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Design and Optimization of the Beam Orbit and Oscillation Measurement System for the Large Hadron Collider

Dissertation Thesis

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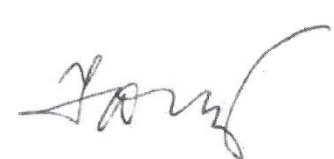
1. Design, prototype, build and optimize the new beam position measurement system for the CERN Large Hadron Collider (LHC). The system should consist of front-end units for processing of Beam Position Monitor (BPM) signals, transmitting the results to the servers over Ethernet network.
2. Develop original BPM signal processing based on the novel compensated diode detection technique. Optimize the processing for beam orbit resolution, accuracy and long-term stability.
3. Develop original BPM signal processing based on the RF diode detectors for measuring small beam oscillations. The data acquisition and digital processing should be synchronous to the LHC beams.
4. Install and commission the final systems for operational use with LHC beams. Perform and analyze the laboratory and beam measurements to characterize and optimize the systems. Optimize the systems for robustness and reliability adequate for machine protection purposes.

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Abstract

The Large Hadron Collider (LHC) accommodates some 100 collimators whose role is to perform beam cleaning and protect the machine from dangerous particle losses. The collimators are mechanical devices consisting of moveable jaws. Precise positioning and control of the jaws is critical for the cleaning efficiency. Therefore, after the first long shut-down, the LHC was equipped with 18 collimators of a new type. Movable jaws of the new collimators have embedded beam position monitors (BPM) which allow their precise positioning with respect to the circulating beam. However, the existing electronic systems for BPM signal processing could not achieve the required resolution and precision of the position measurements.

In addition to the new collimator BPMs, the LHC accommodates more than 1000 BPMs for measuring the transverse positions of the two counter-rotating beams. The standard LHC BPM system uses these BPMs to measure the orbits and oscillations of the beam. The most important BPMs are located next to the LHC experiments. The beam measurements in such locations are the most challenging as the two beams have to be controlled at a fine precision in order to achieve their efficient colliding. Improving the resolution and precision of the position measurements can contribute to the improvement of the machine performance. Performance of the LHC also depends on the magnet optics. Important machine parameters like betatron coupling, beta-beating and phase advance are obtained by exciting the transverse beam oscillations and measuring the amplitudes and phases of the beam response using BPMs. The standard BPM system requires millimetre-order beam excitation to obtain the measurements of a sufficient quality. For machine protection reasons these measurements can be performed only with special beams and dedicated machine set-up.

The main task of this doctoral work was to design, prototype, build and optimize a new electronics system for beam position and oscillation measurements in LHC. The system called DOROS (Diode Orbit and Oscillation) was primarily designed for the new LHC collimators. The same system was also used to provide high-resolution orbit and oscillation measurements in the selected LHC BPMs. The DOROS systems consist of front-ends, each processing signals from up to two BPM sensors composed of horizontal and vertical transverse planes or to two collimators consisting of upstream and downstream BPMs. The RF signals in a front-end are first filtered, amplified and then split into two diode detector sub-systems which work in parallel.

A so called Diode Orbit (DOR) subsystem, based on novel compensated diode detector technique, was designed to perform beam orbit measurements. This thesis describes the analogue processing channels followed by the digital signal processing of the turn-by-turn data and real-time algorithms. The algorithms provide beam based calibration of the channel asymmetries as well as autonomous gain control of the front-end amplifiers. The DOR subsystem was characterized with the laboratory measurements and its performance was demonstrated on a number of beam measurements, showing the achieved sub-micrometre resolution, precision, and long-term stability. The position readings from selected front-ends are also used by the LHC interlock system which terminates operation with beams if the beam positions exceed safe limits.

A so called Diode Oscillations (DOS) subsystem, which is based on direct diode detection technique, was designed and optimised to measure small beam oscillations. This thesis describes both the analogue and digital signal processing in a front-end as well as its synchronization and timing circuits. The sampling of the ADCs can be synchronized to the beam allowing to perform precise measurements of the beam coupling and phase advance.

The front-end units continuously transmit the measurement readings over Ethernet at 25 Hz rate to the system servers synchronously to the LHC timing. Together with the measurement readings the front-ends transmit also statuses and other data important for diagnostics and reliability of the system. At the same time the acquisition data is stored in parallel to the front-end buffers for detailed turn-by-turn and post-mortem signal analysis.

Abstrakt

Kolimátory časticových zväzkov vo Veľkom Hadrónovom Urýchl'ovači (LHC) sú mechanické zariadenia pozostávajúce z pohyblivých čeľustí, ktorých úlohou je chrániť urýchl'ovač pred nebezpečnými stratami častíc, a tak zabrániť jeho poškodeniu. Presné umiestnenie kolimačných čeľustí vzhľadom na cirkulujúci časticový zväzok je teda kritické pre efektivitu kolimácie a čistenia zväzkov. Celkovo sa v LHC nachádza približne 100 kolimátorov v hierarchických zoskupeniach umiestnených v kritických lokalitách urýchl'ovača. Počas prvej dlhej odstávky bol LHC urýchl'ovač vybavený 18 kolimátormi nového typu. Pohyblivé čeľuste týchto kolimátorov majú zabudované senzory merania pozície zväzkov, takzvané BPM, slúžiace na ich presné umiestnenie vzhľadom na cirkulujúci zväzok. Potrebné rozlíšenie a presnosť merania pozície zväzkov nebolo možné dosiahnuť s existujúcimi elektronickými systémami na spracovanie BPM signálov.

Okrem BPM senzorov v nových kolimátoroch je LHC vybavené viac ako 1000 BPM senzormi na meranie transversálnej pozície oboch protiběžných zväzkov. Štandardný LHC BPM systém je elektronický systém, ktorý spracúva signály z týchto senzorov a umožňuje merať orbity a oscilácie zväzkov v LHC. Najdôležitejšie BPM senzory sa nachádzajú v blízkosti LHC experimentov. Pozície oboch cirkulujúcich zväzkov v týchto lokalitách musia byť ovládané s vysokou presnosťou pre dosiahnutie efektivity kolízií v experimentoch. Zvyšovanie dosiahnuteľného rozlíšenia a presnosti systémov merania pozície zväzkov tak môže priamo prispieť ku vylepšeniu výkonnosti LHC urýchl'ovača. Táto tiež úzko súvisí s optimálnym nastavením magnetovej optiky. Dôležité parametre ako napríklad transversálne väzby, beta-beating alebo fázové posuny oscilácií, je možné merať s pomocou budenia transversálnych oscilácií a merania amplitúdovej a fázovej odozvy zväzku s pomocou BPM systémov. Na dosiahnutie adekvátnej kvality merania je potrebné budiť oscilácie zväzkov s rádovo milimetrovými amplitúdami. Spomenuté merania je teda možné vykonávať v LHC len so špeciálnymi nastaveniami urýchl'ovača a typmi zväzkov.

Hlavnou úlohou tejto dizertačnej práce bolo navrhnuť, zostrojiť a optimalizovať nový typ elektronického systému slúžiaceho na merania pozície a oscilácií časticových zväzkov s pomocou BPM monitorov v urýchl'ovači LHC. Tento systém, nazývaný DOROS, bol navrhnutý primárne pre nový typ LHC kolimátorov so zabudovanými BPM senzormi. Identický DOROS systém bol taktiež použitý na presné merania orbity a oscilácií zväzkov vo vybraných BPM senzoroch v LHC nachádzajúcich sa najmä v oblasti experimentov. Oba systémy DOROS pozostávajú z elektronických jednotiek, takzvaných front-endov. Každá jednotka umožňuje spracovávať rádiovfrekvenčné signály a merať pozíciu zväzkov v dvoch BPM senzoroch. Tieto signály sú najskôr filtrované, zosilnené a potom rozdelené na následné spracovanie dvomi paralelnými podsystémami založenými na princípe rádiovfrekvenčných diódových detektorov.

Takzvaný Diode Orbit (DOR) podsystém bol navrhnutý a optimalizovaný na meranie orbity zväzkov. DOR podsystém je založený na novej technike kompenzovaných diódových detektorov. Táto práca sa zaoberá analógovým ako aj digitálnym spracovaním signálov a algoritmami vykonávanými v reálnom čase. Dynamický rozsah signálov je automaticky regulovaný s pomocou autonómneho algoritmu na ovládanie vstupných zosilňovačov. Asymetrie medzi jednotlivými analógovými kanálmi sú minimalizované s pomocou kalibračného algoritmu využívajúceho signály z BPM senzorov. Výsledný DOR podsystém bol charakterizovaný s pomocou laboratórnych meraní. Táto práca taktiež prezentuje niekoľko meraní realizovaných so zväzkami v LHC urýchl'ovači. Rozlíšenie, presnosť a dlhodobá stabilita merania pozície zväzkov bola dosiahnutá rádovo menej ako jeden mikrometer. Niektoré DOROS jednotky sú taktiež využívané takzvaným LHC interlock systémom, ktorý slúži na ukončenie činnosti LHC, ak merané hodnoty pozície zväzku prekročia limity stanovené pre bezpečný chod urýchl'ovača.

Takzvaný Diode Oscillation podsystem, ktorý je založený na princípe priamej diódovej detekcie, bol navrhnutý a optimalizovaný na meranie mikrónových oscilácií zväzkov. Táto práca sa zaoberá analógovým a digitálnym spracovaním signálov oscilácií zväzkov, ako aj súvisiacimi synchronizačnými obvodmi. Vzorkovanie oscilačných signálov v ADC prevodníkoch môže byť synchronizované so zväzkom, čo umožňuje vykonávať presné merania fázového posunu a transversálnej väzby zväzkov.

Každá DOROS jednotka posiela spracované hodnoty jednotlivých meraní prostredníctvom Ethernetového protokolu s frekvenciou 25 Hz. Táto komunikácia je synchronizovaná s časovacími signálmi z LHC. Každý datagram taktiež obsahuje informácie o stave jednotky, ako aj iné dáta slúžiace na diagnostiku a spoľahlivosť systému. Hodnoty z ADC prevodníkov sú zároveň ukladané do lokálnej pamäte nachádzajúcej sa v každej DOROS jednotke. Tieto dáta sú prístupné prostredníctvom Ethernetového protokolu a je ich možné využiť pri takzvanej turn-by-turn alebo post-mortem analýze.

Declaration

I, Jakub Olexa declare that I have authored this thesis independently, that I have not used other than the declared sources / resources, and that I have explicitly marked all material which has been quoted either literally or by content from the used sources.

Bratislava, May 31th, 2018



.....

Signature

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List of abbreviations

AC	Alternating Current
ADC	Analogue – Digital converter
ADT	LHC transverse damper system
AGC	Automatic Gain Control
ALICE	A Large Ion Collider Experiment
AM	Amplitude Modulation
ATLAS	A Toroidal LHC Apparatus experiment
AUCAL	Automatic Calibration
BBQ	Base-Band Tune system
BI	Beam Instrumentation
BLM	Beam Loss Monitor
BPM	Beam Position Monitor
BPMSW	Beam Position Monitor Stripline Warm
BST	Beam Synchronous Timing
CDD	Compensated Diode Detector
CDR	Clock Data Recovery
CERN	“Conseil Européen pour la Recherche Nucléaire”
CMOS	Complementary metal-oxide-semiconductor
CMS	Compact Muon Solenoid experiment
CPU	Central Processing Unit
CRS	Crossed
DAC	Digital-Analog converter
DC	Direct Current
DDR	Double Data Rate memory
DDS	Digital Direct Synthesis
DIP	Dual Inline Package
DIV	Divider
DMA	Direct Memory Access
DOR	Diode Orbit
DOROS	Diode Orbit and Oscillations system
DOS	Diode Oscillations
DPD	Differential Phase Detector
DRDY	Data Ready
DSP	Digital Signal Processing
DTFT	Discrete Time Fourier Transform
ENBW	Equivalent Noise Bandwidth
FESA	Front-End Software Architecture
FFT	Fast Fourier Transformation
FPGA	Field Programmable Gate Array
FS	Full-scale
I2C	Inter-Integrated Circuit
IA	Instrumentation Amplifier
ID	Identification Number
IIR	Infinite Impulse Response

IO	Input / Output
IP	Internet Protocol address
IP*	Interaction Point in an experiment
IPGA	Integrated Programmable Gain Amplifier
IR	Interaction Region
JFET	Junction Field Effect Transistor
JTAG	Joint Test Action Group
LED	Light Emitting Diode
LEP	Large Electron Positron collider
LF	Low Frequency
LHC	Large Hadron Collider
LHCb	Large Hadron Collider beauty experiment
LPF	Low-Pass Filter
LSb	Least Significant bit
LSB	Least Significant Byte
LTG	Local Timing Generator
LVDT	Linear Variable Differential Transformers
MAC	Medium Access Control
MCLK	Master Clock in a sigma-delta digitizer
MSb	Most Significant bit
MSB	Most Significant Byte
MTBF	Mean Time Between Failures
OFC	Orbit Feedback Controller
PCB	Printed Circuit Board
PDL	Programmable Delay Line
PGA	Programmable Gain Amplifier
PLL	Phase Locked Loop
PSD	Power Spectral density
RF	Radio Frequency
RIS	Random Input Sampling
RMS	Root Mean Square
SC	Synchronization Circuit
SFP	Small Form Pluggable
SMA	Sub-Miniature version A
SNR	Signal-to-Noise Ratio
SP3T	Single Pole three Throw
SPI	Serial Peripheral Interface
SPICE	Simulation Program with Integrated Circuit Emphasis
SPS	Super Proton Synchrotron
STC	Synchronization and Timing Circuits
STD	Standard Deviation
STR	Straight
TCP	Transmission Control Protocol
TCPIP	Transmission control protocol/ Internet Protocol
TSG	Test Signal Generator
UDP	User Datagram Protocol

VCO	Voltage Controlled Oscillator
VdM	Van der Meer scan
VREF	Voltage Reference
WBTN	Wide-Band Time Normaliser

List of symbols

a	Polynomial coefficients
A	Signal amplitude
A_{exc}	Equivalent amplitude of the betatron oscillations
c	Polynomial coefficients
C	Capacitance
d	Decimation factor
D	Distance between two BPM electrodes
$DOR.ch.i$	Normalised electrode value measured on the output of the DOR channel i
$ENBW$	Equivalent noise bandwidth
f	Signal frequency
f_{AM}	Amplitude modulation frequency
f_c	Cut-off frequency
f_{ext}	Excitation frequency
f_{rev}	Revolution frequency of the LHC beams
FS	ADC full-scale
f_s	Sampling frequency
f_{tx}	Frequency of the data transmission
g	Gain
h	Impulse response
l_{ADC}	Signal on the ADC input from the left BPM electrode channel processing
L_{ADC}	Samples from the DOR ADC channel processing left BPM electrode signal
L_{cal}	Calibrated samples of left BPM electrode
L_{crs}	Samples from the DOR processing of left BPM electrode signals processed with crossed multiplexer configuration
l_D	Signal from left diode detector
L_{DDS}	Number of samples DDS in one sinewave period
L_{DOR}	Samples from the DOR processing of left BPM electrode signals
l_e	Signal from the left BPM electrode
l_{LP}	Signal from the left input RF filter
l_{RFA}	Signal from left RF amplifier
LSb	Least significant bit
L_{str}	Samples from the DOR processing of left BPM electrode signals processed with straight multiplexer configuration
m,n	Indices
n_{enob}	Effective number of bits
o	Offset
p	Normalised transverse beam position
P	Absolute transverse beam position
p_{ADC}	Estimated resolution of the normalised position as measured by the ADC

p_b	Assumed normalised position of the beam
p_{cal}	Calibrated normalised position
p_{err}	Position error
p_{gerr}	Normalised position error from gain asymmetry
p_m	Measured normalised beam position
p_{oerr}	Normalised position error from offset asymmetry
PSD	Power spectral density
p_x	Normalised transverse beam position in x axis
P_{xcoll}	Corrected absolute beam position as measured by the collimator BPM in the x axis
P_{xcorr}	Corrected absolute beam position as measured by the BPM in the x axis
P_{xlin}	Linear approximation of the absolute beam position as measured by the BPM
p_y	Normalised transverse beam position in y axis
P_{ycorr}	Corrected absolute beam position as measured by the BPM in the y axis
Q	Charge
Q_T	Total capacitor charge during time T
Q_τ	Charge flowing through the forward conducting diode during time τ
R	Resistance
r	Dynamic resistance
r_{ADC}	Signal on the ADC input from the right BPM electrode channel processing
R_{ADC}	Samples from the DOR ADC channel processing right BPM electrode signal
R_{cal}	Calibrated samples of right BPM electrode
R_{crs}	Samples from the DOR processing of right BPM electrode signals processed with crossed multiplexer configuration
r_D	Signal from right diode detector
R_{DOR}	Samples from the DOR processing of right BPM electrode signals
re	Signal from the right BPM electrode
r_{LP}	Signal from the right input RF filter
r_{RFA}	Signal from right RF amplifier
R_{str}	Samples from the DOR processing of right BPM electrode signals processed with straight multiplexer configuration
S_{DDS}	DDS counter step
SNR	Signal-to-noise ratio
SNR_n	Narrowband estimated signal-to-noise ratio
T	Time duration or repetition period
T_{AM}	Amplitude modulation period
T_{clk}	Clock period
t_{err}	Error time difference
$T_{hold-off}$	Programmed delay in the hold-off circuit
Thr_{max}	Maximum threshold value of the beam pulse comparator
$T_{largest\ gap}$	Time of the largest gap in the bunch train
T_{rev}	Revolution period
V	Voltage
V_d	Voltage drop across the diode
V_{DPD}	Signal on the differential phase detector output
V_i	Input voltage
V_{NOR}	Voltage signal on the NOR gate after low-pass filtering
V_o	Output voltage

V_{OR}	Voltage signal on the OR gate after low-pass filtering
V_{PS}	Gate power supply voltage
V_{ref}	Voltage reference
x	Horizontal axis
y	Vertical axis
z	Complex variable
α	Parameter of the infinite impulse response filter controlling its cut-off frequency
ε	Signal level after defined time duration
θ	Normalised digital frequency
θ	Normalised digital cut-off frequency
τ	Time duration

1 Introduction

The main subject of this doctoral thesis is development and optimisation of an electronics system dedicated for precision measurements of the transverse positions of the particle beams in the LHC. The system called DOROS (Diode Orbit and Oscillation) processes electrode signals from beam position monitors (BPM) in order to provide the beam orbit and oscillation measurements. The system was primarily designed for the LHC collimators with embedded BPM electrodes in the movable jaws. This system is critical for safety and reliability of the accelerator. The position measurements allow precision alignments of the collimator jaws with the beams. The readings from the system are also used by the LHC interlock system, which terminates operation with beams if the beam positions exceed safe limits. The DOROS system has been also used to measure beam positions in the most important BPMs around the main LHC experiments. The position measurements have been used to monitor a few LHC beam parameters, important for potential improvements of the machine performance.

The first chapter of this thesis introduces the LHC accelerator and some of its parameters important for this work, such as the beam orbits, betatron oscillations and quantities derived from them. The following subchapters describe the principles and operation of the collimation system and the standard BPM system in the LHC. Motivations and main goals of the thesis are summarized at the end of the chapter.

Signal processing in the DOROS system is based on diode detection techniques. The second chapter is dedicated to theoretical analysis of the diode detectors used for BPM signal processing and position calculations. Principles of the beam position measurements are introduced in the first subchapter. Description of the typical LHC BPM sensors is also provided. The following subchapter is dedicated to the analysis of a diode detector circuit as well as the diode compensation scheme used in the DOROS systems. Principles of its operation are demonstrated on a number of examples. Advantages of the diode detection techniques in the BPM signal processing are summarized at the end of the chapter.

The main part of this work is focused on the design, development and optimisation of the hardware and signal processing of the DOROS system. The third chapter is therefore dedicated to the present architecture and realisation of the LHC systems. Development history and the key milestones are documented in the first subchapter which also marks involvement of the author in the particular realisation stages of the project. The following subchapter provides an overview on the entire system including its main hardware and firmware blocks. The author participated in the full design process with the main focus on the acquisition, digital signal processing and synchronization aspects of the system.

The analogue and digital signal processing in the front-ends was divided into two subsystems which share the same BPM signals. The first subsystem called diode orbit (DOR), is described in a dedicated subchapter. The DOR, based on the compensated diode detectors, has been designed and optimized to measure the beam orbits. Author's contribution to the DOR subsystem was mainly in the hardware optimization and the digital signal processing design. Operation principles of the real-time control algorithms implemented in the FPGA were demonstrated on a number of examples. The following subchapter describes the second subsystem, a so called diode oscillation (DOS), which is based on RF diode detectors optimized for the beam oscillation measurements. The DOS channels share the input circuits and the RF signal processing with the DOR channels. The chapter provides description of the oscillation signal processing with particular focus on the ADC synchronization circuits.

The part of this thesis dedicated to the measurements and results achieved with the DOROS systems is split into four chapters, individually addressing the DOR and DOS subsystems. Performance of the system was optimized and characterized with the experimental results obtained with laboratory measurements. The results obtained from the beam measurements describe the performance of the operational systems in the LHC.

Achieved performance of the DOR subsystem was experimentally evaluated with the laboratory measurements presented in the fourth chapter. Described measurements were focused on several important aspects of the orbit subsystem, namely its linearity, latency, bandwidth, precision, systematics, long-term stability and temperature sensitivity. Each of the topics is introduced with a description of the measurement setup. The main results were then summarized at the end of each subchapter. Author also provides a summary dedicated to the individual topics along with a future outlook.

The fifth chapter of the thesis documents the performance of the DOR subsystem on a number of measurements acquired with the LHC systems. The first subchapter is dedicated to the measurements obtained with the operational DOROS installations on the LHC collimators. Beam measurements were focused on the alignment precision, orbit measurement stability and resolution of the system. The second subchapter is dedicated to the measurements with the DOROS system installed next to the LHC experiments. In these locations the precision measurements are the most challenging. Comparison of the DOROS system to the standard LHC BPM system is also provided. The third subchapter is dedicated to the limitations of the DOR subsystem explained on a number of example measurements. Author presents possible improvements along with a few preliminary results.

Performance of the DOS subsystem evaluated with laboratory measurements is presented in the sixth chapter. Its first subchapter characterizes the operation of the ADC timing circuits and demonstrates achieved performance of the individual synchronization blocks. The second subchapter is focused on evaluating the sensitivity of the DOS subsystem to small oscillation amplitudes. The seventh chapter documents the DOS performance with the beam oscillation signals acquired during the optics measurements in the LHC. Obtained results are then compared to those acquired with the standard LHC BPM system.

The conclusion chapter summarizes the main results and highlights the main contributions of this thesis.

1.1 CERN

CERN (“Conseil Européen pour la Recherche Nucléaire”) is the largest European organisation for nuclear research. The organisation was founded in 1954 after the World War II, unifying scientists from European countries to a peaceful co-operation. The CERN consists of more than 22 member states including countries outside of Europe [1]. The laboratory provides large scale accelerator complex, infrastructure and research facilities built on Suisse-French border near Geneva. Main part of the collaborative work at CERN is focused on the design, development, construction, operation and maintenance of accelerators, detectors and other technical infrastructures. Fundamental research and development of accelerator-related technologies is also useful in many other research fields, such as material sciences, spectroscopy and irradiation, or medical applications [2].

Main driving force behind development of accelerator technology is the particle physics research studying the fundamental constituents of matter. Research particle colliders are designed and built to accelerate and collide particle beams at high energies. High energy research accelerators are typically large, complex and expensive. Recent development in the superconductive technologies allowed building high energy accelerators like the Large Hadron Collider (LHC) at CERN which can achieve beam energies about 7 TeV [3]. Its efficient operation reduces running costs and increases amount of valuable research data. Performance of the LHC is often limited by the performance of its instrumentation.

1.2 The Large Hadron Collider

The Large Hadron Collider is the world’s largest accelerator with the circumference of almost 27 km [3]. It was built in a tunnel of the former Large Electron Positron collider (LEP). Particle beams of positively charged protons or heavy ions are injected into the LHC passing first through a chain of smaller accelerators. The LHC accelerator schematically shown in Fig. 1-1 consists of a ring with two beam lines. The two LHC beams are accelerated to high energies with opposite direction of the beam circulation. Their paths are crossed in the interaction regions in order to collide the particles inside the four main experiments, namely the ATLAS [4], CMS [5], ALICE [6] and LHCb [7]. The physics run in years 2010 - 2013 [8] yielded enough experimental data to discover the Higgs Boson [9].

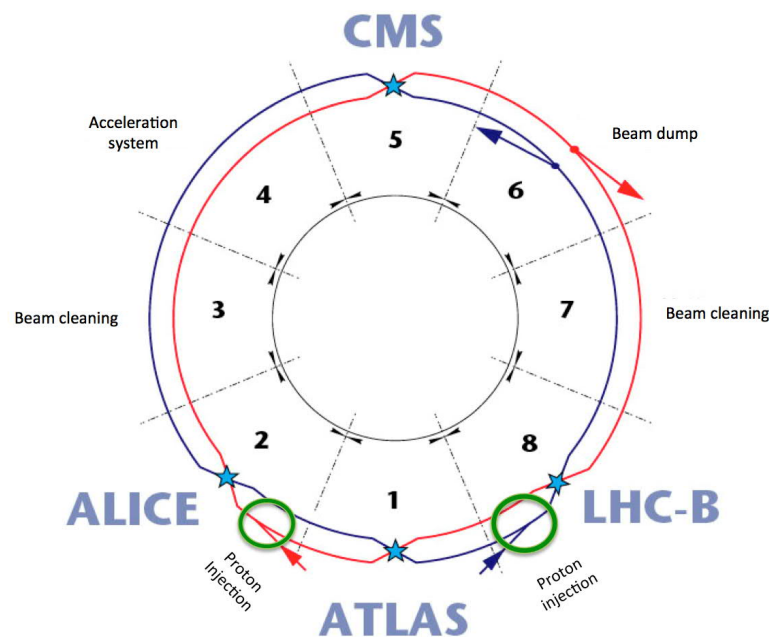


Fig. 1-1. Schematic diagram of the Large Hadron Collider ring with its two beam lines and four main experiments [10].

The LHC ring contains some 1200 superconducting dipole magnets that bend the trajectory of both beams into closed circumference inside of the accelerator vacuum pipes. The dipole magnets can produce magnetic fields of more than 8 T required for beam bending in the horizontal plane. The magnet power supplies deliver current of more than 11 kA controlling the magnetic fields. Some 400 main superconducting quadrupole magnets are distributed on the LHC circumference. The quadrupoles allow controlling the transverse beam sizes in the machine as well as focus the beams in the experiments. Apart of the main dipole and quadrupole magnets, the LHC also contains some 7000 other corrector magnets for controlling beam dynamic effects [3].

The particle beams are bunched and accelerated by a system of superconducting RF cavities. Alternating electro-magnetic fields provide accelerating energy and keep injected particles inside bunches. The beams circulate at revolution frequency f_{rev} of approximately 11.2 kHz defined by the RF system. The revolution period is then equal to some 89.3 μ s. The LHC can store about 2800 bunches in each of its two beam lines [3]. All bunches over the accelerators circumference are organised in a so called bunch pattern. Bunch spacing used in physics operation is typically 25 ns.

Bunch intensity is a parameter defining number of particles in a bunch. Beam intensity defines a sum of all particles in the machine. A so called a pilot bunch has intensity of 5×10^9 elementary charges [3]. Such beam intensity is safe for the machine even during abrupt particle losses at injection energy. A so called nominal bunch has intensity of more than 1×10^{11} which is required for large production of particle collisions in the experiments [3]. Beams consisting of many nominal bunches are therefore used in physics operation.

Particle losses in the machine cannot be completely avoided due to imperfections in the magnet alignments as well as many beam dynamic effects. Excessive beam losses on superconductive can cause magnet quench, an effect of sudden loss of superconductivity. Uncontrolled particle losses pose potential risk of damaging superconducting magnets and other sensitive equipment, rendering the machine inoperable for long time. The LHC collimation system was designed to protect the superconductive magnets and other sensitive equipment by intercepting unstable particles from the beam. The collimation system is a beam cleaning system arranged in multiple stages. Each stage consists of a collimator device installed in the dedicated locations on the LHC circumference. This system is critical for safe and reliable operation of the accelerator.

A typical LHC cycle used for physics operation is composed of a few phases [3]. The cycle begins with preparation of machine for the beam injection. A single pilot is first injected in order to validate the machine settings. The nominal bunches are then injected and distribute on the LHC circumference according to the required bunch pattern. After the injection phase is finished the machine is prepared for the energy ramp. Both beams are then accelerated to the programmed energy. Once the acceleration phase is finished the LHC optics is prepared and adjusted for the beam collisions. Physics data taking period, a so called stable beam phase can last for more than 24 hours. During this time, the intensity of the beam naturally decays. Once reaching the threshold of optimal machine run the beams are ejected into the beam dump and the magnetic fields are ramped down for the next machine cycle.

The primary purpose of the LHC is to produce large number of particle collision events that the main experiments can process. Luminosity is a quantitative measure of the accelerator performance and efficiency in production of the collision events. Optimisation of the luminosity production [11] in the LHC depends on a number of machine settings and beams parameters such as the number of bunches, beam intensity, beam sizes or turnaround time.

Knowledge of the transverse position of a particle beam inside the beam pipe plays a crucial role in the operation, performance, reliability and safety of an accelerator. Systems measuring beam position provide means to control, diagnose and study many important beam parameters such as beam trajectory, orbit, tune and other derived quantities. More than 1000 BPM sensors are required in the LHC to measure the transverse positions of the beams [3]. A BPM sensor is a device composed of a system of sensing electrodes which allow non-invasive measurements of the beam position in the transverse plane. Majority of the BPM sensors in the LHC have four electrodes which allow measuring both horizontal and vertical beam positions in the vacuum pipe [12]. Dedicated electronics systems then process the BPM signals and provide position readings used by a number of monitoring and control systems. Measurement data is also stored in the logging databases for offline analysis.

1.2.1 Beam orbit and trajectory

Position of a bunch can be measured each time it passes through a BPM sensor. A BPM system able to measure positions of individual bunches is referred to as a bunch-by-bunch BPM system. Beam orbit can be defined as an average position of bunches. The averaging can be defined over a given number of bunches and beam revolutions also called turns. The orbits in the LHC are measured as an average of all bunch positions over a few hundred turns. A system optimised for measuring the positions on a one turn basis is referred to as a turn-by-turn system. A system optimized for position measurements over many turns is referred to as a multi-turn system.

As a bunch is advancing through the accelerator ring, the reading of its position can be obtained at each BPM location. Acquired readings from all BPMs on the circumference represent a trajectory of the bunch. Example of a trajectory measurement is displayed in Fig. 1-2. The measurement was acquired by the standard LHC BPM system [13] during the first LHC commissioning in 2008. The figure depicts beam position readings of the horizontal and vertical planes from more than 500 locations on the LHC circumference.

Precise measurements of the LHC orbits are the most important around the interaction regions. There the beam orbits have to be controlled within a few micrometres so that the two beams can collide. Some systems use the position readings for real-time feedback corrections. For example the LHC Orbit Feedback system maintains and controls the stability of the beam orbits in the machine [14]. The orbit measurements are also important for protection of the accelerator. For example, if the orbit exceeds safe limits the beams have to be automatically ejected to prevent potential damage. Performance of the LHC can be therefore limited by the resolution, long-term stability and accuracy of the measurement systems.

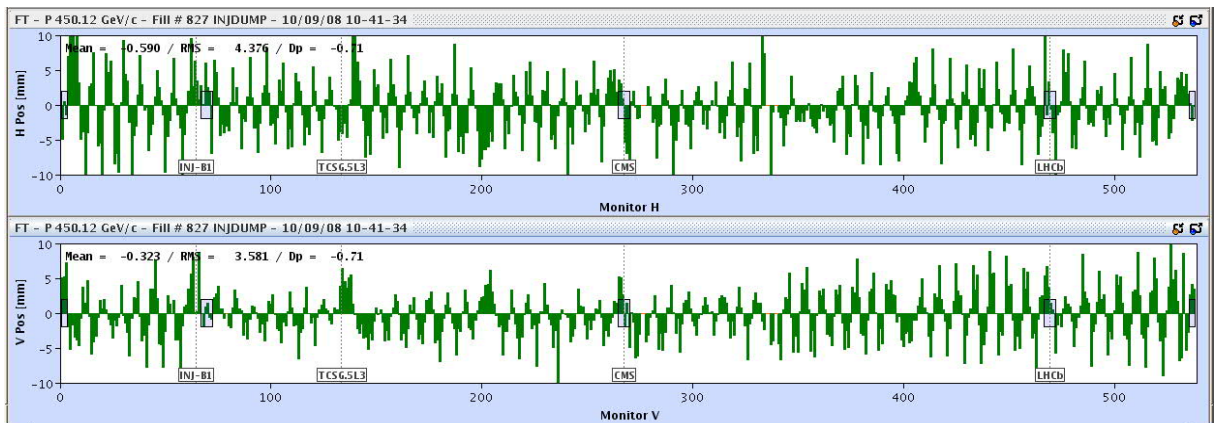


Fig. 1-2. Horizontal and vertical beam trajectories measured during the first LHC commissioning. The measurements on the top (horizontal plane) and bottom (vertical plane) plots were obtained by the standard LHC BPM system. The vertical axis displays the beam position offset in millimetres. The horizontal axis displays the indices of the BPM monitors on the LHC circumference [35].

1.2.2 Betatron coupling and phase advance

In theory of the transverse beam dynamics, each particle undergoes oscillatory-like trajectory around its orbit when passing through the magnet optics. The resulting motion, referred to as betatron oscillations, is defined by the periodically alternating focusing and defocusing magnetic fields in the machine. The frequency of the betatron motion, referred to as the tune, is an observable parameter which is important for the beam stability and accelerator performance. Envelope of the betatron oscillations, a so called betatron function, describes beam size at any point of the circumference. Phase of the betatron oscillations along the circumference, so called phase advance, is an observable parameter which is inversely proportional to the betatron envelope [15].

Measurements of the optics parameters such as the betatron coupling or phase advance are typically based on transverse beam excitations [19]. The natural beam motion has typically very small amplitudes which is changing over time. Therefore such signals are not suitable for precise optics measurements. Exciting the beams with appropriate frequency and amplitude allows reliable measurements of the beam response, necessary for calculating the optics parameters. Principle of such optics measurements is illustratively explained in Fig. 1-3. A dedicated system drives harmonic beam oscillations with amplitude A_{exc} period T_{exc} around the closed orbit. The example depicts the excitation only in the vertical plain. As the bunch passes through the BPM sensor it induces signals on the BPM electrodes. The beam response is then obtained as an amplitude modulation of the BPM signals. Amplitude and frequency of the excitation are set such, that the resulting beam response can be measured by the BPM electronics. The minimum amplitude is usually limited by the required signal-to-noise ratio (SNR) of the measured signals. The maximum excitation amplitude is often limited by the requirements for the optimal and safe operation of the accelerator. The frequency is set around $0.3 f_{rev}$ [3] which is close to the fractional part of the horizontal and vertical tunes [16] of the machine.

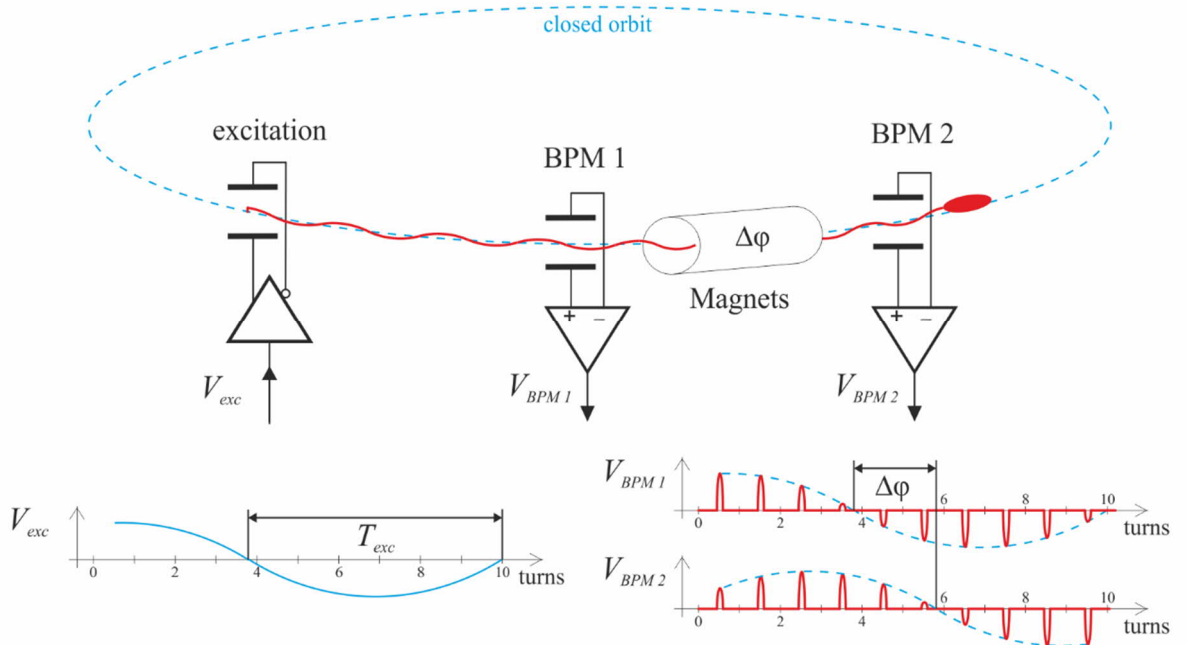


Fig. 1-3. Illustration of excited beam oscillations and the observed response on two BPM sensors. The illustration in the picture is not in scale.

One of the measured beam parameters is the betatron coupling. The beam oscillations can couple between the horizontal and vertical planes due to magnet misalignments and other sources such as the skewed magnets or solenoids [17]. The measurements in the LHC can be performed by exciting the beam at two different frequencies in each plane. The beam response is then measured by a BPM system. The data is used to compute the coupling in each BPM sensor and to apply necessary corrections to the respective magnets. Local corrections of the betatron coupling are important for the beam quality and lifetime [17].

Another measurable LHC parameter is the betatron phase advance, denoted as $\Delta\phi$ in the Fig. 1-3. It can be measured as a phase difference between the measured beam responses obtained at two BPMs in two different locations. The parameter depends on the focusing and defocusing forces of the magnet optics. The most important are the BPMs around the experiments where the optics settings change the most. The strong focusing and defocusing forces in these locations result in large betatron functions [16]. Measurement results can be compared to the numerical models yielding maximum relative deviation of the betatron function, a so called beta-beating. This is an important measure of the quality of the machine setup. The beta-beating also serves as a figure of merit for performance optimisations [18], [19].

The local betatron coupling and phase advance measurements in the LHC are usually using the standard BPM system providing position measurements from many BPM sensors in the LHC. To obtain sufficient SNR the amplitude of the excitation is typically in a millimetre range. The optics measurements are therefore limited to dedicated machine setup and beams typically composed of a few bunches. Duration of these measurements of a few thousand turns is typically limited by the allowed length of the excitation at millimetre amplitudes as well as the size of the memory in the acquisition systems. Experimentally achieved resolution of the phase advance measurements in the LHC is around 1° [20]. The resolution can be improved by using different algorithms, usually based on combining measurements from several BPM sensors [18] [19]. Increasing sensitivity of the position measurements to small beam oscillations can potentially allow to use smaller excitation amplitude, resulting in improved machine safety and quality of the optics measurements. Increasing the acquisition length would allow to compute longer averages and resulting in further improvement of the SNR in the local optics measurements.

1.2.3 Standard LHC BPM system

The standard LHC BPM system is an instrumentation system used to process the BPM signals in the LHC. It was commissioned during the LHC start-up in 2008. The system consists of electronics that equips more than 1000 BPMs in the machine. The signal processing is based on a so called wide-band time normalisation (WBTN) technique [21]. In principle the WBTN converts the individual bunch position into a time difference between two pulses. This operation is realised on analogue front-end cards located close to the BPM sensor in the LHC tunnel. The front-end has processing bandwidth of 70 MHz. The dynamic range of the system is about 35 dB in each of the two possible sensitivity ranges [13]. The bunch position is encoded into a time interval between two laser pulses and transmitted over optical fibres to the acquisition system located on the surface. After receiving the signals from the front-end electronics, the acquisition system converts the time interval into voltage using analogue integration. The resulting voltage is then digitized by a 10-bit self-triggered ADC. The digital data then undergoes further processing in order to obtain position readings.

The system can perform bunch-by-bunch and turn-by-turn acquisitions. The orbit measurements are computed as average positions of all bunches over a few hundred turns. The results are sent at 25 Hz rate to the data concentrators. Position resolution in the orbit mode is around $5\text{ }\mu\text{m}$ [13]. The front-end electronics can be calibrated with 3 different position offsets using test signals generated in the dedicated electronics. The temperature sensitivity of the position measurements is about $20\text{ }\mu\text{m}/^\circ\text{C}$ [13] when applying a software correction algorithm.

1.2.4 LHC collimation system

Uncontrolled particle losses in the machine may damage the magnets and as well as other radiation sensitive equipment. The LHC collimator system is designed to perform beam cleaning by intercepting beam particles in dedicated locations of the machine. The system consists of many collimator devices arranged in several beam cleaning stages with a strict hierarchy. The photograph on the Fig. 1-4 (a) depicts an example of an LHC collimator composed of a pair of 1 metre long moveable jaws inside a tank. The beam enters the collimator through an upstream port and exits through a downstream port, as indicated in the picture with the red arrow. Positioning of the jaws around the beam axis is controlled by dedicated electric motors. As shown in Fig. 1-4 (b), the motors at each side of the collimator allow control of the distance between the jaws. Each jaw position can be set with micrometre resolution. The collimators are designed to withstand high particle losses, high radiation, slow wear-out of the movable parts, and at the same time provide good RF and vacuum properties [22].

Collimator alignment is a process of moving the jaws to a certain distance from the beam. Resulting gap between the jaws is opened for the beam to pass through so that the particles which are outside of the aperture can be absorbed by the collimation surfaces on the jaws. Precise alignment of the jaws is therefore important for proper beam cleaning. A so called Beam Loss Monitor (BLM) based alignment is a method used to align the collimator jaws. It is based on moving each jaw towards the beam until a distinctive particle loss pattern is observed [23]. This happens when the jaw starts to intercept sufficient amount of particles from the beam. Obtained jaw positions are then used as a reference for their further positioning. The method relies on the long-term beam stability and for safety reasons cannot be performed with physics beams. In addition the procedure allows alignment of only a few collimators at a time requiring human supervision which renders the process time consuming. The semi-automated alignment procedure [24] of some 100 moveable collimators requires more than 30 hours [25].

To overcome these limitations the next generation of the LHC collimators features BPM sensors embedded on each side of the jaw [26] [27], as showed in Fig. 1-5. The sensors in a form of retracted button-like electrodes allow to continuously measure the beam position. As a result, both jaws could be aligned using a BPM system without need to create the particle losses as is the case in the BLM-based method. The BPM-based method aims to allow faster and more precise jaw alignments which could be performed even with nominal beam conditions within seconds [28].

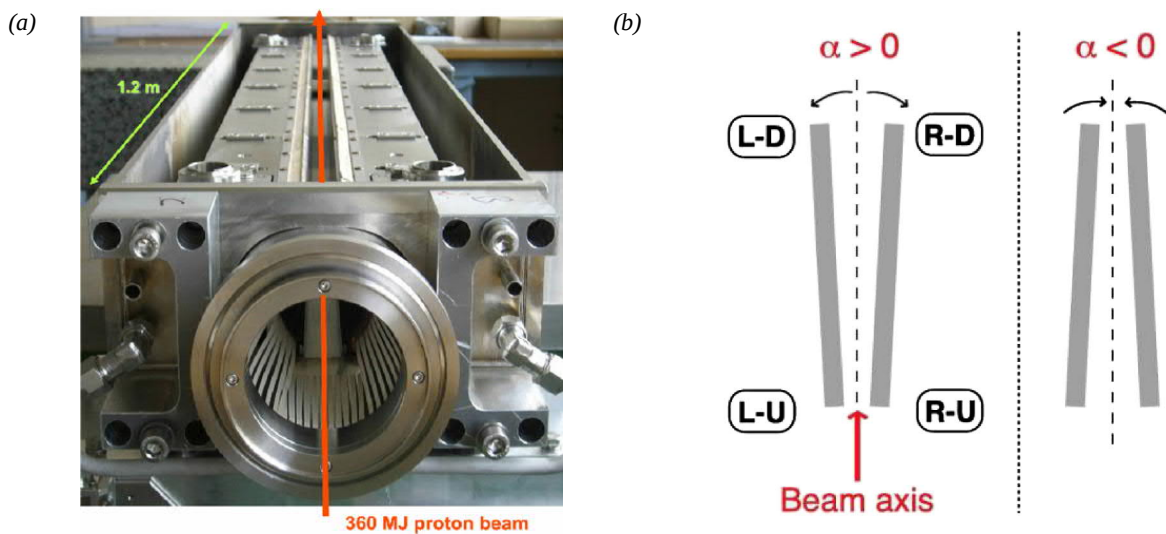


Fig. 1-4. Photograph (a) displays an LHC collimator with two movable jaws enclosed by a 1.2 m long tank. The picture (b) shows an example of the jaw positioning with respect to the beam axis [24]. The red arrow represents a particle beam passing through the collimator.

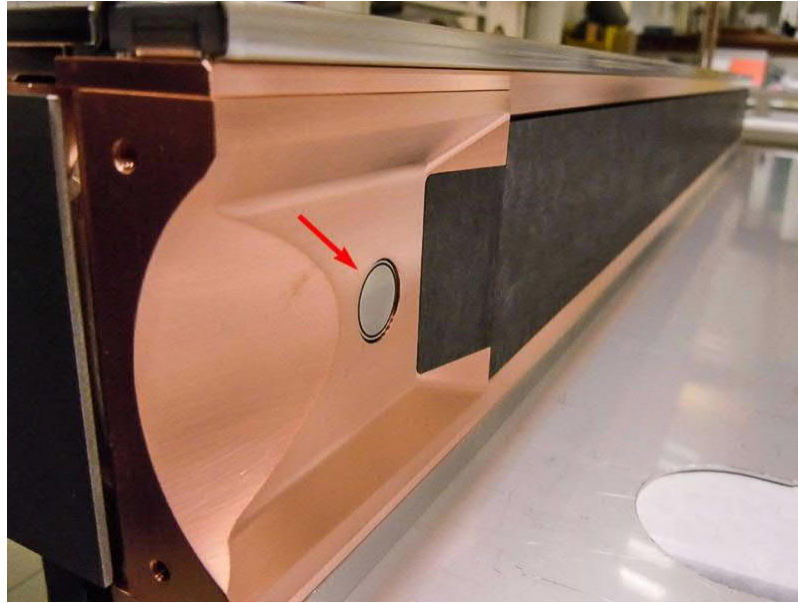


Fig. 1-5. Photograph of a LHC collimator jaw with embedded BPM electrodes. (Courtesy of CERN)

The first prototype of such a collimator was tested in the Super Proton Synchrotron (SPS) accelerator. This special type of BPM application required new type of signal processing system which could meet requirements for precision and robustness. After the prototyping phase some 18 new collimators were built and installed in the LHC during its first long shutdown during 2013 - 2015 [29]. Aligning the collimators with the BPMs has a number of advantages:

- fast and precise collimator alignment within seconds,
- all collimators can be aligned at the same time,
- the alignment can be verified during with all types of beams,
- improved collimation efficiency and performance.

In addition to the collimator alignment, the jaw embedded BPMs can be used for continuous position monitoring during the entire LHC cycle. The position readings from the collimator BPM system are also used by the LHC interlock system. This system can prevent damage to the accelerator by automatically ejecting the beams when the measured orbit values exceed thresholds set for the safe operation [30]. Reliability and robustness of the collimator BPM system are therefore crucial for the operation and safety of the LHC accelerator.

2 Motivations and aims of the thesis

Performance of the beam instrumentation systems has strong impact on achieving high quality of the particle beams and performance of the LHC. The introduction chapter presented multiple fields where improving and optimising the performance of the position measurements can improve the beam diagnostic methods, machine commissioning tools, efficient beam cleaning, interlock protection or feedback systems for real-time corrections.

The main goals of the dissertation work can be summarized in the following points:

- Participation in the design and development of a new beam position measurement system, in particular its front-end electronics. Signal processing in the front-ends is split into two separately optimized subsystems which are dedicated for:
 - Precise and accurate measurements of LHC particle beam orbits based on novel diode detector technique [31], in a so called Diode Orbit subsystem (DOR);
 - Processing of beam oscillations based on dedicated diode detectors, in a so called Diode Oscillation subsystem (DOS), to allow measurements of the local betatron coupling and phase advance with minimal beam excitation.
- Building, testing and installing the new system composed of several front-end units used for operational purposes in the LHC.
- Commissioning, measurement analysis and performance optimisation of the new system in laboratory and with beam signals in the LHC.

The most important application of the Diode Orbit and Oscillation system (DOROS) is to provide precise orbit readings for automatic jaw alignment and interlock operation of the new LHC collimators. Another challenging application of the DOROS system is to provide high-resolution orbit and high-sensitivity oscillation measurements in the important BPMs next to the interaction regions of the LHC experiments. In such locations the new system based on identical front-end hardware can be used to measure various beam parameters, such as the orbit drifts in the interaction points, low frequency orbit oscillations, betatron oscillations, local betatron coupling, phase advance and beta-beating. The results of this work had an important impact on the performance, safety and reliability of the LHC operation.

2.1 Orbit measurement

The most important innovation of the new system is the compensated diode detector technique, to the author's knowledge used for the very first time in beam instrumentation applications. The feasibility of this technique was demonstrated with a prototype that was built and tested in the CERN SPS accelerator [31]. The author took part in the development of the following prototypes. The first version was tested during the years 2011-2013 of the first LHC run and achieved sub-micrometre resolution of the orbit measurements [32]. The final design further improves and optimises the performance of the prototype for operational use in critical applications. The DOR subsystem was therefore designed and optimised for:

- large dynamic range accommodating all LHC beam types,
- sub-micron orbit resolution for small and large beam offsets,
- calibration of the systematic errors of the electronics,
- linear response for every beam offset,
- long-term stability and small systematic errors,
- sufficient measurement bandwidth for observing low frequency oscillations,
- robustness and simplicity of the system,
- compatibility with other instrumentation systems,
- reliable operation even without external timings and prior adjustments.

2.2 Measurements based on the excited beam oscillations

The actual LHC BPM system can be used for reliable measurements of the betatron coupling and beta-beating only with beam excitation in the millimetre range. Such excitation amplitudes are allowed for the LHC only under certain conditions requiring special beams, and dedicated machine settings. Therefore, the optics measurements with physics beams, normally the most important for machine optimisation, were not possible [18] [33].

It became obvious during the early development stages that adding a separate subsystem optimised for turn-by-turn beam oscillation measurements would significantly improve the functionality of the system. In addition the oscillation subsystem does not increase the overall system complexity as it shares most of the hardware resources. The first prototype of the oscillation subsystem was built and tested in laboratory to proof its feasibility. This work has been done as a part of the author's diploma thesis [34] finalized in 2013. It was demonstrated that the oscillation part of the new system can operate with low beam excitation, potentially allowing optics measurements with LHC physics beams. The DOS subsystem was designed and optimised for:

- sensitivity of the subsystem to small beam oscillations,
- synchronisation of the ADC sampling frequency and phase,
- flexibility of the oscillation measurements,
- compatibility with the DOR subsystem.

Achieved performance of the hardware and the digital signal processing was verified with laboratory measurements. Obtained results were confirmed with selected beam measurements obtained from the operational system in the LHC.

3 Diode-based BPM signal processing

3.1 Fundamentals of the beam position measurement

Particles in the LHC beams travel inside the beam pipes emptied to the ultra-high vacuum [3]. The pipes are typically round and metallic, typically made of stainless steel. The beam particles induce image charge of opposite polarity on the pipe walls. The charge density on the pipe surface is defined by the distribution of the particles in a bunch as well as their distance to conductive surface. As the particles propagate at the speed very close to the speed of light the image charge forms a so called image current. The BPM sensors are devices which allow sensing the induced image charge and measure the transverse beam position [36].

The simplest BPM sensors consist of a set of capacitive electrodes. The left picture in Fig. 3-1 shows the transverse cross-section of such a sensor composed of 4 electrodes arranged in the horizontal and vertical pairs. The setup allows measuring the transverse beam position in both axis. The longitudinal cross-section of the sensor is displayed in Fig. 3-1 on the right. The bunch charges are marked in red and the image charges are marked in blue. As seen in the example the transverse position of the bunch is vertically displaced from the centre of the BPM. The position offset causes a difference between the induced charges on the opposing electrodes. The image charge of each bunch is converted on the capacitance of the electrode into a voltage pulse (marked in orange). Difference of the pulse amplitudes A_1 and A_2 from the opposing BPM electrodes is proportional to the transverse position of the bunch. The pulse amplitudes are equal from all electrodes if the bunch was located in the centre of the BPM. The amount of the image charge also depends on the intensity of the bunch. Therefore a sum of the amplitudes A_1 and A_2 yields a quantity which is proportional to the bunch intensity. Depending on the geometry and type of the sensor the pulse amplitudes in the LHC can range between tens to hundreds of volts. Construction of a BPM sensor affects also other parameters like its bandwidth or sensitivity. The most common BPM sensor types used in the LHC are the so called button and stripline BPMs [3].

The electrode signals undergo often quite complex signal processing in order to obtain the normalised position offsets. In general, the normalised position p can be computed on an axis between two opposing electrodes using the corresponding pulse amplitudes A_1 and A_2 as:

$$p = \frac{A_1 - A_2}{A_1 + A_2} \quad (3-1)$$

The equation (3-1) can be used to compute positions in both horizontal and vertical BPM axis. The example showed in the picture would result in a negative p values as the pulse amplitude A_2 would be larger than A_1 .

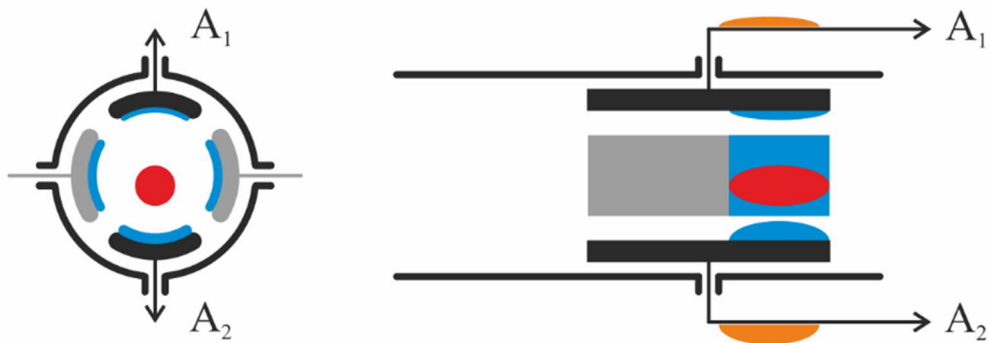


Fig. 3-1. Illustration of a beam position measurement principle. The particle bunch (red) induces image charges (blue) on four electrode plates of a BPM the sensor. The image charge on each electrode is converted into a voltage pulse (orange).

If the beam passes through the centre of the axis the p value would be equal to 0 as both amplitude A_1 and A_2 would be equal. The relationship between the measured normalised position p and real position of the beam in millimetres is described by the geometrical characteristic of the BPM sensor. The characteristics for a button and a stripline BPMs are non-linear. In addition, the normalised positions p measured in the horizontal and vertical axis are mutually dependent. The errors resulting from the cross-plane coupling have to be then corrected in the signal processing.

3.1.1 Button BPM sensor

Majority of the BPM sensors installed in the LHC are of a so called button type [37]. The BPM allows conversion of the image charges into voltage using a button-like capacitive electrodes. The button BPMs are compact, robust and inexpensive with simple mechanical realisation. Schematic drawing of a typical LHC button electrode is showed in Fig. 3-2. The electrode of a 25 mm diameter is located inside the vacuum pipe [37]. A vacuum flange allows its installation onto the beam pipe. An RF feedthrough connects the electrode to a standard RF connector outside of the vacuum. Photograph in Fig. 3-3 shows an installation of an LHC BPM sensor with 4 button electrodes arranged in horizontal and vertical pairs. Such a sensor is typically installed inside the magnet cryostat at temperatures below 25 K and high radiation doses of about 1000 Gy/y [12]. The feedthrough ports and cables have 50 Ω characteristic impedance. The frequency response of a terminated button BPM has a high-pass characteristic with a low-frequency cut-off typically between 100 MHz and 1 GHz. The output pulses from the sensor are partially differentiated as the duration of an LHC bunch is typically smaller than 0.5 ns_{RMS} [3]. The output peak voltage can reach a few tens of volts. The frequency response above 1 GHz is limited by parasitic resonances of the structure [12].

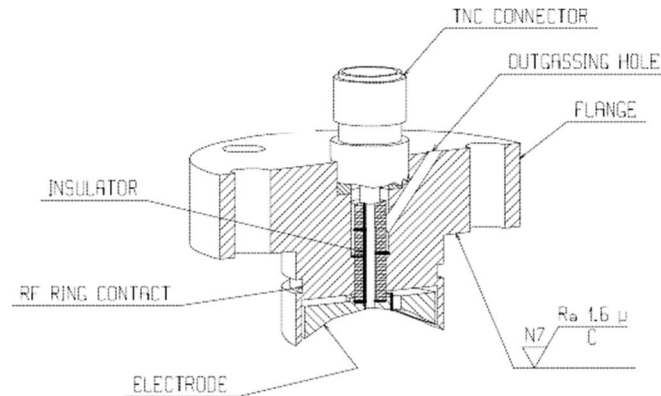


Fig. 3-2. Drawing of a BPM button electrode [37].

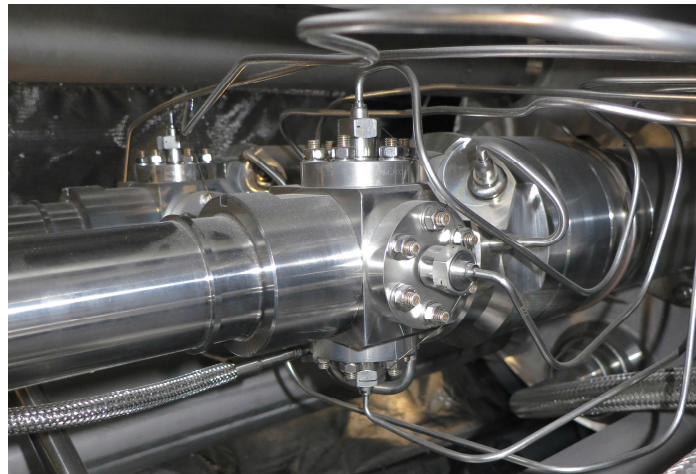


Fig. 3-3. Photograph of a LHC Button sensor. (Courtesy of CERN)

3.1.2 Stripline BPM sensor

Another common type of the BPM sensor in the LHC is a directional coupler, also called stripline. As shown in Fig. 3-4 each electrode is composed of a metallic strip with an RF feedthrough placed on the upstream and downstream sides of each electrode acting as a transmission line. When the bunch enters through the upstream side of the BPM sensor, the induced signals appear on its upstream ports while yielding almost no signals on the downstream ports. The presence of the induced signals on the BPM ports symmetrically exchanges if the beam enters the sensor from the opposite side. One of the important properties of the stripline sensor is its directivity. This feature allows measuring positions of two counter-circulating beams which share a single vacuum pipe. Therefore a typical usage of this type of BPM sensors is in places where the two counter-circulating beams share the same vacuum pipe (e.g. the beam collision areas). The BPM sensors installed in the LHC typically achieve broadband directivity of about 20 dB [3].

The stripline sensors can be also optimised for high sensitivity. Typical amplitudes of the output pulses can be in order of hundreds of volts. A photograph in Fig. 3-5 shows an example of a LHC stripline sensor installed in point 4. When compared to the button BPMs the stripline BPMs are more complex and therefore typically more expensive.

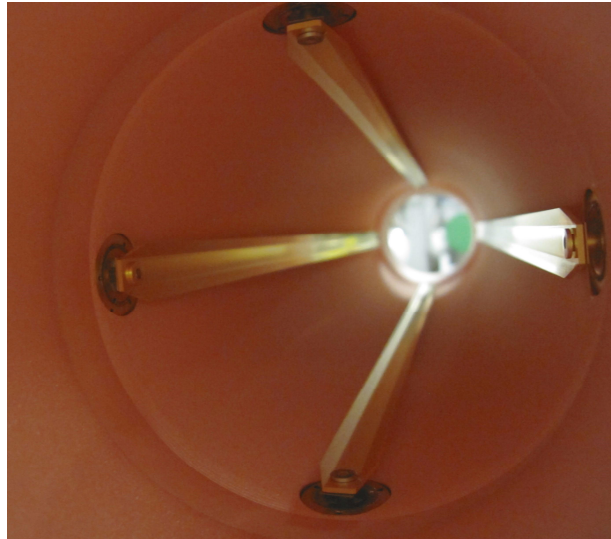


Fig. 3-4. Photograph of stripline electrodes. (Courtesy of CERN)

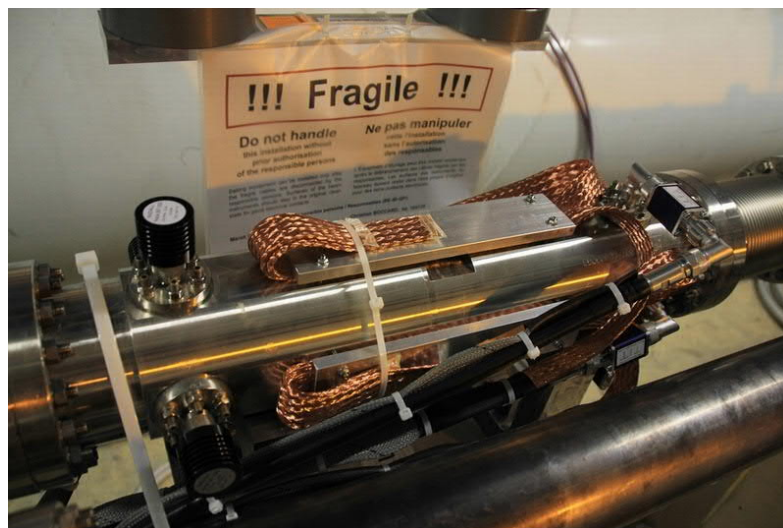


Fig. 3-5. Photograph of a stripline sensor installed in the LHC point 4. (Courtesy of CERN)

3.1.3 BPM position characteristic

Typically the beam diagnostic systems, feedbacks and other applications require the beam position measurements to be expressed in absolute units (e.g. millimetres). The position characteristics of the button and stripline BPMs in the LHC are non-linear functions especially for larger beam offsets from the BPM centre. The BPM characteristics also depend on the beam position offset in the perpendicular axis. If not corrected, this dependency could be a source of additional errors. Therefore a certain correction of the position characteristic needs to be applied which takes into account the geometry of the particular BPM.

As the normalised beam position p_x is a dimensionless number it can be converted to the absolute position Px_{lin} using a linear approximation of a simplified electrostatic BPM model derived in [38]. Index x denotes one of the two axis of a BPM. Index y then denotes the BPM axis in the perpendicular plane. The linear conversion for point-like BPM electrodes can be defined as:

$$Px_{lin} = \frac{D}{4} p_x \quad (3-2)$$

where the parameter D represents the distance between the two electrodes, hereafter referred to as the BPM aperture. It is assumed that the parameter D is equal in both horizontal and vertical planes as the vacuum chamber of the LHC BPMs is typically round. The equation (3-2) does not take into account the beam position offset in the perpendicular axis y . As the parameter D can be expressed in absolute unit the resulting position Px_{lin} is expressed in the same unit. The laboratory measurements presented in this thesis were based on the signal generator and therefore the simple linear approximation (3-2) was used when computing the absolute positions.

In general, the linearization of the position characteristic gives good results for small beam offsets around the centre of the BPM. For larger beam offsets, the BPM non-linearity can be suppressed in real-time processing by using so called correction polynomials. The standard LHC BPM system uses so called 2D cross-term polynomials introduced in 2015 which allow linearization of the BPM characteristic and conversion the normalised position into millimetres. As most of the BPMs consists of horizontal and vertical planes the correction takes both into account. The corrected positions Px_{corr} and Py_{corr} in the in the perpendicular plane can be calculated as [39]:

$$Px_{corr} = c_{50} p_x^5 + c_{30} p_x^3 + c_{10} p_x + c_{32} p_x^3 p_y^2 + c_{14} p_x p_y^4 + c_{12} p_x p_y^2 \quad (3-3)$$

$$Py_{corr} = c_{50} p_y^5 + c_{30} p_y^3 + c_{10} p_y + c_{32} p_y^3 p_x^2 + c_{14} p_y p_x^4 + c_{12} p_y p_x^2 \quad (3-4)$$

where p_x and p_y are the normalised positions in the corresponding planes and the constants $A-F$ are the coefficients of the correction polynomial. Rigorous methods use electromagnetic modelling of specific BPM designs in order to obtain 2D fitting approximations. The coefficient values of the correction polynomials for selected LHC BPMs are summarized in Table 3-1. The values were obtained from electromagnetic simulations [40], [41]. The BPM types listed in the table were used in the beam measurements chapter of this thesis. The polynomial fitting assumed maximum position deviation from the BPM centre of about 40 % of the beam pipe radius. Typically the beam offset does not exceed 25 % of the radius [39]. An example plot in Fig. 3-6 displays the characteristic function of a so called BPMSW stripline with 61 mm aperture which is installed used close to the LHC experiments.

Table 3-1. Correction coefficients for the LHC BPMs consisting of horizontal and vertical plane [39].

Type	Category	Aperture D [mm]	C_{50}	C_{30}	C_{10}	C_{32}	C_{14}	C_{12}
Stripline	BPMSW	61	5.1631	3.0715	17.1037	8.8285	5.3447	1.7684
Button	BPM	49	3.0268	3.4651	13.0207	-1.8265	2.0741	0.5032

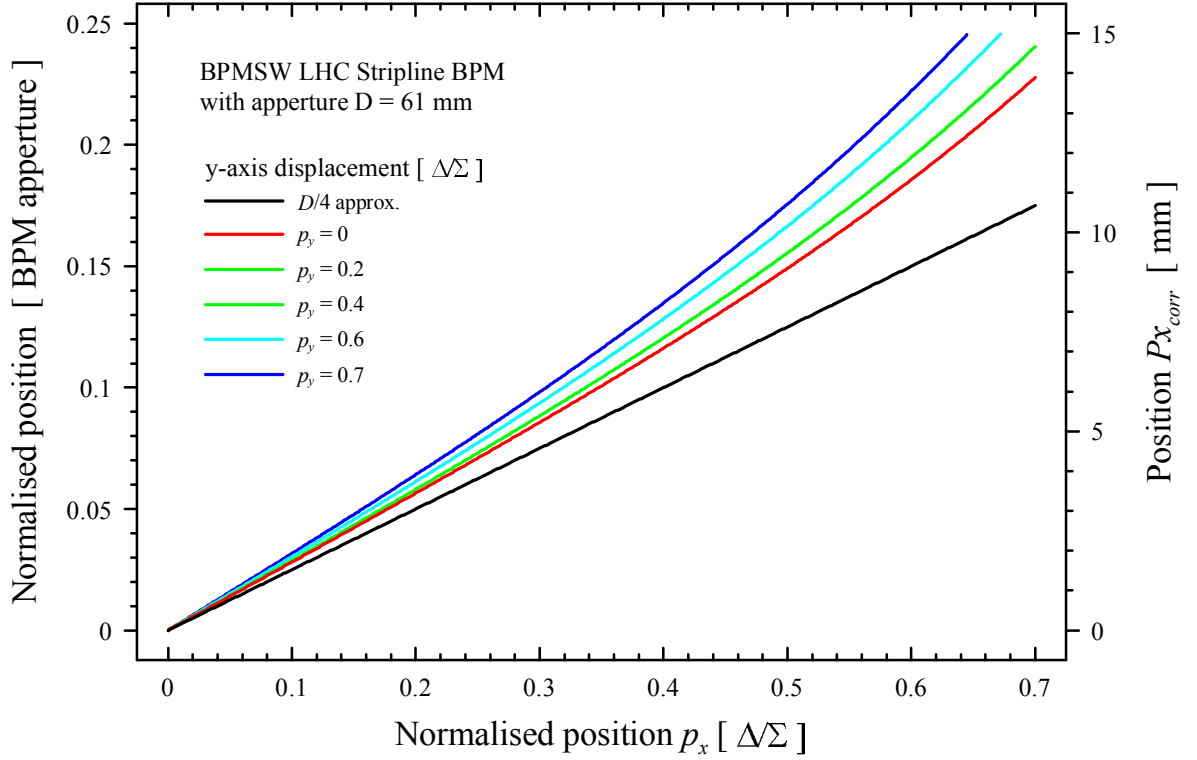


Fig. 3-6. Example plot of the correction function used to convert normalised positions into linearized positions in millimetres for the displacements in both p_x and p_y BPM axis. The coefficients of the polynomials correspond to a BPMSW stripline BPM [39] which can be found in the important locations close to the LHC experiments.

The plot displays the position characteristics described by the polynomial function (3-3) for a number of position offsets in the perpendicular axis. The trace plotted in black represents the conversion characteristic approximated by the linear equation (3-2) using the same 61 mm aperture. Please note that the absolute accuracy of the position measurements based on the correction polynomials relies on the mechanical tolerances of the BPM components.

In case of the collimator BPMs the aperture changes with the collimator gap setting. The geometry of the structure cannot be approximated by a cylinder and therefore the equations (3-3) and (3-4) cannot be applied. In addition the beam position can be directly measured only on the axis between the jaws. A generalized correction polynomial of degree $m-n$ for the collimator BPMs can be expressed as:

$$Px_{coll} = \sum_{a,b=0}^{m,n} c_{a,b} p_x^a D^b \quad (3-5)$$

where p_x is the normalised position on the x axis, D is in this case the distance between the collimator BPM electrodes and c the polynomial coefficients with indices a and b [30]. The parameter D is provided by jaw position readout system [43]. The accuracy of the measurement is better than 20 μm [44].

3.2 BPM signal processing with RF diode detectors

Diode detectors can be used to convert amplitudes of the BPM electrode pulses into slowly varying signals. A schematic diagram of a diode detector is shown in Fig. 3-7 (a). Principle of its operation is demonstrated in Fig. 3-8 (a) on two different input signals consisting of voltage pulses with a repetition period T and amplitudes V_{i1} and V_{i2} . The pulse amplitudes are proportional to the bunch intensity and transverse position. In the LHC the repetition period of a single bunch is approximately $89 \mu\text{s}$ [3]. The response signals on the output of the diode detector is illustrated in Fig. 3-8 (b).

Operation of the detector circuit can be analyzed using a static diode model approximated with a constant forward resistance r and a constant forward voltage V_d . The corresponding voltage-to-current characteristic of a diode is then showed in Fig. 3-7 (b). The diode D conducts when the input voltage V_i is higher than the output voltage V_o seen across the capacitor C and resistor R . Then the voltage difference seen across the diode is more than V_d . The capacitor charging rate is limited by forward resistance r of the diode.

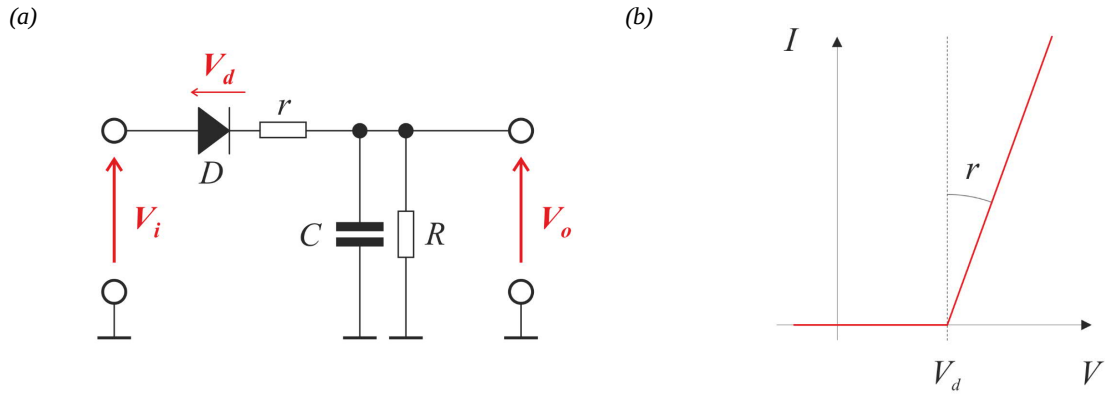


Fig. 3-7. Schematic diagram of a simple RF diode detector showed in the picture (a). Resistor r indicates the dynamic resistance of the diode. Voltage-current characteristic of an ideal diode with a series resistor r is showed in the plot (b).

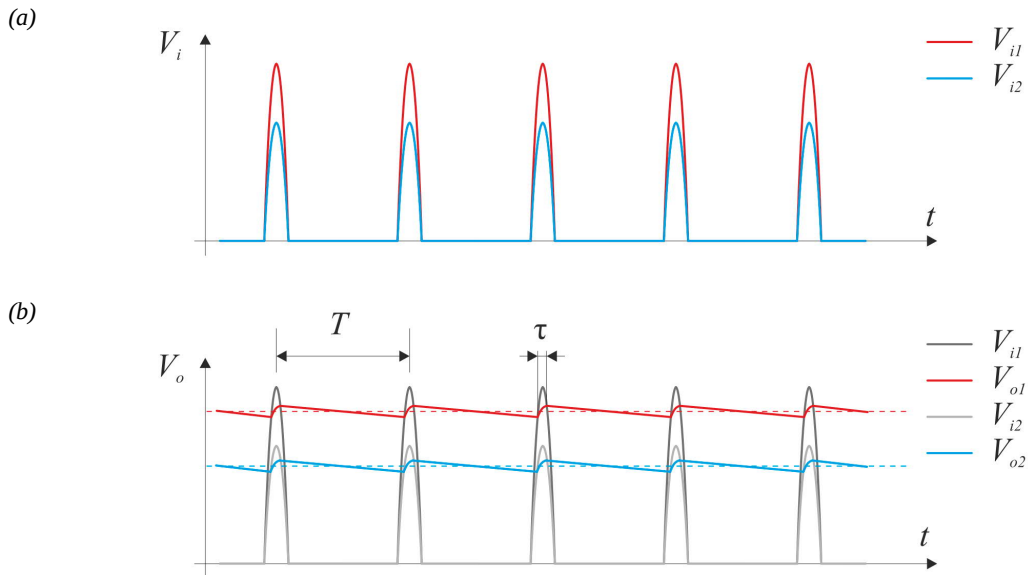


Fig. 3-8. Example illustration of the electrode signals with two amplitudes V_{i1} and V_{i2} applied on the input of the RF diode detector (a). Response of the diode detector in the steady state is shown in the (b) plot. Average voltage is indicated with dashed lines.

In the following analysis it is assumed that the voltage change on the output is small with respect to the input voltage. The latter implies that the charging and discharging currents are constant. The charge ΔQ_τ flowing through the forward conducting diode during time τ can be then expressed as:

$$\Delta Q_\tau = \frac{V_i - V_o - V_d}{r} \tau \quad \text{for } V_i > (V_o + V_d) \quad (3-6)$$

The diode stops conducting once the voltage on the anode drops below the output voltage V_o . The capacitor C is discharged through the load resistor R and the reverse current of the diode. As the reverse current is typically very small (in the order of 10 nA for a typical Schottky diode) it is assumed to be negligible in this analysis. Total charge ΔQ_T which is approximated by a constant discharge current during period T can be expressed as:

$$\Delta Q_T = \frac{V_o}{R} T \quad \text{for } V_i < (V_o + V_d) \quad (3-7)$$

The charging and discharging of the capacitor in the detector is balanced when the signals are in steady state which can be expressed as $\Delta Q_T = \Delta Q_\tau$. The charges can be substituted by the results from (3-6) and (3-7). The resulting equation was rearranged assuming that the input voltage is much larger than the forward voltage drop of the diode. Obtained ratio of the output to input voltages is then expressed as:

$$\frac{V_o}{V_i} = \frac{1}{1 + \frac{rT}{R\tau}} \quad \text{for } V_i \gg V_d \quad (3-8)$$

The voltage ratio depends on the term in the denominator which is composed of the charging and discharging times and respective resistances. The T/τ becomes large when there is a single bunch in the accelerator. The ratio decreases with increasing number of bunches in the accelerator due to the increasing repetition rate. To achieve optimal output voltage $V_o \approx V_i$ the ratio r/R has to be much smaller than T/τ . As seen in the example illustration in Fig. 3-8 (b), the response signals V_{o1} and V_{o2} on the detector output are proportional to the peak amplitude of the respective input pulses V_{i1} and V_{i2} .

In a practical example the values for the parameters for a single bunch can be approximated with $T = 89 \mu\text{s}$, $\tau = 1 \text{ ns}$ and $r = 100 \Omega$. Consequently, for the best output voltage level the R has to be larger than r . In this example the resistance R should be larger than $8.9 \text{ M}\Omega$. The diode detector has to be therefore followed by a stage with a high impedance input. This can be achieved using amplifiers with JFET inputs.

In order to achieve the slowest discharge rate of the detector, the resistor R can be omitted. The capacitor C is then discharged through the reverse current of the diode and the circuit then operates as a peak-detector. Ripple in the output voltage can be expressed as a change ΔV during the capacitor discharge using exponential decay:

$$\Delta V = V_{peak} e^{-\frac{T}{RC}} \quad (3-9)$$

The equation can be simplified when assuming that $T \ll RC$. In practice a typical time constant for a single LHC bunch can be in order of 1 ms. For the physics beam the longest discharge time can be in the order of $1 \mu\text{s}$. The exponential function can be then approximated by a linear decay and the equation (3-9) can be approximated for large peak voltages as:

$$\Delta V \approx V_{peak} \frac{T}{RC} \quad \text{for } T \ll RC \quad (3-10)$$

Two example plots shown in Fig. 3-9 (a) and Fig. 3-9 (b) illustrate the effect of different values of the discharge resistor R on the output signals. Displayed plots allow comparing the response of the diode detector (in red, blue and green) for two pulse repetition periods T_1 and T_2 . The dashed lines indicate the average signal level at the detector output. The plot Fig. 3-9 (a) shows the detector responses for the input pulses (in grey) with repetition period T_1 . Large resistor value R_1 causes slow capacitor discharge such that the output signal in the steady state is found close to the input peak voltage. Decreasing the resistor to value R_2 or R_3 causes an increase in the capacitor discharge rate as well as the output voltage ripple. As a consequence the average output voltage is decreased. The plot Fig. 3-9 (b) shows the detector responses for the input pulses with repetition period T_2 which equals $2T_1$. The example also depicts a transition T_1 into the new steady state with T_2 . The average voltage on the detector output decreases with increased discharge time. The voltage ripple around the average value is increased in accordance with the equation (3-10). The plot also displays the detector signals with smaller resistor values R_1 and R_2 where $R_1 > R_2$. Using a fast discharge rate the output signal may even drop close to zero when the repetition rate of the BPM pulses is too low. In such cases the detector output depends on the pulse duty cycle. This effect also depends on the pulse amplitudes. An example of such scenario could be found with the LHC pilot beams with the maximal repetition period equal to T_{rev} . As can be seen in the plot, the average signal (dashed line) does not drop to zero but stays at some constant level. The dependency on the duty cycle may dominate in the signal which can lead to large errors in the measurements. Such effect can be avoided by using slow detector discharge when measuring low number of bunches grouped on the circumference. The detector is then operated in a so called peak detection mode.

Examples of the diode detector signals with two different bunch patterns are illustrated in Fig. 3-10 (a) and Fig. 3-10 (b). The plot (a) shows an example of a typical distribution of bunches with equal pulse amplitudes. In this scenario the diode is in the conducting state for all bunches in the machine. However, when the amplitude distribution in the pattern becomes asymmetric the detector will measure mostly bunches which produce dominant pulse amplitudes. If the discharge rate of the detector is slow (e.g. the R_1 is large) the remaining bunches producing smaller amplitudes will be ignored and thus will not contribute to the output signal. Such effect can be avoided by using fast detector discharge when measuring BPM signals with important asymmetry in the amplitude distribution. The detector is then operated in a so called averaging mode.

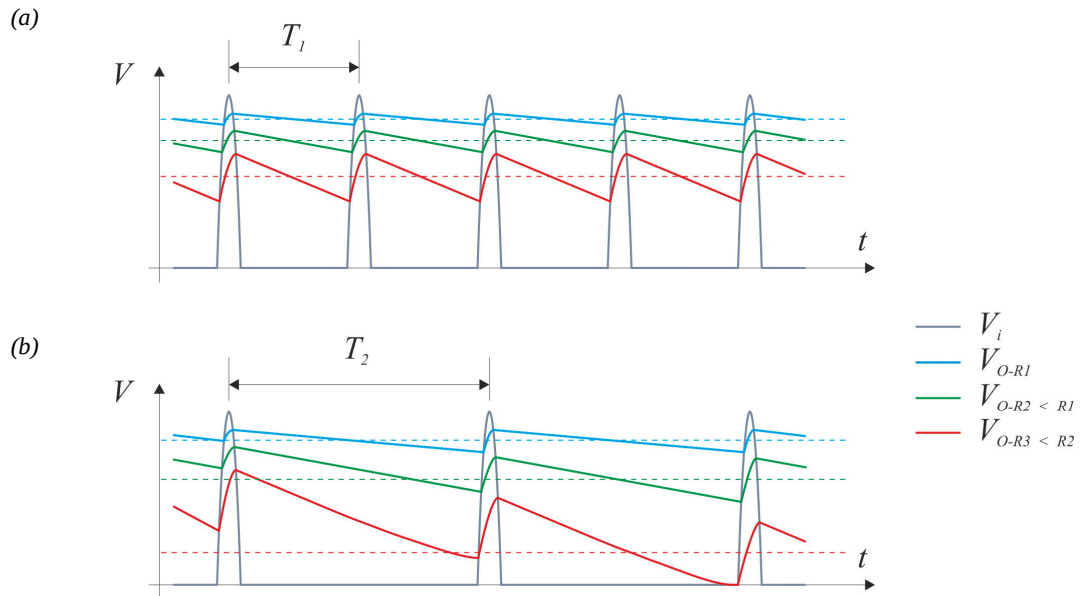


Fig. 3-9. Example illustrations of the diode detector output signals displayed for two bunch repetition periods T_1 and T_2 and different discharge resistors R_1 , R_2 and R_3 . The pulse amplitude is constant in both examples. The relationship between the resistor values is $R_1 > R_2 > R_3$. Average voltage in the steady state is indicated with dashed lines.

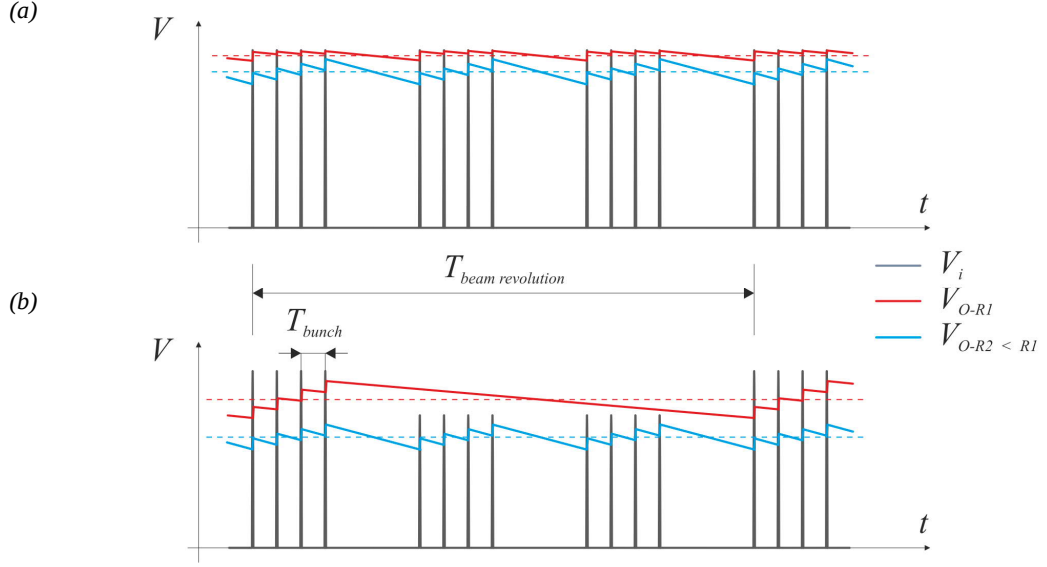


Fig. 3-10. Example of the diode detector signals for different amplitude distribution in the bunch pattern. Bunches repeat every beam revolution T_r with T_{bunch} spacing between them. The plot on the top displays 16 pulses with equal amplitudes. The plot on the bottom illustrates pattern where the amplitudes of 8 pulses in the middle is decreased. Output signals are displayed for discharge resistors R_1 and R_2 . The relationship between the resistors is $R_1 > R_2$. Average voltage in the steady state is indicated with dashed lines.

3.2.1 Beam position measurement

The diode detectors can be used to process electrode signals from a BPM in order to measure the transverse position and oscillations of the LHC beams. Schematic diagram in Fig. 3-11 illustrates a setup of the diode detectors used to process BPM signals V_{i1} and V_{i2} from the two opposing electrodes. The resistors R_T provide impedance matching on the detector inputs. Each electrode signal is then processed by a dedicated diode detector producing output signals V_{o1} and V_{o2} . The voltage drop across the diode is denoted with V_d . As indicated in the picture the two diode detectors are assumed to be identical.

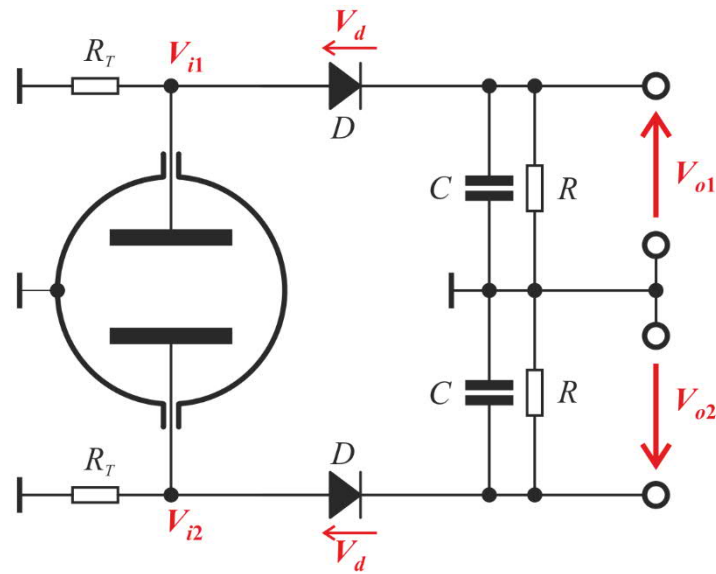


Fig. 3-11. Example schematic of a diode detector setup used to process V_{i1} and V_{i2} signals from two BPM electrodes.

The assumed normalised beam position p_b can be derived from the peak amplitudes of the input signals, as expressed in the equation (3-11).

$$p_b = \frac{V_{i1} - V_{i2}}{V_{i1} + V_{i2}} \quad (3-11)$$

Similarly, the normalised measured position p_m is derived from the signals V_{o1} and V_{o2} on the output of the detectors as:

$$p_m = \frac{V_{o1} - V_{o2}}{V_{o1} + V_{o2}} \quad (3-12)$$

The output signals V_{o1} and V_{o2} are lower than the input signals by the voltage drop V_d across the diode. In the linearized analysis based on a static diode model, the output signals can be expressed as:

$$V_{o1} = V_{i1} - V_d \quad V_{o2} = V_{i2} - V_d \quad (3-13)$$

Relationship between the ideal and measured position can be found by inserting the set of the output voltage equations (3-13) into the equation (3-11). The derived equivalence can be then expressed as:

$$p_m = \frac{V_{o1} - V_{o2}}{V_{o1} + V_{o2}} = \frac{V_{i1} - V_{i2}}{V_{i1} + V_{i2} - 2V_d} \quad (3-14)$$

The diode voltage V_d in the detector signals is cancelled in the nominator. The difference of the two amplitudes is proportional to the position of the beam. Sum of the signals in the denominator does not cancel the V_d which introduces a systematic error in the position measurement. The relative position error p_{err} as measured by the detectors can be then expressed as:

$$p_{err} = \frac{2V_d}{V_{i1} + V_{i2} - 2V_d} \quad (3-15)$$

In practical example of V_{i1} and V_{i2} equal to 5 V and V_d equal to 0.2 V the position error p_{err} is about 5 % which is more than 0.8 mm when projected on a 61 mm BPM aperture. In addition, the parameter V_d depends on the diode current, temperature and production spread. Typical change of the V_d with temperature can be about 2 mV/°C which would translate in this example into a 1 % position error. These limiting factors were addressed by the V_d compensation scheme described in the following chapter.

The simple RF diode detectors can be however designed and optimised for achieving high sensitivity to the relative position change such as the betatron oscillations [45]. Coherent transverse oscillations of the beam appear as a small change in the beam position which can be seen on the BPM electrode signals as an amplitude modulation. The frequencies of the betatron oscillations repeat in the BPM signal spectrum inside the sidebands of every revolution frequency harmonic. Schematic diagram of the oscillation signal processing based on the RF diode detector is illustrated in Fig. 3-12 along with the example signals in important nodes. In the LHC the amplitude of the BPM signals can reach more than 100 V with an amplitude change typically smaller by a few orders of magnitude. Diode detectors can be designed to process such voltages when built with several diodes in series. Schottky RF diodes are suitable for such applications for their fast switching times, low input capacitance and low voltage drop V_d . The detectors then down-convert the small oscillations into the base-band. An important part of the energy of the betatron harmonics appears in the base-band which results in a large betatron signal gain. In order to achieve high sensitivity, the dominating DC part is removed on the series capacitors. The oscillations are then measured after the following operational amplifier and filtering stages. The band-pass filter smooths the charging and discharging of the detector capacitors, attenuates the low frequency background components as well as the revolution frequency components which could potentially saturate the following stages.

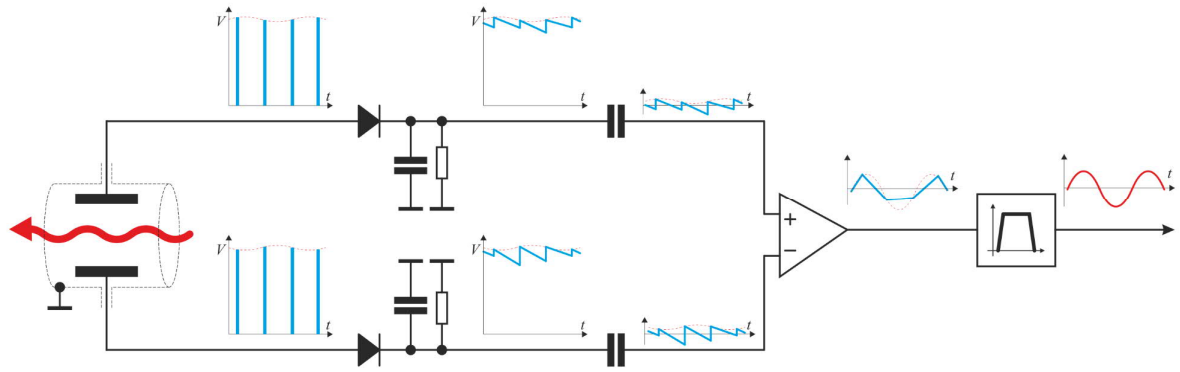


Fig. 3-12. Schematic diagram of an oscillation signal processing based on a diode detector principle. The diagram displays the example signals in the important nodes [46].

Described oscillation signal processing based on the RF detectors was implemented in the LHC base band tune system (BBQ). The unprecedented sensitivity of the system allows to measure frequency of the natural beam oscillations with amplitudes in the order of nanometres and frequency resolution in the order of 10^{-5} [46]. The RF diode detection technique has number of advantages when measuring relative position changes:

- large dynamic range,
- wideband signal processing,
- sensitivity to very small beam signals,
- no external triggering required,
- no biasing required,
- simplicity and robustness of the circuit.

3.3 Compensated diode detectors

The RF diode detectors can be designed to process BPM signals and provide output signals which are proportional to the beam position. As derived in (3-15), using a simple diode detector for orbit measurement would however result in a systematic errors that depend on the diode forward voltage V_d . This voltage drop can be compensated using a circuit similar to the classic diode detector described in a popular electronics book [48]. The circuit showed in Fig. 3-13 is composed of two RF diode detectors connected to the same input signal V_i . A so called 1D diode detector branch is built with a single diode D_1 . Its output voltage V_1 is thus smaller by the diode voltage V_d as indicated in the diagram. A so called 2D diode detector branch has two diodes D_{2a} and D_{2b} in series. Its output voltage V_2 is smaller by $2V_d$. The output signals from the two sub-detectors are processed by two operational amplifiers OA_1 and OA_2 . Voltage drop $2V_d$ is derived from the V_1 by the amplifier OA_1 . It is then added to the output voltage of the amplifier OA_2 which functions as a follower for the V_2 . Output voltage V_o can thus be expressed as:

$$V_o = \left(1 + \frac{R_{oa1}}{R_{oa2}}\right) V_1 - \frac{R_{oa1}}{R_{oa2}} V_2 \quad (3-16)$$

The detector voltages V_1 and V_2 can be approximated as:

$$V_1 = V_i - V_d \quad V_2 = V_i - 2V_d \quad (3-17)$$

The rearranged formula for the output voltage is obtained by inserting the equations (3-17) into (3-16) which can be expressed as:

$$V_o = \left[\left(1 + \frac{R_{oa1}}{R_{oa2}}\right) V_i - \frac{R_{oa1}}{R_{oa2}} V_i \right] - \left[\left(1 + \frac{R_{oa1}}{R_{oa2}}\right) V_d - \frac{R_{oa1}}{R_{oa2}} 2V_d \right] \quad (3-18)$$

As seen in the equation (3-18) the voltage drop is compensated when the amplifier gain is equal to the voltage drop ratio of the 1D and 2D detectors. The resulting output voltage is equal to the input voltage when the feedback resistors are equal, i.e. $R_{oa1} = R_{oa2}$. This is expressed as:

$$V_o = [2V_i - V_i] - [2V_d - 2V_d] = V_i \quad (3-19)$$

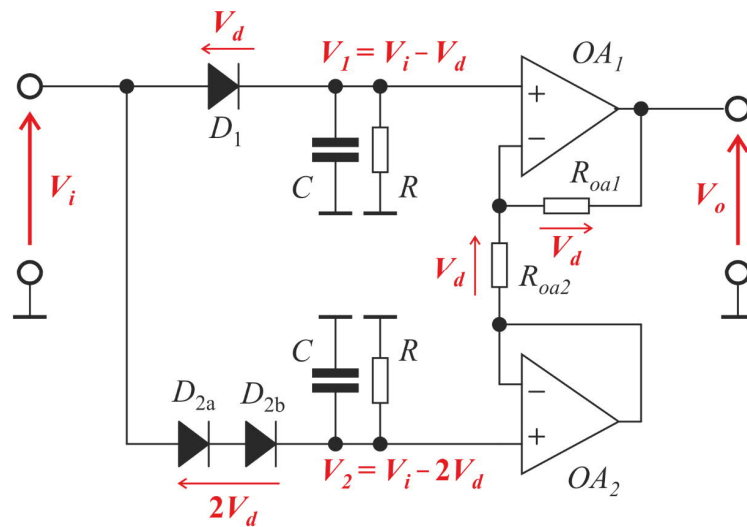


Fig. 3-13. Schematic of a compensated diode detector. Voltage drop over forward biased diode D_1 is compensated by a parallel circuit containing two diodes D_{2a} and D_{2b} . Their voltage is subtracted by an operational amplifier resulting in the output voltage V_o which is equal to the input voltage V_i .

For the compensation scheme to work the diode in the circuit have to have identical characteristics and working conditions. Differences of component spread can be minimized by using three diodes in a single package. The impact of the thermal drifts is also minimized this way. The amplifier feedback resistors R_{oa1} and R_{oa2} should be selected with precision better than the symmetry between the diodes in the circuit. As shown in the Fig. 3-13 the compensation scheme is built with the same R and C component values in both detector branches. Difference in the diode voltage drop between the two branches results in different charging and discharging currents between the two sub-detectors. Example of the input-output voltage characteristic of the compensated diode detector is showed in Fig. 3-14. The plot displays results from the laboratory measurement using $1\text{ M}\Omega$ and $10\text{ M}\Omega$ discharge resistors. As seen in the plot the asymmetry becomes less important when the amplitudes of the input pulses are sufficiently large. The non-linearity of the detector characteristic is also more important when the input pulses are below 0.2 V . The dynamic range of the input signal has to be therefore controlled within certain limits for the best measurement results. The dynamic range can be controlled by a set of amplifiers placed in front of the diode detectors. The slowest time constant of the detector can be achieved by omitting the discharge resistors. The capacitors are then discharged through the reverse currents of the diodes and the leakage currents of the amplifier inputs. In this mode of operation the circuit acts as a peak detector with the longest time constant.

The compensated diode detectors have a number of advantages. Carefully optimized design can be used to build BPM electronics providing beam orbits measurements with sub-micrometre resolution and wide dynamic range. In addition, the technique allows measuring the peak voltages of each BPM electrode separately and compute the beam positions in the digital domain. Digital signal filtering and calibration techniques can further improve the precision and accuracy of the position measurement. The simple construction of the circuit allows to build a robust BPM system with guaranteed operation even without external timing. The technique is thus suitable for safety critical application requiring high position resolution measured over multiple turns.

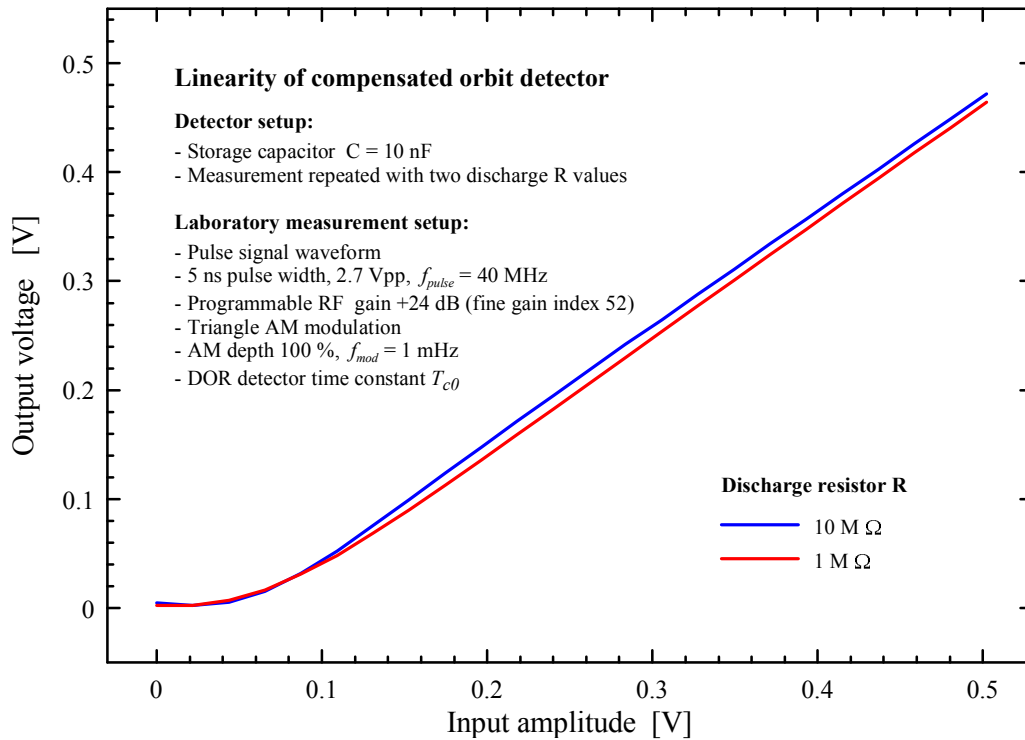


Fig. 3-14. Input-output voltage characteristic of a compensated diode detector zoomed to a 0.5 V amplitude of the input signal. The characteristic was measured with constant capacitor value 10 nF and discharge resistor values of $1\text{ M}\Omega$ and $10\text{ M}\Omega$.

4 Architecture

This chapter describes the architecture of a front-end used in the present LHC DOROS systems. The development history described in the first subchapter marks the main milestones of the project and summarizes its evolution towards the present state. The following subchapters show the overview of the DOROS system followed by detailed description of its most important hardware and software parts.

4.1 History of the development

The idea of using diode detectors in the LHC BPM signal processing emerged at CERN in 2004 during a design of a new system dedicated to measuring the LHC betatron tunes [47]. At that time it was considered to use a well-established method based on a phase-locked loop principle [3], [49], [50]. The PLL technique is based on exciting coherent beam oscillations and measuring the beam response using a BPM system. Amplitude of the excitation had to match the sensitivity of the BPM system and be at the same time compatible with safe operation of the machine and required beam quality. Investigation of the diode detection technique begun after a mistake in the schematic of an input amplifier [47]. It was discovered that a simple diode detector can achieve high sensitivity over wide dynamic range. The LHC tune system emerged as a result of this research. The system was optimised to measure the LHC tunes with precision beyond 1 nm even without driven excitation of the beam [46]. Due to its unprecedented sensitivity, the system became an important tool for machine operation and beam diagnostics. Similar diode-based systems were deployed to measure tunes in other accelerators, also outside of CERN. It was soon realised that the qualities of the simple diode detector could be potentially extended to measure the beam orbits. First research and development project of a new BPM system based on diode detectors began as a part of the LHC collimator project [51]. The electronics and signal processing had to be optimized for high precision measurements using the BPM electrodes embedded in the collimator jaws. The primary goal was to allow automatic alignment of the jaws with beam using the BPM measurements with precision and stability of the orbit measurement at a micrometre level [52]. The compensation scheme in the diode detector offered a suitable solution especially for the centred beams. In such case the residual non-linearity of the orbit measurement have almost no influence on the precision of the beam alignment. In addition the errors arising from the asymmetries between the analogue channels could be calibrated using the multiplexing techniques [53]. Studying the circuit using SPICE simulations at this point of the development was not enough for the required level of precision. The demonstrator of the compensated diode detection principle was built in 2009. A photograph in Fig. 4-1 shows the first prototype built as a modification of the front-end hardware from the LHC tune system.

