Control of an experimental piezoelectric actuators system

J.A.Dzoleko Dongmo

CST 2010.060

Master’s Thesis

Coach: Ir. T. Porck
Supervisor: Prof. Dr. Ir. M. Steinbuch

NXP

Eindhoven University of Technology
Departement of Mechanical Engineering
Control Systems Technology Group

Eindhoven, August, 2010
Abstract

The use of piezoelectric actuator is a growing technology in the domain of nano-motion. However, a problem involving piezoelectric actuators is the open loop hysteresis. This thesis presents the dynamic behavior of a system, which makes use of two piezoelectric actuators coupled in an antagonistic way. To solve the hysteresis exhibited by the system, the configuration and the performance of a current/charge control amplifier are described. To monitor the dynamic behavior of the system in real time, a self-sensing (method that allows using a piezoelectric actuator as sensor and actuator at the same time) method is designed. Furthermore, the current/charge control amplifier is combined to the self-sensing. This combination is done to monitor the system dynamic behavior while getting rid of the hysteresis at the same time. Finally, the use of the self-sensing method for closed-loop feedback control purpose is explored. A positive position feedback is implemented to damp a resonant mode of the piezoelectric actuators system.
Summary

ITEC (Industrial Technology and Engineering Centre) is a division of NXP (Next eXPerience), it is dedicated to design and develop equipment used in the assembly of discrete semiconductors. One of the equipment in the production line is the Phicom wire bonder, which is used for interconnecting the die to the lead frame.

The wire bonder uses at this moment makes use of a resonant piezoelectric operating at a fixed frequency of 60 kHz. In order to have a wire bonder operating from 50 kHz up to 200 kHz, a lightweight push-pull system that makes use of two off-resonant piezoelectric actuators. Both piezoelectric actuators are actuated such that, when one piezoelectric actuator is extending, the other piezoelectric actuator is contracting (then one speak about push-pull or antagonistic system). However, off-resonant piezoelectric actuator tends to exhibit important nonlinearities (hysteresis).

In this thesis, a theoretical modeling of the push-pull system is evaluated to have an idea about the mode (eigenvalues) of the system. A voltage amplifier is designed to actuate high capacitive load such as piezoelectric actuators.

To get rid of the hysteresis, a feedback control configuration is developed to convert the voltage amplifier into a current/charge control amplifier. By using the current control amplifier, the goal is to control the amount of current flowing through the piezoelectric actuator, instead of controlling the voltage across the piezoelectric actuator as done with voltage amplifier. The performances of the current/charge amplifier are evaluated in the frequency domain (FRF) and in the time domain (Hysteresis contour).

A self-sensing method (the same piezoelectric actuator is used as sensor and actuator at the same time) is implemented for bonding process monitoring purpose and feedback control purpose. In the self-sensing method, a bridge network (Wheatstone) is used to extract a signal that is proportional to the mechanical displacement or velocity of push-pull structure. To overcome the piezoelectric capacitance change that might unbalance the bridge, an adaptive filter (LMS algorithm) is added to the self-sensing network to insure the stability.

The current control amplifier is combined with the self-sensing to associate the advantage of both techniques, which is actuating piezoelectric actuators with current control while monitoring the dynamic behavior at the same time.
Finally, the self-sensing is explored for closed loop feedback purpose. A positive position feedback (PPF) controller is implemented to damp a resonant mode of the piezoelectric actuators system (push-pull system).
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Chapter 1

Introduction

1.1 Background

ITEC (Industrial Technology and Engineering Centre) is a division of NXP (Next eXPerience) semiconductors, it is dedicated to design and develop equipment used in the assembly of discrete semiconductors.

The production line of semiconductors involves many processes such as; the product assembly, the packaging and the testing. The line starts with wafers cut in dies. The first step involves the ADAT apparatus, which attach the die to the lead frame connection point by a process of pick and place. The following step involves the Phicom apparatus, which interconnect the die upper part to the lead frame connection point by a process of wire bonding. The next step involves plunger apparatus, where the lead frame is molded into a package. The last step involves the testing of the chip.

The subject of this assignment is related to the Phicom apparatus and the wire bonding process. The wire bonding is used in microelectronics industry as means of interconnecting (via wire) chips, substrates and output pins. The most frequently used method of joining the wire is the ultrasonic welding. The wire bonding process uses combination of ultrasonic vibration (typical around 60 KHz) and force to effectively scrub the interface between wire and substrate, causing a localized temperature rise that generates the diffusion of metals. Although, Phicom does only ball-wedge wire bond, another type of wire bond common in the industry is the wedge-wedge. A complete description of the wire bond is shown in Figure 1.1, where the ball bond is made from the deformation of the free ball air at the tail of the wire and the wedge bond is made from the deformation of the wire and the lead frame surface.
The ultrasonic vibration is typically produced by piezoelectric transducers. The piezoelectric transducer comprises piezoelectric material that converts electrical energy to mechanical energy and vice versa. In the case of producing ultrasonic vibration energy, an electric field is applied to the piezoelectric ceramic to stimulate vibration.

Commonly in the industry, resonant piezoelectric transducers are used to produce ultrasonic vibrations at a specific frequency. A phase lock loop generator is then used to lock the system at the operating frequency. The ultrasonic transducer comprises several elements such as; the piezoelectric discs, the horn, the capillary, the bolt and the clamping. The structure of a resonant piezoelectric transducer is shown in Figure 1.2, where an electrical ultrasonic power is converted to ultrasonic vibration by the piezoelectric disc.

The vibration is a longitudinal compression wave along the body of the ultrasonic piezoelectric. That vibration is amplified by the horn, which lead in an amplified oscillation at the tip of the capillary. The clamping is used for fixation of the piezoelectric transducer to the wire bonder machine. The capillary is used for welding by scrubbing the wire (through its axial hole) and the substrate to form a bond.

1.2 Motivations and objectives

In [Tijs96] the investigations about the influence of higher wire bonding frequencies over the bond quality and the bonding time are evaluated. Those evaluations are done at three distinct frequencies from 60 kHz to 120 kHz. It's come out that, the bond quality and the bonding time is improved at higher frequencies bonding. In the perspective of further
extend the investigations of [Tijs96], ITEC decides to extend the experimentation till a frequency of 200 kHz. A resonant piezoelectric transducer system has the advantage of low driving voltage and low power consumption. Beside those advantages, some disadvantages can be mentioned:

- due to the resonant nature of the system, the system can operate on a single frequency only and the system take time to build up the amplitude,
- the moving mass is relatively heavy.

Due to these disadvantages, the objectives of ITEC is to design a lightweight system that does not make use of a resonance to achieve the required amplitude. The targeted specifications are:

- total transducer system mass of 10 gr,
- selectable operating ultrasonic frequency from 50 KHz to 200 KHz,
- selectable amplitude up to 1 μm in the operating ultrasonic frequency.

### 1.3 Piezoelectric structure description

The system described in this part is designed to provide a repetitive horizontal displacement of one degree on freedom (see Figure 1.3). The system is built around two piezoelectric stack actuators (P1 and P2) mounted in an antagonistic way. Both piezoelectric are facing each other and operating in the push-pull motion. Both piezoelectric actuators are strongly coupled such that when one is extending, the opposing one is contracting. The masses (M1 and M2) are used as counter-masses for the actuator forces. The springs (S1 and S2) are used for the pretension of both piezoelectric actuators. Preload force of 50% is applied to obtain symmetrical push-pull performances in dynamics operations. The holder element (H) is designed to hold the capillary element.

![Figure 1.3: push-pull ultrasonic piezoelectric systems.](image)
1.4 Problem statement and assignment definition

Although, non resonant piezoelectric stacks bring some advantages in term of weight and frequency, but it does imply too some performances limitations such as:

- non-linearity; the piezoelectric stack actuator is known to exhibit hysteresis at high voltage as the relationship between the voltage and the displacement is not linear,
- high power driving; driving piezoelectric stack actuator at high voltages and high frequencies implies a high demand in term of current and power supply,
- high frequencies for sensing quantization; any control technique implying feedback ask for high resolutions data acquisition systems,
- high temperature heating; big temperature heating has tendency to change the capacitance of the load. So any control technique based on the load capacitance is challenging.

Therefore, the problem statement in this assignment resume to:

Design a control system that is able to control the ultrasonic piezoelectric actuators system, by getting rid of the non-linearity inherent to the piezoelectric actuators, by monitoring the dynamic behavior of the piezoelectric system in real time and by optimizing the modal behavior of the piezoelectric actuator system.

From the above problem statement, the assignment for this project can be formulated as:

- Design and implement the electrical power driving strategies for the new transducer system.
- Define the suitable control method to eliminate high harmonics. Due to the non-linear behavior of the piezoelectric actuators, there will be higher harmonics present beside the chosen driving frequency. It should be explored which methods can be used to solve this.
- Develop a method for wire bonding monitoring; for process control, gathering information during bonding is important to get reliable bonds. It should be studied which sensing method best fit in this application.
- Investigate control techniques in order to damp the resonant mode of the piezoelectric system.
1.5 Literature survey in the control of piezoelectric stack actuator

The use of piezoelectric actuator is a growing technology in the domain of nano-positioning and nano-motion. However, a problem regularly related in the literature involving the piezoelectric stack actuator is the open loop hysteresis (see [ACPI] and [IEEE88]). Many methods have been developed in the literature, having purpose of solving the hysteresis of the piezoelectric stack actuators. These methods can be divided into two categories. It can be cited method dealing with open loop control, where no sensor is needed and method dealing with closed-loop feedback control, where sensor is needed.

In the open loop control method, one can refer to the theory of inverse control [Ru06]. The hysteresis is modeled and the open loop controller is done with the inverse of the hysteresis model in a feed-forward manner. The model of the piezoelectric actuator hysteresis can be identified based on Preisach model for hysteresis (see [Ping95] and [Card00]), based on Newton iterative procedure [Ru07] or LMS adaptive algorithm [Ru06]. The controller is implemented through DSP systems, which assumes having a DSP system with a sample frequency much more high than the frequency of the controlled systems.

Another method mentioned in the category of open loop control is the charge control actuation, where a charge-control driver is used to actuate the piezoelectric transducer instead of a voltage driver. The hysteresis is compensated by regulating the amount of charge or current flowing through the piezoelectric actuator (see [Coms81] and [Yi05]). The difficulty of this method is that one has to dive into electronic considerations and design.

In the category of closed loop feedback control method, [Mann00] combines feedback velocity, DSP systems and sensors to control vibrations on a flexible structure. Similarly, [Liu09] uses complementary sensors and robust control technique to implement a tracking control of a nanopositionner. In the same order of idea, [Ping96], [Song05] and [Lean**] combine feed-forward control loop and feedback control loop in a computer-based tracking control to
compensate for the piezoelectric actuator hysteresis. They make use of a displacement sensor and a dSPACE processing system to implement a tracking control. The feedforward controller makes use of the inverse model of the hysteresis (Preisach model) and the feedback controller makes use of conventional PID or lead-lag controller.

![Figure 1.6: feedforward and feedback control for piezoelectric actuator hysteresis.](image)

[Dosc92] developed a self-sensing piezoelectric actuator system, where the properties of the piezoelectric materials are used to make the piezoelectric actuator operates simultaneously as actuator and sensor. Many researchers associate later on that self-sensing technique to known controller technique in a feedback loop control. Thus, [Hago94] investigated the performances of the self-sensing technique under closed loop. Here the self-sensing is combined with several controllers such as Linear Quadratic Gaussian (LQG), positive position feedback (PPF) and strain rate feedback (SRF). Similarly [Jone98] uses self-sensing coupled to a velocity estimator controller for tracking control of a PZT Micropositioner.

![Figure 1.7: piezoelectric actuator with self-sensing control.](image)

1.6 Assignment approach

By considering the literature survey above and the assignment requirements, this assignment will explore the open loop and closed loop control technique in order to control the piezoelectric actuator. Due to the high operating frequencies of the piezoelectric actuators systems, DSP systems and sensors are not conceivable because of the cost and the time consuming that is required. The open loop control will be investigated as it does not make use of any sensor. A charge control driver will be designed and implemented. A self-sensing technique coupled to a controller and a feedback loop is designed and implemented as well.
An overview of the complete assignment is shown in Figure 1.8

1.7 Report outline

This report starts with a complete description of the stack piezoelectric actuator in chapter 2. In chapter 3, control methods related to piezoelectric actuators are discussed and the design of the power voltage driver and charge amplifier driver is presented in chapter 4. The fourth chapter will discuss about the self-sensing design and the LMS algorithm implementation and the feedback control. The results performances of both open loop and closed loop method are evaluated in chapter 5. This report ends with the conclusion in chapter 6.
Chapter 2

Modeling of the piezoelectric actuators structure

The chapter introduces the general theory about piezoelectric materials. It considers the constitutive equation of the piezoelectric single layer and the finite element modeling of the multi-layers piezoelectric stack actuators. An analytical modeling of the complete structure with included piezoelectric stack actuators is derived.

2.1 Model of the structure with included piezoelectric stack actuators

The model of the piezoelectric actuators structure comprises three masses and two piezoelectric actuators as displayed in Figure 2.1. The blocks masses M1 and M3 represent the counter-masses, when M2 represents the lumped mass from holder and capillary. P1 and P2 represent both piezoelectric stack actuators. In the structure, the only forces considered are forces produced by the piezoelectric transducer as the preload forces are setup for dynamics equilibrium. The modeling of the push-pull piezoelectric actuators will limit to the horizontal displacement behavior. It will particularly focus on the push-pull mode despite the fact there exists additional modes such as capillary mode and counter mass mode (for further modeling description see Timm10).

In order to derive the equations that govern the structure, the equations governing single piezoelectric stack actuator is derived by means of constitutive equations and finite element modeling.
2.2 Piezoelectric material

Piezoelectric materials are smart materials that have the ability to respond to various kinds of stimulation. A piezoelectric material exhibits two effects: the direct piezoelectric effect and the inverse piezoelectric effect.

In the direct piezoelectric effect shown on Figure 2.2, an application of mechanical stresses on the piezoelectric body generates electric charge on its surface thus converting mechanical energy to electrical energy. In the inverse piezoelectric effect shown on Figure 2.3, an electric voltage applied to the electrodes leads to the piezoelectric material deformation thus converting electrical energy to mechanical energy. A piezoelectric stack operates in the thickness mode or $d_{33}$ mode as the electric field is parallel to the poling direction and the material deformation parallel to the electric field.

Figure 2.1: model of the piezoelectric actuators structure.

Figure 2.2: direct piezoelectric effect.

