Advanced Staged Control Design for Stellar Interferometry with Experimental Results on SPHERES

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Septembre 2011

SSL # 15-11
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This work is based on the unaltered text of the thesis by Benjamin Fragnière submitted to SUPAERO in partial fulfillment of the requirements for the degree of SUPAERO Aerospace Engineering Diploma at the Institut Supérieur de l’Aéronautique et de l’Espace, SUPAERO.
Advanced Staged Control Design for Stellar Interferometry with Experimental Results on SPHERES

by

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Submitted to the Institut Supérieur de l’Aéronautique et de l’Espace, SUPAERO on September 30, 2011, in partial fulfillment of the requirements for the degree of SUPAERO Aerospace Engineering Diploma

Abstract

The present master’s thesis discusses the control of an optical path length in the context of the space interferometric telescopes. Using the Synchronized Position Hold Engage and Reorient Experimental Satellites (SPHERES), a ground testbed destined to develop the Synthetic Imager Maneuvering Optimization (SIMO) program has been designed by the Space System Laboratory (SSL). The goal was the control of the optical path length going from a laser interferometer to a satellite floating on an air table using an Optical Delay Line (ODL) constituted of piezo and voice-coil actuated mirrors. The main contribution to the project is the design and implementation of the ODL controller, leading to a set of measurements with a 1-DOF floating satellite. A cascaded staged control structure has been designed in order to take full advantage of the different stages of actuators. Furthermore a couple of hardware modifications have been performed and set of recommendations has been proposed for future work in order to reach the performances sought by the SIMO program.

Thesis Supervisor: David W. Miller
Title: Professor
Acknowledgements

This Thesis was performed as part of the SUPAERO (Institute Supérieur de l’Aéronautique et de l’Espace, Toulouse, FRANCE) and EPFL (Ecole Polytechnique Fédérale de Lausanne, Switzerland) double Degree Master’s Thesis. The work on SPHERES was supported by NASA Goddard Space Flight Center Phase 2 SBIR with Aurora Flight Sciences entitled "Synthetic Imaging Maneuver Optimization" (SIMO), under contract NX09CA65C. The author express his profound gratitude to SUPAERO for generously supporting him.
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Introduction

Observation of the universe requires more and more high-resolution telescopes, especially for exo-
planet’s detection. For that reason, telescopes have been transferred from the ground to the space,
overcoming the atmospheric disturbances. The next step of resolution improvement is to increase
the size of the telescope’s aperture, but since this is limited by the launcher, the idea came to use
several independent light collectors flying in formation and acting like a single big telescope. In
this case, the principle is quite different from the classical lens telescopes, since interferences are
used to detect stars. Hence, the required precision for the relative position between vehicles is
in the order of magnitude of nanometers, while the separation between collectors can reach up to
hundred meters. Classical satellite guidance systems are not able of such a precise control. Also
an Optical Delay Line (ODL) has to be inserted between the collecting vehicles and the combiner.
The ODL allows keeping the light path constant from the collectors to the receptors, despite any
vehicle motions and vibrations.

The MIT Space System Laboratory (SSL), in collaboration with Aurora Flight Sciences, built
a ground testbed destined to develop the Synthetic Imager Maneuvering Optimization (SIMO)
program, which aims to help future design of spacecrafts performing the formation flight necessary
for space interferometry missions. In this project an optical path length from a laser interferometer
to a satellite floating on an air table has to be controlled using an Optical Delay Line constituted
of actuated mirrors. The satellite used for that purpose is the Synchronized Position Hold Engage
and Reorient Experimental Satellites (SPHERES) designed by the SSL and operating with its own
metrology, control and propulsion systems. Two stages of actuators try to reject the error brought by
the lack of precision of the satellite’s control system on one hand and by high frequency vibrations
of the optics on the other hand. Furthermore, the rotations of the satellites have to be corrected
in order to keep the laser beam directed at the center of the interferometer. A set of fast steering
mirrors, with sensing of a camera, is used for this purpose (see picture 3-1).

The objective of this Master’s Thesis is the design of a staged controller regulating the optical
path length. The implementation of the controller in the SIMO testbed will be achieved in order to proceed tests and thereby tend to demonstrate that the precisions required for space interferometry missions can be performed on the ground in space-based telescope conditions, that is to say with a floating satellite.
Chapter 1

Space Interferometry Missions

1.1 Interferometry Basics

One of the first interferometers is the Michelson Interferometer (figure 1-1). In this experiment, a spherical light front comes from a the $Q$ point and goes into two slits $O$ and $O'$. After passing the slits, the two new light fronts are hitting a screen and create interferences under the form of fringes on it. The intensity of light on the screen depends on the phase shift between the light coming from the $O$ point and the light coming from the $O'$ point. Thus, as the two beams have the same wavelength, the phase shift depends only on the path-length difference and on the wavelength. For small angles, the path-length difference can be approximated by $\Delta x - x\theta$, with $\theta$ the angle of incident light on the mask and $y$ the separation distance between the optic axis and the considered point on the screen. The phase shift is found after multiplication by the wavenumber $k = \frac{2\pi}{\lambda}$. Finally the total intensity on the screen is proportional to:

$$cos^2(kx(\frac{y}{y} - \theta))$$  \hspace{1cm} (1.1)

Thanks to the property of linearity, a chromatic source can be considered as a superposition of several monochromatic sources, and an incoherent source can be considered as a distribution of several sources. The resulting intensity will be the sum of each intensities.
1.2 Stellar Interferometry

As it will be shown in the following part, the size of a star can be calculated measuring the separation between the slits, $x$, giving the minimum fringe visibility. For that purpose, four photodiode detectors are positioned in the Michelson’s interferometer so that the optical path length difference between the two beams are equal to 0, 1/4, 1/2 and 3/4 of the wavelength (see figure 1.6). If we consider $A$ the complex amplitude of the light coming from O and $A'$ the complex amplitude coming from O', intensity in each detector is defined by:

$$I_1 = |A + A'|^2 = |A + A exp(-i\theta kx)|^2 = 2AA^*(1 + \cos(\theta kx))$$  

$$I_2 = |A + A'|^2 = |A + A exp(i\frac{\pi}{2} - i\theta kx)|^2 = 2AA^*(1 + \sin(\theta kx))$$  

$$I_3 = |A + A'|^2 = |A + A exp(i\pi - i\theta kx)|^2 = 2AA^*(1 - \cos(\theta kx))$$  

$$I_2 = |A + A'|^2 = |A + A exp(i3\frac{\pi}{2} - i\theta kx)|^2 = 2AA^*(1 - \sin(\theta kx))$$
the complex intensity can be described by:

\[
I = (I_1 - I_3) + i(I_2 - I_4) = 2AA^* (2\cos(\theta kx) + 2isin(\theta kx)) = 4AA^* \exp(\theta kx)
\] (1.6)

If we consider a star as a repartition of source points with angle \(\theta\) and brightness \(B(\theta)\), the amplitude \(A(\theta)\) of the light coming from angles \(\theta\) and \(\theta + d\theta\) is given by:

\[
|A(\theta)|^2 = B(\theta)d\theta
\] (1.7)

and thus, using 1.6, the associated complex intensity results in:

\[
I(\theta) = 4B(\theta)\exp(\theta kx)d\theta
\] (1.8)

The total intensity is basically the sum of contributions all over the star:
Figure 1-3: Illustration of the Optical Delay Line in SIM Astrometric Measurement.

\[ I_{\text{total}} = \int_{\text{star}} I(\theta) d\theta = \int_{\text{star}} 4B(\theta) \exp(\theta kx) d\theta \]  \hspace{1cm} (1.9)

The visibility is defined by the complex intensity normalized in order to get one for maximal fringe visibility, and zero when there is no interference:

\[ V(kx) = \int_{\text{star}} \frac{(I_1 - I_3) + i(I_2 - I_4)}{I_1 + I_2 + I_3 + I_4} d\theta = \int_{\text{star}} \frac{4B(\theta) \exp(\theta kx)}{4|A(\theta)|^2} d\theta \]  \hspace{1cm} (1.10)

Finally, the visibility only depends on the wavenumber \( k \) and on the separation between the slits \( x \) the brightness distribution, the distance of the star and its size.

In order to be able to detect stars with a large range of sizes and distances, the two light collectors \( O \) and \( O' \) have to be able to reach a separation \( x \) up to several hundreds of meters. This means that monobloc collectors are not suitable anymore and two separated light collecting vehicles flying in formation becomes the only option. A key point is to keep the distance between the two collectors and the row of detectors constant with a precision level allowing the fringe detection, that means some fractions of the star light wavelength.

As no space vehicle control system can reach that nano-level precision, an Optical Path Delay (OPD) is inserted in one of the path in order to compensate the relative motions of the collectors and thus to keep the two path length equal. Figure 1-3 illustrates the OPD applied to stellar interferometry.
Chapter 2

Previous Work

2.1 System Modeling

2.1.1 Stochastic disturbance model

Disturbances entering the plant have to be characterized in order to predict future performances, and thus to know if the actuators will be able to follow the required precision. A disturbance model will also help for choosing the most appropriate cut-off frequency of the control laws.

Disturbances can be modeled either deterministically or stochastically, depending on their nature. For example, a solar pressure on a satellite or a magnetic field could be modeled as a force constantly applied or related to the position of the satellite. However, high frequency noise, such as vibrations induced by the rotation of a reaction wheel or by thrusters, have to be modeled stochastically, since no deterministic model exists.

