Modelling and Control of Piezoactuators for High Precision Positioning Systems Used in Radioactive Environments

by

Pablo Serrano Galvez

Submitted to the Centro de Electrónica Industrial
in partial fulfillment of the requirements for the degree of

Doctorado en Ingeniería Eléctrica y Electrónica (05F6)

at the

UNIVERSIDAD POLITÉCNICA DE MADRID

July 2019

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Abstract

The future High Luminosity upgrade (HL-LHC) of the Large Hadron Collider (LHC) at CERN requires technical improvements in a wide range of areas such as the collimation system. The Crystal Collimation techniques, which are based on the use of bent crystals to deflect the halo particles, are under study as a possible upgrade due to their improved cleaning performance, particularly promising with ion beams.

However, the success of this technique requires a high positioning performance of the crystal with respect to the beam, which combined with the limitations imposed by the harsh operating environment in particle accelerators (ultra-high vacuum, high mixed-field radiation, high temperature resistance, potential external disturbances, etc.) constitutes a multidisciplinary technological challenge.

The piezo goniometer system developed at CERN for Crystal Collimation is presented and the technological solutions involved are analysed. Its performance and limitations are studied, motivating the three main contributions of the thesis: the study of the radiation effects on the piezoelectric actuator, the proposal of novel absolute angle recovery method for the angular measurement system and the proposal of a phase-locked loop disturbance rejection algorithm for the improvement of the positioning performance.

The radiation effects on the piezoactuator are evaluated with three irradiation tests (gamma and proton irradiation) to simulate similar mixed-field conditions of LHC tunnel. The results show the suitability, under some considerations, of this type of actuator in this specific application and for mixed-field highly radioactive environments in general.

In the field of angular measurement, the dual-interferometer based setup used in the piezo goniometers is presented. A novel Absolute Angle Recovery Method is proposed to overcome the relative angle measurement limitation of the interferometric system. This technique allows the use of a device-specific curve as an absolute reference system to recover the angle in case of interferometer interruption or error, with no need of additional elements such as home switches. The method is simulated (Monte Carlo) and experimentally applied in the rotational stage with excellent

results.

In the field of closed loop control, a phase-locked loop disturbance rejection algorithm is proposed with the capability of attenuating disturbances at frequencies beyond the bandwidth of the original controller. The proposal is simulated including the possibility to update the disturbance observer and the various control techniques are benchmarked. The novel technique is successfully implemented experimentally in a rotational stage used for Crystal Collimation at CERN.

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Acknowledgments

I would like to express my most sincere gratitude to my CERN supervisor, Dr. Mark Butcher, for his never-ending technical support, encouragement and patient with the development and writing of this thesis. I am also pleased to say thank you to Dr. Alessandro Masi and Dr. Mario Di Castro, for their time, technical suggestions and for giving me this unique opportunity at CERN.

I would like to thank also my UPM supervisor, Dr. Jesús Ángel Oliver, for his guidance, help and insightful suggestions despite of the distance.

I will not forget to thank the numerous colleagues that have shared with me these years and contributed in one extent or another to this thesis, with special mention to Dr. Picatoste, Niklas, Lucian, Jorge and Santiago.

A nivel personal me gustaría agradecer enormemente a mis padres, Juan Francisco y Loli, por su ejemplo, apoyo y ayuda incondicional en todos los ámbitos de la vida, así como a toda mi familia.

Y por último, agradecer a la persona más maravillosa que jamás he conocido y el mayor regalo que Dios me ha hecho: Ania. GRACIAS POR SER COMO ERES y hacer que hasta lo más difícil merezca la pena.

Y sobre todo, gracias a Dios por hacerlo todo posible, incluida esta tesis, y la infinidad de regalos no merecidos desde que nací.

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Chapter 1

Introduction

The exploration of fundamental particle physics beyond the limits of current knowledge requires an ever increasing energy in the colliding particles. This aim implies constant advances in the state of the art of particle accelerator technology, posing new challenges never faced before.

Examples of these challenges can be found in nearly all technologies involved in the particle colliders. In some cases, the new requirements are achieved through the improvement of the existing techniques. However, there are other fields such as the beam collimation, where it is necessary to consider novel techniques such as Crystal Collimation.

The success of any novel technique relies initially on the correct study and feasibility assessment of the principle of operation. Although necessary, it is not sufficient for the successful application in practice, which requires a complete technical development for the practical implementation.

This thesis presents the high precision piezogoniometer developed at the European Organisation for Nuclear Research (CERN) for Crystal Collimation and specifically innovative contributions made for its development. The operation of the device is studied focusing on its different parts and performance, and presenting its limitations as the starting point for the three main contributions of the thesis: a study of the radiation resistance of the piezoelectric actuators, a proposal of an absolute angle recovery method for the angular measurement system and a proposal of a phase-locked

loop technique for disturbance rejection beyond the original controller bandwidth.

1.1 Thesis organisation

In this initial Chapter 1, Crystal Collimation is presented, in the framework of Large Hadron Collider (LHC) at CERN, as one of the most promising alternatives to the conventional collimation system.

The piezgoniometer system developed at CERN for Crystal Collimation is presented in Chapter 2. The system is thoroughly described in its different versions, evaluating its performance and remarking the challenges involved. The limitations of the system are identified and the contributions of the thesis are motivated with respect to them.

The Chapter 3 presents the various radiation resistance tests performed on the specific piezoelectric actuator under similar conditions to the LHC tunnel. The results show the suitability of this type of actuator for mixed-field highly radioactive environments such as the LHC.

A novel Absolute Angle Recovery Method (AARM) is proposed in Chapter 4. The objective of this technique is to overcome the relative angle measurement limitation of the interferometric system. The technique is simulated (Monte Carlo) and experimentally applied with good results to a rotational stage similar to the one used in the LHC piezgoniometer.

The last contribution of the thesis is presented in Chapter 5 and consists of a phase-locked loop disturbance rejection algorithm capable of attenuating disturbances at frequencies beyond the bandwidth of the original controller. The proposal is simulated including the possibility to update the disturbance observer and the performance of the studied control techniques are benchmarked. The novel technique is successfully implemented experimentally in the rotational stage used for Crystal Collimation at CERN, showing significant attenuation at frequencies well beyond the controller bandwidth.

Although each chapter includes its own conclusions, the Chapter 6 summarises

the most important points and achievements of the thesis, and presents the future lines of work.

1.2 Circular colliders

A circular collider is a machine where counter-circulating beams at high energy collide at specific points, generally called detectors. The charged particles are accelerated to gain the targeted energy with the effect of resonant electric fields, and they are steered in a circular trajectory using the Lorentz force affecting charged particles when moving within a magnetic field [145].

The operation of circular accelerators are based on three main elements: the Radio Frequency (RF) cavities that provide the energy lost by synchrotron radiation and accelerate the particles by resonant alternation of the electric field along the beam direction, the dipole magnets that produce a vertical magnetic field to steer the particles in the horizontal plane, and a sequence of orthogonal quadrupole magnets that produce a focussing effect on the particle beam.

The general performance of circular colliders can be evaluated with two main parameters that are:

- Luminosity: number of events per second and per cross section of the event.
- Collision energy: the sum of the energies of the colliding particles, corresponding to the case of a perfect centre-of-mass collision.

1.3 The Large Hadron Collider (LHC) and High Luminosity upgrade (HL-LHC) at CERN

The Large Hadron Collider is the most powerful particle accelerator in the world and constitutes the largest ring of the complex at the European Organisation for Nuclear Research in Switzerland. It is a circular collider with a circumference of approximately 27 km at a mean depth of 100 m underground and, as shown in Figure

Figure 1-1: Location of LHC tunnel across France and Switzerland.

1-1, it is situated across the French-Swiss border near Geneva. It has a design beam energy of 7 TeV that results in a total collision energy of 14 TeV when two counter-circulating beams collide in any of its four collision points: ALICE, ATLAS, CMS and LHCb.

The general objective of the LHC at CERN is to reveal the physics beyond the Standard Model by the study of high energy collisions, which reproduce conditions close to those at the instant of the Big Bang. The accelerated hadrons, protons or heavy ions, increase their energy in successive smaller accelerators, passing through the Proton Synchrotron (PS), Super Proton Synchrotron (SPS) and they are finally transferred to the LHC to reach their top energy. The general layout of the CERN complex is presented in Figure 1-2.

The operation of the accelerator works in runs of approximately three years, as shown in the LHC Schedule Figure 1-3, followed by periods of shutdown used for maintenance and upgrade of the equipment.

The Run I (3.5 TeV per beam) and Run II (6.5 TeV per beam) were of great success in terms of physics production. The performance of LHC overcame its design peak luminosity of $10^{-34} \text{cm}^{-2} \text{s}^{-1}$ already in 2016 and the integrated luminosity for ATLAS and CMS has reached 150fb^{-1} at the end of Run II, with successive yearly

The CERN accelerator complex *Complexe des accélérateurs du CERN*

LHC - Large Hadron Collider // SPS - Super Proton Synchrotron // PS - Proton Synchrotron // AD - Antiproton Decelerator // CLEAR - CERN Linear Electron Accelerator for Research // AWAKE - Advanced WAKEfield Experiment // ISOLDE - Isotope Separator OnLine // REX/HIE - Radioactive EXperiment/High Intensity and Energy ISOLDE // LEIR - Low Energy Ion Ring // LINAC - LINEar ACcelerator // n-ToF - Neutrons Time Of Flight // HiRadMat - High-Radiation to Materials // CHARM - Cern High energy AcceleRator Mixed field facility // IRRAD - proton IRRADIation facility // GIF++ - Gamma Irradiation Facility // CENF - CErn Neutrino platForm

Figure 1-2: CERN Accelerator Complex.

Figure 1-3: LHC and HL-LHC luminosity schedule (achieved and foreseen) with the periods of long shutdown (LS) and the starting point of HL-LHC in 2026. Integrated luminosity in blue and peak luminosity in green [120].

records. The accelerator is now (2019) in its Long Shutdown 2 (LS2) where LHC injectors would be updated among several other equipment with the purpose to reach the design collision energy of 14TeV and aim for higher luminosity.

In 2016, the High Luminosity upgrade project for LHC (HL-LHC) was approved [56] with the purpose of achieving instantaneous luminosities of a factor five times larger than the LHC nominal value, with an integrated luminosity of 3000fb^{-1} by 2037 as shown in Figure 1-3. This upgrade is considered as compulsory for scientific progress and implies numerous changes and challenges in accelerator equipment.

1.4 Beam collimation system

The main objective of the LHC Collimation System [7] is the protection of the superconductive magnets and other accelerator equipment from stray particles. These high energy particles can produce serious damage or lead to quenches in the case of the magnets. The protection task is achieved by cleaning the beam halo, which is

formed by all the beam particles outside the core region defined as $\pm 3\sigma$, where σ is the standard deviation of the transverse particle density distribution.

Besides the halo cleaning, the collimator system also provides: passive protection to the machine, cleaning the physics debris generated by collisions in the interaction points, radiation localisation by the concentration of the losses in shielded areas and, in some cases, dedicated protection acting as shield for specific equipment. In addition to this, the Beam Loss Monitors (BLM) in the collimation system can be used as a diagnostic tool, with the possibility to scrape and shape the beam.

The current LHC Collimation system is formed of, approximately, 45 movable collimators per beam ring, with a total of more than a hundred devices considering also the transfer lines. The system is based on a multi-stage setup with primary, secondary and tertiary collimators in cascade, that are mainly located in the betatron cleaning (IR7) and momentum cleaning (IR3) regions as shown in Figure 1-4, and constitutes one of the main sources of impedance to the beam.

Despite various collimator versions, the principle of operation is always based on the passing of the beam between two movable jaws made of very resistant materials, such as carbon-fibre composites or tungsten alloys. As shown in the collimator representation of Figure 1-5, the distance between the jaws can be reduced or increased with the action of four stepper motors, forming the collimator gap that affects the beam halo through diffractive interaction.

The LHC collimation system has presented an outstanding performance, stability and reliability during Run I and II. Its contribution to the accelerator success, that has reached and overcome its expected luminosity among other design parameters, is clear.

The performance of the collimator system can be evaluated with the *local cleaning efficiency* parameters that corresponds to the number of protons lost per unit length at a certain position of the ring, normalised to the total number of losses [120].

The LHC Collimation team has widely studied the effects and implications of the energy increase foreseen by HL-LHC upgrade. The cleaning performance should be improved by the same factor 2 as the beam energy increase [118] for the correct

Figure 1-5: Example of collimator design (half-section) [124].

magnet protection. This increase poses a real challenge to the current collimation system.

The main performance limitation is represented by losses in the Dispersion Suppressors (DS) area [120], which is planned to include a DS collimator as the main improvement for HL-LHC. This solution is expected to provide sufficient cleaning performance and correct layout integration for the case of proton collimation. However, in the case of ion beams, the collimation efficiency is reduced two orders of magnitude due to the nature of the beam. As experienced during Run II, the loss levels with ion beams are very close to the quench limit of the magnets [72]. Apart from the DS collimators, the LHC Collimation team is studying novel techniques such as Halo Diffusion Control systems or Crystal Collimation.

1.5 Crystal collimation physics

Materials with an internal crystalline lattice present particular properties that are exploited in a wide range of applications. The interaction of charged particles with these types of materials shows very interesting mechanisms depending on the relative orientation of the charged particle and the crystal lattice structure, effects that do not take place in amorphous materials.

When the crystalline structure is highly ordered, and under certain circumstances, the charged particles can experience a coherent interaction resulting in its confinement within the electrostatic potential generated by the atoms of the crystalline structure. This interaction, known as channelling, occurs when the transverse momentum of the particle is lower than the height of the potential well. Depending on the crystalline structure the particle can be trapped in between planes, known as planar channelling (CH), or axial channelling (AC).

The channelling effect takes place when the relative angle between the crystalline planes and the particle directions is within a maximum impact angle [95]. This critical angle depends on the particle energy and the crystal material that determines the height of the potential well. Besides that, if the crystal lattice is bent as shown

Figure 1-6: Diagram of a channeling deviation θ_b of a particle in a bent crystal. Particle path in red and crystal planes in blue [120].

(a) Quasimosaic crystal developed by PNPI.

(b) Strip crystal developed by INFN.

Figure 1-7: Examples of quasimosaic and strip crystals used in LHC.

in Figure 1-6, and under special conditions, high energy beam particles could follow the crystal curvature, resulting in the deflection of the beam [19].

Different materials, such as germanium, tungsten or silicon, has been studied to form the bent crystals, with the last option as the most attractive due to the lower manufacturing price. The bend of the crystal, which is optimal at the range of $40 - 60 \mu rad$ for crystal collimation [120], is a challenging process that has led to the study of two different crystal types: strip (ST) and quasi-mosaic (QM) crystals, shown in Figure 1-7, whose main difference is in the use of (110) or (111) planes respectively [120].

Apart from the desired channelling effect, other coherent interactions can occur in a bent crystal depending on the particle direction and energy such as:

- Dechannelling (DC): the particles in channelling might lose the channelling conditions if any interaction with the nuclei or electrons of the crystal modifies its transverse energy.
- Feed-in: opposite to the dechannelling effect, the interaction with nuclei or electron might modify its transverse energy to meet the channelling condition.
- Volume reflection: the particle is reflected by the average potential field, which is more frequent when the trajectory is tangential to the crystalline plane direction in bent crystals.
- Volume capture: in the same trajectory as volume reflection, the particle interaction can cause the loss of transversal energy to meet the channelling conditions.
- Skew plane channelling: the bent crystal has additional symmetries that are diagonal to the planes used for CH. These symmetries can also channel in charged particles, although at lower efficiency (in terms of potential well and deflective angle).

The proposed Crystal Collimation system for LHC [18] implies the substitution of the primary collimator for a bent crystal. The feasibility and implications has been widely studied in [120] and [103]. Crystal Collimation has proven to be a very promising technique and, although still under study, the results obtained during LHC Run I and II were a great success. The main advantages can be summarised as:

- **Beam impedance reduction:** Due to the tight positioning of conventional collimators with respect to the beam, their impact in the beam impedance is high. On the other hand, a small bent crystals of only 2-4 mm has an equivalent bending effect of hundreds of Tesla, which can reduce significantly the beam impedance with respect to conventional collimators.
- **Cleaning efficiency increase:** The high rate of nuclear interaction between halo particles and the standard collimators leads to its relatively low efficiency

when compared with a well-aligned crystal with low number of interactions. The efficiency gain is expected to be in the range of a factor 5 to 10 according to the simulations in [103], with special interest for the case of heavy ion beams.

1.6 Crystal collimation system requirements and challenges

The final requirements for a Crystal Collimator operational device in LHC were assessed at CERN, gathering all the inputs of the different experts involved.

The various coherent crystal-beam interactions presented in section 1.5 were initially considered to achieve the deflection of the particles, specially the volume reflection that would lead to much wider angular acceptance with respect to the CH effect. However, the CH presents a much lower nuclear interaction with the beam, larger single-pass deflection and, as shown in Chapter 2, the technological solution developed at CERN makes it possible to fulfil the required extreme angular precision.

The principle of operation is illustrated in Figure 1-8. The bent crystal is positioned with the linear and rotational axes to achieve the desired channelling effect in halo of the beam.

At 7 TeV, the studies showed an expected critical channelling angle of about $2.3\mu rad$ [103] and therefore much higher resolution, repeatability and angular stability is needed in the positioning system for the success of the technique. Besides that, the resolution and accuracy has to be maintained along the linear axis range in order to correctly maintain the CH during the energy ramp.

In terms of performance, the Table 1.1 summarises all the requirements of the project in terms of control and positioning (linearly and rotationally) of the crystal with respect to the beam.

Additionally, the crystal should be able to physically touch the beam. It means that the crystal and positioning system shall be placed inside a vacuum tank where all components must be ultra-high vacuum (UHV) compatible with the beam vacuum

(a) Linear axis to position the bent crystal in (b) Rotational axis to achieve the critical channelling angle between the bent crystal and the beam direction.

(c) Bent crystal in channelling with the beam halo being deviated.

Figure 1-8: Principle of operation of the Crystal Collimators.

Property	Specification
Linear stroke	$> 50mm$
Linear resolution	$5\mu m$
Linear accuracy	$\pm 20\mu m$
Total angular range	$\pm 10mrad$
Angular resolution	$0.1\mu rad$
Angular accuracy over the entire linear range	$\pm 1\mu rad$
Angular overshoot	10%
Yaw angular settling time	$20ms$
Yaw angular max speed in scan mode	$50\mu rad/s$
Pitch angular accuracy over last 10mm travel	Few urad
Roll angular accuracy over last 10mm travel	Few tens of urad

Table 1.1: Requirements and specifications of piezgoniometers for Crystal Collimation in LHC.

system that ranges from 10^{-10} to $10^{-11}mbar$ [38].

This UHV requirement dramatically restricts the materials choices for the positioning system. Conventional materials are not generally allowed, such as the majority of organic compounds - glues, plastics - or magnetic metals.

Besides that, the commissioning of new devices in LHC implies a large series of validation processes. The vacuum acceptance preparation includes the heating of the device to high temperatures ($180 - 220^{\circ}C$) for several hours while under vacuum. This process, which can be performed several times and is colloquially called as "bake out", leads to the removal of water and residual out-gassing of the components.

Additionally, as will be detailed in chapter 3, the physical location of the piezgoniometers in IR7 of LHC is a highly-radioactive area. According to irradiation simulations for the specific location, the system should be able to withstand 10 MGy mixed-field accumulated dose for the 10 years of expected lifetime.

In summary, apart from the high performance requirements of the application, the system must be compatible with the harsh environment of particle accelerators. In the case of IR7 of LHC, these additional requirements are:

- Ultra-high vacuum compatibility (range 10^{-10} to 10^{-11} mbar [38]).

- High-temperature resistance ($180^{\circ} - 200^{\circ}C$).
- Mixed-field radiation hard (10 MGy expected accumulated dose in 10 years).

In addition to this, from a repeatability point of view, an absolute angular reference system would be very convenient to guarantee the same channelling angle position over time in order to have a practical operation of the crystal collimation technique. Otherwise, the channelling angle position should be confirmed at the beginning of each use.

Chapter 2

Piezo goniometers for Crystal Collimation

The successful operation of the Crystal Collimation system relies on the fulfilment of the requirements presented in section 1.6 in terms of performance and compatibility with the particle accelerator environment at CERN. These conditions represent a technological challenge not solved in any commercially available rotational stage or system.

A general description of the different components and possibilities involved in high precision positioning systems are presented in section 2.1. The in-house (CERN) development of a piezo goniometer device for crystal collimation is detailed in section 2.2, and its limitations and the motivation of the thesis are analysed in section 2.3.

2.1 High-precision positioning systems

Positioning systems have experienced a great development over the last twenty years thanks to the important progress in fields such as material science, nanotechnology, micro electronics or high-precision mechanical manufacturing.

The aim of this section is to provide a quick review of the state-of-the-art regarding the different elements involved in high-precision positioning systems. This review will help to motivate and situate the piezo goniometer system presented in section 2.2.

2.1.1 Motorised-based linear actuators

There are a large number of linear actuators types that are commercially available in the industry. However, when the positioning resolution falls in the sub micron region, the possibilities using electric motors are basically reduced to two main groups:

- **Motorised screw type actuators:** the rotational movement of the motor (stepper, DC or brush-less DC) is transformed into linear displacement through a drive screw mechanism. These stages are available in various accuracy, resolution and range specifications, with ranges of more than a meter in some cases. Great improvements have been achieved in the reduction or suppression of mechanical problems such as backlash.
- **Linear drives:** these direct-drive systems are backlash-free thanks to the absence of drive screws or gear-heads. Their principle of operation allows unlimited travel ranges and their operation can be based on different linear motor types such as linear servo motor, linear stepper motor, voice coil motor or piezoelectric stepping motor, each of them with its particular characteristics.

2.1.2 Piezoelectric transducers

Piezoelectric transducers are the preferred choice in compact high precision positioning systems due to the combination of their high stiffness, compact dimension, vacuum compatibility and virtually infinite resolution, which makes them stand out from the motor-based linear actuators.

2.1.2.1 Piezoelectric effect

The concept of piezoelectricity was initially proposed by C. Coulomb in 1784 when he suggested that electric charges could be created by applying pressure in some materials [13]. However, the first experimental results confirming this theory took place more than a hundred years after, in 1880, when the brothers Pierre and Jacques Curie [33] discovered the existence of crystals in nature (quartz, boracite, Rochelle

salt, etc.) with the capability to produce electric charges when under mechanical stress. The following year G. Lippmann proposed the existence of the inverse effect, in other words, a strain in a material as a response to a voltage difference. The inverse effect was experimentally confirmed by the Curie brothers the following year. The first model explaining this effect was proposed by Lord Kelvin in 1893, but the first applications of these materials (quartz in ultrasonics) did not appear until some years later, in nearly 1920.

The *direct piezoelectric effect* can be described as the capacity of a material to generate an electric charge when an external force is applied. Inversely, the mechanical expansion of a material produced by the application of an electric field parallel to the polarisation of the material is called the *inverse piezoelectric effect*. This non-conventional stimulus-response effect is highly anisotropic and it is present in some crystalline materials whose structure has no centre of symmetry. The direct and inverse mechanism coexist in these materials, giving the possibility for them to be used as sensors (direct effect) or transducers (inverse effect).

The piezoelectric effect in natural monocrystalline materials is very limited and they were quickly surpassed in the 1960s by the development of artificial polycrystalline ceramics, initially with the barium titanate (BaTiO_3) and later with the lead-zirconate-titanate materials, that nearly doubled the piezoelectric response.

The lead-zirconate-titanate material, known as PZT, is formed by nearly same parts of lead zirconate (PbZrO_3) and lead titanate (PbTiO_3) forming a binary solid solution. The extraordinary piezoelectric properties are achieved thanks to their crystal cubic perovskite structure, with general form ABO_3 shown in Figure 2-1.

After the different processes needed in the PZT manufacture, the material presents randomly oriented domains with no piezoelectric properties. The domains are the microscopic regions of the material with net polar orientation. In order to acquire its piezoelectric properties it is necessary treat the material with a *polling* process that reorients the domains and confer a net polar orientation.

The polling process consists initially of the heating of the material to near its Curie Temperature, the temperature at which the crystal structure becomes symmetrical

Figure 2-1: Crystal cubic perovskite structure ABO_3 [73].

(losing its previous polarisation). At this temperature, a strong electric field is applied and maintained during the cooling process, resulting in a remanent general alignment of the dipoles as represented in Figure 2-2.

It is important to remark that most of the piezoelectric materials also present a pyroelectric response, which is the electric charge generation due to temperature variation, although the details of this effect exceed the scope of the thesis.

The constitutive equations for a general piezoelectric material are presented in equations 2.1 and 2.2, where the *IEEE Standard on Piezoelectricity* is followed with the conventional tensor notation, the indices $i, j, k, l = 1, 2, 3$ and summation in all repeated indices:

$$T_{ij} = c_{ijkl}^E S_{kl} - e_{kij} E_k \quad (2.1)$$

$$D_i = e_{ikl} S_{kl} + \epsilon_{ik}^T E_k \quad (2.2)$$

where T_{ij} is the stress tensor, S_{kl} is the strain tensor, c_{ijkl}^E correspond to the elastic constants under constant electric field (Hooke's tensor), e_{ikl} are the piezoelectric constants (C/m^2) and ϵ_{ik}^T are the dielectric constants under constant strain.

Figure 2-2: General polling process in PZT material [46].

The equations 2.1 and 2.2 can be rewritten with T_{kl} and E_k as independent variables as:

$$S_{ij} = s_{ijkl}^E T_{kl} + d_{kij} E_k \quad (2.3)$$

$$D_i = d_{ikl}T_{kl} + \epsilon_{ik}^TE_k \quad (2.4)$$

where following the previous notation s_{ijkl}^E is the tensor of compliance at constant electric field, d_{ikl} correspond to the piezoelectric constant (C/N) and ϵ_{ik}^T is the dielectric constant at constant stress.

Alternatively, the tensor notation can be substituted by the vector notation using:

$$T = \begin{pmatrix} T_{11} \\ T_{22} \\ T_{33} \\ T_{23} \\ T_{31} \\ T_{12} \end{pmatrix}, \quad S = \begin{pmatrix} S_{11} \\ S_{22} \\ S_{33} \\ 2S_{23} \\ 2S_{31} \\ 2S_{12} \end{pmatrix} \quad (2.5)$$

that converts the equations 2.3 and 2.4 into:

$$\begin{pmatrix} S_{11} \\ S_{22} \\ S_{33} \\ 2S_{23} \\ 2S_{31} \\ 2S_{12} \end{pmatrix} = \begin{pmatrix} s_{11} & s_{12} & s_{13} & 0 & 0 & 0 \\ s_{11} & s_{12} & s_{13} & 0 & 0 & 0 \\ s_{11} & s_{12} & s_{13} & 0 & 0 & 0 \\ 0 & 0 & 0 & s_{44} & 0 & 0 \\ 0 & 0 & 0 & 0 & s_{55} & 0 \\ 0 & 0 & 0 & 0 & 0 & s_{66} \end{pmatrix} \begin{pmatrix} T_{11} \\ T_{22} \\ T_{33} \\ T_{23} \\ T_{31} \\ T_{12} \end{pmatrix} + \begin{pmatrix} 0 & 0 & d_{31} \\ 0 & 0 & d_{32} \\ 0 & 0 & d_{33} \\ 0 & d_{24} & 0 \\ d_{15} & 0 & 0 \\ 0 & 0 & 0 \end{pmatrix} \begin{pmatrix} E_1 \\ E_2 \\ E_3 \end{pmatrix} \quad (2.6)$$

$$\begin{pmatrix} D_1 \\ D_2 \\ D_3 \end{pmatrix} = \begin{pmatrix} 0 & 0 & 0 & 0 & d_{15} & 0 \\ 0 & 0 & 0 & d_{24} & 0 & 0 \\ d_{31} & d_{32} & d_{33} & 0 & 0 & 0 \end{pmatrix} \begin{pmatrix} T_{11} \\ T_{22} \\ T_{33} \\ T_{23} \\ T_{31} \\ T_{12} \end{pmatrix} + \begin{pmatrix} \epsilon_{11} & 0 & 0 \\ 0 & \epsilon_{22} & 0 \\ 0 & 0 & \epsilon_{33} \end{pmatrix} \begin{pmatrix} E_1 \\ E_2 \\ E_3 \end{pmatrix} \quad (2.7)$$

Assuming that the piezoelectric material has one-dimensional polarisation, the electrical field is parallel to the polling direction and that, by convention, index 3 is aligned to the polling direction of the material, the equations 2.3 and 2.4 are reduced to:

$$S = s^E T + d_{33} E \quad (2.8)$$

$$D = d_{33} T + \epsilon^T E \quad (2.9)$$

where D is the electric displacement (C/m^2), E is the electric field (V/m), T is the stress (N/m^2), S is the strain, ϵ^T is the dielectric constant (permittivity) at constant stress, s^E is the compliance at constant electric field (inverse of Young's modulus) and d_{33} is the piezoelectric constant (m/V or C/N)

Figure 2-3: Different operating modes of a piezoelectric actuator in positioning systems [46] with the corresponding piezoelectric constant in use.

2.1.2.2 Piezoelectric transducers

The possibilities offered by piezoelectric materials can be understood looking at the wide range of forms and shapes of the transducers developed over the years. The applications range from general purposes devices (such as microphones, spark generators or accelerometers) to industrial applications (ultrasonic transducers) and high-technology systems (aerospace systems, precision positioning and structural vibration control).

The tensor notation in equations 2.3 and 2.4 reflects the variety of modes of deformation in these materials. However, in the case of precision positioning systems the modes of operation are normally restricted to ones presented in Figure 2-3.

According to these deformation modes, the variety of transducers can be classified as:

- **Uni-morphs and bi-morphs actuators:** allow the bending motion based on the expansion (or contraction) difference between opposing layers, which leads to large displacements but with very low force.
- **Quarter-sectored tube-shaped actuators:** allow 2-d and 3-d positioning

Figure 2-4: Examples of piezoelectric actuators from Noliac's catalogue: double bi-morphs (a), quarter-sectored tube-shaped(b) and stack (c) actuators.

based on the tube-shape structure that is poled radially, with the electrodes in the inner and outer surfaces as shown with further details in [46]. They are widely used in scanning probe microscopy.

- **Stack actuators:** allow a linear stroke by the accumulation of thin layers of piezoelectric materials between electrodes. The polarisation direction is perpendicular to the electrode layers and determines the stroke and blocking force direction. They are the preferred choice for high-speed applications thanks to their high stiffness and force output, and can be arranged in a wide variety of forms (single and multi-stack).
- **Shear actuators:** allow a shear displacement by the application of an electric field perpendicularly to the polarization direction, exploiting the shear-strain coefficient d_{15} . This use makes a bipolar operation possible with high force but short displacement compared with stack actuators.

Examples of these configurations can be found in Figure 2-4.

2.1.2.3 Piezoelectric response and non-linearities

Unfortunately, the real response of the piezoelectric material is highly non-linear and the precise control of these transducers implies a real challenge. The main non-linear effects can be summarised as:

- **Hysteresis:** constitutes the larger non-linear effect between the voltage applied and the resulting strain in the actuator. It is a common effect in all ferromag-

netic materials and it was widely studied since its discovery, simultaneously to the piezoelectric effect. The mechanism explaining this behaviour is still under discussion, but there is a consensus associating it to irreversible losses produced in the interaction between similarly oriented electric dipoles when an electric field is applied [78].

The mechanism is better understood using the concept, initially proposed for magnetic materials, of domains of polarization (regions of similar oriented electric dipoles) separated by imaginary boundaries called domain walls. When an electric field is applied, the dipoles start to reorient but present a preference for their original orientation that retard the motion of the domains walls, resulting in the hysteresis response. The internal forces producing this retard are associated with internal frictions, between the dipoles and the domain walls, and material defects.

Besides that, the hysteresis effect presents a “memory-type” effect of the previous state of the material, which constitutes a challenge in high-precision applications and results in the characteristic hysteresis loop shown in Figure 2-5 when the actuator is cycled in its full stroke.

- **Creep effect:** is the low-frequency drift that piezoelectric materials exhibit after the response to a voltage step. This effect, known as creep and shown in Figure 2-6, occurs in the same direction as the dimensional change produced by the voltage step and can be approximated by a logarithmic response with time constant of, typically, a few minutes. It is a rate-dependent deformation experienced at a constant electric field explained as the slow increase of remnant polarisation once the final voltage is achieved.
- **Temperature dependence:** several studies [74] [4] [99] [73] show how the piezoelectric properties vary widely with the temperature and present higher resistance to depolarisation at low temperatures. Consequently, the temperature range of operation in a piezoelectric application has to be considered carefully and temperature compensation included if necessary.

Figure 2-5: Hysteresis effect in piezoelectric materials. Example from piezoelectric actuator used for Crystal Collimation in the LHC.

Figure 2-6: Creep effect in a piezo actuated rotational stage used for Crystal Collimation in the LHC.

In addition to the non-linearities, it is important to remark that the piezoelectric actuators is only one of the various components forming a complete system in an application. For this reason, the whole system has to be studied from a dynamic point of view, analysing all structural resonances that could be excited by specific setpoint signals (triangular waveforms for example) or by external disturbances.

In general, the bandwidth of a piezo-actuated system is limited by its first resonance frequency, and for this reason the scan rates in positioning applications are limited to 1/10th - 1/100th of this frequency. It is possible to increase the bandwidth using stiffer piezo actuators, but at the cost of shorter stroke.

From an electric perspective, a piezoelectric actuator can be considered as capacitive load acting as low-pass filter when combined with the amplifier output impedance and cable inductance. Therefore, the bandwidth of the amplifier plays an important role that, in some cases, can limit the bandwidth of the whole system.

2.1.3 Mechanical design

The mechanical design of a positioning device plays a critical role in the final performance of the system, with a direct impact on its overall accuracy and parasitic errors. As was commented before, the bandwidth of a positioning system is related to its first resonance frequency, which is determined by the combination of actuator and mechanical stage.

The mechanical stiffness and strength of the stage is determined, mainly, by its mechanical design, which also determine the thermal stability of the stage.

2.1.3.1 Flexure-based systems

A flexure hinge is a thin elastic element connecting two rigid bodies that allows a limited rotation through bending or flexing [43] as can be seeing in Figure 2-7. They differ from the conventional mechanical joints in the non-collocation of the centres of rotation of the rigid bodies.

The main advantages with respect to mechanical joints are the lack of friction

Figure 2-7: Flexure hinge circular (a), corner filleted (b) and elliptical (c) [?].

losses, lubrication is not needed and they present no wear nor other non-linear effects such as hysteresis or backlash. Their compactness and micro-scale possibility, since the flexure-stage can be manufactured from a single block, are particularly important for precision positioners.

Despite the limited movement of the flexure hinges, their properties have converted them into the preferred choice for high-precision applications with a wide range of flexure-based systems in the industry. The flexure hinges are designed to maximize the stiffness in the non-actuated directions, minimizing the parasitic movements, while guiding the motion in the actuated direction.

The main failure mechanism for these systems are the yield, when the elastic deformation is exceeded, and fatigue, which should be considered in the material selection and design process.

2.1.4 Position sensors

Due to the presence of cree and hysteresis, closed loop positioning control is required. In this case, a positioning sensor is required to close the loop and guarantee the desired position. There are a wide range of possibilities nowadays for this type of sensor as described below.