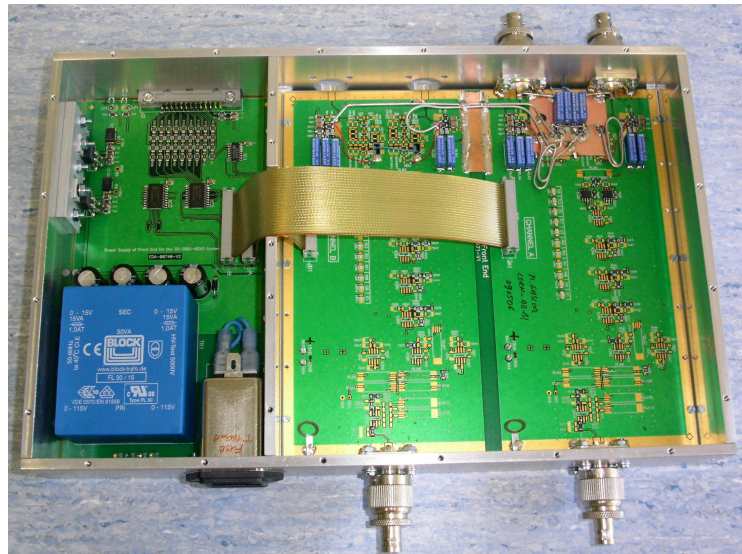


Fig. 4-1. Photograph of the first prototype of the compensated diode detector. The circuit was built as a modification of the front-end hardware from the LHC tune system.

The prototype was equipped with two compensated diode detectors dedicated to the orbit measurements. Each diode orbit (DOR) detector was processing signals from one BPM electrode. The circuit parameters in the prototype were controlled by reed relays. The output voltages from the detectors were measured with a laboratory voltmeter. The prototype provided a proof-of-principle based on laboratory measurements. Promising results started an extensive study aiming to further exploit the potential of the diode technique for precision orbit measurements. The second version of the prototype was built in 2010. The front-end was realised in a 19" 1U rack chassis. The concept was used from another safety critical system designed for detecting beam presence in the machine [54]. The front-end prototype contained the analogue electronics with 4 DOR channels and a power supply. Inputs of the analogue channels were equipped with non-reflective RF filters providing galvanic insulation. Each signal was then amplified by a current feedback RF amplifier followed by a compensated DOR detector stage. Output voltages from each channel were measured with laboratory voltmeter Keithley 2701 with controlled input multiplexer. The setup was evaluated in the laboratory and was used in a few beam measurements obtained with a collimator test setup in the SPS [31]. The collimator BPMs were connected to the prototype over longer coaxial cables. The front-end could be then placed in a radiation-safe area preventing radiation damage to the electronics. Requirements for the radiation tolerance of the components were thus not strict which allowed faster development.

The analogue processing however needed an acquisition and control system. A prototype of the first acquisition system was designed and built in 2011 by the author of this thesis. Photograph in Fig. 4-2 shows the prototype equipped with the DOR signal processing and the acquisition system under development at that time. The dynamic range of the analogue processing was improved by increasing the number of configurable RF amplifier stages. The gain value was set by switching the feedback resistors in each stage using GaAs switches. Additional analogue filters after the diode detectors limited the orbit bandwidth to less than 1 Hz. The data acquisition was built with a 24-bit sigma-delta ADC which provided simultaneous sampling of eight DOR channels at ~11 kHz rate. A simple 32-bit microcontroller provided basic digital signal processing, data streaming, and remote control over Ethernet network.

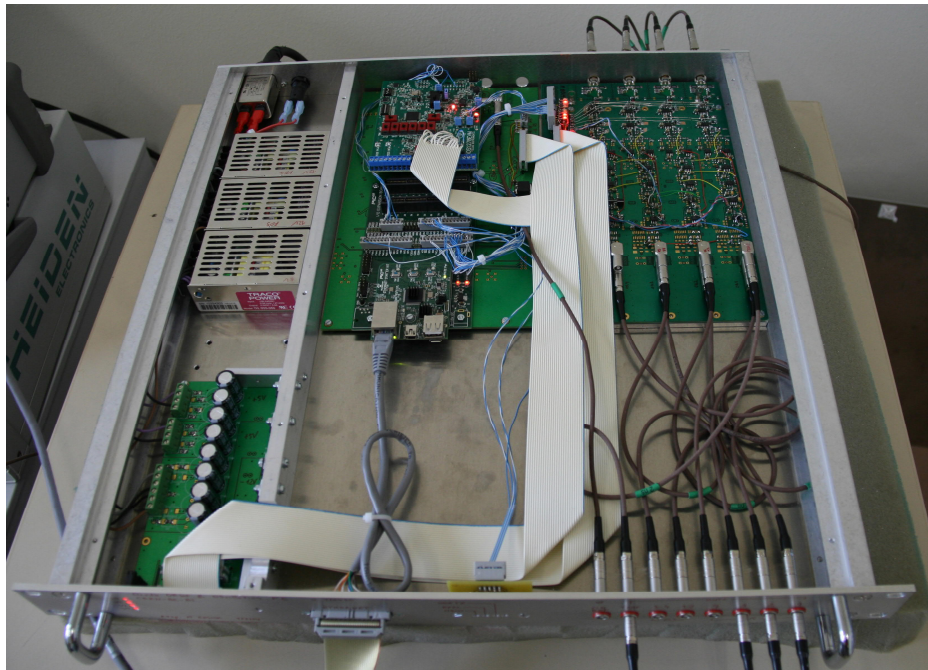


Fig. 4-2. Photograph of the prototype front-end equipped with an analogue processing card and an acquisition system. The analogue board was equipped with 4 orbit channels. Prototype of the acquisition system consisted of an 8-channel 24-bit ADC and a 32-bit microcontroller.

More front-end prototypes built in 2011-2013 were installed on the collimator setup in the SPS as well as a number of selected BPMs in the LHC. The prototypes allowed to measure and analyse beam data for more than two years during the LHC run 1 [55], yielding promising results [56].

At the same time a possibility to measure also the beam oscillations was studied as a part of the author's master thesis [34]. It was realised, that the RF processing stage could be exploited to supply its signals to a second subsystem based on the diode detector optimised for the beam oscillation measurements [57]. The so called Diode Oscillation (DOS) processing was designed to detect small amplitude modulations of the BPM signals in frequency range from some 500 Hz up to approximately 5 kHz. The analogue processing in the laboratory prototype was built with four DOR channels optimized for orbit measurements and two DOS channels optimized for the oscillation measurements. Laboratory measurements demonstrated that a simple DOS processing can achieve excellent sensitivity without affecting the performance of the DOR processing with minimal additional cost and complexity [34]. The oscillation channels became an interesting feature which could allow to measure local beam coupling, beta-beating and phase advance. The system was named after its two main diode detector subsystems and is thus hereafter referred to as the DOROS.

Development of the operational DOROS hardware started in 2013. The analogue processing as well as the signal acquisition and processing were re-designed and optimised for the operational use. It was foreseen to use the DOROS front-ends to equip BPMs of some 18 LHC collimators in the beginning of 2015 after the long LHC shutdown. Design of the operational acquisition system was based on an Altera Cyclone IV FPGA [58]. The platform was chosen based on a good experience from a similar project developed for the Beam Loss Monitoring (BLM) system [59]. The firmware was used as a basis for implementing the Ethernet communication. Prototype of the first FPGA controller board is showed in Fig. 4-3. It was used for firmware development and prototyping. This platform was used to develop a dedicated printed circuit board (PCB) which would feature the required interfaces and dedicated control circuits. The first version of the FPGA board was fully operational and could be used for the first tests with beam in the 2015 LHC run. The timing circuits as well as the 8-channel ADC were implemented on a separate board.

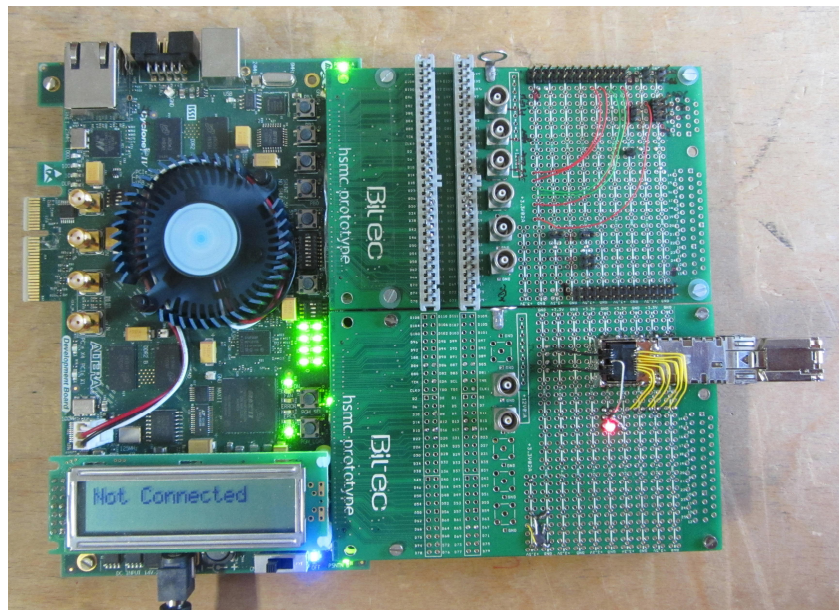


Fig. 4-3. Prototype of the DOROS FPGA controller built with Altera Cyclone IV development kit. The setup was used for Ethernet communication testing and the interface prototyping.

Photograph in Fig. 4-4 displays the latest version of the DOROS front-end built in late 2016. It is based on fourth generation analogue processing board, third generation ADC board and third generation FPGA controller board. The latest hardware updates improve power dissipation and optimize the component values for increased precision and robustness. Detailed description of the DOROS hardware architecture and processing algorithms is provided in the following chapters.

The collimator DOROS system has been installed and commissioned for the beam operation in the beginning of 2015. The installation consisted of 10 DOROS front-ends. The system was successfully used “as is” for two consecutive years of LHC operation [60], [61]. After the hardware upgrades in 2017 the collimator BPM system started to be used also in the interlocked operation. In this mode a software interlock triggers automatic beam dump if the measured orbits exceed programmed thresholds [30]. This operation allows for tighter margins in the collimator settings resulting in increase of the LHC luminosity.

Identical DOROS system (also called standard DOROS system) was also installed in 2015 on selected BPM sensors located around the ATLAS and CMS experiments. The BPM signals were split between the standard LHC BPM system and the DOROS system. In the beginning the DOROS system was used mostly for the R&D purposes, testing the performance of the orbit and oscillation subsystems. In 2016 the system installations were extended to equip more BPMs around the four main LHC experiments in points 1, 2, 5 and 8 [3]. The standard DOROS system was commissioned for beam operation and is being used for a number of applications and beam studies in the LHC such as the so called Van der Meer scans [62] or beam steering [63] in the collision points, to name a few.

More DOROS front-ends were installed in the LHC during the years 2017 and 2018. For the 2018 LHC run the interlocked DOROS system was made redundant with two front-ends. At this time more than 40 operational DOROS front-ends were installed in the LHC and are available for beam measurements [64]. A few more front-ends were also installed in the SPS at CERN for further development and beam studies.

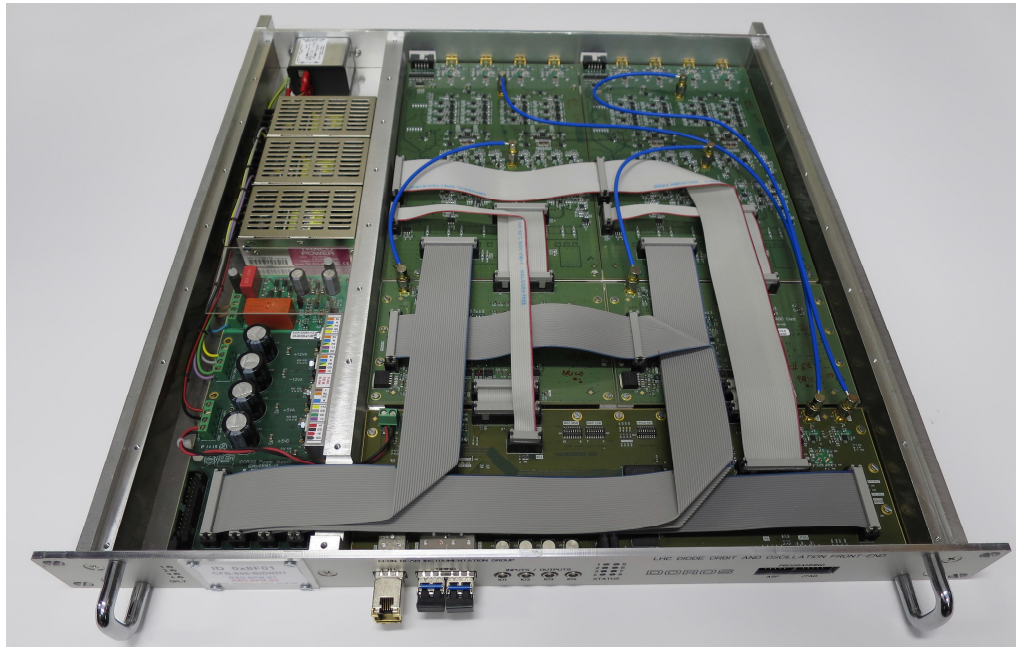


Fig. 4-4. Photograph of a DOROS front-end with removed cover.

4.2 Overview

A DOROS front-end is a stand-alone rack mountable unit built as a 19" 1U box. It consists of six electronics boards and power supply modules as shown in Fig. 4-5. The front-end picture was taken without the top cover and without the internal cabling which would otherwise cover the electronics. The boards are mounted on aluminium bars attached to the chassis. The housing was designed to provide solid structural support as well as good heat dissipation. When closed, the entire box surface serves as a heat sink which can be cooled actively by an external airflow.

Two analogue processing boards A and B can be seen on the right side of the picture. Each board is processing inputs from four BPM electrodes. The inputs can be connected to the upstream and downstream electrode pairs of a collimator or to the horizontal and vertical electrode pairs of a dual-plain BPM. The electrode signals are filtered by the input RF filters, which are followed by the programmable RF amplifiers. The resulting signals are then processed by the respective diode detector subsystems. The orbit subsystem is dedicated and optimized for the beam orbit measurements. The oscillation subsystem is dedicated and optimized for beam oscillation measurements. The low frequency signals from the detectors are further filtered and sent over a flat cable to the ADC board A and B respectively. The ADC board contains an 8-channel 24-bit sigma-delta ADC which is simultaneously sampling the analogue signals from four orbit, two oscillation and two auxiliary channels. The board also contains timing and synchronization circuits as well as other circuits dedicated to the voltage and temperature monitoring. Data from the ADC is sent to the FPGA controller over a dedicated SPI bus.

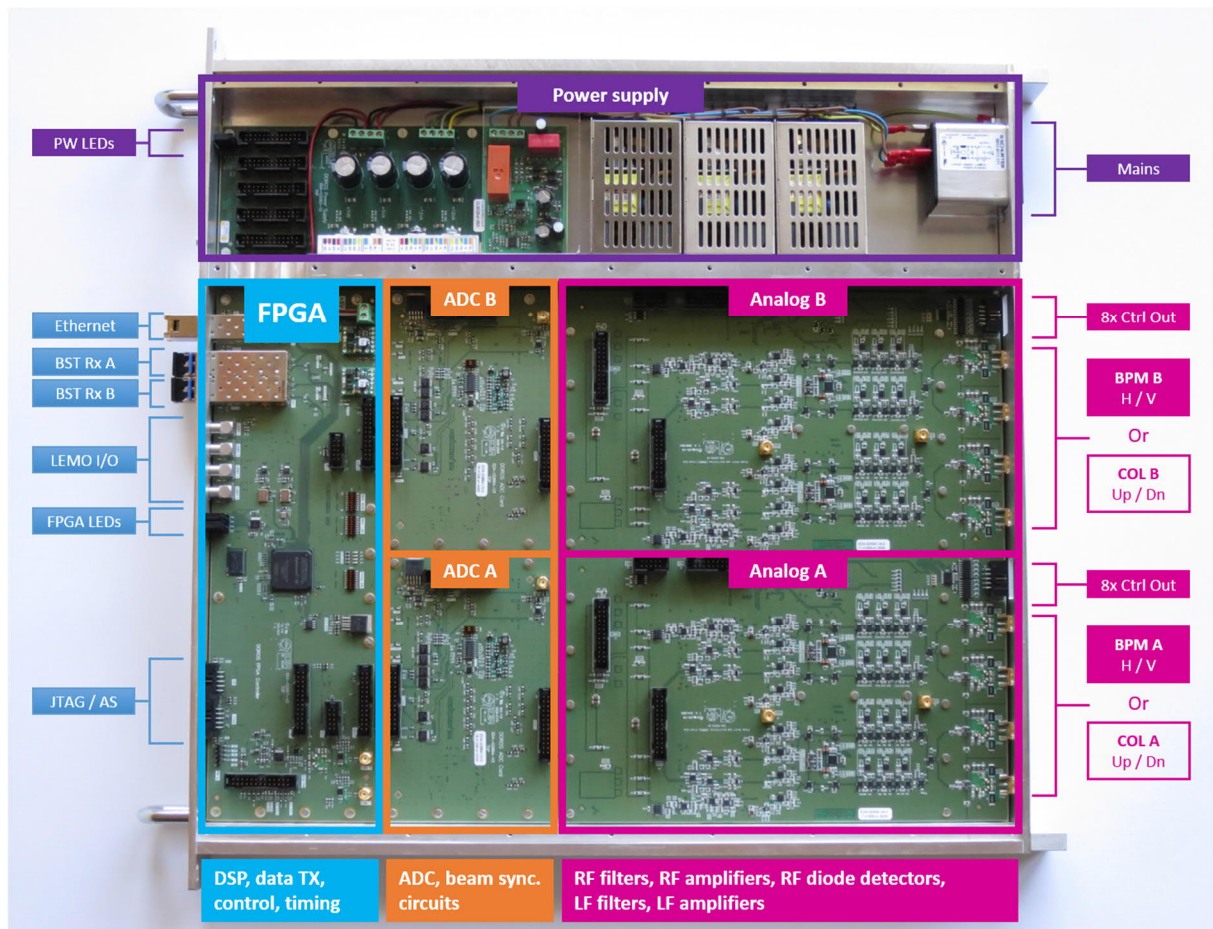


Fig. 4-5. Top view photograph of a DOROS front-end with removed cover and internal cabling.

The FPGA Controller board can be seen in Fig. 4-5 left. The board implements the communication and control interfaces, external timing receivers, on-board oscillators, memories, test signal generator, front-end monitoring and fail-safe recovery. The controller is based on an Altera Cyclone IV FPGA [58] running an embedded system based on a NIOS-II soft-core CPU [65]. The embedded system runs on a Micrium μ C real-time operating system providing TCP/IP stack and multi-threading for networking, front-end controls and maintenance applications [66]. The digital signal processing, timing and clock management and hardware interfaces were implemented in the FPGA logic. Dedicated SFP module provides the Ethernet communication over 1000-BaseT link. The DSP data along with other information, such as the system settings and statuses, is streamed over a UDP protocol to the selected servers on the CERN technical network. The ADC samples can be at the same time buffered in a 1 Gb DDR on-board memory. The readout and control of the DDR buffers is implemented using a TCP protocol. The DOROS project was assigned a dedicated 16-bit MAC address space for the usage on the CERN technical network. The front-end hardware is identical for all DOROS systems, simplifying the front-end installations, management of the spare parts and maintenance of the DOROS systems. Each front-end can be addressed by a unique 16-bit ID number set on the on-board DIP switches. Reprogramming of the FPGA memory is provided over Ethernet network and is based on a dedicated fail-safe technology, developed also by the author. This technology undergoes patenting process. In addition, the FPGA code can be also programmed and debugged over JTAG or Active Serial connectors available on the front-panel.

The LHC timing of each beam is provided over a dedicated fibre optics network. The FPGA board has two dedicated SFP ports and clocking circuits for receiving Beam Synchronous Timing (BST) distribution. The front-end can receive timing from two independent BST sources. The SFP ports can be connected to the FPGA transceivers for gigabit communication if the BST functionality is not required. A BST decoder module in the FPGA allows receiving the timing messages and provides beam reference clocks, triggers and timestamps [68]. The reference clocks are used by the dedicated synchronization circuits generating the ADC clocks synchronized to the beam. The controller board has four LEMO IO connectors with programmable direction for receiving external signals and timing over coaxial cables. The front-end can be also clocked from an on-board oscillator. Local timing derived from on-board crystal oscillators can be used when the external timing is not available. The FPGA controller also supports the White Rabbit protocol [69] used to receive synchronous timing over Ethernet.

The FPGA controls and maintains the internal front-end hardware over eight I2C buses, six SPI buses and one parallel bus. Two optocoupler-isolated connectors in the backplane provide 16 pins for controlling any external hardware, such as fans or RF switches. The controller board also contains a test signal generator for simulating LHC BPM signals. To simulate different beam intensities, the test pulses can be configured with 3 different amplitudes. In order to simulate betatron oscillations the pulse amplitudes can be modulated by a dedicated DAC. The test signals can be injected to the front-end beam inputs using GaAs switches.

The front-end power supplies can be seen in Fig. 4-5 on the top of the picture. There are three AC/DC converters each supplying 25 W power. Two 12 V AC/DCs are dedicated for the positive and negative voltages used mostly by the analogue circuits. The board seen on the left provides the primary linear regulation and filtering. Important analogue circuits are powered from a second regulation stage implemented locally on the boards. The third AC/DC produces 5 V power supply used mostly by the digital circuits such as the Ethernet transceivers, memories or oscillators. The FPGA has dedicated power supplies provided from additional DC/DC regulators.

4.2.1 Signal processing

Diagram in Fig. 4-6 shows signal processing of one BPM electrode pair. The signal processing of the second electrode pair on the analogue board is identical and thus is not displayed in the plot. The processing can be divided into the following parts:

- RF signal processing,
- diode orbit processing (DOR),
- diode oscillation processing (DOS),
- digital signal processing and Ethernet communication and control,
- ADC clocking, timing and synchronization.

As seen in the diagram, the RF part of the signal processing is common for the following blocks. It consists of input circuits that receive the BPM electrode signals, programmable gain amplifiers and the test signal generator. The input stage also contains RF signal multiplexers used for calibrating the asymmetries of the orbit measurement channels. The output of each programmable RF amplifier is used by the DOR and DOS subsystems, as well as the timing circuit generating the beam reference signal. The signals on the inputs of each subsystem are buffered by additional RF amplifiers with fixed gains.

The DOR channels are based on a Compensated Diode Detector (CDD) optimized to measure beam orbits in the frequency range from DC to 100 Hz. Each electrode signal is processed by one CDD. Each DOR channel is digitized by a dedicated ADC channel. The beam positions are computed upon the processed ADC samples. The DOS channel which is based on the RF diode detector is optimized for high sensitivity to small beam oscillations in frequency range from $0.05 f_{rev}$ to $0.5 f_{rev}$ (approximately 0.05 - 0.5 kHz). One DOS channel measuring oscillations in one BPM plain consists of two diode detectors, differential amplifier and filtering stages. Its output is sampled by a dedicated ADC channel.

To accurately measure the beam oscillations the ADC sampling frequency has to be synchronized to the beam revolution. The synchronized f_{ADC} clock, common for all ADC channels, is provided from the synchronization circuits (SC). As showed in the diagram, the reference clock can be provided either from the BST receiver, front-panel LEMO connectors or an on-board oscillator. The SC uses one of these clocks as a reference for the ADC sampling. The beam reference derived from the BPM electrode signals, allows synchronization of the ADC sampling clock to the beam.

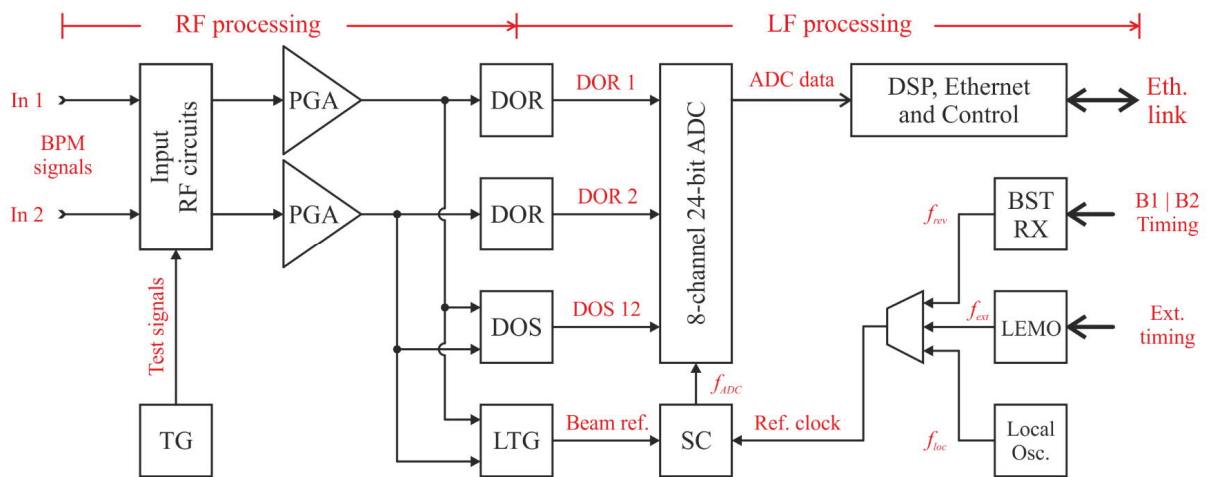


Fig. 4-6. Schematic diagram of DOROS front-end signal processing and synchronization of two BPM electrode channels.
Abbreviations: RF – Radio Frequency, LF – Low Frequency, PA – Programmable RF Gain Amplifier, DOR – Diode Orbit channel, DOS – Diode Oscillation channel, LTG – Local Timing generator, SC – Synchronization circuits, BST RX – Beam Synchronous Timing receiver, TG – Test signal Generator, DSP – Digital Signal Processing, LEMO – external front-panel connectors.

The diagram in the in Fig. 4-7 shows the signal processing, control and communication blocks implemented in the FPGA logic. The ADC samples is transmitted over SPI bus in so called sample frames. A single frame contains serialized samples from all eight channels. The FPGA controller has two dedicated SPI buses allowing to read the frames from both ADCs in parallel.

Readout of each sample frame is initiated upon a data ready (DRDY) signal from the respective ADC. The DRDY signal is synchronous to the selected ADC reference clock. The SPI block receives the frames from the ADCs and provides the de-serialized samples from the ADC channels to the corresponding DSP blocks. Both DOR and DOS subsystems have dedicated processing blocks performing signal filtering and decimation. The output data rate is reduced and adapted to the UDP transmission rate. Detailed description of the DSP processing is provided in the following chapters.

Results from the DOR and DOS subsystems are merged in a single UDP buffer using a dedicated DSP Direct Memory Access (DMA) unit. The UDPs are transmitted over Ethernet at a 25 Hz rate. The sending frequency is compatible with other LHC instrumentation systems, such as the standard LHC BPM system [13]. The transmission can be synchronized to a dedicated Tx trigger received over the BST protocol [68]. The UDPs from the DOROS front-ends are then received and concentrated by a dedicated server machine. Synchronized transmission allows minimizing the spread in the UDP arrival time. The DOROS server can then synchronously provide the processed data to the logging databases as well as to other servers and applications (e.g. Orbit Feedback System [14]). A single UDP datagram contains also other information describing the current state of the front-end such as hardware configuration, PCB temperatures, state-machine settings, statuses or power supply voltages. This information can be used in a server application for monitoring and diagnostics of the system. A copy of each UDP datagram can be streamed to five servers.

The IP addresses of the operational servers are hardcoded in the FPGA code. The remaining IPs of the development and monitoring servers can be configured over Ethernet and stored in a non-volatile memory accommodated on the FPGA board. The memory stores also other information such as the number of power-cycle counts, status reports, error codes and other diagnostics information.

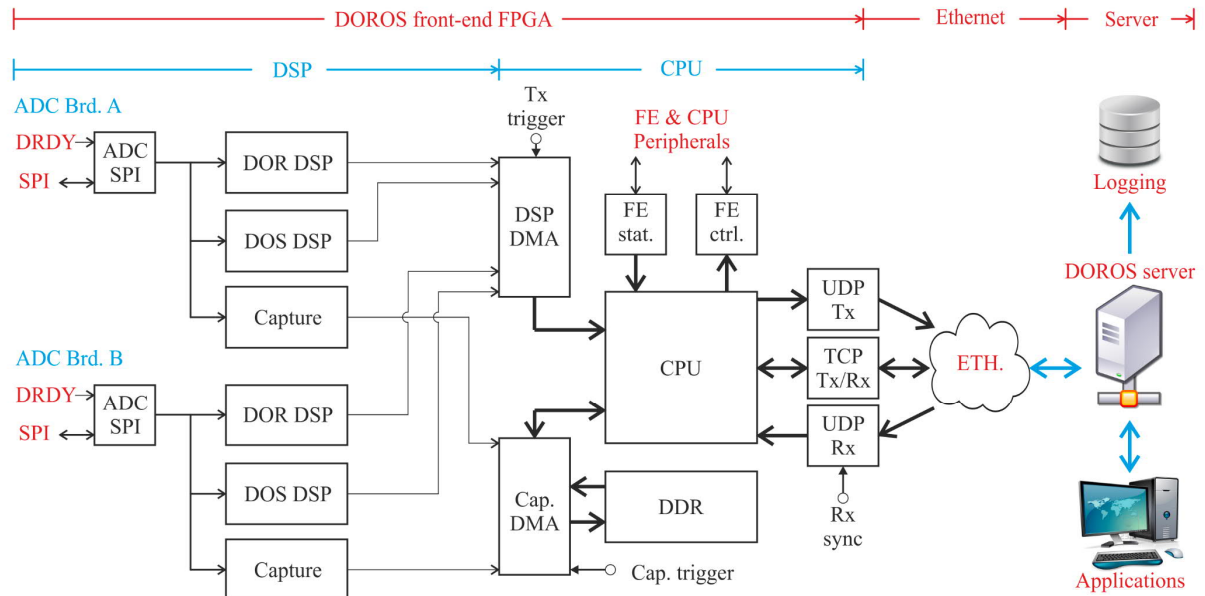


Fig. 4-7. Diagram of DOROS FPGA blocks performing digital signal processing, communication and controls.

Abbreviations: SPI – Serial Peripheral Interface, DOR – Diode orbit processing, DOS – Diode Oscillation Processing, DRDY – Data Ready signal from ADC, DSP – Digital Signal Processing, DMA – Direct Memory Access, CPU – Central Processing Unit, ETH – Ethernet, UDP – User Datagram Protocol for Ethernet data streaming, TCP – Transmission Control Protocol for Ethernet communication

The front-end can also buffer the UDP payload data in so called Post-Mortem buffers located on a dedicated address space mapped in an on-board DDR memory. The buffers allow storing some 4500 UDPs sent at 40 ms rate, which is equivalent to some 3 minutes of the front-end operation or 2 million LHC beam revolutions. The data and control of the buffers is provided over a reliable TCP connection for further offline analysis. The readout can be triggered upon safety critical events, such as the beam injection or beam dump.

The raw ADC samples can be also stored in the local DDR memory. The de-serialized and formatted ADC samples from the SPI modules are buffered at the sampling rate without further digital treatment. A so called capture DMA unit stores the channel data from both ADCs in individual circular buffers mapped in a dedicated address space in the DDR memory. The DMA provides input FIFO buffering and the DDR access arbitration on a “first-come-first-serve” basis. This functionality allows clocking the two ADC boards from independent clock domains. Typically, the ADC board A can be clocked from the LHC beam 1 timing and ADC board B can be clocked from the LHC beam 2 timing. The DMA unit offers two modes of operation:

- Capture – The acquisition is started using hardware or software triggers and finishes after programmed amount of ADC frames. The buffers are then available for readout.
- Freeze – The data is acquired continuously in the circular buffers. New data overwrites the old data after each buffer overflow. A freeze command will stop the buffer pointers and make them available for the readout. The buffering is resumed once the data readout is finished.

The circular buffers store the data from four DOR and two DOS channels for each of the two ADCs. The maximum length of an acquisition is about 1.8 million LHC beam revolutions which is equivalent to some 3.5 minutes. The CPU can control the DMA, trigger the acquisition and access the memory to read the circular buffers after they are stopped. The data is then available “on-request” to the servers through a TCP protocol. The capture functionality is suitable for measurements requiring extended observation bandwidth. The freeze mode can be also triggered after each beam dump for extended post-mortem analysis.

Front-end settings are controlled by dedicated servers over the UDP protocol. A single UDP datagram can contain up to 255 command messages, each consisting of 5 bytes. The command messages are executed in the same order as inserted in the UDP payload. The first byte of the message is interpreted as command ID, which is a unique index identifying the command. The command ID field can address up to 255 different functions to control. The remaining 4 bytes in a message are allocated for control data. The execution of the commands can be synchronized to one of the following sources:

- 25 Hz UDP transmission trigger (provided from BST, local or external source);
- 1 Hz trigger (provided from BST, local or external source);
- Immediately, without waiting for the triggers.

If the selected trigger is not available, the commands will execute after a programmed timeout. The CPU performs a number verifications before the payload is accepted for further processing. This includes checking the information, such as the packet length, signatures or UDP ports. The CPU can also perform a simple IP address filtering. Reception of the command datagram is reported in the UDP data stream providing a simple handshake mechanism. The server can thus monitor the command status without opening additional connections.

4.2.2 DOROS systems

DOROS front-ends, consisting of identical hardware, are configured on the on-board DIP switches to belong to one of the three LHC DOROS systems:

- Collimator system;
- Standard BPM system;
- Development systems (collimator or standard).

Diagram in Fig. 4-8 illustrates an overview of the DOROS front-end integration into the DOROS systems using the CERN infrastructure. In the diagram, the three systems are each represented by one front-end example with corresponding hexadecimal ID number which is configured on the on-board DIP switches. The front-end data and controls are provided over a so called technical network, which is Ethernet network dedicated to the controls and operation of the accelerators at CERN. The ID number is used as a part of the MAC address, which allows addressing the front-ends on the Ethernet network. The ID number also encodes the point and location of a front-end installation. Each DOROS system has assigned dedicated UDP and TCP ports in order to avoid accidental cross-talk. The data concentration and controls are provided by a dedicated server machine which integrates the DOROS system into the CERN controls and operation using a so called FESA framework [70], [71]. As seen in the picture the collimator and standard DOROS systems can send their data to the other server machines. These machines are typically used for development but can be as well used for the server redundancy in case the main operational servers are not available.

The LHC timing is provided in two separate BST networks, one dedicated to beam 1 (blue) and the second to beam 2 (red). As showed in the diagram, the front-ends can be connected to the BPM sensors on separate beams. A front-end can be also connected to the BPM sensors on the same beam requiring the connection only to one corresponding BST. The BST protocol does not provide addressing and therefore the timing is common for all connected front-ends, regardless of the system.

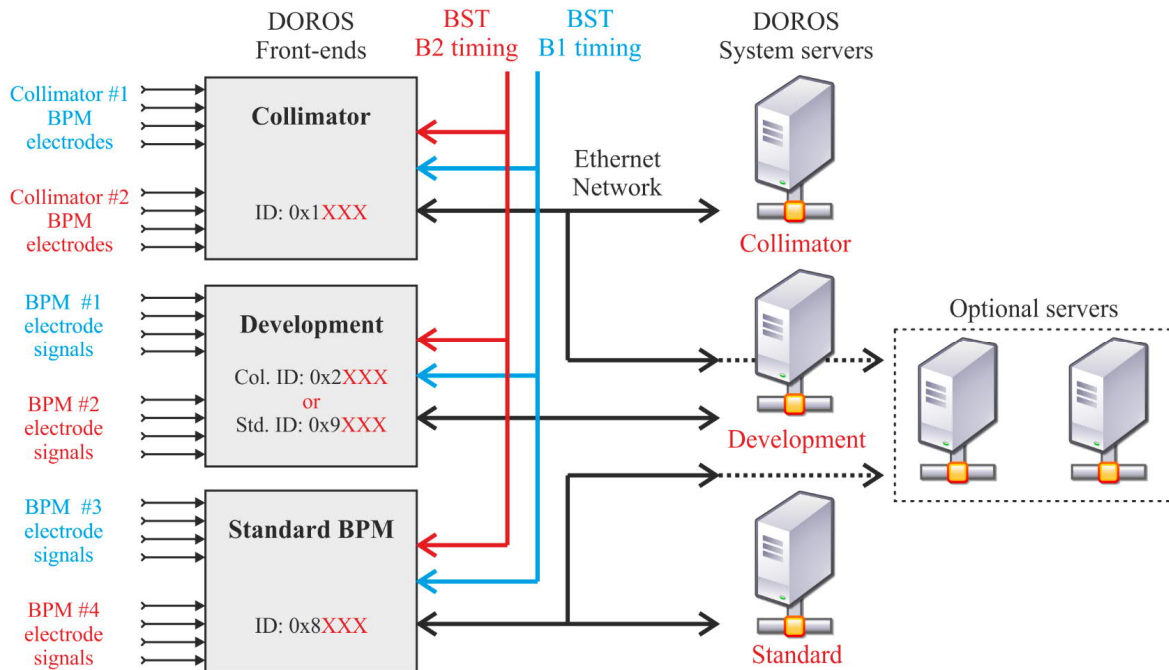


Fig. 4-8. Overview of the DOROS front-end integration into the DOROS systems using the CERN infrastructure.
Abbreviations: BST – Beam Synchronous Timing, BPM – Beam Position Monitor

4.3 Orbit subsystem

Diagram in Fig. 4-9 shows the orbit signal processing of one BPM electrode pair. The waveforms shown on the right of the diagram illustrate the example signals in the most important nodes. The beam current pulses (green), which repeat every revolution period T_r , induce the left l_e (blue) and right r_e (red) electrode signals. The example in the picture shows the case of the beam position offset in the horizontal axis. The signals from the electrodes are processed in the RF input filters. The filters reduce the bandwidth and power of the BPM signals so that they can be processed by the following stages. The dynamic range is controlled by programmable RF gain amplifiers. They allow setting the signal level on the inputs of the diode detectors. The compensated diode detectors (CDD) then convert the short pulses from the RF processing into slowly varying signals around DC. The ripple voltage seen in the detector signals is removed by analogue low-pass filters with the 100 Hz cut-off. The following digitizer stage uses two ADC channels sampling synchronously with the beam revolution frequency. The sampling frequency was optimized for the oscillation processing which requires much larger bandwidth than the orbit measurements.

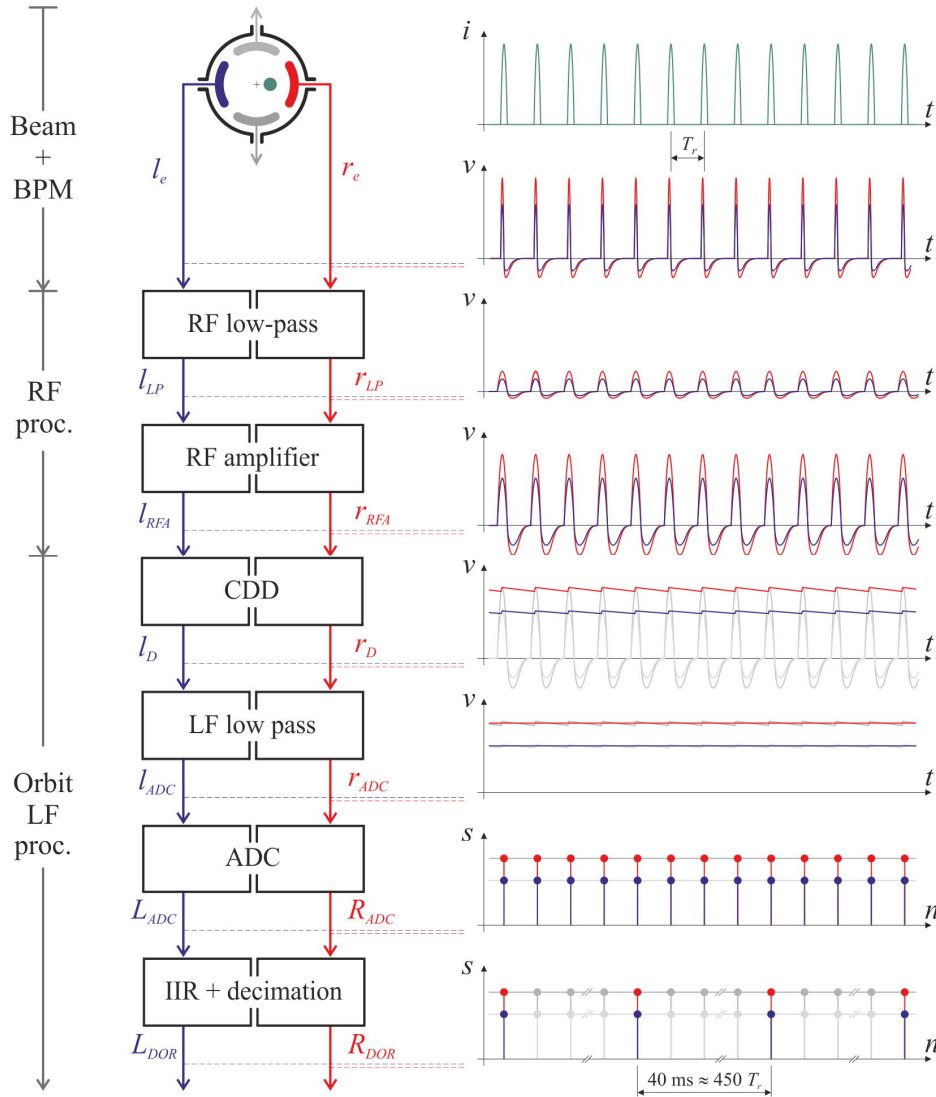


Fig. 4-9. Schematic diagram of orbit signal processing of 2 channels with illustrated signal waveforms shown in the most important nodes. The signal waveforms are given for illustration only. **Abbreviations:** LF – Low Frequency, RF – Radio Frequency, IIR – Infinite Impulse Response filter, CDD – Compensated Diode Detector

To estimate number of quantization bits required to measure the small position changes of the beam, the approximated signals l_{ADC} and r_{ADC} on the ADC inputs can be expressed as:

$$l_{ADC} = \frac{FS + LSb}{2} \quad (4-1)$$

$$r_{ADC} = \frac{FS - LSb}{2} \quad (4-2)$$

where the FS is the full-scale range of the ADC and LSb the smallest quantization step. In this estimation the absolute signal level is set to the middle of the ADC dynamic range and the signal difference, which is proportional to the beam position, is set to $\pm 1/2$ LSb. Inserting the l_{ADC} and r_{ADC} signals into the equation (3-1) yields the smallest change of the normalised position p_{adc} defined as:

$$p_{adc} = \frac{LSb}{FS} = \frac{1}{(2^n - 1)} \quad (4-3)$$

The equation can be simplified by expressing the FS as number of LSb quantization steps. The n then expresses the number of ADC bits. Effective number of bits (n_{ENOB}) defining a dynamic range of an ADC can be estimated based on a well-known equation:

$$n_{ENOB} = \frac{SNR - 1.76 \text{ dB}}{6.02 \text{ dB}} \quad (4-4)$$

where SNR is the signal-to-noise ratio expressing the quality of the input signals expressed in decibels. Assuming that the noise is dominated by the LF part of the signal processing the SNR was estimated to some 120 dB. According to the equation (4-4) the estimated n_{ENOB} is then 19.6 bits. When inserted to equations (4-3) and scaled to millimetres with equation (3-2) using a 61 mm BPM aperture, the orbit resolution can be estimated to some 20 nm. Please note that this estimation does not take into account the low-pass filtering and averaging done in the DSP.

The ADC samples from each electrode are processed by the IIR low-pass filters which operate at the sampling frequency. The digital filtering further improves the precision of the measurement and decreases the signal bandwidth to a configurable cut-off frequency. The output sample rate is then decimated so that the data transmission in the UDP datagrams can be adapted to the 25 Hz. The IIR filter acts as an anti-aliasing filter. However, the relationship between the ADC sampling frequency and the UDP sending rate is not a fixed integer number. The resulting difference between the two time domains is adapted by a variable decimation factor which changes around 450. The IIR filters thus also act as buffers between the two time domains. The processed and decimated DOR samples from the L_{DOR} and R_{DOR} channels are transmitted to the servers which compute beam positions in millimetres.

The signal levels on the CDD inputs have to be controlled within optimal amplitude range. The pulse amplitudes should be large enough so that the detectors can to operate in the most linear part of the characteristic. However, too large amplitudes may saturate the ADC channels. The BPM signals change with the beam intensity (e.g. during bunch injection or beam collisions) as well as the transverse beam position in the BPM. The RF amplifiers in each channel have to accommodate these changes by controlling its gains within range of more than 80 dB. Dedicated control unit in the FGPA automatically sets the RF gains in order to maintain the signal levels within the optimal ADC input range. The RF amplifier in each channel consists of four stages. Each stage can be configured as an attenuator, an amplifier or bypassed using GaAs RF switches. The signal gain before the CDDs is controlled by the configuration of the stages inserted in the RF chain. The short BPM pulses with large peak voltages have to be attenuated and low-pass filtered by the RF filters on the front-end inputs. They are important for protecting the active circuits in the analogue signal processing as well as prevent the amplifier saturation and slew rate distortion.

One of the main advantages of the diode detectors is that they do not require precise clocking for their operation. The diode detectors in the DOR act as sample and hold circuits. The resulting value after low-pass filtering does not depend on the pulse arrival time. This means that the entire RF processing, including the input cables, does not need to be precisely phase matched. With larger BPM signals the processing is also less sensitive to the input perturbations such as noise, reflections or interferences. The peak detection takes into account only the largest pulses. Smaller perturbations are thus automatically suppressed by the detectors.

However, asymmetries between the corresponding analogue channels introduce systematic errors and drifts in the measured positions. Impact of these errors on the position measurements can be demonstrated on the following example. Assuming centred beam position, the pulses in both l_e and r_e electrode signals have equal amplitudes. The relative asymmetry error in the channel gains g_{err} contributes to the position error. The L_{DOR} and R_{DOR} signal from the two DOR chains can be then expressed as:

$$L_{DOR} = R_{DOR} (1 \pm g_{err}) \quad (4-5)$$

The resulting position error p_{gerr} expressed as:

$$p_{gerr} = \frac{g_{err}}{g_{err} \pm 2} \quad (4-6)$$

can be found by inserting the equations (4-5) into the position equation (3-1). Examples of the p_{gerr} for different gain errors between the channels are shown in Table 4-1. As seen in the table, a 1 % gain error g_{err} , typically achievable in practise, would result in a systematic position error of some 76 μm . The g_{err} takes into account all gain asymmetries introduced the analogue processing. The position error p_{gerr} can thus change depending on the configuration of the RF amplifier stages, as well as the thermal drifts of many components in the chain.

The accuracy of the orbit measurements also depends on the assymetric offsets between the channels. Please note that the RF processing does not contribute to the offset asymmetry as the electrode signals before the detectors are AC coupled. Asuming that the assymetry is dominated only by the offsets o_{err} , the channel signals can be expresed as:

$$L_{DOR} = R_{DOR} \pm o_{err} \quad (4-7)$$

The resulting position error p_{oerr} is then expressed as:

$$p_{oerr} = \frac{o_{err}}{o_{err} \pm 2R_{DOR}} \quad (4-8)$$

The error depends on the signal amplitude and can become significant when maintaining the signals at low amplitudes. The Table 4-2 summarizes a few numeric examples for a combination of four different offsets and five signal amplitudes. For example, a 0.1 % offset channel asymmetry, equivalent to a DC voltage offset of some 5 mV, would yield 15 μm position error with signal levels at 0.5 of the ADC full-scale range. Increasing the signal level to 0.7 of the ADC FS would result in decreased error estimation to some 10 μm . Please note that the offset errors in the LF part can be controlled better than the gain errors in the RF part. Precision amplifiers and accurate ADCs can reach offset voltages in order of microvolts. Optimizing the analogue processing to systematically achieve channel symmetry better than 1 % in all RF amplifier configurations would be very difficult. Therefore a dedicated calibration technique was introduced to provide compensation of the channel asymmetries independently of the RF gain. The technique is based on periodic switching of the electrode signals, requiring additional signal processing in the front-end. The automatic calibration became crucial for achieving high accuracy of the orbit measurements. Implementation of the technique is described in the following chapter.

Table 4-1. Examples of a position error estimation for different gain asymmetries between two DOR channels.

Gain asymmetry g_{err} [%]	10	5	3	2	1	0.1	0.01	0.001	0.0001
Norm. pos. p_{gerr} [10^{-3}] *	476.19	243.90	147.78	99.01	49.75	4.99	0.5	0.05	0.005
Abs. pos. P_{gerr} [μm] *†	726.19	371.95	225.37	150.99	75.87	7.62	0.76	0.076	0.008

*Rounded to 3 decimal places, † Referenced to a 61 mm aperture using linear scaling equation (3-2).

Table 4-2. Examples of a position error estimation for different offset asymmetries between two DOR channels.

Norm. sig. R_{DOR} [FS]		Absolute position error P_{oerr} [μm] *†				
		0.1	0.3	0.5	0.7	0.9
Offset asymmetry o_{err}						
	[%] [mV]					
1	50	726.19	250	150.99	108.156	84.254
0.1	5	75.871	25.374	15.235	10.885	8.468
0.01	0.5	7.621	2.541	1.525	1.089	0.847
0.001	0.05	0.762	0.254	0.152	0.109	0.085

*Rounded to 3 decimal places, † Referenced to a 61 mm aperture using linear scaling equation (3-2).

4.3.1 Calibration of the channel asymmetry

Asymmetries between the analogue channels can be compensated using calibration switching techniques. Picture in Fig. 4-10 shows the principle of DOR calibration technique based on the signal multiplexing. Calibration of the asymmetries in the electronics is achieved by switching the input signal paths between the two channels and measuring the electrode signals in each configuration of the switch. The picture displays switching of the electrode signals between the analogue channels A and B. The output of the processing is used to compute beam positions in the respective BPM plane. Please note that the multiplexer is located after the input filters.

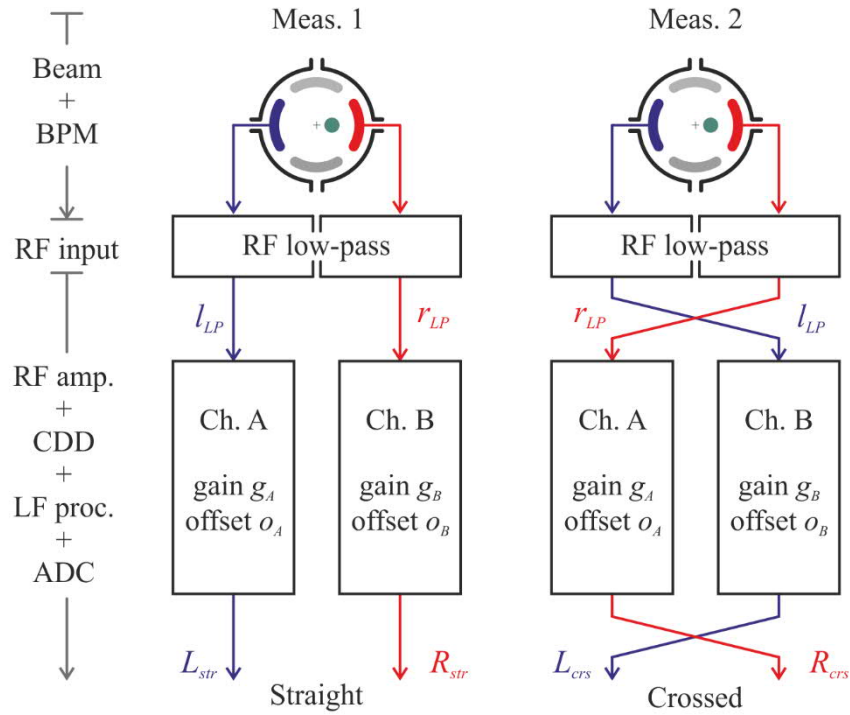


Fig. 4-10. Schematic diagram of the calibration switching to cancel asymmetries in analogue processing of an electrode signal pair. **Abbreviations:** RF – Radio Frequency, LF – Low Frequency, CDD – Compensated Diode Detector.

Assuming linear signal processing in the analogue channels, the output DOR signals from the L_{DOR} and R_{DOR} channels can be defined as:

$$L_{DOR} = g_A l_{LP} + o_A \quad (4-9)$$

$$R_{DOR} = g_B r_{LP} + o_B \quad (4-10)$$

where l_{LP} and r_{LP} are the amplitudes at the output of the RF low-pass filters, g_A and g_B are the gains and the o_A and o_B are the offsets introduced in the two analogue channels. The ideal case without any form of asymmetries between the channels can be expressed as the $g_A = g_B$ and $o_A = o_B = 0$.

As shown in the picture Fig. 4-10, the calibration measurements are performed in two channel configurations. The signal multiplexer is set to straight state in order to pass the electrode signals as shown on the left of the picture. This is also a default setting for cases when the calibration switching is not used. The signal multiplexer allows to swap the electrode signals on the inputs of the two channels as shown on the right of the picture. Signal swapping is equivalent to changing the sign of the position reading. To correct the position sign a de-multiplexer is required at the end of the chain. The output values obtained in the two multiplexer states can be described with the following sets of equations:

$$L_{Str} = g_A l_{LP} + o_A \quad (4-11)$$

$$R_{Str} = g_B r_{LP} + o_B \quad (4-12)$$

$$L_{Crs} = g_B l_{LP} + o_B \quad (4-13)$$

$$R_{Crs} = g_A r_{LP} + o_A \quad (4-14)$$

It is assumed that the channel gains and offsets remain constant during the straight and crossed measurements. The calibration algorithm first uses the four values after de-multiplexing to compute averages of the corresponding straight and crossed measurements. The average electrode values L_{cal} and R_{cal} are expressed as:

$$L_{cal} = \frac{(L_{Str} + L_{Crs})}{2} = \frac{l_{LP} (g_A + g_B) + o_A + o_B}{2} \quad (4-15)$$

$$R_{cal} = \frac{(R_{Str} + R_{Crs})}{2} = \frac{r_{LP} (g_A + g_B) + o_A + o_B}{2} \quad (4-16)$$

The results are then used to compute the calibrated normalised position p_{cal} expressed as:

$$p_{cal} = \frac{R_{cal} - L_{cal}}{R_{cal} + L_{cal}} = \frac{(g_A + g_B)(r_{LP} - l_{LP})}{(g_A + g_B)(r_{LP} + l_{LP}) + 2(o_A + o_B)} \quad (4-17)$$

The equation can be reduced when assuming that the offsets in the signal processing are negligible with respect to the amplified electrode signals. The calculated position is then independent of the gains in the analogue chains:

$$p_{cal} \cong \frac{(r_{LP} - l_{LP})}{(r_{LP} + l_{LP})} \quad \text{for } (g_A + g_B)(r_{LP} + l_{LP}) \gg 2(o_A + o_B) \quad (4-18)$$

Conversion to millimetres can be done by using a simple linear approximation using equation (3-2) or higher order polynomials such as the equations (3-3) and (3-4) which describe the BPM characteristic. The computed position p_{cal} is valid as long as the L_{cal} and R_{cal} components are always computed with channel values obtained from two consecutive states of the RF multiplexer. Please note that the calibration technique requires to switch the input multiplexer periodically in order to continuously remove the position dependence on the gains.

The orbit calibration technique can be demonstrated on a following numerical example. The gains and offsets of the analogue channels are presented in Table 4-3. The example is based on a 3 % gain asymmetry and 0.3 % offset asymmetry. It is assumed in this example that the beam position stays constant during the calibration switching.

Table 4-3. Input parameters of the example computation that demonstrates the calibration technique.

Gain		Offset	
g_A	1.01	o_A	0.001
g_B	0.98	o_B	-0.002
Asymmetry error *	0.03	Asymmetry error *	0.003

*Rounded to 2 decimal places.

The input signals l_{LP} and r_{LP} were set to simulate different positions P before signal processing in the calibration algorithm. The ratio r_{LP} / l_{LP} represents equivalent attenuation value on the r_{LP} channel input. The input values were then inserted into the respective equations to demonstrate the signal levels in the important nodes of the calibration processing. Obtained results were summarized in Table 4-4.

Table 4-4. Example computation demonstrating the channel signals in the important nodes of the calibration processing.

l_{LP}	r_{LP}	$20\log(r_{LP} / l_{LP})$ [dB] **	P [mm] *†	L_{str} *	R_{str} *	L_{crs} *	R_{crs} *	L_{cal} *	R_{cal} *
1	1	0.0	0.000	1.011	0.978	0.978	1.011	0.995	0.995
0.99	1	0.1	0.077	1.001	0.978	0.968	1.011	0.985	0.995
0.9	1	0.9	0.803	0.910	0.978	0.880	1.011	0.895	0.995
0.7	1	3.1	2.691	0.708	0.978	0.684	1.011	0.696	0.995
0.5	1	6.0	5.083	0.506	0.978	0.488	1.011	0.497	0.995
1	0.5	-6.0	-5.083	1.011	0.488	0.978	0.506	0.995	0.497

*Rounded to 3 decimal places, ** Rounded to 1 decimal place, † Referenced to a 61 mm aperture using linear scaling (3-2).

The signal levels were used to demonstrate the channel asymmetries translated into the beam position measurement. Obtained signal levels L_{str} , R_{str} , L_{crs} , R_{crs} from the respective equations were used to compute positions P_{str} and P_{crs} in each multiplexer state. They represent the measurement results which the system would provide if the calibration technique was not applied. The equations (4-17) and (3-2) were then used to compute the calibrated output P_{cal} in millimetres. Position error was evaluated for the P_{str} , P_{crs} and P_{cal} using the input P as a reference. In general the absolute error ΔP_{xx} was expressed as:

$$\Delta P_{xx} = P - P_{xx} \quad (4-19)$$

where the subscript xx is substituted for the str , crs and cal to obtain errors for each selected node in the calibration processing. All position results obtained in this example were summarized in Table 4-5.

Table 4-5. Example computation demonstrating the calibration technique on a few representative input signal cases.

l_{LP}	r_{LP}	$20\log(r_{LP} / l_{LP})$ [dB] **	P [mm] *†	P_{str} [mm] *†	P_{crs} [mm] *†	P_{cal} [mm] *†	ΔP_{str} [μm] **†	ΔP_{crs} [μm] **†	ΔP_{cal} [μm] **†
1	1	0.0	0.000	-0.253	0.253	0.000	253.0	-253.0	0.0
0.99	1	0.1	0.077	-0.176	0.330	0.077	253.1	-253.1	0.0
0.9	1	0.9	0.803	0.549	1.056	0.803	253.4	-253.8	-0.4
0.7	1	3.1	2.691	2.442	2.942	2.693	249.0	-250.9	-1.6
0.5	1	6.0	5.083	4.850	5.321	5.087	232.9	-237.4	-3.4
1	0.5	-6.0	-5.083	-5.321	-4.850	-5.087	237.4	-232.9	3.4

*Rounded to 3 decimal places, ** Rounded to 1 decimal place, † Referenced to a 61 mm aperture using linear scaling (3-2).

As seen from the example results, the channel asymmetries cause important position errors reaching more than 0.5 mm when the calibration is not used. It can be seen in the Table 4-5, that the residual error in the position calibration ΔP_{cal} is fully removed for zero position offset p . This is due to the signal subtraction in the numerator of the equation (4-17). Such accuracy improvement is desired especially for the BPM-based collimator alignment when the beams are close the BPM centre.

As the position p increases to 5 mm (6 dB amplitude signal ratio) the errors ΔP_{str} and ΔP_{crs} increased to more than 0.5 mm. The calibration error ΔP_{cal} also increased to 3.4 μm due to the offsets o_A and o_B seen in the denominator of the equation (4-17). Please note that the offsets of equivalent values but opposite signs would cancel, which would result in zero position error. When comparing the ΔP_{cal} to the ΔP_{str} or ΔP_{crs} the calibration technique improved the precision of the position measurement by an order of magnitude. In addition, the ΔP_{cal} is also independent of the position sign as can be seen in Table 4-5 when the input signals l_{LP} and r_{LP} are swapped.

The described calibration technique cannot distinguish between the asymmetries coming from the gain or offsets. However, under certain conditions these errors can be estimated using the switched electrode signals from the DOR processing. Using the approximated signal levels from equations (4-11), (4-12), (4-13) and (4-14) one can obtain the gain ratio coefficient g_c which can be expressed as:

$$g_c = \frac{g_A}{g_B} = \frac{L_{Str} - R_{Crs}}{L_{Crs} - R_{Str}} \quad (4-20)$$

Similarly, the relative offset coefficient o_c can be expressed as:

$$o_c = o_A - o_B = \frac{g_A}{g_B} = \frac{L_{Crs} R_{Crs} - L_{Str} R_{Str}}{L_{Crs} - R_{Str}} \quad (4-21)$$

In practise the estimation of the g_c and o_c coefficients can be inaccurate, even impossible, when the beam is close to the centre (e.g. aligned collimator jaws). The equations (4-20) and (4-21) lead to large errors if the differences between the channel signals become very small. In such case the small position offsets and the channel asymmetries cannot be distinguished. This effect on the gain coefficient g_c can be seen in Fig. 4-11 obtained with laboratory measurements. The same measurement data was used to estimate the offset coefficient o_c showed in Fig. 4-12. The beam positions were simulated using an input attenuator connected to one of the two front-end inputs. When the input signals become very similar (corresponding to 0 dB attenuator) the estimation can change by more than 10 %. In addition, the spread in the estimated value increases by an order of magnitude. Therefore, using the gain and offset coefficients for calibrating the asymmetries of the electronics could not be considered to be reliable and robust. The calibration based on the periodic input multiplexing has number of advantages:

- improves accuracy of the orbit measurement by compensating important asymmetries independently of the gain settings,
- improves precision and minimizes drifts in the orbit measurement,
- provides minimum residual error around centred beam,
- works with any beam offset.

The frequency of the multiplexer switching will determine the effective bandwidth of the orbit measurement. The highest switching frequency is limited by the time required to stabilize the signals after each switching transient. The lowest switching frequency is limits the measurement bandwidth and latency. The latter is important for systems that use the calibrated DOROS orbit measurements in a feedback loop. Calibration switching frequency typically used during nominal operation is 1 Hz ($\sim 10^3$ LHC beam revolutions). The technique is limited by the following factors:

- asymmetries in the input RF filters are not included in the calibration,
- accuracy is limited by the analogue electronics offsets, as they don't entirely cancel in the denominator of equation (4-17),
- limited measurement bandwidth to low frequencies,
- increased latency of the measurement.

The implementation details of the calibration switching technique are presented in the following chapters.

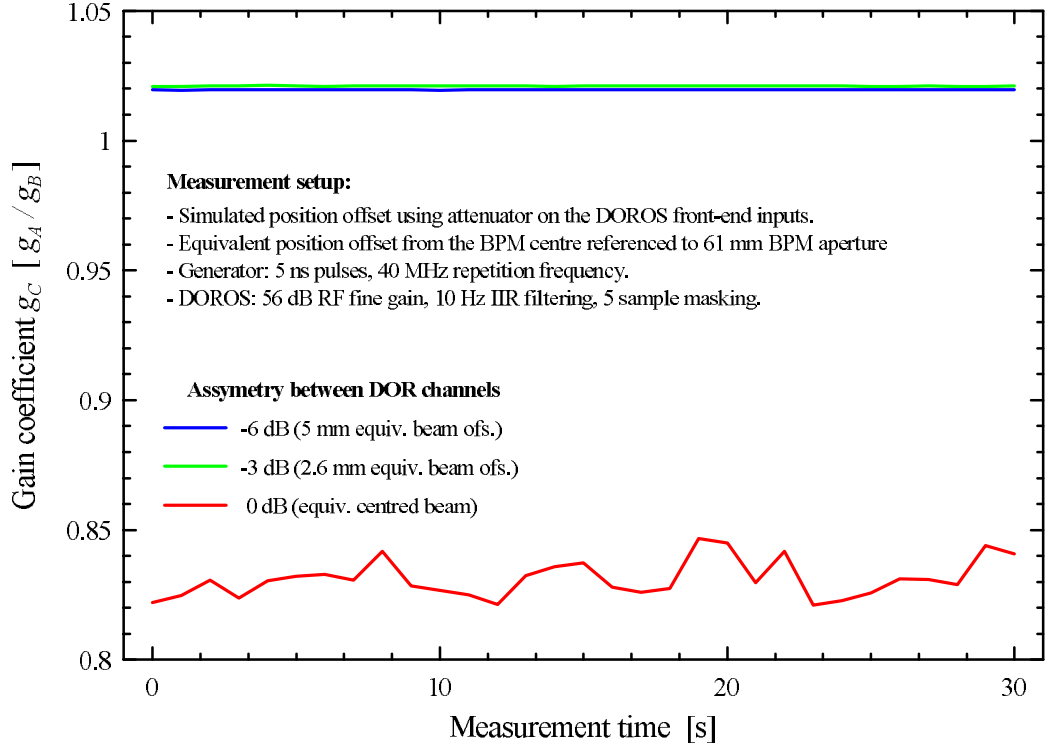


Fig. 4-11. Estimation of the gain coefficient g_c based on laboratory measurements. The signal generator Keysight 33600A simulated BPM signals with 5 ns pulses with 40 MHz repetition rate. The three position offsets were simulated with different attenuators connected to one of the front-end inputs.

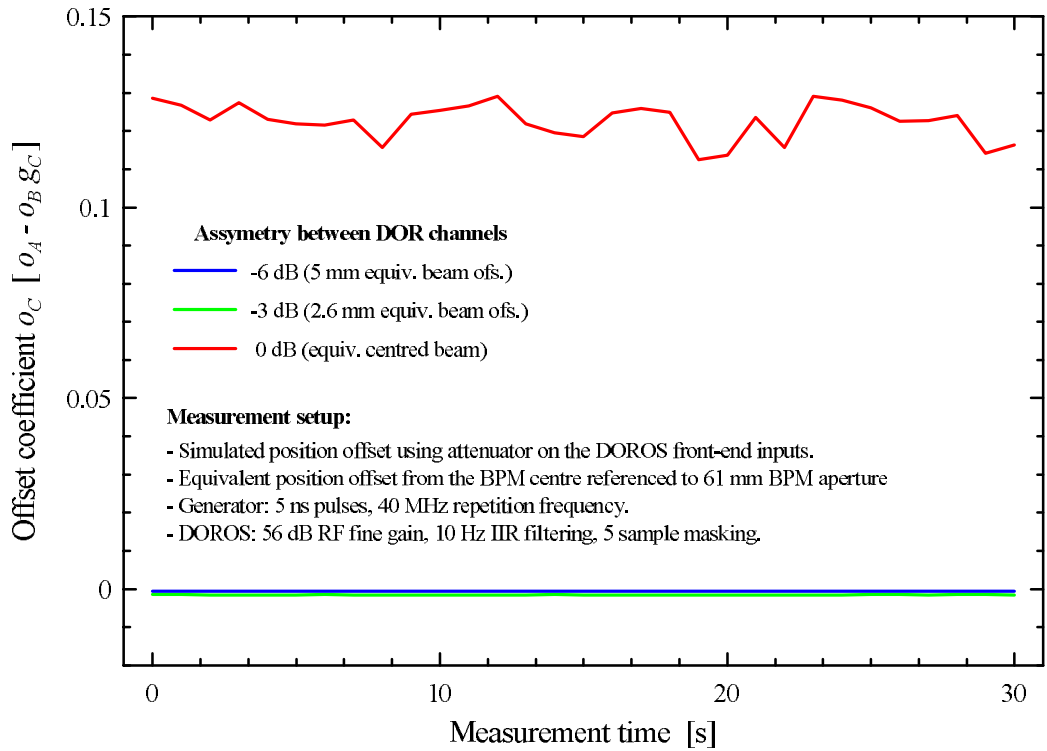


Fig. 4-12. Estimation of the offset coefficient o_c based on laboratory measurements. The signal generator Keysight 33600A simulated BPM signals with 5 ns pulses with 40 MHz repetition rate. The three position offsets were simulated with different attenuators connected to one of the front-end inputs.

4.3.2 Radiofrequency signal processing

The DOROS front-ends are installed in locations with low radiation levels, such as service tunnels and rack galleries. The electrode signals from a BPM are sent to the respective front-end inputs over coaxial cables. Depending on the BPM location in the beam line the length of the BPM cables is typically between 50 and 300 metres. The signals are then received and processed by the RF circuits in the front-end inputs. Block diagram in Fig. 4-13 gives an overview of the RF signal processing of one BPM electrode pair. The RF processing is identical in all electrode pairs in the front-end. The signals are first filtered by the 80 MHz low-pass filters. As the BPM electrodes appear as reactive source impedance the DOROS input filters were designed as non-reflective. The filter provides some -30 dB return loss measured over some 2 GHz frequency range. The filter decreases the amplitudes and broadens the BPM pulses from 1 ns to about 5 ns. With such filtering the used RF amplifiers have enough bandwidth and slew rate to amplify the signals without significant distortion. The transformers shown in the picture provide galvanic insulation to avoid ground loops which otherwise could perturb the transmitted beam signals. The AC coupling provides a low frequency cut-off to filter interferences and noise. In order to measure high intensity beams, the input resistors in a single RF input filter can dissipate up to 2 W. The total dissipation in all input filters in the front-end can thus reach some 16 W. To increase the reliability, the DOROS front-ends are usually installed with additional input attenuators. Most of the RF power is then dissipated outside of the front-end to protect the system and increase its reliability.

The following block in the RF chain is the calibration multiplexer. Asymmetries in the analogue electronics can be reduced by periodic multiplexing of the signal paths between two channel inputs. The multiplexer is realised with a GaAs switch which allows to send signals in a so called straight and crossed configurations. In addition, the switch can be also configured to isolate the inputs from outputs in a high impedance state or connect the four signal ports together. The calibration multiplexers on an analogue board are controlled in parallel. Please note, that the asymmetries between RF filters cannot be compensated by the calibration switching as the multiplexing circuits are located after the input transformers.

The following multiplexer stage is used to select between the BPM signals and the test signal from the FPGA board. The test signal is first split by four before feeding each of the four channels on the analogue board. When in beam mode, the multiplexers connect the BPM signals to the RF amplifiers and the source of the test signal is connected to the load resistors. In addition the test generator is switched off preventing any leakage of the test signal to the beam inputs.

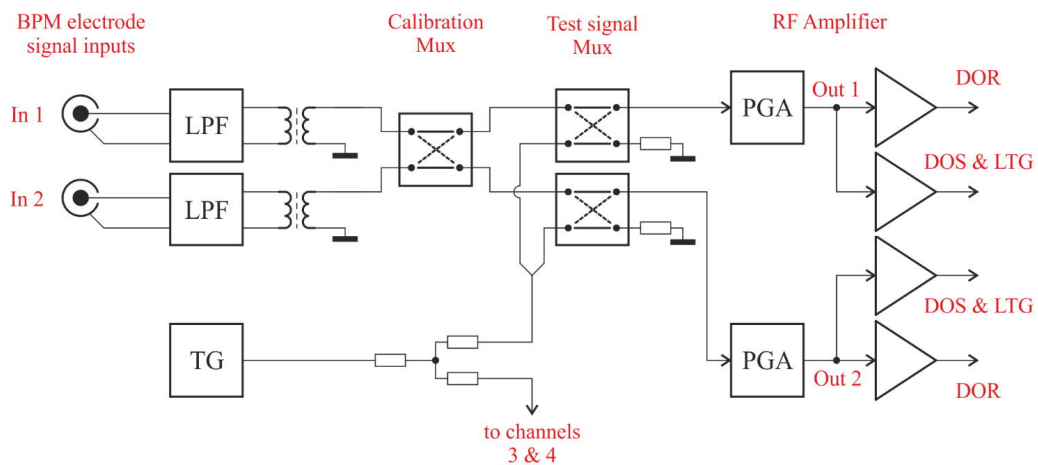


Fig. 4-13. Schematic diagram of the front-end RF processing for 2 channels.

Abbreviations: LPF – Low Pass Filter, TG – Test signal Generator, PGA – Programmable Gain Amplifier.

When in test mode the multiplexers connect the test generator to the amplifiers. This mode can be used when the beam signals are not present on the front-end inputs. The test multiplexers are based on the same GaAs switches as the calibration multiplexer. When required the multiplexers can be also set to high impedance state to increase insulation from the inputs. The test multiplexers on an analogue board are controlled in parallel.

4.3.2.1 Radiofrequency amplifiers

The diagram in Fig. 4-14 shows the RF amplifier blocks which process signals from one BPM electrode pair. The RF amplifier consists of a programmable gain and a fixed gain block. The programmable part of the amplifier consists of four RF stages in the RF path of each channel. Each RF stage can be configured as an amplifier, an attenuator or bypassed. The gain is controlled by configuring and combining the RF stages. The amplifier blocks with fixed gains (marked in Fig. 4-14 as red rectangles) were built with fast current feedback amplifiers. They have excellent noise performance and minimal signal drift due to the amplifier closed-loop operation. The amplifiers also provide large output voltage span which is suitable in the fixed gain part of the amplifier that drives the diode detectors.

As seen in the picture, each RF stage can be also configured as a fixed attenuator (marked in Fig. 4-14 with blue rectangles). The signal paths in an RF stage are controlled with six GaAs switches. Each of the switches $s1$, $s2$ and $s3$ which are single pole three throw (SP3T), are integrated in a single package with three control inputs. Each RF stage contains two such SP3T switches, both addressed and controlled by the FPGA parallel bus. The SP3T switches of the corresponding RF stages are controlled in parallel. Upon the front-end power-up all the GaAs switches are initialized to the high impedance state before the FPGA boots and applies the default settings.

The 4th RF stage contains an integrated programmable gain amplifier (IPGA) which is dedicated for fine gain tuning. The IPGA consists of a digitally controlled attenuator followed by a highly linear transconductance amplifier operating in an open loop. The gain can be controlled from -10 dB to $+14$ dB in smallest steps of 1 dB with ± 0.2 dB maximum error. Absolute gain of the amplifier is set by the load resistance on its output. An output transformer provides AC coupling and converts the differential signals on the output of the amplifier to the single ended signals suitable for the following stages.

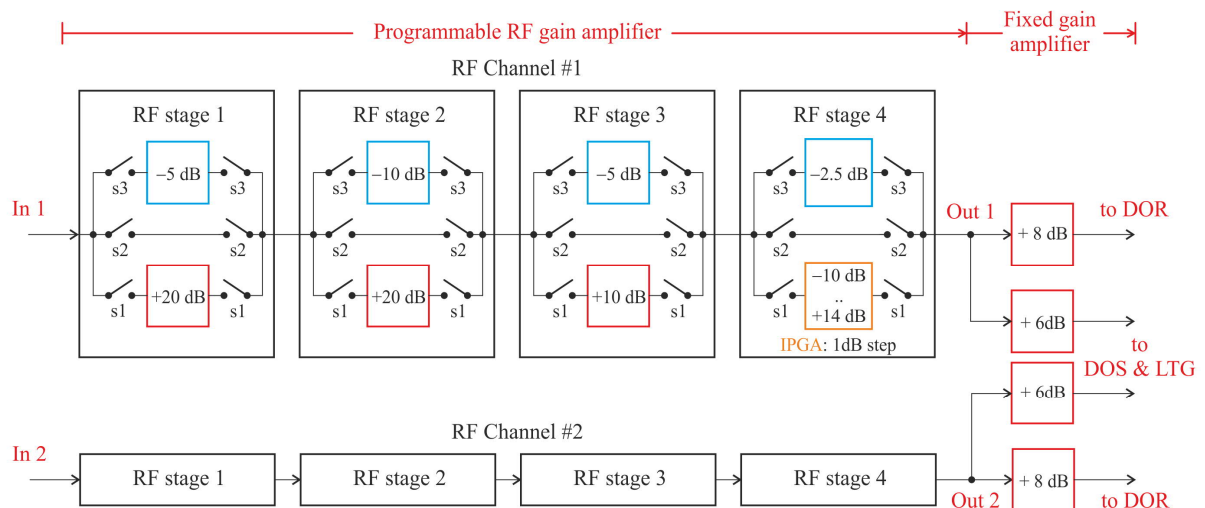


Fig. 4-14. Schematic diagram of the RF amplifiers in front of the diode detectors. Each channel consists of 4 individually controlled stages: a bypass, an amplifier and an attenuator. The last RF stage contains a programmable gain amplifier with 1 dB control step. **Abbreviations:** PGA – Programmable Gain Amplifier, RF – Radio Frequency, DOR – Diode Orbit, DOS – Diode Oscillations, LTG – Local Timing Generator.

Two IPGAs accommodated on a single chip are used to process one electrode pair which decreases the asymmetries between the processing channels. However, the gain of each IPGA can be controlled individually, allowing to introduce asymmetries between the channels by setting different gains. This mode can be used for special operation of the front-end, such as diagnostics of the oscillation processing with the test signals.

The fixed gain amplifiers at the end of each RF chain provide the buffered electrode signals to the following processing blocks. The gains in the fixed amplifier stage were optimized to operate the diode detectors with maximum signal amplitudes.

The layout of the RF amplifier stages was optimized for good thermal connection to the heatsinks as well as symmetry between the paired channels. The IPGA chip for example can dissipate about 1 W with both channels enabled. The IPGA can be disabled when not in use, decreasing the power consumption to a few mW. The differences between the channels can be also minimised by using precision resistors in the amplifier feedback paths and attenuator networks.

The RF gains are configured by the FPGA controller using three schemes:

- fine gain,
- coarse gain,
- asymmetric gain.

The look-up table in the controller memory maps possible gain configurations to a gain index. Each scheme uses its own dedicated table. The tables are used to configure respective registers on the analogue boards, so that the gain can be increased without saturating the amplifier stages.

The fine gain scheme utilizes the IPGA to achieve the minimum gain step of 1 dB. The IPGA amplifiers in the last RF stage are therefore always enabled and used in the RF chain. The remaining stages are used as pre-amplifiers. The look-up table used in the fine scheme is shown in Table 4-6. This mode provides dynamic range control from -30 dB up to $+64$ dB in 103 steps. Gain index 0 represents the state when the signal path in all RF stages is disconnected by the multiplexer switches (high impedance state). Gain transitions occur when the first three RF stages need to be reconfigured. As can be seen in the Table 4-6, there are two equivalent gain realisations around each transition. This helps to introduce an overlap for the gain of the individual RF stages. The main advantage of this scheme is the 1 dB minimum gain step. The disadvantage is the increased noise and sensitivity to temperature variations of the signals due to the open loop amplifier in the IPGA. Another disadvantage is also increased power dissipation on the analogue boards due to the use of the IPGA.

The coarse gain scheme uses only the fixed gain amplifiers and attenuators to achieve minimum gain step of 2.5 dB. Corresponding look-up table is shown in Table 4-7. It provides gain control starting from -22.5 dB up to $+50$ dB in 29 steps. The IPGA is disabled which decreases the power dissipation on the analogue boards. Realisation of the maximal attenuation (indicated as $-\infty$ gain in the table) is shared with the fine gain scheme. The main advantage of this scheme is that all the amplifiers inserted in the chain operate in closed loop, improving the temperature stability and noise.

The asymmetric gain scheme allows to set different gain value in each RF channel. The control is split into the common gain set by the first three RF stages and asymmetric gain controlled only by the IPGA. The common gain is based on the look-up table used for the coarse scheme.

Table 4-6. Configuration look-up table used to setup the RF amplifier gain in fine gain scheme.

Fine gain index	RF 1 [dB]	RF 2 [dB]	RF 3 [dB]	RF 4 (PGA) [dB]	Gain [dB]
0	HiZ	HiZ	HiZ	HiZ	$-\infty$
1	-5	-10	-5	-10	-30
...	-5	-10	-5	...	...
25	-5	-10	-5	14	-6
26	-5	0	-5	4	-6
...	-5	0	-5	...	...
32	-5	0	-5	10	0
...	-5	0	-5	...	...
36	-5	0	-5	14	4
37	0	0	0	4	4
...	0	0	0	...	...
47	0	0	0	14	14
48	0	0	10	4	14
...	0	0	10	...	...
58	0	0	10	14	24
59	0	20	0	4	24
...	0	20	0	...	...
69	0	20	0	14	34
70	0	20	10	4	34
...	0	20	10	...	...
80	0	20	10	14	44
81	20	20	0	4	44
...	20	20	0	...	...
91	20	20	0	14	54
92	20	20	10	4	54
...	20	20	10	...	...
102	20	20	10	14	64

Table 4-7. Configuration look-up table used to setup the RF amplifier gain in coarse gain scheme.

Coarse gain index	RF 1 [dB]	RF 2 [dB]	RF 3 [dB]	RF 4 [dB]	Gain [dB]
0	HiZ	HiZ	HiZ	HiZ	$-\infty$
1	-5	-10	-5	-2.5	-22.5
2	-5	-10	-5	0	-20
3	-5	-10	0	-2.5	-17.5
4	-5	-10	0	0	-15
5	-5	0	-5	-2.5	-12.5
6	-5	0	-5	0	-10
7	-5	0	0	-2.5	-7.5
8	-5	0	0	0	-5
9	0	0	0	-2.5	-2.5
10	0	0	0	0	0
11	-5	0	10	-2.5	2.5
12	-5	0	10	0	5
13	0	0	10	-2.5	7.5
14	0	0	10	0	10
15	-5	20	0	-2.5	12.5
16	-5	20	0	0	15
17	0	20	0	-2.5	17.5
18	0	20	0	0	20
19	-5	20	10	-2.5	22.5
20	-5	20	10	0	25
21	0	20	10	-2.5	27.5
22	0	20	10	0	30
23	20	20	-5	-2.5	32.5
24	20	20	-5	0	35
25	20	20	0	-2.5	37.5
26	20	20	0	0	40
27	20	20	10	-2.5	47.5
28	20	20	10	0	50

4.3.2.2 Test signal generator

The picture in Fig. 4-15 displays the block diagram of the test signal generator along with example waveforms in the most important nodes. Please note that the signal waveforms presented in the diagram are not in scale. The generator can be used for diagnostics of the front-end processing when there is no beam signals on the front-end inputs. The pattern generator was implemented in the FPGA logic as a simple counter with controlled counting period. This circuit allows to setup frequency of the test pulses, simulating equivalently distributed bunches on the LHC circumference with configurable repetition rates. The circuit can be further extended to generate arbitrary bunch patterns including also the abort gap.

Signal from the bunch pattern generator is sent to a hazard circuit, which converts the FPGA signal into short pulses. The conversion is based on a usually undesired effect in the logic circuits when two combinatorial paths that have mismatched propagation delays which normally results in false glitches. The same effect is used in this implementation to generate short pulses. The leftmost NAND showed in the diagram delays the signal in one of the paths. Its propagation delay sets the output pulse duration which is about 1 ns. The following NAND in the chain compares the two paths producing a logical “0” when the FPGA signal changes from logical “0” to “1” (e.g. a rising edge). Otherwise its output stays constant on the logical “1” level. The advantage of this circuit is that FPGA circuits do not require high clock frequencies. The following inverter gate converts the signal to a positive logic. The inverter was realised with the two remaining NAND gates in the chip. Their outputs were connected together to increase the output current which drives the following stages. The inverter was realised with the two remaining NAND gates in the chip. Their outputs were connected together to increase the output current which drives the following stages.

The generated pulses from the hazard circuit are processed in a following RF blocks. The signal passes through a programmable attenuator which controls the pulse amplitudes in three discrete steps. The amplitude attenuation can be set to 0 dB, -20 dB and -40 dB simulating three different intensities of the beam. As the test signals are sent internally inside the front-end the signal from the attenuator are filtered by a non-reflective low-pass filter with 80 MHz cut-off frequency. It was built as a copy of the front-end input filters using the same circuit topology and component values.

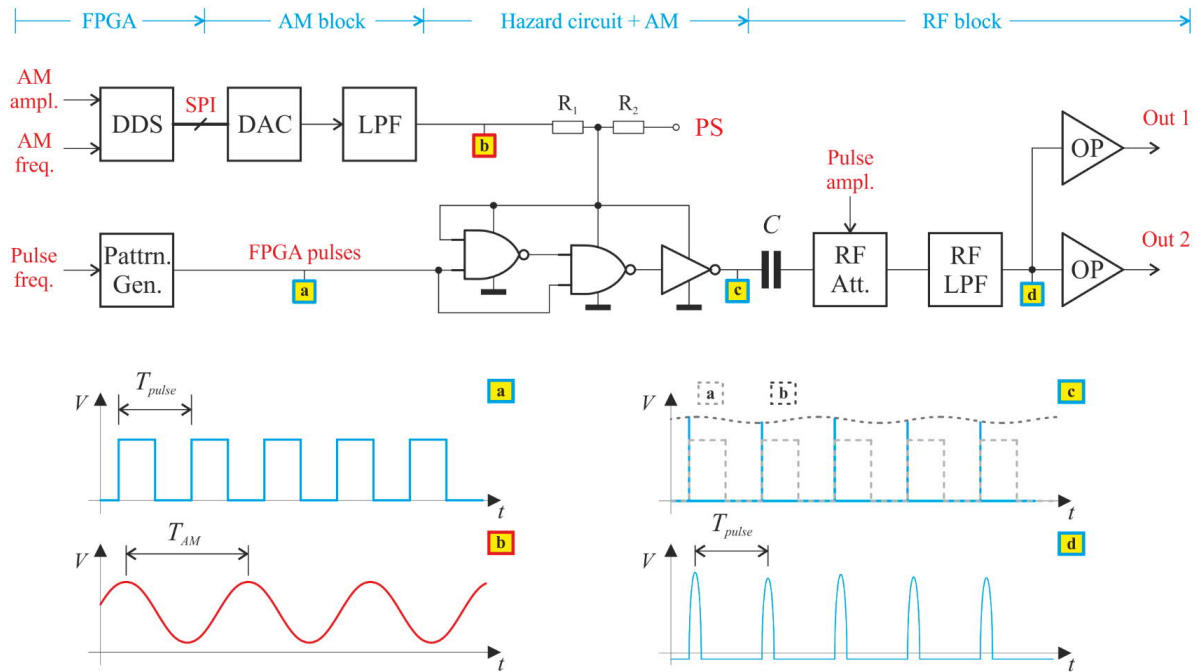


Fig. 4-15. Schematic diagram of the test signal generator with example signal in the most important nodes. **Abbreviations:** LPF – Low Pass Filter, RF – Radio Frequency, OP – Operational amplifier, DDS – Direct Digital Synthesis generator, DAC – Digital/Analogue Converter, PS – Power Supply of the hazard circuit, SPI – Serial Peripheral Interface bus.

Two amplifiers at the end of the RF chain buffer the test signals and provide them to the two analogue boards over coaxial cables inside the front-end. Such layout decouples the signals between the two boards and simplifies the front-end configuration. For example, one of the analogue boards may not be required for the tunnel installation and therefore it may not be installed in the front-end. In this case the output of the respective amplifier is RF terminated. The output stage was realised using operational amplifiers with current feedback providing large signal bandwidth of more than 1 GHz and slew rates over 5 V/ns. Both inputs and output signals on the amplifiers are AC coupled to limit the low frequency noise and interference signals between the boards.

An example of the generated test signals is showed in Fig. 4-16. The signals were measured with a 12-bit oscilloscope HDO6104 which has 1 GHz analogue bandwidth, less than a 0.5 ns rise time and sampling rate of 2.5 GS/s. The measurement displays the output signal with 40 MHz repetition rate and maximum amplitude (attenuator at 0 dB). As seen in the plot, the maximum achieved amplitude was 350 mV. The pulse width was approximately 6 ns when measured at 50 % of the amplitude.

The test signals are used mainly for the DOR channel diagnostics. However, the generator is also equipped with circuits allowing amplitude modulation of the pulses. This functionality can be used to test the oscillation channels in the front-end. For this reason a simple DDS module was implemented in the FPGA to produce the sinewave modulation signals with parametrized frequency and amplitude. The look-up table contains one quarter of the sinewave period which is stored in the FPGA memory. A pointer periodically reads the samples from the table and sends it to a 16-bit DAC over an SPI bus. The modulation frequency f_{AM} can be programmed by changing the pointer increment step. The frequency can be computed as:

$$f_{AM} = \frac{1}{T_{AM}} = f_s \frac{S_{DDS}}{L_{DDS}} \quad S \in \langle 1, L/2 \rangle \quad (4-22)$$

where f_s is the sampling frequency, S_{DDS} is the counter step and L_{DDS} the number of samples in one sinewave period.

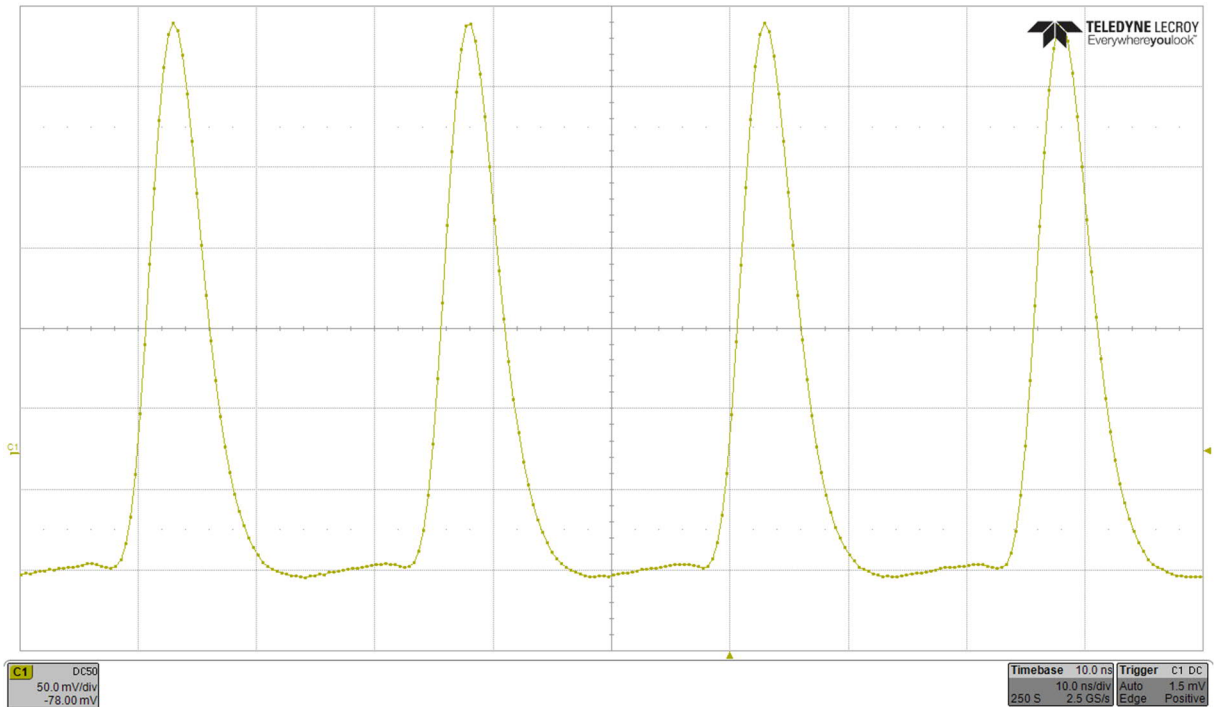


Fig. 4-16. Measurement of the test generator output signals. The generator was set to produce pulses with maximum amplitude (approx. 350 mV) and repetition rate (40 MHz). Displayed signal was acquired with a 12-bit oscilloscope with 2.5 GHz sampling rate, 1 GHz bandwidth and less than a 0.5 ns rise time (according to the datasheet).

The step S_{DDS} can be set in the interval ranging from 1 to $L_{DDS}/2$. The latter sets the maximum output frequency to half of the sampling frequency to satisfy Nyquist criterion. The L_{DDS} in this DDS implementation was chosen to be 100 samples. The DDS look-up table thus stores 25 sinewave samples formatted to the 32-bit scale. Programmable amplitude of the DDS signal allows controlling the modulation depth. The n-bit shifter register in the FPGA divides the amplitude of the DDS output signals by factor 2^n . The resulting samples are converted to 16-bit before they are sent to the DAC. Sampling can be synchronized to the LHC revolution frequency of approximately 11 kHz. According to equation (4-22), the f_{AM} can be controlled in the range from 110 Hz up to 5.5 kHz with 110 Hz steps. This is sufficient for the basic diagnostics of the oscillation subsystem. More advanced measurements with this DDS implementation would require much higher sampling rate requiring much larger look-up table for the same frequency resolution. An online sinewave calculation based on real-time algorithms such as CORDIC would be more efficient in terms of the FPGA resource usage.

Output signal from the DAC is sent to an analogue low-pass filter. It was built with three RC filter stages with consecutively decreasing cut-off frequency. It was foreseen to use the test signal output as a phase reference sinewave synchronized to the revolution frequency of the beam. For this reasons the phase response of the filter was designed to change by about 1° within the modulation frequency range between 0.5 Hz and 5 kHz. The first two sections of the filter remove the spurious DAC switching and other interference signals such as the 40 MHz SPI clock. The last section provides the lowest cut-off frequency of about 150 kHz. The narrowband signal processing in the DOS channels filters and removes aliased modulation components. The sampling frequency of the DAC can be increased up to 1 MHz to improve the efficiency of the low-pass filter.

The amplitudes of the RF pulses in the hazard circuit are modulated by superimposing the signal from the DAC to the power supply voltage of the NAND gates and output buffers. Resistors R_1 and R_2 showed in the diagram set the maximum AM depth of the pulsed signal. The diagnostic tests typically require small modulation depths. The resistor values were set such that the maximum amplitude from the DAC provides around 2 % modulation depth. The DDS can thus provide modulation depths of less than 0.01 %. To test the DOS subsystem the programmable amplifiers have to be configured with asymmetric gains. This is because the modulated test pulses are equally split between the channel pairs. The DOS subsystem demodulates the signals from each electrode on a dedicated RF diode detector. Following differential amplifier extracts the amplitude difference between the two channels. Without the asymmetric RF gain setting the test signals from the TSG would appear on the differential amplifier as common mode signals with equal amplitudes.

4.3.3 Compensated diode detectors and analogue low frequency signal processing

This chapter describes the analogue signal processing in the DOR subsystem starting from the compensated diode detector (CDD) up to the ADC. The Block diagram shown in Fig. 4-17 depicts two analogue channels for processing one pair of the BPM electrode signals. Each contains a compensated diode detector, low frequency processing and an ADC. The input signals are provided from the last stages of the RF amplifier, which drives the diode detectors.

As shown in the picture a CDD circuit is composed of two branches. One branch is composed of an RF detector containing a single Schottky diode D_1 . The second branch contains an RF detector with two Schottky diodes D_{2a} and D_{2b} connected in series. The three diodes in the two branches are accommodated in a single package. The diode detector in each branch is built with storage capacitor C , passive low-pass filter LPF_d and a configurable network composed of resistors R_1 and R_2 and the respective control switches. The LPF_d was implemented as a first order filtering stage with 15 kHz cut-off frequency to limit the signal bandwidth before the following amplifier stage. Two operational amplifiers OA_1 and OA_2 with JFET inputs and feedback resistors R_{oa} complete the CDD circuit. The amplifiers were optimized to achieve high precision and low noise operation. They also provide high input impedance and a small input current in order of pico-amperes. The four CMOS switches accommodated in a single component allow setting the time constant T_c of the CDD by controlling the capacitor discharge rate in both detector branches. The leakage current of the control switches is below a nanoampere when in “off” state. The signals S_1 and S_2 control the three possible T_c configurations. The control signals are sent in parallel to all on-board CDD blocks from a dedicated latch register addressed by the FPGA over parallel bus.

The slowest discharge rate (hereafter referred to as the time constant T_{c0}) can be set by disconnecting all switches in the CDD. When using this mode, the CDD operates as a peak detector. The discharge rate of the capacitors is defined by the leakage currents of the switches and amplifiers as well as the reverse currents of the diodes. The discharge is mostly dominated by the diode reverse currents which are in the order of nanoamperes. The DOR response with the T_{c0} was measured around 0.4 s (expressed as equivalent time constant for e^{-1} signal drop), which is equivalent to more than 4000 LHC beam turns. This time constant is suitable especially for measuring beam signals with low repetition rates as it provides almost no signal ripple in the detectors due to negligible diode currents. It also sets the smallest bandwidth of the CDD. For more details on achieved performance with T_{c0} please refer to the laboratory and beam measurements presented in this thesis.

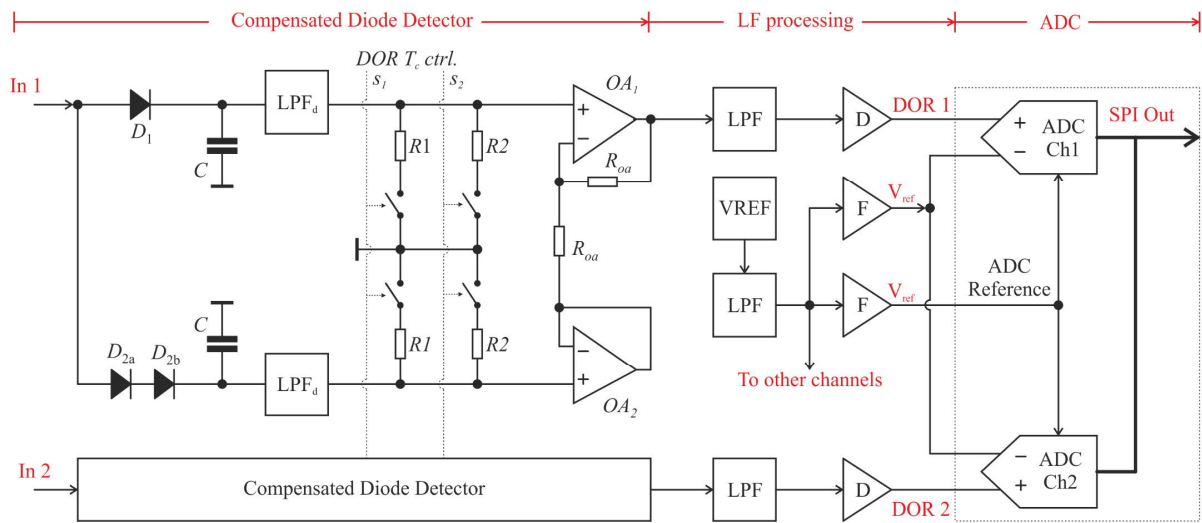


Fig. 4-17. Schematic diagram of diode detector processing and analogue-to-digital signal conversion.

Abbreviations: CDD – Compensated Diode Detector, LF – Low Frequency, LPF – Low-Pass Filter, VREF – Voltage Reference, D – ADC Driver, F – Follower, ADC – Analogue-Digital Converter, DOR – Diode Orbit, T_c – Time constant.

Faster CDD discharge rates can be programmed by connecting resistor R_1 or R_2 in parallel to the capacitors. The so called time constant setting T_{c1} uses S_1 to connect large resistors R_1 achieving discharge rates of about 100 ms, which is equivalent to some 1000 LHC beam turns. Even faster discharge rate can be set with S_2 that connects resistors 10-times smaller resistors R_2 . The corresponding time constant T_{c2} provides discharge rates of some 10 ms. One advantage of the faster modes is that the bandwidth for small signals is much larger than with T_{c0} . Another advantage is that the CDD has faster response time which is suitable for changing the RF gains or switching the calibration multiplexers. The fastest time constant is suitable for measuring only the beam signals with higher repetition rates such as the LHC physics beams. This detector setting yields large measurement errors when used with LHC pilot beams. For more details on the DOR performance with different T_c settings please refer to the laboratory and beam measurements presented in this thesis.

The following LF stages process the low frequency signals from the diode detectors. The filter on each CDD output was implemented as a 2nd order low-pass with Sallen key topology [48] using with low noise precision amplifiers. The cut-off frequency of the filter was chosen to be about 100 Hz removing strong frequency components related to the bunch repetition rate and most of the noise above this frequency. Decreasing the filter cut-off frequency would help to further remove the noise. However, slow dynamic response of the LF filters would limit the performance of the system when calibration switching.

The next stage in the LF processing are the differential ADC drivers located on the ADC board. The drivers were built with low-noise precision amplifiers. Their function is to receive the LF signals sent over flat cable from the analogue board and drive capacitive loads of the ADC input circuits. The two converters showed in the picture Fig. 4-17 are part of a single package which contains 8 simultaneously sampled channels of 24-bit sigma-delta precision ADCs. The differential inputs are compared to the V_{ref} voltage provided from a 2.5 V external reference chip VREF. The voltage is low-pass filtered to 0.01 Hz and buffered with the voltage followers.

The inverting inputs of the ADCs are connected to the buffered V_{ref} in order to utilize the entire ADC dynamic range for sampling only positive DC signals. The output samples from the digitizers are provided in two's complement format [72]. If the input signal is equal to 0 V, the corresponding value of the ADC sample is equal to code 0x800000 (expressed in 24-bit hexadecimal format). If the input signal level is equal to the reference voltage V_{ref} , the corresponding ADC output value is equal to code 0x000000. If the input signal level is equal to 5 V the ADC is saturated and the corresponding ADC output value is equal to code 0x7FFFFFFF. The voltage reference for each DOR pair is individually buffered by a follower. This scheme improves symmetry and reduces possible signal cross-talks between the paired DOR channels.