Under the assumption of a wide-sense stationary stochastic (WSS) process, the behavior of a noise can be characterized by its autocorrelation function,

\[ R_{xx}(\tau) = E\{x(t+\tau)x^*(\tau)\} \]  \hspace{1cm} (2.1)

and by its power spectral sensitivity (PSD),

\[ S_{xx}(\omega) = \int_{-\infty}^{\infty} R_{xx}(\tau)e^{-j\omega \tau}d\tau. \]  \hspace{1cm} (2.2)
The covariance of the noise is then defined by:

\[ E\{ |x(t)|^2 \} = R_{xx}(0) = \frac{1}{2\pi} \int_{-\infty}^{\infty} S(\omega) \, d\omega. \]  

(2.3)

This WSS noise can be modeled as the response to a linear, minimum-phase system when driven by a zero-mean, unit-intensity white noise, \( w(t) \):

\[ x(t) = \int_{-\infty}^{\infty} w(t - \alpha) g(\alpha) \, d\alpha, \]  

(2.4)

\[ S_{xx}(\omega) = G(\omega) G^*(\omega). \]  

(2.5)

where \( g(t) \) and \( G \) are the impulse response of the system and its Fourier transform respectively. Thus, given a WSS process \( x(t) \), it is possible to associate a transfer function \( G \) corresponding to the answer of a system with function transfer \( G \) driven by a white noise. This filter is known as innovation filter. In practice, the PSD of the \( x(t) \) process can be found experimentally, and then the innovation filter \( G \) can be calculated using function 2.5. A disturbance state model can be deduced from this innovation filter:

\[ \dot{x}_d = A_d x_d + B_d w \]  

(2.6)

\[ y = C_d x_d \]  

(2.7)

\begin{figure}[h]
\centering
\includegraphics[width=0.5\textwidth]{figure2-1.png}
\caption{Figure 2-1: Disturbance model integrated into the system model.}
\end{figure}

This expression can be included into the overall dynamical model using state augmentation as shown in figure 2-1. If the disturbance acts as an input to the plant dynamics, the model can be
describe as:

\[
\begin{bmatrix}
\dot{x}_p \\
\dot{x}_d
\end{bmatrix} =
\begin{bmatrix}
A_p & B_u C_d \\
0 & A_s
\end{bmatrix}
\begin{bmatrix}
x_p \\
x_d
\end{bmatrix} +
\begin{bmatrix}
0 \\
B_d
\end{bmatrix} w +
\begin{bmatrix}
B_u \\
0
\end{bmatrix} u
\] (2.8)

\[
y = \begin{bmatrix}
C_p & 0
\end{bmatrix}
\begin{bmatrix}
x_p \\
x_d
\end{bmatrix}
\] (2.9)

Otherwise, if the disturbance acts directly on the output, the model is expressed by:

\[
\begin{bmatrix}
\dot{x}_p \\
\dot{x}_d
\end{bmatrix} =
\begin{bmatrix}
A_p & 0 \\
0 & A_s
\end{bmatrix}
\begin{bmatrix}
x_p \\
x_d
\end{bmatrix} +
\begin{bmatrix}
0 \\
B_d
\end{bmatrix} w +
\begin{bmatrix}
B_u \\
0
\end{bmatrix} u
\] (2.10)

\[
y = \begin{bmatrix}
C_p & C_d
\end{bmatrix}
\begin{bmatrix}
x_p \\
x_d
\end{bmatrix}
\] (2.11)

Once the plant and disturbance have been integrated into this form, the effect of disturbance can be formally analyzed.

### 2.1.2 Actuators characteristics

Actuators are composed by a mechanical and an electrical part. Both of them have to be modeled in order to characterize the whole bandwidth of the actuator. The input \( u \) is an electrical value (voltage or current) and the output \( v = \phi(u) \) is a mechanical displacement. A transfer function can be defined for these linear features, and then converted into a state model. The nonlinearities have to be treated differently; the following part examines the different kinds of nonlinearities that the function \( \phi \) can present.

Saturation is the most important and the most common constraint that can lead to a loss of authority of the actuator. It can be seen as a maximum value reachable by the actuator. Figure 2-2 shows function \( \phi \) in this case.
Resolution is another nonlinearity, it is defined as the minimum physical output of the actuator and is modeled as a small dead band around the zero input level, with a jump in the "normal" value when input signal exceeds the dead band. That behavior can be seen on figure 2-3. This kind of linearity is typical of a mechanical system with friction or an electrical system with a threshold. The minimum duration of a thruster can also be modeled with a resolution.

A minor nonlinearity is the quantization that is due to the roundoff errors introduced by the digital to analog conversion. The $\phi$ function relative to this nonlinearity is plot on figure 2-4. This effect can be modeled by replacing the quantizer with an additive noise [5]. In this model, the input noise is a sequence of random variables $u$ with an uniform probability density function from $-q/2$ to $q/2$ and a variance of $q^2/12$. 

Figure 2-2: Saturation model of actuator.

Figure 2-3: Resolution model of actuator.
2.1.3 Composite model

When pulling together the dynamical model, the innovation filters and the electrical state model we obtain the following augmented state model.

\[
\begin{bmatrix}
\dot{x}_p \\
\dot{x}_d \\
\dot{x}_{a1} \\
\vdots \\
\dot{x}_{an} \\
\end{bmatrix} = 
\begin{bmatrix}
A_p & B_wC_d & B_{u1}C_{a1} & \ldots & C_{an} \\
0 & A_d & 0 & \ldots & 0 \\
0 & 0 & A_{d1} & \ldots & 0 \\
\vdots & \vdots & \vdots & \ddots & \vdots \\
0 & 0 & 0 & \ldots & A_{an} \\
\end{bmatrix} \begin{bmatrix}
x_p \\
x_d \\
x_{a1} \\
\vdots \\
x_{an} \\
\end{bmatrix} + 
\begin{bmatrix}
0 \\
x_d \\
x_{a1} \\
\vdots \\
x_{an} \\
\end{bmatrix} \\
\begin{bmatrix}
0 \\
\vdots \\
0 \\
\end{bmatrix} + 
\begin{bmatrix}
0 \\
\vdots \\
0 \\
\end{bmatrix} w + 
\begin{bmatrix}
0 \\
\vdots \\
0 \\
\end{bmatrix} \tilde{u}
\]

(2.12)

\[
y = \begin{bmatrix}
C_p & 0 & 0 & \ldots & 0 \\
x_p \\
x_d \\
x_{a1} \\
\vdots \\
x_{an} \\
\end{bmatrix} 
\]

(2.13)

where \(p\) index is relative to the plant, \(d\) index to the disturbance - that means the innovation filters - and \(a_i\) index designates the electrical model of the \(i^{th}\) actuator. Furthermore,

\[
B_A = \begin{bmatrix}
B_{a1} & 0 \\
\vdots & \ddots \\
0 & B_{an} \\
\end{bmatrix}
\]

(2.14)

An schema of the overall dynamical model is shown of figure 2-5. This block diagram includes the linear dynamics of the plant, disturbances \(G_w\), actuators frequency response \(G_{A_i}\) and actuators
2.2 Control Laws: Stochastic Linear Quadratic

A specificity of the staged controlled systems is that saturations need much more to be taken into account because they imply actuators with different strokes and bandwidth. This section addresses the question of using the optimal controller when nonlinearities of the actuators are significant.

2.2.1 Stochastic performance analysis for linear systems

This section describes the way a white noise $w$ is forwarded into the output of an uncontrolled system, given by

$$
\dot{x} = Ax + B_1w \\
y = Cx.
$$

The mean of $x$ can be found using the following equation:

$$
\dot{\bar{x}} = A\bar{x} + B_1\bar{w}
$$

The covariance matrix, defined by $\Sigma_{xx}(t) = E\{[x(t) - \bar{x}][x(t) - \bar{x}]^T\}$, is defined by the dynamics:

$$
\dot{\Sigma}_{xx} = A\Sigma_{xx} + \Sigma_{xx}A^T + B_1B_1^T
$$
If $A$ is a Hurwitz matrix, the mean $\bar{x}$ converges to $\bar{x} = -A^{-1}B_1\bar{w}$ and the covariance converges to a constant matrix given by

$$A\Sigma_{xx} + \Sigma_{xx}A^T + B_1B_1^T = 0 \quad (2.19)$$

Hence, the output covariance can be calculated as

$$\Sigma_{yy} = C_1\Sigma_{xx}C_1^T. \quad (2.20)$$

The mean output value is $\bar{y} = -C_1A^{-1}B_1\bar{w}$, which is zero under the assumption that $w$ is zero-mean. The root mean square (RMS) deviation of each output is the square root of the corresponding diagonal element of $\Sigma_{yy}$, and in the case of one output:

$$\sigma_{yy} = \sqrt{C_1\Sigma_{xx}C_1^T}. \quad (2.21)$$

If the system were ideally linear, the technique described above would be enough to predict the performances of a feedback control law. However, they are not appropriate anymore when considering real actuators with non-linearities, such as saturation, quantization or resolution limitation.

