

CHAPTER 6

Predictor-based current limitation method

6.1 Overview

Different exoskeletons and control methods have been designed in Chapter 4 and Chapter 5. The use of all the presented devices involves human-robot interaction. On the one hand, exoskeletons used for passive rehabilitation follow predefined trajectories regardless the interaction force with the subject (although safety mechanisms can be added to stop motion if force is unexpectedly high). On the other hand, exoskeletons and control architectures used for active rehabilitation try to maintain or control certain interaction forces between the human and the device. Exoskeletons which are designed for active assistance normally include force or torque sensors. However, exoskeletons for passive rehabilitation do not necessarily incorporate these kind of sensors. Additionally, these sensors could unexpectedly damage during operation or move from their original positions, generating misleading readings of the interaction force.

When using exoskeletons, subjects apply forces over them, and the response of the controllers is different depending on the application. In active rehabilitation therapies or active assistance, the controller should be able to modify the actuator references in a manner in which the exoskeleton adapts to human intention of motion, maintaining the interaction force under some predefined value (force control, Section 5.3) or generating a certain speed depending on the interaction force (impedance control, Section 5.2). However, in passive rehabilitation, if the user accidentally or voluntarily resists to perform the trajectory imposed by the exoskeleton, the controller will try to overcome the applied disturbance applying more torque, which implies an increase in the current of the actuator. Actuators could be damaged in situations like this, depending on the maximum torque they can apply and the magnitude of the external disturbance. This situation can also arise in active rehabilitation if there is a failure in the force/torque sensor or the force applied by the user is too high. It is for the aforementioned reasons that it is necessary to provide a safety mechanism to protect the actuators from incurring in overcurrent situations which, when prolonged in time, cause overheating of the actuator, generating irreversible damage to it. The actuators of most of the exoskeletons developed in this work are DC motors, so a system for protecting these motors from potentially too high interaction forces was designed.

Through the past decades, different mechanisms for current limitation have been developed, but it still remains a topic of interest, as research conducted in the last few years shows [191]–[195]. Current limitation methods can be classified according to different criteria, such as the implementation method and the influence they have in normal motor operation [196]. According to the implementation, hardware and software methods can be used. Hardware methods implement logic electronics to control the commutation of the switching semiconductors of motor driving circuits. In contrast, software methods modify the voltage command to be sent to the motor before it is transferred to the driving circuit. According to the influence in normal operation of the system, interventionist ([196]–[200]) and regulating ([201], [202]) methods can be distinguished. Interventionist methods only act when the motor current goes beyond a certain threshold. Under safe circumstances, the system works as if no protection system had been added to it. In contrast, regulating methods are just current controllers, which normally are the inner loop of a speed controller. The main drawback of this type of method is that two control loops have

to be adjusted, the speed and the current ones.

The ideal overcurrent protection system would allow motor normal operation under safe conditions, i.e., when the external forces do not generate overcurrent, but would generate some kind of limitation to the voltage applied to the motor when the external force is too high. In this way, the actuator would be capable of applying the maximum torque guaranteed by the manufacturer, but in safe conditions, ensuring that the motor is not overheated. Additionally, easiness of implementation, scalability, modularity and cost are factors which should also be taken into account to design an efficient, robust and adaptable system. For these reasons, a software interventionist current limitation method was designed. The method is based on a predictor which uses the well-known continuous model of the DC motor system. Voltages applied to the motor are dynamically saturated according to the current motor state and its expected evolution during the following control period. In this way, if the motor controller computes a voltage that would generate overcurrent if applied to the motor, the algorithm saturates that voltage to the maximum allowable one which maintains the motor under safe conditions. An improvement of the algorithm that allows the motor current tripping for a configurable period of time was also developed. The proposed method is independent of the motor controller, as it is just added as a block after the controller with the objective to saturate the control action. A general block diagram which illustrates this for a position controller is shown in Fig. 6.1.

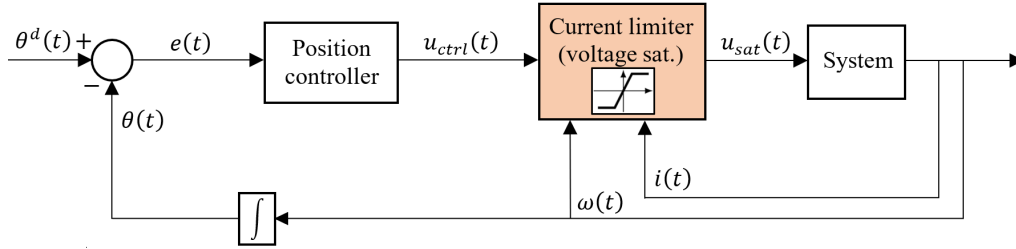


Figure 6.1: Generic block diagram of a position control for a system based on a DC motor. The current limitation algorithm presented in this work can be understood as a saturation block placed just after whatever position controller. The controller output u_{ctrl} can be saturated by the current limitation algorithm.

This work was published in the author’s publication “*Predictor-Based Current Limitation Method for a DC Motor-Actuated Upper-Limb Rehabilitation Exoskeleton*”, IEEE Access, vol. 10, pp. 81684 - 81698, 2022. doi: 10.1109/ACCESS.2022.3194911.

6.2 Algorithm design and analysis

This section presents the analytical design of the proposed algorithm and the evaluation of its performance in simulation. The motor selected for the design and evaluation is the Maxon DCX22S GB KL 48V with a gearbox of 794 to 1 reduction ratio, which is one of the actuators used in the exoskeletons discussed in this thesis. Most motor parameters were obtained from the datasheet. Nevertheless, the motor electrical resistance R and the electric constant k_e were experimentally measured according to the standard procedure in [203] since the precision in both parameters

has an important impact on the algorithm performance. This will be analyzed in depth further on. The motor parameters are gathered in Table 6.1.

Table 6.1: Selected motor model parameters.

Sym.	Description	Value
R	Resistance	18 Ω
L	Inductance	0.881 mH
k_e	Electric constant	0.0359 V/(rad/s)
k_p	Torque constant	0.0359 N/A
J	Moment of inertia	0.2941 kg $\cdot$ m ²
f	Friction coefficient	0.6299 Nm/(rad/s)
n	Reduction ratio	794
i_{sat}	Maximum current	0.4 A

6.2.1 Analytical formulation

The proposed algorithm computes the voltages which, if applied to the motor, generate maximum current in both senses. Using the motor dynamic model, it is possible to predict the current in the motor given a certain state and an input voltage. If the value of the predicted current is set to the maximum allowable current, the voltage which generates that current can be computed. To develop the algorithm it is necessary to first present the DC motor model, which is defined by (6.1) and (6.2):

$$u(t) = Ri(t) + L \frac{di(t)}{dt} + nk_e \omega(t) \quad (6.1)$$

$$nk_p i(t) = f\omega(t) + J\dot{\omega}(t) \quad (6.2)$$

where $u(t)$ is the voltage applied to the motor, $i(t)$ is the motor current, $\omega(t) = \omega_m(t)/n$ is the angular speed of the motor after the gearbox and $\dot{\omega}(t)$ is the angular acceleration also after the gearbox.

The algorithm is executed discretely, as it is implemented over the microcontroller of the system (see Section 3.4). However, the use of a discrete model for the motor is not an option due to the fact that the electrical time constant for the DC motor $\tau_e = L/R$ is around 50 μs , and the control algorithm is executed at 1 kHz, that is to say, every $T_{\text{ctrl}} = 1$ ms. For this reason, the algorithm has to make use of the continuous model of the system.

The algorithm is based on projecting in time how the motor current will evolve for a certain state and applied voltage. First, the time period for which the prediction is computed must be defined. This period of time was called “prediction horizon”, and is noted as t_{ph} . The prediction horizon should be chosen large enough to ensure the motor current has achieved the steady state. The prediction horizon was selected as $t_{\text{ph}} = 5L/R$, which guarantees that the current will have achieved a 99.33% of the steady-state value a time t_{ph} after the prediction is made. The qualitative temporal diagram of the algorithm execution is shown in Fig. 6.2. Each control period T_{ctrl} , the voltages which would generate the saturation current in both senses are computed.

The time-domain expression for the evolution of the motor current has to be obtained. For this purpose, the Laplace transform is applied to (6.1). Considering

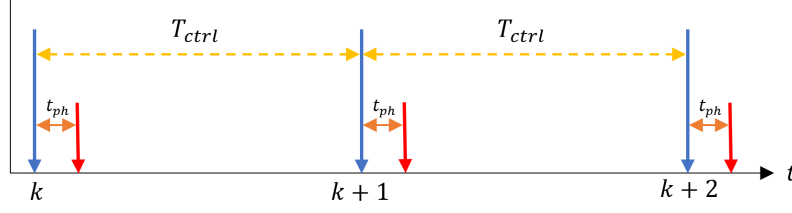


Figure 6.2: Temporal diagram of the current limitation algorithm. Source: Author's publication ©David Pont Esteban et al. 2022; published by IEEE (<http://doi.org/10.1109/ACCESS.2022.3194911>)

the system has the voltage input $u(t)$ and a virtual input $\omega(t)$, it is possible to apply the superposition principle. Although $\omega(t)$ is a state variable, it was considered as an input for the design of the predictor given that it is possible to measure it.

Every control period, the generic motor controller computes the control signal $u(t)$ which, once applied, it remains constant until the next control period. From the predictor's point of view, the voltage is a constant step of amplitude U during t_{ph} . As for the virtual input $\omega(t)$, it is not likely to be constant during t_{ph} since voltage variation will generate speed variation, although that variation will be small during such a short period of time. In this context, the speed was modelled as a step signal with an average speed amplitude, ω_{avg} , during t_{ph} . The estimation of that average speed for prediction at period k is computed considering a uniformly accelerated movement during t_{ph} . The motion acceleration is estimated using the speed associated to the current and previous control periods:

$$\hat{\omega}_{avg}[k] = \frac{1}{2} \left(2\omega[k] + \frac{\omega[k] - \omega[k-1]}{T_{ctrl}} t_{ph} \right) \quad (6.3)$$

Apart from the variation in speed caused by voltage variation applied to the motor, there exists also the possibility that the mechanical load over the motor varies during the prediction horizon. Nevertheless, the dynamics of the electrical part of the system are much faster than the disturbances that can be produced due to load variations. Moreover, if a very brusque disturbance were applied over the motor just after the prediction in period k were made, the method would compensate for it in the next period. The influence over the current limitation performance that the error in the estimation $\hat{\omega}_{avg}[k]$ has is analyzed in more detail in Section 6.2.5. Under the above considerations made for both inputs, separating (6.1) for both inputs and solving both differential equations, expressions (6.4) and (6.5) are obtained:

$$i_u(t) = \frac{U}{R} \left(1 - e^{-\frac{R}{L}t} \right) + C_u e^{-\frac{R}{L}t} \quad (6.4)$$

$$i_\omega(t) = -\frac{\omega_{avg} n k_e}{R} \left(1 - e^{-\frac{R}{L}t} \right) + C_\omega e^{-\frac{R}{L}t} \quad (6.5)$$

being C_u and C_ω constants which depend on the initial conditions. Applying the superposition principle to (6.4) and (6.5)

$$i(t) = i_u(t) + i_\omega(t) = \frac{U - \omega_{avg} n k_e}{R} \left(1 - e^{-\frac{R}{L}t} \right) + C e^{-\frac{R}{L}t} \quad (6.6)$$

with $C = C_u + C_\omega$. To determine the value of C , initial conditions have to be applied to the system. The initial current for each prediction is $i(0) = i_0$. Evaluating (6.1) at

$t = 0$, it can easily be derived that $i'(0) = (U - Ri_0 - \omega_{\text{avg}}k_e)/L$. Substituting the initial conditions in 6.6, the time-domain expression for the motor current evolution is obtained:

$$i(t) = \frac{U - \omega_{\text{avg}}nk_e}{R} \left(1 - e^{-\frac{R}{L}t}\right) + i_0 e^{-\frac{R}{L}t} \quad (6.7)$$

Solving (6.7) for U , evaluating the resulting expression for i_{sat} and $-i_{\text{sat}}$ saturation currents and $t = t_{\text{ph}}$, and using $\hat{\omega}_{\text{avg}}$ as the estimate of ω_{avg} , the voltage saturation values are obtained:

$$\begin{cases} u_1[k] = \frac{R(i_{\text{sat}} - i_0 e^{-(R/L)t_{\text{ph}}})}{1 - e^{-(R/L)t_{\text{ph}}}} + \hat{\omega}_{\text{avg}}[k]nk_e \\ u_2[k] = \frac{-R(i_{\text{sat}} - i_0 e^{-(R/L)t_{\text{ph}}})}{1 - e^{-(R/L)t_{\text{ph}}}} + \hat{\omega}_{\text{avg}}[k]nk_e \\ u_a[k] = \max(u_1[k], u_2[k]) \\ u_b[k] = \min(u_1[k], u_2[k]) \end{cases} \quad (6.8)$$

The time scale is relative to each computation of the saturation values. Time is considered to be $t = 0$ at the beginning of each algorithm execution. Therefore, i_0 corresponds to the measured current at the beginning of each control period.

The algorithm was first validated through simulation using Matlab. Due to the fast dynamics of the motor current, the simulation step was set to $1 \mu\text{s}$. The algorithm was executed periodically every $T_{\text{ctrl}} = 1 \text{ ms}$ and the prediction horizon was set to $t_{\text{ph}} = 5L/R = 245 \mu\text{s}$. The voltage input for the simulation was a square wave of 3.33 Hz, a 50% duty cycle and amplitude of $\pm 24\text{V}$. Two different overcurrent situations were analyzed with the simulation. In the first one, the motor was stalled, no matter how high the voltage applied to it was, so the overcurrent was maintained in time. In the second one, the rotor was freed and it could spin normally, but here, the brusque voltage variations generated current peaks. No load was added to the motor shaft in this situation. The duration of the simulation was set to 1 s, where the motor was stalled the first 0.5 seconds and was subsequently freed for the last 0.5 seconds.

Figure 6.3 presents the voltage applied to the motor in the conducted simulation, along with the evolution of the speed and current, with and without the application of the designed limitation algorithm. The stall situation (from $t = 0 \text{ s}$ to $t = 0.5 \text{ s}$) is first analyzed. When the algorithm is not applied, the current values in the motor alternate between $\pm 1.33 \text{ A}$, which corresponds to what would be expected from (6.1) for the stopped motor, given a $\pm 24\text{V}$ input and a motor resistance of 18Ω . If maintained in time, these current values would end up overheating and damaging the motor. When the motor can rotate freely (from $t = 0 \text{ s}$ to $t = 0.5 \text{ s}$), the steady state current is around a few miliamperes. However, current peaks of a magnitude around 2.57 A are produced when the voltage polarity is reversed. In this situation, although there is no overload over the motor shaft, the term associated to speed in (6.1) represents a large induced voltage, and the current has to substantially increase in absolute value to change the motor rotation direction. Introducing the predictor-based current limitation algorithm in the simulation, the current in the motor never surpasses $i_{\text{sat}} = 0.4 \text{ A}$. As observed in the voltage plot, the applied voltage is saturated by the algorithm such that, in conditions where the predefined $\pm 24 \text{ V}$ input voltage would generate overcurrent, the voltage is saturated to the one that generates a current i_{sat} in the motor. The behavior of the algorithm is the appropriate one in both stall and free rotor movement, chopping the current peaks

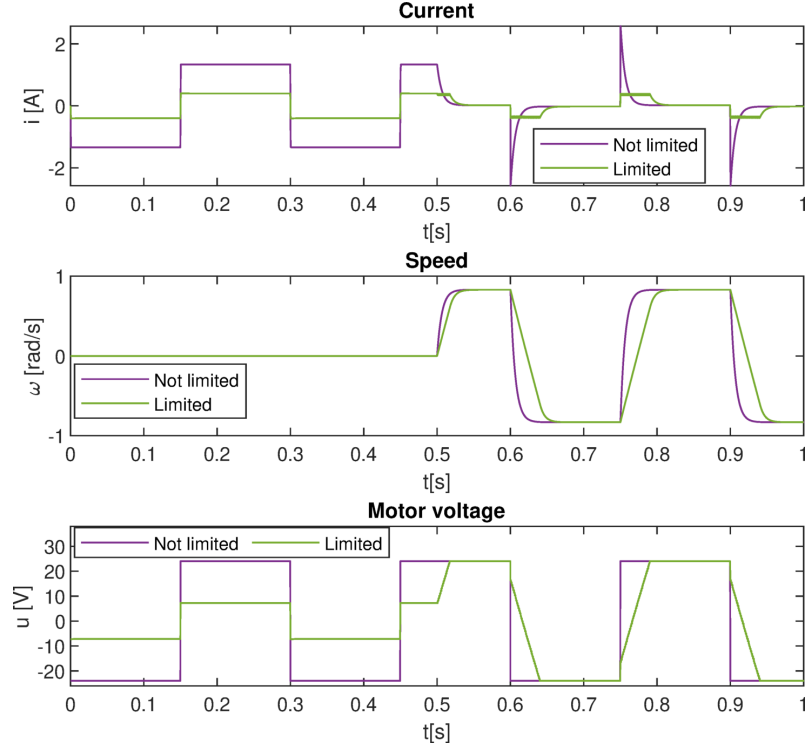


Figure 6.3: Simulation of current, speed and input voltage evolution without current limitation and with the model-based limiter. Motor is stalled until $t = 0.5$ s. From then on, the rotor can move freely.

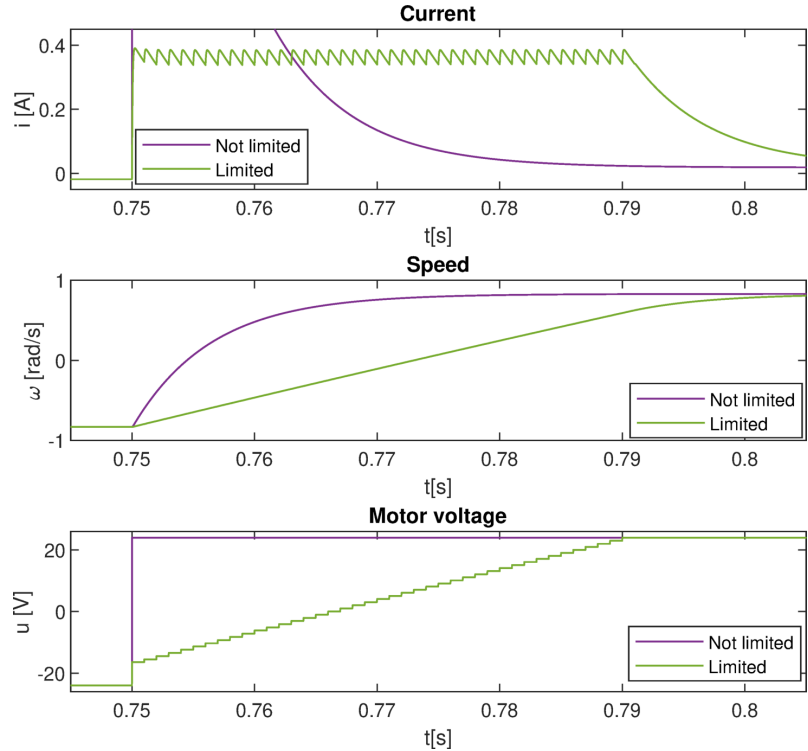


Figure 6.4: Zoom-in of the free rotor situation.

in the latter case. Nevertheless, the short-duration current peaks are not necessarily dangerous for the motor for many applications and, as observed in Fig. 6.3, the

suppression of these peaks slows down the speed evolution of the motor. This will be addressed in the following subsection.

A zoom-in of the simulation for the free rotor situation in Fig. 6.3 is presented in Fig. 6.4. A ripple of period 1 ms (which is the control period) in the current can be observed. Additionally, the current never reaches exactly the value set for i_{sat} of 0.4 A. As observed in the voltage plot, the current limitation algorithm allows the voltage applied to the motor to be slightly increased every control period. That voltage is computed as the voltage that would generate i_{sat} at time t_{ph} . That current is not exactly achieved at t_{ph} because the assumptions made for estimating $\hat{\omega}_{\text{avg}}$ are not perfect. The slow decrease in the current every T_{ctrl} after a new voltage is applied is due to the acceleration of the motor. When increasing positive voltages are applied to the motor in the free rotor condition, the motor speed increases. According to (6.1), for a constant applied voltage, the faster the motor rotates, the lower the current is.