2.1.4.1 Types of positioning sensors

- **Resistive Strain Sensor:** formed by a thin layer of conducting material, generally in a zig-zag form and laminated between two insulated layers. The principle of operation consists of the proportional increase of resistance when the strain gauge is elongated. The sensor is very simple, low-cost and can be

attached to the actuator surface. However, the correct measurement requires an electric bridge due to the small resistance variation and its sensitivity to electrical interference.

- **Piezoresistive Strain Sensor:** two orders of magnitude greater sensitivity than resistive strain sensors using piezoresistive materials. However these sensors have high temperature dependence, slightly non-linear response and low strain range.
- **Piezoelectric Strain Sensor:** piezoelectric transducers can be used in their inverse mode providing high sensitivity with low measurement noise at high frequencies, which convert them into suitable solutions for vibration control. Unfortunately, at low-frequencies their drift prevents their use in static positioning, in addition to their high sensitivity to temperature variations.
- **Capacitive Sensor:** there is a wide range of capacitive sensors based on the change of the capacitance between two electrodes with the geometry of the system. The operation of the sensor can be based on either the movement, lateral or axial, of one of the electrodes or the movement of the dielectric between electrodes. This type of sensor stands out for its linearity, resolution and bandwidth but requires additional electronic equipment for its reading.
- **MEMs capacitive and thermal sensors:** they are based on the same capacitive principle commented before, the MEMs capacitive sensors are smaller in size and therefore require special geometries to achieve acceptable capacitance values. There is a large variety of geometries, optimised for each application, and can include differential detection methods. On the other hand, the thermal sensors are based on the difference in current passing through two micro heaters at constant voltage. According to the overlap (with the heaters) of a moving heat-sink, the heaters achieve different temperatures resulting in a current difference. These last type of sensors are particularly compact with respect to the MEMs capacitive, and with similar or better noise performance.

- **Eddy-current sensors:** they are based on the magnetic induction effect produced by a coil in the proximity of an electrically conductive material. According to Lenz's law, the magnetic field generated by the coil (excited with AC) induces an eddy-current in the conductive target that produces an opposing magnetic field. As a result, the AC resistance of the coil varies depending on the distance to the target material. The electronics involved in the reading are similar to the capacitive sensors, however they are less widespread in positioning applications due to their lower range and temperature sensitivity.
- **Linear Variable Displacement Transformers (LVDT):** they consist of a driving coil that induces a voltage into two sensing coils, all of them mounted in a tube. A ferromagnetic core can move inside the tube producing a voltage difference in the sensor coils. The electronics involved in the reading are similar to capacitive or inductive sensors, and the driving coils are excited with AC with frequencies between 1-5kHz. The reading of this sensor is absolute, with infinitesimal resolution and large range. However, it is sensitive to lateral displacement of the core, with limited bandwidth and the large dimensions of the sensor prevent its use in compact positioning systems.
- **Linear encoders:** they are based on the reading by a head that is sensitive to an encoded pattern contained in a reference scale. The relative movement of the reading-head over the scale is translated into a position. The combination of reading-head and scale can be optical, capacitive or inductive among others. Different techniques were developed to reduce its high sensitivity to dust, contamination of the scale or poor alignment, and to increase its accuracy (up to $0.5\mu m/m$) or to offer absolute reading (in contrast to relative measurement).

2.1.4.2 Laser interferometry

Although laser interferometry constitutes just another type of position sensor, the importance, characteristics and variations justify a dedicated subsection.

The first interferometer was proposed by Michelson in 1880, where a monochro-

Figure 2-8: Michelson interferometer diagram [44], [123].

matic beam was split into two paths as shown in Figure 2-8. In this scheme, half of the beam is reflected in a fixed mirror, whereas the other half is directed to a movable mirror. The reflection from both mirrors are recombined at the sensor, where a characteristic fringe pattern can be observed due to interference.

The intensity of the recombined beam at a fixed point of the sensor increases or decreases with the movement of the mirror according to the constructive or destructive interference created by the path difference. The displacement of the mirror can be determined knowing that a cycle of intensity change correspond to half wavelength of the light (which is known beforehand).

The Michelson's configuration presents a series of limitations that prevent its use in real applications. The phase sensitivity in the maximum and minimum interference is zero, limiting the sub-wavelength displacement measurement, and it is highly sensitive to misalignments or movements of the reference mirror and beam-splitter.

An improved interferometer version was proposed by C. Fabry and A. Perot, where two highly reflective surfaces, facing each other, act as a resonator of light that results in a transmitted and reflected interferometer patterns depending on the distance between the reflecting surfaces. If one of the reflective surface is movable,

Figure 2-9: Fabry-Perot interferometer diagram.

the systems constitutes a Fabry-Perot interferometer shown in Figure 2-9, whereas if both surfaces are fixed the system is called an etalon.

The principle of operation is still the interference of light, but the multiple reflections in the Fabry-Perot resonator result in clearer fringes with respect to the Michelson interferometer, leading to a more accurate detection of half-waveform (of the light source) displacement.

Several Fabry-Perot configurations have been developed over the years but in general they can be classified as plane parallel, confocal or plane-concave setups, according to the form of the reflective surfaces involved (two flat surfaces, two concave surfaces or a flat and concave surface respectively). The details and equations of the Fabry-Perot interferometer can be found in [57].

A great improvement to the original setup was the Heterodyne interferometer proposed in [36] where the phase sensitivity is constant independently of the path length. It is used in most modern interferometers.

The Fabry-Perot setup developed in [80] is particularly interesting, where the resonant cavity is formed between the fibre connector and the reflective surface, separating the optical components from the laser and electronic equipment.

In this case, approximately 4% of the incident light is reflected in the optic fibre

connector forming the reference beam back into the equipment. The transmitted light is collimated in a lens and directed to a tilted reflective surface that returns the light back through the lens, arriving again to the fibre connector but off centred. At this point, the beam is reflected again in the fibre connector and the light travels back through the same path through the lens, it is reflected in the movable surface and back to the lens, finally arriving in the core of the fibre connection that creates a Fabry-Perot resonator with length equal to twice the distance between the fibre connector and mirror.

The reflected intensity is periodically modulated according to the reflective surface distance following the corresponding fringe pattern. Besides that, the light source wavelength is modulated at high-frequency (75 kHz) and, after demodulation in the electronic equipment, forms a quadrature signal with the general intensity. This technique, in combination with the reflected power, is used for direction detection and to achieve sub-wavelength resolution.

The detailed operation of this fibre-optic Fabry-Perot interferometer is presented in [139]. This configuration has lead to several industrial interferometer products that are already on the market and, as will be commented below, it is the positioning sensor used in this thesis.

Laser interferometry has a unlimited theoretical range, offers non-contact measurement and presents the highest accuracy, stability, linearity and resolution of all the position sensing technologies.

However, in terms of cost, it is also the most expensive due to the numerous elements (optical and electronic) involved. Additionally, this technique provides a relative measurement, in contrast with absolute measurement systems. For this reason, if the light beam is interrupted, the system experiences a power cycle or the maximum speed is exceeded (due to mechanical shocks for example), the interferometer position is lost and the system needs to be restarted from a known reference.

In general, the performance of laser interferometry exceeds the alternative solutions and is often used in high-precision positioning applications, including the piezo goniometers studied in this thesis.

2.1.5 Modelling and control

The non-linear behaviour of piezoelectric materials contributes to the difficulties when modelling accurately the plant, which is a key point for most control design techniques.

2.1.5.1 System identification

The Hammerstein structure proposed in [108] presents a widely used decoupling technique for modelling the non-linear response and the dynamics of the system. This separation was proven to be valid in a multitude of applications involving transducers with hysteresis behaviour as in [151] or [152].

The modelling problem can be split therefore into two independent problems focused on a specific behaviour of the material, where the two models will be put in cascade to represent the global actuator.

The non-linear behaviour, mainly rate-independent hysteresis, since creep effect can be compensated through feedback control, can be modelled with different approaches studied originally for ferromagnetic materials such as the experimental polynomial, Maxwell-slip, Duhem, Preisach or Prandtl-Ishlinkii modelling [15].

On the other hand, the remaining part of the Hammerstein structure is associated with the dynamic response of the system that, if assumed to be linear, can be modelled applying standard system identification with white, grey or black box models [96] depending on the prior knowledge of the plant.

2.1.5.2 Control techniques

The number of control approaches applied to positioning systems is very wide. In this subsection, the objective is to present a general overview of the different techniques in order to situate the control technique applied in the piezo goniometers and the proposals of Chapter 5.

The piezoelectric shunt damping technique [47] consists of the connection of an electrical RLC impedance to the piezoelectric actuator causing the reduction of the mechanical resonance modes, that in some cases reach the 20db attenuation of the

first resonance. This technique does not require any feedback sensor and can be applied in addition to open or closed loop controllers.

Feedback control is the most widespread technique used in high-precision positioning systems. Although the presence of a sensor reading is necessary to close the loop, the well developed control theory can be applied to design a controller (using conventional P, PI, PID or general pole-placement techniques) that is robust to model errors with relative simplicity and providing sufficient performance for some applications.

As was commented before, the first mechanical resonance of the system limits the achievable bandwidth for closed-loop control. Different techniques were developed over the years with the purpose of increasing the closed-loop bandwidth and phase-gain margins.

One of these techniques consists of the use of a notch filter to mitigate the main resonance of the system [89], giving the possibility to increase the closed-loop bandwidth even above this frequency [1]. The technique presents a high sensitivity to model errors and if the resonance frequency is not stable can lead to instabilities so it is only applied in cases with high model accuracy or where the feedback controller can be recalibrated dynamically.

Alternatively, damping control techniques increase artificially the damping response of the system, which leads to a possible increase in the feedback gain and closed-loop bandwidth. The main advantage is its low sensitivity to changes in the resonance frequency and it provides better disturbance rejection than the inversion techniques.

There are numerous examples of damping control applied to positioning systems such as the integral resonance control (IRC) [5, 16, 45], positive position feedback (PPF) [40], shunt control [47] or resonant control [129]

On the other hand, optimal control techniques like H_∞ in [128] or LMI controllers in [90] were also applied in positioning systems.

In those applications where the desired trajectories are periodic, which is the case of various scanning applications with triangular waveforms, specific techniques of repetitive control have shown remarkable results such as [83].

Although displacement sensors are the preferred choice to close the feedback loop, load force sensors can be used also. A possibility present in the literature is a simple piezoelectric block placed between the actuator and the moving structure, which presents some benefits such as the high sensitivity, compactness and low-cost. The transducer is then modelled as force actuator, instead of a displacement actuator, leading to a highly damped system that can be easily controlled with high damping performance and robustness as shown in [115, 116]. The force sensor presents the same non-linear behaviour that every piezoelectric transducer and an additional feedback loop, based on a displacement sensor, is needed for the precise control at low-frequencies.

Feedforward control techniques also constitute an alternative, and in some cases a complement, to the feedback structures. For its correct operation a detailed knowledge of the plant is necessary, including its dynamic and non-linear response, that is inverted to compensate the positioning errors.

The Hammerstein structure for piezo modelling commented before, is a useful technique that allows the design of an inversion-based feedforward controller inverting the plant dynamics and hysteresis separately, as in [32, 34].

However, the success of this technique resides in the exact knowledge of the plant, and therefore is highly sensitive to plant variations. For this reason it is applied in some cases in combination with feedback schemes.

The success of iterative feedforward techniques as in [148] is also remarkable, where most recent techniques require minimal previous knowledge of the plant [84] or in iterative learning control techniques [32]. The iterative techniques are not restricted to applications with periodic inputs, but it is in these case where they present better results.

2.1.6 Electrical equipment and considerations

From an electrical point of view, a piezoelectric actuator constitutes a capacitive load. When the actuator is driven in voltage, the response of the amplifier with such a high load can be a real challenge and, in some cases, can limit the maximum bandwidth

of the system. The main characteristics of the amplifier when driving piezoactuators are highly coupled between them and can be summarised in:

- **Bandwidth:** the value given by the manufacturer represent the maximum bandwidth of the device for a given load. However, this value depends on the load and for the case of high capacitances it can be highly reduced or even lead the internal feedback loop of the amplifier to instability.
- **Maximum current:** the required current is proportional to the frequency of operation and the capacitance of the piezoactuator, and therefore it can be a limiting factor.
- **Maximum power dissipation:** it refers to the maximum power dissipated in the output stage of the amplifier due to the difference between the internal supply voltage and the voltage delivered in the load. It reaches its maximum when the amplifier output is short-circuited and determines the maximum current.

The combination of high speed, large displacements (large voltage difference) with high capacitance of the actuator has led to the development of specific dual-amplifiers capable of delivering tens of amps with high bandwidths (in some cases reaching 500kHz).

2.2 Piezogoniometer for Crystal Collimation at CERN

This subsection presents the different parts and components involved in the piezo goniometer device developed at CERN for crystal collimation.

2.2.1 Linear movement

In order to understand the overall structure and functionality of the piezo goniometers, it is important to remark that the system was initially conceived as an experimental prototype device. It means that, unlike other operational devices, its use was originally restricted to the Machine Development (MD) periods (one of the several

Orientation independent angle nomenclature	Angle with respect to beam	
	Horizontal goniometer	Vertical goniometer
Rotation	Yaw	Pitch
Tilt	Pitch	Yaw
Roll	Roll	Roll

Table 2.1: Orientation independent angle nomenclature.

accelerator modes [2]). During these periods, generally of one week every two months, the numerous working teams at CERN can test their equipments and perform the research for the improvement of the accelerator without affecting the physics generation in the collision points.

The piezo goniometer design is presented in Figure 2-10. It is formed of two linear axes where one controls the linear position of the rotational stage with the crystal and the second axis drives a pipe section.

When the crystal is not in use (generally in any other accelerator mode apart from the MD), it can be completely retracted from the beam line to its parking position and the beam pipe section can be positioned at its inner limit, closing the beam pipe and making the device “transparent” to the circulating beam.

On the other hand, when the crystal is used, the beam pipe section is fully retracted and the rotational stage with the crystal is moved forward, closer to the beam line, for its operation.

The same piezo goniometer device can be placed in horizontal or vertical orientation by just the rotation of the tank. Figure 2-11 presents the actuated angle in each case with respect to the beam axis, and an orientation-independent nomenclature is proposed in table 2.1 for practical reasons.

The displacement of the rotational stage with the crystal is achieved through high-precision cross-rolled liner guides (parts in yellow in Figure 2-12). These guides have a nominal parasitic angle tolerance of $1\mu rad/mm$ and are responsible for the parasitic angles (tilt and roll angles defined in table 2.1) during linear axis displacement, so they are carefully measured and adjusted during the validation of the system.

(a) Piezo goniometer tank.

(b) Piezogoniometer section with beam pipe and linear axis.

(c) Linear axis with guides in yellow for the displacement of the rotational stage.

Figure 2-10: Piezo goniometer design.

Figure 2-11: Actuated angle for horizontal and vertical goniometer with respect to the beam axis.

Figure 2-12: Linear axis with cross-rolled linear guides in yellow.

Version	Stiffness	1st resonance	Piezo bakeout temperature	Location
V1	“Low”	≥ 35 Hz	$110^{\circ}C$	B1-H and B1-V
V2	“High”	≥ 130 Hz	$220^{\circ}C$	B2-H and B2-V

Table 2.2: Versions of the rotational stage installed in LHC.

2.2.2 Angular movement

The angular movement of the crystal is achieved through a piezo-actuated rotational stage whose elements, angular measurement system and control techniques are explained in the following subsections.

2.2.2.1 Rotational stage and piezoelectric actuator

The rotational stage is based on a flexural-hinge structure manufactured as one single piece. This mechanism is composed of a stiff frame connected to the rotational head through flexure-hinge blades as shown in Figure 2-13.

The geometry of the design restricts the movement of the head to a virtual axis of rotation, presenting a high stiffness response in any other direction, which minimises the parasitic angles during the angular actuation.

Two generations of the rotational stage were developed at CERN for this project as shown in Figure 2-13(a) and 2-13(b). The second generation was designed with higher stiffness in order to increase the first resonance frequency that, as commented before, is one of the limiting factors for the closed loop control bandwidth. The comparison between the two version is shown in Table 2.2.

The rotational stage is driven by a PZT piezoelectric stack actuator whose expansion produces the rotation of the head through an amplifier structure or a direct lever-mechanism, for version 1 and 2 respectively. This piezoelectric actuation is immune to magnetic fields, with a high force-to-size ratio and with infinite theoretical resolution.

The rotational stage is designed to produce a rotational stiffness that, combined with the selected piezoelectric actuator, guarantees the desired angular range even in

Figure 2-13: Different versions of the rotational stage.

the worst case scenario of the manufacturing tolerances. In the case of the version 2, the system also includes a pre-load spring to maintain the actuator always under a minimum load of 200 N and the expected stroke is shown in Figure 2-14.

The electronic equipment, explained in more detail below, includes a digital-to-analogue converter (DAC) of 20-bits (AD 5791) that, in combination with the piezoactuator, results in a theoretical resolution of $25nrad$. The voltage is amplified using a CEDRAT LC75B Amplifier [28] (for V1 stages) and FALCO WMA-280 Amplifier [39] (for V2 stages) with small signal bandwidths of 660Hz (360mA output current) and 100kHz (300mA output current) respectively.

2.2.2.2 Angular measurement system

The rotational angle is estimated with the dual-interferometer based angular measurement system presented in Figure 2-15. Two linear interferometer heads are fixed to the frame of the rotational stage and a gold-coated mirror is mounted on the rota-

Figure 2-14: Piezoactuator behaviour (blue) under the rotational stiffness produced by the rotational stage V2 (the pre-load is 200N). In the worst case, the $93.07\mu\text{m}$ corresponds to 23.2mrad rotation, which is greater than the desired 20mrad .

tional head of the stage that moves as shown in Figure 2-16. The mirror and crystal are attached to each side of the rotational head, moving all together, whereas the interferometer heads are fixed to the stage frame.

The angle is estimated using the reading of the two linear interferometer axes by simple trigonometry according to:

$$\theta = \tan^{-1} \left(\frac{d_1 - d_2}{D} \right) \quad (2.10)$$

where d_1 and d_2 are the interferometer distances measured by axis 1 and 2 respectively, and D is the distance between axes.

This configuration with two linear interferometric axes allows the use of standard and commercially available solutions. For this application, the high precision AttoCube interferometers FPS3010 and IDS3010 are employed, providing 1pm resolution, 2nm repeatability and 0.14ppm accuracy.

The high performance of the laser interferometry leads to the angular reading

Figure 2-15: Dual-interferometer based angular measurement system with flat reflective surface.

(a) Dual-interferometer based angular measurement system. (b) UHV tank with the rotational stage and angular measurement system. The crystal (not visible in the picture) is below the rotational stage.

Figure 2-16: Rotational stage with angular measurement system used in the piezo goniometers.

Figure 2-17: Hammerstein structure used for system modelling [25].

(with $D = 45mm$) with $0.022nrad$ resolution, $\pm 5nrad$ angular accuracy (evaluated with the uncertainty propagation law) and a reading rate of 240kHz.

As was commented before, laser interferometry provides only a relative measurement that in this case leads to a relative angle estimation, constituting the main limitation of the measurement system. Further details of this limitation and the angular measurement system can be found in chapter 4, where a novel method for absolute angle recovery is proposed.

2.2.2.3 Modelling and control

The plant of the system is formed by the flexure-based rotational stage and the piezoelectric actuator. The Hammerstein structure, commented before, is used for the modelling of the plant as shown in Figure 2-17. This technique allows the decoupling of the non-linear hysteresis behaviour and the dynamics of the system, which are modelled separately.

The hysteresis behaviour is modelled using the well-established rate-independent Maxwell-slip model [65]. The rotational stage is cycled in full voltage with a triangular signal to obtain the major hysteresis loop of the system that is introduced in the Maxwell-slip model identification with 10 elements (10 force coefficients, 10 position coefficients and 3 additional coefficients for a 2nd order polynomial general shape fit).

This hysteresis model identification yields a mathematical structure that reproduces the behaviour of the piezoactuator. But moreover, it gives the possibility to use its parameters to invert this hysteresis behavior, forming an inverted-hysteresis block H^{-1} that is included before the plant in the closed loop algorithm shown in Figure 2-18.

Figure 2-18: Closed loop algorithm used in the piezo actuated goniometers [25].

On the other hand, the linear dynamics of the system are identified through a black box model with fixed number of poles and zeros. Various pseudo random binary sequences (PRBS) are used as the excitation signals of the system. The PRBS signals were generated with a shift register order $n = 15$ and number of repetitions $N = 15$ for averaging of the system response. The amplitude of the PRBS depended on the rotational stage version with an amplitude of 0.2V for the version V1, and amplitude 0.03V for the version V2 due to its higher capacitance. In both cases the voltage amplitude was selected to prevent the saturation of the amplifier.

The response of the system to the PRBS signals was acquired for every point of the PRBS (at 2kHz) and the final system identification was performed in the z-domain with the System Identification Toolbox of Matlab, piezo goniometer with 5 zeros and 6 poles the version V1, and 11 zeros and 12 poles for the version V2.

The version V1 of the rotational stage presents its first resonance around $35Hz$. In this case, the controller is designed using a 2-DOF damping technique proposed in [25] where the high gain of the system resonance is used to obtain a high external disturbance rejection at this frequency.

The feedback controller, shown in Figure 2-18, is designed as a series combination of an proportional-integral-derivative controller PID, a lead filter that improves the phase margin and a notch filter at high frequencies for robustness.

The controller is designed leaving the first resonance un-cancelled, which produces a high attenuation at this resonance as can be seeing in its sensitivity function [25]. In addition to this, the control structure includes a pre-filter F , before the closed loop, to increase the tracking performance of the system.

The final control achieves a phase margin bigger than 50° , $10db$ gain margin and

7Hz bandwidth with the first resonance of the system at 35Hz.

On the other hand, the version V2 of the rotational stage presents its first resonance at a higher frequency, around 135Hz. The same control structure is employed in this case. However, the presence of additional resonances in the vicinity of its main resonance made the damping technique employed in the version V1 not suitable for this case. Instead, the inversion technique of cancelling the main resonance with a wide notch filter is applied, maintaining the PID control block in combination with a lead filter to increase the phase margin and additional notch filters at higher frequencies.

The closed loop controller achieved a phase margin bigger than 58° , gain margin higher than 10db and around 25Hz bandwidth with the first resonance of the system at around 135Hz as shown in Figures 2-19 and 2-20.

The electronic equipment implementing the whole control algorithm is a National Instrument PXI device with a NI-PXI 8115 controller and several multifunction I/O module and FPGA cards. The analogue voltage is generated in an additional cassette containing a single 20-bit DAC (AD 5791), whereas the interferometric positions is read through AttoCube's proprietary serial word protocol (HSSL) in one of the FPGA Cards at 240kHz.

The closed loop control algorithm was first implemented in real-time part of the National Instrument NI-PXI 8115 controller, and then migrated to the FPGA NI-PXI 7854R for higher performance and robustness. The closed loop runs at 2kHz, with the possibility to be increased in the future thanks to the high FPGA performance.

2.2.3 Goniometers in the LHC and SPS

During the run II of the LHC (2017-2018), two new piezongoniometers (horizontal and vertical) were successfully commissioned in Beam 2 of the LHC. It led to a total of 5 operational piezo goniometers for crystal collimation located on the Beam 1 of the LHC (horizontal and vertical), Beam 2 of the LHC (horizontal and vertical) and in the SPS (horizontal). All the goniometers in the LHC are located in the betatron cleaning region (IR7), whereas the SPS goniometer is in the BA5 zone.

(a) System identification (experimental response at different voltage offsets).

(b) Closed loop and complementary sensitivity function.

Figure 2-19: Example of system and closed loop response for goniometer B2H of LHC (BW 24.73Hz, GM 10.7db, PM 58.1°).

(a) Controller margins.

(b) System performance in closed loop.

Figure 2-20: Closed loop performance for B2H of LHC (BW 24.73Hz, GM 10.7db, PM 58.1°).

The performance for these goniometers during the Run 1 and Run 2 of the LHC is summarised in the Section 6.1 of the conclusive chapter.

2.3 Goniometer limitations and motivation of the thesis

It is important to remark that the current version of the piezo goniometers were conceived as experimental MD devices, where the LHC requirements in terms of performance and robustness are not as demanding as for operational devices.

At the beginning of this PhD period, the main points of the system that needed to be studied and were a limitation or concern for its eventual operation as part of the collimation system were:

- **Radiation resistance of the piezoelectric actuator:** Despite of some literature claiming the suitability of the piezoelectric transducers for radioactive environments, and the good results obtained during LHC Run I, there is a lack of radioactive endurance tests performed in the specific piezoactuators used in this application. It is therefore necessary to study the performance of the actuator over its lifetime under the specific mixed-field irradiation conditions present in LHC.
- **Relative measurement of the angle:** the dual-interferometer based angular measurement system presents a very high performance in terms of resolution and accuracy. However, the lack of absolute angular measurement constitutes a real limitation in this application where the location of the Channelling Angle should be guaranteed. The angular measurement system should be studied to provide a method that could solve, or at least mitigate, this key limitation for the eventual use of the Crystal Collimation system during normal operation of LHC, where an absolute angle measurement is needed.
- **External disturbance rejection of frequencies beyond the bandwidth of the controller:** particle accelerators constitute a highly noise environment

where multiple equipments are installed in a confined space and can induce external disturbances in their surrounding. For this reason, the piezo goniometers can be exposed to external vibrations due to the operation of close equipment like vacuum pumps or motors. At low frequencies, the disturbance would be rejected according to the sensitivity response of the feedback control system. However, the disturbance rejection beyond the bandwidth frequency should be studied and, if possible, a method for disturbance attenuation at these frequencies, without affecting the overall performance of the closed loop, should be proposed.

These three points constitute the main field of study of this PhD thesis and they are studied in detail in the following chapters.

Chapter 3

Radiation effects on piezoelectric actuators

3.1 Introduction

The piezogoniometer device presented in the previous chapter was proposed as a solution to the challenging requirements of the Crystal Collimation system in the LHC. The combination of high performance with the harsh operational environment (ultra high vacuum compatible and radiation) limits dramatically the allowed materials involved in rotational stage.

In such hard conditions, the piezoelectric materials constitute a suitable solution due to their ultra high vacuum compatibility (when specifically selected) and potential radiation resistance, apart from their electromechanical properties.

The objective of this chapter is to study the radiation effects on the piezoelectric actuator of the system. After presenting the real LHC radioactive environment, three tests are designed and performed in sections 3.2, 3.3 and 3.4, where the piezoactuators are exposed to radioactive doses similar to the ones in the LHC. The purpose of these tests is to validate the use of this actuator in this specific application and identify potential problems due to radiation effects during operation. The conclusions of each test will be presented at the end of each corresponding section, and general conclusions will be summarised in the section 3.5.

Figure 3-1: LHC Layout showing the eight octants and IRs. Beam 1 and Beam 2 circulate clockwise and anticlockwise respectively.

3.1.1 Overview of the LHC layout

In order to understand the range of radiation levels and nature at the different points of the LHC it is important to consider the basic layout of the accelerator [38].

The 26.7 km circumference of the LHC can be divided into eight arcs and eight straight sectors, denominated interaction regions (IRs), of 538m long each, as seen in 3-1. The collision points and detectors for physics are located in four of these IRs (IR1, IR2, IR5 and IR8), whereas the rest are assigned for machine services as beam cleaning in the collimation insertions (IR3 and IR7), radio frequency cavities (IR4) or beam dumping (IR6).

The arcs and straight sectors are grouped into eight octants of the ring, each of them subdivided in 34 half-cells in left and right directions. Each octant is centered

Figure 3-2: LHC Octant subdivision.

in the straight section and includes previous and subsequent half arcs.

The straight sections are named Long Straight Sections (LSS) and cover the half-cells 1-7 of both directions.

The Dispersion Suppressors, two per octant and placed in the half-cells 8-11 of each direction, are in charge of matching the insertion optics of the LSS to the periodic part of the arc, which occupies the half-cells 12-34 of both directions of the octant as shown in Figure 3-2.

Each subsection can be further subdivided and a more detailed description of the dipole and quadrupole magnets involved in each region can be seen in detail in [38].

Due to the large radiation levels around the LHC ring, there exist different shielded alcoves, distributed around the LHC tunnel, designed to host the majority of the accelerator equipment in the vicinity of the highly radioactive LHC regions.

3.1.2 Radiation environment at LHC

During normal LHC operation, a large variety of beam losses are produced generating a mixed and complex radiation environment. The origin and nature of this radiation differs significantly from point to point of the accelerator [3].

For example, in the vicinity of the interaction points (IR1 and IR5) the main source of losses is related to the particle collision debris (beam-to-beam interaction, usually proton-proton inelastic interactions which generate secondary particles). In

addition, the halo particles are caught in the collimators close to the experiments and generate hadronic and electromagnetic showers that contribute to the radiation production. This particular effect is dominant in the beam cleaning insertions (IR3 and IR7), whereas in the arcs the predominant effect is the production of losses due to beam-gas interaction [98].

3.1.3 Radiation effects on electronics

The effect of a mixed radiation field on electronics can be generally categorised into cumulative effects and stochastic single-event effects [10].

The continuous exposure to ionising particles produce the drift of electric parameters due to degradation of the different components of the electronic element or system, such as the insulators or semiconducting materials. This cumulative dose effect can be subdivided into two main damage mechanisms known as total ionising dose (TID) and displacement damage (DD).

On the other hand, the stochastic interaction of a single high energy ionising particle with a sensitive region of an electronic device can lead to damage in the device. In this case, the damage mechanism is described as a single-event-effect (SEE).

3.1.3.1 Total Ionising Dose Effect (TID)

The total ionising dose (TID) corresponds to the energy deposited in the irradiated material through ionising processes (addition or removal of electrons or other charged particles from atoms).

This energy deposition creates electron-hole pairs in the material and can cause long term effects in the electrical characteristics of the electronic device. Depending on the electrical properties of the material, part of the electron-hole can be recombined or the charges with high mobility (e-) can move toward the anode if a potential is applied. The electron-hole creation can lead to the formation of trapped charges in defect precursor sets or to the formation of interface traps when it reacts with the

hydrogen in the SiO₂.

The effects of TID are particularly important in MOS devices or bipolar components [10]. In the case of MOSFET Transistors, this effect can be used as principle of operation for TID monitoring sensors [24].

3.1.3.2 Displacement Damage (DD)

A different type of cumulative damage, non-ionising in this case, occurs when a high energy particle (protons, neutrons, heavy ions, electrons or, indirectly, photons) produce a displacement of an atom from the lattice of the irradiated material.

The Coulomb interactions and nuclear reactions of the energetic particle with the nucleus of the target material can produce this atomic displacement that creates a vacancy in the original lattice and a subsequent defect in a non-lattice position (Frenkel pair). The evolution of the defects in the material, generally silicon, has been studied in recent years [135] showing the possibility of recombination or evolution to more stable defects. These defects can be arranged differently depending on the material, forming clusters of defects or remaining isolated, but always leading to the degradation of the device.

The study of this mechanism has been particularly important for Charged-Coupled devices (CCD), where DD produces radiation-induced currents and degradation of charge transfer efficiency [113].

3.1.3.3 Single Event Effects (SEE)

When a single high-energy particle strikes a sensitive part (generally the semiconductor material) of an electronic device it can produce an observable change in its state or performance. This type of single event effect (SEE) can occur stochastically at any time, in contrast to the TID and DD effects that are cumulative.

The consequences of the SEE can be divided into “soft” or “hard” depending whether the effect correspond to a loss of information or irreversible physical damage. There exists a wide range of SEE of each type such as upsets (SEU) or transients (SET) for the soft effects, or latch-up (SEL), burnout (SEB) or gate rupture (SEGR)

for hard effects [10].

3.1.4 Radiation effects on piezoelectric materials

Despite the large number of applications where the piezoelectric materials are used in industry, there are few cases where piezoactuators have to withstand harsh environmental conditions similar to particle accelerators (in terms of UHV compatibility and mixed field radiation).

For this reason, the literature about the radiation effects on piezoelectric materials is limited and the use of the piezoelectric material in these studies does not always coincide with the use in this project.

An example of a piezoelectric material application can be found in the ultrasonic transducers employed in nuclear plants, where substantial radiation fields are presents. The literature survey presented in [134] summarises the different radiation endurance tests performed in the recent years on this type of transducer under cumulative gamma dose and neutron fluence.

For the particular case of PZT materials, the results for gamma radiation in [8] shows reliable operation after doses of 1.5 MGy and [75] [76] presents ionization damage threshold of 400 MGy approximately with low temperature and neutron fluence. In the case of neutron radiation, the results in [105] can be remarked with minor degradation in thin films up to $1.5 \times 10^{15} n/cm^2$ or the drop in response amplitude presented in [8]. A fully detailed summary of the test and conclusions on ultrasonic transducers can be found in [134].

Another application of piezoactuators is when operating in an oscillating mode for adjusting and controlling the fundamental resonant mode in the superconducting radio frequency (SFR) cavities of various projects. Initially in [49] and later in [48], piezoactuators from three different manufacturers were irradiated with fast neutrons at liquid helium temperature at CERI cyclotron. The results of the test show no major damage in the piezo samples up to $7 \times 10^{14} n/cm^2$ but slight performance degradation in terms of loss factor increase (5 – 10%).

A more recent radiation test for the same application was presented in [114] with

accumulated dose levels up to 5 MGy with gamma source Co60. The irradiated samples presented a stroke reduction of less than 10% which was considered acceptable for the requirements of that application.

In the field of non-volatile memories, piezoelectric materials (mainly PZT) are used as ferroelectric memories (FeRAMs). The results about ionising radiation response are summarised in [30]. Different studies were performed as in [14], showing good resistance up to 0.05MGy, with minor degradation in the hysteresis loop above this accumulated dose up to 1MGy. The shape of the hysteresis loop was recovered after repeated bias cycling.

In [87], the authors study the irradiation effects on an acoustic sensor using a piezoelectric transducer. Two sensors were irradiated with a neutron fluence of $3.16e19n/cm^2$ and gamma dose up to 0.25MGy. After the irradiation, the optical microscopy, SEM analysis and X-ray diffraction did not identify any radiation induced defect in the piezoelectric material. However a sensor signal degradation was present, which was associated by the authors to changes in the gas interface of the sensor and not directly to the piezoelectric transducer.

In the field of acceleration and shock sensors, the authors in [21] investigated the total dose effects of X-ray and gamma irradiation. The observed ionisation current versus ionizing dose rate presented non-linear dependence and was in good agreement with the proposed simplified model.

Other studies focused on the radiation effects on the bulk material, as in [64], [63], [143], [143] or [9], where reliability of PZT piezoelectric cells was proven for gamma irradiation up to 1.5 MGy and thermal neutron irradiation up $1.6e21n/cm^2$, with variations in the resonance frequency generally smaller than 1%. Another example is [110], where the piezoelectric properties (d_{33} , electrostrictive strain and electromechanical coupling coefficient) in barium stannate titanate piezo ceramic were reduced after exposition to 200Gy of gamma radiation with Co60 source.

Similarly, in [101] the authors study the effects on soft and hard PZT materials of neutron irradiation up to $10e18n/cm^2$. These results show how the electromechanical coupling factor, the dielectric permittivity and the loss factor decreased by 30%,

70% and 50% respectively, with defects in the material produced, mainly but not exclusively, by the oxygen vacancies creation.

