voltage V_{REF} is coupled

¹ Amplifier/Shaper/Discriminator with BaseLine/Restoration [59]

² NXP Semiconductors, BFT92, PNP 5 GHz wideband transistor [61]

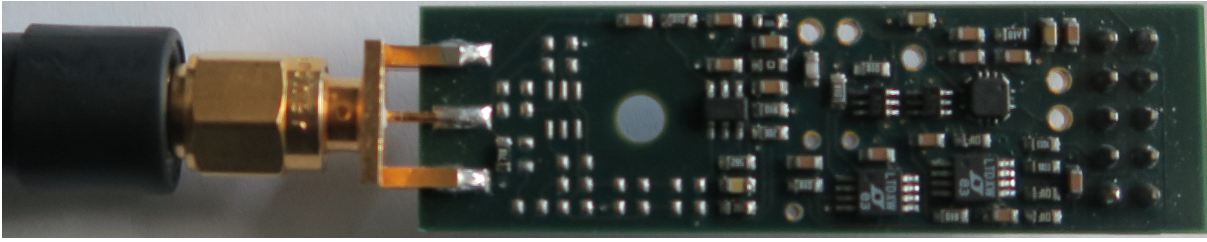


Figure 4.8: Photograph of a unipolar shaper channel with an SMA input connector on the left and the supply voltage and signal output connectors to the motherboard on the right.

to the signal line via R_7 in order to reduce the influence of V_{BLR} on the baseline level. The signal is discretized by a comparator¹ (discriminator) with an LVDS output to the TDC. The comparator threshold V_{THR} can be set externally and has to be adjusted for each BLR voltage setting.

Figure 4.8 shows a photograph of a unipolar shaper board. Up to 24 of these shaping boards are mounted on a motherboard, which provides the supply voltages to the shaper boards generating the voltage levels for all shaper boards with commercial DACs². The threshold voltage for each shaper is generated by one DAC channel while V_{BLR} and V_{REF} are generated by another DAC channel each. The motherboard is controlled by a TDC board, which also reads out the LVDS signals from the comparators.

4.3.4 Bipolar Shaping

Bipolar shaping (cf. [47]) prevents positive baseline shift. A fast signal differentiation by a CR filter with a time constant $\tau \approx 100$ ns, produces a large, but short, undershoot of the same area after each positive input pulse. Figure 4.9 shows the pileup of two subsequent generic input pulses for bipolar shaping. Compared to the AC coupled unipolar shaper, the undershoot is in general significantly shorter but deeper increasing the delay of the threshold crossing of a pulse piling up onto the undershoot of a previous pulse. The pileup effects of several subsequent pulses do not add up to a baseline shift. Hence, bipolar pulse shaping provides good baseline stability, but leads to stronger distortions of the piling up signal pulse amplitudes and threshold crossing times.

Bipolar shaping is the second shaping concept studied for the readout of drift tubes with 15 mm diameter at high rates. The circuit diagram is shown in Figure 4.10. Like in the case of unipolar shaping, the signal output of the preamplifier board is injected via an SMA connector and terminated with $50\ \Omega$ (R_0 and R_1). The first stage is a capacitively coupled (C_0) inverting amplifier³ with a pole-zero network (R_2 , R_3 and C_1) for ion tail cancellation. The second stage differentiates the signal based on $C_1 = 220$ pF. The

¹Analog Devices, ADCMP605 [62]

²ANALOG DEVICES, LTC2637 [63]

³Analog Devices, AD8614 [64]

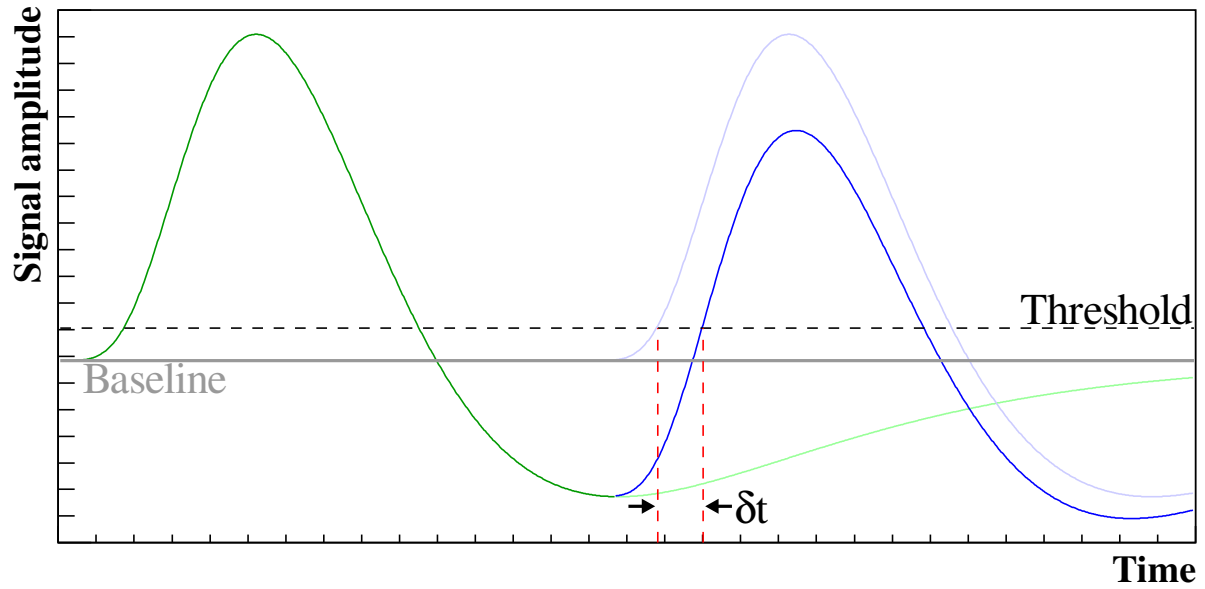


Figure 4.9: Illustration of signal pileup for bipolar shaping. The second signal (blue) adds upon the deep undershoot of the previous hit. Due to this pileup, the output amplitude of the second hit is reduced and the threshold crossing time delayed by δt .

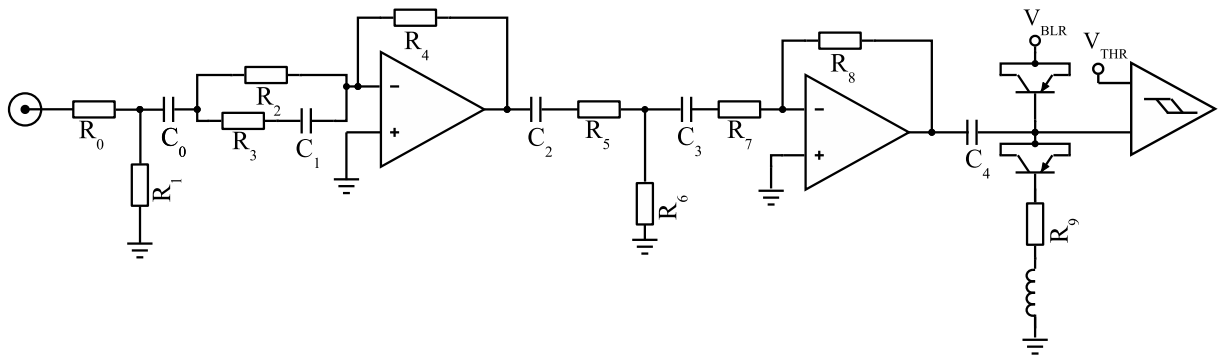


Figure 4.10: Simplified circuit diagram of the discrete bipolar shaper. The detailed diagram with values of the components and additional buffering capacitors is shown in Figure A.2.

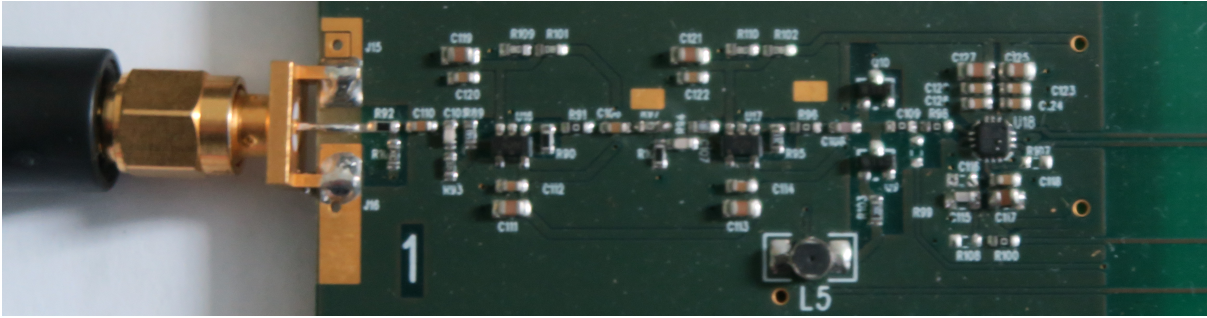


Figure 4.11: Photograph of a bipolar shaper channel with a SMA input connector on the left. Twelve channels are integrated on one board, on which the BLR and threshold voltages are generated.

design and the parameters are chosen to emulate the shaping of the ASD chip and are described in detail in [65]. The BLR stage works like the unipolar one, just a second diode is used to compensate temperature effects. Finally, the signal is discretized by the same comparator like used for the unipolar shaper with LVDS output to the TDC. Referring to the last active stage (cf. shaper stage response in Section 5.1.2.2), this circuit is called bipolar shaper in the following.

Figure 4.11 shows a photograph of a bipolar shaper channel. Twelve channels are integrated on one board with a central LV power supply [66]. The threshold and BLR voltages are generated separately by three octal DAC¹ for each channel. The DAC chips are controlled by a TDC board, which also reads out the LVDS signals from the comparators.

¹Linear Technology, LTC2657 [67]

Chapter 5

Characterization of BLR Readout Electronics

The fundamental behavior of the analog part of the readout electronics has been analyzed with injected pulses in a table-top setup in the lab. Systematic studies were performed to find the optimal operating parameters.

5.1 Studies with Pulse Generators

The electronic pulses from muon and γ rays traversing drift tubes vary vastly in shape, length and amplitude, depending on the energy deposition along the path in the drift tube and on the distance to the wire. To analyze and compare the response of the circuits with unipolar and bipolar shaping stages at different operating points, reproducible input pulses are generated by a pulse generator.

5.1.1 Experimental Setup

First, measurements were performed with δ like input signals produced by a 150 MHz pulse pattern generator¹ with 5 ns minimum pulse length². The shapes of the δ pulses with different amplitudes are shown in Figure 5.1. Additionally, as pulses from the 15 mm diameter drift tubes can be up to almost 200 ns long, 20 ns, 50 ns, 100 ns and 200 ns long rectangular pulses with amplitudes of 10 mV, 20 mV, 50 mV and 100 mV (Figure 5.2) were injected into the readout electronics. Real tube pulses do not have a steep falling edge. To be more realistic, triangular pulses were used to approximate the shape of γ background pulses. Figure 5.3 shows the average of 1000 γ pulses from a ^{137}Cs source, recorded in the Gamma Irradiation Facility GIF⁺⁺ at CERN [68] at low irradiation rate and, therefore, with negligible

¹Hewlett Packard 8110A Pulse Pattern Generator, max. 150 MHz

²Time between the crossings of 10% of the pulse amplitude.

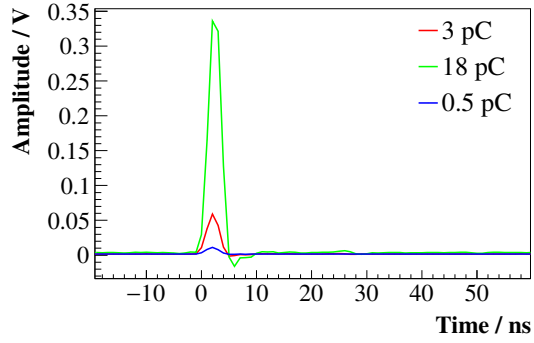


Figure 5.1: Examples of δ input pulses.

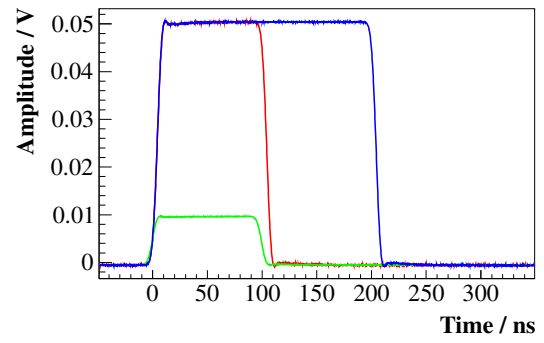


Figure 5.2: Examples of rectangular input pulses.

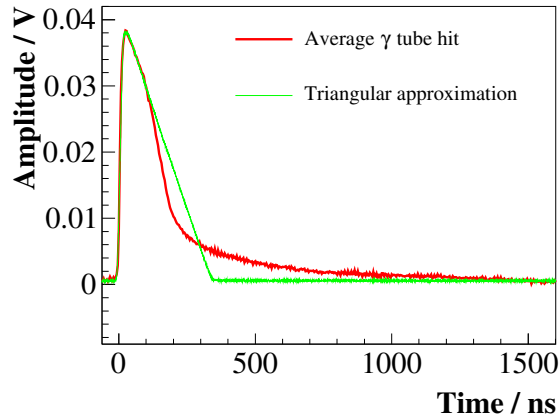


Figure 5.3: Average unshaped γ pulse and its approximation by a triangular pulse with the same rise time, amplitude and charge (area).

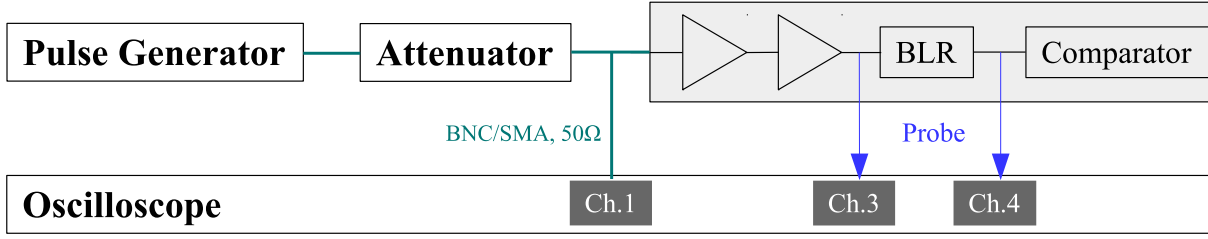


Figure 5.4: Schematic of the pulser test setup. The BNC and SMA connections have a matching impedance of $50\ \Omega$. All oscilloscope channels were set to an impedance of $1\ \text{M}\Omega$. Unipolar and bipolar shapers have been tested with pulses of varying amplitude, length and rate.

space charge effects (see also Section 6.1). The average pulse has a rise time¹ of $t_{\text{rise}} = 16\ \text{ns}$, an maximum amplitude of $V_{\text{max}} = 37.7\ \text{mV}$ and an integrated charge of $C = 126\ \text{pC}$. A triangular pulse with the same values was chosen to approximate it. According to [15], the average muon induced drift tube pulse has a similar shape and roughly 30 % of the amplitude. For the longer pulses, a different generator² was used which produces smaller pre- and post-pulse oscillations due to its smaller frequency range which is sufficient for the generation of the longer pulses with larger rise and falling times than the δ pulses.

For the δ pulse amplitude variations an attenuator was used, which keeps variations of the pulse shape negligible and provides excellent linearity between pulse height and charge. The rectangular pulses were varied in amplitude and length without using the attenuator. The setup is sketched in Figure 5.4, a photograph is shown in Figure 5.5. The input and output impedances of the pulse generator, attenuator and shaper boards are matched to $50\ \Omega$ to avoid signal reflections. The input signals are recorded with an oscilloscope with a high-impedance input (Ch. 1). The response of the second shaper and of the BLR stage are recorded using oscilloscope probes with $10\ \text{M}\Omega$ and $9.5\ \text{pF}$ (Ch. 3 and Ch. 4) to minimize signal distortions. This is especially important for the output of the second shaper stage as it is the input to the BLR stage.

5.1.2 Single-Pulse Response

In this section, the electronics response to inputs with low repetition rates is characterized. Due to the large interval between subsequent pulses, the response to single pulses can be studied without pileup effects to determine the effects of varying input amplitude, pulse length and charge on the output signal. Signal gain, saturation levels, output pulse shape and rise time are determined for the unipolar and bipolar shapers with different BLR voltages.

¹Time between 10 % and 90 % of the maximum amplitude.

²Agilent 33500B Series Waveform Generator, max. 30 MHz

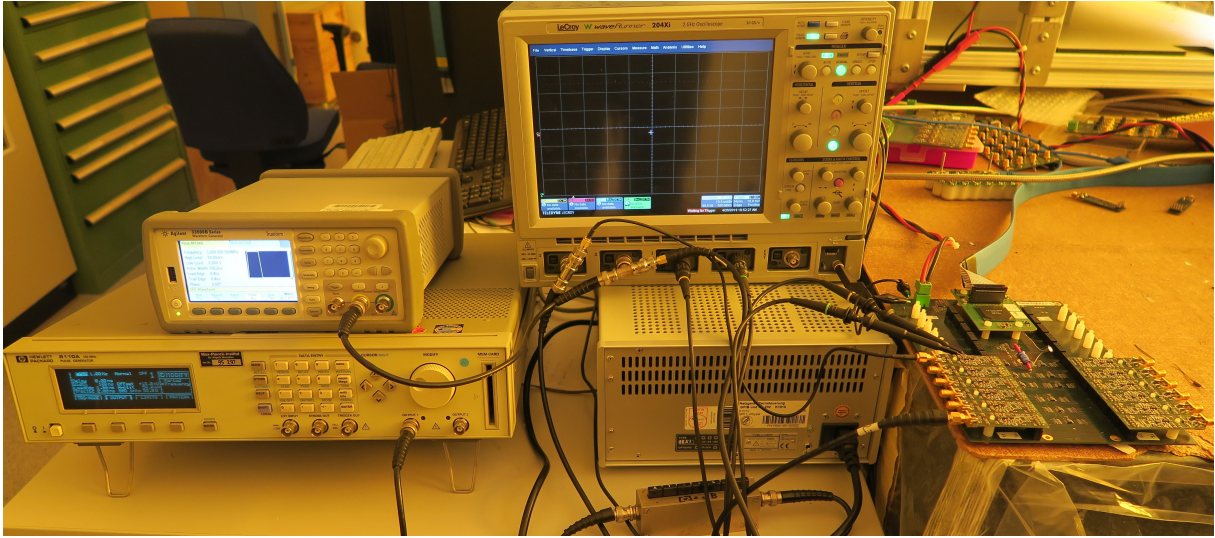


Figure 5.5: Photograph of the pulser test setup with the two pulse generators (left), the attenuator (middle) and the oscilloscope with its two probes on a test board (right).

5.1.2.1 Unipolar Shaper

- Active Shaper-Stage Response to δ pulses

Figure 5.6 shows the response¹ of the second shaper stage of the unipolar shaper to δ pulses of different amplitudes. A comparison of the effects of the different stages is shown in Figure C.1. The unipolar shaper lengthens the pulse from 4.2 ± 0.4 ns to 10.8 ± 0.4 ns at constant gain of 13 for small input charges up to about 5 pC, due to the given bandwidth limitations above 100 GHz. For larger input charges, saturation effects set in decreasing the gain (see Figure 5.7) and increasing the pulse length. Up to 5 pC input charge, corresponding to 100 mV δ amplitude, the shaper shows excellent linearity.

- BLR-Stage Response to δ pulses

The output of the second shaper stage provides the input to the BLR stage (see Section 4.3.3). Figure 5.8 shows the shape of the output of the BLR stage of the unipolar shaper to 3 pC δ pulses for different BLR voltages V_{BLR} . Due to the capacitive coupling of the BLR stage to the shaper stage leading to a differentiation of the pulse, a long undershoot of the same area as the positive signal is expected but

¹To allow for direct comparison between the stages, BLR voltages and shaping types, all the output shapes of the BLR and active stages are shown for AC coupling, i.e. with the baseline set to 0 V. The absolute position of the baseline depends on the BLR voltage.

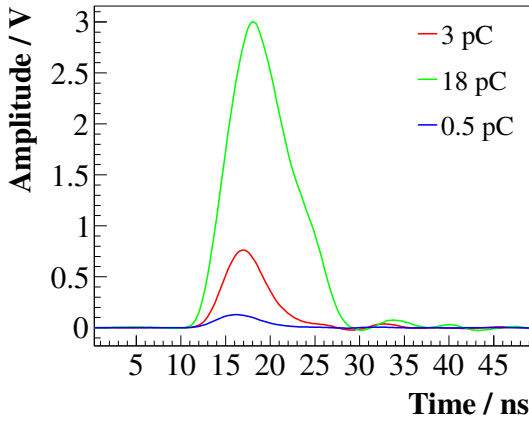


Figure 5.6: Response of the last active stage of the unipolar shaper to δ pulses of different amplitudes with constant rise and fall times.

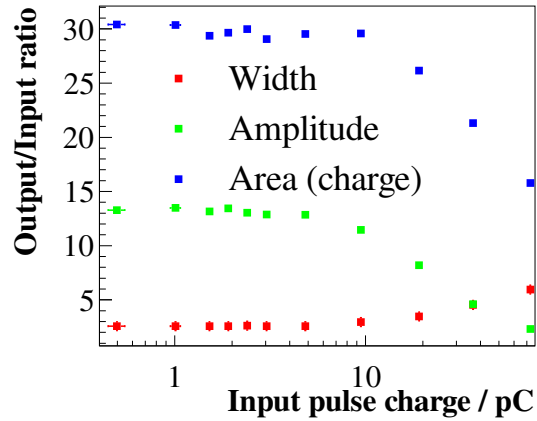


Figure 5.7: Linearity of the last active stage of the unipolar shaper response to δ like input pulses.

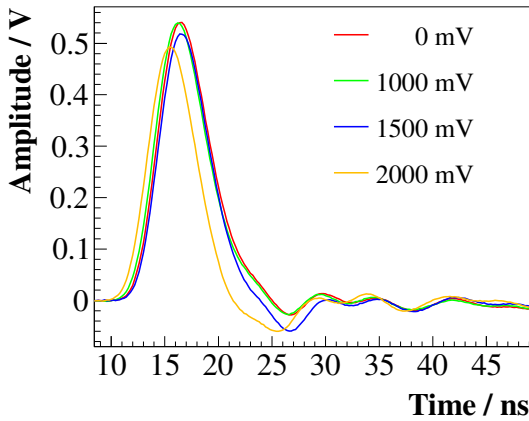


Figure 5.8: Response of the unipolar shaper with BLR stage with different BLR voltages to δ pulses with 3 pC charge.

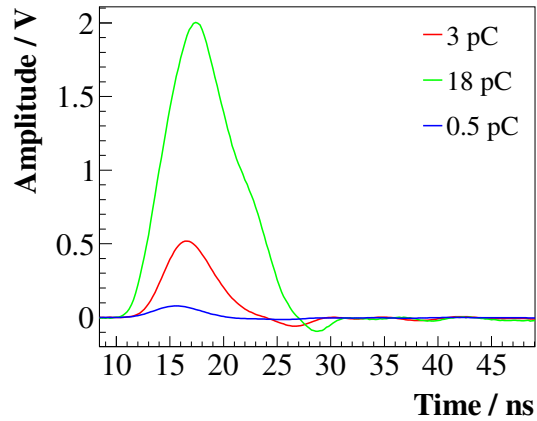


Figure 5.9: Output pulse shape of the unipolar shaper with BLR stage with $V_{BLR} = 1500$ mV for δ pulses of different charge.

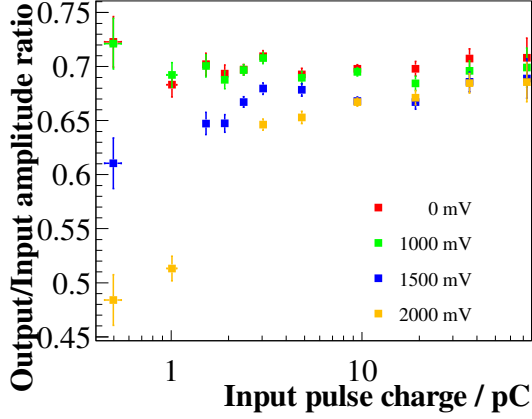


Figure 5.10: Amplitude response of the BLR stage of the unipolar shaper with different BLR voltages to δ like input pulses to the shaper.

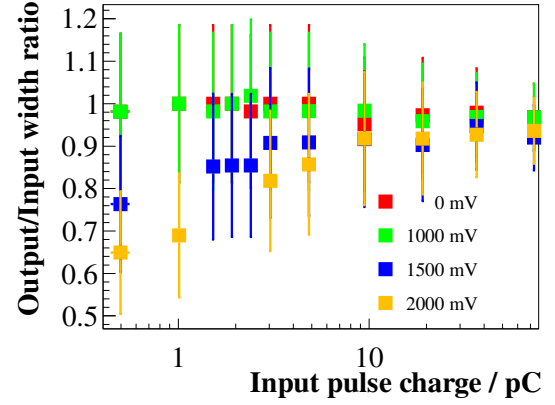


Figure 5.11: Output-to-input pulse length ratio of the BLR stage of the unipolar shaper with different BLR voltages as a function of the shaper input charge. The length of the input δ pulse was 5 ns.

can barely be seen, as it is overlaid by the post pulse oscillations of the pulse generator. The amplitude, however, is reduced due to charge losses through the parasitic capacities of the transistor.

The short and deep undershoot for high $V_{BLR} \geq 1500$ mV is a consequence of the slow reaction of the DAC providing the BLR current to the steep voltage drop of the δ pulse, which does not occur for drift tube signals. Due to the exponential characteristic of the diode, this undershoot increases with the BLR voltage and input amplitude (cf. Figure 5.9).

Figure 5.10 shows the output-to-input amplitudes ratio of the BLR stage for different input charges and BLR voltages¹. For $V_{BLR} \leq 1000$ mV, the pulse attenuation is constant at 70 %. For larger BLR voltages, the amplitudes are reduced more strongly, especially for input charges below 3 pC, as they do not drive the diode completely into the non-conductive region (cf. Section 4.3.3). For $V_{BLR} \leq 1000$ mV, the pulse length is also constant after the BLR stage (cf. Figure 5.11). Higher BLR voltages lead to increased charge losses through the pull down resistor (R_6 in Figure 4.7) leading also to a pulse shortening, especially for small input charges below 3 pC. Both, the amplitude and the shortening, contribute to a pulse area reduction above $V_{BLR} = 1500$ mV (cf. Figure 5.12). The output signal rise time of the unipolar shaper together with the BLR stage, which is crucial for the drift time and spatial resolution, is not deteriorated up to input charges of 5 pC (cf. Figure 5.13).

¹To allow for direct comparison between the BLR stages of the unipolar and the bipolar shaper, the ratios are given with respect to the BLR input, i.e. the output of the last active shaper stage, but as a function of the input to the shaper.

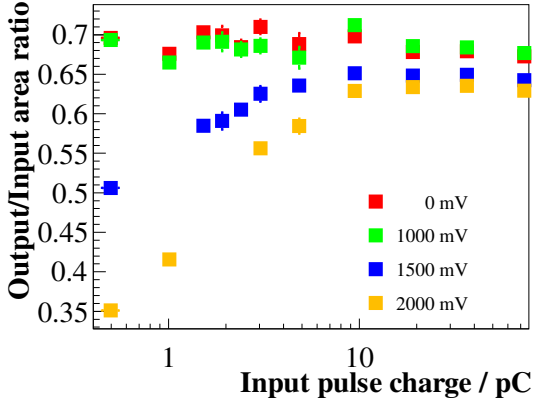


Figure 5.12: Output-to-input pulse area (charge) ratio of the BLR stage of the unipolar shaper with different BLR voltages as a function of the shaper input charge.

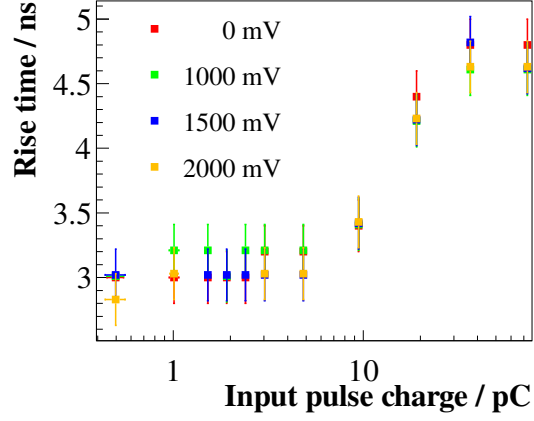


Figure 5.13: Output pulse rise time of the BLR stage of the unipolar shaper with different BLR voltages as a function of the input charge. The rise time of the input δ pulse was 2 ns.

- Active Shaper-Stage Response to Rectangular Pulses

Figure 5.14 shows the response of the second shaper stage of the unipolar shaper to the three rectangular input pulses shown in Figure 5.2. The short fall time of the positive pulse and the undershoot, which increases with the input charge, are caused by the capacitive coupling between the two shaper stages. Due to the large capacitance of 10 nF, this effect is not visible for the δ pulses, only for the longer rectangular pulses. The gain of the unipolar shaper stages of 31.1 ± 0.6 is constant for input amplitudes $U_{in} < 100$ mV, i.e. below saturation (see Figure 5.15), even for large input pulse charges. The difference in the behavior of δ pulses of large input charge (see Figure 5.7) is due to the bandwidth limitations of the circuit. For the highest input amplitude of 100 mV, saturation effects start and the gain is significantly reduced.

- BLR-Stage Response to Rectangular Pulses

The response of the unipolar shaper with BLR stage with $V_{BLR} = 1500$ mV to the three rectangular input pulses of Figure 5.2 is shown in Figure 5.16. Due to the small coupling capacitance of $C_2 = 30$ pF (cf. Figure 4.7) between the last active shaper stage and the BLR stage, the pulse amplitude drops significantly during the duration of the input pulse. The undershoot is efficiently suppressed (compare with Figure 5.14). The undershoot suppression of the BLR stage is shown in Figure 5.17 as a function of the BLR voltage. For zero BLR voltage, the output shape equals the one of simple capacitive coupling.

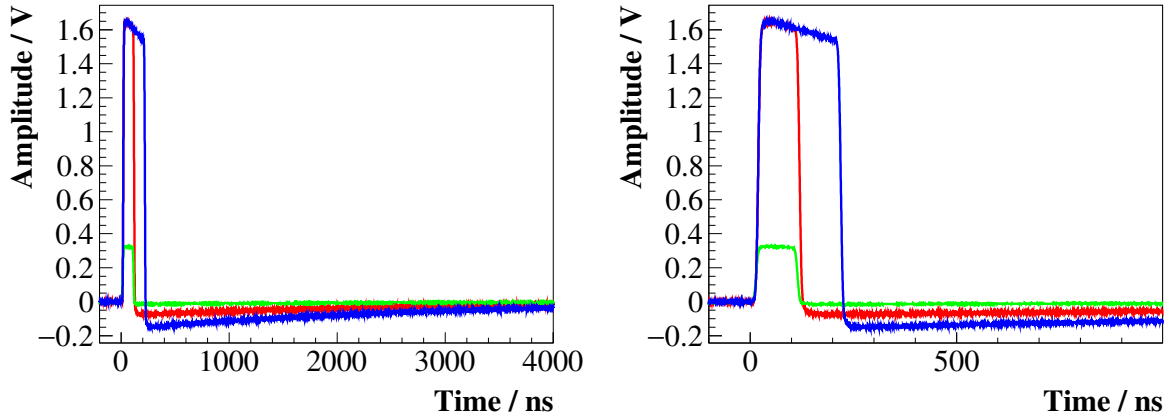


Figure 5.14: Response of the last active unipolar shaper stage to the rectangular input pulses of different amplitudes and widths but equal rise and fall times shown in Figure 5.2.

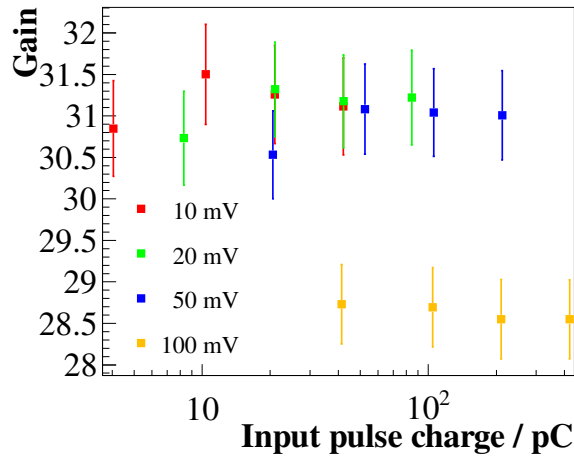


Figure 5.15: Amplitude response of the unipolar shaper stages to rectangular input pulses of different amplitudes as a function of the input charge which is varied via the pulse length.

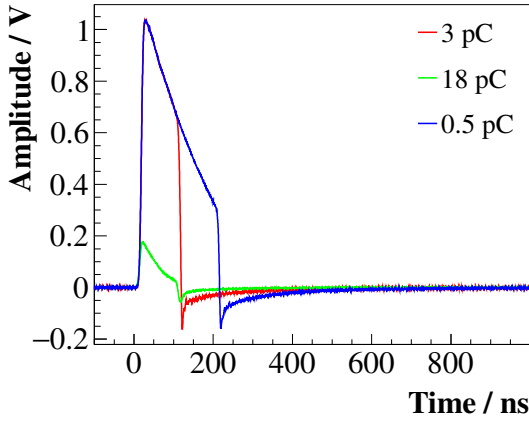


Figure 5.16: Response of the unipolar shaper with BLR stage with $V_{BLR} = 1500$ mV to the rectangular input pulses of different amplitudes and widths but equal rise and fall times shown in Figure 5.2.

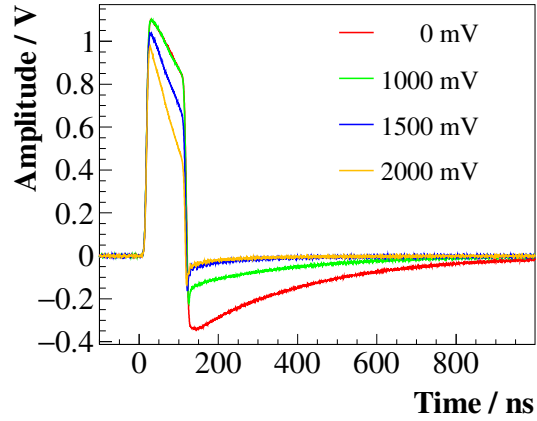


Figure 5.17: Response of the unipolar shaper with BLR stage with different BLR voltages to rectangular input pulses of 100 ns length and 50 mV amplitude.

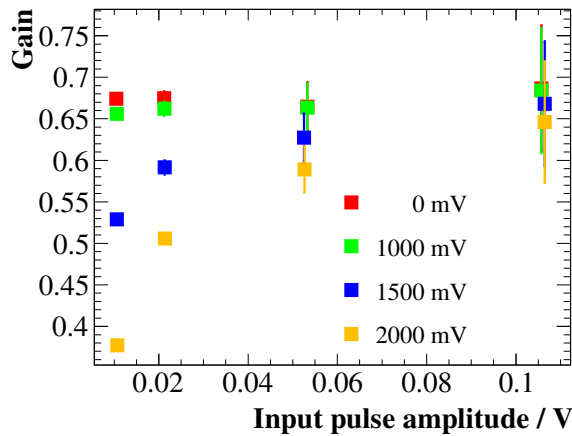


Figure 5.18: Gain of the BLR stage of the unipolar shaper with different BLR voltages for rectangular input pulses of 100 ns width.

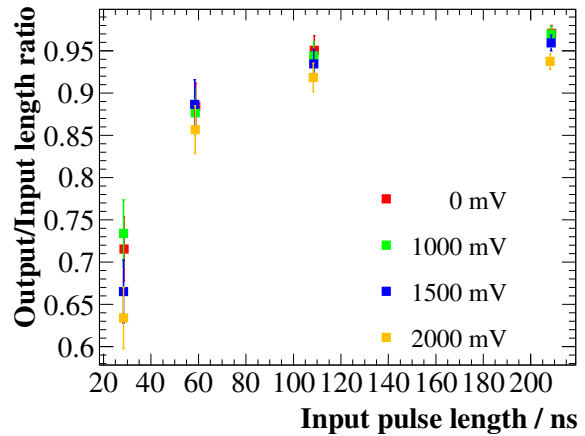


Figure 5.19: Output-to-input length ratio of the BLR stage of the unipolar board with different BLR voltages for rectangular input pulses of 50 mV amplitude.

Already for $V_{\text{BLR}} = 1000 \text{ mV}$, the undershoot amplitude is significantly reduced. For higher BLR voltages, the undershoot is essentially eliminated, but the positive signal amplitude is also attenuated and falls more steeply during the duration of the input pulse. This gain reduction increases with the voltage drop across the diode and, therefore, with V_{BLR} , because the current drain through the pull down resistor R_6 in Figure 4.7 is determined by the exponential voltage-current characteristic of the diode leading significantly to larger current loss at high BLR voltages (see Figure 5.18). Figure 5.18 also shows that for high BLR voltages $V_{\text{BLR}} \geq 1500 \text{ mV}$ small input amplitudes are not sufficient to push the diode into the non-conducting region, which leads to a further gain reduction. The BLR stage also slightly reduces the pulse length especially for short pulses (cf. Figure 5.19).

The extent of the undershoot reduction is shown in Figure 5.20. Without applying a BLR voltage, the signal and undershoot area are equal, because one has just the response of the coupling capacitor. The small deviation from unity for large pulses is a consequence of saturation effects. For $V_{\text{BLR}} = 1000 \text{ mV}$, the increase of the undershoot amplitude with the input charge leads to a stronger reduction of the undershoot area with increasing input charge, but not yet a full compensation. For 1500 mV BLR voltage, the undershoot is reduced to the same level of 10 to 20 % for all, even small, input charges. The exponential diode characteristic leads to a stronger undershoot reduction for larger pulse amplitudes. For $V_{\text{BLR}} = 2000 \text{ mV}$, the charge loss through the pull down resistor is sufficient to fully discharge the capacitor of the BLR stage before the end of input pulses with small amplitudes, reducing the positive pulse area (charge). This leads to an increase of the undershoot-to-signal charge ratio. The effect is already visible for $V_{\text{BLR}} = 1500 \text{ mV}$.

- Response to Triangular Pulses

Figure 5.21 shows the response of the second unipolar shaper stage to the triangular input pulse shown in Figure 5.3. The pulse shape and length are conserved. The capacitive coupling between the two shaper stages leads to an undershoot after the signal, which has a small amplitude because of the large coupling capacitance of $C_1 = 10 \text{ nF}$ (cf. Figure 4.7), but is longer than the time window shown. This small voltage drift is essentially removed by differentiation due to the small $C_2 = 30 \text{ pF}$ coupling capacitor to the BLR stage, leading to a shorter undershoot of larger amplitude. Hence, the time window shown in Figure 5.22 is now sufficient for signal return to the baseline. With zero BLR voltage, the undershoot is drained only by discharging of the capacitor C_2 through the resistors R_6 and R_7 in Figure 4.7. For larger BLR voltages, additional negative charge is drained with the reduction of current flow through the diode (see Section 4.3.3). Already for $V_{\text{BLR}} = 1500 \text{ mV}$, the undershoot is essentially eliminated. The signal amplitude is, however, also reduced with increasing BLR voltage above 1000 mV . The comparison with Figure 5.17 shows that the BLR stage of the unipolar shaper works better if the unrealistically steep falling

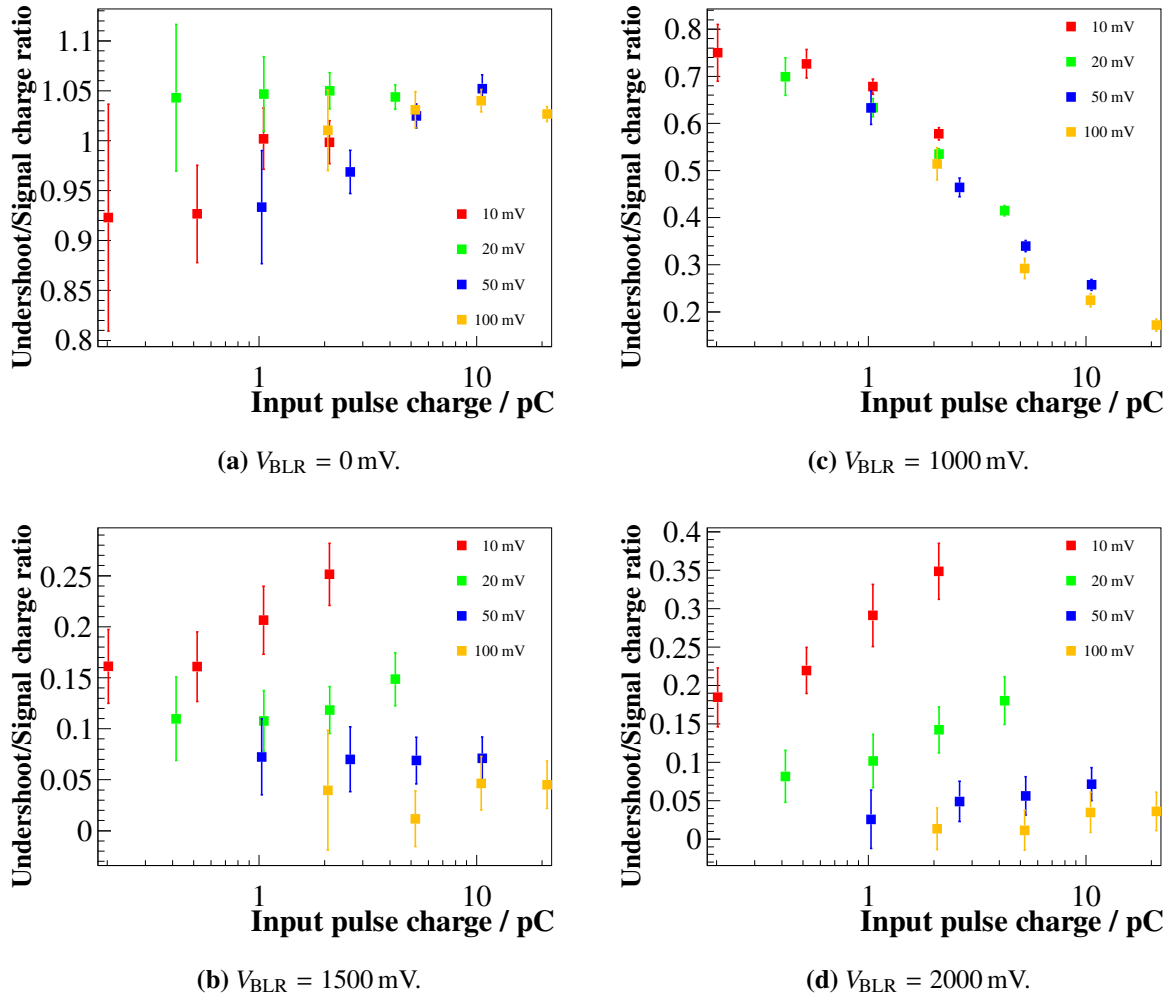


Figure 5.20: Undershoot-to-signal charge (area) ratio of the BLR stage of the unipolar shaper with different BLR voltage settings for different input pulse amplitudes. Changes in input signal length lead to variations of the input charges at a constant input amplitude. Rise and fall times of the input pulses are constant. A summary of the results for all parameter variations is shown in Figure C.3.

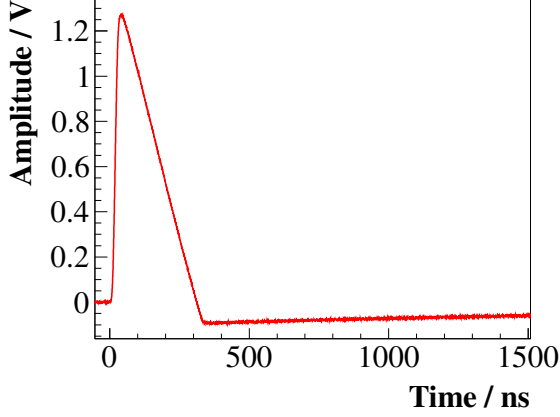


Figure 5.21: Response of the unipolar shaper to the triangular input pulse.

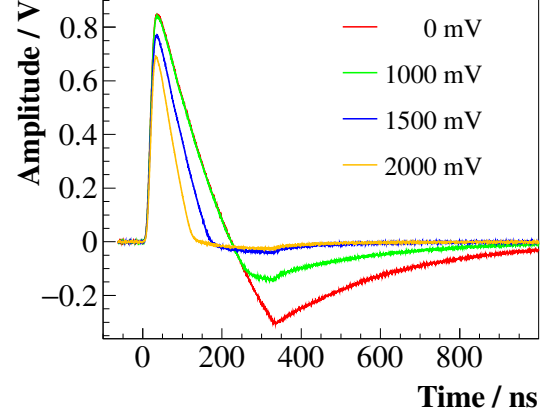


Figure 5.22: Response of the unipolar shaper with BLR stage with different BLR voltages to the triangular input pulse.

edge of the input pulse is replaced by a more realistic slower ramp down. Concerning the pulse shape, small BLR voltage is beneficial, as pulse shape deformations increase with the BLR voltage, especially for $V_{BLR} \geq 1500$ mV.

The coupling capacitor of C_2 (cf. Figure 4.7) between the last active shaper stage and the BLR stage is crucial for the pulse shaping of the unipolar shaper board. A smaller capacitance results in a stronger differentiation of the signal and a smaller undershoot, while a larger capacitance sustains more signal amplitude, leading to a larger undershoot. The discharge of positive through R_6 and R_7 lead to a time constant of $\tau = C_2 \cdot (R_6 || R_7) \approx 200$ ns matching the maximum drift time of the sMDT. Negative charge is drained up to two orders of magnitude faster through the diode, depending on the amplitude. Fine-tuning of C_2 requires consideration of realistic pulse shapes and is only possible with elaborated studies e.g. based on the simulation described in Section 6.8.

