



Control of a battery charger/discharger based on the Sepic/Zeta topology for DC bus regulation

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Resumen

En una microrred de DC que involucra un sistema de almacenamiento basado en baterías, la gestión de la energía primaria es realizada por un cargador/descargador de baterías basado en un convertidor de potencia DC/DC. El cargador/descargador de baterías también se utiliza para regular el voltaje del bus de DC. Uno de los desafíos a nivel de control es regular el voltaje del bus de DC en condiciones de carga y descarga de la batería, pero también bajo diferentes relaciones entre los voltajes de la batería y del bus. Por ello, en este trabajo de investigación se propone un cargador/descargador de baterías basado en el convertidor Sepic/Zeta y controladores adaptativos, que proporcionan flujo de corriente bidireccional, estabilidad del voltaje del bus y características eléctricas satisfactorias. La principal ventaja del sistema de control propuesto es la capacidad de adaptar los parámetros del controlador a cualquier condición de operación, lo que proporciona una solución general para interactuar cualquier batería con cualquier voltaje del bus. Se proponen tres métodos para adaptar los parámetros del controlador, y se evalúa su rendimiento. Este estudio se centra en el procedimiento de diseño tanto del convertidor de potencia como del controlador, donde se realiza un análisis matemático detallado para garantizar el rendimiento y la estabilidad del sistema. Finalmente, la solución propuesta se valida utilizando un prototipo experimental y un caso de aplicación práctico.

Palabras clave— Microrred; Convertidor Sepic/Zeta; Cargador/descargador de baterías; Regulación de bus DC; Control adaptativo.

Abstract

In a dc microgrid that involve a battery storage system, the primary energy management is performed by a battery charger/discharger based on a dc/dc power converter. Moreover, the battery charger/discharger is also used to regulate the voltage of the dc bus. One of the challenges at the control level is to regulate the DC bus voltage under battery charge and discharge conditions, but also under different relations between the battery and bus voltages. For this reason, this research work proposes a battery charger/discharger based on the Sepic/Zeta converter and adaptive controllers, which provides bidirectional current flow, stable bus voltage and satisfactory electrical characteristics. The main advantage of the proposed control system is the capability to adapt the controller parameters to any operation condition, which provides a general solution to interface any battery to any bus voltage. In addition, three methods are proposed to adapt the controller parameters, and its performance is evaluated. This study is focused on the design procedure of both the power converter and the controller, where a detailed mathematical analysis is conducted to ensure the system performance and stability. Finally, the proposed solution is validated using an experimental prototype and a practical application case.

Keywords— Microgrid; Sepic/Zeta converter; Battery charger/discharger; DC bus regulation; Adaptive controller.

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Chapter 1

Introduction

There is a fast growth of renewable energy installations due to increased energy demand and the need to reduce the use of fossil fuels [1]. However, the intermittent nature of renewable energy, for instance, solar and wind power, does not allow to provide energy continuously during the 24 hours of the day, hence it is necessary to store it to use later. For this case, an energy storage system (ESS) based on batteries (BESS) [2, 3, 4], is widely used which, due to its high energy density, can store the energy of the distributed energy sources (DES) and deliver it to the loads, thus forming a microgrid (MG) [5]. These increase the reliability and controllability of the electrical system [6].

In a MG, disturbances associated with changes in loads, component failures, or operational adjustments may occur, affecting the stability of the MG. There are several types of stability problems in MGs. In [7] they are classified into Control and Supply System Stability and Power Balance, in turn these can be due to small or large disturbances, and short or long term phenomena. Another type of stability is DC bus voltage stability, which consists of maintaining the voltage of the dc-link capacitor. This work contributes to this type of stability, under small perturbations and a short term.