Various summary tables can be found in [134], where the results for other piezoelectric materials apart from PZT are also reviewed.

3.1.5 Proposed Irradiation Tests

As was summarised in 3.1.4, most of the literature on this topic is the result of specific tests on different sensors and piezoelectric materials under unique operating conditions, that correspond to each specific application. As a result, the scope of the conclusions of the tests is limited to a specific transducer assembly (with a piezo material composition, insulator composition, internal structure, soldering material, etc.) and operating conditions, which makes difficult the comparison between tests or the extrapolation of results to different transducers.

Therefore, the previous literature can provide an overall picture of what to expect in terms of radiation effects on piezoelectric materials, but the actual degradation, or even complete failure, of a specific actuator assembly has to be individually studied under the conditions foreseen during its operation.

For this reason, three different tests are proposed for the piezoelectric actuators used in the Crystal Goniometers: a Gamma irradiation test up to 10 MGy (offline), a Gamma irradiation test up to 1 MGy (online) and a Proton Irradiation test up to 2 MGy (online). The details and conclusions of each test can be found in sections 3.2, 3.3 and 3.4 respectively. The objective of the gamma irradiation tests is to evaluate the TID effects on the actuator, whereas the proton irradiation test will allow the analysis of potential DD damage. Although no SEE damage is initially expected due to the lack of particularly sensitive zones like silicon layers, the SEE possibility cannot be completely discarded since a failure in any insulating layer can lead to the total failure of the device for example, and therefore SEE could be detected during the three proposed tests.

It is important to remember that crystal goniometers in IR7 of LHC have an estimated total accumulated dose of 10 MGy mixed field during their desired lifetime

of 10 years. The accumulated dose of the tests was selected as a trade-off between the expected total dose of the device in operation, the exposure time and cost of the test. In the case of proton irradiation the activation of the samples after the test was also considered.

3.2 Gamma irradiation tests - Offline - 10MGy Accumulated Dose

3.2.1 Description

This offline test consists of the irradiation of 24 piezo stack actuators up to 10 MGy accumulated dose at BGS Facility of Fraunhofer Institute, Germany, with gamma irradiation source Co-60L. The term “offline” of the test makes reference to the lack of measurements during the irradiation period, so all the characterisation of the devices under test (DUT) will be performed before and after the irradiation period.

The DUT are 26 piezo stack actuators (24 irradiated and 2 control samples) belonging to two different batches labelled as Type A (13 samples) and Type B (13 samples). The main difference between both batches is the use of glue between the piezo actuator blocks in the Type B, whereas in the Type A the blocks are held only by the electrical bus wires and solder on the side of the stack, without additional glue.

The reason to evaluate two different batches is to understand if the presence of glue between blocks, which facilitates the handling of the stack during the mounting process of the actuator in the rotational stage, would present any additional advantage or problem when exposed to ionising radiation with respect to the actuators without glue.

The DUTs were mounted in mechanical holders to maintain a constant pressure on them during the irradiation as can be seen in Figure 3-4 and 3-6.

The specifications of the DUT are:

- Type A (Figures 3-3 and 3-4):

- Manufacturer: Noliac.
 - Model: NAC2013-H30-C09 (NCE51F).
 - Non-glued stack.
 - Piezo stack volume: $5 \times 5 \times 28mm^3$.
 - Kapton wiring (for radiation environments).
 - The sensitive volume is the piezo stack ($5 \times 5 \times 28mm^3$ volume, not including cabling and solder).
 - High melting point solder (Lead (Pb) 93.5%, Tin (Sn) 5%, Silver (Ag) 1.5% - Melting Point $296 - 301^\circ C$).
- Type B (Figures 3-5 and 3-6):
 - Manufacturer: Noliac.
 - Model: NAC2013-H30-C10 (NCE51F).
 - Glued stack with Ceramabond 569.
 - Piezo stack volume: $5 \times 5 \times 30mm^3$.
 - Kapton wiring (for radiation environments).
 - The sensitive volume is the piezo stack ($5 \times 5 \times 28mm^3$ volume, not including cabling and solder).
 - High melting point solder (Lead (Pb) 93.5%, Tin (Sn) 5%, Silver (Ag) 1.5% - Melting Point $296 - 301^\circ C$).

This offline test was performed by Fraunhofer Institute of Technological Trend Analysis (INT) in March-June 2017 (113 days in total). The irradiation took place at Beta-Gamma-Service (BGS) in Whiehl, Germany, with a Co-60 gamma radiation source. The irradiation conditions are summarised in Table 3.1.

3.2.2 Measurements and setup

A full characterization of every DUT was performed at CERN before and after the irradiation test. This characterisation included the following tests (equipment specified in brackets):

Figure 3-3: Type-A Piezo stack actuator.

Figure 3-4: Type-A Piezo stack actuator mounted in the support.

Figure 3-5: Type-B Piezo stack actuator.

Figure 3-6: Type-B Piezo stack actuator mounted in the support.

Number of samples	26
Irradiated samples	24
Control samples	2
Radiation source	Co-60
Max. Accumulate dose	10 MGy (air)
Dose steps	1, 4, 10 MGy (air)
Dose rate	7.7 to 9.0 kGy(air)/h
Start of irradiation	02-03-2017
Stop of irradiation	22-06-2017
Irradiation temperature	$< 50^{\circ}C$

Table 3.1: Gamma irradiation test at Fraunhofer Institute - Offline Test - Irradiation conditions.

1. Leakage current at 5V and 150V (Keithley 2410 Source Measurement Unit (SMU)).
2. Impedance analysis (Cypher Instruments C60 impedance analyser).
3. Stroke test, whose details can be found in A.1.1 (Piezoactuator Characterisation Testbench at CERN A.1).

During this characterisation, the temperature of the laboratory was maintained at $25^{\circ}C$ ($\pm 1^{\circ}C$).

The irradiation test was stopped at the TID steps of 0, 1, 4 and 10 MGy. In each stop, the following measurements were performed in each DUT at the irradiation facility (equipment specified in brackets):

1. Picture.
2. Temperature of the piezo (Fraunhofer temperature sensor).
3. Leakage current Test (Keithley 2410 SMU).
4. Impedance analysis test (Cypher Instruments C60 impedance analyser).

After each irradiation stop and corresponding measurements, four samples of each batch were sent back to CERN for their final characterisation. The accumulated dose of each DUT is presented in Table 3.2.

Offline Test NEO 14-025 II		
	Type A	Type B
Accumulated dose (TID) [MGy]	Piezo stack ID	Piezo stack ID
1	A1	B1
	A2	B2
	A3	B3
	A4	B5
4	A5	B6
	A6	B7
	A9	B9
	A10	B10
10	A11	B11
	A12	B12
	A16 (15-074)	B16 (15-074)
	A17 (15-074)	B17 (15-074)
0 (Control samples)	A13	B13

Table 3.2: Gamma irradiation test at Fraunhofer Institute - Offline Test - DUT accumulated dose distribution.

The setup of the test at the irradiation facility was very simple since it was a offline test where no measurements were necessary during the irradiation period. The DUTs were fixed with aluminium tape onto two trays as shown in Figure 3-7.

These trays were placed in one aluminium basket, shown in Figure 3-8, that was grouped with other samples onto the palette shown in Figure 3-9. Care was taken to leave the DUTs exposed to fresh air to prevent the generation of ozone. In the final application, the actuators will operate under UHV conditions and therefore no risk of ozone accumulation is expected. The final palette was mounted on a conveyor belt that goes around the radioactive source until the desired accumulated dose is reached.

3.2.3 Summary results

The results of the measurements performed at each stop are grouped according to the test type:

Figure 3-7: DUTs mounted onto the trays for irradiation.

Figure 3-8: Ventilated box with DUT trays.

Figure 3-9: DUTs grouped with other samples in a palette.

3.2.3.1 Leakage current test at the different irradiation stops

The leakage current was measured on each DUT with the Source Meter Unit Keithley 2410 while mounted in its holder at each test stop (0, 1, 4 and 10 MGy). For this reason the voltage applied was limited to 5V in order to avoid mechanical damage on the actuator whose expansion was constrained.

The leakage current results for the Type A and Type B actuators are presented in Figure 3-10 and 3-11 respectively. It is important to remember that A13 and B13 are the control samples that were not exposed to radiation. The results show a very low leakage current ($<0.1 \text{ uA}$) in all samples independently of their accumulated dose, with the only exceptions of B16 at 4MGy with (0.2uA) and A17 at 10MGy that reached the compliance limit of 1mA of the measurement equipment.

3.2.3.2 Impedance analysis test

Similarly, the impedance of each DUT was measured at every irradiation stop (0, 1, 4 and 10 MGy) with a Cypher Impedance Analyser that measures the impedance

Figure 3-10: Leakage current results for Type-A Piezo stack actuator at 5V. A13 is the control sample.

Figure 3-11: Leakage current results for Type-B Piezo stack actuator at 5V. B13 is the control sample.

response from 1kHz to 1MHz.

The impedance module results for Batch A are presented for the frequencies of 1kHz and 2kHz in the figures 3-12 and 3-13 respectively. With similar results, the values for the Batch B at 1kHz and 2kHz are presented in the figures 3-14 and 3-15 respectively.

3.2.3.3 Leakage current test without mechanical support

After irradiation period of each DUT, the sample was dismounted from its support and the leakage current was measured at 5V and 150V since there was no mechanical constraint in this case. The results for the Batch A and B are presented in 3-16 and 3-17 respectively. The maximum leakage currents observed were $< 1\mu A$ (with 5V) and $< 12\mu A$ (with 150V) in all cases, with the exception of A17 that reached the compliance limit of 1mA. These low values confirm that none of the samples were short circuited after the irradiation period. The correct operation of A17 was confirmed in the stroke test despite of its relatively high current.

The measurement of the leakage current is very sensitive to environmental factors and the internal state of the piezoelectric material, producing in some cases the negative values in the leakage currents.

3.2.3.4 Stroke Test

The total stroke of the DUT was measured before and after the irradiation period in the piezoactuator characterisation testbench presented in Appendix A.1. The stroke differences are summarised in Figures 3-18 (Batch A) and 3-19 (Batch B) with the pre-irradiation range of all DUTs corrected according to the average variation of the corresponding control sample. The individual stroke analysis can be found in the Appendix A.3.1.

This test is of key importance for the specific application of this project since any failure or stroke reduction of the actuator stroke would have a direct impact on the final angular range of piezo goniometers, which could degrade the performance of the system or even make it completely unusable.

(a) Impedance at 1kHz.

(b) Impedance change (%) at 1kHz.

Figure 3-12: Batch A - Impedance results at 1kHz. A13 is the control sample.

(a) Impedance at 2kHz.

(b) Impedance change (%) at 2kHz.

Figure 3-13: Batch A - Impedance results at 2kHz. A13 is the control sample.

(a) Impedance at 1kHz.

(b) Impedance change (%) at 1kHz.

Figure 3-14: Batch B - Impedance results at 1kHz. B13 is the control sample.

(a) Impedance at 2kHz.

(b) Impedance change (%) at 2kHz.

Figure 3-15: Batch B - Impedance results at 2kHz. B13 is the control sample.

(a) Leakage current at 5V.

(b) Leakage current at 150V.

Figure 3-16: Batch A - Leakage current test with the sample dismounted from the support. A13 is the control sample

(a) Leakage current at 5V.

(b) Leakage current at 150V.

Figure 3-17: Batch B - Leakage current test with the sample dismounted from the support. B13 is the control sample.

The results presented in Figures 3-18 and 3-19 show how all the DUT were operational after the irradiation period. It is important to remark that the number of samples at each radiation dose is only four and, therefore, the statistical value of conclusions is limited. However, it is clear how all DUTs presented similar strokes after the irradiation period, with a stroke reduction trend that never exceeded the 20% even at 10MGy accumulated dose.

Besides the total stroke, it is also interesting to evaluate the radiation effect on the shape of the hysteresis loop, as is suggested by some literature.

The maximum hysteresis value (maximum difference in the hysteresis loop) was measured and the variation (in %) is presented in Figures 3-20(a) (Batch A) and 3-20(b) (Batch B). At the same time, the voltage at which this maximum difference occurs is shown in Figures 3-20(c) (Batch A) and 3-20(d) (Batch B). Finally, the results for the change in the total area of the hysteresis loop are presented in Figure 3-20(e) and 3-20(f) for Batch A and B respectively.

According to these results, there is no clear trend in terms of maximum hysteresis value nor total area change. However, the increase in the voltage where the maximum hysteresis is achieved with respect to the accumulated dose is noticeable, as can be seen in Figures 3-20(c) and 3-20(d).

3.2.4 Conclusions

The irradiation test was performed on 13 piezo samples of each type (non-glued and glued stacks) with a Co60 gamma source. In addition to the irradiated samples, 1 extra sample of each batch was placed in a radiation-free area for control purposes (in this case, sample number 13)

The samples were extracted from the irradiation zone at three different accumulated doses of 1, 4 and 10 MGy. According to the different tests and results presented in the previous sections, the main conclusions are:

- Leakage current test:
 - Small thermal variations of the piezo stack can produce non-negligible

(a) Change of total stroke.

(b) Change of total stroke vs dose.

(c) Change of total stroke (in %) vs dose.

Figure 3-18: Batch A - Stroke test results comparing before and after the irradiation.

(a) Change of total stroke.

(b) Change of total stroke vs dose.

(c) Change of total stroke (in %) vs dose.

Figure 3-19: Batch B - Stroke test results comparing before and after the irradiation.

(a) Change in the maximum hysteresis value vs dose - Batch A.

(b) Change in the maximum hysteresis value vs dose - Batch B.

(c) Change in voltage for maximum hysteresis vs dose - Batch A.

(d) Change in voltage for maximum hysteresis vs dose - Batch B.

(e) Change of hysteresis area vs dose - Batch A.

(f) Change of hysteresis area vs dose - Batch B.

Figure 3-20: Batch A and B - Hysteresis results comparing before and after the irradiation.

		Batch A	Batch B
1 MGy	Min	+7.4%	+5.3%
	Mean	+8.6%	+7.9%
	Max	+11.7%	+12.4%
4 MGy	Min	+13.1%	+8.8%
	Mean	+14.9%	+12.8%
	Max	+18.4%	+19.1%
10 MGy	Min	+16.1%	+12.4%
	Mean	+19.3%	+18.0%
	Max	+23.1%	+23.5%

Table 3.3: Impedance variation (in %) at 1 kHz.

currents that affect the leakage current measurement repeatability. Despite the efforts to measure the piezo stacks in a thermally stable environment, no clear trend can be extracted from the results.

- All piezo stacks withstood the maximum operating voltage of 150 V after the irradiation, including the sample A17 that showed a high leakage current (reaching the compliance value of 1mA for 5 V and 150 V input).
- Impedance analysis :
 - The impedance modulus clearly increased in the irradiated DUTs with respect to the control ones. The impedance variation at 1 kHz is summarised in Table 3.3 and it can be concluded that there was a maximum impedance increase of <25% at 10MGy.
 - The resonance amplitudes and frequencies might be also affected, but no clear trend was identified with respect to the accumulated dose.
 - This impedance increase is similar in both batches.
- Stroke tests:
 - All samples were fully operative after the irradiation period. However, the results showed reduction of the stroke with the accumulated dose. The

	Batch A	Batch B
Mean value	-12.4%	-12.6%
Max value	-21.2%	-21.3%

Table 3.4: Results of stroke variation for the maximum accumulated dose of 10MGy.

maximum stroke loss was $<22\%$ in all cases and the results at 10MGy are summarised in Table 3.4.

- Apart from the stroke reduction, the results suggest a radiation effect on the shape of the hysteresis loop, especially the first half of the hysteresis loop. The maximum hysteresis percentage did not change significantly, with the average value per dose within ($\pm 3\%$). However, the voltage where this maximum hysteresis percentage is reached experienced a shift towards higher values, with a maximum averaged increase of 12.8 V at 10 MGy in Batch A.

3.3 Gamma Irradiation Test - Online - 1MGy Accumulated Dose

3.3.1 Description

This online test consists of the gamma irradiation of 4 piezo stack actuators (2x Type A and 2x Type B) up to 1 MGy accumulated dose at Fraunhofer Institute for Technological Trend Analysis INT, Germany. The term “online” refers to the monitoring via online measurements performed during the irradiation period, in addition to the characterisation tests before and after the irradiation period.

The objective of the test is the study of the gamma radiation effect on the piezoactuators and, unlike the offline test presented in section 3.2, the online measurements will provide a better understanding of the actuator evolution during the irradiation period.

The DUT are similar to the ones presented in section 3.2, with 8 piezo stack

Number of samples	8
Irradiated samples	4
Control samples	2 + 2
Radiation source	Co-60
Max. Accumulate dose	1210 kGy (air)
Dose steps	One single step
Dose rate	0.4005 Gy(air)/h
Start of irradiation	14-12-2016 9:00
Stop of irradiation	18-01-2017 8:52
Irradiation temperature	$22^{\circ} \pm 1.2^{\circ}C$

Table 3.5: Gamma irradiation test at Fraunhofer Institute - Online Test - Irradiation conditions.

actuators belonging to two different batches A (4 samples) and B (4 samples) that differ in the presence of glue between the piezoelectric blocks in batch B.

The DUTs were mounted in the same mechanical holders as in the offline test as shown in Figure 3-4 and 3-6. Further details of the DUTs can be seen in section 3.2.

The irradiation conditions are summarised in Table 3.5, and the temperature conditions can be found in the Appendix A.4.2.

3.3.2 Measurements and setup

According to the radiation facility options, it was decided to distribute the DUT in three different areas: radiation zone, low radiation zone and no-radiation zone. The DUT distribution between the areas is presented in Table 3.6, where the open and shorted cables used for cable length correction, explained later in this section, are also indicated.

The DUTs were placed in each zone according to the following indications:

1. Radiation zone: The DUTs were arranged equidistantly and aligned around the radiation source, as shown in Figure 3-21. In this distribution, the piezo stacks were irradiated from one of their sides, maximizing the dose rate and homogeneity. It is important to remark that the irradiated piezo stacks were maintained in the open air and not enclosed in a container.

(a) Front view with the four DUTs around the source in the center.

(b) Zoom in one of the DUT holders.

Figure 3-21: DUTs mounted around Co-60 Gamma source in the radiation zone.

Number	Item Description	Temperature sensor
1	Radiation zone - Piezo A1	YES
2	Radiation zone - Piezo A2	YES
3	Radiation zone - Piezo B1	YES
4	Radiation zone - Piezo B2	YES
5	Radiation zone - Open cable	NO
6	Radiation zone - Shorted cable	NO
7	Low-radiation zone - Piezo A3	YES
8	Low-radiation zone - Piezo B3	YES
9	Low-radiation zone - Open cable	NO
10	Low-radiation zone - Shorted cable	NO
11	No-radiation zone - Piezo A4	YES
12	No-radiation zone - Piezo B4	YES
13	No-radiation zone - Open cable	NO
14	No-radiation zone - Shorted cable	NO

Table 3.6: Gamma irradiation test at Fraunhofer Institute - Online Test - DUT distribution in the different areas.

2. Low-radiation zone: The control samples in this area were placed on a table and in open air and not enclosed in a container, as can be seen in Figure 3-22.
3. No-radiation zone: The control samples in this area are shown in figure 3-23. They were under the same temperature conditions as the irradiated piezos, keeping also them in the open air and not enclosed in a container.

The no-radiation zone consisted of a separated climatic chamber following the temperature of radiation zone, but no humidity control was possible. For this reason, the low-radiation zone was included since it was in the same enclosure than the radiation zone.

Similarly to the offline test case, a full characterization of every DUT was performed at CERN before and after the irradiation period. This characterisation included the following tests (equipment specified in brackets):

1. Leakage current at 5V and 150V (Keithley 2410 Source Measurement Unit (SMU)).
2. Impedance analysis (Cypher Instruments C60 impedance analyser).

Figure 3-22: DUTs in the low-radiation zone.

Figure 3-23: DUTs in the no-radiation zone.

3. Stroke test (Piezoactuator Characterisation Tesbench at CERN A.1).

During these measurements, the temperature of the laboratory was maintained at 25°C ($\pm 1^{\circ}\text{C}$).

In addition to the previous characterisation, three different measurements were performed remotely online during the irradiation period on each sample:

- The temperature, by means of a temperature sensor attached to the stack.
- The leakage current test at 5V.
- The impedance, by means of an impedance analyser.

For this purpose, the setup presented in Figure 3-24 was deployed in the irradiation facility. Further details of the equipment involved can be found in the appendix A.4.1

A custom program created by CERN performed all the measurements following the sequence:

1. Selection of the element to measure.
2. **Temperature measurement:** Routing of the analogue multiplexer to the corresponding thermometer and measurement of the temperature.
3. **Impedance analysis:** Configuration of the matrix switch to perform an impedance analysis on the selected piezo stack actuator.
4. **Leakage current measurement:** Configure the matrix switch and analogue multiplexer to measure the leakage current with 5V.
5. Return to step 1.

The data was stored locally and, at specific times, it was backed-up remotely at CERN servers.

Active Test Setup

Figure 3-24: Online test setup with the DUT distribution in the different areas.

3.3.3 Results - Online measurements

3.3.3.1 Temperature

The temperature was monitored in every DUT along the full irradiation period. The results presented in Figure 3-25 show how the temperature was successfully maintained within the temperature specifications detailed in the Appendix A.4.2, except for some minutes during the second climatic problem in the facility (13/01/2017 09:04 - 09:07 - Radiation stopped). The problem was resolved within the specified period for contingencies, and therefore, the general temperature conditions are considered as fulfilled.

3.3.3.2 Impedance Analysis

The use of long cables for the connection of the DUT in the different areas imply that the impedance test results include not only the DUT, but the effect of the cables. For this reason, additional "open" and "shorted" cables were included in each region with the purpose of compensating for the corresponding cable characteristics. The Nominal T-Model shown in Appendix A.4.3 was applied.

The results for 1 kHz and 2kHz are presented in Figures 3-26 and 3-27, for the Batch A and B respectively. It can be concluded that there is an initial step increase in the impedance modulus, which could be related not only to the radiation effect but to the internal temperature increase. After this step, the impedance continues increasing until the end of the irradiation period with maximum values of <12% in all irradiated cases.

3.3.3.3 Leakage current results

The leakage current results are presented in combination with the corresponding temperature in the Figures 3-28 to 3-35. In these figures, it can be observed how the temperature slightly affects the leakage current result. All DUTs present low leakage current in absolute values of <20uA, except for some isolated peaks of < 50 uA. The negative sign in the leakage current values might be produced by the high sensitivity

(a) Temperature sample A1.

(b) Temperature sample B1.

(c) Temperature sample A2.

(d) Temperature sample B2.

(e) Temperature sample A3.

(f) Temperature sample B3.

(g) Temperature sample A4.

(h) Temperature sample B4.

Figure 3-25: Temperature monitoring of the DUT.

(a) Impedance vs time at 1kHz.

(b) Impedance vs time at 2kHz.

Figure 3-26: Impedance vs time for batch A.

(a) Impedance vs time at 1kHz.

(b) Impedance vs time at 2kHz.

Figure 3-27: Impedance vs time for batch B.

Figure 3-28: Leakage current and temperature vs time for A1 - Radiation zone.

of the measurement at such low levels to environmental factors, nearby equipment introducing residual currents or even the non-linear behaviour of the piezoelectric actuators. There is no clear trend of the leakage current with the accumulated dose in any of the samples nor evidence of electrical damage in terms of short-circuited DUTs.

3.3.4 Results - Before and after irradiation test

The same DUT characterisation was performed before and after the irradiation period for comparison and the results are presented below.

3.3.4.1 Leakage current Test

The variation in the leakage current results are summarised in the Figures 3-36 and 3-37 for the batch A, and the Figures 3-38 and 3-39 for batch B. The values at 5V before the test are bigger than after the test, but at such a low current levels it

Figure 3-29: Leakage current and temperature vs time for A2 - Radiation zone.

Figure 3-30: Leakage current and temperature vs time for A3 - Low radiation zone.

Figure 3-31: Leakage current and temperature vs time for A4 - No radiation zone.

Figure 3-32: Leakage current and temperature vs time for B1 - Radiation zone.

Figure 3-33: Leakage current and temperature vs time for B2 - Radiation zone.

Figure 3-34: Leakage current and temperature vs time for B3 - Low radiation zone.

Figure 3-35: Leakage current and temperature vs time for B4 - No radiation zone.

might be due to the high sensitivity of the measurement to environmental factors. For the 150V test, the leakage current remained in all cases under $10\mu A$, which is an indication of the healthy status of the DUTs that was confirmed during the stroke test.

3.3.4.2 Stroke test

The results for the stroke test are presented in Figures 3-40 and 3-41, for batch A and B respectively, with the pre-irradiation range of all DUTs corrected according to the average variation of their corresponding control sample. Unlike the results of the 10MGy offline test where the stroke was slightly reduced, in this occasion the total displacement achieved after the irradiation period increased in both batches, with higher expansion in the batch B reaching the 19.7% in one of the DUTs. In all cases, the irradiated samples present a higher increase with respect to the control samples, which suggest that the stroke increase is related to the radiation effect on

Figure 3-36: Leakage current at 5V - Batch A.

Figure 3-37: Leakage current at 150V - Batch A.

Figure 3-38: Leakage current at 5V - Batch B.

Figure 3-39: Leakage current at 150V - Batch B.

the piezoactuator and not due to other external factors.

The shape of the hysteresis loop before and after the irradiation period is presented in Figure 3-42 for A1 and in the Appendix A.4.4 for the rest of DUTs. In the top-right subfigure, the hysteresis loop is normalised in order to compare the shape of the loop, and the path difference (comparing the increasing and decreasing directions) is presented in the bottom-right subfigure. The hysteresis percentage (with respect to the total displacement) is presented in the bottom-left subfigure, where the maximum values are indicated. According to these results, it can be concluded that the radiation has a clear effect in the shape of the hysteresis loop, with an increase of the voltage where the maximum hysteresis occurs. On the contrary, the control samples presented nearly identical hysteresis shapes, as expected, before and after the irradiation period.

The maximum hysteresis value (maximum difference in the hysteresis loop) is summarised in Figures 3-43(a) (Batch A) and 3-43(b) (Batch B). The voltage at which this maximum difference occurs is shown in Figures 3-43(c) (Batch A) and 3-43(d) (Batch B). Finally, the results for the change in the total area of the hysteresis loop are presented in Figure 3-43(e) and 3-43(f) for Batch A and B respectively.

3.3.5 Conclusions

The irradiation test was performed on 2 piezo samples of each type (non-glued and glued stacks) with a Co60 gamma source. In addition to the irradiated samples, the control samples of each batch were placed in a low-radiation area and radiation-free area (4 control samples in total).

The radiation period took place in the following dates:

- 12/12/2016 18:26 - Acquisition system start.
- 14/12/2016 09:00 - Radiation Start.
- 05/01/2017 16:45 - 16:52 - Radiation stopped due to climatic system problem.
- 13/01/2017 09:04 - 09:07 - Radiation stopped due to climatic system problem.
- 18/01/2017 08:52 - Radiation Stop.

(a) Change of total stroke.

(b) Change of total stroke vs dose.

(c) Change of total stroke (in %) vs dose.

Figure 3-40: Batch A - Stroke test results comparing before and after the irradiation.

(a) Change of total stroke.

(b) Change of total stroke vs dose.

(c) Change of total stroke (in %) vs dose.

Figure 3-41: Batch B - Stroke test results comparing before and after the irradiation.

Figure 3-42: Hysteresis variation for A1 - Radiation zone.

(a) Change in the maximum hysteresis value vs dose - Batch A.

(b) Change in the maximum hysteresis value vs dose - Batch B.

(c) Change in voltage for maximum hysteresis vs dose - Batch A.

(d) Change in voltage for maximum hysteresis vs dose - Batch B.

(e) Change of hysteresis area vs dose - Batch A.

(f) Change of hysteresis area vs dose - Batch B.

Figure 3-43: Batch A and B - Hysteresis results comparing before and after the irradiation.

- 19/01/2017 08:21 - Acquisition system stop.

Apart from the initial and final characterisation of the DUT, the samples were measured remotely (temperature, leakage current and impedance) during the irradiation period. The conclusions of the online results presented in section 3.3.3 can be summarised in the following points:

- Thermal variation along the test:
 - The temperature was successfully maintained within the limits established for each zone according to the specifications in appendix A.4.2. The temperature increase in the DUTs due to the irradiation was negligible ($<0.5^{\circ}\text{C}$).
- Impedance test
 - The samples in the radiation zone show a clear general increase (10% approx.) in the whole impedance response with respect to the control samples in the low and no-radiation zone.
 - The increase in the impedance due to radiation is homogeneous for all frequencies, maintaining the resonance frequencies nearly unaltered.
 - The impedance evaluations at 1 kHz and 2 kHz clearly show how the impedance increases only in the samples in the radiation zone, and not in the low or no-radiation zones. At the end of the test (1.2 MGy accumulated dose approx.), the impedance increase reached 10% of the initial value (similar at 1 and 2 kHz).
 - Right after the radiation stopped at the end of the test, the impedance dropped by 1-2% respect to the final value, but it did not reach the initial level, and this effect could be associated with the internal temperature increase due to irradiation.
 - The results does not show a clear difference between non-glued (A) and glued (B) batches.

- The impedance values seems to be slightly affected by the temperature, but the variation is very small compared to the radiation effect.
- Leakage current test
 - Along the the whole radiation test the mean leakage currents were always $< 20 \mu\text{A}$, except for some isolated peaks of $< 50 \mu\text{A}$ in the samples placed in the radiation zone. These values suggest a normal behaviour of all samples (irradiated and non-irradiated) after the test.
 - There is no trend nor clear radiation effect in the leakage current results of the DUTs.
 - According to the radiation zones it can be concluded that:
 - * DUTs in the Radiation Zone (A1, A2, B1, B2): the acquisitions show some higher and isolated peaks ($< 50 \mu\text{A}$), that are independent of the temperature and the radiation dose. The cause of these peaks is unknown, but could be due to interference with additional equipment near the measurement setup.
 - * DUTs in the Low Radiation Zone (A3, B3): The higher temperature variation clearly affects the leakage current measurement, however, the obtained values remain very low, indicating a normal behaviour or the samples.
 - * DUTs in the No Radiation Zone (A4, B4): The thermal fluctuations are also seen in the leakage current values, however the values remain in the normal range.

After the dismounting of the DUTs from their supports, the samples were characterised and the results compared with the ones obtained before the irradiation.

The leakage current results are summarised in tables 3.7 and 3.8 for 5V and 150V respectively, showing low leakage current value that suggest the correct operation of the actuators. It should be noted that they are much lower than during the irradiation when the actuator was mounted in its support.

DUT Type	1 - Radiation zone	2 - Radiation zone	3 - Low radiation zone	4 - Radiation free zone
Batch A	$-0.07\mu A$	$-0.034\mu A$	$-0.065\mu A$	$-0.047\mu A$
Batch B	$-0.045\mu A$	$-0.06\mu A$	$-0.069\mu A$	$-0.064\mu A$

Table 3.7: Leakage current variation (at 5V) comparing before and after the irradiation test.

DUT Type	1 - Radiation zone	2 - Radiation zone	3 - Low radiation zone	4 - Radiation free zone
Batch A	$+3.83\mu A$	$+1.73\mu A$	$+0.2\mu A$	$+0.07\mu A$
Batch B	$+6.58\mu A$	$+0.08\mu A$	$-0.02\mu A$	$+0.02\mu A$

Table 3.8: Leakage current variation (at 150V) comparing before and after the irradiation test

DUT Type	1 - Radiation zone	2 - Radiation zone	3 - Low radiation zone	4 - Radiation free zone
Batch A	$+3.5\%$	$+8.8\%$	$+1.7\%$	$+0\%$
Batch B	$+18.0\%$	$+19.7\%$	$+1.8\%$	$+0\%$

Table 3.9: Stroke variation comparing before and after the irradiation test

- Stroke test:
 - The stroke variation is summarised in Table 3.9. It confirms the resistance of the DUTs to the 1MGy gamma accumulated dose, showing even an increase in the stroke of the irradiated DUTs, with a maximum of $+19.7\%$ in one of the samples.
 - The analysis of hysteresis loops suggest a slight increase in the maximum hysteresis percentage (with a maximum increase of $+3.4\%$ obtained in B2) as well as in the voltage where this maximum hysteresis occurs (with a maximum increase of $+2.1V$ also in B2).

3.4 Proton irradiation tests - Online - 2 MGy Accumulated Dose

3.4.1 Description

The previous Gamma Irradiation Tests achieved large Total Ionizing Dose (TID) in a relative short period of time. Besides that, the gamma radiation allows an easy handling of the samples after the test for characterisation, since the DUTs are not activated.

However, these actuators will work under the CERN accelerator environment, where not only gamma but a mixed field radiation is present. For this reason, although the Gamma Irradiation Tests can provide valuable information about the TID effects of the actuators, it is necessary to evaluate the potential effects of proton irradiation that can lead to Displacement Damage (DD) on the PZT material.

Before the installation of the first piezgoniometers at CERN in 2012, a test with proton irradiation was performed in PIF facility at PSI up to 150kGy, but this low dose was not sufficient to correctly assess any potential DD. (PIF beam energy ranges from 74 MeV to 230 MeV).

For this reason a new irradiation test, up to 2 MGy accumulated dose, was performed at the IRRAD Proton Facility at CERN. The purpose of the test is to evaluate potential displacement damage on the actuators under conditions similar to the accelerator environment.

The 2 MGy accumulated dose was chosen as a trade-off between desired dose of 10 MGy and irradiation time constraints and post-irradiation activation limits.

The devices under test are 8 piezo stack actuators of Type A, similar to the ones used in the previous irradiation tests.

The Figure 3-44 shows an example of the DUT mounted in a custom-designed support, shown in detail in Figure 3-45, that holds and positions the actuator in the beam axis. The blocks holding the DUT are made of polycarbonate (PC) material in order to avoid any metallic part in the beam path, which would be highly activated

Figure 3-44: Example of DUT mounted in its support.

after the test and would cause additional beam distortion downstream.

The IRRAD proton beam has a section of $12\text{mm} \times 12\text{mm}$ (FWHM) approx. and therefore the DUTs are positioned with a 75.17° angle with respect to the beam axis as shown in Figure 3-46. In this orientation, the dose is spread along the whole DUT volume and the effective “thickness” seen by the protons crossing the material will be minimised.

The samples exposed to the beam are placed in a single line with all the supports mounted on a base table, shown in Figure 3-47, that allows the preparation of the DUTs before the positioning in the beam area.

Unlike the gamma irradiation, which is generally produced by an isotropic source that irradiates continuously (like the case of Co60 source), the proton irradiation is generated by protons that have been previously accelerated forming a beam. In the case of the IRRAD Facility, this acceleration is achieved at the Proton-Synchrotron (PS) accelerator at CERN achieving a momentum of $24\text{ GeV}/c$. The primary proton beam at the PS is extracted from its ring and driven to IRRAD facility (located on the T8 beam-line at the CERN PS East Hall) for its use.

The proton beam arrives to the IRRAD facility in spills, whose frequency depends

Figure 3-45: Custom-designed support to hold and position the DUT.

Figure 3-46: Rotation of the DUT by 75.17° to fit in the beam size.