5.1.2.2 Bipolar Shaper

- Active Shaper-Stage Response to δ pulses

Figure 5.23 shows the response of the second stage of the bipolar shaper to δ pulses of different amplitudes, which has the characteristic long undershoot of the same area as the signal. A comparison of the effect of the different stages is shown in Figure C.1, a comparison to the unipolar shaper in Figure C.2. The signal with large amplitude is cut off as a consequence of the amplifier saturation, which occurs for δ pulses

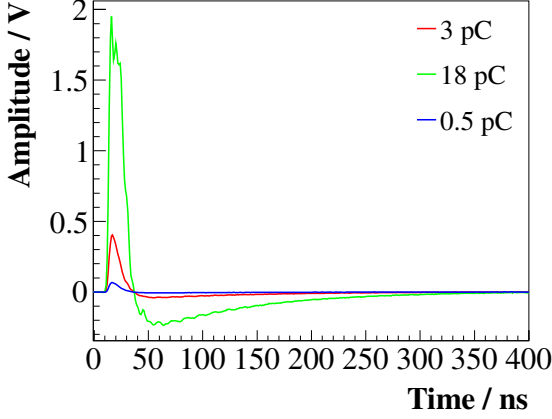


Figure 5.23: Response of the last active stage of the bipolar shaper to δ pulses of different amplitudes with constant rise and fall times.

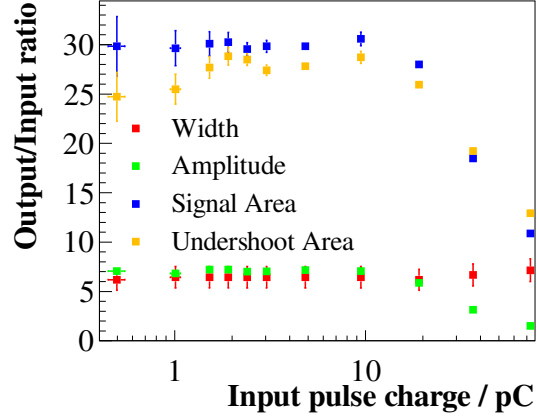


Figure 5.24: Linearity of the last active stage of the bipolar shaper response to δ like input pulses.

of more than 10 pC corresponding to 200 mV amplitude (see Figure 5.24). For small input pulses, the gain of the two bipolar stages shows very good linearity in the amplitude response and a constant pulse broadening from 4.2 ns to 27 ns due to bandwidth limitations. The area of the positive pulse is integrated from the time when the rising edge crosses 10 % of the maximum amplitude to the baseline crossing of the falling edge. The undershoot is integrated from the latter up to 10 % of its maximum amplitude in the tail. As the slope of the undershoot tail is smaller, a larger fraction of its area is neglected in the integration than for the unipolar shaping, which leads to larger pulse compared to undershoot areas for this method. For input pulses larger than 20 pC, saturation reduces the gain and lengthens the pulses further (see Figure 5.23). The gain drop leads to a reduction of the output to input area (charge) ratio for the positive signal and, because of the amplitude reduction, to a smaller undershoot amplitude.

• BLR-Stage Response to δ pulses

Like the unipolar signal, the bipolar signal shape is not significantly changed by the BLR stage, if its control voltage is turned off (Figure 5.25). Based on previous studies [65], which showed that 1800 mV is the optimum V_{BLR} setting, the BLR voltage was set in smaller steps to 1700 mV, 1800 mV and 1900 mV. The resulting output shapes at these BLR voltage settings are similar to each other. In comparison with the case of $V_{\text{BLR}} = 0$ mV, the positive signal amplitude is slightly attenuated and the undershoot strongly reduced. The shape of the δ response for different amplitudes at $V_{\text{BLR}} = 1800$ mV is shown in Figure 5.26.

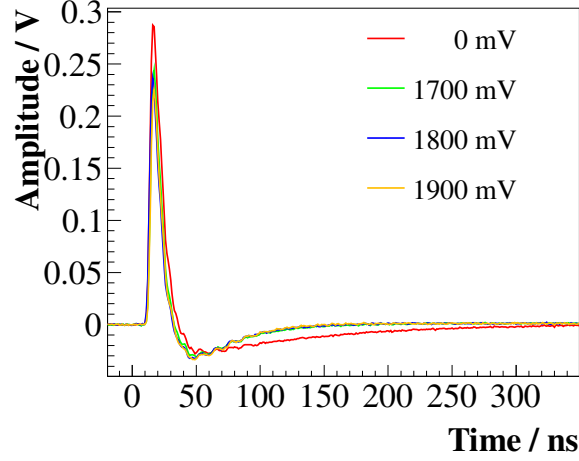


Figure 5.25: Response of the bipolar shaper with BLR stage with different BLR voltages to δ pulses with 3 pC charge.

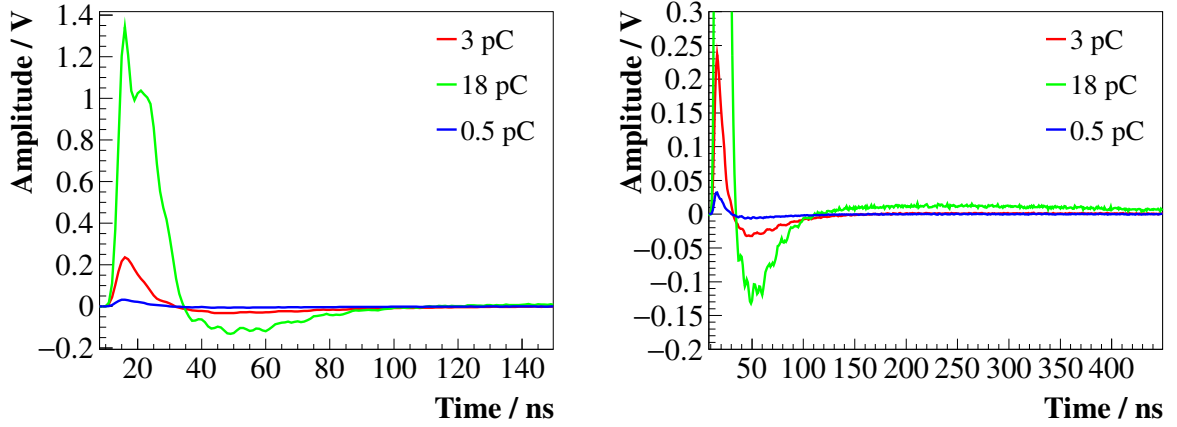


Figure 5.26: Output pulse shape of the bipolar shaper with BLR stage with $V_{BLR} = 1500$ mV for δ pulses of different charge.

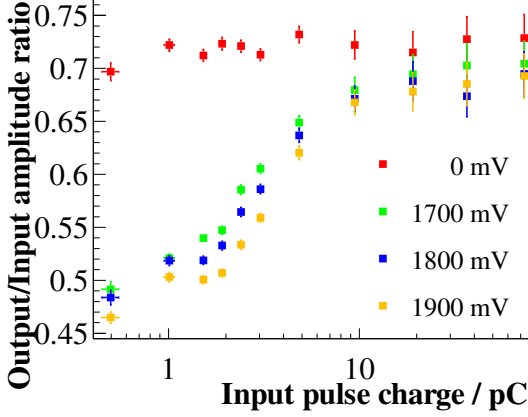


Figure 5.27: Amplitude response of the BLR stage of the bipolar shaper with different BLR voltages to δ like input pulses to the shaper.

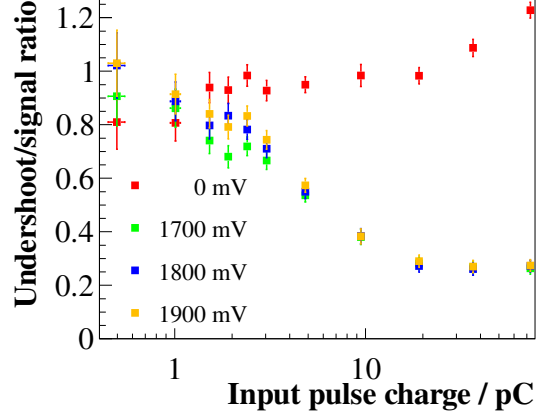


Figure 5.28: Undershoot-to-signal charge (area) ratio of the BLR stage of the bipolar shaper as a function of the input charge for δ like input pulses to the shaper.

For small input pulses the BLR stage reduces the undershoot length efficiently but for the 18 pC δ pulse, a second overshoot occurs after the undershoot, which has already been observed before [69, 70], and been attributed to parasitic capacities on the board [65] or in the test setup [70]. The studies here indicate that this second overshoot is due to the discharge of the capacitor right in front of the BLR stage, which is charged up through the transistor during negative signal amplitudes in the undershoot. Simulations with LTspice [71] prove that the effect of parasitic capacities and inductances is negligible.

For $V_{BLR} > 0$ mV, the BLR stage output amplitude is not linear with the input amplitude, as it is dominated by the exponential current-voltage characteristic of the diode. For $V_{BLR} = 0$ mV, the BLR stage acts like an attenuator, due to parasitic capacities and ohmic loss. Higher BLR voltages lead to a further amplitude reduction, because more signal charge is drained to ground through the transistor (cf. Figure 5.27). This effect is smaller for larger pulses due to the exponential diode characteristics.

The extent of the undershoot removal for δ input pulses is shown in Figure 5.28. For $V_{BLR} = 0$ mV, signal and undershoot have very similar absolute charge. Ohmic losses for small input charges and saturation effects for large input charges only lead to small deviations from equal area. With the BLR voltage turned on, the undershoots are efficiently reduced for medium and large pulses. For small input charges and amplitudes, the conductivity of the diode is not shifted far enough from its operation point to appreciably reduce the undershoot.

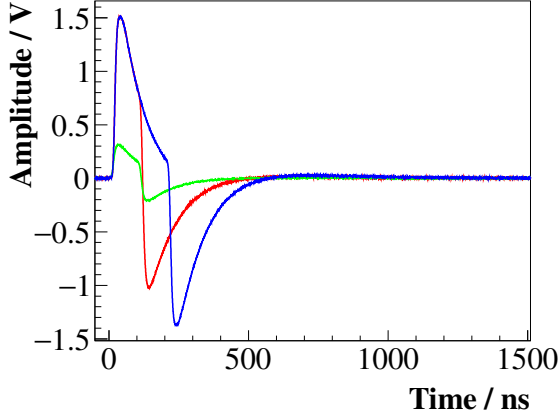


Figure 5.29: Response of the last active bipolar shaper stage to the rectangular input pulses of different amplitudes and widths but equal rise and fall times shown in Figure 5.2.

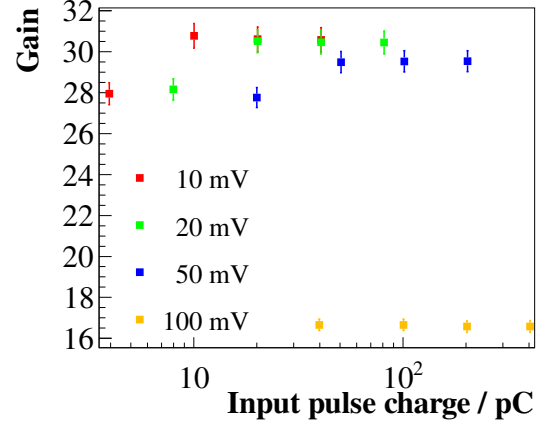


Figure 5.30: Amplitude response of the bipolar shaper stages to rectangular input pulses of different amplitudes as a function of the input charge which varies with the pulse length.

- Active Shaper-Stage Response to Rectangular Pulses

Figure 5.29 shows the typical bipolar response of the second bipolar shaper stage of the bipolar board for the three rectangular input pulses shown in Figure 5.2. The undershoot grows linearly with input pulse length and amplitude, i.e. charge. The linearity of the circuit can also be seen from the input charge dependence of the ratio of maximum output to input voltage (Figure 5.30). Pulses with small input amplitude which are longer than 20 ns are amplified by 30.6 ± 0.6 . For very short pulses, the gain is reduced due to bandwidth limitations (cf. Figure 5.24). For input amplitudes of 50 mV and larger, saturation effects appear. The gain drops to 16.6 ± 0.6 for 100 mV input amplitude. The double differentiation and signal reflections between C_2 and C_3 in Figure 4.10 lead to undershoot-to-signal charge (area) ratio larger than 1 (cf. Figure 5.31), which is compensated by an additional overshoot after the undershoot. For input amplitudes, which drive the circuit far into saturation, the deviation from unity is larger than 10 %. Due to the large length of these reflections, their amplitude is so small, that they are not visible in Figure 5.29 and only detectable, with the setup used, for large input pulse amplitudes. Under realistic operating conditions, the amplitudes of the additional overshoots are smaller than the electronic noise amplitude and, hence, negligible. The length of the undershoot¹ is determined by the time constant $\tau = C_3 \cdot R_7 \approx 120$ ns of

¹Time between the crossings of the baseline and 10 % of the amplitude.

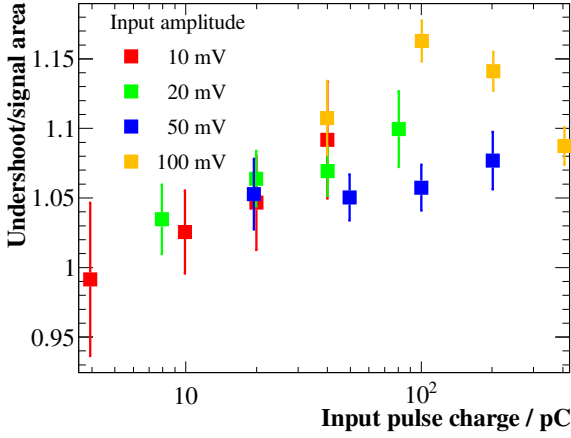


Figure 5.31: Undershoot-to-signal charge (area) ratio of the active stages of the bipolar shaper for different rectangular input pulse amplitudes. Changes of the input signal length lead to variations of the input charge for given input amplitude. Rise and fall times of the input pulses are equal.

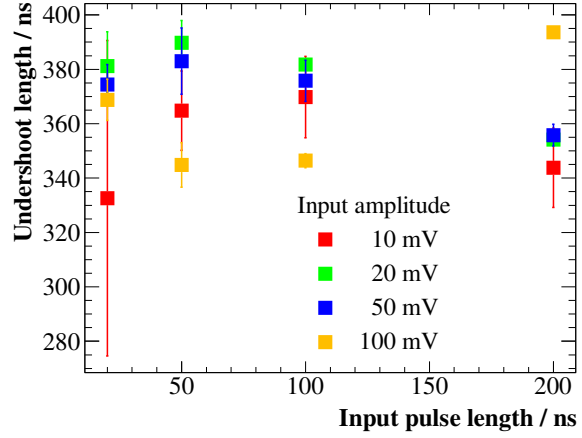


Figure 5.32: Dependence of the undershoot length after the active bipolar shaping on the input pulse length for rectangular input pulses with four different amplitudes.

the bipolar shaper stage. It is about $3 \cdot \tau \approx 360$ ns, independent of the input amplitude and length (cf. Figure 5.32).

• BLR-Stage Response to Rectangular Pulses

Figure 5.33 shows the pulse shape response of the BLR-stage of the bipolar shaper with medium $V_{\text{BLR}} = 1800$ mV to the three rectangular input pulses in Figure 5.2 with shaper response shown in Figure 5.29. While the signal pulse shape is not significantly changed, only reduced in amplitude, the tail of the pulse has a new shape. The length of the undershoot from the last shaper stage is strongly reduced while two new overshoots appear. The first of these overshoots is caused by the regulation of the DAC that delivers the BLR voltage. Due to the sharp voltage drop at the end of the input pulse, the BLR diode quickly draws more current from the DAC causing its voltage to drop. Due to the DAC regulating time constants the undershoot is overcompensated, leading to the first overshoot after the undershoot. The current through the diode charges the capacitor between the bipolar shaper and the BLR stage and only slowly discharges through R_9 (cf. Figure 4.10), leading to the second long overshoot, which is already visible for δ input pulses. Both overshoots and the remaining undershoot increase with amplitude and length of the input pulse.

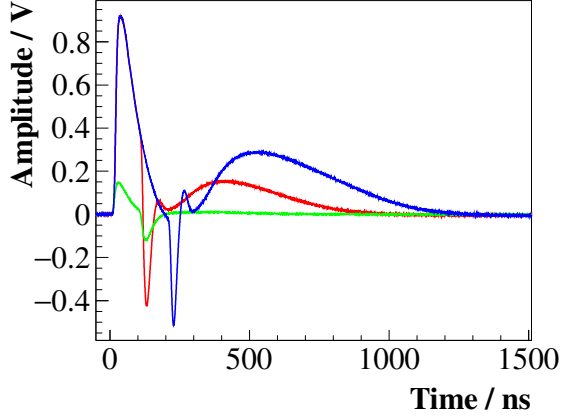


Figure 5.33: Response of the BLR stage of the bipolar shaper to the rectangular input pulses in Figure 5.2 with different amplitudes and widths and equal rise and fall times. The BLR voltage was set to $V_{BLR} = 1800$ mV.

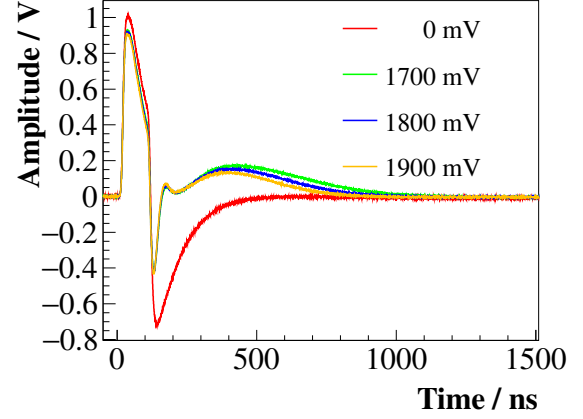


Figure 5.34: Response of the BLR stage of the bipolar shaper to a rectangular input pulse of 100 ns length and 50 mV amplitude for different BLR voltages.

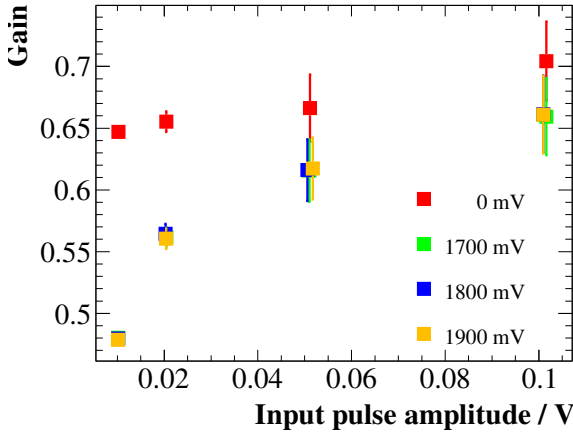


Figure 5.35: Gain of the BLR stage of the bipolar shaper as a function of the pulse amplitude for a rectangular input pulse of 100 ns to the shaper for different BLR voltages.

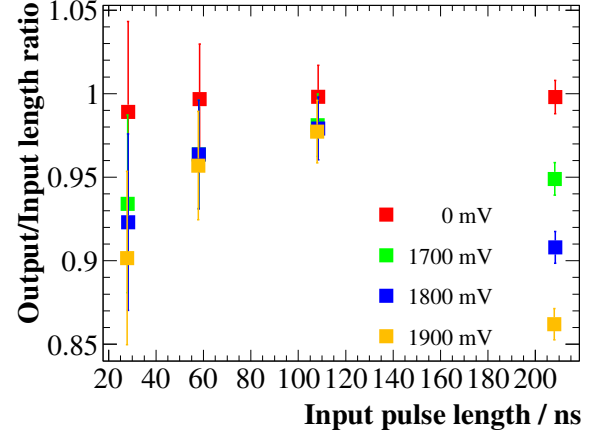


Figure 5.36: Output-to-input length ratio of the BLR stage of the bipolar shaper as a function of the pulse length for rectangular input pulses of 50 mV amplitude and different BLR voltages.

The influence of the BLR voltage on the output shape is shown in Figure 5.34. As for the unipolar case, the BLR stage of the bipolar shaper acts as an attenuator if its supply voltage is turned off. Small variations around $V_{BLR} = 1800$ mV do not change the output shape significantly, but for lower BLR voltages the signal amplitude is higher and the second overshoot has a larger amplitude and length, as less current is drained through R_9 (cf. Figure 4.10).

The ratio of output to input amplitudes (gain) is independent of the input amplitude for zero BLR voltage (Figure 5.35). Only large input amplitudes lead to a slight increase of the gain. When V_{BLR} is turned on, the non-linear behavior of the BLR stage shows up. Input pulses of smaller amplitudes are more strongly attenuated as more current is drained through the main BLR diode and R_9 (cf. Figure 4.10). The gain is insensitive to small variations of the BLR voltage.

The output pulse length is not changed by the BLR stage of the bipolar shaper for zero BLR voltage (cf. Figure 5.36). When V_{BLR} is turned on, the pulse length of short pulses is slightly reduced due to parasitic capacities of the BLR diodes. During long input pulses, the capacitor in front of the BLR stage (C_4 in Figure 4.10) is fully discharged leading to additional reduction of the output pulse length.

The effect of the BLR stage on the undershoot is shown in Figure 5.37. For zero BLR voltage, the When V_{BLR} is turned on, undershoot-to-signal area ratio of the active bipolar shaper stage (cf. Figure 5.31) is reproduced. The deviation from unity is slightly increased, because the capacitances of the two differentiating capacitors, C_3 and C_4 in Figure 4.10, differ less than C_2 and C_3 , which dominate the active bipolar shaper stage. For $V_{BLR} \geq 1700$ mV, the BLR stage reduces the undershoot length and amplitude. Due to the exponential diode characteristics, the reduction is more efficient for large undershoot amplitudes and, therefore, also for larger input pulse amplitudes. For longer input pulse lengths, the capacitor in front of the BLR stage (C_4 in Figure 4.10) is further discharged, leading to a larger undershoot amplitude at the end of the input pulse and, thereby, to a stronger undershoot reduction. For 200 ns long input pulses of amplitudes up to 20 mV, the capacitor in front of the BLR stage is fully discharged before the end of the input pulse limiting the undershoot reduction. Variations of the BLR voltage in the small range around 1800 mV cause no significant changes.

The dependence of the area (charge) of the second overshoot on the input pulse charge to the bipolar shaper is shown in Figure 5.38. Higher BLR voltages cause a larger second overshoot charge. The second overshoot charge increases more than linear with the input charge and increase stronger than the signal charge (cf. Figure 5.39). Large input amplitudes and especially long input pulses lead to strong baseline fluctuations for up to 1 μ s, while the effect of small input pulses is tolerable.

• Response to Triangular Pulses

Figure 5.40 shows the response of the active bipolar shaper stage to a triangular input pulse from Figure 5.3. The bipolar shaping makes the falling edge of the output pulse steeper than for the input pulse and the

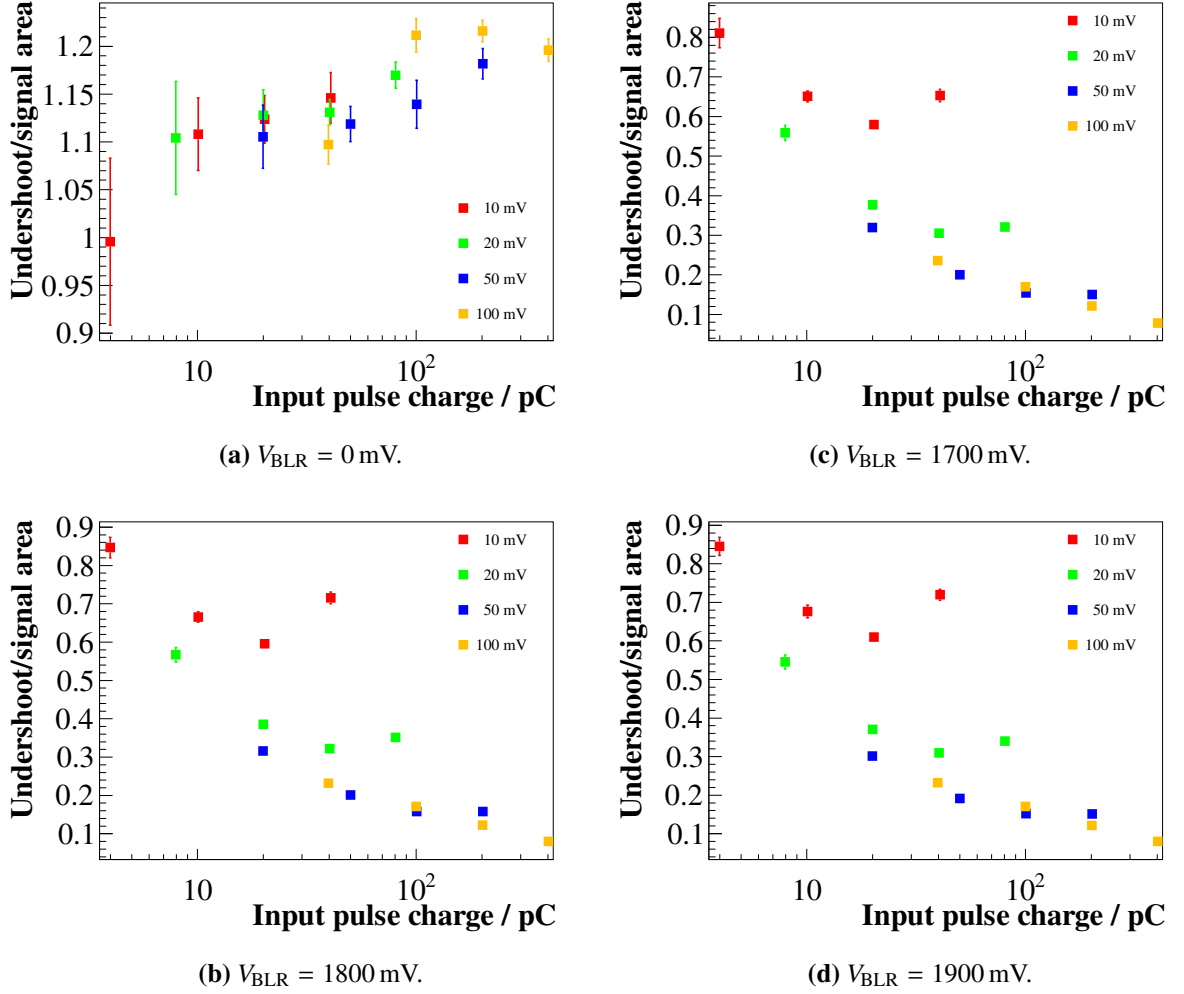


Figure 5.37: Undershoot-to-signal charge (area) ratio of the BLR stage of the bipolar shaper for different input pulse amplitudes and four BLR voltage settings. Changes in input signal length lead to variations of the input charge for constant input amplitude. Rise and fall times were kept constant. A comparison of all measurements is shown in Figure C.4.

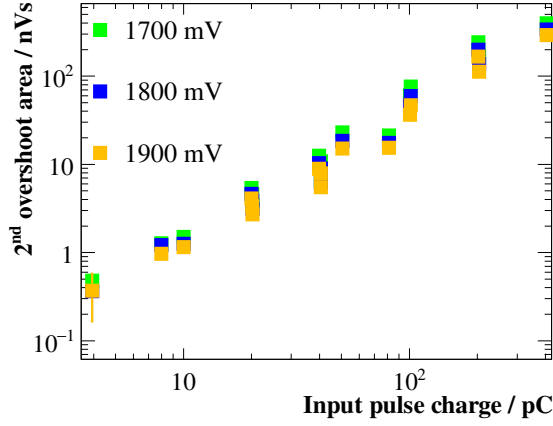


Figure 5.38: Area (charge) of the second overshoot after BLR stage of the unipolar shaper with different BLR voltages as a function of the charge input to the bipolar shaper.

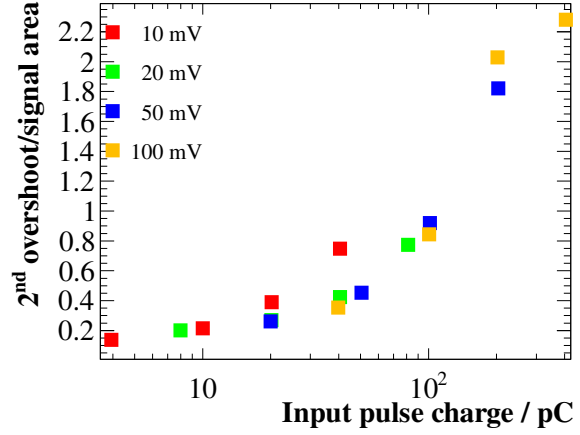


Figure 5.39: Second overshoot-to-signal ratio of the BLR stage of the bipolar shaper for rectangular input pulses with four different amplitudes at $V_{BLR} = 1800$ mV as a function of the charge input to the bipolar shaper which is varied via the pulse length.

undershoot begins while the input pulse is still above zero. The end of the input pulse leads to a kink in the undershoot. The baseline is restored quicker than for a rectangular input pulse of even less charge. This shape is conserved by the BLR stage for zero BLR voltage (Figure 5.41). With the BLR voltage turned on, the undershoot attenuation sets in due to current flow through the diode into the capacitor between the shaper and the BLR stage. This additional charge is slowly drained after the shaper output returned to zero, leading to an overshoot which corresponds to the second overshoot for a rectangular input pulse. Due to the slower voltage drop of the input pulse, the overshoot is smaller than in the case of rectangular input pulses and the regulation of the DAC of the BLR stage does not produce an additional overshoot.

5.1.3 High Rate Effects

In this section, the effect of high input pulse rates on the electronics response is discussed. In the case of long intervals between subsequent input pulses discussed in Section 5.1.2, the pulses do not pile up. High rates, however, lead to interference of pulses which affects the baseline¹, but can also lead to saturation and changes of rise or fall times. In order to study the high rate effects and the effect of the BLR circuit, pulses of different shapes were injected at different rates. As the threshold voltage is adjustable, it can be

¹In Section 5.1.3, all baseline levels given are absolute values. They were obtained by averaging the oscilloscope measurements in a 20 ns long time window just before an input pulse.

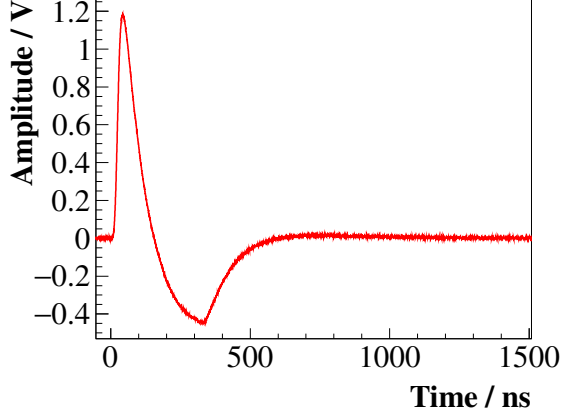


Figure 5.40: Pulse shape response of the active bipolar shaper stage to the triangular input pulse in Figure 5.3.

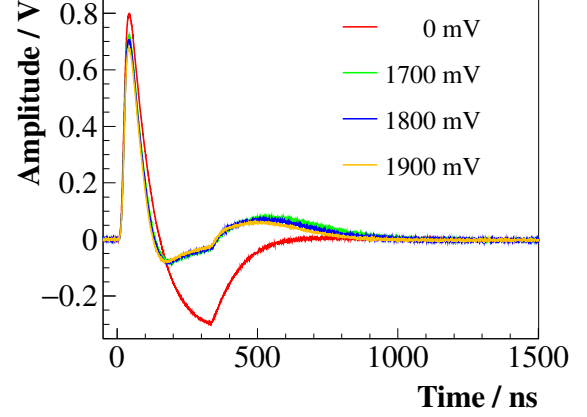


Figure 5.41: Pulse shape response of the bipolar shaper with BLR stage to the triangular pulse of Figure 5.40 for different BLR voltages.

adjusted to the absolute baseline level. The choice of the baseline is described in Section 6.2.1). Variations of the baseline with the input pulse frequency, however, cannot be corrected for and, hence, deteriorate the performance. In reality, the pulses do not arrive periodically like from the pulser, but show fluctuating time intervals. δ pulses with 3 pC charge were chosen as the largest at closest to saturation, in order to obtain the maximum effect. Rectangular pulses with 50 mV imitate γ pulses. The pulse length was set to 100 ns, roughly half of the maximum sMDT drift time. The parameters of the triangular pulses are explained in Section 5.1.1. δ and rectangular pulses were injected with a frequency of up to 5.9 MHz, while the maximum frequency for the triangular pulses is 2.9 MHz, before the input pulses start to overlap.

5.1.3.1 Unipolar Shaper

• Shaper Output

The capacitive coupling of the second shaper stage of the unipolar board prevents a DC current flow. A single pulse of positive polarity is followed by a long undershoot of small amplitude. For a subsequent hit piling up onto this undershoot, this situation corresponds to a lowering of the baseline (cf. Figure 5.42). For δ input pulses, this shift is very small due to the small pulse charge. For rectangular and, even stronger, for triangular pulses, the larger input charge leads to a considerably stronger baseline shift. At the highest rates studied $f_{\text{tri}} > 2.5$ MHz for triangular and $f_{\text{rect}} > 5$ MHz for rectangular pulses, the shaper output pulses start to overlap. Below 200 kHz the pulse frequency dependence of the baseline shift is not linear.

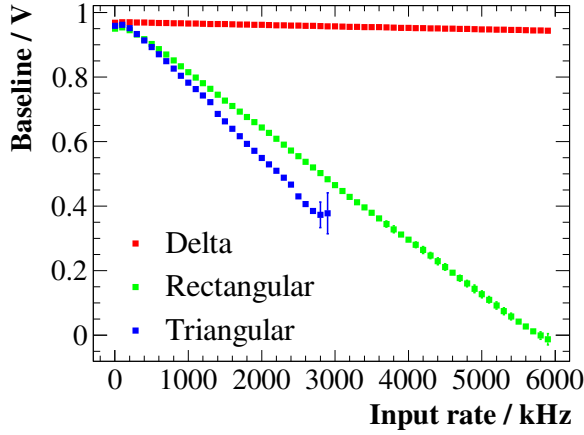


Figure 5.42: Baseline shift after the last active stage of the unipolar shaper as a function of the input rate of pulses shown in Figure 5.1 (δ), Figure 5.2 (rectangular) and Figure 5.3 (triangular).

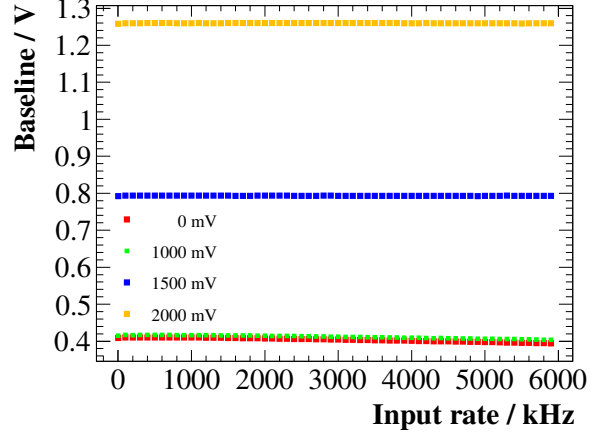


Figure 5.43: Baseline shift after the BLR stage of the unipolar shaper as a function of the input rate of δ like pulses of 3 pC charge shown in Figure 5.1 for different BLR voltages.

Parasitic charge losses prevent a baseline shift below 100 kHz. The strong dependence of the baseline level on the input pulse frequency clearly show that a baseline restoration is necessary for unipolar shaping.

• BLR Stage Output

As the BLR stage stage is AC coupled to the active shape stages via C_2 in Figure 4.7, the baseline of the two stages are independent of each other. The baseline of the BLR stage is dominated by the current through the pull down resistor R_6 and, thereby, the BLR voltage. Figure 5.43 shows the baseline after the BLR stage of the unipolar shaper for different BLR voltages and δ input pulse rates. No baseline shift is observed independent of V_{BLR} for δ input pulses. For larger input charges, baseline shift at higher pulse frequencies becomes visible (cf. Figure 5.44). Above 1 MHz rectangular input pulse rate, the baseline shifts for zero BLR voltage. For $V_{BLR} = 1000$ mV, this shift is already strongly reduced. Higher BLR voltages prevent baseline shift it up to $f \approx 3$ MHz and make it very small above this input rate. For triangular input pulses, the baseline shift behaves in a similar way, but is stronger, due to the larger charge input (Figure 5.45).

Smaller BLR voltages are preferable in with respect to pulse shape and amplitude conservation (cf. Section 5.1.2.1). Therefore, $V_{BLR} = 1000$ mV is chosen as baseline setting. However, higher BLR voltage is needed at very high input rates above 1 MHz.

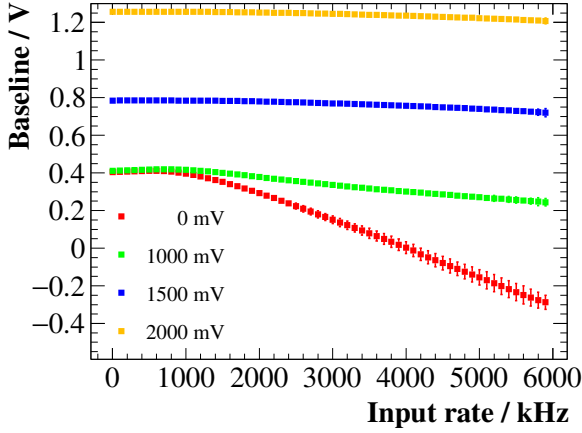


Figure 5.44: Baseline shift after the BLR stage of the unipolar shaper as a function of the input rate of rectangular pulses of 106 pC charge shown in Figure 5.2 and for different BLR voltages.

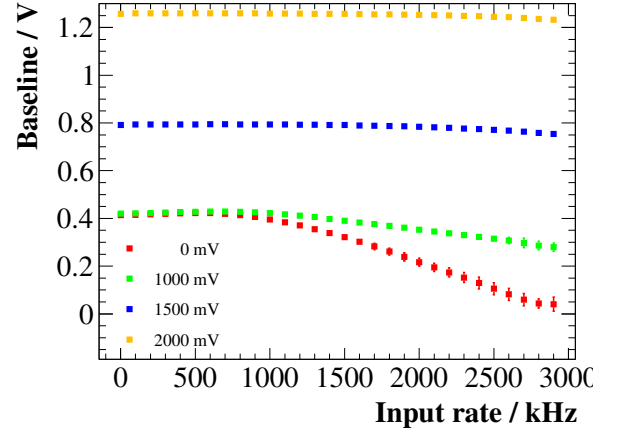


Figure 5.45: Baseline shift after the BLR stage of the unipolar shaper as a function of the input rate of triangular pulses of 127 pC charge shown in Figure 5.3 and for different BLR voltages.

- Baseline Distortions

The BLR voltage V_{BLR} and the reference voltage V_{REF} for the unipolar shaper in Figure 4.7 are generated centrally for up to 24 channels by one DAC¹ output on a motherboard. The voltage levels are buffered centrally with a 100 nF capacitor on the motherboard and additional 10 nF capacitors on each single channel shaper board to avoid wiggling. This capacitive load, however, leads to oscillations under certain conditions due to the regulating time constants of the DAC. The stability of the voltages depends on the V_{BLR} value. The reference voltage depends in addition on the input pulse rate. Figure 5.46 shows the oscillations of the BLR and reference voltage. Without input pulse, the BLR voltage shows no oscillations. Injecting rectangular pulses with 106 pC charge at a frequency of 3 MHz into one channel leads to oscillations of the BLR voltage. The oscillations propagate to the signal baseline at the discriminator input and overlays the electronic pulses. This corresponds to an oscillation of the discriminator threshold deteriorating the drift tube resolution and efficiency.

An active buffer of V_{BLR} and V_{REF} had been omitted by accident on the motherboard for the unipolar shaper. Future motherboards will be equipped with active voltage buffers to ensure stable DC levels at high capacitive loads.

¹ANALOG DEVICES, LTC2637 [63].

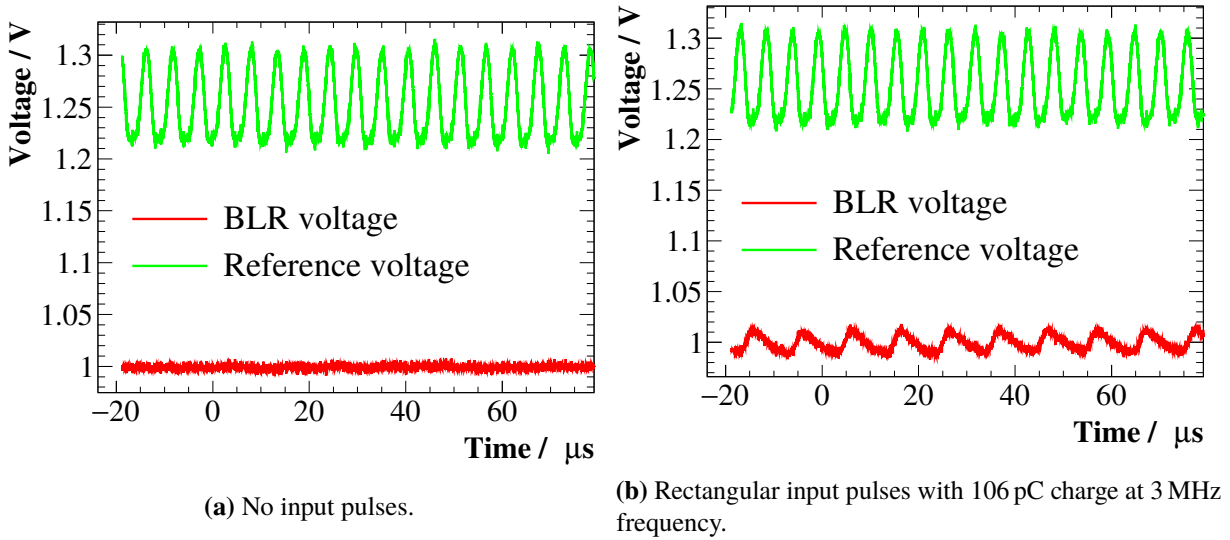


Figure 5.46: Oscillations of the BLR voltage and of the reference voltage (V_{BLR} V_{REF} in Figure 4.7) at $V_{\text{BLR}} = 1000 \text{ mV}$ for different input pulse rates.

5.1.3.2 Bipolar Shaper

• Shaper Output

Due to the undershoot, which has the same area like the positive signal, the baseline is intrinsically stable for bipolar shaping up to medium input rates up to about 1.5 MHz depending on the input pulse charge (cf. Figure 5.47). Only above this frequency, the undershoot of the preceding pulse effectively lowers the baseline for the following signal. This pileup effect is, contrary to unipolar shaping, not only a consequence of pulse frequency, but occurs always for subsection pulses with small time interval (cf. Section 4.3). The higher the rate, the larger becomes the undershoot amplitude onto which the subsequent signal piles up. In the case of triangular input pulses, the baseline shift stops increasing for $f_{\text{tri}} > 2.8 \text{ MHz}$, because the signal starts to pile up onto the positive signal of the preceding pulse. For rectangular input pulses, this is expected to happen only for input pulse frequencies higher than $f_{\text{rect}} > 6 \text{ MHz}$ as they are shorter than the triangular pulses. As the undershoot amplitude is proportional to the input charge, variations of the baseline level are negligible for δ input pulses.

• BLR Stage Output

Like in the unipolar case, an increase of the BLR voltage leads to an increase of the current through the pull down resistor R_9 in Figure 4.10, which increases the baseline of the BLR stage. Figure 5.48 shows the output baseline voltage for different BLR voltages and δ pulse input rates. While zero BLR voltage leads

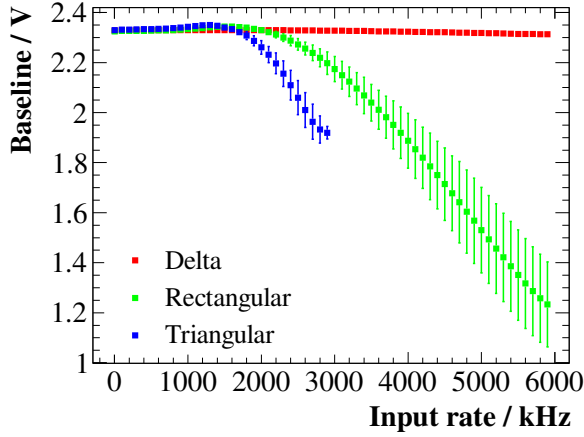


Figure 5.47: Baseline shift after the active bipolar shaper stage as a function of the rate of the pulses shown in Figure 5.1 (δ), Figure 5.2 (rectangular) and Figure 5.3 (triangular).

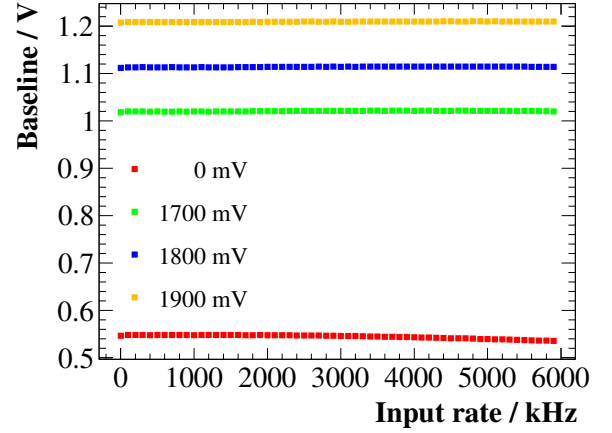


Figure 5.48: Baseline shift after the BLR stage of the bipolar shaper as a function of the input rate of δ like pulses of 3 pC shown in Figure 5.1 for different BLR voltages.

to only a very small baseline shift with increasing rate, the higher BLR voltage settings studied eliminate the shift.