The power fluctuations on the MGs due to generation and demand changes should be compensated, and it is here where the BESS plays an important role. The interface between the BESS and the system bus is a power converter, which manages the energy missing or supports the sources concerning demand [8]. If the bus is DC, then a DC/DC power converter with bidirectional power flow must be used to guarantee the polarity of both battery and DC bus and offers conversion rates that allow step-down and step-up of the voltage. Different power converter topologies, such as buck/boost, Sepic, flyback, and Zeta, have been used to connect BESS to DC buses. Those power converters aim to charge and discharge the BESS and impose characteristics such as voltage, current ripple, current increment rate, and more. For example, a regulated DC bus voltage facilitates the source's power

extraction and guarantees high-quality power to the loads along with the low current ripple. Likewise, low current ripple and low current increment rates increase the battery lifetime.

Many works about bidirectional converters are found during the state-of-the-art review. For example, [9] is a survey from the point of view of topology and the control scheme. That work divides the converters into non-isolated and isolated bidirectional converters, but the analyses do not show their use as an interface between BESS and DC buses.

The most common bidirectional converter is the buck/boost type. It is used for MPPT (Maximum Power Point Tracker) in solar systems [10] and DC bus regulation in MGs with a sliding-mode controller (SMC) [11, 12], with a classical PI controller [13], with a model predictive control (MPC) using a dynamic reference [14]. Similarly, such a solution was also implemented with a hybrid energy storage system using a sliding-mode current controller on a 5th-order averaged model as reported in [15]. Moreover, the work reported in [16] develops a high-gain sliding controller acting on a supercapacitor, and in [17] it is discussed a sliding-mode controller of the Electro-Mechanical Actuator (EMAs) current of an aircraft powered by a supercapacitor. Other works have been based on sliding-mode current controllers operating with supervisory strategies to act on the state-of-charge (SOC) of the battery, its charge/discharge current, the regulation of the high-voltage side, or to maximize the generator current [18, 19, 20, 21, 22, 23]. Besides, some variants exist; first, [24] is a bidirectional converter with coupled inductor used to charge/discharge an EV (Electrical Vehicle) battery and regulate the DC bus voltage together with fuzzy control. The second is a variant integrated by two buck/boost converters in cascade and an auxiliary capacitor that reduces the output current ripple and increases the efficiency, which uses a PI controller to regulate the output voltage in a photovoltaic and energy storage system [25]. The third is a bidirectional converter for the DC bus and battery bank interface in a UPS (Uninterruptible Power Supply) with an auxiliary commutation circuit [26, 27]. The control system is formed by three independent controllers (classical P, PI, and average current), which must control the charge of the battery bank, regulate the output voltage, and ensure the unitary input power factor.

On the other hand, Cuk bidirectional converters have been used to charge/discharge batteries with a PI controller for average inductor current in [28], and DC bus voltage regulation with limiters of current and voltage in [29].

Moreover, the bidirectional flyback dc-dc converter is frequently used to interface BESS and DC links. Ref. [30] presented a multiport flyback with a simple hysteresis voltage control to regulate the DC link voltage of an EV, and [31] shows a flyback converter for mobile phone chargers with current and voltage regulation using a classical PI controller. Besides, [32, 33] use the bidirectional flyback converter as a Li-ion battery cell current equalizer.

This research work is focused on the Sepic/Zeta bidirectional converter and its use as a power manager between a BESS and the DC bus of a MG. This converter is used in different applications, for example, to interface the PEVs (plug-in electric vehicles) to the charger station and to charge/discharge an ESD (Energy Storage Device) in a MG. Nevertheless, there are few works on the Sepic/Zeta bidirectional converter as charger/discharger of BESS.

Ref. [34] proposed an ESDs charging/discharging system designed to regulate the voltage of a DC bus using a SMC controller on a switched model of a Sepic/Zeta bidirectional converter. The DC bus voltage error and the currents of both converter inductors form the switching function, guaranteeing zero in the state-stable voltage error and global stability of the system. One of the three switching function parameters is calculated online for stability purposes. The other two are calculated offline according to closed-loop dynamic to provide null voltage overshoot and a fast bus voltage restoration after transients.

