ON DISTURBANCE ATTENUATION PROPERTIES OF CONTROL SCHEMES FOR EULER-LAGRANGE SYSTEMS: THEORETICAL AND EXPERIMENTAL RESULTS.

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Abstract

In this paper we analyse and experimentally verify the (local) disturbance attenuation properties of some asymptotically stabilizing nonlinear controllers for Euler-Lagrange systems reported in the literature. Our objective with this study is twofold: first, to compare the performance of these schemes from a perspective different from stabilizability; second, to quantify the basic tradeoff between robust stability and robust performance for these designs. We consider passivity-based and feedback linearization schemes developed for the control of DC-to-DC converters and rigid robots. The results are readily checked and analysed for the DC-to-DC converter in the experimental set-up.

1 Introduction

A lot of research in recent years has been devoted to the problem of developing control algorithms for mathematical models of physical systems. In view of its practical interest many researchers have concentrated on mechanical, electromechanical and power electronic systems. Several alternative approaches have been taken to design asymptotically stabilizing controllers for these systems. For instance, passivity concepts have been invoked to control robots, e.g., [20], [18], [12], [3], induction motors [7], and DC-to-DC converters [16]. Feedback linearization is another technique that is used to control these systems, e.g., [19], [11], [15].

In applications these systems are typically subject to external disturbances. For instance, the regulated voltage in converter devices is perturbed by fluctuations in the external voltage source. Our main motivation in this paper is to analyse and compare the disturbance attenuation properties, and simultaneously the robust stability measure, of some (local) nonlinear controllers for Euler-Lagrange systems reported in the literature. We use tools provided by the recent theoretical research on the analysis of the $\mathcal{L}_2$-gain of nonlinear systems, e.g., [9], [21]. Furthermore, we show the results of some experiments that were performed to verify the results.

The remaining of the paper is organized as follows. In Section 2 we consider DC-to-DC "boost" converters, and its controller schemes. We present some of the disturbance attenuation properties of these schemes. In Section 3 we briefly present a similar $\mathcal{L}_2$ gain analysis for rigid robots. We continue in Section 4 with the experimental results on the DC-to-DC converter. We wrap up the paper with some concluding remarks.

2 DC-to-DC Converters

2.1 Model and Stabilization Problem

We consider the switch–regulated "boost" converter circuit of Figure 1. A Pulse Width Modulation (PWM) policy regulating

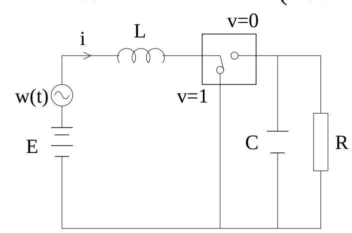

Figure 1: "Boost" converter circuit

the switch position function v, may be specified as follows,

$$v(t) = \begin{cases} 1 \text{ for} & t_k \leq t < t_k + \mu(t_k)T \\ 0 \text{ for} & t_k + \mu(t_k)T \leq t < t_k + T \end{cases}$$

where t_k represents a sampling instant defined by $t_{k+1} = t_k + T,\ k = 0, 1, \ldots$,; the parameter $T > 0$ is the fixed sampling period, also called the duty cycle. The duty ratio function, $\mu(\cdot)$, ranging on the closed interval $[0, 1]$, is the control input to the average PWM model given by [16]

$$\begin{aligned} \dot{z}_1 &= -u\,\frac{1}{L} z_2 + \frac{E + w}{L} \\ \dot{z}_2 &= u\,\frac{1}{C} z_1 - \frac{1}{RC} z_2 \end{aligned} \qquad (1)$$

where $u := 1 - \mu$, and we denote by z_1 and z_2 the average input inductor current and the average output capacitor voltage, respectively. As discussed in [16] this model accurately describes the behavior of the converter provided the switching is sufficiently fast and the capacitor voltage is bounded away from zero, i.e., $x_2 \geq \varepsilon > 0$.

2.2 Control Laws

In this subsection we recall two control laws proposed in the literature to regulate (1). In the absence of external disturbances, i.e., when $w \equiv 0$, they both achieve (local) asymptotic stabilization, that is, they insure that for suitable initial conditions $z_2 \to E_o = \text{const} > E$ with internal stability.