Rectangular pulses (Figure 5.49) inject more realistic charges into the circuit. For zero BLR voltage, the baseline shift is mainly caused by the capacitive coupling between the active bipolar shaper and the BLR stage, where a charge input of increasing rate lowers the baseline (cf. Section 4.3.3). At input rates $f_{\text{rect}} > 3$ MHz, the signals in addition pile up onto the undershoot of the preceding pulse leading to an even faster baseline reduction. With the BLR voltage turned on, the baseline level is controlled by the current through the pull down resistor R_9 , which prevents a baseline shift up to high rates. For rates $f_{\text{rect}} > 1$ MHz, the signal starts to pile up onto the long second overshoot, generated by the BLR stage response to the preceding hit (see Section 5.1.2.2), which leads to a perturbation of the baseline level. For the highest rates $f_{\text{rect}} > 3$ MHz, the time interval between the pulses becomes so short, that pulses pile up onto the undershoots of the previous pulse, which reduces the baseline again. The pulse and undershoot amplitudes are not affected by the pile up and their absolute position follows the baseline evolution closely.

The triangular input pulses lead to similar effects. The overlapping of pulses affects the baseline already at smaller rates (Figure 5.50), due to the larger input charge and length.

Due to the capacitive coupling between the active bipolar shaper and the BLR stage via C_4 in Figure 4.10, zero BLR voltage leads to baseline variations already at small input rates. All BLR voltages studied guarantee baseline stability for input pulse intervals down to $1 \mu\text{s}$. Figure 5.34 and Figure 5.50

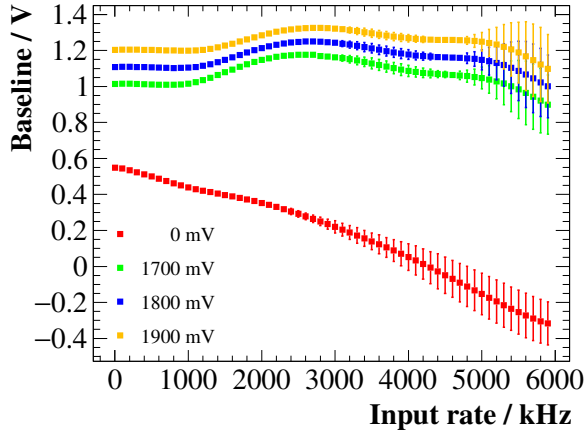


Figure 5.49: Baseline shift after the BLR stage of the bipolar shaper as a function of the input rate of rectangular pulses of 106 pC shown in Figure 5.2 for different BLR voltages.

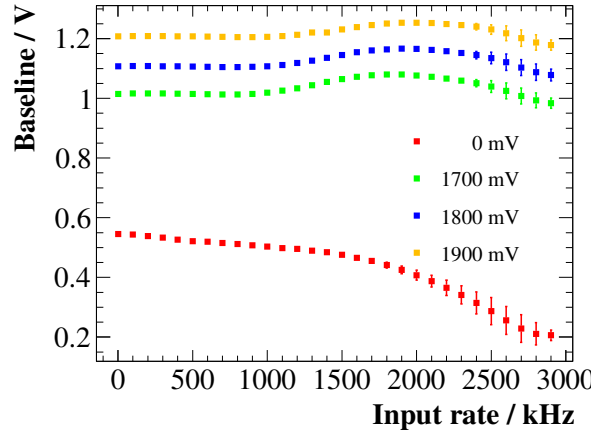


Figure 5.50: Baseline shift after the BLR stage of the bipolar shaper as a function of the input rate of triangular pulses of 127 pC shown in Figure 5.3 for different BLR voltages.

suggest that $V_{BLR} = 1900$ mV is preferable, as the second overshoot is smaller than for the other BLR voltage settings. However, the effect is small and has to be studied under realistic conditions.

5.2 Tube Pulse Response

Muon and γ induced tube pulses are formed by stochastic processes and, therefore, differ strongly in length, amplitude, shape, rise and fall time. In this section, example tube pulses and the electronics response to them are studied. The relevant pulse features are explained and the influence of the BLR voltage is shown.

The input pulses are from a sMDT chamber at nominal operating conditions. Muon signals were recorded with a cosmic muon trigger. A ^{137}Cs source with very low activity was used to generate γ hits at a rate of $f < 1$ kHz such that space charge effects are negligible. The pulses were recorded with an oscilloscope as described in Section 5.1.1.

5.2.1 Unipolar Shaping

Figure 5.51 shows eight tube pulses and the response of the unipolar shaper electronics for different BLR voltages. All signal baselines have been set to 0 V for better comparison. The input amplitude is multiplied by a factor of 10. Signal propagation time through cables leads to a constant time offset of 16 ns between the signal stages.

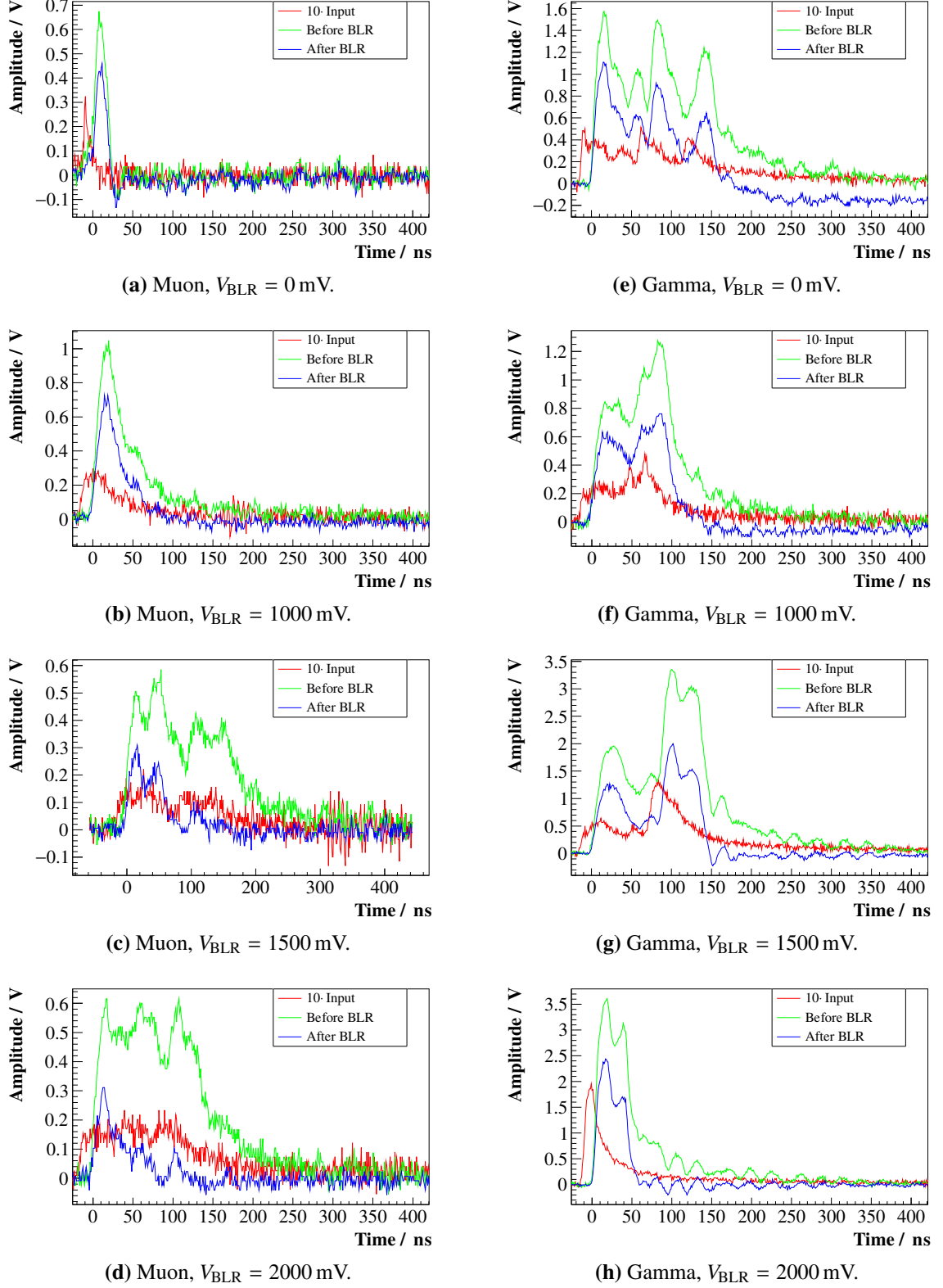


Figure 5.51: Typical muon (left) and γ (right) pulses from a sMDT chamber at nominal operating conditions and the corresponding output pulses of the last active and BLR stage of the unipolar shaper for different BLR voltages. The input amplitude is multiplied by a factor of 10.

The big variations in pulse shape corresponding to the arrival of different ionization charge clusters at the wire can clearly be seen. Bandwidth limitation suppress very sharp signal peaks high frequency noise fluctuations.

On average, γ induced signals have larger amplitude than muon induced ones as they deposit more energy in the detector. This means higher probability of saturating the operational amplifier (e.g. in Figures 5.51g and 5.51h). Hence, the effect of the BLR stage is more pronounced for γ hits. The lowering of the baseline after the BLR stage can clearly be seen here. For $V_{\text{BLR}} > 1000$ mV, this lowering is strongly reduced.

The muon pulse in Figure 5.51b may be a superimposed by the pulse of a δ electron initiated by the muon, which arrived at the wire before the muon signal (cf. Section 2.1). Typical for this case is the pulse shape, where the first peak has a higher amplitude than the following. This muon hit will then be masked by the δ electron hit within the deadtime and not be recorded.

5.2.2 Bipolar Shaping

Figure 5.52 shows four tube pulses and the response of the bipolar shaper electronics for zero and 1800 mV BLR voltage. The systematic difference to the pulses described in Section 5.2.1 is the bipolar output signal shape. The BLR stages with non-zero V_{BLR} reduces the signal amplitude and the bipolar undershoot. Due to higher gain and lower maximum output voltage of the bipolar shaper, compared to the unipolar electronics, most of the γ induced hits drive the electronics into saturation. The γ response of the BLR stage for $V_{\text{BLR}} = 1800$ mV also shows a small second overshoot, which is caused by the discharge of the capacitor by the diode current (see Section 5.1.2.2).

5.3 Conclusion

Both shaper concepts use diodes to stabilize the baseline of a differentiated signal. The strength of the BLR stage is controlled by V_{BLR} . The unipolar shaper differentiates the input signal via the capacitor between the last active shaper stage and the BLR stage (C_2 in Figure 4.7). The bipolar shaper, in contrast, differentiates the input signal before the last amplifier stage (via C_3 in Figure 4.10) resulting in full bipolar pulse shaping. The capacitor between the bipolar shaper and the BLR stage provides AC coupling the two stages with negligible pulse shape modifications.

The BLR stage of the unipolar shaper ensures baseline stability up to 1 MHz pulse rate with $V_{\text{BLR}} = 1000$ mV. At even higher rates, larger BLR voltages are required to stabilize the baseline, which leads to significant signal amplitude reduction. The performance of the unipolar shaper in the beam tests was limited due to the oscillations of the voltage levels on the board defining the signal baseline. They

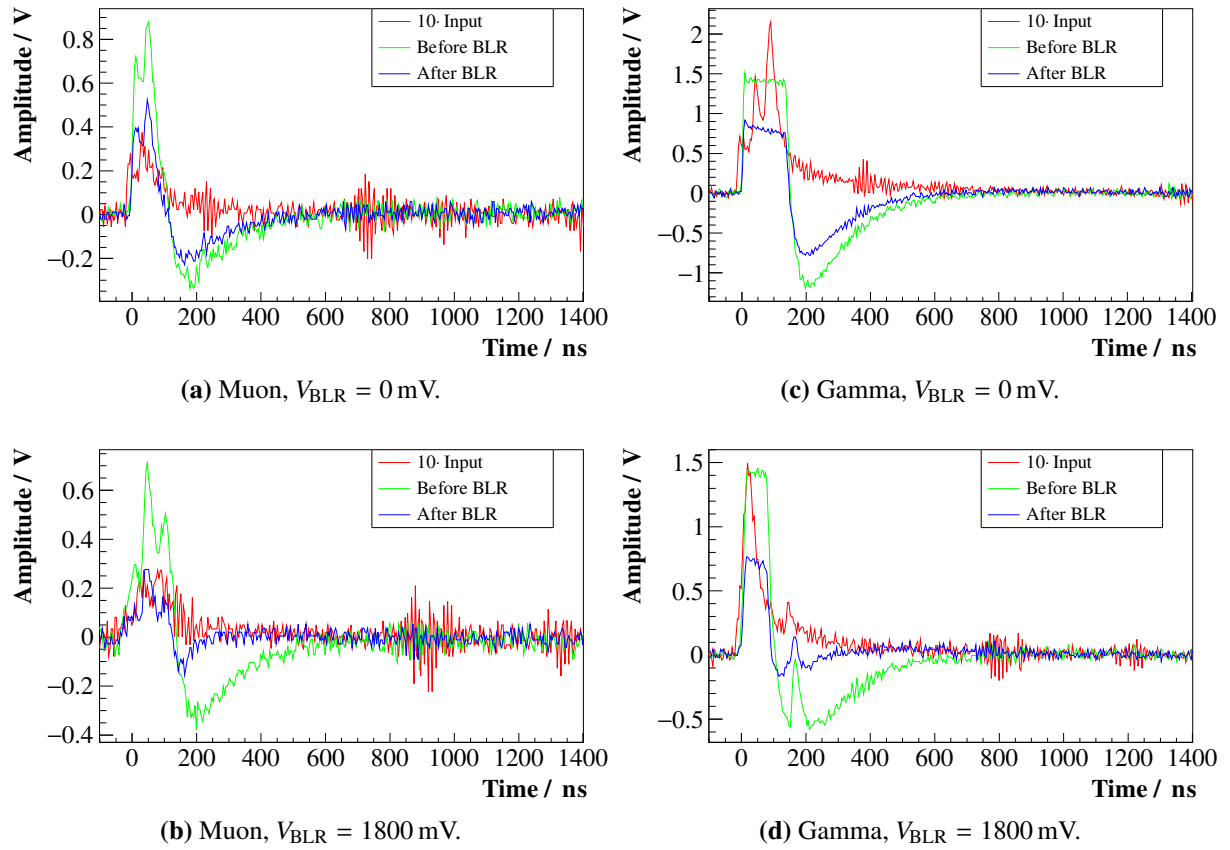


Figure 5.52: Typical muon (left) and γ (right) pulses from a sMDT chamber at nominal operating conditions and the corresponding output pulses of the last active and BLR stage of the bipolar shaper for different BLR voltages. The input amplitude is multiplied by a factor of 10.

required a high threshold to keep the accidental hit rate at a tolerable level leading to reduced spatial resolution and detection efficiency.

The bipolar shaper ensures baseline stability up to 1 MHz pulse rate for all three BLR voltage settings studied allowing for low threshold settings. The additional overshoot after each pulse, which is an property of the combination of the BLR stage with the bipolar shaping stage, shifts the baseline over a longer time interval than the duration of the bipolar undershoot. The amplitude of the overshoot increases with the input pulse charge. Overshoot amplitudes which stay below the threshold, deteriorate the spatial resolution as they add to the slewing of the threshold crossing time. Overshoots crossing the discriminator threshold mask muon hits reducing the muon detection efficiency. A larger BLR voltage setting leads to smaller second overshoots, favoring $V_{BLR} = 1900$ mV.

The effect of the baseline oscillations of the unipolar shaper board and of the second overshoot of the bipolar shaper on the sMDT resolution and efficiency cannot be quantified with the pulser test, but require measurements under realistic conditions. The oscillations of the unipolar shaper can be avoided by an improved board design, while the second overshoot of the bipolar shaper is an intrinsic property of the circuit design. Both discrete shapers have a peaking time of about 5 ns, which is significantly faster than the ASD chip with 15 ns peaking time [41]. Hence, time slewing effects are expected to be smaller for the discrete shapers improving the resolution even with higher noise levels and corresponding higher thresholds as well as small baseline fluctuations.

Chapter 6

sMDT High Rate Performance

The amplifier shaper discriminator circuits presented above were used on a sMDT chamber under different γ irradiation rates in a highly energetic muon beam. This chapter describes the beam test setup, the data analysis methods, the measured and simulated performance of the sMDT chambers with and without BLR electronics under γ irradiation in the Gamma Irradiation Facility at CERN [68].

6.1 Experimental Setup

6.1.1 The Gamma Irradiation Facility at CERN

In the ATLAS experiment at the LHC, muon detectors are exposed to high background irradiation from γ particles and neutrons of on average 1 MeV energy, created in reactions of the particles produced in the proton collisions with the detector and the beam shielding. About ten times higher background radiation fluxes are expected at the HL-LHC and a further up to 20 times increase is expected at a future 100 TeV proton collider. For the test of prototypes and full scale detectors under such irradiation conditions, the new **Gamma Irradiation Facility** GIF⁺⁺ [68] was built at CERN as the successor of the original GIF [72]. It combines a 14 TBq $^{137}_{55}\text{Cs}$ source in a concrete bunker with high intensity charged particle beams from the secondary SPS H4 beam line in the **Experimental Hall North 1** (EHN1). The cesium nuclei undergo β^- decay into an excited state of $^{137}_{56}\text{Ba}$ with a half-life of 30 years with a branching ratio of 94.6 % [73]. The de-excitation of the latter state with a half-life of 153 s produces γ rays of 662 keV energy. Even the high-energy tail of the β spectrum is efficiently shielded by the 1.75 mm thick aluminum casing of the source. β radiation from the source is hence negligible in the experiments in the GIF⁺⁺. The radioactive source is mounted in an irradiator with openings creating two γ emission fields in opposite directions up- and downstream of the muon beam with 74° opening angle each (Figure 6.1). Closest to the irradiator,

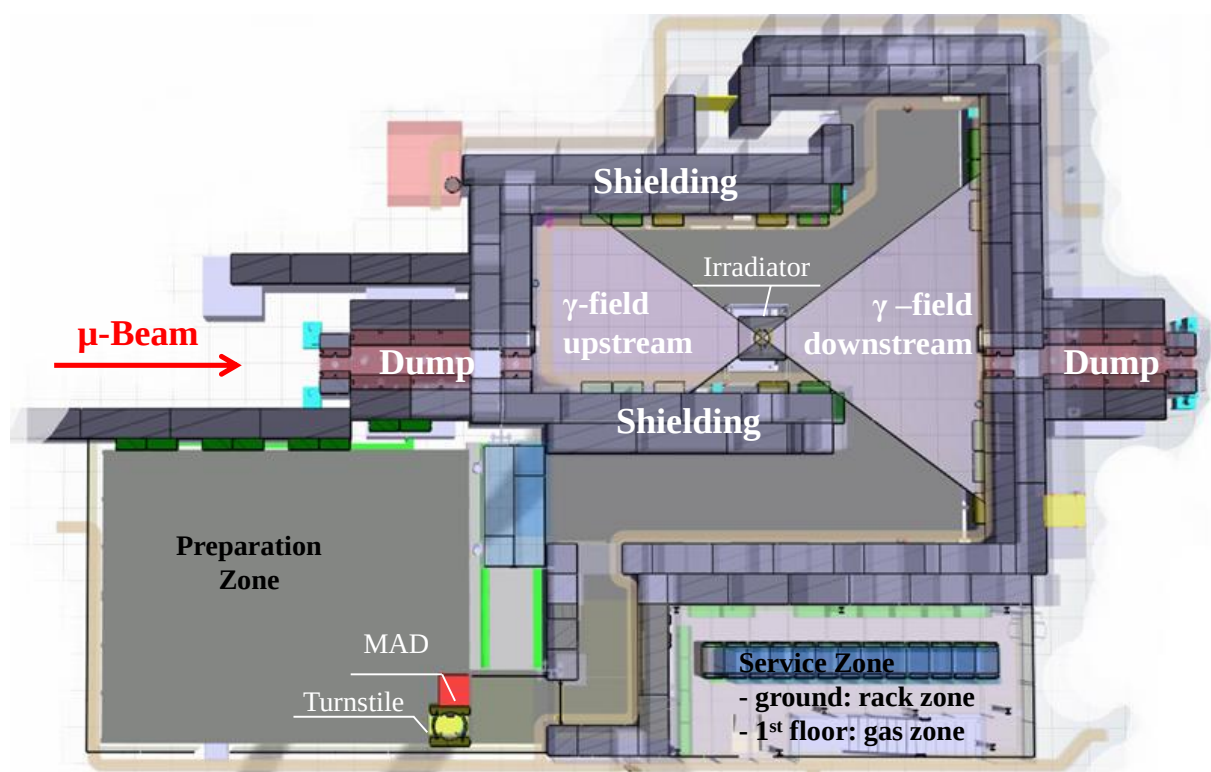


Figure 6.1: Floor plan of the new **Gamma Irradiation Facility** at CERN [68].

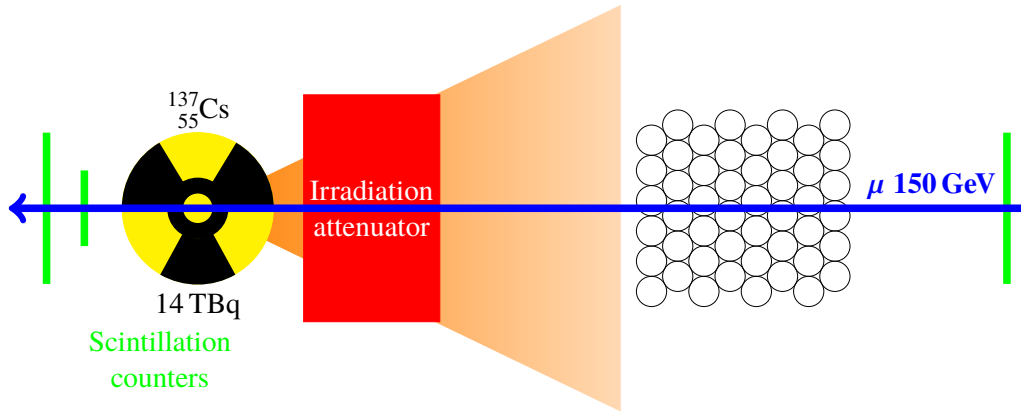


Figure 6.2: Schematic horizontally view perpendicular to the muon beam of the experimental setup in the new **Gamma Irradiation Facility** at CERN.

γ fluxes of $10^5 \text{ ms}^{-1} \text{ cm}^{-2}$ are reached. In order to achieve homogeneous irradiation of large detector areas and uniform γ flux in the cones, an angular correction filter is installed in both irradiation cones. The γ flux can be adjusted independently in the up- and downstream emission cones using a set of lead absorbers [74].

This setup is widely used for aging tests [75] and detector performance studies [76] of mainly gaseous detectors. For the latter purpose, the GIF⁺⁺ is supplied with a secondary muon beam with a nominal muon momentum of $p_\mu = 150 \frac{\text{GeV}}{c}$ from the SPS accelerator complex during about six weeks per year [77]. The source is located at the same height as the beam, but is horizontally displaced by 0.6 m. With the high muon energy, the effects of multiple scattering are negligible. Intensities of up to 2000 muons per proton spill in the beam line enable measurements with high statistics.

6.1.2 The sMDT Test Chamber

A dedicated sMDT test chamber with 96 tubes of 250 mm length in eight layers of twelve tubes was placed in the beam line as close as possible to the source while still being uniformly irradiated. A sketch of the setup is shown in Figure 6.2. The tubes were oriented horizontally. The active wire length of the 250 mm long drift tubes is 220 mm. The chamber was operated at nominal conditions (cf. Table 2.1). A photograph of the setup inside the bunker is shown in Figure 6.3. The gas and HV supplies were located outside of the bunker shielding in the service area to allow for monitoring and control during irradiation. In order to prevent gas impurities, a gas flow corresponding to ten volume exchanges per day was used.

The schematics of the readout electronics is shown in Figure 6.4. The upper half of the the chamber with six times eight tubes was read out by two standard MDT mezzanine cards with three ASD chips and

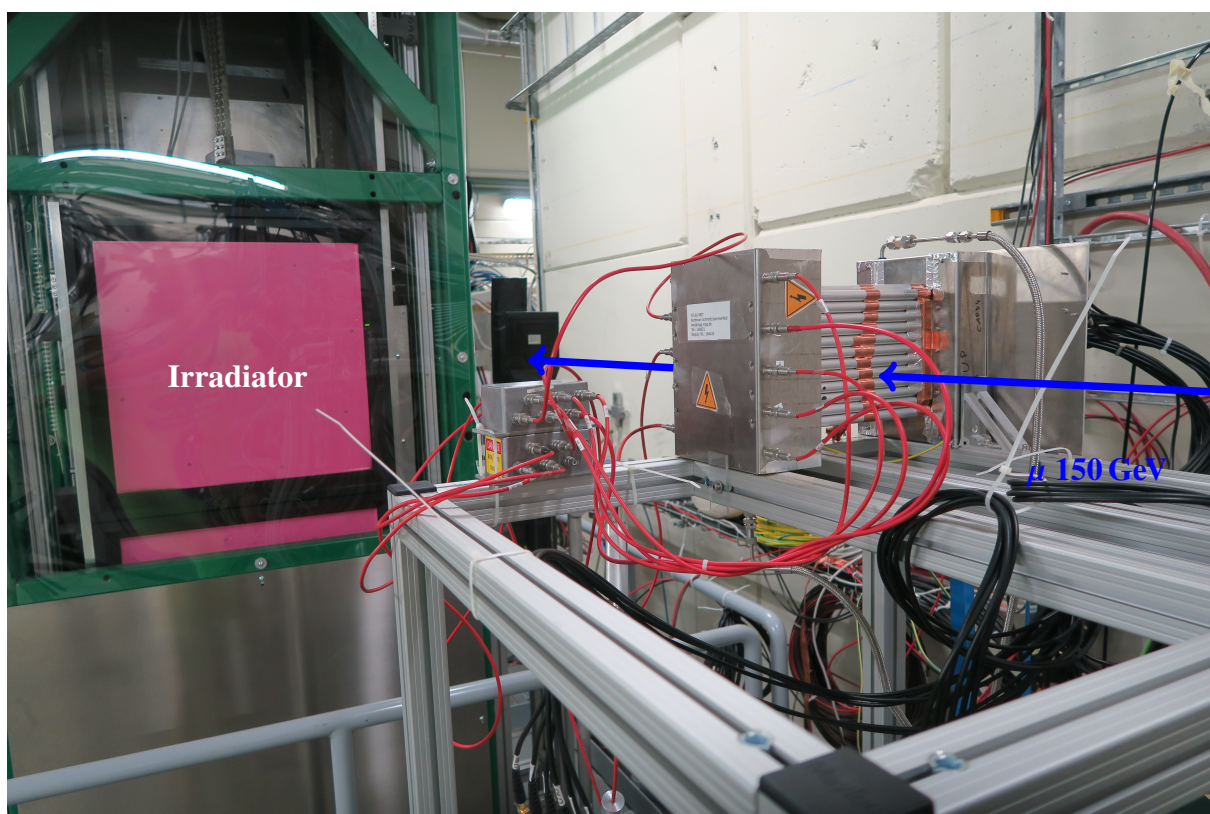


Figure 6.3: Photograph of the experimental setup in the new **Gamma Irradiation Facility** at CERN viewed under an angle of about 45° to the beam axis indicated by the blue arrow.

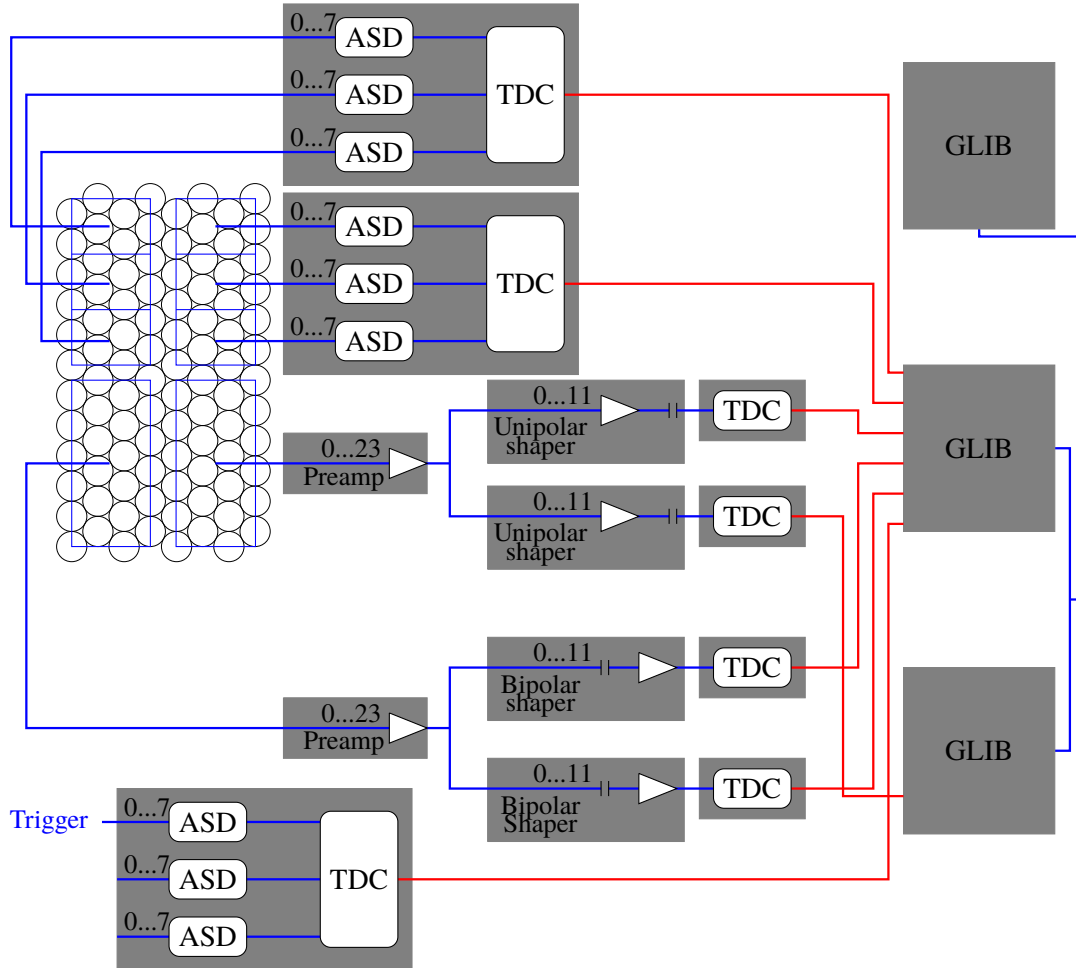


Figure 6.4: Schematics sMDT test chamber readout chain in the GIF⁺⁺. Analog signals are shown in blue, digital data lines in red. The trigger signals are injected into the ASD on a mezzanine board through the ASD feed board, which attenuates and differentiates the NIM level signals. The clock is generated on the top GLIB and distributed to the two lower GLIB. The transmission cables must have the same cable length to the two receiving boards to synchronize the data taking. All GLIB are controlled and read through a RJ45 network.

a HPTDC [78] each (see Chapter 4). The signals of the remaining 48 tubes in the bottom half of the chamber were amplified by two preamplifier boards on the chamber for the discrete shaper boards. One of the preamplifier boards was read out by unipolar, the other by bipolar shapers off the chamber as close as possible. The readout window width of each TDC was set to $1.3 \mu\text{s}$. Instead of a standard ATLAS CSM, Gigabit Link Interface Boards (GLIB) [79] with an adapter [80] to up to six TDC on mezzanine boards [81] were used. One TDC reads out 24 tubes on a standard MDT mezzanine board or twelve tubes with discrete electronics. One mezzanine board was used to read out the trigger times. Hence, a set of three GLIB was needed. In order to ensure reliable low voltage levels, the GLIB were located inside the bunker close to the chamber with remotely controlled low-voltage power supplies.

6.1.3 Trigger System and Time Reference Measurement

At the GIF⁺⁺, two large scintillators shielded from the γ irradiation, one right in front of the bunker and one just behind it, are provided for triggering. In order to select only muons which traverse the small sMDT chamber, a third scintillator with an active area matching the active area of the chamber was set up in the bunker close to the chamber and used in coincidence with the two large scintillators in the trigger. The photomultiplier tube¹ of the small scintillator was shielded from γ irradiation by lead. The mean value of the signals in the large area scintillators t_{upstream} and $t_{\text{downstream}}$ is used as event time. The event time resolution was determined from the standard deviation of $\frac{1}{2}(t_{\text{upstream}} - t_{\text{downstream}})$ and found to be $\sigma = (1.6 \pm 0.2) \text{ ns}$, which is negligible compared to the drift-time resolution. Figure 6.5 shows a diagram of the trigger setup.

6.2 Calibration

This section describes the methods that were used for the calibration of the data recorded in the GIF⁺⁺.

6.2.1 Choice of the Discriminator Threshold

In this thesis, the unipolar shaper circuit (cf. Section 4.3.3) and the bipolar shaper circuit (cf. Section 4.3.4) were studied for varying baseline restoration bias voltage V_{BLR} as alternatives to the current ASD chip at high counting rates. The spatial resolution and muon detection efficiency achieved with these alternative shaping schemes depend on the discriminator threshold and on V_{BLR} of the BLR stage. In the measurements performed in this thesis, the discriminator thresholds of the alternative shaping circuits were set to the

¹Photonics RTC XP 2262 B/01 [82]

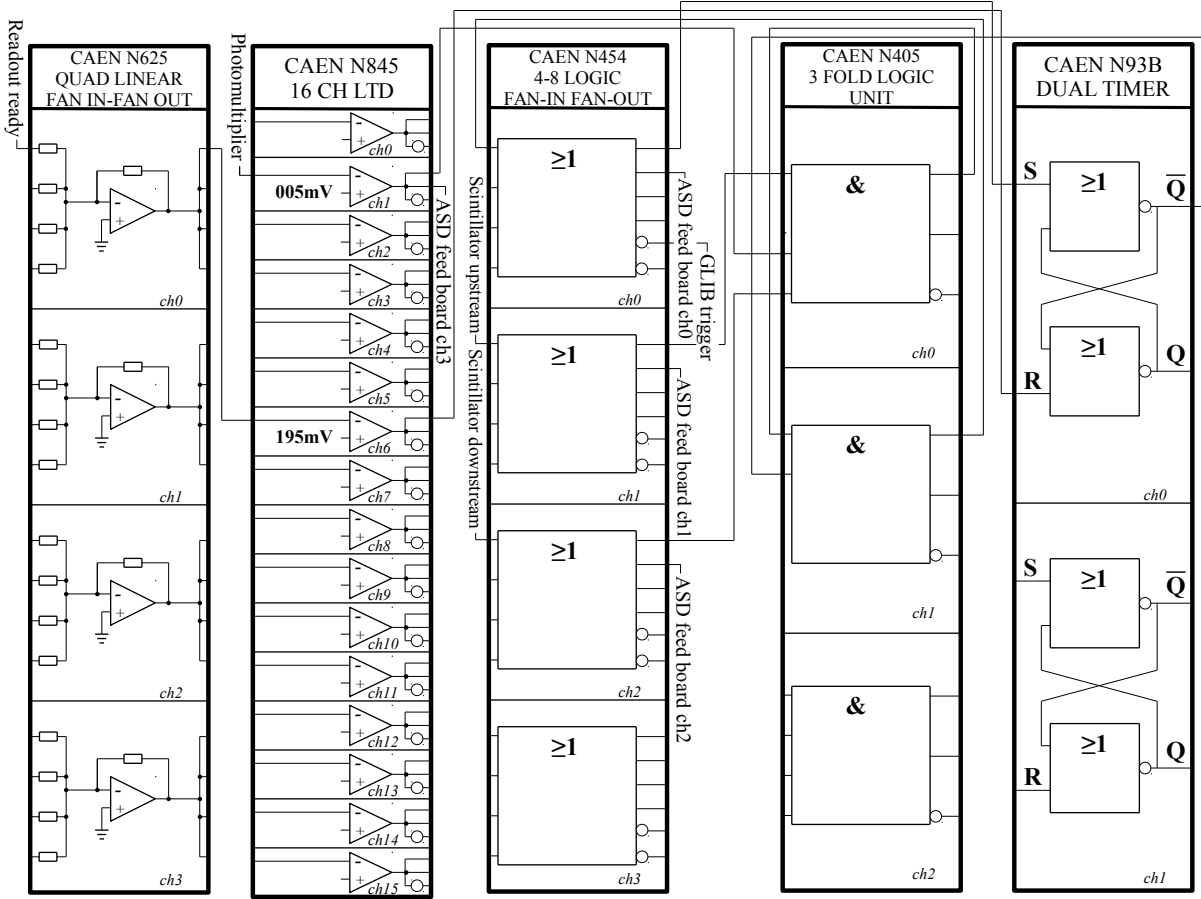


Figure 6.5: Schematic of the trigger logic setup for the beam measurements in the GIF⁺⁺. The scintillator signal inputs are discriminated with a CAEN N845 discriminator. The discriminated scintillator signals are split with a CAEN N454 LOGIC FAN-IN FAN-OUT unit. One branch is used as input to the CAEN N405 coincidence module, the other branch as input to a TDC used for the event-time measurement. The CAEN N93B dual time was used to block the trigger during the readout of the TDC data. It was reset by the readout ready signal, which was inverted with a CAEN N625 LINEAR FAN-IN FAN-OUT unit and discriminated by the CAEN N845 discriminator.

values at which the same accidental hit rate is obtained as with the ASD chip at nominal operating parameters.

The following method was employed for the threshold determination. The signal baseline, i.e. the output voltage of the last shaper stage in the absence of any ionizing particles passing the drift tube, is at a constant value V_b . Due to the thermal noise, caused mainly by the termination resistor of the drift tubes on the high-voltage side, and the intrinsic noise of the amplifier and shaper circuits, the output voltage V_{out} fluctuates around V_b . The width σ_{noise} of this fluctuations can be determined by measuring the counting rate of a tube as a function of the discriminator threshold in the absence of particle flux through the drift tube (cf. Figure 6.6).

Figure 6.7 shows the counting rate as a function of the discriminator threshold for the ASD chip at the nominal hysteresis setting (see Table 6.1). The dependence of the counting rate on the threshold can be described by a normal distribution with mean value V_b and standard deviation σ_{noise} which is a measure for the size of the noise level. As the discriminator input is AC coupled to the output of the last shaping stage of the ASD chip, V_b is close to 0 mV. Due to the even number of shaping stages in the ASD chip, the input pulses to the discriminator of the ASD are negative Section 4.1 requiring a negative threshold. The nominal discriminator threshold of -39 mV corresponds to three times the standard deviation of the counting rate distribution in Figure 6.7.

The counting rates as a function of the discriminator thresholds were also measured for the discrete shapers at different values of V_{BLR} . The mean values and standard deviations of the normal distribution fitted to the measured curves are summarized in Table 6.1.

As the last shaping stages of the discrete shapers are DC coupled to the discriminator, V_b is in general not 0 and rather depends on the value of V_{BLR} (cf. Section 5.1.3). In order to obtain the same accidental hit rates as with the ASD chip at its nominal threshold and hysteresis, the thresholds for the discrete electronics were set to $V_{thr} = V_b + 3 \cdot \sigma_{noise}$ as for the ASD chip (see Table 6.1).

The results of the threshold scans for the discrete electronics with one BLR voltage each is shown in Figure 6.8. The data of the unipolar electronics with $V_{BLR} = 1500$ mV shows a clear asymmetry. The noise rates drops significantly slower at threshold values larger than the baseline than it rises at smaller values. This is a consequence of the non linear characteristic curve of the diode, by which negative amplitudes get dumped stronger than positive ones. The width and maximum rate of the distribution is mainly determined by the baseline oscillations of the DAC output levels (see Section 5.1.3.1) and the hysteresis, limiting the average noise hit rate to $n < 230 \text{ ms}^{-1}$. For the bipolar board, the threshold scan shown is recorded with $V_{BLR} = 1800$ mV. The distribution is significantly less asymmetric, than for the unipolar electronics, because the quiescent current through the diodes is smaller. Average hit rates of more than 700 ms^{-1} show that the hysteresis is on the level of the nominal ASD settings.

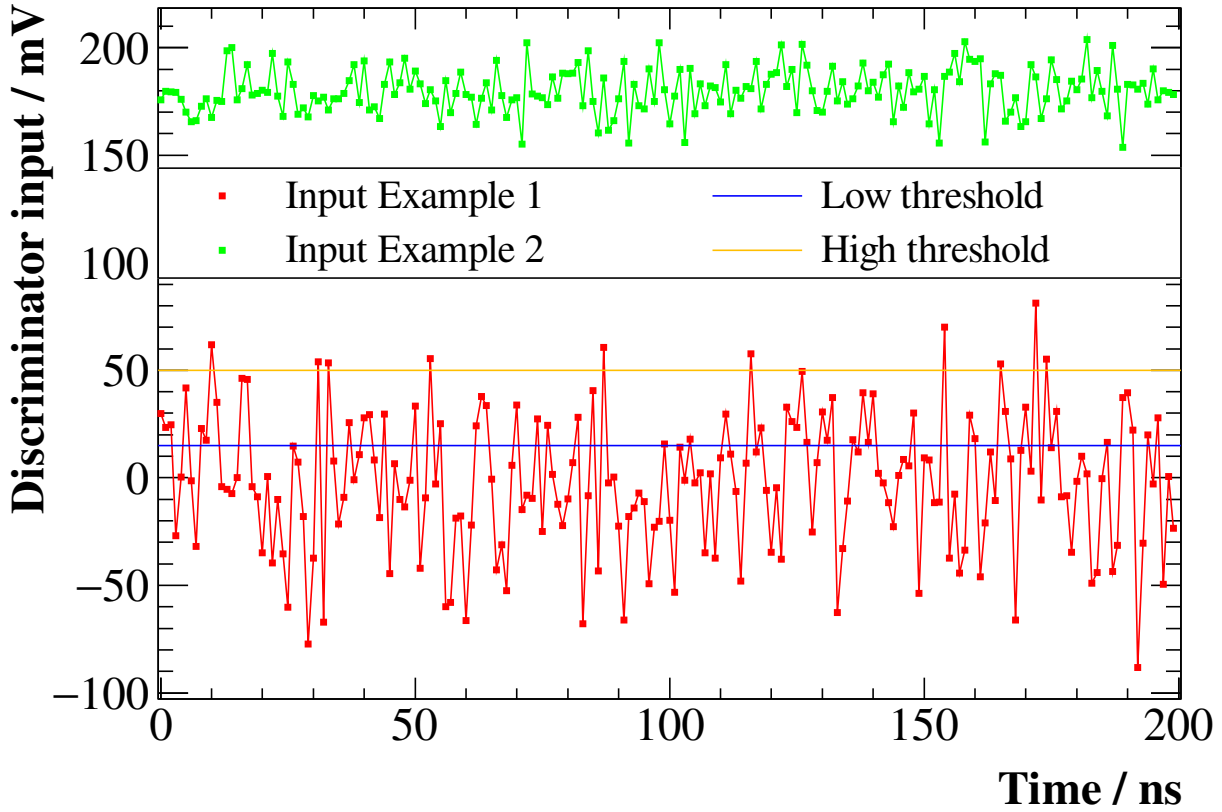


Figure 6.6: Graphical explanation of the threshold scan. The voltage V_{out} of the example shaper stage output (red) fluctuates around a baseline of $V_b = 0$ mV. The low threshold (blue) leads to higher noise counting rate than the high threshold (yellow). Recording the noise counting rate at different thresholds allows for the determination of the baseline voltage V_b and the noise fluctuation width σ_{noise} . The green example output voltage fluctuates around a baseline of $V_b = 180$ mV and has a significantly smaller jitter σ_{noise} .

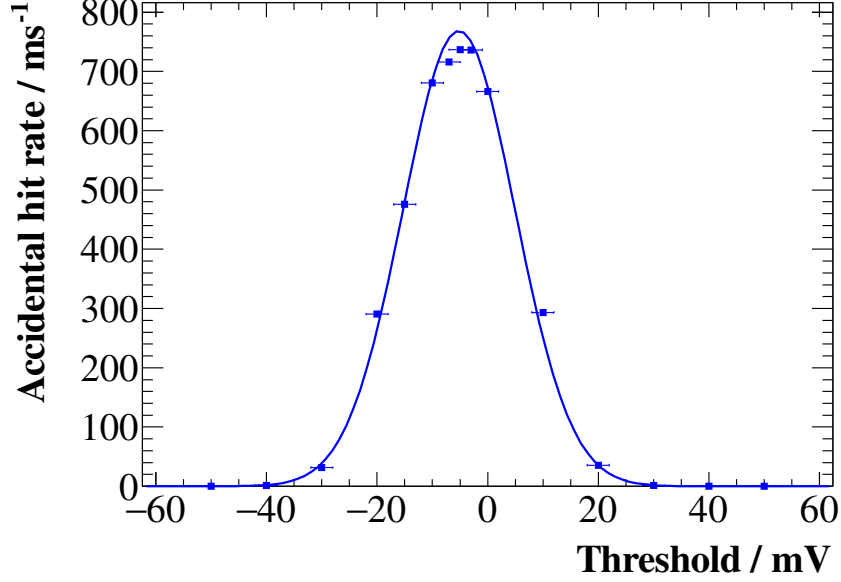


Figure 6.7: Average noise rates of drift tubes with the ASD readout at 8.75 mV hysteresis setting (ATLAS nominal).