6.2.2 Response acceleration

The current limitation algorithm designed in Section 6.2.1 proved to provide in simulation an effective method for limiting the current under the situations of stall rotor and application of brusque voltage variations in free rotor condition. Nevertheless, as previously mentioned, the suppression of the short-time current peaks when motor polarity is reversed is not necessarily harmful for the motor, as long as the peaks are not produced periodically, with a high frequency and for a long period of time,

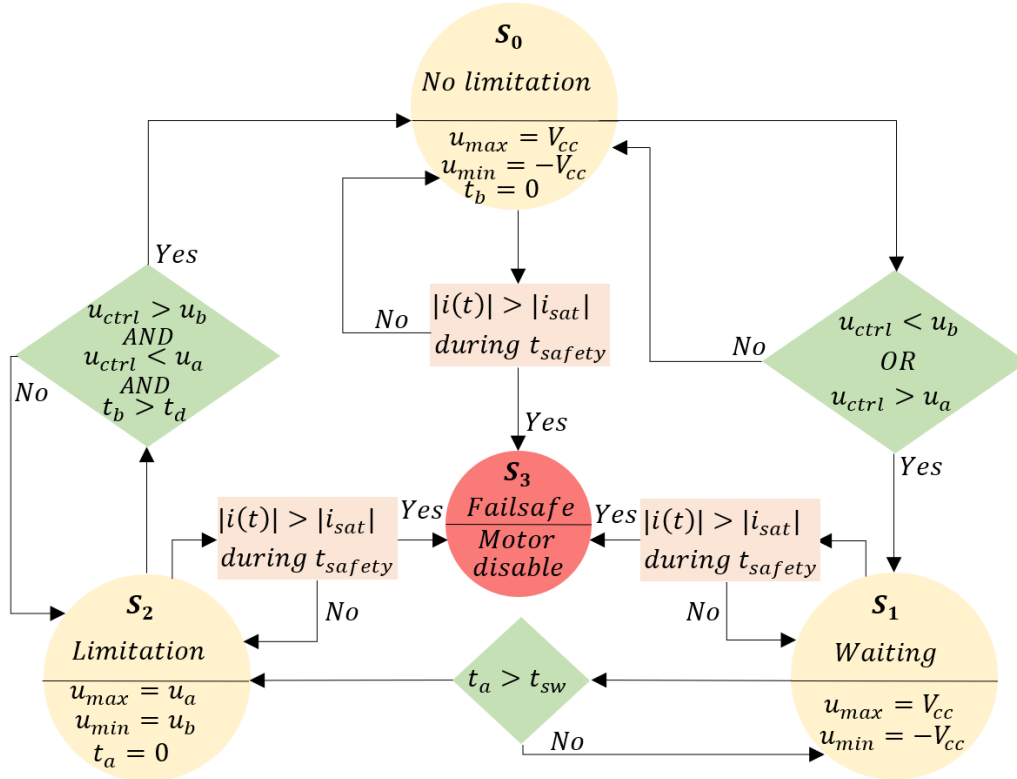


Figure 6.5: Diagram of the proposed state machine, which allows current peaks of a configurable duration t_{sw} . Source: Author's publication ©David Pont Esteban et al. 2022; published by IEEE ([http://doi.org/10.1109/ ACCESS.2022.3194911](http://doi.org/10.1109/ACCESS.2022.3194911))

which is something assumable in many applications. A short-duration current peak in those situations does not generate a temperature increase in the motor. However, if they are repeated frequently enough and for a long enough period of time, overcurrent will generate a temperature increase which will damage the motor. In this context, a state machine that determines if the current should be limited or not was proposed, and is shown in Fig. 6.5. The objective of this state machine is to allow short-current peaks to be produced. This will allow the motor speed to evolve faster when voltage is abruptly varied, but will still prevent overcurrent to be held in time.

The state machine is constituted by four states. In all of the states, the predictor is executed, providing the maximum and minimum allowable voltages to be applied to the motor. In state S_0 , the current is not limited, the saturation voltages are $\pm V_{cc}$. As soon as the voltage which the motor controller determines to be applied is beyond the voltages u_a and u_b , the state machine switches to S_1 . In this state, the voltages applicable to the motor are not saturated yet by the predictor's computations but, if after a configurable time t_{sw} the voltage to be applied to the motor is still beyond the overcurrent thresholds, a state switch to S_2 is produced and the voltages are saturated according to the predictor. If the voltage computed by the controller is within the voltage range allowed by the predictor and a time $t_b > t_d$ has elapsed, then the state machine switches back to S_0 , allowing a new current peak again. The condition of a configurable minimum time t_d before allowing a new current peak was added to limit the frequency at which those peaks can be produced, preventing overheating due to many too frequent current peaks.

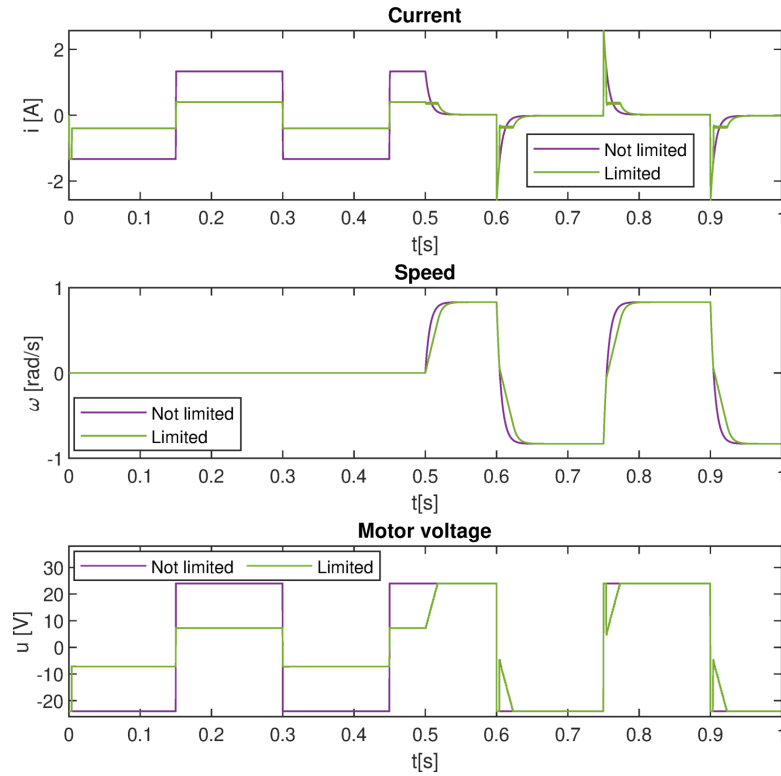


Figure 6.6: Simulation of current, speed and input voltage evolution without current limitation and with the model-based limiter. The state machine described in Fig. 6.5 has been incorporated. Motor is stalled until $t = 0.5$ s. From then on, the rotor can move freely.

The same simulation presented in Fig. 6.3 is shown in Fig. 6.6 but, in this case, the state machine was incorporated. For this simulation, t_{sw} was set to 4 ms and T_{ctrl} was maintained to 1 ms. In the latter case, short-duration current peaks are allowed, and when $t_a > t_{sw}$, the voltage applied to the motor is saturated. It is observed in the current plot that for the free rotor condition, the current evolution for limited and unlimited simulations is the same for the first 4 ms.

Figure 6.7 presents the speeds for the limitation algorithm with and without the inclusion of the state machine. It is observed that the state machine provides a faster response, by allowing the 4 ms current peaks. The selection of the temporal parameters t_{sw} and t_d should be determined according to the motor and the application, considering parameters as the control period, the thermal properties of the motor and the motor controller.

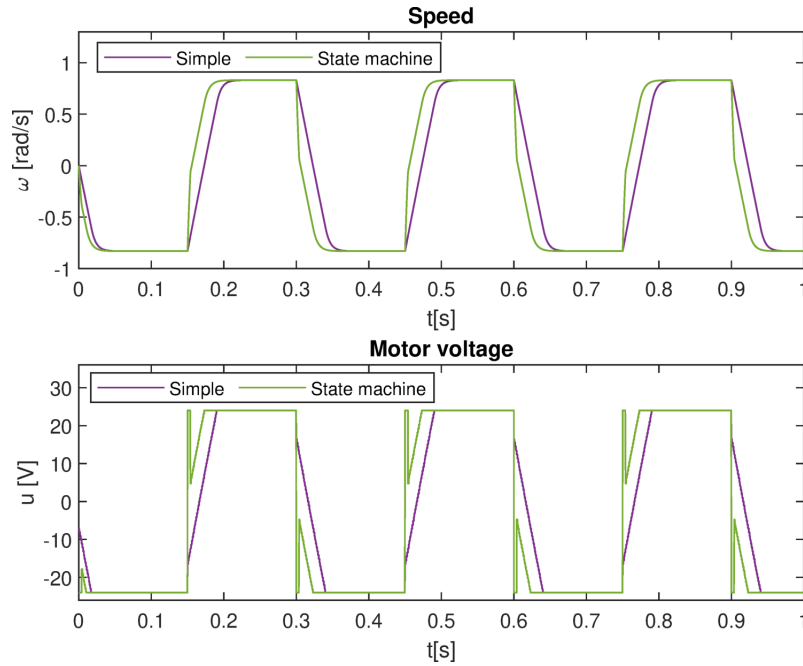


Figure 6.7: Comparison of speed and voltage evaluations with the simple limitation method (without the state machine), and with the inclusion of the state machine described in Fig. 6.5.

6.2.3 Implementation over H-bridge driver

In real applications, DC motors are usually driven by H-bridge drivers. These drivers can apply three different voltages to the motor depending on the control switches: $\pm V_{cc}$ and 0 V. In previous simulations, the effect of employing an H-bridge to drive the motor was not included and it is examined in this subsection. The switching frequency for the PWM in the simulation was set to $f_{PWM} = 40$ kHz, since it is the frequency at which the PWM modules of the designed electronics platform work (see Section 3.4).

The same simulation as in the previous subsections was performed to study the influence of the H-bridge inclusion. Figure 6.8 shows the simulation results with and without H-bridge, applying in both cases the predictor-based current limitation algorithm with the state machine that allows short-duration current peaks. Adding

the H-bridge, the current ripple increases, being 0.11 A in motor free movement and 0.07 A in stall. In addition to the ripple generated by the variation of $\omega(t)$ that was previously discussed, here the fast voltage variation due to the H-bridge nature increases the current ripple, as expected. The average absolute value of the current in the period between 0.3 s and 0.45 s is around 425 mA, which is above $i_{\text{sat}} = 0.4$ A. This increase in the average value of the current in stall conditions can be critical in some applications. The offset over the current threshold can be overcome by slightly decreasing the value of i_{sat} in the algorithm.

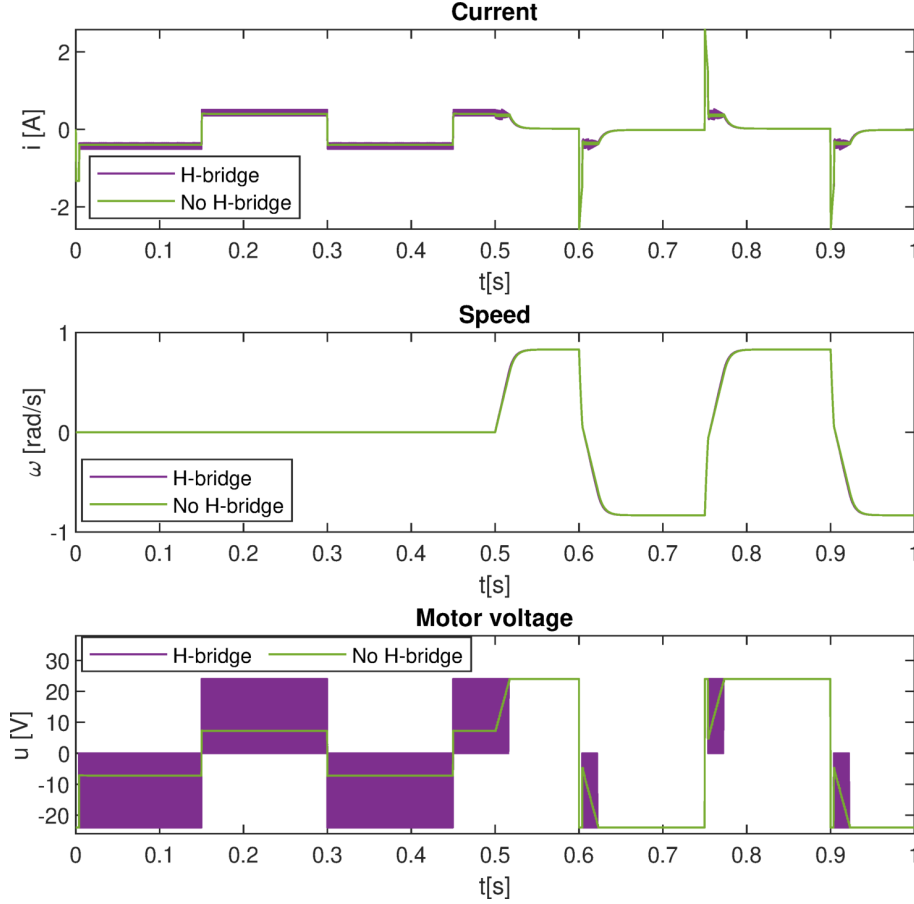


Figure 6.8: Simulation of current, speed and input voltage evolution with and without the inclusion of an H-bridge. The state machine described in Fig. 6.5 has been incorporated. Motor is stalled until $t = 0.5$ s. From then on, the rotor can move freely.

6.2.4 Influence of motor parameter estimation errors

In the mathematical development presented in Section 6.2.1, uncertainty in parameters' values was not considered. The designed algorithm is based in the model of the system, so studying the influence of the error in parameters estimation is crucial to evaluate the performance of the current limitation method.

The nominal values of the motor parameters can be found in Table 6.1. The real values of the parameters involved in this study are the nominal ones plus some value contained in the tolerance band as follows: $k_e = k_{en} \pm 5\%$, $L = L_n \pm 30\%$ and $R = R_n \pm 5\%$. Tolerances for R and k_e were considered of only 5% since they were

accurately experimentally measured parameters. The inductance of the motor was extracted from the datasheet and a conservative value of 30% was considered as its tolerance.

The current error at $t = t_{ph}$ between the nominal expected current and the obtained current due to error in parameter estimations when applying the model-based limitation algorithm is given by:

$$e(t_{ph}) = |i(t_{ph}) - i_n(t_{ph})| \quad (6.9)$$

As imposed by definition in the limitation algorithm in (6.8), the expected current at $t = t_{ph}$ is nominally $i_n(t_{ph}) = i_{sat}$. On the other hand, the prediction horizon was set to $t_{ph} = 5L_n/R_n$. Substituting the expression of u_1 of (6.8) in (6.7) and substituting the resulting expression in (6.9), expression (6.10) is obtained:

$$e\left(\frac{5L_n}{R_n}\right) = \left| \frac{R_n \left(1 - e^{-\frac{5RL_n}{LR_n}}\right)}{R(1 - e^{-5})} (i_{sat} - i_0 e^{-5}) + \frac{\omega_{avg}}{R} n (k_{en} - k_e) \left(1 - e^{-\frac{5RL_n}{LR_n}}\right) + i_0 \left(e^{-\frac{5RL_n}{LR_n}}\right) - i_{sat} \right| \quad (6.10)$$

This expression can be simplified considering $i_{sat} \gg i_0 e^{-5}$ and $i_0 e^{-\frac{5RL_n}{LR_n}} \approx 0$, in the following way:

$$e\left(\frac{5L_n}{R_n}\right) = \left| \frac{R_n \left(1 - e^{-\frac{5RL_n}{LR_n}}\right)}{R(1 - e^{-5})} i_{sat} + \frac{\omega_{avg}}{R} n (k_{en} - k_e) \left(1 - e^{-\frac{5RL_n}{LR_n}}\right) - i_{sat} \right| \quad (6.11)$$

Exploring (6.11), it is immediate to determine that the error is maximized if $\omega_{avg} = \omega_{max}$, $k_e = k_e|_{min}$ and $L = L|_{min}$. The value of ω_{max} corresponds to the maximum speed achievable being the motor current i_{sat} . To obtain this value, i_{sat} is substituted in (6.1) and the obtained expression is solved for ω_{max} :

$$\omega_{max} = \frac{u_{max} - Ri_{sat} - L|_{min} \frac{di(t)}{dt}}{nk_e|_{min}} \quad (6.12)$$

The situation that it is assumed here is that the motor current is i_{sat} for a certain period of time. In this case, $L \frac{di(t)}{dt} = 0$ and (6.12) simplifies to (6.13):

$$\omega_{max} = \frac{u_{max} - Ri_{sat}}{nk_e|_{min}} \quad (6.13)$$

Conversely, the value with which R contributes to the maximization of (6.11) is not as easily obtainable as the values of the other parameters. To obtain the maximum error value, the aforementioned values were substituted in (6.11), and the error function was plotted doing a sweep in the full range of $R = R_n \pm 5\%$, increasing the resistance 0.0036Ω in every step. The value for u_{max} was set to 24 V as it is the voltage to be used in the application. As a result, a maximum error of $e(t_{ph})|_{max} = 72 \text{ mA}$ was obtained. As aforementioned, this error is obtained at maximum speed conditions. A 66.67% of this error is associated with the term which

depends on the estimation of k_e . If the estimation of k_e were perfect or the motor were stopped ($\omega = 0$), the current error associated to the error in the estimation of R and L would be the 33.33% of $e(t_{ph})$, which is around 24 mA. Although the theoretical maximum error due to error in motor parameters estimations is produced at maximum speed, a critical current limitation situation will normally take place in circumstances where an unexpected load is applied to the motor, what will cause the speed to dramatically decrease and, therefore, the error in current limitation will also be lower.

6.2.5 Influence of speed estimation error

For the algorithm design, an average speed during the prediction horizon was exposed as a reasonable assumption. The computation of $\hat{\omega}_{avg}[k]$ was presented in (6.3). The current error generated by the estimation error of $\omega_{avg}[k]$ is addressed.

The currents generated by the limit voltages computed by the algorithm can be obtained substituting in (6.7) the expressions of those voltages presented in (6.8), and are the following ones:

$$i(u_1) = i_{sat} + \frac{nk_e \left(1 - e^{-\frac{R}{L}t}\right)}{R} (\hat{\omega}_{avg} - \omega_{avg}) \quad (6.14)$$

$$i(u_2) = -i_{sat} + \frac{nk_e \left(1 - e^{-\frac{R}{L}t}\right)}{R} (\hat{\omega}_{avg} - \omega_{avg}) \quad (6.15)$$

If the speed estimation were perfect, the currents in (6.14) and (6.15) would be i_{sat} and $-i_{sat}$ respectively, as the voltages applied are the limit ones determined by the algorithm. The errors in both currents at time $t = t_{ph}$ ($\tilde{i}^{u_1}$, $\tilde{i}^{u_2}$) can be computed as:

$$\tilde{i}^{u_1}(t_{ph}) = i(u_1) - i_{sat} \approx \frac{nk_e}{R} (\hat{\omega}_{avg} - \omega_{avg}) \quad (6.16)$$

$$\tilde{i}^{u_2}(t_{ph}) = i(u_2) + i_{sat} \approx \frac{nk_e}{R} (\hat{\omega}_{avg} - \omega_{avg}) \quad (6.17)$$

The errors obtained in (6.16) and (6.17) are equal. Exploring both expressions, it is deduced that the current errors will be lower if the term $\frac{nk_e}{R}$ decreases. Nevertheless, if the reduction ratio n is decreased, the rotor will be able to spin at faster speeds, so the error associated to the term $(\hat{\omega}_{avg} - \omega_{avg})$ would increase. The motor that was used in the development of this work had $\frac{nk_e}{R} = 1.584 \frac{V \cdot s}{\Omega \cdot rad}$.

The farther the speed evolution is from a uniformly accelerated movement during the prediction horizon, the larger the current error due to speed estimation error will be. In this context, the worst case scenario would be the application of a step perturbation over the rotor just after the current limitation algorithm were executed. The perturbation would modify the expected acceleration of the movement during t_{ph} , and a whole control period would still be remaining before another speed estimation were computed.