A. Passivity-based controller

In [16] the following (nonlinear dynamic state feedback) controller that preserves passivity of the closed loop was proposed

$$\begin{aligned} \dot{z}_{2d} &= -\frac{1}{RC}\left\{ z_{2d} - \frac{E_o^2}{E z_{2d}}\left[E + R_1\left(z_1 - \frac{E_o^2}{RE}\right)\right]\right\}, \\ & z_{2d}(0) > 0 \qquad (2) \\ u &= \frac{1}{z_{2d}}\left[E + R_1\left(z_1 - \frac{E_o^2}{RE}\right)\right] \qquad (3) \end{aligned}$$

where $R_1 > 0$ is a design parameter that injects the damping required for asymptotic stability.

B. Feedback linearizing controller

In [15] a (nonlinear static state feedback) controller that linearizes the input-output behaviour of the system was proposed as follows. Consider the circuits total energy that is given by $H := \frac{1}{2}(Lz_1^2 + Cz_2^2)$. Applying the control

$$\begin{aligned} u &= \frac{1}{(\frac{E}{L} + \frac{2}{RC}z_1)z_2}\left\{ \left(\frac{2}{R^2C} - \frac{a_1}{R} + \frac{a_2 C}{2}\right) z_2^2 \right. \qquad (4) \\ & \left. + \left(a_1 E + \frac{a_2 L}{2} z_1\right) z_1 + \frac{E^2}{L} - a_2 H_d \right\} \end{aligned}$$

where $a_1, a_2 > 0$ are the design parameters, and

$$H_d := \frac{E_o^2}{2}\left(C + \frac{L}{R^2E^2}E_o^2\right).$$

yields in the new coordinates $[H, \dot{H}]$ the closed loop linear model given by

$$\ddot{H} + a_1\dot{H} + a_2 H = a_2 H_d \qquad (5)$$

Notice that H_d is chosen such that as $H \to H_d$ we have $z_2 \to E_o$ as desired.

2.3 Disturbance Attenuation Properties

In this subsection we will give the results of studying the disturbance attenuation capabilities of the two controllers given above. Since both controllers achieve local asymptotic stabilizability, the disturbance attenuation capabilities we study are obviously local too. Furthermore, by the Total Stability Theorem (see [8]) the internal stability of the closed loop system implies that the solutions still exists in a neighborhood of the equilibrium for the disturbances $w \in \mathcal{L}_2 \cap \mathcal{L}_\infty^c$, where $\mathcal{L}_\infty^c = \{ w \,|\, \sup_t \| w \| \}$. In order to use this result we have to consider small signal disturbances, i.e., in the remainder we assume without further mentioning that we are dealing with small signal disturbances.

Towards this end, we will evaluate some bounds on the achievable $\mathcal{L}_2$ gain of the closed loop operator from the external disturbance w to the regulated output z_2. The qualifier "achievable" stems from the fact that these bounds are independent of the controller parameters.

It should be remarked that, contrary to what is often the case in the literature, we do *not* treat stability together with the disturbance attenuation.

Preliminary Lemma

First, we present a lemma which establishes an $\mathcal{L}_2$ gain property of the passivity based controller. This lemma will be instrumental for the analysis below.

Lemma 2.1 Consider the system (1) in closed loop with the controller (2), (3). Then, the $\mathcal{L}_2$ gain[1] of the operator $T_{w\tilde{z}_2} : w \mapsto \tilde{z}_2$, where $\tilde{z}_2 := z_2 - z_{2d}$, can be made arbitrarily small with a suitable choice of the design parameter R_1.

Proof

Define $\tilde{z} := z - [\frac{E_o^2}{RE}, z_{2d}]^T$. We find it convenient o write the equations in state space form

$$\dot{x} = f(x) + g(x)w. \qquad (6)$$

where we have defined $x = [x_1, x_2, x_3] := [\tilde{z}_1, \tilde{z}_2, z_{2d}]$, and $f(x)$, $g(x)$ are obtained in an obvious manner. The equilibrium of this system is given by $f(x_0) = 0, x_0 = [0, 0, E_o]$. As "output" signal we take $y = h(x) := x_2$. We know that (e.g. [21], Theorem 2) if for $\gamma > 0$ there exists a smooth nonnegative solution $V(x)$ to the following Hamilton-Jacobi inequality