Table 6.1: Noise amplitude parameters of the ASD chip and the discrete unipolar and bipolar shaper circuits in the GIF⁺⁺ test beam. The baseline and noise width are determined from a Gaussian fit to the threshold scan of the accidental hit rate. The uncertainties are statistical and systematic added in quadrature.

ASD				
Hysteresis / mV	8.75			
Baseline V_b / mV	-5 ± 3			
Noise width σ_{noise} / mV	10.2 ± 0.3			
Nominal threshold / mV	-39			
Discrete unipolar				
V_{BLR} / mV	0	1000	1500	2000
Baseline V_b / mV	385 ± 3	392 ± 3	765 ± 3	1230 ± 3
Noise width σ_{noise} / mV	16.3 ± 0.3	15.7 ± 0.3	17.8 ± 0.4	19.2 ± 0.5
Nominal threshold / mV	437	440	815	1280
Discrete bipolar				
V_{BLR} / mV	0	1700	1800	1900
Baseline V_b / mV	690 ± 3	1021 ± 3	1113 ± 3	1207 ± 3
Noise width σ_{noise} / mV	25.9 ± 0.7	11.7 ± 0.3	11.5 ± 0.3	11.2 ± 0.3
Nominal threshold / mV	770	1060	1150	1240

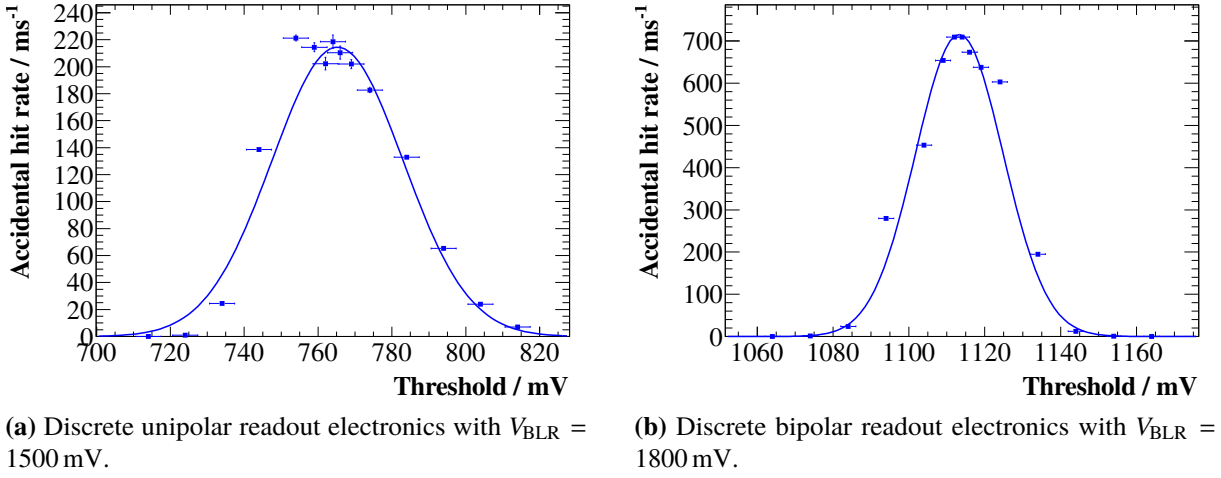


Figure 6.8: Average noise rates of tubes read out by discrete electronics at medium BLR voltages.

6.2.2 Background Rate Measurement

The tube hit rate is determined from the drift tube data. Ideally, for readout without deadtime, the rate of recorded hits n is related to the number of hits H in a readout time interval w and the number of triggers T by

$$n = \frac{H}{w \cdot T}.$$

For random triggers without incident particles, the whole TDC readout window can be used for w . If, instead, the background rate is to be determined from muon beam data, a short time interval in front of the muon drift time spectrum must be used. As discussed in Section 4.2.1, the measured hit rate n has to be corrected for the deadtime t_{dead} in order to obtain the actual rate ν of background hits which according to Equation (4.1), is given by

$$\nu = \frac{n}{1 - n \cdot t_{\text{dead}}}.$$

The minimum deadtime of the ASD chip in ADC mode was determined from the recorded data. Figure 6.9 shows the distribution of time intervals between two successive hits at high γ conversion rate of $\nu = 740 \text{ ms}^{-1}$ and minimum programmable deadtime setting of 17 ns in the ADC mode (see Section 4.1). The starting point of the distribution gives the minimum deadtime of the particular ASD chip. The MDT ASD chips show a deadtime spread and most ASD chips have minimum deadtimes closer to 200 ns [15, 65], which, hence, is used as nominal deadtime for the simulations in Section 6.8.3. The actual ASD deadtime is the sum of the discriminator output pulse length and the programmable deadtime [44] and has, hence, to

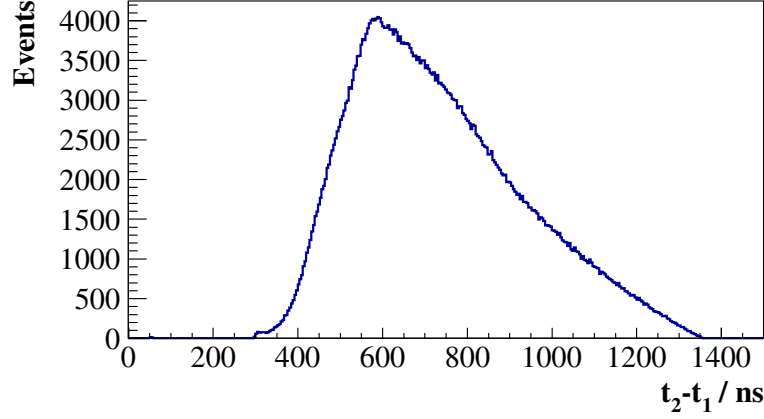


Figure 6.9: Spectrum of the time interval between two successive hits recorded with the ASD chip at minimum deadtime setting at a γ conversion rate of $\nu = 740 \text{ ms}^{-1}$. The actual minimum deadtime equals the time of the first filled bin of the distribution which is 300 ns for this particular chip used.

be averaged over all recorded hits to obtain the effective deadtime. The position of the maximum slope of the rising edge of the distribution in Figure 6.9 corresponds to an average deadtime of about 500 ns.

Another way of determining the expected tube hit rate is from the activity of the source. From the rate of emitted photons, one can calculate the resulting γ flux at a given distance. In the GIF⁺⁺ this value has to be corrected for the attenuation of the angular correction filter. The calculated photon flux for different attenuations is shown in Figure 6.10. A detection probability of 0.7 % for 662 keV photons due to Compton scattering, mainly in the (s)MDT tube wall, [83] is used for the computation of the expected γ conversion flux. The results are in very good agreement with the measured γ hit rates in the tubes. The slight excess at high absorption factors is due to the increasing fraction of conversion of photons scattered first on detector supports and the GIF⁺⁺ shielding walls.

6.2.3 Drift Time Measurement

The threshold crossing time t_{TDC} recorded by the TDC is the sum of four contributions,

$$t_{\text{TDC}} = t_{\text{pass}} + t_{\text{drift}} + t_{\text{wire}} + t_{\text{elec}},$$

which are defined as

t_{pass} The time of the muon passage through the tube when the primary ionization takes place and the electrons start to drift towards the wire. It can be determined from the recorded trigger time t_{trigger}

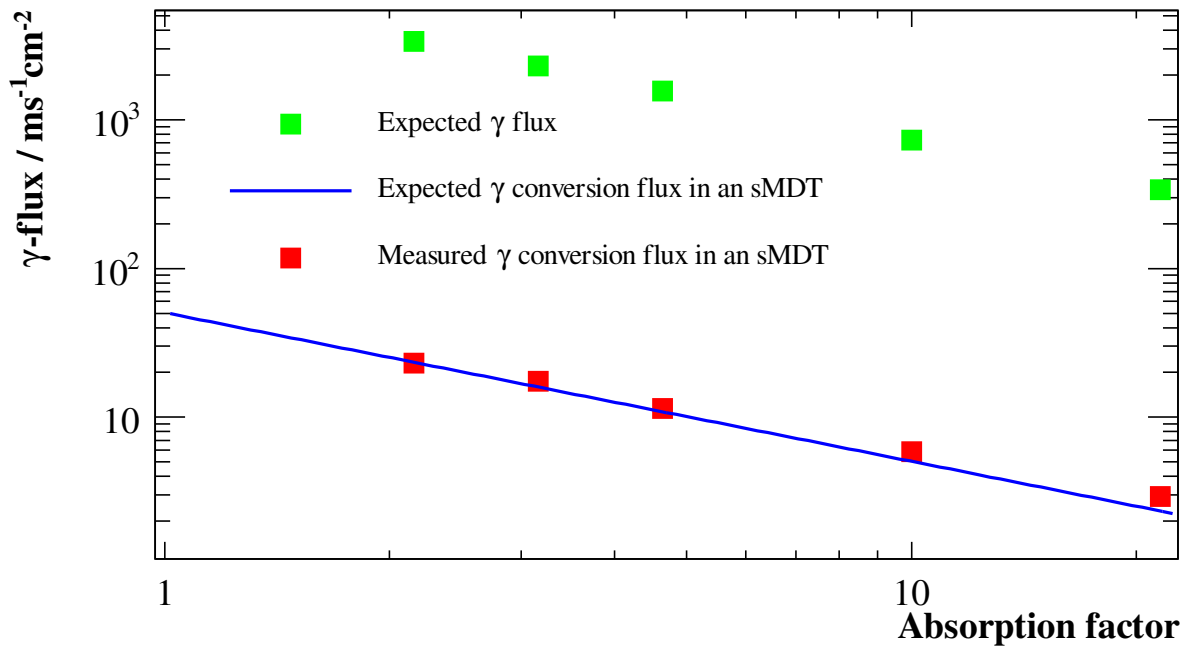


Figure 6.10: Dependence of the γ and γ conversion flux in the sMDT test chamber on the absorption factor of the GIF^{++} $^{137}_{55}\text{Cs}$ source. The expected γ conversion flux is based on a γ detection probability of 0.7 % via conversion in the drift tube walls and gas volume.

via $t_{\text{pass}} = t_{\text{trigger}} + t_{\text{prop}}$ where t_{prop} is the propagation time of the trigger signals from the trigger scintillators through the cables, logic elements and readout electronics to the TDC. t_{prop} is constant for a given electronics setup.

t_{drift} The electron drift time to the anode wire.

t_{wire} The signal propagation time along the wire to the readout electronics. ATLAS drift tubes are up to 6 m long, which corresponds to propagation straggling larger than 20 ns. This has to be corrected for event by event in case of homogeneous muon irradiation. In the GIF⁺⁺ t_{wire} can assumed to be constant as the muon beam is only a few cm wide.

t_{elec} The signal propagation time in the readout electronics which varies from channel to channel and is also affected by the analog electronics settings.

The drift time can then be calculated from the hit time t_{TDC} measured by the TDC as $t_{\text{drift}} = t_{\text{TDC}} - t_{\text{trigger}} + t_0$. The constant term $t_0 = t_{\text{prop}} + t_{\text{wire}} + t_{\text{elec}}$ has to be determined for each tube and electronics setting separately and marks the position of the minimum muon drift time. An automated fitting procedure [84] was used to determine t_0 and the maximum drift time from the drift time spectra of all drift tubes read out by the same electronics (see Figure 6.11 for the ASD readout). The rising edge of the spectrum is fitted by a modified Fermi function

$$G(t) := p_0 + \frac{A_0}{1 + e^{-\frac{t-t_0}{T_0}}}.$$

The parameter T_0 is a measure for the steepness of the rising edge and t_0 the minimum drift time. p_0 accounts for a constant noise and background hit level. A_0 is a normalization factor.

The maximum drift time is determined from a fit of the falling edge of the spectrum with the function

$$H(t) := p_{\text{max}} + \frac{a_{\text{max}} \cdot t + A_{\text{max}}}{1 + e^{\frac{t-t_{\text{max}}}{T_{\text{max}}}}}$$

with slope a_{max} and offset A_{max} , modulated with a Fermi function with negative temperature $-T_{\text{max}}$ at the turning point t_{max} . The difference between background levels p_0 and p_{max} accounts for multiple threshold crossings after the arrival time of the ionization electrons at the wire. The maximum drift time τ_{max} is defined by $t_{\text{max}} + 2 \cdot T_{\text{max}}$. The steepness of the falling edge is limited due to the efficiency loss for short muon trajectories close to the tube wall.

Figure 6.11 shows drift time spectra of sMDT recorded with the ASD chip at nominal ATLAS threshold and hysteresis settings with t_0 subtracted already such that the maximum slope of the rising edge of the spectrum is at 0 ns. Without irradiation, the spectrum shows a small dip after the steep rising

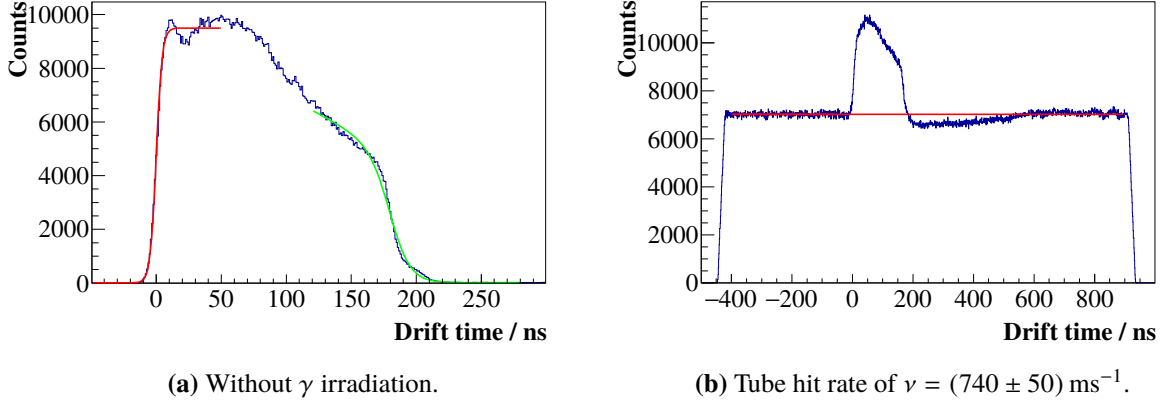


Figure 6.11: Drift time spectra of sMDT without (left) and with $\nu = 740 \text{ ms}^{-1}$ (right) γ background rate. The tubes were read out with the ATLAS ASD chip at nominal ATLAS threshold and hysteresis settings. The left plot shows the fits of the rising and the falling edge for the determination of the minimum and maximum drift times. The right plot shows the constant fit of the plateau of background hits on which the muon signals rest.

edge around 25 ns. This is a consequence of the Ramsauer minimum in the electron-Argon interaction cross section [41] and the resulting reduction of electron drift velocity. The maximum drift time was determined in the run without irradiation to be $\tau_{\text{max}} = (191.9 \pm 0.8) \text{ ns}$. The drift time spectrum under background irradiation is overlaid with a constant background and fit function, which determines the background hit rate (see Section 6.2.2). The drop of the background counting rate just after the maximum muon drift time is an effect of the deadtime of the readout electronics.

Drift time spectra from sMDT read out with the discrete unipolar shaping circuit with $V_{\text{BLR}} = 1500 \text{ mV}$ are shown in Figure 6.12 after the t_0 subtraction. The steepness of the rising edge is very similar to the one for ASD readout, but the dip expected due to the Ramsauer minimum is not visible. As will be explained in detail in Section 6.8.4, the unipolar shaper had to be operated at larger discriminator threshold than the ASD chip in terms of the corresponding number of primary ionization electrons. This leads to a small delay for hits close to the anode wire and a depletion of the drift time spectrum at small drift times. The maximum drift time without irradiation is $\tau_{\text{max}} = 184.0 \pm 1.4 \text{ ns}$, i. e. significantly smaller than in the ASD case. The higher threshold increases the inefficiency at large drift times close to the tube wall due to reduced ionization charge reducing the maximum drift time. The spectrum with background irradiation shows no drop below the average background hit rate behind the muon drift times, as the discrete electronics has no fixed deadtime.

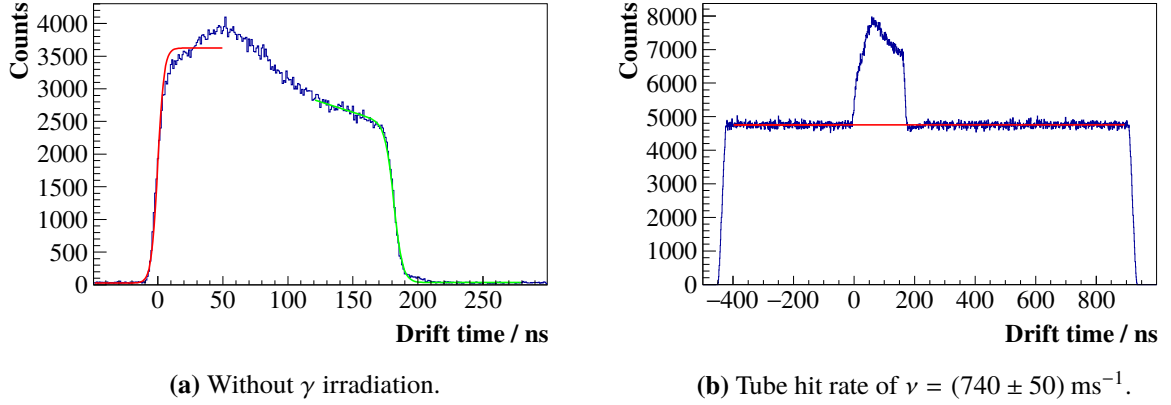


Figure 6.12: Drift time spectra of sMDT without (left) and with $\nu = 740 \text{ ms}^{-1}$ (right) γ background rate. The tubes were read out with the discrete unipolar shaping circuit at $V_{BLR} = 1500 \text{ mV}$ and rather high discriminator threshold required by the noise rate of the board (see Table 6.1). The left plot shows the fits of the rising and the falling edge for the determination of the minimum and maximum drift times. The right plot shows the constant fit of the plateau of background hits on which the muon signals rest.

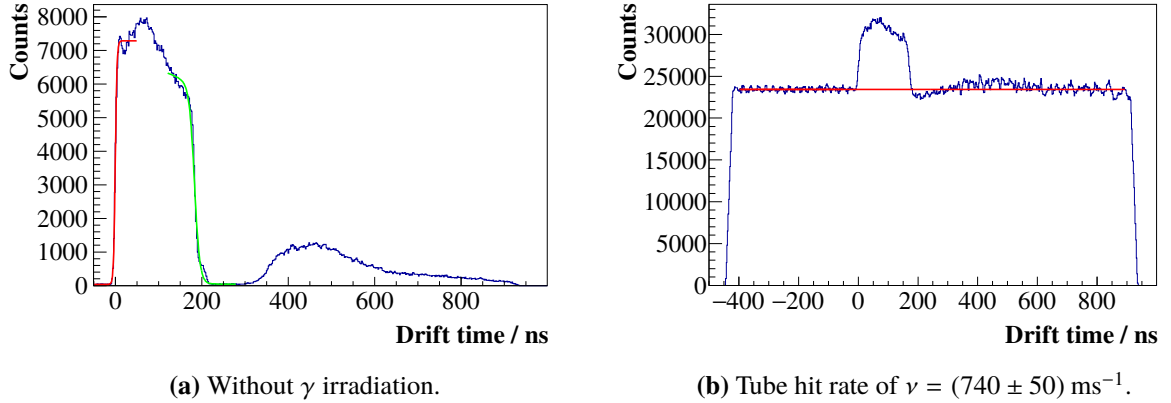


Figure 6.13: Drift time spectra of sMDT without (left) and with $\nu = 740 \text{ ms}^{-1}$ (right) γ background rate. The tubes were read out with the discrete bipolar shaping circuits at $V_{BLR} = 1800 \text{ mV}$. The left plot shows the fits of the rising and the falling edge for the determination of the minimum and maximum drift times. The right plot shows the constant fit of the plateau of background hits, on which the muon signals rest.

Figure 6.13 shows t_0 corrected drift time spectra recorded with the discrete bipolar shaping electronics with $V_{\text{BLR}} = 1800 \text{ mV}$. The rising edge is slightly steeper than for the two other readout electronics and the Ramsauer dip is clearly visible. The maximum drift time of $\tau_{\text{max}} = 191.1 \pm 0.7$ determined without background irradiation is in very good agreement with the ASD case. After the muon drift time spectrum, there is a wide second peak, which reaches up to the end end of the TDC readout window and is due to the additional threshold crossings caused by the overshoot generated by the bipolar BLR stage (compare Section 5.1.2.2). This effect can also be seen in the data with background irradiation. After the muon drift time spectrum, the counting rate drops slightly below the average due to the bipolar undershoot, and the subsequent overshoot causes a counting rate increase afterwards. Roughly 30 % of the muon hits generate a second overshoot large enough to cross the threshold. As the amplitude of the overshoot increases with the signal charge, γ hits generate an additional overshoot with second threshold crossing even more frequently, contributing to masking of muon hits and reduction of the detection efficiency at high counting rates. Comparison of the background hit rates with and without BLR voltage at the same irradiation flux shows that 80 % of the γ induced hits generate a second threshold crossing.

6.2.4 Space-to-Drift-Time Relationship

The space-to-drift-time relationship was determined from the recorded data. A first approximation is obtained by integrating the derivative of the drift time spectrum:

$$\frac{dN}{dt} = \frac{dN}{dr} \frac{dr}{dt} = \frac{dN}{dr} \cdot v_{\text{drift}}(r).$$

Assuming a constant and uniform muon flux across the tube cross section, i.e. $\frac{dN}{dr} = \frac{N}{r_{\text{max}}}$, where N is the total number of recorded muons and r_{max} the maximum drift radius, the space-to-drift-time relation is determined from

$$r(t) = \int_{t_0}^t dt' \frac{dr}{dt'} = \int_{t_0}^t dt' \frac{dr}{dN} \frac{dN}{dt'} = \frac{r_{\text{max}}}{N} \int_{t_0}^t dN.$$

This approximation has a precision of only 0.5 mm, because it ignores the presence of δ electron hits and the inefficiency of the tube close to the tube walls [85]. The space-to-drift-time relation is improved iteratively with the so-called autocalibration method [40]. Muon tracks are reconstructed (see Section 6.3.1) using an approximate space-to-drift-time relation. The residuals of measured drift radii and track radii are used to correct the space-to-drift-time relationship. The method is described in detail in [86].

The space-to-drift-time relationship determined with the integration method from muon hits recorded with the ASD chip at nominal threshold and hysteresis settings was used as starting point for the autocalibration of all runs with different electronics settings and irradiation rates. Figure 6.14 shows the effect of the autocalibration on the space-to-drift time relation. The impact of decreasing drift time

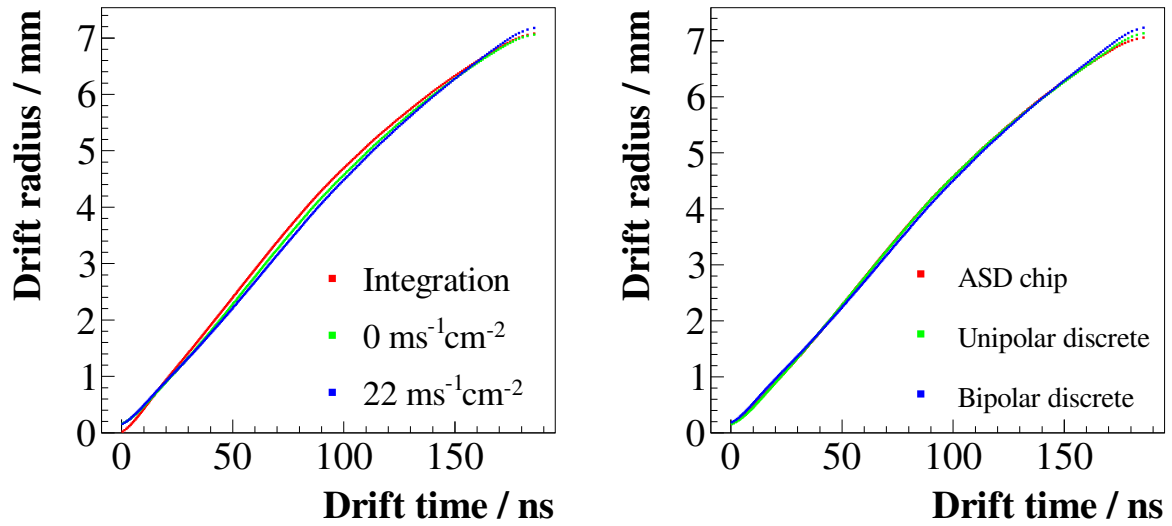


Figure 6.14: Radius-to-drift time relationship of sMDT. The left plot shows the space-to-drift-time relation obtained by integration of the drift time spectrum and the final calibration achieved after two iteration steps of the autocalibration without irradiation and for maximum achieved irradiation. The right plot shows the final corrected relations for the three different readout electronics without background radiation.

resolution near the rising edge of the drift time spectrum is not taken into account by the integration method and has to be corrected by autocalibration. At large radii, the autocalibration method corrects for the detection efficiency loss. The effect is larger for small signal-to-threshold ratios. The comparison of the corrected space-to-drift-time relations for the different readout electronics at large drift times indicates, that not just the unipolar circuit but also the discrete bipolar circuit are used with higher discriminator thresholds than the ASD chip.

6.2.5 Primary Ionization and Gas Gain

According to Equations (2.3) and (2.8), the effective gas gain of a drift tube depends on the space charge in the tube which depends on the tube hit rate and the associated primary ionization charge. Ionization is a statistical process with large fluctuations. In the GIF⁺⁺, the vast majority of hits is caused by Compton scattering of γ particles. The energy of the Compton electron depends on the scattering angle and the energy of the incoming photon. The maximum energy a 662 keV photon from the $^{137}_{55}\text{Cs}$ source can transfer to an electron by Compton scattering is 478 keV. This corresponds to a maximum range of 0.8 mm in aluminum¹, which is twice the tube wall thickness. As a consequence, the Compton electrons do not enter a second tube. According to [87], the average Compton electron energy from 662 keV photons is 253 keV. In Ar/CO_2 (93/7) gas mixture, the Compton electrons have a maximal range of 0.5 m, which means that they are not stopped in the drift gas, but in the tube wall. The average remaining photon energy after Compton scattering is 409 keV. The scattered photon on average transfers 128 keV to an electron by Compton scattering. As the scattering probability in the tube wall is at least 60 times higher than in the gas volume, most Compton electrons are produced in the tube wall. Electrons of 128 keV energy have a maximal range of 0.1 mm in aluminum. Most of these electrons are absorbed in the tube wall. The ones that generate an avalanche in the tube deposit significantly less energy and less primary ionization charge than higher energy electrons. The amount of primary electrons from $^{137}_{55}\text{Cs}$ photons strongly depends on the fraction of scattered photons.

The simulation studies in [18] assume an average energy deposition of 32 keV in a MDT by the ATLAS cavern background leading to 1300 primary ionization electrons on average. According to [17], the average energy deposition by 662 keV photons is 36 keV. Previous measurements with sMDT in the GIF determined an average value of 900 primary ionization electrons [15].

The current I , drawn by a drift tube depends on the number of primary ionization electrons N_p via

$$I = \nu \cdot N_p \cdot G \cdot e,$$

¹Tables with stopping powers and ranges of electrons in various materials can be found at <http://physics.nist.gov/PhysRefData/Star/Text/ESTAR.html>

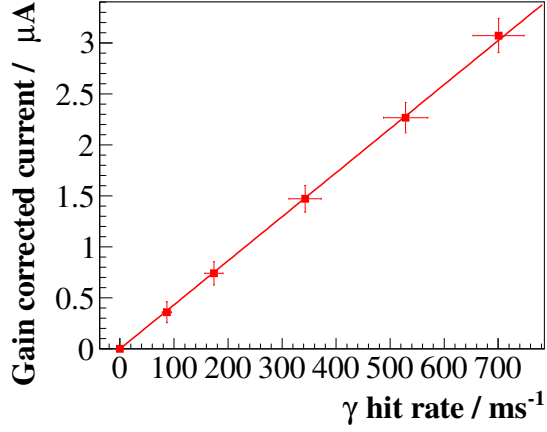


Figure 6.15: Dependence of the gain drop corrected sMDT high voltage current on the hit rate. The gas gain was measured for different γ hit rates based on the charge spectrum measured by the ASD chip in ADC mode.

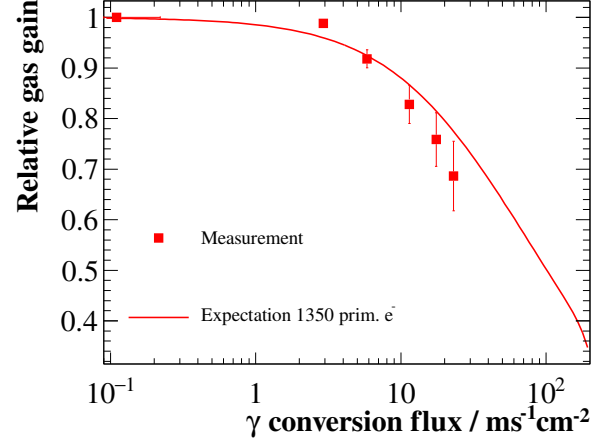


Figure 6.16: Relative sMDT gas gain at nominal operating conditions. The expected curve has large uncertainties, due to the 10% errors on the gas avalanche parameters (see section 6.2.5).

where ν is the tube hit rate, G the gas gain and e the elementary charge. According to Equations (2.3) and (2.8), the gas gain depends on the space charge and, hence, on the tube hit rate.

The relative gas gain, the ratio $\frac{G}{G_0}$ of the gas gain G at a given irradiation rate and the gain G_0 without background irradiation, can be determined directly using the signal charge spectrum measured by the ASD chip in the ADC mode (see Section 4.1). The ADC charge measurement of the charge in the rising edge of the signal pulse is used for time slewing corrections. The signal charge is proportional to the gas gain.

To correct for the nonlinearity of the response of the ASD chip, a charge calibration with inputs of known amplitudes from a pulse generator was performed. To minimize the influence of the pulse shape dependence on the drift radius on the charge measurement, only muon pulses with drift radii $3 \text{ mm} < r < 4 \text{ mm}$ were included in the analysis.

The measured gas gain ratio $\frac{G}{G_0}$ was used to correct tube current for the gain loss at higher rates. The result is shown in Figure 6.15. The fit gives $G_0 \cdot N_p = (2.70 \pm 0.15) \cdot 10^7$. For the nominal gas gain $G_0 = 2 \cdot 10^4$ of non-irradiated tubes, this corresponds to a number of primary ionization electrons of $N_p = 1350 \pm 70$. This number is significantly larger than the result of 900 primary ionization electrons per γ hit in sMDT in [15] which is a consequence of the dependence on the amount of scattered photons in the γ spectrum.

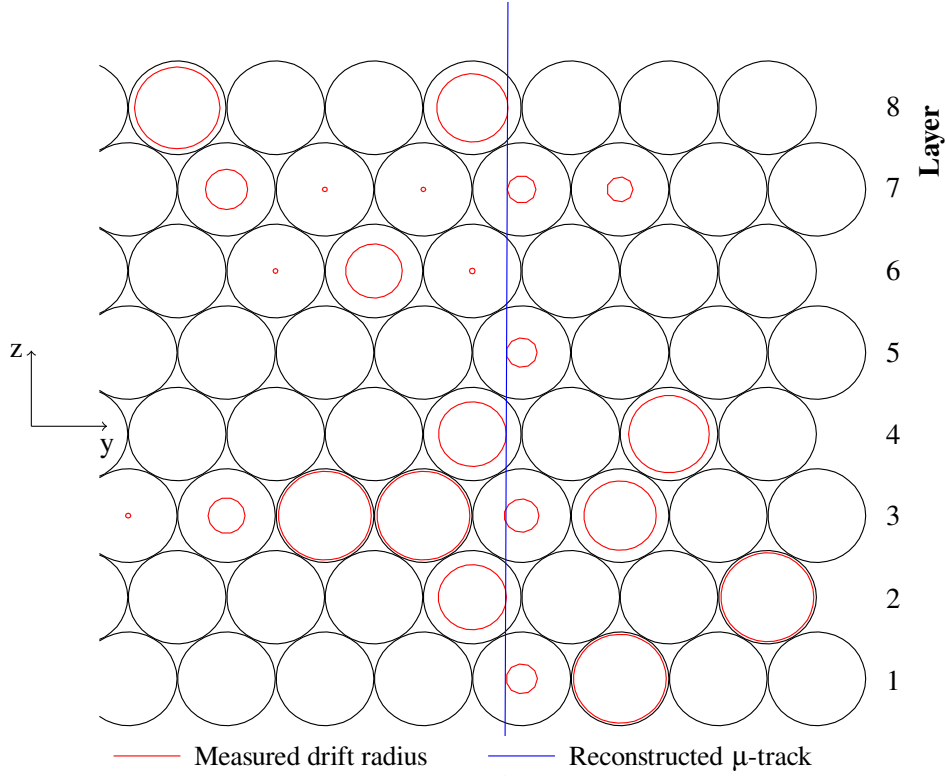


Figure 6.17: Muon event in the sMDT test chamber in the GIF⁺⁺ at 740 ms^{-1} background rate. The hits (red circles of measured drift radii) in the tubes not crossed by the fitted muon track are background hits. In layer six the muon has been masked by a γ hit or a δ electron.

The drop of gas gain due to γ irradiation is shown in Figure 6.16. The measurement data were obtained using the ASD charge measurement as for the current correction. The expectation was calculated using equations (2.3) and (2.8) with the parameter values E_0 and ΔV for the Ar/CO₂ (93/7) gas mixture from [17] and 1350 primary ionizations per γ conversion. The deviation between measurement and the Diethorn prediction shows that the model cannot describe the gain drop properly. The 10 % uncertainty on the parameter values $E_0 = 24 \cdot 10^5 \frac{\text{V}}{\text{m}}$ and $\Delta V = 34 \text{ V}$ for the (s)MDT gas mixture require an independent method for the determination of the gain drop.

6.3 Analysis Methods

6.3.1 Muon Track Reconstruction

A typical event of a muon track in the sMDT test chamber with additional background hits is displayed in Figure 6.17. The muons from the H4 beam line hit the chamber perpendicular to the tube layers. In the chosen coordinate system, this corresponds to small slopes m (compare Section 3.2.5). As the test chamber was operated in the absence of a magnetic field, the tracks are straight lines traversing the whole chamber. The muon trajectory of each recorded event was reconstructed in a two-step procedure [88].

1) Pattern recognition

One hit in each of the outermost tube layers is chosen. The four possible tangents to the circles determined by the measured drift radii are calculated and consecutively processed as track candidates, if their slopes are in the expected range. All tube hits within a road around a track candidate are passed to the next step. In previous studies, it was possible to take into account all possible hit combinations, which also allows for the reconstruction of tracks with large slopes. In order to cope with the large number of recorded events, the constrain that the beam muons hit only one tube per layer, as the tracks have small slope, had to be used. All possible hit combinations with one hit per tube layer were analyzed. If no track with hits in all tube layers was found, the whole procedure was repeated ignoring individual layers consecutively. At least four hits on track are demanded. Only track candidates with the maximum possible number of hits in each event reach the next step.

2) Track reconstruction

The parameters of the track candidates obtained in the pattern recognition step were refined by minimization of the χ^2 function

$$\chi^2 := \sum_{k=1}^n \frac{(r(t_k) - d_k(m, b))^2}{\sigma_k^2},$$

where $r(t_k)$ denotes the drift radius of the k^{th} hit associated with a given track candidate, σ_k the spatial resolution at $r(t_k)$ and $d_k(m, b)$ the distance of a straight line with slope m and intercept b from the anode wire of the hit tube. The two track parameters m and b are the free parameters in the fit. The track candidate with the smallest χ^2 value is selected if there is more than one track candidate in an event.

6.3.2 Spatial Resolution

The drift tube resolution for perpendicular incidence of the muons (track slope $|m| \approx 0$) is determined by analyzing the drift radii measured in two tubes, with the same distance of the wire from the muon trajectory as described in Section 3.2.5. The information of the reconstructed track is only needed to separate muon from accidental background tube hits and to determine on which side of the wire the muon passed.

To allow for comparison with previous studies, the average spatial drift tube resolution at a given γ conversion flux is determined using Equations (3.2) to (3.4), i.e. averaging over the residuals of two track measurements, in the results discussed in Sections 6.4 to 6.7. The averaging method for the simulation studies in Section 6.8, however, averages the drift tube resolution as a function of the drift radius (cf. Equation (6.5)).

The systematic uncertainties in the average tube resolution increase from $3 \mu\text{m}$ without γ irradiation to about $30 \mu\text{m}$ at the highest γ conversion flux independent of the readout electronics. The increased number of outliers to the fitted Gaussian in Equation (3.4) increases the dependence on the fit range, which is the main systematic uncertainty. The errors on the average resolution plots shown in the following are statistical only.

6.3.3 Drift Tube Efficiency

The tube hit efficiency

$$\epsilon_{\text{tube}} = \frac{n_{\text{det}}}{n_{\text{exp}}} \quad (6.1)$$

with the number n_{det} of detected and the number n_{exp} of expected hits is the probability of recording a tube hit when a muon has propagated through the tube gas volume. It does not distinguish muon hits from fake, δ electron or background hits, which require track reconstruction to be identified. Assuming normally distributed drift radius measurement with error $\sigma(r)$, 99.73 % of the measurements will be within $3 \cdot \sigma(r)$. The so-called 3σ -efficiency

$$\epsilon_{3\sigma} = \frac{n_{3\sigma}}{n_{\text{exp}}} \quad (6.2)$$

of measuring a correct muon hit on the reconstructed track is the probability of measuring a drift radius within $3\sigma(r)$ around the actual drift distance determined as the distance of the wire to the reconstructed track. It is the relevant quantity to determine the achievable track reconstruction efficiency.

6.3.4 Time Walk Correction

In the ASD chip and the discrete shaper readout electronics circuits, the discriminator compares input signals to a given threshold. The shaped signals reach their maximum amplitude U_{max} after a time interval

$\Delta t_{\max}$. The rising edge of the input pulse to the discriminator starting at time t_0 can be approximated by a straight line:

$$U_{\text{in}}(t) = \frac{t - t_0}{\Delta t_{\max}} \cdot U_{\max}.$$

The time t_{thr} when the discriminator threshold U_{thr} is crossed then equals

$$t_{\text{thr}} = \frac{U_{\text{thr}}}{U_{\max}} \cdot \Delta t_{\max} + t_0. \quad (6.3)$$

Hence, the threshold crossing time depends on the maximum pulse amplitude $U_{\max}$. This effect is called the time walk of the discriminator threshold crossing time.

The time walk corrected drift radius r' is calculated from the uncorrected radius measurement r and the ADC charge value w_{ADC} according to

$$r' = r + a(r) + b(r) \cdot w_{\text{ADC}} + c(r) \cdot w_{\text{ADC}}^2. \quad (6.4)$$

The correction parameters a , b and c of the correction function are obtained from reconstructed tracks in the recorded data. Figure 6.18 shows the dependence of the track residuals $\Delta r = d - r$, calculated from the track distance d to the wire and the uncorrected drift radius measurement, on the ASD output width w_{ADC} which is a measure of the rising edge charge of the signal pulse. A second order polynomial (Equation (6.4)) is fitted to the means of Gaussian fits to the residual distributions as a function of w_{ADC} to obtain the correction parameters a , b and c . To take variations of pulse shapes and amplitudes into account, the correction parameters are determined separately for each irradiation flux and in 16 drift radius bins.

6.4 ASD Readout

In this section the drift tube resolution and efficiency achievable with the ASD readout is discussed. The data were recorded in the ADC mode with minimum programmable deadtime setting of 17 ns (see Table 4.1 in Section 4.1), which corresponds to an effective minimum deadtime of 300 ns.

6.4.1 Spatial Resolution

Figure 6.19 shows the sMDT spatial resolution as a function of the drift radius. Results are shown without irradiation and under the maximum achieved γ conversion flux of $(22 \pm 1) \text{ ms}^{-1} \text{ cm}^{-2}$ for different discriminator thresholds and ATLAS nominal hysteresis setting of 8.75 mV. Without photon irradiation and nominal threshold setting of -39 mV , the resolution at small drift radii approaches $220 \mu\text{m}$. At large drift radii, it reaches to $70 \mu\text{m}$. With the lower thresholds, the resolutions improves predominantly at small

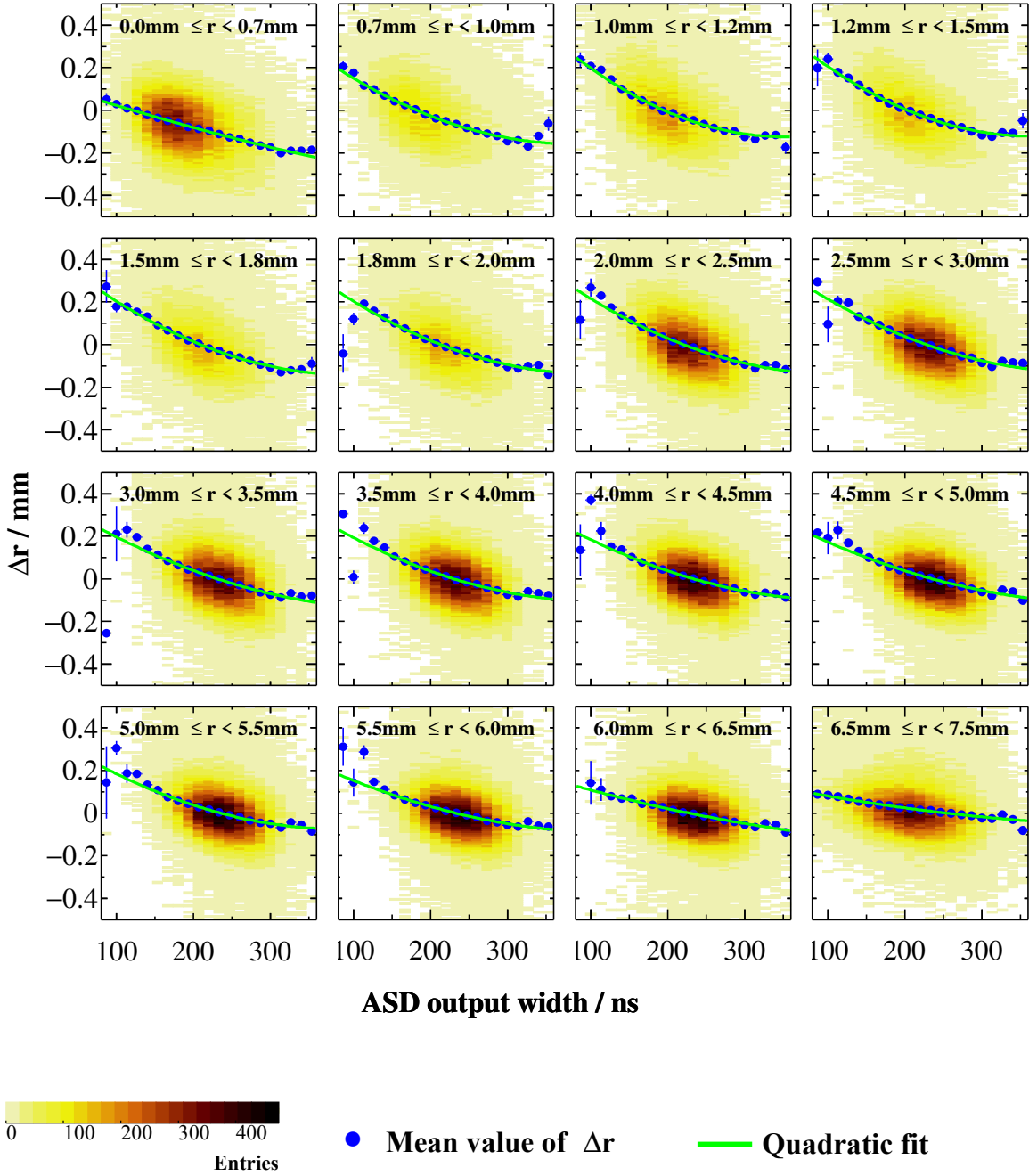


Figure 6.18: Dependence of the discriminator output pulse width which is proportional to the signal rising edge charge measurement on the track residuals Δr . Data of muon tracks recorded without irradiation with ASD readout in ADC mode are shown in 16 regions of drift radii. The blue dots are the fitted means of Gaussian fits to the residual distributions in each width bin. The green line shows a second order polynomial fit to the fitted means used for the time walk corrections (see text).