### 2.2.2 Stochastic performance analysis including non-linear actuators

The closed-loop system in nonlinear case is described by the following equations:

$$dx = Axdt + B_1dw + B_2\phi(u)dt \quad (2.22)$$

$$y = C_1x \quad (2.23)$$

$$u = Kx \quad (2.24)$$

Due to the nonlinearity $\phi(u)$ the close-loop dynamics are nonlinear stochastic differential equations. In order to determine the root mean square (RMS) output deviation we need to find the probability density function (PDF), which is described by the Fokker-Planck equation:

$$\frac{\partial p_x}{\partial t} = -\sum_{i=1}^{n} \frac{\partial}{\partial x_i} \left[ \sum_{j=1}^{n} A_{ij}x_j + \sum_{k=1}^{n} B_{2i} \phi(k)x \right] \frac{\partial p_x}{\partial x_i} + \frac{1}{2} \sum_{i=1}^{n} \sum_{j=1}^{n} \frac{\partial^2(B_1B_1^T)}{\partial x_i \partial x_j} \frac{\partial^2 p_x}{\partial x_i \partial x_j} \quad (2.25)$$

One can note that, in the case of $\phi(u) = u$, we go back to the solution found in equation 2.21. Unfortunately, there is no analytical solution to this equation in the general case, even in the steady-state limit. However, a steady-state solution has been found in the case where the nonlinearity is a
saturation or quantization [8]. The resulting PDF is a piecewise Gaussian function of the form:

\[ p_x(x) = \gamma e^{-\frac{1}{2}x^TQ^{-1}x - \xi \int_0^x \phi(v)dv} \]  

(2.26)

It has been shown [10] that this PDF can be closely approximated by a single Gaussian distribution even with large quantization level. The central question addressed below is to find a way of analytically determine the best Gaussian approximation directly from the system dynamics and nonlinear characteristics of the actuators.

### 2.2.3 Stochastic linearization and stochastic linearized \( H_2 \) controller

The stochastic linearization is a tool used to approximate the nonlinear system by a quasi-linear one. Quasi-linearization is quite different from ordinary linearization. In both approximation strategies an equivalent gain is used to replace the static nonlinearity. However, rather than a constant gain depending only on the equilibrium point, quasi-linearization gain is a function of the statistics of the input to the nonlinear component. A theorem associated to stochastic linearization states: The response \( v = \phi(u) \) of a single-valued nonlinear component \( \phi(u) \), driven by a zero-mean, stationary Gaussian process \( u \) can be approximated by an equivalent linear, time invariant filter \( N \) that minimizes the mean-squared approximation error \( \min \), with:

\[ N(u) = \frac{E\{u\phi(u)\}}{E\{u^2\}} = \frac{E\{u\phi(u)\}}{\sigma_u^2} \]  

(2.27)

Furthermore, it has been shown [6] that:

\[ N(u) = E\{\phi'(u)\} \]  

(2.28)

This theorem can be used to predict the performance of the closed loop system 2.22, but furthermore, when coupled with the LQR or LQG theory, it can lead to the design of an optimal controller taking into account the nonlinearities of the actuators. This stochastic linearized \( H_2 \) controller (SLQR or SLQG) has been developed in [10] and turned the control synthesis into the constrained nonlinear optimization:

\[ K = \arg\min \{ \text{tr}\{C_1\hat{\Sigma}_{xx}C_1^T\} + \text{tr}\{D_{12}K\hat{\Sigma}_{xx}K^TD_{12}^T\} \} \]  

(2.29)
with the steady-state variance defined by the following equations:

\[
(A + B_2N(\hat{\sigma}_u)K)\hat{\Sigma}_{\alpha\alpha} + \hat{\Sigma}_{\alpha\alpha}(A + B_2N(\hat{\sigma}_u)K)^T = -B_1B_1^T \tag{2.30}
\]

\[
\hat{\sigma}_{\alpha_i} = K_i\hat{\Sigma}_{\alpha\alpha}K_i^T \tag{2.31}
\]
Chapter 3

SIMO Testbed

3.1 Overview

The MIT Space System Laboratory, in collaboration with Aurora Flight Sciences, built a testbed destined to help development and prove the feasibility of the Stellar interferometry with independent light collectors. This testbed is part of the Synthetic Imager Maneuvering Optimization (SIMO) program and addresses the problem of the control of the optical path-length difference (OPD), which has to be kept close to zero with a nanometer-level precision. This is required to allow the combiner to interfere the same wavefront coming from the star and thereby create a coherent fringe. The control of the OPD is achieved by an Optical Delay Line (ODL) inserted between the combiner and one of the collectors. An interferometer receiver measures the OPL through to a laser beam going from the combiner to the two collecting satellites. This measurement is used to feed a control loop (known as phasing loop) regulating the ODL.

The testbed can be seen on figure 3-1. It is made up of two SPHERES satellites floating on an air table, assumed to be the star light collectors and one fixed plant, assumed to be the combiner satellite. The SPHERES satellite, capable of large motions, is the first stage of the phasing loop. It is supposed to maintain the desired position as precisely as possible, while the ODL reduces the remaining error from centimeter to nanometer scale. The ODL is connected to the combiner and consists in two stages of actuators: A voice coil actuated mirror for coarse control and a Piezo actuated mirror for fine control. Beside the phasing loop, a pointing loop has to be performed in order to keep the laser beam pointing to the center of the interferometer receiver. This loop is actuated by a Fast Steering Mirror attached to the collecting satellite, while the sensing is performed
by a CCD camera connected to the combiner.

The SIMO testbed can be seen as two quasi-symmetrical sides. Each of them contains a collecting satellite, a pointing system and an OPL measurement system. The only difference is the ODL added on one path. For simplicity reason and as a first step, the system has been reduced to only one side, the path containing the ODL. Once this system is working, there should not be any major issue to append the second side to the system.

**Performances Requirements**

The requirements are defined in order to make the SIMO testbed working as a Michelson interferometer. The first requirement concerns the pointing and allows the measurement to occur. Indeed, the interferometer receiver requires the reference and measurement beam to overlap with a lateral offset tolerance of 1/4 of the beam diameter. This value can be converted into an angular precision for the laser beam. Furthermore, as the laser beam is reflected on the mirrors of the pointing stage, the mechanical precision of the pointing stage has to be half of this value. Considering a beam diameter of 6 mm and a separation between the pointing stage and the interferometer receiver of 1m, the mechanical precision has to be:

\[
\alpha = \arctan\left(\frac{0.006/4}{1}\right) = 0.043^\circ
\]  

(3.1)

The precision requirement for the phasing control is defined by the necessity of observing interference fringes. That means that the Optical Path Length has to be controlled with a precision of some fractions of the light wavelength \(\lambda\) with a minimum of \(\lambda/4\). As the laser wavelength used is \(\lambda = 633\) nm, the OPL precision requirement is:

\[
\Delta OPL = \lambda/4 = 158.25\, nm.
\]  

(3.2)
3.2 Hardware Description

3.2.1 Laser beam

Laser

The laser used in this testbed is a Helium-Neon Agilent 5517B Laser Head. It generates a coherent and collimated light beam divided in two parts of the same power: \( f_1 \) horizontally polarized, and \( f_2 \) vertically polarized with a frequency slightly higher than \( f_1 \). The difference of frequency, known as "Reference Frequency", is measured and sent to the Interferometer Receiver. The part of the beam going directly to the Interferometer Receiver is named "Reference Beam", whereas the one going to the SPHERES is the "Measurement Beam". When measuring the difference between the two beam frequencies and after comparison with the reference frequency, the path-length difference variation can be deduced thanks to the Doppler effect. One can note that motions of the laser head along the beam axis do not disturb the measurement since only the difference between the two beams is taken into account in the receiver.
Interferometer Receiver

The interferometer receiver is an Agilent 10780F Remote Receiver and it is used to convert the Doppler-shifted laser light into an electrical signal. It is composed of two parts connected with optic fiber (see figure 3-2). The lens and polarizer block brings the two beams to the same polarization (45°), otherwise no interference happens, and sends the beam through the optic fiber to the electronics housing, which converts it into an electrical signal. The system is capable of a resolution of 0.15 nm.

Optics

In this testbed, the following optics are used:

- The mirrors can be fixed or attached to actuators. They are used to change the direction of the beam in the first case and to change the optical path length in the second.

- The 50% power beam splitter passes half the power and reflects the other half 90°. Its role in the testbed is to bring part of the beam to the camera used for pointing.

- The polarizing beam splitter separates vertical and horizontal polarizations. One passes, whereas the other is reflected in a 90° angle. One polarizing beam splitter is used as polarizing filter in front of the camera, the reflecting part being not used. Another one is part of the single beam interferometer.

- The single beam interferometer is a polarizing beam splitter together with a retroreflector and two quarter-wave plates (QWP). It can be seen on figure 3-3. The goal is to separate the measurement beam from the reference one, and to bring them back together after the measurement beam has been reflected on the target. The way it works is described in the
The reference beam, horizontally polarized, is reflected, goes through the QWP transforming its polarization into circular, is reflected by the retroreflector, goes through the QWP again, transforming its polarization into vertical, and passes the polarizing beam splitter towards the receiver. The measurement beam, vertically polarized, is transmitted, goes through the QWP and continues towards the Satellite. On its return, after passing a second time the QWP, its polarization is horizontal. It is thus reflected towards the receiver.

### 3.2.2 Phasing

The three stages of the phasing loop are described in this part and are represented on figure 3-4.
Coarse actuator: SPHERES

The SPHERES is the coarsest phasing stage. Its objective is only to keep the desired position and attitude. To achieve this, the satellite uses its own measurement (IMU & metrology) and propulsion systems, and acts independently of the other stages. The attitude and position controller have been designed in previous works [13] and a simple tuning of the gains has to be performed. They are separate PIDs, controlling independently each axes, since the SPHERES is quite symmetrical. The controllers convert position and attitude errors into forces and torques, and the SPHERES determines which thrusters have to fire and for how long [12]. The control period of both position and attitude controllers is 500 ms. The performance in closed-loop is around $\pm 1\text{ cm}$ for position and $\pm 0.5^\circ$ in attitude. Besides, the measurement precision of the SPHERES is $\pm 1.5\text{ mm}$ in position and $\pm 0.5^\circ$ in attitude.