The disturbances that can be applied over an exoskeleton during a rehabilitation therapy are not very demanding in comparison to other applications which involve DC motors. It is necessary to consider that the subjects are in a recovery process, so both the magnitude and the frequency of applied disturbances will be low (a full

arm of a human male is considered to weight around 7 kg and activities of daily living for upper-limbs are in the range of 0.05 Hz to 0.5 Hz [14]).

A simulation was performed in order to evaluate the effect of the disturbances in speed estimation error and, therefore, in current estimation error. The disturbance was modelled as a square wave of an amplitude of 200 N and a frequency of 100 Hz. This disturbance is way more exigent than what could be expected in a rehabilitation context, but modelling this disturbance in an aggressive way can prove that the algorithm can be applied over systems over which the potential disturbances are actually this aggressive. All the simulation parameters were maintained the same as in previous simulations but, in this case, the motor could move freely during the whole simulation. The maximum speed estimation error that was observed was $\tilde{\omega}_{\text{avg}}|_{\text{max}} = 0.013$ rad/s, being the average estimation error $\tilde{\omega}_{\text{avg}} = 0.006$ rad/s. If those values are substituted in (6.16) and (6.17), the maximum and average current limitation errors only represent the 5.25% and 2.38% of i_{sat} , respectively.

6.2.6 Comparison to other methods

In this subsection, the proposed method is compared to two other software interventionist methods, presented in [199] and [200]. First, both algorithms are explained and afterwards, simulations of both and performance results are discussed. The voltage input and temporal parameters for these simulations are the same as the ones of the simulations conducted in Section 6.2.1.

The authors in [199] proposed a discrete approach in which, if the current in the previous period is higher than a certain threshold, the maximum voltage applicable to the motor is decreased with respect to the one applied in the previous control period. Otherwise, the voltage determined by the controller is applied to the motor. This logic is better explained by (6.18):

$$\begin{aligned} &\text{if } |i[k-1]| > 0.8i_{\text{sat}}, \\ &\text{then } u[k] = a \cdot u[k-1] \end{aligned} \tag{6.18}$$

On the other hand, authors in [200] proposed a method to limit the motors start-up current in which the voltage variation between consecutive control periods was limited. If the voltage computed by the motor controller belongs to the allowable range, that voltage is applied to the motor. Otherwise, the voltage associated to the maximum permissible variation with respect to the previous sample is applied. The behavior is described by (6.19):

$$\begin{aligned} &\text{if } u[k] - u[k-1] > \dot{V}_{\text{max}} \\ &\text{then } u[k] = u[k-1] + \dot{V}_{\text{max}} \\ &\text{else if } u[k-1] - u[k] > \dot{V}_{\text{max}} \\ &\text{then } u[k] = u[k-1] - \dot{V}_{\text{max}} \end{aligned} \tag{6.19}$$

The simulation results of the algorithm in [199] are shown in Fig. 6.9. In the original paper, the authors set $a = 0.98$. Nevertheless, it was observed that if the control period was set to $T_{\text{ctrl}} = 1$ ms, the algorithm was not able to properly limit the current. Through simulation, it was found that a much more appropriate value was $a = 0.85$, so this value was chosen for the simulation. It is important to note the impact that the execution frequency of the limitation algorithm has on the necessary

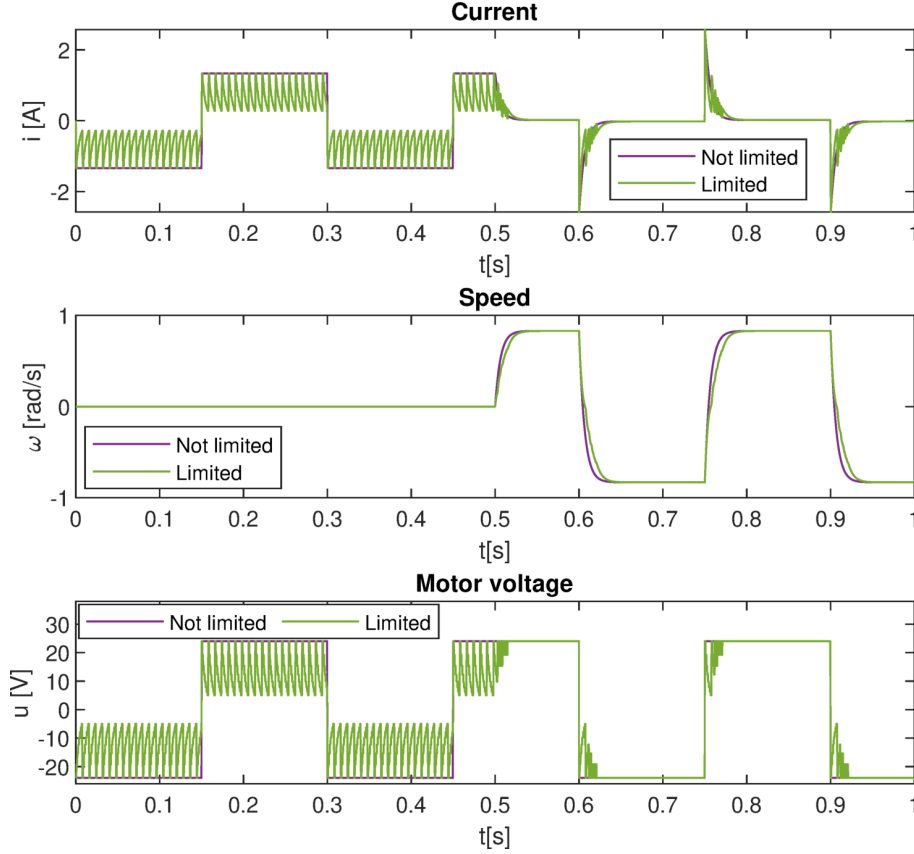


Figure 6.9: Simulation of current, speed and input voltage evolution without current limitation and with the limiter defined by (6.18). Motor is stalled until $t = 0.5$ s. From then on, the rotor can move freely.

value of the a parameter. This is an undesired issue by all means because it hinders the adaptability of the algorithm to different motors and applications. As observed in Fig. 6.9, when the motor is stalled, a remarkable current ripple is appreciated. This is due to the fact that, as soon as the current is below the threshold, the voltage applied to the motor is again the one of determined by the ± 24 V input, which generates the absolute value of the current to increase again, until the threshold is surpassed, and the cyclic behavior restarts. The average absolute value of the current in the stall condition is 0.700 A, widely surpassing the threshold of 0.4 A. The frequency of this ripple could be reduced by adding a minimum time during which the limitation must be maintained, although limitation circumstances had vanished. However, this would overfit the behavior of the algorithm to the stall situation, what would generate the motor not to work at i_{sat} for a reasonable period of time after limitation conditions had disappeared. Under conditions in which the motor can rotate freely, an effective current limitation is not achieved, since, as observed, the high current peaks are still present in the current signal. The rate at which the current is limited highly depends on a , so the obtained performance is highly dependent on this parameter.

For the simulation of the method in [200], two different values for the allowable voltage variation were tested: $\dot{V}_{\text{max}} = 1$ V and $\dot{V}_{\text{max}} = 0.5$ V. Results are shown in Fig. 6.10. The limitation obtained with this method is independent of the state of the motor, as neither the current nor the speed are considered by the algorithm. As the

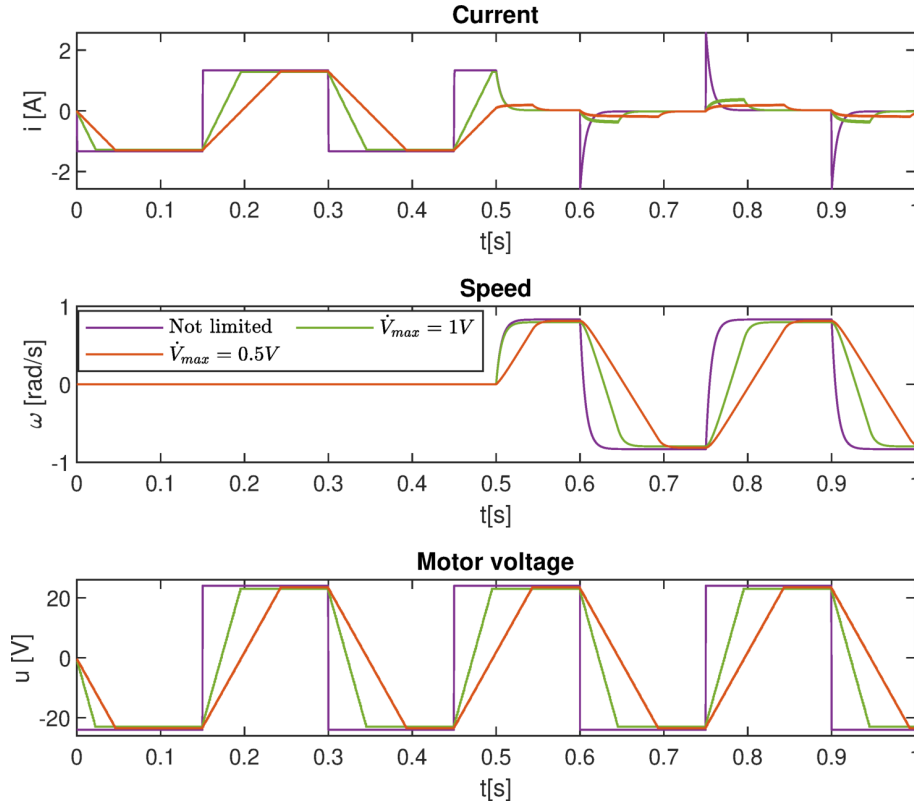


Figure 6.10: Simulation of current, speed and input voltage evolution without current limitation and with the limiter defined by (6.19). Motor is stalled until $t = 0.5$ s. From then on, the rotor can move freely.

authors proposed, this method can be used for limiting the inrush current in motor startup since, in that case, the overcurrent is caused by a too fast voltage variation, which is indeed what this algorithm directly limits. As observed in Fig. 6.10, current peaks are effectively suppressed in free rotor conditions. Nevertheless, in overload situations the current is not effectively limited. This algorithm delays the current increase, but with enough time, the current reaches the same value as without the application of the limitation method, as only the variation of voltage is limited, but the voltage magnitude is not limited at all. An aspect to consider is that the value of the maximum voltage variation $\dot{V}_{\max}$ is determined arbitrarily and will vary for different motor specifications.

Quantitative metrics of the performance of the proposed method and the methods in [199] and [200] are presented in Table 6.2 and Table 6.3. The metrics for the proposed method are the ones referred to the simple version (without the state machine that allows short-duration current peaks) shown in Fig. 6.3. Data in Table 6.2 present the total time during the simulation in which the current has surpassed i_{sat} , and, for that time, the average percentage of power and absolute value of current in relation to what the *perfect* saturation would be: $P_s = i_{\text{sat}}^2 R$ and $|i_s| = i_{\text{sat}}$. As data show, the time during which methods [199] and [200] exceed the current threshold are between 8 and 9 times greater than the time exceeded using the predictor-based method. In addition, even when the threshold is surpassed with the proposed method, power and current exceed very marginally the limits, always below 1%, while methods [199] and [200] widely surpass the saturation thresholds for current

Table 6.2: Performance metrics: current over i_{sat} .

Total time (0s to 1s)			
Method	Sat. time[s]	Power %	Current %
New	0.054	100.61	100.30
[199]	0.450	489.69	204.97
[200] (0.5V)	0.410	795.65	272.70
[200] (1V)	0.456	882.71	291.55
Stall (0s to 0.5s)			
Method	Sat. time[s]	Power %	Current %
New	0.054	100.61	100.30
[199]	0.401	459.64	201.60
[200] (0.5V)	0.410	795.65	272.70
[200] (1V)	0.448	890.99	293.29
Free (0.5s to 1s)			
Method	Sat. time[s]	Power %	Current %
New	0.000	100.00	100.00
[199]	0.049	735.04	232.42
[200] (0.5V)	0.000	100.00	100.00
[200] (1V)	0.008	398.35	189.28

Table 6.3: Performance metrics: current is saturated.

Total time (0 s to 1 s)			
Method	Sat. time [s]	Power %	Current %
New	0.641	96.34	98.06
[199]	0.504	317.67	161.73
[200] (0.5 V)	1.000	335.42	131.53
[200] (1 V)	1.000	414.18	149.94
Stall (0 s to 0.5 s)			
Method	Sat. time [s]	Power %	Current %
New	0.500	100	99.99
[199]	0.450	309.68	162.66
[200] (0.5 V)	0.500	658.4	232.47
[200] (1 V)	0.500	802.40	268.30
Free (0.5 s to 1 s)			
Method	Sat. time [s]	Power %	Current %
New	0.141	83.37	91.22
[199]	0.054	384.25	153.98
[200] (0.5 V)	0.500	12.43	30.59
[200] (1 V)	0.500	25.96	31.58

and power. Table 6.3 provides information on how close to optimal performance is the motor when current is limited with each algorithm. The time during which the current is saturated, as well as the average percentage of power and current for that time in relation with P_s and $|i_s|$, were computed for the four simulations. When

current is limited, the proposed method offers an average 98.06% of i_{sat} , what delivers an average 96.34% of P_s . Observing the metrics separated in stall and free rotor conditions in Table 6.2 and Table 6.3, it is noteworthy that the method in [200] can effectively suppress current peaks due to abrupt voltage variations. However, the current during saturation is very low, around 30% of i_{sat} , so the motor is safe, but it is very far from delivering the full power that it can do when working in the safe region. In the stall condition, methods [199] and [200] perform very poorly.

6.3 Experimental validation

The electronics platform used for the experimental validation of the predictor-based current limitation method is a simplification of the one presented in Section 3.4, with some particularities, as shown in Fig.6.11. The ALICE board was used to control a single DC motor in the conducted experiments. The Jetson Nano recorded the data transferred via SPI from the microcontroller. The unnecessary actuators and sensors from the global electronics platform (IMUs, ADSs, EMGs, force sensors, limit switches, stepper motors and additional DC motors) were removed.

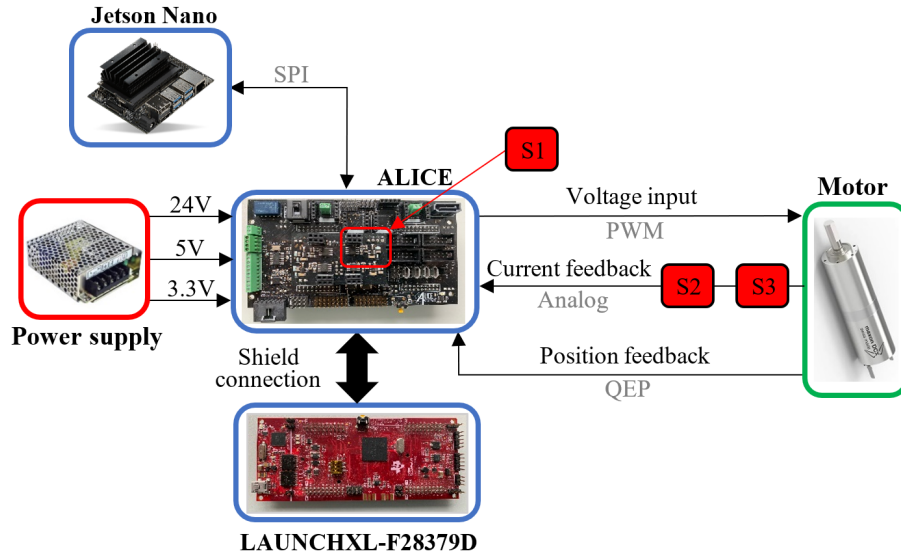


Figure 6.11: System electronics architecture for the validation of the current limitation algorithm.

The current sensors embedded in ALICE's supplementary board were designed to measure currents in the range of -0.5 A to 0.5 A, so two external current sensors (Sparkfun SEN-14544) were added to measure negative and positive currents of higher absolute value, as needed for these specific tests. The current ranges covered by each of the three current sensors employed in this setup are shown in Fig. 6.12. Sensor S2 covers the full positive range (0 A to 3.15 A) and S3 the full negative one (-3.15 A to 0 A). The maximum absolute value was set to 3.15 A since the motor peak current in startup is slightly below 3 A, and in this way a reasonable margin was conceded. The current sensors were calibrated using a precision current source. Different currents were applied to the sensors. For sensor S1, currents between -0.5 A and 0.5 A with steps of 0.1 A were applied. Currents in the range 0 A to 3 A and -3 A to 0 A with steps of 0.3 A were applied to sensors S2 and S3, respectively. For each

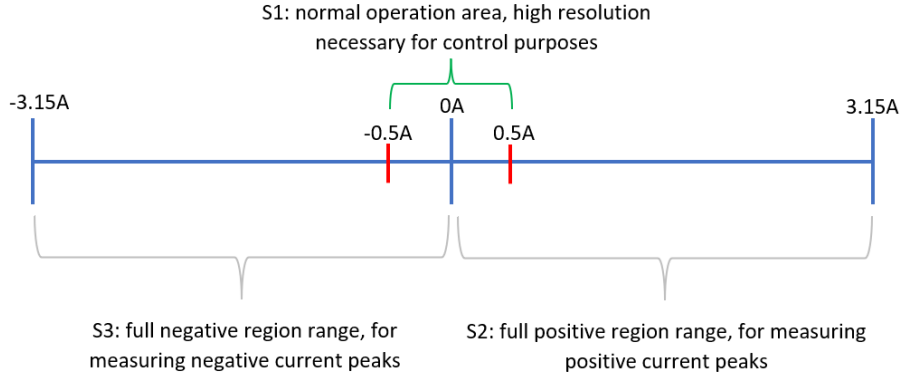


Figure 6.12: Current ranges measured by each of the three current sensors. Sensor S1 is specially designed to measure current in the safe operation range. Sensors S2 and S3 allow measuring current peaks. Source: Author's publication ©David Pont Esteban et al. 2022; published by IEEE (<http://doi.org/10.1109/ACCESS.2022.3194911>)

current, a thousand values were captured and first-order polynomial regressions were computed for those average currents, leading to first-order polynomials as calibration curves. The R^2 factors were the following: $R_{S1}^2 = 0.99995$, $R_{S2}^2 = 0.99998$ and $R_{S3}^2 = 0.99999$.

6.3.1 Stand-alone motor validation tests

These tests were similar to the simulations performed in Section 6.2. Overload (rotor stall) and free-rotor situations were evaluated. The experiments in both situations were performed independently due to physical setup requirements. Motor parameters have been previously shown in Table 6.1.

The free-rotor situation is first addressed. The duration of the experiment was 1 s, during which the applied voltage was a square wave of ± 24 V and a frequency

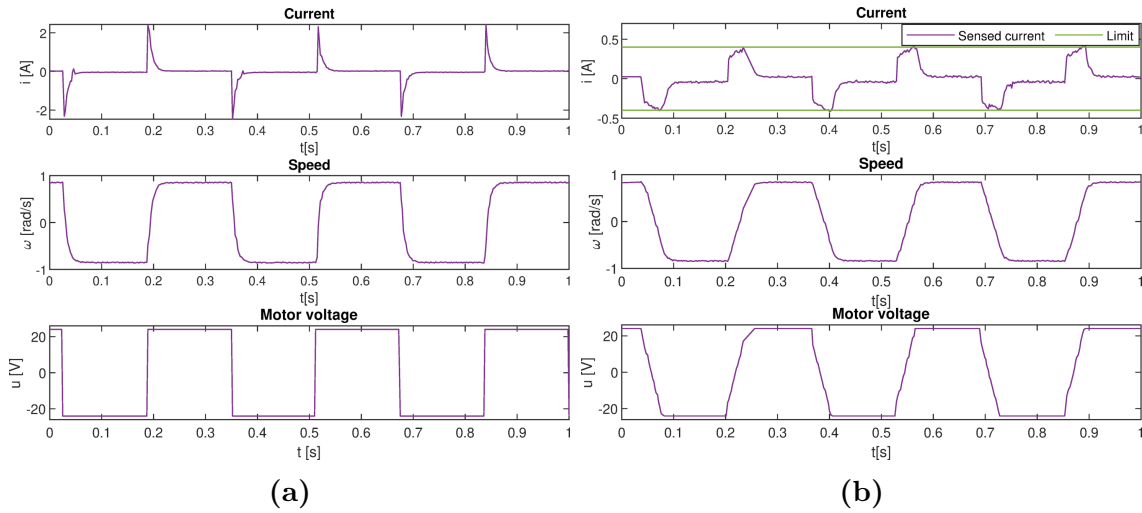


Figure 6.13: Free rotor simulation: (a) current, speed and applied voltage without current limitation, free movement, (b) current, speed and applied voltage with full-time current limitation, free movement.