$$\begin{aligned} & \frac{\partial V}{\partial x}(x) f(x) + \frac{1}{2}\frac{1}{\gamma^2}\frac{\partial V}{\partial x}(x) g(x) g(x)^T \frac{\partial^T V}{\partial x}(x) \\ & + \frac{1}{2} h^T(x) h(x) \leq 0, \quad V(x_0) = 0, \qquad (7) \end{aligned}$$

then the $\mathcal{L}_2$ gain of $T_{w\tilde{z}_2}$ is smaller than or equal to γ. The function $V = \frac{1}{4}R\tilde{z}^T\mathcal{D}\tilde{z}$, where $\mathcal{D} = \text{diag}(L, C)$ is a solution of (7) provided that $\gamma^2 \geq \frac{R}{4R_1}$. Thus, γ can be made arbitrarily small by choosing R_1 arbitrarily large. This concludes the proof. □□□

From the lemma above we see that increasing the damping (R_1) we decrease the effect of the disturbances on the signal $\tilde{z}_2$. On the other hand, from closer observations of the previous lemma, one might be tempted to try a high-gain design, which a more careful analysis reveals not to be a good idea. To see this notice that $\tilde{z}_2 \to 0$ does not imply that $z_2 \to E_o$ as desired, unless $z_{2d} \to E_o$ as well. To study the behavior of the latter consider the signal $\eta := \frac{1}{2}(z_{2d}^2 - E_o^2)$, which satisfies

$$RC\dot{\eta} = -2\eta + \frac{R_1 E_o^2}{E}\tilde{z}_1$$

This equation clearly shows that increasing the damping will induce a "peaking" in η, and consequently a slower convergence of $z_{2d} \to E_o$.

[1] We recall that the $\mathcal{L}_2$ gain of an operator $T : \mathcal{L}_2^m \to \mathcal{L}_2^n$ is defined as

$$\|T\|_{i2} \triangleq \sup_{u \in \mathcal{L}_2} \frac{\|Tu\|_2}{\|u\|_2}$$

where $\|\cdot\|_{i2}$ is the $\mathcal{L}_2$ induced norm, and $\|\cdot\|_2$ the $\mathcal{L}_2$ norm, see e.g. [5]

Lower Bounds

It is important to remark that the "ideal" disturbance attenuation property of the passivity based controller established in lemma 2.1 is with respect to the output signal $\tilde{z}_2$, while the actual regulated output of the system is z_2. Unfortunately, in the following theorem we are able to prove with the help of a result that relates the $\mathcal{L}_2$-gain of the nonlinear system with the $\mathcal{L}_2$-gain of its linearization, that for neither one of the controllers we can actually obtain arbitrarily small disturbance attenuation for z_2.

Proposition 2.2 Consider the system (1) in closed loop with the passivity-based controller (2), (3). Then, the $\mathcal{L}_2$ gain of the operator $T_{wz_2} : w \mapsto z_2$ satisfies the lower bound

$$\|T_{wz_2}\|_{i2} \geq \frac{E_o}{2E}$$

On the other hand, for the linearizing controller (4) we have the lower bound

$$\|T_{wz_2}\|_{i2} > \frac{L^3E_o^3}{E^3CR^2 + 2LEE_o^2}$$

Proof

The proof is based on the linearization of the closed loop systems, and Proposition 6 of [21], which proves that if the linearized system has $\mathcal{L}_2$ gain $> (\geq)\gamma$ then the original nonlinear system also has $\mathcal{L}_2$ gain $> (\geq)\gamma$. □□□

An Upper Bound for the Passivity-Based Controller

The above proposition gives limits on the achievable disturbance attenuation which depend on the system parameters and desired set point, but are independent of the design parameters. In the following proposition we give a formulation for the upper bound on the disturbance attenuation for the passivity based controller.

Proposition 2.3 Consider the system (1) in closed loop with the controller (2), (3) with the design parameter $R_1 < \frac{E}{E_o}$. Then, for all disturbances w such that $z_{2d}(t) > -\frac{1}{2}E_o$, we have the bound

$$\|z_2\|_2 \leq \tilde{\gamma}\|w\|_2$$

where $\tilde{\gamma}$ is the solution to the following optimization problem

$$\tilde{\gamma}^2 \quad := \min_{K_3,K_4} \frac{1}{2R_1} \cdot \frac{K_3K_4^2E_o}{K_4 - 1}$$

with $R_1 \leq \frac{E}{E_o}\left(-\frac{1}{K_3}+1\right)$, $K_3K_4 \geq \frac{RE}{E_o}$, $K_3 > 1$, and $K_4 > 1$.