Medium stage actuator: Voice coil

The medium stage of phasing is performed by a Bruel & Kjaer Mini-shaker Type 4810, actuated by voice coil. An outline of this actuator is shown in figure 3-6. The force applied to the moving part is proportional to the current through the coils. Thanks to the flexures, the position is proportional to the current in the static case. The maximum current in the coils is $\pm 1.8\text{ A}$, leading to a maximum motion of $\pm 3\text{ mm}$. The command is expressed in Volts and sent through a DAC to a Crown DC 300-A series II audio amplifier. In the dynamic case, the equivalent resistance in the voice coil changes. If we want to keep the linearity between the command and the force applied, the voltage has to be adapted in order to keep the current constant. Hence, a feedback loop has been designed
[2] to keep the current constant through the voice coil. The time constant of this loop is around 0.05 ms (20 kHz). Regarding to the working frequencies, it is thus acceptable to consider that the current is proportional to the voltage command. The theoretical gain has been calculated, based on the electrical static gain and the flexures stiffness, leading to $G = 1.96 \text{ mm/A}$. Experimental data show that the displacement is not perfectly linearly proportional to the current, but rather like:

$$G = \frac{\text{displacement}[\text{mm}]}{\text{current}[\text{A}]} = 2.84 - 0.84I$$  \hspace{1cm} (3.3)

The resolution of the actuator is limited by the 16 bits DAC, that means 0.1 $\mu$m. However, the vibrations induced by the ground and the mirror bearing system let it increase.

**Fine stage actuator: Piezo**

The phasing fine stage is a Physik Instrumente S-325 piezoelectric fast steering mirror, shown in figure 3-7. Since this actuator has been designed for precise steering mirror, it contains three piezoelectric linear pistons located within a 25 mm diameter circle separated 120° between each other, every piezo is driven by a Physik Instrumente E-660.OE amplifier. This configuration allows motions of the mirror in tip, tilt and translation perpendicular to the mirror face. In our case, since only the translation is required, the three piezoelectric actuators have to move simultaneously in order to place the mirror in the desired position along the piezo axis. The data-sheet indicates a range of 30 $\mu$m and a resolution of 0.5 nm. Theoretically, it would be enough to give the same
voltage to the three piezo, to obtain a piston movement. However, the characteristics are not exactly the same, and it is necessary to apply slightly different voltage to avoid any tip or tilt. For that purpose, a system using one voltage supply distributed into the three piezo has been designed with variable resistances [2]. A diagram of this voltage supply is shown in figure 3-8.

3.2.3 Pointing

The objective of the pointing system (see figure 3-9) is to keep the laser beam centered in in the interferometer receiver. As the SPHERES attitude controller is not precise enough to accomplish a steady correct pointing, a fine stage has been added. Hence, the Satellite’s mirrors are not fixed on the SPHERES, but they can be precisely rotated. This fine stage is the Fast Steering Mirror (FSM). The sensing for this FSM is provided by a CCD camera or by the SPHERES metrology when the camera’s pictures are not available. This work has been mostly performed in [1], but modifications and updates became necessary due to the testbed configuration changes.

The pointing system outline is:

- The CCD camera images the laser dot on a screen and sends it through USB.
- The position of the beam is extracted from the picture thanks to a centroid calculation and...
sent to the SPHERES through a UART connection.

- The SPHERES compares the centroid position to the desired position of the beam and calculates the command to be sent to the FSM. Furthermore, the Linear Stage position command is calculated using the SPHERES metrology system.

- Data are sent through a UART connection to the PIC 18F8722 Microcontroller, which commands the FSM and the LS.

A diagram of the communication system can be seen in figure 3-10. With this configuration a cable connects the satellite to the camera, which will not be allowed in the final testbed since no mechanical link is possible due to the distance in the mission configuration. Figure 3-11 shows the communication diagram as it is supposed to work finally. The difference is that the centroid is sent to the SPHERES through the wireless SPHERES communication systems, rather than through the UART connection. However, since a cable was already necessary to connect the satellite to the off-board electronics, the configuration with wires was chosen for simplicity.
Coarse pointing stage: SPHERES attitude control

In the same manner as the phasing control, the coarse pointing stage is performed by the SPHERES attitude controller. The SPHERES is supposed to hold a constant attitude. The performance when the attitude closed-loop is run is around 0.5°.

Fine pointing stage: Fast Steering Mirror

The errors in satellite’s attitude are corrected by the fast steering mirrors (FSM) shown on figure 3-12. Those are supposed to make the laser beam going precisely into the Interferometer Receiver. When the FSM are properly aligned, the incoming and returning beams are collinear. The FSM are composed of two mirrors controlling independently the vertical and horizontal angles. They are actuated by voice coil and thus are controlled by a differential voltage. The FSM is mechanically connected to the satellite via the Linear Stage. The FSM is powered and controlled through the SPHERES expansion port. Data is sent through an RS-232 connection to a PIC 18F8722 micro-controller. The PIC receives the UART data (after conversion to TTL levels) and then converts it into the necessary form for the MAX 5523 Digital to Analog Converter (DAC) and outputs the data through the SPI port. The DAC then connects to two LM 4861 amplifiers. The characteristics of the FSM are as follow [9]:

Natural Frequency: 69 Hz
Range: ± 5.3°
Resolution: 0.01°
Camera

The FSM uses measurements of a High Resolution USB 2.0 Board Camera (Videology). The laser beam reflecting from the SPHERES goes through a power beam splitter before entering the Interferometer Receiver. Half of the power hits a white screen, which is observed from behind by the camera. The path can be seen in figure 3-14. The camera has been placed as close as possible to the interferometer receiver in order to represent reliably the position of the beam in the Receiver. The camera needs to see only the beam coming from the satellite, thus the reference beam coming directly from the interferometer has to be hidden from the camera. Hence, a polarizing filter has been used to prevent the reference beam to enter the screen (see setup in figure 3-13). The picture is sent to a computer, where a Matlab code calculates the centroid of the red color intensity, using the formula:

\[
\text{centroid}_x = \frac{\sum_{y=1}^{ny} x I_y}{\sum_{y=1}^{ny} I_y}
\]

\[
\text{centroid}_y = \frac{\sum_{x=1}^{nx} y I_x}{\sum_{x=1}^{nx} I_x}
\]

where \( I \) is the intensity of pixel \((x,y)\), with a total of \((nx,ny)\) pixels in the CCD.

Fine positioning stage: Linear stage

As the FSM size is smaller than the precision that the SPHERES position and attitude controller can achieve, a mechanical system is used to move the FSM in order to keep it inside the laser beam. Thus, the FSM has been fixed on a linear stage able to translate perpendicularly to the laser beam. A Newmark Microslide, Model MS-1-24 shown in figure 3-15, has been chosen to perform this
Figure 3-13: View of the polarization filter, screen and camera setup.

Figure 3-14: Part of the laser beam path, showing the position of the camera.
Figure 3-15: Newmark Microslide linear stage used for FSM fine positioning.

The characteristics of this linear stage are as follows:

Max Velocity: 1 cm/s
Range: ± 1.27 cm
Resolution: 0.02 µm

The stage is driven by a NSC-1S single axis stepper motor controller. The input to the controller is RS-232 and there are two outputs: One provides power to the linear stage’s motor, and the other provides the signal to control the movement.
Chapter 4

System Modeling

This section aims to find a linear model to the two actuators of the Optical Delay Line. A linearization of the voice coil is also achieved and the transfer functions found here will be used for the design of the controllers.

4.1 Voice Coil Actuator

A state model has to be calculated in order to design the controller. Both mechanical part and electrical part have to be characterized. As mentioned in section 3.2.2, the electrical circuit providing the voice coil with the current has a bandwidth of 20 kHz, which is far more than the working frequencies of the actuator. The transfer function from the command to the force can be approximated by a static gain without loss of precision. The mirror’s support is connected to the mini-shaker’s frame with flexures (see figure 3-6). The mechanical part of the voice coil can thus be modeled as a simple one dimension mass-spring system, with forces acting on it. The mechanical parameters of this system (mass, stiffness and damping coefficient) are given by the data-sheet:

\[
\begin{align*}
m_{vc} &= 68g \\
k_{vc} &= 2 \frac{N}{mm} \\
c_{vc} &= 233.2 \frac{Ns}{m}
\end{align*}
\]
The state model and transfer function can be calculated:

\[
\begin{bmatrix}
\dot{x}_{vc} \\
\dot{v}_{vc}
\end{bmatrix}
= \begin{bmatrix}
0 & 1 \\
-\omega^2_{vc} & -2\xi_{vc}\omega_{vc}
\end{bmatrix}
\begin{bmatrix}
x_{vc} \\
v_{vc}
\end{bmatrix}
+ \begin{bmatrix}
0 \\
G\omega^2_{vc}
\end{bmatrix} u
\] (4.1)

\[
F(s) = \frac{G}{1 + \frac{2\xi_{vc}}{\omega_{vc}} s + \frac{1}{\omega^2_{vc}}}
\] (4.2)

with: \(\omega_{vc} = \sqrt{\frac{k_{vc}}{m_{vc}}}\) and \(\xi_{vc} = \frac{c_{vc}}{2\sqrt{k_{vc}m_{vc}}}\). The mirror is not directly attached to the moving part of the voice coil, but a support is used to prevent the mirror to tilt. Indeed, the mini-shaker allows small rotations of the moving part, leading to a misalignment of the laser beam. At the beginning this thesis, a sliding bearing was used to transmit translations of the mirror without transmitting any tilt. It consisted in a 10cm-rod sliding inside a hole.