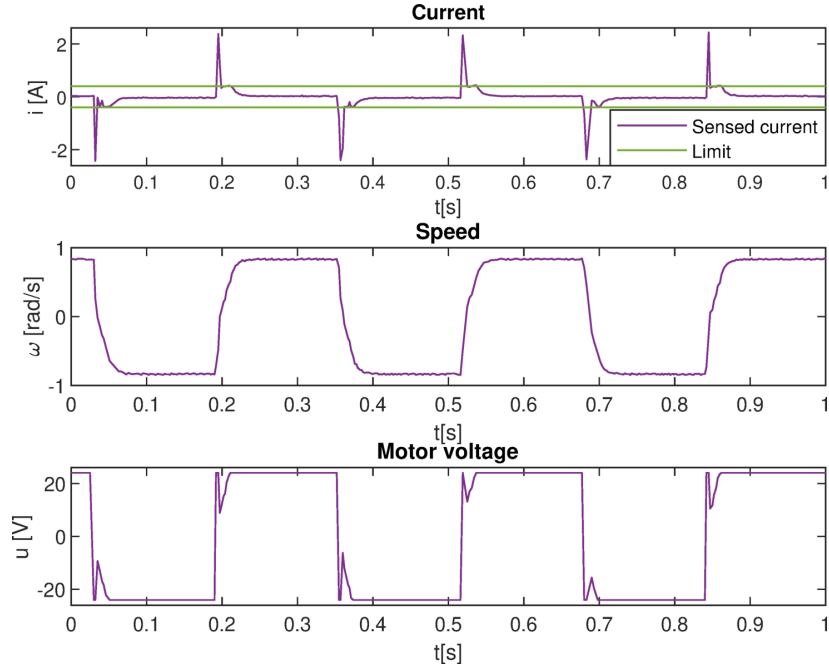


Figure 6.14: Current, speed and applied voltage with current limitation and allowable current peaks, free movement.

of 3.33 Hz. The effectiveness of the basic version of the algorithm in suppressing current peaks can be observed in Fig. 6.13. In Fig. 6.13a, the limitation is not applied. Here, the current peaks can be observed. When applying the limitation (Fig. 6.13b), the current peaks vanish.

The same experiment was performed introducing the predictor-based current limitation algorithm with the state machine that allows short-duration currents peaks. Results are presented in Fig. 6.14. In this case, the duration of the current peaks is shorter than without limitation (around 4 ms, as imposed by design). As observed, the original voltage input is allowed during a certain time when it changes polarity, but after the predefined time during which current is allowed to surpass the threshold, the voltage is saturated and gradually increases as the dynamics of the motor allow it without exceeding i_{sat} .

To stall the motor and test it, a mechanical arrangement to immobilize the rotor was prepared. With the objective of reducing the high torque delivered by the motor when applying 24 V to it, it was decided to conduct this experiment using a 12 V square input voltage signal. The reduction of the amplitude of the voltage reduces the torque delivered by the motor, allowing a simpler mechanical setup. Differently to the simulation, it was not possible to mechanically lock the rotor in such a manner that it is absolutely still since the moment in which voltage is applied. The mechanical setup allows certain movement of the rotor until it is completely stopped. For this reason, the duration of the experiment was set to 20 s and only positive voltages were applied, so that the effect of the slack was reduced. In this way, pulses of 12 V and 0 V of a duration of 3 s were applied to the motor.

Figure 6.15 shows the results obtained without current limitation and with the algorithm that does not allow current peaks to be produced. When not limiting the current, note that the current values for the peaks, as well as for the overload condition, are lower than in simulation because the input voltage was 12 V in this case,

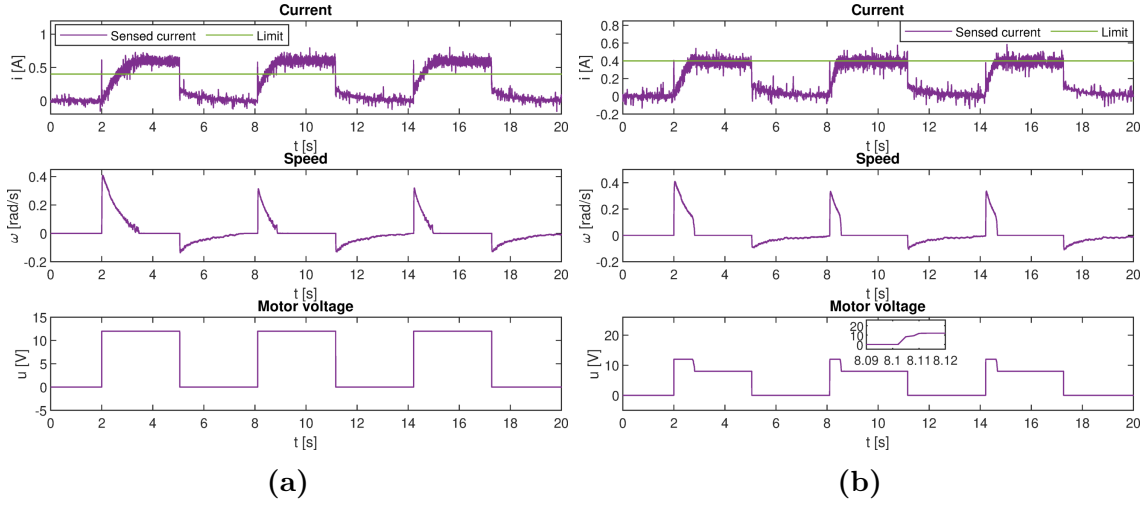


Figure 6.15: Free rotor simulation: (a) current, speed and applied voltage without current limitation, stall condition, (b) current, speed and applied voltage with full-time current limitation, stall condition.

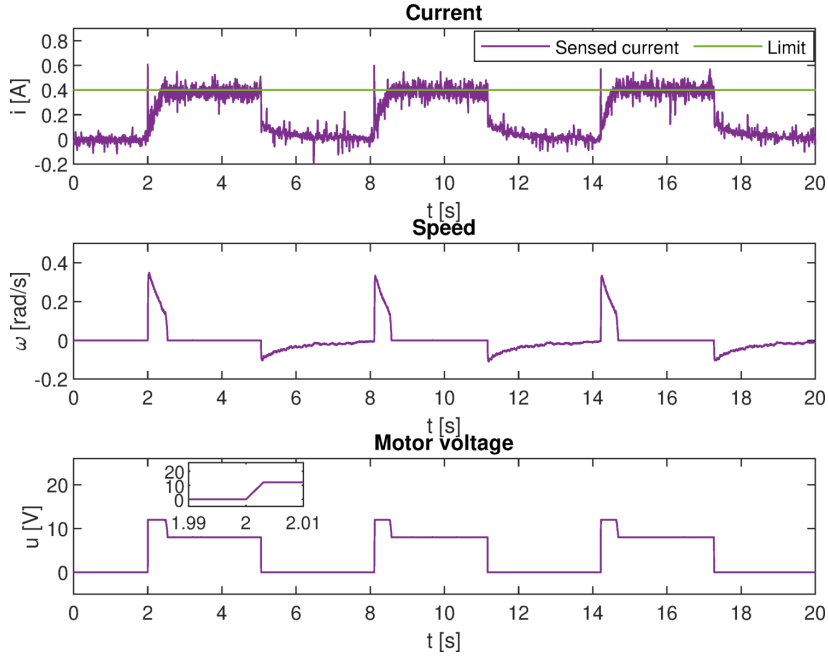


Figure 6.16: Current, speed and applied voltage with current limitation and allowable current peaks, stall condition.

in contrast to the 24 V applied in simulation. The current is effectively limited by the proposed algorithm as Fig. 6.15b shows. If the algorithm with the state machine that allows current peaks is implemented (Fig. 6.16), the current is not limited when the motor starts moving, allowing to deliver full torque while accelerating, but it is limited when the motor is in stall.

To better frame the difference between the performance obtained in simulation and in the experimental tests that were conducted, the percentages of absolute value of the current and power with respect to the allowable maximum values ($|i_s| = i_{\text{sat}}$, $P_s = i_{\text{sat}}^2 R$) for the simulations and the experiments are presented in Table 6.4. Not only it is important to avoid overcurrent, but the current, when limited, should be as

Table 6.4: Performance metrics: real implementations vs. simulation, when current is saturated.

	Limitation	Rotor	Power %	Current %
EXP	Full	Free	77.18	87.38
		Stall	94.51	96.55
	Peaks allowed	Free	82.87	88.09
		Stall	97.47	98.04
SIMU	Full	Free	83.37	91.22
		Stall	100.00	99.99
	Peaks allowed	Free	84.36	91.38
		Stall	100.06	100.04

close as possible to i_{sat} , and this is what these metrics evaluate. The current is farther from i_{sat} in the experimental tests than in simulation. Nevertheless, it is noteworthy the fact that the differences in current between the simulations and the experiments are below 4% for the two versions of the current limitation algorithm. Current over-limitation is more evident in free-rotor conditions. Limiting the current peaks is more difficult since it is a more dynamic situation than having the rotor completely stopped, situation in which the application of a fixed voltage while that condition holds is enough. Finally, the biggest difference in power between experiments and simulations is 6.19%.

6.3.2 Exoskeleton validation tests

The algorithm was validated in a real application involving an exoskeleton. For this purpose, an additional motor was added to ExoFlex to perform shoulder external rotation exercises (Fig. 6.17). This movement was selected because it normally involves higher forces than flexion-extension or abduction-adduction. Subjects wore a textile coupling in their forearm so that, by means of the cable winding or unwinding over the motor's pulley, shoulder rotation was generated.

Four healthy participants were involved in this validation. All experimental procedures were carried out in accordance with the Declaration of Helsinki on research involving human subjects and were approved by the ethical committee of the Universidad Politécnica de Madrid. The subjects were informed about the experiment and gave their consent to participate in it.

The motion generated for these tests involved two trapezoidal speed profiles. The trajectory consisted of an opening movement, in which the cable was wound over the motor pulley and pulled the forearm of the subject; and a closing movement, where the cable was unwound and the subject actively reduced the rotation angle of their arm, going back to the original position. In these experiments, the exact angle of rotation of the shoulder is irrelevant as only the elevation of the motor current and how the limitation algorithm deals with it is studied. For this reason, the trajectory profiles were performed using only motor feedback position, but not articular. The trapezoidal profiles were set between 0 rad and 10 rad, and then back from 10 rad to 0 rad. These values for the motor motion generated an approximate average rotation angle for the shoulder of the participants of around 80 degrees. Each of the four subjects performed the experiment at a different motor speed. Although

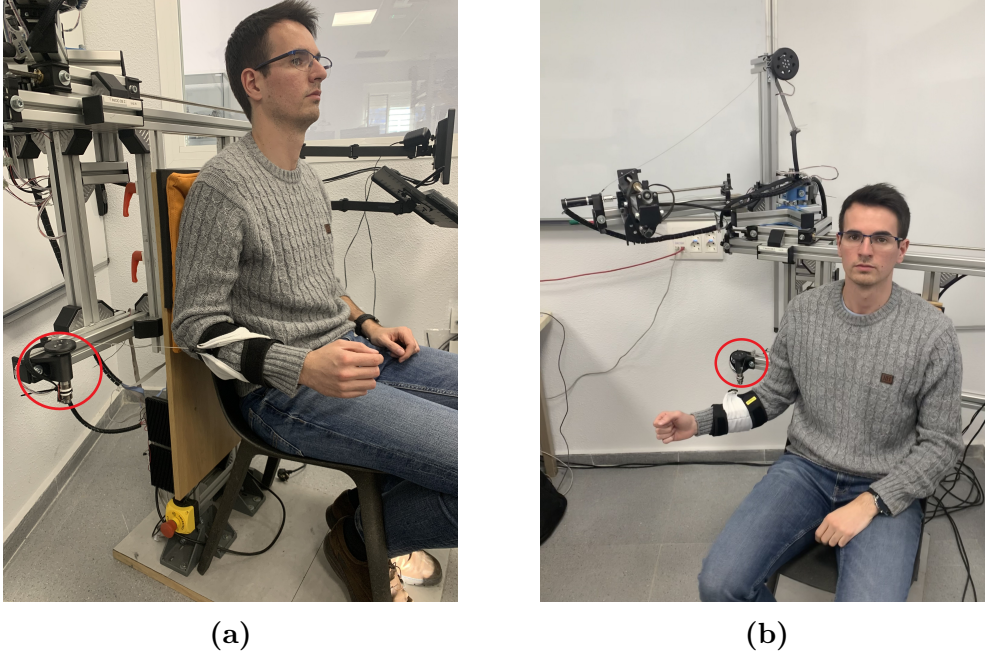


Figure 6.17: Subject performing a shoulder rotation exercise with ExoFlex. The auxiliary motor added to perform shoulder external rotation with ExoFlex is marked with a red circumference.

in Section 6.2.5 it was proven that the error associated to the estimation of speed had little influence in the performance of the current limitation algorithm, it was worthwhile to test the system under different speeds and check that the theoretical results correspond to the real behavior. The maximum speeds for the trapezoidal profiles were chosen as follows: 0.2 rad/s, 0.4 rad/s, 0.6 rad/s and 0.8 rad/s. The

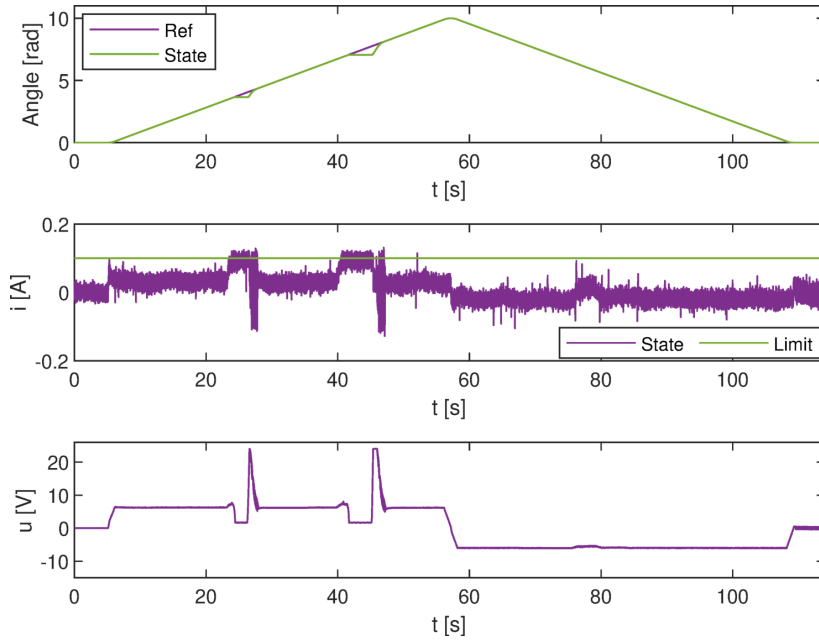


Figure 6.18: Motor position, current and applied voltage with full time current limitation, $\omega_{\max} = 0.2$ rad/s.

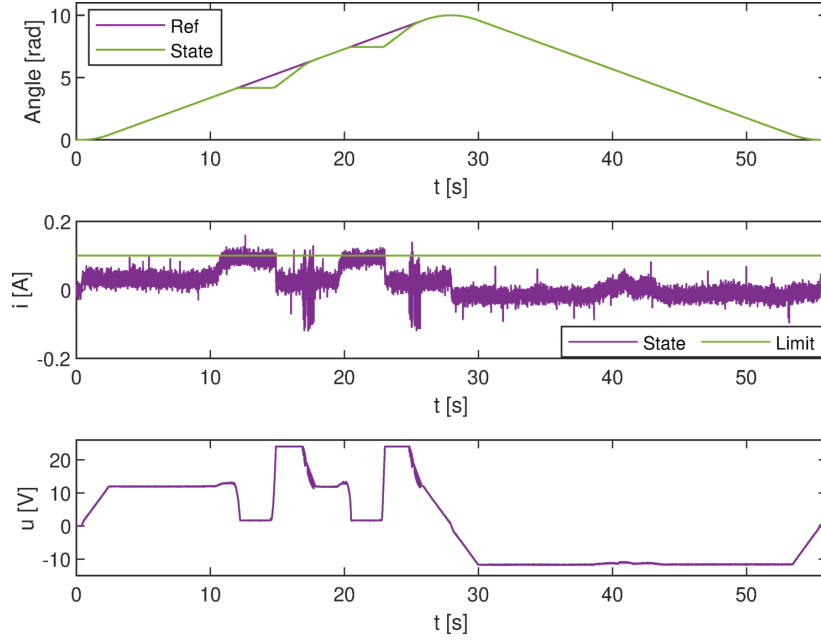


Figure 6.19: Motor position, current and applied voltage with full time current limitation, $\omega_{\max} = 0.4$ rad/s.

acceleration for all the tests was set to 0.2 rad/s^2 . It is necessary to remark that the motor maximum speed is 0.83 rad/s . For these tests, the current threshold was decreased from $i_{\text{sat}} = 0.4 \text{ A}$ to 0.1 A . This decision was made with the objective of reducing the necessary torque that subjects had to apply to generate an overcurrent situation and, therefore, the activation of the algorithm. With the same purpose, the version of the predictor-based current limitation algorithm that was implemented for these experiments was the one that completely limits the current, without letting the generation of short-duration current peaks. The generation of the shoulder rotation movement by the motor was not enough to generate overcurrent, so it was necessary that the subjects introduced disturbances to the system. In the movement opening phase, they were told to offer active opposition to the motor movement until it stopped due to the limitation algorithm at two different moments: the first one, gradually increasing the resistance force during the second third of this phase; the second one, in a brusquer way, during the last third of the phase. Similarly, they were told to accelerate the arm motion during the closing phase at some instant of their choice.

Experimental results for the four speeds are presented in Fig. 6.18 to Fig. 6.21. Looking at the opening phase of any of these figures, while the current is below the 100 mA threshold, the motor follows the predefined trajectory. Once the current reaches 100 mA , the voltage applied to the motor is saturated and the motor stops because the torque exerted by the subject is equal to the one generated by the motor. It is important to remark that it is not that the motor has stopped because no voltage is applied to it, but the motor is generating the maximum allowable torque determined by the 100 mA current, and the subject is doing a physical effort to stop the motor. Once the subject reduces the force, the motor keeps moving, following the commanded trajectory. In regard to the force applied by the subjects during the closing phase, that force is not enough to generate overcurrent. As this force is applied in a way which favours the motor movement, a much greater force

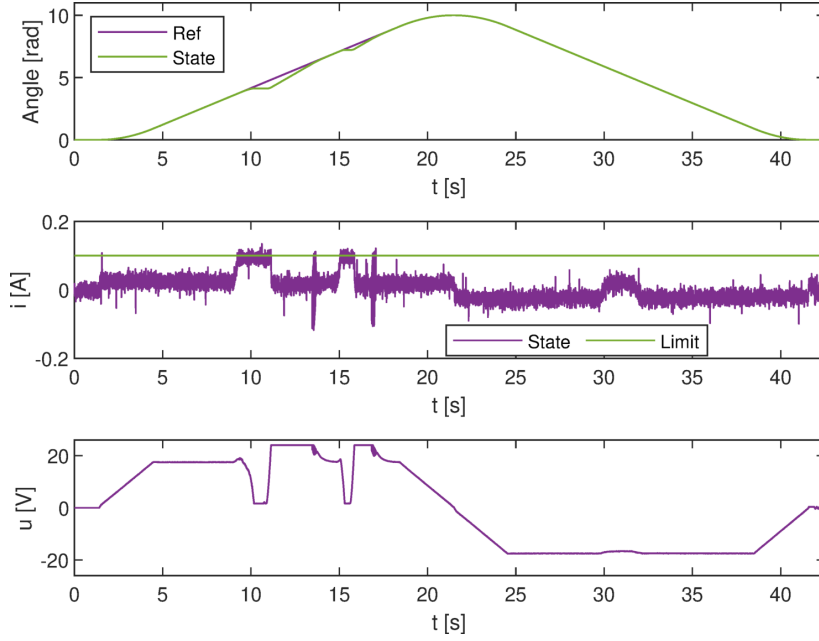


Figure 6.20: Motor position, current and applied voltage with full time current limitation, $\omega_{\max} = 0.6$ rad/s.