The ADC sends its conversion results to the FPGA over an SPI interface. The sample readout is triggered by a data ready signal (DRDY) provided synchronously to the sampling frequency from the ADC. Upon each DRDY trigger the FPGA sends the serial clocks and synchronously reads the ADC data. The data is transmitted over SPI in a form of serialized frames. One frame consists in total 24 bytes which is all eight 24-bit samples from all ADC channels. Received data in the FPGA is then de-serialized and sent to the dedicated DSP channels for further processing.

4.3.4 Digital signal processing

The aim of the DOR digital signal processing is to filter the signals and reduce the data rate such that the resulting samples can be easily transmitted over Ethernet network. The diagram showed in Fig. 4-18 displays two DSP channels processing ADC samples from two DOR channels of one electrode signal pair. The DSP blocks showed in the diagram were implemented in the front-end FPGA. Remaining DOR channels which are not displayed in the picture are processed by identical copies of the DSP chains.

The SPI module on the input of the signal processing chain reads and de-serializes the ADC data frames sent upon the ADC DRDY trigger, synchronised to the f_{rev} . The module then provides the received samples from all eight ADC channels to the respective DSP channels over a parallel bus. Decoder blocks on each DSP input latch samples of the respective channels and forward them for further processing. The following block in the chain converts the received DOS samples from the two's complement into unsigned integer format. The conversion is done by adding half of the ADC full-scale followed by removing the overflow bit after the addition. Next stage in the DSP applies notch filtering to remove the $f_{rev}/4$ frequency harmonics from the signal. This way the notch filter helps to attenuate the frequency components which may be present especially when measuring position of single bunch beams. The next stage are the IIR low-pass filters with controlled cut-off frequency. The low-pass filtering can be adapted to various requirements such as the low cut-off for noise and interference filtering or higher cut-off for faster dynamic response during calibration switching. The filters can be also set to decrease possible anti-aliasing which may occur after the signal decimation in the following decimation block. The filters can be also bypassed or stopped which is mostly used when calibration switching. The controls of the IIR filters between the two analogue boards are separated. The last DSP stage in the FPGA logic decimates the samples to decrease the sampling rate f_{rev} to the f_{tx} rate of the UDP transmission (equal to 25 Hz). The decimation factor d is defined as:

$$d = \frac{f_{rev}}{f_{tx}} \quad (4-23)$$

The factor d is not an integer number as the frequencies of the two frequencies are not harmonically related. Moreover, the f_{rev} cannot be assumed constant during the LHC operation as it changes according to the energy of the particles in the machine.

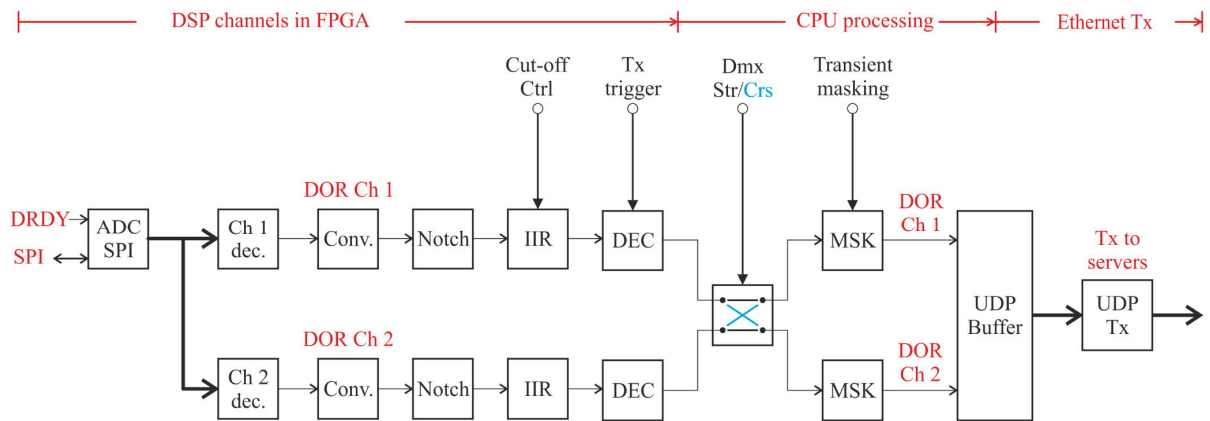


Fig. 4-18. Digital signal processing of two DOR channels performed in a front-end.

Abbreviations: SPI – Serial Peripheral Interface, IIR – Infinite Impulse Response filter, DEC – data rate Decimation, MSK – Masking of the switching transients, UDP – User Datagram Protocol

The decimation modules act as “mailboxes” that latch the samples from the IIR filter upon the UDP Tx trigger. Mismatch between the ADC sampling frequency and the UDP sending frequency is translated into jitter of the transmitted samples. With the LHC f_{rev} of 11245.5 Hz [3] the resulting decimation factor is according to equation (4-23) equal to 449.82. This means that each 449th or 450th sample will be registered in the UDP buffer and sent to the remote servers. This sample count is sent in each UDP together with the DOR data. To avoid aliasing during the decimation process the IIR cut-off can be set such that it matches to the bandwidth of the decimated signal. In addition the sampling frequency of the ADC A and B boards may be locked to the f_{rev} of different beams. The UDP Tx trigger therefore synchronizes the respective DSP channel. Decimated DOR signals after the DSP are registered and further processed by the CPU. Samples and other data from the front-end is gathered upon the Tx trigger and inserted into the UDP buffer. The data is then sent to the DOROS servers.

The calibration switching requires the de-multiplexer and transient masking blocks. The de-multiplexer corrects the sign in the calibrated beam position when the RF multiplexer is in the “crossed” state (indicated as *Crs* in the diagram). After each signal multiplexing the transient masking block excludes a defined number of samples from the calibration processing. The processing block inserts a zero value instead of the DSP data, marking the results as invalid. This data is not taken into account when computing the calibrated positions on the server side. Both de-multiplexing and masking stages are synchronously controlled from a dedicated state-machine running in the FPGA CPU. More information on the implementation of the so called automatic calibration state-machine is provided later in this chapter. When the automatic calibration is not in use the de-multiplexer is in “straight” state (indicated as *Str* in the diagram) and the masking is bypassed. The DSP also allows to invert position sign for each individual electrode pair. This is done by inverting the controls of the selected de-multiplexers in the front-end.

The remaining signal processing and position calculation is performed on the DOROS server side. The DOR data is treated according to the mode of operation. In the simple mode (no calibration switching) the server further averages the electrode data so that it can be sent to other servers and databases such as the logging or collimator control. In the calibration mode the switched electrode signals are processed by the calibration algorithm.

Source of the Tx trigger in the diagram Fig. 4-18 is provided from an output of a digital PLL based on a simple counter implemented in the FPGA. The counter is clocked from a local 40 MHz crystal oscillator. The free-running frequency of the PLL was set to 24.75 Hz providing 1 % margin for the servers receiving the UDPs. The PLL can be locked to the external triggers received in the BST messages. Once the triggers in the BST messages are received, the PLL locks to them and provides the trigger pulses to the DSP and UDP sending modules. Locking bandwidth of the PLL is limited to some 100 Hz. The orbit data is thus automatically sent in the UDP even in case of missing BST fibres or an error in the BST message decoder. The trigger PLL improves the reliability and robustness of the system. The functionality is crucial especially for the interlocked front-ends, as missing UDP data from such a front-end for more than 1 minute triggers asynchronous beam dump.

4.3.4.1 The notch filter

In order to obtain the magnitude and phase characteristics of the notch filter, its transfer function $H_{notch}(z)$ was expressed in Z domain:

$$H_{notch}(z) = 1 + z^{-1} + z^{-2} + z^{-3} \quad (4-24)$$

where z is the complex variable. Using the Discrete Time Fourier Transformation (DTFT) the $H_{notch}(\theta)$ function can be expressed as:

$$H_{notch}(\theta) = 1 + e^{-j\theta} + e^{-2j\theta} + e^{-3j\theta} \quad (4-25)$$

where θ is the normalised digital frequency related to the input signal frequency f as:

$$\theta = 2\pi \frac{f}{f_{rev}} \quad \theta \in \langle 0, 2\pi \rangle \quad (4-26)$$

The roots of the transfer function (4-25) correspond to the notch frequencies located on the unity circle at $\theta = \pi$ (real zero) and at $\theta = \pm\pi/2$ (complex zeros). Resulting magnitude and frequency characteristics of the transfer function are showed in the Fig. 4-19. Horizontal axis was expressed in frequency normalised to the ADC sampling rate. The vertical axis of the magnitude response was not normalised in order to show the processing gain of the filter for DC signals. The DC gain seen in the plot results from the three additions in equation (4-25).

The notch filter was implemented in the FPGA logic with three delay registers and a parallel adder unit. The resolution of the input signals from the channel decoders is 24-bits. The resolution of the output samples after the addition in the notch filter is 26-bits. Each sample is then converted to a 32-bit representation by performing 6-bit left-shift. Obtained results are sent to the following IIR filtering stage.

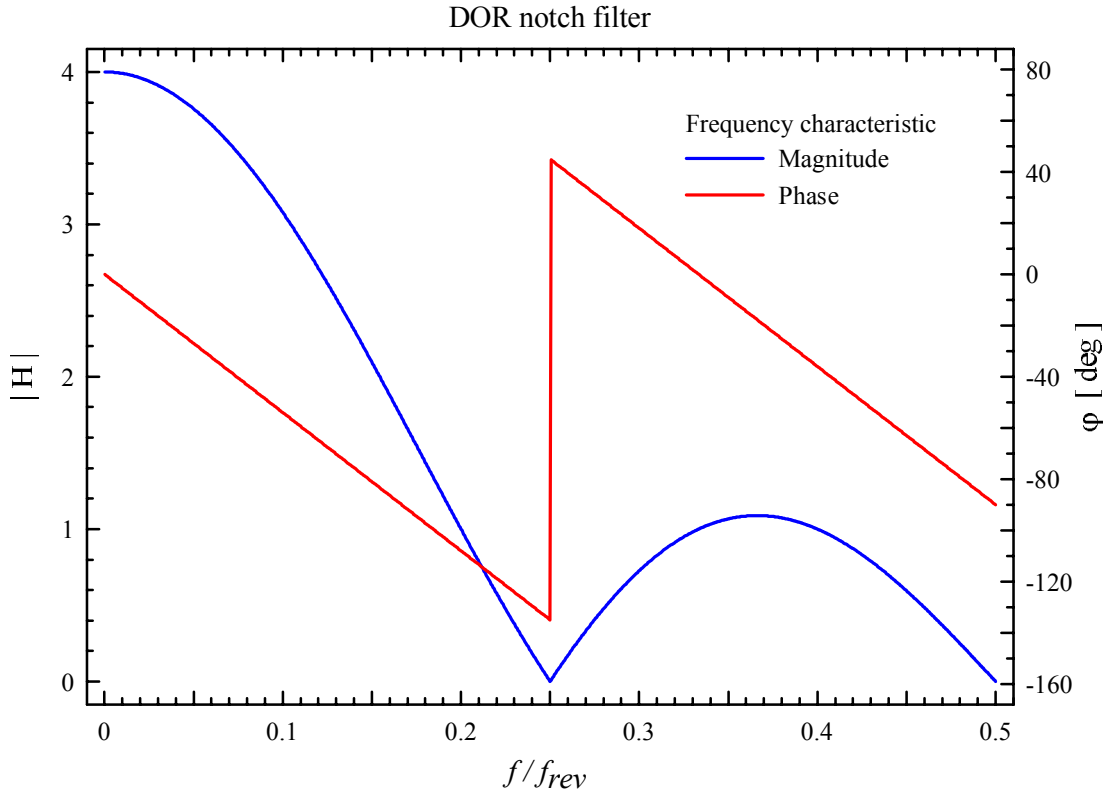


Fig. 4-19. Magnitude and phase characteristics of the notch filter used in the DOR DSP channels.

4.3.4.2 The IIR low-pass filter

The IIR filters of the DOR processing were built as a 1st order low-pass filters. Their cut-off frequency is programmable so that the filtering can be optimised for different filtering requirements. Additional functionality allows to bypass or freeze the filter. The register values can be retained or cleared which is useful when the input signals are multiplexed during calibration switching. The transfer function of the IIR filter $H_{IIR}(z)$ can be expressed as:

$$H_{IIR}(z) = \frac{1 - \alpha}{1 - \alpha z^{-1}} \quad (4-27)$$

where z is the complex variable. The coefficient α is a parameter which allows changing the cut-off frequency of the filter by controlling its single real pole. The function can be rewritten using the DTFT into the following equation:

$$H_{IIR}(\theta, \alpha) = \frac{1 - \alpha}{1 - \alpha e^{-j\theta}} \quad (4-28)$$

where θ is the normalised digital frequency. The characteristic has also a single zero located in the middle of the unity circle. The boundaries of the parameter α are defined on the interval $0 \leq \alpha \leq 1$. When $\alpha = 1$ the pole is located at the centre of the unity circle and the filter has 0 output. When $\alpha = 0$ the real pole is located on the unity circle and the filter acts as an all-pass with gain 1. The distance of the pole from zero increases with the parameter. Dependency of the cut-off frequency on the parameter α was obtained by finding the magnitude of the transfer function at filter gain $1/\sqrt{2}$ (equivalent to -3 dB):

$$|H_{IIR}(\theta, \alpha)| = \frac{1}{\sqrt{2}} \quad (4-29)$$

Solving the equation (4-29) yields normalised cut-off frequency θ_c which is then expressed as:

$$\theta_c = \pm \arccos\left(2 - \frac{1}{2\alpha} - \frac{\alpha}{2}\right) \quad (4-30)$$

The equation is valid for $0 < \alpha \leq 1$. The function which converts the normalised cut-off frequency into parameter α can be found by inserting (4-26) into equation (4-30). Obtained result describes α as a function of the cut-off frequency f_c/f_{rev} for $0 \leq f_c/f_{rev} \leq 0.5$ which is expressed as:

$$\alpha(f_c/f_{rev}) = 2 - \cos\left(2\pi \frac{f_c}{f_{rev}}\right) - \sqrt{3 - 4 \cos\left(2\pi \frac{f_c}{f_{rev}}\right) + \cos\left(2\pi \frac{f_c}{f_{rev}}\right)^2} \quad (4-31)$$

The function is used to compute α parameters used to control the IIR cut-off frequency. The plot of the function (4-31) is displayed in Fig. 4-20. As can be seen, increasing the cut-off frequency results in decreasing the IIR coefficient. Please note that the boundary condition of $f_c = 0.5 f_{rev}$ is still valid as the filter has at this frequency -3 dB gain. The equation (4-31) was also used to compose a look-up table for the IIR cut-off control. The coefficients α were computed with sampling frequency f_{rev} assumed to be equal to 10 kHz. The values in the FPGA look-up table are summarized in Table 4-8.

Resulting magnitude and phase characteristics of the IIR filter are shown in Fig. 4-21 and Fig. 4-22. Displayed characteristics were plotted in logarithmic scale for a few selected α coefficients from Table 4-8. Main purpose of the IIR filter is to attenuate noise and possible interferences from the slowly changing orbit signals. For most of the operation time the DOR filtering typically requires the cut-off frequencies in the range between 10 Hz and 0.1 Hz.

Computing the equivalent noise bandwidth $ENBW_{IIR}$ of the IIR filter allows to express its filtering performance equivalent to a brick filter. The $ENBW_{IIR}$ can be expressed as:

$$ENBW_{IIR} = \int_0^\pi |H_{IIR}(\theta, \alpha)|^2 d\theta \quad (4-32)$$

where θ is the normalised digital frequency and α cut-off parameter defined on the interval $0 \leq \alpha \leq 1$. Integration of the substituted H_{IIR} from equation (4-28) yields the $ENBW_{IIR}$ as a function of the cut-off parameter α :

$$ENBW_{IIR} = \left| \pi \frac{\alpha - 1}{\alpha + 1} \right| \quad (4-33)$$

The $ENBW_{IIR}$ values for different α settings are presented in Table 4-8.

The impulse response of the filter can be found through inverse Z transformation of the transfer function $H_{IIR}(z)$. Normalised impulse response $h(n)$ is then found as a function of sample index n representing the discretized time:

$$h(n) = \alpha^n \quad (4-34)$$

The decay rate of the impulse response is important when multiplexing the signals during the calibration process. Amplitude of the signal step on the IIR input depends on the difference between the electrode signals (proportional to the beam position) as well as the asymmetries between the analogue channels. The equation of the signal decay rate τ_n after a full-scale signal change can be expressed as:

$$\tau_n = \frac{\log(\varepsilon)}{\log(\alpha)} \quad (4-35)$$

where α is the IIR cut-off parameter and ε is the signal level after time duration τ_n expressed in number of ADC samples. As the ADC sampling is synchronized to the f_{rev} , the decay rate can be also expressed as number of beam revolution periods T_{rev} . The τ_n values for the corresponding α parameter were summarized in Table 4-8. The values were computed for a 63 % change in the signal level ($\varepsilon = e^{-1}$) and expressed in equivalent LHC revolution periods T_{rev} . For example, response of IIR filters to a full-scale change in the input signal would decay for more than 15 kT_{rev} if the cut-off is set to 0.1 Hz. Such delay could significantly limit the frequency of the multiplexer switching during the orbit calibration. The filter cut-off has to be thus dynamically controlled during each multiplexer transient to decrease the response time and at the same time provide sufficient low-pass filtering. More details on the dynamic IIR control can be found in the following chapters.

The IIR filter is built with two 32-bit multipliers and one 65-bit accumulator. The multipliers were implemented in FPGA hardware blocks. Registers are clocked at 80 MHz synchronously with the CPU. The output results are rounded to 32-bit values by removing the least significant bits. Obtained results from the IIR filters are then provided to the decimation stage of the DSP. The parameter α is stored in the CPU look-up table represented by a 32-bit integer number. The CPU maps dedicated peripheral registers to provide the parameter α to the respective IIR filters. Decoupled IIR control between the DSP chains of the two analogue boards provides flexibility especially when using the calibration multiplexers.

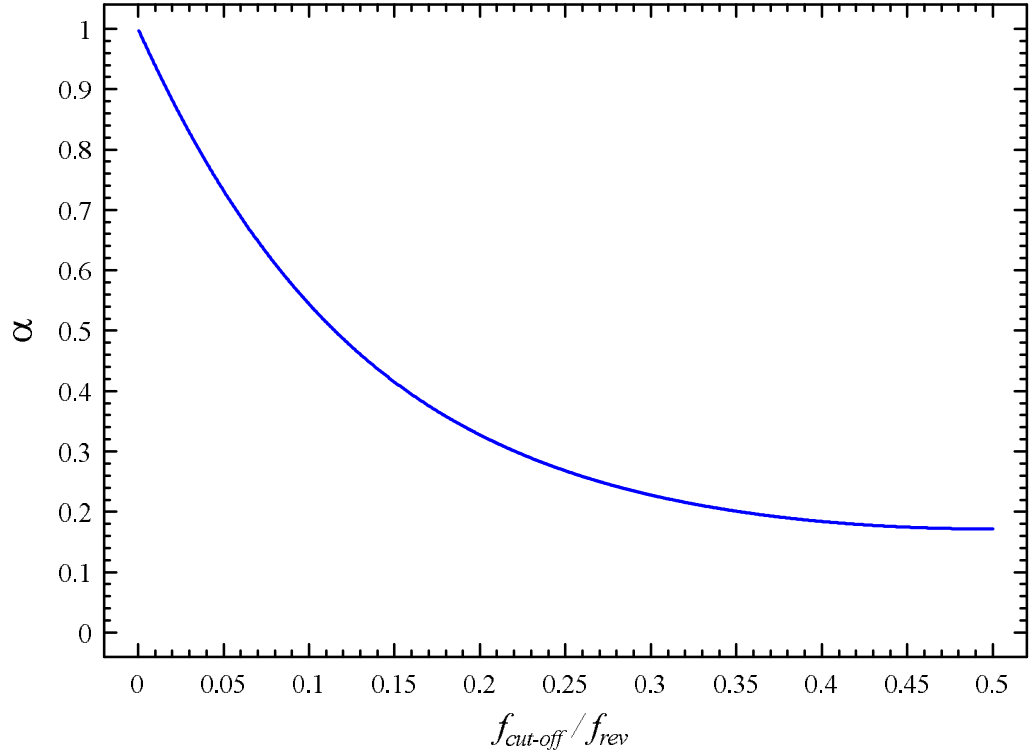


Fig. 4-20. Dependency of the parameter α on the normalised cut-off frequency of the IIR filter. The function is analytically described by (4-31).

Table 4-8. Look-up table used to map the IIR cut-off frequency f_c to the IIR coefficient α . The table also contains the equivalent noise bandwidth and decay time computed for the respective coefficient.

f_c / f_{rev}	f_c [Hz] †	Coefficient α		ENBW		Decay to $\varepsilon = e^{-1}$
		Normalised	32-bit scale	$ENBW / f_{rev}$ *	$ENBW$ [Hz] ‡**	τ_n [LHC T_{rev}]
0	0	1	4294967295	0	0	∞
0.000001	0.01	0.999994	4294940309	0.000002	0.016	159155
0.000002	0.02	0.999987	4294913323	0.000003	0.031	79578
0.000003	0.03	0.999981	4294886338	0.000005	0.047	53052
0.000005	0.05	0.999969	4294832367	0.000008	0.079	31831
0.00001	0.1	0.999937	4294697443	0.000016	0.157	15916
0.00002	0.2	0.999874	4294427607	0.000031	0.314	7958
0.00003	0.3	0.999812	4294157789	0.000047	0.471	5306
0.00005	0.5	0.999686	4293618203	0.000079	0.785	3184
0.0001	1	0.999372	4292269535	0.000157	1.571	1592
0.0002	2	0.998744	4289573470	0.000314	3.142	796
0.0003	3	0.998117	4286879100	0.000471	4.712	531
0.0005	5	0.996863	4281495441	0.000785	7.854	319
0.001	10	0.993737	4268065910	0.001571	15.708	160
0.002	20	0.987512	4241333547	0.003142	31.415	80
0.003	30	0.981328	4214769663	0.004712	47.121	54
0.005	50	0.969075	4162145131	0.007853	78.527	32
0.01	100	0.939121	4033492994	0.015698	156.976	16
0.02	200	0.882057	3788405719	0.031334	313.336	8
0.03	300	0.828665	3559086932	0.046847	468.472	6
0.05	500	0.732269	3145073418	0.077277	772.774	4
0.1	1000	0.544113	2336948484	0.147621	1476.209	2
0.2	2000	0.327376	1406070064	0.253366	2533.659	1
0.3	3000	0.227777	978295213	0.31448	3144.801	1
0.5	5000	0.171573	736899888	0.353553	3535.534	1

* Rounded to 6 decimal places, ** Rounded to 3 decimal places, † f_{rev} assumed to be 10 kHz.

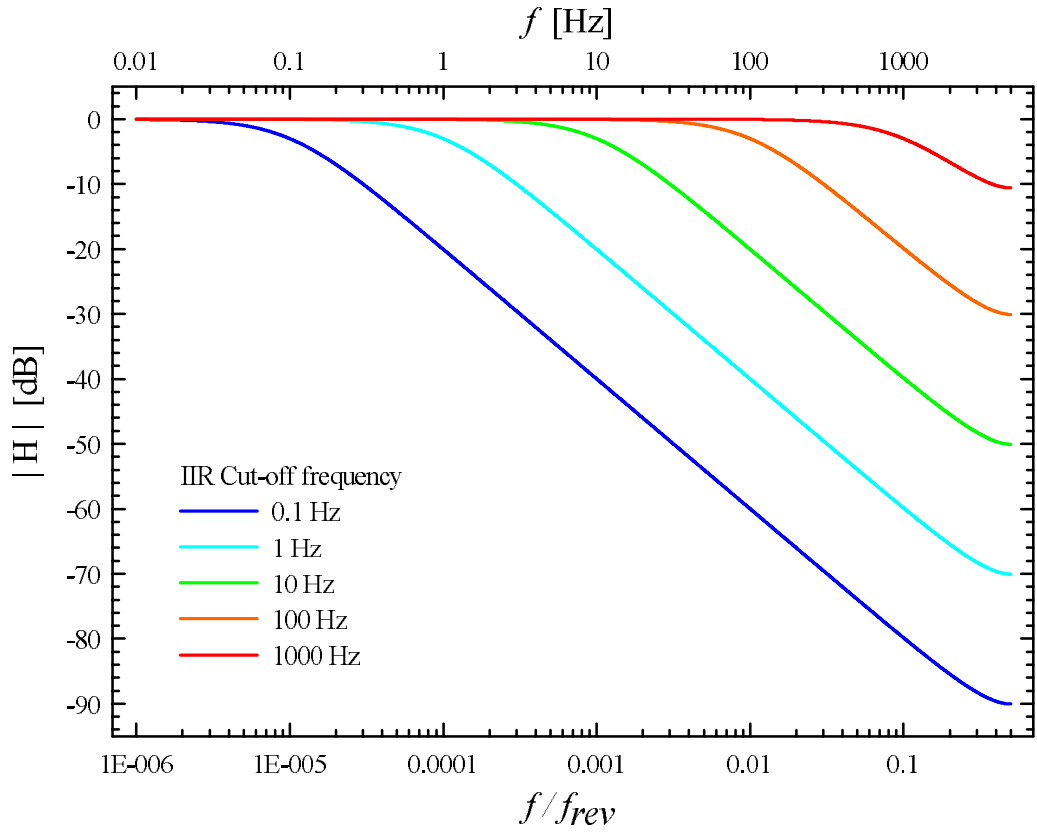


Fig. 4-21. Magnitude frequency responses of the IIR filter plotted for five different IIR cut-off values. The absolute sampling frequency of the filter which is usually synchronous to the LHC revolution was rounded to 10 kHz.

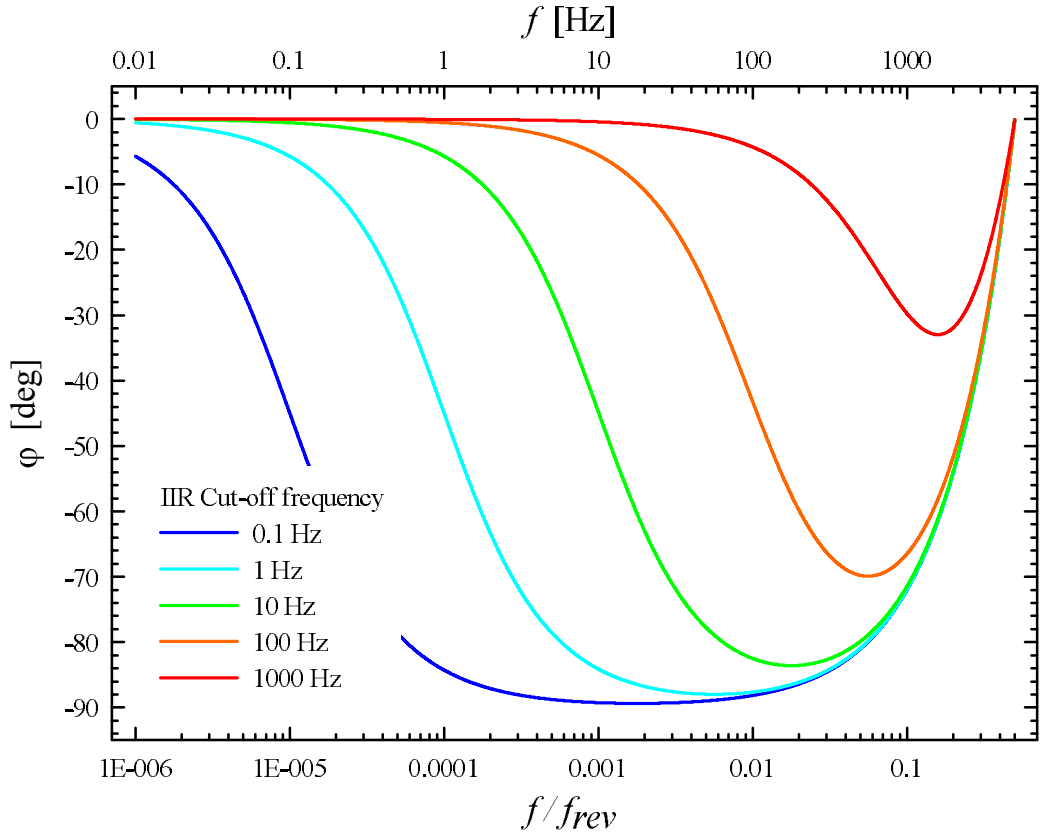


Fig. 4-22. Phase frequency responses of the IIR filter plotted for 5 different f_c values. The absolute sampling frequency of the filter which is usually synchronous to the LHC revolution was rounded to 10 kHz.

4.3.4.3 Automatic gain control

The RF amplifier adjusts the amplitude of the BPM signals so that they can be efficiently processed by the diode detectors. As the signal amplitude changes according to the actual beam intensity and position, the RF gain needs to be adjusted. The RF gain control is provided by the FPGA system configuring the individual RF stages according to a dedicated look-up table.

A so called Automatic Gain Control (AGC) is a state-machine which monitors the DOR signal levels and applies RF gain corrections necessary to keep the signals within programmed thresholds. The state-machine can be globally enabled or disabled depending on the operation mode. Block diagram in Fig. 4-23 shows the AGC implementation in the CPU firmware. The state-machine monitors the signal level and applies the gain correction at a 25 Hz rate synchronously to the UDP Tx trigger. Each of the two analogue boards has a dedicated AGC to decouple the gain control.

The AGC signal regulation is based on the largest signal level of all four electrodes. This is because the gain correction is applied to all RF channels in parallel. Remaining channels with smaller amplitudes follow this adjustment avoiding saturation of the respective ADC channels. The maximum signal level is then compared with a set of programmable thresholds in order to determine if a gain correction needs to be applied. The optimisation goal for the AGC is to keep the signals in the range within the minimum and maximum boundaries. If the signal level is above the maximum threshold Th_{max} the RF gain needs to be decreased to the level below the hysteresis threshold Th_{hmax} , which is defined as $Th_{max} - Hyst$. If the signal level is below the minimum threshold Th_{min} the gains need to be increased to the level above the hysteresis threshold Th_{hmin} , which is defined as $Th_{min} + Hyst$. The AGC then automatically applies increasing or decreasing steps of the RF gain value. The entire process repeats until signal level reaches the optimisation goal set by the thresholds. In order to allow the DOR signals to stabilise after a gain change the AGC waits for a programmed period of time before applying next gain change. The signal monitoring and threshold comparison are disabled during the waiting time. The size of the step as well as the choice of the control look-up table can be configured with dedicated UDP commands. The AGC can also control the detector time constants and cut-off frequency of the IIR filter to improve the transient response of the DOR channel. If enabled, the controller can automatically speed-up the transient response of the DOR channels during each gain step. The speed-up is realized by temporarily applying a faster discharge time constant to the CDDs (typically T_{c2}) and rising the cut-off in the IIR filters (typically 1 kHz). After a configured period of time the CDD and IIR parameters are restored to their previous values. This technique allows to use low frequency cut-off in the DSP chain when the signal levels are stable within the programmed threshold region.

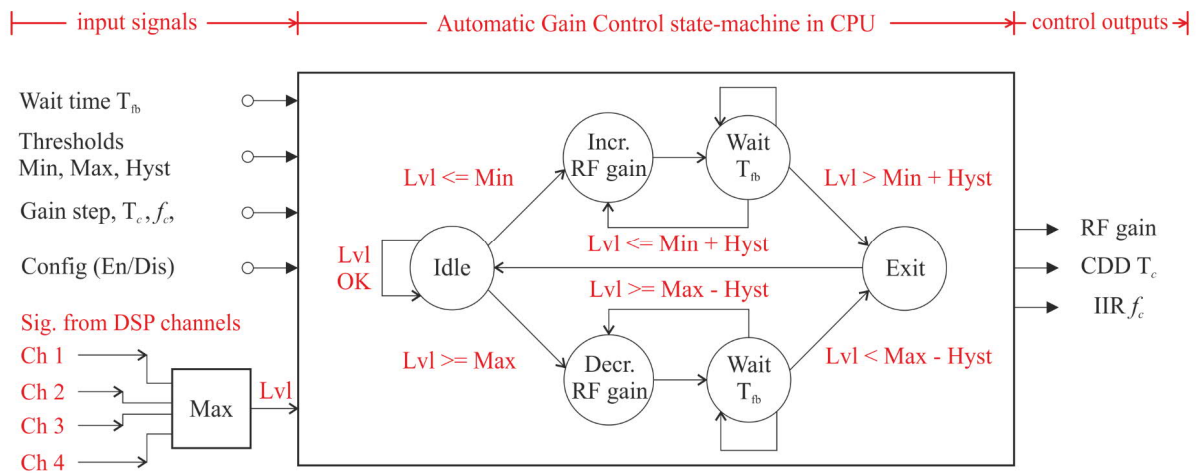


Fig. 4-23. Block diagram of the automatic gain control state-machine implemented in the CPU firmware.

Picture in Fig. 4-24 shows an example of AGC signal control during a typical LHC beam cycle. The plot on the top of the picture displays the maximum signal level from the four BPM electrodes. This signal is used by the AGC for the threshold comparisons. The green region on the left of the plot depicts the threshold range between the Th_{hmin} and Th_{hmax} . This region represents an optimal signal level range and an optimisation goal of the AGC. The yellow region depicts the hysteresis range around the Th_{min} and Th_{max} thresholds. The hysteresis region represents a warning for the AGC that the signals are close to the actual boundary. The AGC does not control the RF gain until the signal level crosses the Th_{min} and Th_{max} thresholds. The red region depicts the signal range in which the AGC adjusts the gains such that the signal level is in the green region within the hysteresis thresholds.

The plot on the bottom of the Fig. 4-24 depicts the beam intensity (green) and the RF gain value (magenta) as a function of time. In this example the beginning of the time axis corresponds to the no-beam condition in the LHC. When there are no BPM signals present on the input, the AGC waits for the first beam injection with the gain set to the allowable programmed maximum. The maximum gain in this example was set such that the AGC has to apply only a small correction during the injection of the low intensity pilot beam.

However, the beam intensity changes the most during the first nominal injection, typically by an order of magnitude. This change results in a large increase in the signal level requiring the AGC to decrease the gain in several gain steps. As seen in the picture, the ADC channel is saturated for a short period of time. The saturation can last for a few seconds before the gains are regulated to optimal values below Th_{max} . This duration strongly depends on the AGC programmed dynamic response. The following injections typically don't significantly increase the maximum signal level. In this example the AGC did not have to further correct the gains as the signal level still remained in the allowable (yellow) region.

Throughout the cycle the AGC may need to regulate the gains also during a change in the beam position. Moreover, the signal level in the collimator system is a function of the BPM aperture which is typically adjusted during the ramp and the optics change before the beam collisions, so called squeeze.

The beam intensity decreases during physics when particles collide in the experiments. In order to compensate the intensity decay the AGC increases the RF gains each time the signal level crosses the Th_{min} threshold, increasing the signal level. The thresholds in the AGC can be configured to regulate the gains always in a single step. This can be achieved by reducing the hysteresis region (yellow) to less than the smallest gain step. The gain regulation compensating the intensity decay has to be typically repeated several times until the beams are dumped at the end of the LHC cycle. The AGC then regulates the RF gain back to the maximum allowed value and waits until the pilot injection in the next cycle.

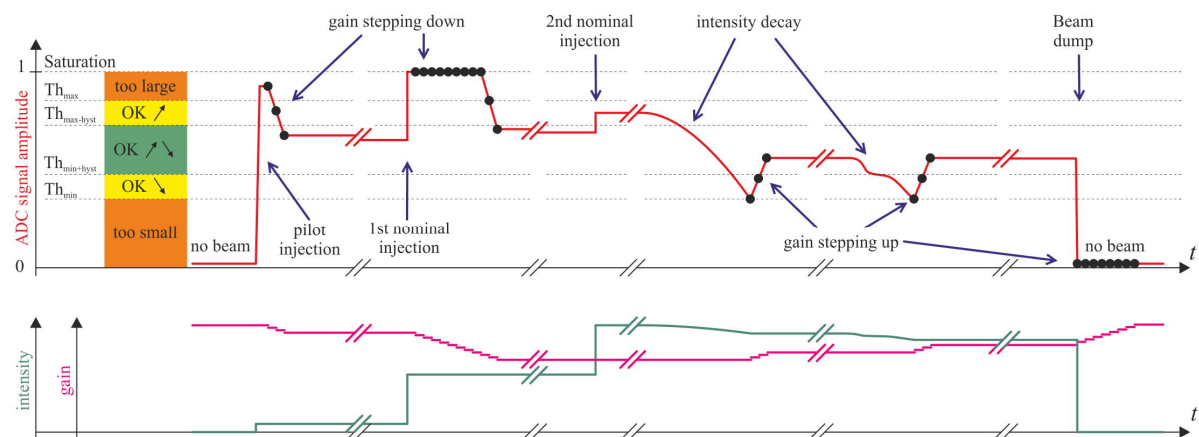


Fig. 4-24. Example of the automatic gain control during one LHC physics cycle.

4.3.4.4 Automatic calibration

The asymmetries between two DOR channels cause significant errors in the corresponding position measurements. The asymmetry of the channel gains can be cancelled by periodic multiplexing of the electrode signals on the inputs of the DOR channels. The diagram in Fig. 4-25 summarizes the main front-end blocks which process one DOR channel pair. The diagram also depicts the parameters used by a so called Automatic Calibration (AUCAL) state-machine which controls the calibration process in the front-end. The AUCAL synchronizes and controls the input multiplexer state, discharge time constant T_c of the CDDs, cut-off frequency of the IIR filters, sign correction in the de-multiplexers and duration of the transient masking. The state-machine allows minimization of the transient duration after each multiplexer switching. This is achieved by sequenced control of the CDD time constants and IIR cut-off frequencies. Fast switching transients allow orbit calibration at higher frequency which decreases the latency of the calibrated orbit measurement. This way the transient masking can be also decreased which improves the precision of the orbit readings after averaging. The AUCAL state-machine is clocked at 25 Hz rate synchronously to the UDP Tx trigger.

Schematic diagram of the calibration state-machine implemented in the front-end CPU firmware is showed in Fig. 4-26. The two analogue boards in the front-end are each controlled by a dedicated AUCAL state-machine with separate settings and DOR chain controls. The input parameters and settings for the state-machine are showed on the left of the picture. The controlled signals which are sent to the respective DOR processing blocks are showed on the right. If disabled, the state-machine is waiting in an idle state. Once enabled the state-machine enters a looped cycle. The straight and crossed states of the state-machine correspond to the state of the input multiplexer and the sign correction de-multiplexer. While in either of this two states, the state-machine temporarily increases the bandwidth of the CDDs and IIRs filters, performing a so called transient speed-up. At the same time the transient DOR samples are masked by sending zero values in the UDP datagrams. The state-machine enters the waiting states after each switching state. The multiplexing period T_{mx} defines the duration of the calibration switching period. While waiting, the nominal CDD and IIR settings are restored and the transient masking is stopped. The duration of each applied setting during the switching can be individually configured. All controlled features can be also individually enabled or disabled. The entire cycle repeats with alternating straight and crossed configuration of the multiplexers. The loop is stopped through the exit state when the state-machine is disabled in any time of its operation. When in exit state the multiplexers are returned to straight position and the nominal CDD and IIR settings are restored.

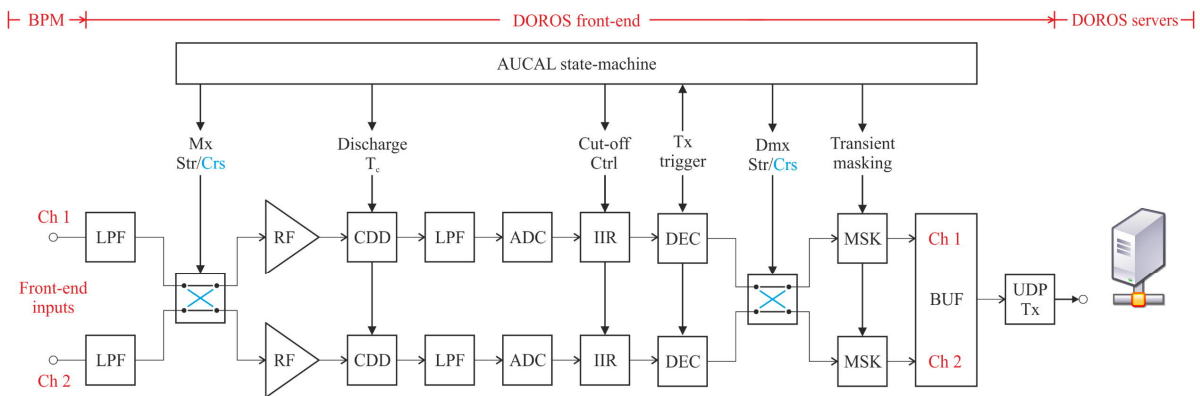


Fig. 4-25. Block diagram of the complete DOR chain with blocks controlled by the automatic calibration controller.

Abbreviations: LPF – analogue Low Pass Filter, CDD – Compensated Diode Detector, RF – Radio Frequency amplifier stage, IIR – digital low pass filters, DEC – sample decimation and UDP synchronisation, MSK – transient sample Masking, AUCAL – Automatic Calibration.

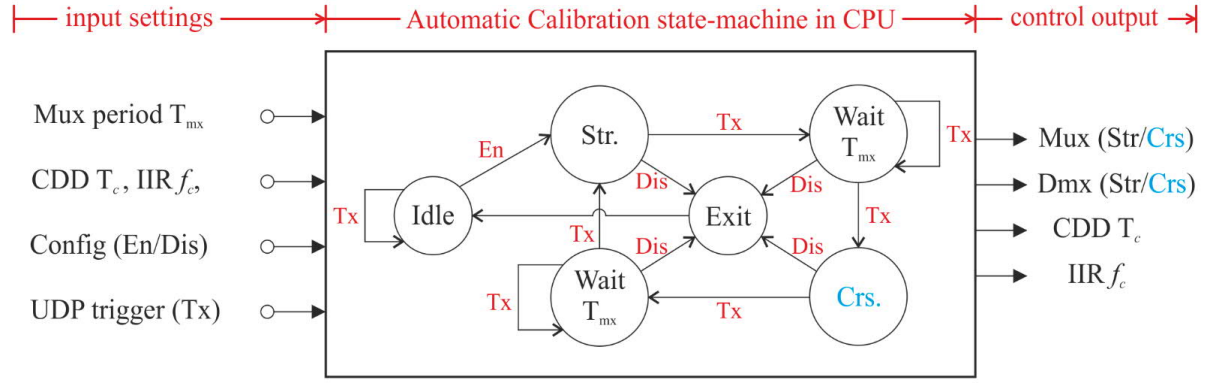


Fig. 4-27. Schematic diagram of the automatic calibration state-machine implemented in the CPU firmware.

The picture in Fig. 4-26 shows an example of the DOR signals and the parameter control with AUCAL set to 1 Hz switching. The horizontal axis corresponds to the indices of the consecutive UDP Tx triggers coming at the 25 Hz rate. The vertical axis in the picture displays the signals from DOR channels L and R along with the parameters controlled by the AUCAL state-machine. This example shows the horizontal BPM plane but the same processing is applied to all planes of the front-end. The picture depicts two consecutive switching periods which correspond to one full cycle through AUCAL state-machine. The trigger with zero index sends the last sample before switching the multiplexers to the crossed state. At that time the IIR cut-off was set to 0.1 Hz and the CDD to the slow discharge T_{c0} . After 1st trigger the multiplexers were switched to crossed state. At this time the DOR samples start to be masked by sending zero values in each UDP. The IIRs are in freeze mode when the registers retain the values stored before the transient. The CDD time constant was set to fast discharge T_{c2} . In the following trigger, the IIR cut-off was set to 1 kHz and the CDD time constant was restored to T_{c0} . The detector capacitors were discharged with faster rate for a duration of 40 ms. The IIR cut-off frequency is restored to 0.1 Hz at the 5th trigger, leaving some 160 ms margin for the transient. The masking is removed at 6th trigger. At this time the samples corresponding to the crossed multiplexer state are sent with 0.1 Hz filtering and slow time constant T_{c0} . The samples are then used by the server algorithm to compute calibrated positions until the 26th trigger when the multiplexers are switched back to the straight state.

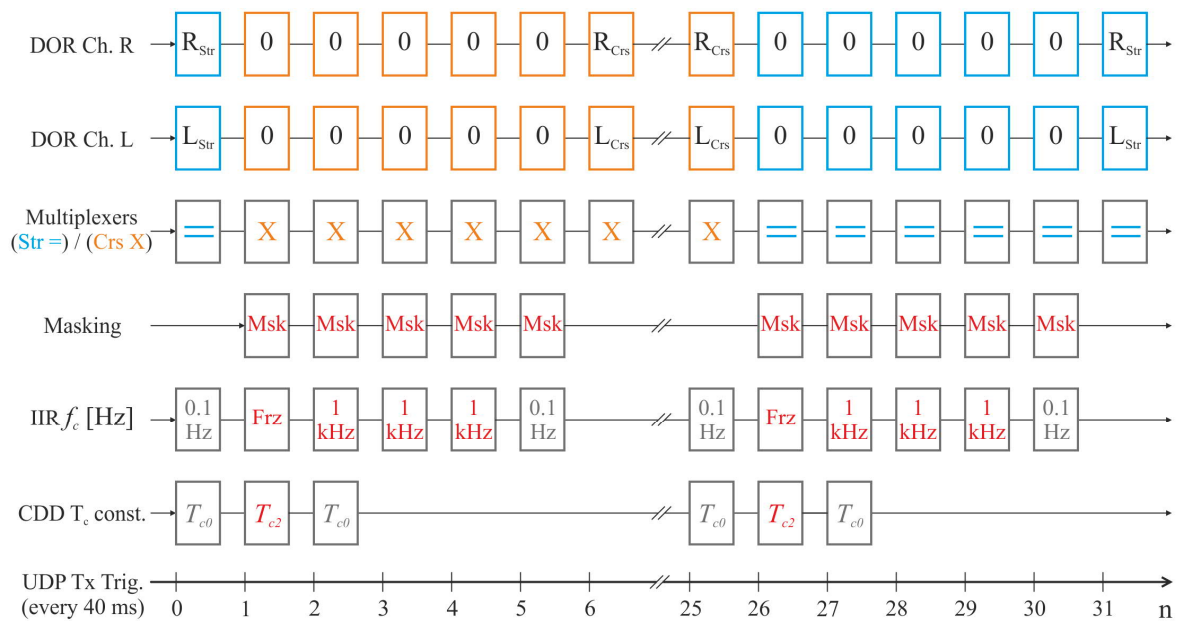


Fig. 4-26. Example of the AUCAL timing when configured for 1 Hz calibration switching.

A short-time change in the detector time constants, a so called detector pre-discharging, allows reducing the capacitor discharge time in the CDD when the signal amplitude L_{str} in the straight mode is larger than amplitude L_{crs} in the crossed mode. However, the pre-discharging increases the capacitor charging time when $L_{str} < L_{crs}$. Please note that the capacitor charging in the CDDs is by design much faster than their discharging, therefore its contribution to the transient duration is less significant. The performance of the orbit calibration using the AUCAL state-machine with the transient speed-up techniques was presented and analysed in the laboratory measurements part of this thesis.

The processed DOR samples are sent to the server in the UDP datagrams together with other data, including the control flags. In order to compute the calibrated orbits, either in real time or in the data post-processing, the receiving side has to run the calibration algorithm. Block diagram in Fig. 4-28 depicts the principle of the signal processing in the calibration algorithm. First the corresponding DOR samples and control flags are fetched from the UDP buffer upon each UDP reception. The multiplexer flag is then used to direct the DOR samples into the straight and crossed channels, each containing a masking and an accumulator blocks. The transient samples are automatically discarded using the masking flag. The *Str* and *Crs* accumulators alternate according to the multiplexer state. Each accumulator thus sums the samples from the respective multiplexer period. Once the samples from one period are summed, the result is copied and normalised by the number *N* of counted samples in the accumulator. The respective accumulators and counters are reset in preparation for the next multiplexer cycle. At this point the data rate is reduced from 25 Hz of the UDP trigger to the multiplexer switching rate. The calibrated electrode signals L_{cal} and R_{cal} are obtained by computing an average of the straight and crossed results. The electrode readings are used to compute the normalised calibrated position p_{cal} . The normalised position is then used to calculate the absolute position P_{cal} using a correction polynomial corresponding to the geometry of the used BPM. To optimize the precision of the numeric values, the UDP samples are kept in 32-bit integer representation and the normalisation block $1/N$ is omitted. Conversion to floating point arithmetic is then realised during computation of the normalised positions by the system server. The averaged electrode signals L_{str} , L_{crs} and R_{str} , R_{crs} can be used to compute the so called raw positions p_{str} and p_{crs} . These values represent positions which are not calibrated and are thus subject to the asymmetries and drifts of the analogue channels. The raw positions can be used to estimate the performance of the calibration algorithm.

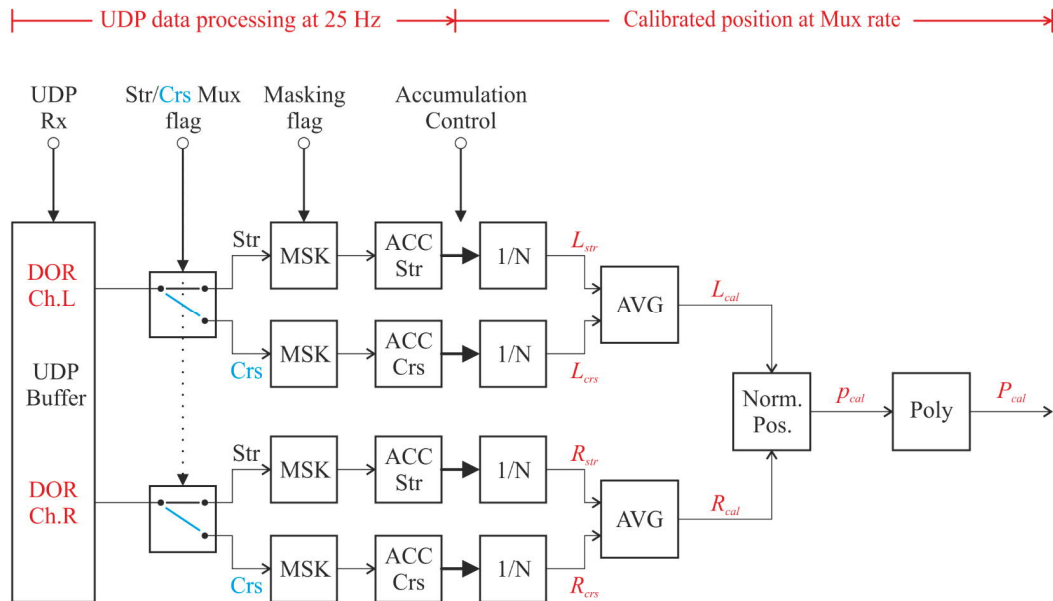


Fig. 4-28. Schematic block diagram of the calibration algorithm performed by the DOROS server.

Abbreviations: ACC – accumulator buffer, MSK – transient sample Masking, DOR – Diode Orbit, AVG – Averaging

A diagram in Fig. 4-29 presents an example of the real-time computation of the calibrated positions in the DOROS server based on the periodic signal multiplexing at 1 Hz rate (equivalent to 25 UDP samples). The horizontal axis corresponds to the sample indices n as they were received from UDP datagrams coming at 25 Hz rate. The vertical axis displays the signals samples from the DOR channels L and R as well as the samples computed in the most important nodes of the calibration algorithm. Please note that this example presents the horizontal BPM plane, but the same processing performed in all BPM planes.

The picture shows the signals during a transition between a so called flat mode when the calibration is disabled and the so called auto-calibration mode with the enabled calibration. The first two samples (indices 0 and 1) were received in the flat mode at the 25 Hz UDP rate. The flat mode in the front-end is changed into auto-calibration mode at the 2nd DOR sample. Each multiplexer period in this example consists of a total of 25 DOR samples. The calibration frequency is equivalent to 1 Hz. The AUCAL state-machine was set to mask five transient samples out of each switching period. These samples (red zero) are then discarded in the algorithm and excluded from the averaging. At the end of each multiplexer period the algorithm produces an average of 20 DOR samples displayed as L_{str} , R_{str} (blue) and L_{crs} , R_{crs} (orange) in the picture. In this example the first calibrated electrode signals L_{cal} , R_{cal} (magenta) and calibrated position P_{cal} (magenta) are computed after the processing of 26th DOR sample. Please note that the obtained results represent in this case a transient between the flat and auto-calibrated modes. As the L_{cal} and R_{cal} were computed only from the averaged DOR samples the corresponding position P_{cal} is not calibrated and therefore should be discarded from further processing. The first valid positions are obtained one multiplexer period later, which in this example corresponds to 1 s latency. The calibrated positions are then continuously published at the 1 Hz rate. Similar transients can be observed when disabling the auto-calibration. However, this case is not showed in this example.

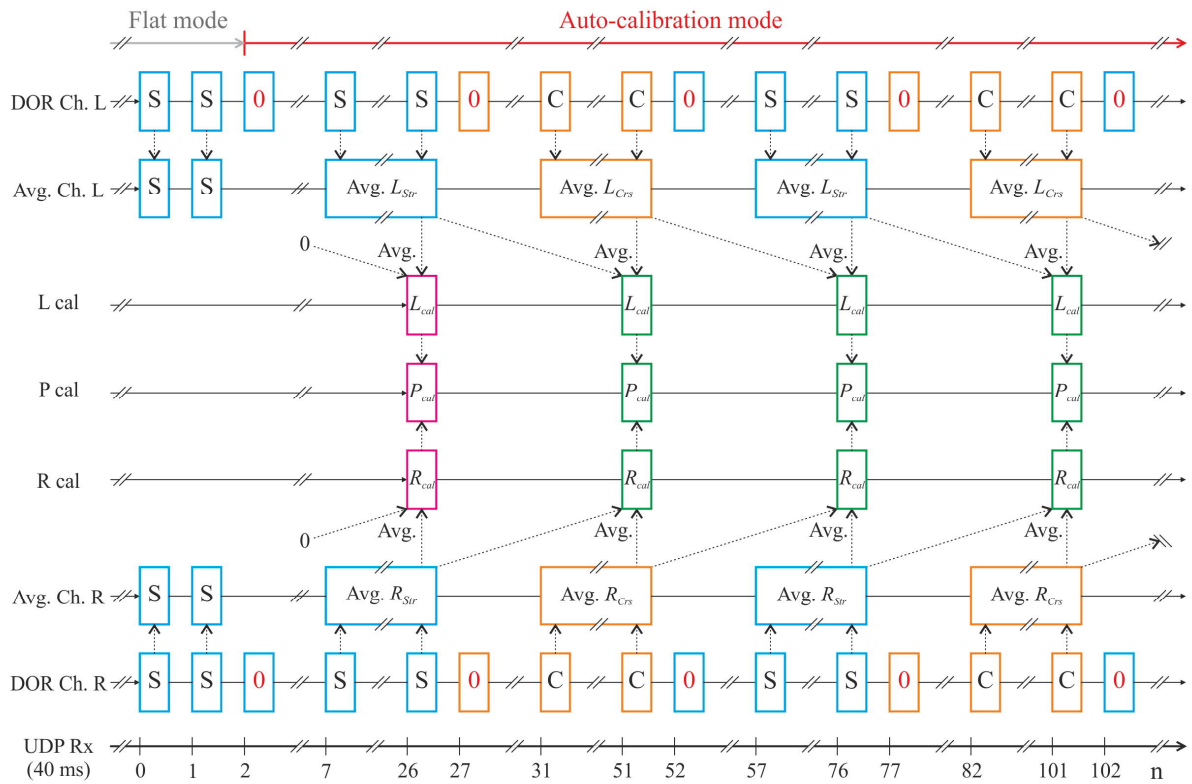


Fig. 4-29. Example depicting the calibration processing of the DOR signals from the L and R channels in the calibration algorithm.

Please note that the algorithm is not perturbed by the UDP losses if at least one unmasked DOR sample is averaged in each multiplexing period. Otherwise if at least one of the straight or crossed averages is missing the algorithm has to reset its processing by clearing the accumulators and counters and waiting for the next multiplexer period. The algorithm reset has then the same transient impact on the calibrated positions as the mode transition showed in the picture.

The programmed transitions between the auto-calibration and the flat mode can occur in the following circumstances:

- Engaging the automatic gain control. The AUCAL needs to be temporarily disabled (typically for a few seconds) to let the AGC state-machine to optimize the signal levels. The AUCAL state-machine is re-enabled after the gain regulation is finished and the DOR signals are stable.
- Performing the capture/freeze acquisition. The flat mode is required when acquiring the orbit and oscillation signals in the DDR memory (typically for a few tens of seconds). The periodic transients could otherwise limit the maximum number of samples which could be used to compute beam spectra.
- Performing diagnostics with test signals. The test pulses are connected to the inputs after the calibration multiplexers. The test signals therefore cannot undergo efficient auto-calibration procedure. The flat mode is then used in this case. When in this mode, the computed position are averaged in order to maintain the same orbit rate as during the calibration switching.

The calibration switching by the AUCAL is by default enabled and configured as a part of the start-up initialization routine of the front-end. The AGC state-machine is also enabled. It waits in the idle state until the signal level needs to be adjusted. The AUCAL state-machine is disabled for the duration of the gain adjustment leaving the orbit computation temporarily paused. The calibration algorithm is restarted and starts publishing calibrated orbits once the AGC is back to the idle state and the AUCAL is enabled. This is important to prevent spurious glitches in the measurements during the automatic gain adjustments.

4.4 Oscillation subsystem

This chapter is focusing on the signal processing and synchronization of the diode oscillation (DOS) subsystem. This subsystem was included in the DOROS front-ends to be able to detect small driven beam oscillations and demodulate them at defined frequencies. The oscillation processing is optimized for the oscillation signals in the frequency range between about $0.05 f_{rev}$ and $0.5 f_{rev}$ (approximately 500 Hz and 5 kHz). Block diagram in Fig. 4-30 displays an overview of the complete oscillation processing chain. The diagram includes the RF low-pass filters and the RF amplifier blocks which are shared between the DOR and DOS subsystems. The decoupling between the subsystems is based on separate RF amplifiers driving the orbit and oscillation diode detectors. Detailed description of the RF low-pass filter and RF amplifier blocks can be found in chapter 4.3.2. The waveforms shown on the right of the diagram illustrate the example signals in the most important nodes of the processing. The beam current pulses (green), which repeat every revolution period T_r , induce the left l_e (blue) and right r_e (red) electrode signals.

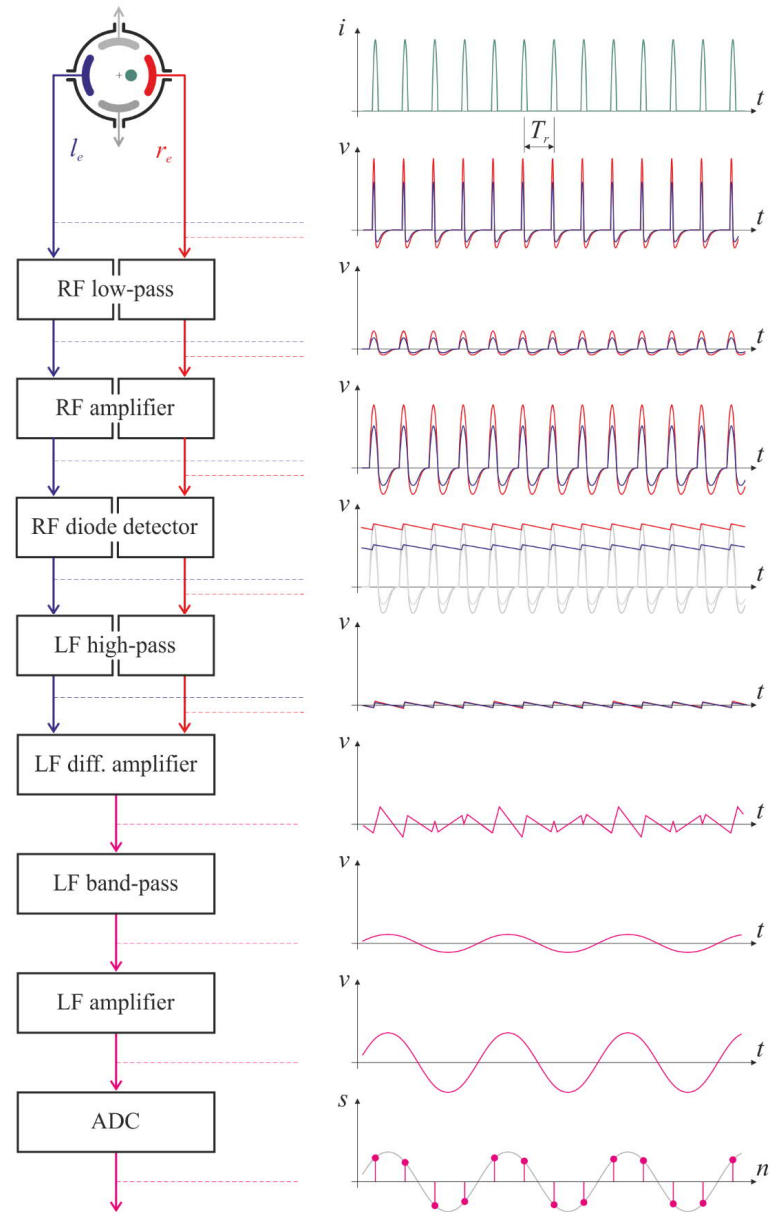


Fig. 4-30. Overview diagram of oscillation signal processing of 2 channels with example signals shown in important nodes.
Abbreviations: LF – Low Frequency, RF – Radio Frequency, IIR – Infinite Impulse Response filter.

As most of the RF circuits are common for both the orbit and oscillation subsystems, some of their settings and features are shared which introduces dependencies. For example, periodic switching of the RF input multiplexers can perturb the operation of the DOS subsystem, especially for acquisitions longer than one multiplexing period. The AUCAL state-machine which controls these multiplexers, has to be temporarily disabled for the duration of the oscillation measurement. Similarly, the AGC control of the RF gains perturbs the DOS channels by introducing short glitches in the signals. Such perturbations have to be masked or taken into account when performing the oscillation measurements. Please note that the oscillation measurements are required less often than the orbit measurements which need to be provided continuously.

The signal processing related only to the oscillation subsystem starts in the dedicated RF diode detectors. The small beam oscillations appear as an amplitude modulation of the BPM electrode signals. The AGC state-machine provides automatic adjustment of the signal amplitude on the diode detector input within its optimal range. The RF diode detector circuit converts this modulation into slowly varying signals superimposed on relatively large DC signal. This DC part, which could saturate the first amplifier stage, is blocked by the following simple high-pass filter. The output signal thus contains information on the relative position change, while the information proportional to the absolute beam position and intensity, converted into the DC component by the detector, is removed from the DOS processing. As seen on the picture, each electrode signal l_e and r_e is up to this point processed in a separate channel. The output signals from the two detectors are then subtracted and amplified by a differential amplifier. The output signal from the amplifier still contains the residual ripple from the diode detectors which is then filtered by the band-pass stage. When the beam consists of a single bunch, the f_{rev} signal component is orders of magnitude larger than the useful oscillation signal and therefore could saturate the LF amplifier and the ADC input circuits. The band-pass filter is thus important for the system dynamic range. The LF amplifier further amplifies the filtered oscillation signals and drives the differential inputs of the ADC. The oscillation samples are used in the following acquisition and DSP channels.

The DOS subsystem also contains the timing and synchronization circuits which are described later in this chapter. They allow synchronization of the ADC sampling to the beam revolution using a timing reference generated locally from the received BPM signals. Synchronous ADC sampling and data processing are important for the LHC optics measurements, such as the beta-beating and phase advance.

4.4.1 Diode detector and low frequency signal processing

Block diagram in Fig. 4-31 displays the DOS analogue processing of one electrode signal pair. Each analogue board contains two such channels with identical processing circuits. The electrode signals are provided from a fixed gain amplifier at the end of the RF chain. These amplifiers drive and decouple the DOS detectors from the DOR subsystem. The levels of the DC signals on the DOS detectors cannot be directly monitored as they are removed in the first high-pass filter. However, the DOS signal levels can be estimated using the DOR subsystem. The largest measurable signal by the DOR subsystem corresponds in the DOS subsystem to a signal level smaller by 2 dB. The AGC thus sets the optimal signal levels also for the DOS subsystem.

The diode detector in each of the two channels was realised as a voltage doubler using two Schottky diodes D_a and D_b accommodated in a single package. Negative pulses carrying the oscillation information can therefore also contribute to the output signals improving the DOS sensitivity. Two capacitors C_{d1} and C_{d2} provide the storage capacitance. The shunt resistor R_s controls the discharge rate of the capacitor C_{d1} . The following LPF_d block contains a passive, high impedance low-pass filter. It was implemented as a second-order low-pass with critical response and a cut-off frequency of 5 MHz. The filter attenuates high frequency components which leak through the parasitic capacitances of the diodes and prevents the saturation of the subsequent active circuits.

The main detector capacitors C_{d2} can be discharged by combination of resistors R_0 , R_1 and R_2 which allow to set the dynamic response and bandwidth of the detectors. As shown in the picture, the signals S_1 and S_2 control the discharge rate by connecting resistor R_1 and R_2 in parallel to R_0 . The slowest time constant T_{c0} is based only on the resistors R_0 with the highest resistance. The slower discharge rate is approximately 3 ms, equal to some 30 LHC beam revolutions. This time constant setting is suitable for LHC beams consisting of a few bunches (e.g. LHC pilot beams). The time constant T_{c1} based on the parallel resistors R_0 and R_1 sets a 10-times faster detector discharge. The fastest discharge rate T_{c2} based on the parallel resistors R_0 and R_2 sets a 100-times faster discharge with the time constant of less than a single turn. This time constant setting is suitable for the LHC beams consisting of at least a few hundred of bunches, like nominal LHC physics beams. The discharge rates of the three DOS time constants are summarized in Table 4-9. The signals from the DOS detectors are then processed by the LF circuits. The capacitors C_{d3} block the DC part of the signal which is related to the beam intensity. Otherwise the DC signals could saturate the following LF stages.

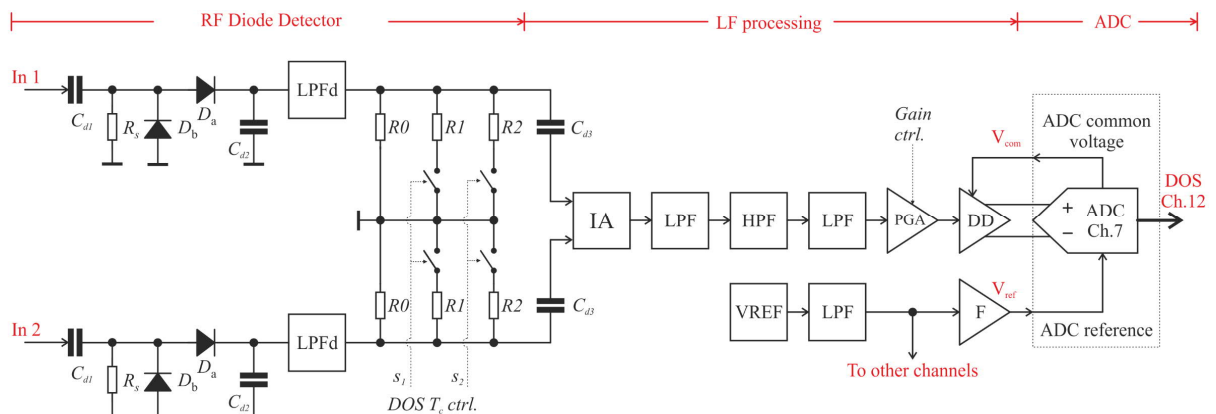


Fig. 4-31. Schematic diagram of DOS diode detectors and low frequency processing of two BPM electrode signals received from the RF amplifier stage.

Abbreviations: LF – Low Frequency, LPF_d – passive Low Pass Filter in the DOS detector, IA – Instrumentation amplifier, LPF – Low-Pass Filter, HPF – High-Pass Filter, PGA – Programmable Gain Amplifier, DD – Differential ADC driver, VREF – Voltage reference source, F – Follower, VCOM – Common mode voltage.

Table 4-9. Summary of the DOS detector time constants.

DOS time constants	T_{c0}	T_{c1}	T_{c2}
Discharge rate [ms] *	3.00	0.30	0.03
Equivalent frequency [Hz] *	53.05	530.52	5305.16
Equivalent number of LHC turns *†	33.77	3.38	0.34

*Rounded to 2 decimal places, † referenced to 11.2455 kHz LHC revolution frequency.

The output signals from the two detectors are then subtracted and amplified in an instrumentation amplifier (IA) built with three low noise, precision operational amplifiers. The IA has the overall gain of 10. The resulting difference of the two electrode signals is proportional to the position change and is not normalised by the intensity. The oscillation signal is denoted in the picture after the corresponding BPM electrode pair (in this case DOS.Ch12).

The three LF filtering blocks in the diagram together form a band-pass filter. The first block in the filtering chain is a 2nd order low-pass Butterworth filter with 5.8 kHz cut-off frequency. Its main purpose is to filter the f_{rev} frequency, its harmonics as well as out-band noise. The following high-pass filter increases the frequency cut-off of the DC blocking to remove the large beam low frequency components, such as the synchrotron oscillation harmonics. It was built as 2nd order Chebyshev filter with 500 Hz cut-off and 1 dB ripple. The last low-pass filter in the chain removes further the f_{rev} and its harmonics. The signal can be further amplified by the programmable gain amplifier (PGA). The PGA gains of 0 dB, 20 dB and 40 dB are controlled by the FPGA.

The resulting oscillation signal is then processed by a differential amplifier that drives the differential ADC inputs. The amplifier also shifts the common mode of the signal by the V_{com} voltage provided from the ADC in order to fully utilize its dynamic range. The ADC normalises its input signal to the doubled voltage reference V_{ref} . The samples are provided in two's component format. The zero voltage corresponds to code 0x000000, expressed in 24-bit hexadecimal format. A positive V_{ref} voltage on the ADC input corresponds to code 0x7FFFFFFF. A negative V_{ref} voltage on the ADC input corresponds to code 0x800000. The samples in the DOS DSP are then formatted as signed 24-bit integers and used in the further DSP.

4.4.2 Digital signal processing

The DOS DSP allow computing an estimate of the peak-peak signal amplitudes for each oscillation channel. Block diagram in Fig. 4-32 displays two DOS processing channels implemented in the FPGA logic. Please note that there are two oscillation channels per analogue board. The first DSP block is the ADC SPI interface which is common for both DOR and DOS subsystems. The ADC sends its data in frames upon the DRDY trigger synchronous to the sampling reference clock. The de-serialized DOS samples are then provided from the SPI module to the respective DSP channels in the FPGA.

The following conversion block changes the signed integer two's complement data into unsigned integers. The conversion is done by adding half of the 24-bit ADC full-scale and ignoring the carry bit from the addition. The resulting 24-bit samples are sent to a comparator module to determine minimum and maximum of the signal. The comparator module compares each input sample with an output register. If the sample is larger (the MAX case) or smaller (the MIN case), than the value is stored in the output register, replacing its previous value. The output register thus stores the minimal and maximal results.

The register value is copied to the UDP buffer and all the comparator registers are cleared upon each Tx trigger, reducing the measurement data rate to 25 Hz. Please note that sampling of both ADCs of a front-end is in general asynchronous, as each ADC may be synchronised to the BST of a different beam. The UDP buffer has to contain data from all DSP channels of both the DOR and DOS subsystems. Therefore the UDP sending is delayed after each Tx trigger until the UDP buffer contains samples from all DSP chains. Maximum spread in the UDP sending rate is about one LHC revolution period (about 90 μ s).

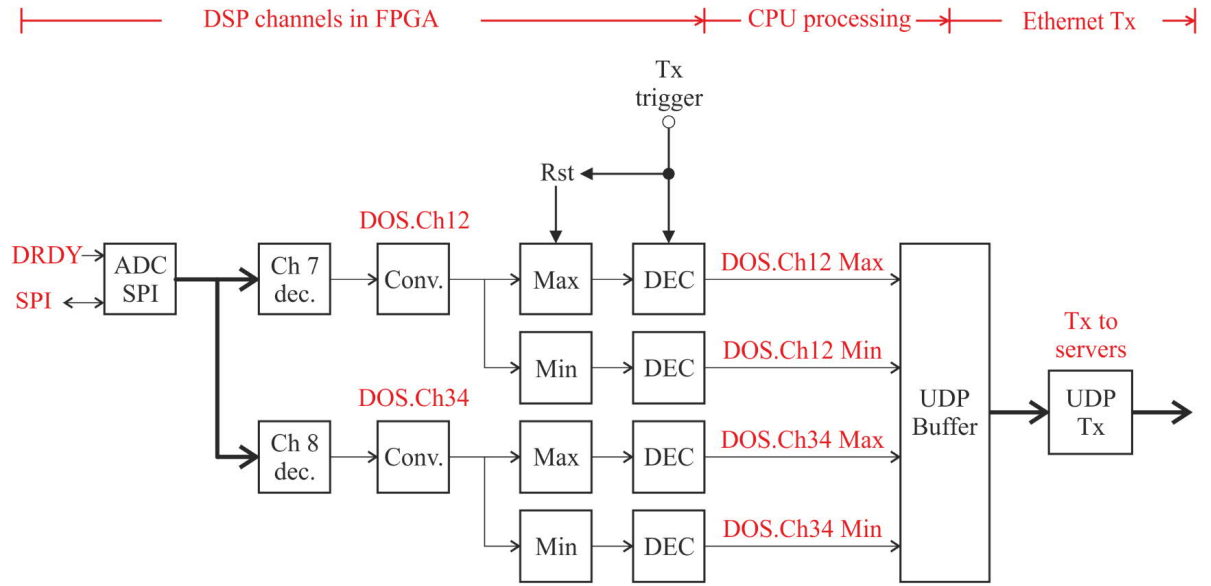


Fig. 4-32. Schematic diagram of the DOS digital signal processing performed in the FPGA.

Abbreviations: SPI – Serial Peripheral Interface, IIR – Infinite Impulse Response filter, RST – Reset, DEC – data rate Decimation, UDP – User Datagram Protocol.

The peak-peak values of the oscillation signals are computed on the DOROS servers by subtracting the transmitted minimum and maximum values. The main purpose of these measurements is to monitor the oscillation signals and prevent their saturation. The implementation of the DSP framework in the FPGA logic foresees adding parallel processing blocks. More complex DSP can perform synchronous detection in order to obtain amplitude and phase of the beam oscillations. This technique was prototyped in the author's master thesis [34].

4.4.3 Front-end synchronization

Synchronization of the ADC sampling to the circulating beam is important for precise DOS measurements. The DOS subsystem was designed to be used to detect and measure the excited beam oscillations and the related beam parameters, such as the betatron coupling or phase advance. The beams during such measurements are harmonically excited to oscillate at frequency f_{exct} typically set close to the tune of the machine in the excitation plane. Since the machine response is at the excitation frequency, processing of the DOS signals can be thus simplified by focusing only on the f_{exct} component. This approach is by far more computationally efficient than analysing the whole available beam spectrum.

An illustration shown in Fig. 4-33 demonstrates principle of the local betatron coupling measurement using signals from a BPM. The measurements are based on exciting coherent transverse oscillations of the beam (green) at a frequency f_H in the horizontal (H) plane (red) and frequency f_V in vertical (V) plane (blue). Since frequencies f_H and f_V are different the beam is excited in both planes at the same time. The observed oscillation signal in one BPM plane contains the component at the excitation frequency and the component coupled from the perpendicular plane. The spectra of the example signals are shown on the right of the picture. The signals from both BPM planes are processed by the same analogue board. The sampling synchronism between the plains is thus guaranteed by design as the ADC samples the H and V DOS signals with the same clock.

The synchronisation of the ADC sampling to the LHC revolution frequency is assured by using the BST signals received by each DOROS front-end. The front-ends contain in addition dedicated circuits, which allow aligning the phase of the ADC clocks to a common phase reference. This feature is important for measuring the LHC beta-beating or phase advance.

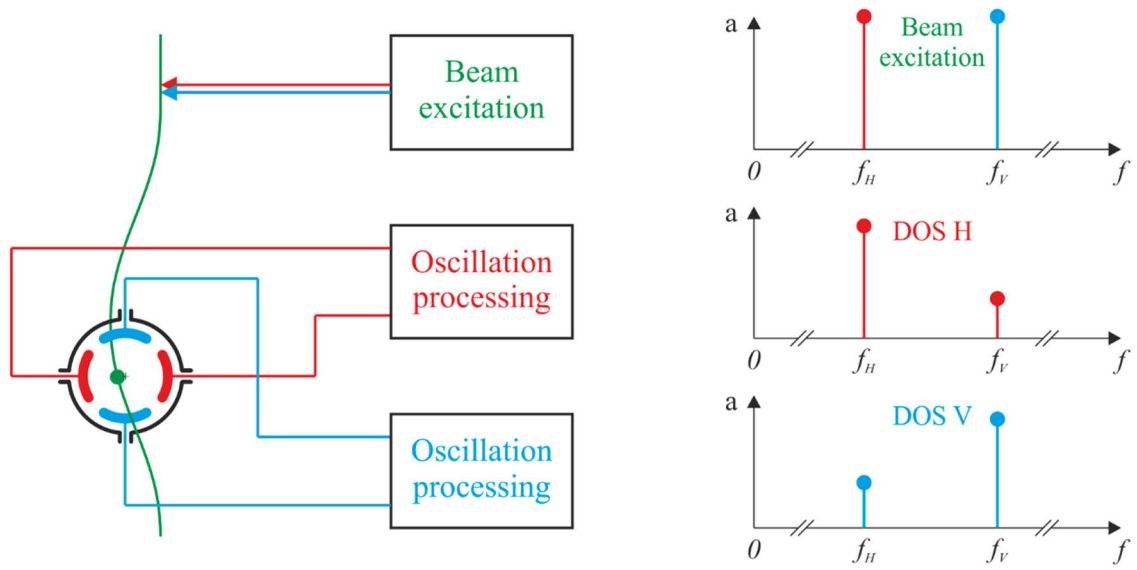


Fig. 4-33. Example illustration of the coupling measurement based on the harmonic beam excitation in horizontal (H) and vertical (V) BPM plane.

The illustration shown in Fig. 4-34 demonstrates the principle of the phase advance measurements, which are based on the beam excitation at a single frequency f_{ext} . The beam oscillations in this illustration are measured by three BPMs connected to three DOROS front-ends, which are in general located in different places. The beam response signal at the excitation frequency in each BPM is demodulated in the corresponding DOS channel. The example signals are displayed on right of the picture. Please note that the illustrated waveforms in the picture are not in scale. The phase difference $\Delta\phi_{01}$ and $\Delta\phi_{12}$ observed between the beam response signals represents the beam phase advances between the respective BPMs. To measure these phase differences it is necessary to sample the oscillation signals from different front-ends at the same phase of the beam revolution. The ADCs in each front-end sample the DOS signals synchronously to the reference clock f_{rev} recovered from the BST distributed over a dedicated fibre optics network.

However, the BST distribution has an arbitrary phase relationship with respect to the circulating beam in different nodes of the fibre network. The phase ϕ_{err} shown in the picture represents an error delay between the beam and the f_{rev} obtained from the BST in the particular location of the front-end. So called local timing circuits in the front-ends use the BPM beam signal pulses to generate a reference signal with a deterministic phase relationship to the circulating beam. The reference signals obtained from the local beam timing is then used for adjusting the phase of the BST clocks with respect to the beam. The adjustment takes into account the beam propagation time between the BPMs. The phase adjusted signal from the BST is then used as the reference clock for the ADCs.

Please note that the amplitude of the BPM signals can change with the beam intensity and position. In addition the signals change also with the gains and calibration multiplexing performed by the state-machines in the front-end. The local beam reference thus cannot be used for continuous ADC clocking in all operational conditions. The ADC PLL circuit requires a permanent presence of the reference signal in order to remain locked. The local timing can be therefore used for the BST phase adjustments only under certain conditions.

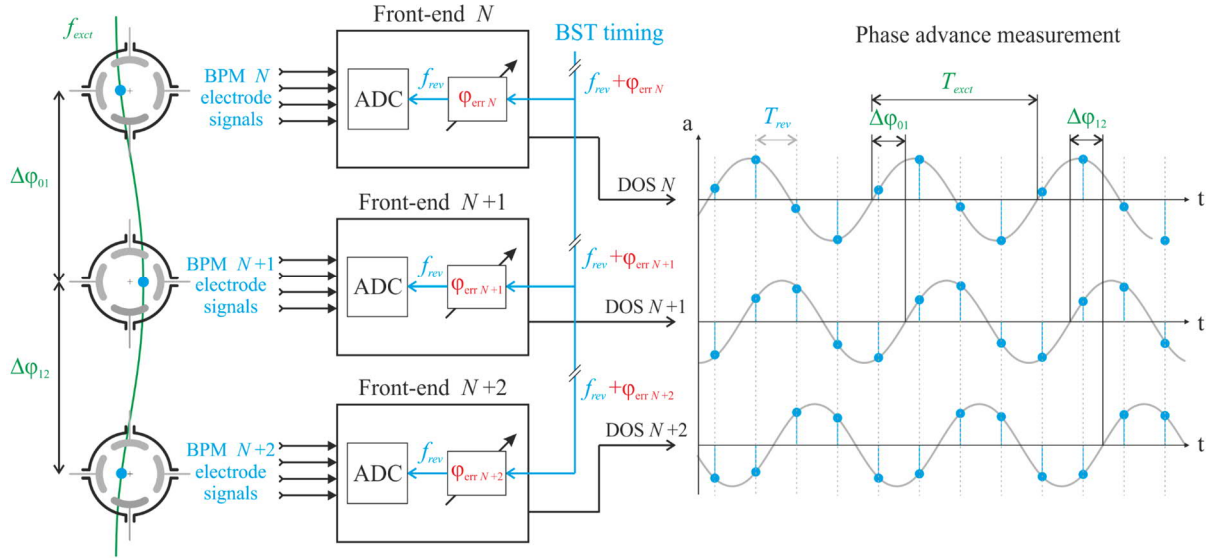


Fig. 4-34. Illustration of the phase advance measurement principle demonstrated on three beam position monitors whose signals are processed by three DOROS front-ends.

4.4.3.1 Clocking and synchronization circuits

Block diagram in Fig. 4-35 displays an overview of the processing and clocking blocks in the DOROS front-end. One front-end can receive BPM signals from two LHC beams which, in general, can have different revolution frequencies. For example, when in one LHC ring are circulating ions and in the second ring protons, the worst-case difference between their revolution frequencies can be about 0.3 Hz, which would spoil phase advance measurements, especially when using synchronous detection. The front-end has therefore two independent timing receivers and synchronization circuits, which allow the ADCs to be clocked synchronously with the corresponding beam timing. The DOR and DOS signals in each ADC are sampled with common clock. The eight channels are clocked from the same source distributed inside the chip with a sub-nanosecond skew in the worst case. As shown in the diagram, the reference clocks for the ADC sampling can be provided from one of the three timing sources selected by a clock multiplexer:

- **BST** – The reference is received from the beam timing master over a fibre network. The clocks can be phase-aligned to the beam reference phase.
- **External** – The reference clocks are provided from the LEMO connectors on the front-panel. This mode is foreseen for testing and special measurements.
- **Local** – A local reference based on a temperature stabilised crystal oscillator on the FPGA board. The worst case error of the local reference is in order of ± 2 ppm. The frequency of this clock cannot be adjusted to the beam.

It is possible to synchronize both A and B processing chains to the reference clocks provided only from one of the two synchronization and timing circuits (STC). Such a setup is used when both A and B analogue boards are connected to BPMs on the same beam, which is often the case in the collimator DOROS system. In such case the STC block automatically selects the timing from the respective BST SFP module. The second SFP module is not used and can be omitted during the front-end installation. The clock multiplexers in the FPGA redirect and distribute the received timing to both ADCs. In case of connecting the SFP modules to the wrong LHC timing (e.g. accidental swapping of the BST fibres on the front-panel) is automatically recognised by the front-end CPU and signalled on the front-panel LEDs. Such feature helps in avoiding connection error when installing the front-ends in the accelerator tunnel. In case both BST signals are missing (e.g. BST fibres not available in the vicinity of the front-end), the CPU automatically selects the local clock until at least one BST signal is available again.

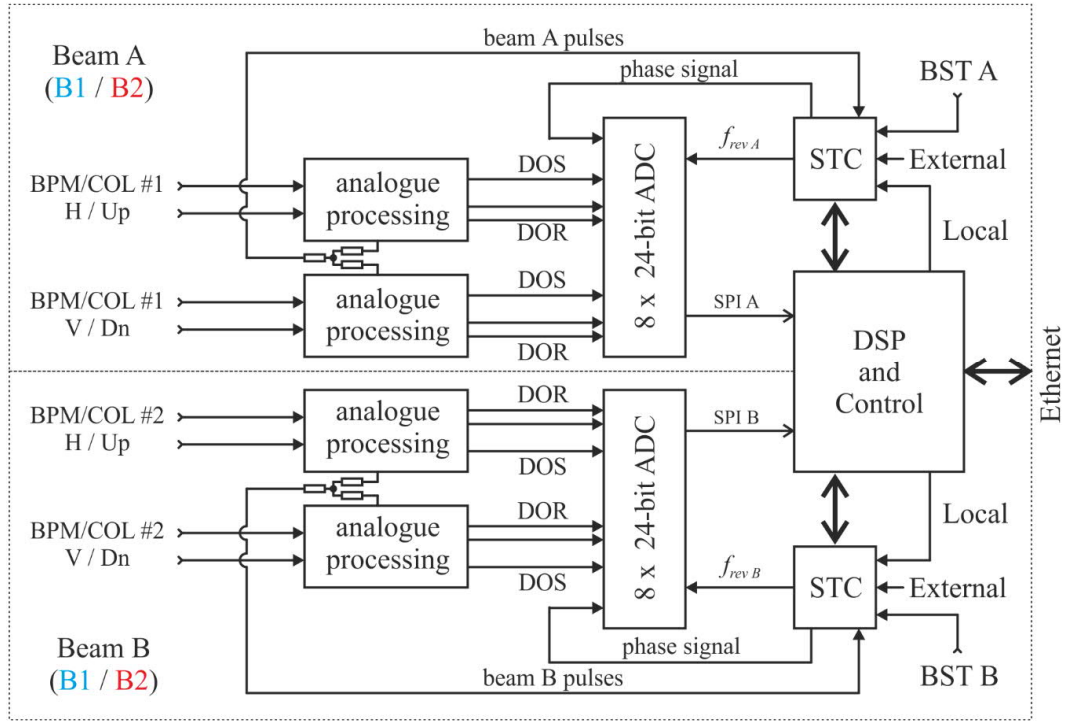


Fig. 4-35. Overview of the processing and clocking blocks in the DOROS front-end.

Abbreviations: SPI – Serial Peripheral Interface, DOR – Diode orbit signal, DOS – Diode oscillation signal, DPD – Differential Phase Detector, STC – Synchronization and Timing Circuits, DSP – Digital Signal Processing.

The block diagram in Fig. 4-36 displays the internal structure of an STC block. The structure can be divided into four main parts:

- the main ADC PLL and the clock multiplexers,
- the BST receiver and the phase-shifters,
- the local timing generator,
- the differential phase detector.

The first part in the block diagram is the main PLL which provides the master clock (MCLK) used by the internal ADC circuits. The architecture of the used ADC requires the MCLK frequency to be 512-harmonic of the sampling frequency f_{sam} and the frequency of the ADC data frames. The data ready (DRDY) signal, which marks the beginning of each frame, is also synchronous to f_{sam} . Its rising edge is delayed by the latency of the ADC data processing. The ADC PLL compensates the ADC latency such that the phase difference between the DRDY clock and the reference clock $f_{ref,Mux}$ is zero. The PLL also produces the MCLK clock by multiplying the reference frequency by 512.

The VCO of the PLL_{ADC} is a varicap-tuned LC oscillator with a JFET transistor. It is designed to provide a centre frequency of the MCLK around 5.76 MHz. The centre frequency can be adjusted within some 7 % by a capacitive trimmer to compensate for the production tolerances of the resonator components. The VCO tuning range is about 2.8 %, much larger than the worst case 25 ppm of the LHC revolution frequency swing. The tuning range must accommodate the drifts and aging of the VCO resonator components. Measured frequencies at the extremities of the PLL tuning ranges are summarized in Table 4-10. The RMS jitter of the MCLK clock was measured to be around 5 ps. The lock signal of the ADC PLL is monitored by the FPGA and the corresponding information is transmitted in the UDP datagrams. The source of the PLL reference clock is provided from a clock multiplexer which is controlled by the CPU. The BST sources are selected as a default reference source.

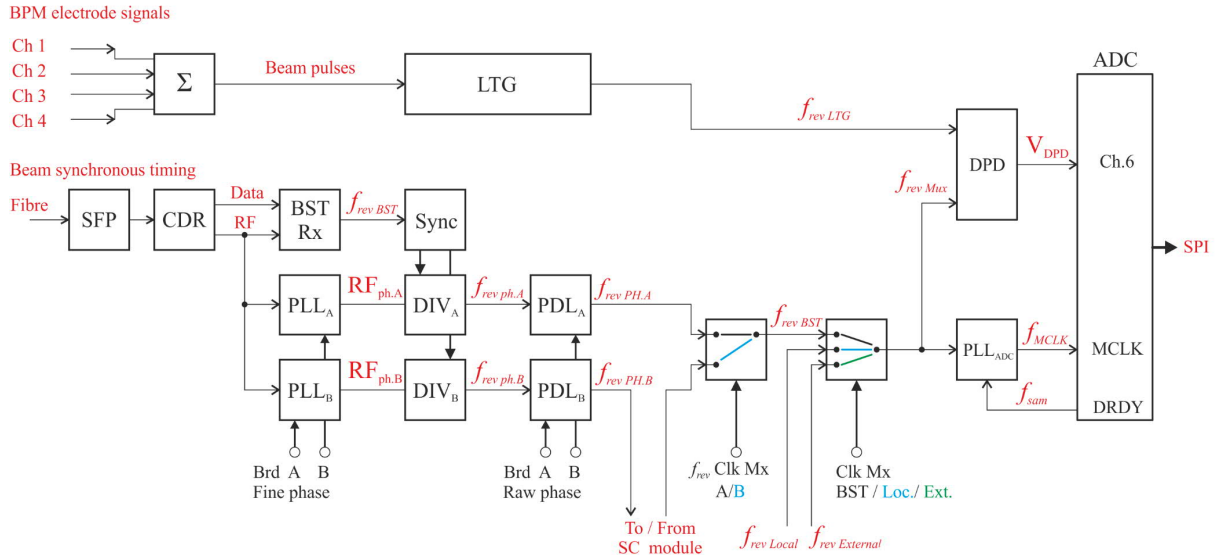


Fig. 4-36. Block diagram of the synchronization circuits in the front-end used to provide the ADC sampling clock reference.
Abbreviations: PLL – Phase Locked Loop, DIV – frequency divider, BST – Beam Synchronous Timing, LTG – Local Timing Generator, MCLK – ADC Master Clock, DRDY – Data Ready signal from ADC, DPD – Differential Phase Detector, PDL – Programmable Delay Line, CDR – Clock and Data Recovery, SFP – Small Form Pluggable module.

If the BST is not available the CPU automatically selects other reference source. The local and external sources for the PLL reference can be also selected upon dedicated UDP command. These sources are typically used for laboratory measurements and diagnostics.

The second part of the synchronization circuitry generates the phase-shifted reference clock based on the BST reference. The BST master transmits the revolution clock along with encoded triggers and other information modulated onto to the 14256th harmonic of f_{rev} beam revolution frequency, that is about 160 MHz. The BST optical signal is received by the dedicated SFP module and sent to the CDR block to obtain the carrier clock (RF). The carrier clock is then sent to the FPGA. The BST Rx module is a standard FPGA module developed by the BI group to receive and decode the BST messages. Decoded $f_{rev BST}$ signal at the output of the BST Rx is synchronized to the RF and used as a reference for the following circuits. The received RF clock is used to derive the phase-shifted f_{rev} which is then used as a reference for the ADC PLL. The RF clock can be also used to derive other clocks, such as the 40 MHz bunch clock (3564th f_{rev} harmonic) defining the 25 ns bunch spacing.

The phase shifting blocks in the diagram are used to align the phase between the BST revolution reference and the corresponding beam-derived signal. As seen in the diagram, two identical chains A and B produce two phase-shifted clocks $f_{rev phA}$ and $f_{rev phB}$, each independently controlled by the front-end CPU. Two PLL modules accommodated in the FPGA synchronize the frequencies of the RF_{phA} and RF_{phB} clocks to the RF reference clock from the BST. The PLLs also introduce a delay between the reference and output clocks which can be controlled in steps.

Table 4-10. Tuning ranges of the main ADC PLL.

	f_{vco} [MHz] *			f_{sam} [Hz] *			Δf_{vco} [MHz]*	Δf_{sam} [Hz]*	$\Delta f / f_{centr.}$ [%]*
	min	centre	max	min	centre	max			
Centre frequency		5.76			11246.09				
Trimmer ¹	5.58		5.99	10898.44		11699.22	0.41	800.78	7.12
Varicap ²	5.66		5.82	11054.69		11367.19	0.16	312.50	2.78

*Rounded to 2 decimal places, ¹measured at 1/2 of the VCO tuning range, ²measured at 1/2 of the VCO trimming range.

The smallest available delay step is about 100 ps. Please note that the smallest step determines the phase alignment error which affects the accuracy of the phase advance measurements. The residual error of the phase alignment $\Delta\varphi_{err}$ can be expressed as:

$$\Delta\varphi_{err} = 360^\circ \Delta t_{err} f_{exct} \quad (4-36)$$

where Δt_{err} is the alignment error and f_{exct} the excitation frequency of the beam. The phase error $\Delta\varphi_{err}$ decreases proportionally to the excitation frequency f_{exct} . For the mentioned alignment error of 100 ps the worst-case phase advance error would be around 0.0001° for the typical measurement frequency of $0.3 f_{rev}$. If the alignment error was one BST clock period of 6.25, then the corresponding phase error would be in order of 0.01° .

The frequency of the RF_{phA} and RF_{phB} clocks is divided in the DIV blocks so that the frequencies of the phase shifted clocks $f_{rev phA}$ and $f_{rev phB}$ match the frequency of the $f_{rev BST}$ from BST Rx. As the DIV blocks are based on simple counters, a synchronizer module removes the phase ambiguity which may occur during the frequency division. The Sync module resets the counters in the DIV block so that their output clocks at zero phase correspond to the revolution frequency $f_{rev BST}$ derived from the BST. As a result, the phase-shift is always related to this clock. The additional programmable delay lines (PDL) at the output of each chain provide a raw phase control in 6.25 ns steps. These blocks were implemented in the FPGA logic and are also independently controlled by the CPU. The resulting clocks $f_{rev PHA}$ and $f_{rev PHB}$ are then phase-controlled copies of the beam synchronous $f_{rev BST}$. The phase-shifted revolution frequency clock $f_{rev PHA}$ is by default sent to the ADC PLL and clock $f_{rev PHB}$ is provided to the second STC block in the front-end. The multiplexer allows to route either of the two reference clocks to both ADC PLLs, for example when both ADCs are locked to the same BST of the same beam.

The third part of the synchronization and timing circuits is a so called Local Timing Generator (LTG) used to generate the reference signal $f_{rev LTG}$ based on the BPM beam signals. Beam signals from each of the four BPM channels are taken after the programmable RF amplifier and summed passively to produce beam pulses with amplitudes independent of the beam position. The signal level is automatically adjusted by the AUG state-machine. The resulting beam pulse signals are sent to the LTG module where they are converted into logic signals by a fast comparator with a programmable threshold. The trigger pulses from the comparator are then used by the LTG logic to synchronise to a selected bunch circulating in the machine. Under certain conditions, the LTG block of each DOROS front-end can synchronise to the same circulating bunch. The revolution frequency of the selected bunch $f_{rev LTG}$ is equal to the reference frequency $f_{rev BST}$ received over the BST. The phase of the reference clock from the LTG output is related to the arrival of the “synchronisation” bunch in the respective BPM. More details on the LTG module functionality and its implementation are provided in the following chapter. Results of the laboratory measurements performed with the LTG circuits are presented in the oscillation measurements part of the thesis.

A phase detector at the end of the synchronization chain measures the phase difference between the BST and LTG reference clocks. The detector output signal is used by the optimisation algorithms which control the phase of the BST clocks such that the measured phase difference is close to zero. A differential phase detector circuit (DPD) was developed to measure fine steps of the PLL phase-shifters. The circuit measures the phase difference of the LTG and BST at the f_{rev} clock frequency. The DPD output DC voltage V_{DPD} is then sampled by one of the ADC channels. A DSP chain, similar to the DOR channels, processes the DPD signal in order to achieve its high resolution readout. The phase results at the output of the DSP are provided together with the DOR data at the 40 ms rate. The phase-to-voltage characteristic of the DPD was optimized to have an extremum when its input signals are exactly in phase. Such characteristic is ideal for optimization algorithms which control the PLL phase-shifters in a slow feedback loop.

The circuit was built to provide precise indication of the zero-phase condition independently of the detector power supply as well as the offset voltages of the ADC and its input circuits. This feature is important for reproducibility and precision of the phase alignment for each DOROS front-end. Further details of the DPD functionality and its implementation are provided in the following chapter. Results of the laboratory measurements performed with the DPD circuit are presented in the oscillation measurements part of this thesis. To increase the speed of the phase alignment a second phase detector, operating in parallel to the DPD, provides faster response. It was implemented in the FPGA logic as a classical clock counter providing the edge-to-edge phase difference every turn. Its phase resolution φ_{res} can be defined as:

$$\varphi_{res} = 360^\circ T_{clk} f_{rev} \quad (4-37)$$

where T_{clk} is the period of the counter and the f_{rev} is the frequency of the measured reference clocks. Achieved resolution φ_{res} of the raw phase detector using the 160 MHz clock from the BST is around 0.025° . The detector is then mainly used for monitoring of the phase alignment and controlling the raw phase shifters.

4.4.3.2 Local timing generator

Block diagram in Fig. 4-37 displays the internal circuits of the local timing generator along with the example signals in its important nodes. As particle bunches pass through the BPM, the pulses arrive at the LTG input with a bunch train pattern repeating every LHC revolution period. The example signal shown in the picture consists of more than 6 bunches that circulate in the machine. An abort gap at the end of the bunch train consists of a space without beam. Its minimal duration of about $3 \mu s$ corresponds to the time required by the kicker magnets to rise their magnetic fields in order to safely eject the beam into the beam dump [3]. This gap provides an asymmetry in the bunch train which is utilized by the LTG circuit. The BPM pulses are converted to logic signals using a fast comparator. The datasheet rise and fall times of the comparator are around 0.6 ns. The comparator threshold voltage is controlled by a 12-bit DAC, which allows to adjust a reliable triggering by the beam pulses.

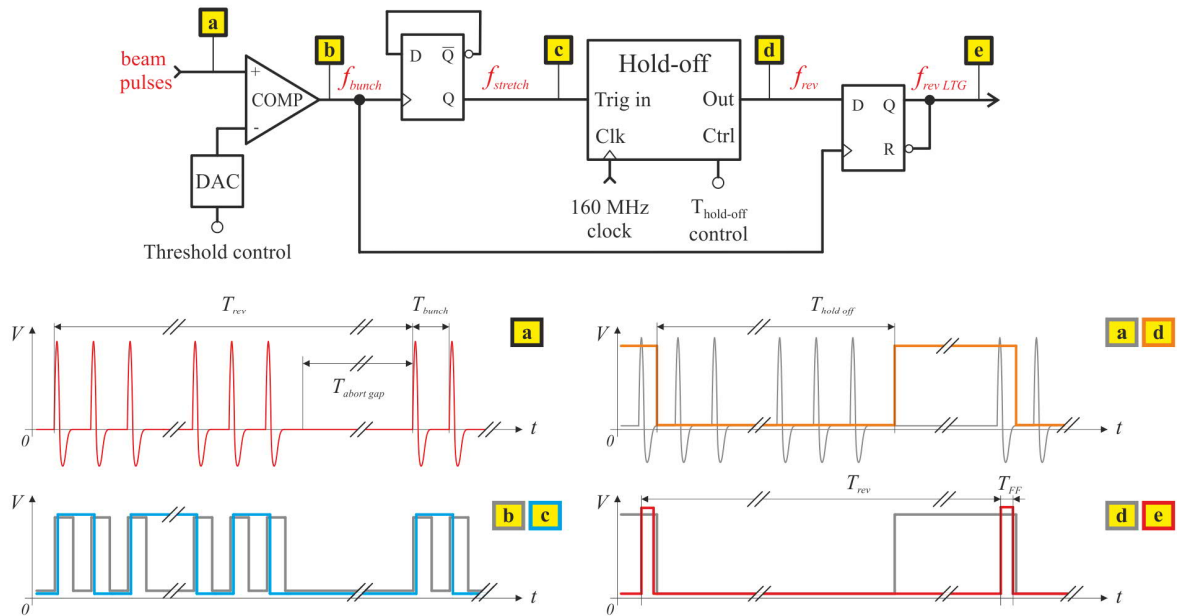


Fig. 4-37. Block diagram of the local timing generator circuits used to provide the beam phase reference.
Abbreviations: DAC – Digital Analogue Converter, LTG – Local Timing Generator, COMP – Comparator

Voltage span on the DAC output allows setting the threshold larger than the BPM pulses. This feature may be used to disable the output signals from the comparator. Dynamic range of the DAC was scaled to the DOR signal level based on laboratory measurements. Obtained results are presented in the oscillation measurements part of this thesis.