A linearity analysis has been performed on this bearing. For this purpose, a low frequency sinus command is sent to the voice coil, and the output amplitude is measured. Figure 4-1 shows the result for several amplitudes commands up to 1 mm. Note that the input is the displacement amplitude when using the gain normalized around 0.5 mm.

![Figure 4-1: Linearity analysis of the sliding bearing.](image)

It appears that this bearing prevents the mirror to move when the command is less than 100 \(\mu m\). This nonlinearity does definitely not fit the requirement of the voice coil precision, since it is supposed to be less than 30 \(\mu m\) as explained in section (see 5.1.1). Figure 4-2 shows the track of the voice coil when a closed-loop is applied. It becomes obvious that the origin of the nonlinearity is a static friction. Given that the stiffness of the voice coil’s flexures are 2 \(N/mm\), and as the mirror
starts to move when the input is more than 100 $\mu m$, it can be deduced that the static friction of the bearing is around 0.2 $N$.

![Figure 4-2: Sinus reference tracking by the voice coil in closed-loop when the mirror is driven by the sliding bearing.](image)

**Voice Coil Linearization**

Several options have been explored in order to avoid this static friction nonlinearity:

- **Rolling-element bearing**

  The first option was to use a commercial bearing with rolling elements as seen in figure 4-3. In this case, the static friction is quite lower compare to the sliding bearing. However, the desired precision of the voice coil is in the order of magnitude of 1 $\mu m$, that means that the static friction has to be around 2 $mN$. Furthermore, it is necessary that the lower stage of control, the piezo, can compensate the lack of precision of the voice coil. That means that no "jump" in the voice coil mirror track is allowed, since they could not be corrected by the piezo. No such bearing can achieve such performance if using sliding or rolling elements.

- **Micro-scale vibrations**

  In some applications, the static friction has been released using micro-scale vibrations. However, two problems prevent this process to be applied. First the required bandwidth can not be reach by the voice coil since limited to 30 Hz. Secondly, this method cannot fulfill the requirement of nanometer precision.
• **Planar spring**

   The only way to avoid any static friction is to use a support whose motions are provided by flexion. The idea came to use a structure based on planar springs as shown in figure 4-5. These spring allow displacements only in one-axis translation with a very low stiffness and a big rigidity regarding to the other directions. Two of them have been connected with a rigid foam in order to have a light bearing preventing any rotation. The final assembly can be seen on figure 4-5.

   A linearity analysis has been performed on this bearing the same way as on the sliding one. The result can be seen on figure 4-6 with a considerable improvement of the linearity. Indeed, the precision is now slightly lower than 1 µm, which is 100 times better than the one of the sliding bearing. The precision fits now the 30 µm overlap requirement of the staged control.

   The new transfer function cannot easily be calculated since the parameters of the new bearing are not available. Hence, the transfer function has been determined by identification. For that purpose, the gain and phase of the voice coil response are measured for a grid of frequencies. The measured transfer function is plot in figure 4-7. The measured transfer function is then approximated by the analytical function that fits the best. The result is a second order transfer function.
Figure 4-5: Bearing using planar spring assembly.

Figure 4-6: Linearity analysis: Comparison between the sliding-based and flexion-based bearings.
described in 4.3. Note that the static gain is not 1, since the command has been normalized around 0.5 mm of amplitude. Due to a minor nonlinearity, the frequency response, measured with input of 0.1 mm, has a slightly different static gain.

\[
F_{vc}(s) = \frac{G_{vc}}{1 + \frac{2\xi_{vc}}{\omega_{vc}}s + \frac{1}{\omega_{vc}^2}s^2}
\]  

(4.3)

with:

\[
\omega_{vc} = 30Hz
\]
\[
G_{vc} = 0.8
\]
\[
\xi_{vc} = 0.28.
\]

### 4.2 Piezo Actuator

The mirror actuated by the piezo is directly attached on the moving support of the piezo. Hence, there is no problem of static friction as seen in figure 4-9. The position is linearly proportional to the command and the transfer function of both the electrical and mechanical parts can be calculated.
Figure 4-8: Curve fitting of the measured voice coil transfer function.

Figure 4-9: Linearity analysis of the piezo actuator.
Calculation of the electrical characteristics of the piezo actuator has been achieved in [2], leading to a bandwidth of 30 Hz. The mechanical part of piezo can be modeled as a simple one dimension mass-spring system, just like the voice coil. The mass and stiffness of the piezo actuator are given by the data-sheet:

\[
\begin{align*}
    m_{pz} &= 28 \text{g} \\
    k_{pz} &= 3474 \frac{N}{mm}
\end{align*}
\]

Consequently, the bandwidth of the mechanics is \( \omega_{pz} = \sqrt{\frac{k_{pz}}{m_{pz}}} = 11 \text{ kHz} \). This frequency is well beyond the working frequencies, since the electrical bandwidth is limited to 30 Hz. Therefore only the electrical dynamics has to be taken into account in the dynamic model.

In order to confirm the bandwidth of the whole actuator and to define the transfer function, an identification has been performed the in the same manner as for the voice coil. The measured transfer function is plot in figure 4-10 and the curve fitting on figure 4-11. The piezo actuator transfer function is best approximated by:

\[
F_{pz}(s) = \frac{A_{pz}}{1 + \frac{1}{\omega_{pz}} s}
\]

with:

\[
\begin{align*}
    \omega_{pz} &= 27 \text{Hz} \\
    A_{pz} &= 1
\end{align*}
\]

Hence, the actuator’s state model can be deduced:

\[
\begin{bmatrix}
    \dot{x}_{pz} \\
    x_{pz}
\end{bmatrix} = \begin{bmatrix}
    -\omega_{pz} & 1 \\
    A_{pz} \omega_{pz}
\end{bmatrix} u
\]

(4.5)
Figure 4-10: Frequency response of the piezo actuator.

Figure 4-11: Curve fitting of the measured piezo transfer function.
Chapter 5

Controller Design

The design of a staged control for the ODL is the main part of this thesis. In this chapter a staged control structure is designed, followed by a description of the process used to tune the different stages of control. The main challenges of this controller is the large ratio between the disturbance amplitude and the error amplitude tolerance. Furthermore, the controller has to deal with a control frequency of only 200 Hz, which is relatively low compared to the frequency of the observed noise, that is to say around 30 Hz. The work will focus on the phasing loop, as the pointing loop has mostly been performed in previous work.

5.1 Phasing Loop

The objective of the phasing loop is to bring the sub centimeter precision level of the SPHERES position control system as close a possible to the 150 nanometer precision required for the SIMO testbed. The SPHERES position and attitude controller is independent of the other stages of control, since using its own metrology system and actuators for keeping position. The two lower control stage are linked by the fact, that the measurement they use - the interferometer receiver - is a combination of their two positions. Indeed the value given by the interferometer receiver is:

\[
OPL_{meas} = \begin{bmatrix} g_{SP} & g_{vc} & g_{pz} \end{bmatrix} \begin{bmatrix} x_{SP} \\ x_{vc} \\ x_{pz} \end{bmatrix} + d_{optics}
\] (5.1)

With:
• $OPL_{\text{meas}}$ is the measurement of the OPL given by the interferometer receiver. Note that the real measurement is the beam path-length including the return, that means twice the OPL. We assume that the $\frac{1}{2}$ correction factor has been applied.

• $x_{\text{sp}}, x_{\text{vc}}, x_{\text{pz}}$ are the position of the SPHERES, voice coil and piezo respectively.

• $g_{\text{sp}}, g_{\text{vc}}, g_{\text{pz}}$ are the geometrical coefficients of the SPHERES, voice coil and piezo respectively. These coefficients convert the position of an actuator into the difference of OPL, depending of the angle of the incoming beam and on the number of hits.

• $d_{\text{optics}}$ are the disturbances due to the optics vibrations.

Note that the interferometer receiver noise is very small since its resolution is 0.5 nm. It will thus be neglected in the following.

5.1.1 Staged control

The specific constraint imposed by the SIMO testbed, in term of controls, is the combination of a large stroke and very small error tolerance. Qualitatively, following a reference in the order of magnitude of 1 centimeter with a tolerance of around 100 nanometer means a controller able of a noise reduction of $10^5$. Furthermore, the bandwidth must be wide enough to reject all kind of disturbances expected to appear, like thruster firing or reaction wheels vibrations. This would be straightforward if a single actuator would exist with such performances. However, it is well known that high stroke actuators usually have a limited resolution and bandwidth, leading to a limited precision in dynamic conditions.