Ref. [35] presented a DC-DC bidirectional Sepic-Zeta converter with a model on averaged state space. The converter manages the power exchange between a BESS and a DC bus, and it is controlled by an adaptive SMC to improve the transient and steady-state behavior. The DC bus voltage error and the currents of both the battery side capacitor and the DC bus side capacitor form the switching function. The results are compared against a classical PID controller, and the overall dynamic performances have been significantly improved, guaranteeing the system's global stability.

Ref. [36] presented a bidirectional converter in a PEV working in three modes: plug-in charging, propulsion, and regenerative. The control system consists of three classical PI controllers: two to regulate the input and battery currents and one to control the motor voltage.

Ref. [37] showed a charging/discharging system for a BESS using a bidirectional DC-DC Sepic/Zeta converter with a PID (proportional-integral-derivative) controller for voltage management of a DC bus. The converter's state-space averaged model is presented to control the switching pulses in both Sepic and Zeta modes. The classical Ziegler-Nichols method is used to design the controller, and the PSO (Particle Swarm Optimization) tool is used to tune it. However, this work does not use robust adaptative control or perform experimental validation.

Some works show a variant of the Sepic/Zeta bidirectional converter with advantages like low switching and conduction losses [38, 39], or low-ripple currents [40]. However, those works do not evidence the application as the interface between a BESS and DC bus with an adaptive control scheme.

This research works proposes a state feedback Linear Quadratic Gaussian (LQG) adaptative strategy, which aims to regulate the MG's DC bus voltage in any system operating condition. For example, the battery can be in charge, discharge, or

standby mode (when there is no exchanged energy between the battery and the DC bus), and the Sepic/Zeta bidirectional converter can be in step-down (buck mode), step-up (boost mode), or with a unitary conversion relation (unitary gain mode). The strategy above is applied to the charger/discharger Sepic/Zeta bidirectional converter, which is modeled with an averaged state-space small-signal model. This model is necessary since the LQG controller requires a linearized model. For this, three control blocks are used: linear quadratic integral (LQI) controller to regulate the DC bus voltage, optimal observer to estimate the system states, and adaptive law to adjust the LQI parameters. The experimental implementation only requires the measurement of two voltages: DC bus and battery.

Works [34] and [35] are similar to the one presented here; however, [34] is a system with a switched model and a sliding-mode controller, and its implementation requires the measurement of two voltages (DC bus and battery) and the currents of both inductors. Similarly, [35] is a system with a state-space averaged modeling and an adaptive SMC. Its implementation requires the measurement of two voltages (DC bus and battery) and four currents: DC bus current, battery current, and capacitors current of both sides of the DC bus and the battery.

Several important contributions were made in this research work. First, a design procedure of a charger/discharger that allows the proper selection of its elements according to specific design criteria. Second, three alternatives are proposed to adaptively adjust the controller parameters, thus allowing the charger/discharger to operate under any operating condition of the batteries (charge, discharge, standby), and the converter (buck, boost, unitary gain). Finally, the experimental implementation of a case to validate the mathematical models, the design procedure, and the proposed control strategy. Those contributions were published in the article [41] (journal of rank Q1).

To achieve the proposed objectives, this research work consists of 6 chapters. Chapter 2 shows the mathematical modeling of the control-oriented system. Chapter 3 provides a design procedure of the charger/discharger, based on design equations and performance indicators. Chapter 4 shows in detail the proposed control strategy and the simulations of the case application. Chapter 5 shows the implementation of a prototype and the experimental results. Finally the Chapter 6 presents the conclusions and future work.

Chapter 2

Modeling of the power stage

This chapter describes the power stage circuit and the mathematical model of charger/discharger, which considers powers losses in the converter elements. The models include a switching non-linear model and a average linear model.