Figure 2.3: inverse piezoelectric effect.
2.3 Piezoelectric Constitutive equations

The piezoelectric material stress, strain, displacement and electric field can be related by a single pair of electromechanical equations. There are many equivalent constitutive equations and the choice should depend on the application. Piezoelectric material constitutive equations standardized in [IEEE88] are formulated as follow:

\[
\{S\} = [s^E]\{T\} + [d]\{E\} \\
\{D\} = [d^T]\{T\} + [e^T]\{E\} \tag{1}
\]

or

\[
\{T\} = [c^E]\{S\} + [e]\{E\} \\
[D] = [e]\{S\} + [e^S]\{E\} \tag{2}
\]

Where \( S \) is the strain tensor, \( T \) is the stress tensor, \( s \) is the compliance tensor, \( d \) and \( e \) are the piezoelectric constants, \( E \) is the electric field, \( D \) is the electric displacement and \( \varepsilon \) is the permittivity. The superscripts \( E \) and \( T \) indicate that the values of the constants are obtained at constant electric field and constant stress respectively. The relation between \( [e] \) and \( [d] \) is \([e] = [d][c]\).

By considering one dimensional geometry and a thickness mode piezoelectric material the set of equations in (2) becomes

\[
T_3 = c_{33}^E S_3 + e_{33} E_3 \\
D_3 = e_{33} S_3 + e_{33}^S E_3 \tag{3}
\]

Having defined the constitutive equations of a single layer piezoelectric material, the equations of a multi-layers piezoelectric stack actuator operating in the thickness mode can be derived.

2.4 Finite element modeling of multi-layers piezoelectric stack actuators

2.4.1 Finite element method formulation

The finite element equations can be derived from the Hamilton principle, in which the Lagrangian and the virtual work are adapted to include the mechanical and the electrical contribution. The potential energy density of the piezoelectric material can be derived as;

\[
H = 1/2 \left[ \{S\}^T \cdot \{T\} - \{E\}^T \cdot \{D\} \right] \tag{4}
\]

as well the virtual work principle can be derived as;

\[
H = 1/2 \left[ \{S\}^T \cdot \{T\} - \{E\}^T \cdot \{D\} \right] \tag{5}
\]
\{F\} is the external force, \{u\} is the displacement, \phi is the electric potential and \sigma is the charge density. From equations (4) and (5) the analogy between the electrical and mechanical variable can be derived (see Table 2.4-1).

<table>
<thead>
<tr>
<th>Mechanical</th>
<th>Electrical</th>
</tr>
</thead>
<tbody>
<tr>
<td>Force {F}</td>
<td>\sigma \text{ charge}</td>
</tr>
<tr>
<td>Displacement {u}</td>
<td>\phi \text{ potential}</td>
</tr>
<tr>
<td>Stress {T}</td>
<td>{D} \text{ electric displacement}</td>
</tr>
<tr>
<td>Strain {S}</td>
<td>{E} \text{ electric field}</td>
</tr>
</tbody>
</table>

Table 2.4-1: analogy between electrical and mechanical variable

In the finite element formulation the displacement vector \{u\} and the electric potential \{\phi\} are related to nodal coordinates \{u_i\} and \{\phi_i\} by means of shape functions \[N_u\] and \[N_\phi\]

\[\{u\} = [N_u]\cdot\{u_i\}\]
\[\{\phi\} = [N_\phi]\cdot\{\phi_i\}\]

The strain field \{S\} and the electric field \{E\} can be expressed according to the nodal coordinates by means of shape functions as:

\[\{S\} = [L]\cdot[N_u]\cdot\{u_i\} = [B_u]\cdot\{u_i\}\]
\[\{E\} = -\nabla[N_\phi]\cdot\{\phi_i\} = [B_\phi]\cdot\{\phi_i\}\]  (6)

[L] is the linear differential operator matrix and \nabla is the gradient operator.

The variation principles governing the piezoelectric material follows from the substitution of \[H \text{ and } \delta W\] into the Hamilton principle (see [[Alli70]]). This results in two equilibrium equations governing the mechanical and electrical behavior of the piezoelectric material.

\[\{M_{uu}\}\ddot{\{u\}} + [K_{uu}]\cdot\{u_i\} + [K_{u\phi}]\cdot\{\phi_i\} = \{f_i\}\]
\[-[K_{u\phi}]\cdot\{u_i\} + [K_{\phi\phi}]\cdot\{\phi_i\} = \{q_i\}\]  (7)

With

\[\{M_{uu}\} = \int_V \rho \cdot [N_u]^T \cdot [N_u] \cdot dV\] mass matrix,
\[\{K_{uu}\} = \int_V [B_u]^T \cdot [e^s] \cdot [B_u] \cdot dV\] elastic stiffness matrix,
\[\{K_{u\phi}\} = \int_V [B_u]^T \cdot [e] \cdot [B_\phi] \cdot dV\] electromechanical stiffness matrix,
\[\{K_{\phi\phi}\} = \int_V [B_\phi]^T \cdot [e^s] \cdot [B_\phi] \cdot dV\] dielectric stiffness matrix,
\[\{f_i\}\] mechanical load force vector,
\[\{q_i\}\] charge vector.
2.4.2 Dynamic equations of free moving multi-layers piezoelectric stack actuator

The piezoelectric stack actuator consists of a number of piezoceramic layers electrically connected in parallel. The polarization is arranged in order to have a material deformation in the same direction for all the layers (see Figure 2.4). In finite element method the piezoelectric stack actuator is decomposed in basic mechanical element such as mass spring.

Assuming uniform electric field within the piezoelectric actuator and zero external force acting on the piezoelectric actuator then from equations (7) the dynamic and sensing equations are as follow (see Error! Reference source not found. for derivation):

\[
\begin{bmatrix}
2m & m \\
 m & 2m
\end{bmatrix}
\begin{bmatrix}
\ddot{u}_1 \\
\ddot{u}_n
\end{bmatrix} +
\begin{bmatrix}
k_v & -k_v \\
-k_v & k_v
\end{bmatrix}
\begin{bmatrix}
u_1 \\
u_n
\end{bmatrix} = 
\begin{bmatrix}
-k_{emv} \\
k_{emv}
\end{bmatrix}
\cdot \phi

\left[-k_{emv} & k_{emv}
\right]
\begin{bmatrix}
u_1 \\
u_n
\end{bmatrix} + C_p \cdot \phi = Q
\]

(8)

\[m = \frac{Apnl}{6}, k_v = \frac{A[e]}{nl}, k_{emv} = \frac{A[e]}{l} \text{ and } C_p = \frac{An[e^s]}{l}\]

If the electrodes of the piezoelectric transducer are short-circuited then the electric potential is \(\phi = 0\). Then the Eigenvalues problem is formulated as;

\[
\begin{bmatrix}
2m & m \\
 m & 2m
\end{bmatrix}
\begin{bmatrix}
s^2 \\
 s^2
\end{bmatrix} +
\begin{bmatrix}
k_v & -k_v \\
-k_v & k_v
\end{bmatrix}
\begin{bmatrix}
u_1 \\
u_n
\end{bmatrix} = 0
\]

(9)

From equation (9) the Eigenfrequencies of the multi-layers piezoelectric stack actuators are then derived as:

\[w_1 = 0 \text{ and } w_2 = \sqrt{\frac{2k_v}{m}}\]

The frequency response of the free clamping piezoelectric actuator is shown in Figure 2.5. It can remark that there are additional zeros in the FRF from the charge Q to the input voltage. Those zeros come from the feedthrough term in the charge equation.
2.5 Analytic modeling of the piezoelectric structure

The analytic model of the structure is derived according to the piezoelectric actuator equations (8). To analyze the system, the structure in Figure 2.1 is decomposed in two operating blocks, which are the extension block and the contraction block. The governing equation of each block is derived according to external inertia forces coming from masses M1, M2 and M3. Assuming sectioning the system in the middle of M2 then \( M^*_2 = \frac{M_2}{2} \)
The motion equation of the extension block is derived according to the voltage potential $\phi_1$ and the electric charge $Q_1$ acting on the extending piezoelectric stack actuator.

$\begin{bmatrix} 2m + M_1 & m \\ m & 2m + M_2 \end{bmatrix} \cdot \begin{bmatrix} \dot{X}_1 \\ \dot{X}_2 \end{bmatrix} + \begin{bmatrix} -k_v & -k_v \\ -k_v & k_v \end{bmatrix} \cdot \begin{bmatrix} X_1 \\ X_2 \end{bmatrix} = \begin{bmatrix} k_{env} \\ -k_{env} \end{bmatrix} \phi_1$

$-\begin{bmatrix} -k_{env} & k_{env} \\ k_{env} & -k_{env} \end{bmatrix} \cdot \begin{bmatrix} X_1 \\ X_2 \end{bmatrix} + C_p \cdot \phi_1 = Q_1$

The motion equation of the contraction block is derived according to the voltage potential $\phi_2$ and the electric charge $Q_2$ acting on the contracting piezoelectric stack actuator.

$\begin{bmatrix} 2m + M_2' & m \\ m & 2m + M_3 \end{bmatrix} \cdot \begin{bmatrix} \dot{X}_2 \\ \dot{X}_3 \end{bmatrix} + \begin{bmatrix} -k_v & -k_v \\ -k_v & k_v \end{bmatrix} \cdot \begin{bmatrix} X_2 \\ X_3 \end{bmatrix} = \begin{bmatrix} k_{env} \\ -k_{env} \end{bmatrix} \phi_2$

$-\begin{bmatrix} -k_{env} & k_{env} \\ k_{env} & -k_{env} \end{bmatrix} \cdot \begin{bmatrix} X_2 \\ X_3 \end{bmatrix} + C_p \cdot \phi_2 = Q_2$

The motion equation of the complete push-pull system results from combining the extension motion and the contraction motion. The push-pull operation is excited with; $\phi_1 = -\phi_2$

$\begin{bmatrix} 2m + M_1 & m & 0 \\ m & 4m + M_2 & m \\ 0 & m & 2m + M_3 \end{bmatrix} \cdot \begin{bmatrix} \dot{X}_1 \\ \dot{X}_2 \\ \dot{X}_3 \end{bmatrix} + \begin{bmatrix} -k_v & -k_v & 0 \\ -k_v & 2k_v & -k_v \\ 0 & -k_v & k_v \end{bmatrix} \cdot \begin{bmatrix} X_1 \\ X_2 \\ X_3 \end{bmatrix} = 0$

$\begin{bmatrix} k_{env} & 0 \\ -k_{env} & k_{env} \\ 0 & -k_{env} \end{bmatrix} \cdot \begin{bmatrix} \phi_1 \\ \phi_2 \end{bmatrix} = 0$

$\begin{bmatrix} k_{env} & -k_{env} & 0 \\ 0 & k_{env} & -k_{env} \end{bmatrix} \cdot \begin{bmatrix} X_1 \\ X_2 \\ X_3 \end{bmatrix} + \begin{bmatrix} C_p & 0 \\ 0 & C_p \end{bmatrix} \cdot \begin{bmatrix} \phi_1 \\ \phi_2 \end{bmatrix} = \begin{bmatrix} Q_1 \\ Q_2 \end{bmatrix}$

If the electrodes of the piezoelectric transducer are short-circuited then the electric potential is $\phi_1 = 0$ and $\phi_2 = 0$. The eigenvalues problem is formulated as;

$w = eig \left( \begin{bmatrix} 2m + M_1 & m & 0 \\ m & 4m + M_2 & m \\ 0 & m & 2m + M_3 \end{bmatrix}^{-1} \begin{bmatrix} -k_v & -k_v & 0 \\ -k_v & 2k_v & -k_v \\ 0 & -k_v & k_v \end{bmatrix} \right)$
Consider the fact that the mass of the piezoelectric actuator can be neglected according to the mass of the counter-mass and the fact that \( M_1 = M_3 \), then the eigenfrequencies of the system are:

\[
\begin{align*}
w_1 &= 0 & \text{rigid body mode; all masses moving in the same direction} \\
\frac{w_2}{M_1} &= \sqrt{\frac{k_v}{k_v}} & \text{counter-masses mode; } M_1 \text{ and } M_3 \text{ moving in opposite direction} \\
\frac{w_3}{M_1} &= \sqrt{\frac{2k_v \cdot M_1 + k_v \cdot M_2}{M_1 \cdot M_2}} & \text{push-pull mode; } M_2 \text{ moving opposite to } M_1 \text{ and } M_3
\end{align*}
\]

Referring to the material properties of the piezoelectric stack and the masses values (see 6.2A.2) the value of the push-pull system eigenvalues are determined as shown in Table 2.5-1.

<table>
<thead>
<tr>
<th>Mode</th>
<th>Frequency</th>
</tr>
</thead>
<tbody>
<tr>
<td>Counter-masses mode</td>
<td>( w_2 = 45 \text{ kHz} )</td>
</tr>
<tr>
<td>Push-pull mode</td>
<td>( w_2 = 173 \text{ kHz} )</td>
</tr>
</tbody>
</table>

Table 2.5-1: Eigenfrequencies values

In the push-pull operation it’s clearly defined from the FRF plot in Figure 2.7 that only the push-pull mode is excited at the tip of the capillary (X2/V).

2.6 Conclusion of the modeling

In this chapter a modeling of a multi-layers stack actuator is derived. From that a model of the push-pull structure is derived in order to define the different eigenfrequencies and their exact positions.
Chapter 3

Current control design

This chapter deals with the design of the current controller system for actuation of the piezoelectric transducer. A notion of hysteresis for piezoelectric actuator is introduced and an analytical modeling for current actuation of piezoelectric stack is derived. A control scheme and electronic design is developed.