The principle of stage control is to use complementary actuators: An upper-stage with large stroke and limited resolution and bandwidth and a lower-stage with high resolution and bandwidth but limited stroke. Thus, each upper control stage only has to avoid saturation of the lower-stage, by bringing the error under the maximum stroke of the lower-stage. It becomes obvious that a necessary condition is that the resolution of an upper-stage overlaps the maximum stroke of its lower-stage. In our case:

1. The SPHERES satellite is used to bring precision down to 1 cm level.

2. The voice coil reduces the precision to 10 $\mu$m level.

3. The Piezo decreases the precision to the final 100 nm level.
Stroke-Resolution Overlap

Before designing any controller, the feasibility of a stage control has to be verified. Figure 5-1a shows the mechanical range of action of the three actuators on a logarithmic scale. The upper limit is the maximum stroke and the lower is the resolution. The interesting value to be taken into account is not the displacement of the mirror, but the actual OPL variation when moving this mirror. Figure 5-1b indicates the resolution and maximum stroke in terms of OPL, which is roughly twice the mechanical motion for the voice coil and the piezo. Indeed, if a mirror moves a certain distance, the beam reflecting on it covers once the distance before reflecting and once after. This is not the case on the SPHERES, where the mechanical displacement corresponds to the OPL variation.

As seen on figure 5-1b, the overlap requirement is not fulfilled between the voice coil and the SPHERES. No stage control law would solve this problem and a hardware change has to be performed. The idea came to double the voice coil maximum stroke by doubling the number of reflections on it. A fix mirror has thus been added in front of the voice coil, as drawn on figure 5-1c. This new configuration allows the overlap requirement to be fulfilled at the price of two drawbacks: First, while doubling the maximum stroke, the resolution gets doubled as well. In this case, this is not a problem, given the large overlap between the voice coil and the piezo. Secondly the noise injected in the OPL by the voice coil is doubled as well. This point was a problem, since the voice coil resonant frequency is around 30 Hz and vibrations coming from the ground tend to excite it, leading to large oscillations of the voice coil mirror. This excitation has been removed by isolating the testbed from the ground thanks to the floating table.

Staged control structure

As explained above, the goal of an upper stage is to decrease the error below the maximum stroke of the lower stage in order to avoid its saturation. This is a first step and can be achieved using independent feedback for each actuator. In this case, the input of every stage is the OPL error and the actuators act in parallel in order to reduce the error. A block diagram of this stage control structure is shown on figure 5-2a. The problem of this simple structure is that the lower stage tends to be continuously saturated, as noticed in [2]. The step response of such a staged control structure with two parallel PIDs has been calculated with Simulink and can be seen on figure 5-2b. The lower stage reaches quickly its saturation and stays on it, while the upper stage corrects alone the remaining error. Hence, the overall performance is the one of the upper stage and the lower stage becomes useless.
Figure 5-1: Strokes and resolutions for staged control.
A new structure, inspired by the work performed in [7] and [3], has been designed with a better saturation avoidance. The action is not parallel anymore, but the role of each stage of control is different. The upper stage tends to correct alone the OPL error, that means without accepting the contribution of the lower stage for corrections it is able to perform by itself. This is achieved by feeding the upper stage controller with the OPL measurement subtracted by the lower stage position. A block diagram of this cascaded structure is drawn on figure 5-3a. Hence, the calculation of the upper stage error gives:

\[ E_{vc} = OPL_{ref} - (OPL_{meas} - P_{pc}) = OPL_{ref} - (P_{vc} + P_{pc} + D_{optics} - P_{pc}) = (OPL_{ref} - D_{optics}) - P_{vc} \]  

(5.2)

That means that the upper stage’s target position is the configuration where the OPL error is zero if the lower stage position would be zero. This tends to reduce the error while naturally centering the lower stage. It is worth noting that this upper-loop is independent of the the lower stage position (if assuming a good knowledge of its position). The error of the lower stage is calculated below:

\[ E_{pz} = OPL_{ref} - OPL_{meas} = OPL_{ref} - (P_{vc} + P_{pz} + D_{optics}) = (OPL_{ref} - P_{vc} - D_{optics}) - P_{pz} = E_{vc} - P_{pz} \]

(5.3)

That means that the lower stage is a closed-loop with the upper stage error as a reference. The
objective of the lower stage is thus to correct only the error that the upper stage is not able to correct (static/drag error, lack of resolution or lack of bandwidth), which is the best correction that a staged control law can achieve. A step response, with the same controller as used on figure 5-2b but with the cascaded structure, is plot on figure 5-3b. The result is better in the sense that the piezo is now centered in the equilibrium state. Comparison between the two step responses is plotted on figure 5-4, showing similar shapes, but the cascaded structure will be more able to correct high frequency disturbances, since the piezo is not in saturation.
Figure 5-5: Block diagram of the cascaded staged control structure with piezo estimator and feedforward.

Figure 5-6: Position of the piezo, overlaid by its estimation, when driven by an open-loop sinus command.

**Estimator and feedforward**

In the case of the SIMO testbed, no direct measurement of the piezo position is available. Hence, the feedback has to be driven by an estimation of this position. The OPL measurement cannot be used for that purpose, since the optics and voice coil disturbances are some orders of magnitude bigger than the piezo displacement. Therefore, the position of the piezo is estimated only using the input command and a model of the actuator. This model is the first order found in section 4.2 with the appropriate saturation appended. As the model is close from the actual actuator, the approximation is good, as it can be seen on figure 5-6. Besides, a feedforward of the reference OPL has been added before the voice coil actuator. This does not affect the closed-loop stability, but allows to follow roughly the reference independently of the controller. Finally, the staged controller structure can be seen in figure 5-5.
5.1.2 Piezo

The piezo controller has been designed by poles placement of the closed-loop transfer function. The reason of not using a LQ design method is that the constraint for this kind of control is not the trade-off between precision and command, since only the precision is sought. The limitation is rather given by the control frequency, 200 Hz, which requires reasonable margins in order to keep both, stability and performances. Hence, a parametrization with bandwidth would be more convenient than one with weights on command and precision. When considering the controller structure showed on figure 5-7a, the closed-loop transfer function is:

\[
CL_{pc} = \frac{K_{pc}F_{pc}}{1 + K_{pc}F_{pc}}
\]  

(5.4)

As the piezo actuator transfer function \( F_{pc} \) is a first order function, a proportional-integral controller is sufficient to place the two poles of the closed-loop transfer function and to prevent any static error. Block diagrams of the actuator and its controller are showed in figures 5-7b and 5-7c. When using a PI controller under the form \( K_{pc} = k_p + \frac{k_i}{s} \), and after injecting the piezo transfer
function $G_{pz}$ as defined in 4.4, the closed-loop function becomes:

$$CL_{pz} = \frac{A_{pz}(k_ps + k_i)}{\omega_{pz}s^2 + (A_{pz}k_p + 1)s + A_{pz}k_i}$$  \hspace{1cm} (5.5)

The closed-loop dynamics can thus be imposed by choosing the appropriate proportional and integral gains. After identification of the denominator of equation 5.5 with the desired characteristic equation $s^2 + 2\chi_{CLS} + \omega_{CL}^2 = 0$, the gains can be calculated depending on the chosen natural frequency and damping of the closed-loop complex conjugate poles:

$$k_p = \frac{1}{A_{pz}} \left( 1 - 2\chi_{CL} \frac{\omega_{CL}}{\omega_{pz}} \right)$$ \hspace{1cm} (5.6)$$

$$k_i = -\frac{\omega_{CL}^2}{A_{pz} \omega_{pz}}$$ \hspace{1cm} (5.7)

### 5.1.3 Voice coil

The gains of the voice coil controller have also been set by poles placement of the closed-loop transfer function. Again, with purpose of preventing any static error, an integrator has to be used in the controller. Hence, a PID controller becomes necessary in order to be able to place the three poles of the closed-loop system (second order actuator and integrator). Block diagrams of the actuator and its controller are showed on figures 5-8b and 5-8c.

Assuming a perfect measurement of the velocity ($v_{vc} = p_{vc}s$), the transfer function of the PID controller under the form showed in figure 5-8c is:

$$u_{PID} = \begin{bmatrix} k_v s - k_p - k_i \frac{1}{3} & k_p + k_i \frac{1}{3} \end{bmatrix} \begin{bmatrix} p_{vc} \\ R_{vc} \end{bmatrix}$$ \hspace{1cm} (5.8)

where $u_{PID}$ is the command sent to the voice coil, $p_{vc}$ is the position of the actuator and $R_{vc}$ is the reference position. After injecting the transfer function of the voice coil actuator found in 4.3, the closed-loop can be calculated as being:

$$CL_{vc} = \frac{k_ps + k_i}{s^3 + (2\xi_{vc} \omega_{vc} - k_v)s^2 + (\omega_{vc}^2 + k_p)s + k_i}$$ \hspace{1cm} (5.9)

The closed-loop dynamics can thus be imposed by choosing the appropriate control gains. The
Figure 5-8: Synthesis model of the voice coil controller.
The target characteristic equation for this 3-poles transfer function is a second order

\[(s^2 + 2\xi_{CL}\omega_{CL}s + \omega_{CL}^2)(s + \alpha\omega_{CL}) = 0\] (5.10)

with \(\omega_{CL}\) and \(\xi_{CL}\) the natural frequency and damping of the closed-loop, and \(\alpha\omega_{CL}\) the natural frequency of the third pole. Choosing specific values of these three parameters yields appropriate controller gains, via the following equations, which are obtained by identification of the coefficients.

\[k_p = 1 - (2\alpha\xi_{CL} + 1) \frac{\omega_{CL}^2}{\omega_{ve}^2}\] (5.11)

\[k_v = \frac{2}{\omega_{ve}} (\chi_{ve} - \chi_{CL}) \frac{\omega_{CL}}{\omega_{ve}} - \alpha \frac{\omega_{CL}}{\omega_{ve}^2}\] (5.12)

\[k_i = \alpha \omega_{CL}^2 \frac{\omega_{ve}^2}{\omega_{ve}^2}\] (5.13)