Proof

The proof is given by careful analyzing and extending the results of Lemma 2.1, see e.g. [14]. □□□

The following remarks are in order:

- Even though we can not solve analytically the optimization problem posed above, standard software can be used to find $\tilde{\gamma}$ for a given system and a damping gain satisfying $R_1 < \frac{E}{E_o}$. It is interesting to note that the latter bound exhibits again the tradeoff between robust stability and robust performance to external disturbances. This stems from the fact that R_1, which relates with the convergence rate as explained in Section 2.3.1, cannot be chosen larger than $\frac{E}{E_o}$ to insure the disturbance attenuation $\tilde{\gamma}$. Furthermore, the expressions above provide some useful guidelines for the selection of the system parameters to enhance the disturbance attenuation properties of the amplifier.
- Notice from (3) that to avoid singularities the controller state z_{2d} should be always positive and bounded away from zero. As discussed in [16] this requirement, which is consistent with the domain of validity of the averaged model, is needed even in the absence of external disturbances. Hence, the assumption made above on the disturbances is by no means restrictive in the present context.

3 Rigid Robots

In this section we briefly consider the problem of attenuation of input disturbances in rigid robots performing a trajectory tracking task. In this case we will provide conditions on the controller tuning parameters such that both, passivity-based and feedback linearization schemes, yield closed loops with arbitrarily good disturbance attenuation properties without compromising the convergence rate. The solution of the Hamilton-Jacobi inequality follows from a similar analysis as in Lemma 2.1, by considering the physics of the system, and adding a term to fulfill the inequality. In other related work (i.e., [13], [1],[2]) the problem of *designing* a controller such that the closed loop satisfies an $\mathcal{H}_\infty$ bound is considered. This in contrast with our approach (see [14]) where the emphasis is on deriving conditions for existing closed loop schemes to achieve such disturbance attenuation properties.

It is well known (e.g. [19]) that the free dynamics of rigid robots (with rotational joints) are described by

$$D(q)\ddot{q} + C(q,\dot{q})\dot{q} + G(q) = u + w \tag{8}$$

where $q \in \mathcal{R}^n$ denotes the joint angular positions, $d_1I \geq D(q) \geq d_2I > 0$ the inertia matrix, $C(q,\dot{q})\dot{q}$ the centrifugal and Coriolis forces, $G(q)$ the gravitational forces, and u the input torques, which we assume are perturbed by some external disturbance w.

The kjth element of $C(q,\dot{q})$ is univocally defined from the elements of $D(q)$ via the Christoffel symbols of the first kind [12] such that

$$\dot{D}(q) = C(q,\dot{q}) + C^T(q,\dot{q}) \tag{9}$$

In the absence of external disturbances, a globally exponentially stable controller that preserves passivity in closed loop is given by

$$\begin{aligned} u \quad = \quad & -D(q)(\ddot{q}_d - \lambda\dot{\tilde{q}}) - C(q,\dot{q})(\dot{q}_d - \lambda\tilde{q}) \\ & -k_1(\dot{\tilde{q}} + \lambda\tilde{q}) - k_2\tilde{q} + G(q) \end{aligned} \tag{10}$$

where $q_d(t)$ is the desired angular trajectory, $\tilde{q} = q - q_d$, and $\lambda, k_1, k_2 > 0$ are design parameters, see e.g. [3]. We are interested here in the choice of these parameters for optimal attenuation of the torque disturbance on the position and speed tracking errors. The solution to this problem is summarized in the proposition below. As mentioned before, a similar result, as well as its extension to flexible joint robots, may be found in the works of [1], [2].