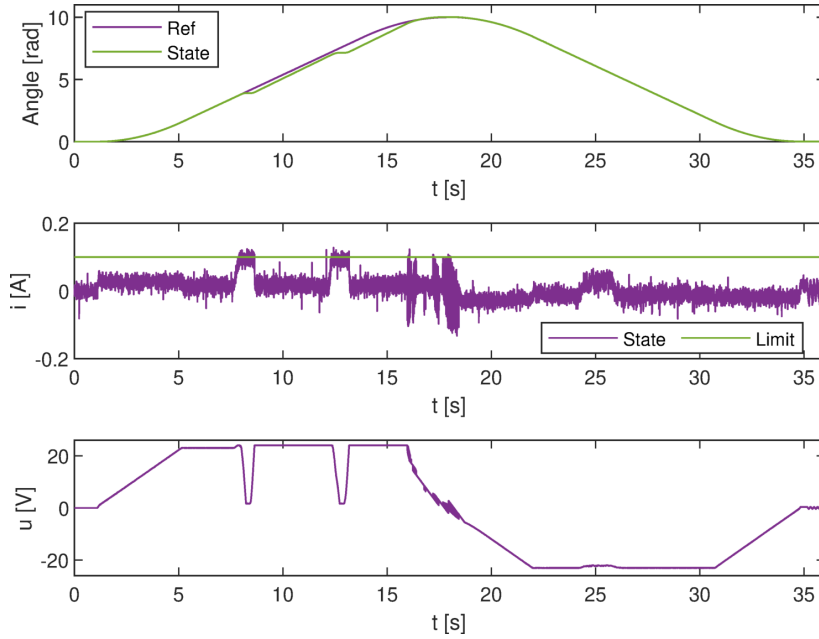


Figure 6.21: Motor position, current and applied voltage with full time current limitation, $\omega_{\max} = 0.8$ rad/s.

should be applied to generate overcurrent. However, having a look at the current waves of any of the tests, it is possible to observe a decrease in the absolute value of the currents in the time interval of the closing phase in which the subjects are applying the external force.

The percentage of the power and absolute value of the current when it is saturated, with respect to the maximum allowable conditions (P_s and $|i_s|$, respectively), are presented in Table 6.5 for the four speeds. As expected, no significant differences in performance for different speeds can be observed. The average percentage for the

Table 6.5: Performance metrics under current limitation: rehabilitation exoskeleton tests.

Max. speed [rad/s]	Power %	Current %
0.2	86.88	92.46
0.4	84.64	91.13
0.6	88.49	93.23
0.8	87.04	92.52
Average	86.76	92.34

current under limitation is 92.34%, and the average power is an 86.76% of the power associated to i_{sat} . These results are comparable with the ones obtained in simulation and in the stand-alone motor tests, so the effectiveness and efficiency of the designed predictor-based current limitation algorithm are validated for exoskeleton applications.

CHAPTER 7

Conclusion

7.1 Conclusions

This dissertation presented the design of five different control architectures (two for passive rehabilitation and three for active rehabilitation) applied to four different upper-limb exoskeletons or assistive robots. A software current limitation method consisting on a predictor based on the DC motor continuous dynamic model was developed for actuator protection. Additionally, a full electronics and real-time software platform was specifically designed for exoskeleton applications.

Control design for upper-limb passive rehabilitation with the ExoFlex exoskeleton was addressed with two substantially different approaches. The first one was a complex design which prioritized RMSE reduction, while the second one intended to be a simpler and easily implementable option. On the one hand, a novel cascade PI-SMC NDO-based motion controller (CPISDO) was designed, implemented and evaluated. The goals behind this controller were to maximize motion tracking accuracy, reduce the influence of model variability and uncertainties, and to reject disturbances. Additionally, the avoidance of subject-specific parameter tuning was also considered in the design. The result was a complex control architecture which resulted to meet all the aforementioned requirements. The effort performed in control design allowed to obtain a robust and accurate performance, compensating for the unmodelled dynamics and user variability. The performance offered by the CPISDO controller was evaluated over 10 healthy subjects and compared to a super-twisting SMC and a PI controller. The novel controller proved to outperform the other two conventional controllers, in terms of tracking accuracy, overshoot and non-oscillatory behavior. Additionally, the obtained RMSEs were also lower than other state-of-the-art approaches. It is necessary to remark that control parameters were not varied from one subject to another in the validation tests. This dramatically reduces the setup time for each patient and represents a big step forward for introducing this technology in rehabilitation centers, as the reduction of this time is critical.

On the other hand, a control architecture based on a geometrical model which removes the need of using on-body sensors for measuring shoulder orientation was proposed. It was observed that, when using IMUs for angular position sensing, they had relative displacement with the body of the subject after performing rehabilitation for a certain time. With the objective of eliminating the dangers derived from this misalignment, the on-body sensors were removed and the shoulder orientation was estimated through the developed geometrical method. This allowed to implement a much simpler control architecture, in which each actuator was individually position-controlled. The main drawback of this approach is the accuracy loss in shoulder orientation estimation due to the imprecisions made at measuring the acromion position, as well as the length of the upper arm and slack present in the textile interface. Nevertheless, the removal of the IMU sensors generated comfort increase for the participants, and the SUS and the questionnaire which addressed the clinicians' usability perception of the device showed that ExoFlex, along with this controller, is a usable tool for passive rehabilitation.

For passive rehabilitation, lots of different control techniques can be implemented, but the goal of the controller is always to provide motion control. Several trajectories with different speed profiles can be set as reference according to the rehabilitation objectives, but the control by itself always consists on motion tracking.

In contrast, many different control targets can be proposed for active rehabilitation, depending on the joint or joints to be rehabilitated, the source of injury or impairment, the stage of the rehabilitation in which the patient currently is and the type of exercise or application. In this context, the controllers for three different active assistance robots were designed.

An impedance controller based on an inner position loop was implemented for exoskeleton-assisted elbow active rehabilitation. As the motion inner control loop, three different controllers were tested: the CPISDO controller, a super-twisting SMC and a PID controller. Two different pure-damping impedances were tested over two subjects. The three controllers presented better performance for higher impedance than for the lower one. This is due to the fact that, for higher impedances, motion speeds are lower and easier to be tracked. It was observed that chattering generated using the sliding-mode controllers remarkably deteriorated the performance of the impedance controller. This was specially noticeable for the CPISDO controller, which was the controller which offered the best impedance tracking for the higher impedance, but was unable to be used with the lower one due to instability. For lower impedances, the super-twisting SMC represented the best option, and the PID offered an overall comfortable and stable behavior. Selecting different inner-loop implementations according to the target impedance value is a reasonable solution. User torque was estimated through a model-based approach using motor current sensing. The human torque estimation has a deep impact in the impedance controller performance, and the employed estimation method should be improved to obtain a more accurate and smooth torque signal. Additionally, the performance of the impedance controller using SMC-based inner position loops could be noticeably enhanced if chattering were better attenuated. All this said, results were promising and the device has potential to be used for elbow active rehabilitation.

For shoulder active rehabilitation, a cable-driven robot to perform internal and external shoulder rotation movements was developed. A PI force controller with gain planning was implemented for this purpose. With this tool, participants can autonomously perform a series of shoulder rotation exercises in different planes with a programmable force reference. The obtained results were encouraging. The capability of this robot to improve angular range of motion for shoulder rotation is supported by the performed clinical trials, with a range of motion increase in 10 out of 12 treated shoulders. It is remarkable that the participants had undergone a rehabilitation session with a physical therapist just before performing the robot-assisted therapy, what increases the significance of ROM improvement after using the device. Moreover, the results of the assessed SUS questionnaire showed that the participants perceived the usability of the device as excellent.

The work conducted over the soft wearable inflatable robot for assistance in ADLs for stroke survivors allowed to provide an example of the importance of having a reliable low-level controller in high-level applications. A gravity compensation controller had been implemented over this device, and its performance was significantly enhanced not only by adding some application-focused refinements to the high-level controller, but also by introducing significant improvements in the low-level one, as cut-off frequency increase for the filters, introduction of hardware considerations in the controller design, or proper parameter tuning. This soft wearable device is currently being tested in a study with stroke survivors to evaluate the provided assistance to this sector of population. The improved gravity compensation controller

will have a positive impact in the performance of the robot assisting stroke survivors in performing ADLs.

One of the most important outcomes of this work was learning that designing very complex control architectures is not always the best solution, in terms of cost and performance. This was not only learnt from the developed controllers in this dissertation, but from reading the literature about control in general and, particularly, control applied to exoskeletons. Application requirements should always be deeply analyzed before starting the design stage. Accuracy and speed might be key factors for some applications, as are for passive rehabilitation, for which designing a complex, robust and precise control architecture was a worthwhile effort. In other cases, comfort, smoothness and compliance might be the key factors, as were for the soft wearable inflatable device for ADL assistance. The mind of a control engineer should always be focused on finding the easiest suitable solution for each application, leaving aside the need one sometimes feels to develop new unnecessarily complex algorithms just for when the application truly requires it.

The actuation system restricts the range of implementable controllers over the system. DC motors are fast actuators which can easily be controlled with high accuracy using high-resolution encoders. Conversely, the pressure of a pneumatic actuator cannot usually be controlled with a precision comparable to the DC motor one, due to not as precise sensors and pumps. Additionally, the actuators for the inflatable device have much slower dynamics than DC motors have, so the attempt to implement, for example, the CPISDO controller over this robot would have failed.

The collaboration with clinicians during the development of the different exoskeletons represented a great support and an invaluable source of knowledge. The possibility of developing this kind of technology in a multidisciplinary group offers a better insight about the critical points to be addressed. Clinicians are the ones who can provide a better knowledge about the requirements for treating each kind of pathology, based not only on their studies, but also on their daily professional experience. On the other hand, engineers can propose solutions for those needs, establishing a discussion with the clinicians on the research plan to be followed and addressing the implementation process.

The development of a hardware and real-time software platform for exoskeleton control is one of the key factors of the success of this thesis. This design was started at the beginning of the author's PhD. It was a challenging process, which took years. The real-time software and the electronics grew as the development of the several exoskeletons presented in this thesis advanced, increasing the number and type of actuators and sensors, as well as the computing and control requirements. However, this was a worthwhile effort, since this platform allowed to obtain a very flexible (it was applied over three exoskeletons), modular and scalable system, which permitted to accurately control the application at all of its levels, from the lowest one to the highest. Off-the-shelf options do not offer this flexibility and synergy between hardware and software. The capability the system has to record data from experimental tests can be used for posterior clinical analysis and supervision of the rehabilitation process, which opens a set of new possibilities for diagnosis and rehabilitation monitoring.

Regarding the software module, the DC motor predictor-based current limitation method proved to effectively limit the current with minimal motor underuse. Not only does the algorithm limit the motor current, but it also keeps it near the prede-

finer threshold. This method was designed to protect the exoskeleton actuators from improper use in human interaction, but it can be applied to other applications. This method represents a good alternative to hardware limitation methods. The main limitation is that motor parameters should be known as accurately as possible for its implementation. However, the parameters can relatively easily be measured and good performance was obtained assuming tolerances of the 5% in most of the motor parameters, as demonstrated in simulation and experimentally.

The developed controllers and systems presented in this dissertation provide a representative sample of the research conducted in this field and a contribution to it. A variety of robots and applications have been exposed, for which control design was methodically, efficiently and rigorously approached.

7.2 Future work

The future work is mainly related to research on active rehabilitation. For the elbow exoskeleton, an improved method to estimate the torque exerted by the human should be investigated. The obtention of a more precise and smooth torque signal would considerably enhance the behavior of the impedance controller. Additionally, testing the controller with different target impedances which involve elastic or inertia-like behaviors is also of interest.

Regarding the soft wearable inflatable device for assistance in performing activities of daily living, investigating other task-focused control techniques apart from the presented gravity compensation control is one of the main next steps. The exploration of the introduction of AI techniques to learn tasks can be a reasonable starting point.

Finally, the conducted experimental tests were mostly focused on evaluating the control performance of the proposed controllers and on studying the perception of both patients and clinicians of the developed exoskeletons. The obtained results were promising and encouraging, so they open the door to performing middle- and long-term clinical studies with a greater population to evaluate the impact of using exoskeletons in a full rehabilitation process.

List of Publications

Journals

- [1] J. L. Samper-Escudero, A. F. Contreras-González, M. Ferre, M. A. Sánchez-Urán, and D. Pont-Esteban, “Efficient multiaxial shoulder-motion tracking based on flexible resistive sensors applied to exosuits,” *Soft Robotics*, 2020.
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List of Acronyms

ADL	Activities of Daily Living.	2
ADS	Angular Displacement Sensor.	32
ALICE	Assistive Limb Control Electronics.	33
CAD	Computer-Aided Design.	84
CCS	Code Composer Studio.	37
CPISDO	Cascade PI-SMC NDO-based Controller.	40
CPU	Central Processing Unit.	37
DC	Direct Current.	24
DoF	Degree of Freedom.	11
ECG	Electrocardiogram.	18
EEG	Electroencephalography.	18
EMG	Electromyography.	9
EOG	Electrooculogram.	18
GPIO	General Purpose Input/Output.	34
GUI	Graphical User Interface.	32
HLC	High-Level Controller.	111
HWI	Hardware Interrupt.	37
I²C	Inter-Integrated Circuit.	34
IMU	Inertial Measurement Unit.	9

LLC	Low-Level Controller.	111
MIMO	Multiple-input Multiple-output.	48
NDO	Nonlinear Disturbance Observer.	41
OT	Occupational Therapist.	116
PMA	Pneumatic Muscle Actuator.	15
PT	Physical Therapist.	24
PWM	Pulse Width Modulation.	35
QEP	Quadrature Encoder Pulse.	33
ROM	Range Of Motion.	28
SCI	Spinal Cord Injury.	18
SEA	Series Elastic Actuators.	18
SISO	Single-input Single-output.	48
SMC	Sliding Mode Control.	15
SPI	Serial Peripheral Interface.	33
SWI	Software Interrupt.	37
TENS	Transcutaneous Electrical Nerve Stimulation.	4
TI	Texas Instruments.	32
UART	Universal Asynchronous Receiver-Transmitter.	34
USB	Universal Serial Bus.	32
VESC	Vedder Electric Speed Controller.	109
WHO	World Health Organization.	3

APPENDIX A

Anatomical background for upper-limbs

This appendix provides basic graphical information of the anatomical background necessary to understand the technical vocabulary related to elbow and shoulder motion. It does not intend to give an extensive anatomical overview of the muscles involved in each movement, as it just focuses on making the reader understand the different types of movements involved for elbow and shoulder which are considered in this work, as well as the anatomical planes of relevance. The assisted shoulder movements (Fig. A.1) are flexion-extension, abduction-adduction and rotation. Arm elevation is also treated, which is a combination of flexion and abduction movements of the shoulder, with an intermediate aperture angle between pure flexion and pure abduction. For the elbow (Fig. A.2), just flexion-extension is assisted, but the pronation-supination movement is also discussed along the document.

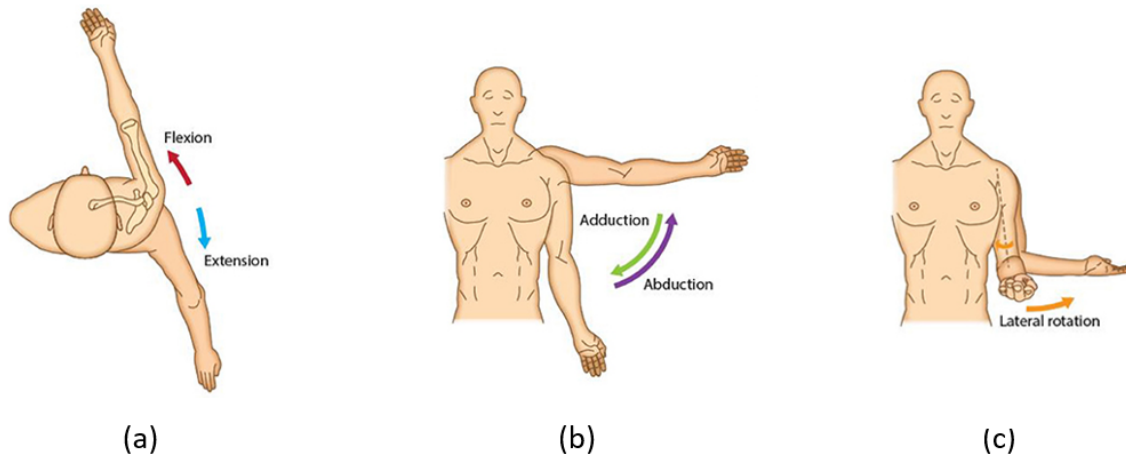


Figure A.1: Shoulder movements: (a) flexion-extension, (b) abduction-adduction, (c) rotation. Images extracted from [204].

Regarding the anatomical planes (Fig. A.3), four main planes are of relevance for this work: the sagittal plane (which divides the body into two equal parts, and is perpendicular to the transverse and frontal planes), the horizontal or transverse plane (which divides the body into superior and inferior sections, and is perpendic-

ular to the frontal and sagittal planes), the frontal or coronal plane (which divides the body into dorsal and ventral sections, and is perpendicular to the sagittal and transverse planes) and the vertical plane or para-sagittal plane (which is a plane parallel to the sagittal plane, placed at an arbitrary distance of it).

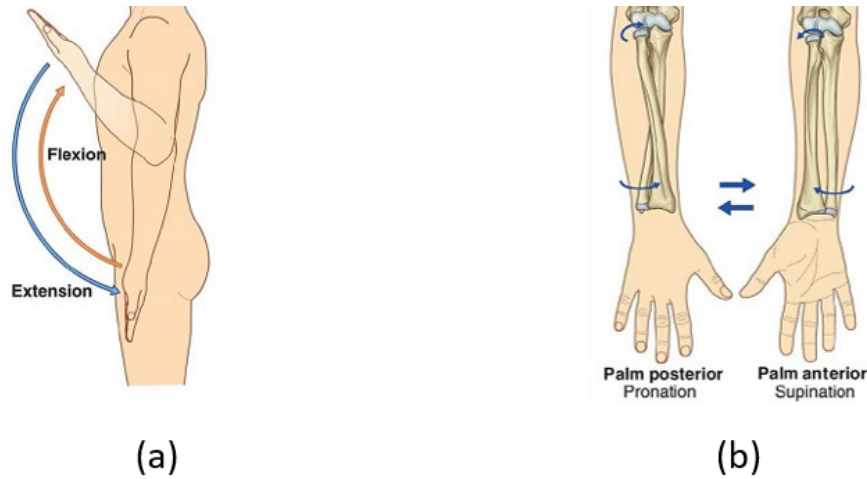


Figure A.2: Elbow movements: (a) flexion-extension, (b) pronation-supination. Images extracted from [205].

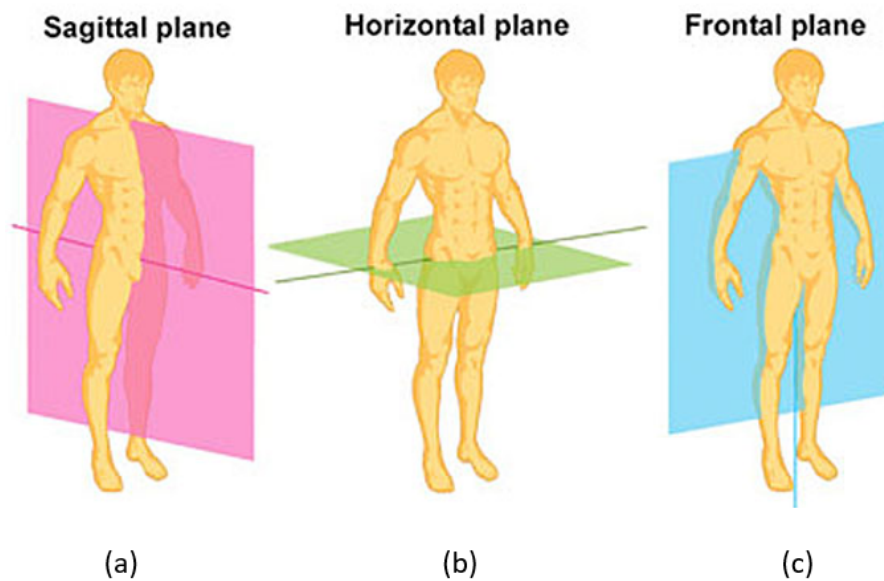


Figure A.3: Anatomical planes: (a) sagittal plane, (b) horizontal or transverse plane, (c) frontal or coronal plane. Image extracted from [206].