5.1.4 Discret-time controller conversion

The controllers have been designed in continuous, but the actual controllers will be implemented on a computer and will thus be discrete-time. The computer used for the control code has a maximum control frequency of 200 Hz. The controllers have been discretized taking this control period, \(\Delta t\), into account. The discrete-time controller implemented is the one given by the following formula [11]:

\[U(n) = k'_p E(n) + k'_v (E(n) - E(n-1)) + k'_i \sum_{j=0}^{n} E(j)\] (5.14)

with \(U(n)\) and \(E(n)\) the command and error at \(n^{th}\) control period, and:

\[k'_p = k_p\]
\[k'_v = \frac{1}{\Delta t} k_v\]
\[k'_i = \Delta t k_i\]

A control frequency of 200 Hz is a very limiting factor, since the sought error attenuation factor is around 100 for each stage. Indeed, the 5 ms induced delay has a considerable influence on the precision for high frequencies. In order to have a more qualitative idea of the loss of precision, the following calculations are performed to determine the error of a perfect reference follower with only a small \(\Delta t\) delay. When considering a sinus reference with amplitude \(A\) and frequency \(f\), the
position is:

\[ p(t) = R(t - \Delta t) = A \cdot \sin(2\pi f(t - \Delta t)) \] (5.15)

The error between the reference and the position is then defined by:

\[ E(t) = A \cdot \sin(2\pi ft) - A \cdot \sin(2\pi f(t - \Delta t)) = [2 \sin \pi f \Delta t] A \cos(2\pi ft - \pi f \Delta t) \] (5.16)

That is to say that the error is equal to the reference with a factor \([2 \sin \pi f \Delta t]\) and a phase delay. A plot of this factor is shown for \(\Delta t = 5\) ms on a semi-logarithmic scale on figure 5-9. The main observation is that, whatever the actuator is, a noise cannot be reduced since its frequency is more than 33 Hz. Also the maximum attenuation of a 3 Hz noise will be \(\frac{1}{10}\). The conclusion is that this control periods prevents the correction of any noise with frequency higher than 33 Hz. The 200 Hz control frequency has thus to be absolutely increased, since it is highly limiting the precision.

### 5.2 Pointing Loop

The pointing loop aims to keep the laser beam pointing into the interferometer receiver. One loop is used for the control of the FSM while another loop controls the linear stage. Here a brief description of the controllers designed by John M. Field is given; for further description refer to [4].
5.2.1 Linear stage

The linear stage has its own non-changeable controller, driven by a reference position. The desired position is calculated using the error in attitude and position given by the SPHERES metrology system and IMU. Geometrical calculations (see figure 5-10), with small angles approximation gives the following formula:

$$U_{LS} = dy \cdot \sin \alpha + dx \cdot \cos \alpha + D_m \cdot \tan d\theta$$  

where $dx$, $dy$ and $d\theta$ are the error in the SPHERE positioning. $\alpha$ is the angle between the laser beam direction and the y-axis and $D_m$ is the distance from the FSM to the center of the SPHERES.

The reference position can be sent as soon as the error in position is known, that means that the maximum control frequency is the IMU measurement frequency, that is to say 20 Hz.

5.2.2 Fast steering mirrors

The FSM has two different modes depending on whether the camera can see the laser beam or not. If the beam is too far away from the screen, the only information about FSM angle is given by the position of the SPHERES. In this case, the position of the FSM is controlled using the rotation error given by the SPHERES metrology system with a proportional controller:

$$u_k = K_p Z_k$$  

where $Z_k$ is the error in the satellite’s vertical rotation, $K_p$ is the proportional gain and $u_k$ is the command at time $k$. When the beam is close enough to its reference position and can fit the screen, the camera is able to extract a centroid from the picture. The centroid error is then used to feed an
integral controller in the form of the following equation:

\[ u_k = u_{k-1} + K_i E_k \]  \hspace{1cm} (5.19)

where \( K_i \) is the integral gain and \( E_k \) is the error between the measured centroid and the reference centroid. Note that the camera detects the presence of the beam on the screen by comparing the level of red-intensity with a threshold value found experimentally.
Chapter 6

Experimental Results

6.1 Controller Tuning

6.1.1 Voice coil

The design of the voice coil PID has been performed using the poles placement described in section 5.1.3. The choice of the three parameters $\omega_{CL}$, $\alpha$ and $\chi_{CL}$ is limited by the fact that the zero of the closed-loop is not controlled and an unstable zero could occur, leading to an inverse initial direction. Furthermore, as the real controller is a discrete-time, the delay margin $\Delta_m$ has to be several times bigger than the control period in order to keep both, the stability and the performances. The design process consisted in two phases: A coarse tuning with the poles placement formulas and robustness calculation, followed by a fine tuning based on the step response of a discrete-time Simulink simulator. Finally, the setting found is:

$$k_p = -0.1, k_v = -0.003, k_i = -45$$

the simulated discrete-time step-response is plotted on figure 6-1 and the closed-loop performances are listen bellow:

$$\omega_{CL} = 28 \ Hz$$
$$\alpha = 0.3$$
$$\chi_{CL} = 0.45$$
$$\Delta_m = 37 \ ms$$
Besides, a hand-tuned low frequency double integrator has been added in order to overcome
the drag-error while following the sinus-reference. Furthermore, an anti-windup has been imple-
menting by limiting the maximum value of the integral and double-integral terms.

6.1.2 Piezo

The design of the piezo PI controller followed the same procedure as the voice coil. After the
tuning process the following gains are used:

\[ k_p = -0.55, \; k_i = -112 \]

Again, the simulated discrete-time step-response is plotted (figure 6-2) and the closed-loop
performances are listen bellow. One can notice that the damping is lower than 0.95, this is due to the
fact that the discretization decreases the performances. An anti-windup has also been implementing
by limiting the maximum value of the integral term.

\[ \omega_{CL} = 22 \; Hz \]
\[ \chi_{CL} = 0.95 \]
\[ \Delta_m = 43 \; ms \]

One can note that the closed-loop frequency of the piezo is lower than the closed-loop fre-
quency of the voice coil. This is not surprising if we compare the frequencies of the actuators in
open-loop: the ratio is almost the same. As the lower stage is supposed to reject the high frequency
noise, the piezo actuator need to have an higher bandwidth. This is actually the case if we consider
the mechanical bandwidth. It is thus necessary to provide the piezo with a decent amplifier whose
bandwidth is adapted to the purpose of rejecting vibration noises.
6.2 Phasing Staged Control

In order to test the phasing stage control independently from the pointing, three tests have been performed with the SPHERES replaced by a fixed mirror. The motion of the SPHERES are replaced by an OPL sinus-reference. The first test has a constant reference, that means that the ODL only needs to reject the noise coming from the optics. The remaining error will be the target noise for the other tests, since a better performance cannot be achieved with a varying reference. In the second test, the OPL has to follow a reference of a couple of millimeters with a low frequency. This is supposed to represent the motions of the SPHERES when controlled by the thrusters. The reference of third test has a higher frequency with an amplitude of a couple of $\mu$m only. This test aims to evaluate the ability of the ODL to reject the vibrations of the different mirrors.

6.2.1 Optics noise reduction

The measured optics noise, that can be seen in figure 6-3a, has a RMS of 500nm. Except an extremely low frequency drift, most of the PSD is located around 30Hz. This main frequency is highlight on figure 6-3b, which is a zoom on the optics noise. This noise comes from the oscillations of the voice coil bearing, excited at its natural frequency by the ground vibrations. Indeed, the linear spring supporting the mirror are very weak in order to let a free motion to the voice coil actuator. Previously to this measurement, the amplitude of this noise has been decreased by a factor 4 by making the whole testbed floating on an air table.

The phasing loop has been run in order to reduce this noise. The result is shown in figure 6-4. The noise RMS has been decreased to 250 nm. This error can not be reduced, since it is composed of frequencies close or higher than 30 Hz and the correction of such high frequencies is limited by the 5 ms delay of the controller discretization, as explained in section 5.1.4. The consequence
of this test is that the control frequency has to be necessarily increased in order to fit the precision requirements.

### 6.2.2 Low frequency references

The second test has been performed with a sinus reference of 0.5 rad/s frequency and 4 mm amplitude. This reference has a speed close to the average speed of the SPHERES when the closed-loop is activated. Figure 6-5a shows the OPL trajectory during 1 minute of control and 6-5b is a zoom on the OPL error. In this figure some peaks of error appear regularly with varying amplitude. One can notice that the maximum peak amplitude occurs when the reference speed is maximal, while there is almost no peak when the reference is constant. Figure 6-6 gives an explanation to these peaks. The control period has been plotted on the same figure and it can be noticed that there are some regular peaks in the control period coincidental with the peaks of error. This is due to the fact that the computer used to control the OPL operates not actually in real-time and the processor is required for other activities every 300 milliseconds. During this timeframe, the control period is not 5 ms, but 16 ms. This has an big effect on the precision of the control, especially when the derivative of the reference is important. In fact, an absence of control during 16 ms, while the reference is varying at 1 mm/s speed, brings an error of 16 μm, which is roughly the actual size of the peak. The Root Mean Square (RMS) of the error is 2400 nm, which is well beyond the requirement. However, the deeper zoom on the error plotted on figure 6-5c gives an idea of what would be this error amplitude without these peaks, that means with real-time computing. The RMS would rather be around 250 nm, which is the amplitude of noise when the reference is zero. This test highlights the necessity of using a real-time operating system for the control of the OPL.