Figure 3-47: Table with DUTs aligned for their positioning in the beam area.

on the PS operation. For the weeks of operation when this test was scheduled, 3 spills of protons (of 5×10^{11} p/spill and 400ms duration) over a CPS of usually 34 basic periods (BP) were foreseen. A BP of the PS is 1.2 seconds then gets 3 spills every 41 seconds approximately and therefore, it is reasonable to assume 1 spill of protons every 14 seconds considering that it is highly improbable to receive all of them one-after-another.

The beam parameters used in our test can be summarised as:

- Momentum of 24 GeV/c.
- Beam spot of $\sim 12mm$ diameter with Gaussian profile (with optics FWHM and beam standard deviation of $\sigma = \sim 5mm$).
- Protons delivered in spills of 5×10^{11} p/spill at a rate of 14 seconds/spill.

The test started the 25th October 2018 and finished the 5th November 2018 (11 days duration) reaching an accumulated dose of 2 MGy and 1.14×10^{16} p/cm² fluence.

After the irradiation period, the DUTs were stored for a cool-down period of 3 months until their radioactivity was safe for the post-irradiation characterisation.

3.4.2 Measurements and setup

The measurements performed on the DUT during the different stages of the test can be divided into three groups according to the moment when the measurement takes place:

3.4.2.1 Measurements before the irradiation test:

A full characterization of the samples was performed in the laboratory before mounting them in their support for the radiation test. This characterisation includes the following tests:

1. Leakage current at 5V and 150V (Keithley 2410 Source Measurement Unit (SMU)).
2. Impedance analysis (Cypher Instruments C60 impedance analyser).

Figure 3-48: IRRAD Online test setup.

3. Stroke test (Piezoactuator Characterisation Tesbench at CERN A.1).

During this characterisation, the temperature of the laboratory was maintained at 25°C ($\pm 1^{\circ}\text{C}$).

3.4.2.2 Measurements during the irradiation test (online test):

Figure 3-48 presents the test setup diagram for the irradiation test. It shows the electronic equipment involved in the measurements and the distribution of the DUT in the different zones. It is important to note that every sample includes a temperature sensor (PT100) and extra shorted and opened ends in the radiation zone for the cable length compensation.

A custom program similar to that used in the gamma online test runs the measurements continuously and saves the data locally and remotely. The variables to be measured in each sample are the temperature, impedance and leakage current, according to the following sequence:

1. Selection of the element to measure.
2. Temperature measurement.
3. Impedance analysis.
4. Leakage current measurement.
5. Return to step 1.

3.4.2.3 Measurements after the irradiation test:

After the cool-down period, all DUTs were characterised with the same tests as at the initial characterisation in section 3.4.2.1.

The testbench was transported to the IRRAD buffer area and the DUT characterisation was performed using lead-glass for personal radiation-protection.

After the final characterisation, the DUTs were considered as waste in accordance with CERN internal regulation EDMS1364231.

3.4.3 Results - Online

3.4.3.1 Temperature

The temperature measured in each DUT remained always within $22.5^{\circ} \pm 2.5^{\circ}$ C, with a maximum temperature increase in the irradiated samples of 4° C as shown in Figure 3-49.

3.4.3.2 Leakage current

The leakage current tests showed normal values during the whole irradiation period, with currents well below 1 mA as can be seen in the appendix A.5.1.

3.4.3.3 Impedance analysis

For case of the impedance analysis test, Fig 3-50 presents the 1kHz and 2kHz response evolution during the irradiation period. The impedance increased only on the irradiated samples, remaining under 25% in all cases for the final 2 MGy accumulated dose.

(a) Temperature evolution along the irradiation test. The temperature sensor in A9 stopped working around the 2nd of November.

(b) Maximum temperature difference.

Figure 3-49: Temperature results.

(a) Impedance at 1kHz.

(b) Impedance at 2kHz.

Figure 3-50: Impedance response evolution along the irradiation period.

3.4.4 Results - Before and after irradiation test

After a cool-down period of three months, the activation level of the irradiated DUTs was $350 \mu\text{Sv/h}$ at contact. The DUTs were then dismounted from their holders and characterised.

3.4.4.1 Leakage current

The results for the leakage current test at 5V and 150V are presented in Figure 3-51. The low current obtained in all cases suggest the correct operation of the DUTs.

3.4.4.2 Stroke test

The stroke variation results are presented in Figure 3-52. All DUTs showed a correct operation and the irradiated samples experienced a stroke reduction of less than 22%, with the pre-irradiation range of all DUTs corrected according to the average variation of the control samples.

The study of the hysteresis response variation, illustrated in Fig 3-53, is summarised in Figure 3-54. The most evident effect is the increase of the voltage value at which the maximum hysteresis occurs as shown in Figure 3-54(c) and 3-54(d). The stroke results for each individual DUT can be found in Appendix A.5.2.

3.4.5 Conclusions

A Proton Irradiation Test on the piezoactuators used for Crystal Collimation at CERN was performed. The test started the 25th October 2018 and finished the 5th November 2018 (11 days duration) reaching an accumulated dose of 2 MGy and $1.14 \times 10^{16} \text{ p/cm}^2$ fluence. The test was performed with a total of 6 DUTs being irradiated and 2 control samples.

The purpose of the test was to evaluate the possible displacement damage caused by the mixed field ionising irradiation present in particle accelerators. The final objective was to understand its effect on the performance of the actuator and assess the consequences for the positioning system.

(a) Leakage current at 5V.

(b) Leakage current at 150V.

Figure 3-51: Leakage current before and after the irradiation test. A5 and A10 are the control samples.

(a) Change of total stroke.

(b) Change of total stroke vs dose.

(c) Change of total stroke (in %) vs dose.

Figure 3-52: Stroke test results comparing before and after the irradiation.

Figure 3-53: Example of hysteresis shape variation (DUT A9).

According to the various tests and results, the main conclusions are:

- Thermal variation along the test:
 - The temperature in the DUTs increased during the irradiation period showing a maximum temperature variation of $<4^{\circ}\text{C}$.
- Impedance test:
 - The samples in the radiation zone show a clear general increase in the whole impedance response with respect to the control samples in the no-radiation zone.
 - The increase in the impedance due to radiation is homogeneous for all frequencies, maintaining the resonance frequencies nearly unaltered.
 - The impedance evaluations at 1 kHz and 2 kHz clearly shows how the impedance increases only in the samples in the irradiated samples. At the end of the test (2 MGy accumulated dose), the observed impedance increase was $<25\%$ of the initial value (similar at 1 and 2 kHz).
 - Right after the radiation stopped at the end of the test, the impedance remained constant.

(a) Change in the maximum hysteresis value.

(b) Change in the maximum hysteresis value vs dose.

(c) Change in voltage for maximum hysteresis.

(d) Change in voltage for maximum hysteresis vs dose.

(e) Change of hysteresis area.

(f) Change of hysteresis area vs dose

Figure 3-54: Hysteresis results comparing before and after the irradiation.

- Leakage current test:
 - All DUTs presented normal leakage current values, with currents well below 1mA.
- Stroke test:
 - All DUTs operated normally after the irradiation period. The results showed a decrease in the final stroke of $< 22\%$ in all irradiated DUTs (maximum decrease of 21.6% in A9).
 - The most remarkable effect in the hysteresis loops is the increase in the voltage where this maximum hysteresis occurs (with a maximum increase of 12.6V in A2).

3.5 General conclusions

The state of the art regarding the radiation effects on piezoelectric materials was reviewed and summarised at the beginning of the chapter. The different effects and mechanisms are reviewed and the most remarkable results were commented. Despite the many studies performed on these materials, the specific material composition of the DUT, the irradiation specifications and the test conditions are, in most of the cases, very specific to the application under study, and therefore it is difficult to make a direct comparison between studies. For this reason, a custom irradiation test is necessary in order to take into account the specific material and irradiation conditions of the application presented in this thesis.

Three different irradiation tests were proposed and performed on the piezoactuators used in the Crystal goniometers. The key objective was to study the resistance of the actuator to the mixed field radiation present in the LHC tunnel, where they are expected to receive 10 MGy accumulated dose in 10 years of operation, the desired lifetime.

The proposed tests were: Gamma irradiation test up to 10 MGy (Offline Test), Gamma irradiation test up to 1 MGy (Online Test) and Proton Irradiation test up to

2 MGy (Online Test). The purpose of the gamma irradiation tests was the evaluation of TID effects on the actuator, whereas the proton irradiation test will allow the analysis of potential DD damage.

In the 10 MGy Gamma irradiation test (Offline Test) a total of 26 DUT were irradiated with 2 control samples in a radiation-free area. The DUTs were composed of two different batches, A and B, that differ in the presence of glue between the piezo blocks or not. Apart from the initial and final characterisation of the DUTs, there were 3 stops at 1, 4 and 10MGy where the DUTs were measured. Some of the DUTs were removed from the radioactive area in each stop, resulting in eight DUTs at each accumulated dose.

The leakage current test presented similar low currents results before and after the irradiation, with the exception of one DUT reaching the compliance value of 1mA. However, the correct operation of all DUTs was confirmed in the Stroke Test with a maximum stroke reduction of $<22\%$ in both batches for the maximum 10MGy accumulated dose. The results suggest a stroke reduction with the accumulated dose, but further tests should be performed for its confirmation due to the low number of samples involved in the test and the large dispersion of the results. The shape of the hysteresis loop was also studied and the results suggest a maximum hysteresis voltage increase. In addition to these results, the impedance test showed a maximum impedance modulus increase of $<25\%$ when evaluating the DUTs at 1kHz and 2kHz.

The 1-MGy Gamma Irradiation Test (Online Test) consisted of the irradiation of 4 DUTs with Co-60 gamma source and 4 control samples separated in two different areas corresponding to low-radiation and radiation-free zones. As in the previous test, two different batches A and B were involved, differing in the presence of glue between the piezoelectric blocks. Apart from the initial and final characterisation of the DUTs, in this occasion additional measurements were performed remotely (temperature, leakage current and impedance) during the whole irradiation period.

The leakage current results presented normal values during the irradiation period and at the completion of the test. The online impedance measurement confirmed the increase of the impedance module with the accumulated dose, reaching a maximum

increase value of $<12\%$ at 1.2 MGy, which is consistent with the results of the 10-MGy Gamma Irradiation Test. The stroke of the DUTs was measured only before and after the irradiation period and the results confirmed the correct operation of all irradiated DUTs. Unlike in the 1 MGy dose step of the 10-MGy Gamma Irradiation Test, where no clear stroke variation was found, in this occasion the results suggest a slight increase of the DUT stroke after the irradiation period, with a stroke increase of $<20\%$ in all DUTs.

Apart from the gamma irradiation tests, a Proton Irradiation Test (Online Test) was performed the end of October 2018 at IRRAD Facility at CERN. In this case, 6 samples were irradiated with protons up to 2 MGy, with 2 additional control samples placed in a radiation-free zone. The samples were characterised for comparison before and after irradiation test and additional measurements were performed remotely on the DUTs during the irradiation period.

The leakage current results presented normal values along the irradiation period and at the completion of the test. The online impedance analysis agreed with the trend in previous irradiation test, with a clear increase of the impedance module with the accumulated dose, presenting a maximum increase of $<25\%$ at the final 2MGy. The stroke results confirmed the correct operation of all irradiated DUTs and showed a stroke decrease of $<22\%$. The trend direction and magnitude of the stroke variation is similar to the 10MGy Gamma Test despite the lower accumulated dose, which can be explained due to the different nature of the irradiation source.

The Table 3.10 summarises the results for the three tests at their final accumulated dose. The results for the impedance and maximum hysteresis voltage variation are consistent between them, with maximum increases of less than 25% and 19.6V respectively in all cases.

The positive sign of the stroke variation in the Gamma Online Test (1MGy) contrasts with the results for the Gamma Offline Test (10 MGy) and Proton Online Test (2 MGy), where the stroke is reduced. Although a stroke increase is not harmful for this application, this result is clearly unexpected and it could be produced due to untracked changes in the DUT (different humidity with respect to the control

Test	Dose (MGy)	Max. Impedance variation (at 1kHz)	Max. Stroke variation	Max. Hysteresis Voltage variation
Gamma (Offline)	10	<+25% <+25%	-21.2% (-12.4%) -21.3% (-12.6%)	+19.6V (+12.8V) +14.6V (+10.5V)
Gamma (Online)	1	<+12% <+12%	+8.8% (+6.1%) +19.7% (+18.9%)	+1.8V (+1.0V) +2.1V (+1.8V)
Proton (Online)	2	<+25%	-21.6% (-18.8%)	+12.6V (+10.9V)

Table 3.10: Summary of the results of the various test performed in the piezoactuator. The figures in brackets “()” correspond to the averaged values and the lines in grey colour correspond to the results for the Batch B (glued stack).

samples) or mechanical changes in the testbench. Further investigation should be performed to motivate this stroke increase. From the final application point of view, the stroke reduction is the key effect threatening the correct operation of the device and therefore, the worst case scenario is represented by the results of the Gamma Offline Test (10 MGy) and Proton Online Test (2 MGy).

Considering the results of the previous gamma and proton irradiation tests, it can be concluded that the studied piezoelectric actuator is suitable for the Crystal Collimation application in the LHC in terms of radiation resistance. All irradiated samples were fully operative after their corresponding tests and the radiation-induced effects, in terms of impedance increase and stroke variation, are acceptable in this specific application. The closed loop control of the system will compensate for the impedance increase or hysteresis variation, and the consequences of a potential stroke reduction are mitigated by considering this reduction in the actuator design process.

Chapter 4

Angular Measurement System with Absolute Angle Recovery Method

4.1 Introduction

High precision absolute angular measurements are required more and more frequently in research and industry. Different techniques are used for angular measurements including electronic sensors [12, 42, 55, 107, 125], angular potentiometers, encoders [11, 121] or optical methods among others. These later methods include interferometric setups [70], optical autocollimators [35], and optical lever schemes with an electronic target [68]. Interferometer setups have clear advantages such as higher precision, accuracy and larger ranges [44, 97, 141].

The application of laser interferometry for angular measurement consists typically in the splitting of the laser beam into two parallel paths that are reflected in a target and form the interferometry pattern back at the sensor. The angle of the target is then estimated using the beam separation and the optical path difference between the two beams.

Although there is not a standard method yet, several approaches have applied the same principle using different elements in the setup like corner-cube retroreflectors [156] [153], glass plates with flat parallel faces [70] [154], homodyne interferometers using different polarization components of a single beam [111] or cat eye retroreflectors

[112].

At the moment, the use of corner-cube retroreflectors is the preferred choice in most of the commercially available solutions for interferometric angular measurement (Renishaw, Zygo or Keysight). However, in some applications the use of these retroreflectors might not be possible [112] or the most suitable due to high sensitivity to translational misalignments .

The main operational limitation of laser interferometry techniques lies in their incremental principle of operation, providing only relative, rather than absolute, measurements [44]. If the laser beam is interrupted, the maximum speed is exceeded or the system is restarted, the position (or angle) read by the interferometer is lost. In applications where it is critical to maintain an absolute reference over time, other absolute angle measurement systems, like optical autocollimators, are more convenient despite their lower performance with respect to interferometric setups.

The advantages for absolute interferometric measurement is clear and different approaches have obtained important results with techniques including frequency combs [127] [31], frequency tuning [85] or the combination of both techniques [140]. Despite their success, its application in industry is still limited due to its complexity and high cost.

The lack of an absolute reference is a common problem to most rotational positioning stages. The commercially available stages are usually comprised of rotational mechanisms including bearings and worm screws that are moved by stepper motors. In these cases, the absolute reference is achieved with the addition of a home reference switch (optical, hall effect or encoder index), achieving bidirectional repeatability in the range of $\pm(0.7 - 100)\mu rad$. For higher angular performance, the mechanical components of the rotational stage are substituted by flexure-hinges [138], leading to custom designed stages that, in most of the cases, have to meet other strict requirements such as ultra-high vacuum, radiation compatibility or high bake-out temperatures. In these cases, the use of non-contact switches is not possible, the repeatability of mechanical switches is limited and a direct mechanical end-stop can damage the flexure-hinges.

In this chapter, a dual-interferometer-based angular measurement system using a flat reflective surface or corner-cube-retroreflectors is proposed and studied. The optical system is analysed and expressions for the measured distances by both interferometric axes are derived explicitly. The non-zero convexity of the characteristic curve through a rotation is proven to be an intrinsic property of the geometry of the system. It provides a fundamental result for the basis of the new absolute angle recovery method (AARM) [130], which is able to provide an absolute reference without the need of any additional elements, and makes the technique generally applicable.

Furthermore, an uncertainty expression for the recovered angle is derived analytically. Besides that, Monte Carlo simulations are preformed to study the stochastic properties of the AARM, evaluating the effect of noise and the number of points on the performance of the recovery.

In addition to this, the AARM is applied to a high-precision rotational stage providing statistically representative experimental results and showing good agreement between the experimental tests and the Monte Carlo Simulations.

The main advantages of the proposed angular measurement setup with two interferometric axes (Fig. 4-1) include the use of conventional and commercially available linear interferometers, the ease of optical alignment and the possibility to measure linear translations of the target simultaneously with the rotation. The proposed absolute angle recovery method overcomes the operational limitations of interferometric systems and makes this specific setup suitable for applications where an absolute angle measurement is required.

The rest of the chapter is organized as follows: the Dual-Interferometer-Based angular measurement system is proposed and studied in Section 4.2. The absolute angle recovery method is presented in Section 4.3 and is evaluated with Monte Carlo simulations in Section 4.4. The performance of the method is experimentally evaluated in Section 4.5. The application of the measurement system and recovery method into the LHC Piezogoniometers is summarised in Section 4.6, followed by the final conclusions in Section 4.7.

Figure 4-1: Dual-interferometer-based measurement system diagram with reflective surface.

The angular measurement system considered in this work is comprised of two independent interferometer axes as shown in Fig.4-1. Both interferometric axes are mounted parallel to each other on a fixed support and the beams are reflected back from a reflective surface or corner-cube-retroreflectors that are fixed to the rotating part of the system. The origin of the reference system is at the centre of rotation.

The geometry of the system presented in Figure 4-1(a) is studied to obtain an analytical expression for the rotational angle θ as a function of the available interferometer distances d_1 and d_2 .

An analytical expression is derived for the different distances between the mirror and the first interferometric axis, where $\mathbf{r}_{1i}$ and $\hat{\mathbf{v}}_{1i}$ are the position and unitary directional vector of this axis, $\mathbf{r}_{1c}$ is the position of the reflection point of the beam at the mirror, $\hat{\mathbf{v}}_{1c}$ is the unitary directional vector of reflected beam, d_m is the distance between the

reflecting surface and the center of rotation and $\hat{\mathbf{n}}$ and $\mathbf{p}_0$ are the unitary normal vector and an arbitrary point of the mirror respectively.

4.2.1.1.1 Distance from head surface to mirror: It can be seen that the distance of the incident beam from the interferometer axis to the mirror, d_{1i} , can be expressed as

$$d_{1i} = |\lambda_1 \cdot \hat{\mathbf{v}}_{1i}| \quad (4.1)$$

where using the general equation of the plane leads to the following expression for λ_1 :

$$\lambda_1 = \frac{-a_0 - n_x \cdot r_{1i,x} - n_y \cdot r_{1i,y} - n_z \cdot r_{1i,z}}{n_x \cdot v_{1i,x} + n_y \cdot v_{1i,y} + n_z \cdot v_{1i,z}} = \frac{d_m - \hat{\mathbf{n}} \cdot \mathbf{r}_{1i}}{\hat{\mathbf{n}} \cdot \hat{\mathbf{v}}_{1i}} \quad (4.2)$$

where $a_0 = -\hat{\mathbf{n}} \cdot \mathbf{p}_0 = -d_m$ and using the scalar product identity in (4.2) and its substitution in (4.1) leads to

$$d_{1i} = \left| \frac{d_m - |\mathbf{r}_{1i}| \cdot \cos(\angle(\hat{\mathbf{n}}, \mathbf{r}_{1i}))}{\cos(\angle(\hat{\mathbf{n}}, \hat{\mathbf{v}}_{1i}))} \right| \quad (4.3)$$

4.2.1.1.2 Distance from mirror (at beam reflection point) to head plane:

Similarly, the distance of the reflected beam from the mirror to the interferometric axis, d_{1r} , can be calculated as

$$d_{1r} = |\beta_1 \cdot (\hat{\mathbf{v}}_{1i} - 2 \cdot (\hat{\mathbf{v}}_{1i} \cdot \hat{\mathbf{n}}) \cdot \hat{\mathbf{n}})| \quad (4.4)$$

where

$$\beta_1 = \frac{\hat{\mathbf{v}}_{1i} \cdot \mathbf{r}_{1i} - \hat{\mathbf{v}}_{1i} \cdot \mathbf{r}_{1c}}{\hat{\mathbf{v}}_{1i} \cdot \hat{\mathbf{v}}_{1c}} \quad (4.5)$$

and using the scalar product in (4.4) we have:

$$d_{1r} = \left| \frac{|\mathbf{r}_{1i}| \cdot \cos(\angle(\hat{\mathbf{v}}_{1i}, \mathbf{r}_{1i})) - |\mathbf{r}_{1c}| \cdot \cos(\angle(\hat{\mathbf{v}}_{1i}, \mathbf{r}_{1c}))}{\cos(\angle(\hat{\mathbf{v}}_{1i}, \hat{\mathbf{v}}_{1c}))} \right| \quad (4.6)$$

4.2.1.1.3 Total distance: The distance measured by the interferometer axes depends on the internal operation of the real device. The interferometers used in this application are for linear measurement without the possibility to know the real path of the beam light. For this reason, the total distance measured by each axis is assumed to be the average between the incoming and outgoing beam paths:

$$d_1 = \frac{d_{1i} + d_{1r}}{2} \quad (4.7)$$

The equations (4.1) - (4.7) were obtained for the interferometer axis 1, but they are equally valid for the interferometer axis 2 if the subindex number is changed accordingly. In this case, similar expressions for d_{2i} , d_{2r} and d_2 can be obtained.

4.2.1.2 Ideal interferometer axis and mirror alignment with non-centered rotation axis

The equations 4.3, 4.6 and 4.7 for d_1 and d_2 can be combined to obtain the analytical expression presented in (4.8) for the angle $\theta = f(d_1, d_2)$:

$$\theta = \tan^{-1} \left(\frac{C \pm \sqrt{C^2 + 4(d_1 + d_2 - 2d_0)(d_1 + d_2)}}{-2(d_1 + d_2)} \right) \quad (4.8)$$

where $C = (D_1 - D_2)$ and it was assumed that the sensors are at the same distance d_0 , the light beam axes are parallel to each other and perpendicular to the reflective surface when it is at $0\mu rad$ and the reflective surface is at the center of rotation ($d_m = 0$).

4.2.1.3 Ideal interferometer axis and mirror alignment

In the case that the center of rotation is fixed at the mid point between the axes ($D_1 = D_2$) in addition to the previous assumptions, it can be seen that the angle θ can be calculated according to the simpler expression in (4.9). In this case, only the geometrical constant d_0 is involved apart from the true distances of the interferometer axes d_1 and d_2 .

$$\theta = \frac{\cos^{-1} \left(\frac{d_0}{d_1 + d_2 - d_0} \right)}{2} \quad (4.9)$$

4.2.1.4 Approximation with $\tan^{-1}$ function

If d_0 is not precisely known or computational cost is critical, as for real time applications, a further approximation is that of (4.10). This relation includes only one constant (distance between axes D) and does not take into account the effect of the rotation of the reflective surface in the interferometer measurement so assumes the incident and reflected beams to be coaxial.

$$\theta = \tan^{-1} \left(\frac{d_1 - d_2}{D} \right) \quad (4.10)$$

4.2.2 Example

4.2.2.1 Expression comparison

The equations (4.8-4.9) and the approximation (4.10) are evaluated using the geometrical parameters of the experimental setup that will be used later in this chapter (full angular range $\theta = \pm 10 \text{ mrad}$, $D = 45 \text{ mm}$, $d_0 = 20 \text{ mm}$).

The geometry of the system is modelled in Matlab and evaluated for the angular range θ of the mirror. For each rotation angle θ of the mirror, the distances d_1 and d_2 are calculated using the equation 4.7. These two distances, which are the readings available in the real application, are then used in equations 4.8, 4.9 and 4.10 to back calculate the rotated angle giving θ_{f8} , θ_{f9} and θ_{f10} respectively (where the subscript

indicates the formula number). The difference between these angles and the original rotated angle θ constitutes the "Angle error" presented in Fig.4-2.

These results present a maximum error of $\pm 1\mu rad$ for the $\tan^{-1}$ approximation for the angular range of $\pm 10 mrad$ (under the assumptions stated in 4.2.1.3).

Although the zero error shown in Fig.4-2 for the formulas 4.8 and 4.9 seems trivial, it actually confirms the correctness of these expressions.

Besides that, the angle error was also evaluated with the centre of rotation at different distances from the two axes ($D_1 = 10mm, D_2 = 35mm$) showing the same angular error for (4.10) and confirming its independence of the position of the centre of rotation.

In a real system, the parallelism of the axis and the alignment of the mirror can only be adjusted within the optical tolerance of the interferometer which is usually specified in the interferometer datasheet. In our case study presented in 4.5, the angular tolerance is $\pm 2^\circ (= \pm 34.907 mrad)$. However, in that specific application where the rotational stage has an angular range of $\pm 10 mrad$, the final tolerance is reduced by this amount leading to a final value of $\pm 24.907 mrad$. The maximum rotation of the heads ($\pm 25 mrad$) in the simulations of 4.2.2.2 was selected considering this final tolerance value.

A calibration operation (with an independent autocollimator) and correction is required for higher measurement accuracy in order to compensate for all systematic effects, including the error introduced by the misalignment of the components. In this scenario, where the calibration is required, the approximation 4.10 presents a clear advantage with respect to more complex analytical expressions in terms of computational load and ease of implementation, which is of high importance if the angle is used in a highly dynamic feedback control loop .

4.2.2.2 Characteristic curve d_2 vs d_1 along the full angular range

The study of the characteristic curve d_2 vs d_1 is of great interest since it is of key importance in the development of the AARM presented in next section. The

Figure 4-2: Angle error of the analytical expressions in Eq.4.9, Eq.4.8 and the approximation in Eq.4.10 along the full angular range $\pm 10\text{mrad}$. Centre of rotation of the mirror at the same distance from both axes ($D_1 = D_2$). The results for Eq.4.9, Eq.4.8 are superimposed.

analytical expression for the distance between the axis and the mirror (Eq. 4.7) allow the simulation of d_1 and d_2 for the angular range of $\theta = \pm 10\text{mrad}$ with different alignments of the axes.

For ease of visualisation of the non-zero-convexity of the characteristic curve d_2 vs d_1 , it is convenient to rotate the coordinate system following the equations for the axis rotation:

$$\begin{aligned} d1_{rot} &= +d_1 * \cos(\theta_r) + d_2 * \sin(\theta_r) \\ d2_{rot} &= -d_1 * \sin(\theta_r) + d_2 * \cos(\theta_r) \end{aligned} \quad (4.11)$$

where θ_r is the angle of rotation of the axis and translate this rotated curve to the origin of the reference system for comparison.

The characteristic curve d_2 vs d_1 (rotated in this section with $\theta_r = -45^\circ$ for illustration purposes) and the error introduced by the equation 4.10 is simulated in the following scenarios:

- i) Changing the parallelism of the axes: Figure 4-3 presents the characteristic

(a) d_2 rotated vs d_1 rotated (translated to the origin) for different angles with zoom in the right end. Same legend as in (a).
(b) Error introduced with the $\tan^{-1}$ approximation. Same legend as in (a).

(c) Variation of θ_1 .

Figure 4-3: Simulation of full angular movement $\pm 10 \text{ mrad}$ for different axis angles ($\theta_1 = \pm 25 \text{ mrad}$) including the ideal case with $\theta_1 = 0 \text{ mrad}$.

curves d_2 vs d_1 when the axes are not maintained parallel with $d_m = 0$. The angle between the axis 1 and the vector normal to the mirror (at $0 \mu\text{rad}$), θ_1 is assumed to be equal to the angle of the second axis but with inverted sign, $-\theta_1$. This simulation includes also the ideal case with $\theta_1 = 0 \mu\text{rad}$.

ii) Changing the perpendicularity of the axes with the mirror: The characteristic curves d_2 vs d_1 when $d_m = 0$, the axes are always parallel but the angle θ_2 with respect to the mirror at $0 \mu\text{rad}$ changes are presented in figure 4-4.

iii) Changing the alignment of only one head: In the case that $d_m = 0$ and only

(a) d_2 rotated vs d_1 rotated (translated to the (b) Error introduced with the $\tan^{-1}$ approximation. Same legend as in (a).

(c) Variation of θ_2 .

Figure 4-4: Simulation of full angular movement $\pm 10 \text{ mrad}$ for different angles between the axes and the mirror ($\theta_2 = \pm 25 \text{ mrad}$).

(a) d_2 rotated vs d_1 rotated (translated to the origin) for different angles. (b) Error introduced with the $\tan^{-1}$ approximation. Same legend as in (a).

(c) Variation of θ_3 .

Figure 4-5: Simulation of full angular movement $\pm 10 \text{ mrad}$ maintaining one axis fixed and varying the angle of the second one ($\theta_3 = \pm 25 \text{ mrad}$).

the second axis changes its alignment (θ_3) with respect to the mirror, the characteristic curves d_2 vs d_1 are presented in figure 4-5.

iv) Changing the distance between the reflective surface and the centre of rotation:

The figure 4-6 presents the results when the alignment of the interferometer axes is ideal and the distance between the reflective surface and the center of rotation varies. The figure 4-6.(b) shows the independence of the $\tan^{-1}$ approximation with respect to d_m variations.

(a) d_2 rotated vs d_1 rotated (translated to the origin) for different d_m values.

(b) Error introduced with the $\tan^{-1}$ approximation. Same legend as in (a). The three curves are superimposed.

(c) Variation of d_m .

Figure 4-6: Simulation of full angular movement $\pm 10 \text{ mrad}$ varying the distance of the reflective surface to the centre of rotation ($d_m = \pm 2mm$).

The figures 4-3(b), 4-4(b), 4-5(b) and 4-6(b) present the angle error introduced by the *atan* approximation in these different scenarios, showing the effect of the axis misalignment on the experimental angle accuracy. The worst case can be observed in figure 4-3.b, with a general error within $\pm 400\mu rad$ in all cases. This can however be corrected via calibration.

The results for the different axis and reflective surface alignments show that, in all cases, the characteristic curve d_2 vs d_1 presents non-zero convexity, a property that will be exploited in the proposal of the next section.

4.2.3 Geometrical study with corner-cube-retroreflector setup

Alternatively to that reflective surface setup, this surface can be replaced by a corner-cube retroreflector. The specific geometry of this optical component, showed in Figure 4-7, guarantees parallelism between the incoming and outgoing beams allowing a higher alignment tolerance of the interferometric axes.

The angular measurement system considered in this subsection is comprised of two independent interferometer axes as shown in Figure 4-1(b). Both interferometric axes are mounted parallel to each other on a fixed support and the beams are reflected back from two corner-cube-retroreflectors that are fixed to the rotating part of the system. The origin of the reference system is at the centre of rotation.

4.2.3.1 Distance calculus

4.2.3.1.1 Distance from head surface to retroreflector face: The distance of the incident beam from the interferometer axis to the retroreflector face, d_i , can be expressed as

$$d_i = |\lambda_1 \cdot \hat{v}_i| \quad (4.12)$$

where using the general equation of the plane leads to the following expression for λ_1 :

Figure 4-7: Internal reflection in a corner-cube retroreflector.

$$\lambda_1 = \frac{-a_0 - n_x \cdot r_{i,x} - n_y \cdot r_{i,y} - n_z \cdot r_{i,z}}{n_x \cdot v_{i,x} + n_y \cdot v_{i,y} + n_z \cdot v_{i,z}} = \frac{d_m - \hat{\mathbf{n}} \cdot \mathbf{r}_i}{\hat{\mathbf{n}} \cdot \hat{\mathbf{v}}_i} \quad (4.13)$$

where $a_0 = -\hat{\mathbf{n}} \cdot \mathbf{p}_0 = -d_m$ and using the scalar product identity in (4.13) and its substitution in (4.12) leads to

$$d_i = \left| \frac{d_m - |\mathbf{r}_i| \cdot \cos(\angle(\hat{\mathbf{n}}, \mathbf{r}_i))}{\cos(\angle(\hat{\mathbf{n}}, \hat{\mathbf{v}}_i))} \right| \quad (4.14)$$

4.2.3.1.2 Distance inside retroreflector: The incident beam is refracted in the retroreflector face according to Snell's law and propagates in the interior of the retroreflector with the following unitary direction vector:

$$\hat{\mathbf{v}}_{r_1} = \frac{n_1}{n_2} (\hat{\mathbf{n}} \times (-\hat{\mathbf{n}} \times \hat{\mathbf{v}}_i)) - \hat{\mathbf{n}} \sqrt{1 - \left(\frac{n_1}{n_2}\right)^2 (\hat{\mathbf{n}} \times \hat{\mathbf{v}}_i)(\hat{\mathbf{n}} \times \hat{\mathbf{v}}_i)} \quad (4.15)$$

The light beam is reflected in the three interior mirrors of the retroreflectors, covering the distances d_{r_1} , d_{r_2} , d_{r_3} and d_{r_4} , that can be calculated as:

$$d_{r_1} = |\lambda_{r_1} \cdot \hat{\mathbf{v}}_{r_1}|, \text{ where } \lambda_{r_1} = \frac{-\hat{\mathbf{n}}_{r_{m_1}} \cdot \mathbf{p}_{0_{m_1}} - \hat{\mathbf{n}}_{r_{m_1}} \cdot \mathbf{r}_{r_1}}{\hat{\mathbf{n}}_{r_{m_1}} \cdot \hat{\mathbf{v}}_{r_1}} \quad (4.16)$$

$$d_{r_2} = |\lambda_{r_2} \cdot \hat{\mathbf{v}}_{r_2}|, \text{ where } \lambda_{r_2} = \frac{-\hat{\mathbf{n}}_{r_{m_2}} \cdot \mathbf{p}_{0_{m_2}} - \hat{\mathbf{n}}_{r_{m_2}} \cdot \mathbf{r}_{r_2}}{\hat{\mathbf{n}}_{r_{m_2}} \cdot \hat{\mathbf{v}}_{r_2}} \quad (4.17)$$

$$d_{r_3} = |\lambda_{r_3} \cdot \hat{\mathbf{v}}_{r_3}|, \text{ where } \lambda_{r_3} = \frac{-\hat{\mathbf{n}}_{r_{m_3}} \cdot \mathbf{p}_{0_{m_3}} - \hat{\mathbf{n}}_{r_{m_3}} \cdot \mathbf{r}_{r_3}}{\hat{\mathbf{n}}_{r_{m_3}} \cdot \hat{\mathbf{v}}_{r_3}} \quad (4.18)$$

$$d_{r_4} = |\lambda_{r_4} \cdot \hat{\mathbf{v}}_{r_4}|, \text{ where } \lambda_{r_4} = \frac{-\hat{\mathbf{n}} \cdot \mathbf{p}_0 - \hat{\mathbf{n}} \cdot \mathbf{r}_{r_4}}{\hat{\mathbf{n}} \cdot \hat{\mathbf{v}}_{r_4}} \quad (4.19)$$

Where $\hat{\mathbf{n}}_{r_{m_1}}$, $\hat{\mathbf{n}}_{r_{m_2}}$ and $\hat{\mathbf{n}}_{r_{m_3}}$ are the unitary normal vectors of three interior mirrors of the retroreflector and the points $\mathbf{p}_{0_{m_1}}$, $\mathbf{p}_{0_{m_2}}$ and $\mathbf{p}_{0_{m_3}}$ correspond to any arbitrary point contained in plane of the same interior mirrors respectively. Besides that, the unitary direction vectors $\hat{\mathbf{v}}_{r_2}$, $\hat{\mathbf{v}}_{r_3}$ and $\hat{\mathbf{v}}_{r_4}$ can be obtained by simple reflection in the three interior mirror surfaces as in Section 4.2.1.1.2.