Proposition 3.1 Consider (8) in closed loop with (10) with the output signal $z := [\tilde{q}, \dot{\tilde{q}}]^T$. For a fixed $\gamma > 0$, assume

$$k_1 > \frac{1}{2}\left(\frac{1}{\gamma^2} + 1 + \lambda\right), \quad k_2 > \frac{1}{2\lambda}(\lambda^2 + \lambda + 1)$$

Under these conditions, the $\mathcal{L}_2$ gain of the operator $T_{wz} : w \to z$ satisfies the bound $\|T_{wz}\|_2 \leq \gamma$. Consequently, arbitrarily good disturbance attenuation is achievable by increasing the gain k_1.

Proof
The proof follows immediately by plugging the quadratic function $V(s, \tilde{q}) := \frac{1}{2}s^T D(q)s + \frac{k_2}{2}|\tilde{q}|^2$ in the Hamilton-Jacobi inequality (7), where we have defined $s := \dot{\tilde{q}} + \lambda.\tilde{q}$, □□□

In contrast with the converter problem, in this case there is no tradeoff between converge rate and disturbance attenuation to be made. This seems to stem from the fact that rigid robots are fully actuated systems, that is, the number of degrees of freedom is equal to number of controls.

Its easy to see that a similar property is enjoyed by the feedback linearization (computed torque) controller.

4 Experimental results

4.1 Configuration for the DC-to-DC converter

The experimental card was assembled using low cost commercial electronic elements placed on a card designed in the laboratory. In Fig. 2 we show the experimental set-up consisting of the boost circuit card that receives control signals from a D/A converter of a DSpace card placed in a PC. Two DC power supplies are necessary, one of them to provide energy to the system (we'll refer to it as the power supply in the rest of the paper), and the other one to feed the electronic part of the card.

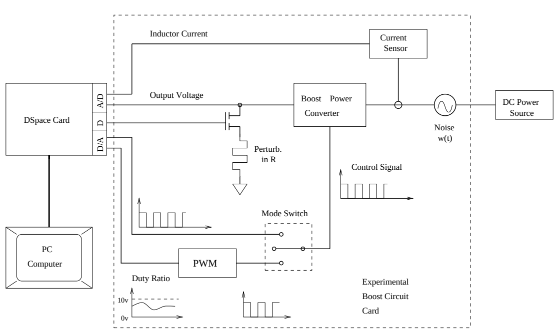

Figure 2: Experimental set-up

4.1.1 Boost circuit card description

The main card is formed by a boost circuit, a pulse width modulation circuit (PWM), a current sensor, a current to voltage converter and a voltage divisor functioning as signal conditioners.

The boost circuit is composed by an inductor, a capacitor, a resistive charge and a switch, the last one is implemented by interconnecting a FET transistor and a rapid diode in a suitable manner, and fed by a DC power supply. The values of the elements can be found in table 1.

Element	Value	Unities
Capacitance	1000	μF
Inductance	170	mH
Resistance	100	Ω
Power supply	10	*Volt*

Table 1: Values of the elements

A current sensor is introduced, which is useful in the control laws. We can connect or disconnect another resistive charge to the output by means of a digital signal coming from the DSpace card, giving us the possibility to introduce disturbances in the resistive charge. A driver has been interconnected with the power supply and the circuit in order to add disturbances to the power source.

4.2 Experiments

The two control schemes that we study in the present article, i.e., the Passivity Based Controller (PBC) and the Feedback Linearizing Controller (FLC), have been implemented. We study the closed loop behaviour for disturbances in the power supply, for variation in the output resistance, and for changes in the desired output voltage. In all the experiments we have as a desired output voltage $V_d = 20$ Volt unless otherwise is indicated.

4.2.1 Typical responses

In Fig. 3 and Fig. 4 we show the typical responses for the system under PBC and FLC, respectively, they include the inductor current $x_1(t)$, capacitor voltage $x_2(t)$ duty ratio $\mu(t)$ and for the PBC also the desired capacitor voltage z_{2d}. The families of curves correspond to different values in the parameters R_1 for the PBC and a_1, a_2 for the FLC.