2.1 Circuital interface

The battery charger/discharger is based on the bidirectional Sepic/Zeta topology, which is reported in figure 2.1. Such a topology was selected to enable the connection of batteries with any voltage relation with the microgrid bus; thus, the battery charger/discharger could be used to connect any commercial battery to any microgrid bus. Figure 2.1 models the battery using a voltage source V_b , while the dc bus (of the microgrid) is modeled using the bus capacitance C_{dc} and the bus current i_o . In such a DC bus model, the bus capacitance is formed by the output capacitances of the sources and the input capacitances of the loads connected to the bus. Similarly, the bus current is the result of the current balance in the bus, hence i_o could be positive (loads require higher current than the current provided by the sources), negative (loads require lower current than the current provided by the sources), or zero (loads require the same current provided by the sources). Finally, the charger/discharger must to regulate the bus voltage V_{dc} , which is needed to provide a safe operation condition to both the sources and loads forming the microgrid.

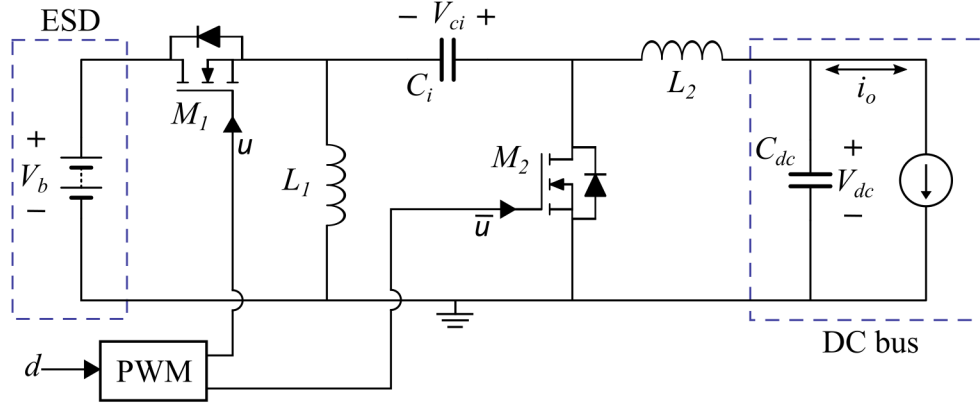


Figure 2.1: Sepic/Zeta based circuitual interface for bidirectional power flow

The Sepic/Zeta bidirectional topology is formed by two MOSFETs (M_1 and M_2) which are complementary activated, two inductors L_1 and L_2 , and an intermediate capacitor C_i . Finally, the circuitual interface includes a PWM circuit to produce the complementary activation signals of the MOSFETs (u and $\bar{u} = 1 - u$) from the desired duty cycle d .

2.2 Control-oriented model

This subsection provides a realistic control-oriented model of the circuitual interface, which takes into account the parasitic resistances in both the MOSFETs and inductors as depicted in figure 2.2.

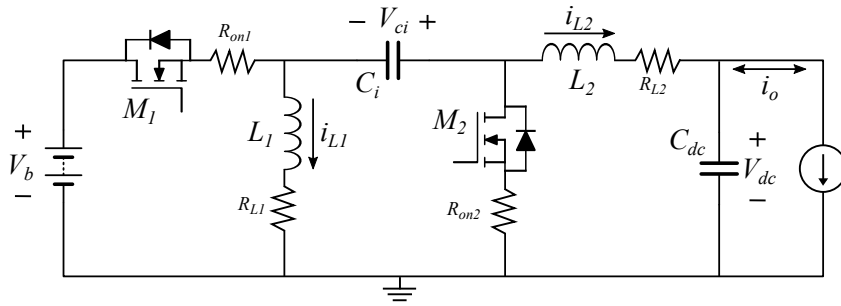


Figure 2.2: Charger/discharger circuit including parasitic losses

The first step to develop a usable control-oriented model is to obtain the differential equations of both inductor currents and capacitor voltages for each possible state of the activation signal u . The circuit obtained from $u = 1$ is depicted in figure 2.3, where the differential equations are:

$$\frac{di_{L1}}{dt} = \frac{V_b - (i_{L1} + i_{L2}) \cdot R_{on1} - i_{L1} \cdot R_{L1}}{L_1} \quad (2.1)$$

$$\frac{di_{L2}}{dt} = \frac{V_{ci} - (i_{L1} + i_{L2}) \cdot R_{on1} + V_b - V_{dc} - i_{L2} \cdot R_{L2}}{L_2} \quad (2.2)$$

$$\frac{dV_{ci}}{dt} = \frac{-i_{L2}}{C_i} \quad (2.3)$$

$$\frac{dV_{dc}}{dt} = \frac{i_{L2} - i_o}{C_{dc}} \quad (2.4)$$

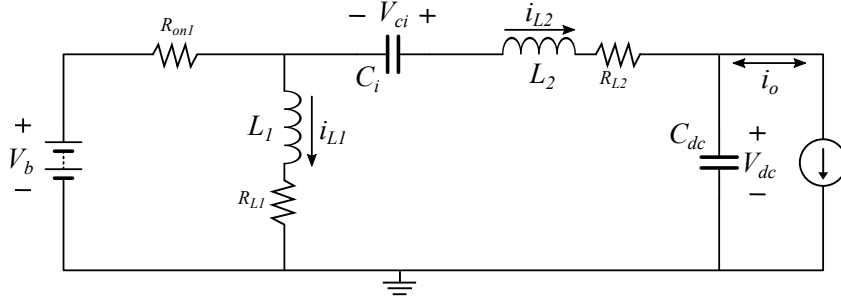


Figure 2.3: Charger/discharger operation for $u = 1$

Similarly, the circuit obtained from $u = 0$ is depicted in figure 2.4, where the differential equations are:

$$\frac{di_{L1}}{dt} = \frac{-V_{ci} - (i_{L1} + i_{L2}) \cdot R_{on2} - i_{L1} \cdot R_{L1}}{L_1} \quad (2.5)$$

$$\frac{di_{L2}}{dt} = \frac{-V_{dc} - (i_{L1} + i_{L2}) \cdot R_{on2} - i_{L2} \cdot R_{L2}}{L_2} \quad (2.6)$$

$$\frac{dV_{ci}}{dt} = \frac{i_{L1}}{C_i} \quad (2.7)$$

$$\frac{dV_{dc}}{dt} = \frac{i_{L2} - i_o}{C_{dc}} \quad (2.8)$$

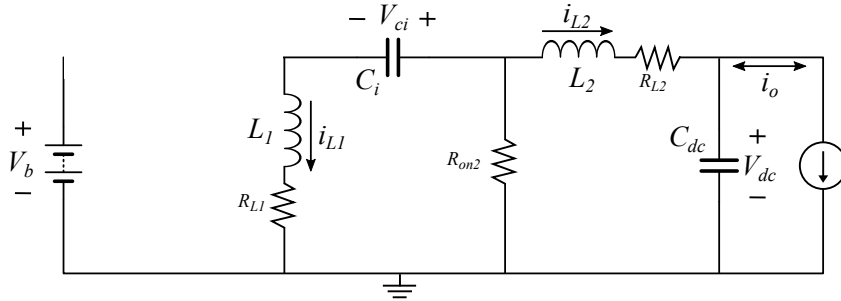


Figure 2.4: Charger/discharger operation for $u = 0$

From the previous differential equations it is formed the switched model, which describes the dynamic behavior of the inductor currents and capacitor voltages depending on the activation signal u :

$$\frac{di_{L1}}{dt} = \frac{(V_b - (i_{L1} + i_{L2}) \cdot R_{on1}) \cdot u + (-V_{ci} - (i_{L1} + i_{L2}) \cdot R_{on2}) \cdot \bar{u} - i_{L1} \cdot R_{L1}}{L_1} \quad (2.9)$$