4.2.3.1.3 Distance from retroreflector face to head surface: The beam is refracted again in the retroreflector external face, and thanks to the specific corner-cube geometry the direction of the outgoing beam is parallel to the incoming beam, $\hat{\mathbf{v}}_r = -\hat{\mathbf{v}}_i$.

The final distance d_r can be calculated as

$$d_r = |\lambda_r \cdot \hat{\mathbf{v}}_r|, \text{ where } \lambda_r = \frac{-\hat{\mathbf{n}}_h \cdot \mathbf{p}_{0_h} - \hat{\mathbf{n}}_h \cdot \mathbf{r}_r}{\hat{\mathbf{n}}_h \cdot \hat{\mathbf{v}}_r} \quad (4.20)$$

where $\hat{\mathbf{v}}_r$ and $\mathbf{r}_r$ correspond to the unitary direction vector and initial point of the outgoing beam from the retroreflector, $\hat{\mathbf{n}}_h = \hat{\mathbf{v}}_i$ is the unitary normal vector to the interferometric head and $\mathbf{p}_{0_h}$ is any arbitrary point contained in the head plane.

4.2.3.1.4 Total distance: The physical path of the beam is the combination of all the distances in and out of the retroreflector. However, it is important to remark that the refractive index of the retroreflector, typically $n_2 = 1.5168$ for N-BK7 material, can be different to the refractive index of the external media n_1 and in this case, the total light path is multiplied by its corresponding refractive index.

Finally, the total distance measured by each interferometric axis is assumed to be

the average between the incoming and outgoing beam paths leading to the expression 4.21.

$$d_{total} = \frac{(d_i + d_r) * n_1 + (d_{r1} + d_{r2} + d_{r3} + d_{r4}) * n_2}{2} \quad (4.21)$$

Expression for a single axis but applicable to both.

4.2.4 Example

4.2.4.1 Expression comparison

The approximation (4.10) is equally applicable to the case of retroreflectors and it can be evaluated using the same geometrical parameters selected for the flat reflective surface: full angular range $\theta = \pm 10 \text{ mrad}$, $D = 45 \text{ mm}$, $d_0 = 20 \text{ mm}$.

The geometry of the system is modelled in Matlab and evaluated for the total angular range θ of the retroreflector structure. For each rotation angle θ , the distances corresponding to the interfometric axes, d_1 and d_2 , are calculated using the equation 4.21. These two distances, which are the readings available in the real application, are then used in equation 4.10 to back calculate the rotated angle giving θ_{ccr} (where the subscript indicates the corner-cube retroreflector setup). The difference between this angle and the original rotated angle θ constitutes the "Angle error" presented in Fig.4-8.

This result presents a maximum error of $\pm 0.5 \mu\text{rad}$ for the $\tan^{-1}$ approximation for the angular range of $\pm 10 \text{ mrad}$ (under the assumptions stated in 4.2.1.3), which is half of the $\pm 1 \mu\text{rad}$ error found for the case of a flat reflective surface. It is important to note the inversion of sign between the two setups. Although the maximum error found with the retroreflector setup suggest a clear improvement with respect to the flat reflective surface setup, the results in section 4.2.4.2 will prove the high sensitivity of this setup to the retroreflector misaligning (with respect to the beam axis), and therefore a calibration operation is still needed in practice to achieve higher measurement accuracy.

The angle error was also evaluated with the centre of rotation at different distances

Figure 4-8: Angle error of the analytical expressions in Eq.4.9, Eq.4.8 and the approximation in Eq.4.10 along the full angular range $\pm 10\text{mrad}$. Centre of rotation of the mirror at the same distance from both axes ($D_1 = D_2$). The results for Eq.4.9, Eq.4.8 are superimposed.

from the two axes ($D_1 = 10\text{mm}$, $D_2 = 35\text{mm}$) showing the same angular error for the approximation 4.10 and confirming its independence of the position of the centre of rotation also in the case of corner-cube retroreflector setup.

4.2.4.2 Characteristic curve d_1 vs d_1

For ease of visualisation of the non-zero-convexity of the characteristic curve d_2 vs d_1 , it is convenient to rotate the coordinate system following the equations 4.11 for the axis rotation.

Unless expressly stated otherwise, it is assumed that the retroreflector vertexes are perfectly aligned with the interferometric axes (at $0\mu\text{rad}$), the refractive index of the retroreflectors is $n_2 = 1.5168$ (corresponding to N-BK7 material) and the refractive index of the external media $n_1 = 1$ (corresponding to vacuum).

The characteristic curve d_2 vs d_1 (rotated in this section with $\theta_r = -45^\circ$ and translated to the centre of reference for illustration purposes) and the error introduced

(a) d_2 rotated vs d_1 rotated (translated to the origin) for different angles. (b) Error introduced with the $\tan^{-1}$ approximation.

(c) Variation of θ_1 .

Figure 4-9: Simulation of full angular movement $\pm 10 \text{ mrad}$ for different axis angles ($\theta_1 = \pm 25 \text{ mrad}$) including the ideal case with $\theta_1 = 0 \text{ mrad}$.

by the equation 4.10 is simulated in the following scenarios :

- i) Changing the parallelism of the axes: Figure 4-9 presents the characteristic curves d_2 vs d_1 when the axes are not maintained parallel (with $d_m = 0$). The angle between the axis 1 and the vector normal to the retroreflector structure (at $0\mu rad$), θ_1 is assumed to be equal to the angle of the second axis but with inverted sign, $-\theta_1$. This simulation includes also the ideal case with $\theta_1 = 0\mu rad$.
- ii) Changing the perpendicularity of the axes with the mirror: The characteristic curves d_2 vs d_1 when $d_m = 0$, the axes are always parallel but the angle θ_2 with respect to the retroreflector structure at $0\mu rad$ changes are presented in figure

(a) d_2 rotated vs d_1 rotated (translated to the origin) for different angles. (b) Error introduced with the $\tan^{-1}$ approximation.

(c) Variation of θ_2 .

Figure 4-10: Simulation of full angular movement $\pm 10 \text{ mrad}$ for different angles between the axes and the retroreflector structure ($\theta_2 = \pm 25 \text{ mrad}$).

4-10.

- iii) Changing the alignment of only one head: In the case that $d_m = 0$ and only the second axis changes its alignment (θ_3) with respect to the retroreflector structure, the characteristic curves d_2 vs d_1 are presented in figure 4-11.
- iv) Changing the distance between the retroreflector structure and the centre of rotation: The figure 4-12 presents the results when the alignment of the interferometer axes is ideal and the distance between the reflective surface and the center of rotation varies. The figure 4-12(b) shows the independence of the $\tan^{-1}$ approximation with respect to d_m variations.

(a) d_2 rotated vs d_1 rotated (translated to the origin) for different angles. (b) Error introduced with the $\tan^{-1}$ approximation. Same legend as in (a).

(c) Variation of θ_3 .

Figure 4-11: Simulation of full angular movement $\pm 10 \text{ mrad}$ maintaining one axis fixed and varying the angle of the second one ($\theta_3 = \pm 25 \text{ mrad}$).

(a) d_2 rotated vs d_1 rotated (translated to the origin) for different d_m values. (b) Error introduced with the $\tan^{-1}$ approximation. Same legend as in (a). The three curves are superimposed.

(c) Variation of d_m .

Figure 4-12: Simulation of full angular movement $\pm 10 \text{ mrad}$ varying the distance of the reflective surface to the centre of rotation ($d_m = \pm 2mm$).

(a) d_2 rotated vs d_1 rotated (translated to the origin) for different refractive indexes.

(b) Error introduced with the $\tan^{-1}$ approximation. Same legend as in (a). The three curves are superimposed.

Figure 4-13: Simulation of full angular movement $\pm 10 \text{ mrad}$ varying the refractive index of the retroreflectors. The surrounded media was considered as vacuum, with refractive index equal 1.

- v) Changing the index of refraction of the retroreflectors: The figure 4-13 presents the results when the alignment of the interferometer axes is ideal and the refractive index of the retroreflectors varies. The figure 4-13.(b) shows the independence of the $\tan^{-1}$ approximation also with respect to variations in the refractive index.
- vi) Changing the X-alignment of the retroreflector: The figure 4-14 presents the results when the retroreflector vertex is displaced by Δx mm. The displacement in Z-direction of the retroreflectors was also simulated and the results show the invariance of the difference distances with respect to this direction.

The figures 4-9(b), 4-10(b), 4-11(b), 4-12(b), 4-12(b) and 4-14(b) present the angle error introduced by the atan approximation in these different scenarios, showing the effect of the axis misalignment or refractive index on the experimental angle accuracy. The worst case can be observed in Figure 4-9, with a general error within $\pm 12500 \mu rad$ in all cases. However, this error can be corrected via calibration as in the flat reflective surface case.

The results for the different axes and retroreflector structure alignments show that, in all cases, the characteristic curve d_2 vs d_1 presents non-zero convexity, similarly to

(a) d_2 rotated vs d_1 rotated (translated to the origin) for different d_m values. (b) Error introduced with the $\tan^{-1}$ approximation. Same legend as in (a). The three curves are superimposed.

(c) Variation of Δx .

Figure 4-14: Simulation of full angular movement $\pm 10 \text{ mrad}$ varying the X-alignment of the retroreflectors.

the results for the flat reflective surface setup. Only in the case that the refractive index of the retroreflector is equal to the external media $n_1 = n_2$, the characteristic curve has zero-convexity and the recovery method that will be proposed in next section wont be applicable, thereby excluding the use of hollow retroreflectors.

4.3 Absolute angle recovery method (AARM)

4.3.1 Description

The method described in this section establishes an absolute angle reference based on the characteristic curve d_2 vs d_1 over the full angular range [130].

The non-zero convexity of the characteristic curve found in section 4.2 results in a specific characteristic curve for each system that depends only on the geometrical alignment of the setup and is independent of the interferometer unit.

This novel method proposes the use of this setup-specific curve as a reference system since any loss of reference of the interferometer, independently of its cause, will result in the addition of an unknown linear offset in each interferometer axis.

The shape of the characteristic curve remains unaltered in both reference systems O and O' (where O is the initial reference system and O' the reference system after the interferometer reset as shown in Figure 4-15). Hence, it is possible to estimate the distance offsets (c_1 and c_2) that would superpose the new curve in O' over the original curve in O .

Using the approximation studied in section 4.2 the angle can be calculated using the distances in both reference systems according to:

$$\theta = \tan^{-1} \left(\frac{d_1 - d_2}{D} \right) = \theta' = \tan^{-1} \left(\frac{d'_1 - c_1 - (d'_2 - c_2)}{D} \right) \quad (4.22)$$

This novel method allows the recovery of the absolute angle in case of any interferometer error. Unlike other techniques that include additional components like reference switches, this method achieves the recovery without requiring any addi-

Figure 4-15: Characteristic curve in the reference systems O and O' (rotated for visualisation purposes).

tional element, which is of great interest in applications with strict requirements such as ultra-high vacuum or radiation compatibility.

A full angular movement, in order to obtain the characteristic curve, in the old and new reference system is needed for the application of the recovery method. It might be an important limitation in some specific applications where this full angular movement might not be allowed, but it is common to many "homing" techniques used for device initialisation.

It is also important to remark that the correct operation of the recovery method assumes that the mechanics of the system does not change between the acquisitions of the characteristic curve and the recovery. The performance of the recovery method would be affected if the mechanics changes over time, which might be caused due to different factors like degradation or thermal effects.

4.3.2 Uncertainty analysis

The uncertainty propagation formula can be applied to both expressions in 4.22 and assuming zero correlation coefficients between the errors of the variables ($\rho_{d_1 d_2} = \rho_{D d_1} = \rho_{D d_2} = \rho_{d'_1 d'_2} = \rho_{D d'_1} = \rho_{D d'_2} = \rho_{d'_1 c_1} = \rho_{d'_1 c_2} = \rho_{d'_2 c_1} = \rho_{d'_2 c_2} = 0$) it can be seen that

$$\delta\theta^2 = \left(\frac{\partial\theta}{\partial D}\right)^2 \delta D^2 + \left(\frac{\partial\theta}{\partial d_1}\right)^2 \delta d_1^2 + \left(\frac{\partial\theta}{\partial d_2}\right)^2 \delta d_2^2 \quad (4.23)$$

$$\begin{aligned}
\delta\theta'^2 = & \left(\frac{\partial\theta'}{\partial D}\right)^2 \delta D^2 + \left(\frac{\partial\theta'}{\partial d'_1}\right)^2 \delta d_1'^2 + \left(\frac{\partial\theta'}{\partial d'_2}\right)^2 \delta d_2'^2 + \\
& + \left(\frac{\partial\theta'}{\partial c_1}\right)^2 \delta c_1^2 + \left(\frac{\partial\theta'}{\partial c_2}\right)^2 \delta c_2^2 + \\
& + 2 \cdot \left(\frac{\partial\theta'}{\partial c_1}\right) \cdot \left(\frac{\partial\theta'}{\partial c_2}\right) \cdot \rho_{c_1 c_2} \cdot \delta c_1 \cdot \delta c_2 \quad (4.24)
\end{aligned}$$

The proposed method estimates the distance offsets c_1 and c_2 and therefore, the last three terms in 4.24 can be identified as the uncertainty introduced by the angle recovery method:

$$\begin{aligned}
\delta\theta_{rec}^2 = & \left(\frac{\partial\theta}{\partial c_1}\right)^2 \delta c_1^2 + \left(\frac{\partial\theta}{\partial c_2}\right)^2 \delta c_2^2 + \\
& + 2 \cdot \left(\frac{\partial\theta}{\partial c_1}\right) \cdot \left(\frac{\partial\theta}{\partial c_2}\right) \cdot \rho_{c_1 c_2} \cdot \delta c_1 \cdot \delta c_2 \quad (4.25)
\end{aligned}$$

The cross correlation coefficient $\rho_{c_1 c_2}$ cannot be assumed as zero since the sensitivity of the recovery method is not equal for all directions. In fact, as it will be illustrated by the Monte Carlo simulations, the estimated c_1 and c_2 (and their error with respect to the real value) are highly correlated with a value that is approximately equal to -1. If the uncertainty in the estimated distances are assumed to be equal

($\delta c = \delta c_1 = \delta c_2$), the equation 4.25 can be evaluated using the partial derivatives as

$$\begin{aligned}
\delta\theta_{rec}^2 &= 2 \cdot \left(\frac{D}{D^2 + (d'_1 - c_1 - d'_2 + c_2)^2} \right)^2 \delta c^2 + \\
&\quad + 2 \cdot \left(\frac{D}{D^2 + (d'_1 - c_1 - d'_2 + c_2)^2} \right)^2 \cdot |\rho_{c_1 c_2}| \cdot \delta c^2 \\
&\leq 2 \cdot \left(\frac{D}{D^2 + (d'_1 - c_1 - d'_2 + c_2)^2} \right)^2 \delta c^2 + \\
&\quad + 2 \cdot \left(\frac{D}{D^2 + (d'_1 - c_1 - d'_2 + c_2)^2} \right)^2 \cdot 1 \cdot \delta c^2 \\
&\leq 4 \cdot \left(\frac{1}{D} \right)^2 \delta c^2
\end{aligned} \tag{4.26}$$

Where the inequalities are obtained for the worst case when the correlation coefficient between the errors is maximum $|\rho_{c_1 c_2}| = 1$ and the rotational axis is in the origin of the first reference system $d'_1 - c_1 = 0$ and $d'_2 - c_2 = 0$. Finally the equation 4.27 provides an expression for the angle uncertainty introduced by the recovery method as a function of the estimated distance offset uncertainty:

$$\delta\theta_{rec}^2 \leq \left(\frac{2}{D} \right)^2 \cdot \delta c^2 \implies \delta\theta_{rec} \leq \frac{2}{D} \cdot \delta c \tag{4.27}$$

4.3.3 Procedure for absolute angle recovery.

The proposed reference recovery method can be summarised in the following steps:

1. Characterisation of the curve d_2 vs d_1 in the initial reference system O :
 - (a) Acquisition of the full range angular movement (from the lower angular limit to the upper limit and back to the initial position).
 - (b) Identification of the characteristic curve d_2 vs d_1 of the rotational stage that will be used as the zero reference.
2. Characterisation of the curve d'_2 vs d'_1 in the new reference system O' :

- (a) Acquisition of the full range angular movement as performed in the zero reference system.
 - (b) Identification of the characteristic curve d'_2 vs d'_1 .
3. Calculation of the offset distances for c_1 and c_2 to match both reference systems:
- (a) Optimisation problem with two parameters (c_1, c_2) to match the characteristic curve d_2 vs d_1 and d'_2 vs d'_1 .
 - (b) Addition of the distance offsets in the angle calculation of O' that will recover the angle as in the reference system O (equation 4.22).

It is important to note that the characteristic curve can be obtained with the system in closed loop angle control. However, the movement to the positive and negative extremes can present windup problems in the closed loop controller. For this reason it is more convenient to measure the characteristic curve in open loop where all the angular range of the system can be exploited.

4.3.4 Optimisation problem

The distance offsets $\mathbf{c}_e = [c_{e1}, c_{e2}]$ can be estimated by minimising the least squares criterion cost function:

$$\mathbf{c}_e = \arg \min_{\mathbf{c}} \int_{r=d_{1min}}^{d_{1max}} \boldsymbol{\epsilon}^T(r, \mathbf{c}) \boldsymbol{\epsilon}(r, \mathbf{c}) dr \quad (4.28)$$

such that $\mathbf{c} = [c_1, c_2] \in \mathbf{D}_1 \times \mathbf{D}_2$, with

$$\begin{aligned} \mathbf{D}_1 &= (d_{1min} - d'_{1max}, d_{1max} - d'_{1min}) \\ \mathbf{D}_2 &= (d_{2min} - d'_{2max}, d_{2max} - d'_{2min}) \end{aligned}$$

and $\boldsymbol{\epsilon}(r, \mathbf{c}) = \mathbf{x}_r - \mathbf{x}'_{r+c_1} + \mathbf{c}$ where $\mathbf{x}$ and $\mathbf{x}'$ correspond to the characteristic curves d_2 vs d_1 and d'_2 vs d'_1 in the reference systems O and O' respectively.

4.4 Evaluation of AARM with Monte Carlo Simulation

In this section, the Monte Carlo Simulation technique is used to validate and evaluate the performance of the proposed AARM. The success of the recovery lies in the capability to match two characteristic curves through an optimisation problem. The complexity of this optimisation in different scenarios makes the Monte Carlo Simulation a suitable technique to evaluate the different cases, obtain numerical estimations of its performance and determine the parameters to be used for a given setup.

The first step is to simulate an interferometer reset which is easily achieved by adding distance offsets to both interferometer axes at a given instant.

The real application of the AARM implies the use of characteristic curves with a finite number of points and certain level of noise in the signals. The Monte Carlo simulation is used to analyse the effect of both parameters in the performance of the recovery.

The simulations presented below are performed according to the following steps for the case of reflective surface setup:

1. Simulation of the mirror movement to generate the characteristic curve d_2 vs d_1 in O using the equation (4.7).
2. Simulation of the interferometer reset (addition of c_1 and c_2 to the characteristic curve d_2 vs d_1) to generate the characteristic curve in O' .
3. Addition of noise (with specific standard deviation and correlation coefficient) to d_1 and d_2 of both reference systems. (The noise is generated independently for both curves).
4. Application of the absolute angle recovery method to estimate c_1 and c_2 . The characteristic curves are fit to a 10th order polynomial before the optimisation.

The last two steps 3 and 4 are repeated 100 times for each scenario to obtain significant statistical results. The components of the system are considered to be perfectly aligned (axes parallelism and perpendicularity with the mirror) and with the same parameters of the experimental setup that will be studied later ($D = 45mm$, $d_0 = 20mm$, $\theta = \pm 10mrad$). The distance offsets used in the simulations are $c_1 = 15 \mu m$ and $c_2 = 25 \mu m$.

The values for the standard deviation of the noise ($\sigma_{noise} = 0.012983\mu m$) and Pearson correlation coefficient ($\rho_{noise} = -0.1034$) used in the Monte-Carlo simulations were obtained directly from the experimental acquisitions of [130] and are very similar to that observed in section 4.5, where the interferometer has a resolution of picometers. The idea was to use noise levels in the simulations similar to the ones observed experimentally.

The effect of the type of noise in the recovery method is analysed evaluating the cases with non-correlated noise and with realistic correlated noise. Moreover, the influence on the performance of the AARM of the number of points and the noise level are studied.

4.4.1 Uncorrelated noise in the characteristic curves

In this simulation, noise of a similar level to the experimental setup of the next section is added to the characteristic curves but maintaining zero correlation coefficient between the added noise to d_1 and d_2 for each curve. Table 4.1 summarises the results that were obtained using the following parameters:

- $\sigma_{noise} = 0.012983\mu m$,
- $\rho_{noise} = 0$ (Pearson correlation coefficient).
- Number of points in the characteristic curve: 340000.

c_1 (μm)			c_2 (μm)			Shift angle (μrad)		
Mean	Error	STD	Mean	Error	STD	Mean	Error	STD
15.001	0.001	0.0041	24.999	-0.001	0.0041	-222.2	0.0	0.18

Table 4.1: Results for Monte Carlo simulation with uncorrelated noise.

4.4.2 Realistic correlated noise in the characteristic curves

The Table 4.2 presents the results with noise and non-zero correlation coefficient, similar to that observed in the experimental setup, added to the characteristic curves. The parameters used in the simulation are:

- $\sigma_{noise} = 0.012983\mu m$,
- $\rho_{noise} = -0.1034$ (Pearson correlation coefficient).
- Number of points in the characteristic curve: 340000.

The results do not show significant difference with respect to the zero correlated noise case.

c_1 (μm)			c_2 (μm)			Shift angle (μrad)		
Mean	Error	STD	Mean	Error	STD	Mean	Error	STD
15.000	+0.000	0.0036	25.000	0.000	0.0036	-222.2	0.0	0.16

Table 4.2: Results for Monte Carlo simulation with correlated noise.

4.4.3 Effect of number of points in the recovery method

The number of points of the characteristic curve is the only parameter that can be controlled in the real application of the AARM. For this reason, the study of the effect of the number of points in the performance of the recovery is of great interest.

Figure 4-16 presents the results of the simulation with uncorrelated noise . It clearly shows an important decrease in the standard deviation of the results when the number of points increases. Similar results were obtained for the case with correlated noise. Equation (4.27) was also evaluated and included in the figure and fits the simulation results closely.

On the other hand, it is important to remark that the increase in the number of points makes the optimisation more computationally intensive. Fortunately in real applications the AARM will typically be applied only in the event of an interferometer reset and therefore the computation time should not be a limiting factor.

4.4.4 Effect of noise standard deviation in the recovery method

The effect of the standard deviation of the sensor noise (added to the characteristic curves) in the recovery performance is studied in this simulation and presented in Fig. 4-17 for the case of non-correlated noise.

The results show how the standard deviation of the estimated distances varies nearly linear with the standard deviation levels of the noise in the characteristic curves, obtaining better results when the noise levels are smaller. Again, the simulations with correlated noise present no significant difference with respect to Fig. 4-17.

Although apparently the values for the standard deviation of c_1 and c_2 seem to be the same in Fig. 4-16 and 4-17, they are actually very close but different. Their similarity is a direct consequence of the high cross-correlation found between c_1 and c_2 , a value that is very close to -1.

4.5 Case Study - Experimental Results

In order to evaluate and confirm the good results obtained in the simulations, the AARM was applied to a high-precision rotational stage used for crystal collimation at CERN.

Figure 4-16: Monte Carlo simulation varying the number of points with $\rho_{noise} = 0$ and $\sigma_{noise} = 0.012983\mu\text{m}$ (100 repetitions for each number of points evaluation). The theoretical shift angle standard deviation is calculated using Eq. 4.27 and the mean of the standard deviation of c_1 and c_2 .

Figure 4-17: Monte Carlo simulation varying σ_{noise} with a fixed number of points (340000) and $\rho_{noise} = 0$ (300 repetitions for each number of points evaluation). The theoretical shift angle standard deviation is calculated using Eq. 4.27 and the mean of the standard deviation of c_1 and c_2 .

(a) Rotational stage from above without interferometric system

(b) Interferometer axes and mirror from the side.

Figure 4-18: Rotational stage setup.

4.5.1 Experimental setup

4.5.1.1 Rotational Stage

The rotational stage was composed of a piezoactuator that applies a force to a pre-stressed flexure-hinge structure producing the rotation of the moving head as shown in Figure 4-18(a) and described in detail in Section 2.2.2.

In this case, the stage presents a nominal angular range of 20mrad with a linear stroke of the piezoactuator of $40\mu\text{m}$ (V1 of the rotational stage). The flat reflective surface is formed by a gold-coated mirror that is attached to the rotational head with the use of a support. The two interferometer axis are fixed to the external structure of the rotational stage (non-moving part) oriented towards the mirror.

This primary measurement system is shown in the right side of Figure 4-18(b), where the reflecting surface is a double-sided mirror with the possibility to mount a secondary measurement system (shown in the left side of Fig. 4-18(b)) whose use is explained below.

4.5.1.2 Angular measurement setup

The interferometer-based system was formed of a 3 axis FPS3010 Attocube interferometer unit with $1pm$ resolution, $2nm$ repeatability (at $10mm$ distance from the head to the mirror), accuracy at ambient conditions of $0.14ppm$ and maximum target velocity of $2m/s$ and with distance D between axes of $45mm$. It allows the angular reading with $< 0.025\mu rad$ uncertainty due to the interferometer system (Type B evaluation [79]), where the resolution, repeatability and accuracy (for maximum $\pm 350\mu m$ displacement) uncertainties are combined using the variance addition rule assuming no cross-correlation between them, perfect alignment of the optical components and excluding the uncertainty in D .

Apart from the primary dual-interferometer system, the angle was measured simultaneously with a secondary dual-interferometer system. The secondary system provides an independent measurement and it is used to evaluate the performance of the recovery method.

In this case, the dual-interferometer system is formed of the same mirror (but opposite face) of the primary interferometer system as can be seen on the left side of Fig. 4-18(b). Both interferometer units had the same specifications with the same distance between axes.

4.5.2 Optimisation problem

As described in the method procedure, a full angular range movement should be performed in each reference system to be compared.

In practice, these two acquisitions might present different number of points, different sample rates or simply the rotational stage might show a different angular range (due to hysteresis or ageing of the actuator). For these reasons, instead of the raw data from the interferometer, the characteristic curves d_2 vs d_1 were fitted before their use in the optimisation problem.

The presented results of this chapter were obtained using a smoothing spline (with a smoothing parameter very close to 1, which is able to fit the small oscillation of

the characteristic curve) instead of the 10th order polynomial used in the Monte Carlo simulations and the results of [130], and the non-linear programming solver *fminsearch()* in MATLAB[®]. Each characteristic curve is fitted to a single curve resulting, approximately, in the average of the slightly hysteretic behaviour found in Figures B-9(a) and 4-19(a).

4.5.3 Performance of the Absolute Angle Recovery Method

The reading of the interferometer was intentionally lost by interrupting one of the laser beams, by resetting manually all the axis positions or simply by restarting the interferometer unit. As commented in section 4.3.1, each position loss resulted in a translation of the angular reference system. The new method was applied to recover the initial angular reference after each interferometer reset.

Two different tests were performed with 5 and 15 repetitions in total respectively. Additionally, during the second test the temperature was maintained within $25.0^\circ \pm 0.2^\circ C$ to simulate the LHC environment where the temperature is controlled

The characteristic curves d_1 vs d_2 for the different reference systems after the losses of position by the interferometer are shown in Fig. 4-19(a) for the Test 2, whereas the results for Test 1 can be found in the Appendix B.6. In both cases, the axes are rotated according to the equations in 4.11, with an angle of rotation calculated from the slope of the linear regression obtained with the first experimental dataset (Ref 0).

Due to the high number of repetitions in Test 2, for visualisation purposes only, the characteristic curves in the reference systems number 0, 2, 6, 9, 12 and 15 are displayed.

The Test 1 was performed with 340000 points in each acquisition and presented mean values for the noise statistics of $\sigma_{noise} = 0.012983\mu m$ and $\rho_{noise} = -0.1034$ (Pearson correlation coefficient), parameters that were used in the Monte Carlo simulation.

Similarly, the Test 2 was performed 204000 points in each acquisition and presented mean values for the noise statistics of $\sigma_{noise} = 0.014\mu m$ and $\rho_{noise} = -0.13$ (Pearson correlation coefficient).

(a) Characteristic curves d_2 vs d_1 (rotated for visualisation purposes) for the different reference systems after the loss of position in the interferometer.

(b) External interferometer acquisition for the rotational stage in closed loop at $0 \mu\text{rad}$ for the different reference systems. Ref 0: Original reference system, Ref 1-15: Angle recovered. The mean and standard deviation are calculated using a 10s acquisition where the angle is stable thanks to the closed loop angle control. The horizontal dashed line marks the origin in the original reference system.

Figure 4-19: Test 2 with 15 repetitions and controlled temperature.

For each reference system (after the angle recovery), the rotational stage was moved in closed loop (to avoid the non-linear creep and hysteresis effects of the piezo actuator) to a reference angle ($0\mu rad$). In this position, the real angle was acquired with the external interferometer system during 7 (Test 1) and 10 (Test 2) seconds ($2kHz$ sampling frequency) and the results for the mean angle with the standard deviation for the Test 2 are presented in Figure 4-19(b), and Figure B-9(b) of the Appendix for Test 1. Since the external interferometer did not experience any error during all the angular recoveries, its angular reading after each recovery (at the fixed position of $0\mu rad$) should be equal to the angle obtained for the first reference system Ref 0.

The standard deviation of the recovered angles for the Test 1 (Ref 1-5) is $0.53\mu rad$ and their average value differs from the initial value (Ref 0) by $0.06\mu rad$, see Appendix B.6 with the detailed results.

The results in the Test 2 are slightly better with standard deviation of the recovered angles (Ref 1-15) of $0.41\mu rad$ and their average value differs from the initial value (Ref 0) by $0.11\mu rad$.

Combining both results in each test as the repetitive and bias uncertainties of the new method, it can be concluded that the angle was successfully recovered within an accuracy of $\pm 1.65\mu rad$ (3σ) for the Test 1 and $\pm 1.29\mu rad$ (3σ) for the Test 2, with $\sigma = 0.55\mu rad$ and $\sigma = 0.43\mu rad$ respectively.

Both results are in very good agreement but it is important to note that the number of repetitions in Test 2 is higher than in Test 1. Besides, the temperature during Test 2 was controlled within $25.0^\circ \pm 0.2^\circ C$, which was not the case during Test 1.

Even though the number of repetitions is small, the experimental performance is in good agreement with the Monte Carlo simulation presented in section 4.4 where the standard deviation of the recovered angle under similar conditions was estimated to be $< 0.25\mu rad$. The difference between results can be explained by the misalignments of the axes and mirror and the presence of hysteresis (introduced by the piezoactuator) in the characteristic curve that are present in the experimental setup but not in the

simulation.

4.6 Application to LHC Piezo Goniometers

4.6.1 Angular measurement system

At the timing of writing, the dual-interferometer-based angular measurement system with flat reflective surface setup is the configuration used in the 5 operational piezo-goniometers for Crystal Collimation in CERN accelerators (4 of them distributed in Beam 1 and 2 of the LHC and another device in the SPS). The reflective surface is formed of a gold coated mirror where three interometric axes are projected. The angle is measured in the rotational axis of the device with axes 1 and 2 (yaw angle). Besides, the parasitic angle in one of the orthogonal directions is also measured with axes 1 and 3 (parasitic pitch angle).

The whole rotational stage and the angular measurement system (comprising the mirror, interferometric heads and optic fibres) is in a tank with ultra-high vacuum and it is continuously exposed to high ionising radiation doses due to its proximity to the particle beam.

The high performance of the laser interferometry allows the use of the angular measurement for closed loop control of the rotational angle. The use of optic fibres allows the placement of the interferometer electronic unit in one of the access tunnels, in a radiation free area, together with the rest of electronic equipment.

From an optical point of view, one of the most sensitive components in the optic path is the "feed-through". It allows optical connections between the inside and outside of the ultra-high vacuum tank. However, as it was observed in the installed devices, this component highly attenuates the intensity of the interferometric signal, which can lead to malfunction of the interferometer unit due to poor alignment signal intensity.

The use of corner-cube retroreflectors (with its higher range of operation and alignment tolerance) instead of the reflective surface is under study for future piezo-

goniometers. This change would allow the movement of the interferometric heads out of the ultra-high vacuum tank, removing with this the need of any "feed-through" and the subsequent increase in intensity of the optical signals.

The new piezogoniometers with corner-cube retroreflectors are under development and they are expected to be installed in the LHC in the near future.

4.6.2 Absolute angle recovery method

In the current piezogoniometers installed at CERN it is of key importance to maintain the absolute angular reference in order to know the Channelling Angle of the device.

In the event of an interferometer error (due to poor alignment signals, movements exceeding the maximum allowed speed or optical beam interruptions), the absolute angle recovery method presented in this chapter is used to recover the reference.

In other cases, the interferometer unit is turned off due to unexpected or planned (for UPS Testing) power-cuts. In these cases, the AARM is also used for the recovery of the angle when the interferometer is powered back on.

Finally, the AARM is frequently applied during the different stages of the piezogoniometer validation for the final commissioning in the accelerators. This process includes several steps where it is necessary to completely remove all connections, including the optic fibres. The channelling crystal is aligned with respect to the tunnel angular reference system and the recovery method is use to estimate the orientation of channelling crystal.

4.7 Conclusion

A dual-interferometer-based angular measurement system has been proposed and studied for two different configurations including a flat reflective surface and corner-cube retroreflectors. Analytical expressions for the distances measured by the interferometer axes and for the calculation of the angle were obtained and evaluated.

The characteristic curve d_2 vs d_1 was simulated for different alignments of the components and showed non-zero convexity in all cases except for the configuration of

corner-cube retroreflectors with the same refractive index as the external media. The error introduced in the calculus of the angle using a simpler approximation was found to have a maximum of $1\mu rad$ and $0.5\mu rad$ error (for the reflective surface and corner-cube retroreflector setup respectively) for an angular rotation of $\pm 10 mrad$, when the axis separation is $45 mm$, distance from axis to mirror is $10 mm$ and assuming perfect alignment of the components.

A new absolute angle recovery method for dual-interferometer-based rotational stages was proposed. The method is based on the use, as an absolute reference, of the characteristic curve of the rotational stage. The intrinsic convexity of the characteristic curve allows the use of this technique for absolute angle recovery, providing a method that depends uniquely on the geometry of the system.