APPENDIX B

Current sensing board design

In this appendix, the designed supplementary PCB to ALICE board for current sensing is described. The board (Fig. B.1) includes four current sensing circuits, each of them based on the Allegro ACS723LLCTR-05AB-T Hall sensor. It was decided to design the current sensing module as a separate stackable PCB for two main reasons: first, to increase system flexibility, providing the possibility of using the current sensing board externally to the ALICE board in a different setup; in second place, to allow to replace it if any of the channels is somehow damaged. The current sensing board includes a set of jumpers which allow to connect the current sensors to the motor drivers or to connect all the current sensors in series to calibrate them through an external current source.

The Allegro Hall sensor can measure currents in the range of ± 5 A, and has a sensitivity of 400 mV/A. The safe operation area of the employed DC motors is in the range of ± 0.4 A, but a range of ± 0.5 A was considered in the design to allow a reasonable margin. The ADC of the microcontroller can measure voltages between 0 V and 3 V, so the sensor output signal must be in that range to optimize resolution.

The schematic for the proposed current sensing circuit for one channel is shown in Fig. B.2. The schematic consists of three modules: the Hall sensor, the amplification

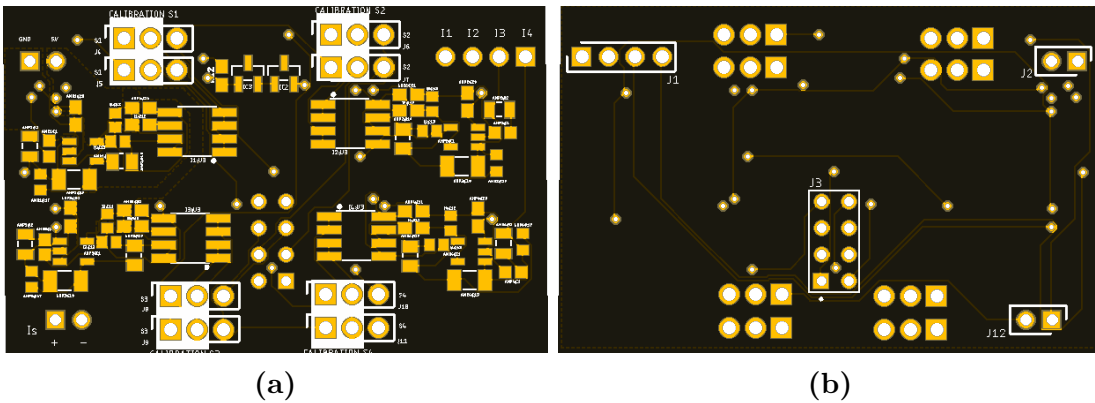


Figure B.1: Current sensing board CAD model without assembled components: (a) top view, (b) bottom view.

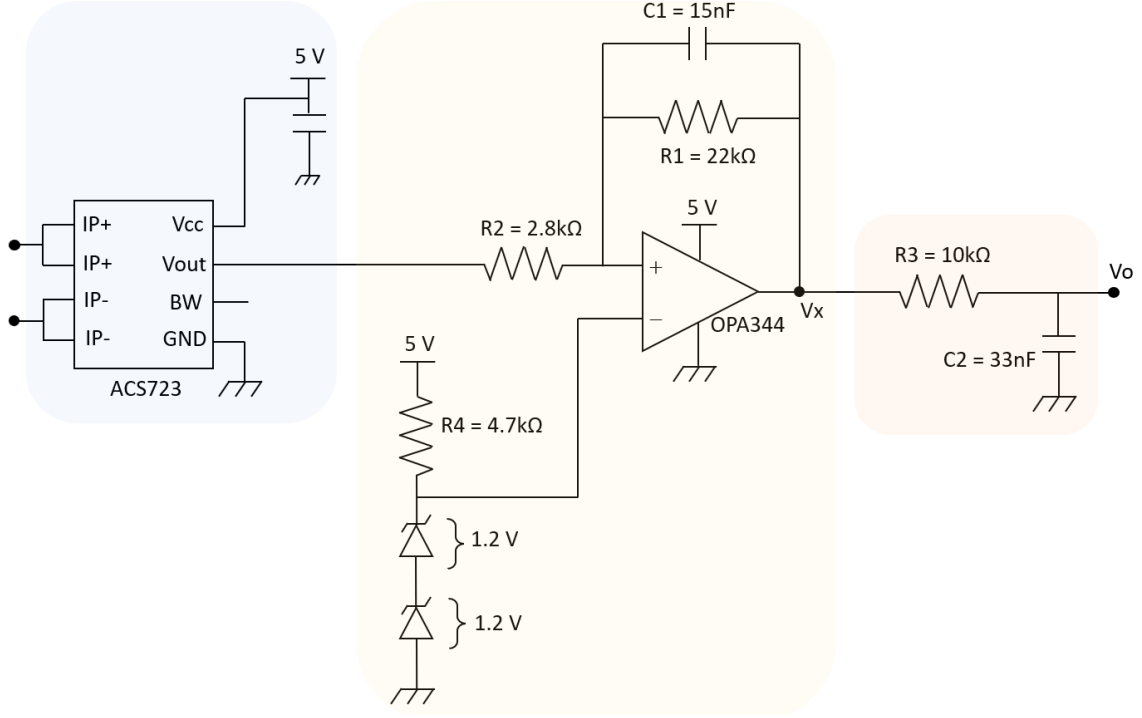


Figure B.2: Schematic for one channel current sensing circuit. The module shaded in blue is the Hall sensor, the yellow one is the amplification module and the orange one is the output filter. The current sensing board includes four of these circuits.

stage and an output RC filter. The amplification stage also contains a first-order filter. Given that the ADC sampling is performed at 1 kHz, according to the Nyquist-Shannon sampling theorem [207], the cut-off frequency was set to 482 Hz for both filters. It was considered to introduce a lower cut-off frequency. However, hardware-filtering the signal removed flexibility of the sensing system, so the possibility of reducing the cut-off frequency was implemented through digital filters in the software module.

In the Laplace domain, the circuit output voltage is:

$$V_o = -\frac{R_1}{R_2} \frac{V_{out}}{(1 + R_3 C_2 s)(1 + R_1 C_1 s)} + \left(\frac{1}{1 + R_3 C_2 s} + \frac{R_1}{R_2} \frac{1}{(1 + R_3 C_2 s)(1 + R_1 C_1 s)} \right) V^+ \quad (\text{B.1})$$

At frequencies below the cut-off frequency, the output voltage can be approximated by:

$$V_o = -\frac{R_1}{R_2} V_{out} + \left(1 + \frac{R_1}{R_2} \right) V^+ \quad (\text{B.2})$$

To obtain the theoretical calibration curve which relates the value captured by the 12-bit ADC (ADC_{REG}) with the current i , the voltage mapping to the ADC range (B.3) and the ACS723 output (B.4), extracted from the datasheet, have to be considered:

$$V_o = \frac{3}{2^{12} - 1} \text{ADC}_{\text{REG}} \quad (\text{B.3})$$

$$V_{out} = si + 2.5, \quad s = 0.4 \frac{\text{mV}}{\text{A}} \quad (\text{B.4})$$

Solving (B.2) for V_{out} , substituting it in (B.4) and substituting (B.3) in the resulting expression, the current as a function of the captured ADC values is:

$$i(\text{ADC}_{\text{REG}}) = \frac{1}{s} \left(-\frac{R_1}{R_2} \left(\frac{3}{2^{12}-1} \text{ADC}_{\text{REG}} - \left(1 + \frac{R_1}{R_2} \right) V^+ \right) - 2.5 \right) \quad (\text{B.5})$$

However, for proper calibration, ADC values applying a set of predefined currents through a precision current source were collected. Data were fitted to a first-order polynomial.

An evaluation of the influence of the tolerance of the PCB components over the sensor performance was addressed. The main concern was to ensure that the motor currents (± 0.4 A) fitted in the ADC range (0 V to 3 V). The selected values and tolerances for the components of the study were $R_1 = 22 \text{ k}\Omega \pm 1\%$, $R_2 = 2.8 \text{ k}\Omega \pm 0.1\%$ and $V^+ = 2.4 \text{ V} \pm 0.3\%$. Table B.1 presents the critical combinations of these values and proves how the motor currents properly fit in the ADC range. The first row represents the nominal combination.

Table B.1: Analysis of the influence of parameter tolerance over the sensor measurement range.

Component value			Output voltage for each current [V]			
$R_1[\Omega]$	$R_2[\Omega]$	$V^+[\text{V}]$	i=0.5A	i=-0.5A	i=0.4A	i=-0.4A
22000	2800	2.400	0.043	3.186	0.357	2.871
22220	2797	2.407	0.081	3.259	0.399	2.941
21780	2802	2.393	0.006	3.114	0.316	2.803
22000	2800	2.393	-0.047	3.130	0.270	2.812
21780	2802	2.407	0.132	3.240	0.443	2.929

CPISDO supplementary material

The content included in this appendix was extracted from the author's publication "*Robust motion control architecture for an upper-limb rehabilitation Exosuit*", IEEE Access, vol. 10, pp. 113631–113648, 2022. doi: 10.1109/ACCESS.2022.3217528.

C.1 Model equations and parameters

The most important equations for modelling ExoFlex v1 have been described in Section 4.2.1. In this section, the model equations which are necessary for performing the controllability analysis are provided.

$$f_{5is}(x_1, x_4) = \frac{-d_5 l_a \sin(\alpha_s) n k_p}{r d_s} \quad (C.1)$$

$$f_{5ie}(x_1, x_4) = \frac{d_5 l_f \sin(\alpha_e) n k_p}{r H_{21}} \quad (C.2)$$

$$f_{2is}(x_1, x_4) = \frac{l_a \sin(\alpha_s) n k_p}{r d_s} - \frac{H_{12}}{d_s} f_{5is} \quad (C.3)$$

$$f_{2ie}(x_1, x_4) = -\frac{H_{12}}{d_s} f_{5ie} \quad (C.4)$$

$$f_{03}(\mathbf{x}) = -\frac{1}{L} (n k_e k_{1s} x_2 + R x_3) \quad (C.5)$$

$$f_{06}(\mathbf{x}) = -\frac{1}{L} (n k_e k_{1e} x_5 + R x_6) \quad (C.6)$$

$$b = 1/L \quad (C.7)$$

with

$$k_{1s}(x_1) = -\frac{1}{r} \frac{dl_s(x_1)}{dx_1}, \quad k_{1e}(x_1, x_4) = -\frac{1}{r} \frac{\partial l_e(x_1, x_4)}{\partial x_4} \quad (C.8)$$

$$d_5(x_1, x_4) = \left(\frac{H_{22}}{H_{21}} + \frac{l_f \sin(\alpha_e) J k_{1e}}{r H_{21}} - \frac{H_{12}}{d_s} \right)^{-1} \quad (C.9)$$

$$d_s(x_1, x_4) = H_{11} + l_a \sin(\alpha_s) Jk_{ls}/r \quad (C.10)$$

$$H_{11}(x_4) = m_1 l_{c1}^2 + I_1 + I_2 + m_2 (l_{at}^2 + l_{c2}^2 + 2l_{at}l_{c2} \cos(x_4)) \quad (C.11)$$

$$H_{12}(x_4) = H_{21}(x_4) = I_2 + m_2 (l_{at}l_{c2} \cos(x_4) + l_{c2}^2) \quad (C.12)$$

$$H_{22} = m_2 l_{c2}^2 + I_2 \quad (C.13)$$

Functions f_{2a} and f_{5a} in (4.19) and (4.20) are nonlinear functions of x_1, x_2, x_4 and x_5 , which do not depend on the motor currents x_3 and x_6 and do not have further interest for the mathematical content of this work. Model parameters of the above functions are described in Table C.1 along with their values. Figure 4.1 should be consulted to obtain geometrical context.

Table C.1: ExoFlex v1 specific model parameters

Symb.	Description	Value
l_f	Distance from elbow to cable anchor point	0.08 m
m_1	Upper-arm mass	2.423kg
l_{c1}	Distance from shoulder to upper-arm CoM	0.153 m
m_2	Forearm mass	2.260kg
l_{c2}	Distance from elbow to forearm CoM	0.21 m
I_1	Upper-arm inertia	0.0751 kgm ²
l_{at}	Upper-arm total length	0.305 m
I_2	Forearm inertia	0.0738 kgm ²
x_{ps}	Distance in x from shoulder to pulley D	0.072 m
J	Actuation system lumped inertia	0.2942 kgm ²
O_{sy}	Shoulder y distance to origin	0.15 m
k_p	Motor torque constant	0.0359 N/A
E_x	Distance in x from E to O_s	0.03 m
k_e	Motor electric constant	0.0359 $\frac{V}{(\text{rad/s})}$
y_{ps}	Distance in y from shoulder to pulley D	0.07 m
R	Motor resistance	16.6 Ω
O_{sx}	Shoulder x distance to origin	0.05 m
L	Motor inductance	0.881 mH
E_y	Distance in y from E to O_s	0.03 m
n	Motor reduction ratio	794
l_a	Distance from shoulder to cable anchor point	0.18 m

C.2 System controllability

Proof 1 Assuming that there do not exist disturbances and uncertainties over the model in (4.13) and splitting $\mathbf{g}$ in the columns for the control inputs u_1 and u_2 :

$$\dot{\mathbf{x}}(t) = \mathbf{f}_0(\mathbf{x}) + \mathbf{g}_1 \cdot u_1(t) + \mathbf{g}_2 \cdot u_2(t) \quad (\text{C.14})$$

Vector fields $\mathbf{f}_0$, $\mathbf{g}_1$ and $\mathbf{g}_2$ determine the way the system state can evolve in the state space $\mathbf{X} \subset \mathbb{R}^6$. Using the Lie Brackets of these fields, the system accessibility from a point $\mathbf{x}_0 \in \mathbf{X}$ can be studied [208]. The Lie Bracket of two vector fields $\mathbf{f}$ and $\mathbf{g}$ is a new vector field which is the derivative of $\mathbf{g}$ along the flow generated by $\mathbf{f}$ and is denoted as $[\mathbf{f}, \mathbf{g}](\mathbf{x})$ (C.15).

$$[\mathbf{f}, \mathbf{g}](\mathbf{x}) = \frac{\partial \mathbf{g}}{\partial \mathbf{x}}(\mathbf{x})\mathbf{f}(\mathbf{x}) - \frac{\partial \mathbf{f}}{\partial \mathbf{x}}(\mathbf{x})\mathbf{g}(\mathbf{x}) \quad (\text{C.15})$$

If the accessibility distribution $\mathbf{Q}$ spans $\mathbb{R}^n$ (i.e. the rank of $\mathbf{Q}$ equals n), with

$$\begin{aligned} \mathbf{Q} = & [\mathbf{g}_1, \dots, \mathbf{g}_m, [\mathbf{g}_i, \mathbf{g}_j], \dots, [\mathbf{ad}_{\mathbf{g}_i}^k, \mathbf{g}_j], \dots, [\mathbf{f}, \mathbf{g}_i], \dots, \\ & [\mathbf{ad}_{\mathbf{f}}^k, \mathbf{g}_i], \dots], k = 1 \dots n-1, i = 1 \dots m, j = 1 \dots m \end{aligned} \quad (\text{C.16})$$

being n the number of state variables and m the number of inputs, then the system is locally accessible from $\mathbf{x}_0$ [209]. If the system is locally accessible from every $\mathbf{x}_0 \in \mathbf{X}$, then it is controllable in $\mathbf{X}$ [210]. Here, we intend to study the controllability in $\mathbf{X}_c \ni \forall x_1, x_4 \mid x_1 + x_4 < 180^\circ, x_1 \in [0^\circ, 110^\circ], x_4 \in [0^\circ, 150^\circ]$. As for the articular speeds and motor currents $x_2 \in [\omega_{\min}, \omega_{\max}]$, $x_3 \in [i_{\min}, i_{\max}]$, where $\omega_{\min}$, $\omega_{\max}$, $i_{\min}$ and $i_{\max}$ depend on the actuator limitations. At a theoretical level, $\omega_{\max} = i_{\max} = \infty$ and $\omega_{\min} = i_{\min} = -\infty$. So, to prove the system controllability in $\mathbf{X}_c$, it is necessary to find 6 linearly independent rows of $\mathbf{Q}$ ($n = 6$) when evaluating all the vector fields in every $\mathbf{x}_0 \in \mathbf{X}_c$.

Taking advantage of the system's particular structure (constant input vectors, decoupled inputs and the way every f_i depends on $\mathbf{x}$), the analytical controllability analysis of this 2-input 6th order system is feasible (see Section 4.2 and Appendix C.1). The first two vector fields which will be considered are $\mathbf{g}_1$ and $\mathbf{g}_2$, which clearly are linearly independent for every $\mathbf{x}_0$. The two next proposed fields are $[\mathbf{f}_0, \mathbf{g}_1]$ and $[\mathbf{f}_0, \mathbf{g}_2]$. Using expression (C.15), the new vector fields are

$$[\mathbf{f}_0, \mathbf{g}_1] = \left[0, -b \frac{\partial f_2}{\partial x_3}(\mathbf{x}), b \frac{R}{L}, 0, -b \frac{\partial f_5}{\partial x_3}(\mathbf{x}), 0 \right] \quad (\text{C.17})$$

$$[\mathbf{f}_0, \mathbf{g}_2] = \left[0, -b \frac{\partial f_2}{\partial x_6}(\mathbf{x}), 0, 0, -b \frac{\partial f_5}{\partial x_6}(\mathbf{x}), b \frac{R}{L} \right] \quad (\text{C.18})$$

It seems that $[\mathbf{f}_0, \mathbf{g}_1]$ could be expressed as a linear combination of $[\mathbf{f}_0, \mathbf{g}_2]$, $\mathbf{g}_1$ and $\mathbf{g}_2$. This will only happen if the second and fifth components of (C.17) and (C.18) are linearly dependent. If there exists a state $\mathbf{x}_0$ so that the following conditions are met

$$\begin{cases} \frac{\partial f_2}{\partial x_3}(\mathbf{x}_0) = k_1 \frac{\partial f_2}{\partial x_6}(\mathbf{x}_0) \\ \frac{\partial f_5}{\partial x_3}(\mathbf{x}_0) = k_1 \frac{\partial f_5}{\partial x_6}(\mathbf{x}_0) \end{cases} \quad k_1 \in \mathbb{R} \quad (\text{C.19})$$

then $[\mathbf{f}_0, \mathbf{g}_1](\mathbf{x}_0)$ and $[\mathbf{f}_0, \mathbf{g}_2](\mathbf{x}_0)$ are linearly dependent. Otherwise, they are linearly independent. By developing the conditions from (C.19) according to the model in (4.16) to (4.20) and Appendix C.1, it can be seen that they only depend on x_1, x_4

and k_1 . Computing the derivatives, dividing the first equation by the second one and moving all the terms in the left-hand side, expression (C.20) is obtained.

$$z(x_1, x_4) = \frac{f_{2is}(x_1, x_4)}{f_{5is}(x_1, x_4)} - \frac{f_{2ie}(x_1, x_4)}{f_{5ie}(x_1, x_4)} = 0 \quad (\text{C.20})$$

If we instead divide the second equation in (C.19) by the first one and follow the same procedure, we obtain

$$q(x_1, x_4) = \frac{f_{5is}(x_1, x_4)}{f_{2is}(x_1, x_4)} - \frac{f_{5ie}(x_1, x_4)}{f_{2ie}(x_1, x_4)} = 0 \quad (\text{C.21})$$

Fig. C.1 shows the partial derivative terms in (C.19). Terms $\frac{\partial f_5}{\partial x_3}$ and $\frac{\partial f_2}{\partial x_6}$ evaluate to 0 in some points, so z and q will have singularities on those points, as Fig. C.2 shows. If q had not been studied, the singularity in z could have been misunderstood as a zero in q and vice versa, but the plot of both functions shows that they never cross 0. On no account neither (C.20) nor (C.21) have a solution, so $[\mathbf{f}, \mathbf{g}_1]$ and $[\mathbf{f}, \mathbf{g}_2]$ are linearly independent for every state $\mathbf{x}_0 \in \mathbf{X}_c$.