$$\frac{di_{L2}}{dt} = \frac{(V_{ci} - (i_{L1} + i_{L2}) \cdot R_{on1} + V_b) \cdot u - (i_{L1} + i_{L2}) \cdot R_{on2} \bar{u} - V_{dc} - i_{L2} \cdot R_{L2}}{L_2} \quad (2.10)$$

$$\frac{dV_{ci}}{dt} = \frac{i_{L1} \cdot \bar{u} - i_{L2} \cdot u}{C_i} \quad (2.11)$$

$$\frac{dV_{dc}}{dt} = \frac{i_{L2} - i_o}{C_{dc}} \quad (2.12)$$

The previous model accurately describes the behavior of the bidirectional Sepic/Zeta structure, but this model does not includes the PWM circuit. Considering that duty cycle corresponds to the averaged value of the activation signal u inside the switching period T_{sw} , as given in (2.13), a new modeling approach including the PWM operation is obtained by averaging the switched differential equations within the switching period.

$$d = \frac{1}{T_{sw}} \cdot \int_0^{T_{sw}} u \, dt \quad (2.13)$$

Such an averaged model is obtained by replacing the binary signal u with the continuous duty cycle d . In as much as the MOSFETs support the same current, the ON resistances of the MOSFETs are considered with the same value ($R_{on1} = R_{on2} = R_{on}$), which leads to the following averaged state-space system:

$$\begin{bmatrix} \dot{x}_1 \\ \dot{x}_2 \\ \dot{x}_3 \\ \dot{x}_4 \end{bmatrix} = \begin{bmatrix} -\frac{R_{on}+R_{L1}}{L_1} & -\frac{R_{on}}{L_1} & \frac{-(1-d)}{L_1} & 0 \\ -\frac{R_{on}}{L_2} & -\frac{R_{on}+R_{L2}}{L_2} & \frac{d}{L_2} & \frac{-1}{L_2} \\ \frac{(1-d)}{C_i} & \frac{-d}{C_i} & 0 & 0 \\ 0 & \frac{1}{C_{dc}} & 0 & 0 \end{bmatrix} \cdot \begin{bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{bmatrix} + \begin{bmatrix} \frac{V_b}{L_1} \\ \frac{V_b}{L_2} \\ 0 \\ 0 \end{bmatrix} \cdot d + \begin{bmatrix} 0 \\ 0 \\ 0 \\ \frac{-1}{C_{dc}} \end{bmatrix} \cdot i_o \quad (2.14)$$

$$y = \begin{bmatrix} 0 & 0 & 0 & 1 \end{bmatrix} \cdot \begin{bmatrix} x_1 \\ x_2 \\ x_3 \\ x_4 \end{bmatrix} \quad (2.15)$$

where the states are:

$$x_1 = i_{L1} \quad x_2 = i_{L2} \quad x_3 = V_{ci} \quad x_4 = V_{dc} \quad (2.16)$$

Figure 2.5 presents a comparison between the previous averaged non-linear model and the circuital (switched) simulation performed in MATLAB. The magnitude and phase of the bode diagrams confirm the high accuracy provided by the averaged model in the reproduction of the circuit behavior. Moreover, the right-side of the figure shows the dynamic response of both the model and the circuit to a step from $d = 0.5$ to $d = 0.6$, where the main difference is the ripple oscillations present in the circuital (switched) simulation, which are not present in the averaged model. Therefore, for control-oriented purposes, the averaged model provides a satisfactory representation of the real circuit, including the parasitic losses.