3.1 Nonlinear behavior of the piezoelectric actuator

The extension and the contraction of the piezoelectric material is the result of applying positive or negative voltage (see section 2.2). But as piezoelectric materials are dielectrics, they are subjected to dielectric hysteresis ([Main95]) that leads to a nonlinear relationship between the applied voltage and the resulting mechanical displacement and excitation of undesirable higher frequencies harmonics (see Figure 3.1). The hysteresis does not affect the whole input range. The level of distortion will vary according to the maximum value of the input voltage and the frequency of the input voltage. The hysteresis is caused by internal energy losses within the piezoelectric materials (dielectric). The piezoelectric ceramic lead zirconate titanate (PZT) shows significant amount of hysteresis due to their low Q factor. For linear operation of piezoelectric actuator, low voltages values can be applied. But to use the piezoelectric actuator at their maximum potential, a charge or current control actuation is an approach to significantly reduce the hysteresis behavior ([Main95], [Coms81]).
3.2 Modeling of the piezoelectric structure under Current or charge actuation

The objective of the charge actuation is to deliver a known amount of charge. From equations (8), the equation governing the free clamped piezoelectric transducer (Figure 2.4) under current control can be derived.

\[
\begin{bmatrix}
[M_{uu}] \cdot \ddot{\{u\}} \\
\frac{P}{\varepsilon_s \cdot n} \\
\end{bmatrix} + \left( [K_{uu} + [K_{u\varphi}] \cdot [K_{\varphi\varphi}]^{-1} \cdot [K_{\varphi u}]) \right) \cdot \{u\} = -[K_{u\varphi}] \cdot [K_{\varphi\varphi}]^{-1} \cdot \{Q\} \tag{3.1}
\]

\[
\begin{bmatrix}
2m & m \\
\frac{k_q}{m} & -k_q & k_q \\
\end{bmatrix} \begin{bmatrix}
\ddot{u}_1 \\
\ddot{u}_n \\
\end{bmatrix} + \begin{bmatrix}
k_q & -k_q \\
-k_q & k_q \\
\end{bmatrix} \begin{bmatrix}
u_1 \\
u_n \\
\end{bmatrix} = \begin{bmatrix}
k_{emq} \\
-k_{emq} \\
\end{bmatrix} \cdot \{Q\} \tag{3.2}
\]

\[k_{emq} = \frac{[e]}{\varepsilon_s \cdot n} \text{ and } k_q = \left( k_v + \left( \frac{n \cdot k_v \cdot d_{33}}{C_p} \right)^2 \right)\]

Under Laplace domain \( Q = I/sC_p \), where I is the current and s the Laplace operator.

Following a similar reasoning as in section 2.5, the analytic equation governing the piezoelectric actuator structure under charge control is;

\[
\begin{bmatrix}
2m + M_1 & m & 0 \\
m & 4m + M_2 & m \\
0 & m & 2m + M_3 \\
\end{bmatrix} \begin{bmatrix}
\dddot{X}_1 \\
\dddot{X}_2 \\
\dddot{X}_3 \\
\end{bmatrix} + \begin{bmatrix}
k_q & -k_q & 0 \\
-k_q & 2k_q & -k_q \\
0 & -k_q & k_q \\
\end{bmatrix} \begin{bmatrix}
X_1 \\
X_2 \\
X_3 \\
\end{bmatrix} = \begin{bmatrix}
k_{emq} & 0 \\
-k_{emq} & k_{emq} \\
0 & -k_{emq} \\
\end{bmatrix} \begin{bmatrix}
\phi_1 \\
\phi_2 \\
\phi_3 \\
\end{bmatrix} \tag{3.3}
\]

Considering the fact that the mass of the piezoelectric actuator can be neglected according to the mass of the counter-mass and the fact that \( M_1 = M_3 \), then the eigenfrequencies of the system are;
Comparing the dynamics of the piezoelectric actuator system under charge control actuation to the one of voltage actuation, one remark is there is no change in term of mode. The counter-mass mode and push-pull mode are still present. The only change to notice is the shift of the push-pull and counter-masses mode frequency as displayed in Table 3.2-1. Therefore the current or charge control driver can be designed as one is aware that the structural dynamic will not change significantly.

<table>
<thead>
<tr>
<th>Mode</th>
<th>frequency</th>
</tr>
</thead>
<tbody>
<tr>
<td>Counter-masses mode</td>
<td>$w_2 = 55 \text{ kHz}$</td>
</tr>
<tr>
<td>Push-pull mode</td>
<td>$w_2 = 201 \text{ kHz}$</td>
</tr>
</tbody>
</table>

Table 3.2-1: eigenfrequencies under charge control

### 3.3 Current/charge control system scheme

The basic goal of a current/charge control driver is to regulate the amount of current/charge flowing through the piezoelectric actuator. Basically the piezoelectric stack actuator is modeled electrically as a capacitor. Therefore an electrical impedance network can be used to sense the current/charge through the piezoelectric actuator. A basic concept for the current/charge controller (see Figure 3.2) can be a non-inverting operational amplifier configuration where an impedance network $Z_S$ is used to sense the current/charge flowing through the piezoelectric actuator impedance $Z_P$. One speaks about a current controller when the sensing device $Z_S$ is resistor impedance and about a charge controller when the sensing device $Z_S$ is capacitor impedance. The high loop gain and the voltage driver HVA work to make the sensing voltage $V_S$ equate the reference voltage $V_{REF}$. The current going through the piezoelectric load is $I_p = V_S / Z_S$.

![Figure 3.2: basic configuration of a current/charge controller.](image)
All the designing procedure in this report will be done with a sensing resistor instead of capacitor as a current is easily monitored than a charge practically. Nevertheless a summarizing result will be given as well in case of using capacitor sensor. Assuming a resistor $R_s$ as sensing element and $C_p$, a capacitive load (piezoelectric actuator), the transfer function from the current flowing through the load $I_p$ and the reference input voltage $V_{REF}$ is;

$$\frac{I_p}{V_{REF}} = \frac{Aol \cdot C_p \cdot s}{1 + Aol \cdot R_s \cdot C_p \cdot s + R_s \cdot C_p \cdot s}$$

The current control bandwidth is then limited from the lower bound $w = R_s C_p$ to the upper bound defined by the bandwidth of the voltage driver (Aol). The typical transfer function response from the reference voltage $V_{REF}$ to the capacitive load current $I_L$ is shown on Figure 3.3.

![Figure 3.3: transfer function from reference voltage to the capacitive load current.](image)

The use of basic configuration however implies some limitations when driving highly capacitive load. The circuit does not incorporates any DC feedback between the voltage driver output and the sensing load, any small offset voltage within the circuit result in a net offset across the capacitive load. Another limitation comes from the fact that at high frequencies, highly capacitive load has low impedance and this lead to an obligation of using very low impedance sensing resistor to avoid dissipation of the sensor. By using sensing resistor with low impedance the lower bound $w = R_s C_p$ of the current control bandwidth is limited.

To overcome the limitations mentioned above, a current control system is designed where the circuit includes two feedback loops (cascade control). The cascade control includes a fast inner loop and a slow outer loop. The inner loop ensures the tracking of the reference current signal and the outer loop regulates the DC offset at the output of the voltage driver (see Figure 3.4).

The inner loop comprise the voltage driver HVA which is a high voltage amplifier, the capacitive load $Z_p$ which is the piezoelectric actuator, the sensing resistor $Z_s$ and a controller block $K_C$ to shape and increase the bandwidth of the current control system. The outer loop adds a controller block $K_V$ to shape the low frequency behavior of the voltage driver output.

Before diving in the design of the current controller driver for the piezoelectric actuator, the high voltage amplifier (HVA) has to be designed for dynamics and static operations.
3.3.1 Static requirement of the voltage amplifier circuit

Powers, voltages and currents rating of the Op Amp are defined according to the piezoelectric actuator operating conditions. The relevant characteristics for the electrical operation are:

- the driving frequency is up to 200kHz,
- the driving voltage is up to +/- 50V (piezoelectric actuator stroke is 2.2 μm at 100V),
- the capacitive load is up to 50nF.

Slew rate requirement for the Op Amp
The maximum slew rate is computed using the highest frequency and the largest voltage swing. The required slew rate to track a sinusoidal signal at maximum frequency and voltage is as follow:

\[
S.R \geq 2 \pi f V_p = 2 \pi 200k \times 40 \times 10^{-6} = 62.9V/\mu s
\]

Output current requirement for the Op Amp
As the piezoelectric behave as a pure capacitor then the impedance is

\[
X_C = \frac{1}{2\pi f C_L} = \frac{1}{2 \pi \times 200k \times 50n} = 15.91\Omega
\]

The maximum current occurs at highest frequency with capacitive load

\[
I_M \geq \frac{V_p}{X_C} = \frac{50V}{15.91} = 3.2A
\]

Power dissipation requirement for the Op Amp
A reactive element (capacitor) does not dissipate power; however the op amp driving the reactive load will dissipate power. 
As the op amp output voltage swing is ±50V ;
\[ P_D = \frac{V_p^2}{2X_C} \left[ \frac{4}{\pi} - \cos \theta \right] \]

Worst case for purely capacitive load \( \cos \theta = 0 \) thus

\[ P_D = \frac{V_p^2}{2X_C} \left[ \frac{4}{\pi} \right] = \frac{4 \times 50^2}{2 \times \pi \times 15.91} = 100W \]

**Op Amp characteristics**

The Op Amp is selected such that it fills the rating computed above. The MP108 Op Amp from APEX appears to fill those requirements and selected to drive the piezoelectric actuator (see 6.2C.3).

### 3.3.2 Dynamics of the higher voltage amplifier

#### 3.3.2.1 Voltage amplifier transfer function

The voltage amplifier is built around a non-inverting configuration (see Figure 3.5). The resistor network \( R_I \) and \( R_F \) define the gain of the amplifier. The block diagrams of the amplifier shows the Op Amp open loop gain \( A_{ol} \) and the feedback network \( \beta \).

\[ \beta = \frac{R_I}{R_I + R_F} \]

The transfer function from the output to the input is:

\[ \frac{V_{OUT}}{V_{IN}} = Acl = \frac{A_{ol}}{1 + A_{ol} \cdot \beta} \]

For high Op Amp open loop gain \( A_{ol} \beta >> 1 \) therefore

\[ \frac{V_{OUT}}{V_{IN}} = Acl = \frac{1}{\beta} \]

![Figure 3.5: non-inverting Op Amp amplifier and its corresponding block diagram.](image)
3.3.2.2 Dynamics of the voltage amplifier without capacitive load

The open loop gain of the amplifier is roughly modeled with a first order transfer function in case of resistive load. The Aol second pole is generally far away from the system bandwidth (see Error! Reference source not found.).

\[ A_{ol} = \frac{K \cdot p}{s + p} \]

where K is the open loop DC gain and p is the pole (break frequency) of the open loop. The loop gain of the voltage amplifier contains dynamics similar to the open loop Aol with shifted pole and controlled amplification gain \((1/\beta)\). The single pole will introduce a phase shift of 90° before the 0 dB crossover point. The relation between Aol, \(1/\beta\) and \(A_{ol}\beta\) is shown in Figure 3.6.

![Figure 3.6: loop gain and open loop dynamics of the voltage amplifier without capacitive load.](image)

3.3.2.3 Dynamics of the voltage amplifier with capacitive load

The Op Amp output impedance \(R_o\) and the loading capacitive load \(C_L\) introduce a pole to the dynamics of the open loop Aol. In this configuration the open loop Aol is modeled as a second order system;

\[ A_{ol} = \frac{K \cdot p_1 \cdot p_2}{(s + p_1)(s + p_2)} \]

The loop gain \(A_{ol}\beta\) contains two poles, accumulating about 180° of phase shift before the 0 dB crossover point. This fact can lead to the voltage amplifier stability issues.
3.3.3 Voltage amplifier compensation for stability and performances

Several techniques exist for capacitive load effect compensation [[Green]]. One can denote the following compensation techniques:

- The isolation resistance compensation is a technique where a resistance is added between the Op Amp output and the capacitive load. But this technique introduces an error in the output voltage across the capacitive load.
- The capacitor feedback compensation is a technique where a capacitor is added parallel to the feedback resistance. This technique reduces as well the overall noise within the system but it reduces the bandwidth of the system.
- The noise gain compensation is a technique that adds a capacitor in series with a resistor to the feedback input of the Op Amp. This technique does not affect the system bandwidth but increase the overall noise gain of the Op Amp circuit.

Due to the limited bandwidth of the Op Amp used for the voltage amplifier, the noise gain compensation is implemented (see Figure 3.8). The loop gain of the system is:

\[
A_{ol, \beta} = \frac{K \cdot p_1 \cdot p_2}{(s + p_1) \cdot (s + p_2)} \cdot \frac{R_F \cdot (R_N \cdot C_N \cdot s + 1)}{(R_N \cdot R_I + R_N \cdot R_F + R_I \cdot R_F) \cdot C_N \cdot s + (R_I + R_F)}.
\]
The noise gain compensation introduces a pole \( P_N \) and a zero \( Z_N \) to the system loop gain. That pole is derived as;

\[
P_N = \frac{1}{R_N \cdot C_N}.
\]

The zero is derived as;

\[
Z_N = \frac{(R_I + R_F)}{(R_N \cdot R_I + R_N \cdot R_F + R_I \cdot R_F) \cdot C_N}.
\]

The pole \( P_N \) is defined such that the rate of closure between the open loop \( \text{Aol} \) and the feedback network \( 1/\beta \) is 20 dB/decade (see Figure 3.9). The closed loop gain of the system is not affected by the compensation circuit.

\[
\frac{V_{\text{OUT}}}{V_{\text{IN}}} = \frac{R_I + R_F}{R_I}.
\]
3.3.4 High voltage amplifier implementation

Referring to the Figure 3.8 the value of resistors RF and RI and chosen big enough to avoid dissipation and power consumption and less enough to minimize the closed loop transfer function peaking. The gain of the amplifier is chosen quite small to minimize the bandwidth loss as $f_{3dB} \approx f_{GBW} / A_v$. The voltage gain is;

$$A_v = \frac{RI + RF}{RI} = 6.$$ 

From Spice simulation the compensating network is dimensioned such to damp the resonance resulting from driving a capacitive load. Figure 3.10 shows how a resonance is damped with the noise gain compensation technique. The complete high voltage amplifier design is demonstrated in 6.2 Appendix.C with schematic and Spice simulations.

3.4 Current controller circuit design

The purpose of this part is to convert the high voltage amplifier to a current control amplifier. The system is divided in two control loop: the tracking control loop where the load current is controlled and the DC control loop where the DC current across the load is controlled.

3.4.1 Process transfer function

For the current control configuration a resistor is used to sense the current flowing through the capacitive load. The current flowing through the sensing resistor increases as the frequency increases;

$$I_L = \frac{V_L}{Z_L} = V_L \cdot C_p \cdot s.$$ 

Therefore the value of the sensing resistor is chosen quite small to limit its dissipation at high frequencies.
The transfer function of the process is extracted by deriving the dynamics from the input of the high voltage amplifier $V_{IN}$ to the sensing resistor voltage $V_S$ (see Figure 3.12). That transfer function is expressed as;

$$\frac{V_S}{V_{IN}} = \frac{R_s \cdot C_p \cdot s}{R_s \cdot C_p \cdot s + 1} \cdot Acl.$$  

Where $Acl$ is the voltage amplifier closed loop with load as combination of capacitive load and sensing resistor. $Acl$ is modeled as;

$$Acl = \frac{V_L + V_S}{V_{IN}} = \frac{K}{1 + \frac{s}{w_p \cdot Q_p} + \frac{s^2}{w_p^2}}.$$  

Where $K$ is the close loop gain, $w_p$ is the resonance frequency and $Q_p$ the quality factor. The bode plot shown in Figure 3.11 confirms the differentiator character of the process dynamics at frequencies below;

$$f = \frac{1}{2\pi R_s C_p}.$$  

Figure 3.11: bode plot of the process dynamics.