### 6.2.3 High frequency references

The third test has been run for high frequency and low amplitude reference. The OPL track for a sinus of 10 microns amplitude and a frequency of almost 10 Hz can be seen in figure 6-7. The result is unsatisfactory, considering that the error goes up to ±2000 nm, that is to say to the amount of 40% of the reference. According to the figure 5-9, the 5 ms delay on a 10 Hz frequency reference leads to an error of more than 30% of the reference amplitude. It can be concluded that the major part of this error is due to the low frequency of control.
(a) Noise of the optics during 100s.

(b) Zoom on the optics noise with highlight on the 30Hz component.

(c) Power Spectral Density of the optics noise.

Figure 6-3: Noise of the optics.
6.3 Pointing Control

The two actuators were able to perform their task independently, but a disfunction occurred when they were acting together. The control of the linear stage is partially successful, considering that the correction of the SPHERES positioning error is effective, allowing the laser beam to hit the FSM most of the time. Nevertheless, an unexpected issue appeared when the pointing loop combined both, the linear stage and the FSM. The mirrors of the FSM are connected to their support with very weak springs, they are consequently very sensitive to disturbances. This sensitivity becomes a problem when the linear stage accelerates quickly, transmitting torques to the mirrors and leading to a misalignment of the beam. The NSC-1S Newmark Systems controller used to drive the linear stage has been chosen for its good precision provided by an high bandwidth. Hence, for each reference position received at the control frequency (20 Hz), the linear stage jumps to its new position, injecting micro-impulses into the FSM. These impulses tend to excite the natural frequency of the FSM (69 Hz). The induced vibrations cannot be corrected, since their frequency is well beyond the frequency of the centroid measurement provided by the camera, which is the only available measurement of the FSM position.

As it is not possible to access the gains of the NSC-1S Newmark Systems controller, the only way of reducing the energy of the impulses would be to decrease the length of the step between two reference positions. The following shows however that it is not possible to have a sufficient speed for the linear stage without inducing excessive vibrations. The minimum distance per step can be derived when assuming a maximum rotation speed of the SPHERES during closed-loop control of 0.02 rad/s and a maximum control frequency of 20 Hz (that is to say every time the IMU data is
Figure 6-5: Staged control trajectory when reference is a sinus with 4 mm amplitude and 0.5 rad/s frequency.
Figure 6-6: Staged control trajectory with an emphasis on the correlation between the error peaks and the control period peaks. Note that the OPL has been scaled down in order to be more readable.

Figure 6-7: Staged control trajectory when reference is a sinus with 10 microns amplitude and 60 rad/s frequency.
Note that 192 mm correspond to the distance from the FSM to the center of the SPHERES. Besides, experiment showed that with 0.1 mm per step the induced vibration still prevents good measurements of the OPL.

It can be concluded that the Newmark Microslide linear stage motor driven by the NSC-1S Motion Controller is not an appropriate actuator, since it is not capable of smooth motions. A new definition set of requirements can be defined, with performances fitting better with the objective of the actuator:

- The current resolution of 0.02 µm can be released up to the order of magnitude of 0.1 mm.
- The actuator has to be able to reach a speed of 0.5 mm/s.
- The bandwidth of the actuator has to be well below 69 Hz in order to prevent excitation of the FSM.

6.4 Floating SPHERES

The final objective of the SIMO testbed is to run both – the phasing and the pointing loops – while the satellite is floating and controlled. However, as explained above, the linear stage prevents the pointing loop to be precise enough because of the vibrations of the fast steering mirrors induced by the linear stage’s jerks. Hence, the linear stage cannot be used while getting measurements. An option has thus been explored of dividing the test time into two phases switching regularly from one to another. One phase is a measurement phase during which the linear stage is switched off while the phasing loop and the FSM control occur. The other phase is the desaturation and capture phase during which, the linear stage first places the FSM centered in the laser beam and then the FSM brings the beam back inside the interferometer. However, it turned out that the desaturation phase has to be activated too frequently and no measurement can be recorded. The frequent need of desaturation is exacerbated by the presence of the wires bringing the power and commands from the amplifier to the pointing stage (see figure 3-10). These wires link the SPHERES to an electrical test-board placed on the ground and lead to a set of forces and torques disturbing the control of the satellites.
Hence, it can be concluded that this linear stage cannot be used to keep the FSM centered in the beam. The extent of the test has to be reduced taking this insight into account. The floating satellite’s motions have been limited in one-dimension translation. For this purpose, a bar has been fixed on the floating table. A diminutive tilted position of the air table lets the satellite laying slightly against the bar. In this condition the satellite is only controlled along the bar's direction. This test tends to demonstrate the ability of the phasing loop to correct the macroscopic motions of the satellite in floating and controlled conditions. Even if the satellite’s rotation is limited by the bar, small rotations occur which makes the control of the FSM, especially during the thrusters firing time. Nevertheless, this test does not validate the pointing loop since the rotation and lateral shift of the satellite is highly limited.

The result of this test is shown on figures 6-8. The 12 seconds test shows that the ODL is able to correct the displacement of 3 mm. Figures 6-8b and 6-8c show the same phenomena as seen on the results with fixed mirrors: A precision of about 500 nm can be achieved disturbed by huge peaks due to the absence of real-time control, increasing the RMS error up 1200 nm.

Another conclusion of this test regarding to the validation of the controller is demonstrated in figures 6-8a and 6-8b. The track of the voice coil and piezo commands attest the efficiency of the staged control structure, since most of the large correction is performed by the voice coil, while the piezo rejects the high frequency, low amplitude error. It is worth noting that the piezo keeps an average position of zero, thanks to the constant centering action of the voice coil, as designed in section 5.1.1.
(a) Overview of the OPL and the piezo and voice coil commands.

(b) Zoom on the OPL and the command of the piezo.

(c) Highlight of the OPL only.

Figure 6-8: Results of the control of the OPL with a one-dimension floating SPHERES.
Chapter 7

Conclusion and Future Work

In the present master’s thesis the control of optical path length in the context of the space interferometric telescopes has been performed. The major outcome of this thesis is the collection of data arose from tests, demonstrating the feasibility of controlling the optical path length from the ground to a floating object with a 500 nm precision while this object is slightly moving with thrusters. This is a first step for the completion of a full ground demonstrator for the space interferometry missions. Furthermore, the limiting features have been determined and a solution has been proposed for each of them.

The main contribution to the project is the design and implementation of the ODL controller. In this thesis, a cascaded staged control structure has been designed with the objective of centering the piezo. One of the main issues of the previous controller implemented in SIMO for the ODL was the constant saturation of the lower-stage. Hence, the overall performance was reduced to the one of the upper stage. Thanks to this new structure, the error is now distributed to the two controllers in the following manner: The upper stage reject all the noise that it is able to reject alone, while the lower stage only rejects the remaining noise, non-controllable by the voice coil. This design allows a natural desaturation of the lower stage and an optimal use of the specific performances of each actuator. Furthermore, a tuning methodology for the two actuators has been proposed by poles placements. After implementation in SIMO, tests have been performed, validating with hardware-in-the-loop the efficiency of the designed control law.

Besides, a couple of hardware modification have been performed in order to fulfill the requirement of overlap between the resolution of an upper stage and the stroke of the lower stage. The voice coil actuator has been linearized by replacing its support with a frictionless one. The increase of precision filled the gap between the voice coil and the piezo. The overlap between the voice coil
and the SPHERES has been completed by adding an extra-mirror across from the voice coil actuator (see figure 6-5), doubling the voice coil stroke. Moreover, a new configuration of the optics and sensors allowed an increase of the pointing system precision (see figure 3-14).

A set of recommendations are proposed for the future work, in order to improve the phasing-loop and to make the pointing-loop working. It can be summarized by the following items:

- According to the cascaded structure, the lower stage aims to rejects the high frequencies noise, like optics vibrations induced by the thruster or by the ground, while the upper stage corrects the low frequency displacements of the satellite. However, the current piezo actuator has roughly the same bandwidth as the voice coil because of its amplifier. The amplifier of the piezo has to be changed in order to access the large bandwidth of the piezo and take full advantage of the cascaded control structure.

- The main limitations, noticed in the previous tests, to the ODL precision is related to the control period. The control is currently accomplished by a Windows computer without real-time operation and the minimum control period reached is 5 ms with a peak up to 20 ms every 300 ms. It has been shown that this frequency is too short for the rejection of mechanical vibrations with the precision required for an ODL. This precision will be reached only when using at least a real-time operating system, or even a micro-controller for optimal performances.

- The current pointing-loop cannot be run since the FSM vibrates when the linear stage is actuated. Its motor is driven by a NSC-1S Motion Controller and is highly precise due to its high closed-loop bandwidth. The consequence is a displacement by jumps, leading to an excitation of the FSM and a loss of measurements. Calculation and experiment showed, that when reducing the size of the steps at the minimum required speed, induced vibrations still prevent good measurement to appear. This stage has to be replaced by another one with smoother displacements. A new set of performance requirements has been defined.

Once the linear stage have been changed, it will be possible to control the ODL with the floating SPHERE and thereby to permit the validation of the pointing loops. After the completion of the above recommendations, the precision achieved has to be tested in order to validate the phasing loop with the 150 nanometers precision sought for the SIMO program. The last step will be to include the part relative to the second collecting satellite in order to run the final test consisting in using the ODL for keeping the two optical paths at the same length and thereby run a demonstrator for space interferometry missions in full conditions.
Bibliography