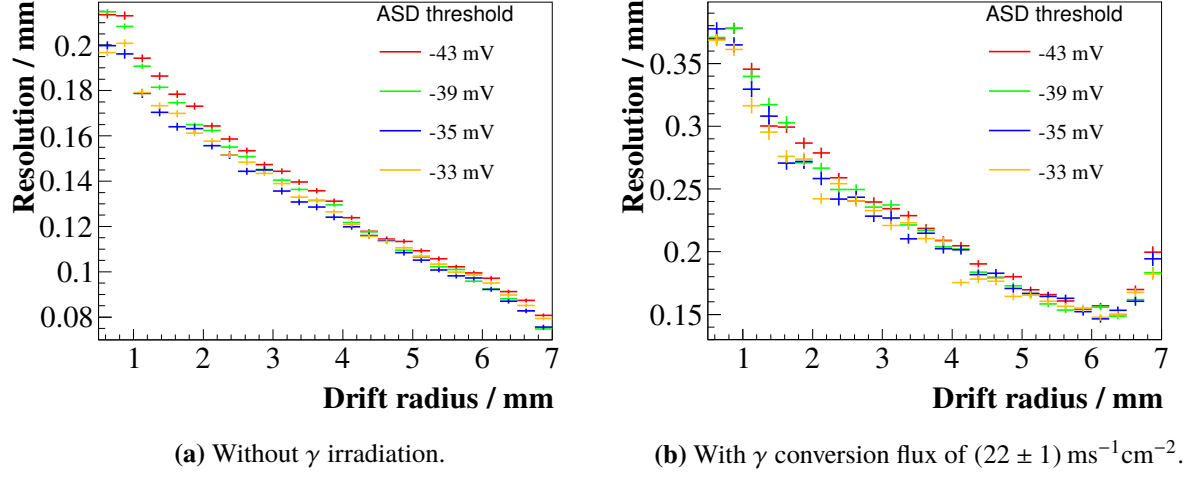


Figure 6.19: Dependence of the sMDT spatial resolution on the drift radius for ASD readout with different threshold settings without irradiation (a) and under the maximum recorded γ flux (b). The hysteresis was set to 8.75 mV.

drift radii by up to $20 \mu\text{m}$. At large drift radii, the threshold value has less impact on the resolution. With increasing photon irradiation rate, the resolution deteriorates. At a γ conversion rate of $(740 \pm 50) \text{ ms}^{-1}$, corresponding to a flux of $(22 \pm 1) \text{ ms}^{-1} \text{ cm}^{-2}$, where the gas gain drops below 70 % of the nominal value due to space charge (compare Section 6.2.5), the resolution at small drift radii deteriorates to $380 \mu\text{m}$ at the nominal threshold and reaches $150 \mu\text{m}$ at a drift radius of 6 mm. At larger drift radii, space charge fluctuations start to degrade the resolution. The effect of 10 mV threshold variation at small drift radii is smaller than in the case of no background irradiation.

The effect of increasing γ irradiation rate on the resolution can be seen in Figure 6.20. The decrease of gas gain due to space charge leads to increasing resolution deterioration, which increases at small drift radii. At the highest irradiation rates, also the deteriorating effect of space charge fluctuations sets in above a drift radius of 6 mm.

Figure 6.21 shows the mean sMDT spatial resolution as a function of the γ conversion flux at different threshold settings. The resolution deteriorates steadily up to the maximum conversion flux of $22 \text{ ms}^{-1} \text{ cm}^{-2}$ reached in the GIF^{++} . This means that the resolution degradation is dominated by the gas gain drop while the contribution of space charge fluctuations is small. The improvement of the resolution by lowering threshold saturates around -34 mV when the smearing of the threshold crossing time due to time walk equals the one caused by noise fluctuations. The mean tube resolutions are shown with statistical uncertainties only. The systematic uncertainties on the mean tube resolution increase from $3 \mu\text{m}$ without

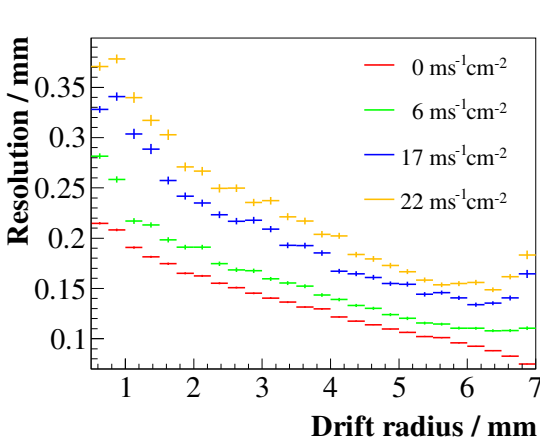


Figure 6.20: Dependence of the sMDT spatial resolution on the drift radius at different γ fluxes for ASD readout. The threshold is -39 mV and the hysteresis 8.75 mV.

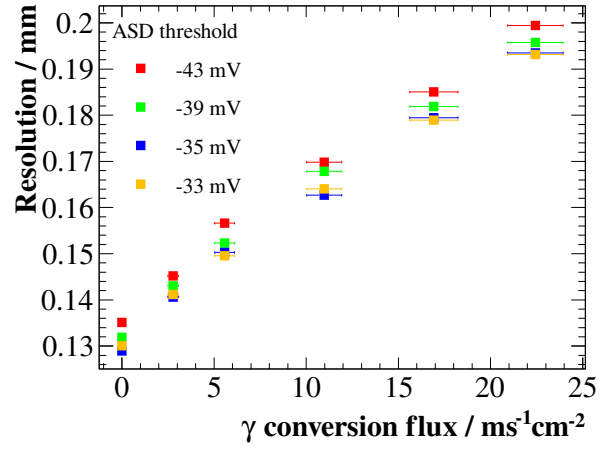


Figure 6.21: Dependence of the sMDT spatial resolution on the γ fluxes at different threshold settings for ASD readout and hysteresis setting of 8.75 mV.

γ irradiation to about $30 \mu\text{m}$ at the highest γ conversion flux due to the increased number of outliers from the track reconstruction independent of the readout electronics settings. Figure 6.22 shows the effect of the hysteresis settings on the mean tube resolution as a function of the γ irradiation. It can be seen that the discriminator threshold increases effectively with the hysteresis voltage.

Figure 6.23 shows the effect of the time walk corrections (see Section 6.3.4) on the resolution. The resolution dependence on the drift radius shows that the corrections are especially effective for the smaller signal amplitudes at small drift radii, where the corrections improve the resolution by up to $50 \mu\text{m}$ in the case without background irradiation and by even up to $80 \mu\text{m}$ at the highest background rates. For increasing drift radius, the improvement is getting smaller, because the signal amplitudes become larger and the steepness of the signal rising edge is less affected by statistical fluctuations. As the gas gain and hence the signal amplitude decrease with increasing space charge, the corrections are more effective at high irradiation rates. The impact of space charge fluctuations on the spatial resolution is somewhat reduced by the time walk corrections.

6.4.2 Muon Efficiency

Figure 6.24 shows the 3σ -efficiency as a function of the drift radius for different γ hit rates. The data were recorded with -39 mV threshold and a hysteresis of 8.75 mV. Without γ irradiation, the 3σ -efficiency is almost 100 % for muons close to the wire. At larger drift radii, muon hits are masked with increasing

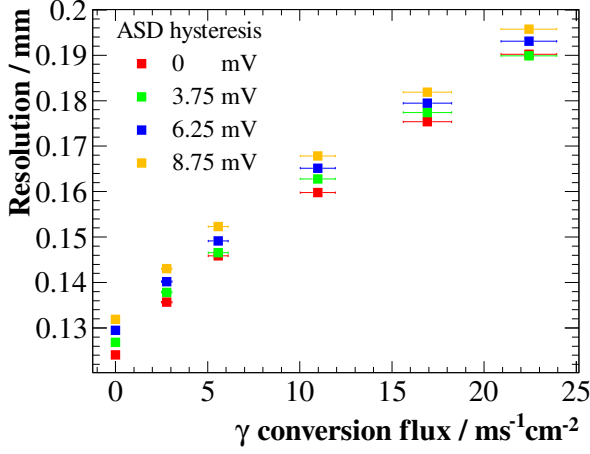


Figure 6.22: Dependence of the sMDT spatial resolution on the γ flux for ASD readout with different hysteresis settings and nominal threshold setting of -39 mV.

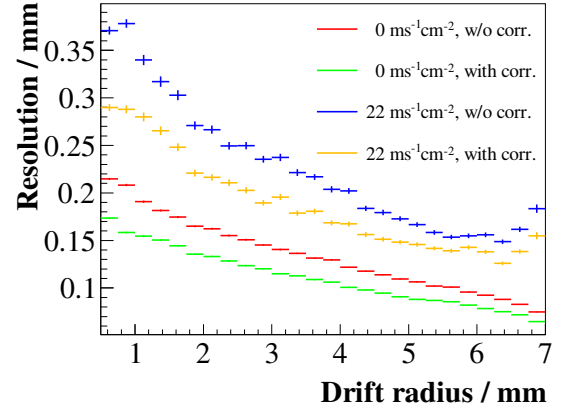


Figure 6.23: Dependence of the sMDT spatial resolution on the drift radius for ASD readout with and without time walk correction at different γ conversion fluxes.

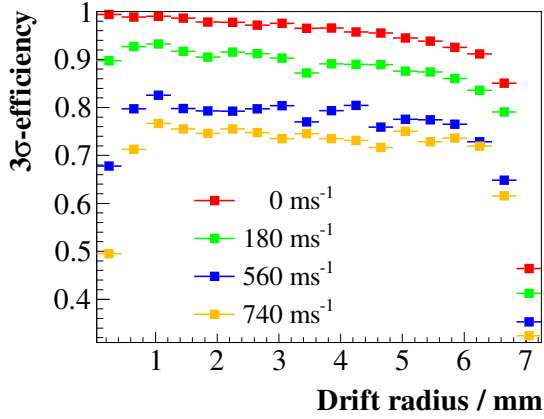


Figure 6.24: Dependence of the 3σ muon efficiency on the drift radius for different tube hit rates for ASD readout.

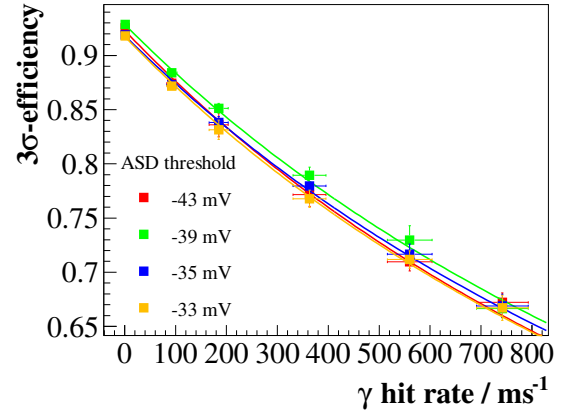


Figure 6.25: Dependence of the 3σ muon efficiency on the γ hit rate for ASD readout with different threshold settings.

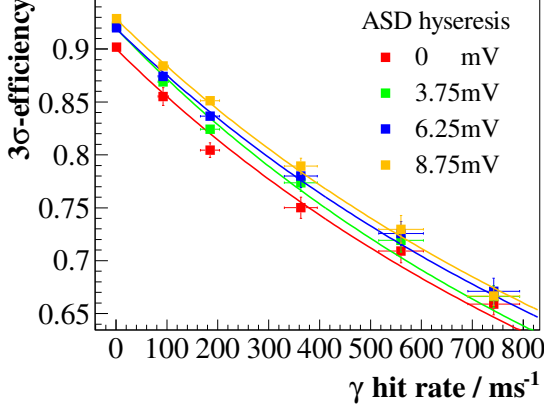


Figure 6.26: Dependence of the 3σ -efficiency on the γ hit rate for ASD readout at the nominal threshold setting of -39 mV for varying hysteresis settings.

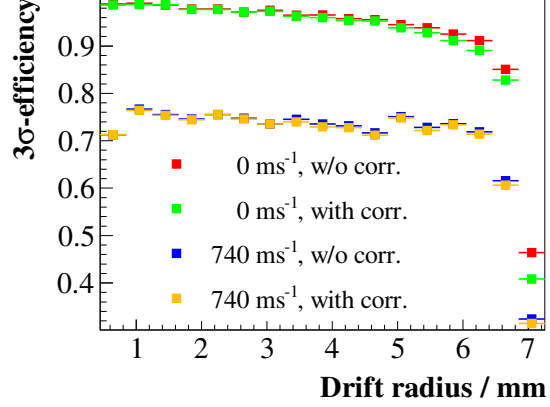


Figure 6.27: Dependence of the 3σ -efficiency on the drift radius for ASD readout with and without time slewing correction for zero and maximal γ hit rate.

probability by δ electrons. Above 6.5 mm radius, the muon path inside the tube is getting too short to ensure sufficient primary ionization for a threshold crossing and the 3σ -efficiency drops steadily towards the tube wall. With increasing γ irradiation rate, more muon hits are masked by γ hits, which decreases the efficiency. At the highest irradiation rates, the drop of the gas gain leads to additional inefficiency at small radii.

The dependence of the average 3σ -efficiency on the γ hit rate is shown in Figure 6.25 for different thresholds at a hysteresis setting of 8.75 mV. Fitting Equation (4.3) for nominal ASD threshold and hysteresis setting to the data gives an effective deadtime of $t_{\text{dead}} = (500 \pm 30)$ ns. This value is consistent with the value obtained in Section 6.2.2 and shows that the deviation of the hit multiplicity m (see Section 4.2.1) from one is less than 10 % for the nominal electronics settings. The efficiency in the limit without irradiation from the fit is $\epsilon_0 = (93 \pm 1)$ %, in agreement to the result in [15]. Reducing the threshold slightly reduces the 3σ -efficiency, because muon hits are more likely to be masked by noise hits. Threshold values above the nominal value lead to loss of muon signals with small amplitudes which increasingly occur at high irradiation flux because of gain loss. Compared to [15], the γ counting rates and masking effects are lower in spite of the higher γ flux because of the small tube length.

The effect of the hysteresis setting at the nominal threshold of -39 mV on the 3σ -efficiency is shown in Figure 6.26. When the hysteresis voltage is lowered from its nominal value of 8.75 mV, more muons are masked by electronics noise hits reducing the 3σ -efficiency independently of the irradiation rate. In

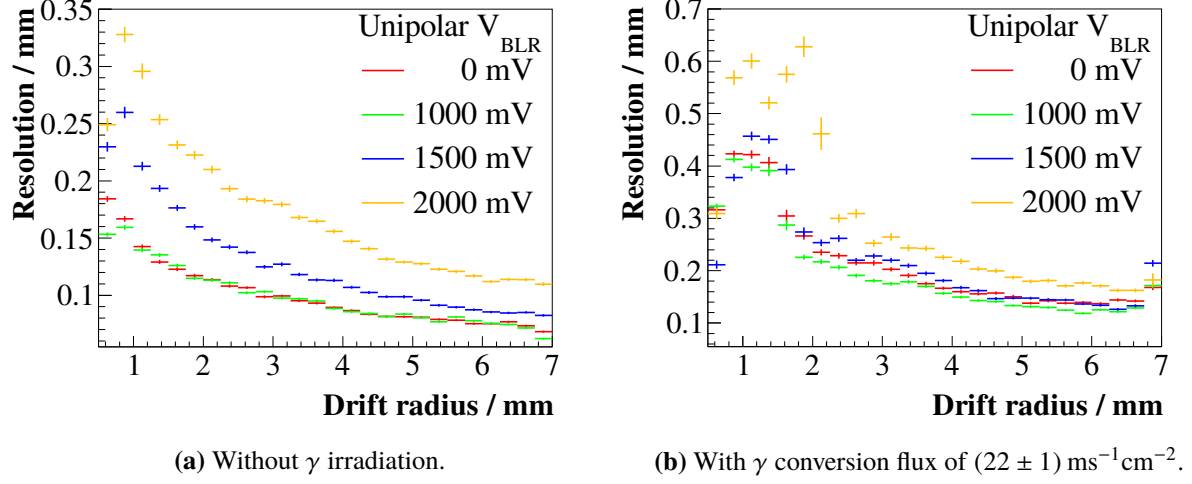


Figure 6.28: Dependence of the sMDT spatial resolution on the drift radius at zero and maximum recorded γ flux for the unipolar shaper readout with different V_{BLR} settings. The drop of the resolution close to the wire is an artifact of the track reconstruction (see text).

addition, the hit multiplicity m is increased by 10 % which leads to a steeper decrease of the efficiency with the γ rate, because of the longer effective deadtime.

The effect of time walk corrections on the 3σ -efficiency is shown in Figure 6.27 for two γ hit rates. Due to the improvement of the resolution, more muon hits are masked by δ electrons at large drift radii. This effect reduces the average 3σ -efficiency by roughly 0.5 % independent of the irradiation rate.

6.5 Unipolar Shaper Readout

6.5.1 Spatial Resolution

Figure 6.28 shows the the spatial resolution as a function of the drift radius for the unipolar shaper readout with different BLR voltages and the nominal thresholds adopted for each BLR voltage setting (see Table 6.1) at zero and maximum γ irradiation. Without irradiation and for low BLR voltage ≤ 1000 mV, the unipolar readout provides a resolution of $170 \mu\text{m}$ close to wire and of better than $70 \mu\text{m}$ at large drift radii. Increasing BLR voltages deteriorate the resolution due to the increasing signal attenuation by the BLR stage. At the high γ irradiation rates, raising the BLR voltage to 1000 mV leads to a significant improvement of the resolution compared to $V_{\text{BLR}} = 0$ mV for all drift radii. For 1500 mV BLR voltage, the resolution becomes only slightly worse than for 0 mV when the baseline stabilization and pulse amplitude loss effects approximately compensate each other. For $V_{\text{BLR}} = 2000$ mV the effect of signal attenuation

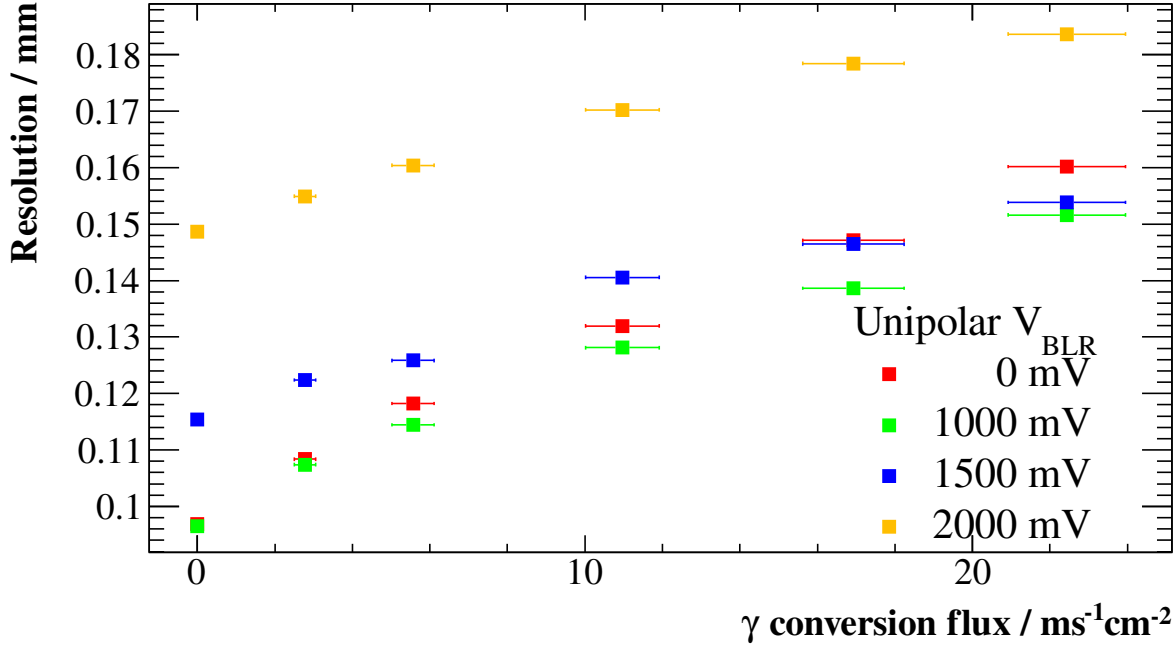


Figure 6.29: Dependence of the mean sMDT spatial resolution on the γ conversion for unipolar shaper readout with different BLR voltage settings at nominal threshold.

(cf. Section 5.1.2.1) is already too strong to be able to benefit from the improved baseline stability. The resolution drop at very small drift radii is an artefact of inefficient track reconstruction close to wire (cf. Figure 6.31) and much stronger than for the ASD readout due to the higher discriminator thresholds (cf. Section 6.8.4).

A comparison of the drift tube resolutions as a function of the γ conversion flux for four BLR voltage settings is shown in Figure 6.29. It shows increasing improvement of the resolution due to the baseline restoration at high irradiation rates for moderate BLR voltages around $V_{\text{BLR}} = 1000$ mV and no deterioration at low γ rates. gives the best resolution. Higher BLR voltages up to 1500 mV may become be beneficial for the resolution at the highest rates.

The effect of ideal BLR is illustrated by applying an artificial deadtime of $\Delta t = 400$ ns between adjacent hits in the data analysis suppressing pileup effects. For $V_{\text{BLR}} \leq 1000$ mV this leads to a significant improvement of the spatial resolution at high rates of similar magnitude (see Figure 6.30). For $V_{\text{BLR}} = 1500$ mV, the improvement is significantly smaller and zero for $V_{\text{BLR}} = 2000$ mV due to the increased BLR strength.

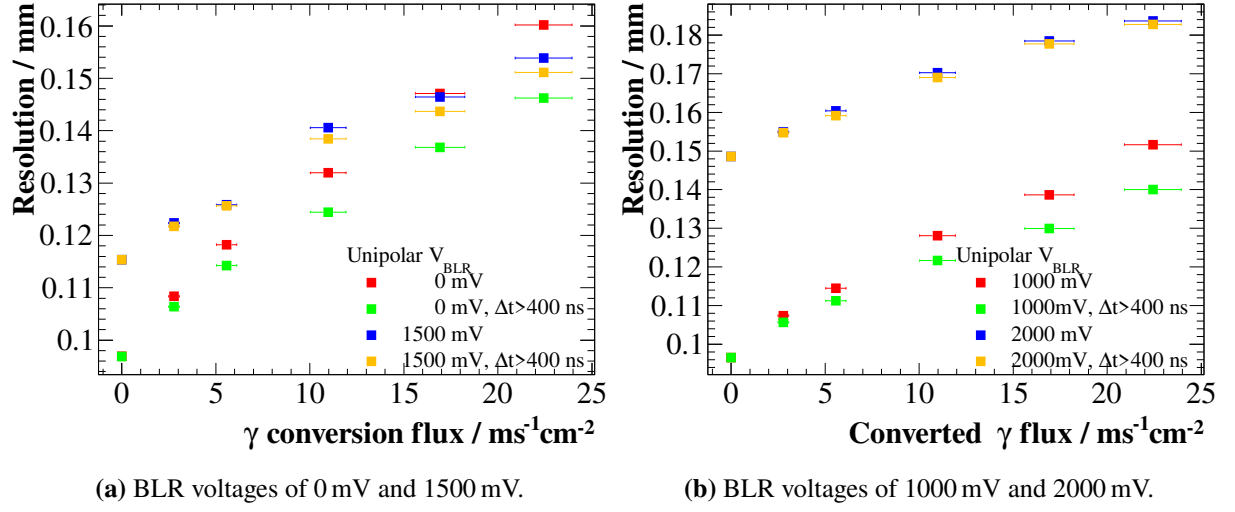


Figure 6.30: Effect of a minimum time difference requirement of $\Delta t > 400$ ns between adjacent pulses on the mean sMDT spatial resolution as a function of the γ conversion flux for the unipolar shaper readout at different BLR voltages.

6.5.2 Muon Efficiency

Figure 6.31 shows the 3σ -efficiency as a function of the drift radius without γ irradiation and for maximum achieved γ hit rate for the nominal threshold adopted for each BLR voltage setting. Without irradiation, BLR voltages up to 1500 mV give the same result as for the ASD readout reaching 100 % efficiency close to the wire and decreasing slowly with increasing drift radius due to δ electrons. For $V_{\text{BLR}} = 2000$ mV, the increased signal amplitude reduction by the BLR stage leads to significant efficiency loss at small drift radii, as signals with small amplitudes cross the threshold too late to be identified as matching muon hits. At the γ hit rate of $(740 \pm 50) \text{ ms}^{-1}$, the unipolar shaper electronics achieves significantly higher efficiencies than the ASD readout already at zero BLR voltage, due to the not fixed and effectively smaller deadtime. For $V_{\text{BLR}} = 1000$ mV, improved baseline stability increases the 3σ -efficiency. Higher BLR voltages lead to increasing efficiency losses near the wire. A BLR voltage of 1500 mV, however, leads to better resolution than $V_{\text{BLR}} = 1000$ mV for drift radii $r > 5.5$ mm, where the baseline stability has the largest effect (see Section 6.5.1).

The 3σ -efficiency dependence on the γ hit rate is shown in Figure 6.32 for the different BLR voltage settings and the corresponding nominal thresholds. Fitting the data with Equation (4.3) in Section 3.2 gives an effective deadtime of $m \cdot t_{\text{dead}} = (340 \pm 20) \text{ ns}$ for $V_{\text{BLR}} = 0$ mV and $V_{\text{BLR}} = 1500$ mV, where m is the threshold crossing multiplicity per hit. Setting the BLR voltage to 1000 mV gives $m \cdot t_{\text{dead}} = (250 \pm 20) \text{ ns}$ significantly improving the 3σ -efficiency at high rates. For $V_{\text{BLR}} \geq 1500$ mV, the 3σ -efficiency decreases

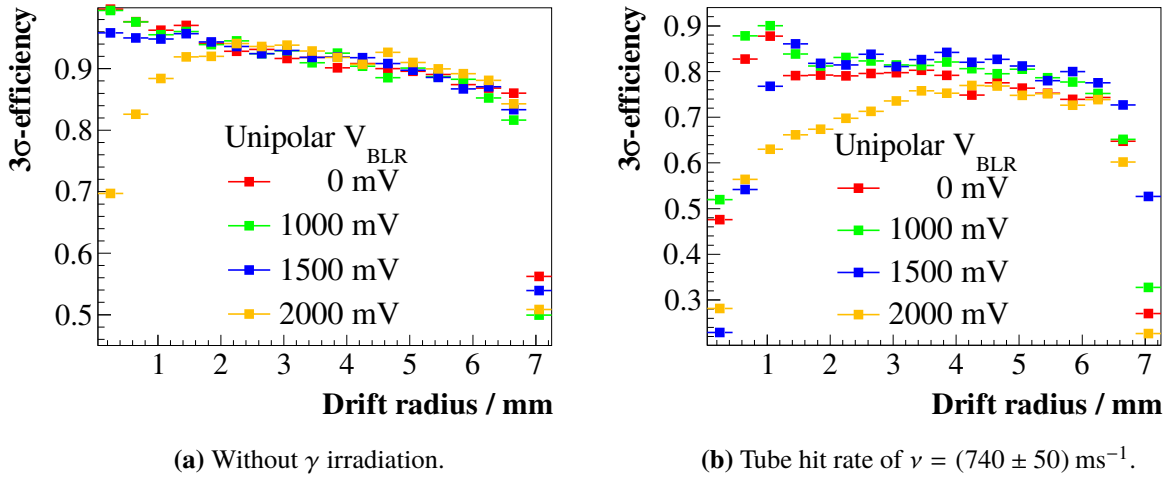


Figure 6.31: Dependence of the 3σ -efficiency on the drift radius with the discrete unipolar readout at different BLR settings without and with the maximum recorded γ irradiation.

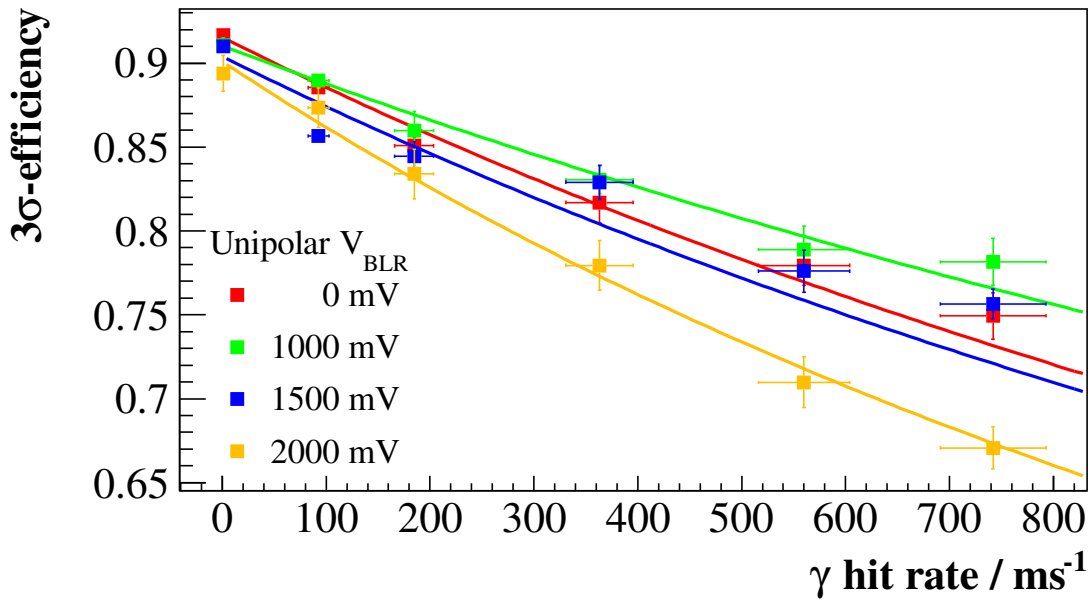


Figure 6.32: Dependence of the average 3σ -efficiency with the discrete unipolar readout on the γ hit rate for different BLR voltages settings at the nominal thresholds settings.

again. The fit to the efficiencies for $V_{\text{BLR}} = 1500$ mV is affected by an outlier at low γ irradiation and the efficiency at high rates indicate that the resulting effective deadtime is an upper limit and is closer to the one of $V_{\text{BLR}} = 1000$ mV for even higher rates. Especially the BLR value of 2000 mV leads to a significant loss of 3σ -efficiency due to the damping of the signal amplitude, leading to an effective deadtime of $m \cdot t_{\text{dead}} = (460 \pm 40)$ ns. The large deviations of the 3σ -efficiency from the fitted expectation curve at several irradiations show that the assumption of a constant effective deadtime is not sufficient to fully describe the efficiency but models the trend of the efficiency to allow for a direct comparison between readout electronics and their settings. As $V_{\text{BLR}} = 1000$ mV gives the best resolution and 3σ -efficiency for the unipolar readout electronics, the optimum working point should be close to this value.

6.5.3 Dependence on the Discriminator Threshold

Measurements were also performed at lower discriminator threshold settings to study the achievable resolution and efficiency for the case of lower noise levels expected for implementation of the discrete circuit as an ASIC.

The achieved resolutions are shown in Figure 6.33. Lower thresholds reduce the resolution deterioration caused by time walk (cf. Equation (6.3)), improving the average sMDT resolution.

The achieved 3σ efficiencies are shown in Figure 6.34. Lower thresholds reduce the 3σ efficiency, as higher noise counting rates mask more muon hits and improved resolutions cut a larger fraction of hits caused by δ electrons.

The resolution improvement and efficiency loss at lower thresholds show that the threshold choice based on the noise counting rate (cf. Section 6.2.1) leads to reliable results.

6.6 Bipolar Shaper Readout

6.6.1 Spatial Resolution

Figure 6.35 shows the spatial resolution as a function of the drift radius without irradiation and for maximum irradiation rate at the nominal threshold settings (Table 6.1) and different BLR voltages. The BLR voltage has a significantly smaller influence on the resolution compared to the unipolar readout electronics. The resolution without irradiation does not significantly deteriorate for $V_{\text{BLR}} > 0$ mV. It is, like for the unipolar shaper readout, significantly better than for the ASD chip readout, because of the shorter peaking time. At maximum irradiation, the BLR stage with $V_{\text{BLR}} > 0$ mV improves the resolution for medium and large drift radii. Even for the highest BLR voltage, no effect of signal attenuation and high thresholds is visible leading to better resolution at maximum irradiation than for the unipolar shaper readout.

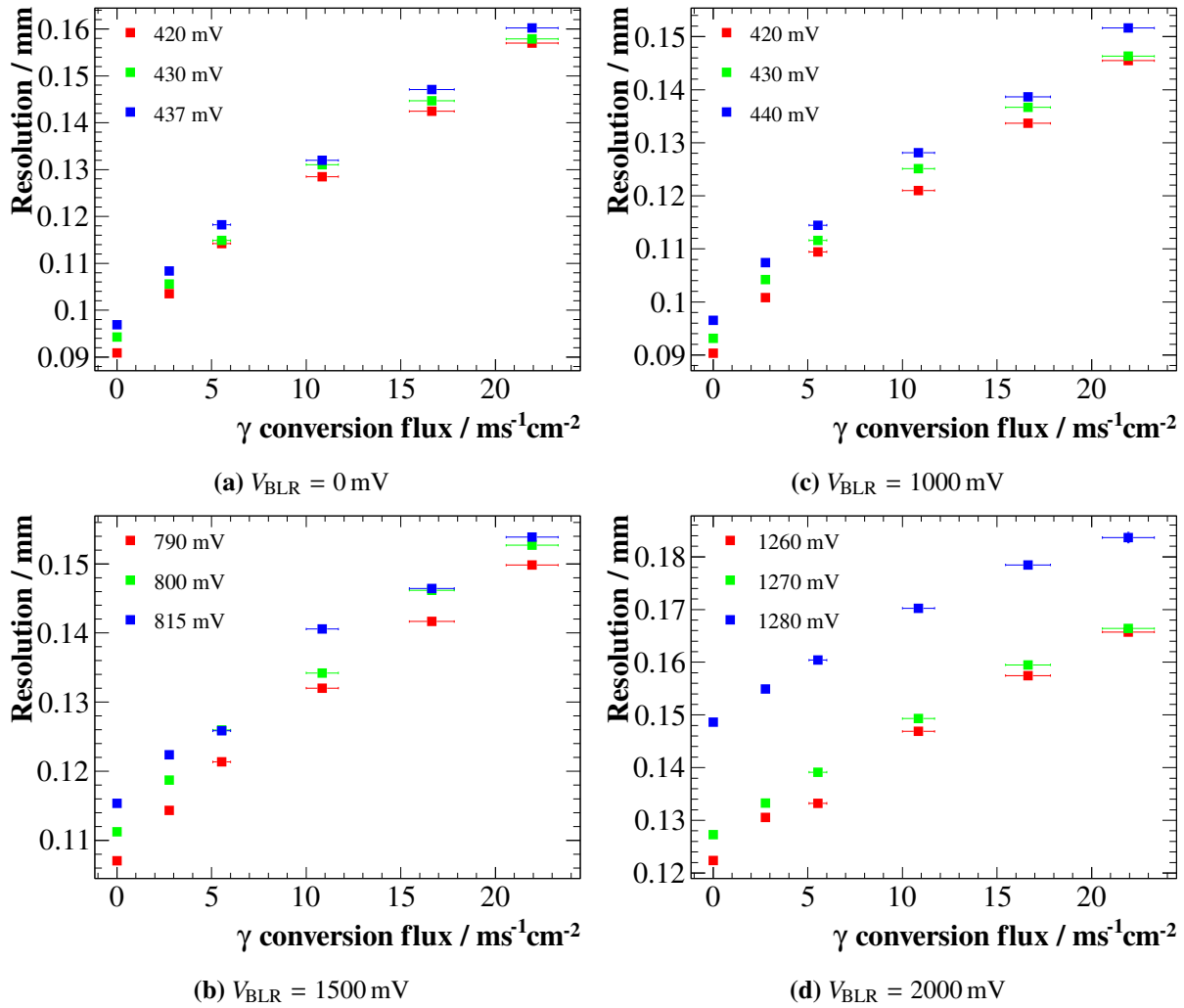


Figure 6.33: Dependence of the sMDT resolution on the γ -flux for unipolar shaper readout with the different investigated BLR voltages and thresholds.

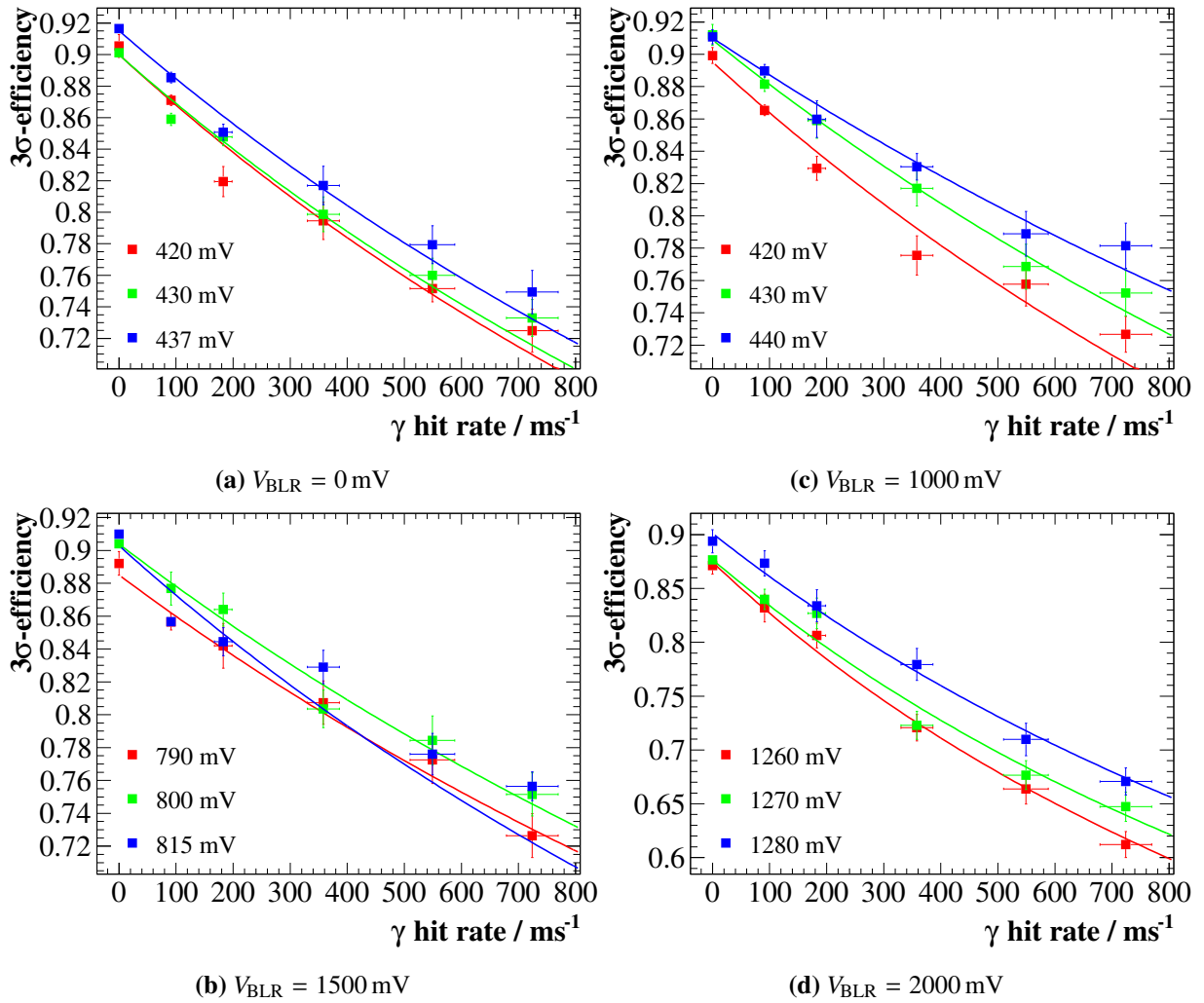


Figure 6.34: Dependence of the average 3- σ efficiency on the γ hit rate for unipolar shaper readout with the different investigated BLR voltages and thresholds.

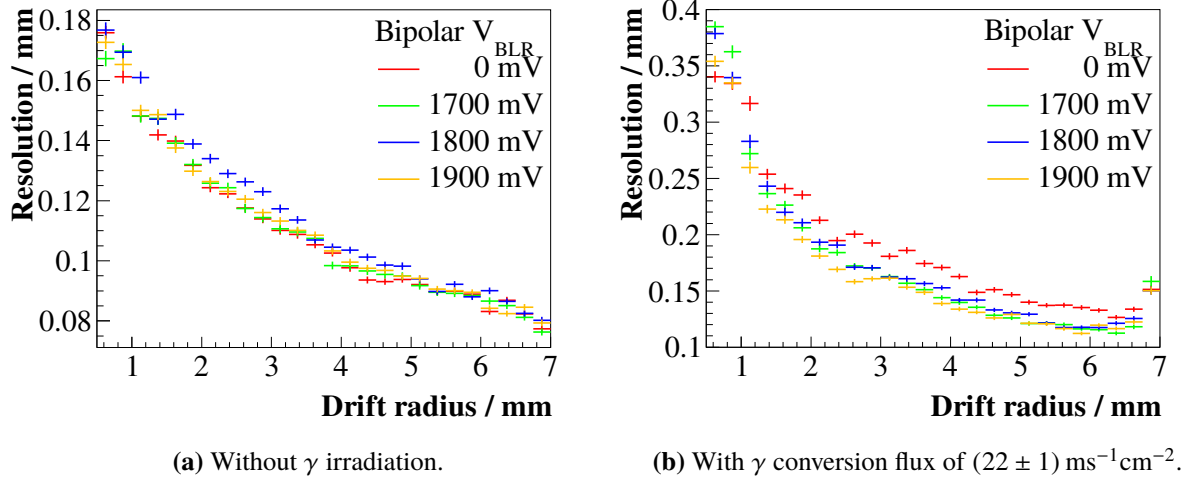


Figure 6.35: Dependence of the spatial resolution on the drift radius with the discrete bipolar readout at different BLR settings without and with the maximum recorded γ flux.

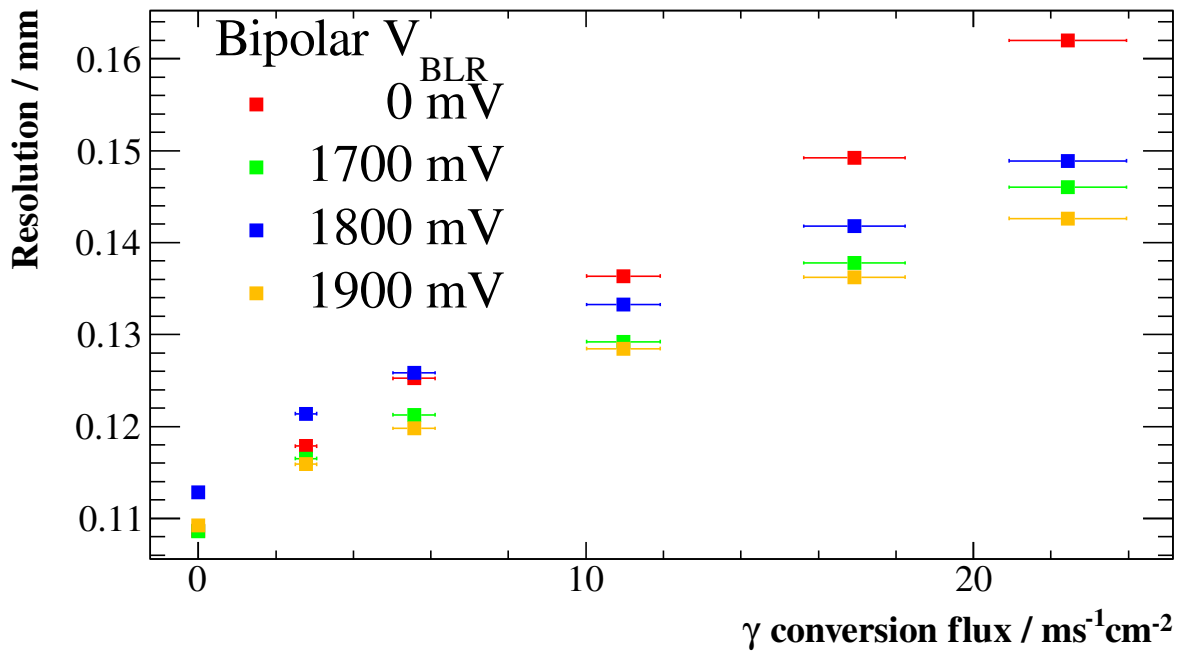


Figure 6.36: Dependence of the average sMDT resolution on the irradiation rate for the discrete bipolar readout with nominal signal thresholds settings and different BLR voltages.

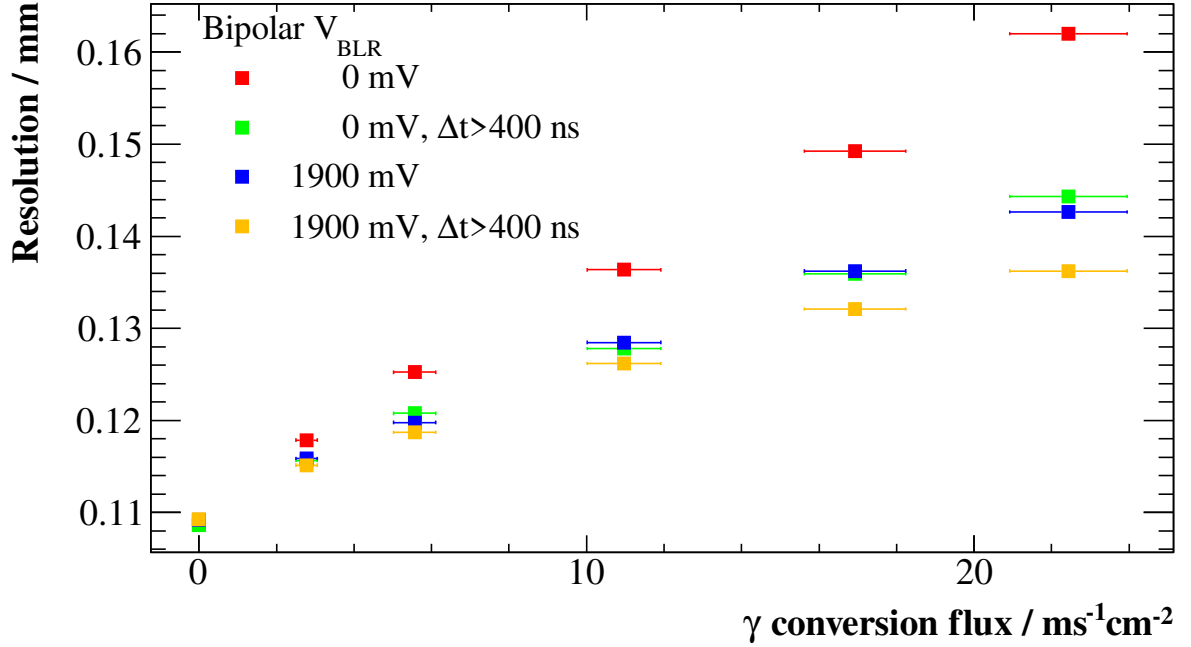


Figure 6.37: Effect of a minimum time difference requirement of $\Delta t > 400$ ns between adjacent pulses on the average sMDT resolution as a function of the γ conversion flux for the bipolar electronics with $V_{\text{BLR}} = 0$ mV and maximum BLR voltage used.