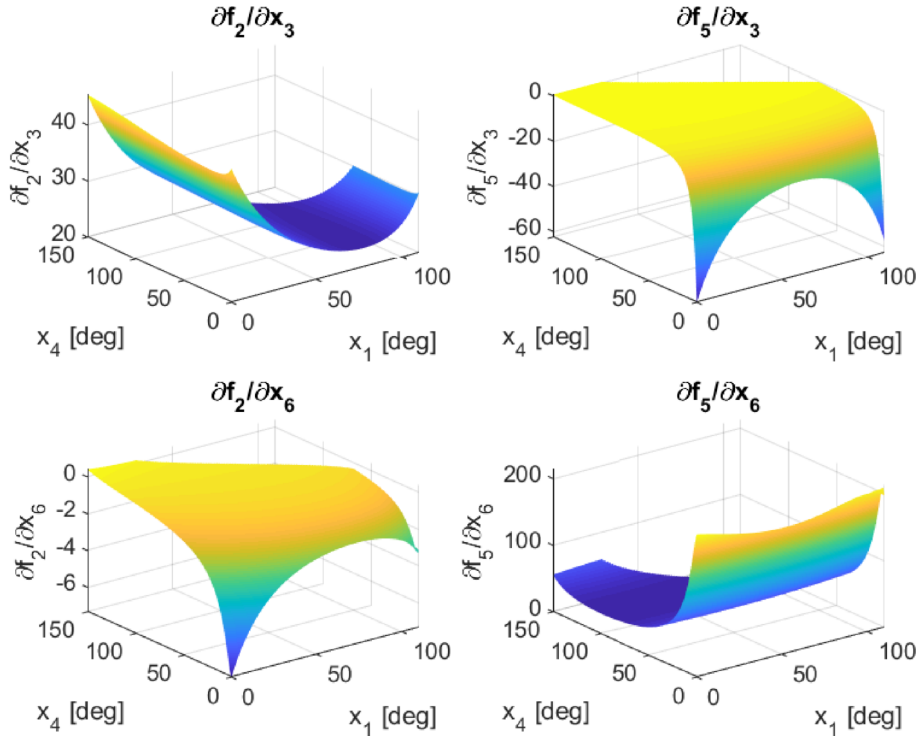


Figure C.1: Derivative terms of z and q for $x_1 \in \mathbf{X}_c$ and $x_4 \in \mathbf{X}_c$. Source: Author's publication ©David Pont Esteban et al. 2022; published by IEEE (<http://doi.org/10.1109/ACCESS.2022.3217528>).

Remark 2 If $\frac{\partial f_2}{\partial x_3}(\mathbf{x}_0) = 0$ and $\frac{\partial f_5}{\partial x_3}(\mathbf{x}_0) = 0$ then $[\mathbf{f}_0, \mathbf{g}_1] \parallel \mathbf{g}_1$ and if $\frac{\partial f_2}{\partial x_6}(\mathbf{x}_0) = 0$ and $\frac{\partial f_5}{\partial x_6}(\mathbf{x}_0) = 0$ then $[\mathbf{f}_0, \mathbf{g}_2] \parallel \mathbf{g}_2$. In that case, $[\mathbf{f}_0, \mathbf{g}_1]$ and $[\mathbf{f}_0, \mathbf{g}_2]$ would not be linearly independent of $\mathbf{g}_1$ and $\mathbf{g}_2$. However, note that Fig. C.2 shows that $\frac{\partial f_2}{\partial x_3}(\mathbf{x}_0) \neq 0, \forall \mathbf{x} \in \mathbf{X}_c$ and $\frac{\partial f_5}{\partial x_6}(\mathbf{x}_0) \neq 0, \forall \mathbf{x} \in \mathbf{X}_c$, proving that the aforementioned situation cannot arise, and therefore, the linear independence of the four vector fields evaluated in every $\mathbf{x}_0 \in \mathbf{X}_c$ is guaranteed.

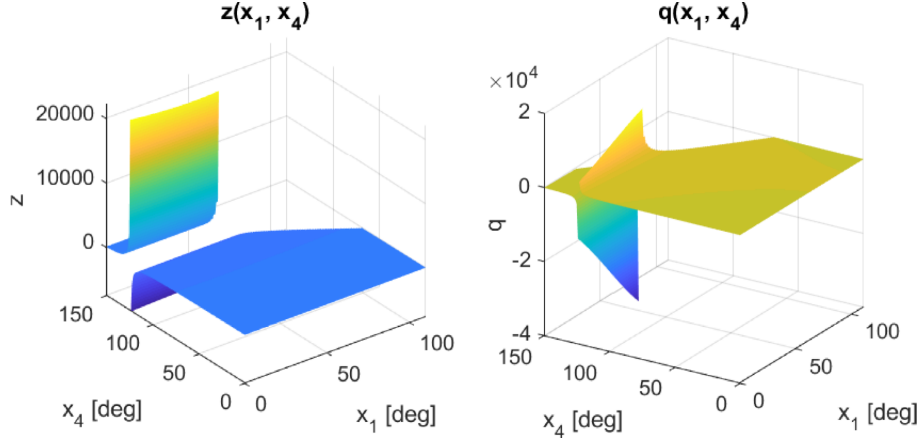


Figure C.2: Computation of z and q for $x_1 \in \mathbf{X}_c$ and $x_4 \in \mathbf{X}_c$. Source: Author's publication ©David Pont Esteban et al. 2022; published by IEEE (<http://doi.org/10.1109/ACCESS.2022.3217528>).

In order to obtain the two last independent columns of $\mathbf{Q}$, vector fields $\mathbf{h}_1(\mathbf{x}) = [\mathbf{f}_0, [\mathbf{f}_0, \mathbf{g}_1]]$ and $\mathbf{h}_2(\mathbf{x}) = [\mathbf{f}_0, [\mathbf{f}_0, \mathbf{g}_2]]$ are proposed

$$\mathbf{h}_1(\mathbf{x}) = \left[b \frac{\partial f_2}{\partial x_3}(\mathbf{x}), y_1(\mathbf{x}), y_2(\mathbf{x}), b \frac{\partial f_5}{\partial x_3}(\mathbf{x}), y_3(\mathbf{x}), y_4(\mathbf{x}) \right] \quad (\text{C.22})$$

$$\mathbf{h}_2(\mathbf{x}) = \left[b \frac{\partial f_2}{\partial x_6}(\mathbf{x}), y_5(\mathbf{x}), y_6(\mathbf{x}), b \frac{\partial f_5}{\partial x_6}(\mathbf{x}), y_7(\mathbf{x}), y_8(\mathbf{x}) \right] \quad (\text{C.23})$$

with $y_i(\mathbf{x})$ being functions composed of products and sums of the terms of $\mathbf{f}_0(\mathbf{x})$ and its partial derivatives which are not relevant for this analysis. Note that the first and fourth components of $\mathbf{h}_1(\mathbf{x})$ and $\mathbf{h}_2(\mathbf{x})$ correspond to the second and fifth components of $[\mathbf{f}_0, \mathbf{g}_1]$ and $[\mathbf{f}_0, \mathbf{g}_2]$. Plus, $\mathbf{g}_1, \mathbf{g}_2, [\mathbf{f}_0, \mathbf{g}_1]$ and $[\mathbf{f}_0, \mathbf{g}_2]$ have first and fourth null components $\forall \mathbf{x}_0 \in \mathbf{X}$. Therefore, applying to $\mathbf{h}_1(\mathbf{x})$ and $\mathbf{h}_2(\mathbf{x})$ the

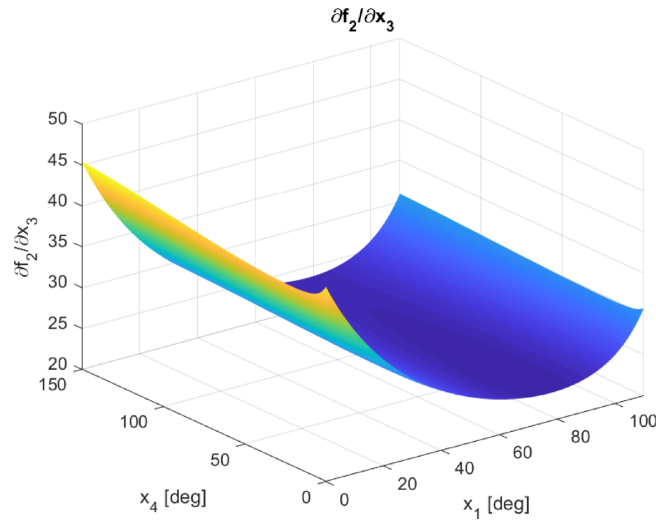


Figure C.3: Computation of $\frac{\partial f_2}{\partial x_3}$ for the full articular range. Source: Author's publication ©David Pont Esteban et al. 2022; published by IEEE (<http://doi.org/10.1109/ACCESS.2022.3217528>).

analysis performed over $[\mathbf{f}_0, \mathbf{g}_1]$ and $[\mathbf{f}_0, \mathbf{g}_2]$, it is proven that the system is locally accessible from every $\mathbf{x}_0 \in \mathbf{X}_c$, which implies that the system is controllable in $\mathbf{X}_c$.

Finally, to study the shoulder controllability in all its range $x_1 \in [0^\circ, 110^\circ]$, no matter the elbow state, the same analysis is applied to the shoulder subsystem (C.24)

$$\dot{\mathbf{x}}_s(t) = \mathbf{f}_s(\mathbf{x}) + \mathbf{g}_1 \cdot u_1(t) \quad (\text{C.24})$$

with $\dot{\mathbf{x}}_s = [\dot{x}_1, \dot{x}_2, \dot{x}_3]$ and $\mathbf{f}_s = [x_2, f_{02}, f_{03}]$. Choosing the vector fields $\mathbf{g}_1$, $[\mathbf{f}_s, \mathbf{g}_1]$ and $[\mathbf{f}_s, [\mathbf{f}_s, \mathbf{g}_1]]$ one obtains the reduced controllability matrix $\mathbf{Q}_s$.

$$\mathbf{Q}_s = \begin{bmatrix} 0 & 0 & b \frac{\partial f_2}{\partial x_3} \\ 0 & -b \frac{\partial f_2}{\partial x_3} & y_9(\mathbf{x}) \\ b & b_L^R & y_{10}(\mathbf{x}) \end{bmatrix} \quad (\text{C.25})$$

Figure C.3 shows that $\frac{\partial f_2}{\partial x_3} \neq 0, \forall x_1 \in [0^\circ, 110^\circ], \forall x_4 \in [0^\circ, 150^\circ]$, what makes $\mathbf{Q}_s$ full rank. Therefore, the shoulder position is controllable beyond $\mathbf{X}_c$, no matter the elbow state.

C.3 Stability analysis of the NDO

Proof 2 Choose the Lyapunov function candidate W as

$$W = \frac{1}{2} \tilde{\mathbf{p}}^\top \tilde{\mathbf{p}} \quad (\text{C.26})$$

Given that W is positive definite, in order to guarantee the asymptotic stability, its derivative must be negative definite, so (C.27) must hold.

$$\dot{W} = \tilde{\mathbf{p}}^\top \dot{\tilde{\mathbf{p}}} = \tilde{\mathbf{p}}^\top (\dot{\mathbf{p}} - \dot{\mathbf{p}}) = \tilde{\mathbf{p}}^\top (\dot{\mathbf{p}} - \mathbf{L} \mathbf{g}_d \tilde{\mathbf{p}}) \quad (\text{C.27})$$

Analyzing (C.27) for each of the six elements of $\mathbf{p}$ and substituting $\mathbf{g}_d = \mathbf{I}_6$ according to the model in (4.13) to (4.16), these errors can be bounded as follows in (C.28)

$$\begin{aligned} \tilde{p}_i (\dot{p}_i - L_{ii} \tilde{p}_i) &\leq \|\tilde{p}_i\| \dot{p}_{i,\max} - L_{ii} \tilde{p}_i^2 \\ &= \|\tilde{p}_i\| (\dot{p}_{i,\max} - L_{ii} \|\tilde{p}_i\|) \end{aligned} \quad (\text{C.28})$$

From (C.27) and (C.28) we have that if (C.29) holds

$$\dot{p}_{i,\max} - L_{ii} \|\tilde{p}_i\| < 0 \quad (\text{C.29})$$

then the nonlinear observer is asymptotically stable. Therefore, $\dot{W} < 0, \forall \|\tilde{p}_i\| > \dot{p}_{i,\max}/L_{ii}$, which means that the lumped disturbance and uncertainty estimation error $\tilde{p}_i$ for each state variable x_i is globally uniformly ultimately bounded and the nonlinear observer is globally asymptotically stable, as W is radially unbounded ($\|\tilde{\mathbf{p}}\| \rightarrow \infty \Rightarrow W \rightarrow \infty$).

C.4 Reachability of the sliding mode

Proof 3 Choose the Lyapunov function candidate V as

$$V = \frac{1}{2} \boldsymbol{\sigma}^\top \boldsymbol{\sigma} \quad (\text{C.30})$$

Note that the sliding manifold $\boldsymbol{\sigma}$ in (4.34), and therefore the Lyapunov function candidate V , depend on the state variable vector $\mathbf{x}(t)$ as well as on the reference vector $\mathbf{x}^d(t)$ and the estimation of lumped disturbance and uncertainty vector $\hat{\mathbf{p}}(t)$, provided by the NDO. Taking the derivative of (C.30)

$$\dot{V} = \boldsymbol{\sigma}^\top \dot{\boldsymbol{\sigma}} = \sigma_s \dot{\sigma}_s + \sigma_e \dot{\sigma}_e \quad (\text{C.31})$$

If $\dot{V}$ is negative definite, i.e. $\dot{V} \leq 0$, $\forall \boldsymbol{\sigma}_0 \neq \mathbf{0}$, the sliding mode is reached ($\boldsymbol{\sigma} = \mathbf{0}$) [211]. V is negative definite if (C.32) and (C.33) hold.

$$\dot{V}_s = \sigma_s \dot{\sigma}_s < 0, \quad \forall \sigma_s \neq 0 \quad (\text{C.32})$$

$$\dot{V}_e = \sigma_e \dot{\sigma}_e < 0, \quad \forall \sigma_e \neq 0 \quad (\text{C.33})$$

Substituting (4.33) and (4.34) in (C.32), it is proven that each control input u_i acts only over the sliding manifold σ_i , so there is no need to design an input to surface decoupling diffeomorphism [212], nor a coordinate transformation in the control space [213], necessary when this condition is not met. In this way, each control input can be designed independently, as if we were working with two SISO systems, and that is why u_1 and u_2 were designed separately one from the other.

Substituting (4.35) and (4.36) in (4.33), then substituting (4.33) in the derivative of (4.34) and substituting that derivative in (C.32) and (C.33), expressions (C.34) and (C.35) are obtained. Accurate design of the control actions and sliding manifolds has allow the cancellation of most of the terms.

$$\sigma_s \left[\sigma_1 (p_1 - \hat{p}_1) + \sigma_2 (p_2 - \hat{p}_2) + p_3 - k_s \psi(\sigma_s) \right] < 0 \quad (\text{C.34})$$

$$\sigma_e \left[\sigma_4 (p_4 - \hat{p}_4) + \sigma_5 (p_5 - \hat{p}_5) + p_6 - k_e \psi(\sigma_e) \right] < 0 \quad (\text{C.35})$$

As analyzed in Appendix C.3, the estimation errors for lumped disturbances $\tilde{p}_i$ and uncertainties maximize when $\dot{\hat{p}}_i = \dot{p}_{i,\max}$. Considering this worst-case situation, substituting $\tilde{p}_i$ by the maximum estimation errors and considering the maximum values for p_3 and p_6 , the following expressions are obtained:

$$\sigma_s \left[\sigma_1 \frac{\dot{p}_{1,\max}}{L_{11}} + \sigma_2 \frac{\dot{p}_{2,\max}}{L_{22}} + p_{3,\max} - k_s \psi(\sigma_s) \right] < 0 \quad (\text{C.36})$$

$$\sigma_e \left[\sigma_4 \frac{\dot{p}_{4,\max}}{L_{44}} + \sigma_5 \frac{\dot{p}_{5,\max}}{L_{55}} + p_{6,\max} - k_e \psi(\sigma_e) \right] < 0 \quad (\text{C.37})$$

Selecting

$$k_s > \sigma_1 \frac{\dot{p}_{1,\max}}{L_{11}} + \sigma_2 \frac{\dot{p}_{2,\max}}{L_{22}} + p_{3,\max} \quad (\text{C.38})$$

$$k_e > \sigma_4 \frac{\dot{p}_{4,\max}}{L_{44}} + \sigma_5 \frac{\dot{p}_{5,\max}}{L_{55}} + p_{6,\max} \quad (\text{C.39})$$

expressions (C.36) and (C.37) turn into

$$\dot{V}_s \leq -c_s |\sigma_s| < 0, \quad \forall \sigma_s \neq 0 \quad (\text{C.40})$$

$$\dot{V}_e \leq -c_e |\sigma_e| < 0, \quad \forall \sigma_e \neq 0 \quad (\text{C.41})$$

with

$$c_s = \begin{cases} k_s - \sigma_1 \frac{\dot{p}_{1,\max}}{L_{11}} - \sigma_2 \frac{\dot{p}_{2,\max}}{L_{22}} - p_{3,\max} & \text{if } \sigma_s > 0 \\ k_s + \sigma_1 \frac{\dot{p}_{1,\max}}{L_{11}} + \sigma_2 \frac{\dot{p}_{2,\max}}{L_{22}} + p_{3,\max} & \text{if } \sigma_s \leq 0 \end{cases} \quad (C.42)$$

$$c_e = \begin{cases} k_e - \sigma_4 \frac{\dot{p}_{4,\max}}{L_{44}} - \sigma_5 \frac{\dot{p}_{5,\max}}{L_{55}} - p_{6,\max} & \text{if } \sigma_e > 0 \\ k_e + \sigma_4 \frac{\dot{p}_{4,\max}}{L_{44}} + \sigma_5 \frac{\dot{p}_{5,\max}}{L_{55}} + p_{6,\max} & \text{if } \sigma_e \leq 0 \end{cases} \quad (C.43)$$

Therefore, it is proven that the control actions (4.36) and (4.37) along with the NDO defined by (4.30) drive the system to the sliding mode. Consequently, the analysis of error dynamics in Section 3.4 is valid.

Remark 3 In equations (C.38) and (C.39), σ_i and L_{ii} are design parameters. However, the thresholds for $\dot{p}_{i,\max}$ and $p_{j,\max}$ are dependent on the system and the application. Although obtaining accurate values for these thresholds might be in many occasions a difficult task or even impossible, establishing conservative thresholds is rarely unfeasible. In this case, the gains will be larger than strictly necessary, but the reaching condition will be guaranteed. Simulation and experimental testing can help to obtain more accurate thresholds.

A more relaxed situation arises if the lumped disturbances and uncertainties evolve slowly ($\dot{\hat{p}}_i \approx 0$), as the estimations converge to the real values ($\tilde{p}_i = p_i - \hat{p}_i = 0$). In this situation, simply choosing $k_s > p_{3,\max}$ and $k_e > p_{6,\max}$, expressions (C.34) and (C.35) simplify to (C.44) and (C.45)

$$\sigma_s [-k_s \psi(\sigma_s)] = -k_s |\sigma_s| < 0, \quad \forall \sigma_s \neq 0 \quad (C.44)$$

$$\sigma_e [-k_e \psi(\sigma_e)] = -k_e |\sigma_e| < 0, \quad \forall \sigma_e \neq 0 \quad (C.45)$$

assuring, under these conditions, that the sliding mode is reached.

Remark 4 In this analysis, the physical parameters of the subject play a role in the control actions (4.35) and (4.36). However, its validity is not dependent on the specific values of the arm parameters, as they are considered in both the model in (4.16) and the control actions. If the parameters are not properly measured, the absolute values of vector $\mathbf{p}(t)$ will increase, but as far as $\mathbf{p}(t)$ is properly estimated by the NDO and the control switching gains k_e and k_s are sufficiently large, the proposed controller will reach the sliding mode and, therefore, the system will be moved to the desired state. This unequivocally means that the controller is robust to user parameters variability, external disturbances and uncertainties.