Unlike other techniques that include additional components like reference switches, this method can achieve similar angular recovery performance without requiring any additional element, which is of great interest in applications with strict requirements such as ultra-high vacuum, radiation compatibility or high reliability.

The recovery method requires a full angular movement inspection in each new reference system, which can be a limitation in critical applications where a full angular movement might not be allowed. Besides that, the performance of the method can be significantly affected if the mechanics of the rotational stage varies over time, which might be caused by different factors (degradation of the actuator, thermal effects, etc.) affecting the system.

The performance of the recovery method is evaluated with Monte Carlo simulations. The results show how the uncertainty of the recovery decreases significantly with the number of points of the characteristic curve and increases linearly with the noise level of the distance measurements.

The performance of the angle recovery technique was experimentally evaluated on a high-precision rotational stage at CERN. The angle was recovered 15 times with success after the loss of the reference position (due to beam interruption, position reset or device restart) of the interferometer. An independent interferometer system showed an angular accuracy of $\pm 1.29\mu rad$ (3σ) in the absolute angle recovery for this specific

rotational stage. These results are published in [132]. They are in good agreement with the previous results, published in [130], and the Monte Carlo simulations.

The dual-interferometer-based angular measurement system and the proposed absolute angle recovery method are successfully applied to the five piezgoniometers installed in the CERN LHC and SPS accelerators.

Chapter 5

Phase Locked Loop Disturbance Rejection Algorithm with Observer Update

5.1 Introduction

High precision positioning systems are vital in many applications including scanning tunneling microscopes (STM) [155], atomic force microscopes (AFM) [142] and semiconductor lithography. In AFM, for instance, high precision positioning is required to control the vertical position of the scanning probe.

In all these systems, the presence of external disturbances, like vibrations due to nearby machinery, can highly compromise their performance. The spectrum of these disturbances can be wide but, due to their origin, in most cases it is possible to identify their main components, whose attenuation would have the highest impact on the system performance.

According to classical control theory [51], the response of a linear feedback controlled system to external disturbances depends on the disturbance frequency content and the system sensitivity function. Although the feedback loop attenuates all disturbance frequencies below the system bandwidth, little improvement is achieved beyond

this frequency.

The disturbance rejection problem is widely studied in the literature with different techniques depending on the targeted frequency band attenuation, and special focus on repetitive disturbances. When the disturbance spectrum is known, techniques based on the Internal Model Principle (IMP) [50], which incorporates a model of the main disturbance in the controller itself, or the inclusion of specifically tuned notch filters in the controller, have shown significant results.

Active Disturbance Rejection Control (ADRC), a term first used in the English literature in [61], was proposed as a promising alternative since the explicit disturbance model is not required [59], but it is based on the direct and active estimation (through an observer) and rejection of a generalised disturbance. The number of different observers in the literature is high [117], and they usually require a nominal mathematical model. A very successful approach is the use of the extended state observer (ESO) [144] [102], which is summarised and implemented also in this work. The ADRC is successfully applied in many applications such as motion control problems [60] or high precision positioning systems [77] [136], among others.

In the ideal case of a single frequency disturbance with known plant and disturbance models, the sensitivity function for the ADRC with ESO is zero at the disturbance frequency, as proven in [102], achieving its total cancellation. However, the Bode integral constraints [51] apply in this direct connection of the observer, which creates an additional feedback loop. The good disturbance rejection performance therefore comes at the cost of a sensitivity function increase for other nearby frequencies [53] that, as it will be shown in this work, can lead to a deterioration of the general tracking performance in some cases.

In order to overcome this limitation, a Feedforward Disturbance Rejection with Switching Scheme approach (*FDRSS*) was proposed in [53] [52] as an add-on algorithm with the capability of single or multi-frequency disturbance rejection, even beyond the original controller bandwidth, while preserving the system's nominal closed-loop characteristics.

The *FDRSS* was originally introduced to compensate for repetitive disturbances

due to disk decentering in hard disk drives. However, in more general applications where the disturbances are independent of the system or they drift in phase or frequency, the *FDRSS* produces oscillations, known as the beat-effect [127], in the cancellation performance [37].

More sophisticated adaptive control techniques are necessary when the frequency of the periodic disturbance is uncertain or time-variant [109]. The frequency estimation can be achieved through several approaches such as non-parametric power spectral estimation [82], extended Kalman filter frequency estimation [86] [20], adaptive notch filtering [93] [119] or phase locked loop algorithms [23] [146].

The estimated disturbance frequency is then applied to numerous adaptive schemes presented in literature that can be classified, according to the knowledge of the system, as: unknown plant and disturbance model [41], unknown plant model with known disturbance [137] or, the more common scenario, known plant and unknown disturbance model [88]. They can also be categorised according to their indirect or direct adaptive scheme [109]. In the first case, the disturbance frequency is estimated and used to adapt a disturbance cancellation controller maintaining the independent design of the disturbance estimation and cancellation scheme [109] [106]. On the other hand, the direct approach directly adapts the controller for unknown disturbance rejection in an integrated algorithm [147] [67] [22] [88] [94].

In high precision positioning systems, the control challenge arises due to the first mechanical resonance of the system that limits the achievable bandwidth for closed-loop control. Much effort has been made over the years to develop techniques with the purpose of increasing the closed-loop bandwidth and phase-gain margins [46]. In this scenario, the benefits of disturbance rejection approaches that can be used as an add-on algorithm for a nominal controller such as in the *FDRSS*, represent a clear advantage as long as no beat effect is present.

The contributions of this work have their starting point at the *FDRSS* and can be summarised as:

- Application of the *FDRSS* to a more general case where the disturbance frequency drifts over time to illustrate the beat-effect problem in a real application.

- Proposal of a new control algorithm, as an improvement of the FDRSS, where the switching scheme is substituted by a Phase Locked Loop with amplitude detection after the disturbance observer (PLLDR). This scheme maintains the benefits of the FDRSS (the selected frequency is highly attenuated while keeping the sensitivity function of the nominal controller unaltered for other frequencies) and eliminates the beat-effect problem, making this technique more generally applicable.
- Proposal of a novel self-tuning algorithm, based on the previous PLLDR, where the disturbance observer is updated (at a low rate) with the disturbance frequency estimated in the PLL (PLLDR with observer update). This update leads to the attenuation of a disturbance not only at the preselected frequency, but while the disturbance remains within a certain frequency range.
- The different approaches (nominal controller, direct observer, FDRSS, PLLDR and PLLDR with observer update) are simulated for the system used in the experimental part. The disturbance rejection response is analysed via the sensitivity function and the performances are studied for the cases when the disturbance frequency is known, the disturbance frequency is not known accurately and when the disturbance drifts in frequency.
- The proposed algorithms are experimentally validated on a rotational stage used for Crystal Collimation in the Large Hadron Collider (LHC) at the European Organization for Nuclear Research (CERN). The response of the different approaches to disturbances (artificially generated with a shaker) beyond the bandwidth of the nominal controller are studied and compared. The results show excellent attenuation performance of the proposed methods.

Following the nomenclature of other literature such as [88], the second proposal, PLL with observer update, is considered a self-tuning scheme. In this case, the controller parameters are frozen during the computation of the new observer parameters (with the possibility of being calculated off-line), which are updated at a low rate compared to the nominal controller. For this reason, the proposed control algorithm

cannot be categorized among the adaptive control techniques (in neither their direct nor indirect version), where the parameters of the controller (or controller and plant model) are updated at each control iteration, following the given technique's adaptive laws.

As commented before, the use of a PLL in disturbance rejection algorithms is a known technique. However, the use of a PLL structure as a decoupling element to avoid the direct feedback of the disturbance observer presented in this work is, to the authors' knowledge, a novel technique. The simulation and experimental results confirm the suitability of this method as an add-on solution to existing controllers, capable of single frequency disturbance rejection beyond the nominal controller bandwidth and with easier design and computational cost than adaptive control techniques. Compared to the schemes with a direct observer connection, the proposed techniques preserve the closed loop characteristics of the nominal system (apart from at the selected attenuation frequency), similarly to the *FDRSS* but with a suppression of the beat-effect.

The chapter is organized as follows: in Section 5.2 the basic theory for *FDRSS* is summarised. The novel approaches *PLLDR* and *PLLDR with observer update* are presented in Section 5.3. The results have been divided into two sections. The simulation results are presented in Section 5.4 while the experimental results are shown in Section 5.5. In Section 5.7 the work is summarized and conclusions are drawn.

5.2 Background

The theory for the feedforward disturbance rejection with switching scheme, [53], is summarised in this section.

5.2.1 Observer structure for disturbance estimation

Let the system be described by a continuous time linear state space representation as shown in Figure 5-1. Disturbances entering the system on the output, d_o , are

Figure 5-1: State space representation of the plant and disturbance.

modelled as input disturbances, d_i , to allow cancellation on the input as seen in equation (5.1a), where it is assumed that the plant and disturbance are linear and therefore $d_o(s) = G(s)d_i(s)$. In this case, the system can be described as follows:

$$\dot{\mathbf{x}}_s(t) = \mathbf{A}_s \mathbf{x}_s(t) + \mathbf{B}_s(u(t) + d_i(t)) \quad y(t) = \mathbf{C}_s \mathbf{x}_s(t) \quad (5.1a)$$

and the disturbance as

$$\dot{\mathbf{x}}_d(t) = \mathbf{A}_d \mathbf{x}_d(t), \quad d_i(t) = \mathbf{C}_d \mathbf{x}_d(t) \quad (5.2)$$

where $\mathbf{x}_s$ and $\mathbf{x}_d$ are the system and disturbance state vectors in the plant input, and $\mathbf{A}_s, \mathbf{B}_s, \mathbf{C}_s, \mathbf{A}_d$ and $\mathbf{C}_d$ are the known system and disturbance matrices respectively. Assuming that the disturbance is sufficiently well described as a sinusoid, the one-sided Laplace transform of $\frac{1}{w} \sin(wt)$ is $1/(s^2 + w^2)$ and yields the state space equations in (5.3) with zero input for the sinusoidal disturbance.

$$\mathbf{A}_d = \begin{bmatrix} 0 & 1 \\ -w^2 & 0 \end{bmatrix} \quad \mathbf{C}_d = \begin{bmatrix} 1 & 0 \end{bmatrix} \quad (5.3)$$

In practice, the disturbance frequency, w , can normally be estimated from the Fast Fourier Transform of the system output with the system in regulation.

The zero order hold (ZOH) time discretisation leads to the state space representation of the combined system in equations (5.4).

$$\mathbf{x}_{zs}[n+1] = \mathbf{A}_{zs}\mathbf{x}_{zs}[n] + \mathbf{B}_{zs}(u_z[n] + d_{zi}[n]) \quad (5.4a)$$

$$y_z[n] = \mathbf{C}_{zs}\mathbf{x}_{zs}[n] \quad (5.4b)$$

$$\mathbf{x}_{zd}[n+1] = \mathbf{A}_{zd}\mathbf{x}_{zd}[n] \quad (5.4c)$$

$$d_{zi}[n] = \mathbf{C}_{zd}\mathbf{x}_{zd}[n] \quad (5.4d)$$

where $u_z[n]$ and $d_{zi}[n]$ are assumed to be piecewise constant during each sampling period T_s and n is the time instant [54].

The disturbance can be estimated by the discrete predictive observer [102] described by

$$\hat{\mathbf{x}}[n+1] = \mathbf{A}\hat{\mathbf{x}}[n] + \mathbf{B}u[n] + \mathbf{K}(\mathbf{y}[n] - \mathbf{C}\hat{\mathbf{x}}[n]) \quad (5.5)$$

where $\mathbf{A}$, $\mathbf{B}$ and $\mathbf{C}$ are the augmented system matrices and $\mathbf{K}$ is the observer gain:

$$\mathbf{A} = \begin{bmatrix} \mathbf{A}_{zs} & \mathbf{B}_{zs}\mathbf{C}_{zd} \\ \mathbf{0} & \mathbf{A}_{zd} \end{bmatrix}, \mathbf{B} = \begin{bmatrix} \mathbf{B}_{zs} \\ \mathbf{0} \end{bmatrix} \quad (5.6)$$

$$\mathbf{C} = \begin{bmatrix} \mathbf{C}_{zs} & \mathbf{0} \end{bmatrix}, \quad \hat{\mathbf{x}} = \begin{bmatrix} \hat{\mathbf{x}}_{zs} \\ \hat{\mathbf{x}}_{zd} \end{bmatrix}, \quad \mathbf{y} = y_z \quad (5.7)$$

and the disturbance is estimated as $\hat{d}_{zi}[n] = \mathbf{C}_{zd}\hat{\mathbf{x}}_{zd}[n]$.

The observer gain should be tuned (placing the eigenvalues of $\mathbf{A} - \mathbf{K}\mathbf{C}$) to achieve a trade-off between the convergence rate in the state reconstruction and the sensitivity to measurement noise. If the noise characteristics are known, an optimal choice of $\mathbf{K}$

Figure 5-2: Block diagram of the *FDRSS* approach including an observer and a feedback controller.

can be calculated using Kalman theory [51]. It is shown in [52] that the disturbance rejection will be achieved at every sampling point in steady state.

Despite the excellent disturbance rejection performance for the selected frequency, the new feedback loop formed by the direct observer connection affects the entire system response by reducing the sensitivity function at the selected frequency (as desired) but increasing it at other frequencies due to the water-bed effect [51]. This effect limits the applicability of this algorithm as a “plug-in” solution and, as will be shown in the simulation and experimental results, might lead in some cases to a general performance degradation with respect to the original controller, even with the correct attenuation of the selected frequency.

5.2.2 Feedforward Disturbance Rejection with Switching Scheme

A solution to the previous limitations was presented in [53] by including the switching scheme shown in Fig. 5-2, where F , C and G represent the prefilter, feedback controller and system respectively.

The switching mechanism is defined to switch in an integer number (m) of disturbance periods (T_d) after the observed states have converged. This signal is then replicated in the delay loop and subtracted from the input signal as illustrated in the

block diagram. In practice, the original continuous delay block ($e^{-st_1} = e^{-smT_d}$) is replaced by its discrete equivalent ($z^{-\alpha}$), where $\alpha = mT_d/T_s$ is the number of samples of the delay, and the switching on and off time have to be set in advance, hence T_d must be known.

It is important to note that the number of delay samples, α , has to be an integer number and therefore the number of cycles, m , has to be selected accordingly. Depending on the relation between T_s and T_d , the integer number condition might not be possible to meet, or it might require an impractically large m . Using a large number of cycles m would reduce the mismatch consequences but it constitutes a clear limitation of this technique.

If the estimated disturbance, which is being replicated, and the real disturbance match completely, they cancel and the disturbance is rejected. However, in practice the frequencies of the estimated and real disturbances can be slightly different due to a wrong selection of the disturbance frequency or a drift in the real disturbance. In this case, the combination of both signals result in a constructive and destructive interference, known as the beat effect [127], producing an undesired oscillation in the system output. The amplitude modulation frequency of the output is twice the beat-effect frequency $|f_1 - f_2|/2$, where f_1 and f_2 are the frequencies of the interfering signals.

5.3 Proposed approaches

In this section an approach including a phase locked loop is proposed as an improvement to the *FDRSS*, by solving the beat effect. Additionally, a second proposal including an observer update is presented.

Figure 5-3: Block diagram of the *PLDR* approach including a PLL with amplitude detection.

5.3.1 Proposed method 1: Phase Lock Loop Disturbance Rejection Algorithm

The proposed solution is the replacement of the switching scheme by a phase locked loop with amplitude detection as shown in Fig.5-3.

In order to grasp the operation and novelty of the proposal it is important to understand the signal arriving to the PLL reference input. If the disturbance observer is properly designed and has converged, its output is a signal whose main component, though not unique, corresponds to the estimated disturbance at the preselected frequency. The PLL then acts as a perfect single-frequency filter, allowing only this frequency component to pass through and eliminating all other components. Unlike other linear options, such as band-pass filters, the PLL guarantees a single, in-phase sinusoid at its output and therefore does not affect the rest of the frequency response.

In this way, the PLL provides the decoupling of the observer with respect to the original closed loop system, avoiding the main drawback of the approach with direct observer feedback, the sensitivity function degradation due to the water-bed effect.

The amplitude detection module uses a circular buffer containing, at least, one period of the estimated disturbance. The amplitude is estimated by calculation of the maximum and minimum values of the elements in the rotational buffer. It is then

(a) Charge-pump PLL diagram.

(b) Charge-pump PLL with frequency band selector in blue.

Figure 5-4: Charge-pump PLL diagram and frequency band selector implementation.

used to modulate the PLL output.

Different PLL approaches have been studied in recent years. The phase is commonly detected by means of two quadrature signals that are generated from the original input signal. The different techniques for the quadrature generation include the use of the Park transform [6], the Hilbert transform [58, 81], the delay block including 90 degree delay with respect to the original signal [133] or the more recent technique of the Second Order Generalized Integrator (SOGI) [29].

5.3.1.1 Charge-Pump PLL

In this work, the PLL is implemented using the Charge-Pump approach that is widely used in different applications [100] [91] [104] thanks to its theoretical zero static phase offset and its simple and effective design methodology. This approach also allows the decoupling of different parameters like the lock range, damping factor or loop bandwidth, which is a clear advantage in the design process.

A general Charge-Pump PLL is presented in figure 5-4(a) and it is formed by a phase frequency detector (PFD), a charge-pump (CP), a passive loop filter (LF), and a voltage controller oscillator (VCO).

The PFD, which can be implemented digitally with two D-flip flops, compares the positive edges of the reference clock and the VCO output and generates a pair of digital pulses corresponding to the phase/frequency error between them. These digital pulses are converted to an analogue signal in the CP that is subsequently filtered in the LF and feeds the VCO. The negative feedback of the VCO output leads to zero phase/frequency error, and the LF is designed to guarantee the stability of the closed loop.

The stability of this type of PLL has been studied by different authors for different orders of the passive loop filter. The second-order case was studied in [62, 71].

The third-order PLL, which is implemented in this work, was studied in [69] where an exact analysis of the dynamics of the system using state equations was presented. It can be seen in [69] that the gain of the PFD with CP is equal to $I_{cp}/2\pi$ and the transfer function for the loop filter has the following expression,

$$F(s) = \frac{s + \frac{1}{RC_1}}{C_2 s \left(s + \frac{1}{R \left(\frac{C_1 C_2}{C_1 + C_2} \right)} \right)} \quad (5.8)$$

where C_1 , C_2 , R and I_{cp} are the capacitors, resistance and current, respectively, of the equivalent analogue circuit shown in Figure 5-5.

The authors in [69] proved that, with the optimal choice of the capacitor ratio C_1/C_2 , the phase margin is relatively immune to process variations and derived the expressions for phase margin (5.9) and capacitor relationship (5.10)

$$\Phi_M = \arctan \left(\sqrt{\frac{C_1}{C_2} + 1} \right) - \arctan \left(\frac{1}{\sqrt{\frac{C_1}{C_2} + 1}} \right) \quad (5.9)$$

$$C_1 = 2C_2 \left(\tan^2 \Phi_M + \tan \Phi_M \sqrt{\tan^2 \Phi_M + 1} \right) \quad (5.10)$$

The previous equations allow, for an application requiring a given loop bandwidth

Figure 5-5: Third order Charge-Pump PLL analogue equivalent [69].

and phase margin, the calculus of the variables C_1 , C_2 and I_{cp} , and where the R is selected based on noise constraints which completes the PLL design.

5.3.1.2 Interpolation technique to overcome dead-zone of the PLL

The main limitation of the charge-pump PLL is the presence of dead zone, which is the range of phase difference that does not result in charge/discharge of the PFD. When the PLL is locked, the input and output of the PLL can cross the zero level at the same sampling instant even if both signals are not perfectly in phase due to the discretization of the system. In this scenario, the PFD is not charged/discharged and therefore no phase correction is achieved.

In this specific application of disturbance cancellation, the dead-zone would cause a phase difference between the estimated disturbance (by the observer) and the disturbance compensation (PLL output) re-injected in the system, causing a clear degradation of the disturbance rejection performance of the PLL with respect to the direct observer approach.

The dead-zone problem can be reduced by generating a small signal of equal width "up" and "down" even if there is no phase frequency detected between two inputs, however a small jitter is produced [150].

However, the solution proposed in this thesis is the inclusion of an interpolation module at the entrance of the PFD. This new element acts only in the event of phase locked and when the two input signals cross the zero level at the same instant, that it is when the PFD fails to generate the start of the charge/discharge pulse. In this case, the input signals are interpolated using the new and previous values to determine which signal crosses first the zero line and the new points are substituted by the interpolated points around the zero level. This approach allows the detection of any phase difference between the two inputs in each rising edge of the cycle, even when the phase is locked as can be seen in fig. 5-6(a).

This technique completely suppresses the dead-zone, however it includes a jitter in the frequency since, in every cycle of the input signal, there is always a pulse (of at least one iteration period) of charge or discharge in the PFD. This effect is highly reduced by establishing a correction factor, which can be understood as a dual-gain structure, that will only act when the frequency is locked according to the function

$$r(i_{ref}, i_{out}) = \begin{cases} \max(0.01, d), & d < \epsilon \quad \& \quad Interp.active \\ 1, & \text{otherwise} \end{cases} \quad (5.11)$$

where $d = i_{ref} - i_{out}$, with i_{ref} and i_{out} being the PLL input and output respectively. The parameter ϵ establishes a threshold for the application of this correction factor that acts only when the interpolation is executed. The performance of this technique is presented in fig. 5-6(b).

5.3.1.3 Frequency band selector

The inclusion of the PLL in the disturbance rejection algorithm provides additional information that can be used to improve the performance of the system.

The actual disturbance frequency, which is known thanks to the PLL locking operation, can be used in a frequency band selector. This structure multiplies the final PLL amplitude by a factor 0 or 1 depending on whether the frequency is within the selected frequency band or not.

(a) Interpolation technique for phase difference detection. At the instant $t = 0.855s$, both the original setpoint and the PLL output cross the zero value causing the non-detection of the phase difference. Alternatively, the interpolated approach shifts these values around the zero level making that only one of them crosses the zero at $t = 0.855s$ and therefore triggering the charge (or discharge) PFD output as shown in 5-6(b).

(b) Comparison of PFD performance for original PFD, PFD with interpolation and PFD with interpolation and correcting factor.

Figure 5-6: PFD performance with the interpolation and correcting factor.

In order to avoid a discontinuity in the amplitude, the frequency band selector includes a sigmoid block which is also used in other applications [122]. It allows the gradual transition between the values 0 and 1 around the cut-off frequencies and forms a frequency band selector for the disturbance cancellation action.

It is important to note that out of the selected frequency band, the observer and PLL continue working normally. However, the estimated disturbance reinjected into the system is asymptotically zero (due to the frequency band selector with the sigmoid function), acting as a disconnection of the disturbance cancellation.

5.3.1.4 SNR threshold module

In order to avoid the use of the PLL output as disturbance cancellation when the PLL has not locked or the observer has not converged, the inclusion of a signal-to-noise (SNR) threshold module is proposed. This element, connected after the amplitude detector block, calculates the signal to noise ratio of the estimated amplitude and transform this value into a factor from 0 to 1 (using again a sigmoid function again) which multiplies the amplitude, as can be seen in Fig.5-7. The number of points in the calculus of the SNR should be chosen to include several periods of the estimated disturbance.

The amplitude detector and PLL work continuously, however this elements allows the application of the disturbance cancellation only when the SNR of the amplitude is above a fixed threshold.

5.3.2 Proposed method 2: Phase Locked Loop Disturbance Rejection Algorithm with Observer Update

As mentioned previously, in real applications the selected frequency to be cancelled might not correspond to the exact disturbance frequency, leading to a reduced disturbance cancellation performance of the direct observer or *PLLDR* approach.

The final proposal of this work is the update of the extended observer with the actual disturbance frequency obtained in the PLL as can be seen in the upper switch

Figure 5-7: Signal-to-noise limiter diagram.

Figure 5-8: Block diagram of the *PLDR with observer update*.

of Fig.5-8. This approach allows the system to adapt to a disturbance that could drift in frequency over time, leading to a worse disturbance cancellation performance.

The proposed *PLDR with observer update* approach presents clear advantages with respect to the previous approaches and can lead to an improved disturbance cancellation due to its ability to identify the frequency of the disturbance and update the extended observer accordingly.

The feedback of the frequency produces, however, an additional closed-loop with possible stability issues. It is therefore only applicable when the frequency drift is

significantly slower than the dynamics of the system and the update of the frequency is performed at a much slower rate than the general control loops.

5.4 Simulation results

Simulations are performed to demonstrate the effectiveness of the proposed methods. The sensitivity functions of the different approaches are presented in 5.4.2 where the disturbance rejection performance is evaluated. A fixed 50 Hz, 50.1 Hz and variable 45-55 Hz disturbances are then simulated and the results are presented in 5.4.3.

The plant model and controller used in the simulations are the same that were identified and designed for the experimental rotational stage (5.13a)-(5.13e) presented in 5.5.1 and 5.5.2 (7th order plant model, a closed loop bandwidth of 22 Hz and 79° and 13dB phase and gain margins respectively and a 2kHz sampling rate). The input of the plant is a voltage signal and the output is an angle. The observer is implemented in state space form with the following matrices.

$$\mathbf{A}_{\text{obs}} = \mathbf{A} - \mathbf{K}\mathbf{C} \quad \mathbf{B}_{\text{obs}} = [\mathbf{B}, \mathbf{K}] \quad (5.12a)$$

$$\mathbf{C}_{\text{obs}} = \text{diag}(\mathbf{C}_{\text{zs}}, \mathbf{C}_{\text{zd}}) \quad \mathbf{D}_{\text{obs}} = \text{zeros}(2, 2) \quad (5.12b)$$

$$\hat{\mathbf{x}} = \begin{bmatrix} \hat{\mathbf{x}}_{\text{zs}} \\ \hat{\mathbf{x}}_{\text{zd}} \end{bmatrix} \quad \hat{\mathbf{y}} = \begin{bmatrix} \hat{y}_z \\ \hat{d}_{zi} \end{bmatrix} \quad \mathbf{u} = \begin{bmatrix} u \\ y_z \end{bmatrix} \quad (5.12c)$$

The observer gain $\mathbf{K}$ was calculated using Kalman theory (function `dlqe` in Matlab), with a high measurement noise ($\mathbf{R}$) to process noise ($\mathbf{Q}$) ratio, trusting in the predefined model. The values provided to the `dlqe` function were $R = 1$, $Q = 10^{-7}$ and $G = [1, 0, 0, 0, 0, 0, 0, 10, 0]^T$, parameters that were tuned manually to obtain acceptable convergence time of the observer. The parameters used in the simulation correspond to the values implemented in the experimental case of section 5.5.

The PLL used in the simulations is the same as the one implemented experimen-

tally in section 5.5. It is a third-order charge-pump PLL designed with a bandwidth of 10 Hz, phase margin of 60° , quiescent frequency equal to the selected frequency to be cancelled and including a frequency band of ± 3 Hz (allowing the cancellation in this range around the quiescent frequency) and a SNR threshold of 4 (with a moving window of 1600 points). In the *PLDDR with observer update* approach, the observer is updated at 1 Hz.

The simulations are performed in MATLAB[©] (Simulink) with double precision.

5.4.1 PLL design for *PLDDR* algorithm

The PLL is designed making use of the equations 5.8, 5.9 and 5.10 for a fixed phase margin of 60 degrees. The results presented in Fig.5-9 show the response of the PLL to a step signal of 1Hz for different bandwidths.

It can be seen how the increase in the bandwidth of the PLL leads to a faster settling time. However, once the PLL is locked the jitter in the frequency is also greater for higher bandwidths. The solutions proposed in 5.3.1.2 allow the design of the PLL with a high bandwidth (10Hz) with the absence of frequency jitter as it can be seen in the last response of fig.5-9. This PLL is used in the following simulations and experimental results.

5.4.2 Sensitivity function comparison

The disturbance rejection performance of each approach is well described by its sensitivity function in closed loop. Figure 5-10 presents the simulation of the sensitivity function of each approach for a 50 Hz cancellation.

In these results, as well as in sections 5.4 and 5.5, the term “Controller” in the figure legend refers to the nominal feedback controller with no additional disturbance rejection technique.

For the cases of the feedback controller only and the direct observer, the system is linear and the sensitivity functions are calculated in a conventional way. However, the inclusion of the PLL in the *PLDDR* and *PLDDR with observer update* approaches

Figure 5-9: PLL Performance comparison including the final $w_{ugb} = 10\text{Hz}$ with Interpolation and Dual gain. The input signal is a step signal passing from 50Hz to 60Hz at $t = 60\text{s}$.

makes the system nonlinear and an analytical expression can no longer be derived. In these cases, time domain simulations with a combined input of a low-power white noise and a single frequency component (SFE-WN), which varies at each simulation over the whole frequency range, are performed to evaluate the sensitivity functions. The disturbance observer is designed to be highly sensitive to the preselected frequency compared to other frequencies. If the applied disturbance frequency is different to the preselected one, the observer would estimate a very low amplitude disturbance but at this test frequency. In this scenario, a free PLL would then lock to this test frequency and the loop would be closed yielding the same water-bed effect as the direct observer approach. In reality there is always background white noise present. For the illustration of the proposed algorithm's effectiveness, it is therefore more interesting to include this low power white noise disturbance in addition to the single frequency component under test. In this case, the white noise is enough to maintain the observer and PLL focused on the preselected frequency to be cancelled, and not on the frequency under evaluation. This leads to the sensitivity response of the *PLDDR*

in Fig.5-10 that shows the decoupling effect of the PLL. It is important to note that if the PLL is limited with the frequency band selector the same result is achieved even without the presence of white noise.

The density of evaluated frequencies is increased at the most interesting regions, like the resonance frequency of the system and the selected frequency to be rejected.

The zoom at the selected frequency to be rejected, Fig.5-10(b), clearly shows how the direct observer and *PLLDR* approaches present a peak attenuation at this frequency. This attenuation is of the same magnitude in both cases and it is not present with the original feedback controller. As expected for the *PLLDR with observer update* approach, it does not present an attenuation peak but a flat-bottom zone around the selected frequency which attenuates the disturbance in the selected frequency band (± 3 Hz in this case). It is important to note that the flat-bottom zone of the *PLLDR with observer update* approach does not mean that all the frequencies in this region will be attenuated at this level. Only a single (or main) disturbance frequency moving in this region will be attenuated according these values, so it could be understood as a single peak attenuation that is able to move in this flat-bottom region.

Looking at the main resonance frequency of the system, Fig.5-10(c), it can be seen how the *PLLDR* and *PLLDR with observer update* approaches maintain the sensitivity function unaltered with respect to the original controller. It presents a clear improvement with respect to the direct observer approach whose attenuation at this frequency is reduced due to the water-bed effect caused by the increase of attenuation at the selected frequency.

5.4.3 Disturbance rejection performance at 50Hz

The disturbance rejection performance of the different approaches are benchmarked in simulation for the case of a single 50 Hz disturbance. A second simulation is presented where the selected frequency is 50 Hz but the real disturbance is slightly different at 50.1 Hz. Finally the case when the disturbance frequency moves from 45 Hz to 55 Hz in steps of 2.5 Hz is simulated. The standard deviation of the angular

(a) Sensitivity functions of the different approaches.

(b) Zoom at the selected frequency to be cancelled (50Hz).

(c) Attenuation at the resonance frequency of the system showing the lower attenuation of the direct observer.

Figure 5-10: Sensitivity functions for 50 Hz cancellation. The *PLLDR* and *PLLDR with observer update* were simulated with a frequency band selector of ± 3 Hz for the SFE, and with no frequency band selector in the low power white noise (SFE WN - Power 10^{-3}) case.

output shown in the results is calculated using a moving window of 1s.

Figure 5-11(a) presents the results for the simulation with a single disturbance, 50 Hz, and an amplitude of $\pm 3\mu\text{rad}$, that is injected at the system output. The results show a clear improvement with respect to the original controller and present similar attenuation between the different approaches.

In order to illustrate the beat-effect problem of the *FDRSS* approach, the previous simulation is repeated with the injected disturbance at 50.1 Hz instead of 50 Hz. As expected, the Fig.5-11(b) shows the undesired beat-effect in the *FDRSS* approach and how the other techniques successfully attenuate the disturbance. The improved performance of the *PLLDR with observer update*, that is able to identify the correct disturbance frequency and update the observer accordingly, is apparent.

Finally, the case of a moving disturbance from 45 Hz to 55 Hz in steps of 2.5 Hz is simulated and presented in Fig.5-11(c). It can be concluded that the attenuation

of the PLL with observer update is clearly superior to the rest of the approaches due to its ability to identify and adapt to the actual disturbance. The frequency band selector in the *PLLDR with observer update* case operates correctly inside the ± 3 Hz band, whereas there is no difference between this approach and the standard feedback controller out of this band.

5.5 Experimental validation

The different control approaches are experimentally evaluated on a high precision piezo-actuated rotational stage used for Crystal Collimation [126] in the Large Hadron Collider (LHC) at the European Organization for Nuclear Research (CERN).

5.5.1 Rotational Stage

The rotational stage used in the experimental evaluation is shown in Figure 5-12(a). It is composed of a pre-stressed flexure-hinge structure actuated by a piezoelectric stack. The flexure-hinge based structure avoids sliding parts and thereby enhances precision by reducing the number of non-linear effects (e.g. backlash and friction). A piezoelectric actuator (PEA) is used to generate the rotational movement by interacting on an amplifying lever that is attached to the rotational head a few millimetres from the centre of rotation (marked as a "X" in the picture). This amplifying structure gives the stage a rotational range of 20 mrad from a nominal linear range of 130 μm . This combination results in a typical resonant structure common to most piezoactuated systems. Different control approaches has been proposed in the recent years searching for a high bandwidth and resonance damping of the structure [26, 66, 92, 149].

The angular position is measured by a dual-axis interferometer system described in [131]. It comprises two optic fibre based interferometer axes pointing toward a reflective surface that is fixed to the rotational head as shown in Fig.5-12(b).

(a) Simulation of 50 Hz disturbance. Activation and deactivation of the disturbance cancellation at $t = 10s$ and $t = 70s$ respectively.

(b) Simulation of 50.1 Hz disturbance maintaining the selected frequency to be cancelled at 50 Hz. Activation and deactivation of the disturbance cancellation at $t = 10s$ and $t = 70s$ respectively.

(c) Simulation of disturbance frequency drift from 45 Hz to 55 Hz in steps of 2.5 Hz at $t = 20s$, $t = 40s$, $t = 60s$ and $t = 80s$.

Figure 5-11: Simulation of the disturbance rejection performance for 50 Hz cancellation with different methodologies when varying the applied disturbance frequency.

5.5.2 Controller Setup

The controller of the system implemented in this work is a 2-DOF structure [26] composed of a feedback controller block $C(z)$ and a prefilter $F(z)$, Fig.5-13. The linearised rotational stage is represented as $G(z) = H^{-1}(z)G_0(z)$. The linear and non-linear dynamics of the system are modelled separately using a Hammerstein structure, shown to be applicable in [27]. To compensate for the non-linear dynamics a Maxwell-Slip Hysteresis Compensator ($H^{-1}(z)$) with 10 elements is designed [17]. The linear dynamics of the system is identified through a black box model with output error structure, with 7 poles and 6 zeros. The system is excited with a pseudo random binary sequence (PRBS) and the response is used for the system identification in the z-domain.