### 3.4.2 Tracking control of the capacitive load current

For tracking control (see Figure 3.12) the current $I_L$ flowing through the capacitive load should be proportional to the reference voltage $V_{REF}$. The ideal transfer function from the flowing current to the applied reference is;

$$\frac{I_L}{V_{REF}} = \frac{1}{R_s}.$$  

Thus the transfer function from the sensing voltage $V_S$ to the applied reference voltage should be equal to 1.
After the derivation of the process dynamics, the capacitive load current tracking resumes to study the stability and the performances of the system under the closed loop (see Figure 3.13). The controller $K_c$ should be tuned such that the sensing voltage $V_s$ is equal to the reference voltage under the defined bandwidth.

An easy approach for the controller should be a double integrator. But this approach will increase the offset within the system circuit. A solution is to limit low frequencies amplitude and increase high frequencies till the high voltage roll off frequencies. A suitable choice can be a double lag controller. The controller is expressed as:

$$K_c = G \cdot \left( \frac{1}{2 \cdot \pi \cdot f_p} s + 1 \right)^2 \cdot \left( \frac{1}{2 \cdot \pi \cdot f_z} s + 1 \right)^2 = 10^4 \left( \frac{1}{2 \pi \cdot 2 \cdot 10^6 s + 1} \right)^2 \cdot \left( \frac{1}{2 \pi \cdot 2 \cdot 10^7 s + 1} \right)^2.$$

$G$ is the gain, $f_p$ is the frequency of the pole and it represent the lower bound frequency of the current controller $f_z$ is the frequency of the zeros and used to limit high frequencies gain. The frequency response of the controller is shown in Figure 3.14.
The electronic implementation of the lag controller is built around a low voltage Op Amp. The electronic schematic of the lag controller is represented in Figure 3.15. The transfer function is then expressed as:

\[
\frac{V_{OUT}}{V_{IN}} = -\sqrt{KC} = -\frac{R_{FL}}{R_{IL}} \frac{R_LC_L s + 1}{(R_L + R_{FL})C_L s + 1}
\]

The closed loop transfer function of the current tracking control is derived as:

\[
\frac{V_S}{V_{REF}} = \frac{P \cdot C}{1 + P \cdot C} = \frac{\text{process} \cdot K_C}{1 + \text{process} \cdot K_C}
\]

The resulting closed loop Bode plot shows \( V_S/V_{REF} = 1 \) in the bandwidth moving from 1 kHz up to 500 kHz. The closed loop is stable with a phase margin of 80° and a gain margin of 9dB (see Figure 3.16).
The frequency response of the output of the voltage amplifier \( V_L/V_{REF} \) is shown in Figure 3.17. The gain is quite big at low frequencies. Thus any offset in the loop will be amplified (i.e. supposing an inherent Op-Amp offset of 10 mV the output \( V_L \) will have a theoretical offset of \( 10e^{-3} \cdot 10e^{5} = 100V \)). To overcome that situation another control loop is implemented to drop that gain at low frequencies.

### 3.4.3 DC offset control across the capacitive load

In order to control the DC offset across the capacitor a slow loop is added in cascade to the tracking control loop (see Figure 3.13). The overall current control block displayed in Figure 3.4 can be re-sketched as shown Figure 3.18 where \( P1 = \text{Process} = V_S/V_{IN} \) and \( P2 = V_L/V_S \). The new process becomes:

\[
\text{process2} = (V_S/V_{REF}) \times (V_L/V_S) = V_L/V_{REF}.
\]
The controller block \( K_v \) should be tuned such that there is no DC offset at the output \( V_L \). The operating frequency range of the DC offset control loop should not interfere with the operating frequency range of the current control loop.

According to the new process (see Figure 3.17) A PI controller is used to decrease the gain at low frequencies and the integration cut-off frequency is chosen quite small according to the current control lower bound frequency (this to avoid interaction or conflict between current control and offset control).

The PI controller is expressed as:

\[
K_v = k \cdot \frac{s + 2\pi \cdot f_I}{s} = 10^{-5} \cdot \frac{s + 2\pi \cdot 10}{s}.
\]

The frequency response of the PI controller is represented in Figure 3.19.

The electronic design of the PI controller is represented in Figure 3.17, where the transfer function is:

\[
\frac{V_{OUT}}{V_{IN}} = -K_v = -\frac{R_{FI}}{R_II} \cdot \frac{s + \frac{1}{R_{FI} \cdot C_I}}{s}.
\]
The output DC transfer is expressed as:

\[ \frac{V_L}{V^*_{REF}} = \frac{P_2 \cdot (V_{REF}/V_S)}{1 + K_V \cdot P_2 \cdot (V_{REF}/V_S)}. \]

The frequency response of current control loop and offset control are displayed respectively in Figure 3.22 and Figure 3.22. One can notice that the current control loop is only affected by the DC offset control loop at frequencies below the integration cut-off frequency. By consequent the current control remains unchanged in the previously defined bandwidth. The stability of the current controller remains unchanged as well.
3.5 Conclusion of the current control design

The electronic design of the current controller is implemented in order to control the amount of current flowing through the piezoelectric actuator. The electronic design of the current controller can clearly be divided in four blocks. The first block includes the plant made of the voltage amplifier and the piezoelectric actuator load, the second block includes the current tracking controller, the third block includes the DC offset controller and the final block is a differentiator that allows to close the different loops.

The implementation is realized by using low cost Op amp and conventional resistors and capacitors. The complete schematic of the current controller driver can be found in Error! Reference source not found..
Chapter 4

Self-sensing control of the piezoelectric actuator structure

4.1 Self-sensing piezoelectric actuator

Self-sensing piezoelectric actuator refers to the fact that a single piezoelectric transducer serves simultaneously as actuator and sensor. Referring to the equations (7) governing the dynamics behavior of the piezoelectric transducer, the sensor equation contains the mechanical response \( [K_{\theta}] \cdot \{u\} \) and a feedthrough charge term \( [K_{\phi}] \cdot \{\phi\} \) caused by the voltage applied. It’s obvious that a good arrangement will be to have sensor equation without the feedthrough charge term. This will result to sensing equation containing only the mechanical response.

4.1.1 Self-sensing theory

Dosch and al [[Dosc92]] developed a method to compensate the feedthrough charge term. That method makes use of a bridge configuration (Wheatstone bridge) where an additional leg is added to compensate the electric term. The basic self-sensing bridge configuration is shown in Figure 4.1.
In the bridge the piezoelectric actuator is represented by their electrical models which are capacitor $C_p$ and voltage generator $V_m$. The capacitor $C_m$ represents the model of the piezoelectric actuator, $Z_S$ represents the sensing impedance, $Z_R$ represents the reference impedance and $U_a$ represents the control voltage. The transfer function of the self-sensing bridge configuration is:

$$U_m = \frac{Z_S \cdot C_p \cdot s}{Z_S \cdot C_p \cdot s + 1} \cdot V_M + \left( \frac{Z_S \cdot C_p \cdot s}{Z_S \cdot C_p \cdot s + 1} - \frac{Z_R \cdot C_m \cdot s}{Z_R \cdot C_m \cdot s + 1} \right) \cdot U_a.$$  \hfill (4.1)

If $C_p \approx C_m$ and $Z_S \approx Z_R$ then;

$$U_m \approx \frac{Z_S \cdot C_p \cdot s}{Z_S \cdot C_p \cdot s + 1} V_M.$$  \hfill (4.2)

The approximation comes from the fact that in a practical situation it is difficult to match impedances. The sensing voltage $U_m$ is expressed only according to the mechanical response $V_M$.

### 4.1.2 Piezoelectric structure dynamics under self-sensing

To study the properties of the self-sensing system, the piezoelectric actuator structure represented in Figure 2.6 is combined to the self-sensing bridge configuration represented in Figure 4.1. For simplicity equation (10) can be rewritten as follows;

$$M\ddot{x} + Kx = -\Theta u_p$$

$$-\Theta^T x + C_p u_p = q$$  \hfill (4.3)

$M$ is the mass of the structure, $K$ is the stiffness of the structure, $\Theta$ is the electromechanical term and $u_p = \phi$ is the voltage across the piezoelectric actuator.

The sensing impedance can be either a capacitor or a resistance. If the sensing impedance is a resistor $Z_S = R_S$ and $Z_R = R_R$ then one peaks about strain rate sensing (velocity). In case of sensing capacitor $Z_S = C_S$ and $Z_R = C_R$ then one speaks about strain sensing (displacement).
4.1.2.1 Strain self-sensing circuit

The strain self-sensing circuit is a fully capacitor bridge (CC bridge). The sensing capacitor senses the charge through the piezoelectric actuator. Thus from the charge and the actuating voltage, the capacitance of the piezoelectric actuator can be derived from equations (4.3). The complete computation procedure of that transfer function is done in Error! Reference source not found. The capacitance is expresses as:

\[
\frac{Q(s)}{U_p(s)} = \frac{\Theta^T \cdot M^{-1} \cdot \Theta}{s^2 + M^{-1} \cdot K} + C_p = C_{\text{Mechanical}} + C_{\text{Electrical}} = C_M(s) + C_p(s)
\] (4.4)

The capacitance contains contribution from mechanical part \( C_m(s) \) and contribution from purely electrical capacitance \( C_p(s) \). The frequency response of the piezoelectric actuator capacitance is displayed in Error! Reference source not found.

Having the capacitance, the transfer function of the self-sensing bridge can be derived as:

\[
\frac{U_m}{U_a} = \frac{C_M(s) + C_p(s)}{C_M(s) + C_p(s) + C_S} = \frac{C_m}{C_m + C_R}
\] (4.5)

The transfer functions include the mechanical capacitance \( C_M(s) \), which complies with the fact that the mechanical behavior can be retrieved with CC self-sensing.

4.1.2.2 Strain rate self-sensing circuit

The strain rate self-sensing circuit is a capacitor resistor bridge (RC bridge), the sensing resistor sense the current through the piezoelectric actuator. From the current and the actuating voltage the admittance of the piezoelectric actuator is derived from equations (4.3). The admittance is then expresses as:

\[
\frac{I(s)}{U_p(s)} = \frac{s\Theta^T M^{-1} \Theta}{s^2 + M^{-1} K} + sC_p = sC_{\text{Mechanical}} + sC_{\text{Electrical}} = Y_M(s) + Y_p(s)
\] (4.6)
The admittance contains contribution from mechanical part $Y_M(s)$ and contribution from electrical part $Y_p(s)$. The frequency response of the admittance is displayed in Error! Reference source not found..

From the admittance the transfer function of the self-sensing bridge can be derived as expressed as:

$$\frac{U_m}{U_a} = \frac{R_S \cdot \left( Y_M(s) + Y_p(s) \right) - R_R \cdot Y_m}{\left( 1 + R_S \cdot \left( Y_M(s) + Y_p(s) \right) \right) \cdot \left( 1 + R_R \cdot Y_m \right)}$$

(4.7)

The transfer functions include the mechanical admittance $Y_M(s)$ which complies with the fact that the mechanical behavior can be retrieved with RC self-sensing.

### 4.1.3 Self-sensing stability

The theory on self-sensing relies on the fact that the matching capacitance should be equal to the piezoelectric actuator capacitance. But practically it’s relatively difficult to satisfy that condition. The mismatch between the piezoelectric actuator capacitance and the matching capacitor does not affect the poles of the system, but has tendency to migrate the zeros of the system as demonstrated in 6.2B.2.

For the matching capacitor $C_m$ capacitance lower than the piezoelectric capacitance $C_p + C_M$ there will be alternating couple of poles and zeros with an extra pair of zeros at the end.

For the matching capacitor $C_m$ capacitance greater than the piezoelectric capacitance $C_p + C_M$ there will be alternating couple of zeros and poles with an extra pair of zeros at the beginning.

Figure 4.4 and Figure 4.5 illustrate the migration of the poles and zeros under sensing resistor and sensing capacitor.
4.1.4 Electronic design of the self-sensing circuit

The design of the self-sensing can be divided in two blocks where the first block is the sensing bridge and the second block is the difference amplifier. The design of the bridge is straightforward as one has only to define the impedances values. However the matching leg requires an extra demand in term of current. In order to limit the dissipation of the power voltage driving the piezoelectric actuator, the matching capacitor is chosen ten times smaller than the piezoelectric actuator capacitance. To balance the bridge the following parameterization is defined;
The sensing resistor is chosen ten times bigger than the piezoelectric actuator sensing resistance. The same analogy can be done when sensing with a capacitor. The difference amplifier is an instrumentation amplifier builds around three low power operational amplifiers (see Figure 4.6). The transfer function of the instrumentation amplifier is:

\[
\frac{U_m}{U_s - U_r} = \left(1 + \frac{2R_f}{R_g}\right).
\]

Rg is used to define the instrumentation gain.

As stated above in 4.1.3 the self-sensing technique relies on the fact there is a perfect balancing between the piezoelectric leg and the matching capacitor leg. In case of imbalance there is a migration of zeros leading to corruption of the sensed signal and instability of the self-sensing system. In case of imbalance the instrumentation amplifier does not bring any kind of compensation. It's why it crucial to implement any kind of adaptive system to compensate for imbalance within the self-sensing bridge.

4.2 Adaptive algorithm for piezoelectric actuator self-sensing

A traditional method to remove the feedthrough term in self-sensing has be implemented in section 4.1.4, however that method is sensitive to capacitance change as demonstrated in section 4.1.3. Many situations can contribute to the mismatch evaluation of the piezoelectric actuator capacitance:

- the capacitance of the piezoelectric change over time and with the temperature,
- the capacitance change according to the actuating frequency (see 6.2B.3),
- the capacitance is prove to be difficult to measure due to the hysteresis ([Cole94]).

An alternative to the traditional method has been developed in [Clark96] and [Cole94], where the feedthrough capacitance is compensated digitally by adaptive algorithm. That
adaptive algorithm make use of known techniques such as least mean square (LMS) algorithm or recursive least square (RLS) algorithm.