A following flip-flop converts the short output pulses from the comparator into edges. The frequency of the comparator pulses is divided by two so that it can be easily sent over a flat cable to the FPGA. The circuits in the FPGA inputs are then sampling the $f_{stretch}$ signal at both rising and falling edges. Resulting signal is used to trigger a hold-off circuit. A bunch pulse starts the hold-off counter which immediately sets its output low. At the same time the triggering input is disabled for the duration of $T_{hold-off}$ which can be configured by the front-end CPU. A dedicated counter in the FPGA measures the hold-off duration in 6.25 ns steps. Once the counter has reached the programmed delay $T_{hold-off}$, the output of the hold-off circuit is set to high and the trigger input is enabled for the beam pulses.

The LTG locks to the first bunch after the largest gap in the bunch train when the hold-off time is set as:

$$T_{hold-off} = T_{rev} - T_{largest\ gap} + \Delta T \quad (4-38)$$

$$(T_{rev} - \Delta T) > T_{largest\ gap} \geq T_{abort\ gap} \quad (4-39)$$

where T_{rev} is the beam revolution period (about 88.9 μ s), $T_{largest\ gap}$ is the time of the largest gap in the bunch train and ΔT is a small margin interval. The boundaries of the $T_{largest\ gap}$ are defined in equation (4-39). For example, if the bunch train consists of many bunches spaced by 25 ns (e.g. physics beams) the $T_{largest\ gap}$ is set to the time duration of the abort gap. If the beam consists of a single bunch (e.g. a pilot beam) the $T_{largest\ gap}$ is set to $T_{rev} - \Delta T$. The interval ΔT sets a margin of a few nanoseconds leaving some time for the hold-off circuits to enable its triggering input. The positive edge of the resulting output signal has beam synchronous period T_{rev} and phase relationship fixed to a bunch which is common for all DOROS front-ends. The last flip-flop in the chain synchronizes the FPGA output back to the beam clock domain. The positive edge of the hold-off signal is then converted to a pulse with width T_{FF} defined by the delay in the flip-flop reset circuitry.

The LTG circuit is triggered by a selected bunch passing through the respective BPM, so the bunch time-of-flight is taken into account in the LTG reference clocks. This feature is important when measuring the beam betatron phase advance from one BPM to another. However, the delay introduced by the coaxial cables connecting the BPMs to their DOROS front-ends is also incorporated in the reference clock. Differences in the cable propagation times for particular BPMs translate into a systematic error of the phase advance measurements. Such errors can be calibrated by measuring the propagation delays in the cables.

Please note that the LTG circuit may not lock to the same bunch if there are several gaps in the bunch train which are equal to $T_{largest\ gap}$. For example, if beam is composed of two equally spaced bunches on the LHC circumference the LTG will produce correct output frequency equal to f_{rev} . However, the phase reference in each front-end may differ by 180° depending on the circuit initial conditions during beam injection or after a circuitry reset. The LTG may be therefore used for phase alignment only with suitable beam patterns that satisfy the unique $T_{largest\ gap}$ condition. A pilot beam injected in the beginning of each cycle consists of only a single bunch. Setting the $T_{hold-off}$ to the abort gap length guarantees in this case a stable and un-ambiguous phase reference. The BST phase alignment can be thus automatically checked and realigned if necessary during this part of the LHC cycle.

4.4.3.3 Differential phase detector

The diagram in Fig. 4-38 shows the differential phase detector (DPD) circuit along with example waveforms in its most important nodes. In order to achieve an accurate zero-phase alignment of two reference clocks with f_{rev} frequency, the DPD was optimized to resolve the 100 ps phase steps over an 89 μ s clock period. The time difference between the respective signal edges is converted into a DC voltage digitized by one of the 24-bit ADC channels. The circuit architecture was also optimized to achieve high accuracy and long term stability of the measurement. The DPD has a number of advantages:

- the phase-to-voltage characteristic has an extremum when the input clocks are in phase,
- the measurement at zero-phase condition is independent of the power-supply and ADC reference voltages,
- by design the detector operates with 50 % duty-cycle clocks and is free of the “dead zone” problem [73].

The 50 % duty cycle of the reference clock signals is guaranteed by the two D flip-flops at the DPD inputs operating as simple frequency dividers. The following DPD circuits thus operate with signals at twice lower frequency with period $2T_{rev}$. The time difference between the respective clock edges, denoted as Δt , is defined on an interval:

$$\Delta t \in \left\langle -\frac{2T_{rev}}{2}, \frac{2T_{rev}}{2} \right\rangle \quad (4-40)$$

The flip-flops drive the inputs of an OR gate followed by two XOR gates working in a buffer and an inverter configurations. Such setup guarantees an equal propagation delay for both outputs. Together the OR and XOR gates form logical functions equivalent to OR and NOR gates. The NOR function converts the time delay Δt into duty cycle of the output signal.

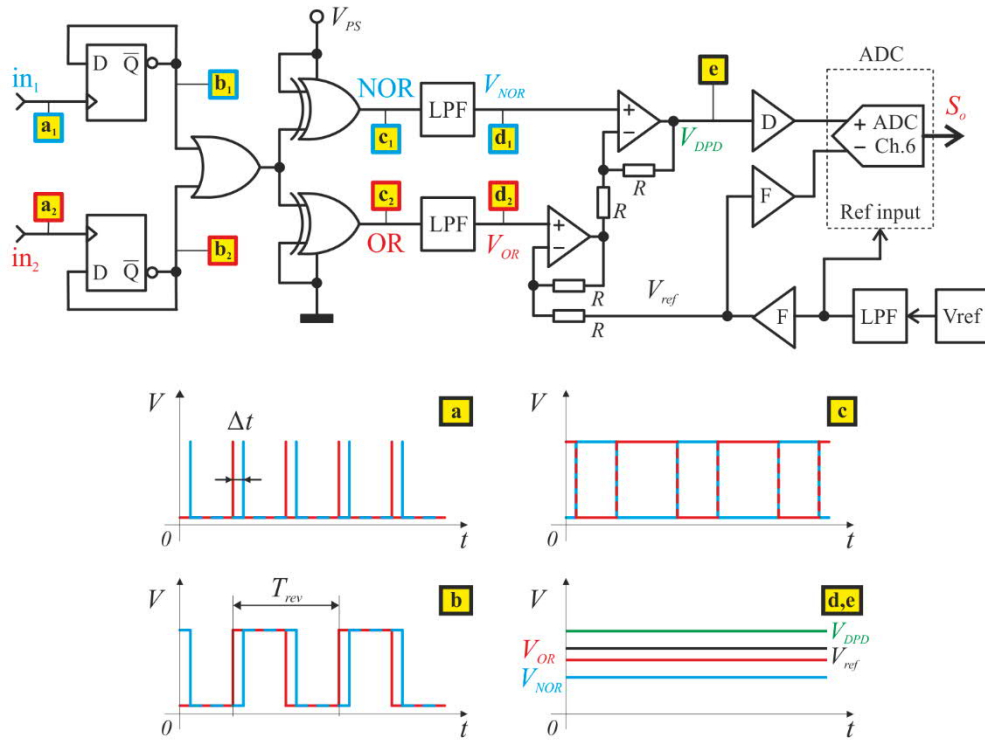


Fig. 4-38. Block diagram of the differential phase detector circuit with example signals in its important nodes.

Abbreviations: DPD – Differential Phase Detector, D – ADC driver, F – Follower, PS – Power Supply, LPF – Low-Pass Filter.

The signal from the NOR is then low-pass filtered, resulting in the average voltage V_{NOR} changing linearly with Δt as:

$$V_{NOR} = V_{PS} \left(\frac{1}{2} - \frac{|\Delta t|}{2 T_{rev}} \right) \quad (4-41)$$

where V_{PS} is the power supply voltage of the gates. As the used gates are from the CMOS family, the output voltage of the XOR gates can be approximated as either 0 V (corresponding to logical 0) or the power supply V_{PS} (corresponding to logical 1). Identical processing is performed with the signal from the OR, yielding the low-pass filtered voltage V_{OR} :

$$V_{OR} = V_{PS} \left(\frac{1}{2} + \frac{\Delta t}{2 T_{rev}} \right) \quad (4-42)$$

The low-pass filters, identical for both circuit branches, are built as three simple RC sections. When both input clocks are exactly in phase, i.e. $\Delta t = 0$, the output signals of the NOR and OR have equal 50 % duty cycles and V_{NOR} is equal V_{OR} . Please note that at condition $\Delta t = 0$ the voltage V_{NOR} has a minimum and the voltage V_{OR} a maximum. The following amplifier stage amplifies the V_{NOR} and V_{OR} voltages obtained from the low-pass filters and subtracts them to obtain the output voltage V_{DPD} :

$$V_{DPD} = V_{ref} - 2V_{OR} + 2V_{NOR} \quad (4-43)$$

The amplifier gain of two is set by the precision resistors R with equal values. The dynamic range of the ADC is between 0 V and $2V_{ref}$, therefore the circuit adds V_{ref} so that the DPD operates in the middle of the ADC dynamic range. By inserting V_{NOR} and V_{OR} voltages obtained from (4-41) and (4-42) the simplified DPD output voltage can be expressed as:

$$V_{DPD} = V_{ref} - 4V_{PS} \frac{|\Delta t|}{2 T_{rev}} \quad (4-44)$$

The detector voltage equals to V_{ref} when both input signals are exactly in phase (the case $\Delta t = 0$). In this condition the V_{DPD} is independent of the power supply voltage, contrary to the “single-ended” voltages V_{NOR} and V_{OR} . The amplifier gain compensates the sensitivity loss caused by the frequency division in the input flip-flops. The following stage buffers the DPD voltage, which is then converted by the ADC into normalised digital signal S_O changing from 0 to 1:

$$S_O = \frac{1}{2} - 2 \frac{V_{PS}}{V_{ref}} \frac{|\Delta t|}{2 T_{rev}} \quad (4-45)$$

For $\Delta t = 0$ the ADC signal $S_O = 0.5$ and it is independent of the gates power supply V_{PS} (about 3.3 V) and the reference voltage V_{ref} (very close to 2.5 V). These voltages set only the slope of the phase-to-voltage conversion, whose actual value is not important for the precise phase alignment. Please note that at the $\Delta t = 0$ condition the phase-to-voltage characteristic reaches an extremum. Due to the input flip-flops the characteristic has an ambiguity between the minimum and the maximum. The circuit was optimized to operate around the maximum (equal to $S_O = 0.5$) when the flip-flops operate synchronously. Otherwise the detector characteristic converges to a minimum which is outside of the ADC input dynamic range. To remove the ambiguity of the detector the input frequency dividers can be synchronised to operate in-phase using the asynchronous reset inputs of the flip-flops (not shown in the picture). The reset is synchronously controlled from the FPGA. Stability and precision of the zero-phase alignment is achieved by minimizing asymmetries between the NOR and OR circuit branches. The input flip-flops are part of one chip, similarly the XOR gates, the operational amplifiers and the followers. To minimise slow drifts the circuit uses precision operational amplifiers and passive components with optimised thermal stability. Small footprint of the circuit minimises the temperature gradients on the PCB.

5 Orbit subsystem results based on laboratory measurements

5.1 DOR detector linearity

The presented measurements were performed with beam signals simulated by pulses from a generator. The pulse amplitude was changed to simulate beam intensity decay typical for long LHC physics fills. Relative amplitude and position changes were calculated on a server receiving signal samples processed by the measured DOROS front-end. The linearity of the orbit detectors is studied by comparing the measured amplitudes to the reconstructed linear modulation of the input signal pulses.

5.1.1 Measurement setup

Linearity of the DOROS orbit subsystem was measured using the laboratory setup, whose simplified diagram is shown in Fig. 5-1. Beam signals were simulated by a waveform signal generator Keysight 33600A, providing 5 ns pulses to the beam inputs of the DOROS front-end. The real beam pulses from the LHC BPMs are much shorter, in the order of 1 ns, however, the used 5 ns pulse width was the shortest available. As explained in the chapter 4.3.2, on the inputs of the DOROS front ends there are non-reflective 80 MHz low pass filters, which increase the pulse length to some 5 ns. This is why 5 ns pulses are considered as acceptable for the presented measurements. The repetition frequency of the pulses was 40 MHz to simulate the 25 ns bunch spacing of typical LHC beams. The amplitude variation of the pulses was achieved by their amplitude modulation using a slow triangle waveform. The AM depth was set to 100 % to scan the entire ADC dynamic range. The programmable RF gain was set to some +24 dB (fine gain index 52). The amplitude was set such that the ADC could be saturated just before reaching the maximum of the modulated signals. The chosen pulse amplitude of some 2.7 Vpp was far enough from the generator limit of 4 Vpp. The modulation frequency of 1 mHz produced some 500 calibrated samples for each modulation slope. This is equivalent to some 10 mV amplitude increase of the generator signal for each measurement point. The calibrated signal amplitudes were provided by the measured DOROS front-end at a 1 Hz rate.

The signal from one generator output was split between the four front-end inputs by a passive splitter as shown in Fig. 5-2. This arrangement allowed using standard SMA attenuators while minimising asymmetries between the processing channels. The RF gain stages in the front-end were configured to provide a 24 dB gain by combining a single 10 dB stage and 14 dB gain of the IPGA. More details can be found in chapter 4.3.2.1.

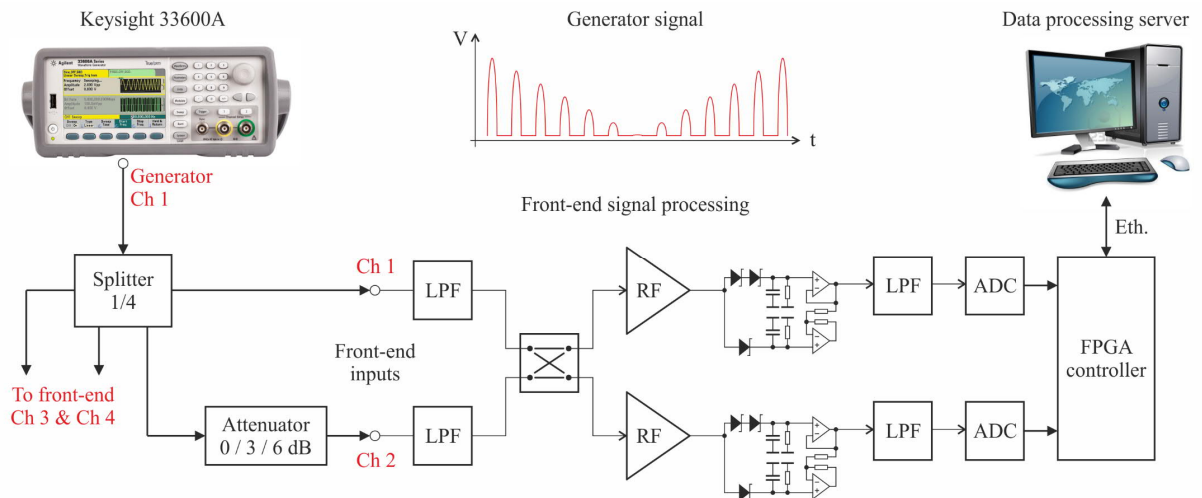


Fig. 5-1. Diagram of the laboratory setup used to measure linearity of the diode detector characteristic. The signal generator Keysight 33600A was used to generate pulses with amplitude modulation. **Abbreviations:** LPF – Low Pass Filter, RF – Radio Frequency, Eth – Ethernet communication.

The DOROS orbit processing is characterised in the following measurements at three simulated beam positions. The positions were selected by using passive RF attenuators connected to the even inputs of the front-end. The 3 dB and 6 dB attenuators gave simulated positions of 2.6 mm and 5.1 mm respectively, assuming a linear position approximation on 61 mm BPM aperture according to equation (3-2). Beam offsets of a similar order are common for LHC BPMs near experiments. Centred beam was simulated using equal signals on the front-end inputs.

In Fig. 5-3 is shown an example of front-end input signals measured with a 12-bit oscilloscope Teledyne-Lecroy HDO6104. Channels 1 and 3 were connected directly to the splitter outputs, while channel 2 and 4 were connected through 6 dB and 3 dB attenuators, respectively. The oscilloscope had a 1 GHz analogue bandwidth and a 2.5 GS/s sampling rate increased to 125 GS/s by using the Random Input Sampling (RIS) mode.

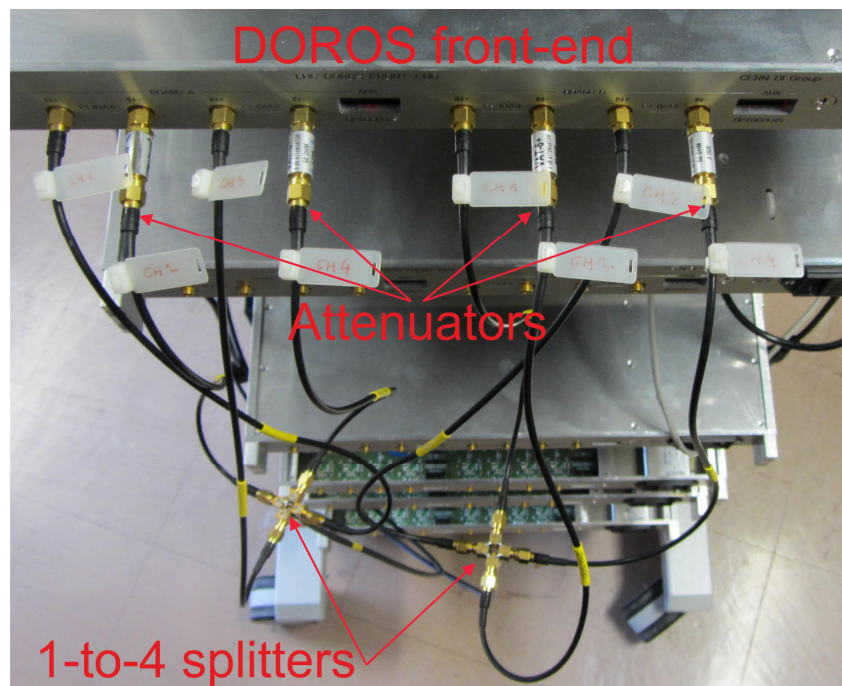


Fig. 5-2. A photograph of the measurement setup in the laboratory. Two 1-to-4 splitters were used to split signals from two generator channels to all eight front-end inputs. Attenuators on the even front-end inputs simulated beam position offset.

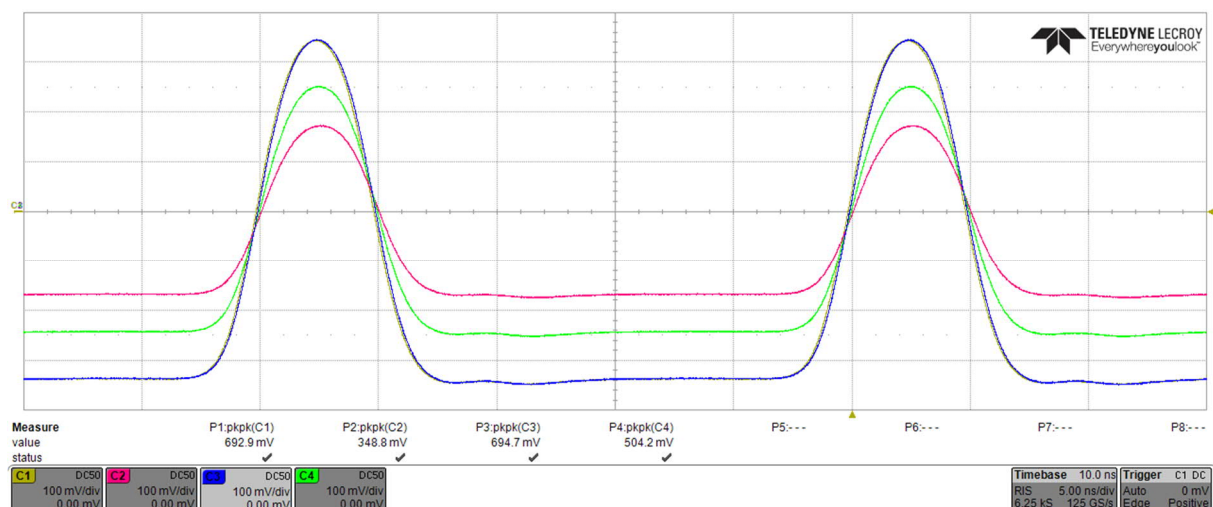


Fig. 5-3. Output signals of the 1-to-4 splitter measured with the oscilloscope. The C1 and C3 signals were connected directly to the splitter. The C2 signal (red) was connected to the splitter output through a 6 dB attenuator and C4 (green) through a 3 dB attenuator.

5.1.2 Linearity of a compensated diode detector

The plot in Fig. 5-4 shows an example of an amplitude scan used to evaluate the linearity of a compensated diode detector. The presented channel 1 signals were obtained with the setup of Fig. 5-1. The measured channel was connected to the signal splitter with no attenuator. In order to measure the whole detector dynamic range the generator signal amplitude was going 6 % above the ADC full-scale, that is some 30 mV.

The calibration multiplexer was switching the inputs of the DOR channels at the 1 Hz rate. “Straight” and “crossed” signals in the plot correspond to the two configurations of the multiplexer. They were obtained by de-multiplexing the signals and removing the transient part in each period. The calibrated signal in the plot is a result of averaging the neighbouring samples measured in straight and crossed configurations. As explained in chapter 4.3.1, such averaging results in compensating the asymmetry of the channel pair after the input multiplexer. The relative error between the straight and crossed signals was measured to be in the order of 1 % for amplitudes close to the ADC full-scale.

The analysis was done upon data corresponding to the first positive ramp of the scan. This corresponds to the first 8 minutes of the measurement. The following plot in Fig. 5-5 depicts the calibrated signal in the range corresponding to the full ADC dynamic range. The horizontal axis is scaled in the equivalent generator amplitude normalized by the ADC full-scale. The minimum point on the axis corresponds to zero voltage on the ADC input. The maximum point on the axis corresponds to the sample value measured close to the ADC full-scale. The remaining points on the axis were linearly distributed assuming linear amplitude change of the triangle modulation.

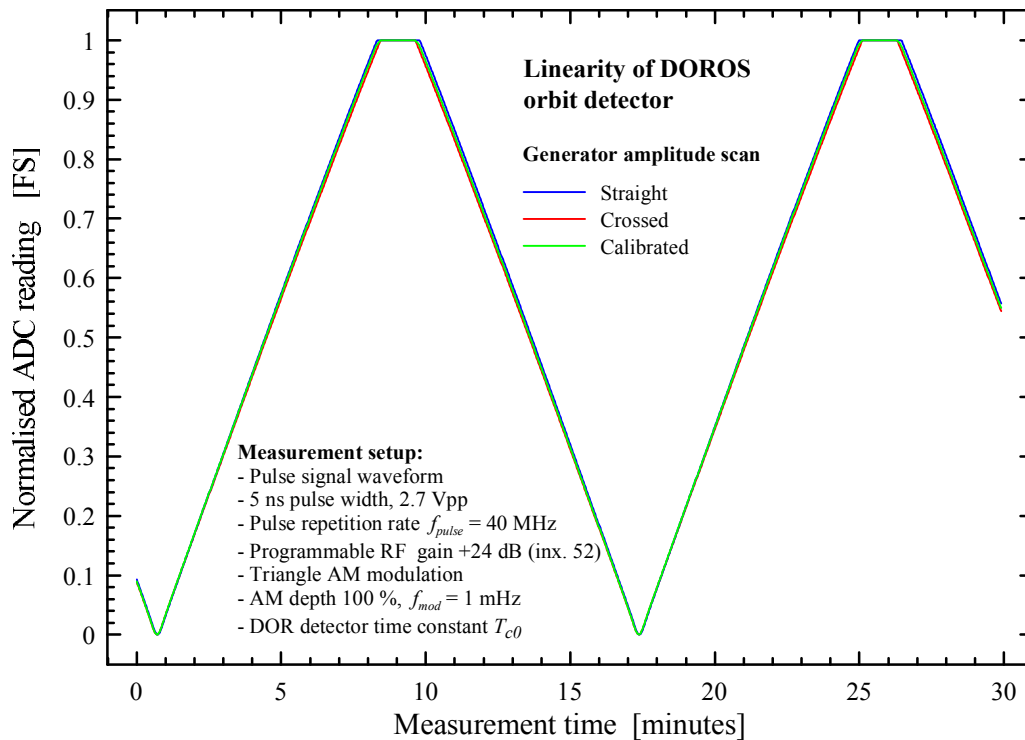


Fig. 5-4. Amplitude scan of a DOROS orbit channel. The generator was set to simulate LHC beams with 5 ns pulses, 40 MHz repetition rate and amplitude 2.7 Vpp. The pulse amplitude was changed with triangle AM modulation with 100 % depth and 1 mHz frequency. Displayed straight and crossed signals were obtained from the two states of the input signal multiplexer. The calibrated signal was computed as a moving average of two consecutive samples measured in each multiplexer state. The signal processing was configured to 1 Hz switching rate, 24 samples masked after each multiplexer switching and 10 Hz cut-off frequency of the IIR low-pass filter. The detectors were configured to time constant T_{c0} .

The linearity of the data set was analyzed using the linear regression method. The analysis was focused on the data with amplitudes typical for beam operation, namely 0.3 - 0.8 of the ADC full-scale. The linear fitting was also optimized for this range. The measurement reproducibility was verified by repeating the linear regression analysis for other ramps of the measurement. Relative differences in the regression linear coefficients for both positive and negative ramps were measured to be in order of 0.1 % with respect to the first measurement.

The difference between the measured characteristic and the linear fitting is displayed in Fig. 5-6. The measurement was done for the three detector time constants. The fitting parameters were summarized in the plot. It can be seen in the plot that the most linear characteristic was obtained with time constant T_{c0} , corresponding to the slowest discharge of the detector capacitor. The characteristic of the detector deviates further from the fitted line when the discharge increases. The deviation of 1.5 % was measured with time constant T_{c2} corresponding to the fastest discharge. Such deviation is equivalent to some 80 mV over the 5 V dynamic range of the ADC.

The measured detector characteristics were approximated with polynomial fitting. The optimization was performed for the signals with amplitudes in the range 0.3 - 0.8 FS. The deviation from the measured characteristics are shown in Fig. 5-7 for time constants T_{c0} and T_{c2} . The linear function is plotted for comparison. It can be seen that the second order polynomial has fitting errors below 300 ppm equivalent to 15 mV on the ADC input. Higher order polynomials decreased the fitting error mostly around the lower part (between 0.2 and 0.3) and higher part (between 0.8 and 0.9) of the characteristic. Increasing the fitting to a third order polynomial decreased the error in the boundary ranges by a factor of 2. Obtained fitting coefficients for the three detector time constants are listed in Table 5-1, Table 5-2 and Table 5-3 respectively.

The polynomial coefficients depend on the actual detector time constant and their change is in the order a few percent. Using a single set of coefficients to approximate the detector characteristic for all three time constants would yield fitting errors in the same order of magnitude. In practice such errors would be too large for linearization of the detector characteristic. Therefore correction of the diode detector characteristic has to have its own set of coefficients for each time constant setting. The so called linearization polynomials were computed by performing inversed polynomial fitting on the measured data. The resulting polynomials can be used to linearize the characteristic of the compensated diode detectors. Obtained linearization coefficients for the three detector time constants are listed in Table 5-4, Table 5-5 and Table 5-6 respectively.

In the discussed measurements any nonlinearity of the modulation contributes to the final results. The modulation nonlinearity appears as a nonlinear distribution of the measurement points on the horizontal axis. To remove this error the linearity of the amplitude modulation would have to be measured with precision much better than 1 %. Such measurements were not possible with available equipment. Contribution of the generator nonlinearity to the actual measurement of the DOR characteristic was verified with a 12-bit oscilloscope. Description of the measurement and obtained results are presented in chapter 5.1.4.

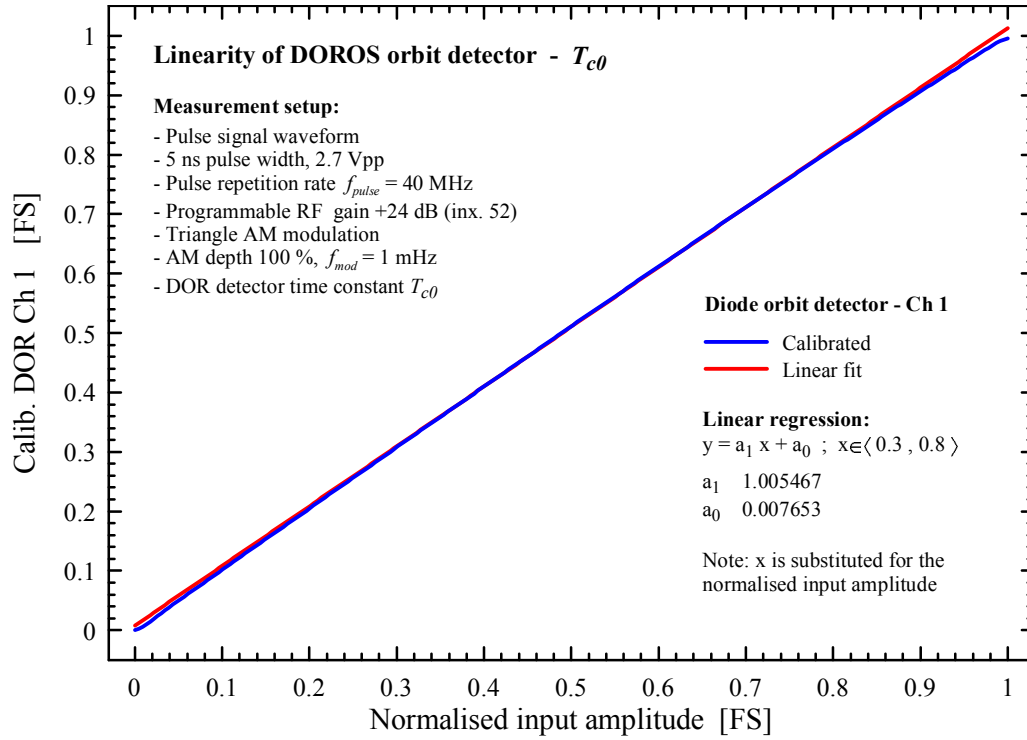


Fig. 5-5. Amplitude scan of a compensated diode detector in a DOROS channel. The horizontal axis scale is normalized to the generator amplitude corresponding to the maximal ADC reading. The diode detectors were configured with time constant T_{c0} . The generator was set to produce 5 ns pulses with 40 MHz repetition rate, simulating BPM signals obtained with LHC physics beams. The pulse amplitude on the generator output was set to 2.7 Vpp. The input signal amplitude was swept using a triangle AM modulation with 100 % depth and 1 mHz frequency. The signal processing was configured to 1 Hz calibration switching, 24 samples masked after each multiplexing and IIR low-pass filtering of the samples with a 10 Hz cut-off.

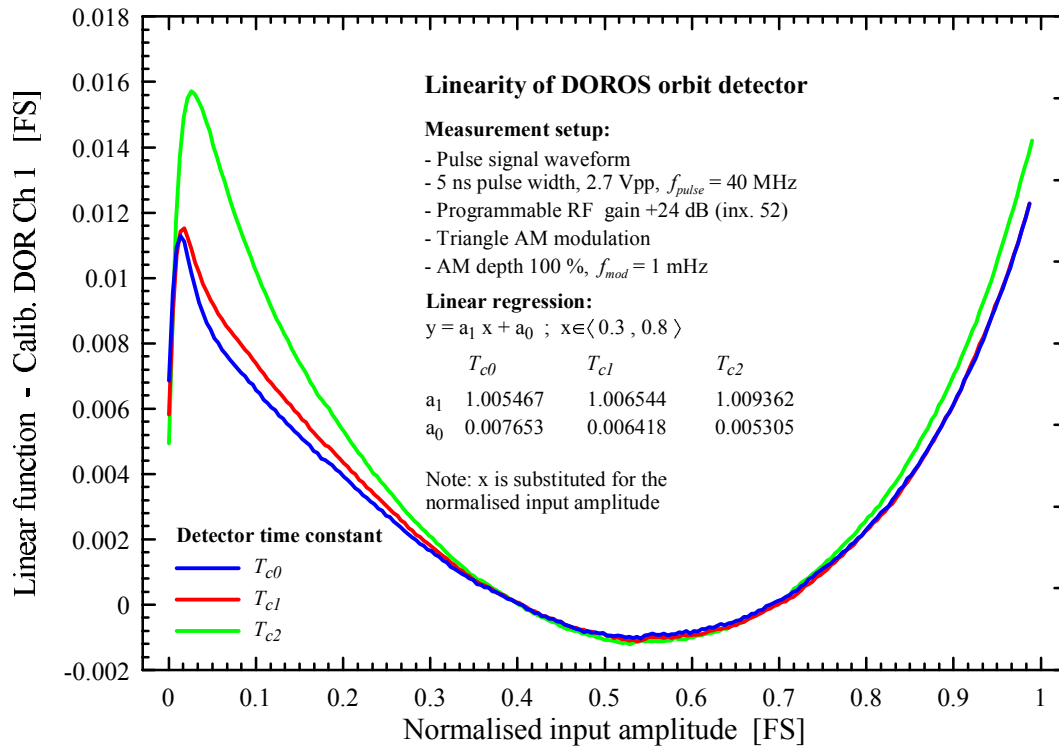


Fig. 5-6. Deviation of the calibrated detector characteristic from the corresponding linear fitting function obtained for the three detector time constants. The linear fitting of each time constant measurement was performed on data in range 0.3 - 0.8 of the ADC full-scale. Time constant T_{c0} corresponds to the slowest capacitor discharge with leakage currents. The time constants T_{c1} and T_{c2} correspond to faster capacitor discharge with 10 M Ω and 1 M Ω resistors, respectively.

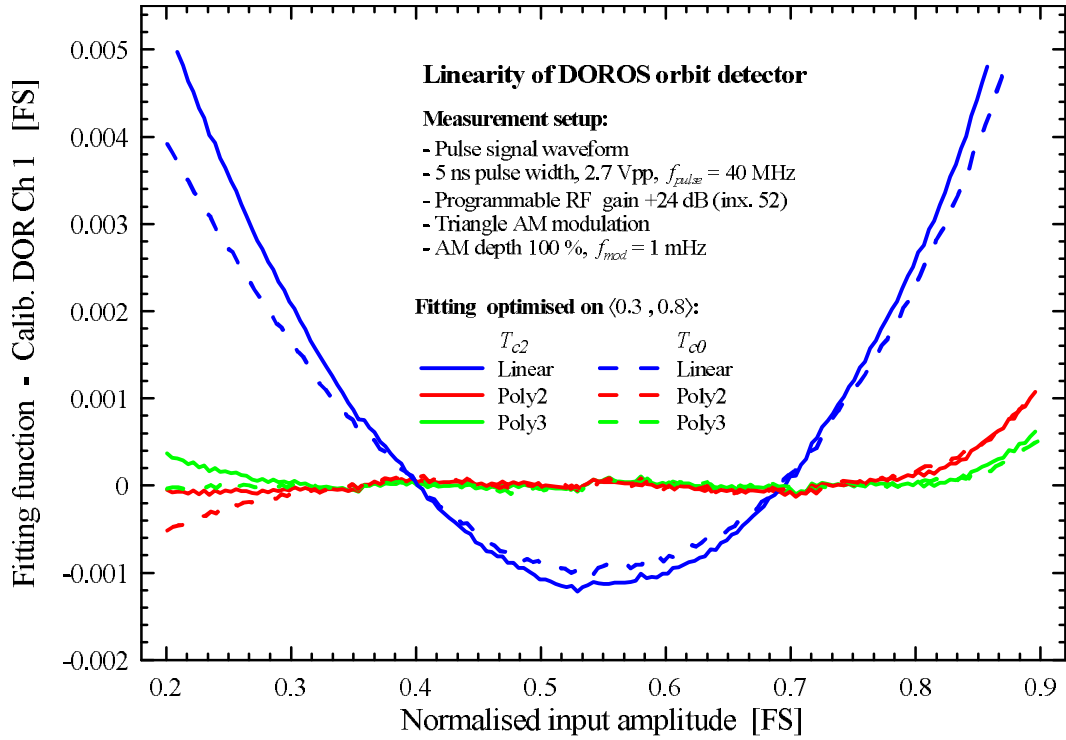


Fig. 5-7. Deviation of the measured detector characteristic from fitting polynomials of the 3rd and 5th orders for T_{c0} and T_{c2} time constants. Corresponding linear fitting is shown for comparison. The polynomial fitting was optimized for data in the range 0.3 - 0.8 of the ADC full-scale. Time constant T_{c0} corresponds to the slowest capacitor discharge and time constant T_{c2} to the fastest. The generator was producing 5 ns pulses with 40 MHz repetition rate and 2.7 Vpp amplitude. The input signal amplitude was swept using a triangle AM modulation with 100 % depth and 1 mHz frequency. The front-end signal processing was configured to 1 Hz calibration switching rate, 24 samples masked after each multiplexing and the IIR low-pass filtering with a 10 Hz cut-off. The linear fitting was performed on data in range 0.3 - 0.8 of the ADC full-scale.

Table 5-1. Coefficients of polynomial fitting of the detector characteristic with T_{c0} time constant.

Polynomial order	T_{c0} Fitting function [†] $y = a_5 x^5 + a_4 x^4 + a_3 x^3 + a_2 x^2 + a_1 x + a_0$					
	a_0	a_1	a_2	a_3	a_4	a_5
2 nd	-0.005046	1.056002	-0.046669	0	0	0
3 rd	-0.002546	1.040780	-0.017291	-0.018087	0	0
4 th	-0.002808	1.042923	-0.023629	-0.010036	-0.003718	0
5 th	0.009096	0.920644	0.464591	-0.958175	0.893317	-0.331367

*Results rounded to 6 digits, [†]x was substituted for the normalised input amplitude.

Table 5-2. Coefficients of polynomial fitting of the detector characteristic with T_{c1} time constant.

Polynomial order	T_{c1} Fitting function [†] $y = a_5 x^5 + a_4 x^4 + a_3 x^3 + a_2 x^2 + a_1 x + a_0$					
	a_0	a_1	a_2	a_3	a_4	a_5
2 nd	-0.007037	1.059729	-0.048722	0	0	0
3 rd	-0.003984	1.041260	-0.013334	-0.021612	0	0
4 th	-0.006794	1.064110	-0.080467	0.063035	-0.038772	0
5 th	-0.001452	1.009555	0.135942	-0.354270	0.353046	-0.143576

*Results rounded to 6 digits, [†]x was substituted for the normalised input amplitude.

Table 5-3. Coefficients of polynomial fitting of the detector characteristic with T_{c2} time constant.

Polynomial order	T_{c2} Fitting function [†] $y = a_5 x^5 + a_4 x^4 + a_3 x^3 + a_2 x^2 + a_1 x + a_0$					
	a_0	a_1	a_2	a_3	a_4	a_5
2 nd	-0.009921	1.069645	-0.055410	0	0	0
3 rd	-0.007784	1.056704	-0.030556	-0.015230	0	0
4 th	-0.010526	1.079024	-0.096246	0.067815	-0.038166	0
5 th	-0.007525	1.048358	0.025573	-0.167595	0.183483	-0.081492

*Results rounded to 6 digits, [†]x was substituted for the normalised input amplitude.

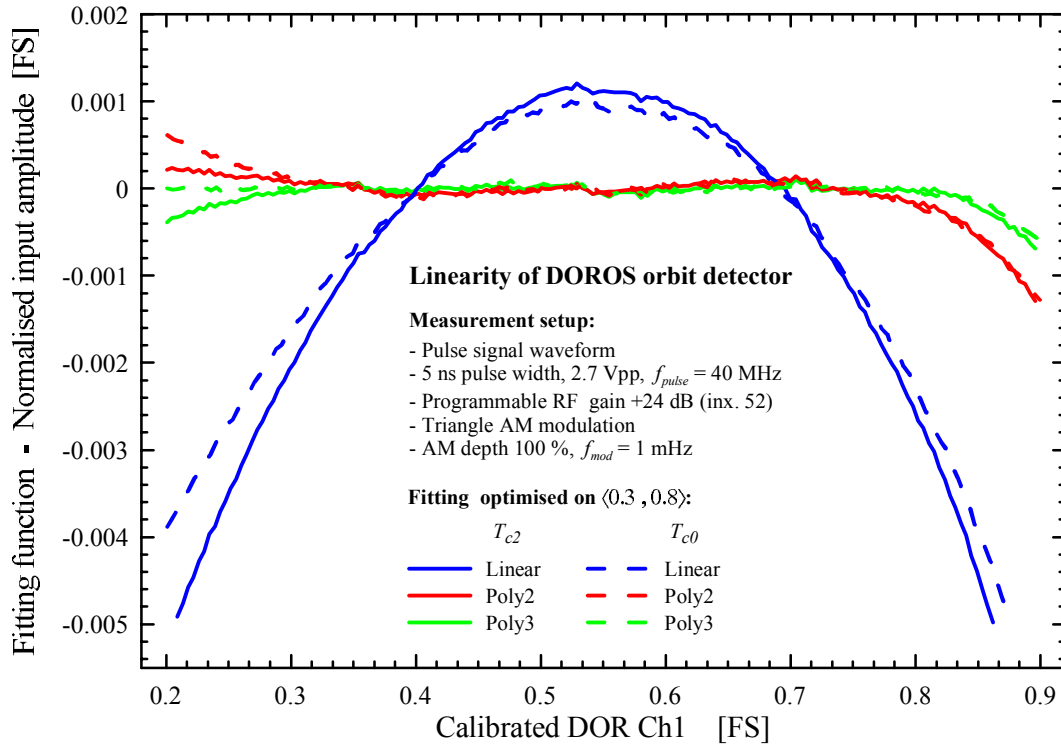


Fig. 5-8. Deviation of the corrected detector characteristic from the normalized input amplitude on the diode detector. The characteristic was corrected by the polynomial functions fitted to the DOR data in range 0.3 - 0.8 of the ADC full-scale. The time constant T_{c0} corresponds to the slowest capacitor discharge through leakage currents. The constant T_{c2} correspond to the capacitor discharge rate through 1 M Ω . The generator was producing 5 ns pulses with 40 MHz repetition rate and 2.7 Vpp amplitude. The input signal amplitude was swept using a triangle AM modulation with 100 % depth and 1 mHz frequency. The front-end signal processing was configured to 1 Hz calibration switching rate, 24 samples masked after each multiplexing and the IIR low-pass filtering with a 10 Hz cut-off. The linear fitting was performed on data in range 0.3 - 0.8 of the ADC full-scale.

Table 5-4. Coefficients of the polynomial fitting to correct the diode detector characteristic with T_{c0} time constant.

Polynomial order	T_{c0} Fitting function [†] $y = a_5 x^5 + a_4 x^4 + a_3 x^3 + a_2 x^2 + a_1 x + a_0$					
	a_0	a_1	a_2	a_3	a_4	a_5
2 nd	0.005363	0.943950	0.045907	0.000000	0.000000	0.000000
3 rd	0.002162	0.963073	0.009651	0.021950	0.000000	0.000000
4 th	0.002745	0.958386	0.023266	0.004950	0.007721	0.000000
5 th	-0.009996	1.086809	-0.480224	0.965737	-0.886018	0.324784

*Results rounded to 6 digits, [†]x was substituted for the calibrated DOR amplitude.

Table 5-5. Coefficients of the polynomial fitting to correct the diode detector characteristic with T_{c1} time constant.

Polynomial order	T_{c1} Fitting function [†] $y = a_5 x^5 + a_4 x^4 + a_3 x^3 + a_2 x^2 + a_1 x + a_0$					
	a_0	a_1	a_2	a_3	a_4	a_5
2 nd	0.007283	0.940477	0.047771	0.000000	0.000000	0.000000
3 rd	0.003482	0.963090	0.005135	0.025644	0.000000	0.000000
4 th	0.006805	0.936515	0.081957	-0.069724	0.043033	0.000000
5 th	0.000783	0.996979	-0.154020	0.378202	-0.371162	0.149540

*Results rounded to 6 digits, [†]x was substituted for the calibrated DOR amplitude.

Table 5-6. Coefficients of the polynomial fitting to correct the diode detector characteristic with T_{c2} time constant.

Polynomial order	T_{c2} Fitting function [†] $y = a_5 x^5 + a_4 x^4 + a_3 x^3 + a_2 x^2 + a_1 x + a_0$					
	a_0	a_1	a_2	a_3	a_4	a_5
2 nd	0.010086	0.931092	0.053874	0.000000	0.000000	0.000000
3 rd	0.007064	0.949088	0.019873	0.020509	0.000000	0.000000
4 th	0.010217	0.923848	0.092955	-0.070426	0.041150	0.000000
5 th	0.006566	0.960547	-0.050475	0.202364	-0.211718	0.091558

*Results rounded to 6 digits, [†]x was substituted for the calibrated DOR amplitude.

5.1.3 Linearity of the diode orbit measurement

Measurements presented in this chapter quantify the linearity of the DOROS orbit subsystem with beam signals simulated in the laboratory by a generator.

Data signal processing is similar to the one used on the preceding chapter. Simulated beam orbits are calculated using the difference and the sum signals of two calibrated DOR channels, normally processing signal from one pair of BPM electrodes. The positions presented in the following plots were computed using the equation (3-2) for linear approximation of the BPM characteristic. Please note that the positions were referenced to the 61 mm aperture of the Q1 BPMs (directional couplers of BPMSW type). The horizontal axis was projected to the equivalent generator amplitude on the input normalized to the ADC full-scale. The position offsets were simulated using external attenuators of 3 dB and 6dB values. The centered beam was simulated with no attenuators on the front-end inputs.

The plot in Fig. 5-9 presents the position change as a function of the normalized input amplitude. The characteristics were measured with the simulated centered beam for the three detector time constants. The position change was referenced to the ADC full-scale which corresponds to value 1 on the horizontal axis. As seen, the position changes up to 1 μm for the presented range 0.2 - 1 FS. During operation with beam the signal levels are typically maintained within the range of 0.4 - 0.8 FS. The amplitude on the horizontal axis is also proportional to the beam intensity. This means that a change in the intensity of the beam is translated to a false change in the position. As seen on the plot, the position would change by some 0.2 μm if the intensity would change within the operating range by 50 %. Position differences caused by changing the detector time constants are smaller than 0.3 μm . The resolution of the discussed measurements is some 0.05 μm peak-peak.

The systematic position errors increase by some two orders of magnitude once the simulated beam offset is 2.6 mm, as shown in Fig. 5-10. Depending the used time constant the orbit changes by up to 45 μm when the signal amplitude drops by 50 % within the operating range of 0.4 - 0.8 FS. Such amplitude change can be caused by a beam intensity decay during an LHC physics fill. Therefore, if the position systematic errors are not compensated in the operational systems, then the errors can reach the levels measured in the laboratory. The smallest position error was measured with the time constant T_{c0} . The relative improvement measured at 0.4 FS is some 7 μm (8 % of the absolute error) with respect to T_{c1} and some 25 μm (30 % of the absolute error) when compared to T_{c2} . However, when operating the signals at 0.8 FS a significant improvement was achieved only with respect to the T_{c2} , yielding some 13 μm difference (30 % of the absolute error).

For the third measurement the simulated beam position was increased to 5.1 mm by attenuating one of the input signals by 6 dB. The measurement results are shown in Fig. 5-11 for the three time constants. It can be seen that the position changes by up to 80 μm for a 50 % amplitude drop. Increasing the position offset by factor 2 increased the position systematic errors by a similar factor.

Similarly to the previous measurements, the smallest position error was measured with the time constant T_{c0} . The relative improvement at 0.4 FS yielded some 16 μm (12 % of the absolute error) with respect to T_{c1} and some 56 μm (40 % of the absolute error) when compared to T_{c2} . The improvement measured at 0.8 FS yielded some 17 μm (30 % of the absolute error) with respect to T_{c2} . The presented results show that the systematic errors introduced by the orbit detectors operated at low amplitudes with the time constant T_{c0} are similar to the corresponding errors for T_{c2} when the detectors are operated close to the ADC saturation.

All measured characteristics can be compared on a single plot showed in Fig. 5-12. The characteristics for different time constants are shown with dash lines and for different position offsets are marked with different colors. The plot presents all measurement results with the same scale on the vertical axis between 0 and 400 μm .

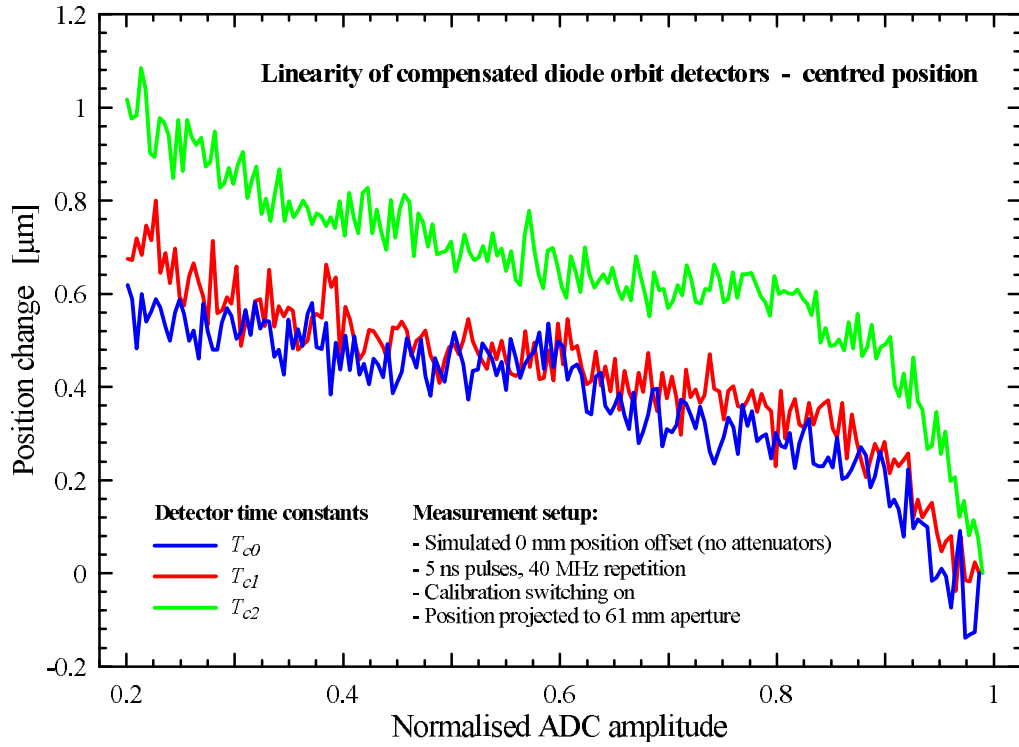


Fig. 5-9. Calibrated position as a function of the generator amplitude normalized to the ADC full-scale. The centered position was simulated by applying equal signals on the two DOR channels.

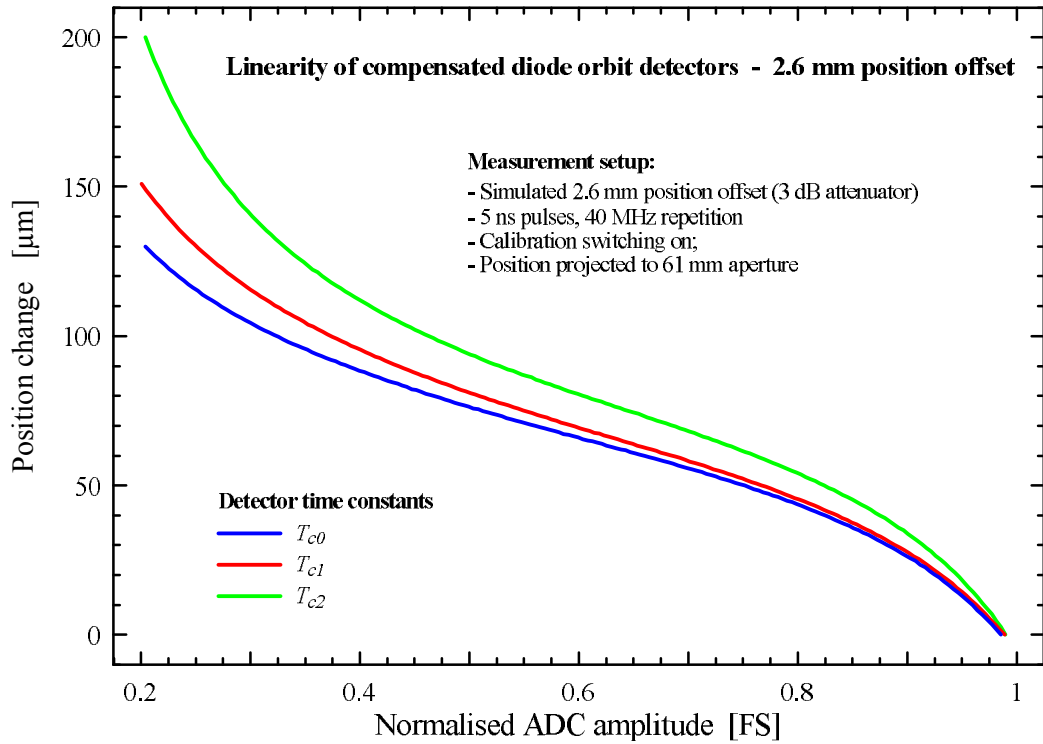


Fig. 5-10. Calibrated position as a function of the generator amplitude normalized to the ADC full-scale. The position offset of 2.6 mm was simulated by inserting a 3 dB attenuator in the path of the generator signal connected to DOR channel 2.

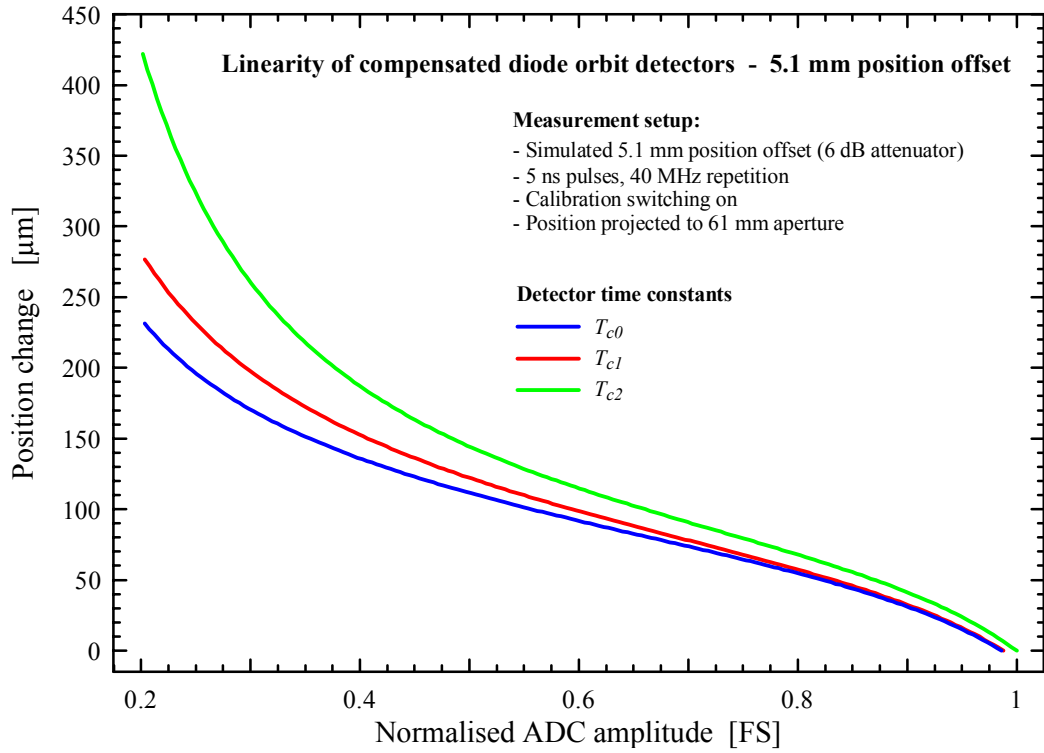


Fig. 5-11. Calibrated position as a function of the generator amplitude normalized to the ADC full-scale. The position offset of 5.1 mm was simulated by inserting a 6 dB attenuator in the path of the generator signal connected to DOR channel 2.

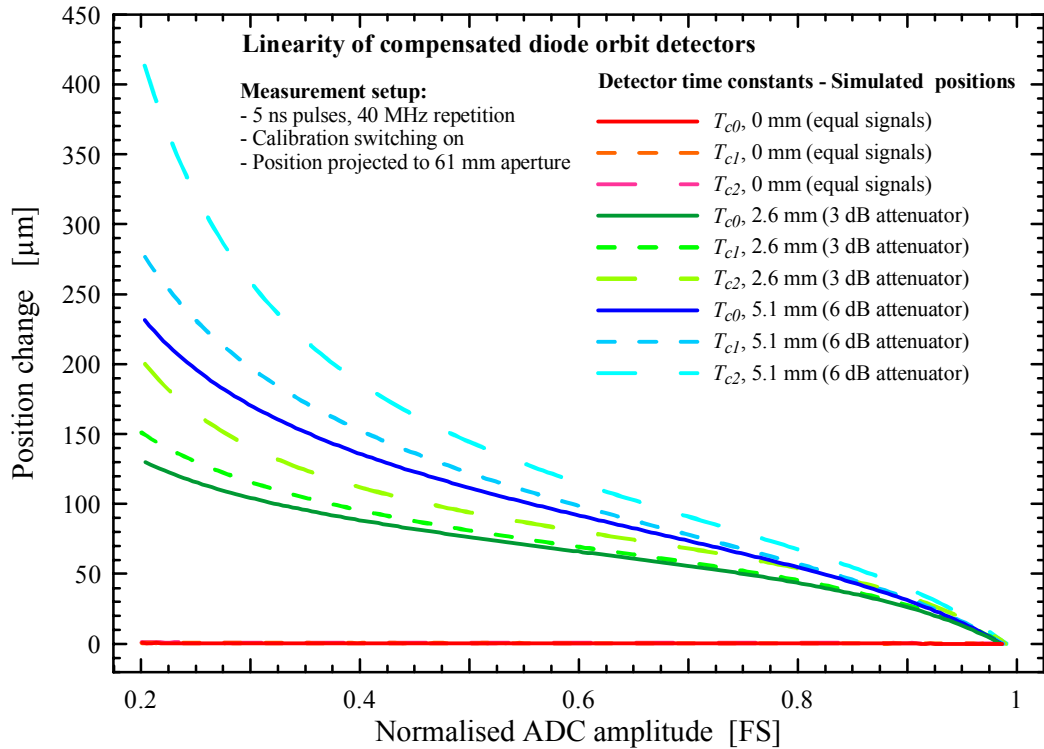


Fig. 5-12. Calibrated position as a function of the generator amplitude normalized to the ADC full-scale. The computed positions are referenced to a 61 mm aperture BPM. The position offsets were simulated by inserting attenuators in the path of the generator signal connected to DOR channel 2. Detector time constant T_{c0} corresponds to the slowest capacitor discharge through leakage currents. The constants T_{c1} and T_{c2} correspond to the capacitor discharge with 1 M Ω and 10 M Ω resistors. The generator was set to simulate LHC beams with 5 ns pulses, 40 MHz repetition rate and amplitude 2.7 Vpp. The amplitudes of the pulses were scanned using a triangle AM modulation with 100 % depth and 1 mHz frequency.

5.1.4 Linearity of the generator amplitude modulation

The linearity measurements of the DOROS channels presented in the previous chapter assumed that the triangular amplitude modulation of the generator used to generate the input signals is linear. In this chapter this assumption is verified by measuring the pulse amplitudes using DOROS channels in parallel with a 12-bit oscilloscope. Diagram of the measurement setup is shown in Fig. 5-13. It is based on the same components which were used in previous measurements. The goal of these measurements was to confirm, that the residual nonlinearity of the position measurements comes from the DOROS system, not from the generator. A photograph of the measurement setup in the laboratory is shown in Fig. 5-14.

The generator was setup with pulses having 5 ns width and 40 MHz repetition rate. The generator amplitude was set to 2.7 V_{pp} to reach the ADC full-scale with 24 dB gain in the programmable RF amplifiers. The amplitude modulation was set to the triangular shape with 100 % modulation depth and 250 μ Hz frequency. The frequency was set such that one linear ramp used for a full amplitude scan takes about 30 minutes. The generator signal was then split using SMA one-to-two “galvanic” adapters to produce four copies of the same signal. The 6 dB attenuator connected to channel 2 simulated the 5.1 mm position offset. The used oscilloscope inputs were set with AC coupling and 1 M Ω impedance. An example of the signals measured with the oscilloscope is shown in Fig. 5-15. The plot displays both the channel 1 signal and the channel 2 signal after the 6 dB attenuator.

The amplitude was periodically measured on each measurement device every 2 minutes during the modulation ramp. This way some 20 discrete measurements were obtained across the whole dynamic range of the DOROS ADCs. The amplitude of the generator pulses was changing between each measurement by 5 %, which is equivalent to a voltage change of some 130 mV.

The amplitude estimation in the oscilloscope was based on standard deviation (STD) computed upon 1 s of the buffered data. To improve the precision of the measurement, the STD values were averaged for a duration of 1 minute. The relative peak-peak error of the averaged STD results was below 0.3 %, yielding better precision than a peak-peak measurement available in the oscilloscope.

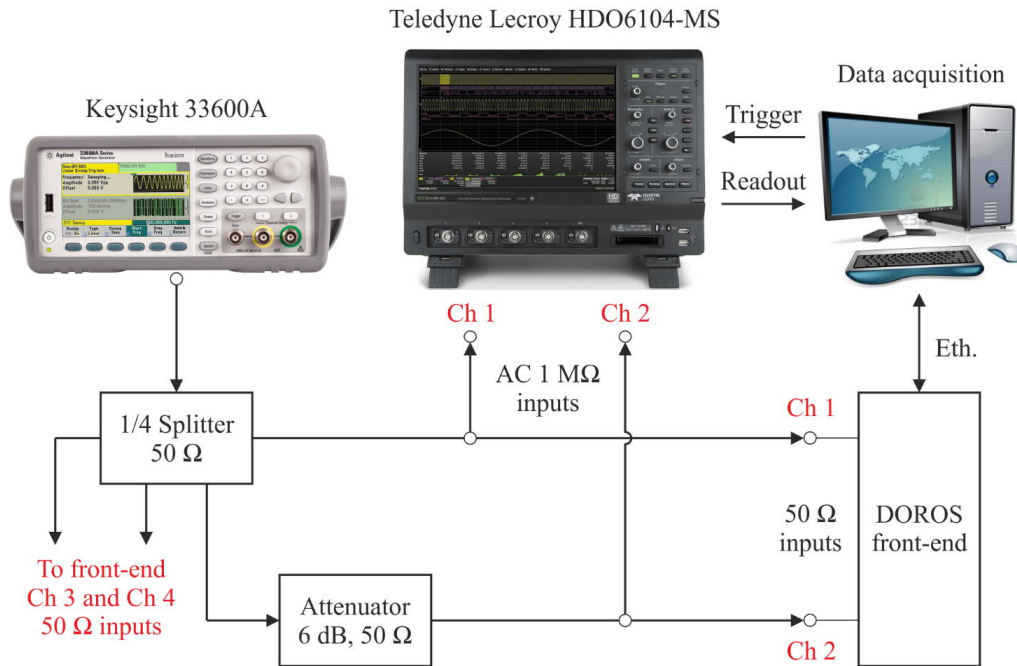


Fig. 5-13. Laboratory setup used to measure amplitude of the modulated pulse signals from the signal generator. The amplitudes were measured in parallel with both, the DOROS front-end and the 12-bit oscilloscope.

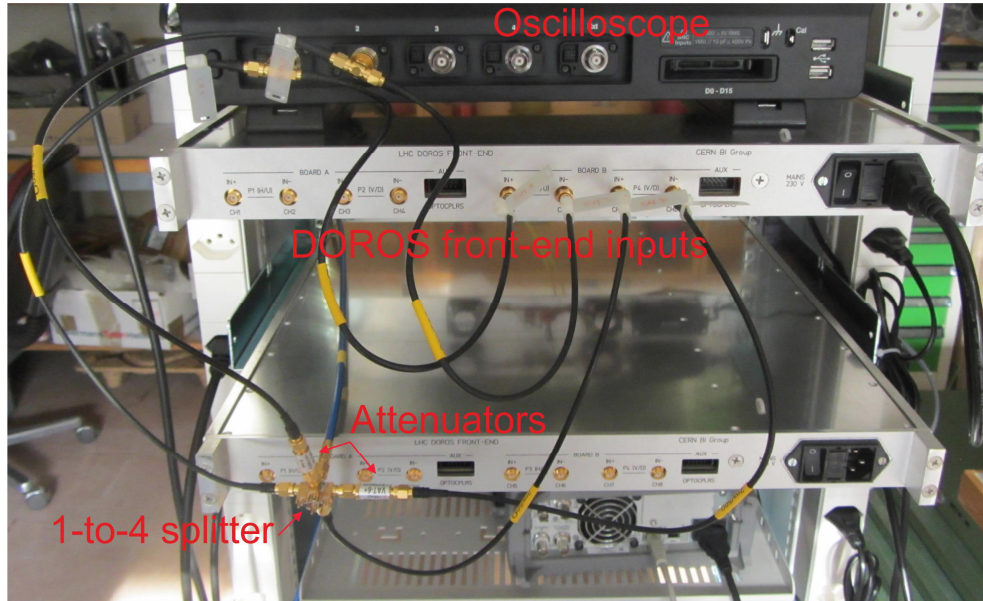


Fig. 5-14. Photograph of the laboratory setup used to measure amplitude of the modulated pulse signals from the signal generator. The amplitudes were measured with DOROS front-end and 12-bit oscilloscope connected in parallel.

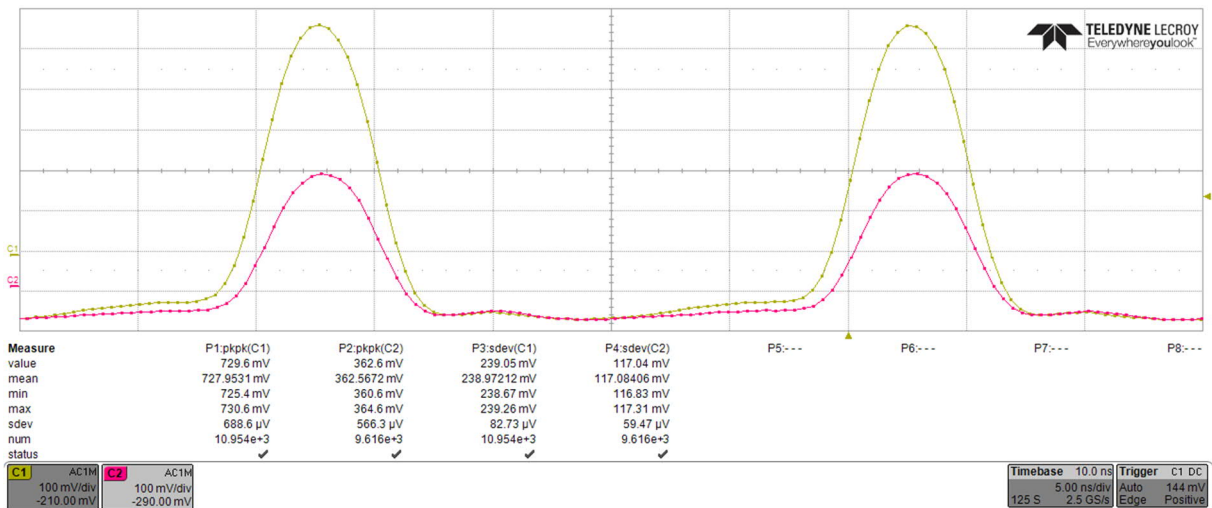


Fig. 5-15. An example of the generator signals measured with the 12-bit oscilloscope. The scope was acquiring in parallel to the DOROS front-end. The channel 1 measurement (C1 in yellow) represents the signal from the splitter. The channel 2 measurement (C2 in magenta) shows the signal from the splitter followed by a 6 dB attenuator.

The DOROS front-end was used to measure the amplitude and position in parallel to the oscilloscope. The amplitude measurement after the diode detectors was based on computing the calibrated DOR samples with 1 Hz multiplexing. The diode detectors were set to T_{c2} time constant. As shown in the chapter 5.1.3, such setup with the 6 dB amplitude difference between the two channels yields the most non-linear characteristic of the diode detectors.

Obtained amplitudes were then used to compute positions in each point of the amplitude ramp. Similarly to the previous measurements the positions were referenced to the 61 mm BPM aperture. Obtained results from both the oscilloscope and the DOROS are plotted in Fig. 5-16 for their comparison. The horizontal axis was scaled to fit to the results from the channel 1 oscilloscope measurement. The vertical axis displays the relative position change in micrometers. Displayed position change is relative to the last position value measured before reaching the DOROS ADC saturation.

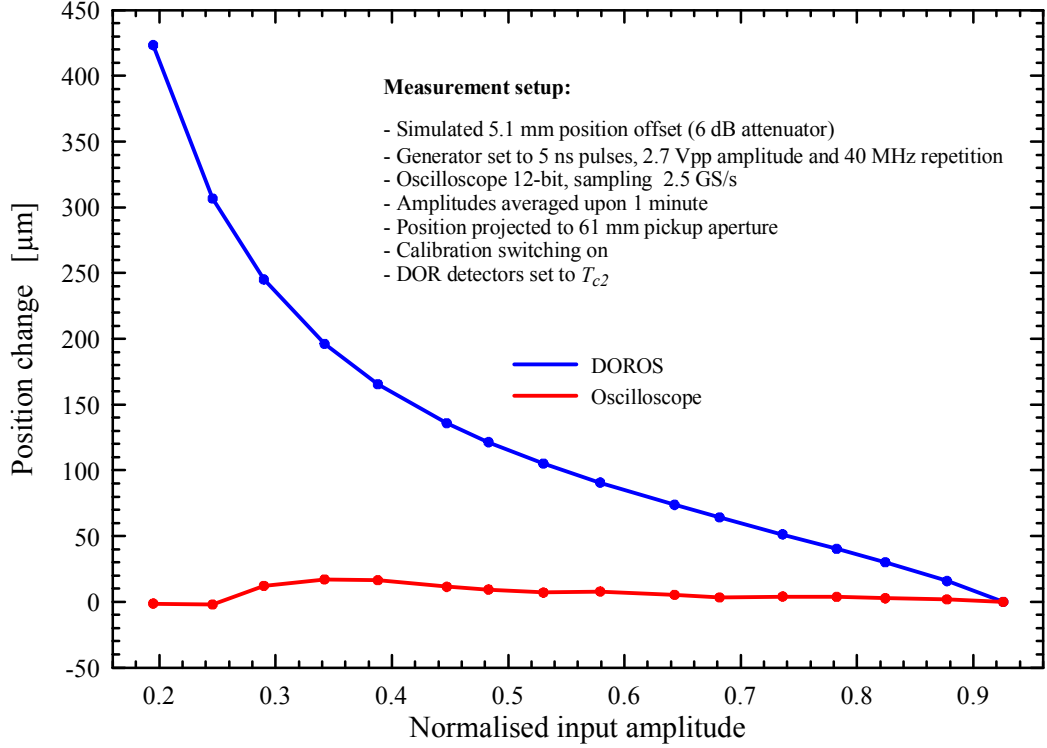


Fig. 5-16. Positions calculated upon generator signals measured in parallel with both, the 12-bit oscilloscope and the DOROS front-end. The horizontal axis was scaled to fit to the results from the channel 1 oscilloscope measurement. Displayed position change is relative to the last position value measured before reaching the DOROS ADC saturation.

As seen on the plot the positions calculated from the oscilloscope data vary by less than 20 μm . This is equivalent to a 4 ppm error on a 5.1 mm position offset. The corresponding measurement obtained at the same time with the DOROS front-end is some 420 μm which is equivalent to a position error of about 80 ppm. The presented measurements confirm that the linearity of the generator amplitude modulation ramp was much better than the linearity of the DOROS processing channels.

5.1.5 Summary and outlook

Measurement results characterizing the linearity of the compensated diode detectors of the DOROS orbit subsystem were presented in this chapter. The measurements were performed with a laboratory setup with a signal generator simulating BPM signals.

The results showed, that when the simulated beam is centered and the pulses from the electrodes have equal amplitudes, the influence of the signal amplitude changes on the measured positions is very small. In such conditions the diode detectors of each channel pair operate around very similar part of the detector characteristic, reducing the influence of the residual nonlinearity of the detectors. As a result the position errors are very small. The measurements yielded some 1 μm position change when the signal amplitude changed by more than 50 %. The periodic calibration switching helped to reduce these effects further by suppressing most of the asymmetry between the two channels processing signals from the BPM electrode pair. Such measurement conditions are typical for collimator BPMs where the position of the beams is typically well centered between the collimator jaws accommodating the embedded BPM electrodes.

Systematic errors of the beam position measurements become more important with increasing beam offsets. When the simulated beam had 5.1 mm offset, the resulting position change was more than 100 μm for the 50 % amplitude drop.

Such large beam offsets are typical for BPMs next to the LHC experiments. In such conditions the signals for the opposing BPM electrodes have much different amplitudes, causing the corresponding diode detectors to operate at quite distant part of their characteristics. Then the residual nonlinearities of the characteristics are converted into systematic position errors changing with the signal amplitudes driven by the beam intensity decay.

The presented systematic position errors for larger beam offsets become a limitation of the system, especially for precision measurements in the LHC interaction points where the offsets can reach some 8 mm. There are a few solutions that allow minimizing such systematic errors. The linearity errors can be minimized by automatically controlling the RF gain such, that the amplitude drop is compensated and the maximum electrode signals are held at constant levels. The front-end architecture allows to change gains in 1 dB steps using the programmable gain amplifiers. The automatic gain control can compensate the intensity drop each time the amplitude decreases by more than 12 %. This minimizes the amplitude change on the characteristics of the diode detectors and consequently minimizes the changes in position readings caused by the residual nonlinearities. The smallest position change estimated from the measured characteristics was achieved when controlling the signal levels in range 0.6 - 0.8 FS. The control can be achieved by setting the minimum threshold in this region and setting the hysteresis of the gain control to minimum. The hysteresis region is set to avoid gain oscillations in all operational scenarios. The residual position error was estimated from the measurements to be around 25 μm . This error is limited by the smallest allowable gain step in the programmable RF amplifier.

On the market there are a few RF programmable gain amplifiers which offer gain steps of 0.5 dB or even 0.25 dB. Their performance and potential application in the DOROS front-ends is under study. Another option would be to use a variable gain elements such as the voltage controlled gain amplifiers which can be controlled by a DAC to achieve even smaller steps. However, typical bandwidth of such amplifiers is below 100 MHz and as such is not sufficient for their use in the DOROS front-ends.

The measurement results shown, that the smallest systematic errors can be achieved when using the time constant T_{c0} with the slowest detector discharge. When the signals are operated around 0.8 FS, the T_{c0} can decrease the position errors by more than 20 % when compared to the results obtained with the T_{c2} . However, using only T_{c0} time constant is not optimal in all operational conditions. An example of such a case is a physics fill lasting typically several hours. During this time the diode detectors start to measure the orbit of so called non-colliding bunches, whose nominal intensity does not decay due to the collision losses. This effect was measured in the LHC. The results are provided in the chapter 6.2.4. The problem was solved by using the faster discharge constant T_{c2} to operate the detectors in average peak mode.

The author attempted to optimize the DOROS compensated detector circuit to improve its linearity. The research performed so far indicates that it is difficult to significantly improve the present detector performance, especially with time constant T_{c0} .

Potentially the linearity of the orbit measurement could be improved by electrical centering of the BPM electrode signals. This could be achieved by inserting attenuators with calibrated values to equalize the amplitudes of the beam signals on the front-end inputs. The observed artificial offset in the measured signals could be then translated into signal attenuation and removed after the digital processing. However, this approach requires the beam offset in the BPM to remain constant. Unfortunately this assumption is not true in practice.

Other approach towards improving the linearity of the orbit measurements is to linearize the detector characteristic using polynomials. This approach relies on very precise linearity measurements, much better than 1 %. Therefore such measurements could be performed only in laboratory conditions. This chapter presented the linearization polynomials obtained for the three detector time constants. These polynomials were applied to linearize the orbit measurements of the standard DOROS system with promising preliminary results. More details can be found in chapter 6.2.4.

5.2 DOR detector step response

Duration of the transient effects in the DOR processing is an important parameter for measurements which require low latency between the input signal change and its valid measurement result. The effects are mostly dominated by the programmable time constants in the diode detectors. This chapter is focused on measuring response of a single DOR channel to a full-scale amplitude step. As the response depends on the repetition rates of the input signal the measurements were setup to represent two typical cases used in the LHC. Measured transients were used to estimate the duration with different detector time constants.

5.2.1 Measurement setup

Diagram in Fig. 5-17 depicts the laboratory setup used to measure the response of one of the DOR channels to an amplitude step. Waveform generator Keysight 33600A provided 5 ns pulse signal split into 8 front-end inputs. After the non-reflective input filters the pulse width is stretched to about 6 ns. The pulse repetition rate was set to 12 kHz to simulate a single bunch in the LHC. The maximum amplitude was set to 3 Vpp. Beams with 25 ns bunch spacing were simulated with 40 MHz pulse repetition. The amplitude was increased to about 3.8 Vpp to compensate the drop of the pulse base line. The screenshot shown in Fig. 5-18 presents an example of the droop transient measured on the input of the compensated diode detector. The measurement was acquired with an 8-bit oscilloscope (Lecroy WaveSurfer 64x). The transient duration of the droop measured by the oscilloscope was estimated to about 0.5 ms, as marked on the plot with cursors.

In order to generate the amplitude step the pulses simulating bunches had their amplitude modulated by a square waveform with 100 % modulation depth. The low signal level corresponds to the minimum available voltage on the generator output. The high signal level corresponds to the 100 % of the amplitude modulation. The DOR chain was configured to have 20 dB RF gain in order to amplify the signal to about 95 % of the ADC full-scale. The step response was acquired using the capture mode. It allowed to store a few millions of ADC samples acquired in the local DDR memory at the f_{rev} rate with stored every 20th sample (equivalent to at 550 S/s data rate). The data was then read by the acquisition server for offline analysis.

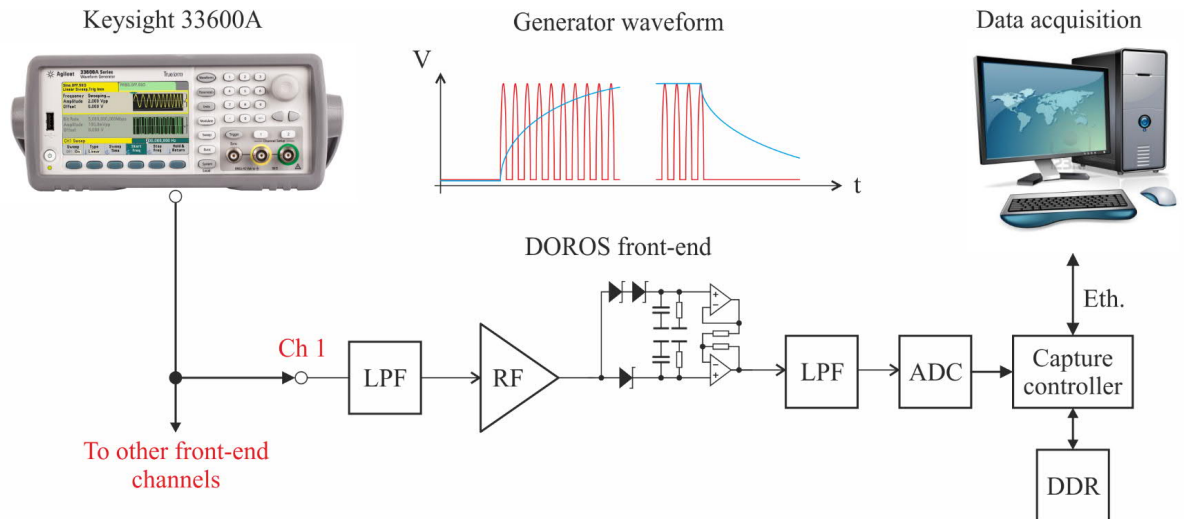


Fig. 5-17. Diagram of laboratory setup used to measure step response of DOR channel. The signal generator Keysight 33600A was used to provide pulse input signal with 5 ns width and 3 Vpp amplitude. The signal was switched “on” and “off” by amplitude modulation with 100 % depth. Pulse repetition rates of 12 kHz and 40 MHz were used to simulate bunch spacing typically used in LHC. **Abbreviations:** LPF – Low Pass Filter, RF – Radiofrequency, DDR – Dual Data Rate memory in the front-end.

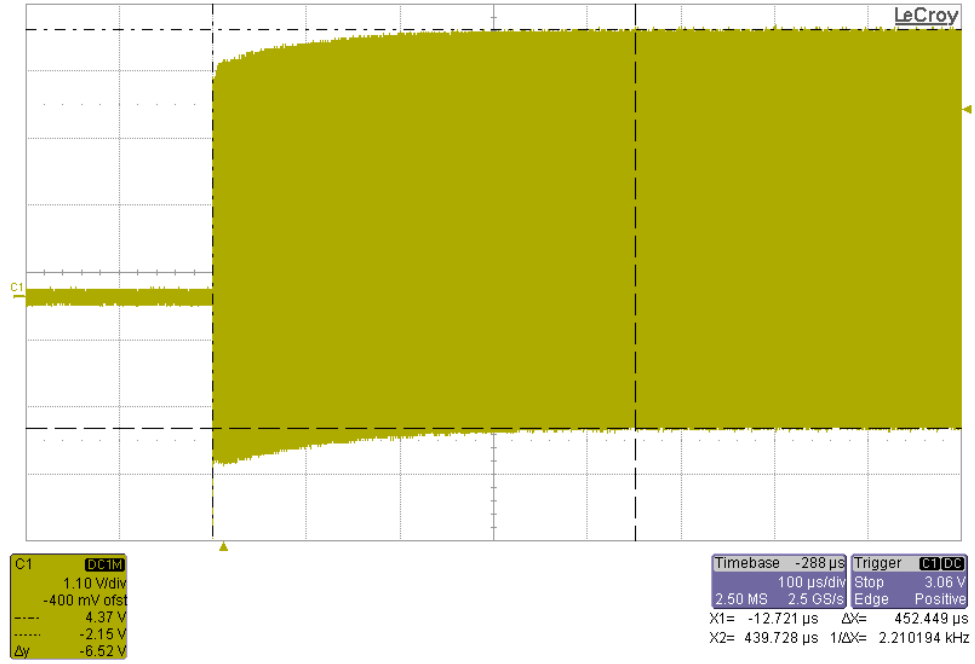


Fig. 5-18. Measurement of the transient effect acquired during switch “on” of the front-end input signal. The signal generator Keysight 33600A provided pulses with 5 ns width and 40 MHz repetition rate. The signal was switched “on” by amplitude modulation with 100 % depth. The effect was measured on the input of the compensated diode detector.

5.2.2 Step response of a DOR channel

The results of the first part of the measurements shown in Fig. 5-19 is focused on the response of the DOR channel to the positive step of the amplitude modulation. The signal step at 0.1 on the plot corresponds to enabling the generator pulses. This situation is similar to a first beam injection into an accelerator. Obtained characteristics were used to estimate the charge time of the orbit detector capacitors. As seen on the plot the transient response depends on the selected detector time constant as well as the pulse repetition rate.

With the 12 kHz repetition rate simulating one bunch circulating in the LHC, the signal level at the DOR output depends on the selected detector time constants. The amplitude for T_{c1} is smaller with respect to T_{c0} by 20 %. The effect becomes even larger for T_{c2} , where the amplitude decreases by more than 60 % with respect to T_{c0} . With T_{c1} and T_{c2} the capacitor discharge rate is too large for the very small duty cycle of the charging pulses. For only few bunches circulating in the LHC only time constant T_{c0} should be used operationally, otherwise large orbit errors will occur. Further details can be found in the following chapter.

The DOR signal levels with the 40 MHz pulse repetition rate, simulating the LHC physics beams, change with the time constants by less than 2 %. Therefore when the LHC operates with physics beams any of the time constants can be selected. The equivalent time constants of the charging transients are listed in Table 5-7. The constants were measured at the time required to reach 63 % of the steady state, as used to measure the time constant of an exponential rise.

Table 5-7. Summary of the DOR charging and discharging rate measurements.

Exponential decay rate	DOR charging (63.212 % of steady state amplitude)		DOR discharging (63.212 % of decay amplitude)	
	12 kHz	40 MHz	12 kHz	40 MHz
T_{c0} [ms] *	23.3	2.3	403.4	387.2
T_{c1} [ms] *	17.4	2.1	81.9	83.8
T_{c2} [ms] *	9.2	2.8	13.1	16

*Results rounded to 1 digit.

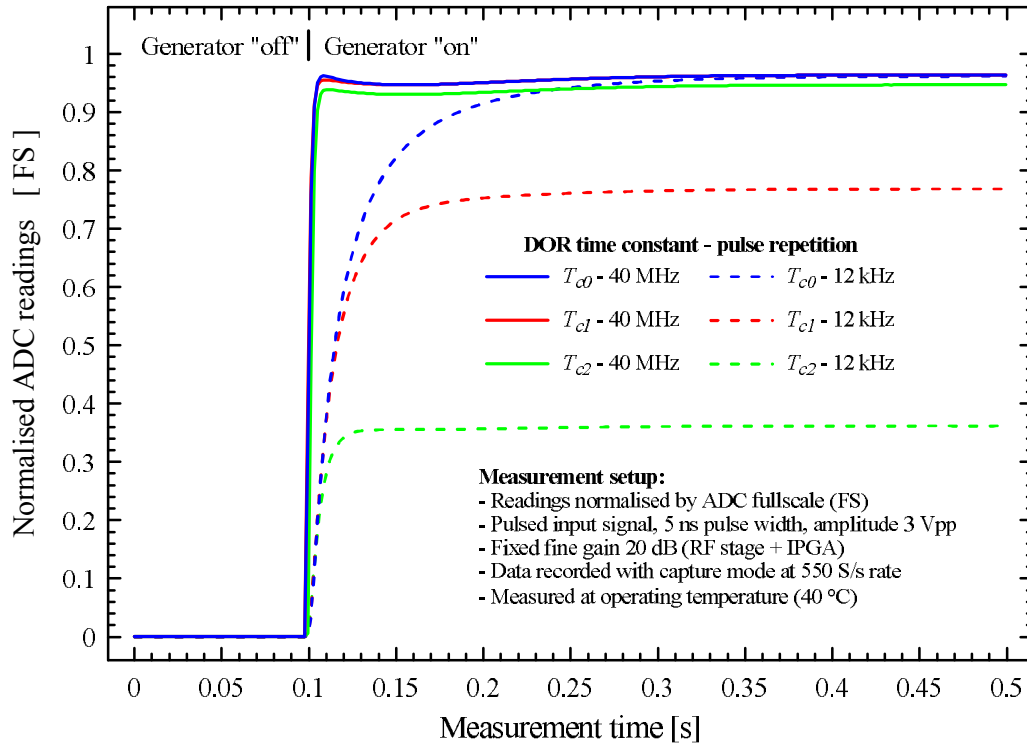


Fig. 5-19. Response of DOR channel to positive amplitude step of the input pulses. Signal generator Keysight 33600A provided 5 ns pulses with 3 Vpp and 12 kHz (measurements shown in dashed lines) or 40 MHz repetition rates (measurements showed in solid lines). The input signal was “switched on” using 100 % amplitude modulation in the generator. Data was recorded using capture mode acquiring the ADC data f_{rev} rate, storing every 20th sample.

The slowest rate of some 20 ms was measured with time constant T_{c0} for 12 kHz pulses. The fastest rate of some 9 ms was measured with time constant T_{c2} . The charging rate increases by more than factor 3 for T_{c2} and factor 10 for T_{c0} when the 40 MHz pulse repetition rate is used. The transient response has also a slower residual component, which can be seen as a low ripple on the steady state part of the signal. This is observed especially with the 40 MHz pulse repetition rate, resulting in faster charging. At this pulse repetition the impact of the selected time constant on the duration of this residual slow transient becomes less significant. The observed slow transients are caused by the asymmetry of the charging time for the single-diode and double-diode detectors in the compensated detector.

The second measurement shown in Fig. 5-19 is done one on the falling slope of the detector signal, which corresponds to switching off the generator pulses at 0.1. Obtained characteristics were used to estimate the discharge times of the detector capacitors, which are listed in Table 5-7. The detector discharge depends on the selected time constant. The response duration was estimated with an exponential decay rate. This corresponds to the time required for the signal to decrease from the initial steady state by approximately 63 %. Please note that the discharge of the detector capacitors in the T_{c0} configuration is dominated by reverse currents of the diodes. The measurement results show the worst case discharge time of about 0.4 s. The measurement spread between the 8 channels was about 10 %. For time constant T_{c0} the discharge of the detector capacitors is done only with the reverse current of the detector Schottky RF diodes. The presented measurement can be used to estimate the discharge current. This current discharging 10 nF capacitors was estimated to about 70 nA. The corresponding discharge resistance assuming a simple RC network model was estimated to about 60 MΩ at 90 % of the ADC full-scale (equivalent to input voltage of 4.5 V). For comparison, in T_{c1} and T_{c2} time constants the discharge resistors are 10 MΩ and 1 MΩ, respectively. For T_{c1} time constant the discharge rate was estimated to about 80 ms. The T_{c2} provided rate of less than 16 ms. Please note that the response duration becomes longer once the signal level is below 5 % of the ADC full-scale (equivalent to input voltage of 0.25 V), however, such low amplitudes are not used in operational conditions.

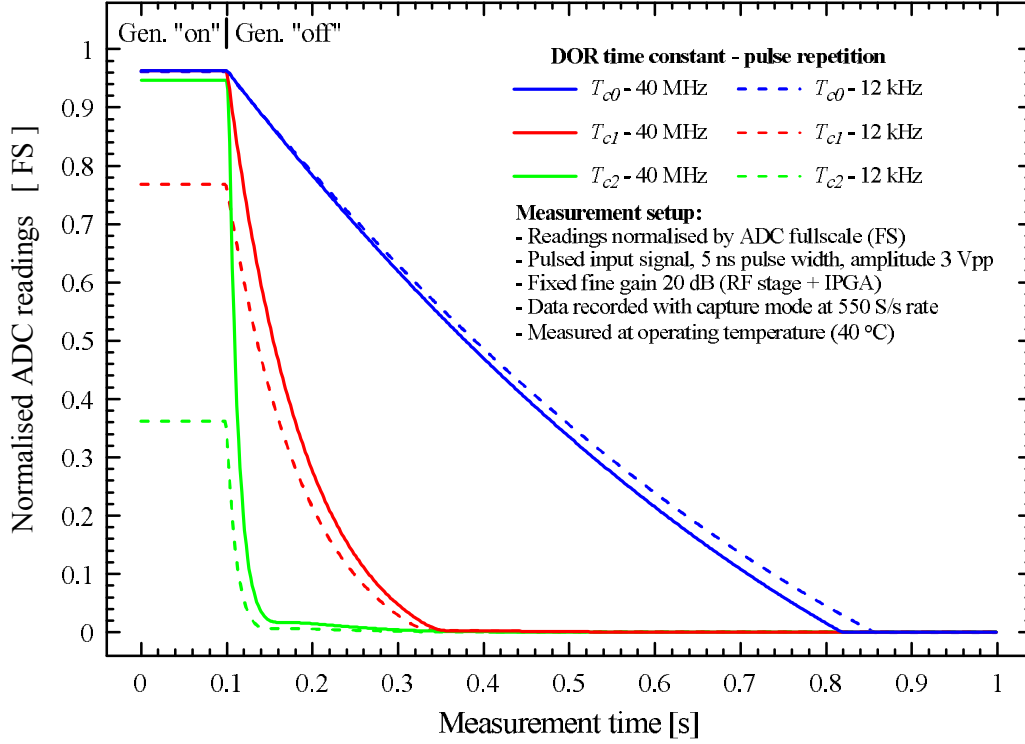


Fig. 5-20. Response of DOR channel to negative amplitude step of the input pulses. Signal generator Keysight 33600A provided 5 ns pulses with 12 kHz and 40 MHz repetition rates. The input signal was “switched off” using 100 % amplitude modulation in the generator. Data was recorded using capture mode acquiring the ADC data f_{rev} rate, storing every 20th sample.

5.2.3 Summary and outlook

This chapter presented measurements of the transient response of the DOR channels to a full-scale amplitude step on the inputs. The measurements were focused on simulating two extreme operational cases, namely the LHC with a single circulating bunch and full machine physics beam.

Fast transient response is important for feedback systems requiring small latencies, such as the LHC orbit feedback system [14]. Orbit calibration used in the DOROS system requires periodic switching the beam input signals with the input multiplexers and measure the signal levels during the steady states after the switching transients are already gone. The switching rate is then limited by the switching transients. For more details on the used strategies for the calibration switching please refer to chapter 4.3.4.4. Transient response of the orbit processing depends on the repetition rate of the input pulses and the programmable time constants of the detectors. Analysis of the transients was performed separately for the rising and falling slope of the output signal, caused by respectively enabling and disabling the input pulse signals. The measurements were used to estimate and analyse the transient times. The slowest measured transient was observed with T_{c0} detector time constant and simulated single bunch operation. It was the case for both, rising and falling slopes of the output signal.

The measurements demonstrated that the steady state signal level at low pulse rate depends strongly on the selected detector discharge resistor. The detector output signal can decrease by more than 60 % when using the fastest discharge T_{c2} with single bunches. Such settings must be avoided in operational use when the bunch repetition rate becomes comparable to the detector discharge time. However, such conditions can be used for testing the channel symmetry of the newly produced hardware or for operational diagnostics. When the orbit channels are tested with simulated single bunches it is easy to discover any anomalies in the detector time constants, as even small asymmetries translate into different voltages at the detector outputs.

5.3 DOR frequency response

The capture mode in the DOROS front-ends allows acquiring DOR signals synchronously to the LHC revolution frequency. The spectral content of the acquired beam orbit signals can be observed with the bandwidth limited by the analogue filtering of the DOR processing. The dynamics of the DORS response can be configured by selecting the time constant of the orbit detectors. This chapter is focused on measuring the amplitude characteristics and bandwidth at different stages of the DOR processing.

5.3.1 Measurement setup

The amplitude characteristics were measured with two laboratory configurations as depicted in diagrams of Fig. 5-21 and Fig. 5-22. The beam signals in both setups were simulated using a 16-bit arbitrary waveform generator (Keysight 33522A) connected to one front-end input. The generator produced sinewave signal with the frequency of 20 MHz simulating a BPM electrode signal with 50 ns LHC bunch spacing. The 1.5 Vpp signal was harmonically modulated in amplitude with depth of 1 % (15 mVpp modulation amplitude). Such modulation is equivalent to a position change of approximately 150 μm assuming BPMs with 61 mm aperture (Q1 LHC BPMs).

For the discussed measurements the configurable RF amplifier stages in front of the compensated diode detector were bypassed. The setup shown on diagram in Fig. 5-21 allowed measuring signals directly at the output of the compensated diode detectors. The response of the orbit analogue chain consisting of the compensated diode detectors followed by the analogue low-pass filter was measured at the output of the filter. The filter limits the signal bandwidth before the ADC to some 100 Hz. The amplitude characteristics were evaluated in the range from 1 Hz to 10 kHz by setting 16 frequency values for the modulation signal and measuring corresponding amplitudes with an oscilloscope.

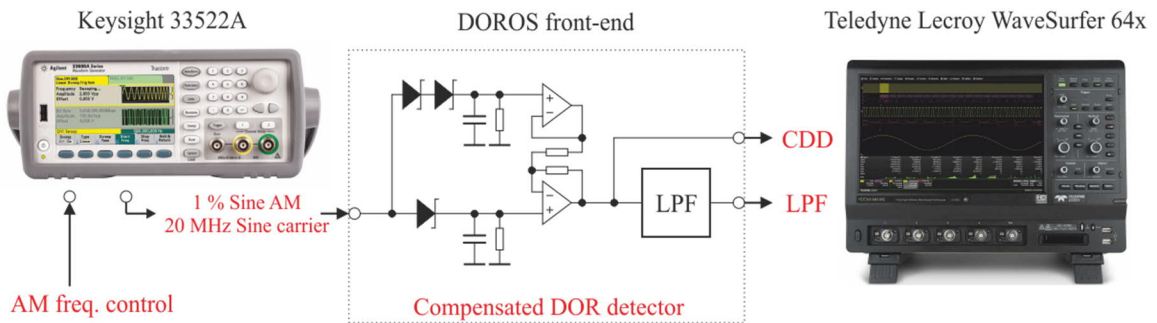


Fig. 5-21. Diagram of the laboratory setup used to measure frequency response of compensated diode detector (CDD). Frequency response of the DOR analogue processing was measured at the output of the low-pass filter (LPF).

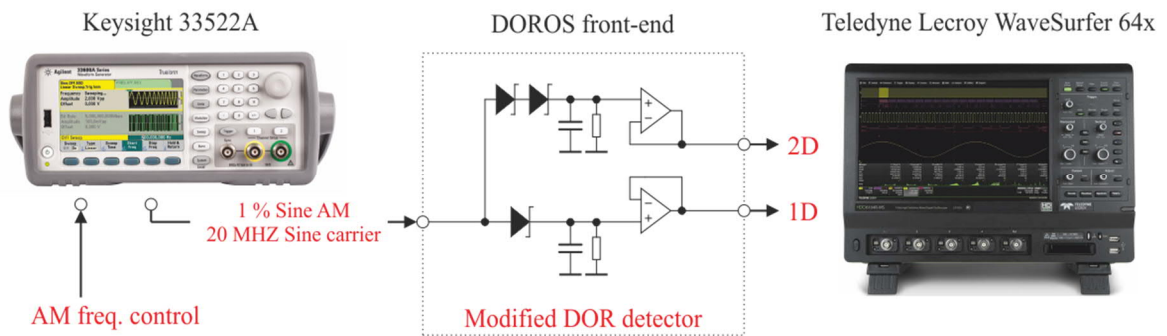


Fig. 5-22. Diagram of the laboratory setup used to measure frequency response of the 1 diode (1D) and 2 diode (2D) branches. The operational amplifiers of the compensated diode detector were modified to buffer the outputs of each detector branch.

The frequency response on the CDD output is a result of the combined responses of two diode sub-detector branches. One of the sub-detectors consists of a single diode (1D) and the second of two diodes (2D) connected in series. Their outputs are processed by two operational amplifiers in a configuration allowing the compensation of the voltage drop on the diodes. The setup shown on diagram in Fig. 5-22 was used to measure frequency response of each CDD branch separately. The JFET amplifiers in the CDD were modified to a follower configuration in order to split and buffer the sub-detector outputs. The output of each follower was then connected to input of the oscilloscope (Lecroy Wave Surfer 64x).

5.3.2 *Frequency response of a DOR channel*

The dynamic properties of the orbit compensated diode detector (CDD) are defined by their three configurable time constants. The time constants define the discharge rate of the detector capacitors. Consequently, the detector time constants together with the following active low-pass filter (LPF) define the amplitude characteristics and the bandwidth of the orbit measurements.

The plot in Fig. 5-23 displays the frequency response of the orbit channel, consisting of the CDD and the LPF (solid traces). The plot also shows the frequency response measured at the CDD output (dashed traces) without the LPF. Results of the amplitude measurements were normalised to the value measured at the lowest modulation frequency of 1 Hz. Please note that the CDD response measurement with the slowest time constant T_{c0} has a 2 dB peak at about 10 Hz. The roll-off starts at about 20 Hz. The response is similar to the characteristic measured at the LPF output. The CDD response obtained with time constant T_{c1} has 1 dB peaking at about 40 Hz with roll-off starting at about 100 Hz. In this case the low-pass filter starts contributing to the frequency response. The effect can be observed as the decreased amplitude peaking and lowered the cut-off frequency. The measurement was repeated with the fastest time constant T_{c2} , resulting in the CDD cut-off frequency of about 500 Hz. In these conditions the amplitude peaking was not measurable. The output characteristic is dominated by the low-pass filter, whose cut-off frequency is below that of the CDD.

The following measurements shown in Fig. 5-24 were focused on the amplitude characteristics of the two detectors of the CDD scheme. The measurements were repeated for the three detector time constants. The displayed amplitudes were normalised to the values measured at 1 Hz. Please note that both detectors were built with the same passive components and they differ only by the number of the diodes accommodated in a single package.

The lowest cut-off frequency of about 20 Hz was measured on the output of the 1D detector with the T_{c0} time constant. The cut-off frequency of the 2D detector was measured to be about 10 Hz. Please note that the discharge rate with this time constant is dominated by the leakage currents of the detector diodes. It can be observed that the difference between the 1D and 2D cut-off frequencies is decreasing when using faster time constants T_{c1} and T_{c2} , when the discharge is defined by the resistors.

The following figures Fig. 5-25, Fig. 5-26 and Fig. 5-27 display the frequency responses of the 1D and 2D sub-circuits in comparison to the characteristics obtained for the CDD and the following low-pass filter. The amplitudes at different frequencies are displayed as measured by the oscilloscope. It can be seen in the plots that the observed peak in the CDD frequency response depends on the difference between the cut-offs of the two particular detectors.