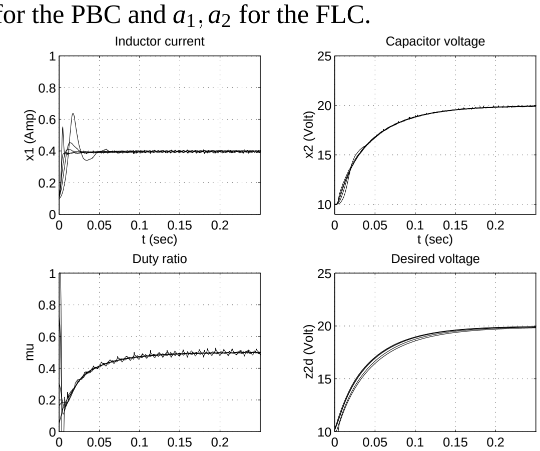

Figure 3: Typical responses for PBC

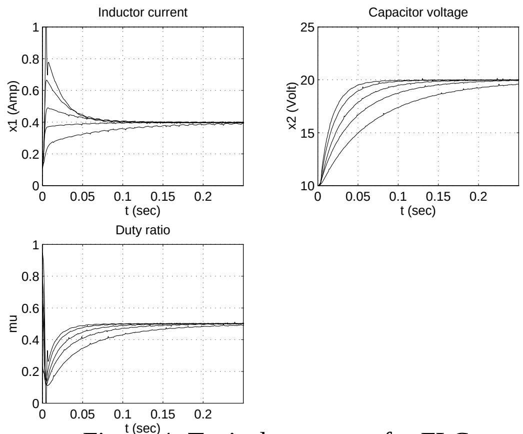

Figure 4: Typical responses for FLC

4.2.2 Frequency response to periodic disturbances in the source

In this experiment we consider the frequency responses of the output voltage under periodic perturbations introduced in the power supply. Hence, we add to the E source a perturbation $w(t) = A_w sin(2\pi f t)$ for different values of f, and for $A_w = 3$ Volts. We can conclude form the experiment that these frequency responses are similar to low pass filters.

In the case of the PBC, the cutoff frequency goes from less than $0.1Hz$ to $8.1Hz$ whereas for the FLC it goes from less than $0.1Hz$ to $4.5Hz$. This frequency depends on the given power converter parameters and we can try to vary it by changing the design parameters R_1 and a_1, a_2 respectively. For PBC big values of R_1 produce small values of this frequency, and for the FLC we can see that fast poles also reduce the cutoff frequency. It is important to remark that the cutoff frequency is relatively small compared with the natural line frequency noise, which means that the rejection of this kind of natural perturbations is assured.

4.2.3 Tracking a desired voltage signal

The controllers are designed for regulation, but in some applications it is desirable for the closed loop system to follow a time-varying output signal $V_d(t)$. The experiment is performed with signals of the form $V_d(t) = V_{d0} + A_{Vd} sin(2\pi f t)$, and we initiate in V_{d0}. This means we only consider the alternating part of the response.

From the frequency responses for the tracking problem, we obtain the cutoff frequencies for both controllers, i.e., for the PBC $f_c = 3.1Hz$ which is fixed independently of the value of the parameter R_1 and for the FLC it could go from $f_c = 2.0Hz$ to $f_c = 10.5Hz$, where the largest value can be obtained by chosen a_1, a_2 in such a way that the corresponding damping coefficient is small and the natural frequency high.

The typical tracking responses of the closed loop systems have been obtained (but are not shown due to space limitations), where we have chosen parameters that give similar typical responses. The value of the frequency for the desired signal is set to $f = 2.0Hz$ in both cases. It can be observed that in this particular case the response of the FLC scheme has an smaller delay with respect to the desired signal.

4.2.4 Response to an $\mathcal{L}_2$ disturbance signal

In this experiment we add an $\mathcal{L}_2$ disturbance signal to the power supply. We do this by means of a squared signal from a signal generator. Characteristics of this signal are: amplitude $3Volt$ and duration $0.1sec$. In Fig. 5 and 6 we can see the response of the circuit variables again for different parameters values when this disturbance $w(t)$ is applied.