4.2.1 Least mean square algorithm

Adaptive filtering can be considered as a process in which the parameters used for the processing of the signals changes according to some criterion. Usually the criterion is the estimated mean squared error or the correlation. The adaptive filters are time-varying since their parameters are continually changing in order to meet the required performances.

The traditional scheme of the adaptive filtering is shown in Figure 4.7, where \( k \) is the iteration number, \( s(k) \) denotes the desired signal, \( n(k) \) define the input signal, \( y(k) \) is the adaptive filter output and \( e(k) \) the error signal. Using the error an adaptive algorithm (weighting filter \( w(k) \)) adjusts the response characteristics by minimizing the measure of the error.

![Figure 4.7: adaptive filter scheme.](image)

The filter output (linear combiner) is derived as;
\[
y(k) = x(k) \cdot w(k)
\]
(4.8)

The error signal is derived as;
\[
e(k) = d(k) - y(k)
\]
(4.9)

The instantaneous power in the output signal, given by the cost function in equation 4.10 can be minimized using a steepest descent method.
\[
J(k) = E\{e(k)^2\}
\]
(4.10)

The cost function \( J(k) \) is a quadratic form, which means the global minimum of \( J(k) \) uniquely exists.

It demonstrated by [Widr85] that, by using the steepest descent method the LMS update rule for the filter can be expressed as;
\[
w(k+1) = w(k) + \mu \cdot \nabla J(k) = w(k) + 2\mu \cdot x(k) \cdot e(k)
\]
(4.11)

\( \nabla \) is the gradient term and \( \mu \) is the convergence rate.

[Widr85] demonstrate as well that the filter will converge with a convergence rate of \( \mu \).
4.2.2 Adaptation of the LMS algorithm to piezoelectric actuator self-sensing

The adaptation of LMS algorithm to the self-sensing of piezoelectric actuator was implemented by [Clark96]. The developed concept uses LMS algorithm approach to balance the piezoelectric actuator leg to matching capacitance leg (see Figure 4.8). This concept slowly compensates the piezoelectric actuator capacitance change over the time.

By analogy to the LMS algorithm [Clark96] came out with the following equation;

\[ U_m(k) = U_S(k) - U_R(k) \cdot V_{DSP}(k) \]  
(4.12)

Here \( V_{DSP} \) denotes the updating adaptive filter. The value of \( V_{DSP} \) is determined by minimizing the quadratic cost function of the following equation;

\[ J(V_{DSP}(k)) = U_m^2(k) \]  
(4.13)

The updating filter \( V_{DSP} \) is determined as;

\[ V_{DSP}(k+1) \cong V_{DSP}(k) - \mu N(J(V_{DSP}))\Delta t \cong V_{DSP}(k) + 2\mu \cdot \Delta t \cdot U_m(k) \cdot U_R(k) \]  
(4.14)

If \( V_{DSP}(k) \) converges, then the LMS algorithm compensates the mismatch between the piezoelectric actuator capacitance and the matching capacitance. The sensing bridge including the adaptive filter is represented in Figure 4.9.

\[ U_m(k) = U_S(k) - U_R(k) \cdot V_{DSP}(k) \]

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(4.14)

If \( V_{DSP}(k) \) converges, then the LMS algorithm compensates the mismatch between the piezoelectric actuator capacitance and the matching capacitance. The sensing bridge including the adaptive filter is represented in Figure 4.9.

![](image1.png)

**Figure 4.8:** Adaptive compensation for piezoelectric actuator self-sensing

![](image2.png)

**Figure 4.9:** Self-sensing using digital adaptive filter.
Although the adaptive systems implemented by [Clark96] solve the problem of varying capacitance, but it makes use of digital adaptive filter. This leads to the need of DSP systems for the implementation of such filters. The piezoelectric actuator push-pull system is drove at high frequencies (up to 200 kHz). DSP systems capable to handle such frequencies are costly. A solution to this problematic will be to convert the digital adaptive filter to an analog adaptive filter.

### 4.2.3 Analog adaptive filter for self-sensing of piezoelectric actuators

Continuous time implementation of the LMS algorithm is possible by simply converting iterative algorithm into continuous integral (see [Caru00]). One way to obtain an analog LMS is to transform the updating adaptive filter. Thus the filter equation in 4.11 can be rewritten in z-transform as;

$$z \cdot W(z) = W(z) + 2\mu \cdot X(z) \cdot E(z) \Rightarrow W(z) = \frac{2\mu \cdot X(z) \cdot E(z)}{z - 1}$$

(4.15)

The relation between the z-transform and the discrete-time Laplace transform is;

$$z = e^{r_s} = 1 + T \cdot s + \frac{(T \cdot s)^2}{2!} + \frac{(T \cdot s)^3}{3!} + \cdots$$

(4.16)

$T$ is the sampling frequency and the second term is the expansion of the exponential. Assuming $T \to 0$, as is the case in continuous time, higher than first–order term can be neglected. This yields to the following approximation;

$$W(s) = \frac{2\mu \cdot X(z) \cdot E(z)}{e^{r_s} - 1} \approx \frac{2\mu \cdot X(z) \cdot E(z)}{T \cdot s}$$

(4.17)

Supposing zero initial condition, the integration of the Laplace-transform expression in 4.17 yields to the following;

$$w(t) = 2\mu \int_0^t x(\tau) \cdot e(\tau) \cdot d\tau$$

(4.18)

The formulation of the discrete time linear combiner of equation 4.8 in analog time is straightforward ([Widr85]).

$$y(t) = x(t) \cdot w(t)$$

(4.19)

Having transformed the discrete-time adaptive algorithm to analog adaptive algorithm and by expressing the output of the analog LMS as $V_{ADAPT}(t)$, the self-sensing error equation 4.12 can be reformulated as;

$$U_m(t) = U_s(t) - U_r(t) \cdot V_{ADAPT}(t)$$

(4.20)

The self-sensing updating filter 4.14 can be reformulated as;

$$V_{ADAPT}(t) \approx 2\mu \int_0^t (U_m(t) \cdot U_r(t)) \cdot dt$$

(4.21)

The basic application of the analog adaptive self-sensing for piezoelectric actuator is shown in
4.2.4 Matlab simulation of the analog adaptive self-sensing

To confirm the theoretical results the self-sensing scheme is simulated in Simulink. In the Simulink block diagram (see Figure 4.11) a sinusoidal signal VM is introduced to represent the mechanical signal that should be sensed. VM can roughly be estimated from equation (8);

\[ Q = k_{\text{env}} \cdot u + C_p \cdot \phi = C_p V_M + C_p \cdot \phi \]

\[ V_M = (k_{\text{env}} \cdot u)/C_p \]

Considering a linear operation of the piezoelectric actuator the mechanical displacement is \( u = 0.5\mu m \) from \( \pm 12 \) \( V \) actuation. \( V_M = 0.77 \) \( V \) (Refer to 6.2A.2 for piezoelectric properties) and due to the voltage divider in the bridge the sensed \( V_M \) is 0.05V. The Simulink block is simulated with the parameters in Table 4.2-1.

The time response of the adaptive self-sensing is displayed in Figure 4.12. It can easily be seen that the convergence rate is about 1 msec and at 4.8 msec the LMS output \( U_m \) is perfectly tracking the mechanical signal VM.
Table 4.2-1: simulink parameters

<table>
<thead>
<tr>
<th>symbol</th>
<th>function</th>
<th>values</th>
</tr>
</thead>
<tbody>
<tr>
<td>HVA</td>
<td>High voltage Amplifier gain</td>
<td>6 V/V</td>
</tr>
<tr>
<td>Cp</td>
<td>Piezoelectric actuator capacitance</td>
<td>[20 nF  30nF]</td>
</tr>
<tr>
<td>Cm</td>
<td>Matching capacitance</td>
<td>2.5 nF</td>
</tr>
<tr>
<td>CS</td>
<td>sensing capacitance in the piezo leg</td>
<td>330 nF</td>
</tr>
<tr>
<td>CR</td>
<td>sensing capacitance in the matching leg</td>
<td>33 nF</td>
</tr>
<tr>
<td>2*mu</td>
<td>Convergence rate parameter</td>
<td>1e3</td>
</tr>
<tr>
<td>Vref</td>
<td>Actuating signal</td>
<td>A=2V; f=100 kHz</td>
</tr>
<tr>
<td>VM</td>
<td>Mechanical signal</td>
<td>A=0.05V; f=110 kHz</td>
</tr>
</tbody>
</table>

After compensating for the bridge circuit imbalance, the obtained measurement signal can be used for feedback control purpose.
4.3 Feedback control of the push-pull system

With an appropriate choice of the controller, the feedback control can be used to damp the resonance modes of the push-pull system. In this project, to achieve the damping of resonant modes, a positive position feedback method is used.

4.3.1 Positive position feedback

One property of the structures with collocated actuators and sensors is the alternating pole-zero patterns close to the imaginary axis of the open loop. This allows deriving the transfer function of the structure as a sum of resonant modes. The positive position feedback can then be used to control independently each resonant mode of the structure. The positive position feedback (PPF) was introduced for the first time by Goh85. Stable closed-loop system can be achieved by positively feed-in back the position signal. The PPF control presents some advantages:

- the PPF control does not require a system model in its design. The PPF control only needs the resonance mode frequencies and the low frequency gain,
- The PPF control architecture is not sensitive to spillover (disturbances amplification by the closed loop system relative to the open-loop gain),
- With PPF control, each resonant mode can be damped independently without affecting other resonant modes,
- the PPF controller can be implemented with relatively simple analog circuits.

The transfer function of the PPF controller is presented as:

$$ C(s) = \frac{g \cdot w_f^2}{s^2 + 2 \cdot \xi \cdot w_f \cdot s + w_f^2} $$

where $w_f$ is the natural frequency, $\xi$ is the damping ratio and $g$ is the gain of the PPF controller.

The PPF controller is well known in electronic as a second order low pass filter. The controller is tuned to resonate at one of the structure frequency mode. The controller phase at the resonant frequency is $90^\circ$ lag. In the positive feedback, that $90^\circ$ phase lag results in a selective damping to the structure resonance mode.
4.3.2 PPF control of the push-pull system

To be able to damp the resonant modes of the push-pull system, the PPF controller is combined to self-sensing method. The complete scheme of the PPF control with self-sensing is displayed in Figure 4.14. Each PPF controller is dedicated for damping a specific resonant mode. To damp many modes, one has to put in parallel several PPF controllers in the loop. In this case the first PPF controller for instance is used to damp the push-pull mode of the piezoelectric actuators system. The second PPF could be used to damp the counter-mass mode. The natural frequency $\omega_f$ of the first PPF controller is fixed at the push-pull resonant mode and the damping ratio is tuned to nearly match the damping ratio of push-pull resonant mode.

4.3.3 Matlab implementation of the PPF control

In the modeling of the push-pull system described in section 2.5, the peak of the push-pull resonant mode is large. This comes from the fact that, the damping of the system was not accounted in the modeling. To have something more realistic a damping term is added to the modeling equation (10).
Considering the damping term \( C \), the equation (10) can be rewritten as:
\[
M \cdot \ddot{x} + C \cdot \dot{x} + K \cdot x = -\Theta \cdot u_p
\]
\[\quad - \Theta^T \cdot x + C_p \cdot u_p = q\]

The push-pull displacement transfer function and the self-sensing transfer function result in a response with less peaking of the push-pull mode. The transfer function \( G \) from the self-sensing output \( U_m \) to the actuation input \( U_a \) (see Figure 4.10) is displayed in Figure 4.15.

![Figure 4.15: response of the self-sensing with damping accounted.](image)

The PPF controller \( C \) is designed with the following parameter:

<table>
<thead>
<tr>
<th>values</th>
<th>Natural frequency ( w_f )</th>
<th>Damping ratio ( \zeta )</th>
<th>gain</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>173 kHz</td>
<td>0.1</td>
<td>20</td>
</tr>
</tbody>
</table>

The closed-loop transfer function of the PPF control is:
\[
T = G/(1 - G \cdot C)
\]

The transfer function of the controller \( C \), of the open-loop \( G \cdot C \) and the closed-loop \( T \) are displayed in Figure 4.16.

![Figure 4.16: response of the controller, open-loop, and closed-loop](image)
4.4 Conclusion of the self-sensing

In this chapter a basic self-sensing circuitry has been implemented to retrieve the mechanical behavior of the push-pull system. To compensate for capacitance change of the piezoelectric actuator an analog adaptive algorithm has been added to the basic self-sensing circuitry.

The electronic implementation of the analog adaptive self-sensing is done by using low-discrete components. The analog multiplier is based on four quadrant multipliers. The integrator is based on operational amplifier as explained in section 3.4.3. All active components such as multipliers and operational amplifiers are selected such that their operating frequency bandwidth can handle the actuating frequency of the piezoelectric actuator push-pull system. The complete electronic circuitry of the adaptive self-sensing can be found in Error! Reference source not found..

Furthermore, the self-sensing method is used for closed-loop feedback control. A PPF control is implemented to damp the push-pull mode of the piezoelectric actuators structure.
Chapter 5

Experimental measurement and results

At the first instance in this chapter the test setup is presented. At the second instance the FRF measurements of the piezoelectric actuator push-pull system are done. The FRF measurement of the current controller driver are completed and compared to the FRF of the simulation. The performance of the current controller is evaluated by estimating the percentage of hysteresis reduction. At the third instance the FRF of the self-Sensing are completed and the performances of the sensors are evaluated.

5.1 Test setup

To be able to drive the push-pull system and record the frequency and time behavior of the push-pull system, a test setup is organized in that purpose. The test setup can resume to scheme presented in Figure 5.1.
To drive the piezoelectric actuator a signal generator is required to produce the driving sinusoidal signal. The signal from the generator is then amplified through a voltage amplifier or current control amplifier. The displacement of the capillary tip is then captured by a laser probe, which is amplifier and treated though a laser decoder device. The output of the laser decoder device is then recorded for FRF analysis through network analyzer or for time response through an oscilloscope.