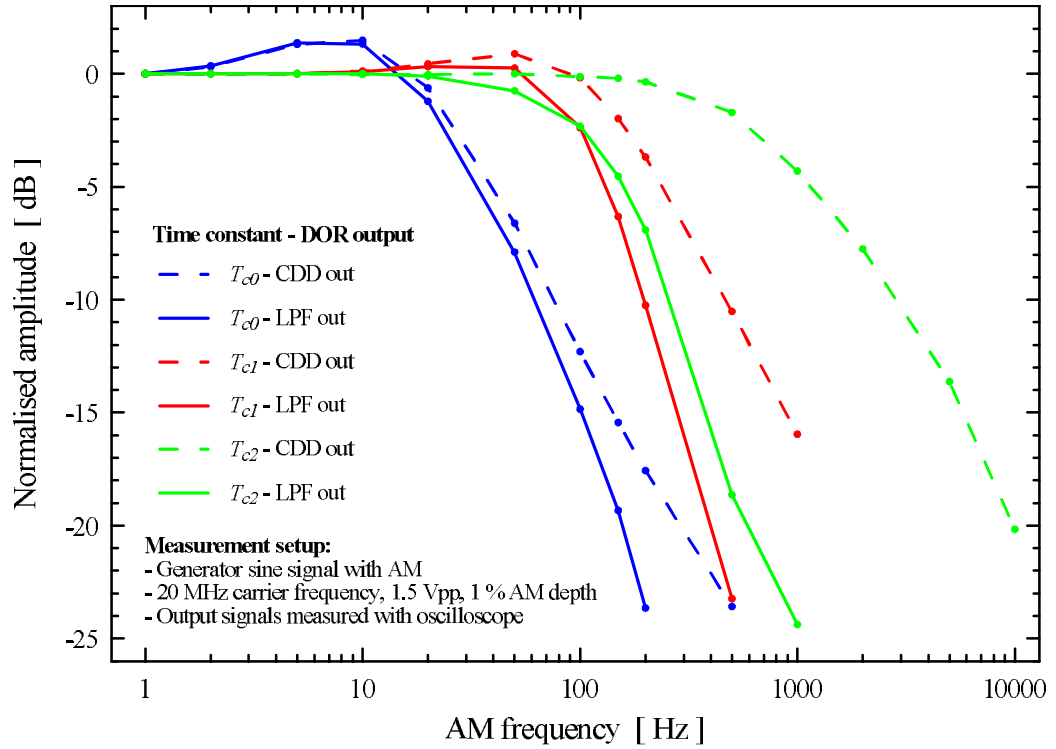


Fig. 5-23. Measured frequency response of the Compensated Diode Detector (CDD) and low-pass filter (LPF) after the CDD. The cut-off frequency of the LPF is 100 Hz. The generator Keysight 33600A provided 20 MHz sinewave signal with sine amplitude modulation (AM) waveform. The AM frequency was controlled in steps. Amplitudes of signals were measured with oscilloscope (Lecroy Wave Surfer 64x) for each frequency. Measurements were repeated for the three detector time constants. Displayed results were normalised to the value measured at 1 Hz.

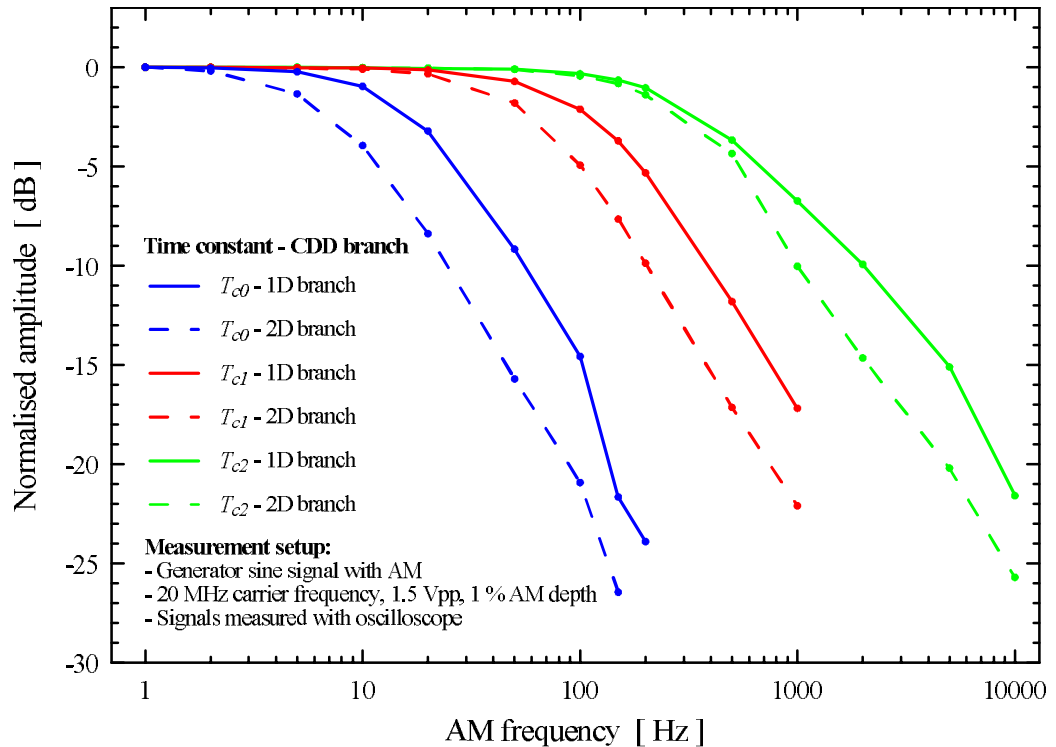


Fig. 5-24. Frequency response of the individual sub-detectors in the Compensated Diode Detector (CDD) circuit. The 1D branch in the CDD is the sub-detector built with one diode and the 2D branch is built with two diodes in series. The generator Keysight 33600A provided 20 MHz sinewave signal with sine amplitude modulation (AM) waveform. The AM frequency was controlled in steps. Amplitudes of signals were measured with oscilloscope (Lecroy Wave Surfer 64x) for each frequency. Measurements were repeated for the three detector time constants. Displayed results were normalised to the value measured at 1 Hz.

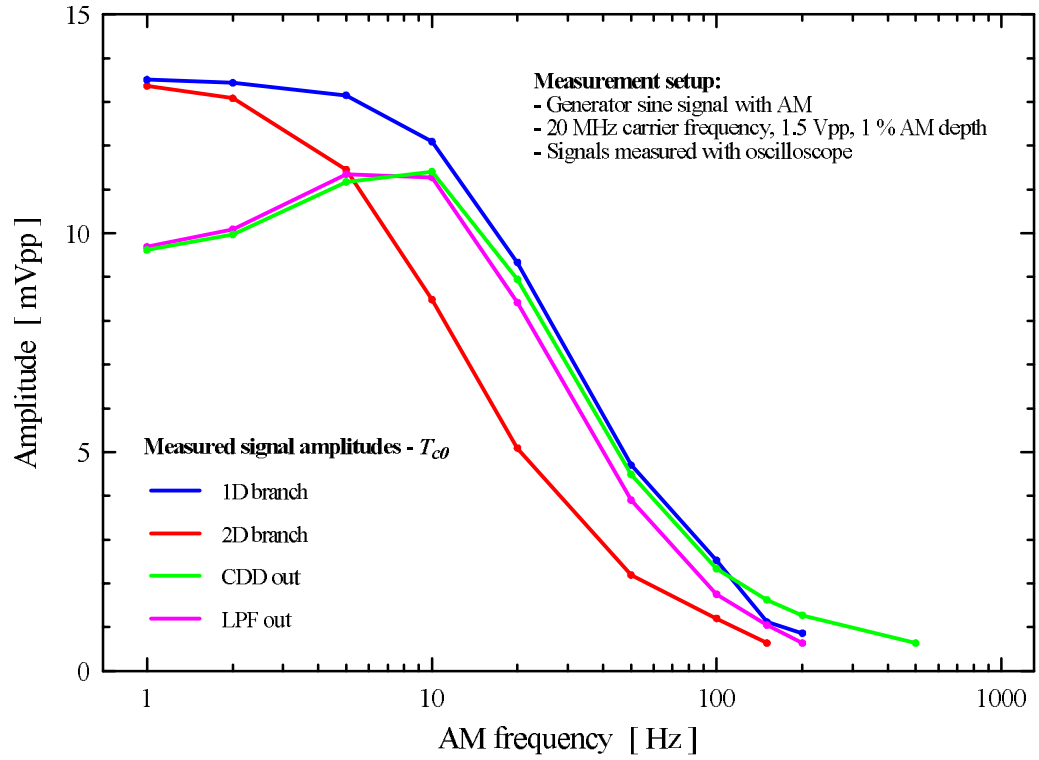


Fig. 5-25. Frequency characteristics of the Compensated Diode Detector (CDD), its sub-detector branches (1D with single diode and 2D with two diodes in series) and output of the analogue processing chain after the low-pass filter (LPF). The measurement results were obtained for the time constant T_{c0} with the slowest discharge rate.

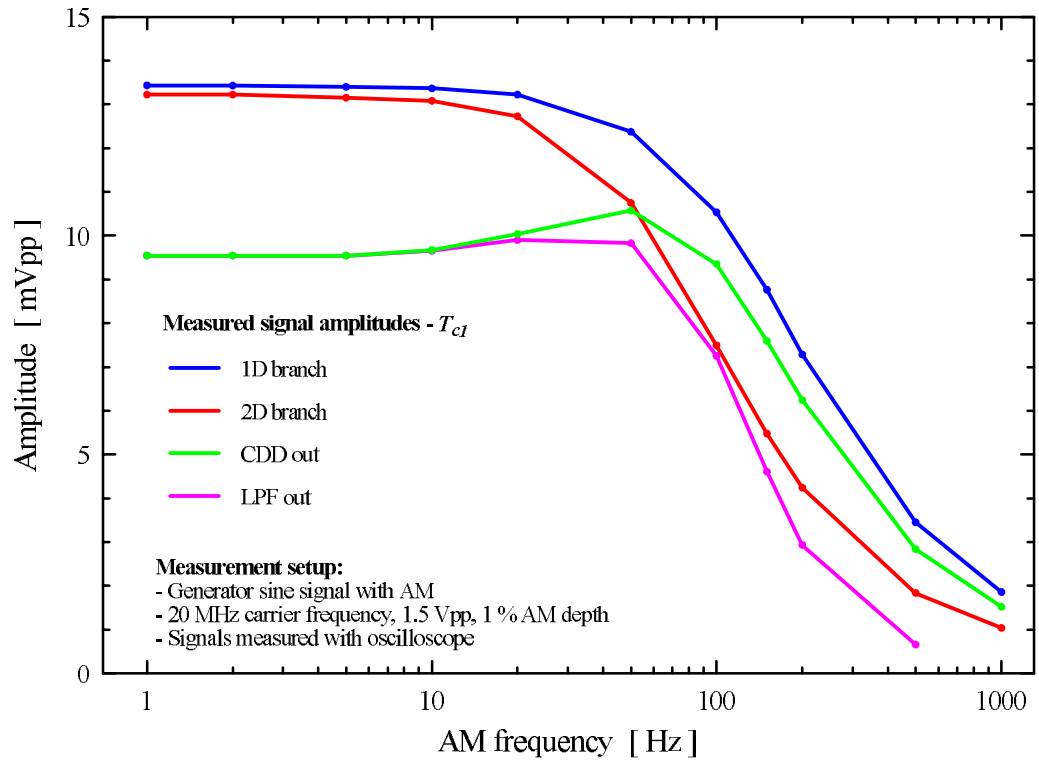


Fig. 5-26. Frequency characteristics of the Compensated Diode Detector (CDD), its sub-detector branches (1D with single diode and 2D with two diodes in series) and output of the analogue processing chain after the low-pass filter (LPF). The measurement results were obtained for the time constant T_{c1} with the discharge rate set by 10 M Ω resistor.

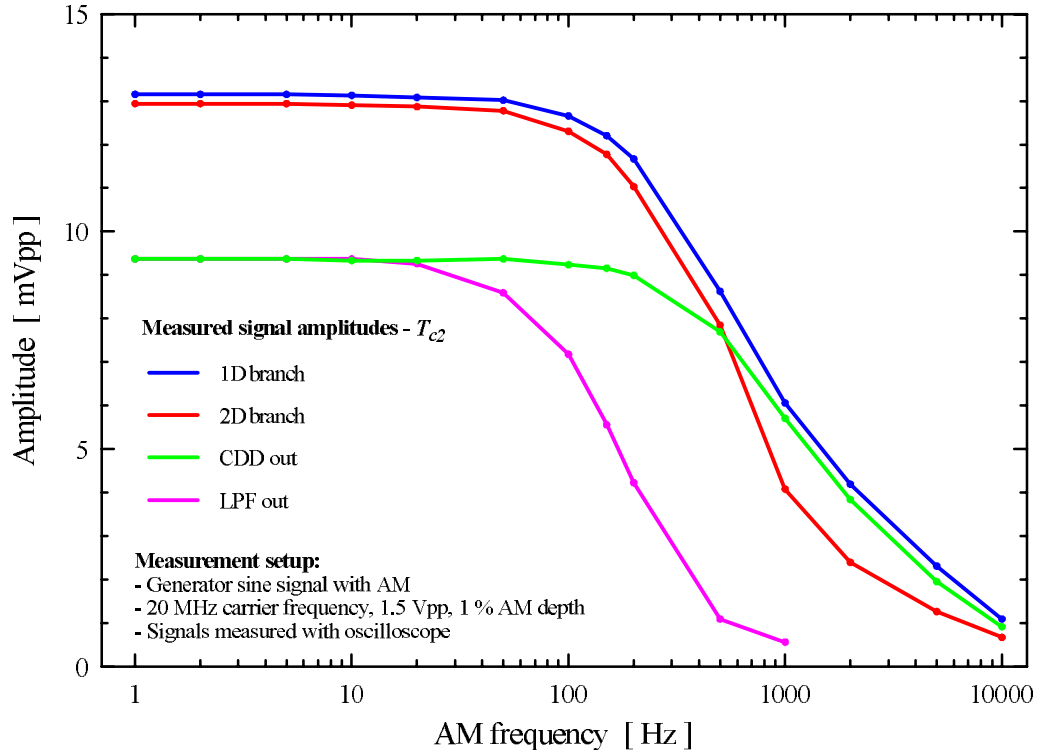


Fig. 5-27. Frequency characteristics of the Compensated Diode Detector (CDD), its sub-detector branches (1D with single diode and 2D with two diodes in series) and output of the analogue processing chain after the low-pass filter (LPF). The measurement results were obtained for the time constant T_{c2} with the fastest discharge rate set by 1 M Ω resistor.

5.3.3 Summary and outlook

The capture mode allows measurements with bandwidth ranging from DC to some 5 kHz using both, the DOR and DOS channels. The actual cut-off frequency depends on the selected time constant. The bandwidth of the DOR chain is limited by a low-pass analogue filter to 100 Hz.

Time constant T_{c0} provides the smallest measurement bandwidth up to 10 Hz. Due to its slow discharge rate this time constant is typically used for input pulses with low repetition rates. The amplitude characteristic, dominated by the CDD response, showed 2 dB peaking around the cut-off frequency. Further investigations showed that the peaking originates from differences between the amplitude characteristics of the CDD 1D and 2D detectors. The measured cut-off of the branch with a single diode detector was 20 Hz. The cut-off frequency of the branch with two diode detector was 10 Hz.

Time constant T_{c2} provides the largest CDD bandwidth, up to some 500 Hz. With this time constant the bandwidth difference between the two CDD detectors is small, resulting in flat frequency response up to the cut-off. Time constant T_{c2} is used for beams with higher repetition rates. The amplitude characteristic measured at the DOR output is dominated by the output low-pass filter with 100 Hz cut-off. The T_{c2} setting can be used for orbit measurements requiring observation with larger bandwidths like (e.g. orbit oscillations).

5.4 Automatic calibration switching

Accuracy of the orbit measurements delivered by the DOROS system depends on the symmetry of the analogue channels processing signals from the opposing BPM electrodes. Any gain or offset asymmetry between the channels results in a systematic error of the measured orbit. The asymmetry can be strongly reduced by using the automatic calibration technique. This chapter is dedicated to present and discuss measurements on the performance of the orbit processing operating in the calibration mode. The performance is characterized as a function of a few system parameters, like the RF gain or the handling of the transients introduced by the automatic calibration.

5.4.1 Switching transients

The automatic calibration technique is based on periodic multiplexing of the BPM signals between the two processing DOR channels. The multiplexer allows passing the filtered BPM signals directly to their channel pair (so called straight configuration) or to swap their order (so called crossed configuration). As the signals of either BPM electrode can have quite different amplitudes, the switching process typically causes important transients in the acquired data.

These effects define an upper limit to the multiplexing frequency and latencies in the orbit measurement. The transient duration depends on a few factors, like the detector time constant or the amplitude differences of the switched signals. The bandwidth of the orbit measurements can be changed by choosing one of the three configurable detector time constants, controlled by the system FPGA. The following measurements provide an overview of the DOR detector response dynamics during the calibration switching process. In order to reduce the transients caused by the calibration switching, the author developed a special technique, so called speed-up, described in detail in chapter 5.4.1.3.

5.4.1.1 Measurement setup

The used laboratory setup is depicted in Fig. 5-28. Signal generator Keysight 33600A provided pulse signals with 5 ns width and 2.5 Vpp amplitude. The measurements were done with the pulse repetition rates of 12 kHz and 40 MHz, simulating respectively a single LHC bunch and the full machine. The signal was split to four front-end inputs with a passive splitter. A 6 dB attenuator was used on one of the inputs to simulate a beam offset of 5.1 mm (when referenced to a 61 mm BPM aperture). Such offsets are typical for the LHC BPMs located close to the experiments.

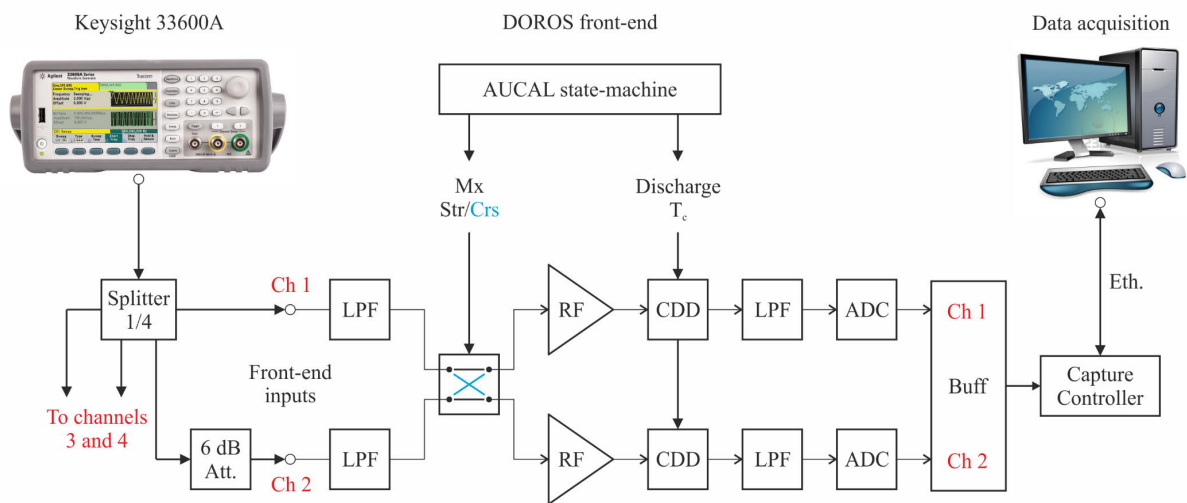


Fig. 5-28. Diagram of the laboratory setup used to measure the transient effects of the DOR chain during calibration switching. **Abbreviations:** AUCAL – Automatic Calibration, LPF – Low Pass Filter, CDD – Compensated Diode Detector, RF – Radio Frequency.

The front-end RF stages were configured with 10 dB gain, amplifying the signals to the amplitude of 0.6 FS. Such amplitudes are representative for DOROS front-ends operating with LHC beams. The switching frequency of the auto-calibration multiplexer was set to 1 Hz. Such setting results in calibrated orbit readings at a 1 Hz rate with 1 s latency. The measurement data was acquired using the front-end capture mode. In this mode the ADC samples are decimated by the factor of 20, resulting in the effective sampling rate of some 550 S/s. The decimated samples were stored in the local DDR memory without any additional digital filtering. The acquisition was controlled by a capture trigger sent over a TCP command.

5.4.1.2 Transient duration

The described setup was used to perform laboratory measurements with multiplexer switching. Their results are shown in Fig. 5-29. The switching allows reducing the asymmetry in the analogue processing of the orbit signals. The plot displays the channel 1 output signal measured with 12 kHz and 40 MHz pulses, simulating LHC beam signals, with three detector time constants. The parameters were chosen to show switching transients in simulated conditions representative for different beam operation scenarios. The two consecutive transients seen in the plot were caused by the change of the multiplexer state. The falling edge of the first transient corresponds to the multiplexer configuration change from the straight to the crossed mode. Please note that the sign correction de-multiplexer was not used in this measurement setup. In the crossed state of the multiplexer the input of channel 1 measures the generator signal passing through the attenuator. The rising edge of the second transient corresponds to the multiplexer switching back to its previous state. Please note that the transients would become small in case of input signals with similar amplitudes, when the beam is close to the BPM centre. Such scenario is typical for orbit measurements in the collimators, where the jaws are placed symmetrically around the beam. The transient durations summarized in Table 5-8 were estimated by computing time required to reach the steady state within 0.01 % of the ADC full-scale.

Obtained results for transients caused by both multiplexer state changes are shown in Fig. 5-30. The sign correction, which would be done by the de-multiplexer in the DOR DSP, was in this case done by computing absolute values. The absolute positions were scaled in millimetres using equation (3-2) referenced to a 61 mm BPM aperture. The plot shows the influence of the switching transients in the DOR channels on the position measurement readings. The difference between the rising and falling transients seen in the plot was caused by different charging and discharging rates of the DOR detectors. Especially in case of time constant T_{c0} the falling transient can be slower by more than factor 2 in comparison to the corresponding rising transient. In addition the measured response depends on the detector time constant. The results show that the duration of both transients can be decreased by an order of magnitude when using T_{c2} . In order to increase the multiplexing frequency the demonstrated effect is exploited in the speed-up technique to cope with the slow response of the detector. The measurement results show that the detector response becomes slower for low pulse repetition rates such as the 12 kHz case, simulating the LHC operating with a single bunch. Please note that using the fast discharge constant T_{c2} causes large orbit error with low repetition rates. Under such circumstances this setting must not be used during beam operation. As seen in the plot the difference between the straight and crossed orbit measurements exceeds 10 %.

Table 5-8. Estimated transient duration on DOR channel 1 during 1Hz calibration switching.

Input pulse repetition	Multiplexer transition Crossed - to - Straight (transient rising edge)		Multiplexer transition Straight - to - Crossed (transient falling edge)	
	12 kHz	40 MHz	12 kHz	40 MHz
T_{c0} [s] *	0.20	0.15	0.40	0.35
T_{c1} [s] *	0.15	0.05	0.25	0.10
T_{c2} [s] *	0.05	0.02	0.05	0.02

*Results rounded to 2 digits.

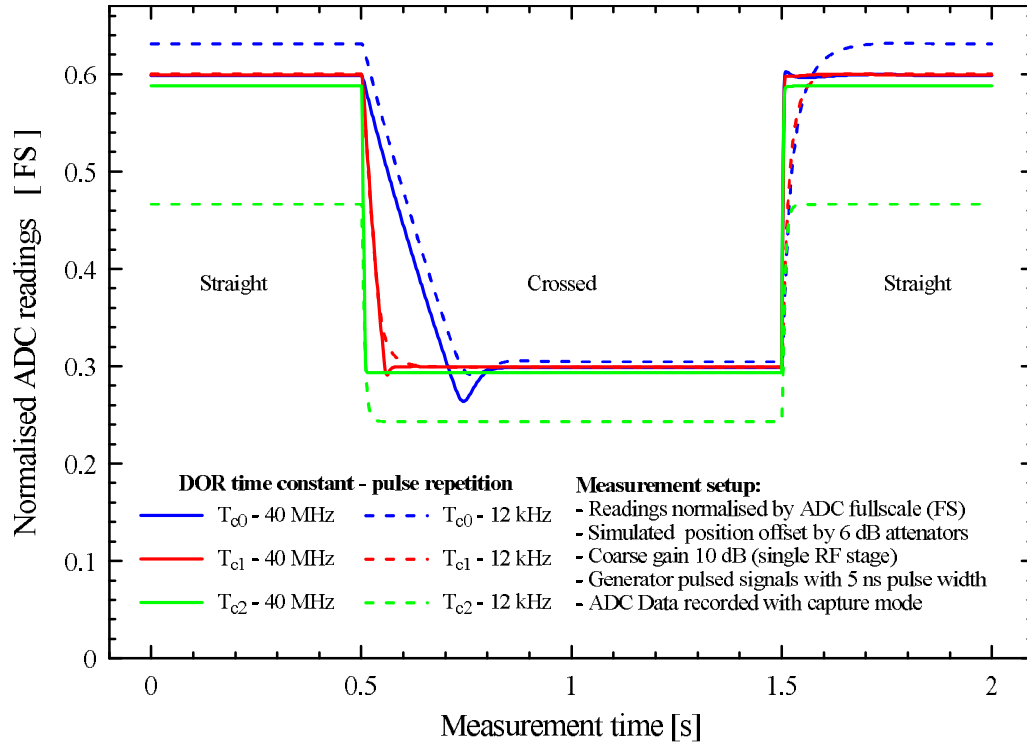


Fig. 5-29. Response of DOR channel 1 to calibration switching. The input multiplexer switches the input signals between the straight and crossed input configurations. Signal generator Keysight 33600A provided 5 ns pulses with 2.5 Vpp amplitude and 12 kHz (measurements shown in dashed lines) or 40 MHz repetition rates (measurements showed in solid lines). During the “state” channel 1 is connected to the channel 2 input with a 6 dB attenuator causing the change in the signal level. The measurements were repeated for the three detector time constants. The DOR channels were multiplexed at 1 Hz rate. Data was recorded using capture mode acquiring the ADC data f_{rev} rate, storing every 20th sample.

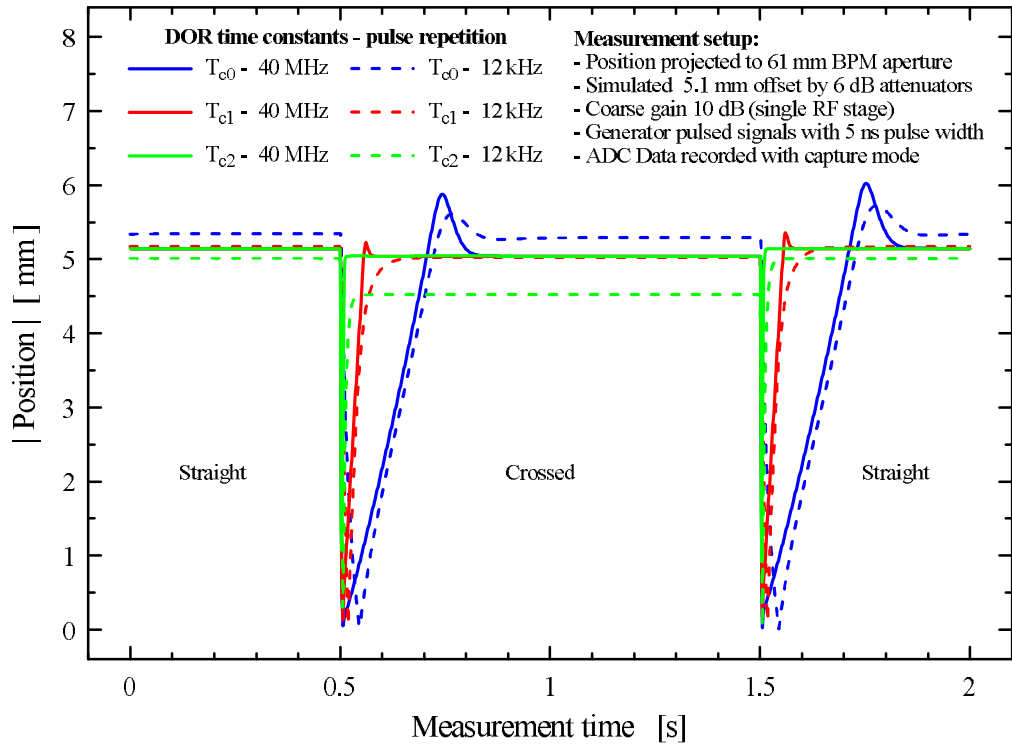


Fig. 5-30. Position measurement during the calibration switching. The absolute value of the position signal was computed using the DOR channel signals showed in Fig. 5-29. The positions were referenced to a 61 mm BPM aperture. The transients correspond to the input multiplexer switching. The position offset of around 5.1 mm was simulated by a 6 dB attenuator connected on one of the front-end inputs.

5.4.1.3 Detector speed-up technique

The slowest detector response was measured with time constant T_{c0} . In this configuration the detector capacitors are discharged at the lowest rate. The capacitor discharge is defined mostly by the reverse currents of the detector RF Schottky diodes, typically in the order of 10 nA. The time constant T_{c0} is therefore suitable for operation with low pulse rates such as single bunches in the LHC. The DOR response without using the speed-up technique is depicted in Fig. 5-31. The measurement was obtained with the setup described earlier in chapter 5.4.1.1. The plot displays signals from the two DOR channels during one multiplexer transition. The calibration multiplexer was switching from straight to crossed configuration at 0.1 s. The repetition rate of the generator pulses was 12 kHz. Difference in the signal levels seen in the plot was caused by the 6 dB attenuator on the input of channel 2. The DOR signals were then used to compute the position signal showed in Fig. 5-32. The difference between the positions measured during the steady states is proportional to the asymmetry between the respective DOR channels. The transient in the position signal reached the steady state within 0.01 % of the ADC full-scale in around 0.4 s. Such transient duration would limit the switching frequencies to less than 2.5 Hz. In addition, long transient masking would limit the number of the steady-state samples that can be taken into account for the position calculation. Precision of the position measurements with different transient masking lengths was measured in the following chapters.

The speed-up technique was developed to decrease the transient duration especially when using T_{c0} . It is based on pre-discharging of the detector capacitors with the fast time constants T_{c1} or T_{c2} applied for a short period of time after each switching. It employs the fact that for T_{c0} the detector charging is much faster than its discharging. Parameters of the speed-up such as the pre-discharge duration and time constants are controlled by the AUCAL state-machine. The plots in Fig. 5-33 and Fig. 5-35 show the DOR signal transients with the speed-up technique using the pre-discharging with T_{c1} and T_{c2} , when measuring with detector time constant T_{c0} . The corresponding position signals are displayed in Fig. 5-34 and Fig. 5-36. As seen in Fig. 5-33 the DOR signals with T_{c0} setting were pre-discharged at 0.1 s with T_{c1} for duration of 40 ms. The falling transient seen in channel 2 was estimated around 0.2 s. It can be seen that the pre-discharge duration was not long enough for the channel 2 to reach the steady state. This difference in the signal levels had to be further discharged by the slow T_{c0} . The rising transient seen in channel 1 was increased by the speed-up technique to about 0.25 s. The resulting transient duration in the position signal showed in Fig. 5-34 was thus estimated around 0.25 s. The speed-up technique with T_{c1} allowed increasing the calibration switching frequency to about 3.5 Hz providing 10 % margin for the measurement in the steady-state.

The DOR signals with T_{c2} pre-discharging are displayed in Fig. 5-36. The duration of both the falling and rising transients was estimated around 0.15 s. As seen in the plot, the signal level after the 40 ms pre-discharging is below the steady state measured with T_{c0} . This residual difference between the signal levels has to be processed with the slower T_{c0} which adds onto the transient duration. However, the detector charging with the T_{c0} is faster than its discharging. The resulting transient duration in the position signal showed in Fig. 5-36 was measured around 0.15 s. The speed-up technique with T_{c2} allowed increasing the calibration switching frequency to about 5 Hz providing 10 % margin for the measurement in the steady-state.

The plot in Fig. 5-37 shows the change in the channel 1 DOR signals during two switching transients. The first transient in 0.5 s corresponds to the detector charging and the second transient in 1.5 s corresponds to the detector discharging. The measurement displayed in blue was obtained with time constant T_{c0} without speed-up. The measurement with the speed-up technique using the fast time constants T_{c1} (red) and T_{c2} (green) were plotted for comparison. The acquired signals were referenced to the beginning of the measurement. The corresponding change in the position signal is showed in Fig. 5-38. As seen in the plot the shortest transient duration was achieved with the T_{c2} pre-discharge. Please note that the transient duration would decrease further for the input pulses with higher repetition rates.

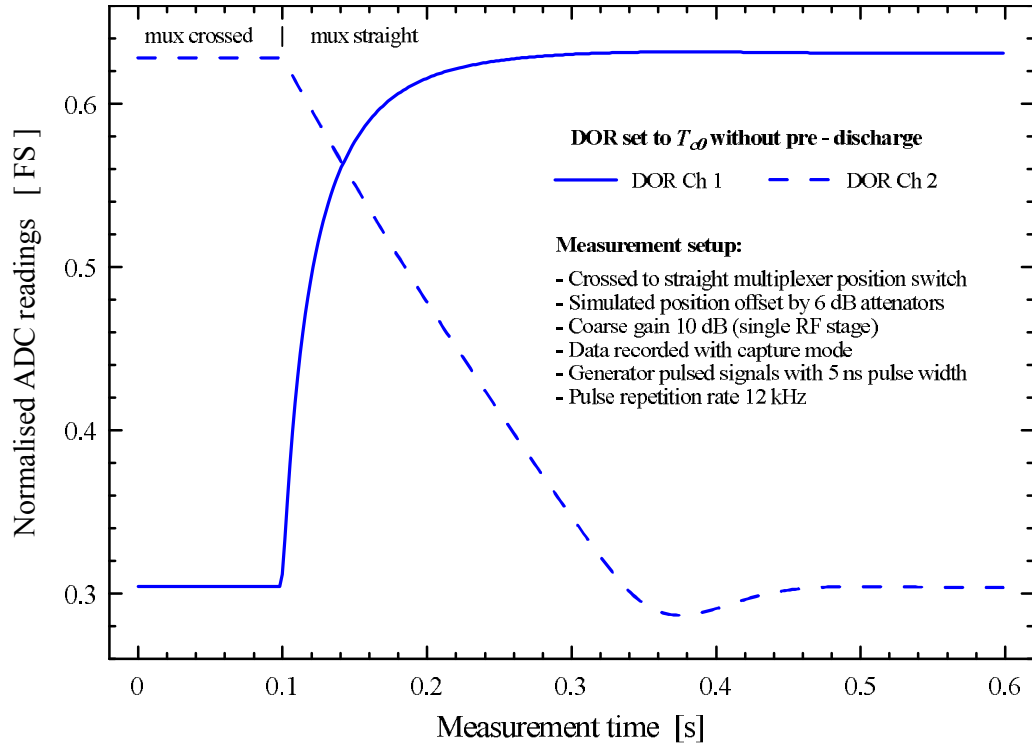


Fig. 5-31. Response of the DOR channel 1 and 2 signals to the calibration switching the time constant T_{c0} and no transient speed-up. Signal generator Keysight 33600A provided 5 ns pulses with 2.5 Vpp amplitude and 12 kHz repetition rate. The generator signal was split and sent to the front-end inputs. The input signal in the channel 2 was attenuated by a 6 dB attenuator. The input multiplexer was switched from “crossed” to “straight” state in 0.1 s. The measurement was recorded using capture mode acquiring the ADC data f_{rev} rate, storing every 20th sample.

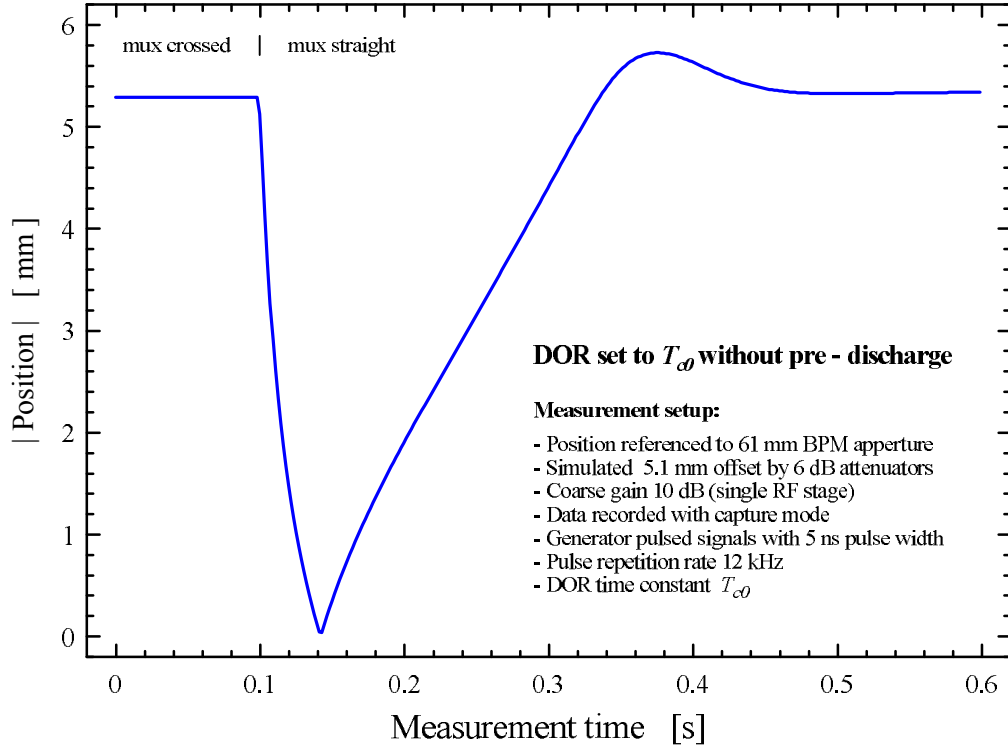


Fig. 5-32. Response of the position signal to the calibration switching transition without the speed-up technique. The detectors were configured with T_{c0} time constant. The absolute position values were computed from the channel 1 and 2 signals showed in Fig. 5-31. The positions were scaled to millimetres using a 61 mm BPM aperture. The position offset of some 5.1 mm was simulated by a 6 dB attenuator connected to a front-end input.

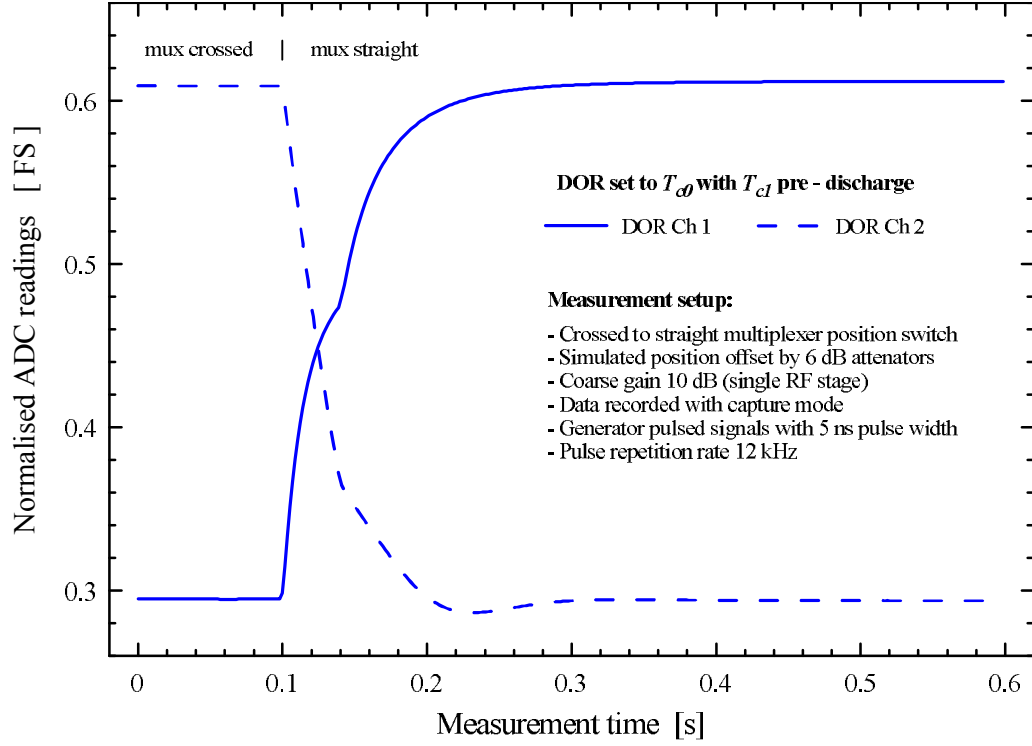


Fig. 5-33. Response of the DOR channel 1 and 2 signals to the calibration switching using the T_{c1} detector pre-discharge. The input multiplexer was switched from “crossed” to “straight” state in 0.1 s. At the same time the detector time constant is switched from T_{c0} to T_{c1} for a duration of 40 ms. The detectors are set back to the nominal time constant T_{c0} in 0.14 s. Signal generator Keysight 33600A provided 5 ns pulses with 2.5 Vpp amplitude and 12 kHz repetition rate. The generator signal was split and sent to the front-end inputs. The input signal in the channel 2 was attenuated by a 6 dB attenuator. The measurement was recorded using capture mode acquiring the ADC data f_{rev} rate, storing every 20th sample.

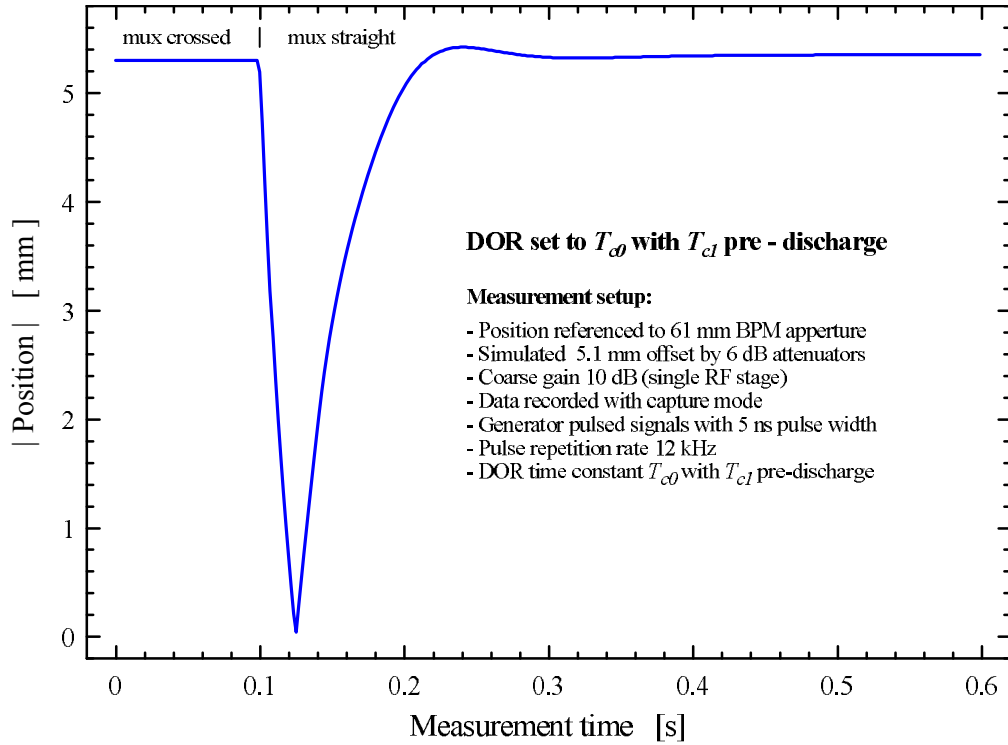


Fig. 5-34. Response of the position signal to the calibration switching transition with the speed-up technique using the T_{c1} for detector pre-discharge. The absolute position values were computed from the channel 1 and 2 signals showed in Fig. 5-33. The positions were scaled to millimetres using a 61 mm BPM aperture. The position offset of some 5.1 mm was simulated by a 6 dB attenuator connected to a front-end input.

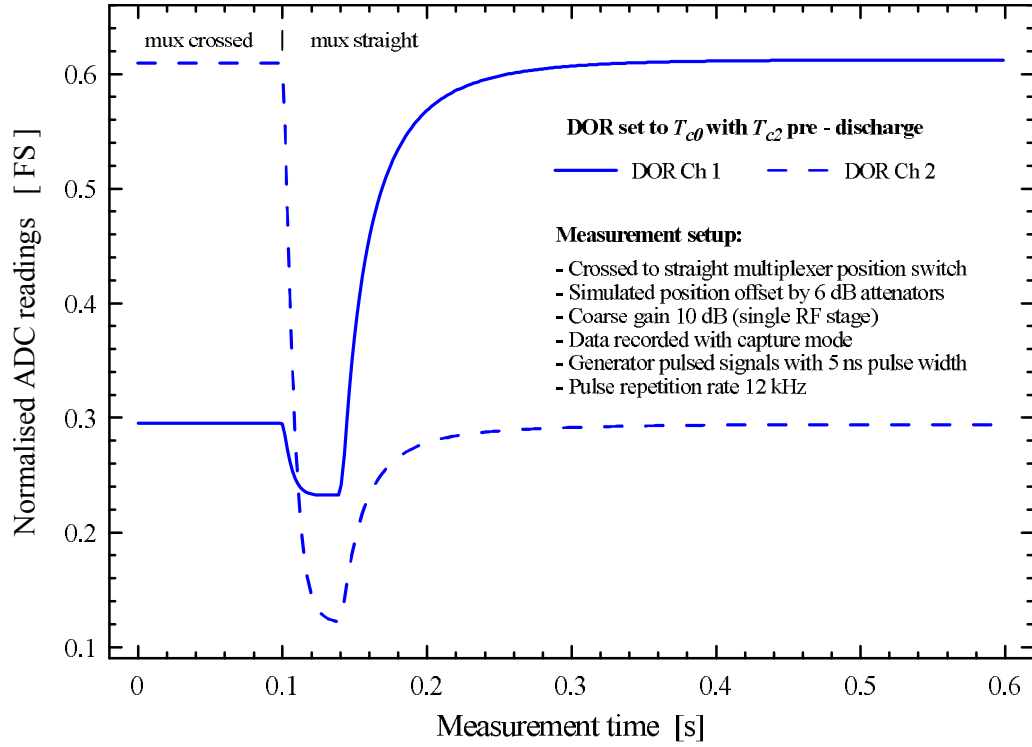


Fig. 5-35. Response of the DOR channel 1 and 2 signals to the calibration switching using the T_{c2} detector pre-discharge. The input multiplexer was switched from “crossed” to “straight” state in 0.1 s. At the same time the detector time constant is switched from T_{c0} to T_{c2} for a duration of 40 ms. The detectors are set back to the nominal time constant T_{c0} in 0.14 s. Signal generator Keysight 33600A provided 5 ns pulses with 2.5 Vpp amplitude and 12 kHz repetition rate. The generator signal was split and sent to the front-end inputs. The input signal in the channel 2 was attenuated by a 6 dB attenuator. The measurement was recorded using capture mode acquiring the ADC data f_{rev} rate, storing every 20th sample.

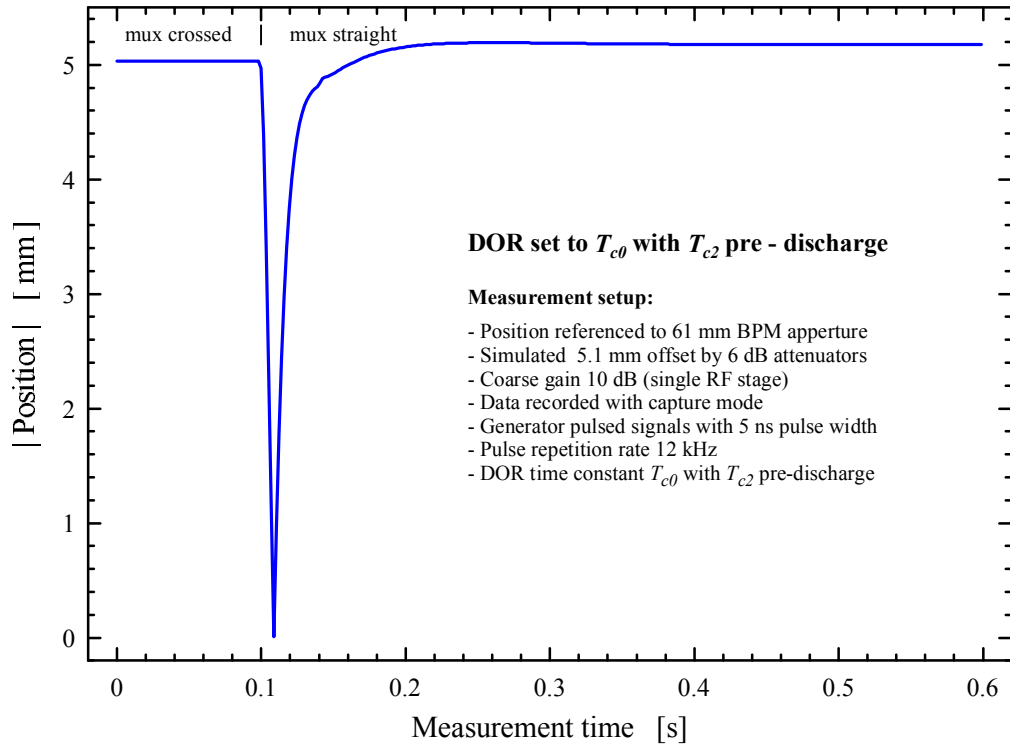


Fig. 5-36. Response of the position signal to the calibration switching transition with the speed-up technique using the T_{c2} for detector pre-discharge. The absolute position values were computed from the channel 1 and 2 signals showed in Fig. 5-35. The positions were scaled to millimetres using a 61 mm BPM aperture. The position offset of some 5.1 mm was simulated by a 6 dB attenuator connected to a front-end input.

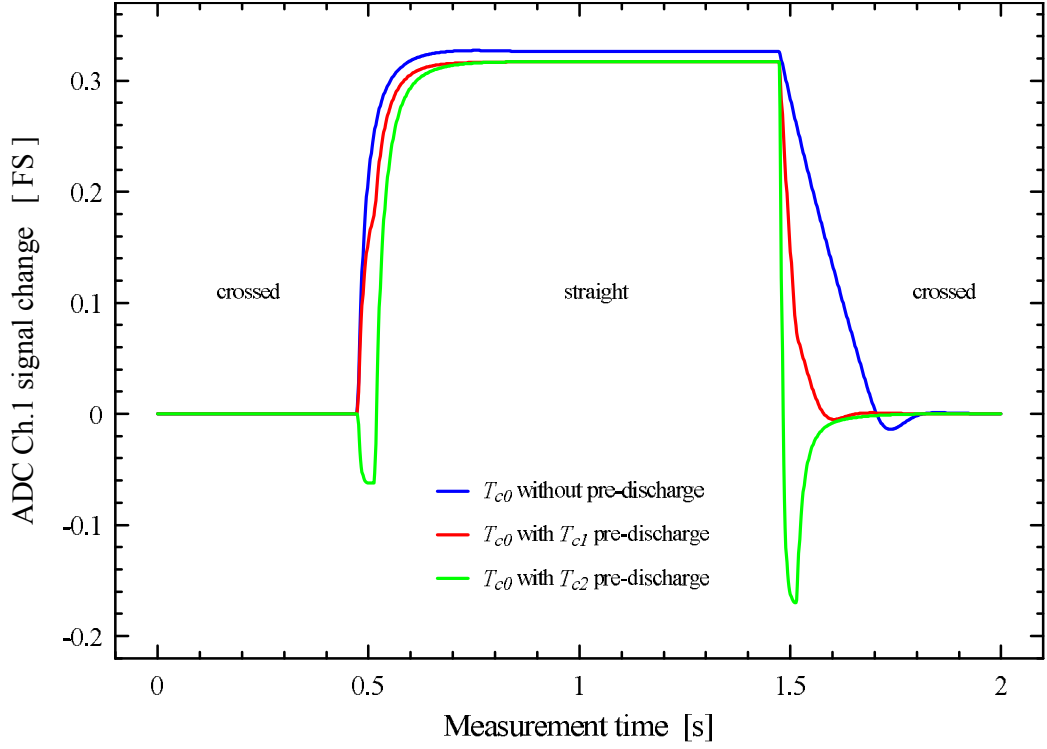


Fig. 5-37. Comparison of a DOR channel 1 responses to the calibration switching with different transient speed-up settings. The multiplexer switching was applied in 0.5 s and 1.5 s. The detector pre-discharge with T_{c1} (red) and T_{c2} (green) was applied for duration of 40 ms after each switching. Obtained DOR signals were referenced to the beginning of the measurement. Signal generator Keysight 33600A provided 5 ns pulses with 2.5 Vpp amplitude and 12 kHz repetition rate. The generator signal was split and sent to the front-end inputs. The input signal in the channel 2 was attenuated by a 6 dB attenuator. The measurement was recorded using capture mode acquiring the ADC data f_{rev} rate, storing every 20th sample.

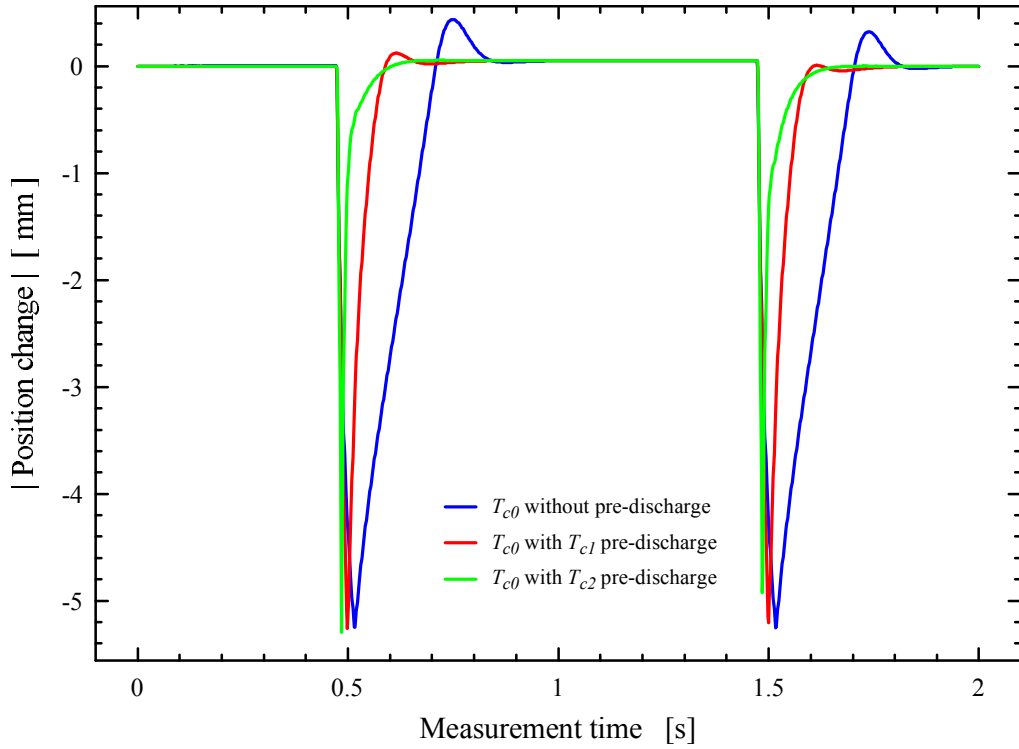


Fig. 5-38. Comparison of the position measurement response to calibration switching with different transient speed-up settings. The absolute position values were scaled to millimetres using a 61 mm BPM aperture and then referenced to the beginning of the measurement. The changes in the position signal correspond to the channel multiplexing showed in Fig. 5-37. The position offset was simulated by a 6 dB attenuator connected to a front-end input. The detectors were pre-discharged with T_{c1} and T_{c2} for duration of 40 ms after each switching.

5.4.1.4 Summary and outlook

Slow transient response of the DOR signals during the calibration switching limits the maximum switching frequency of the calibration which affects the latency of the calibrated position measurements. Additionally, long transient masking decreases number of valid samples averaged in the calibration algorithm which affects the precision of the position measurements. This chapter presented measurements of the transient response of the DOR channels with calibration switching.

The first part of this chapter was focused on measuring the transient response of the DOR signals caused by the multiplexer switching with different detector constants and simulated beam conditions. Both the charging and discharging transients of the DOR detectors were analysed. The longest duration of the switching transient was measured with the slow detector discharge using time constant T_{c0} . The measurement setup was set to simulate beams with the bunch repetition frequency comparable to the LHC revolution and large simulated position offsets of some 5 mm. The computed position signal reached the stable state in about 0.4 s, yielding maximum switching frequency of about 2.5 Hz. The equivalent transient masking of the data transmitted over the UDP protocol would have to be set to 10 samples.

The second part was focused on measurements with the speed-up technique. The aim of this technique is to decrease the detector transient duration by a temporal use of faster time constants. Presented measurement results with T_{c2} pre-discharge showed that the speed-up technique can decrease the transient duration to about 0.15 s. The calibration switching frequency could be increased to about 5 Hz. The equivalent transient masking would be decreased to five 40 ms samples. It was observed that the fixed 40 ms duration used for the detector pre-discharging is sufficient for regular DOROS operation. The optimal timing could be achieved with dynamic control for adapting the discharge duration to the actual signal levels.

It was observed that fast pre-discharging with time constant T_{c2} allowed decreasing only the falling transient (capacitor discharge) at the expense of increasing duration of the rising transient (capacitor charge). The speed-up technique is thus limited mostly by the capacitor charging, which cannot be dynamically controlled. Please note that the presented results were measured using the DOR detectors with 10 nF capacitors.

The response time of the detector can be further improved by decreasing the capacitor value. Comparison of the DOR channel response with 10 nF and 1 nF capacitors in the detectors is showed in Fig. 5-39 for the falling and Fig. 5-40 for the rising transients. The signals were measured with the time constant T_{c0} and T_{c1} settings without using the detector pre-discharge. When using smaller 1 nF capacitors the duration of both rising and falling transients with time constant T_{c0} was smaller than 40 ms. This result is comparable to the actual 25 Hz processing rate of the AUCAL state-machine. The calibration switching could thus reach frequency of 12.5 Hz and decrease the latency of the calibration processing to 80 ms. Increasing the response of the compensated diode detectors consequently decreases their filtering capabilities which can affect the precision of the position measurements. The implementation and use of the faster switching modes in the orbit calibration is currently under studies.

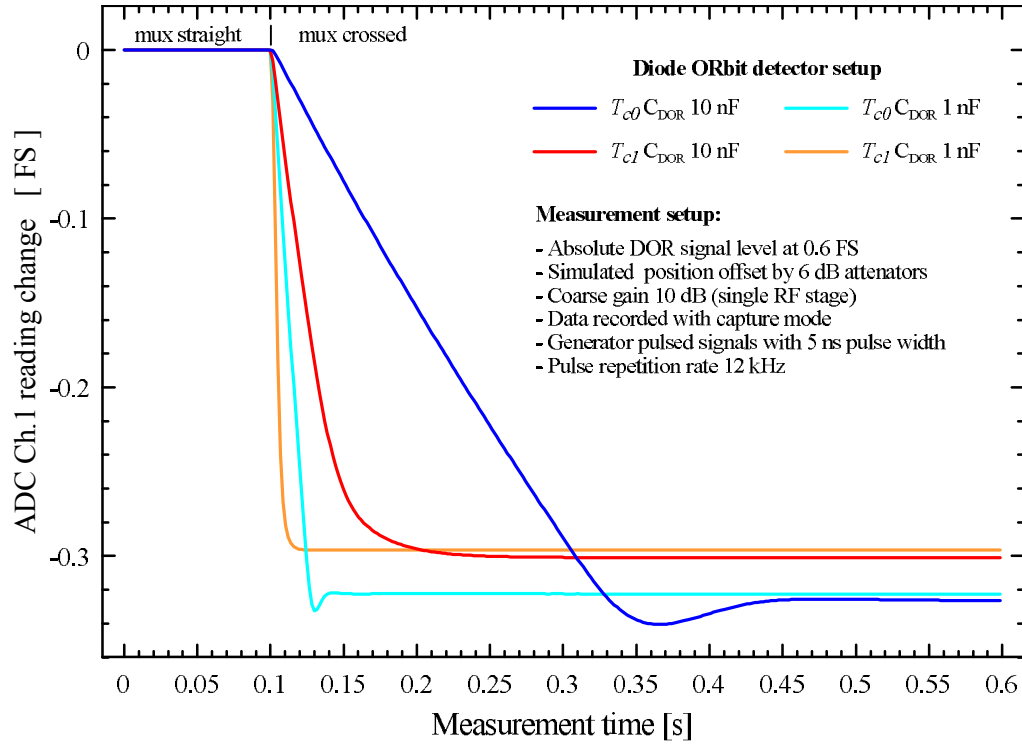


Fig. 5-39. Comparison of falling (discharge) DOR transients caused by calibration switching with 10 nF and 1 nF capacitors in the compensated diode detectors. The DOR signals were referenced to the beginning of the measurement. Signal generator Keysight 33600A provided 5 ns pulses with 2.5 Vpp amplitude and 12 kHz repetition rate. The measurement was recorded using the capture mode acquiring the ADC data at f_{rev} rate, storing every 20th sample.

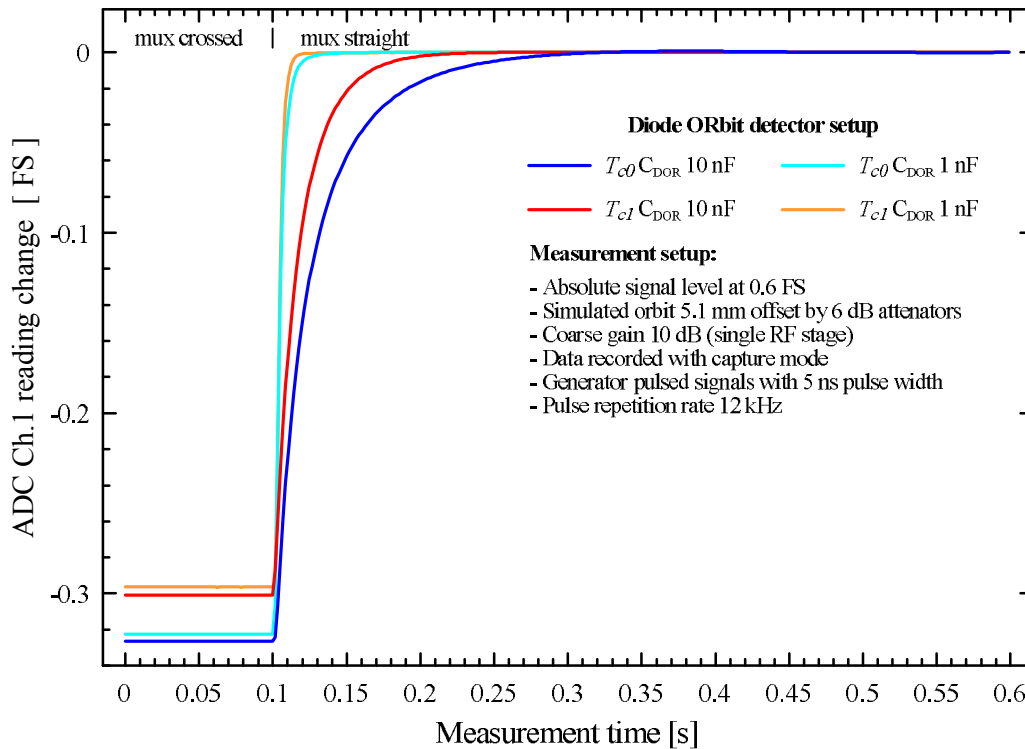


Fig. 5-40. Comparison of rising (charge) DOR transients caused by calibration switching with 10 nF and 1 nF capacitors in the compensated diode detectors. The DOR signals were referenced to the beginning of the measurement. Signal generator Keysight 33600A provided 5 ns pulses with 2.5 Vpp amplitude and 12 kHz repetition rate. The measurement was recorded using the capture mode acquiring the ADC data at f_{rev} rate, storing every 20th sample.

5.4.2 Transient masking

Discussed transient effects in the DOR processing caused by calibration switching can be masked by the AUCAL state-machine. The masked samples are excluded from processing in the calibration algorithm. As shown in the previous chapter, unmasked transients can cause large changes in the position signal resulting in degraded precision, accuracy and reliability of the measurements. The longest transient duration defines the minimum number of samples being masked by the state-machine. The beam setup, position offset and the time constant setting in the DOR detectors influence the optimal masking and the calibration performance. This chapter is focused on measuring the systematic errors and noise of the calibrated position signals measured with different transient masking lengths.

5.4.2.1 Measurement setup

The laboratory setup used for position measurements with the transient masking is shown in Fig. 5-41. Signal generator Keysight 33600A provided 8 ns pulse signal split into eight front-end inputs simulating a beam in the BPM centre. The transient masking was evaluated for the generator and the measured DOROS front-end set to simulate the conditions typical for the LHC pilot and physics beams.

The LHC physics beams were simulated using a 40 MHz pulse repetition rate. The amplitude of the pulses was set to 3 Vpp to achieve the signal level around 50 % of ADC full-scale with a single RF amplifier stage with 10 dB gain. The LHC pilot beams were simulated with 12 kHz pulse repetition rate and the same amplitude. Low intensity of such beams was simulated with 40 dB attenuation of the generator output. The RF amplifier was configured to 50 dB gain to operate around 50 % of ADC full-scale.

The AUCAL state-machine was configured to provide the calibration switching with 1 s period, that is every 25 output samples. The transient speed-up was set to use T_{c2} pre-discharging for 40 ms (1 output sample) followed by T_{c0} used for the steady-state signals. The cut-off frequency of the IIR low-pass filters was increased to 1 kHz during the transients for 160 ms (4 output samples) and decreased to 10 Hz for the steady state signals. The calibration de-multiplexer was synchronously switched with the input multiplexers to correct the position sign changes. The transient masking in the front-end was fixed to 200 ms, excluding 5 out of 25 output samples for each switching period. The data was sent over the UDP protocol to the acquisition server recording the transmitted DOR data. Additional transient masking and computation of the calibrated positions was done in the offline post-processing scripts. The data analysis of different masking values was performed on the same data set to ensure equal measurement conditions for the comparisons.

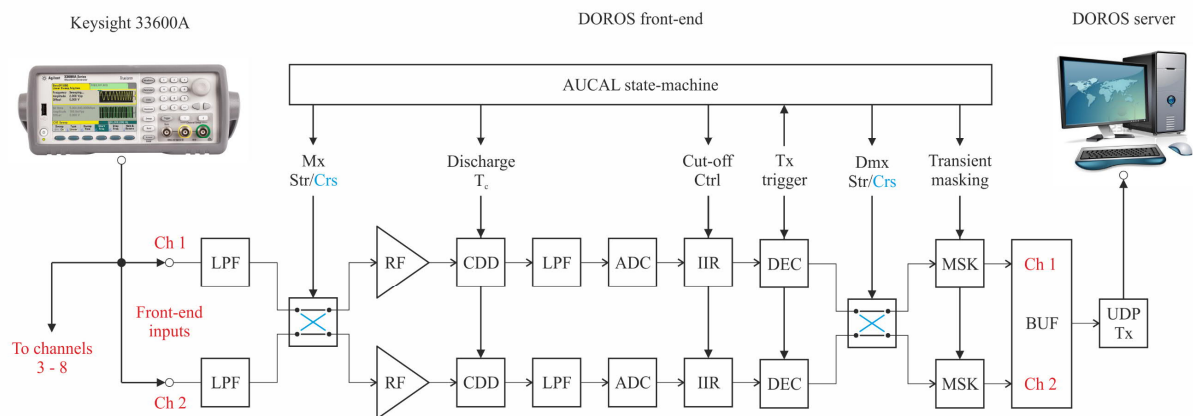


Fig. 5-41. Diagram of the laboratory setup used to measure the performance of the transient masking.

Abbreviations: LPF – analogue Low Pass Filter, CDD – Compensated Diode Detector, RF – Radio Frequency amplifier stage, IIR – digital low pass filters, DEC – sample decimation and UDP synchronisation, MSK – transient sample Masking, AUCAL – Automatic Calibration.

5.4.2.2 Transient masking measurements

The plot in Fig. 5-42 shows the DOR signals from channels 1 and 2 measured with 40 MHz pulse repetition rate and a 10 dB programmable RF gain. Displayed signals were obtained by separating the data corresponding to the straight and crossed multiplexer states and removing the masked samples. The measurements obtained with 5 samples masking is shown in light colours. This setting excludes 20 % of the samples in each multiplexer period. Shown in the dark colours, the transient masking of the measurement data was then increased in the post-processing to 15 samples. This setting excludes 60 % of the samples in each multiplexer period.

The differences seen between the signal levels measured in straight and crossed states are proportional to the asymmetries in the RF processing, mostly dominated by the RF amplifiers. The signal routing through individual RF channels is indicated in the legend for each measured signal. The change in signal level measured between the straight and cross configuration was in the order of 1 %. This asymmetry can be translated into a 150 μm position offset when scaling to a 61 mm BPM aperture. As can be seen in the plot, the masking did not have significant effect on the systematic errors and can be considered negligible.

The acquired DOR signals were then used in the post-processing algorithm to compute calibrated positions displayed in Fig. 5-43. As described in chapter 4.3.4.4, the calibration algorithm averages the valid samples in each calibration period reducing the data rate to 1 S/s. Computed position values were scaled to micrometres using equation (3-2) referenced to a 61 mm BPM aperture. The plot displays the calibrated position signals computed with 5 S, 15 S and 24 S transient masking. The corresponding sample averaging in the algorithm for each signal is indicated in the legend of the plot. As can be seen, the sample masking has negligible effect on the calibrated positions.

The mean position offset quantifying the accuracy of measuring the simulated centred beam was estimated to be around 33 μm . This error can be translated to a 0.4 % residual asymmetry between the input signals which could not be calibrated by the signal switching. Such error can be explained by the signal asymmetry in front of the calibration multiplexers, dominated by the input RF filters and distribution of the signals from the generator.

Noise of the calibrated position was estimated during two measurements, lasting 1 minute and 10 minutes, by computing standard deviations (STD) and the peak-peak values. Estimated noise results obtained for different transient masking applied in the post-processing and three calibration averaging lengths were summarized in Table 5-9. Both masking and averaging lengths were expressed as percentage of the switching period. The table also provides computed ratio between the peak-peak and standard deviation values. As seen in the table, the noise in calibrated position signal decreases by some 40 % when increasing the averaging from 1 to 20 samples. The gain in the signal-to-noise performance is smaller than would be expected when averaging white noise. It means that the number of masked samples can be increased to improve the margin for the transient effects at the expense of relatively small noise increase.

Table 5-9. Calibrated position noise analysis with 10 dB RF gain and 40 MHz pulse repetition.

Transient masking	24 samples (96 % T_{mux})		15 Samples (40 % T_{mux})		5 Samples (20 % T_{mux})	
Calibration averaging	1 sample (4 % T_{mux})		10 Samples (60 % T_{mux})		20 Samples (80 % T_{mux})	
Noise estimation length	1 min	10 min	1 min	10 min	1 min	10 min
Peak-peak [μm] *	0.28	0.44	0.22	0.4	0.21	0.34
STD [μm_{RMS}] *	0.07	0.08	0.06	0.06	0.05	0.06
Peak-peak / STD ratio *	3.93	5.81	3.93	6.18	4.23	5.69

* Rounded to 2 decimal places.

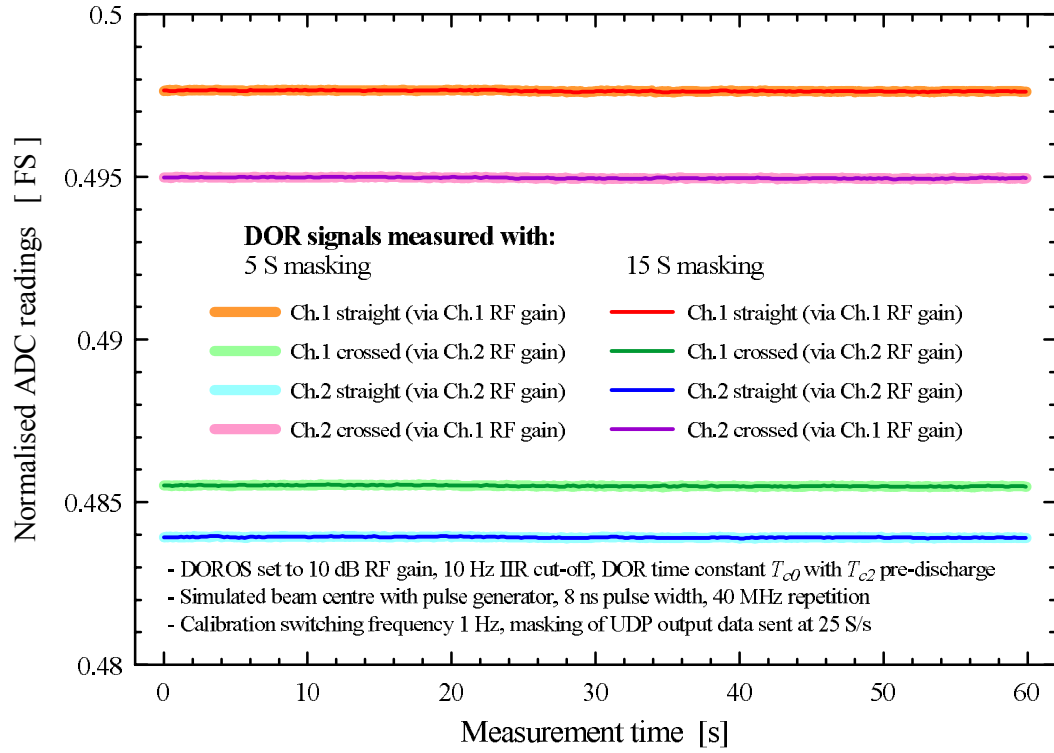


Fig. 5-42. Measured signals from DOR channels 1 and 2 after the calibration switching. Displayed signals were obtained by separating the data belonging to the straight and crossed multiplexer states and removing the masked samples. The transient masking of 5 samples (light colour) and 15 samples (dark colour) was applied in the post-processing of the same acquisition data. The generator Keysight 33600A was set to provide 8 ns pulses with 40 MHz repetition rate and 3 V_{pp} amplitude. The signal was split between the front-end channels simulating zero position offset.

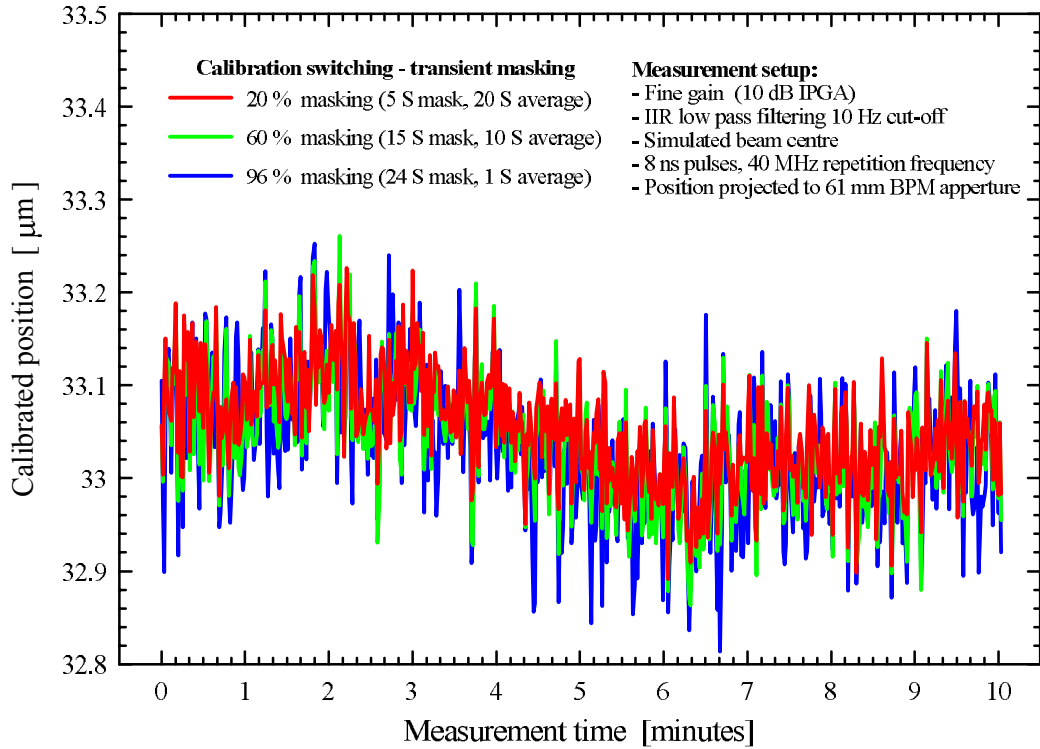


Fig. 5-43. Calibrated position signal with transient masking settings applied in the post-processing of the same acquisition data. The generator was set to 8 ns pulses, 40 MHz repetition rate and 3 V_{pp} amplitude simulating typical BPM signals with LHC physics beam and zero position offset. The programmable RF amplifier in the DOROS front-end was set to 10 dB gain using the Integrated Programmable Gain Amplifier (IPGA) stage.

The plot in Fig. 5-44 shows the DOR signals from channels 1 and 2 measured with 12 kHz pulse repetition rate and a 50 dB programmable RF gain. It can be seen that the masking by 5 samples did not completely remove the calibration transients. The residual change in the DOR signals was measured around 0.2 % of the ADC full-scale in both multiplexer configurations. The channel asymmetry between the two DOR channels was measured to be in order of 1 %, a similar value to the result of the previous measurement with 10 dB RF gain. Note that the channel asymmetry changes with the changing configuration of the programmable RF amplifier. The additional 40 dB attenuator used in this measurement does not influence the measured asymmetry as the attenuator is connected to the generator output common for both DOR channels.

The plot in Fig. 5-45 shows the computed calibrated positions from the switched DOR signals. The unmasked transients are not seen as the calibration algorithm computes moving average of the data acquired over the consecutive multiplexer states. The mean position offset quantifying the accuracy of the measured simulated centred beam was estimated to some 22 μm . The dependency of the position offset on the transient masking was evaluated by comparing the average positions computed over the first minute of the measurement. Comparison of results obtained with different masking lengths yielded around 20 nm change in the absolute position. However, it can be seen in the plot that the masking parameter influences the noise performance of the measurements.

Noise in the position signal can be estimated by computing the standard deviation and peak-peak values upon the 1 minute and 10 minutes of the measurement. The obtained values computed for the three masking parameters are summarized in Table 5-10. Please note that the analysis was applied to the same acquired data. As can be seen in the table, increasing the averaging from 1 to 20 samples reduced the estimated noise levels by factor 3. Increasing the averaging length in the calibration algorithm can thus significantly improve the precision of the measurement with low repetition rates.

It was also observed that the noise levels increased in comparison to the previous measurement with the 40 MHz pulse repetition. The plot in Fig. 5-46 shows the standard deviations of the position signal computed for different masking settings using the first minute of the measured data, corresponding to 60 calibrated positions. The plot vertical axis displays the standard deviation values in logarithmic scale. The horizontal bottom axis is expressed in the number of masked samples and the horizontal top axis is scaled in numbers of averaged samples. The plot also summarizes results measured with 40 MHz pulse repetition and 10 dB RF gain as well as the second measurement with 12 kHz pulse repetition and 50 dB RF gain. The plot allows to compare the standard deviations obtained with the 1 dB step fine RF gain scheme (solid) and the 2.5 dB step coarse scheme (dashed). As explained in chapter 4.3.2.1 the fine scheme uses the programmable amplifier (IPGA) with open loop. The coarse scheme uses only the RF stages with the IPGA bypassed. The results obtained with 5 sample and 24 sample masking are summarized in Table 5-11. It was observed that decreasing number of the masked samples improved the noise performance by factor 3 mostly for the high RF gains settings, regardless of the gain scheme. Changing the masking parameter with low gains had less than 2 % influence on the measurement precision. It was also observed that the noise performance for the low RF gains can be further improved when using coarse gain scheme. Improvement of factor 5 was achieved by using coarse gain configuration with minimum masking of 5 samples, the low gain of 10 dB and 40 MHz pulse repetition.

Table 5-10. Calibrated position noise analysis with 50 dB RF gain and low pulse repetition frequency.

Transient masking	24 samples (96 % T_{mux})		15 Samples (40 % T_{mux})		5 Samples (20 % T_{mux})	
Calibration averaging	1 sample (4 % T_{mux})		10 Samples (60 % T_{mux})		20 Samples (80 % T_{mux})	
Calculation period	1 min	10 min	1 min	10 min	1 min	10 min
Peak-peak [μm] *	4.47	6.88	2.59	3.45	1.57	2.82
STD [$\mu\text{m RMS}$] *	1.06	1.16	0.52	0.54	0.34	0.40
Peak-peak / STD ratio *	4.21	5.95	4.95	6.37	4.61	7.07

* Rounded to 2 decimal places.

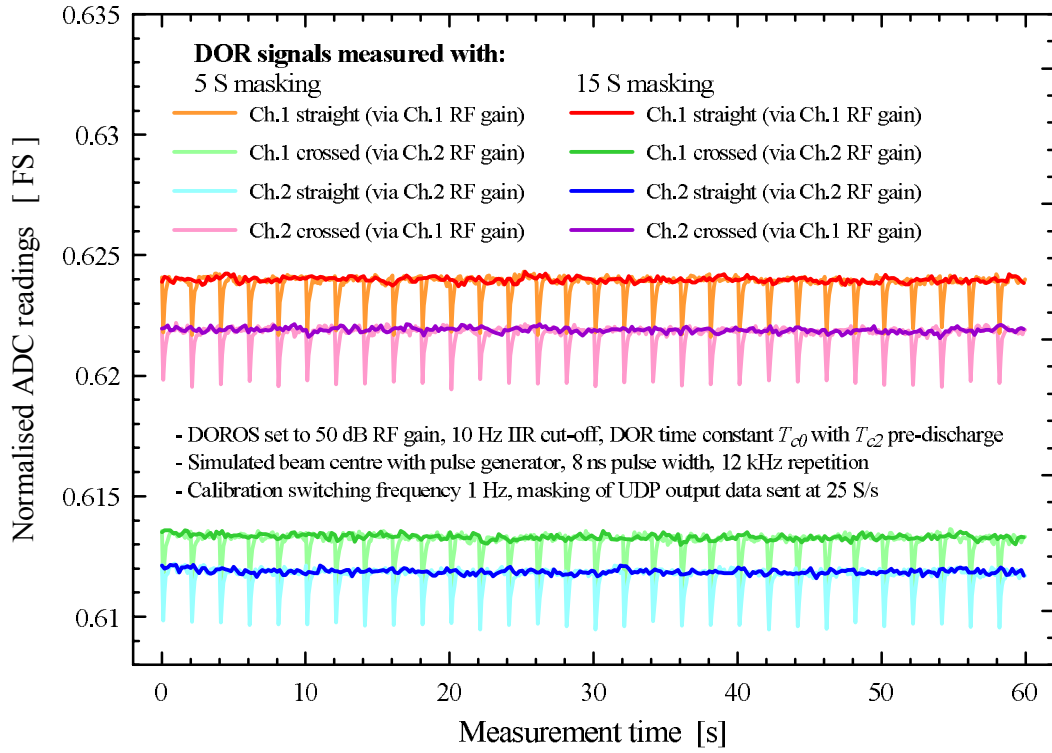


Fig. 5-44. Measured signals from DOR channels 1 and 2 after calibration switching. The signals were obtained by separating the data corresponding to the straight and crossed multiplexer states and removing the masked samples. The transient masking of 5 samples (light colour) and 15 samples (dark colour) was applied in the post-processing. The input signal was 8 ns pulses with 12 kHz repetition rate and 3 Vpp amplitude attenuated by a 40 dB attenuator on the generator output. The signal was split between the front-end channels simulating zero position offset. The measured channels were operated with 50 dB RF gain.

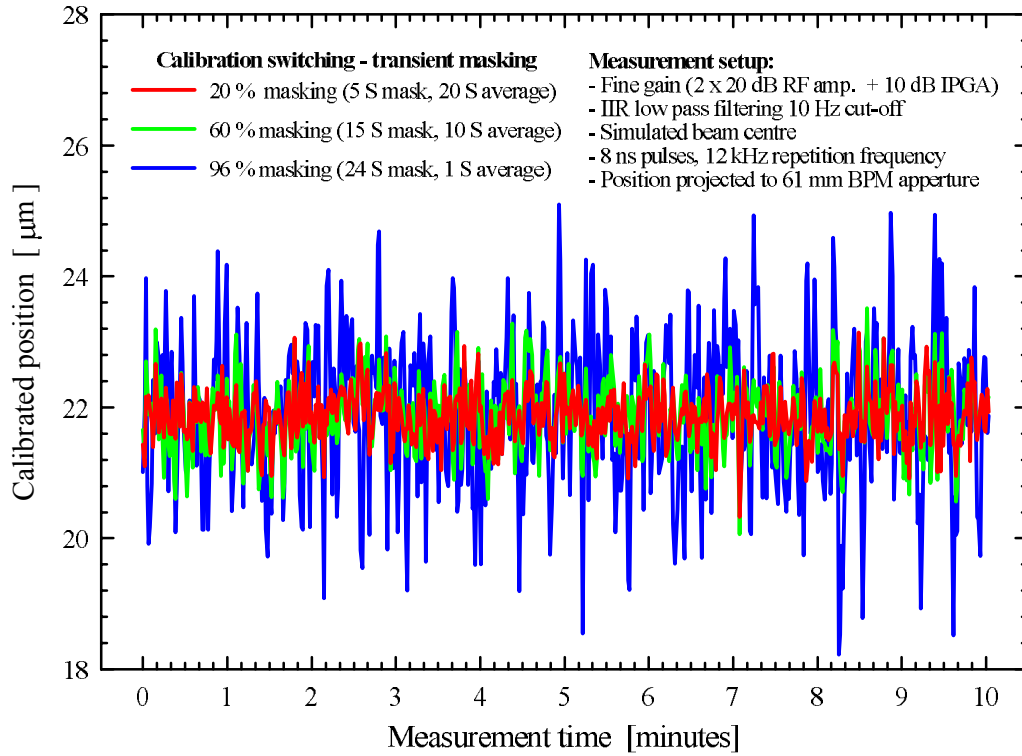


Fig. 5-45. Calibrated position signal with three transient masking settings applied in the post-processing of the same acquisition data. The generator was set to 8 ns pulses, 12 kHz repetition rate and 3 Vpp amplitude simulating typical BPM signals with LHC physics beam and zero position offset. The pulses were attenuated by a 40 dB attenuator on the generator output. The programmable RF amplifier in the DOROS front-end was set to 50 dB gain.

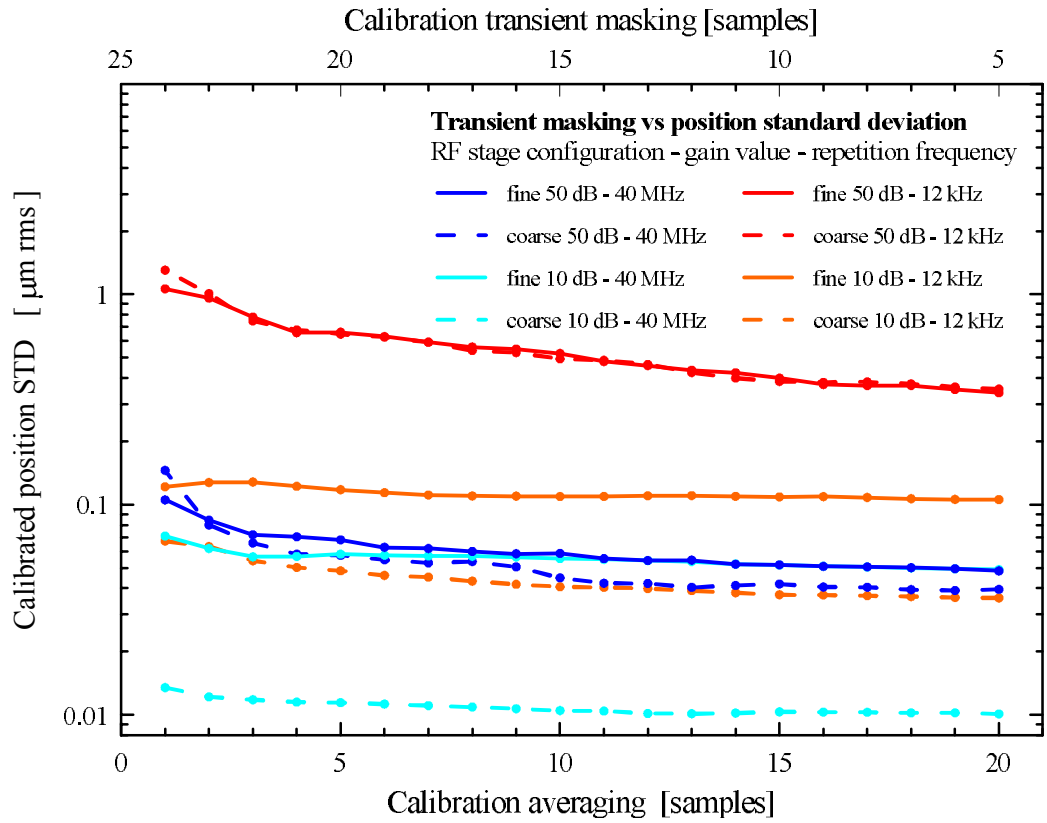


Fig. 5-46. Standard deviation computed upon 60 calibrated positions (1 minute measurement) displayed as a function of transient masking. The horizontal axis was expressed in the number of masked samples (bottom axis) and number of averaged samples (top axis). The generator Keysight 33600A was set to provide 8 ns pulses with 3 Vpp amplitude. The repetition rate was set to 12 kHz (red and orange) and 40 MHz (blue and cyan). The measurement were repeated with 10 dB RF gain (cyan and orange) and 50 dB RF gain (blue and red) with a 40 dB attenuator on the generator output. The measurement were obtained repeated with fine (solid line) and coarse (dashed line) RF gain schemes. The fine scheme uses the programmable amplifier (IPGA) with open loop. The coarse scheme uses only the RF stages with the IPGA bypassed.

Table 5-11. Summary of the calibrated orbit noise for 50 dB and 10 dB RF gains and pulse repetition frequencies of 40 MHz and 12 kHz.

RF gain [dB]		50 dB				10 dB			
Gain stage configuration		Fine		Coarse		Fine		Coarse	
Calibration averaging [samples]		1 S	20 S	1 S	20 S	1 S	20 S	1 S	20 S
Transient masking [samples]		24 S	5 S	24 S	5 S	24 S	5 S	24 S	5 S
Standard deviation [μm RMS] *	Pulse repetition 40 MHz	0.11	0.05	0.15	0.04	0.07	0.05	0.01	0.01
	Pulse repetition 12 kHz	1.06	0.34	1.3	0.35	0.12	0.11	0.07	0.04

* Rounded to 2 decimal places.

5.4.2.3 *Summary and outlook*

The measurements presented in this chapter showed the impact of the transient masking on the accuracy and precision of the position measurement. Presented results were obtained with two generator settings simulating centred LHC beams, representative for the collimator system. Measurements were done to simulate the high intensity physics beams and the low intensity pilot beam. The simulated beam positions were computed with 1 Hz calibration switching and the transient masking applied in post-processing of the acquired data. The measurement results obtained with the fine gain and coarse gain schemes were also compared at the end of this chapter.

It was observed that the transient masking at high pulse repetition rates simulating LHC physics beams did not influence the systematic errors of the calibrated position measurements. The calibrated absolute position measured with the fine gain was around 33 μm with the noise estimated by the data standard deviation of some 0.07 μm_{RMS} . Minimizing the masking to 5 samples decreased the estimated standard deviation to some 0.05 μm_{RMS} improving the noise performance by 40 %. The measurement results were dominated by the low frequency drifts estimated to some 0.2 μm . The noise performance could be further improved when using the coarse gain scheme which bypasses the IPGA. The noise of the calibrated positions was estimated to some 0.01 μm_{RMS} , that is improvement of factor 5 with respect to the fine gain scheme.

At low repetition rates simulating LHC pilot beams the absolute calibrated position was measured with the laboratory setup to be around 22 μm . The measured position was changing with the used transient masking only by some 20 nm. The noise of the measurement with maximum masking was estimated to 1 μm_{RMS} . It was possible to decrease the noise to some 0.3 μm_{RMS} , improving the noise by factor 3 with respect to the results obtained with the masking of just 5 samples. However, the switching transients were not fully removed with the 5 sample masking, leaving some 0.2 % change in the DOR signals. Using the fine and coarse gain schemes did not have a significant impact on the noise performance.

5.4.3 Precision and accuracy

The DOROS front-ends provide automatic control of the RF gain to adapt the large dynamic range of the BPM signals to the limited range of linear operation of the compensated diode detectors. As explained in chapter 4.3.2.1, each electrode signal is amplified by a combination of four configurable RF stages. The stages can be controlled using so called fine and coarse gain schemes, which provide 1 dB and 2.5 dB gain steps, resulting in different measurement performance. The asymmetries between the channels depending mostly on the gain configuration, are minimised by the calibration switching technique. The technique also reduces systematic measurement errors and minimizes low frequency drifts.

The measurements presented in this chapter were focused on evaluating the precision and systematic errors of the calibrated position signals with different RF gain settings. The measured data was used in the post-processing to compute the calibrated position signals. The same data was also used to compute the raw position signals showing the measurement performance without the calibration switching. Comparison of the resulting signals allows performance evaluation of the calibration switching with different gain values and gain schemes. The measurements were focused on the low repetition rates.

5.4.3.1 Measurement setup

The laboratory setup used in the following measurements was identical to the setup showed in Fig. 5-41 described in chapter 5.4.2.1. The generator Keysight 33600A provided 8 ns pulse signal with 3 V_{pp} amplitude. The setup was configured to simulate a single bunch in the LHC with repetition rate of 12 kHz. The signal which was split and provided to eight front-end inputs simulating a centred beam. The 10 dB RF gain allowed setting the DOR signal level to 50 % of ADC full-scale. Measurements with the high gain setting used additional 40 dB attenuator connected on the generator output. The RF amplifier was then configured to 50 dB gain to set the signal level back to 50 % of ADC full-scale. The DOR processing was configured with the T_{c0} time constant, T_{c2} pre-discharging, 15 sample masking and 10 Hz cut-off frequency for the IIR filters. Obtained position signals were scaled to millimetres using equation (3-2) referenced to a 61 mm BPM aperture.

The measurements were repeated with the two schemes used to control the RF gains. The fine scheme provides 1 dB gain steps using the integrated programmable gain amplifier (IPGA). The IPGA is an RF amplifier operating in an open loop and providing up to 20 dB gain. Its datasheet bandwidth is 700 MHz. The 10 dB gain used in the measurements consisted only of the IPGA gain stage. The 50 dB gain adds the two additional 20 dB stages, each built with one current-feedback amplifier.

The coarse scheme provides 2.5 dB steps achieved by combining resistive attenuators and the RF stages. The IPGA amplifier is bypassed and disabled when using this scheme. Both 10 and 50 dB gain configurations used in the following measurements consisted only of the RF stages with operational amplifiers.

5.4.3.2 Orbit measurements

The plot in Fig. 5-47 shows the position signals measured with 12 kHz pulse repetition and 10 dB gain configured using the fine scheme. The raw position signals shown in the plot were obtained by separating the measured samples corresponding to the straight (str.) and crossed (crs.) multiplexer states and averaging the valid samples from each switching period. The calibrated position signal was obtained from the calibration algorithm which uses both the straight and crossed samples. Please note that the 0.5 Hz rate of the raw signals was increased to match the 1 Hz rate of the calibration by repeating the last value. The position change was referenced to the beginning of the measurement by subtracting the first samples from the respective signals. The measurement and signal processing was repeated with the coarse gain scheme. The resulting position signals are showed in Fig. 5-48.

Table 5-12 summarizes the average values of the calibrated and raw position signals. The values were computed upon 10 minute measurements for both the fine and coarse gain schemes. It can be seen that the calibration switching is important for reducing the systematic errors caused by the asymmetries between the two DOR channels. The absolute position error measured without the calibration was ranging between $-10\text{ }\mu\text{m}$ and $180\text{ }\mu\text{m}$. The worst case result was obtained with fine gain scheme where the error reached more than $\pm 130\text{ }\mu\text{m}$. The residual position error after the calibration processing was measured around $21\text{ }\mu\text{m}$. The difference of the calibrated positions measured with the two gain schemes was less than $2\text{ }\mu\text{m}$.

It can be seen that the raw position signals are dominated by a low frequency noise and drifts of the channel electronics. Table 5-12 summarizes also standard deviation (STD) and peak-peak values computed for the 10 minute measurements of the raw and calibrated position signals measured with the fine and coarse gain schemes. The noise of the raw position signals measured with the fine scheme was estimated to some $1.3\text{ }\mu\text{m}_{\text{RMS}}$. The calibration switching decreased the noise by factor 8, yielding the STD of some $0.15\text{ }\mu\text{m}_{\text{RMS}}$. The noise of the raw signals with the coarse scheme was measured around $0.8\text{ }\mu\text{m}_{\text{RMS}}$. The calibration switching decreased the noise to less than $0.07\text{ }\mu\text{m}_{\text{RMS}}$ yielding improvement of factor 11. The improvement in comparison to the fine scheme was more than factor 2.

The peak-peak estimation in the raw signals is dominated by low frequency drifts ranging between $3\text{ }\mu\text{m}$ and $5\text{ }\mu\text{m}$. The calibration reduced this noise to less than $1\text{ }\mu\text{m}$. As seen in Fig. 5-48 the residual drift was estimated to less than $0.2\text{ }\mu\text{m}$ when using the coarse scheme. Short bursts of noise with peak-peak amplitude up to some $0.3\text{ }\mu\text{m}$ were also observed. An example of one such burst can be seen between the 7th and 8th minute of the measurement in Fig. 5-48. The bursts in the plot are correlated with the fast changes in the raw signals lasting between 20 to 30 seconds.

The measurements were repeated with increased RF gain to 50 dB. The same post-processing was applied to the acquired data in order to obtain the raw and calibrated position signals. Obtained results with the fine gain scheme were displayed in Fig. 5-49 and with the coarse scheme in Fig. 5-50. Similarly to the previous measurements the first position samples were subtracted from the signals to display only the relative position changes. Average positions of both raw and calibrated signals were computed upon the 10 minute measurement and summarized in Table 5-13. The absolute position error measured without the calibration was ranging between $-105\text{ }\mu\text{m}$ and $150\text{ }\mu\text{m}$. The worst case result was obtained with the fine gain scheme where the error reached more than $\pm 100\text{ }\mu\text{m}$. The residual position error after the calibration processing was reduced to around $21\text{ }\mu\text{m}$. When comparing obtained results in Table 5-12 and Table 5-13 it can be seen that the position error changed when increasing the RF gain from 10 dB to 50 dB. The most significant position change of some $50\text{ }\mu\text{m}$ was observed with the coarse gain setup. The change seen in the fine gain setup was around $30\text{ }\mu\text{m}$. The systematic errors in the calibrated position signal changed by less than $0.5\text{ }\mu\text{m}$ in both measurements with the fine and coarse gain schemes.

Table 5-12. Performance of the raw and calibrated position signals measured with 10 dB gain.

10 dB gain	Average position over 10 minutes [μm] *			Standard deviation over 10 minutes [μm_{RMS}] *			Peak-peak over 10 minutes [μm] *		
	Str.	Crs.	Calib.	Str.	Crs.	Calib.	Str.	Crs.	Calib.
fine	178.13	-133.02	22.67	1.34	1.34	0.16	5.97	5.8	1.03
coarse	52.24	-10.18	20.76	0.79	0.81	0.07	3.36	3.12	0.42

* Rounded to 2 decimal places.

Table 5-13. Performance of the raw and calibrated position signals measured with 50 dB gain.

50 dB gain	Average position over 10 minutes [μm] *			Standard deviation over 10 minutes [μm_{RMS}] *			Peak-peak over 10 minutes [μm] *		
	Str.	Crs.	Calib.	Str.	Crs.	Calib.	Str.	Crs.	Calib.
fine	149.29	-105.61	21.87	1.58	1.65	0.54	11.36	12.99	3.45
coarse	-0.95	43.52	21.28	1.18	1.21	0.54	7.17	6.38	3.58

* Rounded to 2 decimal places.

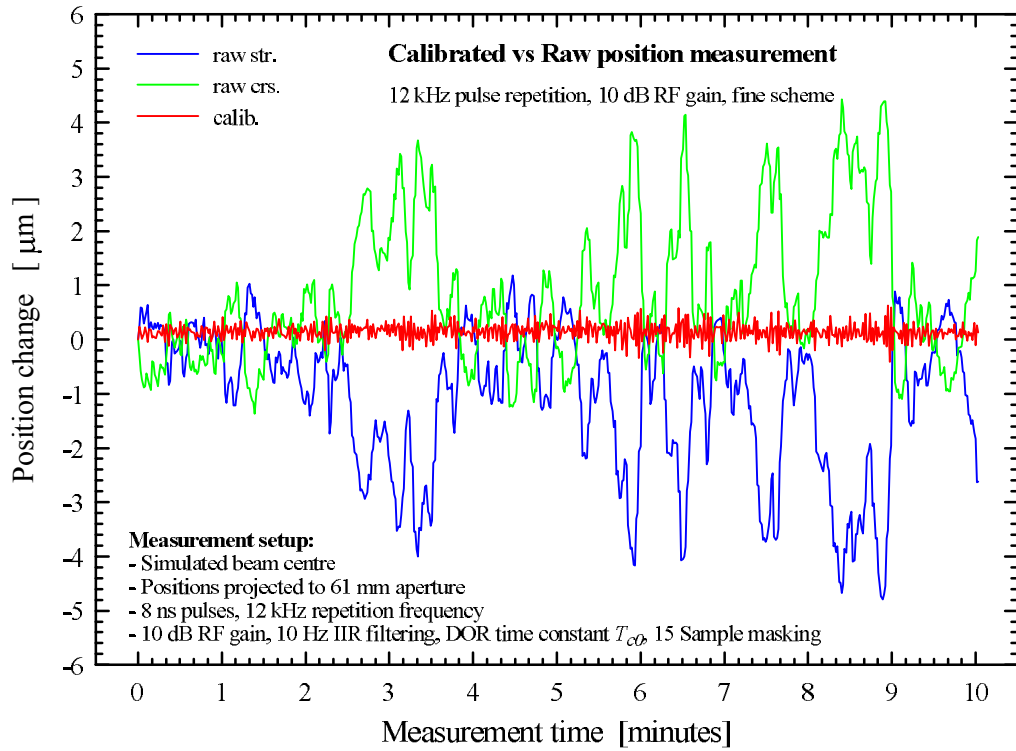


Fig. 5-47. Measurement of the relative change of the calibrated position signal and the straight and crossed raw position signal. The generator Keysight 33600A was set to provide 8 ns pulses with 3 Vpp amplitude and 12 kHz repetition rate. The signal was split to 8 front-end inputs. The 10 dB gain in the fine scheme consists only of the IPGA. Other amplifiers in the RF chain were bypassed. The DOR processing was configured with detector time constant T_{c0} with T_{c2} pre-discharging, 15 sample masking and 10 Hz low-pass IIR filter cut-off.