The controller block $C(z)$ is a series combination of a 2-pole controller, a notch filter and a lead filter, which stabilizes the system, increases the phase margin and makes the system more robust to high frequency modes respectively. Finally, to enhance the tracking performance, a prefilter $F(z)$ can be added to the system. The system model and controller parameters are presented in equations (5.13a)-(5.13e)

Figure 5-12: Experimental setup. (a) Piezo-actuated rotational stage used in crystal collimation. The crystal is mounted below the stage, attached to the rotational head. (b) Angular measurement system formed by two interferometer axes and a mirror that is mounted on the rotational head. (c) Piezoelectric shaker to generate external disturbances.

forming a closed loop system with a 22 Hz bandwidth and 79° and 13dB phase and gain margins respectively that runs at 2 kHz (control loop and sampling frequency).

Figure 5-13: Block diagram of the present control loop, including controller, prefilter and hysteresis compensator.

$$G(z) = \frac{\begin{pmatrix} 0.02638z^{-1} + 0.04911z^{-2} + 0.02637z^{-3} \\ + 0.02132z^{-4} + 0.03163z^{-5} + 0.02z^{-6} \end{pmatrix}}{\begin{pmatrix} 1 - 1.075z^{-1} - 0.6308z^{-2} + 1.437z^{-3} \\ - 0.7495z^{-4} - 0.3769z^{-5} + 0.6524z^{-6} \\ - 0.08247z^{-7} \end{pmatrix}} \quad (5.13a)$$

$$F(z) = 1, \quad C_c(z) = \frac{1.546 - 3.044z^{-1} + 1.543z^{-2}}{1 - 1.390z^{-1} + 0.390z^{-2}} \quad (5.13b)$$

$$C_{lead}(z) = \frac{2.521 - 1.966z^{-1}}{1 - 0.445z^{-1}} \quad (5.13c)$$

$$C_{notch}(z) = \frac{\begin{pmatrix} 0.475 + 0.512z^{-1} + 0.407z^{-2} \\ + 0.533z^{-3} + 0.360z^{-4} \end{pmatrix}}{\begin{pmatrix} 1 + 0.290z^{-1} + 0.473z^{-2} \\ + 0.397z^{-3} + 0.128z^{-4} \end{pmatrix}} \quad (5.13d)$$

$$C(z) = C_c(z)C_{lead}(z)C_{notch}(z) \quad (5.13e)$$

The resulting closed loop complementary sensitivity and sensitivity functions are shown in Figure 5-14.

The control loop is implemented on a National Instruments PXI with a real-time operating system. The PXI applies a control voltage between 0 and 7 V which is

Figure 5-14: Bode diagram of the resulting closed loop complementary sensitivity and sensitivity functions.

amplified by a linear amplifier with a gain of 20.

5.5.3 Experimental Results

All approaches were benchmarked experimentally with single-frequency rejection of a real environmental 50 Hz disturbance (identified as the largest disturbance in the operational environment) and two artificial disturbances of 90 and 250 Hz added to the system by a shaker. The shaker, shown in Fig.5-12(c), consists of a piezoelectric actuator mounted in a pre-stressed structure with a mass attached. It is clamped to the testbench base, transmitting its vibration to the rotational stage. A signal generator is configured to produce a sinusoidal voltage at the desired frequency. The signal is then amplified and applied to the shaker. When generating the 90 and 250 Hz disturbances, the amplified voltage applied to the shaker had an amplitude of 0.6V and 0.2V respectively.

All acquisitions were taken consecutively with one cancellation methodology work-

ing at a time. The delay block in the *FDRSS* was configured with 40, 400 and 8 delay points to include an integer number of cycles of the 50 Hz, 90 Hz and 250 Hz disturbance cases respectively. The results for the observer update correspond to the case when the frequency of the disturbance model is updated every 1s.

5.5.3.1 Disturbance cancellation performance

The frequencies of 50 Hz, 90 Hz and 250 Hz were selected in order to evaluate the performance of the different approaches when rejecting frequencies well above the original controller bandwidth (22 Hz) and at frequencies that are lower and higher than the resonance frequency of the system (67 Hz).

The cancellation performance in closed loop operation for the different methodologies, including the direct feedback connection of the observer, are presented in Fig.5-15. These results correspond to the values of a moving standard deviation (1s time window) of the angle when cancelling a 50, 90 and 250 Hz disturbance component, respectively. In the three cases, the *FDRSS* suffers from the beat effect while the rest of the methods manage to compensate for it. In these cases, the directly connected observer achieves similar performance to approaches including a PLL despite its effect on the closed loop characteristics. It is important to note also that the cancellation of the 250 Hz component is still efficient even though the disturbance was in a region far outside the closed loop bandwidth.

The attenuation over the whole cancellation period is summarized in Table5.1, computed as the percentage of the angular standard deviation relative to the standard controller.

All the approaches, with the exception of the *FDRSS*, present significant attenuation of the disturbance with similar results between them. In this case of a known disturbance frequency, the *PLLDR Observer Update* performs slightly worse than the *PLLDR*, which might be due to the combination of numerical effects in the estimation of the frequency, higher noise sensitivity or plant model errors.

Additionally, in the simulation presented in Fig.5-11(a), a single 50 Hz disturbance was applied to the system, whereas the experimental results in Fig.5-15(a)

correspond to real background noise whose main disturbance, but not unique, correspond to the 50 Hz component. For this reason, a direct quantitative comparison is not possible but the experimental results confirm the expected attenuating behaviour and illustrate the problem of the *FDRSS*, whose response in the real example of Fig.5-15(a) is very similar to the one obtained in Fig.5-11(b).

(a) Cancellation of 50Hz real background disturbance. The results for *PLDR* and *PLDR Observer update* nearly overlap during the activated period.

(b) Cancellation of 90Hz disturbance generated with shaker.

(c) Cancellation of 250Hz disturbance generated with shaker.

Figure 5-15: Disturbance rejection performance with the different methodologies, showing the activation ($t = 10$ s) and the deactivation ($t = 70$ s).

Disturbance freq. (Hz):	50	90	250
Approach	<i>% with respect to standard controller</i>		
Direct Observer	46.07	54.55	67.39
<i>FDRSS*</i>	170.40	112.75	81.62
<i>PLLDR</i>	36.43	50.48	66.75
<i>PLLDR Obs Update</i>	39.02	55.75	67.61

Table 5.1: Disturbance rejection performance with respect to the standard controller. Performance in percentage (of the standard deviation) with respect to the standard controller. *Note that the values in this row display an average of the cancellation response i.e. it is not showing the maximum standard deviation that occurs during constructive interference.

5.5.3.2 Drifting disturbance cancellation

In addition, a similar test was performed in order to evaluate the robustness of the different approaches against frequency drift. Figures 5-16, 5-17 and 5-18 show the cancellation responses and the performance, summarised in the Tables 5.2, 5.3 and 5.4, for the cases of a disturbance drifting in the range 45-55 Hz, 85-95 Hz and 245-255 Hz in steps of 2.5 Hz, with selected frequencies to be cancelled of 50 Hz, 90 Hz and

250 Hz respectively.

In this case, all the disturbances were generated artificially with the shaker with the same amplified voltage amplitudes for the 90 Hz and 250 Hz cases, and 6V for the 50 Hz case. These amplitudes were selected to produce a similar effect on the system between them and to be in line with the angular standard deviation experienced in the LHC application ($<3\mu\text{rad}$).

The results show how the *PLLDR Observer Update* performs best due to its ability to auto-tune the observer frequency. At the initial and final frequencies of these drift tests, only the *PLL Observer update - No freq limit nor SNR* presents a significant attenuation. The results for the other approaches are very close to each other and in some cases overlap.

The experimental results in Fig.5-16 confirm the attenuating behaviour predicted for the different approaches in the simulation of Fig.5-11(c). Unlike in the simulation case, the use of the shaker as a disturbance source implies that, even if the amplitude of the sinusoidal voltage is constant, the response at each frequency can be different and therefore the applied disturbance is not constant during the frequency drift.

5.5.3.3 Mismatch between selected frequency and actual disturbance

The case when the selected frequency does not match the actual disturbance is of special interest. In this case, the direct observer approach might present a clear performance degradation with respect to the standard controller due to the water-bed effect. This degradation would be particularly visible when the actual disturbance frequency is close to that of the system resonance as was concluded looking Fig.5-10(c).

The mismatch between selected and actual disturbance frequencies is experimentally evaluated by selecting a cancellation frequency of 50 Hz but applying a 70 Hz disturbance. As expected, the results in Fig.5-19 show the degradation in the direct observer case, while the *PLLDR* and the *PLLDR with observer update* (both cases with frequency limit of ± 3 Hz) performance remains very close to the conventional controller.

Figure 5-16: Disturbance rejection performance for disturbance moving in frequency from 45Hz to 55Hz with steps of 2.5Hz at $t = 20s$, $t = 40s$, $t = 60s$ and $t = 80s$.

Disturbance freq. (Hz):	45	47.5	50	52.5	55Hz
Approach	% with respect to standard controller				
Direct Observer	94.7	91.8	48.0	90.9	98.4
PLLDR	99.5	95.4	47.8	93.7	100.1
PLLDR Obs.Up. ¹	93.6	71.7	47.3	33.3	100.8
PLLDR Obs.Up. ²	71.5	76.3	50.3	36.2	28.4

Table 5.2: Disturbance rejection performance for disturbance moving in frequency from 45Hz to 55Hz. Performance in percentage (of the standard deviation) with respect to the standard controller. ¹. Frequency limit = 3Hz, SNR=4. ².No frequency limit nor SNR.

5.5.3.4 Stability during rotational movement

Finally, an experimental evaluation of the cancellation stability during an angular movement is shown in Fig.5-20. The disturbance applied in this case is at 50 Hz and the results show how the disturbance rejection performance is not altered by the positive and negative angular movement at $t = 30s$ and $t = 50s$ in any of the control approaches.

Figure 5-17: Disturbance rejection performance for disturbance moving in frequency from 85Hz to 95Hz with steps of 2.5Hz at $t = 20s$, $t = 40s$, $t = 60s$ and $t = 80s$.

Disturbance freq. (Hz):	85	87.5	90	92.5	95
Approach	<i>% with respect to standard controller</i>				
Direct Observer	90.7	80.1	34.7	107.1	104.9
<i>PLLDR</i>	101.7	84.5	24.5	107.6	102.0
<i>PLLDR Obs.Up.</i> ¹	101.5	26.4	31.7	36.2	102.0
<i>PLLDR Obs.Up.</i> ²	20.8	26.4	31.8	36.2	37.2

Table 5.3: Disturbance rejection performance for disturbance moving in frequency from 85Hz to 95Hz. Performance in percentage (of the standard deviation) with respect to the standard controller. ¹. Frequency limit = 3Hz, SNR=4. ².No frequency limit nor SNR.

In the majority of these experimental results, the applied disturbance was generated artificially and therefore the frequency to be cancelled was known in advance. In practice, this frequency is not known and a Fourier Transform of the output should be performed to assess the principal disturbance frequency, that will be selected for attenuation, as was done for the 50 Hz case in Fig.5-15(a).

Figure 5-18: Disturbance rejection performance for disturbance moving in frequency from 245Hz to 255Hz with steps of 2.5Hz. Standard deviation with disturbance moving in frequency from 245 Hz to 255 Hz in steps of 2.5 Hz at $t = 20s$, $t = 40s$, $t = 60s$ and $t = 80s$.

Disturbance freq. (Hz):	245	247.5	250	252.5	255
Approach	% with respect to standard controller				
Direct Observer	91.6	83.4	52.5	110.9	105.6
PLLDR	98.7	83.7	51.6	109.2	98.7
PLLDR Obs.Up. ¹	98.0	51.0	51.7	50.8	98.3
PLLDR Obs.Up. ²	50.5	51.0	51.5	50.8	51.8

Table 5.4: Disturbance rejection performance for disturbance moving in frequency from 245Hz to 255Hz with steps of 2.5Hz. Performance in percentage (of the standard deviation) with respect to the standard controller. ¹. Frequency limit = 3Hz, SNR=4. ².No frequency limit nor SNR.

5.6 Application to LHC Piezo Goniometers

As it was commented in the previous chapter, currently there are four piezo goniometers in LHC and another one in SPS accelerator. The study of the different disturbance rejection alternatives and the development of the technique proposed in this chapter was motivated in the continuous aim to improve the overall performance

Figure 5-19: Mismatch between the selected frequency to be cancelled (50 Hz) and the actual disturbance (70 Hz) applied to the system. Activation of each disturbance rejection approach at $t = 10s$ and deactivation at $t = 70s$.

Figure 5-20: Cancellation performance for a 50Hz disturbance, showing the activation ($t = 10s$), a relative movement of $1000 \mu rad$ ($t = 30s$), a relative movement of $-1000 \mu rad$ ($t = 50s$) and the deactivation ($t = 70s$).

of the devices for Crystal Collimation.

Along the development of this disturbance rejection technique, other improvements have also been developed in parallel and implemented into the LHC piezo goniometer devices. Among these improvements it is important to remark the migration of closed loop controller from the Real-Time host to a FPGA Card and the installation of the new stepper motor drive with field oriented control technique [124], also developed at CERN, for the linear movement of the rotational stage.

From a mechanical point of view of the rotational stage, a new high-stiffness stage (V2) has been developed to increase significantly the first resonance frequency. This change has allowed the design of a higher bandwidth controller, increasing from 10Hz to 30Hz approx. the closed loop bandwidth.

The combination of these improvements have increased the overall tracking performance of the piezo goniometers at CERN accelerators, achieving acceptable standard deviation values of the angular position for the current requirements.

However, the future upgrades of CERN LHC will require improved performances also for the piezo goniometers. Since the disturbance rejection techniques proposed in this chapter can be used as an additional 'plug-in' module to the existing controller, the final implementation of this technique in the LHC and SPS could take place in future in order to reduce the effect of the main disturbances.

5.7 Conclusions

The limitations of the FDRSS technique, originally developed for hard disk drives, are illustrated when the external disturbances are independent of the system. In such cases, the de-phasing between the estimated and real disturbance will inevitably appear over time, resulting in an undesired oscillation known as beat-effect.

A new disturbance rejection approach including a phase locked loop is proposed in this chapter to avoid this problem and preserve the single-frequency disturbance attenuation performance while maintaining the closed-loop characteristics of the nominal controller as in the FDRSS. The switching structure of FDRSS is now substituted

by a PLL with amplitude detection that acts as decoupling element after the disturbance observer. The beat effect is therefore prevented, even for small differences between the preselected cancelling frequency and the real disturbance or when the frequency of the disturbance drifts slowly.

The use of the PLL allows the implementation of a frequency band selector and a signal-to-noise threshold module, which allow the automatic activation and deactivation of the disturbance rejection action.

Furthermore, this chapter proposes a new PLL Disturbance Rejection with observer update approach. In this case, the observer is updated with the disturbance frequency estimated in the PLL. This technique allows the attenuation of a disturbance that can vary in frequency over time and constitutes a great improvement with respect to previous techniques.

Both techniques can be considered as ‘plug-in’ algorithms that can be added to existing controllers to improve out-of-bandwidth single-frequency disturbances while maintaining the response of the original controller at other frequencies. Furthermore, this characteristic makes them easily scalable for multi-frequency cancellation by the addition in parallel of single-frequency blocks.

All the approaches are simulated and experimentally evaluated with real noise and observer model mismatch in a high-precision rotational stage, used for Crystal Collimation in the LHC at CERN, showing excellent attenuation performance.

The experimental results, in agreement with the simulations, prove the suppression of the beat effect in both of the proposed schemes with respect to the *FDRSS*. In the cases with a fixed disturbance frequency at 50Hz, 90Hz and 250Hz, the results show a clear reduction of the standard deviation to less than 40%, 56% and 68% of the standard controller performance. It is important to remark that all three selected frequencies are beyond the bandwidth of the standard controller.

Furthermore, the proposed PLL Disturbance Rejection with observer update correctly adapts to the disturbance frequency. The experimental results show an improved frequency range of attenuation (of ± 3 Hz in the studied case) around the pre-selected frequency, presenting a clear advantage with respect to the rest of the

techniques in the cases where the disturbance frequency varies or is not precisely known in advance. The final selection of the technique to be used should consider the performance improvement versus the computational cost of the algorithm implementation.

The future LHC upgrades will require higher requirements affecting the Crystal Collimation equipment. Therefore, the disturbance rejection technique presented in this chapter could be considered for its final implementation into the piezo goniometers installed at LHC and SPS at CERN in order to increase the tracking performance of the devices.

The proposed technique for disturbance rejection, together with the simulation and experimental results, was submitted for publication in the journal IEEE Transactions on Mechatronics. It is currently under revision it is expected to be published along 2019.

Chapter 6

General conclusions

In this thesis, the high precision piezo goniometer developed at CERN for Crystal Collimation has been presented, detailing the specific innovative contributions involved. The technological challenges posed by Crystal Collimation have been summarised as:

- High precision positioning requirements.
- Ultra high vacuum and high temperature compatibility.
- Radiation resistance.
- Need of an absolute angular reference.

The piezo goniometer with its different components and versions has been described in detail. The performance of the system has been studied and its limitations have been identified. They have served as the motivation and starting point for the three main contributions of the thesis:

1. A radiation resistance study of the piezoelectric actuator.
2. An angular measurement system with absolute angle recovery method.
3. A Phase Locked Loop disturbance rejection algorithm for frequencies beyond the bandwidth of the controller of the system.

Three different irradiation tests have been performed with the objective to understand the radiation effects on the piezoelectric actuator under the mixed field radiation environment of the LHC tunnel, where an accumulated dose of 10MGy is expected in the 10 year lifetime of the goniometer.

The results of the Gamma Irradiation Tests (online test up to 10 MGy and offline test up to 1MGy) have proven the resistance of the actuator to TID effects. All samples were fully operative after the irradiation period, with a final stroke never below 80% of its original value, a reduction that is taken into consideration in the design process. The impedance results showed an increase of less than 25% in all the 34 DUTs which, as well as the observed hysteresis variations, can be compensated by the closed loop control of the system.

A third Proton Irradiation Test (online test up to 2 MGy) was performed. The purpose of evaluating the DD effects on the actuator. The results have shown a stroke reduction of less than 20% and an impedance increase of less than 25%.

The suitability of this piezoelectric actuator for Crystal Collimation in the LHC is therefore confirmed. The radiation-induced effects, in terms of impedance increase and stroke variation, are acceptable for this specific application. The closed loop control of the system can compensate for the impedance increase or hysteresis variation, and the consequences of a potential stroke reduction are mitigated by considering this reduction in the device design.

In the field of precision angular measurements, a dual-interferometer-based system has been proposed and studied for two different configurations including a flat reflective surface and corner-cube retroreflectors. The relative measurement limitation of laser interferometry has been overcome with the proposal of a novel absolute angle recovery method.

The method is based on the use, as an absolute reference, of the characteristic curve of the optical system, exploiting the convexity of this curve and providing a solution that depends uniquely on the geometry of the system.

The performance of the method has been evaluated with Monte Carlo simulations studying the effects of noise and the number of points in the acquisition of the characteristic curve. The technique has been applied to a rotational stage used for Crystal Collimation and has shown an angular accuracy of $\pm 1.29 \mu rad$ (3σ) in the absolute angle recovery for this specific rotational stage.

The dual-interferometer-based angular measurement system and the proposed ab-

solute angle recovery method are successfully applied to the five piezo goniometers installed in the CERN LHC and SPS accelerators.

The last contribution of the thesis has been the proposal of a phase-locked loop disturbance rejection algorithm capable of attenuating single disturbances at frequencies beyond the bandwidth of the original controller. The PLL keeps the estimated and real disturbances in phase. The undesired beat effect is therefore suppressed, even for small differences between the preselected cancelling frequency and the real disturbance, or when the frequency of the disturbance slowly drifts.

The use of a PLL has opened other possibilities such as the integration of a frequency band selector or a signal-to-noise threshold module, which allows the automatic activation and deactivation of the disturbance rejection action.

In addition, a novel PLL Disturbance Rejection with observer update algorithm has been proposed. In this case, the observer is updated with the disturbance frequency estimated in the PLL. This technique allows the optimal attenuation of a disturbance that can vary in frequency over time and constitutes a great improvement with respect to the previous technique.

Both techniques can be considered as ‘plug-in’ algorithms that can be added to existing controllers to attenuate out-of-bandwidth single-frequency disturbances, without affecting the original controller response for the rest of the frequency spectrum.

The proposed techniques have been studied and benchmarked in simulation with different disturbance frequencies and time variations. They have been also experimentally applied in the lab to a rotational stage used for Crystal Collimation at CERN, and the results have shown significant attenuation of disturbances at frequencies beyond the controller bandwidth (standard deviation is reduced to the range of 32 – 75% with respect to the original controller for the evaluated cases).

Figure 6-1: Beam Loss Monitor versus crystal angle showing the channelling effect for the B2H piezo goniometer in the LHC with proton beams.

6.1 Piezo goniometers operational experience in the LHC

The piezo goniometers in the LHC were intensively used during Run II, with successful operation in more than 10 so-called machine development periods. The channelling angle was always found in all four goniometers and at the same angular position (except when the interferometer system had to be restarted, in which case the Absolute Angle Recovery Method presented in Chapter 4 was applied and the variation in the angular position was expected). The angle of the crystal was, in all cases, correctly controlled with a standard deviation below $1 \mu\text{rad}$ for static operation (no linear axis movement) allowing the clear measurement of the channelling angle, as the example shown in Figure 6-1.

An important milestone was achieved during the MD of 12/09/2018 when all four piezo goniometers in LHC were launched synchronously in a combined movement (rotational and linear axis) to maintain the CH angle during the RAMP-UP mode

Figure 6-2: B2-H ramp-up performance during MD 12/09/2018 with angular standard deviation lower than $1 \mu\text{rad}$ (Yaw angle acquired at 2KHz and STD evaluated with sliding window of 2000 samples).

of the accelerator. The standard deviation of the angle during the whole profile was less than $6 \mu\text{rad}$ in all devices, with the best performance in the latest goniometer version achieving $< 1 \mu\text{rad}$ as shown in Figure 6-2.

The excellent performance and reliability of the piezo goniometers allowed its first use for real collimation during physics production in September 2018. Until this moment, the crystal collimators were always used during machine development periods, but on this occasion the LHC operators proposed the long-operation use of the crystals for a High-Beta Test during physics production in the experiments. The results, from the physics point of view, were very promising and the acquired data

are under study for future publications.

As commented in section 1.4, one of the limitations of the standard collimation system was heavy ion particles, in which case the crystal collimation could theoretically provide higher cleaning efficiency. The last two months of run II was dedicated to heavy ion physics. During this period, the piezo goniometers were used intensively. They were kept ready for operation (closed loop) during 28 days (24/7) for MDs and physics production as any other operational collimator. The performance of the devices was as expected and the physics results are under analysis.

6.2 Future work

The complexity of the piezo goniometer system leads to several lines of work for future improvement of the system. Some of them have already started:

- A new piezo goniometer version with CCRs instead of the plane mirror allows the location of interferometric heads out of the UHV tank. However, the radiation effects on the optical components involved, such as darkening or a change in the refractive index of the viewport and CCRs, has to be studied. An irradiation test is under preparation and it will take place during the second quarter of 2019 at the Fraunhofer Institute.
- The tiles forming the piezoelectric actuator are connected in parallel, which implies that if any any of them are short-circuited, the whole actuator is unusable. This problem has never happened in the operational goniometers, but it has been experienced in the lab. Possible failure-mitigation techniques, compatible with the UHV requirements, should be studied.
- Accelerated ageing and temperature cycle tests should be performed on the piezoactuators to model and predict potential problems over the lifetime of the goniometers.
- Study the possibility to include a force sensor in the piezoelectric actuator for force-feedback control or estimate the force from the measured signals.

- Develop algorithms for optimal measurement fusion and error detection based on redundant measurements.
- Develop and implement a bump-less structure to allow the opening and closing of the feedback loop without abrupt changes in the angle.
- Explore other alternatives for a “home index” in the angular measurement system that are compatible with the LHC accelerator requirements. It would lead to a real absolute reference as an alternative to the AARM.
- Study the applicability of online-diagnostic techniques to monitor the potential degradation of the actuator to allow preventive actions if necessary.
- Study the automatic controller-tuning techniques that could be applied to the system.
- Develop a standard for the measurement of piezoelectric actuator blocking force.

6.3 Publications

The results of the thesis have been published, or in some cases are still in the reviewing process for its publication, in the following conferences and journals:

- **JRC journal publications:**

- Serrano Gálvez, P., Butcher, M. and Masi, A. (2019). A Dual-Interferometer-Based Angular Measurement System With Absolute Angle Recovery Method. *IEEE Transactions on Instrumentation and Measurement*, 68(3), 864-873. <https://doi.org/10.1109/TIM.2018.2857598>

- **Conference publication:**

- Serrano Gálvez, P., Butcher, M. and Masi, A. (2017). An Absolute Angle Recovery Method for Rotational Stages. *2017 IEEE International Instrumentation and Measurement Technology Conference (I2MTC) (pp. 1-6). Turin*. <https://doi.org/10.1109/I2MTC.2017.7969798>
- Serrano Gálvez, P., Solis Paiva, S., Guillermain, E., Butcher, M., Di Castro, M., Masi, A.. Proton Irradiation Test on a Piezoelectric Actuator. *Radiation and its effects on components and Systems* (Accepted for RADECS 16th-20th September, 2019).
- Sola Merino, J., Serrano Galvez, P., Butcher, M., Di Castro, M., Masi, A., Vibration reduction techniques for stepper motors at standstill in high precision applications. *Electrical Machines and Industrial Drives* (Accepted for IECON 14th-17th October, 2019).

• **Paper submitted but still under review (Journal):**

- Serrano Galvez, P., Ericson, N., Oliver Ramirez, J. A., Butcher, M. and Masi, A. (2019). A Phase-locked Loop Disturbance Rejection Algorithm with Observer Update with Application to a Piezo-actuated Rotational Stage. *IEEE Transactions on Mechatronics*.

Appendices

Appendix A

Radiation Tests

A.1 Piezoactuator Characterisation Testbench

The piezoactuator characterisation testbench was developed with the objective to carry out most of the measurements involved during the research and development activities of the piezo goniometer project. These activities include the acceptance tests on the actuator at reception from the manufacturer, the characterisation of the piezo stacks for the research activities such as the radiation tests or the various measurement before the installation in the final rotational stage.

The testbench consists of five main parts that can be described as:

- **Mechanical structure:** it is shown in Fig.A-1 and is form by the base, the rods, the low-friction blocks, the locks and the spring.
- **Motor assembly:** it includes a servo motor (SANMOTION R R2AA04010FCP00M6), a gear-head and a torque limiter that fits on top of the mechanical structure as shown in Fig.A-1.
- **Sensors and auxiliary devices:** the system is prepared to work with up to two capacitive sensors (Fogale MCC-10), accelerometers (PCB Piezotronics 356B18) and a load cell (Althen SG-KP-12E-420). Each sensor needs its corresponding electronic equipment for its correct operation. The auxiliary equipment for the capacitive sensors is shown in A-1.

- **PXI chassis and modules:** A NI PXIe-8133 controller with a NI LabVIEW Real-Time operating system host the software developed at CERN to operate the different components of the testbench and acquire the reading of the sensors. Apart from the main controller, the PXI chassis includes a series of cards to increase the versatility of the testbench:

- PXIe-4357 + TB-4357: for PT100 Temperature sensor.
- PXIe-4496: for accelerometer sensors.
- PXI-4461: for load cell and current sensors.
- PXI-4070: for DMM multimeter.
- PXIe-4330 + TB-4330: for strain gauge sensor.
- PXI-7841R + SCB-68: FPGA module for motor control.
- PXI-4461: Capacitive sensor, piezo excitation signal (generation and feedback acquisition).

This complete setup offers the possibility of more than seven different tests combining the various actuators and sensors. The motor allows the variation of the load applied on the piezoactuator in static (constant load) or dynamic mode (load on the piezoactuator varying in time).

A.1.1 Stroke Test Measurements

The testbench presented in A.1 is powered at least 1 hour before the start of the measurements. During this time, the control temperature of the lab is set to 25°C and the DUTs are left out of any closed container for their temperature stabilisation.

In the Stroke Test, the springs, load cell and the motor assembly are removed from the mechanical structure. The expansion of the DUT produces the lift of the block with two capacitive sensors that measure the distance with respect to the base holding the DUT. After the calibration of the capacitive sensors for the length of the DUTs, each sample is measured according to the following three main steps:

Figure A-1: Mechanical structure and motor assembly of the piezoactuator characterisation testbench.

Figure A-2: Peripheral devices- National Instruments PXI chassis and modules (bottom-left), motor servo amplifier and mechanical brake power supply (top-left) and stack of linear amplifier for piezo actuators and capacitive measurement modules (right).

1. Full range test (0-150V at 1 Hz - 10 cycles), to verify that the voltage is being applied correctly and the measurement is as expected.
2. Full range test (0-150V at 1 Hz - 100 cycles), to put the DUT always in the same initial state.
3. Full range test (0-150V at 1 Hz - 10 cycles).

The last step is repeated a total of 5 times, which forms the raw data that are combined for the results of the test. In each repetition, the DUT is repositioned in the testbench.

A.1.2 Uncertainty estimation

A.1.3 Stroke Change

The two main uncertainty sources affecting the experimental testbench are:

- Repeatability uncertainty: The positioning of the piezo stack in the centre of the test bench is a manual process. The misalignment of the sample with respect to the vertical axis of the test bench introduces an uncertainty that was experimentally evaluated in a “Repeatability Test”, where a single piezo stack was measured 30 times re-positioning the sample in the centre of the test bench for each measurement. The results showed a repeatability uncertainty of:

$$u_{rep} = 0.35\mu m \quad (\text{A.1})$$

- Capacitive sensor misalignment: The alignment of the capacitive sensor with the base of the test bench can introduce non-negligible variations in the value of the measured stroke. The uncertainty introduced by this misalignment was studied in the “Effect of the Capacitive Sensor Alignment in the Stroke Test” document with the final result of:

$$u_{cap} = 0.75\mu m \quad (\text{A.2})$$

The final uncertainty for a stroke measurement is the combination of the two previous uncertainties:

$$u_{stroke} = \sqrt{u_{rep}^2 + u_{cap}^2} = 0.828\mu m \quad (A.3)$$

In practice, the stroke test is always repeated 5 times in order to statistically reduce this uncertainty. The uncertainty of the mean with n-measurements is:

$$u_{stroke_mean} = \frac{u_{stroke}}{\sqrt{n}} \quad (A.4)$$

The uncertainty in the stroke change can be computed according to the uncertainty propagation law as:

$$u_{stroke_diff} = \sqrt{u_{stroke_mean}^2 + u_{stroke_mean}^2} \quad (A.5)$$

A.1.4 Stroke Change in %

The stroke change in percentage was calculated according to:

$$stroke_diff_ \% = \left(\frac{r_f}{r_i} - 1 \right) * 100 \quad (A.6)$$

The corresponding uncertainty can be calculated using the propagation errors theory:

$$u_{stroke_diff_ \%} = \sqrt{\left(\frac{100}{r_i} \right)^2 * u_{stroke_mean}^2 + \left(\frac{r_f * 100}{r_i^2} \right)^2 * u_{stroke_mean}^2} \quad (A.7)$$

$$= \sqrt{\left(\frac{1}{r_i} \right)^2 + \left(\frac{r_f}{r_i^2} \right)^2} * 100 * u_{stroke_mean} \quad (A.8)$$

A.1.5 Hysteresis change in %

According to the “Repeatability Test” evaluation it can be seen that

$$u_{hyst_ \%} = 0.66\% \quad (\text{A.9})$$

The final value for hysteresis percentage is calculated taking the mean of n-measurements. The uncertainty of the mean can be calculated as

$$u_{hyst_ \%_mean} = \frac{u_{hyst_ \%}}{\sqrt{n}} \quad (\text{A.10})$$

For the case of the hysteresis change in %, the uncertainty is calculated as

$$u_{hyst_ \%_diff} = \sqrt{u_{hyst_ \%_mean}^2 + u_{hyst_ \%_mean}^2} \quad (\text{A.11})$$

A.1.6 Voltage change for maximum hysteresis %

According to the “Repeatability Test”, it can be seen that

$$u_{hyst_ V\%} = 0.27V \quad (\text{A.12})$$

If the final voltage value is calculated taking the mean of n-measurements, the uncertainty in the mean correspond to

$$u_{hyst_ V\%_mean} = \frac{u_{hyst_ V\%}}{\sqrt{n}} \quad (\text{A.13})$$

The uncertainty of the voltage difference is estimated as

$$u_{hyst_ V\%_diff} = \sqrt{u_{hyst_ V\%_mean}^2 + u_{hyst_ V\%_mean}^2} \quad (\text{A.14})$$

A.1.7 Area change

According to the “Repeatability Test” it can be seen that

$$u_{area} = 26.5\mu m * V \quad (A.15)$$

When the area is calculated as the mean of n-measurements, the uncertainty in the final value correspond to

$$u_{area_mean} = \frac{u_{area_mean}}{\sqrt{n}} \quad (A.16)$$

$$u_{area_diff} = \sqrt{u_{area_mean}^2 + u_{area_mean}^2} \quad (A.17)$$

A.2 Impedance Test uncertainty evaluation

The uncertainty of the impedance analyser Cypher C60 and the agreement between the two units available at CERN (one unit was sent to Fraunhofer while the second one stayed at CERN during the offline gamma irradiation test) was investigated in the internal report “Impedance Analyser Comparison”.

In this test, a single piezo sample was measured 20 times alternating the Cypher C60 unit. The results showed a good agreement between both units and a repeatability uncertainty in the impedance measurements of:

$$u_{z_rep} = 0.25\Omega m \quad (A.18)$$

The previous value was used to estimate the corresponding uncertainty for impedance difference in %:

$$Z_{diff_ \%} = \left(\frac{|Z_f|}{|Z_i|} - 1 \right) * 100 \quad (A.19)$$

Figure A-3: Stroke test analysis - DUT A1.

$$u_{z_diff_ \%} = \sqrt{\left(\frac{100}{Z_i}\right)^2 * u_{z_rep}^2 + \left(\frac{z_f * 100}{Z_i^2}\right)^2 * u_{z_rep}^2} = \quad (A.20)$$

$$= \sqrt{\left(\frac{1}{Z_i}\right)^2 + \left(\frac{Z_f}{Z_i^2}\right)^2} * 100 * u_{z_rep} \quad (A.21)$$

A.3 Gamma Irradiation Test 10MGy - Offline Test

A.3.1 Stroke test

The stroke test analysis for the DUTs A1-A17 and B1-B17 are presented in Figures A-3-A-15 and A-16-A-28 respectively.

A.4 Gamma Irradiation Test 1 MGy Active Test

A.4.1 Equipment

The instrumentation, components and cables involved in the Active Test are:

Figure A-4: Stroke test analysis - DUT A2.

Figure A-5: Stroke test analysis - DUT A3.

Figure A-6: Stroke test analysis - DUT A5.

Figure A-7: Stroke test analysis - DUT A6.

Figure A-8: Stroke test analysis - DUT A7.

Figure A-9: Stroke test analysis - DUT A9.

Figure A-10: Stroke test analysis - DUT A10.

Figure A-11: Stroke test analysis - DUT A11.

Figure A-12: Stroke test analysis - DUT A12.