5.2 Measurement results under voltage Amplifier

The performance of the voltage amplifier is evaluated, the FRF measurement of the push-pull piezoelectric actuator system is presented and the hysteresis exhibited by the push-pull system under voltage actuation is presented as well. To summit the push-pull system in the same measurement conditions for voltage amplifier and current control amplifier a double voltage amplifier is used to actuate the push-pull system. Each voltage amplifier is actuating one piezoelectric actuator of the push-pull system. To obtain push-pull actuation the second piezoelectric actuator is connected reversely according to the first piezoelectric actuator (see Figure 5.2).
5.2.1 FRF results of the voltage amplifier

Here the dynamics performances of the voltage amplifier are analyzed. The voltage amplifier is first tested with a capacitor load of 25 nF at the output under the condition were the voltage amplifier is not compensated for capacitive load and under the condition were the voltage amplifier is compensated for capacitive load. With a noise gain compensation technique, it’s clearly seen that the resonance created by the capacitive load is damped. The price to pay for that compensation it’s that there is a phase loss of 7 degrees (phase limitation in case of adding a feedback loop) at 200 kHz with compensation instead of 2 degrees loss with no compensation (See Figure 5.3).

The capacitor load is replaced with the piezoelectric actuator (27 nF) and the noise gain compensation yields to same result as for the capacitor.

Figure 5.3: FRF of the voltage amplifier with no compensation (left) and with noise gain compensation (right)
5.2.2 Dynamics of the Push-pull piezoelectric actuators system under voltage amplifier

The push-pull mode is measured at the tip of the capillary and the resulted FRF is shown in Figure 5.4. The push-pull mode is the resonance peak at 127 kHz, the capillary mode is the resonance peak at 49 kHz and the counter mass is at 21 kHz. Compared to the model the amplitude amplification is 5 nm/V instead of 11 nm/V and the resonance peak are shift down according to the model. Those differences come from the lack of symmetry and all sort of internal loss among the piezoelectric actuator (see Timm10). It can be noticed that at 10 kHz the phase loss is already important (~40 degrees), however this phase loss does not come from the push-pull system itself but from the laser vibrometer instrument.

![Bode Diagram](image)

Figure 5.4: frequency response at the top of the capillary

5.2.3 Hysteresis Measurement of the push-pull system under voltage amplifier

The hysteresis is measured by using the test setup in Figure 5.1. A sinusoidal voltage is applied at the input of the high voltage amplifier (HVA) to drive the piezoelectric actuator. The output of the displacement decoder is recorded in order to evaluate the displacement according to the input. Before evaluating the hysteresis at the tip of the capillary (where the displacement results from interaction between two piezoelectric actuators), the hysteresis of a single operating piezoelectric actuator is evaluated using a wire-clamp system (see Error! Reference source not found.). From the hysteresis plot of the clamped piezoelectric actuator one realize that the hysteresis is becoming larger as the actuation voltage increases. Another observation is that the hysteresis becomes even more significant as the frequency increases (see 6.2G.2). Considering the push-pull system, the same conclusion can be made from the nonlinearities exhibited by the tip of the capillary.
Nevertheless a combination of high voltage (>10 Volts) with a high frequency actuation (>100 kHz) of the push-pull system shows a different shape of nonlinearity (saturation phenomenon). The conclusion that comes out from that observation is the push-pull systems is adding additional nonlinearity to the hysteresis exhibited by the piezoelectric actuators.

Figure 5.5: nonlinearity exhibited by the tip of the capillary under voltage actuation

5.2.4 Measurements results of the current control amplifier

The performances of the current control amplifier are evaluated in the frequency domain by analyzing the stability of the frequency response of $V_s/V_{REF}$, $V_l/V_{REF}$ and in the time domain by estimating the percentage reduction of hysteresis. Once again the push-pull system is actuated with to identical current control amplifier (see Figure 5.6). This makes it possible to treat the hysteresis of each piezoelectric actuator independently of the other piezoelectric actuator.
5.2.5 FRF results of the current amplifier

The performances of the current amplifier are evaluated from the open loop FRF (see Figure 5.7) and the tracking control closed loop FRF (see Figure 5.8). Due to the high gain in the open loop, the open loop gain is derived from the sensitivity measurement.

\[ PC = \left( \frac{1}{S} \right) - 1. \]

The stability is firstly evaluated from gain and phase margin of the open as:

<table>
<thead>
<tr>
<th>parameter</th>
<th>result</th>
</tr>
</thead>
<tbody>
<tr>
<td>Gain margin</td>
<td>7 dB</td>
</tr>
<tr>
<td>Phase margin</td>
<td>60 degrees</td>
</tr>
</tbody>
</table>

In comparison to the capacitor load situation (see 6.2H.3) the presence of the electrical resonance of the piezoelectric actuator limits the performances of the controlled systems (lower bandwidth and greater error). The response of the tracking control using a piezoelectric actuator as load is presented in Figure 5.8, where the bandwidth of the current amplifier goes from 5 kHz to 250 kHz. At 200 kHz the tracking control has a phase loss of 20° degrees which can be assumed to be important if someone wants to add
a feedback loop for other purpose. In complement to the tracking control, the FRF of the offset control is displayed as in Figure 5.9.

![Bode Diagram](image1.png)

**Figure 5.8: FRF response of the tracking control Vs/Vref**

![Bode Diagram](image2.png)

**Figure 5.9: offset control VL/Vref**

### 5.2.6 FRF of the Capillary tip under current control amplifier

The response of the capillary tip with current control actuation is shown Figure 5.10. Within the current control bandwidth the FRF magnitude with current control actuation is identical to FRF magnitude with voltage amplifier actuation. At lower frequency the FRF with current control actuation is 90° phase lag according to the FRF with voltage amplifier. This is explained by the fact in current control actuation; the capillary tip displacement is the result of controlling the current going through the piezoelectric actuator. As the piezoelectric actuator behaves as a capacitor, the current is 90° phase shifted according to the voltage across the piezoelectric actuator. Another current control effect is the shifting of the push-pull mode peak from 127 kHz to 131 kHz as noticed in section 3.2.
5.2.7 **Hysteresis measurements under current control actuation**

In this situation the push-pull system is actuated with the current control driver and both the input actuation and the capillary tip displacement are recorded. Sequences of measurement are done at 10 kHz, 30 kHz and 100 kHz to evaluate the performances of the current control. The capillary tip displacement is quasi-linear at 30 kHz whereas at 100 kHz the hysteresis still important. Nevertheless the saturation phenomenon at 100 kHz is solved with current control actuation.

![Bode Diagram](image)

Figure 5.10: FRF of the capillary tip with current control actuation

![Nonlinearities](image)

Figure 5.11: nonlinearities exhibited by the capillary tip under current control and voltage actuation
5.3 Measurement results of the self-sensing

In this part the measurement of the push-pull FRF is done using self-sensing method. The self-sensing of the piezoelectric actuator with capacitor sensor allows measuring the strain (position) acting on the piezoelectric actuator. Unlike the method with laser beam which measure displacement only at the tip of the capillary the self-sensing will measure the displacement at its edges (middle of the capillary holder and middle of the counter mass). Therefore the laser vibrometer is used to make a differential measurement between the top of the counter-mass and the top of the capillary holder. The FRF of the scaled self-sensing with capacitor sensor and the FRF of the scaled laser vibrometer displacement output are shown in Figure 5.12 and one can notice that the push-pull mode is clearly the highest peak. The capillary mode and the counter mass mode are measured as well. The counter-mass mode (first resonance peak) and the push-pull mode (third resonance peak) from laser vibrometer and self-sensing are perfectly matched. The self-sensing method adds extra zeros for the capillary mode (second resonance peak) as noticed in section 4.1.3.

![Bode Diagram](image)

**Figure 5.12: FRF of the self-sensing with capacitor sensor**

The time response of the self-sensing with capacitor is shown in Figure 5.14. The actuating signal (Vref) at the input of the amplifier is compared to the signal at the output of the self-sensing (Um). One can notice that the output of the self-sensing reproduce the actuating signal without any distorsion.

The self-sensing of the piezoelectric actuator with resistor sensor allows measuring the strain rate (velocity). The FRF of the self-sensing with resistor sensor and the FRF of the laser vibrometer velocity output are done in Figure 5.13 and there is any structural...
difference with the one of capacitor sense than adding a derivative operator. One can notice a phase loss of 5 degrees with strain rate self-sensing.

![Bode Diagram](image)

**Figure 5.13:** FRF of the self-sensing with resistor sensor

![Time Response](image)

**Figure 5.14:** Time response of the self-sensing with capacitor

### 5.3.1 Self-sensing with current control

Till now a current control has been designed to reduce the hysteresis effect and a self-sensing circuit has been designed to sense mechanical dynamics. The goal in this part is to take advantage of both circuitries by combining them. Interconnecting the current controller to the strain rate self-sensing is straightforward as a matching leg has to be incorporated to the current controller to form a RC bridge. To complete the scheme the LMS algorithm is added to the bridge network. The complete scheme is shown in **Figure 5.15**.
The frequency response of the self-sensing with current control is displayed in Figure 5.16. At low frequencies there is an amplitude and phase difference according to the laser FRF. This is mainly explained by the fact that there is any feedback in the matching leg and due to the leakage coming from the matching capacitor Cm there is a signal difference between the matching leg and the piezoelectric leg. Nevertheless that mismatch is not affecting the circuitry at the project targeted bandwidth (50 kHz to 200 kHz). The push-pull mode is at 131 kHz as the actuation deals with current control.
5.4 Measurements results of the PPF control of the piezoelectric actuators structure

Due to the mismatch (in term of amplitude and in term of resonance frequencies) between the theoretical modeling estimation and the measurement result of the mechanical behavior of the push-pull system (see section 5.2.2), the differential mechanical behavior across the piezoelectric actuator is modeled from measurements and a new PPF controller is implemented to damp the push-pull mode.

The FRF model of the differential dynamic across the piezoelectric actuator is displayed in Figure 5.17. The target is to model the three resonance modes as close as possible according to the measurement response.

A PPF controller is then implemented to damp the push-pull mode. The natural frequency of the controller is set to match the frequency of the push-pull mode at 127 kHz (see 6.2 Appendix.I for electronic implementation). The response of the PPF closed-loop system is displayed in Figure 5.18. From the result the push-pull mode is damped by 24 db without altering the other resonance modes.
Afterward, the PPF controller used for the model is implemented practically in order to damp the push-pull mode. The response of the PPF controlled system is displayed in Figure 5.19.

The push-pull mode is damped only by 8 db. This is explained mainly by the fact that, it was not possible to produce practically, a PPF controller with sharp peak at 127 kHz (see Figure 5.20. This limitation comes from the bandwidth of the Op-Amp that was used to implement the PPF controller.
Figure 5.20: result of the practically implemented PPF controller.
Chapter 6

Conclusions and recommendations

6.1 Conclusions

At the first place, a current controller has been addressed theoretically and experimentally, the current controller has been addressed in the purpose of reducing the hysteresis inherent to piezoelectric actuators. From the experimentation, the current controller was able to track the reference in a bandwidth from 5 kHz up to 250 kHz. With the use of that current controller, there was a quasi-linear operation (pure sinusoidal movement) of the capillary tip at 30 kHz. However, this quasi-linearity could not be reached at 100 kHz and this can be explained by two reasons. The first reason is coming from current controller limitation at very high frequencies. The current controller cannot allow high open loop gain around the piezoelectric actuator resonance frequency. The second reason comes from the push-pull systems dynamics, as the hysteresis exhibited by the tip of the capillary is more accentuated according to the hysteresis exhibited at the tip of the piezoelectric actuator.

At the second place, a self-sensing method has been addressed theoretically and experimentally. The self-sensing method has been addressed to sense the mechanical behavior of the push-pull systems. The self-sensing method can be used as sensor for feedback control or for monitoring for any horizontal displacement within the push-pull system. However, the capillary mode cannot fully be monitored with this method as the capillary mode is the result of rotational and translation movement. Two kinds of self-sensing has been evaluated and the conclusion that comes out is the strain self-sensing (displacement) shows better performances (less phase loss at low frequencies) than the strain rate (velocity) self-sensing.

At the third place, a self-sensing with current control amplifier has been experimented by combining the current controller and the self-sensing method. With that configuration, it’s possible to control linearly the piezoelectric actuator while in the same time sense the
mechanical behavior. In the bandwidth requirement (50 kHz up to 200 kHz) this configuration shows the same performances as one cited above (quasi-linearity at 30 kHz and strain rate self-sensing).

Finally, a PPF controller is combined to the self-sensing to make a closed-loop feedback control. The PPF control method allowed damping of the push-pull mode by 8 dB. The final conclusion is that, the self-sensing method is reliable enough to be used for monitoring purpose or feedback control purpose.

### 6.2 Recommendations

The current controller can further be optimized:

- by using a piezoelectric actuator with highest resonance frequency. With resonance frequency away from the current control bandwidth, large gain can be introduced to the tracking control loop. This will result in a better tracking control performances and by consequent a better hysteresis reduction.
- By using a voltage amplifier with much larger bandwidth. Therefore, a notch filter can be used to avoid excitation of the piezoelectric actuator resonance frequency.

Still in the purpose of canceling hysteresis, one should explore other methods:

- Open-loop control, where an FPGA or other digital devices can be combined with feedforward control and hysteresis model.
- Closed-loop method with negative feedback control.

To fully monitor the capillary mode, an additive sensor should be used. In that purpose, the integration of sensor like piezoelectric patches to the push-pull system should be explored.

The self-sensing with current control can further be improved by solving the leakage problem. One solution could be to add an extra current feedback amplifier for the matching leg. The goal is to bring both legs of the bridge network to operate in the same condition.