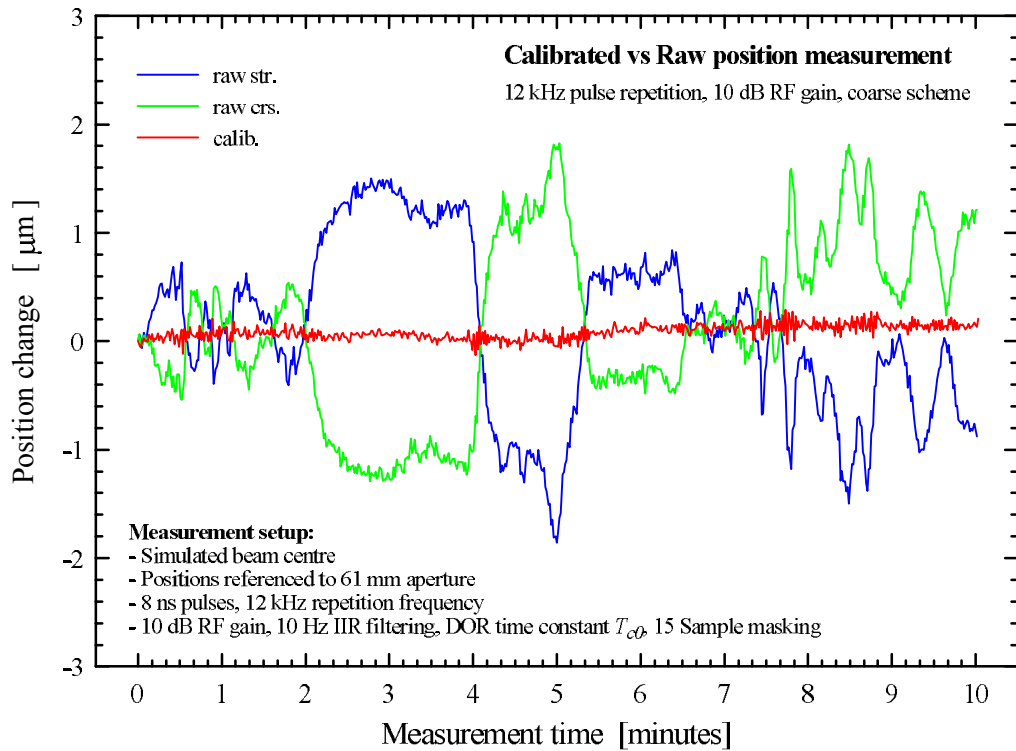


Fig. 5-48. Measurement of the relative change of the calibrated position signal and the straight and crossed raw position signal. The generator Keysight 33600A was set to provide 8 ns pulses with 3 Vpp amplitude and 12 kHz repetition rate. The signal was split to 8 front-end inputs. The 10 dB gain in the coarse scheme consists of a single operational amplifier. The IPGA was disabled and bypassed. The DOR processing was configured with detector time constant T_{c0} with T_{c2} pre-discharging, 15 sample masking and 10 Hz low-pass IIR filter cut-off.

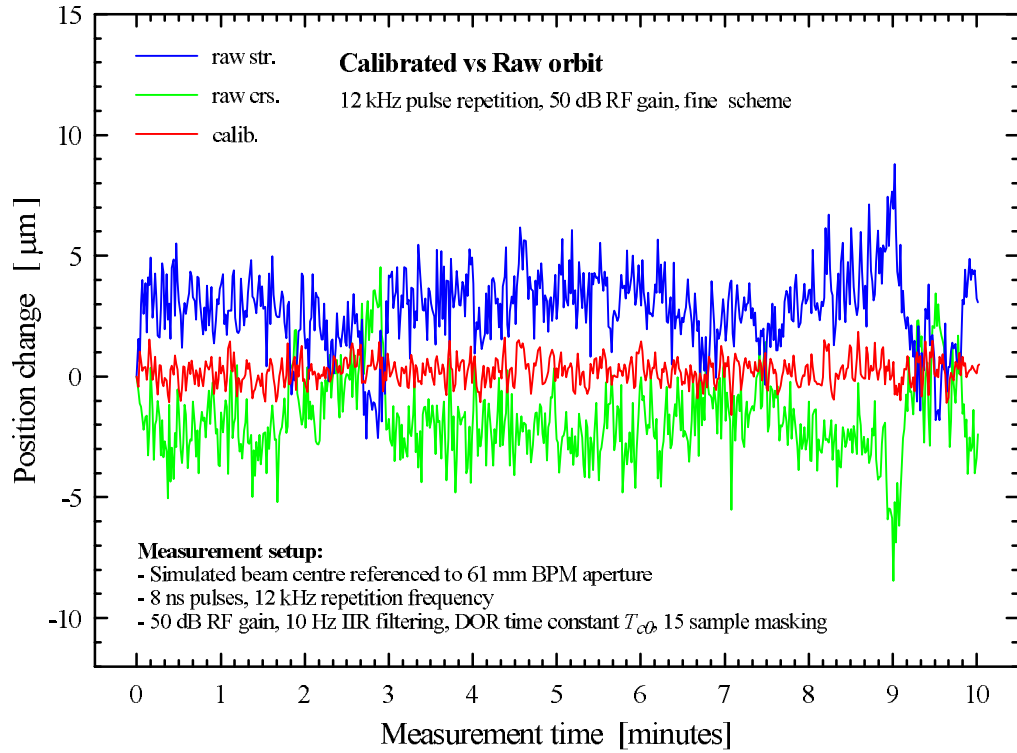


Fig. 5-49. Measurement of the relative change of the calibrated position signal and the straight and crossed raw position signal. The generator Keysight 33600A was set to provide 8 ns pulses with 3 Vpp amplitude and 12 kHz repetition rate. The pulses were attenuated by a 40 dB attenuator on the generator output. The signal with was split to 8 front-end inputs. The 50 dB gain in the fine scheme consists of the IPGA and two 20 dB stages in the RF chain. The DOR processing was configured with detector time constant T_{c0} with T_{c2} pre-discharging, 15 sample masking and 10 Hz low-pass IIR filter cut-off.

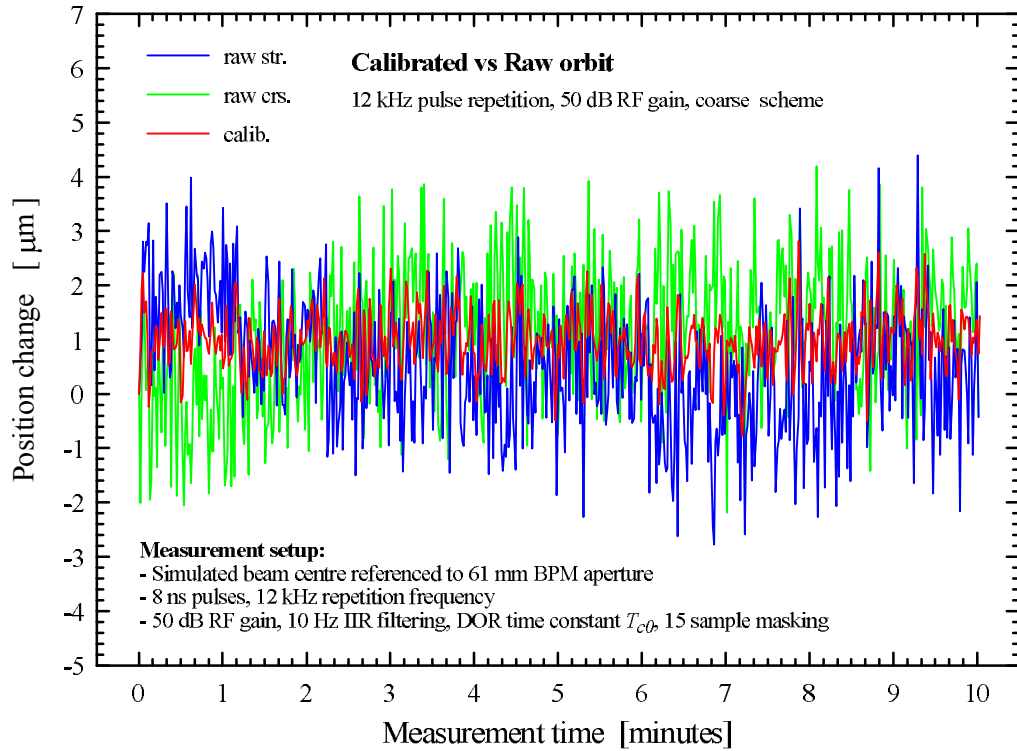


Fig. 5-50. Measurement of the relative change of the calibrated position signal and the straight and crossed raw position signal. The generator Keysight 33600A was set to provide 8 ns pulses with 3 Vpp amplitude and 12 kHz repetition rate. The pulses were attenuated by a 40 dB attenuator on the generator output. The signal with was split to 8 front-end inputs. The 50 dB gain in the coarse scheme consists of one 10 dB and two 20 dB stages in the RF chain. The DOR processing was configured with detector time constant T_{c0} with T_{c2} pre-discharging, 15 sample masking and 10 Hz low-pass IIR filter cut-off.

The Table 5-13 summarizes the STD and peak-peak values computed upon the 10 minutes of the position signals measured with the 50 dB gain. The peak-peak of the signals was measured around 12 μm for the fine and 7 μm for the coarse gain. As observed in Fig. 5-49 the fine gain measurement was dominated by a low frequency drift. The peak-peak noise of the calibrated signals was in both gain schemes estimated to some 3 μm . The STD of raw position signals was estimated to some 1.6 μm_{RMS} when measured with the fine scheme and to 1.2 μm_{RMS} when measured with the coarse scheme. Choosing the coarse gain over the fine gain would improve the STD of the raw position signals by some 30 %. The calibration algorithm would decrease the estimated noise level in both measurements by more than factor 2. The plot in Fig. 5-51 shows the calibrated position signals measured with the 12 kHz pulse repetition. The plot summarizes the signals measured with the 10 dB and 50 dB RF gain settings as well as the fine and coarse gain schemes.

The performance of the calibration switching was also measured with 40 MHz pulse repetition setting in the generator to simulate the 25 ns LHC bunch spacing typical for physics beams. The rest of the measurement setup and data processing described in chapter 5.4.3.1 remained unchanged. The plot in Fig. 5-52 shows the change of the calibrated position signals measured with the 10 dB And 50 dB gains as well as fine and coarse schemes. The change is referenced to the average position values measured in each signal. The Table 5-14 summarizes the average position values, peak-peak amplitudes and standard deviations computed over the first minute and 10 minutes of both the 12 kHz and 40 MHz measurements. It was observed that increasing the pulse repetition from 12 kHz to 40 MHz changed the calibrated positions by some 10 μm . This change was systematic for all measurement settings. Such change in the position could be explained by a frequency dependent asymmetry of around 0.1 % between the input filters. The calibration switching of the multiplexers located after the input RF filters does not compensate such errors.

Comparison of the STD values shows that the calibrated position measurement using the fine gain scheme have worse noise performance over the coarse scheme. This can be seen especially at low gain when only one gain stage was active in the RF amplifier. The largest noise improvement of factor 6 was achieved with coarse gain when measuring 40 MHz generator signals. The noise level was estimated over 1 minute to some 0.01 μm_{RMS} and 0.06 μm peak-peak. The difference seen in comparison to the 10 minute estimation was caused by the low frequency drift.

The performance improvement of the calibration switching with respect to the raw signals was evaluated by computing the STD and peak-peak ratios. Obtained ratios represent how much the position noise decreases when the calibration switching is used demonstrating the performance improvement. The improvement factors obtained from STD ratios of all the measurement configurations presented in this chapter are summarized in Table 5-15. The best performance improvement of factor 11 was measured with the low repetition rates and low RF gain setting. The measurement setup was similar to the LHC pilot beam with a low intensity bunch. The smallest performance gain was measured with the low RF gain setting and high pulse repetition. Such conditions are typically used during LHC physics fills with many high intensity bunches in the machine.

Table 5-14. Calibrated orbit signals measured with 50 dB and 10 dB gains and 12 kHz pulse repetition rate.

Pulse repetition frequency	RF Gain Configuration	Gain value									
		10 dB					50 dB				
		Avg. cal.pos. [μm] *	STD [μm_{RMS}] *		Peak-peak [μm] *		Avg. cal.pos. [μm] *	STD [μm_{RMS}] *		Peak-peak [μm] *	
			1 min	10 min	1 min	10 min		1 min	10 min	1 min	10 min
12 kHz	Fine	22.67	0.11	0.16	0.53	1.03	21.87	0.52	0.54	2.59	3.45
	Coarse	20.76	0.05	0.07	0.33	0.42	21.28	0.49	0.54	2.47	3.58
40 MHz	Fine	33.04	0.06	0.06	0.22	0.4	33.69	0.06	0.06	0.29	0.42
	Coarse	32.44	0.01	0.02	0.06	0.11	32.38	0.04	0.05	0.18	0.3

* Rounded to 2 decimal places.

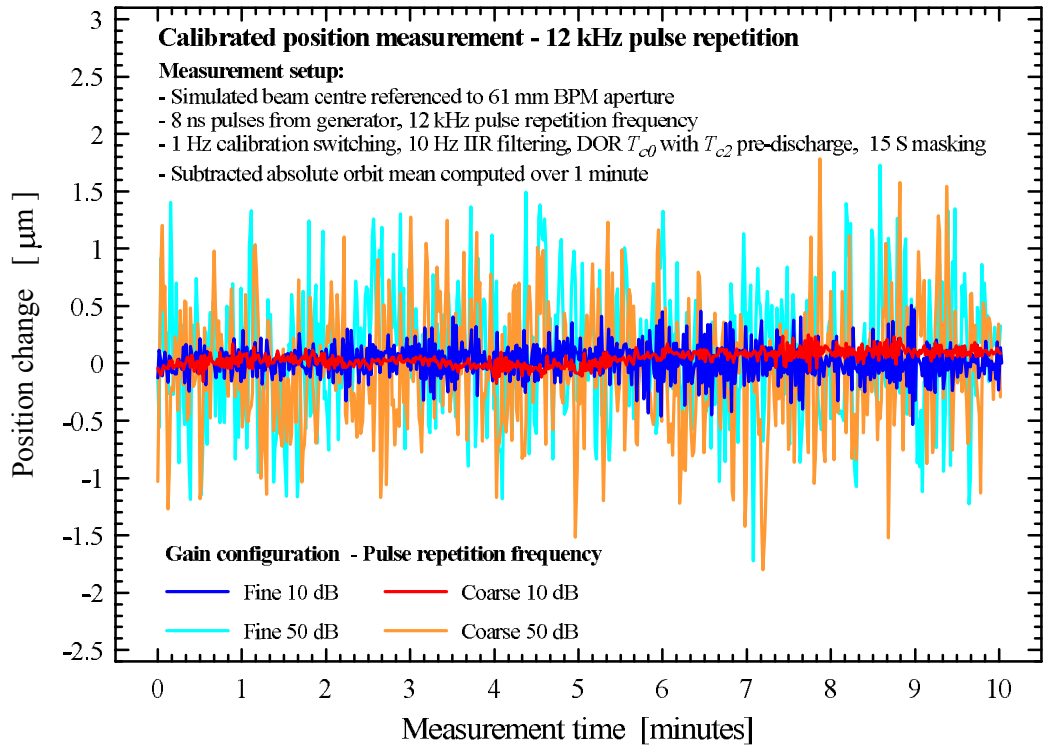


Fig. 5-51. Measurement of the relative change of the calibrated position signal with 12 kHz pulse repetition. The generator Keysight 33600A was set to provide 8 ns pulses with 3 Vpp amplitude. The signal were measured with 10 dB and 50 dB RF gains, each configured with both fine and coarse gain schemes. The signal with was split to 8 front-end inputs to simulate centred beam. The DOR processing was configured with 1 Hz calibration switching, detector time constant T_{c0} with T_{c2} pre-discharging, 15 sample masking and 10 Hz low-pass IIR filter cut-off.

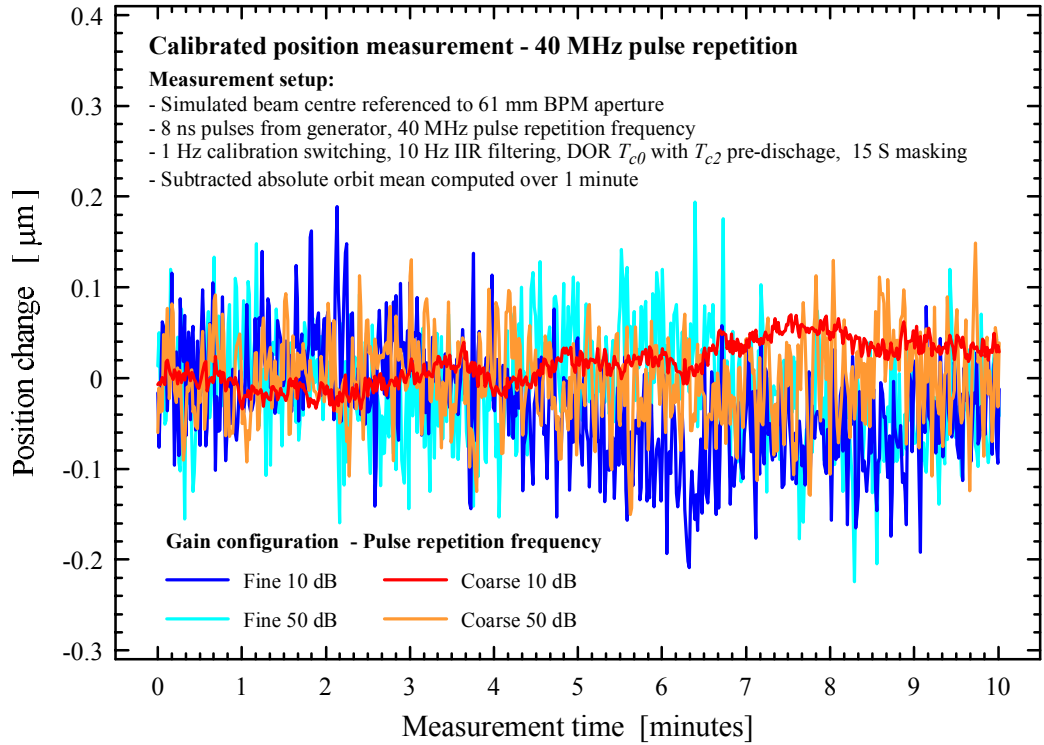


Fig. 5-52. Measurement of the relative change of the calibrated position signal with 40 MHz pulse repetition. The generator Keysight 33600A was set to provide 8 ns pulses with 3 Vpp amplitude. The signal were measured with 10 dB and 50 dB RF gains, each configured with both fine and coarse gain schemes. The signal with was split to 8 front-end inputs to simulate centred beam. The DOR processing was configured with 1 Hz calibration switching, detector time constant T_{c0} with T_{c2} pre-discharging, 15 sample masking and 10 Hz low-pass IIR filter cut-off.

Table 5-15. Calibration improvement factors.

Pulse repetition frequency	RF Gain Configuration	Calibration switching improvement factor *							
		STD Raw Str. / Calib.				Peak-peak Raw Str. / Calib.			
		10 dB		50 dB		10 dB		50 dB	
		1 min	10 min	1 min	10 min	1 min	10 min	1 min	10 min
12 kHz	Fine	2.9	8.4	2.1	2.9	2.8	5.8	2.1	3.3
	Coarse	5.6	11.2	1.9	2.2	3.4	8	1.6	2
40 MHz	Fine	1.7	2.8	2.2	2.7	1.9	2.3	2.2	2.1
	Coarse	2.2	7.5	2.1	3.6	1.7	4.7	1.9	3

* Rounded to 1 decimal place.

5.4.3.3 Summary and outlook

The measurements presented in this chapter demonstrated the performance of the calibration switching algorithm in comparison to the raw position signals. The laboratory setup simulated the zero beam position offsets in the BPM. Both the calibrated and raw positions were computed from the same data to assure equal conditions for comparisons. The analysis was focused on evaluation and comparisons of the systematic errors and precision of the position signals measured with different RF gain settings and pulse repetition rates.

The raw position signals demonstrate the measurement accuracy achieved without the calibration switching. The systematic offset errors of the raw signals were measured to be around $\pm 150 \mu\text{m}$ when using the integrated programmable gain amplifier in the RF stage. Using the coarse gain scheme instead decreased the systematic errors to around $\pm 50 \mu\text{m}$. These errors change with each RF gain setup. The calibration switching reduced the offset errors to some $20 \mu\text{m}$ when measuring the 12 kHz pulses and to $30 \mu\text{m}$ when measuring 40 MHz pulses. The results can be translated into asymmetry between the DOR signals in the order of 0.1 %. Such asymmetry can be explained by the input filters or the cables and splitters of the measurement setup which are not calibrated by the switching. Achieved results were independent of the RF gain setup, which is important for achieving consistent accuracy over the whole dynamic range. Further improvements of the measurement would require another set of multiplexer switches placed close to the BPM signal source (such as the signal generator in the laboratory). Such setup could potentially allow to distinguish between the offset errors coming from the cables from the asymmetry in DC offsets from the low-frequency part of the DOR channels.

The precision of the measurement was estimated using the standard deviations and peak-peak values. Without the calibration switching the low frequency noise and drifts in the electronics over 10 minutes of the measurement could reach some $10 \mu\text{m}$ peak-peak and about $1.5 \mu\text{m}_{\text{RMS}}$. Most of the raw position measurements were dominated by low frequency drifts which would be difficult to remove by a simple filtering. The calibration algorithm reduced the drifts in the position signals by more than factor 2. The most significant improvements up to factor 8 were achieved when using the low gain settings. Such measurement conditions are typically used when measuring the LHC physics beams consisting of many high intensity bunches. The standard deviation of the noise level was estimated to less than $0.1 \mu\text{m}_{\text{RMS}}$.

It was also observed that the precision of the calibrated signals was better when bypassing the IPGA in the RF chain. The penalty for using the coarse gain scheme is reduced gain control step to 2.5 dB. However, the differences between the two gain schemes become less significant when using higher RF gains. Such conditions are typically used when measuring the LHC pilot beam consisting of a single low intensity bunch.

5.5 Sensitivity to induced temperature change

The gain and offset asymmetries between two DOR channels depend on the precision and spread of the component values used in the analogue part. The components in the channels are also subject to gradients and their slow fluctuations of the temperature inside the front-end. This dependence also causes drifts and systematic errors in the measured positions. The periodic switching of the channels minimizes such drifts by following the change in the asymmetry. The sensitivity of the DOR processing to the induced temperature changes is presented in this chapter. The measurements were performed by exposing a front-end unit to the air flow from an external ventilation unit. Measurement of both the raw and calibrated positions demonstrate the impact of the calibration switching on the measurement precision. The positions were also measured with opened front-end exposing the DOR channels to the natural air convection as well as forced air flow. The aim of these measurements was to measure the temperature sensitivity to conditions typically found in systems with direct airflow cooling.

5.5.1 Measurement setup

The laboratory setup described in chapter 5.4.2.1 was also used in the following measurements. The setup is based on the signal generator Keysight 33600A used to simulate BPM signals with 8 ns pulse width and 40 MHz repetition simulating the typical LHC physics beams. The signal was split to the eight front-end inputs to simulate zero position offset in the BPM. The generator drifts had thus minimal influence on the position signals. The front-end RF stage was configured to use the integrated programmable gain amplifier set to 10 dB gain and the standard AUCAL settings described in chapter 5.4.2.1. The temperature change was induced using the setup shown in Fig. 5-53. The setup consisted of two ventilator units located in rack slots above and below the front-end box. The ventilation unit on the bottom is necessary for regular operation of the front-end. It provided constant airflow during entire time of the measurement. The ventilation unit located above was used to generate the changes by remotely switching its fans on and off. Dedicated temperature sensors inside the front-end allow continuous monitoring of the average temperatures on each PCB. The readings from all temperature are acquired at 1 Hz rate and transmitted in the UDP payload for online monitoring and diagnostics. Accuracy of the temperature sensor is ± 0.4 °C with resolution of some 0.008 °C.



Fig. 5-53. Photograph of the laboratory setup used to measure the sensitivity of the DOR subsystem to the temperature change induced by the ventilation unit located above the front-end. This unit was remotely controlled with the front-end optocouplers. The ventilation unit located on the bottom provided constant airflow necessary for regular operation of the front-end.

5.5.2 Sensitivity measurements

The plot in Fig. 5-54 displays the average temperature measured on the analogue board. The data from the DOR channels located on the same analogue board was acquired at the same time in parallel to the temperature readings. The raw electrode signals from the DOR channel 1 and 2 are showed in Fig. 5-55. They were computed in the post-processing algorithm which separates the data belonging to the straight and crossed multiplexer states and removes the masked samples of each switching transient. The vertical axis on the left was normalised to the ADC full-scale. The axis shown on the right of the plot was scaled to represent the voltage levels present on the ADC inputs.

The first part of the measurement depicts 30 minutes of the steady-state temperature conditions in the front-end. The measurement started when the front-end reached operating temperatures after several hours of operation. The steady-state temperature was measured around 45 °C. The peak-peak temperature change during the 30 minutes was measured to about 0.2 °C. The change in DOR signals was measured in the order of 200 ppm of the ADC full-scale, which is equivalent to some 1 mV change over 5 V ADC dynamic range. The front-end cover was removed after the 30 minutes of the measurement exposing the electronics to the laboratory environment. The temperature decreased by some 9 °C. As a response to the temperature drop the raw DOR signals changed by more than 2.5 % which is equivalent to almost 70 mV.

After 60 minutes of the measurement the ventilation unit located above the front-end was switched on to force airflow over the exposed electronics. The board temperature decreased by additional 5 °C. The overall temperature changed by 15 °C since the start of the measurement. Please note that such temperature change is large in comparison to a few degree fluctuations measured in the DOROS front-ends installed in the LHC. Corresponding change in the raw DOR signals was measured about 1.7 % which is equivalent to about 50 mV. Absolute and relative DOR signal change was measured for each induced temperature change. Obtained results were summarized in Table 5-16. Please note that the generator used to generate the input signals was placed away from the test setup and thus was not affected by the induced change.

The raw signals were then used to compute raw positions shown in Fig. 5-56. Displayed raw position signals for straight (blue) and crossed (cyan) multiplexer state would be measured if the calibration switching was not used. The position change was referenced to the beginning of the displayed measurement. The values were scaled to a 61 mm BPM aperture using equation (3-2). The change in the calibrated position signals (red) was computed from the same measurement data and shown in the same plot for comparison. Absolute positions during the stable temperature conditions of the measurement were measured around 216 μm for the straight signal and $-147 \mu\text{m}$ for the crossed signal. The calibration switching reduced the position offset to around 35 μm .

Table 5-16. Absolute and relative change of the raw electrode signals as a response to the temperature change.

Measurement time [minutes]	Temp. change [°C]*	DOR Ch1 / Ch2 Raw signal change							
		Channel 1				Channel 2			
		Str.		Crs.		Str.		Crs.	
		Abs. [mV] *	Rel. [%]**	Abs. [mV] *	Rel. [%]**	Abs. [mV] *	Rel. [%]**	Abs. [mV] *	Rel. [%]**
Stabilised (0 - 30 minute)	0.2	0.9	0.02	0.9	0.02	1.0	0.02	0.9	0.02
FE opened (30 - 60 minute)	9.3	68.7	2.68	71.1	2.84	70.6	2.83	68.2	2.68
FE opened + fan (60 - 75 minute)	5.8	46.2	1.77	48.2	1.89	47.8	1.88	45.6	1.76
Entire meas. (0 - 75 minute)	15	115.8	4.47	120.2	4.75	119.4	4.73	114.7	4.46

* Rounded to 1 decimal place, ** Rounded to 2 decimal places.

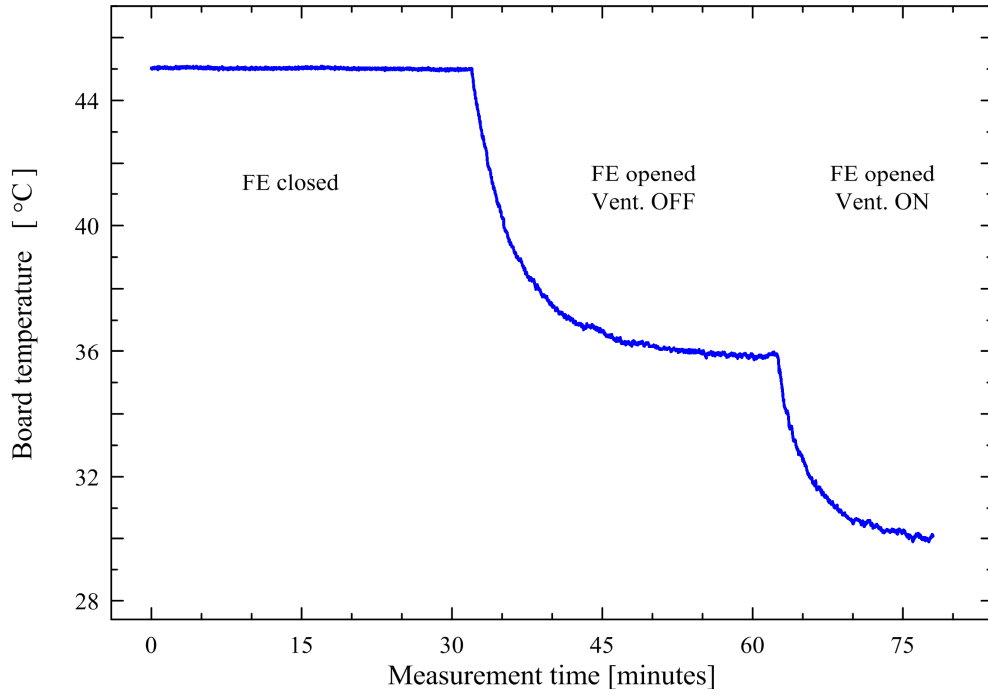


Fig. 5-54. Average temperature measured on the analogue card inside the DOROS front-end (FE). First 30 minutes were measured during the stable operation of the FE. The first temperature change was induced in 30th minute by exposing the electronic components in the front-end to natural air convection in the laboratory. The second temperature change was induced in 60th minute by turning on the ventilation unit located above the opened front-end.

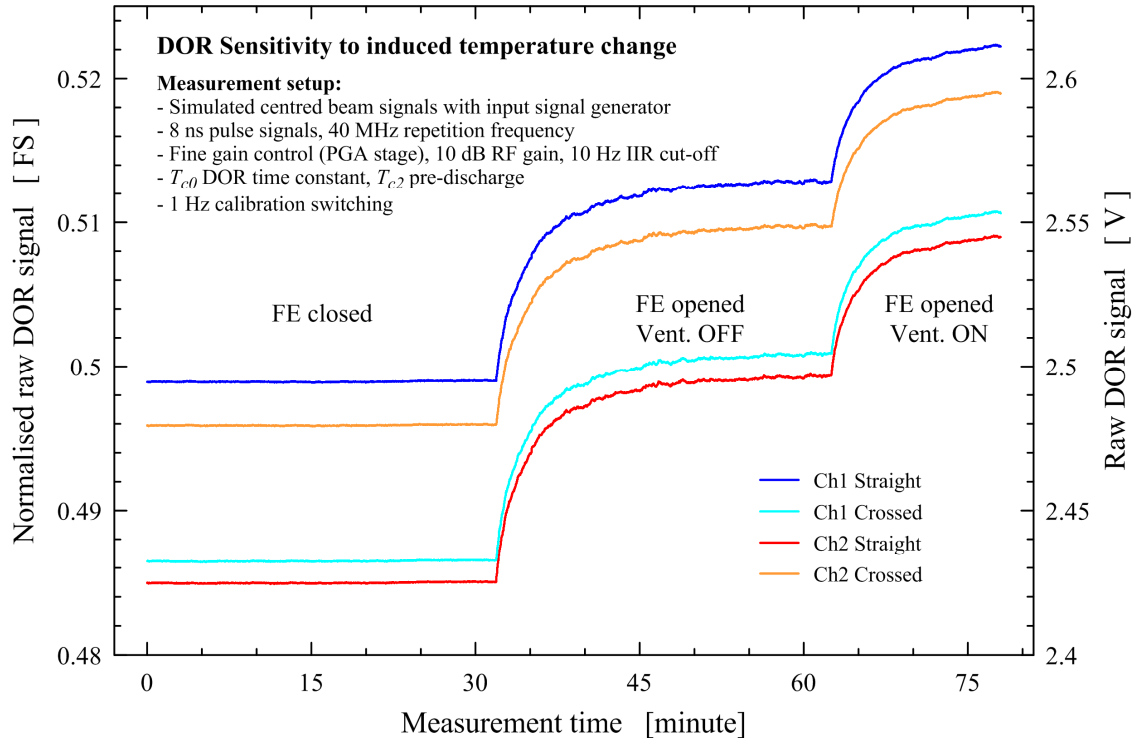


Fig. 5-55. Measured raw DOR signals during induced temperature changes. The straight and crossed raw signals were obtained after de-multiplexing the switched DOR signals. The first temperature change corresponds to opening the front-end cover. The second temperature change corresponds to turning on the ventilation unit located above the electronics. Signal generator Keysight 33600A provided 8 ns pulses with 3 V_{pp} amplitude and 40 MHz repetition simulating the LHC physics beam signals. The generator signal was split and provided to eight front-end inputs. The RF chain was set to use the integrated programmable amplifier with 10 dB gain. The DOR channels were configured with 1 Hz calibration switching, 10 Hz low-pass filtering in the IIRs, T_{c0} time constant with T_{c2} pre-discharging and 5 sample masking.

Table 5-17 summarizes the changes of the raw and calibrated position signals computed for each temperature change throughout the measurement. It can be seen that the raw positions drifted by more than 20 μm due to the temperature change in the front-end. The calibration switching allowed to decrease this error to some 1.5 μm , yielding improvement of more one order of magnitude. The plot in Fig. 5-57 depicts the calibrated position signal in greater detail. Please note that this measurement was acquired with only a single RF amplifier stage.

The position noise was estimated by computing peak-peak upon stable regions of each measurement stage. Computed values are summarized in Table 5-18. The peak-peak noise of the raw position signals was estimated to about 1.4 μm during the stable temperature. The calibration switching reduced this noise to some 0.4 μm , improving the measurement precision by more than factor 3. It can be seen that the noise levels of both the raw and calibrated orbits increased after opening the front-end approximately by factor 2. After switching on the ventilation unit the noise level of both raw and calibrated orbits decreased to the level when it is comparable to the values measured during stable conditions. This effect was caused by the air movement forced by the fans which as parallel to the DOR channels, minimising localized temperature differences between the two channels. The calibration switching in this case improved the precision by factor 2.

Estimated temperature sensitivity of the raw and calibrated position measurements is summarized in Table 5-19. As seen in the table, the sensitivity of position measurements to temperature changes without using the calibration switching was estimated to approximately 1 $\mu\text{m}/^{\circ}\text{C}$. The calibration switching allowed to decrease the sensitivity to some 0.1 $\mu\text{m}/^{\circ}\text{C}$, improving the performance by one order of magnitude. This is equivalent to a temperature dependent asymmetry between two input signals in the order of 10 ppm/ $^{\circ}\text{C}$.

Table 5-17. Position errors induced by temperature changes.

Measurement time span [minutes]	Temperature change [$^{\circ}\text{C}$] *	Position change **		
		Raw [μm]		Calib. [μm]
		Str.	Crs.	
Stabilised (0 - 30 min.)	0.2	0.02	0.21	0.06
FE opened (30 - 60 min.)	9.3	11.65	12.86	0.73
FE opened + fans (60 - 75 min.)	5.8	8.54	10.26	0.67
Entire meas. (0 - 75 min.)	15.3	20.21	23.33	1.46

* Rounded to 1 decimal place, ** Rounded to 2 decimal places.

Table 5-18. Estimated peak-peak position noise during induced temperature changes.

Measurement time span [minutes]	Temperature peak-peak [$^{\circ}\text{C}$] *	Position peak-peak **		
		Raw [μm]		Calib. [μm]
		Str.	Crs.	
Stabilised (0 - 30 min.)	0.2	1.45	1.44	0.36
FE opened (55 - 60 min.)	0.3	2.59	2.82	0.72
FE opened + fans (73 - 78 min.)	0.4	1.18	1.03	0.49

* Rounded to 1 decimal place, ** Rounded to 2 decimal places.

Table 5-19. Normalised sensitivity of the position measurement to the temperature change.

Measurement time span [minutes]	Normalised temperature sensitivity		
	Raw [$\mu\text{m} / ^{\circ}\text{C}$] *		Calib. [$\mu\text{m} / ^{\circ}\text{C}$] *
	Str.	Crs.	
FE opened (30 - 60)	1.27	1.4	0.08
FE opened + fans (60 - 75)	1.53	1.84	0.12

* Rounded to 2 decimal places.

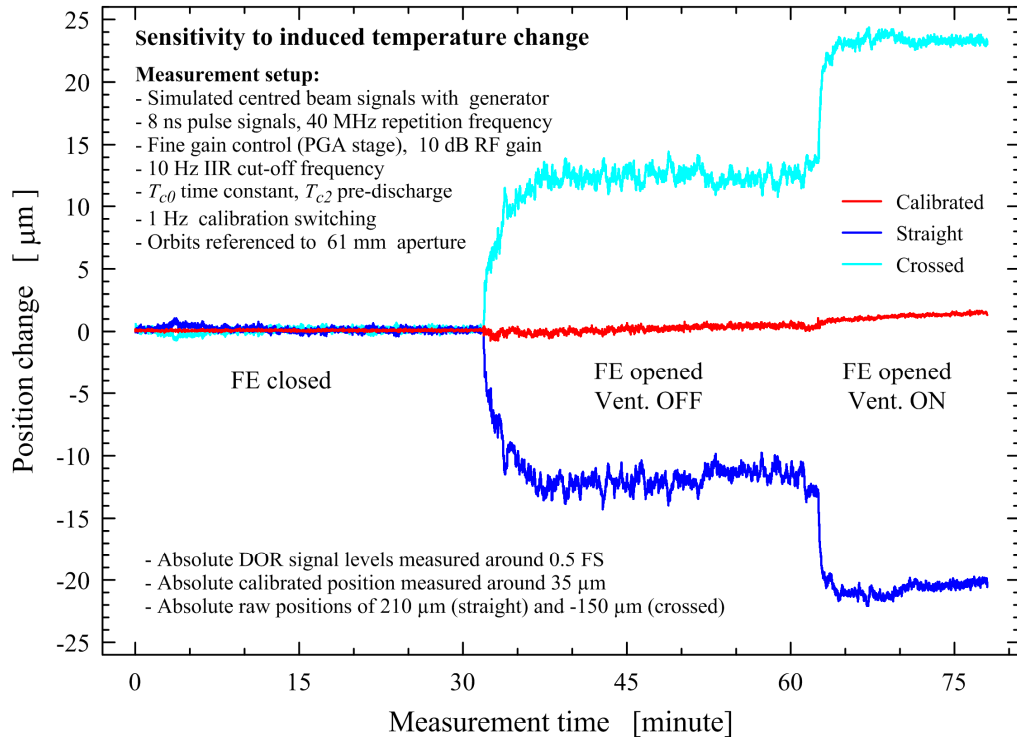


Fig. 5-56. Sensitivity of the orbit measurement to induced temperature changes in the front-end. The straight and crossed position signals were computed from the de-multiplexed DOR signals. The calibrated orbit signal was calculated from the same data. The first temperature change corresponds to opening the front-end cover. The second temperature change corresponds to turning on the ventilation unit located above the electronics. Signal generator Keysight 33600A provided 8 ns pulses with 3 Vpp amplitude and 40 MHz repetition. The RF chain was configured to use only the integrated programmable amplifier with 10 dB gain. The DOR channels were configured with 1 Hz calibration switching, 10 Hz low-pass filtering, T_{c0} time constant with T_{c2} pre-discharging and 5 sample masking. Computed positions were referenced to 61 mm BPM aperture.

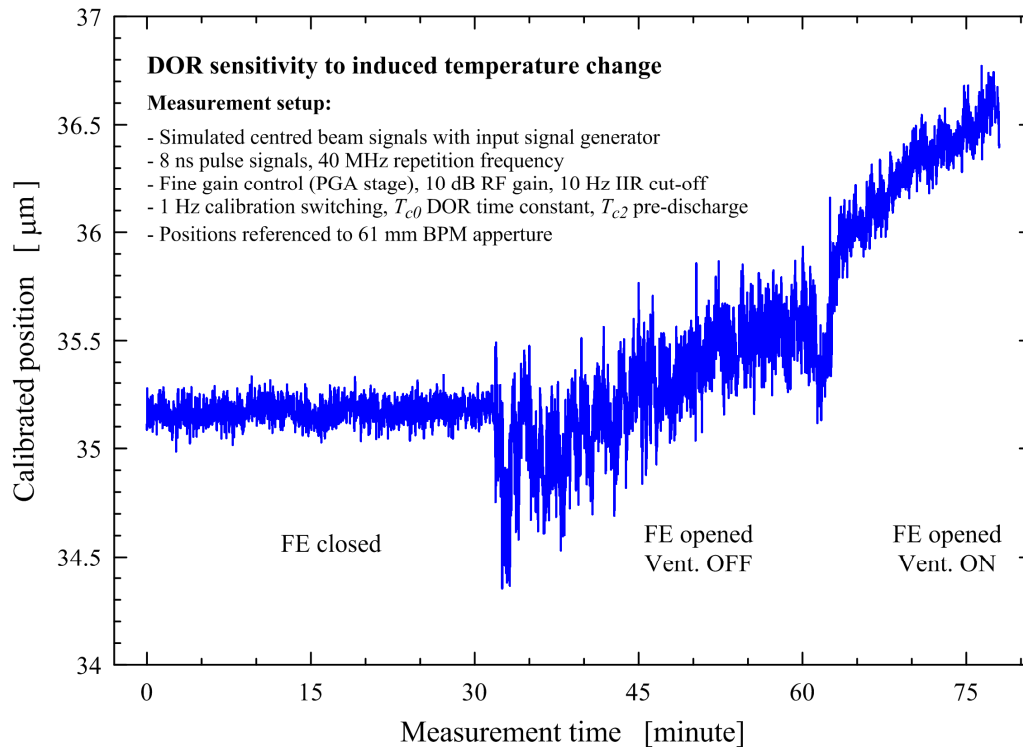


Fig. 5-57. Sensitivity of the calibrated orbit measurement to induced temperature changes in the front-end. The signal change was computed by subtracting the first samples of the measurements. Displayed positions were referenced to 61 mm aperture. The first temperature change corresponds to opening the front-end cover. The second temperature change corresponds to turning on the ventilation unit located above the electronics.

5.5.3 *Summary and outlook*

The results of the laboratory measurement presented in this chapter evaluate the sensitivity of the calibrated position signals to the induced temperature change. The signal changes were induced by opening the front-end unit and exposing its electronics to temperature fluctuations. The opened front-end was then exposed to the air flow forced by an external ventilation unit located above the electronics. The average temperature measured on the analogue board decreased during the measurement by 15 °C.

The measurements showed that without the calibration switching the position may change by more than 20 μm . Observed change was approximately 10 % of the measured position offset error. The temperature sensitivity was estimated to 1 $\mu\text{m}/^\circ\text{C}$. The calibration switching at 1 Hz rate allowed to reduce the signal change to around 1 μm . The sensitivity was reduced to 0.1 $\mu\text{m}/^\circ\text{C}$, improving the measurement performance by one order of magnitude.

The noise of the position signals was estimated by computing the peak-peak values in each stage of the measurement. It was observed that the noise of the raw positions increased from 1.4 μm to almost 3 μm when opening the front-end unit. It was also observed that the noise decreased to about 1 μm when the channel pair was exposed to the parallel airflow from the ventilation unit. Similar phenomenon was also seen in the computed peak-peak noise of the calibrated position signal.

Minimising the thermal gradients between the channels helps to decrease the dependency of the position measurements on the temperature change. Placing the electronics inside systems with a direct airflow cooling may result in significant performance degradation, especially when the motor speed of the fans is controlled in a feedback loop.

6 Orbit subsystem results based on beam measurements

6.1 Collimator BPM system

The first DOROS front-ends were installed during the first LHC long shutdown (LS1) to measure beam orbits in 18 LHC collimators with the jaw-embedded BPM sensors. The system was first time commissioned with beams in the beginning of 2015 and was used operationally since then. Measured beam positions from the system are primarily used for precise alignments of the collimator jaws. A so called BPM-based alignment allowed to significantly reduce the time overhead spent on the collimator setup. This chapter presents the achieved performance of the system which is demonstrated on a number of measurements acquired the LHC beams.

6.1.1 BPM signal measurement

Picture in Fig. 6-1 displays an example of the electrode signals from the upstream and downstream collimator BPMs measured by a 12-bit oscilloscope HDO6104. The measurement was acquired during the LHC physics operation at the top energy of 6.5 TeV and some 2000 bunches. The bunch spacing of the measured beam was 25 ns which is equivalent to a distance of some 7.5 m on the LHC circumference. The signal was induced on the capacitive electrodes and sent through a 200 m long coaxial cables to 3 dB attenuators placed on input of the oscilloscope. The channels C1 and C2 were measuring the pulses on the upstream BPM where the beam enters the collimator. The channels C3 and C4 were connected to the downstream BPM where the beam exits the collimator. As seen in the plot, the amplitude of the pulses was measured around 3 V. By comparing the amplitudes of the individual pulses it can be seen that the beam is passing close to the centre of each BPM. The time difference seen between the upstream and downstream BPM was measured around 4 ns. As the LHC beams propagate very close to speed of light the distance between the two collimator BPMs can be estimated to 1.2 m.

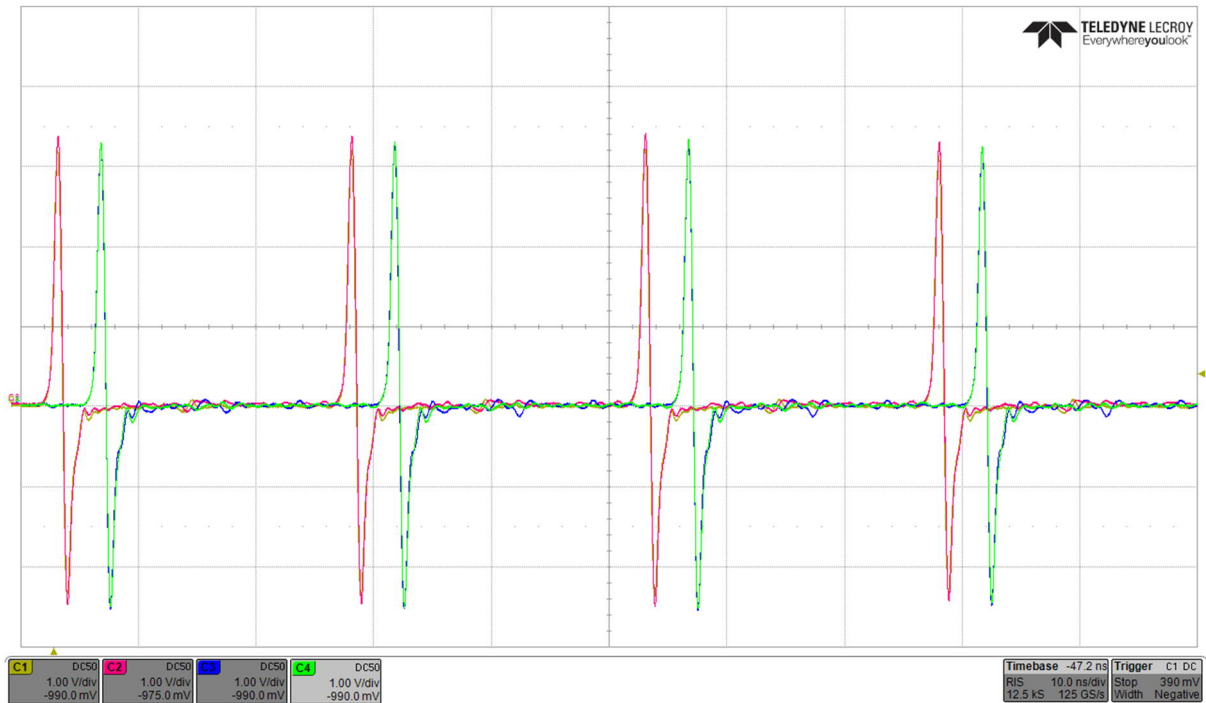


Fig. 6-1. Oscilloscope measurement of BPM signals from a TCTP collimator located in LHC point 5 (CMS). The BPM signals were connected through 3 dB attenuators to a 12-bit oscilloscope HDO6104 with 2.5 GS/s sampling and 1 GHz signal bandwidth. The C1 and C2 channels were measuring the BPM electrodes on the upstream side of the collimator. The C3 and C4 channels were measuring the downstream BPM electrodes. The plot displays 4 consecutive bunches of a regular LHC physics beam at nominal beam intensity. The vertical scale was set to 1 V/div and the horizontal scale to 10 ns/div.

6.1.2 Collimator alignment

Picture in Fig. 6-2 shows example of the BPM signals acquired during an automatized BPM-based alignment of a TCSP.A4R6.B1 collimator using the DOROS system. The picture on the right sketches the two jaws of the collimator around the beam (red). The BPMs are placed in the left (L) and right (R) corners of the upstream (U) and downstream (D) side of the collimator jaws (grey). The first two plots on the top show the calibrated electrode signals and the corresponding calibrated positions measured by the DOROS. The absolute position of the individual jaws shown in the two plots below was measured by the Linear Variable Differential Transformers (LVDT) in the collimator [44].

The alignment algorithm moves the left and right jaws such that the electrode signals are equalised. The right electrodes are moved towards the beam and the left electrodes are moved away from the beam keeping the gap between the jaws constant. As seen in the plot the jaws are first aligned with the beam at the upstream collimator port. After some 5 seconds the alignment continues by aligning the jaws at the downstream port. The BPM-based alignment shown in the Fig. 6-2 was completed in less than 30 s which is by more than two orders of magnitude improvement when compared to the BLM method. The position difference between the two collimator ports seen at the end of the alignment was found to be a tilt error caused by misaligned collimator tank with respect to the beam axis [74].

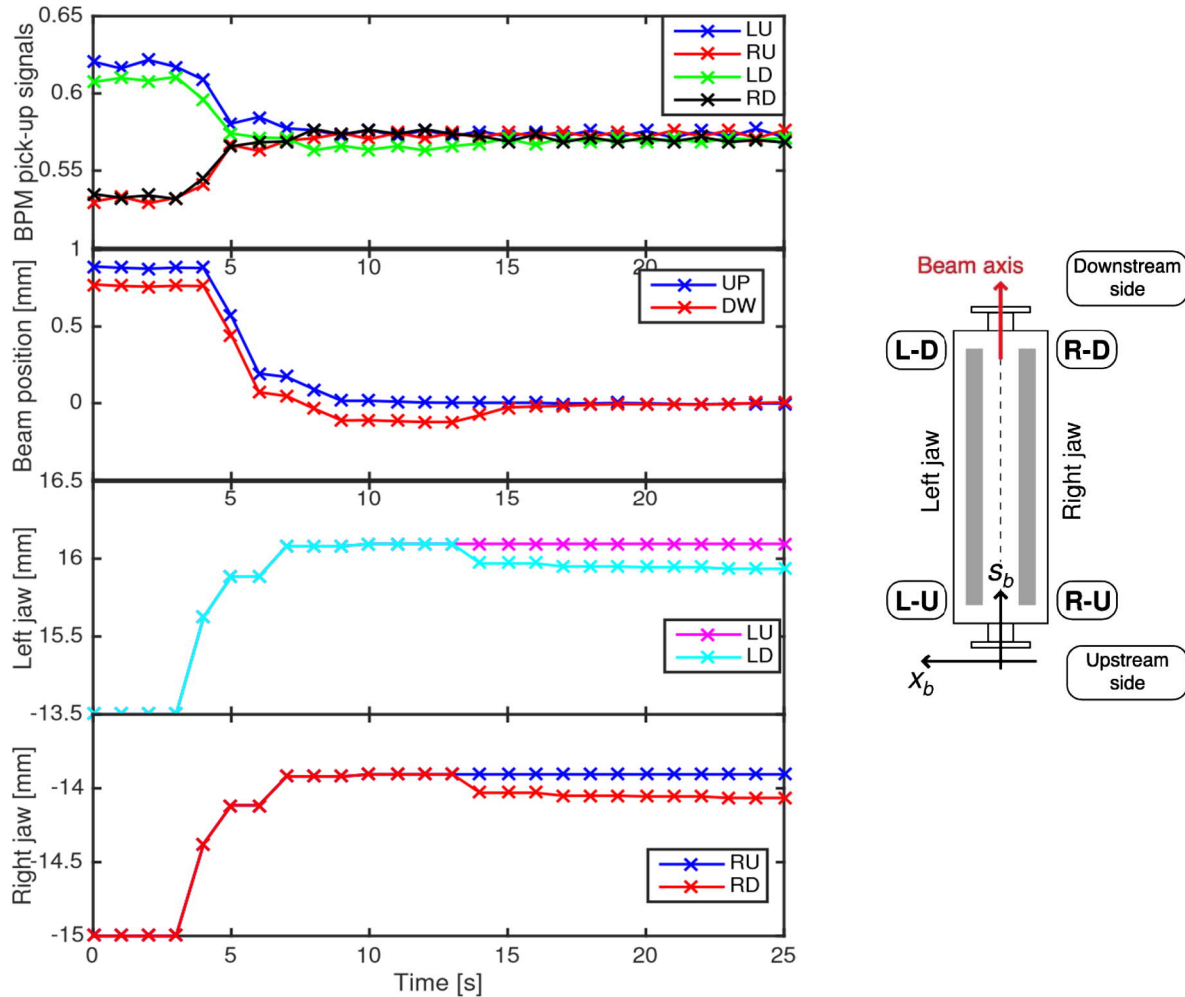


Fig. 6-2. Example of the collimator alignment based on the beam position measurement technique. The position was measured at the upstream (UP) and downstream (DW) collimator ports with BPMs embedded in the jaws. The BPM electrodes located in left-up (LU), left-down (LD), right-up (RU) and right-down (RD) jaw corners were connected to a DOROS front-end. The top plot shows the calibrated electrode signals normalised to the ADC full-scale. Below is the plot showing corresponding beam positions at the two collimator ports. The two plots on the bottom show the absolute position of each collimator corner obtained from linear variable differential transformers (LVDT) in the collimator [44],[74].

6.1.3 Resolution and long-term stability

The collimator BPMs were used to estimate the resolution and long-term stability of the orbit subsystem based on the beam measurements. As the beam trajectory is not affected when passing through 1 m long collimator, the difference between beam orbits measured in the upstream and downstream ports can be considered negligible. Subtracting the beam positions acquired at the upstream and downstream BPMs removes the coherent part of the measurement, mostly the position change and drifts of the beam. The remaining signal is proportional to the incoherent noise and error drifts of the measurement system. The method allows to evaluate the resolution and stability of the measurement system assuming that the angle at which the beam enters the collimator does not change over time.

The plot in Fig. 6-3 displays the calibrated orbit signals in the upstream (dark blue) and downstream (light blue) collimator BPM measured during 11 hours of stable beams in the LHC. The collimator measured the vertical plane of the beam 2 on the right side of the ATLAS experiment (IP1). The beam consisted of some 1000 nominal bunches at energy of 6 TeV. The electrode signals from the collimator BPMs were processed by the DOROS front-end using four channels on a single analogue card. The calibrated signals were computed in the DOROS server and stored in the logging database.

The absolute position of the beam with respect to the centre of each BPM was approximately 0.9 mm. Position difference measured between the upstream and downstream ports was around 5 μm . The positions in the beginning of the measurement were subtracted from the respective signals to obtain the orbit change. The corresponding axis with 10 μm span is shown on the left of the plot. As can be seen the beam orbits measured on the two ports are strongly correlated. The position change seen at 5th and 9th hour corresponds to a so called Van der Meer scan performed on the beam at that time [75].

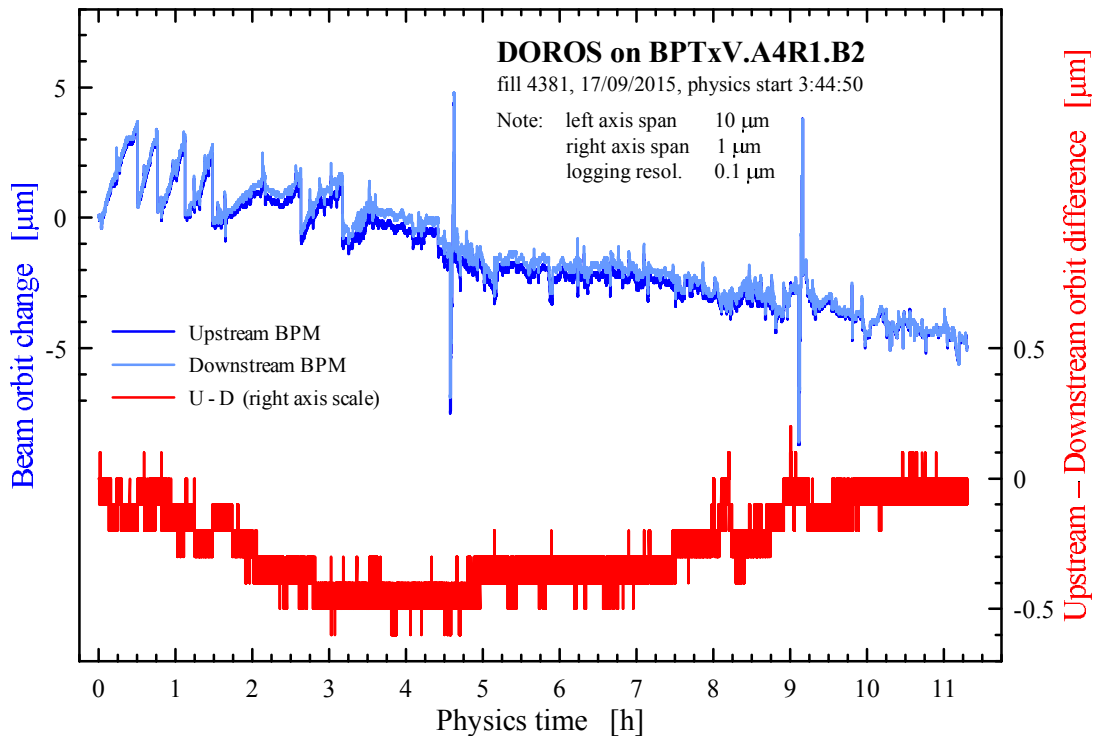


Fig. 6-3. Calibrated orbit signals measured in the collimator BPMs during 11 hours of LHC stable beam with 1000 nominal bunches at 6 TeV energy. Displayed orbit change was measured in the upstream (dark blue) and downstream (light blue) of the vertical collimator in beam 2. The corresponding vertical axis is shown on the left side of the plot. Difference between the two orbits (red) removes the coherent beam position part of the signal and leaves the drifts and incoherent noise of the DOROS front-end. The corresponding vertical axis is shown on the right side of the plot. Displayed data was obtained from logging database with resolution limited to 0.1 μm . The DOROS front-end was configured with time constant T_{c0} with T_{c2} pre-discharge, 5 S masking 10 Hz low-pass IIR filtering cut-off and 1 Hz calibration switching.

The orbit difference between the upstream and downstream collimator ports is also showed in Fig. 6-3 for comparison (red). The corresponding axis is located on the right of the plot. As can be seen, the orbit difference measured over 11 hours did not drift more than 0.5 μm . This can be also expressed as change of the beam angle estimated to some 10^{-5} degrees over 1 m distance between the two collimator BPMs. The steps seen on the difference signal are caused by rounding the computed positions by the DOROS server to 0.1 μm before they are stored in the database. Presented results show that the resolution of the calibrated orbit measurements is better than 0.1 μm . Contribution of the system to the noise and drifts of the orbit measurement is smaller than the position jitter of the beam.

6.1.4 Summary and outlook

This chapter presented example measurements demonstrating the BPM-based alignment of the collimator jaws with beam. The BPM-based method allowed to speed up the collimator alignment by more than two orders of magnitude when compared to the BLM-based method. The BPM readings were also used to estimate the resolution and long-term stability of the DOROS orbit measurements with the LHC beams. The results showed that the DOROS system can offer orbit resolution better than 0.1 μm which is the current limit in the data logging. The orbit measurements were dominated by the position jitter of the beam.

Currently, the DOROS collimator system operates on more than 20 LHC collimators with embedded BPMs. The system was used throughout the LHC runs in 2015, 2016 and 2017 for precise beam-based alignments as well as for several beam studies [29], [30], [74] and [76]. One of the studies was focused on the fill-to-fill reproducibility of the measurements during standard LHC operation [30]. Based on the achieved performance of the system it was decided to use the calibrated orbit readings from the collimator DOROS system for the LHC machine protection. The LHC interlock system automatically terminates operation with beams if the beam positions measured by the DOROS exceed safe limits. This protection mechanism was commissioned and put in operation during the LHC run in 2017. The interlocked operation in late 2017 was without a single false beam dump which would stop the LHC operation for hours.

Other studies involving the collimator DOROS system are focusing on the so called wire collimators and their precise alignment for compensating the beam-beam effects during the collisions in the experiments [77]. High resolution of the system is important for recent studies using the collimator BPMs to measure quadrupolar term in the beam positions and compute the related beam size.

6.2 Standard DOROS system

The DOROS system was also installed on a few selected BPMs in the LHC. There the so called “standard” DOROS system measures orbits and oscillations of the beams. Both hardware and firmware are identical to the collimator system. Most of the DOROS front-ends were installed on the most important BPMs around interaction points. In the beginning of 2015 four front-ends were installed to equip the so called Q1 BPMs located on the left and right side of the ATLAS experiment in interaction point (IP1) and the CMS experiment located in IP5. The Q1 BPMs measure the two beams merged in a common beam pipe. The remaining Q1 BPMs around the ALICE (IP2) and LHCb (IP8) along with a few more selected BPMs in the LHC were equipped later in 2015 and 2016. In 2018 the installation of the LHC standard DOROS system consists of about 20 front-ends. The most important front-ends were installed in parallel to the standard LHC BPM system which is based on the wide-band time normalisation technique (WBTN). The picture in Fig. 6-4 displays one such installation located close to the CMS experiment. The electrode signals from the BPM were split between the systems using paired reactive splitters. They have bandwidth up to 2 GHz and provide some 30 dB insulation between the ports. Attenuators were placed on the inputs of the splitters to limit the power sent to the measurement systems. The attenuation value in this installation was 10 dB. The same scheme was used for other front-end installations of the standard DOROS front-ends in the LHC. Such measurement setup also allows to compare the performance of the two BPM systems with the same beam signals.

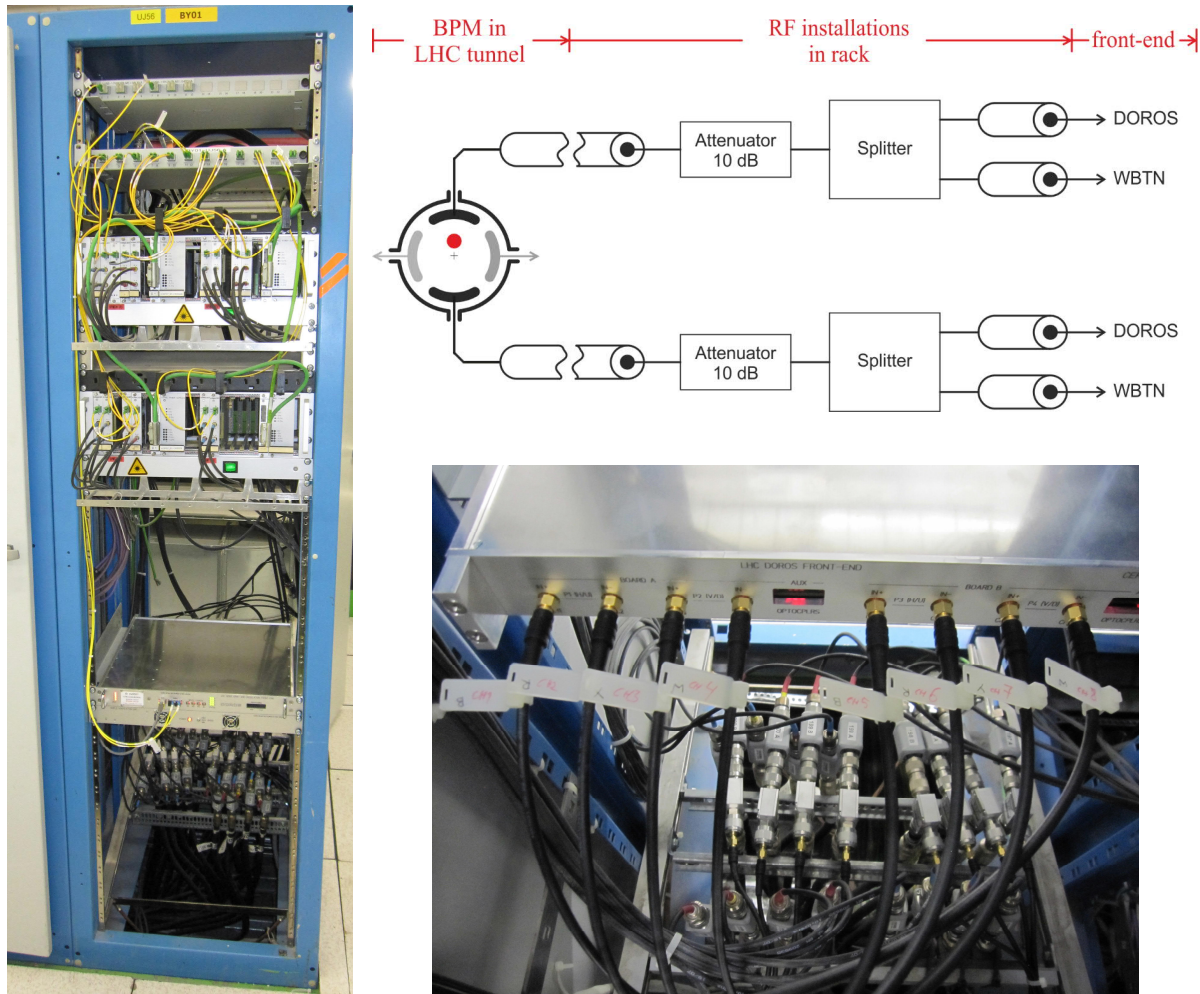


Fig. 6-4. DOROS front-end installation close to the CMS experiment in the LHC point 5. As described on the top left picture, the BPM electrode signals are split between the DOROS and the Wide-Band Time Normaliser (WBTN) systems measuring the beam positions. The photograph on the right bottom shows the rear view of the RF splitter and cabling installation in the rack.

6.2.1 Orbit measurements during Van der Meer scans

The standard DOROS system is used for a number of beam measurements and studies which target performance and reliability improvements of the LHC. The so called Van der Meer scans (VdM) are important for optimising the collision rates in the experiments. The VdM scans are based on scanning the transverse positions of the two counter-circulating beams and measuring the relative collision rates in an interaction region [75]. The beam positions are scanned in steps of a few micrometres to find the optimal transverse overlap of the two beams. The typical transverse size of a beam during the collisions is in order of $20 \mu\text{m}_{\text{RMS}}$ [3]. To minimise uncertainties in the VdM scans the beam orbits have to be measured with micrometre resolution, long-term stability and reproducibility.

Plot in Fig. 6-5 shows an example of an orbit measurements of the beam 1 (blue) and beam 2 (red) obtained during a VdM scan in the ATLAS interaction region (IP1). The position readings shown in the plot were acquired with both the DOROS (dark colour) and WBTN (light colour) connected to the Q1 stripline BPM located at the right side of the experiment. Both beams with energy of 6.5 TeV consisted of some 50 nominal bunches. The position scan was performed in 24 steps. Each step lasted about 1 minute to obtain sufficient measurement statistics. The scan range was $\pm 300 \mu\text{m}$ around the reference orbit.

The position change is referenced to the absolute positions measured in the beginning of the scan. The Absolute positions that were subtracted from the measurements are summarized Table 6-1. Difference between the measured values is also presented in the table. Measurement discrepancy between the two BPM systems was in order of some $400 \mu\text{m}$ for the B1 measurement and some $50 \mu\text{m}$ for the B2 measurement. The measured difference is equivalent to some 5 % amplitude asymmetry between the input signals.

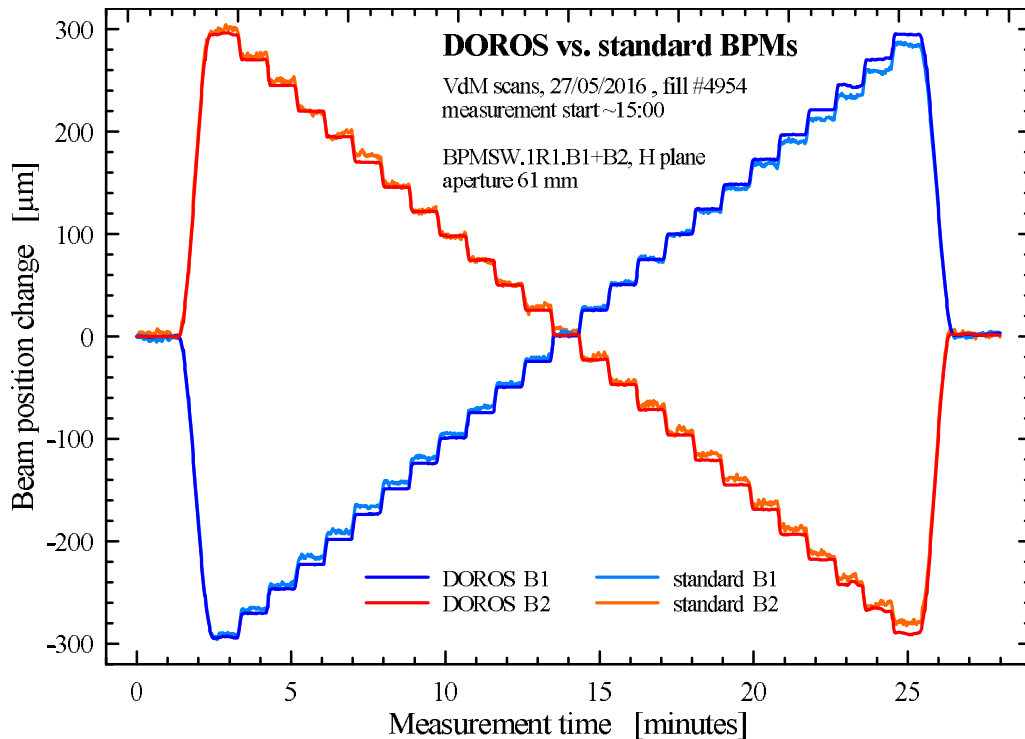


Fig. 6-5. Beam position change measured with DOROS and the standard LHC BPM systems during a Van der Meer scan performed in the ATLAS interaction point (IP1). The positions of both beams were measured in the horizontal plane of the Q1 stripline BPM located on the right side of the IP1. The DOROS front-end was set to the slowest detector time constant T_{c0} with T_{c2} pre-discharge, 5 S transient masking, 10 Hz low-pass IIR cut-off and 1 Hz calibration switching.

The plot in Fig. 6-6 displays the beam positions in the middle of the VdM scan. The plot is zoomed to ± 2 minutes around the 14th minute of the measurement shown in Fig. 6-5. Please note that the displayed signals are relative to the beginning of the scan at $t=0$. The data from the centre of the plot (± 15 seconds around the 14th minute) was used to compute standard deviations (STD) and peak-peak values to estimate the measurement noise. The performance difference between the two systems was evaluated with STD ratios. Obtained results summarized in Table 6-2 show, that the DOROS system achieved more than 6 times better noise performance when compared to the standard LHC BPM system.

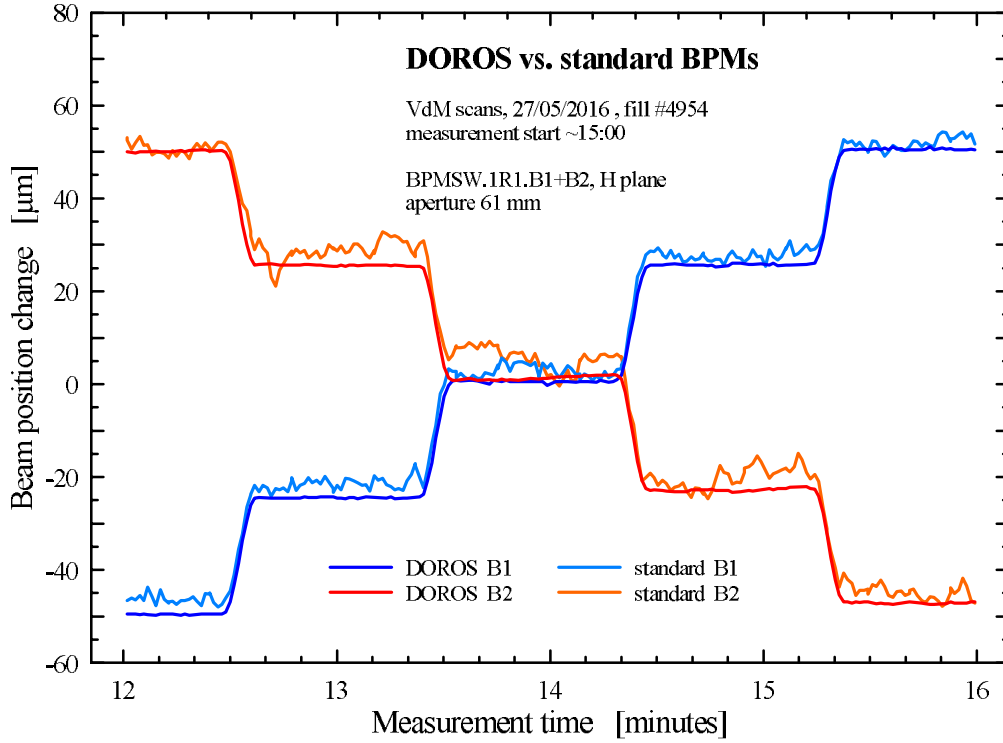


Fig. 6-6. Zoomed part of the beam position measured during the Van der Meer scan in the ATLAS interaction point (IP1). The positions of both beams were measured in the horizontal plane of the Q1 stripline BPM located on the right side of the IP1. The DOROS front-end was set to the slowest detector time constant T_{c0} with T_{c2} pre-discharge, 5 S transient masking, 10 Hz low-pass IIR cut-off and 1 Hz calibration switching.

Table 6-1. Absolute position of the colliding beams measured on the left and right side of the ATLAS interaction point.

	Absolute position [μm]*	
	B1	B2
WBTN	-490	-793
DOROS	-879	-739
WBTN – DOROS	389	54

*Results rounded to 1 μm .

Table 6-2. Standard deviation and peak-peak values of the beam orbits measured upon 30 seconds around the centre of a Van der Meer scan (14th minute).

	Orbit STD [$\mu\text{m RMS}$] * †		Orbit peak-peak [μm] ** †	
	B1	B2	B1	B2
WBTN	1.2	2.5	5.5	9.9
DOROS	0.2	0.3	0.9	1.2
WBTN/ DOROS	6.0	8.3	6.1	8.3

*Results rounded to 0.1 $\mu\text{m RMS}$, **Results rounded to 0.1 μm , † Computed upon 30 s of measurement.

6.2.2 Orbit measurements during stable beams

The long-term stability of the calibrated orbit measurements from the DOROS system was estimated using the data acquired during 37 hour-long stable beams in the LHC. This was one of the record-long LHC runs achieved in year 2016. Beam 1 consisted of some 250 proton bunches and the beam 2 consisted of almost 200 ion bunches at energy of 4 TeV. The measurements were focused on the proton beam whose intensity decreased only by 2.4 % since the beginning of the stable beams. Such conditions were found suitable for long-term orbit measurement without the need to change the RF gain settings in the DOROS front-end.

The calibrated orbit signals acquired during the run are shown in Fig. 6-7 for the horizontal plane and Fig. 6-8 for the vertical plane. Both plots show the measurement from the Q1 BPMs located on the left and right side of the ATLAS experiment in point 1 (IP1). Measurement from both BPMs can be used to project the beam position to the middle of the IP1. The projected position shown in the plot (green) was obtained by computing an average position of the left and right positions. The distance between the two BPMs is approximately 40 m [3]. Please note that the left and right BPM signals were processed by two independent DOROS front-ends.

Table 6-3 summarizes the absolute positions acquired at the beginning of the measurement in IP1. The measured offsets were used to illustrate the trajectory of the beam through the interaction region showed in Fig. 6-9. Please note that the picture is not in scale. The vertical beam positions in the BPMs were measured around 3 mm and the horizontal around 1 mm. The projected beam position was marked by red point in the middle of the picture. The beam offsets projected to the middle of the IP1 were measured around 0.2 mm in vertical and 70 μm in horizontal plane.

The projected orbits were also computed from the Q1 BPMs located on the left and right side of the CMS experiment in point 5 (IP5). The two experiments in IP1 and IP5 are apart by 3 octants on the LHC circumference (distance of some 10 km) [3]. Measured absolute position values were summarized in Table 6-3. The vertical beam positions in the BPMs were measured around 0.2 mm and the horizontal around 2.5 mm. The projected positions in the IP5 were measured around 0.1 mm in vertical and 40 μm in horizontal plane.

The plots in Fig. 6-10 and Fig. 6-11 display the projected orbit signals in the horizontal and vertical plane of IP1 and IP5. The position changes in each signal is referenced to the beginning of the measurement. As seen in the plot the orbit signals in both interaction points were changing by less than 15 μm over the 37 hour-long measurement. It can be seen that the change of the projected orbit in the two IPs are correlated.

The following plot in Fig. 6-12 displays the projected orbits zoomed to the beginning of the measurement. It can be seen that the peak-peak noise over the first 3 minutes was around 0.5 μm . The orbit peak-peak change seen between the 3rd and 7th minute of the measurement increased to some 4 μm . It was observed that the projected orbits follow decaying oscillation-like pattern in both IPs and both horizontal and vertical planes with amplitudes of about 1 μm . Investigation throughout the acquired data in 2016 showed that similar orbit oscillations could be observed with the standard DOROS system on both LHC beams. Preliminary results show that the 3 μm orbit changes observed in Q1 BPMs around LHCb experiment (IP2) caused 18 % luminosity variation in the experiment. Further investigations are carried out to identify the source of the jitter and improve performance of the LHC.

Table 6-3. Absolute beam position measurement around the ATLAS (IP1) and CMS (IP5) interaction regions.

Abs. position [mm]*	B1 Horizontal			B1 Vertical		
	1L	1R	IP projection	1L	1R	IP projection
IP1	0.685	-1.183	-0.249	3.145	-3.284	-0.070
IP5	-2.626	2.853	0.114	-0.141	0.209	0.034

*Results rounded to 1 μm .

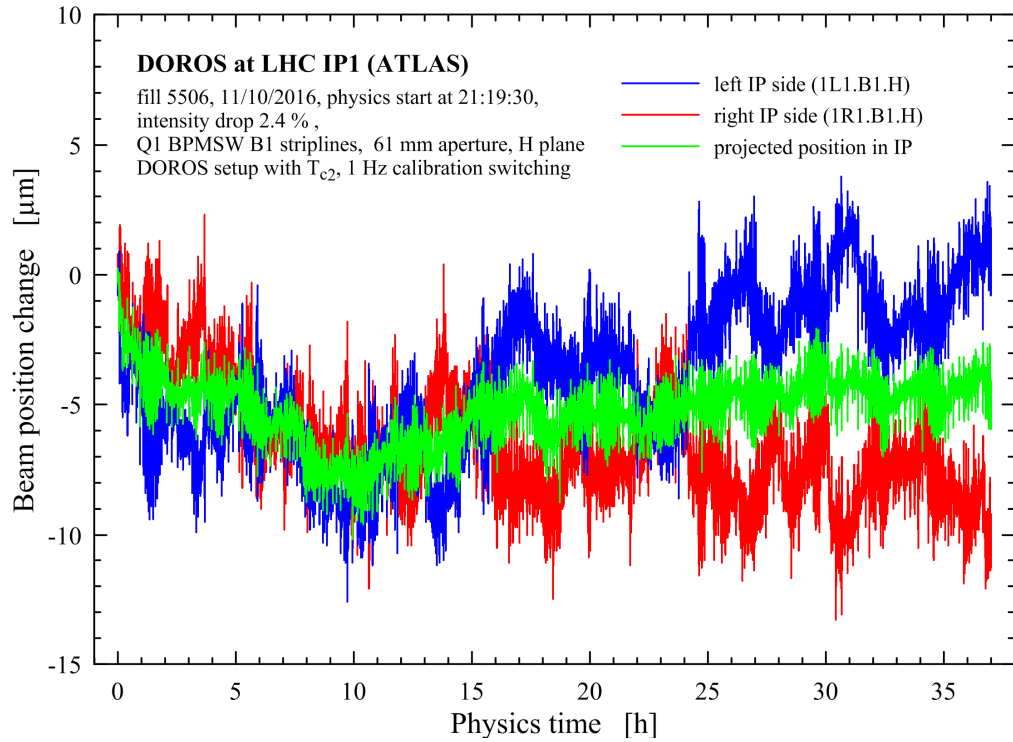


Fig. 6-7. Measurement of the calibrated and the projected horizontal orbits of the beam in the ATLAS interaction region. The position projection to the middle of the IP1 was calculated as an average of the positions measured in left and right Q1 BPMs. The measurement was acquired during a 37 hour LHC run with proton-ion beams. The plot displays orbits of the proton beam with some 250 nominal bunches. During the run the beam intensity decreased by 2.4 %. The DOROS front-ends were set to 1 Hz calibration switching, time constant T_{c2} , 5 sample masking and 10 Hz IIR low-pass cut-off.

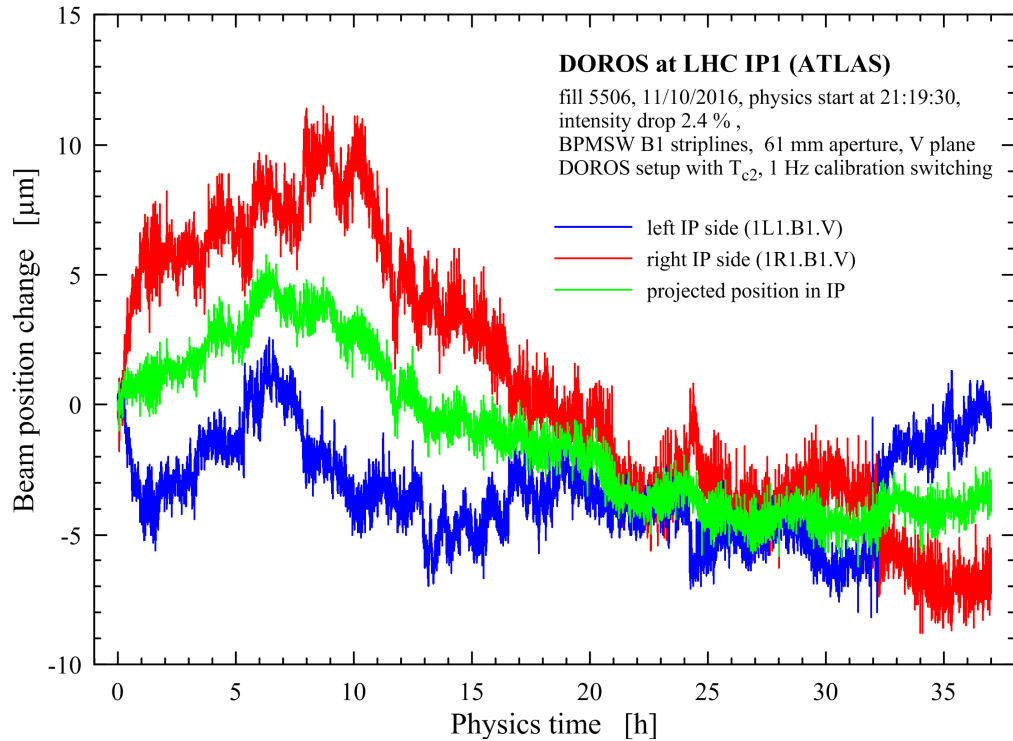


Fig. 6-8. Measurement of the calibrated and the projected vertical orbits of the beam in the ATLAS interaction region. The position projection to the middle of the IP1 was calculated as an average of the positions measured in left and right Q1 BPMs. The measurement was acquired during a 37 hour LHC run with proton-ion beams. The plot displays orbits of the proton beam with some 250 nominal bunches. During the run the beam intensity decreased by 2.4 %. The DOROS front-ends were set to 1 Hz calibration switching, time constant T_{c2} , 5 sample masking and 10 Hz IIR low-pass cut-off.

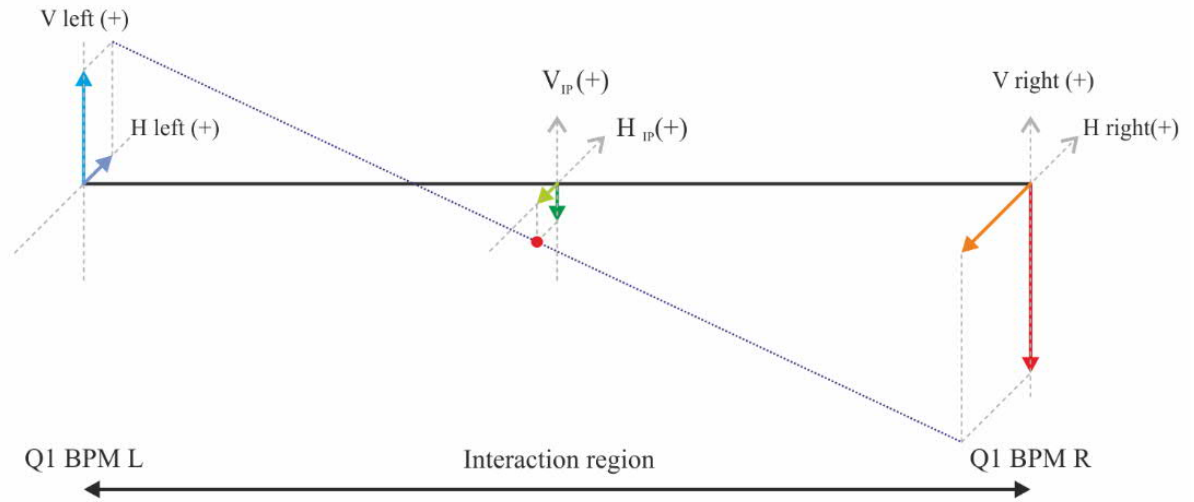


Fig. 6-9. Diagram depicting the beam offset projection in the ATLAS interaction region (IP1). Horizontal and vertical offsets showed in the left and right side of the diagram (blue and red arrows) represent absolute orbit measurement measured by the DOROS system (values not in scale). The projected offsets to the middle of the IR1 are represented by the green arrows in both horizontal and vertical axis (values not in scale).

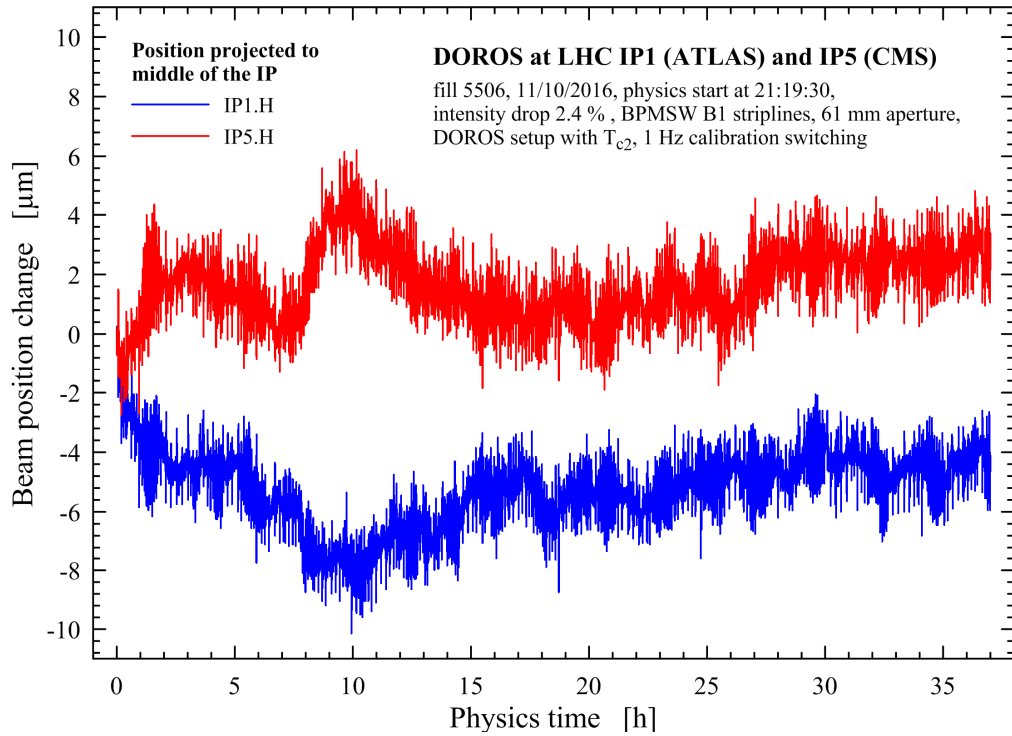


Fig. 6-10. Orbit measurement in horizontal plane of the ATLAS (IP1) and CMS (IP5) interaction regions obtained during a 37 hour physics operation with proton-ion collisions. The plot displays orbits of the proton beam with intensity drop of 2.4 % over the measurement time. The position projection to the centre of the IPs was calculated as an average of the positions measured in left and right BPMs around the experiments. The DOROS front-ends located at IP1 and IP5 were configured with time constant T_{c2} , 1 Hz calibration switching, 5 sample masking, 10 Hz IIR low-pass cut-off.

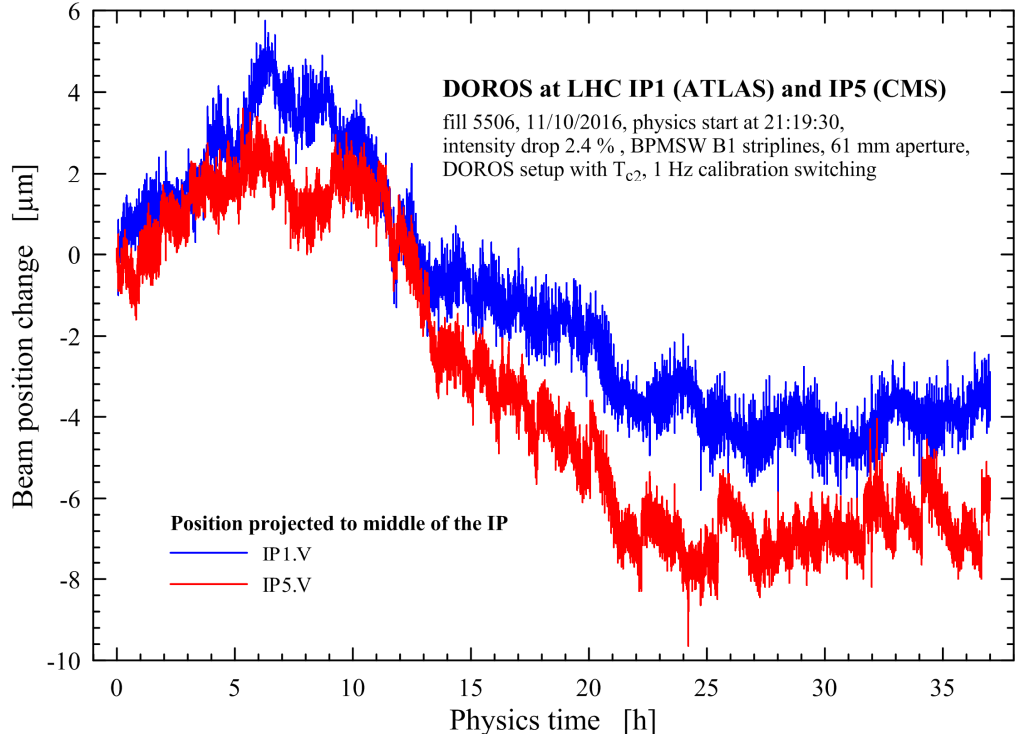


Fig. 6-11. Orbit measurement in the vertical plane of the ATLAS (IP1) and CMS (IP5) interaction regions obtained during a 37-hour LHC physics run with proton-ion collisions. The plot displays orbits of the proton beam with intensity drop of 2.4 % over the measurement time. The position projection to the centre of the IPs was calculated as an average of the positions measured in left and right BPMs around the experiments. The DOROS front-ends located at IP1 and IP5 were configured with time constant T_{c2} , 1 Hz calibration switching, 5 sample masking, 10 Hz IIR low-pass cut-off.

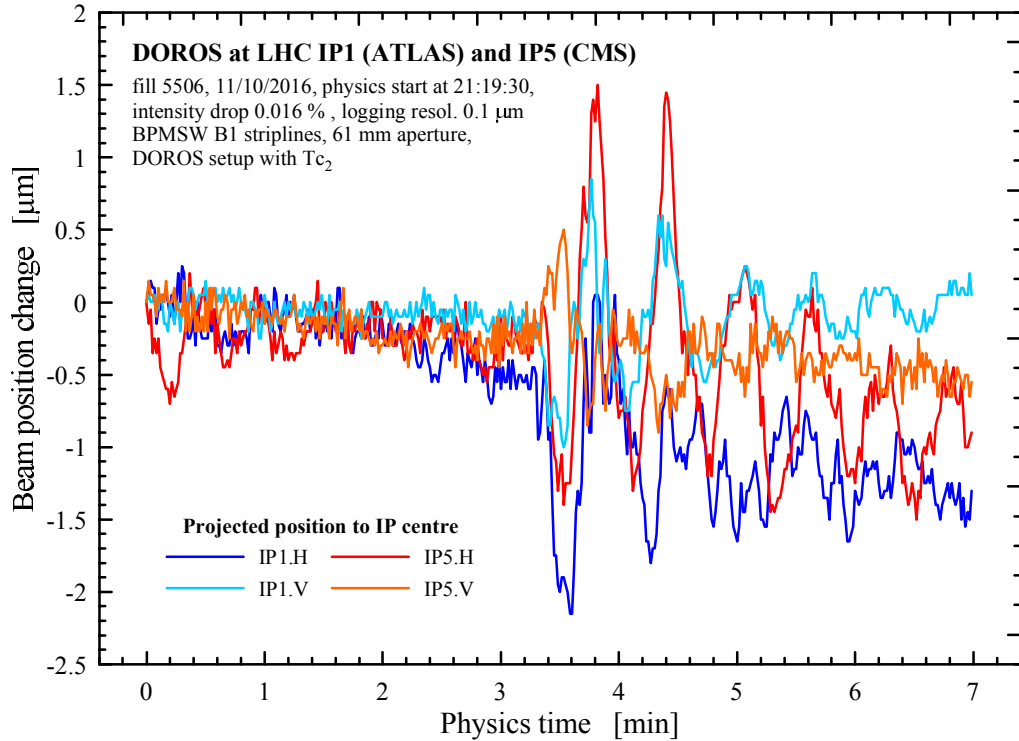


Fig. 6-12. Orbit measurement in both the vertical and horizontal planes of the ATLAS (IP1) and CMS (IP5) interaction regions zoomed to the beginning of the LHC physics run with proton-ion collisions. The plot displays orbits of the proton beam with intensity drop of 0.016 % over the measurement time. The position projection to the centre of the IPs was calculated as an average of the positions measured in left and right BPMs around the experiments. The DOROS front-ends located at IP1 and IP5 were configured with time constant T_{c2} , 1 Hz calibration switching, 5 sample masking, 10 Hz IIR low-pass cut-off.

6.2.3 Low frequency orbit oscillations

The capture mode in the DOROS front-ends allows to acquire the ADC samples at the turn-by-turn rate and store them in the local buffers for offline analysis. In this mode, the orbit signals can be observed with bandwidth from DC extended up to some 100 Hz. The spectral analysis of the captured signals is can be performed without the calibration switching. This feature became interesting for machine diagnostics requiring high sensitivity to low frequency beam oscillations. Such oscillations decrease the performance of the LHC by decreasing the beam lifetime as well as affecting the collision rates in the experiments.

The plot in Fig. 6-13 presents magnitude spectra of the orbit signal measured in the BPM on the right side of the ATLAS experiment (IP1). The measurement was acquired at the energy of 6.5 TeV with the LHC beam 1 consisting of more than 1000 nominal bunches. The capture was set to store data for 1 million beam turns, which is approximately 1.5 minutes of turn-by-turn data acquisition. The DOR data used for computing the beam orbits was sampled at f_{rev} rate and decimated by factor 20. The orbit spectra shown in the plot based on 50 kS FFT with Hann windowing. Both the vertical and horizontal axis were displayed in logarithmic scale. The vertical scale represents orbit oscillation amplitudes expressed in micrometres. The horizontal axis displays frequencies ranging from 1 Hz up to the half of the decimated sampling frequency (equivalent to approximately 280 Hz). The measurement was repeated with the three DOR time constants. It can be seen in the plot that increasing the discharge rate from T_{c0} to T_{c2} increased the sensitivity to frequencies above 20 Hz due to increased bandwidth of the diode detectors.

Some of the spectral lines seen between 5 Hz and 20 Hz were identified as Eigen frequencies of the mechanical structure of the Q1 magnets [78]. The results were however not conclusive and more measurements are required. Similar type of measurements allow to monitor the orbit spectra on the periodic basis and correlate the results with possible driving terms to identify its source. Identifying and mitigating such sources is important for the LHC performance improvements.

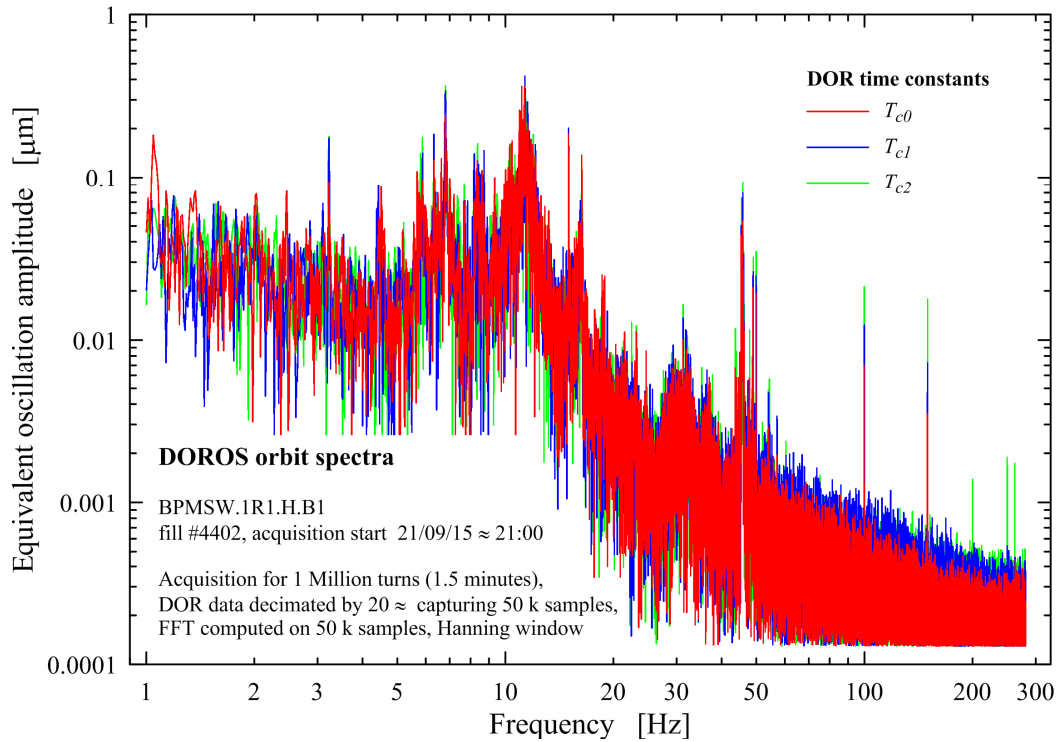


Fig. 6-13. Magnitude spectra of the beam orbit signals measured on the right side of the ATLAS experiment during the LHC stable beams. The measurement was repeated with the three DOR detectors time constants. The time constant T_{c0} corresponds to the slowest discharge rate of the detector capacitors. The time constants T_{c1} and T_{c2} correspond to faster discharge rates. The signals were acquired for 1 million turns (equivalent to 1.5 minutes) and decimated by factor 20.

6.2.4 Limitations of the diode orbit subsystem

The laboratory measurements in chapter 5.1 showed that the non-linear position errors depend on the beam offset, detector time constant and amplitude of the input signals. This chapter presents the limitations of the DOROS system caused by the non-linear characteristic of the compensated diode detectors. The standard DOROS system installed around the experiments has to often measure large beam offsets in order of millimetres. The presented measurements in this chapter were acquired in the Q1 BPMs around the ATLAS interaction point (IP1). There the two beams may have position offsets larger than 4 mm.