Figure A-13: Stroke test analysis - DUT A13.

Figure A-14: Stroke test analysis - DUT A16.

Figure A-15: Stroke test analysis - DUT A17.

Figure A-16: Stroke test analysis - DUT B1.

Figure A-17: Stroke test analysis - DUT B2.

Figure A-18: Stroke test analysis - DUT B3.

Figure A-19: Stroke test analysis - DUT B5.

Figure A-20: Stroke test analysis - DUT B6.

Figure A-21: Stroke test analysis - DUT B7.

Figure A-22: Stroke test analysis - DUT B9.

Figure A-23: Stroke test analysis - DUT B10.

Figure A-24: Stroke test analysis - DUT B11.

Figure A-25: Stroke test analysis - DUT B12.

Figure A-26: Stroke test analysis - DUT B13.

Figure A-27: Stroke test analysis - DUT B16.

Figure A-28: Stroke test analysis - DUT B17.

1. NI PXIe-1078: PXI with Windows 7. It will execute the LabView software developed at CERN in charge of performing the different measurements continuously. The NI PXIe-1078 contained the following cards:
 - Network connection for remote access.
 - NI PXI-4070: For the leakage current measurement.
 - NI PXI-4461: For the signal generation and the signal acquisition.
2. Agilent 34970A Data Acquisition/Switch Unit. It will implement the analogue multiplexer and perform the temperature and voltage measurements, with the help of 2 different modules:
 - Agilent 34901A 20-Channel General Purpose Multiplexer: for the temperature and voltage measurements.
 - Agilent 34904A 4x8 Two-wire Matrix Switch (x2): for the analogue multiplexer implementation.
3. Cypher Instruments C60 impedance analyser: will be used to perform the online impedance analysis on the piezo actuators.

4. Cable NER48: High radiation resistant cable with the following characteristics:

- 24 twisted pairs of 0.5mm^2 copper wires.
- High radiation resistance.
- 17 mm max diameter.
- Bending radius: minimum 10 x diameter: 170 mm.

This cable is used to connect all the elements under radiation, that is the piezo actuators and the temperature sensors, to the above instrumentation in the radiation free area. The connection diagram is illustrated in Figure ??.

5. Piezo stack actuators and mechanical supports distributed in the different areas as:

- 4 samples in the radiation area (2x Type-A and 2x Type-B).
- 2 control samples (1x Type-A and 1x Type-B) in the low radiation area
- 2 control samples (1x Type-A and 1x Type-B) in the radiation free and environmentally stable area, together with the instrumentation.

All samples were provided by CERN already mounted in their mechanical supports.

6. Temperature sensors: A PT100 was attached to each piezo actuator sample in order to register its temperature. Additionally a temperature sensor were installed near the installation to measure ambient temperature changes.

A.4.2 Temperature Conditions

The requested temperature conditions for the Active Test can be summarised in the following points:

- The mean air temperature, as measured by two PT100 sensors installed by INT and CERN in the irradiation room TK1000B near the devices under test

(DUT), will be in the range of $20^{\circ}C$ to $24^{\circ}C$. The variation of the temperature from the mean during the period of irradiation will be limited to $\pm 1.2^{\circ}C$.

- Due to the energy imparted by the radiation the temperature in the DUT may increase after the beginning of the irradiation.
- The difference between the mean temperature of the irradiation room and the mean temperature of the climate controlled chamber shall not exceed $\pm 3^{\circ}C$. (The difference depends on the heat production of the devices placed inside the chamber and can only be confirmed after those have been installed.)
- The temperature variation inside the climate chamber from the mean will be limited to $\pm 1.5^{\circ}C$ as measured by two PT100 sensors installed by INT and CERN.
- When the door of the irradiation room or the door of the climate chamber is opened the variation of the temperature might exceed $\pm 1.5^{\circ}C$ in either place, as the air conditioning of both places is connected. The recovering of the temperature may need several hours.
- In the case of an interruption of the air conditioning system it shall be repaired within 48 hours during which the agreed temperature ranges might not be realized.

A.4.3 Cable length correction for Impedance analysis

According to the Nominal T-Model shown in A-29 for a medium transmission line, the total impedance can be calculated as:

$$Z_T = Z_1 + \frac{Z_3 \cdot (Z_2 + Z_L)}{Z_3 + Z_2 + Z_L} \quad Z_{open} = Z_1 + Z_3 \quad Z_{closed} = Z_1 + \frac{Z_3 \cdot Z_2}{Z_3 + Z_2} \quad (A.22)$$

Where Z_T is the total impedance measured by the impedance analyser, Z_L is the impedance of the DUT being measured and Z_{open} and Z_{closed} correspond to the open

Figure A-29: Nominal T-Model for cable modelling.

and short circuited impedance of the cable.

The analytical expression in A.23 for the load impedance can be achieved considering $Z_1 = Z_2$ and substituting the value of Z_{open} in the first expression in A.22.

$$Z_T = Z_1 + \frac{Z_3 \cdot (Z_2 + Z_L)}{Z_{open} + Z_L} \quad \Rightarrow \quad Z_L = \frac{(Z_T - Z_{closed}) \cdot Z_{open}}{(Z_{open} - Z_T)} \quad (\text{A.23})$$

The expression for Z_L in A.23 is used to correct all the impedance measurements using their corresponding values for Z_{open} and Z_{closed} .

A.4.4 Stroke test

The stroke test analysis for the DUTs A2-A4 and B1-B4 are presented in Figures A-30-A-36 respectively.

A.5 Proton Irradiation Test 2 MGy Active Test - Online results

A.5.1 Leakage current results

The Figures in A-37 present the leakage current results for the DUTs A1-A10.

Figure A-30: Hysteresis variation for A2 - Radiation zone.

Figure A-31: Hysteresis variation for A3 - Low radiation zone.

Figure A-32: Hysteresis variation for A4 - No radiation zone.

Figure A-33: Hysteresis variation for B1 - Radiation zone.

Figure A-34: Hysteresis variation for B2 - Radiation zone.

Figure A-35: Hysteresis variation for B3 - Low radiation zone.

Figure A-36: Hysteresis variation for B4 - No radiation zone.

A.5.2 Stroke test

The stroke test analysis for the DUTs A1-A10 are presented in Figures A-38-A-45 respectively.

(a) A1

(b) A2

(c) A3

(d) A5

(e) A6

(f) A7

(g) A9

(h) A10

Figure A-37: Leakage current measured online.

Figure A-38: Stroke test analysis - DUT A1.

Figure A-39: Stroke test analysis - DUT A2.

Figure A-40: Stroke test analysis - DUT A3.

Figure A-41: Stroke test analysis - DUT A5.

Figure A-42: Stroke test analysis - DUT A6.

Figure A-43: Stroke test analysis - DUT A7.

Figure A-44: Stroke test analysis - DUT A9.

Figure A-45: Stroke test analysis - DUT A10.

Appendix B

Absolute Angle Recovery

B.1 Geometrical relationships

B.1.1 Intersection between line and plane

Let $\mathbf{s}$ be the line generated by the directional vector $\mathbf{v}$ and containing the point $\mathbf{r}$ as in figure B-1. The line is defined by its parametric equation

$$\mathbf{s} = \mathbf{r} + \lambda \mathbf{v}, \quad \forall \lambda \in \mathbb{R} \quad (\text{B.1})$$

On the other hand, the general equation for the plane $\mathbf{w}$ is

$$a_0 + a_1 w_x + a_2 w_y + a_3 w_z = 0 \quad (\text{B.2})$$

where the $\mathbf{n} = (a_1, a_2, a_3) \neq \mathbf{0}$ is any vector perpendicular to the plane and the constant a_0 is defined as $a_0 = -\mathbf{n} \cdot \mathbf{p}_0$ with $\mathbf{p}_0$ any arbitrary point contained in the plane.

The point of intersection $\mathbf{s}_c = \mathbf{r} + \lambda_c \mathbf{v}$ between the line and the plane is found easily by substituting B.1 in B.2 and obtaining an expression for the parameter λ_c :

$$a_0 + a_1(r_x + \lambda_c v_x) + a_2(r_y + \lambda_c v_y) + a_3(r_z + \lambda_c v_z) = 0$$

Figure B-1: Intersection between the line $\mathbf{s}$ and the plane $\mathbf{w}$.

$$\begin{aligned}
 \implies (a_1 v_x + a_2 v_y + a_3 v_z) \cdot \lambda_c &= \\
 &= -a_0 - a_1 r_x - a_2 r_y - a_3 r_z \\
 \implies \lambda_c &= \frac{-a_0 - a_1 r_x - a_2 r_y - a_3 r_z}{a_1 v_x + a_2 v_y + a_3 v_z} = \frac{-a_0 - \mathbf{n} \cdot \mathbf{r}}{\mathbf{n} \cdot \mathbf{v}} \quad (\text{B.3})
 \end{aligned}$$

$$\lambda_c = \frac{\mathbf{n} \cdot \mathbf{p}_0 - \mathbf{n} \cdot \mathbf{r}}{\hat{\mathbf{n}} \cdot \mathbf{v}} \quad (\text{B.4})$$

B.1.2 Reflection of an incident beam on a plane

Given an incident beam with directional vector $\mathbf{v}_i$ and a plane with perpendicular unitary vector $\hat{\mathbf{n}}$ as presented in figure B-2, the reflected beam $\mathbf{v}_f$ can be calculated as:

$$\mathbf{v}_f = \mathbf{v}_i - 2(\mathbf{v}_i \cdot \hat{\mathbf{n}})\hat{\mathbf{n}} \quad (\text{B.5})$$

B.1.3 Refraction of light: Snell's law

Given an incident beam with directional unitary vector $\hat{\mathbf{v}}_i$ in a media with refractive index n_1 as presented in figure B-3, the refracted beam $\hat{\mathbf{v}}_o$ in a media with refractive

Figure B-2: Reflection of an incident beam on a plane.

Figure B-3: Snell's law in vectorial form.

index n_2 can be calculated using the vectorial form of Snell's law as:

$$\hat{v}_o = \frac{n_1}{n_2} \left(\hat{N} \times \left(-\hat{N} \times \hat{v}_i \right) \right) - \hat{N} \sqrt{1 - \left(\frac{n_1}{n_2} \right)^2 \left(\hat{N} \times \hat{v}_i \right) \left(\hat{N} \times \hat{v}_i \right)} \quad (\text{B.6})$$

B.2 Beam path length calculus

The dual-interferometer based measurement system is presented in figure B-4. In this section the distances d_i and d_f are calculated analytically.

B.2.2 Distance from mirror to head

The point of intersection $\mathbf{r}_c$ can be calculated using B.4. The direction of the reflected beam is obtained with the equation B.5, that substituting the values from figure B-4 takes the form

$$\hat{\mathbf{v}}_c = \hat{\mathbf{v}}_i - 2(\hat{\mathbf{v}}_i \cdot \hat{\mathbf{n}})\hat{\mathbf{n}} \quad (\text{B.10})$$

The intersection of the reflected beam with the plane of the head takes place at

$$\mathbf{r}_f = \mathbf{r}_c + \lambda_f \hat{\mathbf{v}}_c \quad (\text{B.11})$$

and therefore the reflected distance, d_r , is

$$d_r = \|\mathbf{r}_f - \mathbf{r}_c\| = |\lambda_f \hat{\mathbf{v}}_c| \quad (\text{B.12})$$

$$\implies d_r = |\lambda_f| \cdot |\hat{\mathbf{v}}_c| = |\lambda_f| \quad (\text{B.13})$$

where, using B.4 with $\hat{\mathbf{v}}_i$ and $\mathbf{r}_i$ as the normal vector and arbitrary point of the plane of the head respectively, λ_f can be expressed as

$$\lambda_f = \frac{\hat{\mathbf{v}}_i \cdot \mathbf{r}_i - \hat{\mathbf{v}}_i \cdot \mathbf{r}_c}{\hat{\mathbf{v}}_i \cdot \hat{\mathbf{v}}_c} \quad (\text{B.14})$$

The use of the scalar product identity in B.14 leads to the final equation for the reflected distance:

$$d_r = \left| \frac{|\mathbf{r}_i| \cdot \cos(\angle(\hat{\mathbf{v}}_i, \mathbf{r}_i)) - |\mathbf{r}_c| \cdot \cos(\angle(\hat{\mathbf{v}}_i, \mathbf{r}_c))}{\cos(\angle(\hat{\mathbf{v}}_i, \hat{\mathbf{v}}_c))} \right| \quad (\text{B.15})$$

B.2.3 Total distance

The interferometer unit used in this measurement system is a standard linear interferometer with no compensation for the rotation of the target. For this reason, the distance measured by the interferometer is just the mean between the incident and

Figure B-5: Dual-interferometer based measurement system with perfect alignment of the interferometer axes.

reflected path lengths.

$$d = \frac{d_i + d_r}{2} \quad (\text{B.16})$$

B.3 Analytical expression for $\theta = f(d_1, d_2)$

B.3.1 Perfect alignment of the heads and centre of rotation in the mid point between heads

The dual-interferometer based measurement system studied in this section is presented in figure B-5. As shown in the diagram, the setup is evaluated under the following assumptions:

- Interferometer axes perfectly perpendicular to the mirror at $0\mu\text{rad}$ ($\hat{\mathbf{v}}_{i_1} = \hat{\mathbf{v}}_{i_2} = (-1, 0, 0)$) and at the same distance from the centre of rotation ($\mathbf{r}_{i_1} = (d_0, D/2, 0)$ and $\mathbf{r}_{i_2} = (d_0, -D/2, 0)$).
- Mirror rotating in the X-Y plane with centre of rotation at the origin of the reference system .

The analytical expression obtained in B.9 and B.15 can be applied to the axis 1, leading to

$$d_{i_1} = \left| \frac{\sqrt{d_0^2 + \left(\frac{D}{2}\right)^2} \cdot \cos(\theta_1 - \theta)}{\cos(180 - \theta)} \right|$$

$$d_{r_1} = \left| \frac{\sqrt{d_0^2 + \left(\frac{D}{2}\right)^2} \cdot \cos(\theta_1 + 180) - \frac{D}{2\cos(\theta)} \cdot \cos(-90 + \theta)}{\cos(180 + 2\theta)} \right| \quad (\text{B.17})$$

The previous equations can be simplified by renaming the constants as $A = \sqrt{d_0^2 + \left(\frac{D}{2}\right)^2}$ and $B = \sqrt{d_0^2 + \left(\frac{D}{2}\right)^2} \cdot \cos(\theta_1)$ and using the trigonometrical property of the cosine:

$$\cos(\alpha \pm \beta) = \cos \alpha \cos \beta \mp \sin \alpha \sin \beta \quad (\text{B.18})$$

In this case, B.3.1 and B.17 can be rewritten as

$$d_{i_1} = \left| \frac{A \cdot \cos(\theta_1 - \theta)}{-\cos(\theta)} \right|$$

$$d_{r_1} = \left| \frac{B + \frac{D}{2} \cdot \tan(\theta)}{\cos(2\theta)} \right| \quad (\text{B.19})$$

Similarly, for the axis 2 it can be seen that

$$d_{i_2} = \left| \frac{\sqrt{d_0^2 + \left(\frac{D}{2}\right)^2} \cdot \cos(-\theta_1 - \theta)}{\cos(180 - \theta)} \right|$$

$$= \left| \frac{A \cdot \cos(\theta_1 + \theta)}{-\cos(\theta)} \right| \quad (\text{B.20})$$

$$d_{r_2} = \left| \frac{\sqrt{d_0^2 + \left(\frac{D}{2}\right)^2} \cdot \cos(\theta_1 + 180) - \frac{D}{2\cos(\theta)} \cdot \cos(90 + \theta)}{\cos(180 + 2\theta)} \right|$$

$$= \left| \frac{B - \frac{D}{2} \cdot \tan(\theta)}{\cos(2\theta)} \right| \quad (\text{B.21})$$

The total distances measured by the two interferometer axes are expressed as

$$2 \cdot d_1 = \frac{A \cdot \cos(\theta_1 - \theta)}{\cos(\theta)} + \frac{B + \frac{D}{2} \cdot \tan(\theta)}{\cos(2\theta)} \quad (\text{B.22})$$

$$2 \cdot d_2 = \frac{A \cdot \cos(\theta_1 + \theta)}{\cos(\theta)} + \frac{B - \frac{D}{2} \cdot \tan(\theta)}{\cos(2\theta)} \quad (\text{B.23})$$

where the absolute value bars are removed considering that $\theta = \pm 10 \text{ mrad}$, $\theta \ll \theta_1$ and $B \gg \pm D/2 \tan \theta$. The previous two equations can be added leading to

$$\begin{aligned} 2 \cdot (d_1 + d_2) &= \frac{A \cdot [\cos(\theta_1 - \theta) + \cos(\theta_1 + \theta)]}{\cos(\theta)} \\ &\quad + \frac{B + \cancel{\frac{D}{2} \cdot \tan(\theta)} + B - \cancel{\frac{D}{2} \cdot \tan(\theta)}}{\cos(2\theta)} \\ &= \frac{2 \cdot A \cdot \cos(\theta_1) \cancel{\cos(\theta)}}{\cancel{\cos(\theta)}} + \frac{2 \cdot B}{\cos(2\theta)} \\ &= 2 \cdot B + \frac{2 \cdot B}{\cos(2\theta)} \end{aligned} \quad (\text{B.24})$$

Equation B.24 can be rearranged giving a final expression for the angle as a function of both distances and the constant B :

$$\theta = \frac{\cos^{-1} \left(\frac{B}{d_1 + d_2 - B} \right)}{2} \quad (\text{B.25})$$

B.3.2 Perfect alignment of the axes and centre of rotation displaced from the mid point between heads

In this case, the system in figure B-6 is studied. The centre of rotation is displaced with respect to the mid point between the heads and the interferometer axes are assumed to be perfectly aligned as in the previous configuration.

Applying the results in B.9 and B.15 to the axis 1, it can be seen that

Figure B-6: Dual-interferometer based measurement system with perfect alignment of the interferometer axes with centre of rotation displaced from the middle point.

$$\begin{aligned}
 d_{i_1} &= \left| \frac{\sqrt{d_0^2 + (D_1)^2} \cdot \cos(\theta_1 - \theta)}{\cos(180 - \theta)} \right| \\
 &= \left| \frac{A_1 \cdot \cos(\theta_1 - \theta)}{-\cos(\theta)} \right|
 \end{aligned} \tag{B.26}$$

$$\begin{aligned}
 d_{r_1} &= \left| \frac{\sqrt{d_0^2 + D_1^2} \cdot \cos(180 + \theta_1) - \frac{D_1}{\cos(\theta)} \cdot \cos(-90 + \theta)}{\cos(180 + 2\theta)} \right| \\
 &= \left| \frac{-A_1 \cos(\theta_1) - D_1 \cdot \tan(\theta)}{-\cos(2\theta)} \right| \\
 &= \left| \frac{B_1 + D_1 \cdot \tan(\theta)}{\cos(2\theta)} \right|
 \end{aligned} \tag{B.27}$$

where $A_1 = \sqrt{d_0^2 + \left(\frac{D_1}{2}\right)^2}$ and $B_1 = \sqrt{d_0^2 + \left(\frac{D_1}{2}\right)^2} \cdot \cos(\theta_1)$.

And similarly, for the axis 2:

$$\begin{aligned}
d_{i_2} &= \left| \frac{\sqrt{d_0^2 + (D_2)^2} \cdot \cos(\theta_2 - \theta)}{\cos(180 - \theta)} \right| \\
&= \left| \frac{A_2 \cdot \cos(\theta_2 - \theta)}{-\cos(\theta)} \right|
\end{aligned} \tag{B.28}$$

$$\begin{aligned}
d_{r_2} &= \left| \frac{\sqrt{d_0^2 + D_2^2} \cdot \cos(180 + \theta_2) - \frac{D_2}{\cos(\theta)} \cdot \cos(90 + \theta)}{\cos(180 + 2\theta)} \right| \\
&= \left| \frac{-A_2 \cos(\theta_2) - D_2 \cdot \tan(\theta)}{-\cos(2\theta)} \right| \\
&= \left| \frac{B_2 - D_2 \cdot \tan(\theta)}{\cos(2\theta)} \right|
\end{aligned} \tag{B.29}$$

where $A_2 = \sqrt{d_0^2 + \left(\frac{D_1}{2}\right)^2}$ and $B_2 = \sqrt{d_0^2 + \left(\frac{D_1}{2}\right)^2} \cdot \cos(\theta_2)$.

The distances measured by both interferometer axes are summarized in the equations

$$2 \cdot d_1 = \frac{A_1 \cdot \cos(\theta_1 - \theta)}{\cos(\theta)} + \frac{B_1 + D_1 \cdot \tan(\theta)}{\cos(2\theta)} \tag{B.30}$$

$$2 \cdot d_2 = \frac{A_2 \cdot \cos(\theta_2 - \theta)}{\cos(\theta)} + \frac{B_2 - D_2 \cdot \tan(\theta)}{\cos(2\theta)} \tag{B.31}$$

Both expressions B.30 and B.31 can be added and developed as

$$\begin{aligned}
2 \cdot (d_1 + d_2) &= \frac{A_1 \cdot \cos(\theta_1 - \theta) + A_2 \cdot \cos(\theta_2 - \theta)}{\cos(\theta)} \\
&\quad + \frac{B_1 + B_2 + (D_1 - D_2) \cdot \tan(\theta)}{\cos(2\theta)}
\end{aligned} \tag{B.32}$$

The first term of the left side of the equation can be developed and simplified as

$$\frac{A_1 \cdot \cos(\theta_1 - \theta) + A_2 \cdot \cos(\theta_2 - \theta)}{\cos(\theta)} = \tag{B.33}$$

$$\begin{aligned}
&= \frac{A_1 \cos \theta_1 \cos \theta + A_1 \sin \theta_1 \sin \theta}{\cos \theta} + \\
&\quad + \frac{A_2 \cos \theta_2 \cos \theta + A_2 \sin \theta_2 \sin \theta}{\cos(\theta)} \quad (\text{B.34})
\end{aligned}$$

$$\begin{aligned}
&= \frac{(A_1 \cos \theta_1 + A_2 \cos \theta_2) \overset{2d_0}{\cancel{\cos \theta}}}{\cancel{\cos \theta}} + \\
&\quad + \frac{(A_1 \overset{D_1}{\cancel{\sin \theta_1}} + A_2 \overset{-D_2}{\cancel{\sin \theta_2}}) \sin \theta}{\cos \theta} \quad (\text{B.35}) \\
&= 2d_0 + (D_1 - D_2) \tan \theta
\end{aligned}$$

Using that $B_1 + B_2 = 2B = 2d_0$ the equation B.32 can be rewritten as

$$\begin{aligned}
2(d_1 + d_2) &= 2d_0 + (D_1 - D_2) \tan \theta \\
&\quad + \frac{2B + (D_1 - D_2) \tan \theta}{\cos(2\theta)} \quad (\text{B.36})
\end{aligned}$$

The denominator can be expanded using the trigonometric relationship:

$$\cos(2\theta) = \frac{1 - \tan^2 \theta}{1 + \tan^2 \theta} \quad (\text{B.37})$$

and then equation B.36 converts into

$$\begin{aligned}
2(d_1 + d_2) - 2d_0 - (D_1 - D_2) \tan \theta \\
= \frac{2B \cdot (1 + \tan^2 \theta) + (D_1 - D_2) \tan \theta (1 + \tan^2 \theta)}{1 - \tan^2 \theta} \implies \quad (\text{B.38})
\end{aligned}$$

$$\begin{aligned}
&\implies 2(d_1 + d_2 - d_0) - (D_1 - D_2) \tan \theta \\
&\quad - 2(d_1 + d_2 - d_0) \cdot \tan^2 \theta + \cancel{(D_1 - D_2) \cdot \tan^3 \theta} \\
&= 2B + 2B \cdot \tan^2 \theta + (D_1 - D_2) \tan \theta + \cancel{(D_1 - D_2) \cdot \tan^3 \theta} \quad (\text{B.39})
\end{aligned}$$

$$\begin{aligned}
&\implies 2(d_1 + d_2 - d_0) - 2B + \\
&\quad + (-2(D_1 - D_2)) \tan \theta + \\
&\quad + (-2(d_1 + d_2 - d_0) - 2B) \cdot \tan^2 \theta = 0
\end{aligned} \quad (\text{B.40})$$

$$\begin{aligned}
&\implies (d_1 + d_2 - 2d_0) + \\
&\quad - (D_1 - D_2) \cdot \tan \theta + \\
&\quad - (d_1 + d_2) \cdot \tan^2 \theta = 0
\end{aligned} \quad (\text{B.41})$$

$$\begin{aligned}
&\tan \theta = \\
&\quad \frac{(D_1 - D_2) \pm \sqrt{(D_1 - D_2)^2 + 4(d_1 + d_2 - 2d_0) \cdot (d_1 + d_2)}}{-2 \cdot (d_1 + d_2)} \quad (\text{B.42})
\end{aligned}$$

Finally, the renaming of the constant $(D_1 + D_2)$ as C leads to the final equation:

$$\theta = \tan^{-1} \left(\frac{C \pm \sqrt{C^2 + 4(d_1 + d_2 - 2d_0) \cdot (d_1 + d_2)}}{-2 \cdot (d_1 + d_2)} \right) \quad (\text{B.43})$$

where, apart from the interferometer distances, there are two constants $C = D_1 - D_2$ and d_0 from the geometry of the system.

B.4 Uncertainty Analysis of " $\tan^{-1}$ approximation"

Using the approximation studied in section 4.2.1.4 the angle can be calculated using the distances in both reference systems according to:

$$\theta = \tan^{-1} \left(\frac{d_1 - d_2}{D} \right) = \tan^{-1} \left(\frac{d'_1 - c_1 - (d'_2 - c_2)}{D} \right) \quad (\text{B.44})$$

B.4.1 Uncertainty analysis

The uncertainty propagation formula can be applied to both terms:

$$\begin{aligned} \delta\theta^2 = & w_{d_1}^2 \delta d_1^2 + w_{d_2}^2 \delta d_2^2 + w_D^2 \delta D^2 + \\ & + 2 \cdot w_{d_1} \cdot w_{d_2} \cdot \rho_{d_1 d_2} \cdot \delta d_1 \cdot \delta d_2 + \\ & + 2 \cdot w_{d_1} \cdot w_D \cdot \rho_{d_1 D} \cdot \delta d_1 \cdot \delta D + \\ & + 2 \cdot w_{d_2} \cdot w_D \cdot \rho_{d_2 D} \cdot \delta d_2 \cdot \delta D \end{aligned} \quad (\text{B.45})$$

$$\begin{aligned}
\delta\theta^2 = & w_{d'_1}^2 \delta d'_1{}^2 + w_{d'_2}^2 \delta d'_2{}^2 + w_D^2 \delta D^2 + \\
& + w_{c_1}^2 \delta c_1{}^2 + w_{c_2}^2 \delta c_2{}^2 + \\
& + 2 \cdot w_{d'_1} \cdot w_{d'_2} \cdot \rho_{d'_1 d'_2} \cdot \delta d'_1 \cdot \delta d'_2 + \\
& + 2 \cdot w_{d'_1} \cdot w_D \cdot \rho_{d'_1 D} \cdot \delta d'_1 \cdot \delta D + \\
& + 2 \cdot w_{d'_2} \cdot w_D \cdot \rho_{d'_2 D} \cdot \delta d'_2 \cdot \delta D + \\
& + 2 \cdot w_{c_1} \cdot w_{c_2} \cdot \rho_{c_1 c_2} \cdot \delta c_1 \cdot \delta c_2 + \\
& + 2 \cdot w_{c_1} \cdot w_D \cdot \rho_{c_1 D} \cdot \delta c_1 \cdot \delta D + \\
& + 2 \cdot w_{c_2} \cdot w_D \cdot \rho_{c_2 D} \cdot \delta c_2 \cdot \delta D + \\
& + 2 \cdot w_{d'_1} \cdot w_{c_1} \cdot \rho_{d'_1 c_1} \cdot \delta d'_1 \cdot \delta c_1 + \\
& + 2 \cdot w_{d'_1} \cdot w_{c_2} \cdot \rho_{d'_1 c_2} \cdot \delta d'_1 \cdot \delta c_2 + \\
& + 2 \cdot w_{d'_2} \cdot w_{c_1} \cdot \rho_{d'_2 c_1} \cdot \delta d'_2 \cdot \delta c_1 + \\
& + 2 \cdot w_{d'_2} \cdot w_{c_2} \cdot \rho_{d'_2 c_2} \cdot \delta d'_2 \cdot \delta c_2
\end{aligned} \tag{B.46}$$

where $w_x = \frac{\partial\theta}{\partial x}$, and ρ_{xy} is the correlation coefficient of the error in x and y .

The uncertainty associated with the recovery method can be identified as all the additional terms presented in the uncertainty propagation equation for the case of

the new reference system:

$$\begin{aligned}
\delta\theta_{rec}^2 = & w_{c_1}^2 \delta c_1^2 + w_{c_2}^2 \delta c_2^2 + \\
& + 2 \cdot w_{c_1} \cdot w_{c_2} \cdot \rho_{c_1 c_2} \cdot \delta c_1 \cdot \delta c_2 + \\
& + 2 \cdot w_{c_1} \cdot w_D \cdot \cancel{\rho_{c_1 D}} \cdot^0 \delta c_1 \cdot \delta D + \\
& + 2 \cdot w_{c_2} \cdot w_D \cdot \cancel{\rho_{c_2 D}} \cdot^0 \delta c_2 \cdot \delta D + \\
& + 2 \cdot w_{d'_1} \cdot w_{c_1} \cdot \cancel{\rho_{d'_1 c_1}} \cdot^0 \delta d'_1 \cdot \delta c_1 + \\
& + 2 \cdot w_{d'_1} \cdot w_{c_2} \cdot \cancel{\rho_{d'_1 c_2}} \cdot^0 \delta d'_1 \cdot \delta c_2 + \\
& + 2 \cdot w_{d'_2} \cdot w_{c_1} \cdot \cancel{\rho_{d'_2 c_1}} \cdot^0 \delta d'_2 \cdot \delta c_1 + \\
& + 2 \cdot w_{d'_2} \cdot w_{c_2} \cdot \cancel{\rho_{d'_2 c_2}} \cdot^0 \delta d'_2 \cdot \delta c_2
\end{aligned} \tag{B.47}$$

For the partial derivative we find that

$$w_{c_1} = \frac{-D}{D^2 + (d'_1 - c_1 - d'_2 + c_2)^2} \tag{B.48}$$

$$w_{c_2} = \frac{+D}{D^2 + (d'_1 - c_1 - d'_2 + c_2)^2} \tag{B.49}$$

These expressions can be substituted into the previous equation for $\delta\theta_{rec}^2$ and considering $\delta c_1 = \delta c_2 = \delta c$ it can be seen that:

$$\begin{aligned}
\delta\theta_{rec}^2 = & 2 \cdot \left(\frac{D}{D^2 + (d'_1 - c_1 - d'_2 + c_2)^2} \right)^2 \delta c^2 + \\
& - 2 \cdot \left(\frac{D}{D^2 + (d'_1 - c_1 - d'_2 + c_2)^2} \right)^2 \cdot \rho_{c_1 c_2} \cdot \delta c^2
\end{aligned} \tag{B.50}$$

The key parameter to understand the uncertainty introduced by the recovery relies in the correlation coefficient $\rho_{c_1 c_2}$. It is important to remark that this analysis uses the approximation formula 4.10, and therefore, the shape of the characteristic curve is not included in this equation.

If we perform several recoveries, the correlation coefficient $\rho_{c_1 c_2}$ represents how much correlation exists between the error estimation of c_1 and c_2 of each recovery. Looking at the shape of the real characteristic curve d_1 vs d_2 it can be concluded that the dispersion of the pairs (c_1, c_2) around the real offset (c_{real_1}, c_{real_2}) will not be symmetrically distributed around this final value, but they will be spread around the characteristic curve. This fact implies that the correlation coefficient $\rho_{c_1 c_2}$ cannot be assumed as 0, and therefore this term cannot be cancelled.

The real value of $\rho_{c_1 c_2}$ is given experimentally by the performance of the recovery method, which depends on different factors such as the number of points or the standard deviation of the noise in the distance signals. However, the expected value will be very close to -1 since the error of the estimated offset is expected to follow the characteristic curve shape.

In this case, the equation B.50 leads to:

$$\begin{aligned}
\delta\theta_{rec}^2 &= 2 \cdot \left(\frac{D}{D^2 + (d'_1 - c_1 - d'_2 + c_2)} \right)^2 \delta c^2 + \\
&\quad + 2 \cdot \left(\frac{D}{D^2 + (d'_1 - c_1 - d'_2 + c_2)^2} \right)^2 \cdot |\rho_{c_1 c_2}| \cdot \delta c^2 \\
&\leq 2 \cdot \left(\frac{D}{D^2 + (d'_1 - c_1 - d'_2 + c_2)^2} \right)^2 \delta c^2 + \\
&\quad + 2 \cdot \left(\frac{D}{D^2 + (d'_1 - c_1 - d'_2 + c_2)^2} \right)^2 \cdot 1 \cdot \delta c^2 \\
&\leq 4 \cdot \left(\frac{1}{D} \right)^2 \delta c^2
\end{aligned} \tag{B.51}$$

B.5 Correlation coefficient of Monte Carlo simulations and d_1 vs d_2

B.5.1 Effect of number of points in the recovery method

Figure B-7 present the values of the correlation coefficients for the Monte Carlo simulation varying the number of points of Section 4.4.3.

(a) c_2 vs c_1 .

(b) Correlation coefficient vs Number of points.

Figure B-7: Additional results for simulation of Fig. 4-16.

B.5.2 Effect of standard deviation of the noise in the recovery method

Figure B-8 present the values of the correlation coefficients for the Monte Carlo simulation varying the standard deviation of the noise of Section 4.4.4.

B.6 Experimental results - Test 1

The Test 1 was performed with 340000 points in each acquisition and presented mean values for the noise statistics of $\sigma_{noise} = 0.012983 \mu m$ and $\rho_{noise} = -0.1034$ (Pearson

(a) c_2 vs c_1

(b) Correlation coefficient vs Standard deviation of the noise.

Figure B-8: Additional results for simulation of Fig. 4-17.

correlation coefficient), parameters that were used in the Monte Carlo simulation. The results for Test 1 are presented in Figure B-9.

In this position, the real angle was acquired with the external interferometer system during 7 seconds ($2kHz$ sampling frequency) and the results for the mean angle with the standard deviation are presented in Figure B-9(b).

The standard deviation of the recovered angles for the Test 1 (Ref 1-5) is $0.53\mu\text{rad}$ and their average value differs from the initial value (Ref 0) by $0.06\mu\text{rad}$.

Combining both results as the repetitive and bias uncertainties of the new method,

it can be concluded that the angle was successfully recovered within an accuracy of $\pm 1.65 \mu rad$ (3σ) for the Test 1, with $\sigma = 0.55 \mu rad$.

(a) Characteristic curves d_2 vs d_1 (rotated for visualisation purposes) for the different reference systems after the loss of position in the interferometer.

(b) External interferometer acquisition for the rotational stage in closed loop at $0\mu\text{rad}$ for the different reference systems. Ref 0: Original reference system, Ref 1-5: Angle recovered. The mean and standard deviation are calculated using a $7s$ acquisition where the angle is stable thanks to the closed loop angle control.

Figure B-9: Test 1 with 5 repetitions.

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