The PPF control technique used in this project can further be improved by designing PPF controller with sharp peak. This can be done by choosing Op-Amp with better performances.
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## Nomenclature

<table>
<thead>
<tr>
<th>Symbol</th>
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<tr>
<td>$u$</td>
<td>nodal displacement</td>
</tr>
<tr>
<td>$S$</td>
<td>strain</td>
</tr>
<tr>
<td>$T$</td>
<td>stress</td>
</tr>
<tr>
<td>$E$</td>
<td>electric field</td>
</tr>
<tr>
<td>$\varepsilon$</td>
<td>permittivity</td>
</tr>
<tr>
<td>$e$</td>
<td>piezoelectric constant</td>
</tr>
<tr>
<td>$M$</td>
<td>mass</td>
</tr>
<tr>
<td>$K$</td>
<td>stiffness</td>
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<tr>
<td>$f$, $F$</td>
<td>force</td>
</tr>
<tr>
<td>$\beta$</td>
<td>damping</td>
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<tr>
<td>$c$</td>
<td>elastic stiffness</td>
</tr>
<tr>
<td>$s$</td>
<td>elastic compliance</td>
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<tr>
<td>$v$, $V$</td>
<td>voltage</td>
</tr>
<tr>
<td>$q$, $Q$</td>
<td>charge</td>
</tr>
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<td>$\phi$</td>
<td>electric potential</td>
</tr>
<tr>
<td>$C_p$</td>
<td>piezoelectric capacitance</td>
</tr>
<tr>
<td>$C_S$</td>
<td>sensing capacitor in the piezo leg</td>
</tr>
<tr>
<td>$C_m$</td>
<td>matching capacitor</td>
</tr>
<tr>
<td>$C_R$</td>
<td>sensing capacitor in the matching leg</td>
</tr>
<tr>
<td>$R_S$</td>
<td>sensing resistor in the current amplifier</td>
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<tr>
<td>$\rho$</td>
<td>mass density</td>
</tr>
<tr>
<td>$A$</td>
<td>cross-sectional area</td>
</tr>
<tr>
<td>$n$</td>
<td>number of layers</td>
</tr>
<tr>
<td>$l$</td>
<td>piezoelectric thickness</td>
</tr>
<tr>
<td>$L$</td>
<td>piezoelectric stack length</td>
</tr>
<tr>
<td>$W$</td>
<td>work</td>
</tr>
<tr>
<td>$[\bullet]$</td>
<td>matrix</td>
</tr>
<tr>
<td>${\bullet}$</td>
<td>vector</td>
</tr>
<tr>
<td>$\delta$</td>
<td>virtual</td>
</tr>
<tr>
<td>$k_{env}$</td>
<td>electro-mechanical stiffness under voltage actuation</td>
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</table>
\( k_{emq} \) - electro-mechanical stiffness under charge actuation
\( k_v \) - stiffness under voltage actuation
\( k_q \) - stiffness under charge actuation
\( f_{3dB} \) - frequency at -3dB
\( f_{GBW} \) - gain bandwidth frequency
\( A_y \) - voltage gain
HVA - high voltage amplifier

<table>
<thead>
<tr>
<th>Superscript</th>
<th>meaning</th>
</tr>
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<tbody>
<tr>
<td>E</td>
<td>-measured under constant field (short circuit)</td>
</tr>
<tr>
<td>S</td>
<td>-measured under constant strain</td>
</tr>
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</table>
Appendix.A  Piezoelectric stack modeling

A.1 Finite element modeling of multi-layers piezoelectric stack actuator in thickness mode

Assuming $E_{33} = -\frac{V}{t}$, $[B_\phi]$ can be computed as $[B_\phi] = -[1/t]$. Assuming simple linear discretization (see [Buch94]) where $N_1 = \frac{l-u}{l}$ and $N_2 = \frac{u}{l}$, then $[N_u] = \begin{bmatrix} l-u & u \\ l & l \end{bmatrix}$ and $[B_\phi] = \begin{bmatrix} \frac{\partial N_1}{\partial u} & \frac{\partial N_2}{\partial u} \end{bmatrix} = \begin{bmatrix} -1 & 1 \\ l & l \end{bmatrix}$.

Therefore the following matrix can be rewritten as:

$$
[M_{uu}] = \int_0^L \rho[N_u]^T[N_u]dV = \int_0^L A\rho \begin{bmatrix} \frac{L-u}{L} & \frac{L-u}{L} \\ \frac{L-u}{L} & \frac{L-u}{L} \end{bmatrix} dL = A\rho l n [2 1] \\
[K_{uu}] = \int_0^L [B_\phi]^T[\varepsilon^E]B_\phi dV = \int_0^L A[\varepsilon^E] \begin{bmatrix} -1 & 1 \\ \frac{L}{1} & \frac{1}{L} \end{bmatrix} dL = \frac{A[\varepsilon^E]}{nl} [1 -1] \\
[K_{u\phi}] = \int_0^L [B_\phi]^T[\varepsilon^S]B_\phi dV = \int_0^L A[\varepsilon^S] \begin{bmatrix} \frac{1}{L} \\ \frac{1}{L} \end{bmatrix} dL = \frac{A[e]}{l} \begin{bmatrix} 1 \\ -1 \end{bmatrix} \\
[K_{\phi\phi}] = \int_0^L [B_\phi]^T[\varepsilon^S]B_\phi dV = \int_0^L A[\varepsilon^S] \begin{bmatrix} -1 & 1 \\ \frac{1}{L} & \frac{1}{L} \end{bmatrix} dL = -\frac{A[n]}{l}$$

Assuming boundaries of zero external forces acting on the piezoelectric stack, rewriting equations (7) results in the following formulation,

$$\begin{align*}
\frac{A\rho l n}{6} \begin{bmatrix} 2 & 1 \\ 1 & 2 \end{bmatrix} \begin{bmatrix} \dot{u}_1 \\ \dot{u}_7 \end{bmatrix} + \frac{A[\varepsilon^E]}{nl} \begin{bmatrix} 1 & -1 \\ -1 & 1 \end{bmatrix} \begin{bmatrix} u_1 \\ u_7 \end{bmatrix} + \frac{A[e]}{l} \begin{bmatrix} -1 \\ 1 \end{bmatrix} \phi &= 0 \\
-\frac{A[e]}{l} \begin{bmatrix} -1 & 1 \\ u_1 & u_7 \end{bmatrix} + \frac{A[n]}{l} \phi &= Q
\end{align*}$$
### A.2 Piezoelectric stack and masses properties

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
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</thead>
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<tr>
<td>Dimensions A x B x TH</td>
<td>2 x 2 x 2 (mm)</td>
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<tr>
<td>Nominal displacement</td>
<td>2.2 μm @ 100V</td>
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<tr>
<td>Blocking force</td>
<td>120 N</td>
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<tr>
<td>Electrical capacitance</td>
<td>25 nF</td>
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<tr>
<td>Resonant frequency</td>
<td>&gt;300 kHz</td>
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Table A-1: piezoelectric stack actuator technical data

<table>
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<tr>
<th>Parameter</th>
<th>Value</th>
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<tr>
<td>Density (ρ)</td>
<td>7.8 (g/cm³)</td>
</tr>
<tr>
<td>Permittivity (ε_{33}/ε₀)</td>
<td>1750</td>
</tr>
<tr>
<td>Coupling factor (k)</td>
<td>0.69</td>
</tr>
<tr>
<td>Piezoelectric charge constant (d_{33})</td>
<td>400x10⁻¹² (C/N)</td>
</tr>
<tr>
<td>Elastic constants (S^{ε}_{33})</td>
<td>20x10⁻¹² (m²/N)</td>
</tr>
</tbody>
</table>

Table A-2: material properties of the piezoelectric stack

<table>
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<tr>
<th>Parameter</th>
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<tr>
<td>Piezo actuator</td>
<td>64 mg</td>
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<tr>
<td>Capillary</td>
<td>100mg</td>
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<tr>
<td>Counter mass</td>
<td>1400 mg</td>
</tr>
<tr>
<td>Capillary holder</td>
<td>73 mg</td>
</tr>
</tbody>
</table>

Table A-3: masses overview
Appendix.B  Self-sensing formulation

B.1 Self-sensing transfer function derivation

For simplification the matrix vector and index notation is omitted then the equation becomes

\[ M\ddot{x} + Kx = -\Theta u_p \]
\[ -\Theta^T x + C_p u_p = q \]  (B.1)

The first equation of B.B.1 becomes in the Laplace domain as;

\[ X(s) = \frac{-M^{-1}\Theta}{s^2 + M^{-1}K} U_p(s) \]  (B.2)

This equation is the transfer function from the piezoelectric displacement to the voltage acting on the piezoelectric actuator.

From the second equation of B.1 the admittance of the piezoelectric actuator is derived;

\[ -\Theta^T x + C_p u_p = q = \frac{I}{s} \]  (B.3)

The admittance is then;

\[ \frac{I(s)}{U_p(s)} = \frac{s\Theta^T M^{-1}\Theta}{s^2 + M^{-1}K} + sC_p = sC_{\text{Mechanical}} + sC_{\text{Electrical}} = Y_M(s) + Y_p(s) \]  (B.4)

The mechanical capacitance is expressed as;

\[ C_{\text{Mechanical}} = \frac{\Theta^T M^{-1}\Theta}{s^2 + M^{-1}K} \]

The electrical capacitance is expressed as;

\[ C_{\text{Electrical}} = C_p \]

\( Y_M(s) \) is the mechanical admittance and \( Y_p(s) \) is the electrical admittance.

Referring to the self-sensing bridge configuration the relation between the current flowing through the piezoelectric actuator and the control voltage can be derived

\[ I(s) = \frac{U_s(s)}{Z_s(s)} = \frac{U_a(s) - U_p(s)}{Z_s(s)} \Rightarrow \frac{U_p(s)}{U_a(s)} = \frac{1}{1 + Z_s(s)(Y_M(s) + Y_p(s))} \]  (B.5)

The transfer function from the output displacement to the control input is done from equations B.2 and B.5.

\[ \frac{X}{U_a} = \frac{X}{U_p \cdot U_a} = \frac{-Y_M}{s\Theta^T \cdot 1 + Z_s(s)(Y_M(s) + Y_p(s))} \]  (B.6)

The transfer function from the sensor voltage to the control input is derived from equation B.5.
\[ U_S = U_a - U_p \Rightarrow \frac{U_S}{U_a} = \frac{Z_S(Y_m(s) + Y_p(s))}{1 + Z_S(Y_m(s) + Y_p(s))} \quad (B.7) \]

The transfer function from the measurement voltage to the control voltage is derived from equation B.7.

\[ U_m = U_S - U_R \Rightarrow \frac{U_m}{U_a} = \frac{Z_S(Y_m(s) + Y_p(s)) - Z_R Y_m}{(1 + Z_S(Y_m(s) + Y_p(s))) (1 + Z_R Y_m)} \quad (B.8) \]

**B.2 Structure properties under balancing conditions**

The poles and the zeros of the self-sensing systems are evaluated according to the balancing conditions.

Let the mechanical admittance be rewritten as:

\[ Y_M(s) = \frac{s\Theta^T M^{-1} \Theta}{s^2 + M^{-1} K} = \frac{s n(s)}{d(s)} \quad (B.9) \]

The transfer function of the self-sensing bridge B.8 can be rewritten as:

\[ \frac{U_m}{U_a} = \frac{(Z_S C_p - Z_R C_m) s d(s) + s Z_S n(s)}{(1 + s Z_R C_m) [(1 + s Z_S C_p) d(s) + s Z_S n(s)]} \quad (B.10) \]

- **Poles analysis**

  The poles of the system are derived from the denominator of transfer function B.10 as:

  \[(1 + s Z_R C_m) = 0 \quad \text{and} \quad [(1 + s Z_S C_p) d(s) + s Z_S n(s)] = 0 \]

  Considering that the contribution of \( n(s) \) is negligible then the self-sensing bridge will add two poles \( s = -\frac{1}{Z_R C_m} \) and \( s = -\frac{1}{Z_S C_p} \).

- **Zeros analysis**

  The zeros of the system are derived from the numerator of the transfer function B.10 as:

  \[ (Z_S C_p - Z_R C_m) s d(s) + s Z_S n(s) = 0 \Rightarrow s Z_R C_m \alpha + s Z_S \frac{n(s)}{d(s)} = 0 \]

  where \( \alpha = \frac{Z_S C_p}{Z_R C_m} - 1 \) is the balancing parameter. The bridge network is perfectly balanced when \( \alpha = 0 \).

\[ s Z_R C_m \alpha + s Z_S \frac{n(s)}{d(s)} = 0 \Rightarrow Z_R C_m \alpha = -Z_S \frac{n(s)}{d(s)} \Rightarrow Z_R C_m \alpha = -Z_S \frac{\Theta^T M^{-1} \Theta}{s^2 + M^{-1} K} \quad (B.11) \]

The zeros of the system is the intersection between \( Z_R C_m \alpha \) and \(-Z_S \frac{\Theta^T M^{-1} \Theta}{s^2 + M^{-1} K}\).
\[ \alpha = -\frac{1}{C_m s^2 + M^{-1}K} \Theta^T M^{-1} \Theta \] with \( Z_S = Z_R \). From numerical computation one notice that

For \( \alpha < \beta = -\frac{1}{C_m M^{-1}K} \Theta^T M^{-1} \Theta \Rightarrow C_m < C_M(s) + C_P(s) \);

then there is an alternate poles-zeros with extra zeros at the end

For \( \alpha > \beta = -\frac{1}{C_m M^{-1}K} \Theta^T M^{-1} \Theta \Rightarrow C_m > C_M(s) + C_P(s) \);

Then there is an alternate zeros-poles with extra zeros at the beginning

Thus the zeros of the structure are migrating according to the balancing parameter

Figure B-1: zeros migration according to the balancing parameter under resistor sensing.

Figure B-2: zeros migration according to the balancing parameter under capacitor sensing.
B.3 Piezoelectric actuator Capacitance measurement

Two sequences of capacitance measurements are implemented. The first sequence measures the capacitance of the piezoelectric actuator according to the actuating frequency change. The second sequence measures the capacitance according to the temperature change.

The capacitance of the piezoelectric actuator is measured practically with the aid of a precision LCR meter (HP 4284A). The frequency of the LCR meter is swept from 10 kHz to 250 kHz and the corresponding capacitance is recorded.

The first sequence of measurements is done at an operating piezoelectric actuator temperature of 26° Celsius. One remarks that the capacitance of the piezoelectric actuator is decreasing as the frequency is increasing (see Figure B-3).

![Capacitance vs frequency at 26 degrees Celsius](image)

Figure B-3: capacitance change according to frequency

The second sequence of measurements is done at 10 kHz and the temperature of the operating piezoelectric actuator is increased. In this situation the capacitance of the piezoelectric actuator increases as the temperature increases (see Figure B-4).

![Capacitance vs temperature](image)

Figure B-4: capacitance change according to temperature
Appendix C  High voltage amplifier design

C.1 High voltage amplifier with MP108 Op Amp

The voltage gain is $A_v = \frac{RI + RF}{RI} = 6$. A resistor RL is dimensioned to limit the output current to $6A$. A resistor RL is dimensioned to limit the output current to $6A$.

C.2 Simulation results with MP108 Op Amp

The Bode plot in Figure C-2 shows the open loop Aol, loop gain $Aol.\beta$ and feedback network $1/\beta$ with resistive load. The second pole of the open loop Aol is far away from the gain bandwidth. Thus Aol can be modeled as first order system.
The Bode plot in Figure C-3 shows the open loop $A_{ol}$, loop gain $A_{ol} \beta$ and feedback network $1/\beta$ with capacitive load. The second pole is close to the closure frequency. This will lead to more peaking, phase loss and eventually to instability as the rate of closure is greater than 20dB/decade.