The plot in Fig. 6-14 shows the DOR signals from two DOROS front-end measuring the BPM electrodes on the left (blue and cyan) and right side (red and orange) of the IP1. The corresponding vertical axis is located on the right of the plot. The measurement was acquired during a 24-hour LHC physics operation with more than 2000 bunches in each beam. During this time the two beams were colliding in the experiment. Beam intensity (green) measured by the Fast Beam Current Transformers [80] is also shown in the plot with corresponding vertical axis displayed on the left. The values were normalised by the intensity measured at the beginning of the measurement. It can be seen that the beam intensity dropped during the 24-hour measurement by some 42 %. However, the intensity proportional DOR signals followed the intensity change only during the first 4 hours of the measurements. For the remaining 20 hours of the measurement the DOR signals changed by less than 2 % without any correction in the automatic gain control. At the end of the 24-hour measurement the deviation of the DOR signals from the beam intensity was around 30 %. It was discovered that during the 20 hour period the DOROS system measured the orbits of bunches which don't collide in the experiment (the so called non-colliding bunches).

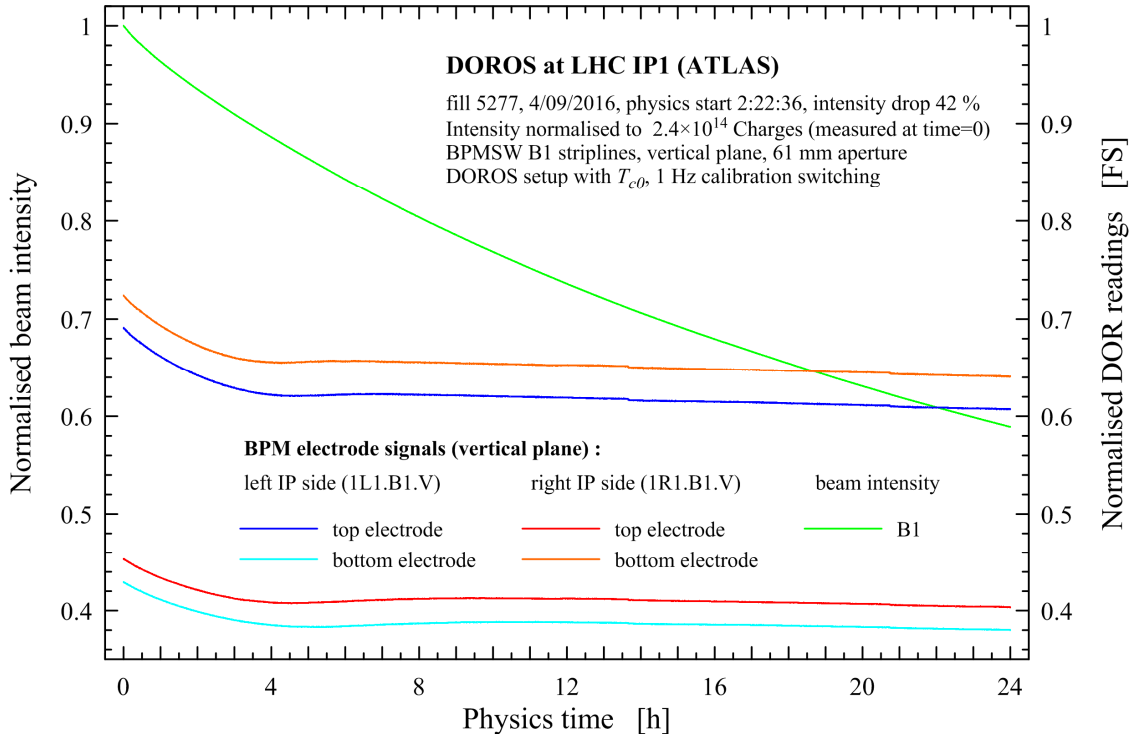


Fig. 6-14. DOR electrode signals acquired from Q1 BPMs on the left and right side of the ATLAS interaction point during 24 hours beam collisions in the experiment. The beam consisted of more than 2000 bunches. The plot displays the BPM signals from vertical plane (electrode positions marked as top and bottom). The corresponding vertical axis is showed on the right side of the plot. The normalised beam intensity is plotted with the BPM signal with the corresponding vertical axis located on the left of the plot. The DOROS front-ends were operating with 1 Hz calibration switching, DOR time constant T_{c0} , T_{c2} pre-discharging, 5 S masking and 10 Hz IIR cut-off frequency.

The plot in Fig. 6-15 displays intensity of individual bunches measured at the beginning, at 5th hour and at the end of the 24-hour physics operation in the LHC. The vertical axis represents the number of charges in a bunch. The horizontal axis represents the bunch index ranging from 1 to 3564 in the bunch train. The first 12 bunches in the beginning of the in the bunch train are non-colliding. The difference between the non-colliding bunches and the rest of the beam was very small in the beginning of the measurement. Differences between the peak amplitudes in the BPM signals are therefore also small. The DOR diode detectors with slow discharge constant T_{c0} take most of bunches in the beam into account and measure their average positions. However, after 5 hours of the beam collisions in the experiment the average intensity of the colliding bunches decreased by more than 10 %. The intensity of the non-colliding bunches decreases only by some 2 %. Due to the slow discharge rate the DOR detectors then start to take into account bunches with maximum intensity. The effect becomes more apparent after 24 hours when the average intensity between the colliding and non-colliding bunches decreased by more than 40 %. At this time the diode detectors measure only the non-colliding bunches while the remaining bunches in the beam are not taken into account. More details about the diode detection principle can be found in chapter 3.2.

Resulting effect on the orbit measurement is shown in Fig. 6-16. The plot displays the change of the calibrated orbits measured in BPMs located on the vertical plane on the left (blue) and right (red) side of the Interaction Point (IP*). The plot also shows the projected position to the middle of the IP* displayed (green) computed as an average of the left and right positions. The absolute positions values from the beginning of the measurement are summarized in Table 6-4. It can be seen that the position measurements on both sides of the experiment change by $\pm 80 \mu\text{m}$. The change occurs around 5th hour of the measurement when the diode detector start to filter the colliding bunches. The resulting change on the projected orbit was measured around $20 \mu\text{m}$. The error is comparable to a typical beam size of some $20 \mu\text{m}_{\text{RMS}}$ in the IP1 during collisions [3]. The projected orbit stabilised around 12th hour once the diode detectors started to measure the non-colliding bunches. The orbit spikes seen on the projected orbit were identified as the beam adjustments to optimise the collision rate in the experiment.

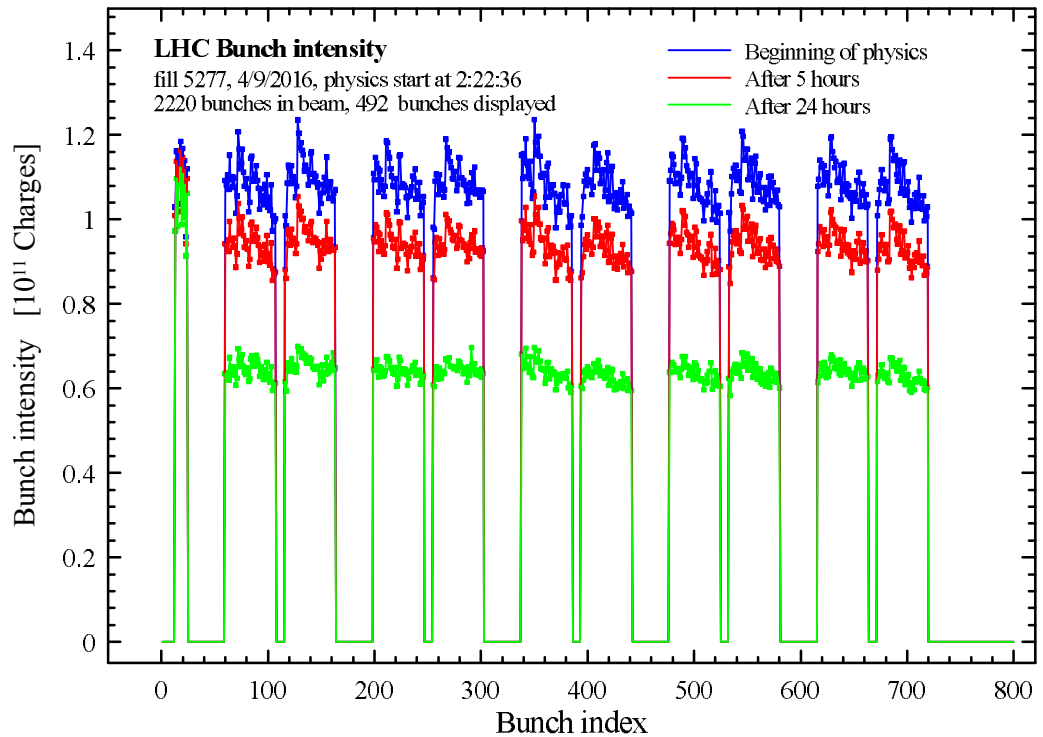


Fig. 6-15. Intensity of first 492 bunches in the bunch train. The plot displays three measurements acquired with Fast Beam Current Transformers [80] at different time during one LHC physics run. The beam consisted of 2220 bunches (492 displayed) spaced by a minimum of 25 ns.

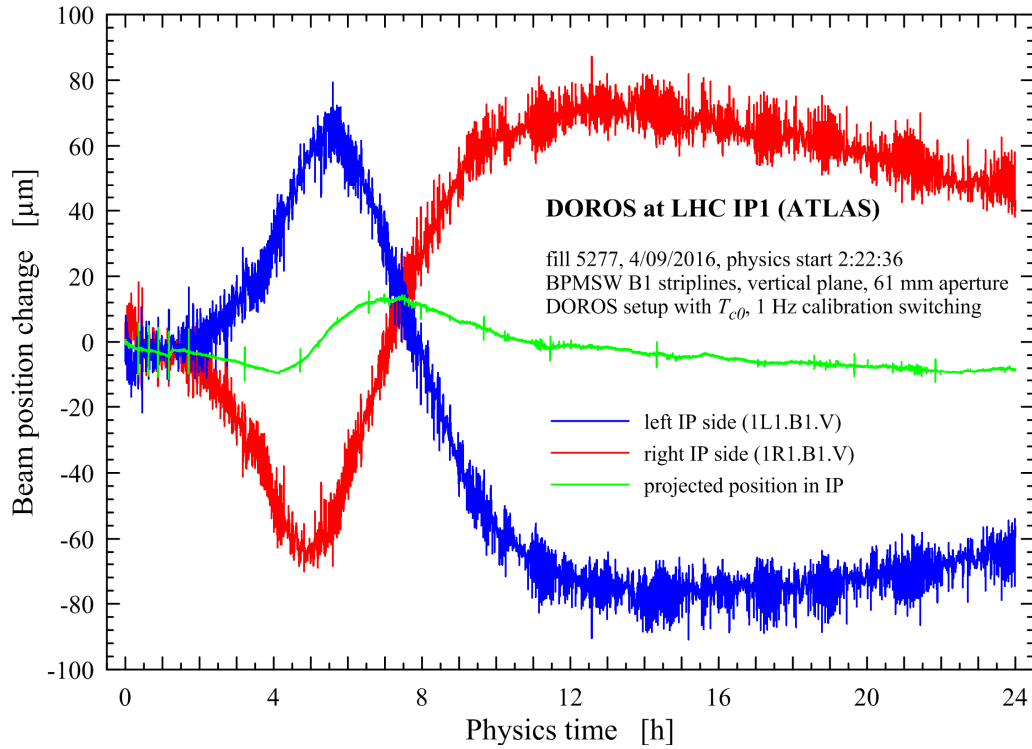


Fig. 6-16. The 24 hour orbit measurements during LHC physics operation in ATLAS interaction point (IP1). The plot shows calibrated orbits acquired from the left (blue) and right (red) side of the IP as well as projected position to the middle of the IP (green). The diode detectors were operating with slow discharge constant T_{c0} . The DOROS system started to measure orbit of the non-colliding bunches at about 4th hour of the measurement with an intensity difference between bunches of about 8 %. Projected position was calculated as an average of the left and right position measurements.

Table 6-4. Absolute positions measured in Q1 BPMs around ATLAS interaction point (Fill 5277).

IP1	B1 Horizontal			B1 Vertical		
	1L	1R	IP projection	1L	1R	IP projection
Abs. position [mm]*	0.7314	-0.9949	-0.13175	4.0377	-3.9771	0.0303

*Rounded to 0.1 μm .

The time constant T_{c0} which is the preferred setting for the beam injections cannot be universally used for any LHC beam operation. The false drift in the orbit measurement perturbs especially systems such as the orbit feedback system which rely on the stability and reproducibility of the measurement. A solution to this problem was to automatically set the DOR diode detectors during the LHC physics operation to fast discharge with T_{c1} or T_{c2} .

The plot in Fig. 6-17 shows the electrode signals and beam intensity signals during a similar measurement with the same BPMs located around IP1. In this measurement the DOR detectors were configured with fast discharge constant T_{c2} . The beam intensity decreased over the 26-hour long physics operation by some 24 %. In this case the automatic gain control (AGC) compensated the amplitude drop of the BPM signals by increasing the RF gains in 1 dB steps. The DOR signal levels were controlled in the range 0.6 - 0.8 FS set by the AGC thresholds. Please note that the gain adjustment is not synchronised between the front-ends and thus can occur when the maximum electrode reading is below the minimum threshold. Minimal gain steps allow for minimal amplitude change in the DOR signals.

The effect of the gain change on the orbit measurement is displayed in Fig. 6-18 for the horizontal and in Fig. 6-19 for the vertical BPM plane. The plot presents the orbit signals obtained from the two sides of the IP* as well as the projected position. The absolute positions were subtracted from the signals. Obtained values are presented in Table 6-5.

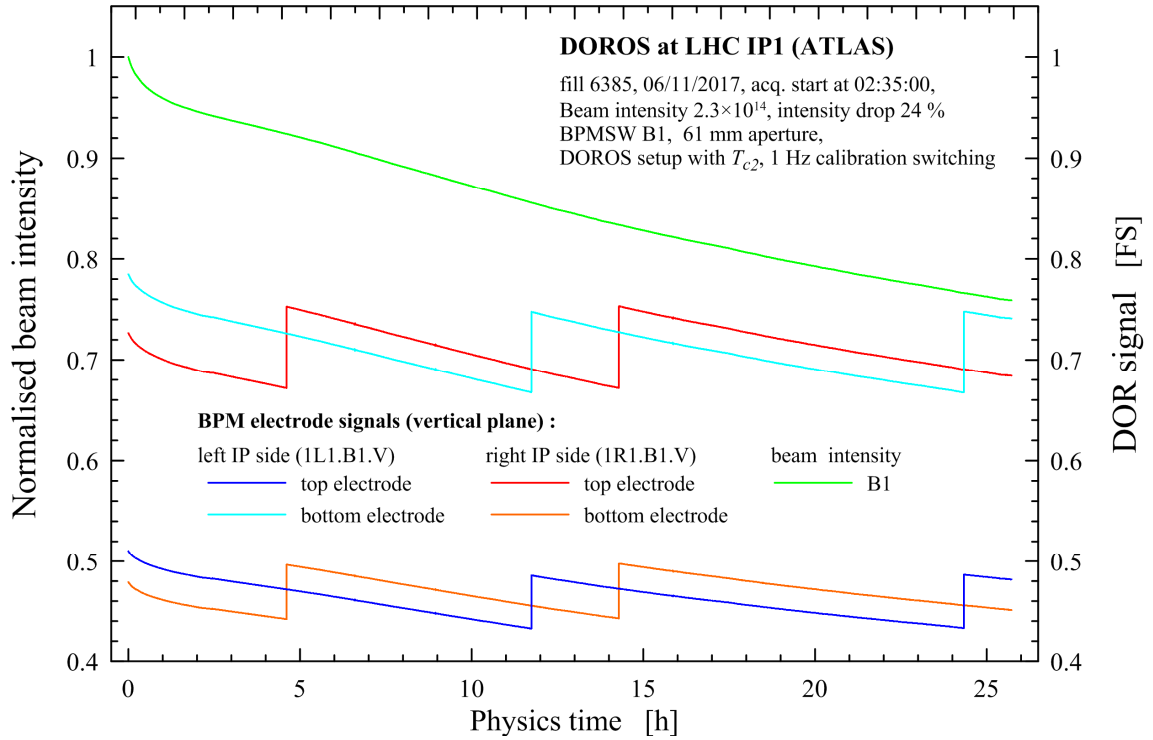


Fig. 6-17. DOR electrode signals acquired from Q1 BPMs on the left and right side of the ATLAS experiment during 26 hours of beam collisions in the experiment. The beam consisted of some 2000 bunches. The plot displays the BPM signals from vertical plane (electrode positions marked as top and bottom). The corresponding vertical axis is showed on the right side of the plot. The normalised beam intensity is plotted with the BPM signal with the corresponding vertical axis located on the left of the plot. The DOROS front-ends were operating with 1 Hz calibration switching, DOR time constant T_{c2} , 5 S masking and 10 Hz IIR cut-off frequency.

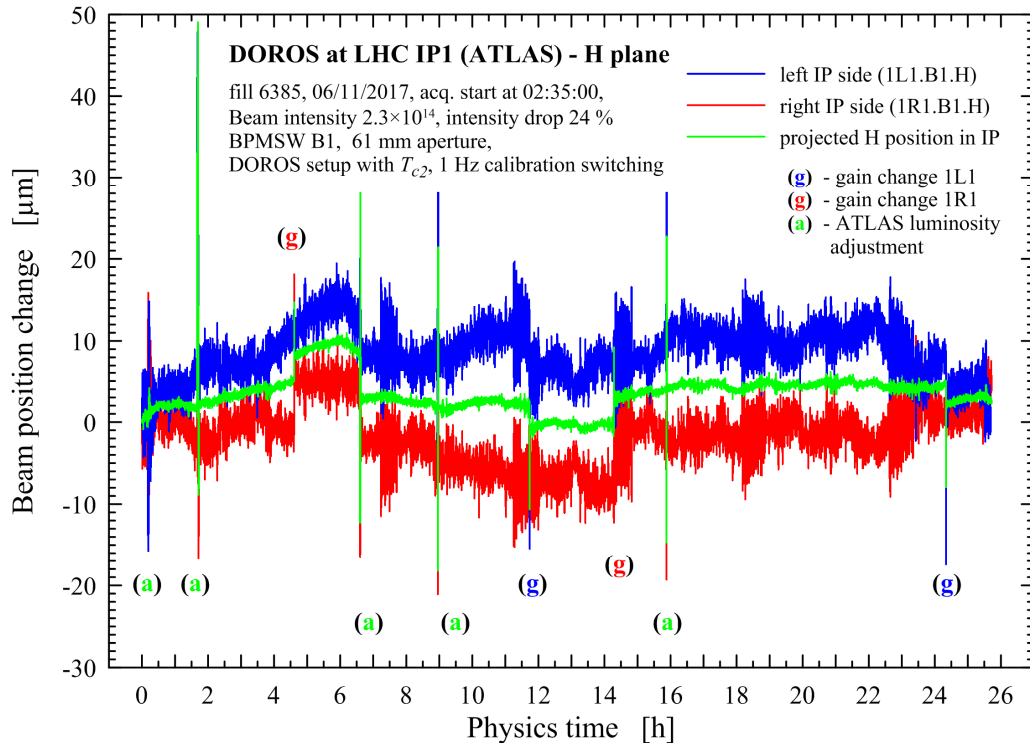


Fig. 6-18. The 26 hour orbit measurements in the horizontal plane during LHC physics operation in ATLAS experiment (IP1). The plot shows calibrated orbits acquired from the left (blue) and right (red) side of the IP as well as projected position to the middle of the IP (green). The intensity drop during the measurement was around 24 %. The diode detectors were operating with fast discharge constant T_{c2} . Projected position was calculated as an average of the left and right position measurements.

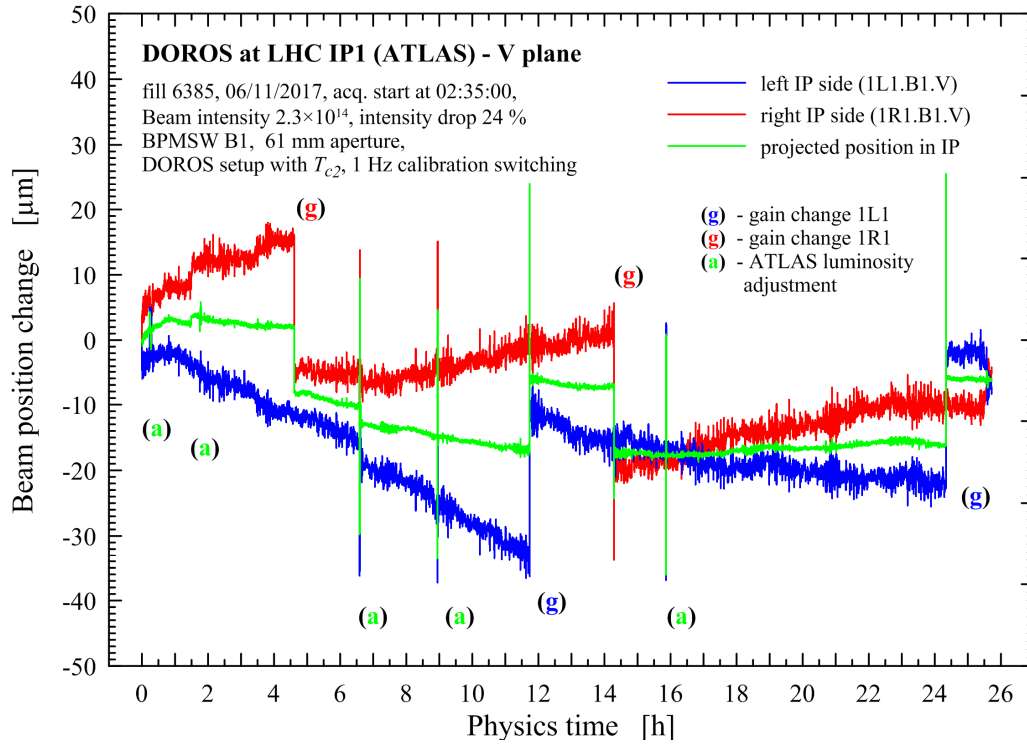


Fig. 6-19. The 26 hour orbit measurements in the vertical plane during LHC physics operation in ATLAS experiment (IP1). The plot shows calibrated orbits acquired from the left (blue) and right (red) side of the IP as well as projected position to the middle of the IP (green). The intensity drop during the measurement was around 24 %. The diode detectors were operating with fast discharge constant T_{c2} . Projected position was calculated as an average of the left and right position measurements.

Table 6-5. Absolute positions measured in Q1 BPMs around ATLAS interaction point (Fill 6385).

IP1	B1 Horizontal			B1 Vertical		
	1L	1R	IP projection	1L	1R	IP projection
Abs. position [mm]*	1.0322	-1.1301	-0.0489	-3.6547	3.5439	-0.0554

*Rounded to 0.1 μm .

Some of the position jumps seen in both the vertical and horizontal plane correspond to the adjustments performed at that time. These were marked with (a) in the plots to distinguish from other effects. The 1 dB gain steps during the automatic gain adjustments are marked in both plots with (g). The increase in the signal amplitudes from 0.68 to 0.78 FS on the ADC inputs resulted in some 20 μm orbit discontinuities visible mostly in the vertical plane. There the position offsets of the beam 1 were larger than 3 mm. Observed orbit errors were caused by the non-linear characteristic of the compensated diode detectors. The characteristic of the DOR detectors with time constant T_{c2} and similar position offset was measured and presented in chapter 5.1. The 3rd order polynomial for T_{c2} was chosen to correct the non-linear errors shown in this analysis. The polynomial with coefficient from Table 5-6 was applied to each DOR reading in the post-processing to produce corrected electrode signals. Obtained values were then used to compute the normalised positions according to equation (3-1). The positions were converted to millimetres using the 2D polynomials (3-3) and (3-4) with coefficients summarized in Table 3-1 to correct the BPM characteristic.

Obtained corrected orbits from the left and right side of the IP* as well as the projected orbit in middle of the IP* are shown in Fig. 6-20 for the horizontal and Fig. 6-21 for the vertical BPM plane. Absolute positions for the corrected orbits are summarized in Table 6-6. It can be seen that the correction polynomials obtained in the laboratory were sufficient to minimize the orbit discontinuities during the gain steps. The orbit drift on each side of the experiment were measured around 10 μm in the horizontal and 20 μm in the vertical plane. The projected position in both planes drifted by some $\pm 5 \mu\text{m}$.

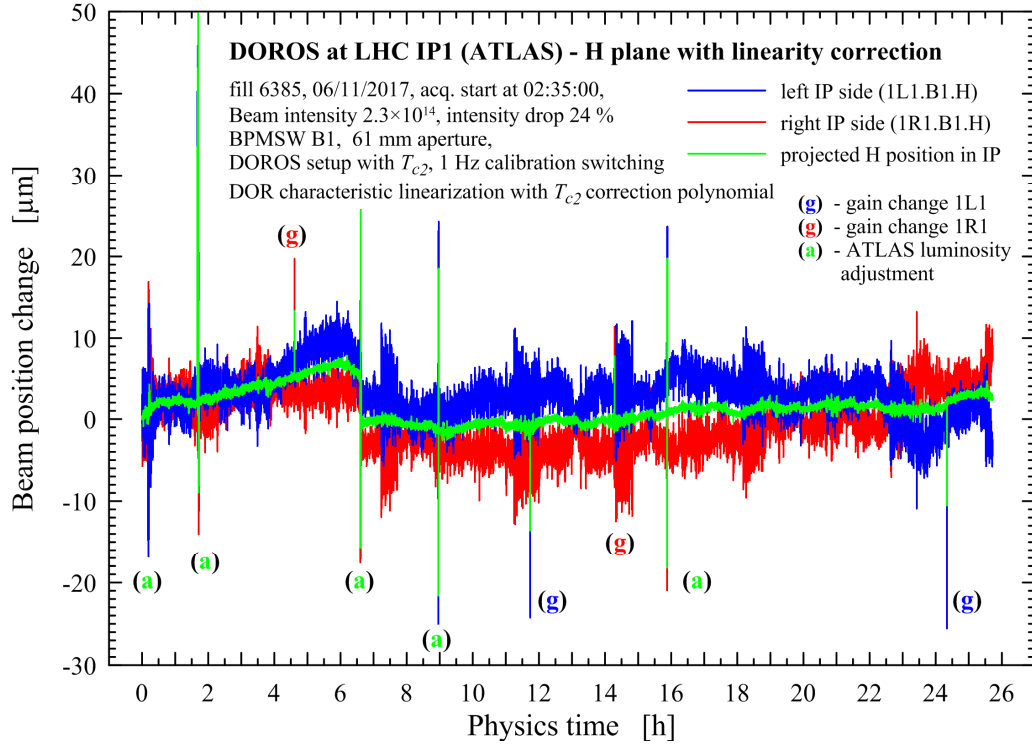


Fig. 6-20. The 26 hour orbit measurements in the horizontal plane during LHC physics operation in ATLAS experiment (IP1). The plot shows calibrated orbits with corrected non-linearity of the DOR detectors. The orbits were measured in left (blue) and right (red) side of the IP as well as projected position to the middle of the IP (green). The diode detectors were operating with fast discharge constant T_{c2} . Projected position was calculated as an average of the left and right position measurements.

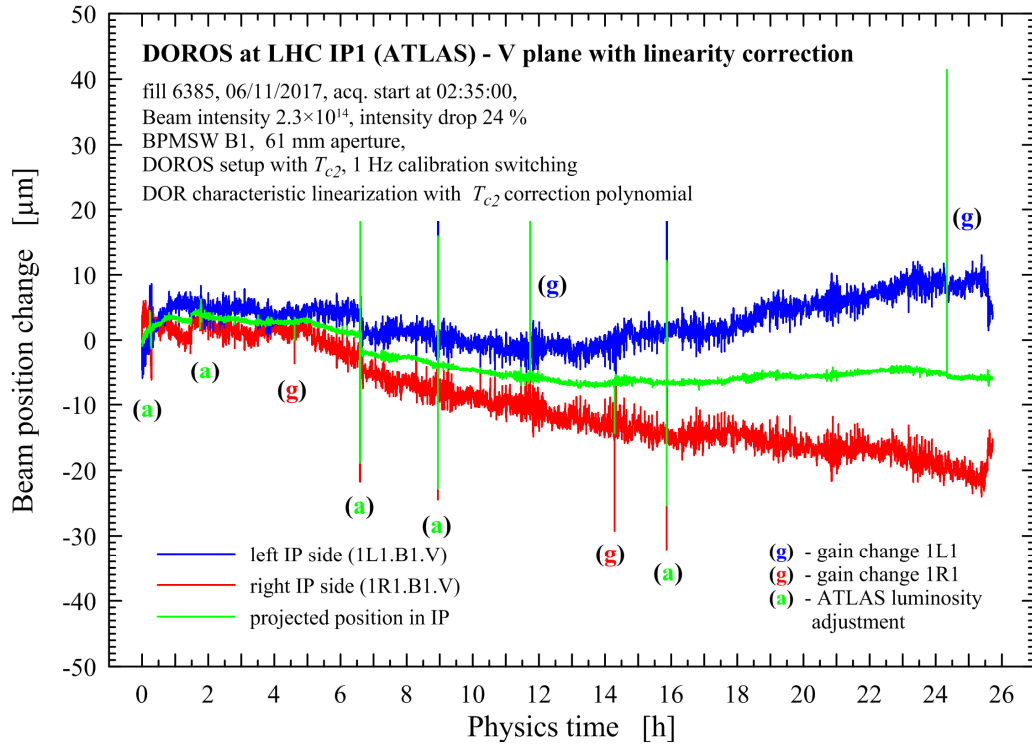


Fig. 6-21. The 26 hour orbit measurements in the vertical plane during LHC physics operation in ATLAS experiment (IP1). The plot shows calibrated orbits with corrected non-linearity of the DOR detectors. The orbits were measured in left (blue) and right (red) side of the IP as well as projected position to the middle of the IP (green). The diode detectors were operating with fast discharge constant T_{c2} . Projected position was calculated as an average of the left and right position measurements.

Table 6-6. Absolute corrected positions measured in Q1 BPMs around ATLAS interaction point (Fill 6385).

IP1	B1 Horizontal			B1 Vertical		
	1L	1R	IP projection	1L	1R	IP projection
Abs. position [mm]*	1.0509	-1.1457	-0.0474	-3.7222	3.5955	-0.0633

*Rounded to 0.1 μm .

The remaining transients during the gain changes were caused by a transition between the automatic gain control and the automatic calibration switching. The calibration switching is switched off during these transitions and the gain is automatically adjusted. The invalid positions would be normally discarded by the DOROS server and the previous valid value would be published. Such functionality was not available at the time of the measurement.

6.2.5 Summary and outlook

This chapter presented beam orbit measurements acquired with the standard DOROS system. As of 2018 the system consists of about 20 front-end units installed on selected BPMs located in LHC points 1 (ATLAS), 2 (ALICE), 4 (RF cavities), 5 (CMS), 6 (beam dump) and 8 (LHCb). The most important DOROS front-ends were installed on the BPM sensors located next to the LHC experiments. There, the orbit measurements are the most challenging as high resolution, long-term stability and low systematic errors are required for achieving the best LHC performance. Since 2015 the standard DOROS system was used for numerous LHC beam measurements, studies and diagnostics. The performance and limitations of the DOROS system were demonstrated on orbit measurement examples acquired from selected Q1 BPMs around the ATLAS and CMS. The electrode signals in these locations were split between the DOROS and the standard LHC systems allowing to evaluate and compare their measurement results.

The comparison of the orbit measurements between the two systems was shown on an example of the Van der Meer position scans in the ATLAS experiment. This type of measurements is important for optimising the beam collisions. The largest discrepancy between the absolute beam positions was measured around 400 μm which is equivalent to 0.6 % error on the BPM aperture. Part of the orbit measurement was used to estimate and compare the precision of the two systems. In this comparison the DOROS system achieved improvement of more than factor 6 in both the standard deviation and peak-peak noise estimation. However, the measurements were limited by the noise of the beam itself. These results initiated more detailed studies targeting the performance of the standard LHC system.

Long-term stability of the DOROS system was demonstrated on the orbit measurements acquired during a 37 hour-long LHC physics operation with proton and ion beams. This particular cycle was selected due to the very low intensity decay of the proton beam. The beam orbits were measured in the Q1 BPMs located on the left and right side of the ATLAS and CMS experiments. The left and right beam positions were used to compute the projected position to the middle of the interaction region where the two beams collide. Measured orbit signals changed by less than 15 μm . It was observed that the beam position drifts are correlated between the ATLAS and CMS interaction regions. A more detailed comparison of the orbit measurements revealed periodic beam motion with amplitudes in the order of micrometres and frequency in the order of 0.01 Hz. It was observed that the oscillations have impact on the collisions in the experiments.

The DOROS system can measure orbits with increased bandwidth in the turn-by-turn capture mode. Capturing a few million beam revolutions in the LHC allows to observe beam oscillations in the bandwidth from DC up to some 100 Hz. Long acquisitions enable to reach FFT resolution better than 0.5 mHz. A few ongoing studies are focusing on observing the orbit spectra in the low frequency range to investigate and identify potential sources of instabilities and beam losses. Such investigations are important for the high luminosity upgrades improving the performance and reliability of the LHC.

Based on the achieved performance presented in [78] the standard DOROS system will be used to periodically acquire orbit spectra during each fill. The 1 Hz calibration switching would have to be switched off during each data capturing. Obtained spectrograms will allow monitoring of the beam activity throughout the LHC cycles.

The last part of this chapter is focused on limitations of the DOROS system measuring the beam orbits during the typical physics operation of the LHC. The first measurement demonstrated the dependency of the orbit measurement on the intensity distribution of the individual bunches when slow detector discharge T_{c0} is used. In this regime the diode detectors operate in peak detection mode, ignoring bunches with smaller intensity. When the intensity of the colliding bunches decreased by more than 8 % the system started to measure the orbits of the 12 non-colliding bunches in the beam. This effect induced a false position change of some $\pm 80 \mu\text{m}$ in the Q1 BPMs and some $20 \mu\text{m}$ when projecting the orbit in the middle of the IP*. This problem can be solved by using the fastest detector discharge T_{c2} . In this regime the diode detectors operate in the averaging mode. This was demonstrated on a second example of the orbit measurements from the same BPMs during a similar LHC fill. It was shown that the DOR signals were changing according to the decreasing beam intensity. The signal drop was compensated by the automatic gain control increasing the RF gains in the front-ends by 1 dB. As the time constant T_{c2} is not suitable for measuring the LHC pilot beams, this parameter should be automatically adapted according to the number of bunches in the machine and the operation mode. This is important especially during the injection process. Ultimately, the dependency of the orbit measurement on the intensity distribution between bunches can be solved by gating the RF signals on the diode detector inputs. The orbit measurement would then depend on precision and reliability of the gating signals. Such solution would also increase complexity of the system with potential impact on its reliability and robustness. Proposed orbit gating is currently under studies.

The second measurement also demonstrated the non-linear effects of the DOR characteristic on the precision of the orbit signals. The non-linearity errors became apparent especially in the vertical plane of the Q1 BPMs where the beam offsets were larger than 3 mm. Each 1 dB increase in the RF gains caused some $20 \mu\text{m}$ step in orbit measurements. Such jumps significantly limit the achievable precision of the orbit measurement. Similar result was achieved with the laboratory measurements presented in chapter 5.1. Proposed linearization polynomials presented in the same chapter were used to correct the non-linear effects observed in the beam measurements. Obtained results showed that all the orbit steps caused by the gain change were minimized in this measurement to less than $1 \mu\text{m}$. Please note that the achieved improvement depends on the matching between the laboratory measured polynomials and the actual DOR characteristic. Preliminary results showed that the 3rd order correction polynomials presented in this thesis can reduce the non-linear errors by more than factor 5 in operational LHC conditions. Based on achieved results the non-linearity correction is performed as a part of the position computation in the DOROS server. The same correction coefficients are used for all DOROS front-ends. Further studies are carried out to improve the precision of the orbit measurements by decreasing the minimal gain step and improving the precision of the correction coefficients.

The standard DOROS system was used for a number of machine dedicated studies. Many of them were targeting the LHC high luminosity upgrade [81]. An example of one such study is the luminosity control and the so called β^* levelling. These methods require measurements of the beam orbits around the LHC experiments with high resolution, precision and fill-to-fill repeatability. The standard DOROS system was used for a few measurements whose results were presented as part of a doctoral thesis [63].

In view of the LHC upgrades the standard DOROS system was used to evaluate the directivity of the Q1 stripline BPMs. These studies are important for future BPM designs. Comparison of the measurement results from the DOROS system and the standard LHC system was presented in [79].

7 Oscillation subsystem results based on laboratory measurements

7.1 Synchronisation circuits

The DOROS front-ends accommodate circuits which allow to synchronize the ADC sampling clocks to the beam revolution. As described in chapter 4.4.3, the front-ends receive the f_{rev} reference clocks from the BST fibre distribution. However the phase relationship between the BST clocks and the beam is not guaranteed. Therefore in addition to the BST each front-end contains a local timing generator (LTG) circuit which produces f_{rev} reference based on the BPM signals. Phase of the BST clocks can be then aligned to the LTG reference using PLL-based phase shifters and a state-of-the-art differential phase detector (DPD). This chapter presents laboratory measurements used to characterise the LTG and DPD circuits and evaluate the precision of the phase alignment.

7.1.1 Scaling of the LTG input threshold

The local timing generator uses the pulses from the four BPM electrodes to produce the beam synchronous reference clock. The four electrode signals are first processed by the RF filters and amplifiers. The resulting signals are then summed in a passive combiner to generate pulse signal whose amplitude is independent of position change. Bandwidth of the signal is limited by the input RF filters to some 80 MHz bandwidth. Amplitude of the signal is proportional to the bunch intensity. The resulting beam pulses have to be converted to the beam clock which can be processed by the following logic circuits. This operation is done by a fast comparator whose threshold voltage can be set by a 12-bit DAC. In order for the LTG to operate reliably, the threshold level has to be adjusted according to the pulse signal amplitude. The beam signal level can be estimated using the output of the DOR channels which shares the input signals with the LTG.

The laboratory setup shown in Fig. 7-1 was used to measure the scaling of the LTG threshold with respect to the DOR signal level. The diagram depicts the signal processing on one of the two channels.

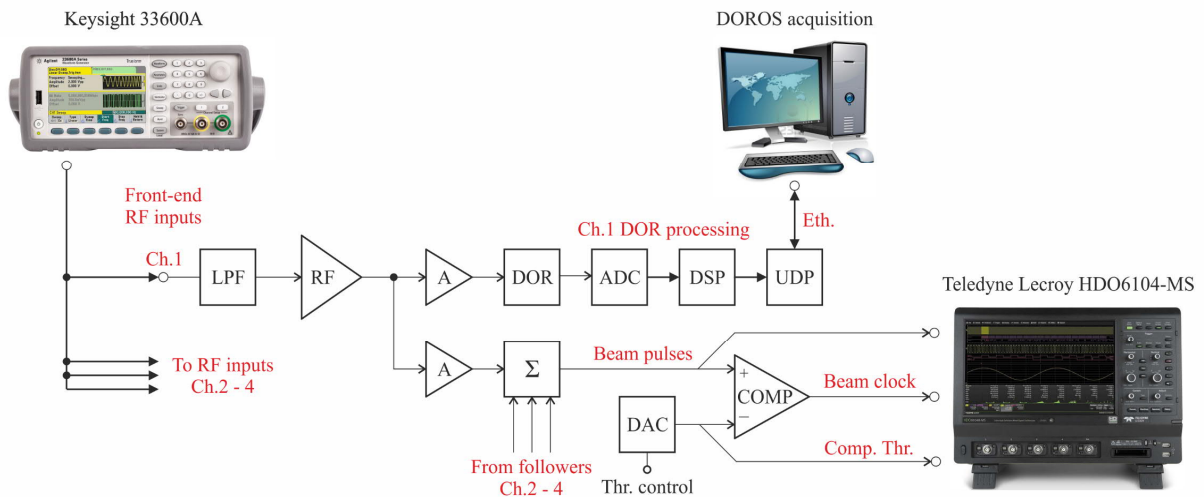


Fig. 7-1. Diagram of laboratory setup used to measure the scaling factor between the comparator threshold controlled by the DAC and the processed DOR signals. The signal generator Keysight 33600A was used to simulate beam signals pulses with 11.24 kHz repetition rate, 5 ns width and 1 Vpp amplitude. The maximum DAC threshold was found by measuring the simulated beam pulses using a 12-bit oscilloscope HDO6104 with high impedance inputs, 1 GHz bandwidth and 2.5 GS/s sampling. **Abbreviations:** LPF – Low Pass Filter, RF – RF amplifier, A – fixed gain Amplifier, DOR – Diode Orbit channel, COMP – Comparator, Thr – Threshold.

The waveform generator Keysight 33600A was used to simulate the 5 ns beam pulses with repetition rate equal to 11.2 kHz which is similar to the LHC f_{rev} . The signal was then split between the 4 RF inputs of the front-end. The RF gain was configured to some 25 dB so that the 1 V generator amplitude corresponds to a signal level of about 0.95 FS, close to the ADC saturation (equivalent to some 4.75 V). Please note, that the DOR channels and the signal combiner on the input of the LTG are decoupled by fixed gain amplifiers. A 12-bit oscilloscope HDO6104 measures the amplitude of the resulting beam pulses on the output of the combiner. The remaining oscilloscope channels were used to monitor the comparator output and its threshold voltage controlled by the DAC. Resolution of the DAC is 0.001 FS which according to the datasheet corresponds to 5 mV.

The scaling was evaluated for the maximum threshold voltage allowing the comparator to produce stable signals. This threshold voltage was found by increasing the DAC value until the signal on the comparator output became unstable or constant 0 V. The DAC value was then decreased by one step followed by the 1 minute-long measurements of different parameters parameter. The pulse amplitude was measured with the oscilloscope using the amplitude function. The threshold voltage was measured with oscilloscope using the mean function. The DOR signal level was measured as an average of the raw 25 Hz data acquired by the DOROS server. The measurement was repeated for a few generator amplitudes scanning the dynamic range of the DOR ADC with fixed RF gain.

Obtained results were summarized in Table 7-1. The amplitude of the beam pulses on the comparator input can be estimated by averaging the signal levels from the four DOR channels. Linear regression was used to find the scaling factor between values set in the threshold DAC and the average signal level measured by the DOR. The DC offset was estimated in the order of 10^{-4} and was considered to be negligible. The residual fitting errors were estimated in the same order of magnitude. The function which allows to compute the maximum threshold value for the DAC using the DOR signals is expressed as:

$$Thr_{Max} = 0.98 \frac{1}{4} \sum_{i=1}^4 DOR_{Ch.i} \quad (7-1)$$

where i is the DOR channel index and $DOR_{Ch.i}$ is the normalised electrode value. The actual threshold voltage measured by the oscilloscope was smaller by some 8 % when compared to the Thr_{max} scaled to volts. The maximum threshold voltage was measured to be approximately 90 % of the average pulse amplitude measured by the oscilloscope. The DAC threshold is usually operated at $\frac{1}{2} Thr_{max}$.

Changes of the comparator threshold or the beam pulse amplitude can cause a false phase change of the LTG reference clock. This effect was quantified in the laboratory using the measurement setup shown in Fig. 7-2. The signal generator Keysight 33600A produced 5 ns pulses with 1 V amplitude and 2.9 ns rise and fall times. The repetition rate was synchronized to the f_{rev} clock provided from the BST decoder. A passive splitter was used to provide the generator signal to 4 front-end inputs. The RF amplifier was set to some 28 dB so that the DOR signal levels can reach approximately 0.9 FS. The maximum threshold value Thr_{max} was computed from the DOR signals according to the equation (7-1). The DAC was then set to $\frac{1}{2} Thr_{max}$ and the phase difference between the reference clocks $f_{rev\ BST}$ and $f_{rev\ LTG}$ was zero-aligned by the fine phase shifter (PS). The differential phase detector was used to measure the phase shift between the reference clocks.

Table 7-1. Maximal comparator thresholds scaled to the ADC readings of the DOR channels.

Generator amplitude [V]	Avg. DOR signal level [ADC FS] *	Comparator input pulse amplitude [mV] †	Comparator DAC max. threshold	
			[DAC FS] **	[mV] †
1	0.977	495	0.096	440
0.75	0.728	391	0.072	331
0.5	0.480	251	0.047	218
0.25	0.230	128	0.023	110

*Rounded to 3 decimal places, **DAC resolution 0.001 FS (equivalent to 5 mV), † measured by oscilloscope.

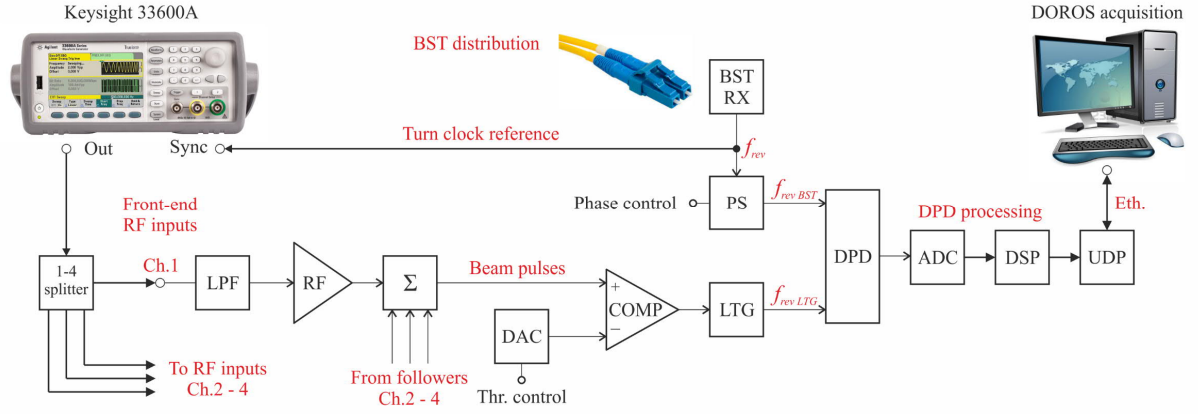


Fig. 7-2. Diagram of the laboratory setup use to measure the dependency of the reference phase from local timing generator on the input comparator threshold. The signal generator Keysight 33600A was used to simulate beam signals pulses with 5 ns width 1 Vpp amplitude and repetition rate synchronized to the LHC beam synchronous timing. **Abbreviations:** LPF – Low Pass Filter, RF – RF amplifier, F – Follower, DOR – Diode Orbit channel, COMP – Comparator, Thr – Threshold, PS – Phase Shifter, BST – Beam Synchronous Timing, DPD – Differential Phase Detector.

The threshold DAC was used to scan the threshold voltage of the comparator. The scan started at the minimum DAC value of 0.001 FS which was then linearly increased by 0.001 steps until the Thr_{max} value was reached. The phase difference was measured in each DAC step lasting about 6 s. The total time required to obtain measurement of a single scan was approximately 10 minutes. The scan was then repeated in reverse order to increase the statistics and demonstrate the reproducibility of the measurement.

The dependency of the LTG phase reference on the comparator threshold is showed in Fig. 7-3. The horizontal axis on the bottom represents the threshold DAC values normalised by the Thr_{max} . The horizontal axis on the top of the plot is scaled to the equivalent DOR signal level according to equation (7-1). The vertical axis on the left represents the DPD measurement scaled to the equivalent time difference between the LTG and BST reference clocks. The scaling is derived from the equation (4-45) using ideal power supply voltages ($V_{PS} = 3.3$ V and $V_{ref} = 2.5$ V). The vertical axis on the left represents the delay as a phase difference $\Delta\phi$ between the reference clocks. The 360° phase difference would correspond to one period of the reference clock T_{rev} . Please note that the zero value on the vertical axis corresponds to the maximum of the voltage-to-phase characteristic of the DPD.

It can be seen that a 70 % change in the DAC threshold or the DOR signal level shifts the reference phase of the LTG clock by an equivalent of 1 ns (equivalent to some 0.004°). Typically, the automatic gain control reduces the maximum change in DOR signals change to some 10 %. The timing circuits are operated with aligned reference clocks at the $\frac{1}{2} Thr_{max}$ DAC threshold. The equivalent delay between the reference clocks would be around 0.13 ns which is equivalent to some 0.0005° phase shift at T_{rev} . This can be also translated as a 0.0001° error in the phase advance measurement when using a 3 kHz excitation frequency. The perturbation is three orders of magnitude smaller than the required phase advance resolution.

Reproducibility of the measurement was evaluated by subtracting the measured values from the increasing (blue) and decreasing (blue) scan of the DAC. The largest difference of some 20 ps was measured around $\frac{1}{2} Thr_{max}$. Please note that the entire measurement took some 20 minutes to complete.

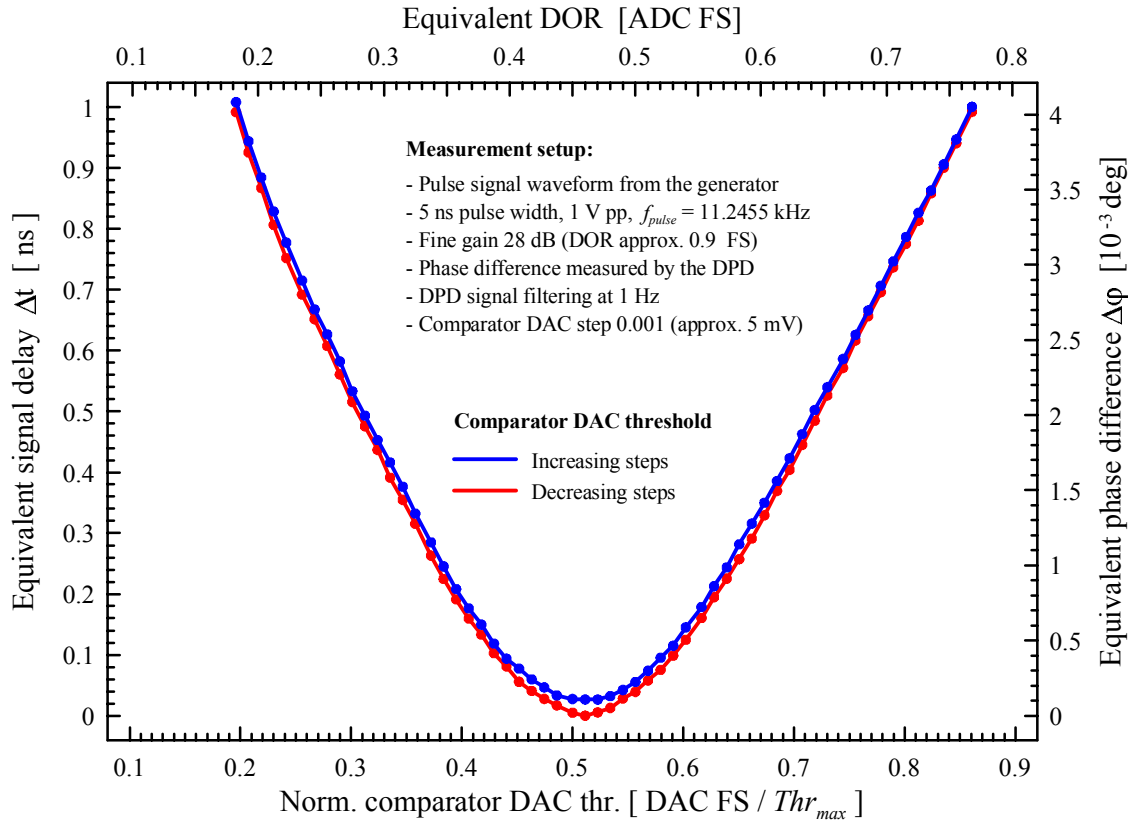


Fig. 7-3. Measurement of the phase shift in the LTG reference clock during a scan of the comparator threshold. The f_{rev} clocks from the LTG and BST were zero-aligned at the comparator threshold corresponding to 50 % of the input pulses. The Thr_{max} was estimated from measured DOR signal using (7-1). The DAC threshold was linearly increased 0.001 FS step (blue). The measurement was repeated with decreasing steps (red). The DPD signal processing was set to 1 Hz filtering. Zero value on the vertical axis correspond to the maximum of voltage-to-phase characteristic of the DPD.

7.1.2 Reference clock phase adjustment and stability

The phase difference between the LTG and BST reference clocks can be zero-aligned using the so called raw and fine phase shifters. As explained in chapter 4.4.3.1, the raw shifter is used for adjusting large phase offsets with some 6 ns step. The fine phase shifter based on an FPGA-integrated PLL is used for fine phase trims. The differential phase detector was designed to resolve the smallest steps and provide accurate detection of the zero-phase alignment. The measurement setup shown in Fig. 7-2 was used to measure the resolution of the DPD with the smallest available phase steps provided from the fine phase shifter.

The plot in Fig. 7-4 shows the DPD output signal measured during a phase scan using the fine phase shifter. Each change in the DPD signal corresponds to a 100 ps step. Reproducibility of the measurement was verified with two consecutive measurements using the increasing (blue) and decreasing (red) phase steps. The horizontal axis on the bottom corresponds to the measurement time during the increasing phase steps. The horizontal axis on the top corresponds to the measurement with decreasing steps which was reversed in time to overlap the two signals. The vertical axis on the left corresponds to the change in the DPD signal scaled to the ppm of the ADC full-scale (ppmFS). The zero ppmFS on the axis corresponds to the maximum value measured with the DPD. The vertical axis on the right was scaled to represent equivalent time difference between the two reference clocks. The scaling is derived from the equation (4-45) using ideal power supply voltages ($V_{PS} = 3.3$ V and $V_{ref} = 2.5$ V). The condition $\Delta t = 0$ corresponds to the zero-phase alignment achievable with the DPD. Please note that the actual maximum of the voltage-to-phase characteristic of the DPD is close to the zero on the vertical axis.

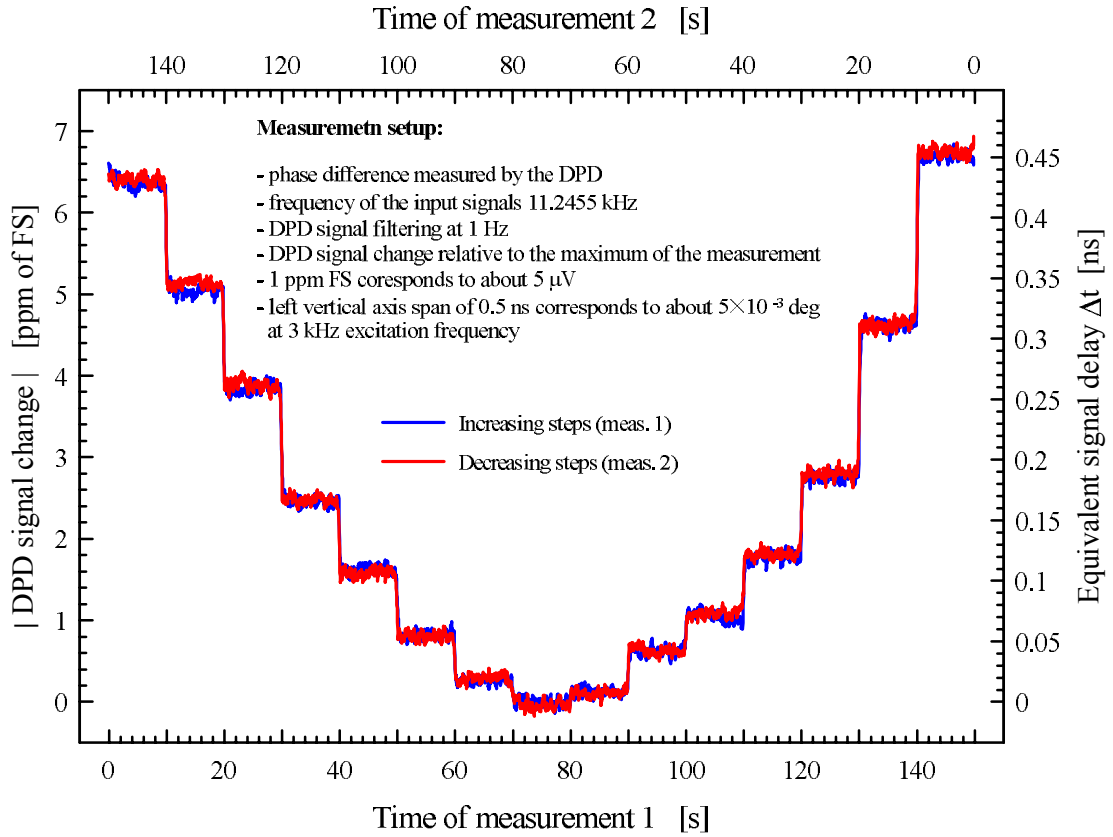


Fig. 7-4. Measurement of the reference clock alignment measured by differential phase detector. The phase shift of the reference clocks was controlled in 100 ps steps using a fine phase shifter based on FPGA PLL. The first measurement shows the linear phase scan with increasing steps (blue). The measurement was then repeated in opposite direction with decreasing steps (red). The cut-off of the IIR low-pass filter in DPD signal processing was set to 1 Hz. Zero value on the vertical axis correspond to the maximum of voltage-to-phase characteristic of the DPD.

A 10 s part of both measurements (equal to 250 S) around the 0 ppmFS was used to compute the standard deviation and peak-peak in order to evaluate the resolution of the circuit. Obtained results from both measurements are summarized in Table 7-2. The standard deviation yielded less than 4 ps_{RMS}. The peak-peak estimation of the DPD signal change yielded some 0.3 ppmFS which is equivalent to a clock jitter of some 20 ps. This is 5-times less than the resolution of the fine phase shifter.

It was observed that the size of the phase steps measured by the DPD is uneven. The change in the phase signal decreases when approaching maximum of the DPD characteristic. The non-linearity in the phase signal is systematic and does not depend on the direction of the phase step. Please note that the measured non-linearity is not critical for the application in the DOROS system. This is because the maximum of the DPD characteristic, where the two reference clocks are in phase, is not affected. The alignment error comes from the smallest available step of the phase shifter and long-term drifts of the DPD signal. However, the non-linearity needs to be taken into account when using the DPD circuit for measuring absolute phase differences.

Table 7-2. Noise measurement of the differential phase detector.

Measurement (70 - 80 s) †	Standard deviation *			Peak-peak *		
	[ppmFS]	[ps]	[10 ⁻⁶ deg] **	[ppmFS]	[ps]	[10 ⁻⁶ deg] **
Inc. (meas. 1)	0.058	3.279	3.541	0.277	18.650	20.142
Dec. (meas. 2)	0.058	3.928	4.242	0.288	19.371	20.921

* Rounded to 3 decimal places, ** equivalent phase at 3 kHz frequency, † close to zero-aligned phase difference.

The following measurements were focused on evaluating long-term stability and sensitivity of the DPD circuit to the induced temperature change. Diagram in Fig. 7-5 displays the laboratory setup used to evaluate the performance the DPD. During these measurements the DPD was operated at the maximum of the voltage-to-phase characteristic when the two input clocks are zero-aligned. As shown in the diagram, this condition is guaranteed by connecting the two DPD inputs to a same f_{rev} clock signal provided from a signal generator. The temperature change was induced using ventilation units arranged the same configuration as shown in Fig. 5-53 from the chapter 5.5. The ventilation unit located below the front-end was turned on during the measurements. The ventilation unit located above was used to generate the temperature change by remotely switching its fans on and off. The measurements were acquired from both DPDs in the front-end to increase the statistics. The temperature of the boards containing the DDP circuits was monitored with dedicated temperature sensors. Acquired temperature values and the processed DPD samples from the two boards were transmitted from the front-end to the DOROS server for further processing.

The plot in Fig. 7-6 shows the temperature measurement (top) along with measurement of the two DPD signals (bottom) acquired for 8 hours. The measurement started with stabilised temperature conditions. After some 4 hours the ventilation unit was switched on to induce the temperature change. The vertical axis on the left represents the change of the DPD signals scaled to the ppmFS. First sample was subtracted from each DPD measurement. Please note that 1 ppmFS is equivalent to some 5 μ V. The axis on the right was scaled in equivalent change in the time delay.

The estimated standard deviation and peak-peak change of the DPD signals is summarized in Table 7-3. The estimated peak-peak noise of the DPD signals yielded some 0.45 ppmFS which is equivalent to 30 ps jitter. This result can be also translated as a phase fluctuation of about 0.00003° at 3 kHz. At about 4th hour of the measurement ventilation unit on the top of the front-end was switched on. Consequently the PCB temperature decreased by some 3 $^\circ$ C. This temperature change caused the drift observed in the DPD signals. The DPD 1 signal decreased by some 0.8 ppmFS which is equivalent to 60 ps shift in the delay and the DPD 2 decreased by 0.2 ppmFS which is equivalent to 15 ps. The measurement was repeated with exchanged positions of the PCBs in the front-end. This resulted in the corresponding exchange of the measured change. This results indicate that the measured responses are linked to each individual DPD circuit, rather than a temperature gradient inside the front-end.

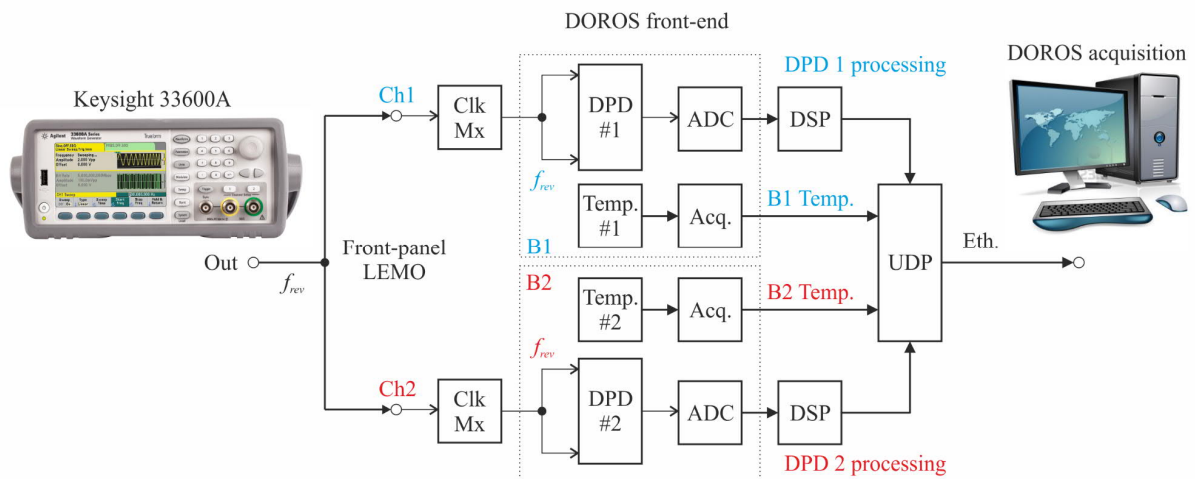


Fig. 7-5. Laboratory setup used to measure the long term stability and sensitivity of the DPD circuit to temperature change. The setup uses both DPD circuits available in the front-end. The reference clock $f_{rev} = 11.2455$ kHz was provided from a Keysight 33600A generator. The temperature of each of the two boards (B1 and B2) was measured by a dedicated temperature sensor. **Abbreviations:** Clk Mx – Clock Multiplexer, DPD – Differential Phase Detector, Temp.– An on-board Temperature sensor, Eth. – Ethernet protocol.

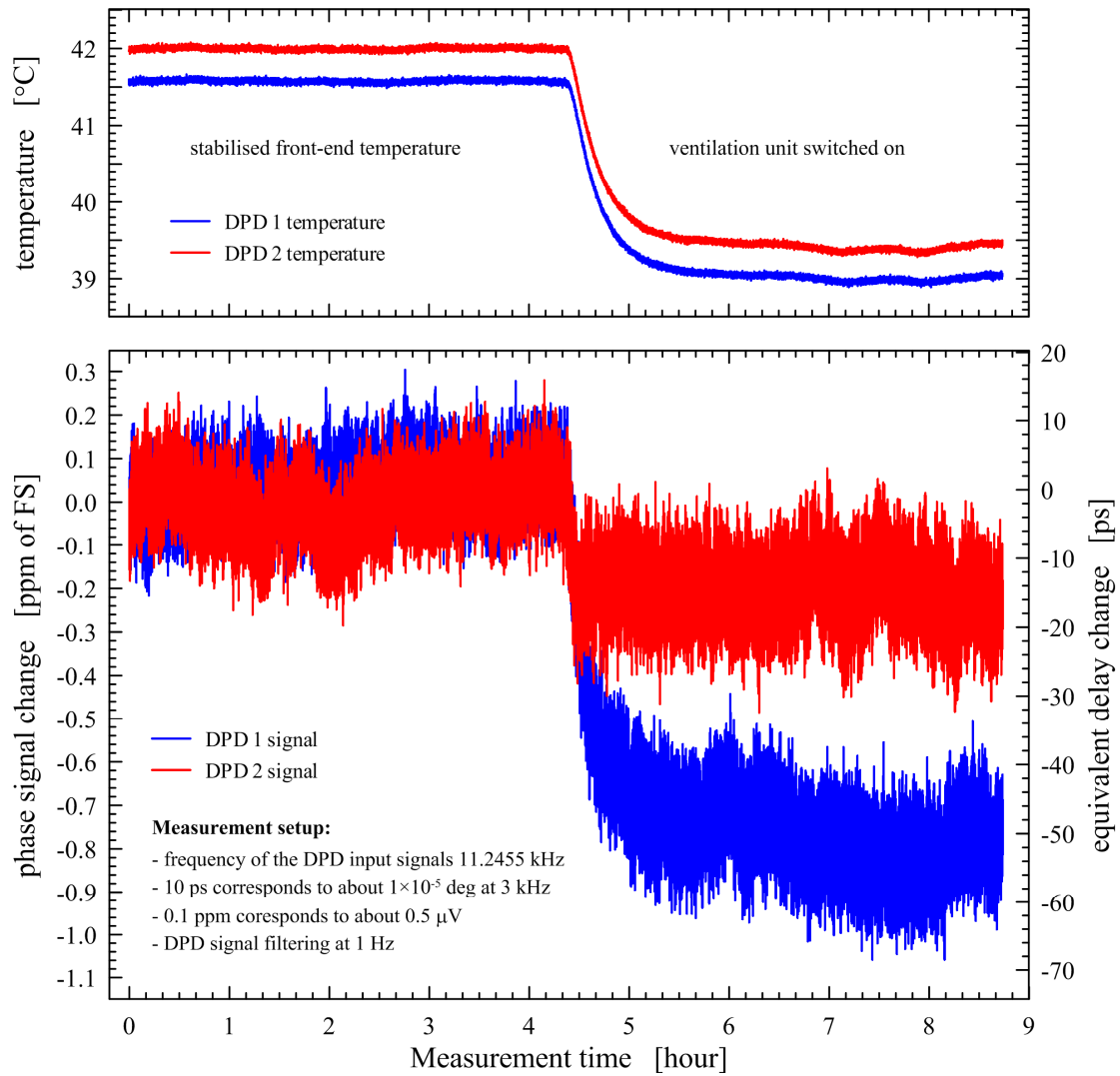


Fig. 7-6. Output signals of two differential phase detectors (DPDs) as digitised by their ADC when both detector inputs were driven by the same signal. The change in temperature was caused by the air flow from the ventilation unit located above the closed front-end box. The upper plot shows the temperatures of the boards accommodating the DPD circuits.

Table 7-3. Noise estimations of the differential phase detector signal.

Meas. time	DPD 1				DPD 2			
	Peak-peak *†		STD *†		Peak-peak *†		STD *†	
	[ppmFS]	[ps]	[ppmFS RMS]	[ps RMS]	[ppmFS]	[ps]	[ppmFS RMS]	[ps RMS]
1 min	0.228	15.402	0.045	3.023	0.234	15.795	0.051	3.407
10 min	0.367	24.784	0.058	3.906	0.334	22.568	0.06	4.065
1h	0.397	26.78	0.058	3.946	0.41	27.66	0.059	3.999
4 h	0.459	30.976	0.058	3.933	0.512	34.56	0.067	4.508
7-8 h	0.418	28.242	0.059	3.975	0.426	28.76	0.058	3.933

* Rounded to 3 decimal places, † measured with zero-aligned input clocks.

Reproducibility of the measurements was confirmed by repeating them several times. Thermal sensitivity of the worse DPD 1 was measured in the order of 0.3 ppmFS/°C which can be also expressed as a voltage change of 1.6 μ V/°C or equivalent time delay change of 22 ps/°C. This result is comparable to the thermal stability of the ADC channel measuring the DPD signal. As indicated in the ADC datasheet a typical offset drift is about 0.8 μ V/°C and the gain drift in the middle of the input range is about 3 μ V/°C. Please note that the observed drift may be also influenced by the ADC reference (1 μ V/°C), the operational amplifiers (0.2 μ V/°C) and their resistors (25 ppm/°C).

It was noticed that the DPD performance was significantly changed once the front-end case was open. The effect was investigated in more detail. The plot in Fig. 7-7 shows a consecutive sequence of five measurements, each lasting about 15 minutes, when the front end case was alternated between being covered with the top plate and open. It was observed that once the case is open, the estimated peak-peak noise in the DPD signals increased from the initial 0.4 ppmFS to some 2.2 ppmFS for the worse DPD 2, that is by more than 5 times. The noise returned to its previous value right after the case was again closed. When the case was open for the second time the air flow was forced by the ventilation unit. This time the noise of the DPD 2 increased to some 8.6 ppmFS. The resulting noise was 21 times larger than during the nominal conditions with closed front-end box. In this case, the resolution of the DPD would be limited to about 0.5 ns, which is equivalent to some 5 phase shifter steps. The best performance of the DPD and its related circuits is therefore achieved if the front-end is closed, preventing exposure to the air flow.

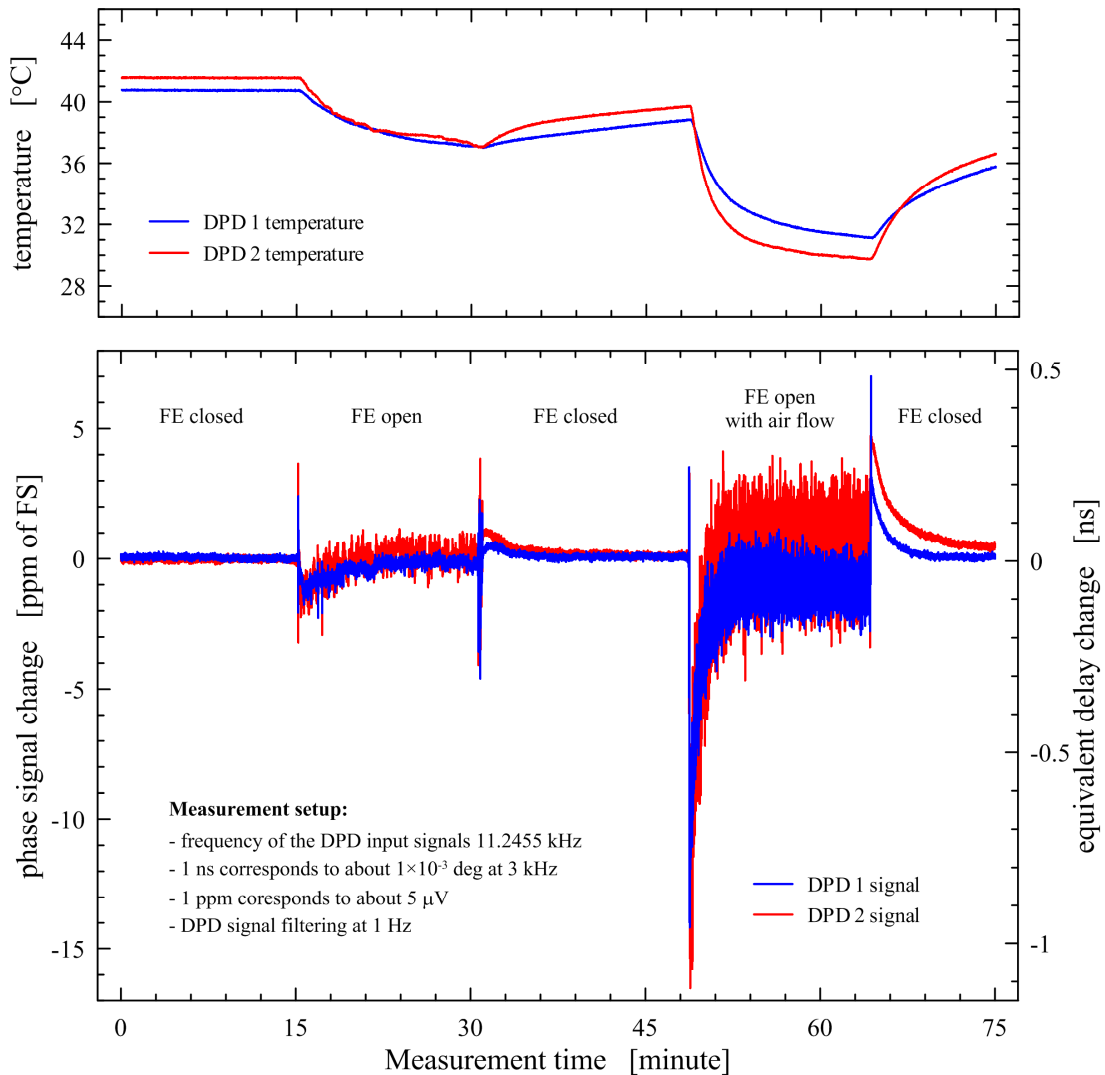


Fig. 7-7. Output signals of two differential phase detectors (DPDs) as digitised by their ADC when both detector inputs were driven by the same signal. The changes in temperature were caused by opening and closing the front-end (FE) case cover. When the cover was raised for the second time the air flow was forced by the fans from the top ventilation unit. The upper plot shows the temperatures of the boards accommodating the DPD circuits.

7.1.3 Summary and outlook

Measurements presented in this chapter were focused on the performance of the phase alignment system and some of its most important circuits. The timing system is crucial for the diode oscillation subsystem measuring precise phase advance and beta-beating of the excited beams. It was shown that the threshold level on the LTG comparator input can be optimised based on the orbit signals measured by the DOR subsystem. The threshold level is important for reliability of the locally generated timing reference used to synchronize the ADC sampling to the beam. It was shown that a change in the threshold level introduces a delay in the reference clock in the maximum range of approximately 1 ns. Proper setting of the threshold level according to the DOR signals improves robustness and immunity of the LTG reference clock to interferences. At the same time the threshold DAC can be used to control and minimize the systematic errors in the reference phase. Measurement results showed that it is possible to do the error adjustment with precision better than 20 ps, which is 5 times smaller than the available resolution of the fine phase shifters in the FPGA. The DAC could be thus used for fine trimming during the phase alignments.

The second part of this chapter was focused on evaluating the performance of the differential phase detector circuit. The phase detector was built and optimized to measure the small phase difference between two f_{rev} clocks for the purpose of their phase alignment using the phase shifters. Presented measurements demonstrated that the DPD circuit is capable of detecting the zero-phase condition with resolution well exceeding the minimum phase step of 100 ps. Peak-peak noise of the DPD signals was estimated at the zero-phase condition to some 30 ps. The nominal performance can be reached with the circuit enclosed inside the front-end chassis. High measurement stability allows to use the DPD in the application involving long term phase monitoring.

The adjustment of the phase was performed manually, which typically requires several minutes to complete. Automated procedure would allow to decrease this time to just a few seconds rendering the phase alignment process more efficient and applicable to any number for front-ends in the machine.

7.2 DOS detector sensitivity

This chapter is focused on evaluating the performance of the oscillation subsystem with the laboratory measurements. The most important parameter of the DOS processing is the sensitivity to small amplitudes at a dedicated frequency (typically the beam excitation frequency). Achieving high signal-to-noise ratio (SNR) with small beam excitation is important for high quality of the measurement as well as the machine safety. Presented measurements were focused on an oscillation signal at a single frequency generated synchronously to the sampling clock of the ADC.

7.2.1 Measurement setup

The laboratory setup showed in Fig. 7-8 was used to evaluate the performance of the oscillation subsystem. Waveform generator Keysight 33600A was used to simulate the betatron oscillations of the beam by using the amplitude modulation of the pulses. The modulation represents the beam excitation. The second channel of the generator was used to produce the reference clocks for the ADC sampling at $f_{rev} = 11.2455$ kHz. The amplitude modulation was set to a sinewave function synchronized to the same clock. The smallest available modulation depth in the generator is 0.01 % which would be equivalent to 1.5 μm oscillation on a 61 mm BPM aperture. The modulation frequency was synchronized with the ADC sampling.

The generator signal is provided to the front-end inputs through a variable attenuator and a splitter. The attenuator allowed to decrease the generator amplitude without modifying the modulation depth setting on the generator. The signals from the splitter were provided to all positive input channels on the front-end RF input. The negative inputs were terminated. The AM signal is demodulated by the diode detector and passed through the instrumentation amplifier to the low-frequency signal processing. The calibration switching was turned off during the oscillation measurements. The ADC data was acquired at the f_{rev} rate and stored to the DDR memory upon a trigger command received from the DOROS server. The data was then transmitted to the acquisition server. Measurement data was then further processed and analysed by offline tools. Please note that using only the positive inputs in this measurement setup decreases the sensitivity of the DOS subsystem by a factor two. The generator also contributes to the measured noise levels. When measuring the oscillations from BPM signals, the coherent noise of the beam pulses in each electrode pair would be rejected as common mode signal on the inputs of the DOS instrumentation amplifiers.

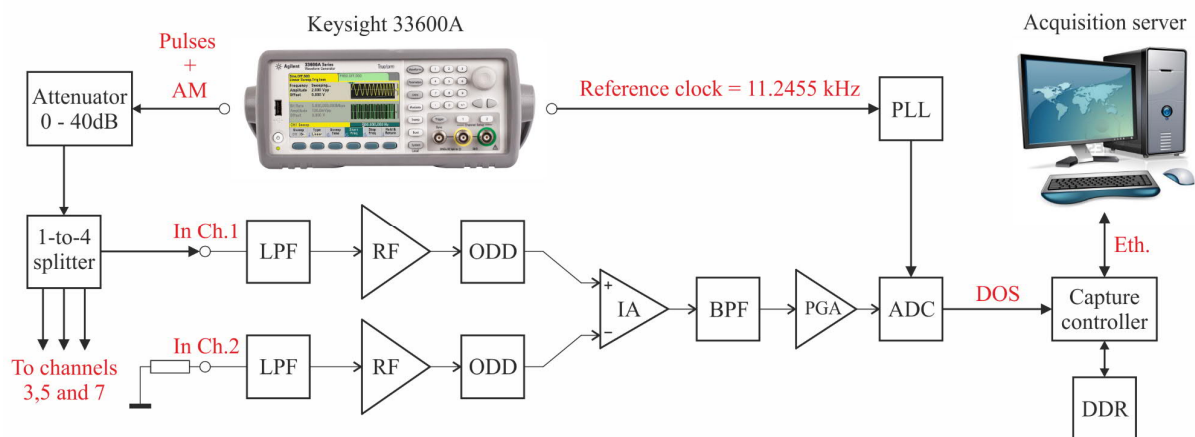


Fig. 7-8. Laboratory setup for measuring the sensitivity of the oscillation subsystem to the amplitude modulation of the pulsed signals from a generator. The first channel of the Keysight 33600A signal generator was used to provide the modulated pulses with 5 ns width and 1 V_{pp} amplitude. Minimum available modulation depth is 0.01%. The second generator channel was used to synchronize the ADC sampling to the generator. **Abbreviations:** AM – Amplitude Modulation, LPF – Low-Pass Filter, RF – Radio Frequency amplifier stage, ODD – Oscillation Diode Detector, IA – Instrumentation Amplifier, BPF – Band-Pass Filter, PGA – Programmable Gain Amplifier, DOS – Diode Oscillation signal, ETH – Ethernet network

7.2.2 Estimation of the signal-to-noise ratio

The signal-to-noise ratio (SNR) of the DOS signals was estimated based on the power spectral density (PSD) spectra obtained from the acquired data. The plot in Fig. 7-9 displays the PSD spectra measured with the three DOS time constants. The signal generator was configured to simulate a pilot beam with 11.2 kHz pulse repetition, 1 Vpp amplitude and 0 dB attenuator. The AM signal was set to 0.1 % depth at a frequency of $0.25 f_{rev}$, equal to approximately 2.8 kHz. These settings were chosen to represent typical beam excitation close to the LHC tunes [3]. The PSD was computed from 10 kS data acquired for approximately 1 s. Rectangular windowing was then applied to the data set before computing the PSDs. The measurements were repeated with the three available discharge rates in the DOS detectors. The DOS time constant T_{c0} corresponds to the slowest discharge and the T_{c2} to the fastest discharge. Please refer to chapter 4.4 for further details about the DOS time constants.

The measurement was then repeated with two different pulse repetition rates. The lowest pulse repetition rate $f_{pulse} \approx 11.2$ kHz represents a single bunch. The highest repetition rate $f_{pulse} \approx 40.1$ MHz represents a beam with 25 ns bunch spacing. The plot in Fig. 7-10 shows the PSD spectra of the DOS signals measured with the T_{c0} time constant at high and low pulse frequencies. In order to visualize the overlapping spectral peaks in the plot, the two spectres were shifted by 2 FFT bins. Both peaks seen at the $0.25 f_{rev}$ correspond to the AM signal demodulated by the diode detectors. This measurement was also repeated for the T_{c1} and T_{c2} .

As seen in PSD plots in Fig. 7-9 and Fig. 7-10, the signal power is concentrated in a single FFT bin. It can be also observed that the noise power is not equally distributed over the whole DOS bandwidth. In addition, the shape of the noise floor changes with the detector time constant. The SNR was therefore estimated for narrowband signals using the equation:

$$SNR_n = 10 \log_{10} \left(\frac{PSD_{signal}}{PSD_{average\ noise}} \right) \quad (7-2)$$

Where PSD_{signal} is the signal power in a single bin and the $PSD_{average\ noise}$ is the average noise power averaged over ± 100 bins around the PSD_{signal} . The bandwidth of the SNR_n estimation is equivalent to some 224 Hz. This method was used to estimate the SNR_n values from all the measured PSDs. Obtained results were summarized in Table 7-4. Variation of the SNR_n results between the DOS channels was measured to be smaller than 0.5 dB.

In order to verify the SNR_n scaling with the length of the processed data, each PSD spectra was computed upon 10 kS, 100 kS and 1 MS. This is equivalent to the acquisition time of approximately 1 s, 10 s and 100 s respectively. Please note that the DOS data acquisition was adapted to match the longest required measurement. The data for the shorter acquisitions is just a selected data subset of the long acquisition. Obtained results were summarized in Table 7-4. It can be seen that the estimated SNR_n can be improved by acquiring and processing the DOS signals for longer time. According to the measurement results from the table, increasing the FFT length by factor 10 improves the estimated SNR by some 10 dB independently of the time constant or the pulse repetition rate. The improvement follows the well know equation of the processing gain in an FFT.

It can be seen in the plot Fig. 7-10 that the DOS sensitivity depends on the pulse repetition rate which can be also translated as the number of bunches in the beam. The SNR_n estimated at the high repetition rate was by some 30 dB higher than the SNR_n estimated at the low repetition rate. Please note that the difference between the equivalent number of bunches is some 4 orders of magnitude. To achieve equivalent improvement with the same AM settings, the measurement time of a single bunch beam would have to be increased by a factor 1000. In this case the measurement time would increase from 1 s to some 15 minutes which is often not possible when exciting the beam.

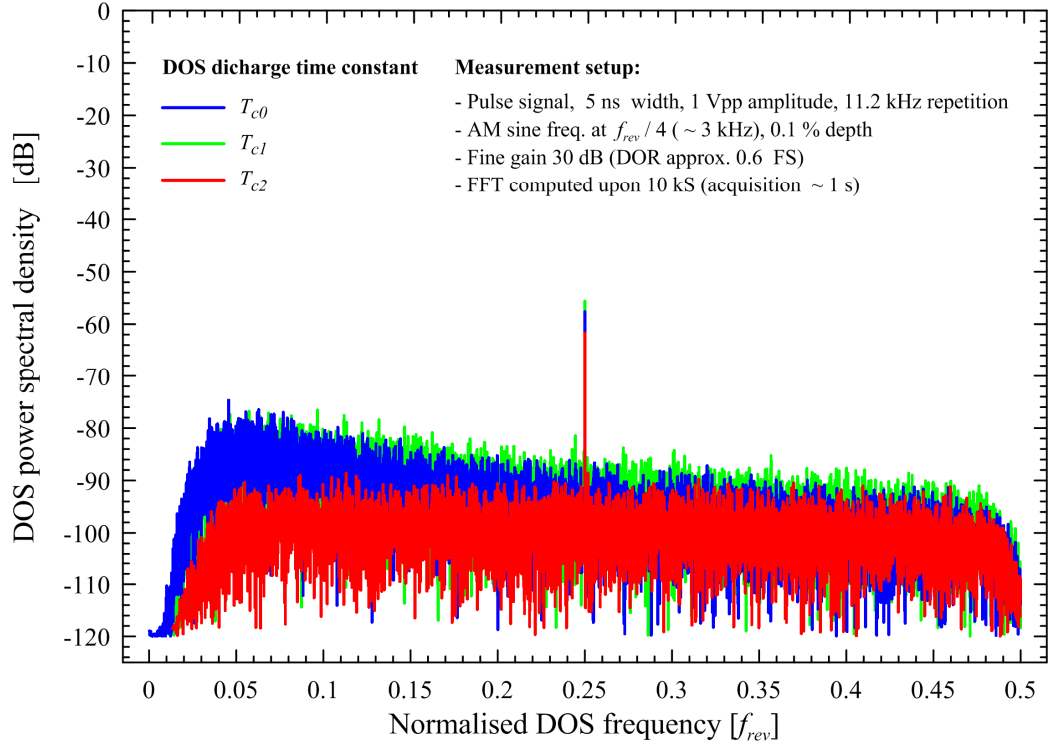


Fig. 7-9. Power density spectra of the DOS signals measured with the three configurable discharge time constants of the diode detector. The amplitude modulation of the 5 ns pulses from the generator was set to 0.1 % depth at frequency of $0.25 f_{rev}$ (approximately 2.8 kHz).

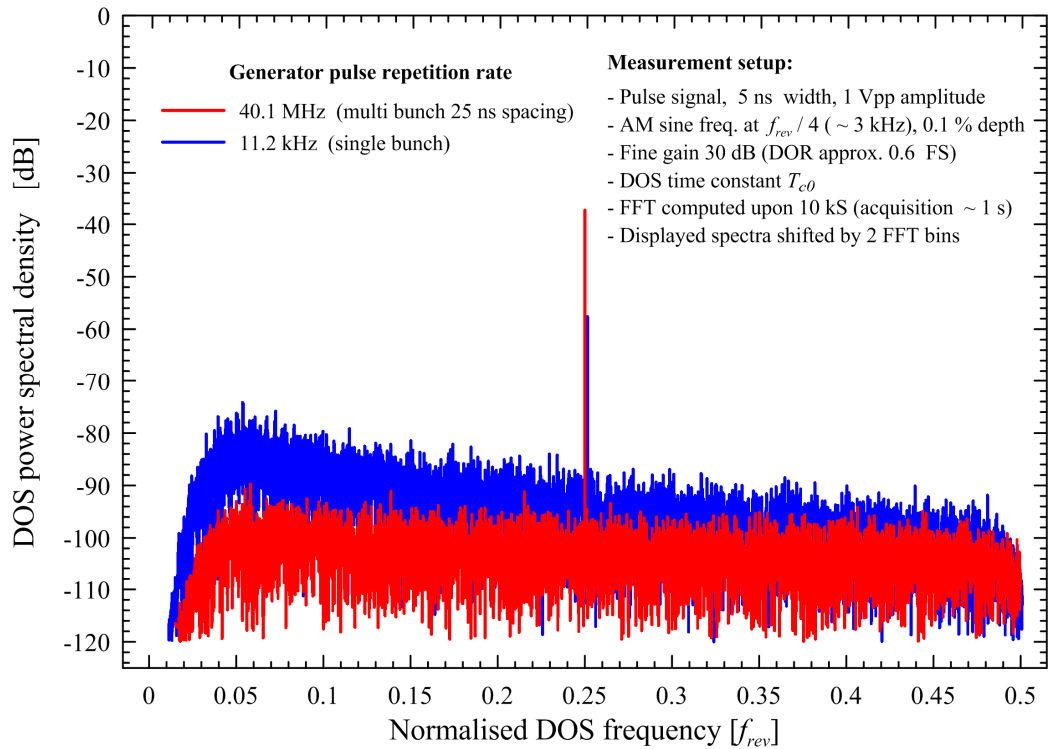


Fig. 7-10. Power density spectra of the DOS signals measured with the two different pulse repetition rates. The lowest pulse repetition rate $f_{pulse} \approx 11.2$ kHz represents a single bunch and the highest rate $f_{pulse} \approx 40.1$ MHz represents a beam with 25 ns bunch spacing. The DOS detector was set to T_{c0} . The signal generator provided 5 ns pulses with a sinewave AM configured to 0.1 % depth and frequency of $0.25 f_{rev}$ (~ 2.8 kHz).

Table 7-4. Estimated sensitivity of the DOS signals to a pulse AM signal of 0.1 % depth and frequency of $0.25 f_{rev}$.

Pulse repetition	FFT length [kS]	Acq. time [s] ** †	Estimated DOS SNR _n [dB] *		
			T_{c0}	T_{c1}	T_{c2}
11.2 kHz	10	0.89	35.7	36.2	35.6
	100	8.93	46.1	45.9	45.3
	1000	89.29	55.9	56.0	55.7
40.1 MHz	10	0.89	65.1	64.6	65.1
	100	8.93	74.6	74.8	75.0
	1000	89.29	84.8	84.8	84.9

* Rounded to 1 decimal place, ** Rounded to 2 decimal places, † assumed ADC sampling frequency of 11.2 kHz.

When comparing the estimated values in the Table 7-4 it can be seen that the DOS sensitivity does not significantly change with the different time constants. The same observation was made at both measured pulse repetitions at the normalised AM frequency of $0.25 f_{rev}$. For this reason the measurement was repeated for a number of AM frequencies to scan the DOS bandwidth in range from $0.05 f_{rev}$ up to $0.45 f_{rev}$. The measurements were performed at the 11.2 kHz pulse repetition simulating a single bunch in the LHC. The plot in Fig. 7-11 displays the obtained parts of the PSD spectra measured for each AM frequency with the DOS time constant T_{c0} . The plot in Fig. 7-12 displays similar measurement repeated with the DOS time constant T_{c2} . Each part of the PSD spectra was displayed in a bandwidth of ± 100 bins around the AM frequency. This is equivalent to a bandwidth of some 224 Hz. Please note that the same bandwidth is used for estimating the SNR_n. It was observed that the AM signals, measured as the spectral peaks in each PSD, change with the AM frequency. This is apparent especially in the measurement with time constant T_{c0} where the signal level was changing depending on the AM frequency by more than 10 dB. The same measurement at faster time constant T_{c2} yielded change in the signal level smaller than 2 dB. It was also noted that the measured AM signal level followed the shape of the noise floor.

The measurement data was used to estimate the SNR_n at each AM frequency. The plot in Fig. 7-13 displays the SNR_n results obtained from the measurements with T_{c0} and T_{c2} time constants and 11.2 kHz pulse repetition. The plot also shows the results from all DOS channels available in the front-end for comparison. It can be seen that the DOS sensitivity changed across the AM frequency range by less than 1.5 dB. The same measurement was also repeated at 40.1 MHz pulse repetition. Obtained results are shown in Fig. 7-14. The estimated DOS sensitivity changed in this measurement by more than 2 dB. The maximum was observed at the frequency range between 0.2 and $0.3 f_{rev}$.

Differences in DOS sensitivity measured with the T_{c0} and T_{c2} was in the order of 0.5 dB. This is comparable to the SNR_n variation seen between the DOS channels. In this case the sensitivity improvement gained from the DOS time constants can be considered negligible. Please note that the generator allows to simulate only the uniform intensity distribution for all bunches. In this case the DOS detectors with any time constant work in the regime when all the pulses contribute to the output signal. Differences in the sensitivity between the DOS time constants can become significant with important intensity variation between bunches in the LHC beam. In addition, the laboratory setup does not allow to measure the oscillations differentially. As a result the noise of the generator may dominate in the measurement result instead of the system noise of the DOS channels. This can be verified with a more complex measurement setup which allows to modulate the pulses with in-phase and inverted phase sinewave. Such setup is out of the scope of this thesis.

The DOS sensitivity is evaluated in the following measurements with different AM depth which is equivalent to changing the amplitude of the beam excitation. The modulation depth parameter was controlled in the generator settings. The plot in Fig. 7-15 displays the estimated SNR_n for the 11.2 kHz pulse repetition (blue) as well as the 40.1 MHz pulse repetition (red). The estimation was performed on a 10 kS data acquisition, which is equivalent to ~ 1 s. The vertical axis displays the estimated SNR_n expressed in decibel units. The horizontal axis on the bottom of the plot shows the modulation depth parameter in the generator in logarithmic scale.

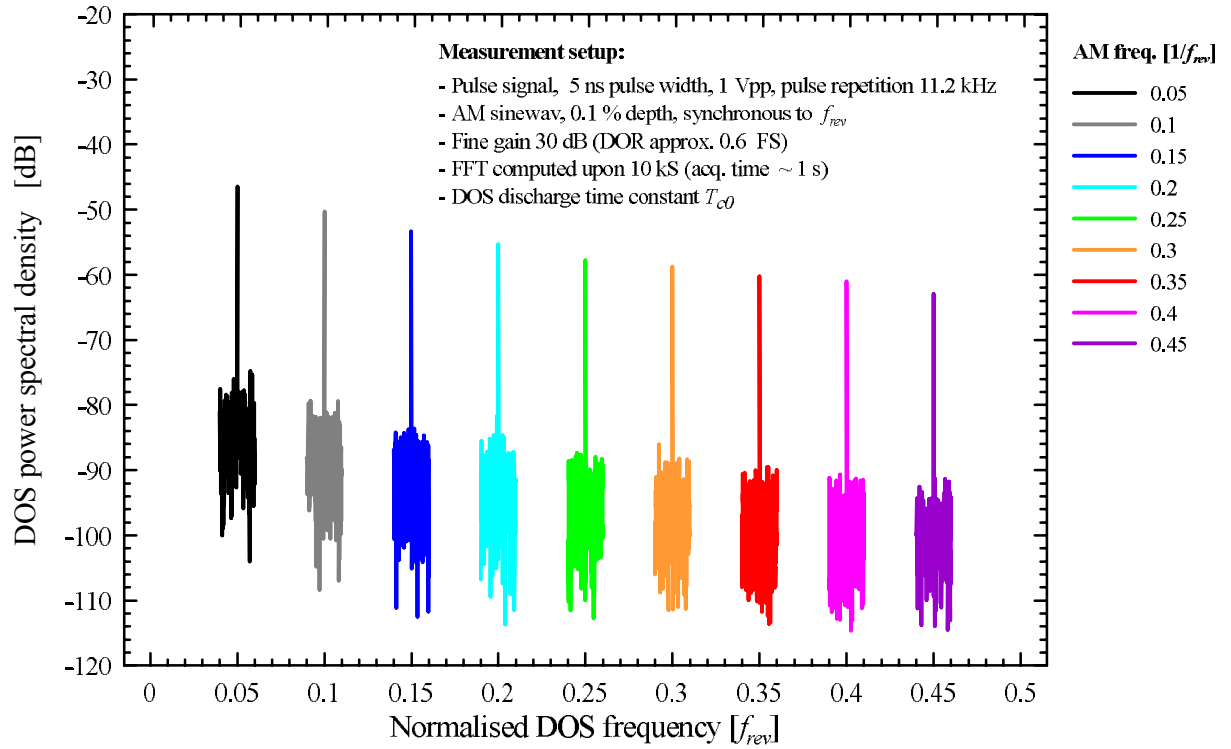


Fig. 7-11. Power density spectra (PSD) of the DOS signals measured at different frequency of the generator sinewave modulation simulating the excited beam oscillations. Each measurement is highlighted with different colour in the plot. Displayed parts of the PSDs were focused on ± 100 bins around the respective AM frequency of each peak (equivalent bandwidth of some 224 Hz). The DOS detector was configured with slow discharge time constant T_{c0} . The signal generator produced 5 ns pulses with sinewave AM at 0.1 % depth which was synchronized to the revolution frequency f_{rev} .

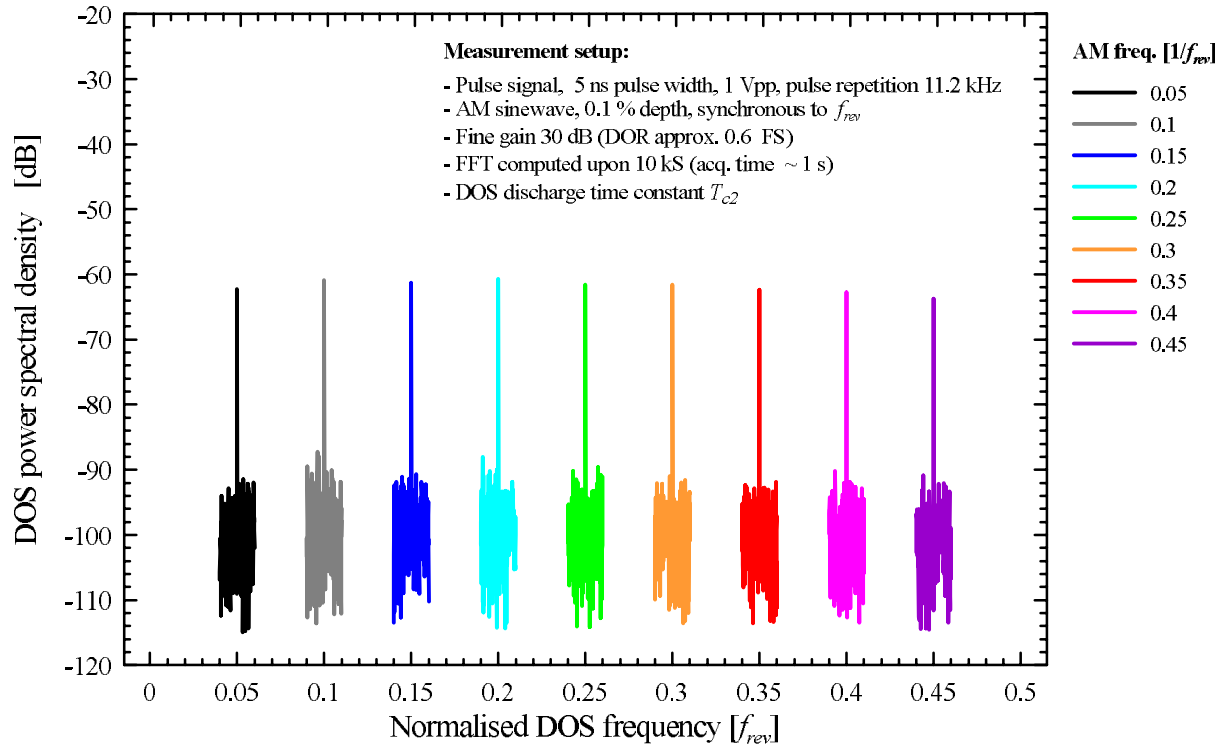


Fig. 7-12. Power density spectra (PSD) of the DOS signals measured at different frequency of the generator sinewave modulation simulating the excited beam oscillations. Each measurement is highlighted with different colour in the plot. Displayed parts of the PSDs were focused on ± 100 bins around the respective AM frequency of each peak (equivalent bandwidth of some 224 Hz). The DOS detector was configured with fast discharge time constant T_{c2} . The signal generator produced 5 ns pulses with sinewave AM at 0.1 % depth which was synchronized to the revolution frequency f_{rev} .

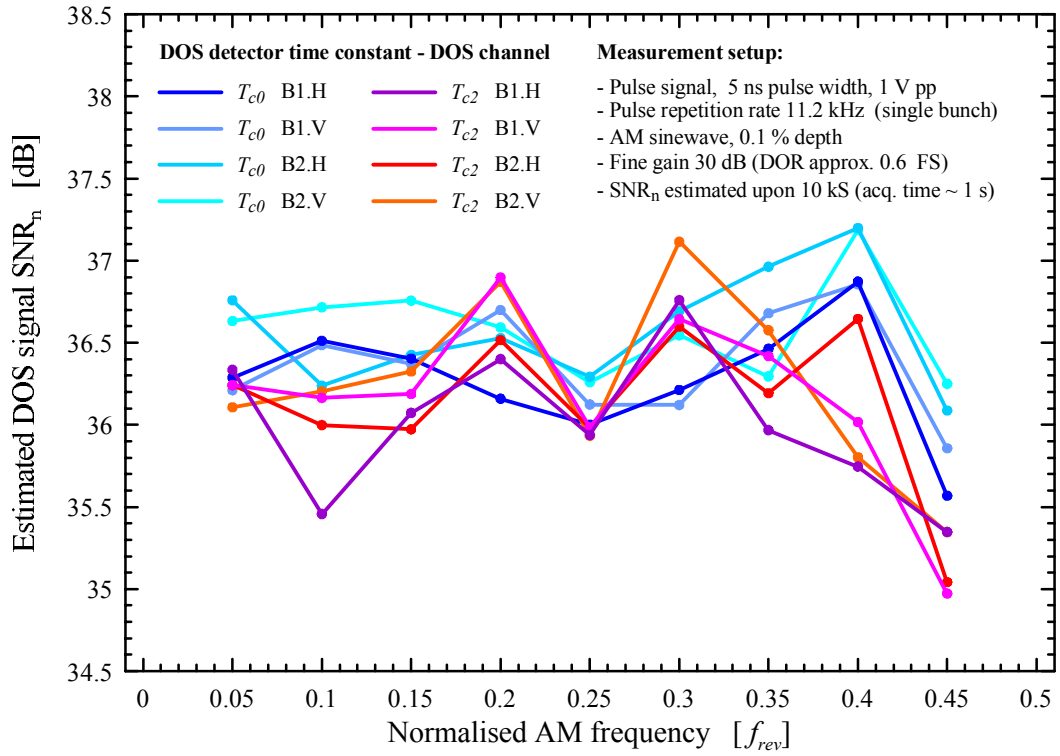


Fig. 7-13. Estimated signal-to-noise ratio of the DOS signals measured at 11.2 kHz pulse repetition with different AM frequencies. The AM frequency scan was measured with DOS the slowest times constant T_{c0} as well as the fastest time constant T_{c2} . The plot displays results from all four DOS channels in the DOROS front-end. The signal generator connected to the front-end inputs produced 5 ns pulses with sinewave AM at 0.1 % depth which was synchronized to the revolution frequency f_{rev} .