A comparison of the tube resolutions as a function of the γ conversion flux is shown in Figure 6.36 for the four different BLR voltages. For $V_{\text{BLR}} = 0$ mV, high hit rates cause a baseline drift, which is compensated by applying a BLR voltage $V_{\text{BLR}} > 0$ mV (compare Section 5.1.3.2). The effects of the three applied BLR voltages cannot be distinguished within the systematic uncertainty of $3 \mu\text{m}$ without γ irradiation to about $30 \mu\text{m}$ at the highest γ conversion flux. The maximum BLR voltage $V_{\text{BLR}} = 1900$ mV used tends to give the best resolution.

The signal pileup effects and potential improvements by BLR are clearly demonstrated by applying an offline cut $\Delta t > 400$ ns in the time interval between successive hits. For $V_{\text{BLR}} = 0$ mV, this leads to a significant improvement of the spatial resolution, as signal pileup effects are suppressed (see Figure 6.37). For $V_{\text{BLR}} = 1900$ mV, the difference is significantly smaller, showing the effect of the BLR stage. For $V_{\text{BLR}} = 1700$ mV and $V_{\text{BLR}} = 1800$ mV, the effect is on the same level.

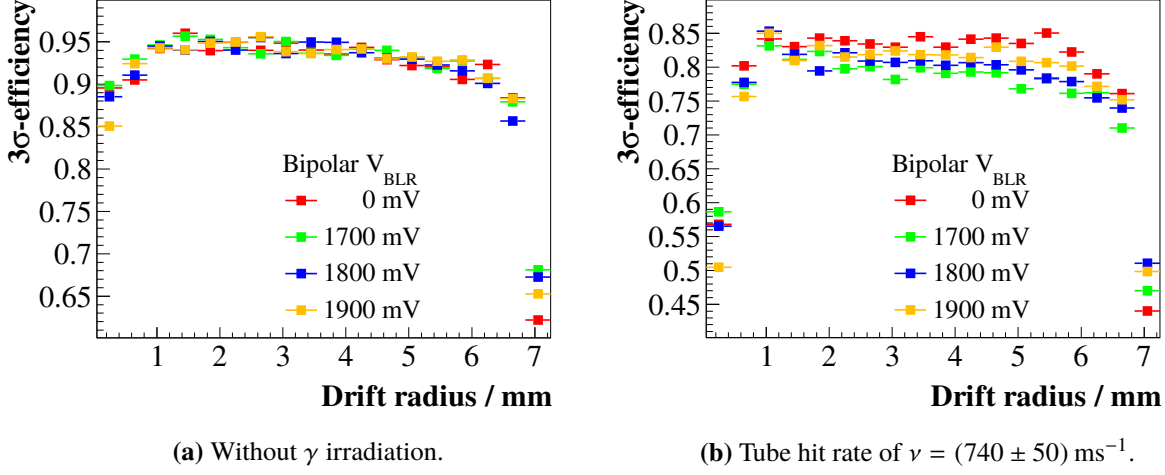


Figure 6.38: Dependence of the 3σ -efficiency on the drift radius for bipolar shaper readout with different BLR voltage settings without irradiation and with maximal γ irradiation.

6.6.2 Muon Efficiency

Figure 6.38 shows the 3σ -efficiency as a function of the drift radius with the bipolar shaper readout with different BLR voltages and nominal thresholds adopted for each BLR voltage setting. Without irradiation, the 3σ -efficiency for drift radii $r > 3$ mm has a similar shape like for the ASD chip and unipolar shaper readout. The drop close to the wire is a consequence of the delayed threshold crossing of signals with small amplitudes, which are not identified as muon hits and are hence lost in the 3σ -efficiency.

Without irradiation, no significant effect of the BLR voltage setting can be seen. At high irradiation rate, the 3σ -efficiency is reduced due to masking of muon hits by preceeding γ hits and the drop at small drift radii is more pronounced because of gas gain loss. Due to the additional overshoots (see Section 5.1.2.2), the 3σ -efficiency at high γ hit rates is reduced for $V_{\text{BLR}} > 0$ mV but less with increasing BLR voltage.

The 3σ -efficiency dependence on the γ hit rate is shown in Figure 6.39 for the different BLR voltage settings and the corresponding nominal thresholds. Fits of Equation (4.3) to the data give a 3σ -efficiency without irradiation of $\epsilon_0 = (93 \pm 1) \%$ for all BLR voltages. The large deviations of the 3σ -efficiency from the fitted expectation curve at several irradiations show that the assumption of a constant effective deadtime is not sufficient to fully describe the efficiency but models the trend of the efficiency to allow for a direct comparison between readout electronics and their settings. For $V_{\text{BLR}} = 0$ mV, the effective deadtime is $m \cdot t_{\text{dead}} = (260 \pm 20)$ ns. Applying a BLR voltage leads to additional overshoots (compare Section 6.2.3), which increase the effective deadtime. For $V_{\text{BLR}} = 1700$ mV, the effective deadtime

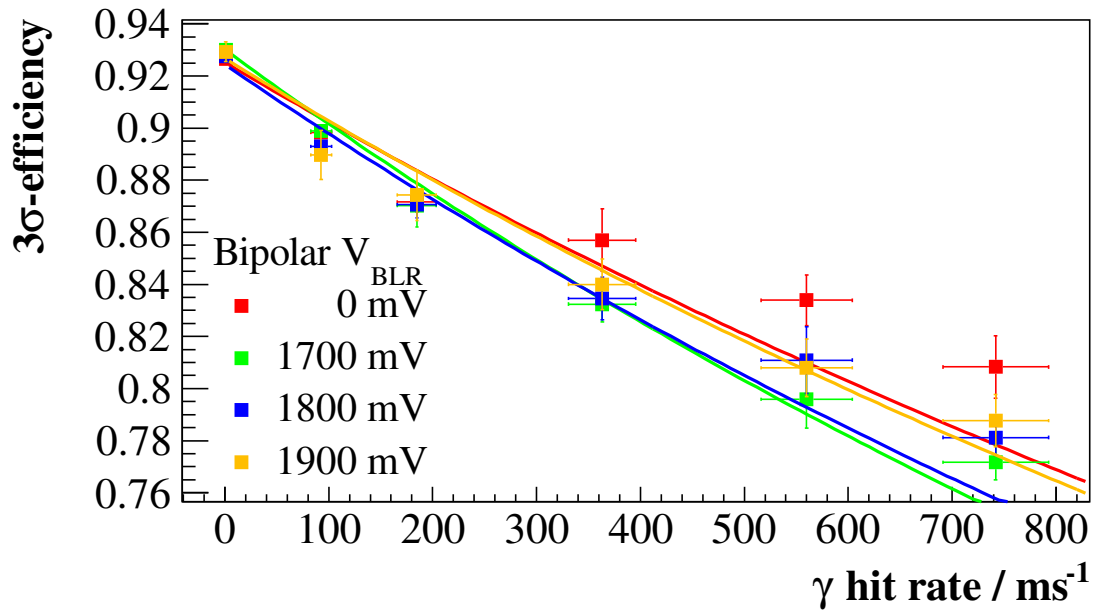


Figure 6.39: Dependence of the average 3σ -efficiency on the γ hit rate for the bipolar shaper with different BLR voltage settings and the corresponding nominal thresholds.

is $m \cdot t_{\text{dead}} = (320 \pm 20)$ ns. Higher BLR voltages reduce the width and amplitude of the additional overshoots, leading to effective deadtimes of (300 ± 20) ns and (270 ± 20) ns for $V_{\text{BLR}} = 1800$ mV and $V_{\text{BLR}} = 1900$ mV, respectively.

The $V_{\text{BLR}} = 1900$ mV operating point is preferred because of the improvement in resolution at high rates. The 3σ -efficiency shows a small preference for $V_{\text{BLR}} = 0$ mV but is not significant. If operated with $V_{\text{BLR}} \geq 1700$ mV the studies with generator pulses in Section 5.1.2.2 tends to prefer higher BLR voltages due to the smaller secondary overshoots (Figure 5.38). Hence, $V_{\text{BLR}} = 1900$ mV is used as nominal setting for the discrete bipolar readout in the following.

6.6.3 Dependence on the Discriminator Threshold

Measurements were also performed at lower discriminator threshold settings to study the achievable resolution and efficiency for the case of lower noise levels expected for implementation of the discrete circuit as an ASIC.

The achieved resolutions are shown in Figure 6.40. Like for the unipolar shaper, lower thresholds reduce the resolution deterioration caused by time walk (cf. Equation (6.3)), improving the average sMDT resolution.

The achieved $3\text{-}\sigma$ efficiencies are shown in Figure 6.41. Lower thresholds reduce the $3\text{-}\sigma$ efficiency, as higher noise counting rates mask more muon hits and improved resolutions cut a larger fraction of hits caused by δ electrons. With $V_{\text{BLR}} > 0$ mV, lower thresholds increase the number of threshold crossings of the second overshoot (compare Section 5.1.2.2), increasing the effective dead time and reducing the $3\text{-}\sigma$ efficiency at high γ hit rates.

The resolution improvement and efficiency loss at lower thresholds show that the threshold were chosen well, based on the noise counting rate (cf. Section 6.2.1).

6.7 Comparison of the Electronics Performances

In this chapter, the resolution and efficiency of sMDT with ASD chip and discrete unipolar and bipolar readout electronics with their nominal thresholds are compared.

Figure 6.42 shows the resolution for ASD chip readout and discrete unipolar and bipolar shaper readout with $V_{\text{BLR}} = 0$ mV. The ASD chip readout without time slewing corrections provides significantly worse resolution than discrete shaper readout with BLR due to the smaller bandwidth and resulting longer signal rise time of the ASD chip. With time slewing corrections, the difference is smaller but still significant due to shorter rise time of the discrete shaper circuits and BLR at high rates. The unipolar electronics shaper provides the best resolution without γ irradiation.

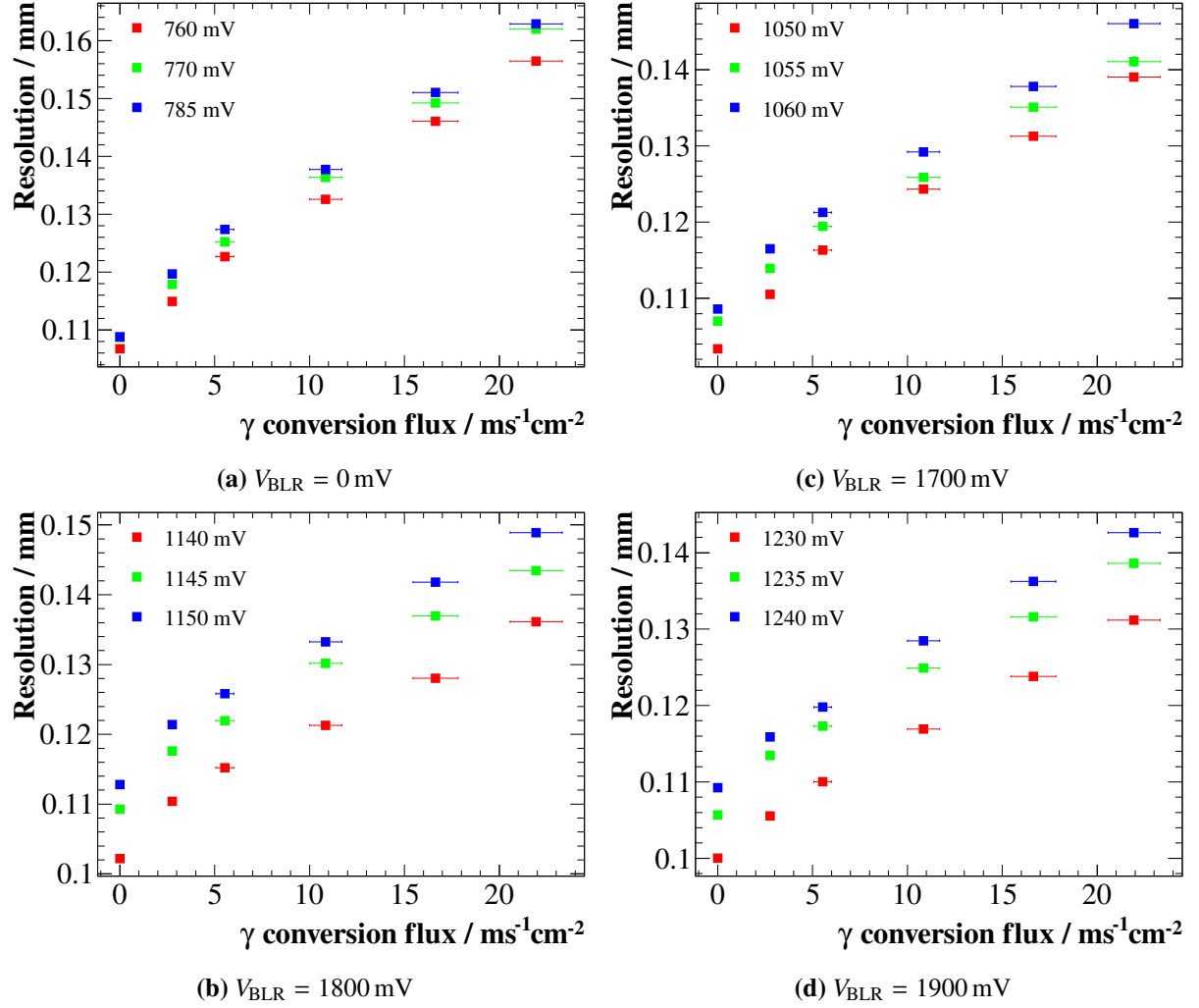


Figure 6.40: Dependence of the sMDT resolution on the γ -flux for bipolar shaper readout with the different investigated BLR voltages and thresholds.

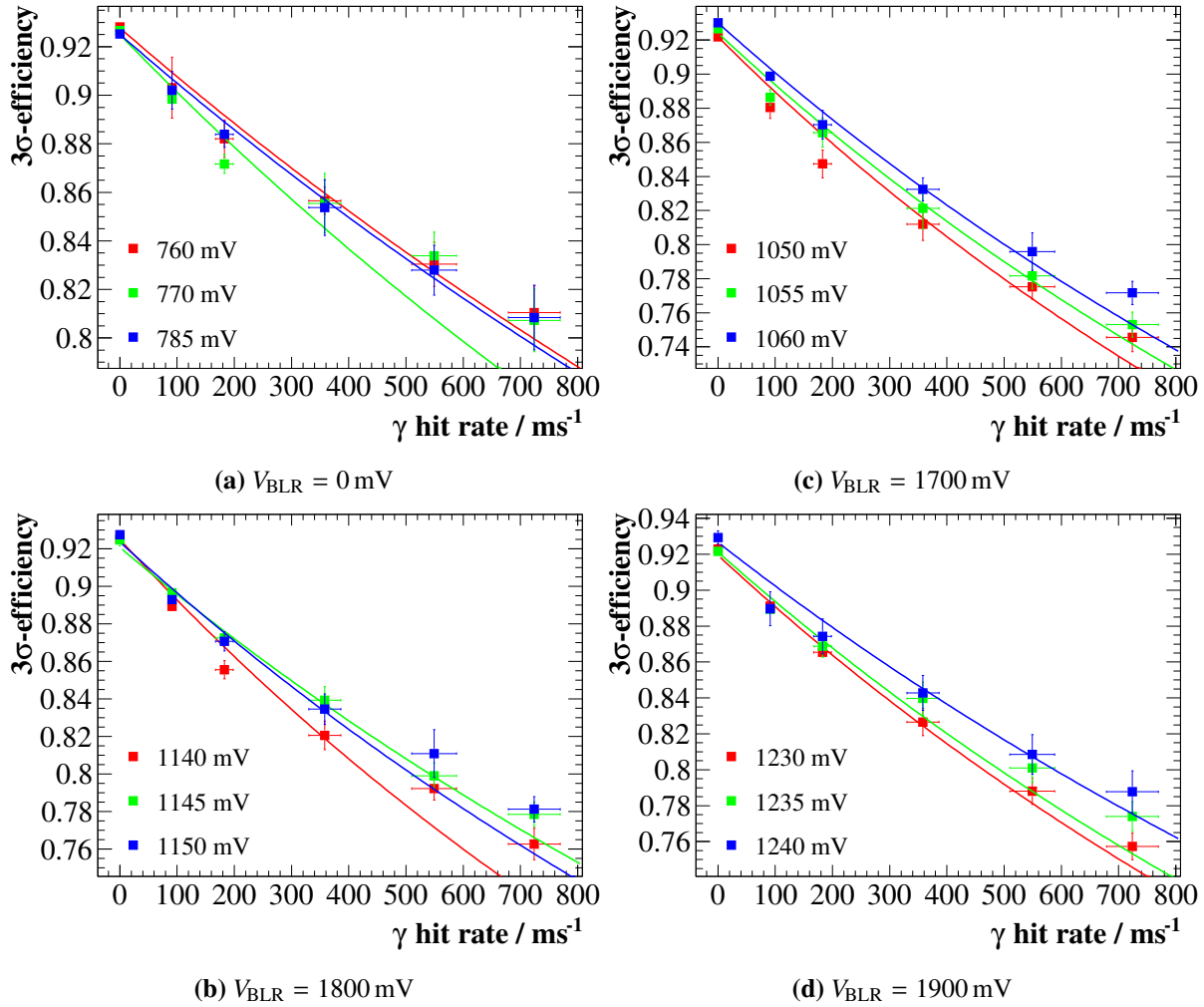


Figure 6.41: Dependence of the average 3σ efficiency on the γ hit rate for unipolar shaper readout with the different investigated BLR voltages and thresholds.

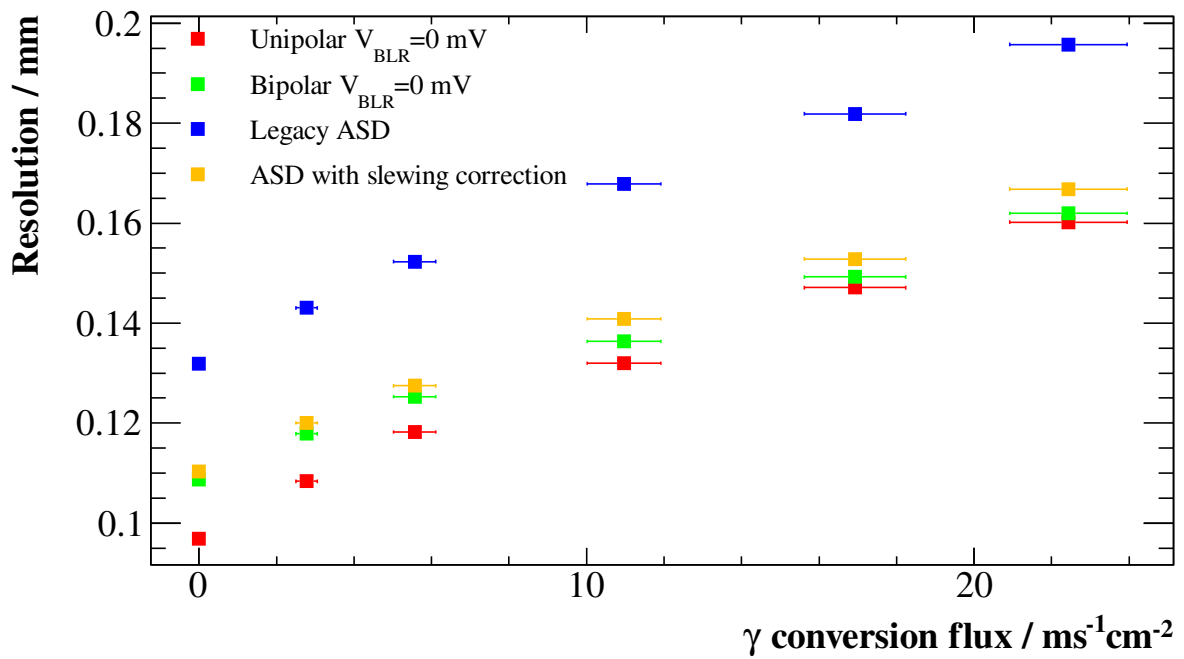


Figure 6.42: Comparison of the average resolution according to Equation (3.4), page 47, as a function of the γ conversion flux with the ASD chip and discrete uni- and bipolar readout electronics with $V_{\text{BLR}} = 0 \text{ mV}$. The resolution for ASD chip readout with time slewing corrections based on the ADC charge measurement is also shown.

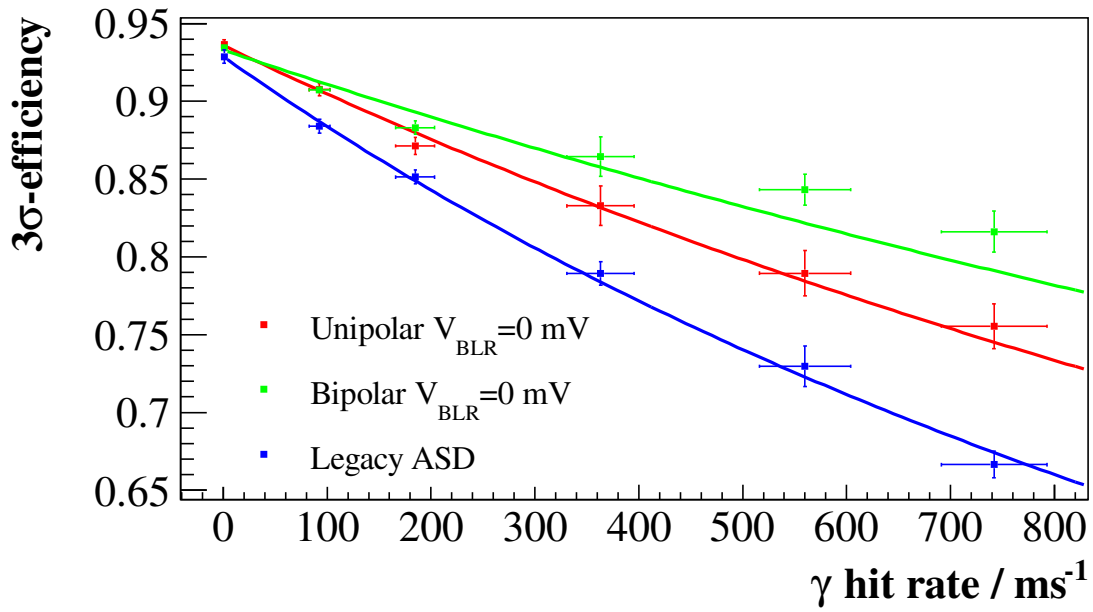


Figure 6.43: Comparison of the dependence of the averaged 3σ -efficiency on the tube hit rate with the different readout electronics for $V_{\text{BLR}} = 0$ mV. To allow for direct comparison, the resolution of the ASD readout was used to calculate the 3σ -efficiency.

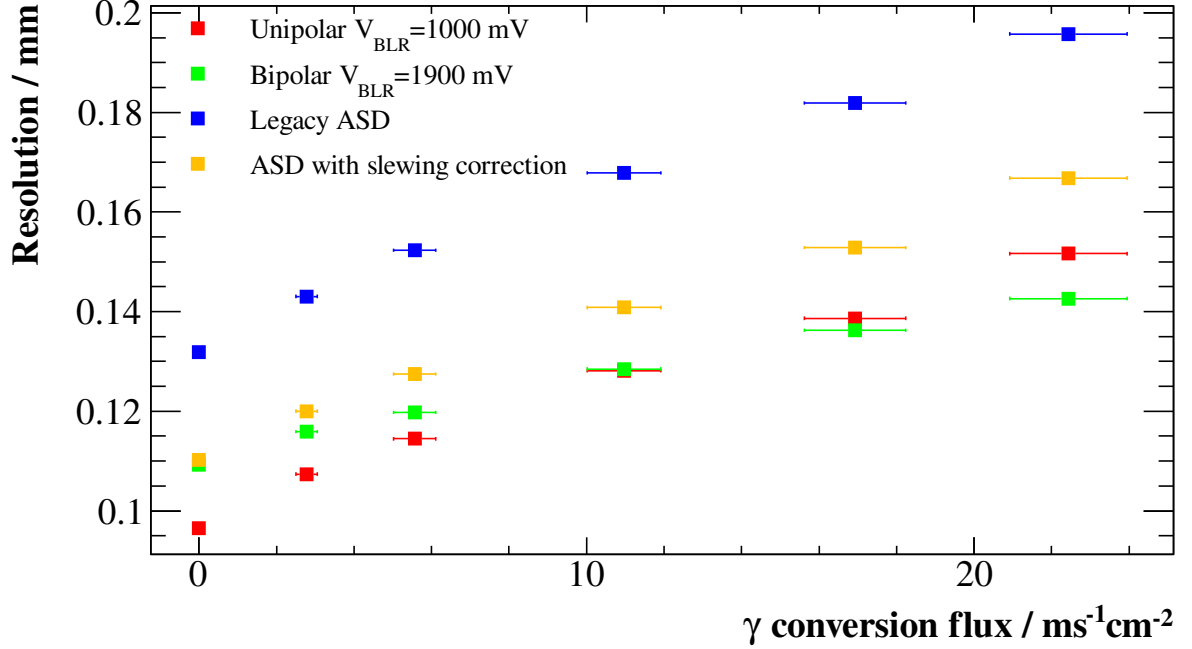


Figure 6.44: Comparison of the average resolution according to Equation (3.4), page 47, as a function of the γ conversion flux with the ASD chip and discrete uni- and bipolar readout electronics at their nominal operating parameters. The resolution for ASD chip readout with time slewing corrections based on the ADC charge measurement is also shown.

The dependence of the sMDT 3σ -efficiency on the γ hit rate is shown in Figure 6.45 for ASD chip readout and discrete unipolar and bipolar shaper readout with $V_{\text{BLR}} = 0$ mV. The same resolution as a function of drift radius provided by the ASD chip readout is used for the determination of the 3σ -efficiency for all readout electronics (see Section 6.3.3). The efficiency drop for the ASD chip readout in ADC mode corresponds to an effective deadtime $m \cdot t_{\text{dead}} = (500 \pm 30)$ ns with a minimum deadtime of 300 ns. The discrete shaper circuits operate in time-over-threshold mode leading to an effective deadtime of (340 ± 30) ns for the unipolar and of (240 ± 20) ns for the bipolar shaper. The discrete shapers with BLR with V_{BLR} lead to a strong improvement of the 3σ -efficiency increasing with the γ counting rate due to the smaller deadtime.

Figure 6.44 shows the resolution for the three readout electronics at their nominal operation parameters. The ASD chip readout without time slewing corrections provides significantly worse resolution than discrete shaper readout with BLR due to the smaller bandwidth and resulting longer signal rise time of the ASD chip. With time slewing corrections, the difference is smaller but still significant due to shorter rise time of the discrete shaper circuits and BLR at high rates. The unipolar electronics shaper provides the

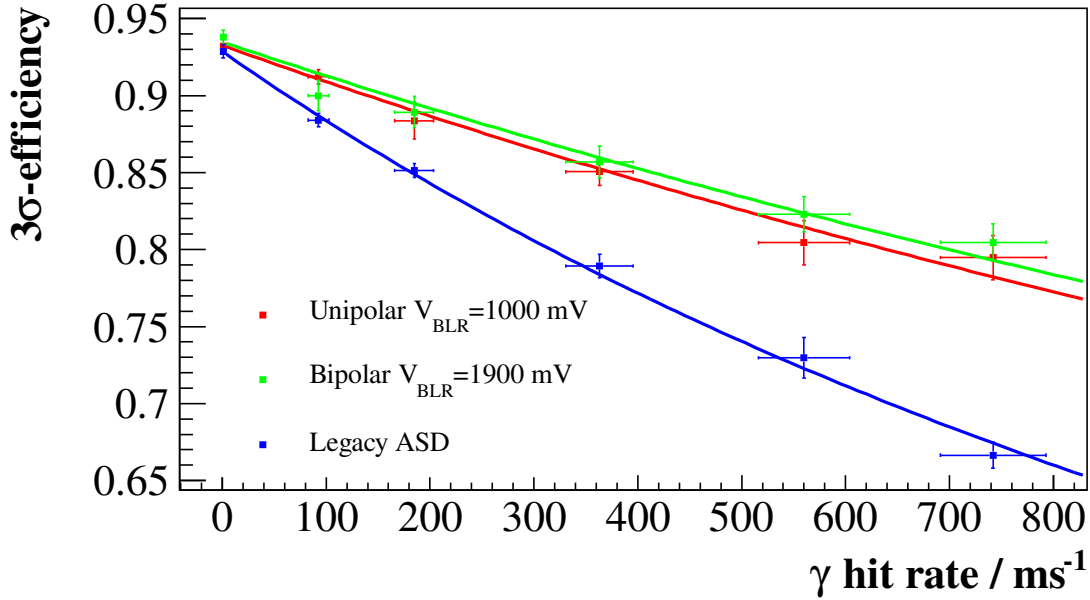


Figure 6.45: Average 3σ -efficiency of sMDT as a function of the γ hit rate for the different readout electronics at their nominal operating parameters. To allow for direct comparison, the resolution determined with ASD chip readout was used to determine the 3σ -efficiency.

best resolution without γ irradiation, but the degradation with increasing rate is faster than for the discrete bipolar shaper readout as a consequence of the increasing baseline distortions due to missing voltage buffers on the boards (compare Section 5.1.3.1).

The dependence of the sMDT 3σ -efficiency on the γ hit rate is shown in Figure 6.45 for the three readout electronics at their nominal operating parameters. The same resolution as a function of drift radius provided by the ASD chip readout is used for the determination of the 3σ -efficiency for all readout electronics (see Section 6.3.3). The efficiency drop for the ASD chip readout in ADC mode corresponds to an effective deadtime $m \cdot t_{\text{dead}} = (500 \pm 30)$ ns with a minimum deadtime of 300 ns. The discrete shaper circuits operate in time-over-threshold mode leading to an effective deadtime of (260 ± 20) ns for the unipolar and of (240 ± 20) ns for the bipolar shaper. The discrete shapers with BLR with nominal V_{BLR} lead to a strong improvement of the 3σ -efficiency increasing with the γ counting rate. The main improvement is due to the smaller deadtime.

6.8 Simulation of High Rate Effects

As described in the previous chapters, high γ hit rates deteriorate the performance of the MDT and, to much lesser extend, of the sMDT muon precision tracking chambers due to gas gain drop, space charge fluctuations, masking of muon hits by background hits and signal pileup in the readout electronics. To better quantify the impact of these effects on the spatial resolution and efficiency and to interpret the measurements in the Gamma Irradiation Facility GIF⁺⁺ for the different readout electronics, detailed simulation studies with the ATLAS MDT ASD and the discrete unipolar and bipolar shaper electronics have been performed using the Garfield++ program [89] and compared with the measurement data.

The ASD chip response was simulated for nominal threshold and hysteresis settings. The discrete unipolar and bipolar shaper responses was simulated for $V_{BLR} = 0$ mV and nominal thresholds. An overview of the electronics settings is given in Table 6.1.

6.8.1 Simulation Analysis

In Sections 6.4 to 6.7, the average spatial resolution of the tubes for given γ conversion flux is determined using Equation (3.4), here called “mean resolution”, to allow for direct comparison of the resolution and efficiency with the results of previous measurement campaigns. For the same reason, the 3σ -efficiency was used in the previous Sections 6.4 to 6.7, which also is the relevant quantity for the chamber tracking efficiency. The track residual distributions show tails which are not fully described by the mean resolution leading to systematic uncertainties caused by a dependence of the width on the fit range for Equation (3.4).

In this section, the average single tube spatial resolution $\langle\sigma\rangle$ is calculated according to

$$\langle\sigma\rangle = \sqrt{\frac{1}{r_{\max} - r_{\min}} \int_{r_{\min}}^{r_{\max}} dr (\sigma(r))^2}, \quad (6.5)$$

with the width $\sigma(r)$ of the track residual distribution at drift radius r . This resolution average is called “RMS resolution” and is better suited to compare the measurements to the simulation.

Due to the efficiency drop close to the tube wall and the steep radial dependence of the resolution near the wire, the averaging is here limited to the range between $r_{\min} = 0.5$ mm and $r_{\max} = 6.5$ mm in the analysis of measurement and simulation data. The RMS resolution is larger than the mean resolution. The difference in the results obtained with the two analysis methods are shown in Figures 6.46 to 6.48 for the three types of readout electronics.

For better comparison of measurement and simulation, the tube hit efficiency is used in this section instead of the 3σ -efficiency.

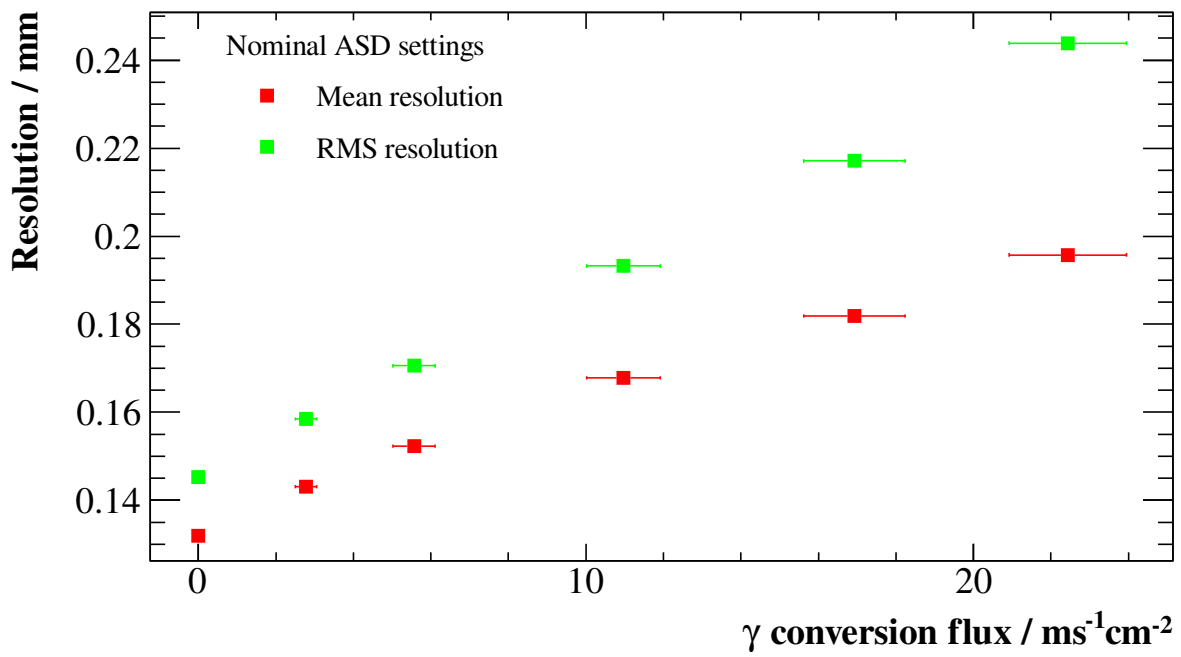


Figure 6.46: Comparison of averaging methods of the sMDT resolution over the tube radius as a function of the γ flux for ASD readout at nominal hysteresis and threshold settings. Mean resolution refers to Equation (3.4) in Section 3.2.5, RMS resolution to Equation (6.5) in Section 6.8. The difference is a measure of the systematic uncertainties in the average drift tube resolution.

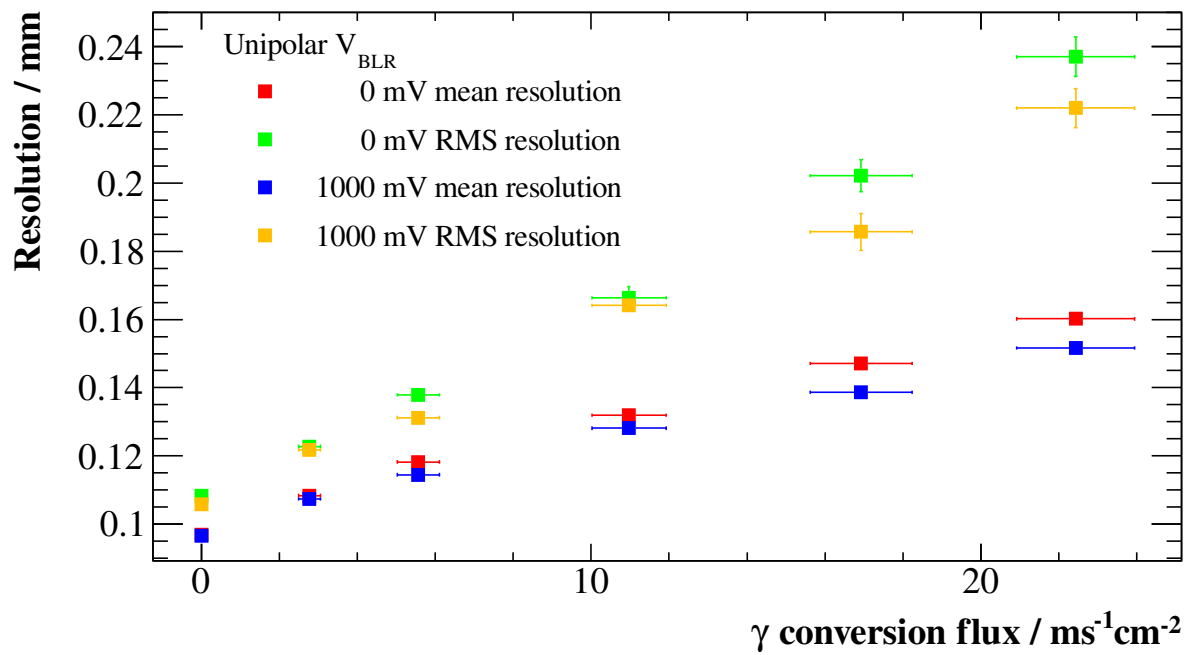


Figure 6.47: Comparison of averaging methods of the sMDT resolution over the tube radius as a function of the γ flux for discrete unipolar readout at two BLR voltages and nominal threshold settings. Mean resolution refers to Equation (3.4) in Section 3.2.5, RMS resolution to Equation (6.5) in Section 6.8. The difference is a measure of the systematic uncertainties in the average drift tube resolution.

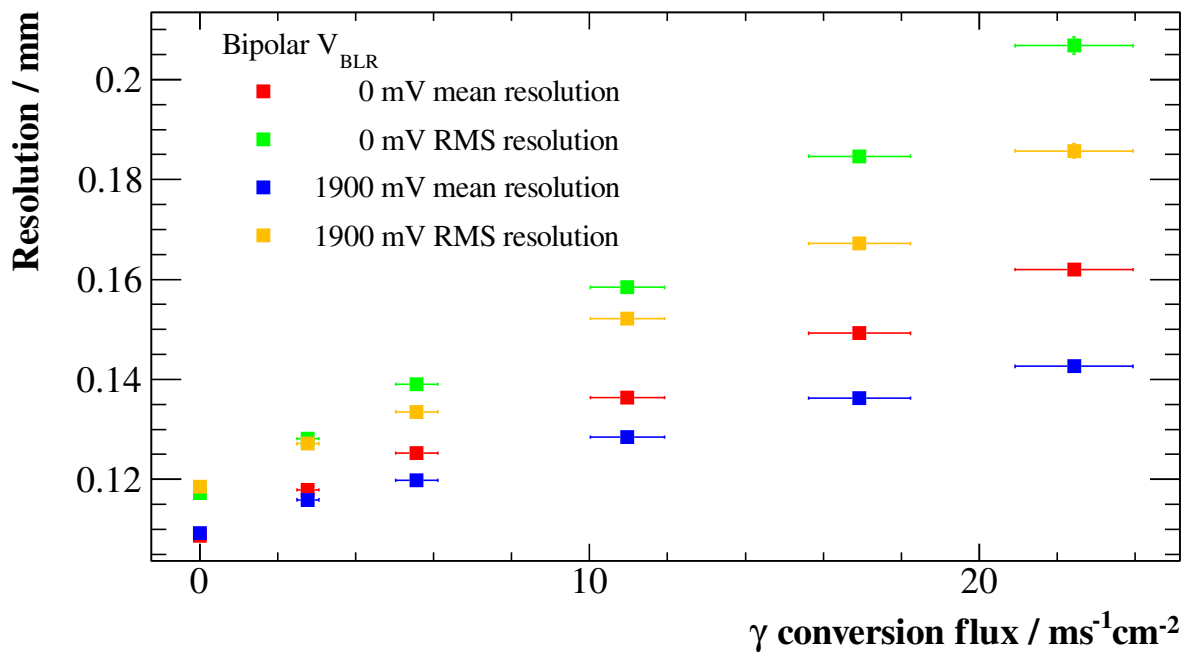


Figure 6.48: Comparison of averaging methods of the sMDT resolution over the tube radius as a function of the γ flux for discrete bipolar readout at two BLR voltages and nominal threshold settings. Mean resolution refers to Equation (3.4) in Section 3.2.5, RMS resolution to Equation (6.5) in Section 6.8. The difference is a measure of the systematic uncertainties in the average drift tube resolution.

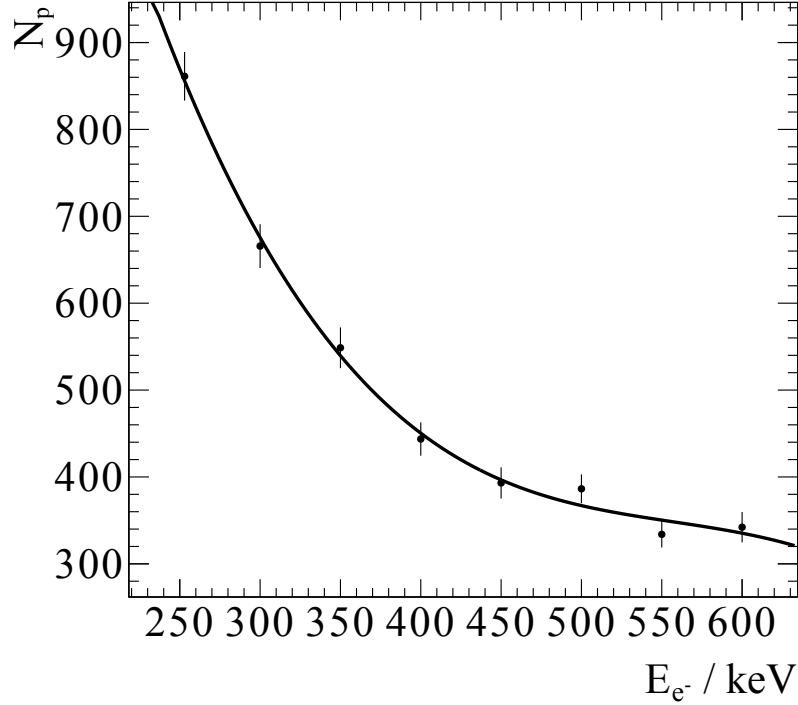


Figure 6.49: Dependence of the average number of primary ionization electrons N_p generated by Compton electrons in Ar/CO_2 (93/7) on the electron energy E_{e^-} from the Garfield++ simulation program [89]. The curve is a fourth order polynomial fitted to the simulated points in order to extrapolate to energies below 200 keV.

6.8.2 Drift Tube Pulse Simulation

For the simulation of the hits induced by the γ background radiation, the energy distribution of the Compton electrons created in the tube wall had to be determined. The electron energy distribution from Compton scattering of the 662 keV photons from the $^{137}_{55}\text{Cs}$ source at the GIF++ has been calculated using the Klein-Nishina formula [90]. The energy deposition of Compton electrons with energy E_{e^-} in Ar/CO_2 (93/7) gas at 300 kPa was simulated with the Garfield++ program [89] to obtain the number of primary ionization electrons N_p (cf. Figure 6.49).

Due to the low-energy cutoff in the Garfield++ simulation, N_p cannot be calculated for electron energies below 200 keV. The extrapolation of the Garfield++ results to energies below 200 keV was performed using a fit of a fourth order polynomial to the simulated data points.

The average number of primary ionization electrons $\langle N_p \rangle = \int_{E_{\min}}^{E_{\max}} N_p(E_{e^-}) f(E_{e^-}) dE_{e^-}$ generated in the energy interval $36 \text{ keV} < E_{e^-} < 662 \text{ keV}$ per Compton electron in the Ar/CO_2 (93/7) drift gas was

computed using the Klein-Nishina formula for the energy spectrum $f(E_{e^-})$ of the Compton electrons and the fit curve $N_p(E_{e^-})$ from Figure 6.49. The lower edge of the energy interval was tuned to match $N_p = 1350$ from Section 6.2.5.

The Garfield++ simulation package is further used to simulate the unamplified drift tube signals from the Compton electrons and from 150 GeV muons, the nominal muon energy in the test beam. Compton electron signals and muon signals are superimposed to the tube signals for γ background counting rates per tube of 0 ms^{-1} , 90 ms^{-1} , 180 ms^{-1} , 360 ms^{-1} , 560 ms^{-1} and 740 ms^{-1} . The raw signals are convolved with the δ response function of the readout electronics without BLR stage to obtain the shaped signals and to determine their threshold crossing times. The simulated data are analyzed in the same way described in Section 6.3 as the measurement data recorded in the GIF⁺⁺.

The shaped signal amplitudes are described in numbers of primary electrons by the simulation. The discriminator thresholds in the simulation corresponding to the threshold voltages of the readout electronics are determined iteratively for each readout circuit to match the measured resolution without γ background irradiation. With the determined discriminator threshold, the noise amplitude, which is added as flat white noise distortions to the simulated raw tube signals before the electronics simulation, is determined to match the measured noise count rate given in Table 6.1.