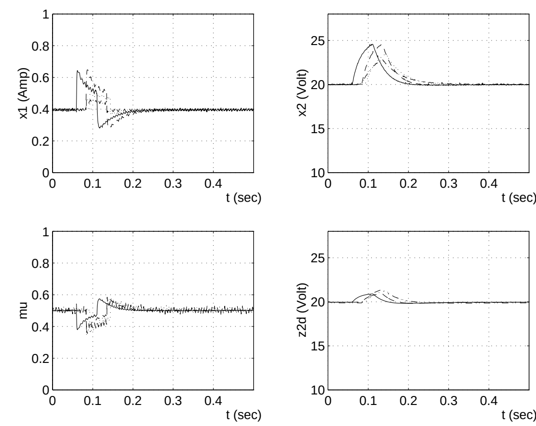

Figure 5: Response to a $w(t) \in \mathcal{L}_2$ for PBC

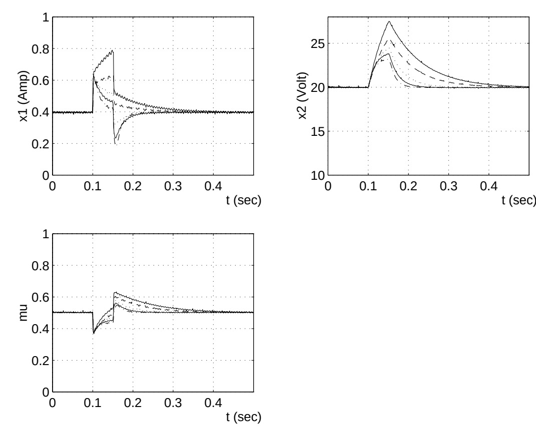

Figure 6: Response to a $w(t) \in \mathcal{L}_2$ for FLC

Signal $w(t)$ is applied after the output has arrived to the equilibrium point. Hence, in order to compute the $\mathcal{L}_2$ norm of the output, we only have to consider the error signal $\tilde{x}_2(t)$ caused by the disturbance.

From these experiments we obtain the following gains

$$\gamma_{PBC} = \frac{\|\tilde{x}_2\|_2}{\|w\|_2} = 0.6506 \rightarrow 0.9910 \quad \text{PB Controller}$$

$$\gamma_{FLC} = \frac{\|\tilde{x}_2\|_2}{\|w\|_2} = 0.6855 \rightarrow 1.9787 \quad \text{FL Controller}$$

where, in PBC, the smallest value corresponds to a large R_1, and in FLC, the smallest value of γ corresponds to a large a_1 and an small a_2.

On the other hand, substituting the parameters values given in table 1 into the equations for the lower bounds given in Proposition 2.2, we obtain for the PB and FL controllers, respectively,

$$||T_{w\tilde{z}_2}||_{i2} > \frac{E_0}{2E} = 1$$
$$||T_{w\tilde{z}_2}||_{i2} > \frac{L^3E_0^3}{E^3CR^2 + 2LEE_0^2} = 0.0035$$

As we can see the values obtained for the FL controllers fit within these bounds. On the other hand, the values for the PB controllers do not fit, hence the applied disturbance may be not large enough in an $\mathcal{L}_2$ sense.

The upper bound for the disturbance attenuation for the case of PBC is obtained as the solution of an optimization problem. This yields $\tilde{\gamma} = 36.8356$, $R_1 = 0.4269$, $K_1 = 100.1178$, and $K_2 = 0.0685$.

In Fig. 7 we present a 3D graphic for $\tilde{\gamma}^2$ as a function of parameters K_3 and K_4 that fits the conditions given in the optimization problem for the set of parameters in the real circuit. In this graphic the optimal point is marked with a star and is located at the perimeter.

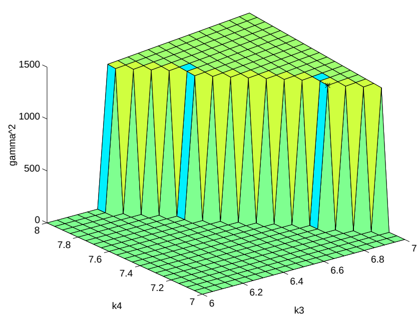

Figure 7: Function $\tilde{\gamma}(K_3, K_4)$ and optimal point $(*)$

5 Concluding Remarks

The long term motivation of the present study is to provide a framework to compare, from a perspective different from stabilization, various existing controllers proposed for Euler-Lagrange systems. A similar research, albeit specialized to robots with flexible joints, was reported in [4], where the comparison was based on continuity properties and adaptivity. As alternative performance indicator we propose here to adopt the robustness to external disturbances, which is measured via the $\mathcal{L}_2$ norm of the corresponding closed loop operator.

Current research is under way to extend this study to other systems and controllers. In particular, we are interested in carrying out the analysis for backstepping-based controllers and induction motors.

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