APPENDIX D

Geometrical method supplementary material

D.1 ExoFlex v2 parameters and actuator limits

Table D.1 presents the geometrical parameters of ExoFlex v2 structure. Table D.2 contains position and speed limitations for the degree of freedom controlled by each actuator.

Table D.1: ExoFlex v2 physical parameters.

Parameter	Value [mm]
k_{AOX}	176.0
k_{AOY}	-45.0
k_{23X}	-70.0
k_{23Y}	155.0
k_{23Z}	132.5
k_{45Y}	45.0
k_{45Z}	-72.0

Table D.2: ExoFlex v2 motor limits associated to the controlled degree of freedom.

Motor	Minimum position	Maximum position	Maximum speed
M1	$l_{c,min} = 0.0$ mm	-	$v_{1,max} = 47.3$ mm/s
M3	$M_{3,min} = 0.0$ deg	$M_{3,max} = 90.0$ deg	$v_{3,min} = 72.2$ deg/s
M4	$M_{4,min} = 0.0$ deg	$M_{4,max} = 85.0$ deg	$v_{4,min} = 22.6$ deg/s
M5	$M_{5,min} = 0.0$ mm	$M_{5,max} = 248.0$ mm	$v_{5,min} = 74.4$ mm/s
M6	$M_{6,min} = 0.0$ mm	$M_{6,max} = 685.0$ mm	$v_{6,min} = 200.0$ mm/s
M7	$M_{7,min} = 0.0$ mm	$M_{7,max} = 170.0$ mm	$v_{7,min} = 200.0$ mm/s

D.2 Forward kinematics

The homogeneous transformation matrices for the kinematics of ExoFlex v2 are the following ones. A diagram of all the reference frames involved in the kinematics is shown in Fig. 4.12.

$${}^{\text{ABS}}\mathbf{T}_0 = \begin{bmatrix} 1 & 0 & 0 & k_{\text{AOX}} \\ 0 & 1 & 0 & k_{\text{AOY}} \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix} \quad (\text{D.1})$$

$${}^0\mathbf{T}_1 = \begin{bmatrix} 1 & 0 & 0 & d_6 \\ 0 & 1 & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix} \quad (\text{D.2})$$

$${}^1\mathbf{T}_2 = \begin{bmatrix} \cos(\theta_3) & \sin(\theta_3) & 0 & 0 \\ -\sin(\theta_3) & \cos(\theta_3) & 0 & 0 \\ 0 & 0 & 1 & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix} \quad (\text{D.3})$$

$${}^2\mathbf{T}_3 = \begin{bmatrix} 1 & 0 & 0 & -|d_7 + k_{23X}| \\ 0 & 1 & 0 & |k_{23Y}| \\ 0 & 0 & 1 & |k_{23Z}| \\ 0 & 0 & 0 & 1 \end{bmatrix} \quad (\text{D.4})$$

$${}^1\mathbf{T}_2 = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & \cos(\theta_4) & -\sin(\theta_4) & 0 \\ 0 & \sin(\theta_4) & \cos(\theta_4) & 0 \\ 0 & 0 & 0 & 1 \end{bmatrix} \quad (\text{D.5})$$

$${}^4\mathbf{T}_5 = \begin{bmatrix} 1 & 0 & 0 & 0 \\ 0 & 1 & 0 & |d_5 + k_{45Y}| \\ 0 & 0 & 1 & -|k_{45Z}| \\ 0 & 0 & 0 & 1 \end{bmatrix} \quad (\text{D.6})$$

D.3 Subset of achievable distances d for a given θ

It is necessary to determine the subset of allowable distances d between shoulder (H) and end-effector (Y) given a certain shoulder aperture angle θ . In this way, if a non-achievable d is imposed, this will be saturated so that M6 and M7 are within their respective ranges.

The range of $\theta \in [0^\circ, 90^\circ]$ is discretized with a resolution of 1 degree, and the range for $d_6 \in [M_{6,\min}, M_{6,\max}]$ is also discretized with a resolution of 5 mm. The resulting distance $d_{7,ij}$ is computed according to equation (4.68) and, if $d_{7,ij} \in [M_{7,\min}, M_{7,\max}]$, the distance d resulting from the configuration determined by θ_i , $d_{6,j}$ and $d_{7,ij}$ is saved as a valid configuration. When all the workspace is explored, a range of achievable distances d as a function of θ is obtained (Fig. D.1). These curves should be used to properly select the distance d for the application of the Algorithm 1 presented in Section 4.3.2. A surface formed by θ , d_6 and d representing the admissibility of the corresponding $d_{7,ij}$ is plotted in Fig. D.2.

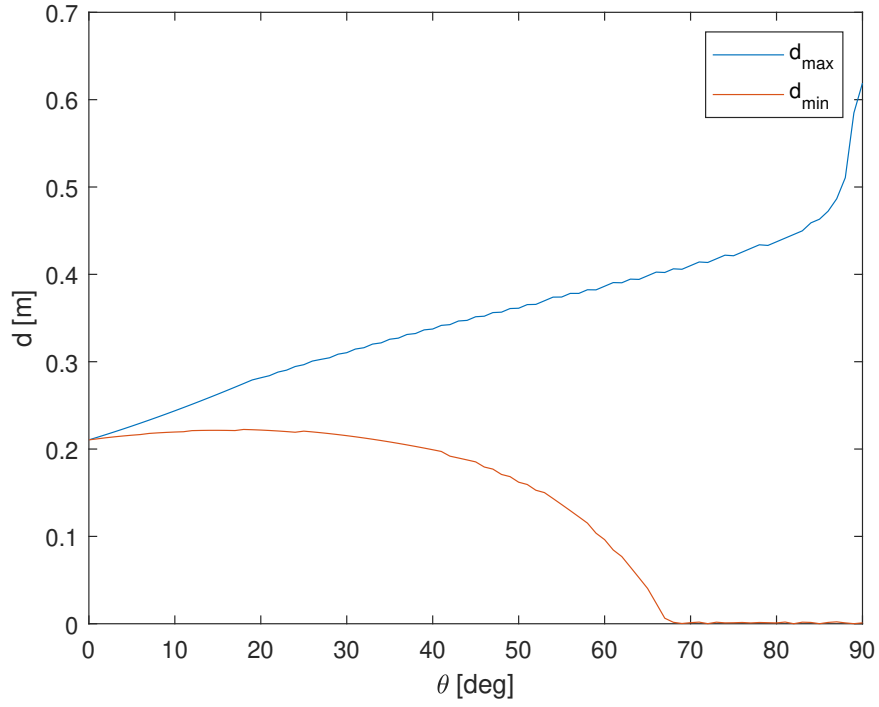


Figure D.1: Feasible distances d as function of the shoulder aperture angle θ . The area between the blue and orange lines contains the admissible distances for θ .

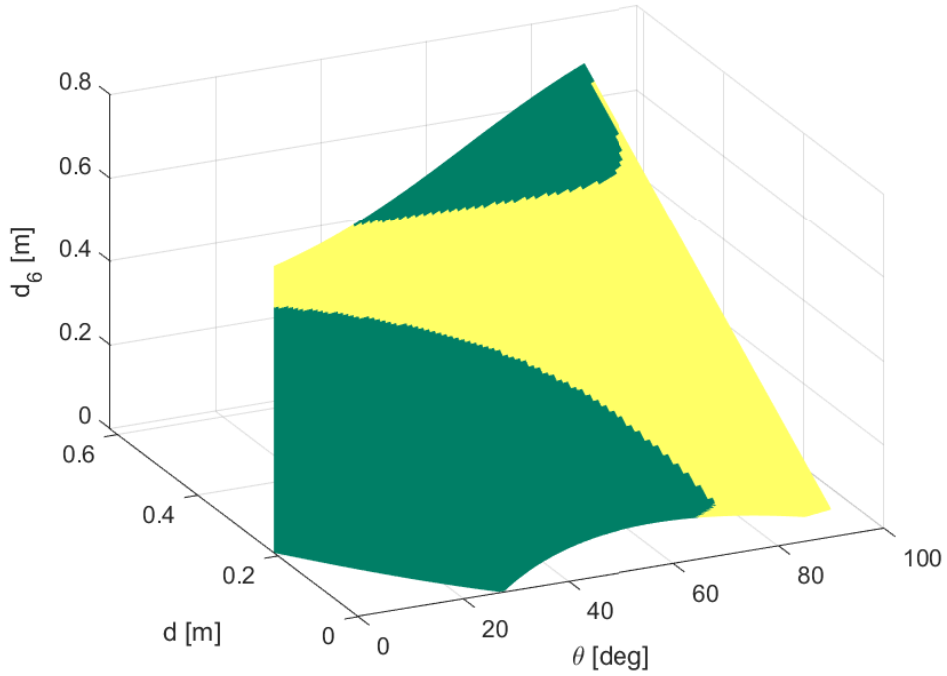


Figure D.2: Admissibility of $d_{7,ij}$ as a function of θ , d_6 and d . The yellow area is the subset of admissible configurations.

Tension sensor design

This appendix introduces the preliminary design of a low-cost force sensor to measure cable tension for cable-driven exoskeletons. The designed sensor is based on the FlexiForce A101 piezoresistive transducer. The mechanical design as well as the designed signal conditioning circuit are presented.

E.1 Mechanical design

The FlexiForce A101 is an ultra-thin and flexible transducer of applied pressure to electrical resistance. Its dimensions are $7.6 \text{ mm} \times 15.6 \text{ mm}$, and it can measure forces up to 44 N (almost 4.5 kg). Motion in cable-driven exoskeletons is generated through tension forces, as cables cannot push. To convert the tension in the cables into measurable compression by the sensor, the following mechanical design was proposed.

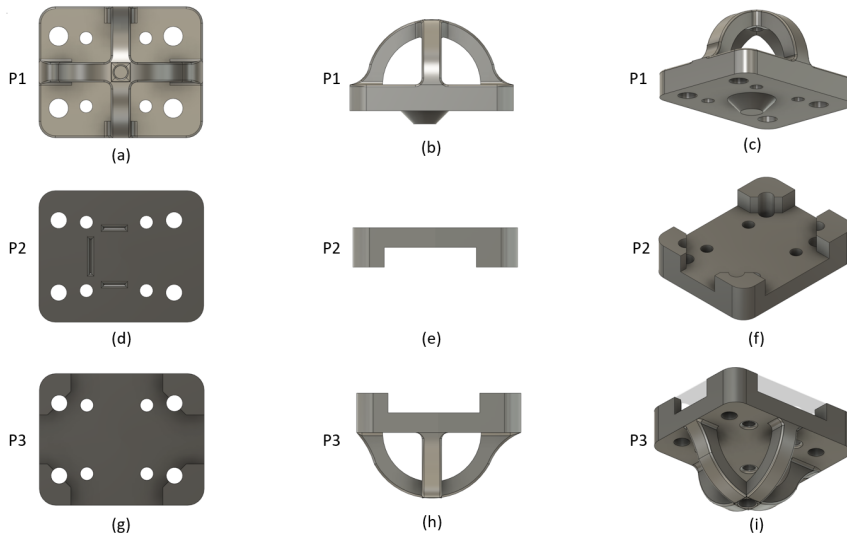


Figure E.1: Cable tension sensor parts. Each row corresponds to a single part. From left to right, top, lateral and 3D views of each part are shown.

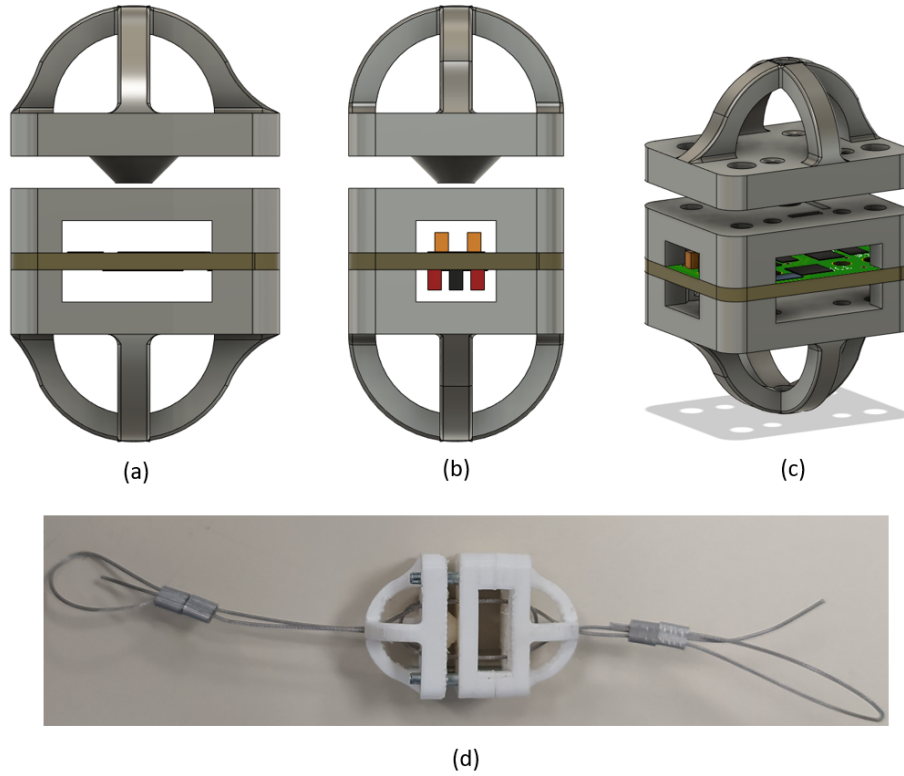


Figure E.2: Sensor assembly: (a) frontal view, (b) lateral view, (c) 3D view with PCB and without cables, (d) 3D view with mounted screws and cable routing.

The design consists of three different 3D-printed parts (Fig. E.1). The assembled sensor is shown in Fig. E.2. The dimensions of the sensor are $20.0 \text{ mm} \times 25.0 \text{ mm} \times 40.5 \text{ mm}$. The A101 is placed between parts P1 and P2, and the signal conditioning PCB is allocated between parts P2 and P3. The transducer is stuck with ultra-thin double sided tape to the flat surface of P2. Three marks which frame the sensor placement area were added to improve the sticking task. The use of a very thin tape is mandatory to reduce the influence of this compressible material in force measurements, as the FlexiForce manufacturer recommends [214]. The protruding conical frustum of P1 was introduced to concentrate the force in the A101 sensing area. Each of the parts have eight holes on their surfaces. Same holes were drilled in the designed PCB, as explained in the following subsection. The four outer holes act as conducts for four screws which serve as a structural support for the sensor, avoiding the misalignment of the parts during operation. Parts P2 and P3 are rigidly attached through the inclusion of a nut on each screw over the P2 surface. No nuts were added to P1 to allow relative displacement with respect to P2 and let normal sensor operation. The four inner holes are used to route the cable arrangement (Fig. E.2d), which transforms tension into compression. Positioning the sensor vertically, the cable is introduced through the hole in the P1's handle-like structure, then enters through one of the cable holes of P1, goes through the aligned holes of P2, the PCB and P3, goes up through the opposing holes of P3, PCB, P2 and P1, and again inside of the hole of the handle-like structure of P1. The two sides of the cable are then coupled together with a compressed aluminum bush, generating a connection interface to the exoskeleton cable. The same process is symmetrically repeated starting on the P3 side using the two remaining conducts.

When a tension force is applied between the two terminals of the sensor, parts P1 and P2 experiment a compression force which is measured by the transducer. The handle-like structures of P1 and P3 were introduced to align the cable terminals with the center of the sensing area, so that force is applied perpendicularly.

E.2 Electronic design

The circuit in Fig. E.3 was designed to convert the resistance variation of the transducer into voltage variation. The manufacturer provides an example conditioning circuit [215], but the requirement of negative voltages motivated the design of an alternative. The proposed circuit consists of a non-inverting amplifying stage, a differential amplifier and an RC filter. The output of the circuit, for frequencies below the cut-off frequency (set to around 160 Hz) is:

$$V_o = V_{\text{ref}} \frac{R}{R_{\text{sens}}} = 1.2 \frac{4120}{R_{\text{sens}}} = \frac{4944}{R_{\text{sens}}} \quad (\text{E.1})$$

To determine the value of R1, that resistor was substituted by a potentiometer. Then, maximum load (44 N) was applied to the transducer and the potentiometer was adjusted to obtain a signal output of 3 V, which is the maximum voltage mea-

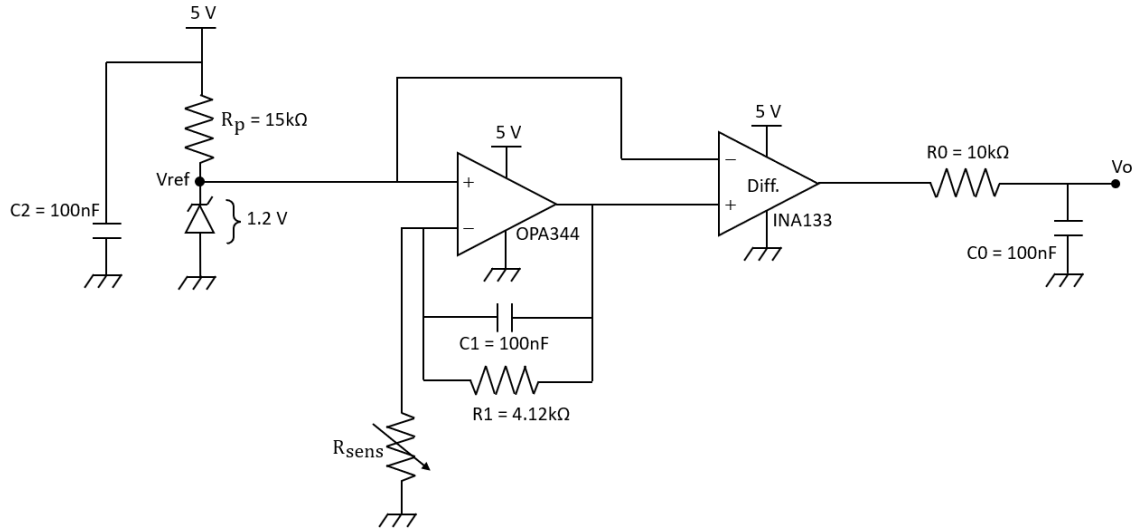


Figure E.3: FlexiForce A101 conditioning circuit schematic.

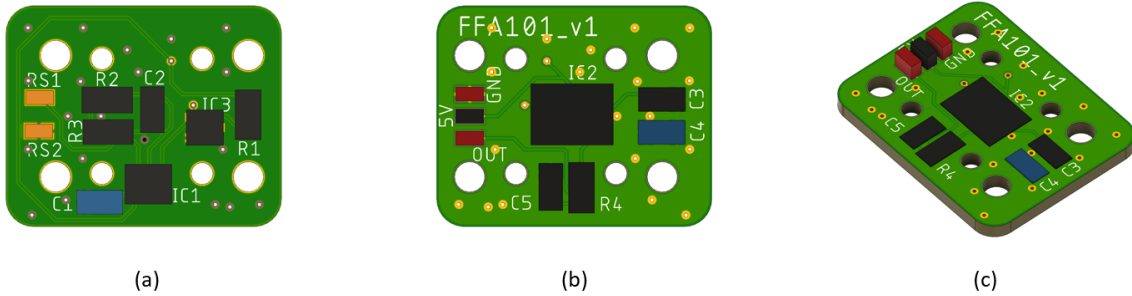


Figure E.4: FlexiForce A101 conditioning circuit PCB: (a) bottom view, (b) top view, (c) 3D view.

surable by the microcontroller's ADCs. The resistance value of the potentiometer was then measured with a multimeter. The closest commercially available value to the measured resistance was obtained through two high-precision resistors in series, which are represented in Fig. E.3 as a single resistor for simplicity. In this way, the resolution of the sensor is maximized, fitting the whole force range into the measurable voltage range.

The electronics were built over a prototyping board to check the behavior, prior to designing the PCB (Fig. E.4). As observed in (Fig. E.2), the length and width of the PCB are equal to the ones of the nylon parts. The inclusion of the holes for the screws and cables in the PCB allowed to obtain a compact design, in which the electronics are tightly integrated.

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