The gain loss due to space charge is always taken into account by scaling the pulse amplitudes with the measured gain drop. The ratio $\frac{G}{G_0}$ of the gas gain G relative to the nominal value G_0 without irradiation was determined from the GIF⁺⁺ muon beam with varying γ conversion fluxes using the pulse charge measurement of the ASD chip as described in Section 6.2.5.

Using a series of γ pulses instead of a flat baseline for a muon pulse simulates signal pileup. The probabilities of *gamma*-hits in 20 consecutive time intervals of 200 ns is calculated according to the given irradiation rate. The γ pulse shapes are randomly chosen from the recorded pulse library. For the simulation of the pileup effects, the 4 μs long pulse combination generated is processed by the Garfield++ electronics simulation.

In order to study the effect of baseline restoration, an idealized signal tail cancellation is added after the last active shaper stage for each readout electronics, adjusting the analog shaper output to the baseline 20 ns after the crossing of the baseline from the signal to the corresponding undershoot.

6.8.3 sMDT Performance with the ASD Chip

The δ response function is used in Garfield++ to describe the amplification and shaping response of the ASD ASIC. As the ASD analog signal of the ASD chip cannot be measured directly, the simulated δ response function [44] is used for the simulation. To take into account that the actual peaking time of the

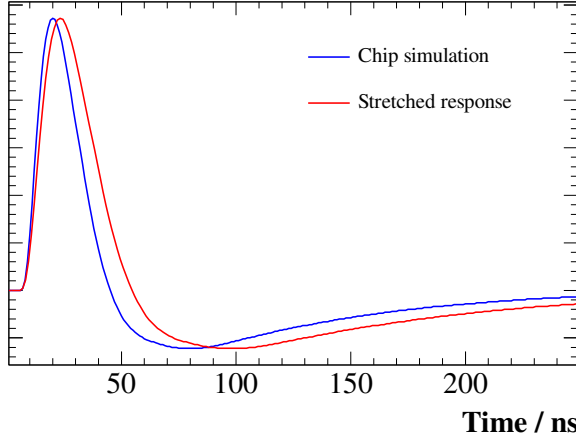


Figure 6.50: δ response function of the ASD chip. The nominal response function (blue) was stretched (red) to match the measured peaking time [44].

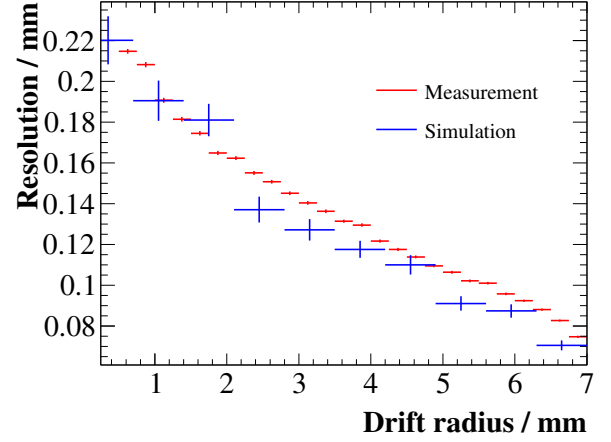


Figure 6.51: Simulated and measured dependence of the spatial resolution on the drift radius for ASD chip readout without γ irradiation.

ASIC chip is 15 ns [41] instead of 12 ns of the δ response function, the simulated δ response was stretched as shown in Figure 6.50.

To match the measured average RMS drift tube resolution without irradiation, $\langle\sigma_{meas}\rangle = 142.0 \pm 0.3 \mu\text{m}$, the discriminator threshold in the simulation was set to an equivalent of 24.4 primary ionization electrons, which gives $\langle\sigma_{sim}\rangle = 138 \pm 3 \mu\text{m}$. The RMS noise level, which reproduces the accidental hit rate in the measurement setup corresponds to 6.2 primary ionization electrons.

Figure 6.51 shows the measured and simulated sMDT resolution as a function of the drift radius without γ irradiation, which are in good agreement.

The charge measurement of the MDT ASD chip was compared to the simulated pulse height distribution in the case of no irradiation. The ADC charge measurement has to be calibrated. For this purpose, test pulses of constant shape but varying amplitude were differentiated with a 1.8 pF capacitor and injected into an ASD chip with the same configuration as used in the muon beam measurement (cf. Table 4.1). The width w_{ADC} of the LVDS output pulse encoding the charge measurement was recorded with an oscilloscope. A function

$$h(w_{\text{ADC}}) = p_0 + p_1 w_{\text{ADC}} + p_2 \exp(p_3 w_{\text{ADC}}),$$

with free parameters $p_0 \cdots p_3$, was fitted to the charge measurements as shown in Figure 6.52 and used to correct the signal amplitude depending on the measured ASD output signal width. The dependence on the drift radius of the rising edge charge within 18.5 ns of the shaped signal corresponding to the ASD output

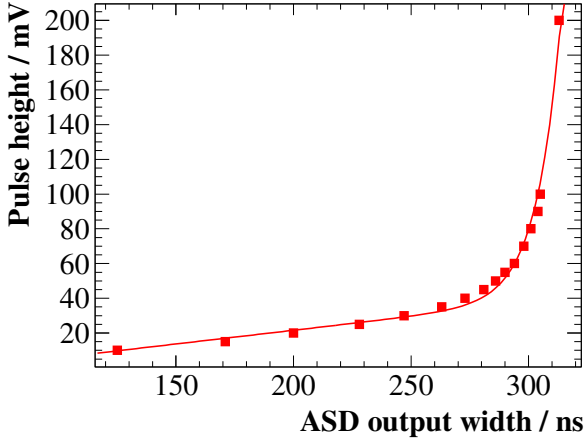


Figure 6.52: Calibration of the charge measurement of the ASD chip for 18.5 ns integration gate length. The input pulses were differentiated with a 1.8 pF capacitor.

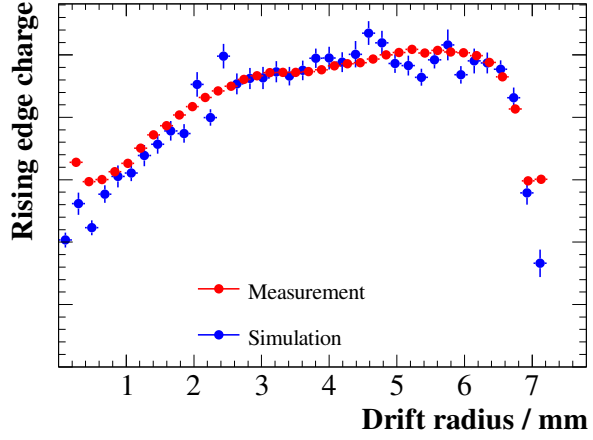


Figure 6.53: Dependence of the pulse charge measurement height on the drift radius for measured data and for the Garfield++ simulation without γ irradiation.

width is shown in Figure 6.53 in comparison with the Garfield++ simulation results, which agrees well with the measurements.

When operated in the time-over-threshold (TOT) mode (cf. Section 4.1), the width of the ASD chip output encodes the signal pulse width. Figure 6.54 shows a comparison of the measured and the simulated TOT distribution. While the simulated spectrum is rather broad with a mean of $0.13 \mu\text{s}$, the measured spectrum is narrower with a tail towards large values and a mean of $0.11 \mu\text{s}$. This discrepancy arises from imperfections in the simulation of the pulse shape. Deficiencies in the description of the pulse tail have already been observed in a previous simulation based on the FORTRAN Garfield version [15]. As the tube efficiency strongly depends on the effective deadtime determined by the pulse width with the falling edge, the simulation is not expected to describe the tube efficiency reliably. The resolution, on the other hand, mainly depends on the signal rising edge and is hence more reliable.

The simulation dependence of the RMS resolution on the γ conversion flux including gain drop and signal pileup effects agrees well with the measurements as shown in Figure 6.55.

Figure 6.56 shows the effects of gain drop, signal pileup and baseline restoration on the average resolution (RMS) as a function of the γ conversion flux with the MDT ASD chip readout. The resolution deterioration at high γ conversion fluxes is mainly caused by signal pileup even in only 0.25 m long drift tubes. Neglecting the pileup contribution by making the tube length infinitely small in the simulation, the resolution deterioration is only due to gain drop and much smaller. The resolution deterioration caused by signal pileup (cf. Section 4.3.1) can be suppressed by ideal baseline restoration, which has also been

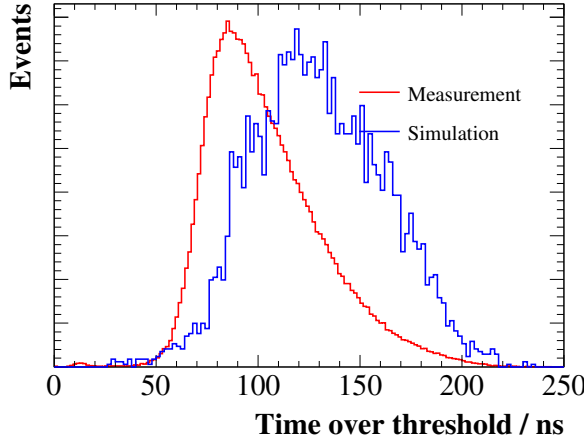


Figure 6.54: Simulated and measured time-over-threshold spectra of γ hits recorded with the ASD chip readout γ irradiation normalized to the same number of events.

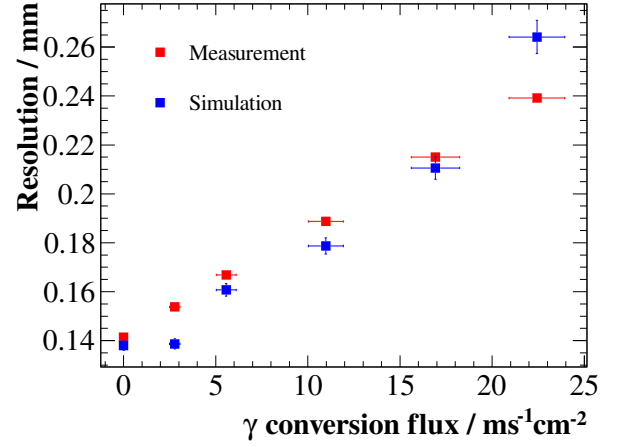


Figure 6.55: Simulated and measured dependence of the RMS resolution on the γ conversion flux for ASD readout and 0.25 m tube length.

simulated for MDT ASD chip shaping and confirms the results of previous simulation studies in [15]. Pileup suppression becomes even more important for longer tubes.

Figure 6.57 depicts the tube hit efficiency as a function of the γ conversion rate. Neglecting deadtime and signal pileup effects in the simulation, efficiency loss caused by gain drop is negligible even at the highest irradiation despite high space charge at given γ hit rate in the only 0.25 m long tubes. Adding the effect of deadtime of 200 ns and signal pileup in the simulation significantly reduces the efficiency at high rates. As mentioned in Section 6.2.2, ASD chips with 300 ns minimum deadtime were used in the GIF⁺⁺ muon beam measurement, leading to a larger efficiency drop at high background rates than in the simulation. The actual ASD deadtime depends on the output pulse width determined by the charge measurement, which is different for each pulse and channel and can hence not be simulated reliably. The efficiency gain by BLR is negligible, as signals piling up on a undershoot still cross the discriminator threshold.

6.8.4 sMDT Performance with the Discrete Unipolar Shaper

The transfer function of the discrete unipolar shaper electronics (cf. Section 4.3.3) has been measured (cf. Section 5.1.2.1) and simulated using the **Simulation Program with Integrated Circuit Emphasis** (SPICE) program [71]. The two response functions are shown in Figure 6.58. The signal peaking time of the unipolar shaper is 5 ns, one third of the peaking time of the MDT ASD chip. The rising edge is very

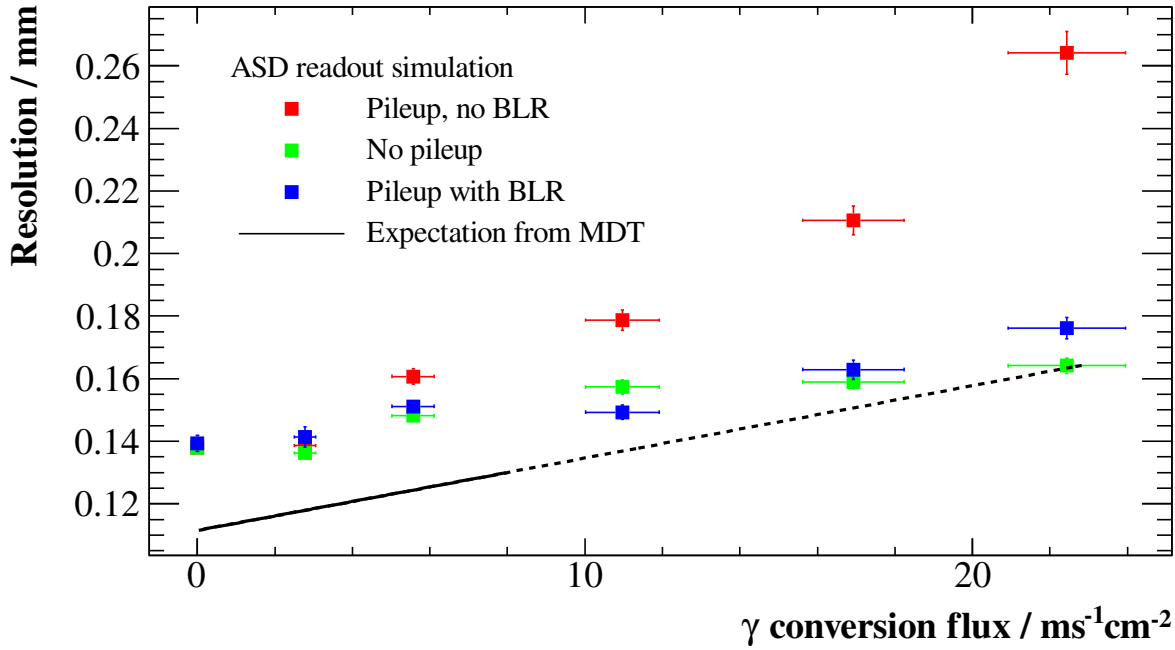


Figure 6.56: Simulated dependence of the average resolution (RMS) on the γ conversion flux for ASD chip readout with and without signal pileup and ideal BLR. Pileup is simulated for 0.25 m tube length and “No Pileup” corresponds to infinitely short tubes. The remaining resolution deterioration at high rates without pileup is due to gain drop and relatively small. The black line shows an expectation using the average mean resolution of 30 mm diameter drift tubes in the drift radius range $r < 7.1$ mm and scaling the flux by a factor of ten to account for the one order of magnitude higher rate capability of 15 mm diameter drift tubes in terms of gain drop [15]. It is the optimum sMDT resolution which can be expected. The dashed line shows a linear extrapolation to higher rates. The deviation to the simulation arises from time walk corrections and lower threshold applied to the MDT measurement data on which the expectation is based.

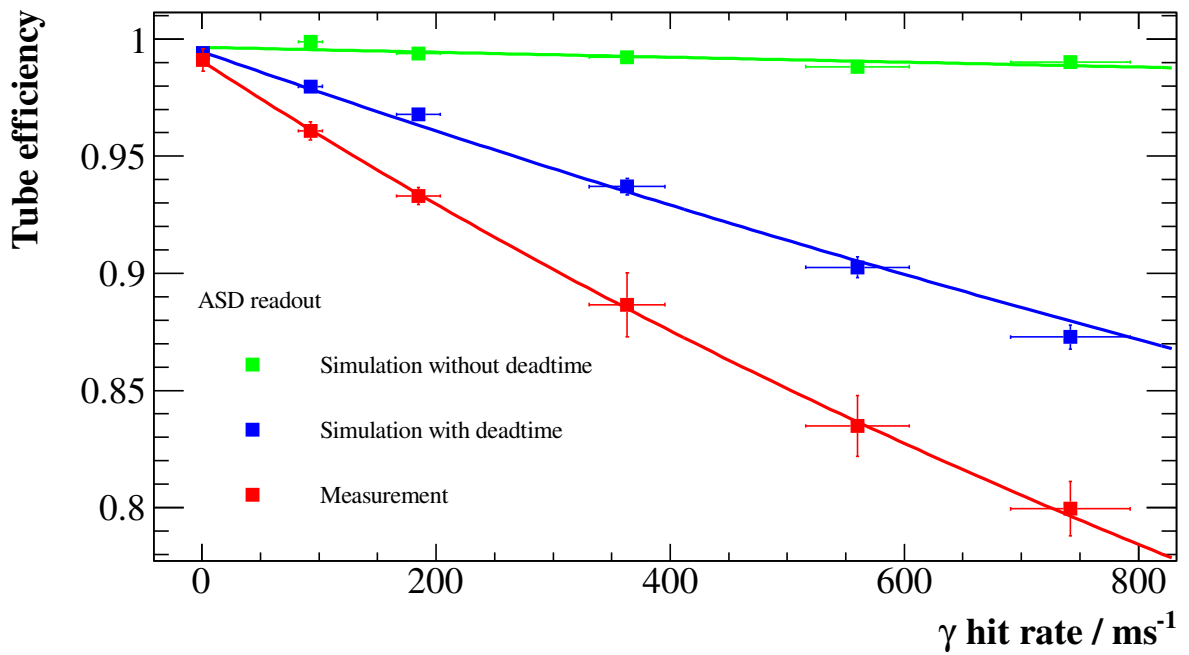


Figure 6.57: Simulated and measured dependence of the tube hit efficiency on the γ hit rate for MDT ASD chip readout with minimum deadtime of 200 ns and 0.25 m tube length with fits to the data points. The simulation without deadtime simulates only gain drop but no pileup and deadtime effects.

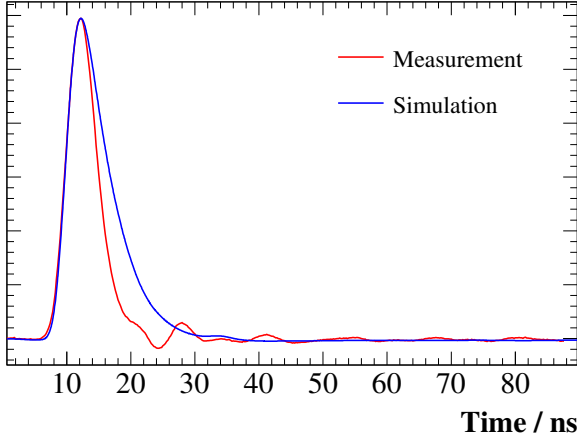


Figure 6.58: Simulated and measured δ response function of the unipolar shaper circuit without BLR stage.

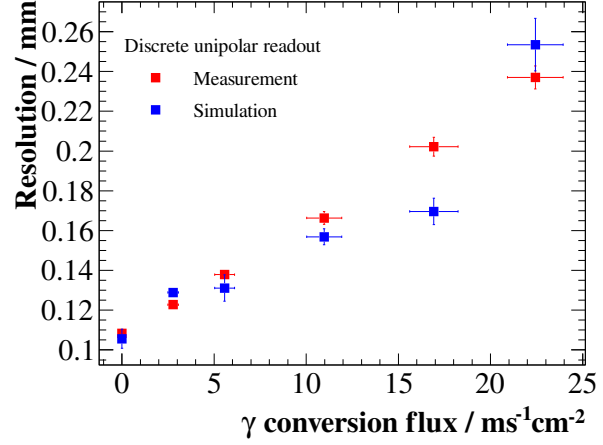


Figure 6.59: Simulated and measured dependence of the average resolution (RMS) on the γ conversion flux for the discrete unipolar shaper with $V_{\text{BLR}} = 0$ mV and 0.25 m tube length.

well described by the simulation. The tail cancellation of the unipolar shaper is underestimated in the simulation. The well defined SPICE simulated response curve was chosen for the performance simulation.

The threshold for the simulation was determined in the same way as for the ASD chip to be 43 primary ionization electrons. Although almost twice as high as for the ASD chip, the average tube resolution achieved with the discrete unipolar shaper readout is better because of the considerably smaller signal peaking time. The simulated dependence of the resolution on the γ conversion flux agrees well with the measurement (cf. Figure 6.59). The noise width for the simulation was determined in the same way as for the ASD chip to be 11 primary ionization electrons, slightly lower than the measured noise level (cf. Table 6.1), because the BLR stage, which is not included in the simulation, reduces the signal-to-noise ratio.

Figure 6.60 shows the effects of signal pileup and baseline restoration on the average resolution (RMS) as a function of the γ conversion flux with the discrete unipolar readout electronics. Like for the MDT ASD chip readout without BLR, the resolution deterioration at high γ conversion fluxes is mainly caused by signal pileup, while the effect of gain drop is relatively small. The idealized simulated BLR comes close to the simulation without signal pileup, which is simulated by infinitely small tube length. Figure 6.61, however, shows that the effect of the BLR stage of the unipolar shaper does not reach the ideal case as can be explained from the discussion in Section 6.5.1 (Figure 6.30).

Figure 6.62 shows the tube hit efficiency as a function of the γ conversion rate for the discrete unipolar readout on the γ conversion rate. Like for the ASD readout, the effect of gain drop due to space charge is

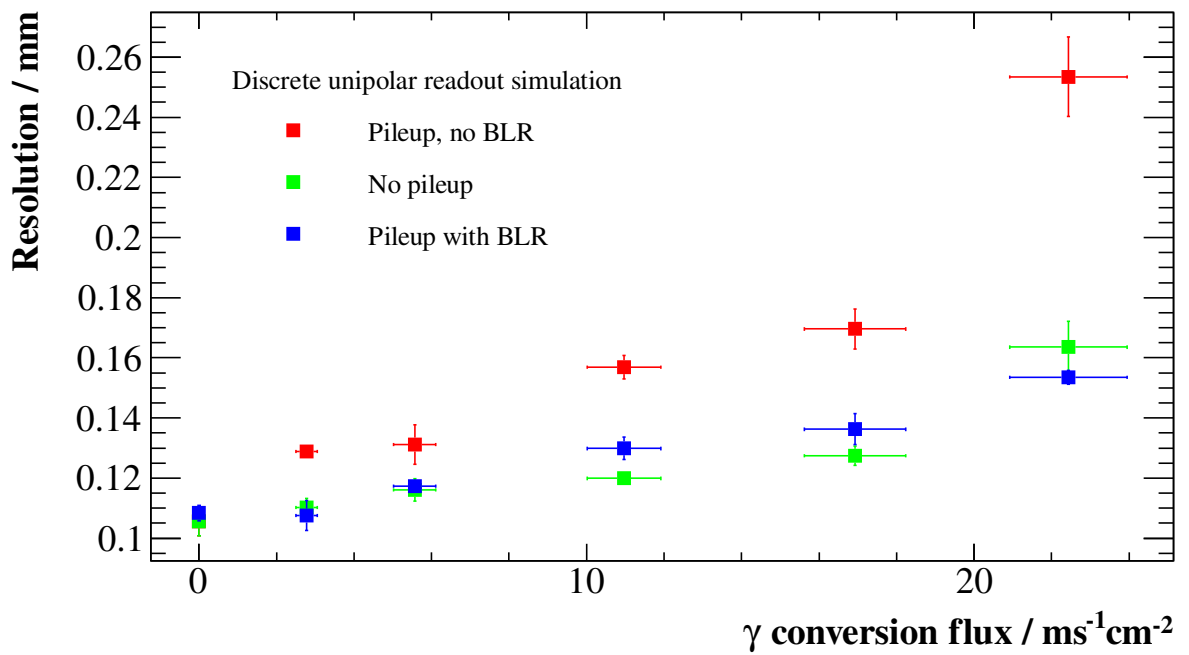


Figure 6.60: Simulated dependence of the average resolution (RMS) on the γ conversion flux for the discrete unipolar readout on the γ conversion flux with and without signal pileup and ideal BLR. Pileup is simulated for 0.25 m tube length and “No Pileup” corresponds to infinitely short tubes. The remaining resolution deterioration at high rates without pileup is due to gain drop and relatively small.

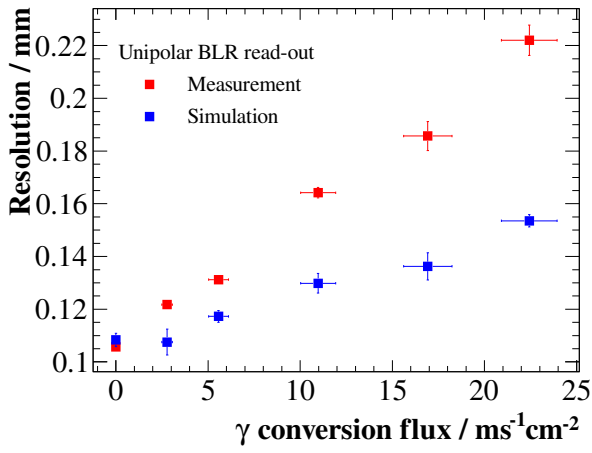


Figure 6.61: Simulated and measured dependence of the average resolution (RMS) on the γ conversion flux for the discrete unipolar shaper with $V_{\text{BLR}} = 1000$ mV and 0.25 m tube length.

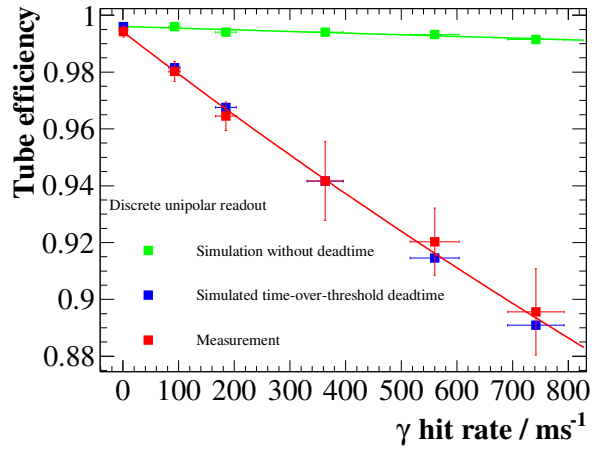


Figure 6.62: Simulated and measured dependence of the tube hit efficiency on the γ hit rate for discrete unipolar readout and 0.25 m tube length. The simulation without deadtime simulates only gain drop but no pileup and deadtime effects. The simulated deadtime is calculated using 159 ns average time-over-threshold as deadtime. The red and green curves are fitted to the data points.

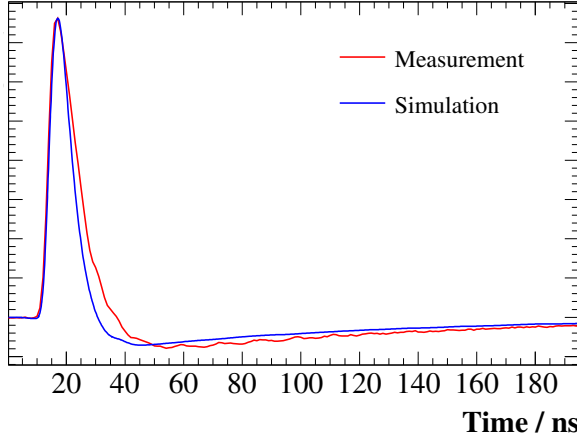


Figure 6.63: Comparison of the simulated and measured δ response shape of the last active stage of the bipolar shaper.

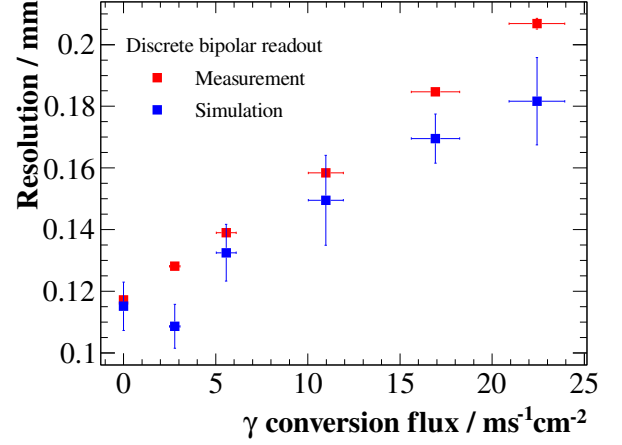


Figure 6.64: Simulated and measured dependence of the average resolution (RMS) on the γ conversion flux for the discrete bipolar shaper with $V_{BLR} = 0$ mV and 0.25 m tube length.

negligible (green curve). The deadtime estimation from the Garfield++ simulation time-over-threshold was used for the calculation of the efficiency. Despite imperfections of the simulation regarding the pulse shape and the underestimation of the signal tail cancellation of the unipolar shaper, the measured and predicted tube hit efficiency dependencies on the γ hit rate agree very well as do the corresponding average time-over-threshold or deadtime values of 159 ns and 150 ± 20 ns, respectively. In addition, the time-over-threshold values show that the signal reliably drops below the discriminator threshold after a pulse, proving sufficient signal tail cancellation.

6.8.5 sMDT Performance with the Discrete Bipolar Shaper

As for the unipolar shaper, the transfer function of the discrete bipolar shaper has been measured (cf. Section 5.1.2.2) and simulated with SPICE. A comparison of the measured and simulated δ -responses is shown in Figure 6.63. Like for the unipolar shaper, the rising edges of simulated and the measured response functions agree very well with a signal peaking time of 5 ns. The tail cancellation of the bipolar shaper is slightly overestimated by the simulation. The SPICE simulated δ response is used for the Garfield++ simulation like for the unipolar shaper.

The threshold for the simulation was determined to be 39 primary ionization electrons, which is very similar to the value for the unipolar readout. The RMS noise width, which reproduces the accidental hit rate in the measurement, corresponds to 8 primary ionization electrons for the bipolar shaper. To the

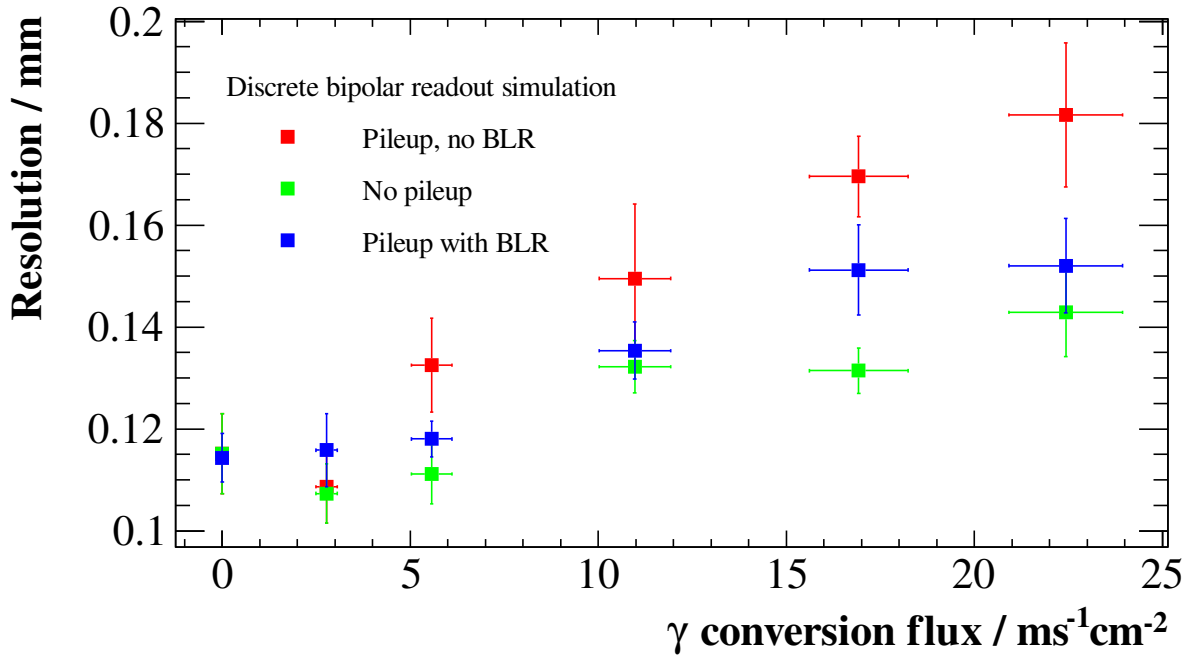


Figure 6.65: Simulated dependence of the average resolution (RMS) on the γ conversion flux for the discrete bipolar readout on the γ conversion flux with and without signal pileup and ideal BLR. Pileup is simulated for 0.25 m tube length and “No Pileup” corresponds to infinitely short tubes. The remaining resolution deterioration at high rates without pileup is due to gain drop and relatively small.

same extend as the BLR stage of the unipolar readout, the BLR stage of the bipolar shaper reduces the signal amplitude and the signal-to-noise ratio. Without γ irradiation, the average spatial resolutions of the unipolar and bipolar shapers are very similar due to the equal signal peaking times. Figure 6.64 shows the average spatial resolutions (RMS) of the bipolar shaper as a function of the γ conversion flux from measurement and simulation. The slightly worse simulated resolution is due to the worse tail cancellation. In measurement and simulation without BLR, the bipolar shaper provides faster tail cancellation and, consequently, better resolution at high rates than the unipolar shaper.

The effects of signal pileup and baseline restoration on the average resolution (RMS) for the discrete bipolar readout electronics are shown in Figure 6.65. Due to the equally short signal rising time, the effect of gain drop on the resolution at high rates (green points) is similar small as for the unipolar shaper readout. With the idealized BLR in the Garfield++ simulation, pileup effects deteriorate the resolution by the same amount as for the unipolar shaper and the MDT ASD chip readout.

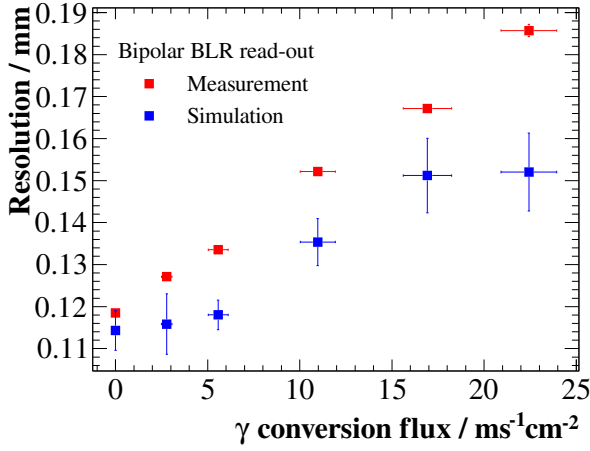


Figure 6.66: Simulated and measured dependence of the average resolution (RMS) on the γ conversion flux for the discrete bipolar shaper with $V_{\text{BLR}} = 1900$ mV and 0.25 m tube length.

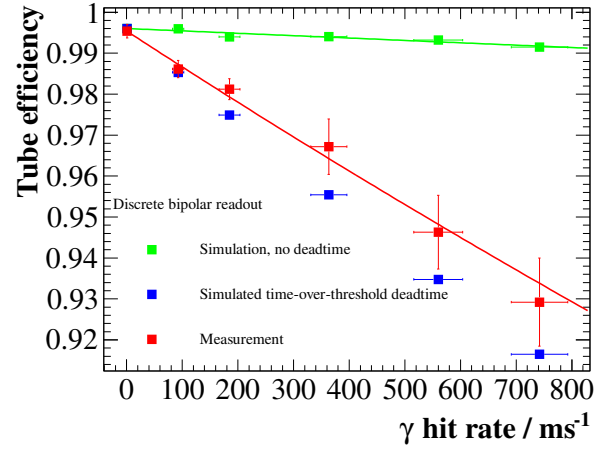


Figure 6.67: Simulated and measured dependence of the tube hit efficiency on the γ hit rate for discrete bipolar readout and 0.25 m tube length. The simulation without deadtime simulates only gain drop but no pileup and deadtime effects. The simulated deadtime is calculated using 117 ns average time-over-threshold as deadtime. The red and green curves are fitted to the data points.

The improvement of the measured resolution at high γ hit rates with $V_{\text{BLR}} = 1900$ mV does not reach the ideal BLR case (cf. Figure 6.66) as expected from Section 6.6.1 (Figure 6.37), but is more efficient than for the unipolar case.

Figure 6.67 shows the dependence of the tube hit efficiency on the γ conversion rate for the discrete bipolar shaper readout. Like for the unipolar readout, the effect of gain drop due to space charge on the efficiency is negligible (green curve). Due to a slightly lower threshold, the simulation overestimates the average time-over-threshold compared to the measurement with 117 ns and 89 ± 11 ns, respectively. Consequently, the simulation underestimates the efficiency but is higher than for the unipolar shaper readout.

6.9 Summary

The achieved sMDT performance for the different readout electronics are summarized in Table 6.2.

The MDT ASD chip has a signal peaking time of 15 ns compared to 5 ns of the discrete unipolar and bipolar shaper. The required larger bandwidth increases the RMS noise width from 6 primary electrons to 8 for the bipolar shaper. The unipolar shaper noise width is 11 primary electrons because of additional

Table 6.2: sMDT performance comparison for the different readout electronics at the nominal operating parameters summarized in Table 6.1. The resolutions refer to the measured mean resolution according to Equation (3.4) shown in Figure 6.44. The 3σ -efficiencies $\epsilon_{3\sigma}$ and effective deadtimes are measurement results shown in Figure 6.45. The minimum deadtime is the minimum time between two discriminator threshold crossings recorded by the TDC.

Parameter	Unit	MDT ASD chip	Discrete unipolar shaper	Discrete bipolar shaper
Signal peaking time	ns	15	5	5
Noise RMS	primary e^-	6	11	8
Threshold	primary e^-	24	43	39
Resolution without irradiation	mm	0.132	0.097	0.109
Resolution at $22 \text{ ms}^{-1} \text{ cm}^{-2}$	mm	0.196	0.152	0.143
$\epsilon_{3\sigma}$ without irradiation	%	93	93	93
$\epsilon_{3\sigma}$ at 740 ms^{-1}	%	67	79	80
Average time-over-threshold (simulated)	ns	130	159	117
Minimum deadtime (measured)	ns	300	0	0
Effective deadtime (measured)	ns	500	260	240

baseline distortions caused by missing voltage buffers (cf. Figure 5.46 in Section 5.1.3). The ASD discriminator threshold corresponds to 24 primary electrons. The larger noise width requires increased discriminator thresholds of 43 and 39 primary electrons for the unipolar and bipolar shaper, respectively. The average sMDT spatial resolution for MDT ASD chip readout is 0.132 mm without irradiation and deteriorates to 0.196 mm due to gain drop and space charge fluctuations at the highest achieved γ conversion flux of $22 \text{ ms}^{-1} \text{ cm}^{-2}$. In spite of the higher thresholds, the resolution without irradiation for the discrete unipolar and bipolar readout electronics are 0.097 mm and 0.109 mm, respectively. This resolution improvement is due to the faster signal peaking time. At $22 \text{ ms}^{-1} \text{ cm}^{-2}$, the resolution for the unipolar shaper is 0.152 mm and even only 0.143 mm for the bipolar shaper. While the unipolar shaper readout achieves the best resolution without irradiation, the faster signal tail cancellation and baseline stability of the bipolar shaper achieves the best resolution at the highest γ irradiation rates.

The muon efficiency $\epsilon_{3\sigma}$ without irradiation is 93 % independent of the readout electronics. At high γ irradiation rates, the efficiency is reduced due to the electronics deadtime. For the ASD chip in the ADC mode, the deadtime is determined by the charge measurement encoded into the discriminator output width of at least 200 ns. For the discrete unipolar and bipolar shaper readout, the deadtime is determined by the time-over-threshold (TOT). In the simulation the deadtime is 130 ns for the ASD chip, 159 ns for the discrete unipolar shaper and 117 ns for the discrete bipolar shaper readout. The effective deadtime takes

into account also multiple threshold crossings and baseline distortions. In the measurement, the effective deadtime was 500 ns for the ASD chip readout in ADC mode, determined by the charge measurement. The discrete unipolar and bipolar shaper electronics use the time-over-threshold mode with effective deadtimes of 260 ns and 240 ns, respectively, significantly improving the efficiency at high γ irradiation rates.

Chapter 7

Performance of Drift Tube Detectors with the new MDT ASD ASIC

To keep the ATLAS muon trigger rate constant after the luminosity upgrade of the LHC, the selectivity of the trigger system has to be increased [91]. This can be achieved by using the high spatial resolution of the MDT chambers at the first level trigger [92]. This requires triggerless readout electronics and fast track reconstruction in the MDT chambers. Prototype MDT barrel trigger processors have been successfully tested in the GIF⁺⁺ [93, 94]. The mezzanine boards of all MDT chambers in the ATLAS muon spectrometer have to be replaced for operation at HL-LHC. Therefore, a new version of the ASD chip has been developed. Prototype chips have been tested on MDT and sMDT chambers in the GIF⁺⁺.

Table 7.1 summarizes the main performance parameters of both MDT ASD chip generations. The signal peaking time is 12 ns compared to 14.4 ns of the legacy MDT ASD, reducing the deteriorating effect of time walk on the resolution (cf. Section 6.3.4). The signal amplification is increased from $10 \frac{\text{mV}}{\text{fC}}$ to $21 \frac{\text{mV}}{\text{fC}}$ in the new chip is, improving the efficiency at low gas gains. Smaller noise and threshold spread between channels of the new MDT ASD chip allow for operation lower thresholds.

Table 7.1: Main performance parameters of the legacy and new MDT ASD chips [95].

Parameter	Unit	Specification	Legacy ASD	New ASD
Signal peaking time	ns	15	14.4	12
Amplification	$\frac{\text{mV}}{\text{fC}}$	8.9	10	21
Noise RMS	mV	8.5	8	4
Threshold spread	mV		12...16	4
Power consumption	mW		300	360

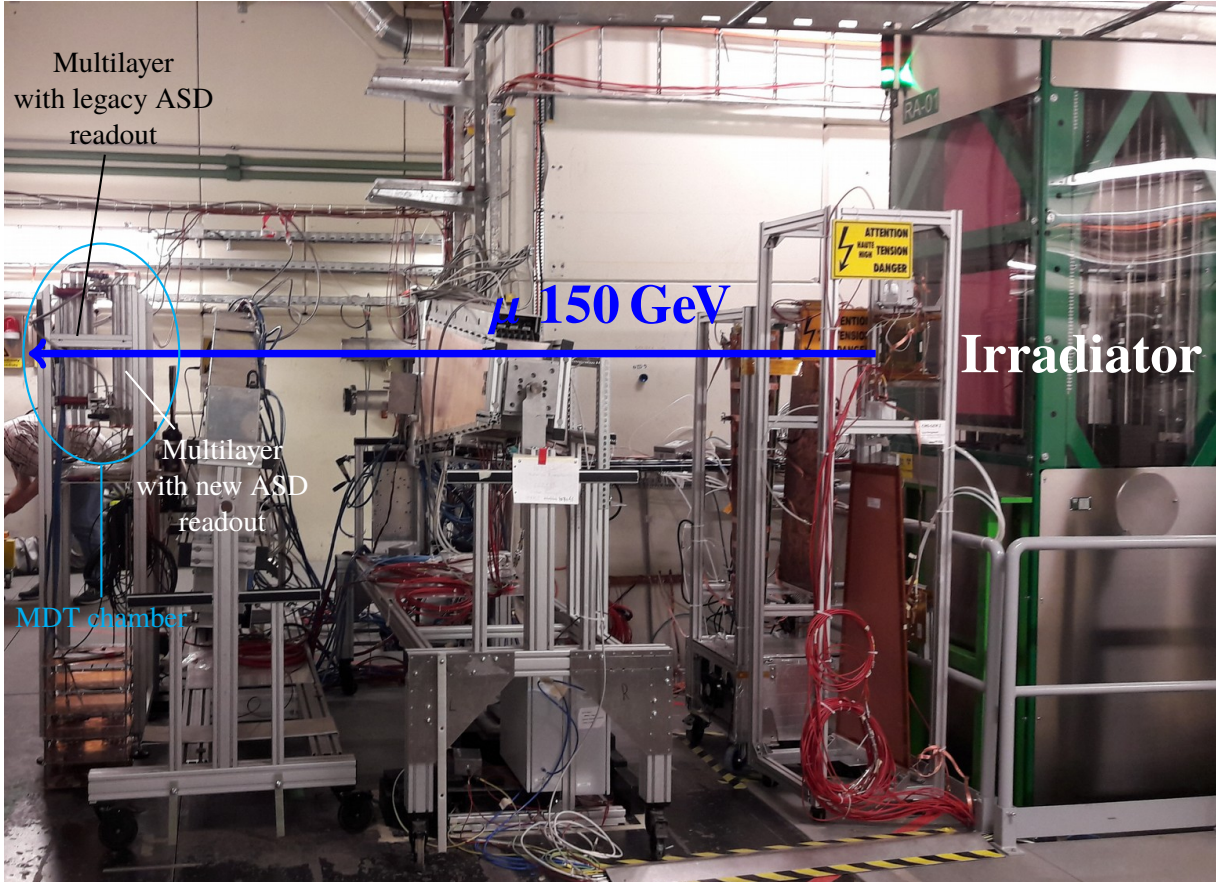


Figure 7.1: MDT chamber setup in the new **Gamma Irradiation Facility** at CERN viewed vertically to the beam axis indicated by the blue arrow.

7.1 Experimental Setup

The performance of the new ASD MDT chip was studied on a MDT chamber with 30 mm diameter drift tubes and on a sMDT chamber with 15 mm diameter drift tubes in the *GIF⁺⁺*. The setup is analogous to the one described in Section 6.1. Both chambers consisted of 48 tubes, half of which were read out with two 24-channel mezzanine boards, one with three legacy and the other with three new ASD chips.

The sMDT test chamber consisted of 48 tubes with a length of 0.95 mm in eight layers. It has been built in 2011 for a different test program [15] and could provide only 22 working tubes for the legacy ASD and 15 for the new ASD. The setup is equivalent to the one sketched in Figure 6.2.