The Bode plot in Figure C-4 shows the open loop $A_{ol}$, loop gain $A_{ol} \beta$ and feedback network $1/\beta$ with capacitive load and noise gain compensation the rate of closure is brought back to 20dB/decade. This leads to less peaking and better phase margin. Figure C-5 shows the difference between the closed loop without compensation and with compensation.
Figure C-4: Op amp response with capacitive load and compensation

Figure C-5: closed loop transfer function without compensation and with compensation
C.3 Datasheet of the MP108FD Op Amp

FEATURES
- LOW COST
- HIGH VOLTAGE - 200 VOLTS
- HIGH OUTPUT CURRENT - 10 AMPS
- 100 WATT DISSIPATION CAPABILITY
- 300kHz POWER BANDWIDTH

APPLICATIONS
- INKJET PRINTER HEAD DRIVE
- PIEZO TRANSDUCER DRIVE
- INDUSTRIAL INSTRUMENTATION
- REFLECTOMETERS
- ULTRA-SOUND TRANSDUCER DRIVE

DESCRIPTION
The MP108FD operational amplifier is a surface mount constructed component that provides a cost effective solution in many industrial applications. The MP108FD offers outstanding performance that rivals much more expensive hybrid components yet has a footprint of only 4 sq in. The MP108FD has many optional features such as four-wire current limit sensing and external compensation. The 300 kHz power bandwidth and 10 amp output of the MP108FD makes it a good choice for piezo transducer drive applications. The MP108FD is built on a thermally conductive but electrically insulating substrate that can be mounted to a heat sink.

EQUIVALENT CIRCUIT DIAGRAM

INKJET NOZZLE DRIVE
The MP108FD’s fast slew rate and wide power bandwidth make it an ideal nozzle driver for industrial inkjet printers. The 10 amp output capability can drive hundreds of nozzles simultaneously.

EXTERNAL CONNECTIONS
### ABSOLUTE MAXIMUM RATINGS

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
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<tr>
<td>SUPPLY VOLTAGE, +V&lt;sub&gt;G&lt;/sub&gt; to -V&lt;sub&gt;G&lt;/sub&gt;</td>
<td>200V</td>
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<td>SUPPLY VOLTAGE, +V&lt;sub&gt;G&lt;/sub&gt;</td>
<td>+V&lt;sub&gt;G&lt;/sub&gt; +15V&lt;sup&gt;6&lt;/sup&gt;</td>
</tr>
<tr>
<td>SUPPLY VOLTAGE, -V&lt;sub&gt;G&lt;/sub&gt;</td>
<td>-V&lt;sub&gt;G&lt;/sub&gt; -15V&lt;sup&gt;6&lt;/sup&gt;</td>
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<td>12A, within SOA</td>
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<td>INPUT VOLTAGE</td>
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<td>TEMPERATURE, junction&lt;sup&gt;2&lt;/sup&gt;</td>
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### SPECIFICATIONS

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<td>OPEN LOOP @ 15Hz</td>
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<td>degrees</td>
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<td></td>
<td>V</td>
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<td>VOLTAGE SWING</td>
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<td>+V&lt;sub&gt;G&lt;/sub&gt; -10</td>
<td>V</td>
<td></td>
<td></td>
</tr>
<tr>
<td>VOLTAGE SWING</td>
<td>I&lt;sub&gt;D&lt;/sub&gt; = 10A</td>
<td>-V&lt;sub&gt;G&lt;/sub&gt; +10</td>
<td>V</td>
<td></td>
<td></td>
</tr>
<tr>
<td>VOLTAGE SWING</td>
<td>I&lt;sub&gt;D&lt;/sub&gt; = 10A, +V&lt;sub&gt;G&lt;/sub&gt; -V&lt;sub&gt;G&lt;/sub&gt; +10V</td>
<td>V</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>CURRENT, continuous, DC</td>
<td></td>
<td>10</td>
<td></td>
<td></td>
<td>A</td>
</tr>
<tr>
<td>SLEW RATE, A&lt;sub&gt;g&lt;/sub&gt; = 20</td>
<td></td>
<td>150</td>
<td>170</td>
<td></td>
<td>V/μS</td>
</tr>
<tr>
<td>SETTLING TIME, to 0.1%</td>
<td>2V Step</td>
<td>1</td>
<td></td>
<td></td>
<td>μS</td>
</tr>
<tr>
<td>RESISTANCE</td>
<td>No load, DC</td>
<td>5</td>
<td></td>
<td></td>
<td>Ω</td>
</tr>
<tr>
<td>POWER BANDWIDTH 180V&lt;sub&gt;G&lt;/sub&gt;</td>
<td>C&lt;sub&gt;D&lt;/sub&gt; = 10pF, +V&lt;sub&gt;G&lt;/sub&gt; = 100V, -V&lt;sub&gt;G&lt;/sub&gt; = -100V</td>
<td>300</td>
<td></td>
<td></td>
<td>kHz</td>
</tr>
<tr>
<td>POWER SUPPLY</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>VOLTAGE SWING</td>
<td>±15</td>
<td>±50</td>
<td>±100</td>
<td></td>
<td>V</td>
</tr>
<tr>
<td>CURRENT, quiescent</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td>mA</td>
</tr>
<tr>
<td>THERMAL</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>RESISTANCE, AC, junction to case&lt;sup&gt;5&lt;/sup&gt;</td>
<td>Full temperature range, f &lt; 60Hz</td>
<td>1</td>
<td></td>
<td></td>
<td>°C/W</td>
</tr>
<tr>
<td>RESISTANCE, DC, junction to case</td>
<td>Full temperature range, f &lt; 60Hz</td>
<td>1.25</td>
<td></td>
<td></td>
<td>°C/W</td>
</tr>
<tr>
<td>RESISTANCE, junction to air</td>
<td>Full temperature range</td>
<td>13</td>
<td></td>
<td></td>
<td>°C/W</td>
</tr>
<tr>
<td>TEMPERATURE RANGE, case</td>
<td>-40</td>
<td></td>
<td></td>
<td>85</td>
<td>°C</td>
</tr>
</tbody>
</table>

### NOTES:

1. Unless otherwise noted: T<sub>0</sub>=25°C, compensation C<sub>D</sub>=100pF, DC input specifications are value given, power supply voltage is typical rating.
2. Long term operation at the maximum junction temperature will result in reduced product life. Derate internal power dissipation to achieve high MTBF.
3. Doubles for every 10°C of case temperature increase.
4. +V<sub>G</sub> and -V<sub>G</sub> denote the positive and negative supply voltages to the output stage. +V<sub>G</sub> and -V<sub>G</sub> denote the positive and negative supply voltages to the input stages.
5. Rating applies when the output current alternates between both output transistors at a rate faster than 60Hz.
6. Power supply voltages +V<sub>G</sub> and -V<sub>G</sub> must not be less than +V<sub>G</sub> and -V<sub>G</sub> respectively.
Appendix.D Current controller for piezoelectric actuator

D.1 Electronic Schematic design for the current controller
D.2 Current controller simulations results

The bode plot of the current tracking transfer function $I_p/V_{REF}$ is represented in Figure D-1 where the performances are defined as follows:

<table>
<thead>
<tr>
<th>bandwidth</th>
<th>Gain margin</th>
<th>Phase margin</th>
</tr>
</thead>
<tbody>
<tr>
<td>10kHz to 300 kHz</td>
<td>6.3 dB</td>
<td>60 degrees</td>
</tr>
</tbody>
</table>

![Figure D-1: bode plot of the current tracking transfer function.](image1)

The bode plot of the offset control transfer function $V_L/V_{REF}$ is represented in Figure D-2 where at low frequencies (<1 Hz) the output voltage is damped.

![Figure D-2: bode plot of the offset control transfer function.](image2)
Appendix.E  Electronic circuitry of the adaptive self-sensing
Appendix. F  Piezoelectric clamping device

The piezoelectric clamping device is used to test each piezoelectric actuator individually. The time and FRF results of the piezoelectric actuator are first estimated in piezoelectric clamp configuration and later estimated in the push-pull system. The piezoelectric clamping device is represented Figure F-1. F represents the external force acting on the piezoelectric actuator, M represents the mass of the clamping structure, x represents the displacement, T represents the kinetic energy and V the strain energy of the structure.

![Piezoelectric clamping structure](image)

Figure F-1: Piezoelectric clamping structure.

From energy conservation and Lagrangian formulation stated in section 2.4 the equations governing the clamping structure are;

\[ M \ddot{x} + (K_a + K)x = K_a nd_{33} V + F \]
\[ C_p V + nd_{33} K_a x = Q \]

Practically the displacement of the clamping structure is measured by using the differential functionality of the laser beams measurement device.
Appendix.G Measurements results of the high voltage amplifier

G.1 Dynamic results of the voltage amplifier

The voltage amplifier is first tested with a capacitor of 25 nF as load. The FRF is firstly recorded when there is no compensation added to the amplifier. The compensation technique damp the resonance but reduce the cross over bandwidth and increase the phase shift at 200 kHz from 2 degrees to 5 degrees.

![Figure G-1: FRF of the voltage amplifier with no compensation.](image)

Next the FRF is recorded with added noise gain compensation (Figure G-2).

![Figure G-2: FRF of the voltage amplifier with noise gain compensation.](image)
Later on the capacitor is replaced by the piezoelectric actuator and the FRF measurement are done to confirm the result got when the load is a capacitor. From the FRF measurement one can realize that there is no fundamental difference between the situation when the load is a capacitor and the situation when the load is a piezoelectric actuator. However the inherent resonance of the piezoelectric actuator at 600 kHz should not be confused with the resonance caused the capacitive load at 1.8 MHz.

Figure G-3: FRF of the voltage amplifier with piezoelectric actuator and no compensation.

Figure G-4: FRF of the voltage amplifier with piezoelectric actuator and compensation.
G.2 Hysteresis result of a single clamped piezoelectric

The piezoelectric clamping device in this situation is used to have an idea of the hysteresis exhibited by a single piezoelectric actuator. As stated early the hysteresis does increase with voltage and frequency (see Figure G-5).

![Hysteresis Graphs](image)

Figure G-5: hysteresis exhibited by a clamped piezoelectric actuator.
Appendix.H Measurements results of the current control amplifier

H.1 FRF of the Process dynamics

Referring to section 3.4.1, before starting any control loop the process dynamics ($V_s/V_{in}$) has to be evaluated. In the current controller the process dynamics is the current flowing through the sensing resistor $V_s$ over the voltage amplifier input $V_{in}$. The process dynamics is shown in Figure H-1. At 600 kHz there is a resonance peak which is the electrical resonance exhibited by the piezoelectric actuator.

![Figure H-1: process dynamics.](image)

H.2 FRF of the controller

The response of the controller used in the practical implementation is shown Figure H-2
However a high pass filter with roll off frequency at 100 Hz is added to the PI controller to get rid of a 50 Hz hum corrupting the output voltage of the current amplifier that is used to actuate the piezoelectric actuator. The response of the modified controller is shown in Figure H-3.

**Figure H-3: PI controller FRF response.**

H.3 Tracking control response

To test the controlled system a capacitor is used as load (instead of piezoelectric actuator) and the resulting FRF of the tracking control loop is shown in Figure H-4. The current control bandwidth goes from 2 kHz to 350 kHz. The phase loss at 200 kHz is 18° degrees.
**H.4 Hysteresis measurements of a clamped piezoelectric actuator under current control**

Two sequences of measurements are done to evaluate the hysteresis of a single clamped piezoelectric actuator under current control. In first test both the voltage amplifier and the current amplifier are actuated at 1 kHz. The output of the laser decoder is recorded at the same time as the actuating input VREF. After the input scaling of the current amplifier according to the voltage amplifier the hysteresis is done by plotting the displacement according to the actuating input. At 1 kHz there is a linear operation of the clamped piezoelectric actuator. At 100 kHz the hysteresis is reduced with a percentage of 65 %.
Appendix.I  Electronic implementation of the PPF controller

\[ f_c = \frac{1}{2\pi\sqrt{R1C1R2C2}} \]

Transfer function:

\[ \frac{V_{out}(s)}{V_{in}(s)} = \frac{G(2\pi f_c)^2}{s^2 + 2\zeta(2\pi f_c)s + (2\pi f_c)^2} \]

\[ Q = \frac{1}{2\zeta} \]

\[ G = \frac{R3 + R4}{R3} \]
Appendix.J Measurement equipments

J.1 Scopes

Handyscope HS3 is the instrument used as function generator and oscilloscope along the project. It’s a PC based data acquisition instrument from TIEPIE. Its main technical data is shown Table J-1

<table>
<thead>
<tr>
<th>Parameter</th>
<th>indication</th>
</tr>
</thead>
<tbody>
<tr>
<td>Number of inputs</td>
<td>2</td>
</tr>
<tr>
<td>Number of outputs</td>
<td>1</td>
</tr>
<tr>
<td>Voltage range</td>
<td>0.2V up to 80V</td>
</tr>
<tr>
<td>Frequency range</td>
<td>10 Hz up to 10 MHz</td>
</tr>
<tr>
<td>Measuring points</td>
<td>Up to 131060</td>
</tr>
</tbody>
</table>

Table J-1: HS3 technical data.

J.2 Network analyzer

The HP 4395A is the network analyzer used to make and record the frequency response. Its main technical data is shown in

<table>
<thead>
<tr>
<th>Parameter</th>
<th>indication</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency range</td>
<td>10 Hz up to 500 MHz</td>
</tr>
<tr>
<td>resolution</td>
<td>1mHZ</td>
</tr>
<tr>
<td>Voltage range</td>
<td>10 mV up to 1V</td>
</tr>
</tbody>
</table>

Table J-2: HP 4595A technical data

J.3 Laser vibrometer

The fiber-optic vibrometer Polytec OFV-552 is used to measure the vibration. It determines the velocity and displacement from the Doppler shift of back scattered laser light. For more literature about it look at [www.polytec.com/eur/158_1049](http://www.polytec.com/eur/158_1049). It is coupled to a vibrometer controller OFV-5000 for velocity and displacement high resolution.
J.4 Thermistor

The thermistor used to record the temperature is a Sensordata NTC thermistor.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>indication</th>
</tr>
</thead>
<tbody>
<tr>
<td>Temperature range</td>
<td>-40°C to 105°C</td>
</tr>
<tr>
<td>accuracy</td>
<td>± 0.1°C</td>
</tr>
<tr>
<td>Thermal time constant</td>
<td>5 sec.</td>
</tr>
<tr>
<td>Resistance</td>
<td>10 kOhm</td>
</tr>
</tbody>
</table>

Table J-3: thermistor technical data