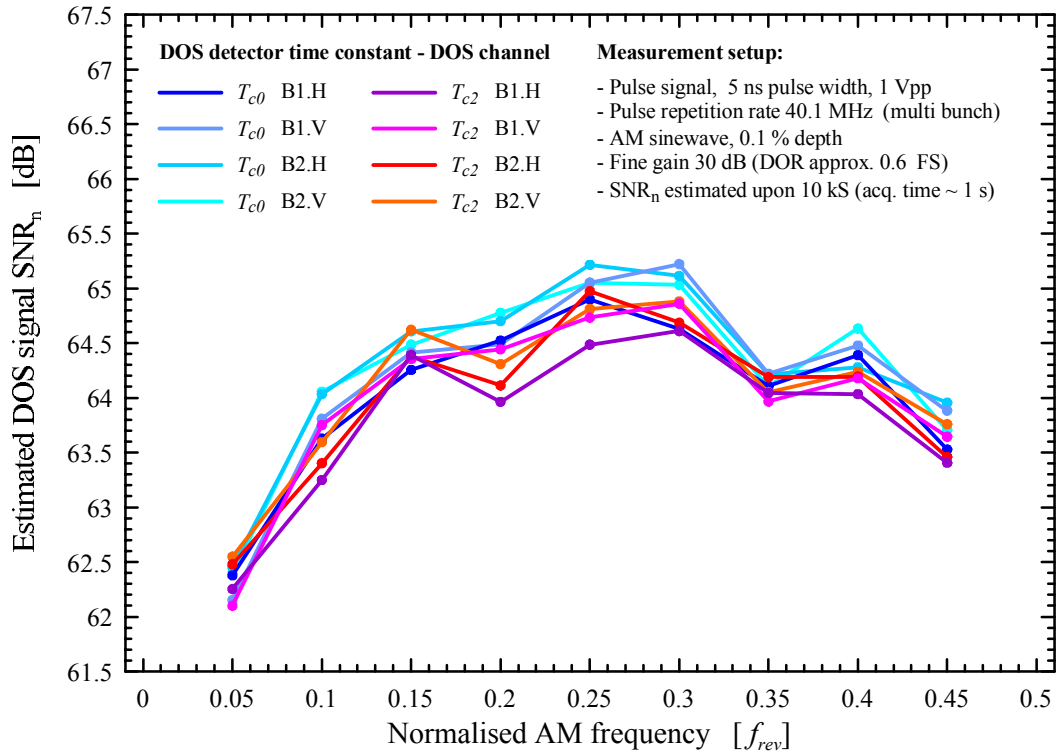


Fig. 7-14. Estimated signal-to-noise ratio of the DOS signals measured at 40.1 MHz pulse repetition with different AM frequencies. The AM frequency scan was measured with DOS the slowest times constant T_{c0} as well as the fastest time constant T_{c2} . The plot displays results from all four DOS channels in the DOROS front-end. The signal generator connected to the front-end inputs produced 5 ns pulses with sinewave AM at 0.1 % depth which was synchronized to the revolution frequency f_{rev} .

The horizontal axis on the top of the plot was scaled in the equivalent amplitude of the betatron oscillations (denoted as A_{exc}) expressed in micrometres. The conversion between the modulation depth and the A_{exc} can be expressed as:

$$A_{exc} = \frac{1}{2} \frac{D}{4} \frac{AM_{depth} [\%]}{100} \quad (7-3)$$

where D is the BPM aperture. In these measurements the conversion assumed to a 61 mm BPM aperture. The formula was derived using the equation (3-2) assuming linear position change. Factor $\frac{1}{2}$ in the equation (7-3) is necessary only for the purposes of this laboratory measurement which uses one of the two DOS differential inputs.

As seen in Fig. 7-15 a 10 μm oscillation of a single LHC bunch would yield a SNR_n of about 36 dB. Increasing the oscillation amplitude to 100 μm would result in SNR_n increase to some 56 dB, yielding improvement by some 20 dB. This is equivalent to increasing the data acquisition by two orders of magnitude that is from 10 kS (~ 1 acquisition time) to 1 MS (~ 100 acquisition time). It can be also seen in the plot, that the same SNR_n was achieved with 40.1 MHz repetition and just 2 μm oscillation amplitude. With this measurement setup it was possible to reach more than 90 dB SNR_n when the generator AM depth was reaching close to 1 %, equivalent to some 70 μm oscillations.

Dependency of the DOS sensitivity to the repetition rate was measured with DOS detectors set to time constant T_{co} . The generator AM was set to 0.1 % depth and 0.25 f_{rev} frequency. The plot in **Fig. 7-16** shows the estimated SNR_n in decibels. The horizontal axis on the bottom of the plot corresponds to the repetition frequency of the generator pulses showed on logarithmic scale.

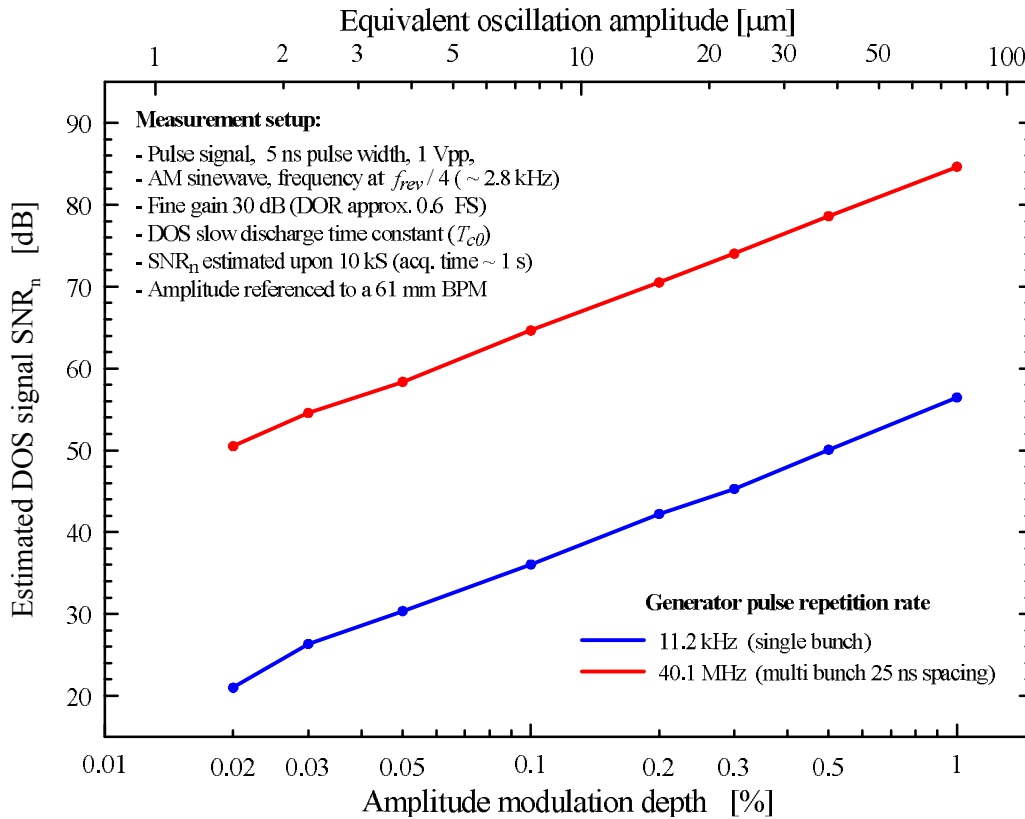


Fig. 7-15. Estimated DOS signal-to-noise ratio for different amplitude modulation depth set on the signal generator. The 11.2 kHz repetition rate of the generator pulses represents an LHC beam with a single bunch and the 40.1 MHz rate represents a physics beam with 25 ns bunch spacing. The DOS detector was set to slow time constant T_{co} . Frequency of amplitude pulse modulation was set to 0.25 f_{rev} (~ 2.8 kHz).

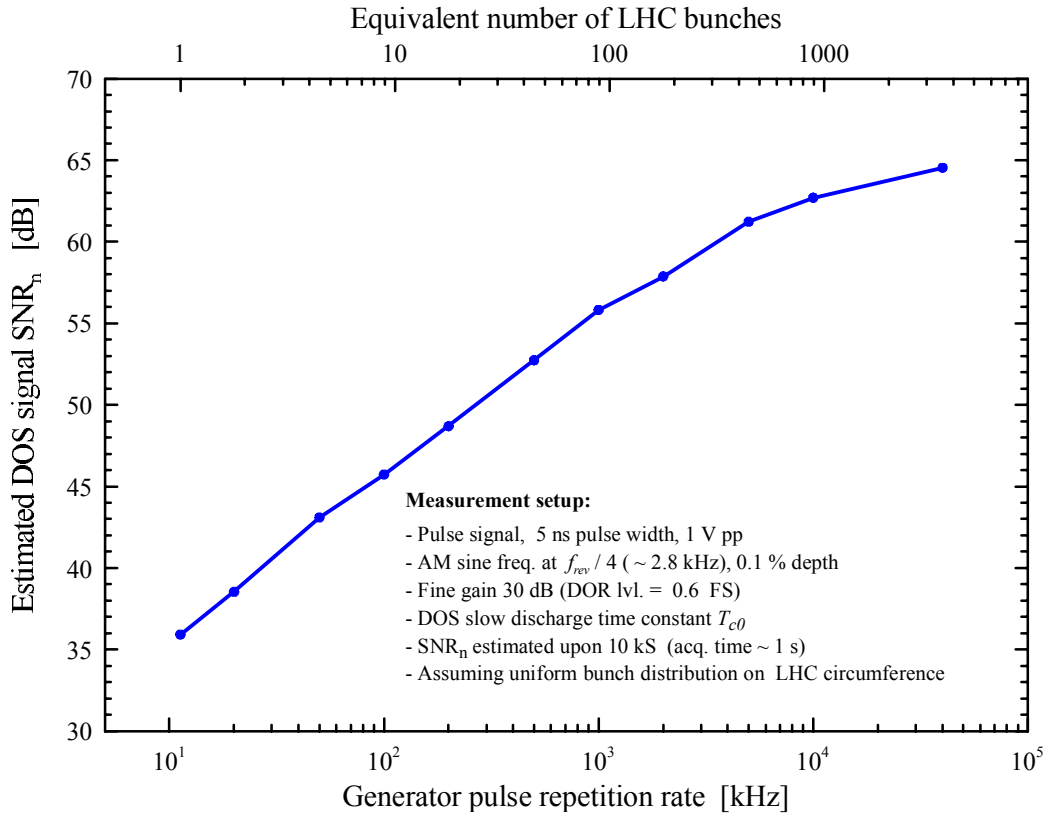


Fig. 7-16. Estimated DOS signal-to-noise ratio for different pulse repetition frequency set on the signal generator. The lowest 11.2 kHz pulse repetition represents a single bunch. The 40.1 MHz repetition frequency represents a beam with 25 ns bunch spacing. The equivalent number of bunches on the top horizontal axis assumes uniform bunch distribution on the LHC circumference. The DOS detector was set during the measurement to T_{c0} . The amplitude modulation of the generator was set to 0.1 % depth and frequency of $0.25 f_{rev}$ (~ 2.8 kHz).

The horizontal axis on the top of the plot corresponds to the equivalent number of bunches in the beam, assuming uniform distribution on the LHC circumference. A pulse repetition of 11.2 kHz corresponds to 1 bunch. A pulse repetition of 40.1 MHz would correspond to 3564 bunches. Please note that the LHC beams contain less than 2808 bunches. This is because the bunch trains always contain an area without bunches, so called abort gap, which is required for safe beam extraction from the accelerator [3].

It can be seen in **Fig. 7-16** that increase of the pulse repetition frequency from 11.2 kHz to 10 MHz results in increase of the estimated SNR_n by some 30 dB, that is some 10 dB per decade. Further increase in the frequency from 10 MHz up to 40 MHz results in SNR_n increase of about 10 dB, that is some 4 dB per decade. Please note that the average signal level on the DOS detectors also changes with the repetition rate. The change was measured with the DOR channels to be around 10 % of the ADC full-scale. This effect may be caused by the AC coupling in the RF stage.

The following measurement was focused on evaluating the DOS sensitivity with different RF amplifier gain. The signal generator simulated a single LHC bunch with 5 ns and 11.2 kHz repetition frequency. The generator AM was set to 0.1 % depth and $0.25 f_{rev}$ frequency. This is equivalent to some 10 μ m beam oscillation. Amplitude of the pulses was set to 4 Vpp, which is the maximum available value provided by the generator. The RF amplifier was then configured to +20 dB to set the signal levels measured by the DOR detectors to around 0.5 of the ADC full-scale. External attenuators were used to change the amplitude of the input signal in 10 dB steps. This measurement method allowed to decrease the amplitude of the generator pulses while keeping the modulation depth constant. The signal attenuation was then compensated with equivalent increase of the RF gain to maintain constant signal level on the diode detector inputs.

The plot in Fig. 7-17 displays the SNR_n values estimated for the different RF gains. The horizontal axis on the top of the plot corresponds to the attenuation value connected to the generator output. The horizontal axis on the bottom corresponds to equivalent RF gain setting used to compensate the signal attenuator.

The lowest DOS sensitivity was estimated at high gain settings which are typically required with the low intensity pilot beams. The SNR_n was estimated to less than 25 dB when using more than 50 dB RF gain. High intensity beams typically used for physics would require much lower RF gain, typically between 20 and 30 dB. In this case, the SNR_n was estimated around 35 dB. Please note that maximal variation of the SNR_n estimation between the DOS channels was smaller than 1 dB. The signal levels measured by the DOR subsystem varied between the measurements by less than 10 % of the ADC full-scale. As the calibration switching was turned off during the measurement the asymmetries of the RF channels were not compensated.

The 3 dB discrepancy in the SNR_n value seen between 30 dB and 40 dB RF gain can be explained by additional insertion loss in the channel. Such losses depend on the number of the RF stages being used. The error in the low gain settings was measured to be in the order of 5 dB using the oscilloscope measurements in each RF stage. Please refer to the chapter 4.3.2.1 for more details on the RF stage configuration at different gain settings.

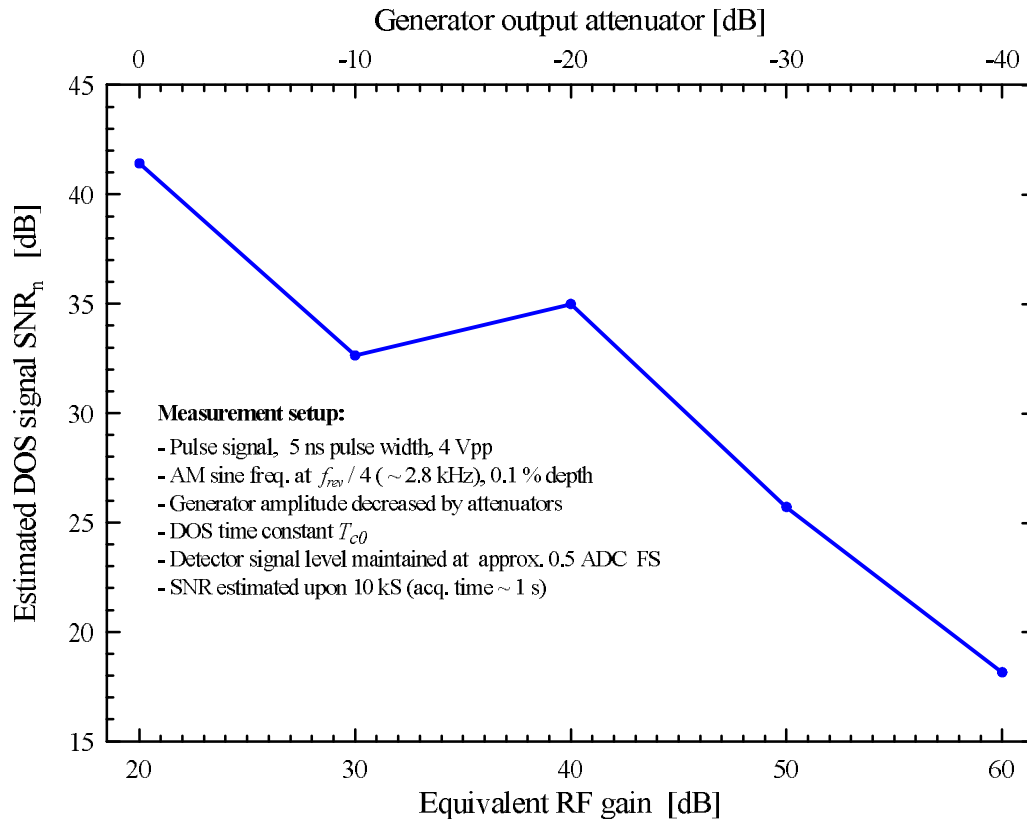


Fig. 7-17. Estimated DOS signal-to-noise ratio for different gain setting in the RF amplifier. The generator pulses were attenuated using external attenuators. The attenuation was then compensated by the RF amplifier. The signal level measured by the DOR subsystem was maintained at approximately 0.5 ADC full-scale. The DOS detector was set to T_{c0} . The signal generator provided 5 ns pulses with 11.2 kHz repetition and amplitude modulation. The modulation was set to 0.1 % depth and frequency of $0.25 f_{\text{rev}}$ (~ 2.8 kHz).

7.2.3 Summary and outlook

The oscillation subsystem allows to measure small amplitude oscillations of the BPM signals. The main purpose of the DOS subsystem is to detect the excited beam oscillations at a single frequency. Measurements presented in this chapter were focused on characterisation of the DOS channels with a laboratory generator. The generator simulate the BPM signals during a number of different beam conditions. The measurement results showed, that the signal-to-noise ratio of the narrowband DOS signals depends on a few parameters:

- the pulse amplitude modulation depth, which represents the oscillation amplitude,
- the pulse repetition rate, which represents the bunch spacing the bunch pattern,
- the generator signal amplitude, which represents beam intensity,
- the FFT length, which represents duration of the beam oscillation (e.g. beam excitation).

It was estimated that a single nominal bunch in the LHC, with some 10 μm oscillation would yield signal-to-noise ratio of the oscillation measurement of some 40 dB if the measurement took approximately 1 s. By increasing the number of bunches to the full capacity of the accelerator, the same signal-to-noise ratio could be reached with less than 1 μm oscillation. This means that the beam excitation during the optics measurements can be decreased by an order of magnitude. To improve the signal-to-noise ratio even further, the excitation can be applied for longer period of time.

The laboratory measurements also showed that the noise floor in the frequency domain changed depending on the DOS detector time constants and the pulse repetition frequency. The measurements showed that the oscillation amplitude also followed the noise floor changes. The resulting signal-to-noise ratio estimated from the measurements did not significantly change with the oscillation frequency or the detector time constant. Observed effect may have an impact on measurements which require flat amplitude response of the DOS detectors, such as the coupling measurements. It was observed that the fastest DOS time constant T_{c2} provided the flattest signal spectra, minimising the relative amplitude differences for changing oscillation frequencies.

8 Oscillation subsystem results based on beam measurements

The oscillation subsystem in the DOROS front-ends can be used to measure beam oscillations. To measure the local betatron coupling or the phase advance, the beams are excited under special conditions on a dedicated frequency. This chapter is focused on the measurements of excited beam oscillations with the standard DOROS system installed on the most important BPMs around the LHC experiments. The measurement setup in these BPMs allows to compare the performance of the DOS subsystem with the standard LHC BPM system (WBTN) which is typically used for the optics measurements.

The first beam measurements with the DOS subsystem were performed during the local beam coupling measurements in 2015. At that time the LHC beams were typically excited by means of an AC dipole magnet. The device can produce beam oscillations in the order of a millimetre. The excitation from the AC dipole could be applied for a few thousand turns which was enough to reach sufficient SNR with the WBTN system. Additionally, the excitation could not be synchronized to the beam revolution frequency f_{rev} . Beam parameters such as the intensity, number of bunches and energy had to be limited so that the accelerator can be operated in safe conditions.

Another way to excite the beam in the LHC was through the transverse damper system (ADT). The main advantage with respect to the AC dipole is that the excitation could be synchronized with the revolution frequency and applied for arbitrary number of beam turns. However, the ADT can produce smaller excitation amplitudes than the AC dipole. Sufficient SNR with the WBTN system could be achieved only with the ADT with excitation amplitudes close to maximum. One of the main advantages of the WBTN system is that the oscillation measurements are directly scaled in millimetre unit. This is not the case for the DOS subsystem which removes the position proportional part of the spectra (DC) in order to gain sensitivity to small oscillations at higher frequencies ($0.05 - 0.5 f_{rev}$). To overcome this limitation, the relative amplitudes of DOS measurements can be calibrated using the estimated amplitudes from the WBTN system. The BPM setup was also used to estimate the amplitude of the DOS oscillation signals in micrometres.

8.1 DOS sensitivity to small beam oscillations

An example of beam oscillation spectra is shown in Fig. 8-1 in the horizontal plane and Fig. 8-2 in the vertical BPM plane. The data was obtained during the local beam coupling measurement in the LHC. The signals were acquired from both the DOROS and the WBTN systems connected to the same stripline BPM located on the right side of the ATLAS experiment (IP1). The beam consisted of a single pilot bunch at the injection energy of 450 GeV. The ADT system was used to excite harmonic beam oscillations simultaneously in both transverse planes. The excitation frequency was set to $f_H = 0.27 f_{rev}$ horizontal plane and $f_V = 0.32 f_{rev}$ in the vertical plane. The amplitude was set to $A_V = 1$ and $A_H = 1$, which corresponds to the maximum amplitude allowed by the setup. The data was acquired by both BPM systems for 50 kTurns (~ 5 s acquisition), which is the maximum acquisition limit set by the WBTN system. The DOS subsystem was set to the fastest discharge constant T_{c2} . The computation of the displayed magnitude spectra was based on the FFT with rectangular windowing. As seen in both plots, the beam response was concentrated in a single FFT bin.

The oscillation amplitudes from the DOROS system were first normalised by the maximum measured magnitude in the spectra. The corresponding axis is shown on the right side of the plot. Obtained normalised amplitudes of the DOS spectra were then scaled to micrometres using the maximum oscillation amplitude measured by the WBTN system. The corresponding axis is shown on the left side of the plot. Please note that the displayed results were not scaled by the actual betatron function at the particular position of the BPM on the LHC circumference. Obtained results therefore don't correspond to the actual excitation amplitude applied in the ADT, as the betatron function at its location may be different.

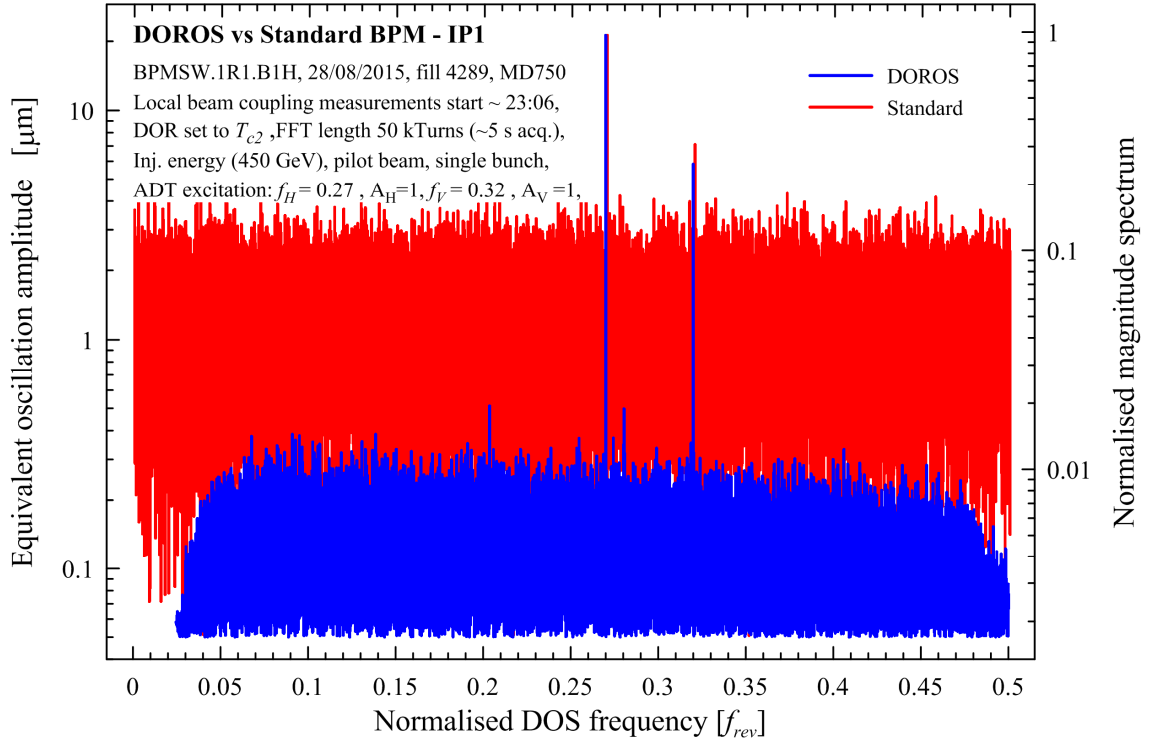


Fig. 8-1. Magnitude spectra obtained during the local betatron coupling measurements in horizontal BPM plane on the right side of ATLAS experiment (IP1). The beam was transversely excited by the transverse damper (ADT) with maximum available amplitude in both horizontal and vertical axis. The FFT was computed on 50 kS. Measured magnitudes were scaled to the maximum amplitude estimated with the WBTN system.

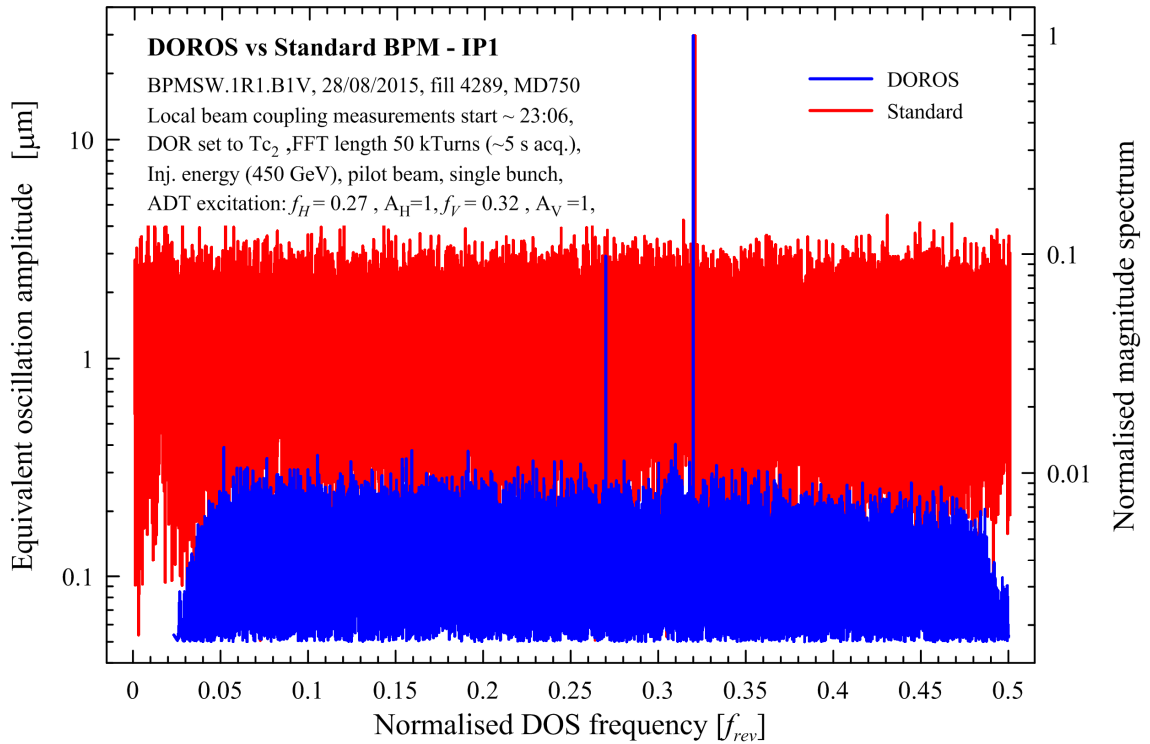


Fig. 8-2. Magnitude spectra acquired during the local betatron coupling measurements in vertical BPM plane on the right side of ATLAS experiment (IP1). The beam was transversely excited by the transverse damper (ADT) with maximum available amplitude in both horizontal and vertical axis. The FFT was computed on 50 kS. Measured magnitudes were scaled to the maximum amplitude estimated with the WBTN system.

As can be seen in the plots, the achieved sensitivity of the DOS subsystem allowed to detect and measure the excited oscillations as well as the coupling products from the perpendicular plane. The noise floor of the DOS measurement was estimated to around 0.1 μm . The noise floor of the WBTN system was estimated to be around 1 μm . The measured signals were used to estimate the signal-to-noise ratio of both the excited and coupled oscillations using the method described in 7.2.2. The estimation bandwidth was chosen to $\pm 0.01 f_{rev}$, (approximately ± 100 Hz) around each spectra peak. Obtained SNR_n values from the two measurement systems were summarized in Table 8-1. The same table also contains the equivalent amplitudes in micrometres measured by the WBTN system. These values were used for the scaling the normalised DOS values in micrometres. The difference between the two systems was estimated to more than 20 dB. The coupling component in the horizontal plane could be measured only with the DOROS system. The estimated SNR_n in this case was about 27 dB.

Table 8-1. Comparison of the estimated SNR_n for the WBTN and DOROS systems.

Local betatron coupling	WBTN				DOROS	
	Osc. amplitude [μm] *		SNR_n [dB] * †		SNR_n [dB] * †	
Norm. f_{exc} [f_{rev}]	H plane	V plane	H plane	V plane	H plane	V plane
	0.27	0.32	0.27	0.32	0.27	0.32
H plane osc.	21.24	-	23.21	14.01	44.56	33.46
V plane osc.	-	29.71	-	26.36	27.76	48.23

* Rounded to 2 decimal places, † SNR_n estimated upon 50 kS with average noise power in ± 100 Hz range around the peak.

An example measurement of the DOS sensitivity with the three time constants was acquired during the LHC optics validation after technical stop in 2016. Measurement results were obtained with the DOROS front-end connected to the stripline BPM located on the left side of the ATLAS experiment. The beam consisted of a single pilot bunch at the injection energy of 450 GeV. The beam was excited in both transverse planes using the ADT system. The excitation frequency was set to $f_H = 0.26 f_{rev}$ horizontal plane and $f_V = 0.32 f_{rev}$ in the vertical plane. The excitation frequency in the horizontal plane was set closer to the LHC tune by 1 % as opposed to the previous measurement. The amplitude was set to $A_V = 1$ and $A_H = 1$, which corresponds to the maximum amplitude allowed by the setup. The DOS acquisition was set to 100 kTurns (~ 10 s acquisition time).

Acquired oscillation signals were used to compute the PSD spectra showed in Fig. 8-3 for the horizontal plane and in Fig. 8-4 for the vertical plane. It can be seen that the shape of the noise floor changes with the DOS time constants. Estimated SNR_n were summarized in Table 8-2 together with the magnitude values measured at the respective excitation frequencies. The amplitude of the beam oscillations follow this change which results in very small SNR_n change of less than 0.5 dB in horizontal plane and about 1.3 dB in the vertical plane. The fast detector discharge T_{c2} provided the most flat amplitude response which is suitable for the computing the beam coupling in this case.

Table 8-2. Estimated SNR_n and amplitudes from three beam measurements with the different DOS detector time constants.

Local betatron coupling	SNR_n [dB] * †				Normalised amplitude [ppm of FS] *			
	H plane		V plane		H plane		V plane	
Norm. f_{exc} [f_{rev}]	0.26	0.32	0.26	0.32	0.26	0.32	0.26	0.32
DOS T_{c0}	49.04	28.30	21.02	44.83	113.52	0.66	0.18	31.39
DOS T_{c1}	48.56	28.53	20.17	44.46	164.28	1.12	0.25	46.87
DOS T_{c2}	48.34	28.89	19.34	44.13	42.12	0.43	0.05	15.57

* Rounded to 2 decimal places, † SNR_n estimated upon 100 kS with average noise power in ± 100 Hz range around the peak.

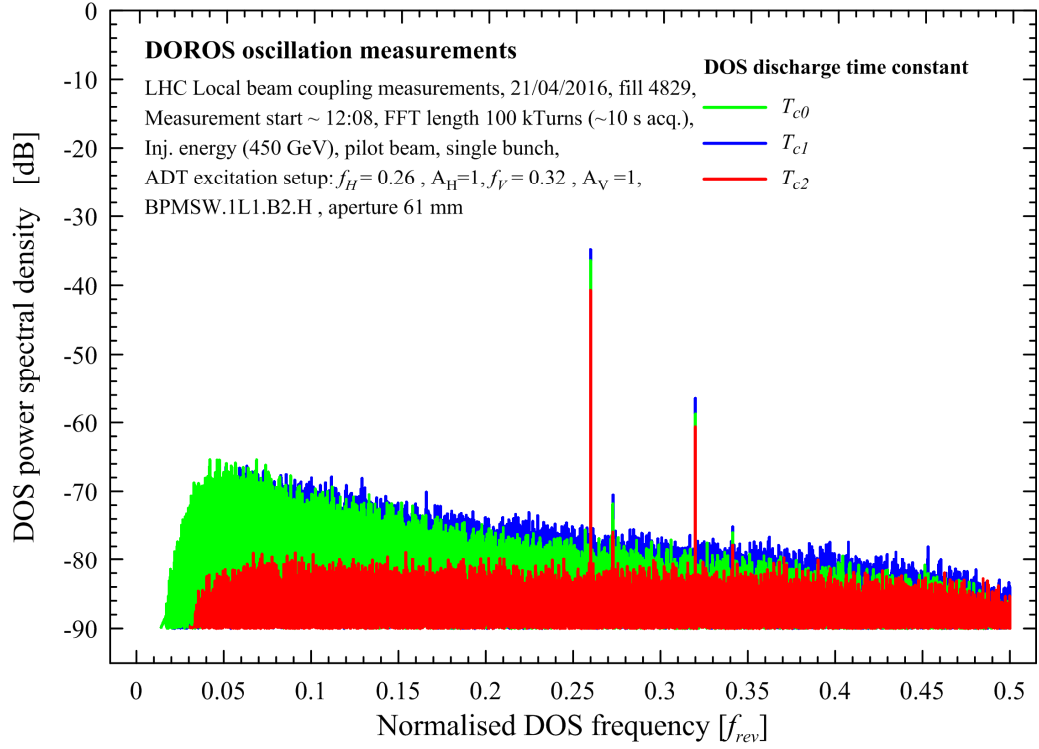


Fig. 8-3. Power spectral density of the DOS signal measured with beam in the horizontal axis of the BPM located in left of PT1. A single low intensity bunch was excited using the ADT system set to the maximum amplitude. The measurement was repeated for the three detector discharge constants.

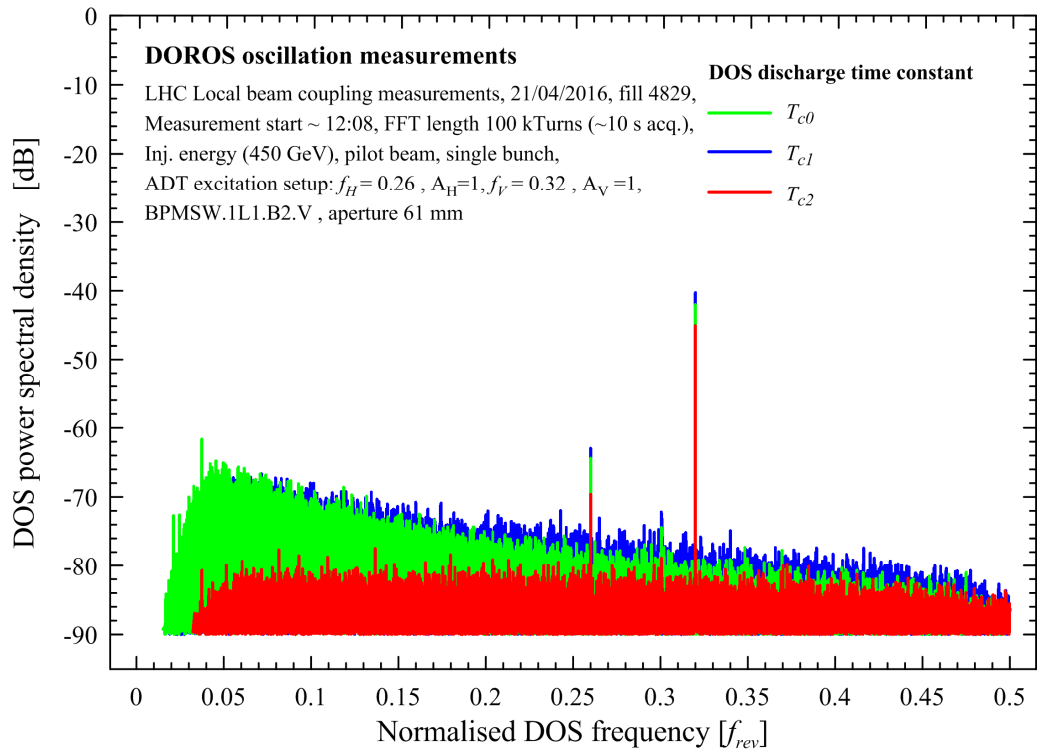


Fig. 8-4. Power spectral density of the DOS signal measured with beam in the vertical axis of the BPM located in left of PT1. A single low intensity bunch was excited using the ADT system set to the maximum amplitude. The measurement was repeated for the three detector discharge constants

8.2 Summary and outlook

This chapter presented the performance of the DOROS system measuring excited oscillations of the LHC beams. Discussed results were obtained during the local betatron coupling measurements performed with a single pilot bunch at the injection energy of 450 GeV. The beam was excited with the transverse damper system at the frequencies close to the LHC tunes. Both the excitation and the ADC sampling of the beam signals were synchronous to the beam revolution which allowed precise detection of the beam oscillations and simplified the data analysis. Obtained results were compared to the standard LHC BPM system (WBTN) measuring the beam oscillations with the same BPM sensor.

As the DOS part of the DOROS system measures only the position signal changes, the WBTN system was used to estimate amplitude of the oscillations in micrometres. The estimation was based on 5 s acquisition (50 kTurn) with the maximum available excitation amplitude. Obtained results were used to scale the magnitude spectra measured by the DOS subsystem in micrometres. Noise floor of the DOS subsystem was estimated to an equivalent amplitude of 0.1 μm achieving an order of magnitude improvement with respect to the WBTN system. The SNR of the DOS subsystem in the used measurement conditions was estimated to about 40 dB, improving the measurement by some 20 dB in comparison to the WBTN.

The oscillation measurements were used to compute and correct the local beam coupling in the LHC. Obtained results of the preliminary test were published in [82], [83] and [84], showing good agreement between the DOROS and WBTN systems. The coupling measurements are typically performed during dedicated time, scheduled only for a few times per year. The beam measurements with the DOS subsystem were thus focused on its sensitivity. Based on the promising results it is planned to perform the local coupling measurements using small excited oscillations at higher beam energies and a few high intensity bunches in the LHC. Other beam oscillation measurements with the DOS subsystem which would require the zero-phase alignment, such as the phase advance and beta-beating, are currently under study.

One of the limitations of the DOS subsystem is that it cannot be used for long acquisitions during regular operation with the orbits calibration switching. Multiplexing of the electrode signal on the RF amplifier inputs perturbs the DOS signals with the switching transients. The FFT length is then limited to less than one multiplexing period which is typically less than 1 s (10 kTurns). Consequently, the frequency resolution and sensitivity of the oscillation measurements are limited in such mode of front-end operation. Disabling the calibration switching for the duration of the DOS acquisition would cause discrepancies in the orbit measurements due to asymmetries between the DOR channels. Mitigation techniques used to minimise such orbit discrepancies are currently under study.

Another limitation of the DOS subsystem is that the measured oscillation amplitudes cannot be directly scaled in millimetres. In the presented measurement examples the WBTN system was used to obtain the scaling factors. However, the scaling can be obtained with sufficient quality only with large beam excitation amplitudes which may be limiting factor due to the required quality of the beam, measurement conditions as well as safety of the machine. In addition, the oscillation amplitude also scales with the betatron function of the LHC magnet optics depending on the location of the BPM on the ring [19]. This has to be taken into account when obtaining the calibration coefficients for the systems which don't share the same BPM signals with the WBTN system, such as the DOROS collimator system. Further development is thus focused on using the DOR subsystem measuring the absolute beam positions for calibrating the oscillation amplitudes measured by the DOS subsystem.

9 Summary of results and future outlooks

The main subject of this doctoral thesis was to develop and optimise an electronics system called DOROS. The BPM signal processing in this system is based on two diode detector types which were separately optimised for beam orbit and oscillation measurements of the LHC particle beams. As described in chapter 4.1, the DOROS project started as a simple circuit built in 2009 showing that the diode detectors could be used for BPM signal processing. The following prototyping and first beam measurements showed that a BPM system based on the diode detector techniques could reach resolution of the position measurements better than a micrometre. In the beginning of the doctoral project in 2013 it was decided to develop and deploy a fully operational system dedicated to precise orbit measurements for automatic alignment of the LHC collimator jaws with embedded BPM electrodes. These are critical devices assuring safety and reliability of the LHC. At the same time it was foreseen to test the system on a few selected BPMs around the experiments in the LHC to see if such approach could be used for precision measurements. There the resolution and stability of the position readings is important for a number of measurements, crucial for optimising the luminosity in the experiments and pushing forth the performance limits of the LHC machine. The first DOROS front-ends were installed and commissioned for beam operation in 2015 and more were added over the years of operation. The hardware and firmware in the DOROS front-ends is identical to that of the collimator system which simplifies the maintainability of the two systems.

As of beginning of 2018, the current collimator DOROS system consists of 20 front-end units equipping 22 LHC collimators with the jaws-embedded BPMs located in LHC points 1, 2, 5, 6, 7 and 8. The system is used for precise beam-based alignment of the jaws. This technique allowed to reduce the collimator setup time from several hours to just a few seconds. The system has been used during many machine studies [29], [30], [74] and [76], which were targeting the performance and reliability improvements of the LHC beam cleaning system. Other studies are focusing on the so called wire collimators and their precise alignment for compensating the beam-beam effects during the collisions in the experiments [77]. In addition, the position readings from a few selected DOROS front-ends are used by the LHC interlock system to terminate the LHC operation with beams if the beam positions measured in the collimators exceed safe limits. This functionality became operational since September 2017 without a single malfunction which would trigger a false beam dump and prevent beam injection into the LHC. A lot of effort was put into optimising the system reliability and robustness on both the hardware and software layers. In addition the interlocked system was made redundant with two DOROS front-ends for each collimator BPM to prevent blocking the LHC operation in case of a hardware failure. Each front-end performs a number of sanity checks, including its FPGA-embedded system running an real-time operating system as well as all Ethernet communications with the data concentrators and control servers. In case of a detected hardware or software failure, the front-end can automatically reset and resume operation within some 10 s. The current MTBF estimate of the DOROS system reaches more than 60 years.

The current standard DOROS system consists of 21 front-ends which equip 40 BPMs in the LHC. The system has been installed on the most important Q1 BPMs around the four LHC experiments as well as on a few selected Q5, Q6 and Q7 BPMs located in LHC points 1, 2, 4, 5, 6, and 8. The signals from these BPMs were split between the standard LHC BPM and the DOROS system allowing to evaluate and compare their performance. The standard DOROS system is operationally used for optimisation measurements such as the Van der Meer scans or local betatron coupling measurements. The system was also used during many LHC beam studies such as β^* levelling [63], stripline BPM directivity [79] or beam oscillations at frequencies below 100 Hz [78]. Most of these studies are targeting the LHC high luminosity upgrade necessary for increasing its discovery potential. A recent beam study is focusing on a non-invasive measurements of the quadrupolar moment during the beam energy ramp using the BPM signals. Both the standard and collimator DOROS systems have become important beam instrumentation tools used for the LHC beam diagnostics and performance optimisation.

The main contributions and results of this dissertation work can be summarized as follows:

- **Design, development and optimisation of a new BPM signal processing system based on diode detectors.**

As described in chapter 4, the DOROS system consists of front-ends realised as 19" 1U rack units which can process signals from eight BPM electrodes. The BPM signals are first filtered and amplified before they can be processed by the diode detectors. The RF amplifier consists of four stages controlled by the FPGA to provide variable dynamic range of about 90 dB controlled in 1 dB steps. The beam orbit and oscillation signal processing is then addressed by two separate subsystems, each containing specifically optimised diode detector processing. The analogue signals from each subsystem are digitized by an 8-channel 24-bit sigma-delta ADCs and processed in DSP channels implemented in the FPGA logic. The results are then transmitted over Ethernet network to the respective servers. The data transmission at the 25 Hz rate is synchronous to the external triggers received over the LHC BST networks. The ADC sampling is synchronized to the LHC beam revolution frequency f_{rev} of approximately 11.2 kHz using the BST clocks phase-aligned to the local clock reference generated from the BPM signals. At the same time the raw ADC data is stored at the f_{rev} rate in the circular buffers in the local DDR memory. The data is then available for capture or post-mortem analysis. The FPGA system controls the functionality of the front-end such as the calibration switching, automatic gain control, test signal generator, synchronization circuits, status monitoring or the fail-safe reprogramming and recovery system.

- **Design and optimisation of signal processing dedicated to the beam orbit measurements in the LHC based on compensated diode detector technique.**

The most important innovation of the new system is the compensated diode detector technique, to the author's knowledge used for the very first time in beam instrumentation applications. This technique is used in the so called DOR subsystem in the front-end for measuring beam orbits and low frequency beam oscillations. A typical BPM sensor in the LHC can produce pulses with time duration of about 1 ns, repeating every 25 ns. Depending on the bunch intensity and the BPM type, the peak amplitudes can change by some four orders of magnitude, typically between 0.1 V and 100 V. Measurement of a 1 μ m position change on a 100 mm aperture BPM would require resolving a change of the peak amplitude in an order of 10^{-5} , which is equivalent to a voltage change of 10 μ V on 0.1 V peak amplitude. As described in chapter 3, the RF diode detectors have a number of advantages for measuring small beam position changes:

- fast response,
- sensitivity,
- no external triggering required,
- simplicity and robustness.

It was shown in chapter 3.3 that beam positions computed from the RF diode detector signals depend on the voltage drop seen across the diode detector which can change according to the current flowing through the diode, temperature or the production spread. The compensation scheme minimizes these dependencies, reducing the temperature drifts and offset errors in the measured beam positions. Precision and accuracy of the measurement are influenced by the gain and offset asymmetries between the respective analogue channels. In order to minimise such asymmetries using only the beam signals, a so called automatic calibration switching method was developed. As explained in chapter 4.3.1, this technique is based on periodic signal multiplexing between the front-end analogue channels followed by further signal treatment in the DSP and a calibration algorithm computing the calibrated positions on the remote server. The automatic calibration switching minimizes the systematic position errors which depend on the RF gains as well as drifts in the analogue electronics. The calibration switching is important especially for the collimator jaw alignments or the long-term measurements during the LHC stable beams.

The dynamics of the diode detectors can be changed with their time constants controlled by the FPGA system. Three DOR time constant settings T_{c0} , T_{c1} and T_{c2} are available. The T_{c0} is the slowest and the T_{c2} the fastest capacitor discharge rate in each detector. These settings affect number of parameters such as the detector bandwidth, the frequency characteristic or the transient response during the calibration switching. Slow transient response of the diode detectors limits the maximum switching frequency and the minimum latency of the calibrated measurements.

Transient response with different time constant settings were analysed in chapter 5.2. The duration of the transient with the slowest time constant T_{c0} was measured to be some 0.4 s which would limit the maximum calibration frequency to less than 2.5 Hz. A transient speed-up method was developed to decrease the transient duration by means of automatic control in the FPGA system. The speed-up is based on temporal increase of the detector discharge rate after each transient which is achieved by dynamic control of the time constants. The data affected by the transients is then rejected in the position computation algorithm. The laboratory measurements presented in chapter 5.4.1 showed that this technique helps to decrease the transient duration to some 0.15 s. Consequently, the calibration switching rate with the slowest time constant can be increased to more than 5 Hz.

Frequency characteristics and the bandwidth of the compensated diode detectors were measured in laboratory and presented in chapter 5.3. It was shown that the time constant T_{c0} sets the smallest measurement bandwidth, up to some 10 Hz. Due to its slow discharge rate, this time constant setting should be used for input pulses with low repetition rates, such as the LHC pilot beams consisting of a single bunch. Time constant T_{c2} provides the largest bandwidth of the DOR detectors, up to some 500 Hz. However, the measurement bandwidth is limited by the output analogue filters to 100 Hz. This setting is more suitable for higher pulse repetition rates, such as the LHC physics beams which typically consist of more than thousand bunches. As opposed to the T_{c0} , the frequency response of the detectors with T_{c2} is flat and therefore more suitable for measurements such as orbit oscillations.

Precision and accuracy of the calibrated orbit measurements was evaluated in laboratory and presented in chapter 5.4.3. The measurements were performed with a signal generator simulating the LHC BPM signals. The analysis was focused on evaluating the systematic errors and noise of the position signals measured with different RF amplifier settings and pulse repetition rates. The quality of the calibrated measurements was compared to the raw measurements without calibration. The position computation assumed 61 mm BPM aperture. The offset errors of the absolute positions was estimate to be around $\pm 150 \mu\text{m}$ when the calibration was not used. Such offset error would be translated as alignment error of the collimator jaws induced by the position measurement system. It was also observed that, depending on the RF gain settings, these errors changed by more than 50 %. The calibration switching reduced these offset errors to less than $30 \mu\text{m}$ which is equivalent to around 0.1 % asymmetry between two DOR channels. Obtained results were also independent of the RF gains. It was shown that the calibration switching helps to improve the resolution and decrease the low frequency drifts caused by slow changes of the gain asymmetries. The position drifts were reduced by the calibration switching by more than factor 2. The position noise was estimated in the calibrated measurements to less than $0.5 \mu\text{m}_{\text{RMS}}$. For favourable conditions the noise level was estimated to be less than $0.05 \mu\text{m}_{\text{RMS}}$. This result could be achieved with high pulse repetition rates on the signal generator and low gain settings in the RF amplifier. Similar settings are typically used during physics beams with nominal intensity bunches.

Sensitivity of the position measurements to the induced temperature change were evaluated and presented in chapter 5.5. After 4 hours of stable measurement conditions, the temperature changes were induced by opening the front-end unit and exposing its electronics to the temperature fluctuations. The following measurement was performed with a forced air flow over the electronics with an external ventilation unit. The temperature change during the test reached 15°C , which is much more than the typical operating conditions observed in the LHC tunnel. It was showed, that the temperature sensitivity without the calibration switching can reach about $1 \mu\text{m}/^\circ\text{C}$. Applying the calibration switching at 1 Hz rate reduced the dependency of the measurement on the temperature change by an order of magnitude.

Performance of the DOR subsystem was evaluated with beams using the DOROS systems in the LHC. As presented in chapter 6.1, the upstream and downstream BPM electrode readings of the collimator DOROS system were used to estimate the noise and long-term stability with the LHC beams. The orbit resolution estimated over 11 hours of measurements with physics beams was better than $0.1\ \mu\text{m}$. The offset drift most probably caused by the DOROS electronics was estimated to less than $0.5\ \mu\text{m}$. Long-term stability of the DOR subsystem was also demonstrated with the standard DOROS system acquiring orbits measurements during a 37 hour LHC physics fill. The beam orbits were measured in the BPMs located on the left and right side of the ATLAS and CMS experiments. The projected orbit signals to the middle to the experiments changed during this measurement by less than $15\ \mu\text{m}$. It was observed that the beam position drifts are correlated between the ATLAS and CMS interaction regions. The position results from the standard DOROS system were used in comparison with the standard LHC BPM system. Presented measurement examples were acquired during the Van der Meer position scans in the ATLAS experiment. The largest discrepancy between the absolute beam positions was measured around $400\ \mu\text{m}$ which is equivalent to $0.6\ \%$ error on the BPM aperture. Part of the orbit measurement was used to estimate and compare the precision of the two systems. In this comparison the DOROS system achieved improvement of more than factor 6 in both the standard deviation and peak-peak noise estimation. This comparison assumed that the estimated noise was dominated by the measurement system. Further beam measurements showed that the standard DOROS system measures oscillatory-like beam motion with amplitudes in the order of a few micrometres and frequency in the order of $0.01\ \text{Hz}$. It was also observed that the measured oscillations have negative impact on the collision rates in the experiments.

As the compensated diode detectors consist of non-linear elements, the linearity of the position measurement has been measured and analysed. The laboratory measurements presented in chapter 5.1 showed that when the simulated beam is centered and the pulses from the electrodes have equal amplitudes, the influence of the signal amplitude changes on the measured positions is very small. This is typically the case for the collimator system where the BPM-embedded jaws are aligned symmetrically with respect to the beam. In such conditions the diode detectors of each channel pair operate with very similar signals, reducing the influence of the residual nonlinearity of the detectors. In the least favorable conditions when the diode detector operates with T_{c2} and the signal amplitude changes by $50\ \%$ of the ADC full-scale, the position change was measured to be less than $1\ \mu\text{m}$. The calibration switching helps to suppress most of the asymmetries between the two channels processing signals from the BPM electrode pair and thus also improve the linearity of the measurement with centered beams.

The systematic errors of the beam position measurements become more important with increasing beam offsets. In such conditions the signals for the opposing BPM electrodes have much different amplitudes, causing the corresponding diode detectors to operate at quite distant part of their characteristics. Then the residual nonlinearities of the characteristics are converted into systematic position errors changing with the signal amplitudes driven by the beam intensity decay. Depending the detector time constant and position offset the position error for a $50\ \%$ amplitude drop could reach more than $100\ \mu\text{m}$. The linearity errors were decreased by compensating of the amplitude drop using the automatic gain control with $1\ \text{dB}$ gain steps in the RF amplifier. The automatic gain control can reduce the signal change on the detector inputs to about $12\ \%$. The position error was reduced to some $25\ \mu\text{m}$ when using the time constant T_{c0} and setting the automatic gains to maintain the signal levels within $0.6 - 0.8\ \text{ADC FS}$ range. It was also showed that the smallest systematic errors can be achieved with the slowest detector discharge T_{c0} achieving more than $20\ \%$ position error reduction in comparison to the T_{c2} . However, using only the T_{c0} time constant is not optimal in all operational conditions. An example of such a case is a physics fill lasting several hours. During this time the diode detectors start measuring the orbit of non-colliding bunches, whose nominal intensity decays slower. The problem was solved by using the faster discharge constant T_{c2} to operate the detectors in the averaging mode.

Further improvements can be achieved by linearization of the diode detector characteristic using the correction polynomials obtained from the laboratory measurements. Chapter 5.1.2 presents the fitting and correction coefficients computed up to the 5th order polynomials for all three detector time constants. Performance of the linearization method and its impact on the precision of the orbit measurement was evaluated with the LHC beams. Without additional corrections the non-linear errors became apparent especially in the BPM planes where the beam offsets were larger than 3 mm. Each 1 dB increase in the RF gains caused some 20 μm discontinuity in the orbit measurements which significantly limited the operational precision. Proposed linearization polynomials obtained from the laboratory measurement in the chapter 5.1.2 were used to correct the non-linear effects observed in the beam measurements. Obtained results of the 3rd order polynomial correction showed that all the orbit discontinuities caused by the gain change were minimized to less than 1 μm . The non-linearity correction is currently performed as part of the position computation in the DOROS server. The same correction coefficients are used for all DOROS front-ends. This simplification results in the error increase to some 5 μm when comparing the performance of all standard DOROS front-ends.

In addition to the calibrated orbit processing, the DOROS systems can acquire the DOR signals from each BPM electrode using the data capture in the front-end memory. The main advantage is that the capture mode allows to acquire the raw ADC samples at the turn-by-turn rate for a few millions of LHC beam revolutions. The increased measurement bandwidth is suitable for observing orbit oscillations at higher frequencies than the calibration switching. Depending on the time constant of the diode detectors the observation bandwidth can be increased to some 100 Hz. For such measurements the calibration switching has to be temporarily turned off to measure the oscillations with fine frequency resolution in the frequency domain. An example of the orbit spectra measured with the three time constants was presented in chapter 6.2.3. This mode became useful for diagnostics of low frequency beam spectra [78].

- **Design and optimisation of signal processing and synchronization of the beam oscillation measurements.**

The so called DOS subsystem in the DOROS front-ends was designed and optimised to perform measurements of the excited beam oscillations and related beam and machine optics parameters, such as betatron coupling or phase advance. The beams during these measurement are excited at a frequency close to the fractional LHC tunes. Processing of the DOS signals can be thus focused only on the excited oscillation components rather than on the whole frequency spectrum. The narrow-band signal processing allows to increase the sensitivity of the system and simplify the data processing. As described in chapter 4.4.1, the oscillation signal processing of a BPM electrode pair is based on two RF diode detectors. An AC coupling removes the dominant DC part of the signals related to the beam intensity and thus increase the dynamic range. The oscillation parts of the BPM signals are then subtracted on a differential amplifier, allowing to measure the relative position changes in frequency range from approximately 500 Hz up to some 5 kHz. The DOS signals are sampled by the multichannel 24-bit sigma-delta ADC shared with the DOR subsystem. The DOS data is then processed by dedicated DSP channels transmitted at 25 Hz rate over Ethernet. This design did not significantly increase the overall system complexity as most of the hardware resources are shared with the DOR subsystem, such as the RF filters and amplifiers. The proper signal levels on the input of the DOS detectors are guaranteed by the automatic gain control used by the DOR subsystem. The laboratory measurements presented in chapter 7.2 were focused on characterisation of the DOS performance with a signal generator simulating the BPM signals. The beam oscillations were simulated with amplitude modulation of the pulses. The measurement results showed, that the SNR of the narrowband DOS signals depends on a few parameters, such as the:

- pulse amplitude modulation depth, which represented the oscillation amplitude,
- pulse repetition rate, which represented the bunch spacing,
- generator signal amplitude, which represented beam intensity,
- and FFT length, which also represented duration of the beam measurement.

Large beam excitations applied for very long time lead to increasing beam size resulting in decreasing efficiency of the collisions. It was estimated with laboratory measurements, that a single nominal bunch in the LHC with excited oscillations of only 10 μm amplitude for 1 s duration would yield some 40 dB SNR in the DOROS oscillation measurements. By increasing number of the simulated bunches to the full capacity of the LHC, the same SNR could be reached within the 1 s duration with less than 1 μm oscillation amplitude.

The standard DOROS system was used to measure excited beam oscillations in the LHC in order to evaluate the local betatron coupling. The beam excitation was chosen such that both the DOROS and the standard LHC BPM system can detect and measure the beam oscillations. As the DOS part of the DOROS system measures only the position change, the standard BPM system was used to estimate amplitude of the oscillations in micrometres which was then used to scale the DOS results. The noise floor of the DOS subsystem was estimated to 0.1 μm . In comparison to the standard BPM system, the DOS achieved sensitivity improvement by an order of magnitude. This improvement was important in a few cases when the sensitivity of the standard system was not sufficient to detect the beam oscillation. Obtained results from the local beam coupling analysis [82], [83], [84] showed very good agreement between the two BPM systems.

Timing circuits described in chapter 4.4.3 allow synchronization of the DOROS front-ends to a common reference clock derived from the BPM signals synchronous to the beam. The local timing generator circuit was designed and optimized to derive the LHC revolution clocks using a fast comparator and a hold-off circuit implemented in the FPGA logic. A beam synchronous reference clock is generated locally in each DOROS front-end from a selected bunch in the bunch train, common for all front-ends. It was shown in chapter 7.1.1 that the threshold level of the input comparator can be optimised based on the orbit signals measured by the DOR subsystem. A change in the threshold level controlled by a DAC introduces a delay in the reference clock maximum about 1 ns. Proper setting of the threshold level according to the DOR signals improves robustness of the produced beam reference clock.

The beam-based clock reference is not available for all beam conditions and front-end settings. Therefore, another clock reference is received from the BST network with guaranteed operation in all beam conditions. However, the phase synchronism with beams is not guaranteed causing systematic errors in the phase advance measurements. Therefore, the phase of the revolution clocks received from the BST can be adjusted in 100 ps steps by a PLL-based phase shifter. Such small phase differences are difficult to be reliably measured by classical counting methods. As described in chapter 4.4.3.3, a dedicated phase detector circuit has been developed and optimized to detect small phase difference between two f_{rev} clocks and allow their precise phase alignment. Its phase-to-voltage characteristic was designed to have an extremum when its two input signals are exactly in phase. In this condition all its digital signals are of 50 % duty cycle so that the circuit characteristic does not have a dead zone. This feature allows a precise indication of the zero-phase condition, which is independent of the detector power supply and the offset of its ADC readout. The laboratory measurements in chapter 7.1.2 demonstrated that the differential phase detector circuit is capable of detecting the zero-phase condition with resolution well exceeding the minimum phase step of 100 ps. The peak-peak noise of the phase measurement was estimated at the zero-phase condition to less than 30 ps. The nominal performance can be reached only with the circuit enclosed inside the front-end. The temperature sensitivity of the phase detector signal was estimated to 1.6 $\mu\text{V}/^\circ\text{C}$ which can be expressed as an equivalent time delay of around 22 ps/ $^\circ\text{C}$. The signals from the phase detector can be thus used for the long-term monitoring of the phase alignment.

At the time of finishing this work, further studies and developments are performed on the DOROS systems. One of the developments targets a radiation tolerant version of the DOROS electronics. Such system could be installed in the LHC tunnel where the ionising doses may reach levels harmful for regular electronics. Length of the coaxial cables from the BPM sensors that would allow to place the electronics to safe locations may become an important factor for the overall system cost and performance. The study investigates the tolerance of the individual components to the ionising radiation and its impact on the measurement performance.

Another study looks into developing a version of the DOROS system optimised for the SPS accelerator. There the beam revolution frequency is four times higher than in the LHC, which would require to increase the ADC sampling as well as the bandwidth of the input RF filters and amplifiers. In addition the SPS cycle is much faster. A typical cycle can last only a couple of seconds depending on the type and purpose of the SPS beam. Therefore the bandwidth of the signal processing including the digital signal processing has to be adapted to provide more orbit samples especially for short SPS cycles.

The author also proposes a few studies which can address the limitations of the current DOROS systems:

- **Linearity**

The systematic position errors caused by the residual non-linearity of the compensated diode detectors become a limitation of the system, especially for larger beam offsets. These errors can be minimized by automatically maintaining the signal levels at a quasi-constant level when the beam intensity decreases over long LHC cycle. The front-end architecture allows to change gains in 1 dB steps using the programmable gain amplifiers compensating a 12 % amplitude drop. One of the improvements could be to use amplifiers with smaller programmable gain step. Variable gain elements, such as the voltage controlled gain amplifiers controlled by a DAC, can achieve even smaller steps. However, typical bandwidth of such amplifiers is below 100 MHz and as such are not suitable for the RF signal processing in the DOROS front-ends. Another option could be to use so called RF PIN diodes as a variable resistor controlled from a DAC. The PIN diode added at the end of the RF amplifier chain could increase resolution of the gain control beyond the 1 dB step.

- **Scaling of the oscillation signals**

Another limitation of the DOR subsystem is that it measures the absolute orbits based on many beam turns. On the other hand, the DOS subsystem can perform high sensitivity turn-by-turn measurements, which are the base for many beam measurements. However, the DOS can measure only the relative position change without the intensity normalisation. The DOS subsystem readings have to be thus cross-calibrated with the standard LHC BPM system. This method is limited only to a certain beam settings due to required large excitation amplitudes. The most representative beam settings such as the physics beams with many high intensity bunches are excluded from such method due to the machine protection reasons. One of the possibilities could be to use the compensated diode detectors with extended bandwidth to perform both the orbit and oscillation measurements. The beam oscillations measured by the DOS subsystem could be normalised by the intensity proportional signal measured in the orbit subsystem using the same diode detector. This method could minimise the uncertainties of the cross-calibration process and potentially extend the number of applications of the DOROS system.

- **Calibration switching**

Gain asymmetries in each channel pair processing one BPM plane, can be minimized by the periodic calibration switching. However, the asymmetries between the channel pairs cannot be calibrated this way as the same electrode signal is not measured with all four channels. A quadruple switching could allow to calibrate also the asymmetries between the BPM plains. Such functionality may be important when measuring the quadrupolar moment and derived quantities such as the beam size or emittance.

- **Bunch gating**

The compensated diode detectors can measure the average beam position of all bunches over many LHC beam revolutions. The bunch gating of the RF signals in front of the diode detectors can allow orbit measurements of the selected bunches in the beam. The gating can be also used to exclude selected bunches from the signal processing. As shown in chapter 6.2.4, the compensated diode detectors are limited to use only the fast time constants during the physics operation to avoid position drifts caused by the changing intensity difference between the of the colliding and non-colliding bunches.

10 Conclusions

It can be concluded that both the collimator and standard DOROS systems, which were described in this doctoral thesis, have become an important part of the LHC beam instrumentation. This thesis demonstrated that a BPM measurement system based on simple diode detector circuits can be built and optimised for achieving submicron resolution for long-term beam orbit measurements as well as sensitivity to micrometre-order beam oscillations. The DOROS systems have been commissioned and operationally involved in many beam measurements used to optimise the performance, reliability and safety of the LHC accelerator. This dissertation work was realised as a part of the beam instrumentation research and development at CERN.

11 Resumé

Hlavnou úlohou tejto dizertačnej práce bolo vyvinúť a optimalizovať elektronický systém s názvom DOROS. Spracovanie BPM signálov v tomto systéme je založené na dvoch druhoch diódových detektorov, ktoré boli oddelene optimalizované na meranie orbity a oscilácií časticových zväzkov v urýchľovači LHC. Projekt DOROS začal v roku 2009 prototypom jednoduchého obvodu, ktorý ukázal, že diódové detektory môžu byť použité na spracovanie BPM signálov. Následný vývoj prototypov a prvé merania so zväzkami ukázali, že BPM systém založený na technológii diódových detektorov môže dosiahnuť rozlíšenie merania polohy zväzkov lepšie ako jeden mikrometer. V roku 2013 bolo rozhodnuté vyvinúť a inštalovať plne funkčný systém, ktorý by umožnil presné zarovnávanie čelustí LHC kolimátorov pomocou integrovaných BPM senzorov. Takzvaný kolimátorový DOROS systém je kritický pre bezpečný chod a spoľahlivosť urýchľovača LHC. Idea použitia diódových detektorov na presné merania pozície zväzkov bola súčasne testovaná na niekoľkých vybraných BPM senzoroach v okolí LHC experimentov. Rozlíšenie a stabilita meraných hodnôt v týchto lokalitách LHC predstavujú dôležité parametre pre množstvo meraní, ktoré sú kritické pre optimalizáciu luminozity jednotlivých experimentov a prekonávanie limitov LHC. Prvé DOROS zariadenia boli nainštalované a uvedené do prevádzky v roku 2015. Takzvaný štandardný DOROS systém bol potom postupne v priebehu rokov rozširovaný o ďalšie jednotky v rôznych lokalitách LHC. Hardvér a firmvér je identický vo všetkých DOROS zariadeniach v LHC, čo umožňuje ich jednoduchšiu údržbu.

Súčasný kolimátorový DOROS systém sa skladá z 20 jednotiek, ktoré umožňujú merať pozíciu zväzkov v 22 LHC kolimátoroch s integrovanými BPM senzormi. Tento systém sa používa na presné zarovnanie čelustí vzhľadom na meranú pozíciu zväzku. Čas potrebný na nastavenie kolimátorov bol pomocou DOROS systému rádovo zredukovaný z niekoľkých hodín na niekoľko sekúnd. Systém bol použitý počas viacerých štúdií [29], [30], [74] a [76] zameraných na zlepšenie výkonnosti a spoľahlivosti kolimácie zväzkov v LHC. Ďalšie štúdie sú zamerané na presné zarovnanie tzv. “wire” kolimátorov, ktoré slúžia na kompenzáciu efektov vznikajúcich počas kolízií zväzkov v experimentoch [77]. Merané hodnoty z niekoľkých vybraných DOROS jednotiek sú taktiež používané takzvaným LHC interlock systémom, ktorý slúži na ukončenie operácie LHC v prípade, ak meraná pozícia zväzku v kolimátoroch prekračuje bezpečné limity. Tento ochranný mechanizmus bol uvedený do prevádzky v septembri 2017. Akákoľvek dlhodobá porucha DOROS jednotky môže zbytočne prerušiť a prípadne zablokovať operáciu LHC. Z tohoto dôvodu bolo vynaložené úsilie na zabezpečenie dostatočnej spoľahlivosti a odolnosti systému na hardvérovej ako aj softvérovej úrovni. Kritické časti systému boli implementované redundantne pomocou dvoch nezávislých DOROS jednotiek, aby sa zabránilo blokovaniu prevádzky LHC v prípade zlyhania hardvéru. Každá DOROS jednotka taktiež automaticky monitoruje svoj stav, vrátane FPGA systému s implementovaným procesorom s operačným systémom v reálnom čase, ako aj všetku Ethernetovú komunikáciu s koncentračnými a riadiacimi servermi. V prípade detekcie poruchy je jednotka automaticky resetovaná a jej operácia je obnovená do približne 10 sekúnd. Súčasný odhad MTBF systému DOROS dosahuje viac ako 60 rokov.

Súčasný štandardný DOROS systém pozostáva z 21 jednotiek, ktoré boli inštalované v LHC a slúžia na merania pozície zväzkov v 40 BPM senzoroach. Tento systém bol inštalovaný na najdôležitejších BPM senzoroach v LHC, vrátane BPM Q1 nachádzajúcich sa pri kvadrupólových magnetoch v okolí štyroch experimentov, ako aj na niekoľkých vybraných BPM Q5, Q6 a Q7 nachádzajúcich sa v LHC bodoch 1, 2, 4, 5, 6 a 8. Signály z týchto BPM boli rozdelené medzi štandardný LHC BPM systém a systém DOROS, čo umožňuje vyhodnotiť a porovnať merané výsledky z týchto dvoch systémov. Štandardný DOROS systém sa využíva na optimalizačné merania v LHC, akými sú napríklad takzvané Van der Meer skeny alebo meranie lokálnej betatrónovej väzby. DOROS systém bol tiež použitý počas viacerých štúdií, akými sú napríklad takzvaný β^* levelling [63], direktivita BPM stripline senzora [79] alebo merania nízkofrekvenčných oscilácií zväzkov [78].

Väčšina týchto štúdií je zameraných na vylepšenia LHC, ktorých cieľom je zvýšenie a optimalizácia luminozity v experimentoch. Nedávna štúdia sa začala zaoberať neinvazívnymi meraniami kvadrupolárneho momentu zväzkov pomocou BPM signálov. Takéto merania sú dôležité najmä počas akcelerácie v LHC, kedy dochádza k zvyšovaniu energie zväzkov. Štandardné a kolimátorové DOROS systémy sa stali dôležitými nástrojmi na meranie a diagnostiku zväzkov ako aj na optimalizáciu výkonnosti LHC urýchľovača.

Hlavné prínosy tejto dizertačnej práce možno zhrnúť do nasledujúcich bodov.

- **Návrh, vývoj a optimalizácia nového systému spracovania BPM signálov založeného na báze diódových detektorov.**

Systém DOROS opísaný v kapitole 4 pozostáva z hardvérových zariadení realizovaných ako 19 "1U jednotky, ktoré dokážu spracovať signály z dvoch BPM senzorov pozostávajúcich zo štyroch elektród. Signály z BPM senzora sú najskôr filtrované a zosilnené predtým, než sú spracované diódovými detektormi. Implementovaný rádiový frekvenčný zosilňovač pozostáva zo štyroch blokov ovládaných pomocou FPGA. Dynamický rozsah signálov tak môže byť ovládaný v rozsahu 90 dB s 1 dB krokom. Signály orbity a oscilácií zväzkov sú potom spracovávané oddelene s pomocou dvoch samostatných podsystémov obsahujúcich špecificky optimalizované diódové detektory. Analógové signály z oboch podsystémov sú potom digitalizované pomocou 8-kanálového 24-bitového sigma-delta prevodníka. Digitálne signály sú spracovávané v paralelných DSP kanáloch implementovaných v FPGA logike. Spracované výsledky sa potom posielajú prostredníctvom Ethernetovej siete na príslušné servery. Opakovacia frekvencia posielania datagramov rovná 25 Hz je synchronná s externým LHC časovaním (takzvaným BST) distribuovaným prostredníctvom optických vlákien. Vzorkovanie ADC pri frekvencií približne 11.2 kHz je synchronizované s obletom LHC zväzku. Synchronizácia vzorkovania medzi DOROS jednotkami je založená na stabilnom hodinovom signáli z BST, ktorý je možné fázovo zarovnať s referenčným hodinovým signálom generovaným zo samotných BPM signálov. Jednotlivé ADC vzorky sú ukladané taktiež v nespracovanej forme do DDR pamäti nachádzajúcej sa v každej DOROS jednotke. Tieto dáta sú k dispozícii serverom prostredníctvom TCPIP protokolu a umožňujú merať a analyzovať oscilácie zväzkov. Systém FPGA riadi aj ostatné funkcie DOROS jednotky, akým je napríklad kalibrácia merania orbity, automatické riadenie zisku vstupných RF zosilňovačov, generátor testovacieho signálu, riadenie synchronizačných obvodov, monitorovanie stavu alebo aj preprogramovanie FPGA na diaľku prostredníctvom Ethernetovej siete.

- **Návrh a optimalizácia spracovania signálu na meranie orbity LHC zväzkov s pomocou kompenzovaných diódových detektorov.**

Najdôležitejšou inováciou nového systému je metóda kompenzovaných diódových detektorov, ktorá podľa autorovho poznania bola prvýkrát použitá v aplikácii na meranie pozície časticových zväzkov. Táto technika sa používa v tzv. DOR podsystéme na meranie orbity a nízkofrekvenčných oscilácií zväzkov. Typický BPM senzor v LHC môže produkovať impulzy s časovým trvaním približne 1 ns, ktoré sa opakujú každých 25 ns. V závislosti od intenzity zväzku ako aj typu BPM senzora sa môžu amplitúdy výsledných pulzov meniť približne o štyri rády, typicky v rozmedzí 0.1 V a 100 V. Meranie 1 μm zmeny pozície zväzku v transverzálnej rovine pomocou BPM senzora s 100 mm apertúrou by bolo potrebné detekovať zmeny amplitúdy pulzov v oblasti rádovo 10^{-5} , čo je ekvivalentné 10 μV zmene napätia BPM signálu s 0.1 V amplitúdou. Použitie RF diódových detektorov na meranie takýchto malých zmien signálov má niekoľko výhod:

- rýchla odozva,
- vysoká senzitivita,
- nevyžaduje externé časovanie,
- jednoduchosť a robustnosť.

V kapitole 3.3 bolo ukázané, že pozícia zväzku vypočítaná zo signálov na výstupe RF diódových detektorov závisí od poklesu napätia na diódach. Toto napätie sa mení v závislosti od teploty, prúdu, ako aj od výrobného procesu, v dôsledku ktorých sa mení aj výsledná chyba merania. Kompenzované detektory minimalizujú závislosť výstupných signálov od napäťového poklesu na dióde, čo je dôležitým predpokladom pre maximalizáciu presnosti a stability meraní. Presnosť merania je taktiež ovplyvnená asymetriami medzi príslušnými analógovými kanálmi. Aby bolo možné minimalizovať chyby vyplývajúce z takýchto asymetrií, bola vyvinutá tzv. automatická kalibračná metóda, založená len na BPM signáloch. Táto technika opísaná v kapitole 4.3.1 využíva periodické multiplexovanie signálov na vstupe príslušných DOR kanálov. Po následnom spracovaní signálov v analógovej a digitálnej časti DOROS jednotky je možné vypočítať kalibrované hodnoty pozície z prijatých dát na serveri pomocou kalibračného algoritmu. Automatické kalibrovanie umožňuje minimalizovať systematické chyby merania, ktoré závisia od zisku RF zosilňovačov, ako aj od teplotných a iných zmien v analógovej časti. Kalibrácia merania je kritická pre presné nastavovanie čelustí kolimátorov, ako aj pri meraní pozície zväzkov počas kolízií v LHC experimentoch.

Dynamiku diódových detektorov je možné meniť pomocou časových konštánt, ktoré je možné spínať pomocou FPGA systému. V každom DOR detektore sú k dispozícii tri nastavielne časové konštanty, konkrétne T_{c0} , T_{c1} a T_{c2} . Nastavenie T_{c0} určuje najpomalšiu a T_{c2} najrýchlejšiu rýchlosť vybíjania pamäťových kondenzátorov v detektoroch. Tieto nastavenia ovplyvňujú viacero parametrov, akými sú napríklad šírka frekvenčného pásma, frekvenčná prenosová charakteristika alebo prechodová odozva počas kalibračného multiplexovania signálov. Pomalá prechodová odozva diódových detektorov obmedzuje maximálnu frekvenciu multiplexovania, čo má za následok zvýšenie latencie kalibrovaných meraní.

Prechodové odozvy detektora v jednotlivých nastaveniach časovej konštanty boli merané a analyzované v kapitole 5.2. Dĺžka trvania prechodu s najpomalšou konštantou T_{c0} bola odhadnutá na približne 0.4 s, čo by obmedzilo maximálnu frekvenciu multiplexovania signálov na menej ako 2.5 Hz. Z tohoto dôvodu bola vyvinutá nová metóda urýchlenia prechodových javov využívajúca automatické riadenie časových konštánt pomocou FPGA systému. Táto metóda je založená na urýchlení vybíjania detektora pomocou dočasného nastavenia T_{c1} alebo T_{c2} konštanty pri každom multiplexovaní vstupných signálov. Merané dáta, ktoré sú ovplyvnené prechodovými javmi, sú neskôr vyradené kalibračným algoritmom na výpočet kalibrovanej pozície. Laboratórne merania uvedené v kapitole 5.4.1 ukázali, že táto technika pomáha znižovať trvanie prechodov na približne 0.15 s. V dôsledku toho môže byť frekvencia multiplexovania s konštantou T_{c0} zvýšená na viac ako 5 Hz.

Frekvenčné prenosové charakteristiky a šírka frekvenčného pásma kompenzovaných diódových detektorov boli ukázané na laboratórnych meraniach v kapitole 5.3. Najmenšie frekvenčné pásmo merania sa pohybuje od DC do približne 10 Hz v prípade nastavenia T_{c0} konštanty. Vzhľadom na pomalé vybíjanie kondenzátorov v detektoroch by sa toto nastavenie malo používať pre vstupné signály s nízkymi opakovacími frekvenciami, akými sú napríklad pilotné zväzky v LHC. Nastavenie konštanty T_{c2} poskytuje naopak šírku frekvenčného pásma až do 500 Hz. Výsledné frekvenčné pásmo merania je však obmedzené analógovými filtermi na 100 Hz. Nastavenie T_{c2} je teda vhodnejšie pre vyššie opakovacie frekvencie vstupných signálov, akými sú napríklad zväzky využívané pri interakciách v LHC experimentoch. Frekvenčná odozva s nastavením T_{c2} je na rozdiel od T_{c0} plochá, a preto je vhodnejšia pre merania nízkofrekvenčných oscilácií zväzkov.

Presnosť kalibrovaných signálov bola vyhodnotená pomocou laboratórnych meraní prezentovaných v kapitole 5.4.3. Tieto merania boli realizované pomocou signálového generátora, ktorý simuloval signály z BPM senzora v LHC. Analýza bola zameraná na vyhodnotenie systematických chýb a šumu, ktoré boli merané s rôznymi nastaveniami vstupných RF zosilňovačov ako aj frekvenciou opakovania pulzov. Kvalita kalibrovaných meraní bola porovnaná s meraniami bez kalibrácie. Výpočet pozície sa vzťahuje na 61 mm apertúru BPM senzora. Bez použitia kalibrácie bola absolútna odchýlka merania v meracom zapojení simulujúcom nulovú odchýlku odhadnutá na približne $\pm 150 \mu\text{m}$.

Takáto odchýlka merania by sa napríklad prejavila ako chyba nastavenia kolimátorových čelustí vzhľadom na zväzok. Bolo tiež ukázané, že v závislosti od nastavenia zisku v RF zosilňovači sa tieto chyby menili o viac ako 50 %. Kalibračná metóda zredukovala tieto odchýlky na menej ako 30 μm , čomu zodpovedá približne 0.1% asymetria medzi dvoma analógovými kanálmi. Tento odhad chyby zároveň zahŕňa asymetrie, ktoré nie je možné kalibrovať s pomocou multiplexovania signálov, akými sú napríklad vstupné RF filtre alebo aj samotné pripojenie generátora k DOROS jednotke. Dosiahnuté výsledky kalibrovaných meraní boli tiež nezávislé od nastavenia zisku v RF zosilňovači. Bolo ukázané, že kalibračná metóda pomáha znižovať šum ako aj pomalé zmeny medzi analógovými kanálmi, ktoré boli zmenšené o viac ako 2-krát. Hladina šumu v meraní kalibrovannej pozície bola odhadnutá na menej ako 0.5 μm_{RMS} . Za priaznivých podmienok merania bola táto hladina nižšia ako 0.05 μm_{RMS} . Tento výsledok bol dosiahnutý pri laboratórnom nastavení s vysokými opakovacími frekvenciami generovaných pulzov a relatívne nízkymi hodnotami zisku vo vstupnom RF zosilňovači. Takáto konfigurácia simuluje LHC zväzky s nominálnou intenzitou, ktoré sú používané pri kolíziách v experimentoch. Senzitivita merania pozície zväzkov na indukovanú zmenu teploty bola meraná a vyhodnotená v kapitole 5.5. Po 4 hodinách merania v stabilných podmienkach bola vnútorná elektronika v DOROS jednotke vystavená vonkajším teplotným zmenám. Následná zmena teploty v jednotke bola vykonaná pomocou externej ventilačnej jednotky umožňujúcej prietok vzduchu ponad exponovanú elektroniku v DOROS jednotke. Celková dosiahnutá zmena teploty počas tohoto laboratórneho testu dosiahla 15 °C, čo je oveľa viac, ako bolo pozorované v DOROS systémoch pri typických prevádzkových podmienkach v LHC tuneli. Bolo ukázané, že senzitivita merania pozície na teplotné zmeny môže dosiahnuť približne 1 $\mu\text{m}/^{\circ}\text{C}$ v prípade, ak nie je použitá kalibračná metóda. Meranú senzitivitu merania bolo možné znížiť o jeden rád s použitím kalibračnej metódy pri 1 Hz frekvencií multiplexovania vstupných signálov.

Výkonnosť DOR pod systému bola vyhodnotená s pomocou LHC zväzkov. Ako je uvedené v kapitole 6.1, na odhad šumu a dlhodobú stabilitu merania boli použité údaje z kolimátorového DOROS systému. Rozlíšenie merania orbity bolo odhadnuté na menej ako 0.1 μm počas 11 hodín nominálnej operácie LHC. Zmena odchýlky merania počas celej dĺžky merania bola odhadnutá na menej ako 0.5 μm . Dlhodobá stabilita pod systému DOR bola ukázaná taktiež na meraniach orbity zväzkov so štandardným DOROS systémom realizovaných počas 37 hodín nominálnej operácie LHC. Orbity zväzkov boli merané v BPM senzoroach umiestnených na ľavej a pravej strane ATLAS a CMS experimentov. Extrapolované pozície zväzkov do stredu oboch experimentov sa menili menej ako 15 μm počas celej dĺžky merania. Pozorovaný priebeh meranej orbity v oboch experimentoch bol navzájom korelovaný. Výsledky merania so štandardným DOROS systémom boli porovnané so štandardným LHC BPM systémom. Prezentované výsledky meraní boli získané počas takzvaných Van der Meer skenov v ATLAS experimente. Najväčší rozdiel medzi absolútnymi hodnotami merania pozície zväzkov z oboch BPM systémov bol približne 400 μm , čomu zodpovedá 0.6 % chyba v rámci 61 mm BPM apertúry. Toto meranie bolo taktiež použité na odhad a porovnanie presnosti týchto dvoch BPM systémov. V tomto porovnaní dosiahol systém DOROS 6-krát menší šum ako štandardný LHC BPM systém za predpokladu, že šum meracieho zariadenia dominuje nad šumom pozície samotného časticového zväzku. Detailnejšia analýza meraní ukázala, že štandardný DOROS systém dokáže zaznamenať nízkofrekvenčné vibrácie zväzku s amplitúdou rádovo niekoľkých mikrometrov a frekvenciou rádovo 0.01 Hz. Bolo tiež pozorované, že tieto oscilácie majú negatívny vplyv na počet kolízií v experimentoch.

Linearita merania predstavuje dôležitý parameter, ktorý bol meraný a analyzovaný v laboratórnych podmienkach pomocou signálového generátora. Výsledky týchto meraní, ktoré boli prezentované v kapitole 5.1 ukázali, že v prípade, keď majú pulzy z BPM elektród rovnaké amplitúdy, je vplyv zmeny týchto amplitúd na výsledné nameranie pozície veľmi malý. Takéto podmienky merania sú pozorované v kolimátoroch, keď sú čeluste so zabudovanými BPM elektródami symetricky umiestnené vzhľadom na meraný zväzok a signály z týchto senzorov majú rovnaké amplitúdy. Diódové detektory DOR pod systému potom pracujú s veľmi podobnými signálmi, čím sa znižuje vplyv ich nelinearity na meranie.

V najmenej priaznivých podmienkach, kedy diódové detektory pracujú s nastavením T_{c2} a amplitúda signálu sa mení v 50 % rozsahu ADC, bola meraná zmena pozície v dôsledku nelinearity menej ako 1 μm . Kalibrácia merania pomáha potlačiť väčšinu asymetrie medzi kanálmi, čím tiež zlepšuje linearitu merania. Pri zvyšujúcich sa odchýlkach pozície zväzkov od stredu medzi dvomi BPM elektródami sa v dôsledku nelinearity zvyšujú systematické chyby merania orbity. V takýchto prípadoch majú signály z protiahlych BPM elektród rozdielne amplitúdy, čo spôsobuje, že príslušné diódové detektory pracujú v odlišných bodoch prevodovej charakteristiky. Reziduálna nelinearita charakteristiky je tak konvertovaná do systematickej chyby merania, ktorá závisí od amplitúdy BPM signálov. Táto amplitúda závisí od intenzity zväzku, ktorá sa môže v rámci jedného LHC cyklu výrazne meniť. V závislosti od nastavenia časovej konštanty v detektoroch ako aj od pozície zväzku môže chyba merania pri 50 % poklesu amplitúdy dosiahnuť viac ako 100 μm . Takúto chybu merania je možné znížiť kompenzáciou poklesu amplitúdy BPM signálov s pomocou automatického ovládania zisku v RF zosilňovači s krokom 1 dB. Zmena amplitúdy signálov na vstupoch detektorov je tak zredukovaná na približne 12 %. Pri použití časovej konštanty T_{c0} a automatického udržiavania úrovni signálu v rozsahu 0.6 – 0.8 ADC FS bolo dosiahnuté zníženie chyby merania na približne 25 μm . Bolo taktiež ukázané, že najmenšie chyby vyplývajúce z nelinearity diódových detektorov možno dosiahnuť s nastavením T_{c0} , kedy je vybíjanie kondenzátorov v detektore najpomalšie. Optimalizáciou časovej konštanty je možné dosiahnuť približne 20 % redukcii nelinearity v porovnaní s T_{c2} . Limitovanie nastavenia časovej konštanty iba na T_{c0} počas celého LHC cyklu však nie je vždy optimálne. Príkladom, kedy nastavenie T_{c0} nie je vhodné pre meranie pozície predstavuje prípad viac hodinových kolízií zväzkov experimentoch. Od určitého momentu začnú diódové detektory merať orbitu časticových grúp (takzvaných bunch), ktorých menovitá intenzita klesá pomalšie ako je priemerná intenzita grúp podieľajúcich sa na kolíziách. Tento problém bol vyriešený s pomocou nastavenia T_{c2} , ktoré umožňuje znížiť senzitivitu diódových detektorov na zmeny intenzity medzi časticovými grupami v rámci jedného obletu zväzku.

Ďalšie zlepšenia linearity merania bolo možné dosiahnuť pomocou linearizácie prechodovej charakteristiky diódových detektorov pomocou takzvaných korekčných polynómov. Kapitola 5.1.2 prezentuje fitovacie a korekčné koeficienty polynómov až do 5. rádu. Tieto boli získané z laboratórnych meraní pre tri možné nastavenia časových konštánt detektora. Metóda linearizácie prechodovej charakteristiky a jej vplyv na presnosť merania bola vyhodnotená s pomocou meraní zväzkov štandardným DOROS systémom v LHC. Chyby linearity bolo možné pozorovať najmä v meraniach, kde bola pozícia zväzku vzhľadom na stred BPM senzora väčšia ako 3 mm. Pri každom zvýšení zisku v RF zosilňovači o 1 dB dochádzalo ku približne 20 μm zmene meranej orbity v dôsledku nelinearity detektora. Polynómy získané z laboratórneho merania v kapitole 5.1.2 boli potom aplikované na korekciu pozorovaných zmien orbity. Bolo ukázané, že linearizačná metóda dokáže minimalizovať pozorované chyby merania na menej ako 1 μm pomocou polynómu 3. rádu. Vzhľadom na tento výsledok bola táto linearizačná metóda implementovaná ako súčasť výpočtu pozície na DOROS serveri v reálnom čase. Táto implementácia využíva rovnaké korekčné koeficienty pre všetky DOROS jednotky. Takéto zjednodušenie sa však prejavilo pri porovnaní výsledkov zo všetkých štandardných DOROS jednotiek vo zvýšení reziduálnej chyby merania na približne 5 μm .

DOROS systémy umožňujú okrem kalibrovanej orbity zväzkov zároveň merať DOR signály v širšom frekvenčnom pásme pomocou ADC vzoriek ukladaných do DDR pamäti. Jednotlivé vzorky sú ukladané pri vzorkovacej rýchlosti ADC prevodníka, ktorá je synchronná s frekvenciou obletu LHC zväzkov. Uložené dáta je možné pozorovať a analyzovať po dobu niekoľkých minút, čo predstavuje rádovo niekoľko miliónov obletov zväzku v LHC. Zvýšená šírka pásma merania na približne 100 Hz je vhodná na pozorovanie oscilácií orbity v rádovo vyšších frekvenciách, ako by bolo možné pozorovať s pomocou kalibračnej metódy. Pri takomto type meraní musí byť multiplexovanie vstupných signálov dočasne vypnuté, aby bolo možné počítať spektrá signálov s dostatočným frekvenčným rozlíšením. Príklady takýchto meraní, ktoré sú prezentované v kapitole 6.2.3, boli užitočné pri diagnostike nízkofrekvenčného spektra orbity zväzkov [78].

- **Návrh a optimalizácia spracovania signálov a synchronizácie merania oscilácií zväzkov.**

Takzvaný DOS podsystem v každej DOROS jednotke bol navrhnutý a optimalizovaný na detekciu buđených transversálnych oscilácií zväzkov a meranie súvisiacich parametrov urýchľovača, akými sú napríklad betatrónova väzba alebo fázový posun oscilácií. Časticové zväzky sú počas týchto meraní vybudené na frekvencii blízkej k vlastným osciláciám zväzkov v LHC. Z tohoto dôvodu môže byť spracovanie DOS signálov zamerané len na konkrétne frekvenčné komponenty. Spracovanie signálov v úzkom frekvenčnom pásme umožňuje zvýšiť citlivosť systému a taktiež zjednodušiť spracovanie dát. Kapitola 4.4.1 opisuje spracovanie signálov oscilácií zväzkov v DOS podsysteme. Detekcia oscilácií v každej BPM rovine je založená na dvoch RF diódových detektoroch. Dominantná DC časť signálov, ktorá je úmerná intenzite a orbite zväzku, je odstránená pomocou blokovacích kondenzátorov, čím je umožnené zvýšiť dynamický rozsah a senzitivitu merania. Signály z oboch diódových detektorov sú potom odčítané s pomocou diferenciálneho zosilňovača, čo umožňuje merať relatívne zmeny pozície. Následná nízkofrekvenčná časť obmedzuje výsledný frekvenčný rozsah merania na pásmo od 500 Hz až po približne 5 kHz. Signály DOS sú potom vzorkované s pomocou viac-kanálového 24-bitového sigma-delta ADC prevodníka, ktorý je spoločný s podsystemom DOR. Jednotlivé vzorky z kanálov DOS sú spracovávané v DSP kanáloch a výsledky sú posielané spolu s ďalšími dátami na DOROS serveri prostredníctvom Ethernetovej siete. Keďže väčšina signálového spracovania, ako napríklad RF filtre a RF zosilňovače je zdieľaná s podsystemom DOR, implementácia podsystemu DOS výrazne nezvýšila celkovú zložitnosť systému. Automatické ovládanie RF zosilnenia využívajúce signály DOR garantuje správne nastavenie úrovne signálu aj na vstupe DOS detektorov. Laboratórne merania uvedené v kapitole 7.2 boli zamerané na charakterizáciu podsystemu DOS pomocou signálového generátora simulujúceho signály z BPM senzora. Oscilácie zväzkov boli simulované pomocou amplitúdovej modulácií pulzov. Výsledky merania ukázali, že odstup signál-šum v meraných úzkopásmových signáloch závisí od viacerých parametrov, ako napríklad:

- hĺbky amplitúdovej modulácie, ktorá predstavuje amplitúdu oscilácií zväzkov,
- opakovacej frekvencie pulzov, ktorá predstavovala rozstup jednotlivých časticových grúp,
- amplitúde pulzov, ktorá predstavuje intenzitu zväzku,
- dĺžke FFT, ktorá súvisí s trvaním merania.

Excitované oscilácie zväzkov s veľkými amplitúdami a príliš dlhým trvaním nepôsobia priaznivo na kvalitu zväzkov v dôsledku čoho je znižovaná účinnosť kolízií v experimentoch. Laboratórne merania ukázali, že je možné s pomocou DOROS systémov dosiahnuť 40 dB odstup signál-šum, ak je amplitúda buđených oscilácií iba 10 μm a ak meranie trvalo aspoň 1 s. Toto laboratórne meranie simulovalo zväzok skladajúci sa len z jednej časticovej grupy vo zväzku. Pomocou toho istého laboratórneho zapojenia bolo ukázané, že zvýšením počtu časticových grúp na plnú kapacitu LHC (približne 2800 grúp) by bolo možné dosiahnuť rovnakého odstup šumu s amplitúdou oscilácie menšou ako 1 μm . Takáto amplitúda meraných oscilácií by zároveň predstavovala rádovo nižšiu perturbáciu kvality zväzku počas excitácie.

Štandardný DOROS systém bol použitý na meranie excitovaných oscilácií zväzkov slúžiacich na merania lokálnej betatrónovej väzby v LHC. Amplitúda budenia zväzku bola zvolená tak, aby systém DOROS ako aj štandardný LHC BPM systém dokázali detekovať oscilácie zväzku. Keďže DOS podsystem umožňuje merať iba zmenu pozície a nie jej absolútnu hodnotu, bol na odhad absolútnej hodnoty amplitúdy oscilácií v mikrometroch použitý štandardný BPM systém. Pomocou zameraného škálovacieho koeficientu bolo možné odhadnúť hladinu šumu v DOS na 0.1 μm . V porovnaní so štandardným BPM systémom bola dosiahnutá citlivosť podsystemu DOS vyššia o jeden rád. Toto zlepšenie bolo dôležité v prípadoch, keď citlivosť štandardného systému nebola dostatočná na detekciu oscilácií zväzkov, kedy je získané meranie neplatné. Analýza výsledkov merania lokálnej betatrónovej väzby ukázala veľmi dobrú zhodu medzi oboma BPM systémami [82], [83], [84].

Časovacie obvody opísané v kapitole 4.4.3 umožňujú synchronizáciu DOROS jednotiek s referenčným hodinovým signálom odvodeným z BPM signálov. Tento referenčný signál s frekvenciou približne 11.2 kHz je tak intrinzičky synchronný s obletom zväzku. Obvod časovacieho generátora bol navrhnutý a optimalizovaný na odvodenie synchronizačného hodinového signálu z BPM signálov pomocou rýchleho komparátora a tzv. hold-off obvodu implementovaného v FPGA logike. Synchronizačný signál je vytvorený v každej DOROS jednotke a slúži predovšetkým ako referencia fázy pre jednotlivé ADC prevodníky. V kapitole 7.1.1 bolo ukázané, že porovnávacía úroveň vo vstupnom komparátore môže byť optimalizovaná pomocou subsystému DOR. Týmto spôsobom je možné zvýšiť robustnosť výsledných referenčných hodín. Porovnávaciu úroveň je možné nastavovať pomocou DAC prevodníka, ktorým je zároveň možné riadiť oneskorenie referenčných hodín v rozmedzí približne 1 ns.

Synchronizačný signál odvodený z BPM signálov však závisí od konfigurácie zväzkov ako aj od nastavenia DOROS jednotky. Spoľahlivosť tohto referenčného signálu za každých okolností tak nie je možné garantovať. Z tohoto dôvodu má každá DOROS jednotka prístup k externému hodinovému signálu BST, ktorého spoľahlivosť je garantovaná počas celého LHC cyklu. Fáza BST signálov však vzhľadom na oblet zväzku nie je zaručená, čo spôsobuje chyby pri meraniach fázového posunu oscilácií zväzkov. Z tohoto dôvodu bol implementovaný mechanizmus umožňujúci posuv fázy referenčných BST hodín v 100 ps krokoch. Výsledné fázové rozdiely dvoch 11.2 kHz hodinových signálov počas jednotlivých krokov je náročné spoľahlivo merať pomocou klasických metód. Z tohoto dôvodu bol vyvinutý obvod diferenčného fázového detektora opísaný v kapitole 4.4.3.3, ktorý bol optimalizovaný na detekovanie malých fázových rozdielov medzi dvoma nízkofrekvenčnými hodinovými signálmi, čím je potom umožnené ich presné fázové zarovnanie. Prevodová charakteristika tohto detektora bola navrhnutá tak, aby mala extrémum v prípade, keď sú oba vstupné hodinové signály presne vo fáze. V tomto bode charakteristiky majú všetky hodinové signály v detektore 50 % striedu, čo zvyšuje presnosť detekcie a výrazne redukuje závislosť meraného fázového rozdielu na rýchlosti spínania hradiel v detektore. Prezentovaný fázový detektor umožňuje presnú detekciu nulového fázového rozdielu, ktorý nie je závislý od napájacieho napätia detektora ani od offsetu na vstupe ADC. Laboratórne merania v kapitole 7.1.2 ukázali, že diferenčný fázový detektor je schopný detekovať fázové rozdiely s rozlíšením presahujúcim minimálny ovládací krok fázového posunu. Špičkový šum meraného fázového rozdielu z detektora bol odhadnutý s pomocou laboratórneho merania na menej ako 30 ps. Tento výsledok bolo možné dosiahnuť iba ak bol fázový detektor uzavretý vnútri DOROS jednotky. Senzitivita fázového detektora na teplotné zmeny bola odhadnutá na $1.6 \mu\text{V}/^\circ\text{C}$, čo môže byť taktiež vyjadrené ako časový rozdiel medzi dvoma hodinovými signálmi rovný približne $22 \text{ ps}/^\circ\text{C}$.

V čase dokončovania tejto kandidátskej práce sa na systémoch DOROS uskutočňoval ďalší vývoj. Jedna zo štúdií sa zameriava na nasledujúcu generáciu DOROS elektroniky tolerantnej voči radiácií. Takýto systém by mohol byť inštalovaný na miestach v LHC, kde môže hladina radiácie dosahovať úroveň, ktorá predstavuje problém pre štandardnú elektroniku. Typickými miestami sú práve v blízkosti samotných kolimátorov alebo v blízkosti magnetov v oblúkových sektoroch LHC, kde dipólové magnety ohýbajú trajektóriu zväzkov. Dĺžka koaxiálnych káblov z BPM senzorov, ktorá by umožnila inštalovať senzitívnu elektroniku v bezpečnejších lokalitách, by v týchto prípadoch mohla predstavovať dôležitý faktor v celkových cenových nákladoch na inštaláciu. Spomenutá štúdia skúma senzitivitu jednotlivých komponentov v DOROS jednotke na ionizujúce žiarenie ako aj vplyv na celkovú presnosť merania a spoľahlivosť systému. Ďalšia štúdia sa zaoberá vývojom verzie systému DOROS, ktorá by bola optimalizovaná na meranie orbity v SPS akcelératore. Frekvencia obletu zväzku v SPS je štyrikrát vyššia ako v LHC, čo by vyžadovalo zvýšenie vzorkovacej frekvencie ADC ako aj rozšírenie frekvenčnej šírky pásma vo vstupných RF filtroch a zosilňovačoch. Okrem toho je SPS cyklus omnoho kratší. V závislosti od typu a účelu zväzku môže typický SPS cyklus trvať len pár sekúnd. Frekvenčná šírka pásma v nízkofrekvenčnej časti spracovania signálov ako aj DSP musia byť adaptované tak, aby bolo možné zvýšiť počet meraní orbity zväzkov, najmä počas krátkych SPS cyklov.

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