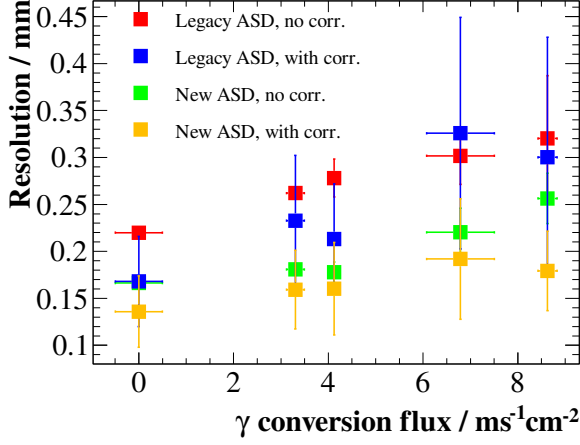


Figure 7.2: Comparison of the average resolutions of 950 mm long 15 mm diameter sMDT with the legacy and new ASD readout electronics with and without time walk corrections. The error bars include the systematic uncertainties.

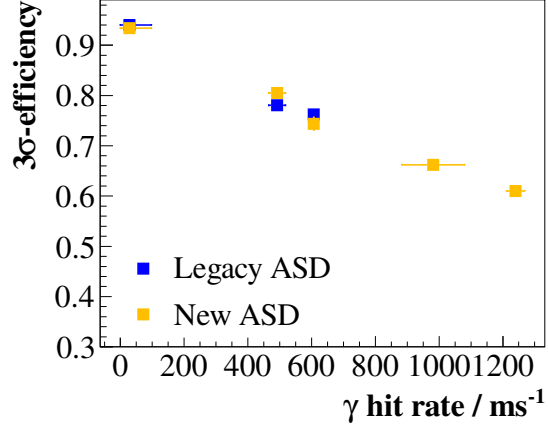


Figure 7.3: Comparison of the average 3σ -efficiency of 950 mm long 15 mm diameter drift tubes with the legacy and the new ASD chip readout.

The MDT test chamber consisted of two separated multilayers read out by one mezzanine board each described. Each multilayers consists of three layers of eight 0.5 m long drift tubes. A photograph of the setup inside the *GIF*⁺⁺ bunker is shown in Figure 7.1. More details are described in [96].

The chambers were operated at nominal operation conditions with Ar/CO₂ (93/7) gas mixture at 300 kPa and 3 080 V and 2 730 V, respectively (cf. Table 2.1). Both generations of the MDT ASD chips were operated in the ADC mode with minimum deadtime and the nominal ATLAS threshold of -39 mV and hysteresis of 8.75 mV.

7.2 sMDT Chamber

Figure 7.2 shows the dependence of the mean resolution on the γ conversion flux. Due to oxidation of electrical connections and repeated mechanical stress in several test campaigns (see [15, 53, 65, 93]) the noise level was significantly higher than e.g. in the test setup for the discrete electronics, leading to signal loss and deteriorated resolution. The direct comparison of both ASD versions, however, clearly show the beneficial effects of increased electronic gain and signal rise time also in a high noise environment. The improvement of the average resolution by time walk corrections is typically 20 μm with the legacy ASD chip (cf. Figure 6.23 and [15]) and also the new ASD chip.

Figure 7.3 shows a comparison of the dependence of the average 3σ -efficiency on the γ hit rate for the two ASD chip versions. For hit rates below 700 ms^{-1} , the measurements with the legacy and with the new chip overlap and agree with each other and with the results in Section 6.4.2 within errors. For higher γ hit rates, the maximum data rate of the mezzanine card with the legacy ASD chip was exceeded, leading to data loss (compare Section 4.2.1). As the mezzanine card with the new ASD had to read out fewer tubes, it could cope with the higher γ hit rates without data loss.

7.3 MDT Chamber

The electronic count rate is determined from the number of hits in the short time interval in front of the muon drift time spectrum as described in Section 6.2.2 and has to be corrected for deadtime according to Equation (4.3). Due to the maximum drift time of $0.7 \mu\text{s}$ in 30 mm diameter drift tubes, one γ conversion can cause multiple threshold crossings. Hence, the average γ hit multiplicity m is required for the determination of the γ conversion rate. For this purpose the hits in tubes which are not traversed by a muon track are selected. The first threshold crossing of a tube event must be within a 50 ns long time interval in the middle of the readout window to ensure that only γ hits are considered, which are completely within the readout window. The number of hits after the 50 ns time interval is then used to determine the average γ hit multiplicity m . It is 1.4 for both MDT ASD chip versions.

The average primary charge produced by a γ hit and the resulting Townsend avalanche with negligible space charge and gas gain G_0 is determined using the method described in Section 6.2.5. It is $(5.68 \pm 0.3) \text{ pC}$ and about 40 % larger than in sMDT. The dependence of the relative gas gain, the ratio $\frac{G}{G_0}$ of the gas gain G at a given irradiation rate ν and the gain G_0 without background irradiation in a MDT is shown in Figure 7.4. As described in Section 6.2.5, it is determined from the signal charge measurements in the ADC mode of the ASD chip. Using the average primary charge, the relative gas gain can also be determined from the high voltage tube current $I(\nu)$

$$\frac{G}{G_0} = \frac{I(\nu)}{5.68 \text{ pC} \cdot \nu}.$$

The results of both methods are in agreement. While previous measurements only covered MDT gain drops to about 80 % at a γ conversion flux below $1000 \text{ s}^{-1} \text{ cm}^{-2}$, the presented studies from data recorded in the GIF⁺⁺ at unprecedented γ conversion flux of $3000 \text{ s}^{-1} \text{ cm}^{-2}$ show the resolution and efficiency down to 60 % of the nominal MDT gas gain.

A comparison of the resolutions as a function of the γ conversion flux with the two readout chips is shown in Figure 7.5. Without γ irradiation, the resolution of the 30 mm diameter drift tubes is better than the sMDT resolution because of the larger maximum drift radius, but deteriorates much faster with

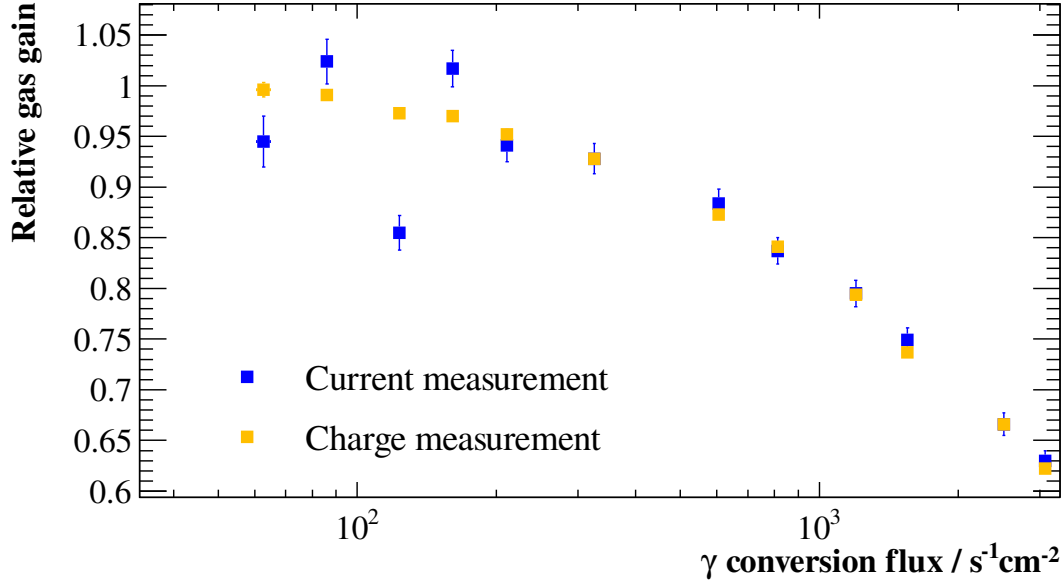


Figure 7.4: Relative MDT gas gain at nominal operating conditions. The gain drop can be determined from the high voltage tube current and the pulse charge measurement of the ASD chip in ADC mode. Nominal threshold of -39 mV was used in both cases.

increasing γ conversion flux. At small γ conversion flux, the new ASD chips lead to about $10 \mu m$ better resolution due to the shorter signal rise time and larger preamplifier gain. The γ conversion fluxes measured in the multilayers with different chip versions differ due to different distance to the γ source. At low γ irradiations, the resolution deterioration due to gain drop and space charge fluctuations increases linearly with the γ conversion flux. At about 75 % of the nominal gain, the resolution deteriorates significantly stronger at a γ conversion flux of about $1000 s^{-1}cm^{-2}$.

Figure 7.6 shows the drift tube efficiency (see definition in Section 6.3.3) of the 30 mm diameter drift tubes with the two chip versions. Up to a γ conversion flux of $1000 s^{-1}cm^{-2}$, corresponding to a tube hit rate of $150 ms^{-1}$, the efficiency loss is caused by signal pileup and is the same for both ASD chip versions. At higher irradiation rates, the gain drop caused by space charge additionally suppresses signals with small amplitudes. The larger gain of the new ASD chip reduces this loss, leading to larger efficiency than with the old ASD chip at γ conversion fluxes above $1000 s^{-1}cm^{-2}$. At the highest rates, the hit losses are again dominated by signal pileup. Acceptable efficiency above 95 % for 30 mm diameter drift tubes up to γ conversion fluxes of $1500 s^{-1}cm^{-2}$ corresponding to a gas gain of about $1.6 \cdot 10^4$. Only at even higher rates and lower gas gains the MDT efficiency drops strongly for both ASD chip versions.

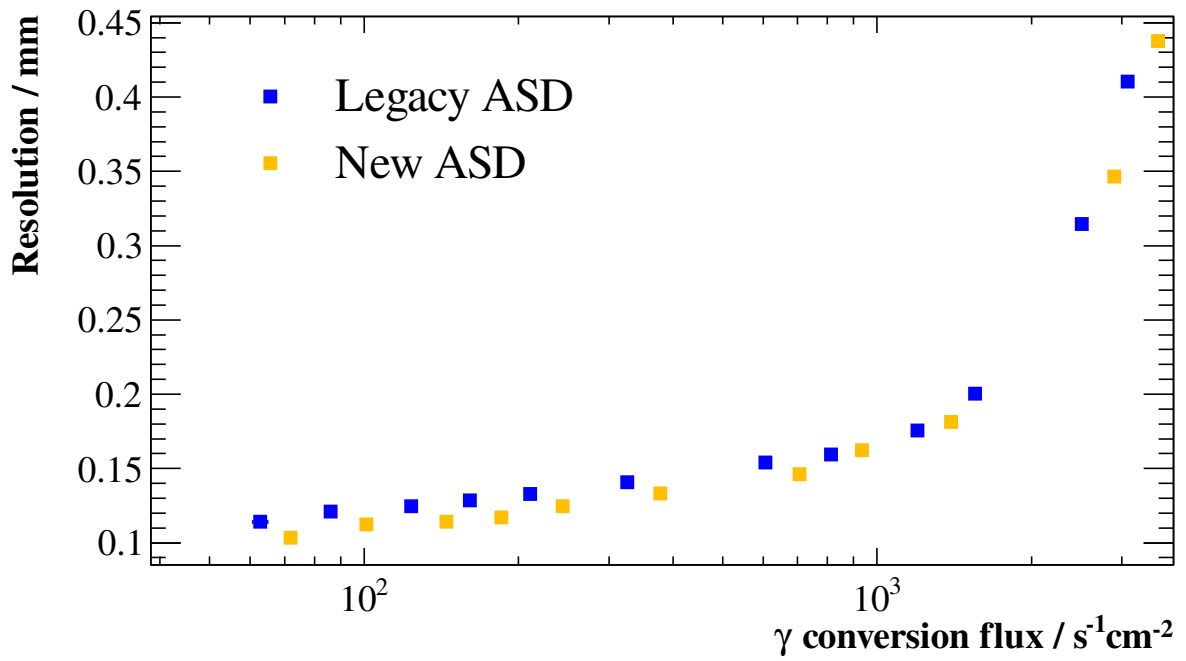


Figure 7.5: Comparison of the average resolutions of 0.5 m long 30 mm diameter drift tubes with the legacy and the new ASD chip without time walk corrections. With the new ASD chip, the resolution is improved at all γ irradiations.

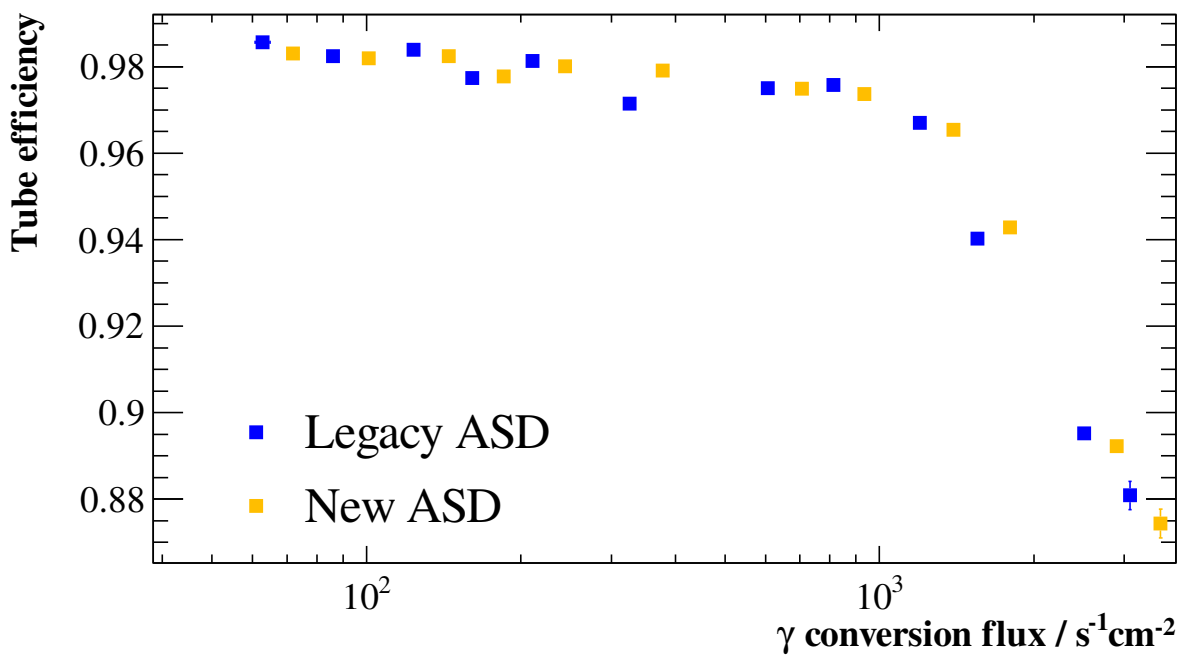


Figure 7.6: Comparison of the average tube efficiency of 0.5 m long 30 mm diameter drift tubes with the legacy and the new ASD chip.

The maximum γ conversion flux expected in MDT chambers in the ATLAS Muon Spectrometer at the HL-LHC is $400 \text{ s}^{-1}\text{cm}^{-2}$. The measurement campaign in the GIF⁺⁺, in which γ conversion fluxes above $3500 \text{ s}^{-1}\text{cm}^{-2}$ were achieved, proved that 30 mm diameter drift tubes provide spatial resolutions of 0.167 mm with the legacy ASD chip and 0.165 mm at $1000 \text{ s}^{-1}\text{cm}^{-2}$ γ conversion flux. At $1500 \text{ s}^{-1}\text{cm}^{-2}$, the resolution deteriorated to 0.193 mm and 0.197 mm for the two MDT ASD chip versions. The efficiency at $1000 \text{ s}^{-1}\text{cm}^{-2}$ was 97.2 % with both chip versions and at $1500 \text{ s}^{-1}\text{cm}^{-2}$ 94.4 % were achieved with the legacy ASD chip and 95.9 % with the new ASD chip. Due to the increased signal amplification and faster signal peaking time with at the same time lower noise rate and threshold spread allowing for 10 mV lower nominal thresholds, resolution and efficiency can be further improved with the new ASD chip.

Summary

In order to increase the rate of proton-proton collisions at the **Large Hadron Collider (LHC)** for precision measurements of the Higgs boson and top quark and for searches for new physics, the instantaneous luminosity will be upgraded to $\mathcal{L} = 7 \cdot 10^{34} \text{ cm}^{-2} \text{ s}^{-1}$ for the future **High-Luminosity LHC (HL-LHC)**. In order to increase the collision energy and the reach for physics beyond the Standard Model a new FCC-hh proton proton collider is considered with a center-of-mass energy of 100 TeV and instantaneous luminosities up to $\mathcal{L} = 3 \cdot 10^{35} \text{ cm}^{-2} \text{ s}^{-1}$. The unprecedented resulting particle fluxes drastically increase the demands on the radiation hardness and rate capability of the particle detectors and electronics.

In the ATLAS muon spectrometer, Monitored Drift Tube (MDT) chambers are used for precise muon track reconstruction. They consist of six to eight layers of 30 mm diameter aluminum drift tubes filled with Ar/CO_2 (93/7) gas at 300 kPa pressure and operated at a gas gain of $2 \cdot 10^4$. The vast majority of hits in the drift tubes is due to the cavern background consisting of γ particles and neutrons with average energies about 1 MeV. The spatial resolution and muon detection efficiency are deteriorated by increasing tube hit rates, which scale with the instantaneous luminosity. In the muon system of the baseline detector for the FCC-hh hadron collider, γ conversion fluxes up to $30 \text{ ms}^{-1} \text{ cm}^{-2}$ are expected, including a safety factor of three.

New small-diameter **Muon Drift Tube (sMDT)** chambers with reduced tube diameter of 15 mm have been developed to ensure efficient and precise track reconstruction in all parts of the ATLAS muon spectrometer after the HL-LHC upgrade. They are also very well suited for the muon detector of a future 100 TeV FCC-hh hadron collider. Compared to 30 mm tube diameter, the effects of space charge, i.e. gain drop due to shielding of the wire potential and stochastic variations of the electric field and drift velocity caused by fluctuations of the space charge density, are reduced by an order of magnitude in the sMDT chambers. They are also exposed to half the γ hit rate due to their smaller cross section.

Due to the smaller tube diameter, sMDT chambers fit into regions of the ATLAS muon spectrometer in the barrel middle (BM) previously not instrumented. The twelve so called BMG chambers were the first sMDT chambers to be installed in ATLAS for LHC run 2 in 2016. An extensive test and quality control

program ensured reliable operation in the detector and provided input for design optimizations for the upcoming sMDT chamber construction for HL-LHC.

The sMDT chambers currently in operation in the ATLAS experiment are read out with the same electronics as the MDT chambers. For HL-LHC this MDT chamber front-end electronics has to be replaced and new ASD chips had to be developed in order to cope with the ten times higher first-level trigger rates and to allow for the implementation of a highly-selective (s)MDT barrel first-level muon trigger which will also be necessary for FCC-hh. In addition, as has been observed in previous studies and confirmed by the test beam measurements presented in this thesis, the MDT ASD front-end readout chip is not optimal for the operation of sMDT chambers at much higher background rates than encountered in the ATLAS muon system for instance at a future FCC-hh hadron collider. The spatial resolution and muon efficiency is substantially deteriorated due to pileup of muon on background hits at high γ hit rates. It has been studied in this thesis, to which extent this signal pileup can be suppressed by amplifier-shaper electronics with baseline restoration (BLR). Two discrete amplifier-shaper versions with BLR have been investigated and compared with the old and new MDT ASD chips. One circuit adopted bipolar shaping like the ATLAS ASD chip, the other one unipolar shaping, which becomes feasible with BLR.

Both circuits had in common that the shaping time was only about one third of the shaping time of the ASD chip. Such short shaping times were disfavored in the past because of higher noise level at larger bandwidth. Indeed, the tests showed that the discriminator thresholds of the discrete circuits had to be raised by about a factor of two compared to the ASD chip. However, they also showed significantly better spatial resolution with the discrete circuits are achieved compared to the ASD chip due to the shorter peaking time. This is also true for the new MDT ASD chip for HL-LHC which has three quarters of the peaking time of the legacy ASD chip and larger amplifier gain. The improved signal rise time reduces slewing of the threshold crossing time.

In the case of the discrete bipolar shaper, no degradation of the tube efficiency due to the increased threshold was observed. A minor efficiency degradation for muon tracks close to the tube walls was found for the unipolar shaper. Without baseline restoration, both discrete shaper circuits provide similar spatial resolution up to the highest γ conversion fluxes of $22 \text{ ms}^{-1} \text{ cm}^{-2}$ reached for the first time at the new **Gamma Irradiation Facility** GIF⁺⁺ at CERN, showing the sMDT chambers can be operated reliably in terms of resolution and efficiency in all parts of the ATLAS muon spectrometer at HL-LHC and the FCC-hh muon detector with more than 100 % safety margin. The studies prove that unipolar shaping with strong ion tail cancellation can be used for drift tube detectors even at very high background fluxes. The baseline restoration stage leads to a small improvement of the spatial resolution for the unipolar shaper and a substantial improvement for the bipolar shaper and significantly reduces the deadtime, increasing the detector efficiency. The improvement did not reach the optimum expected from simulation. Improvements

are possible by reducing the noise and optimizing the shaping time constants, in particular the capacitances of the discrete circuits. The optimization also has to be done in view of implementing the discrete circuit as ASIC.

Appendix A

The Discrete Shaping Circuits

This appendix shows the complete schematics of the discrete amplifier-shaper circuits with BLR stage and discriminator.

In the conceptual circuit diagram of the unipolar and bipolar shaper in Figure 4.7 and Figure 4.10, respectively, the decoupling capacitors, the low pass filters of operating voltages, generated by digital-to-analog converters (DAC), the power supply of the active components the DC level defining voltage dividers of the two active shaper stages and the discriminator hysteresis have been left out for simplicity.

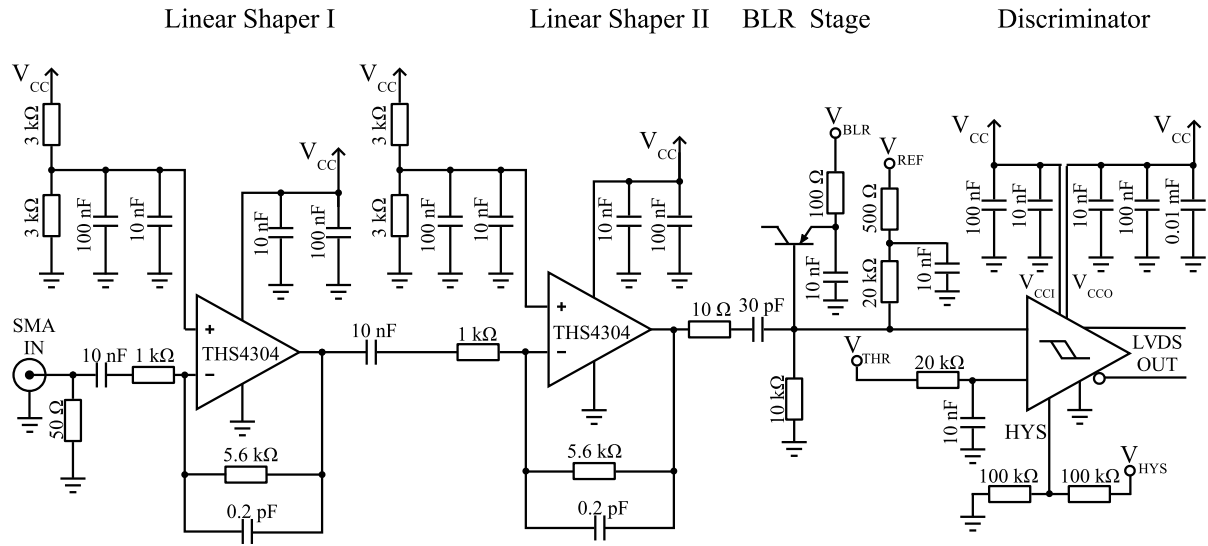


Figure A.1: Full schematics of the discrete unipolar circuit.

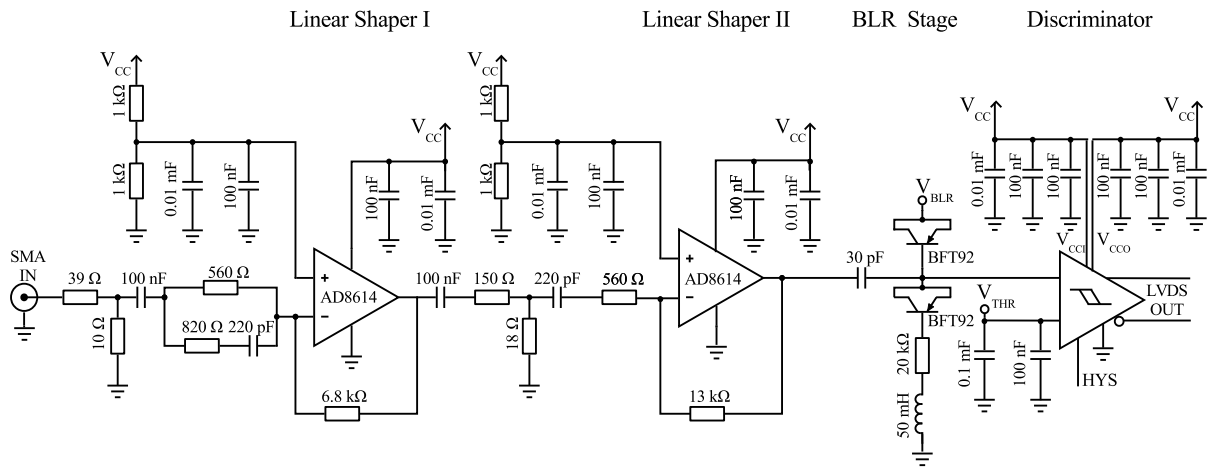


Figure A.2: Full schematics of the discrete bipolar circuit.

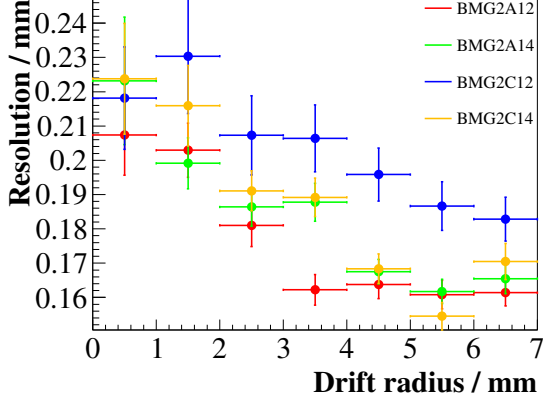
Appendix B

BMG sMDT Chamber Performance

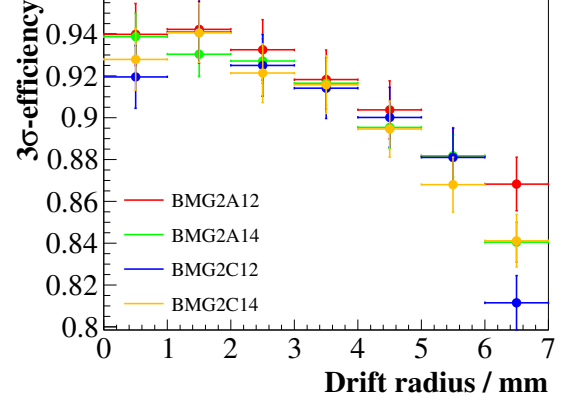
This appendix shows details of the performance of the twelve BMG chambers for the middle layer of the ATLAS muon spectrometer tested in the cosmic ray setup described in section 3.2.4.

Figure B.1 shows the dependence of the spatial drift tube resolution on the drift radius for each chamber. As described in section 3.2.5, multiple scattering adds $\sigma_{\text{MS}} = (0.11 \pm 0.01)$ mm to the tube resolution. The average drift tube resolution of all chambers is shown in Figure 3.20 together with the expected value from simulation. The average resolution over all drift radii is shown in Figure 3.21 for each of the twelve BMG chambers.

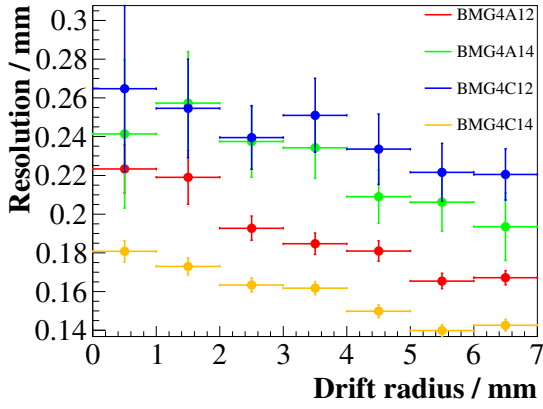
The BMG tube hit efficiencies discussed in section 3.2.6 are complemented in Figure B.2 by the 3σ -efficiencies (cf. section 6.3.3) based on the measured resolution shown in Figure B.1. The dependence on the drift radius is shown for each chamber.



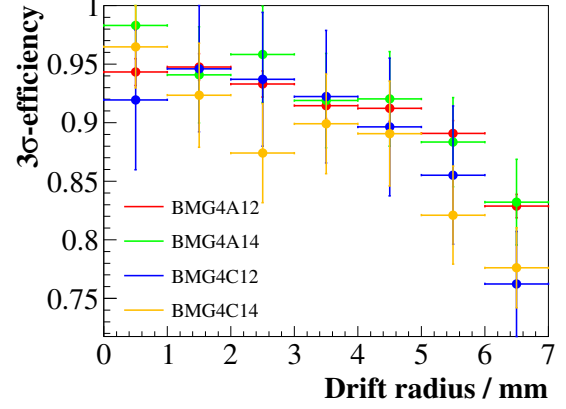
(a) BMG-2 chambers.



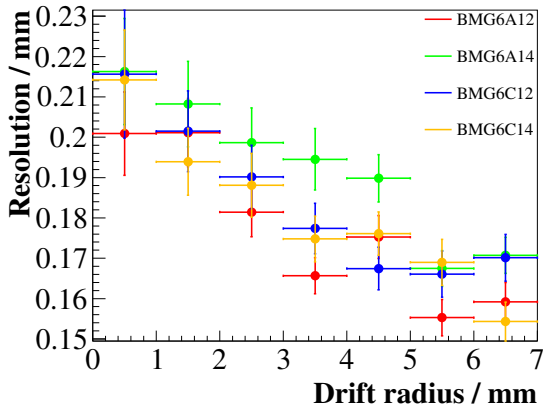
(a) BMG-2 chambers.



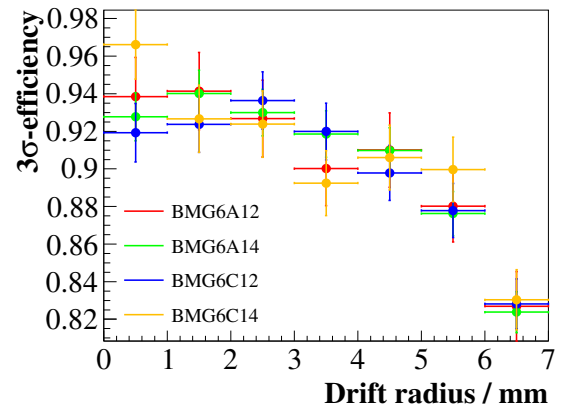
(b) BMG-4 chambers.



(b) BMG-4 chambers.



(c) BMG-6 chambers.



(c) BMG-6 chambers.

Figure B.1: Dependence of the spatial resolution of the BMG on the drift radius for each chamber.

Figure B.2: Dependence of the 3σ -efficiency of the BMG chamber on the drift radius for each chamber.

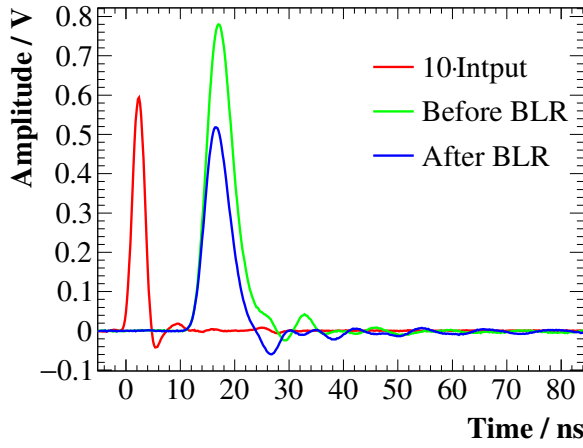
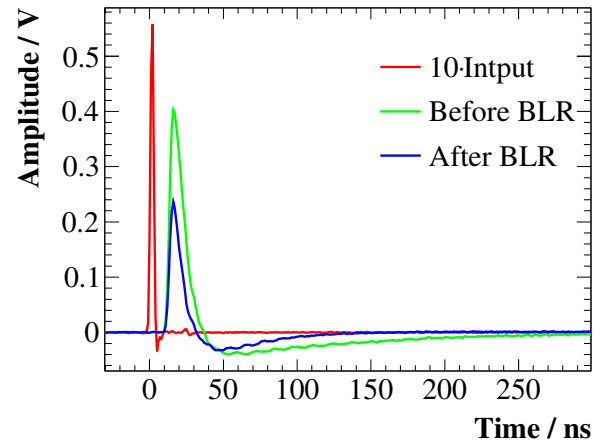
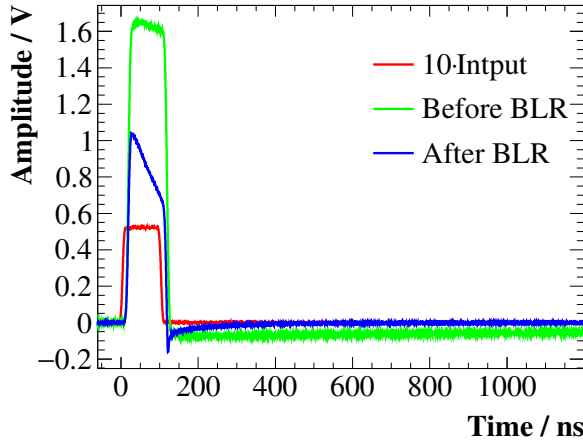
Appendix C

Pulse Generator Response of the Discrete Shapers

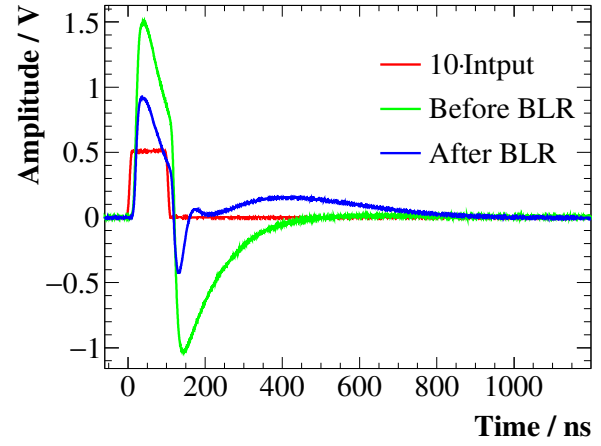
This appendix contains additional plots of the response studies of the uni- and bipolar shaper to the generator input pulses discussed in section 5.1.1.

Figure C.1 shows a direct comparison of the pulse evolution with the unipolar (left) and bipolar (right) shaper board. The amplitudes of the input pulses (red) of the three types (top: δ , middle: rectangular, bottom: triangular) are scaled by a factor of ten, to allow for a direct comparison with the output of the last shaper stage, i.e. before the BLR (green) and the shaper output after the BLR (blue). The delay between the signal input and the shaper response results from the signal propagation time through the measurement cables (cf. Figure 5.4). The pulse evolution of these pulses is compared between the unipolar and bipolar shaper in Figure C.2. The pulse shown were used to study pileup effects (cf. section 5.1.3).

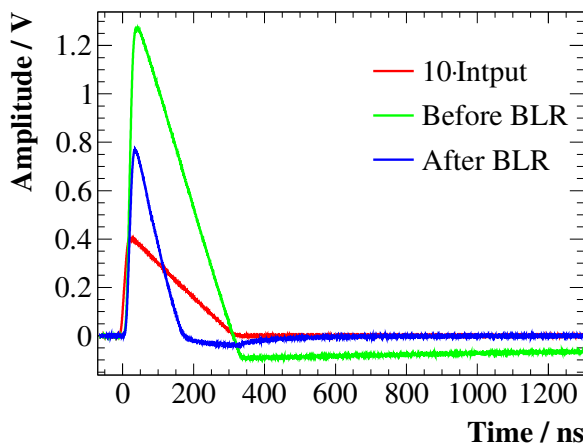
Figure C.3 summarizes the plots in Figure 5.20 to allow for a direct comparison. It shows the undershoot-to-signal ratio of the unipolar board, i.e. after the the BLR stage. The input pulses had rectangular shape with amplitudes of 10 to 100 mV and 20 to 200 ns width. With $V_{\text{BLR}} \geq 1000$ mV, the circuit is strongly nonlinear. For $V_{\text{BLR}} \geq 1500$ mV, the large BLR current leads to a fast discharge of the capacitor of the BLR stage, which results in an increased ratio for wide pulses and hence a deviation between pulses of the same input charge. Analog, Figure C.4 shows a summary of the bipolar shaper response to rectangular pulses in Figure 5.37.

(a) δ pulse into the unipolar shaper.(d) δ pulse into the bipolar shaper.

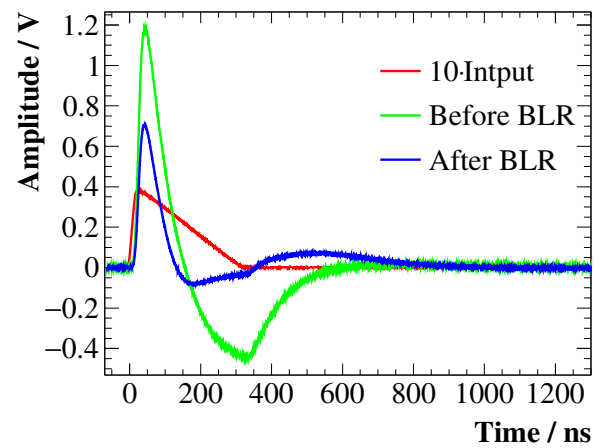
(b) Rectangular pulse into the unipolar shaper.



(e) Rectangular pulse into the bipolar shaper.

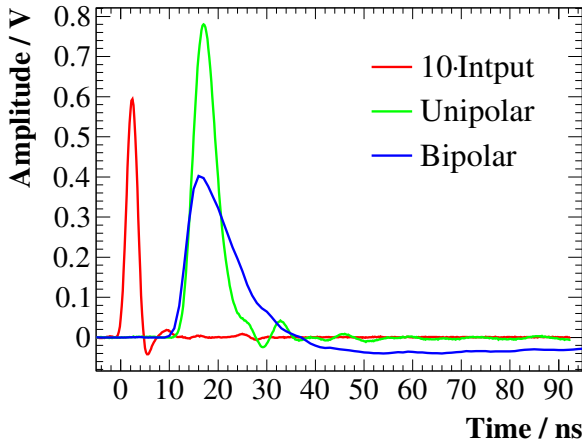


(c) Triangular pulse into the unipolar shaper.

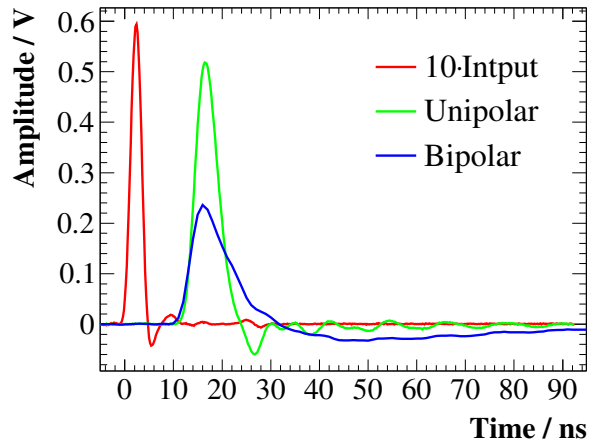


(f) Triangular pulse into the bipolar shaper.

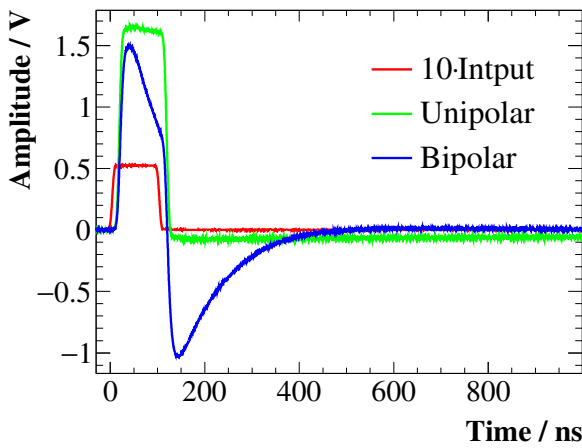
Figure C.1: Pulse shape evolution of the three types of generator pulses for the uni- and bipolar shaper.



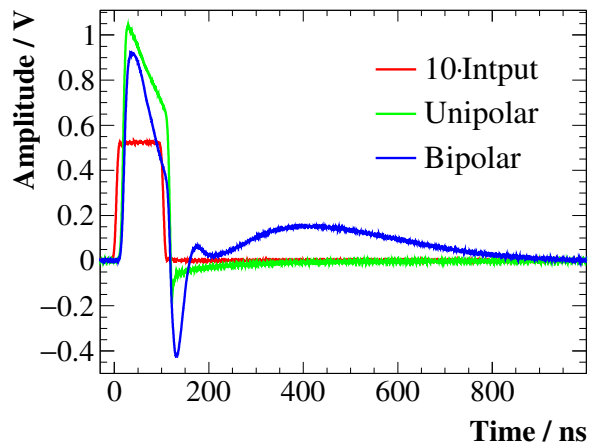
(a) Response of the last active shaper stage to a δ input pulse.



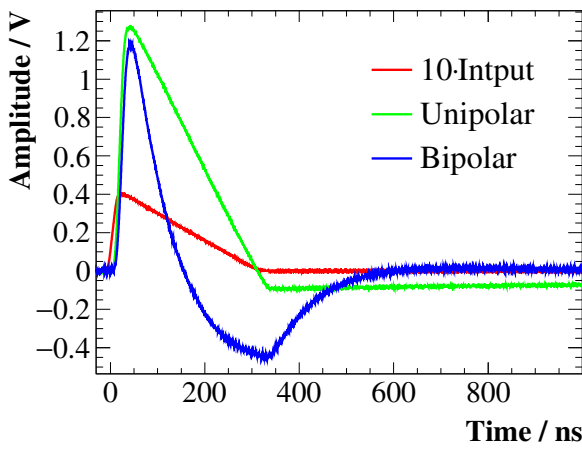
(d) Response of the BLR stage to a δ input pulse to the shaper.



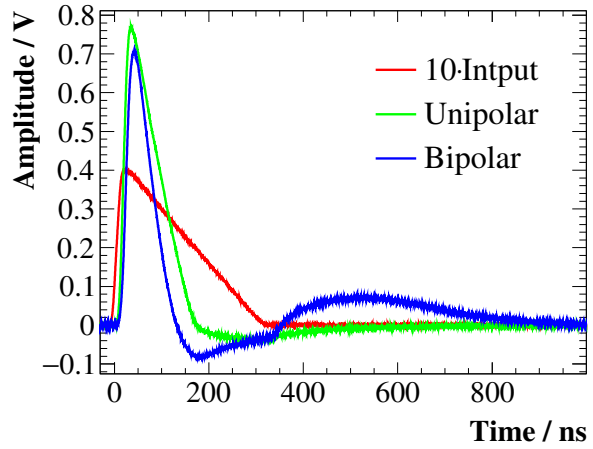
(b) Response of the last active shaper stage to a rectangular input pulse.



(e) Response of the BLR stage to a rectangular input pulse to the shaper.



(c) Response of the last active shaper stage to a triangular input pulse.



(f) Response of the BLR stage to a triangular input pulse to the shaper.

Figure C.2: Comparison of the unipolar and bipolar shaper response to the three types of generator pulses.

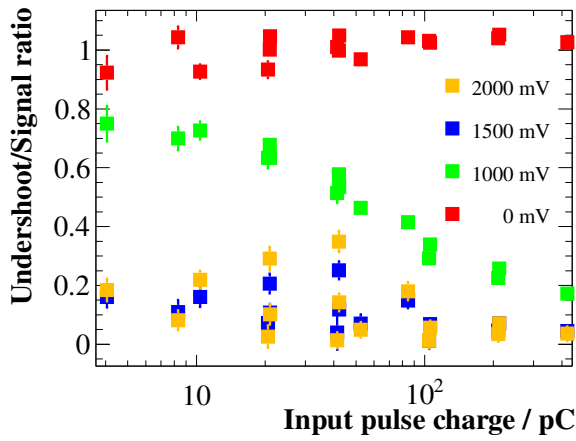


Figure C.3: Undershoot-to-signal ratio of the unipolar board for different BLR voltages. The input width and amplitude was varied.

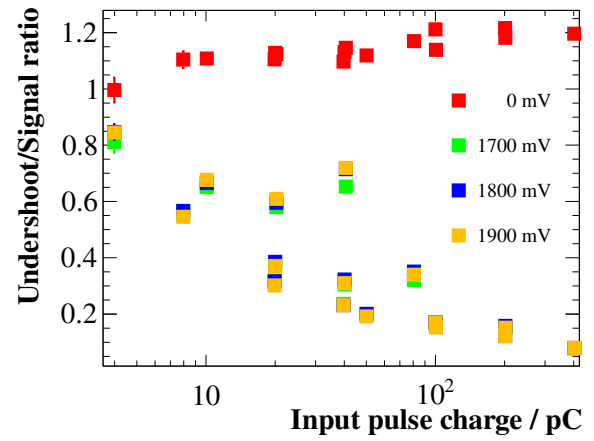


Figure C.4: Undershoot-to-signal ratio of the bipolar board for different BLR voltages. The input width and amplitude was varied.

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