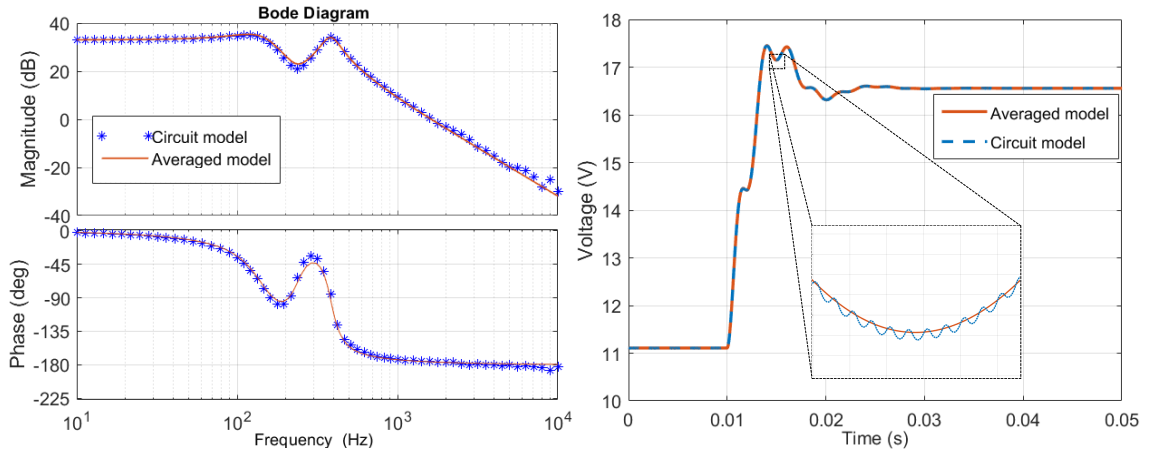


Figure 2.5: Bode plot and step response of Sepic/Zeta converter models

In steady-state conditions the derivatives of the state-space model are equal to zero, which leads to the steady-state conditions reported in (2.17)-(2.20), where d_e corresponds to the duty cycle at a specific operating point; those equations consider equal the ON resistances of both MOSFETs, which is realistic since both MOSFETs support the same peak current.

$$V_{dc_e} = V_b \cdot \left(\frac{d_e}{1 - d_e} \right) - i_o \cdot \left(\frac{R_{L1} \cdot d_e^2}{(1 - d_e)^2} + R_{L2} + \frac{R_{on}}{(1 - d_e)^2} \right) \quad (2.17)$$

$$V_{ci_e} = V_b \cdot \left(\frac{d_e}{1 - d_e} \right) - i_o \cdot \left(\frac{R_{L1} \cdot d_e + R_{on}}{(1 - d_e)^2} \right) \quad (2.18)$$

$$i_{L1e} = \frac{i_{L2e} \cdot d_e}{1 - d_e} \quad (2.19)$$

$$i_{L2e} = i_o \quad (2.20)$$

It is observed that the averaged model is a non-linear system (2.14)-(2.16), thus it can be used to design linear controllers by obtaining the small-signal representation given in (2.21) and (2.22), where A_m (state matrix), B_m (input matrix) and C_m (output matrix) are the matrices of (2.14) evaluated at the steady-state conditions, C_m is the vector of (2.15) and $D_m = 0$. Finally, $\tilde{x}$, $\tilde{y}$ and $\tilde{d}$ are the small-signal changes on the state vector, output and duty cycle, respectively.

$$\dot{\tilde{x}} = \underbrace{\begin{bmatrix} -\frac{R_{on}+R_{L1}}{L_1} & -\frac{R_{on}}{L_1} & \frac{-(1-d_e)}{L_1} & 0 \\ -\frac{R_{on}}{L_2} & -\frac{R_{on}+R_{L2}}{L_2} & \frac{d_e}{L_2} & \frac{-1}{L_2} \\ \frac{(1-d_e)}{C_i} & \frac{-d_e}{C_i} & 0 & 0 \\ 0 & \frac{1}{C_{dc}} & 0 & 0 \end{bmatrix}}_{A_m} \cdot \tilde{x} + \underbrace{\begin{bmatrix} \frac{V_b+V_{ci_e}}{L_1} \\ \frac{V_b+V_{ci_e}}{L_2} \\ \frac{-i_{L1e}-i_{L2e}}{C_i} \\ 0 \end{bmatrix}}_{B_m} \cdot \tilde{d} \quad (2.21)$$

$$\tilde{y} = \underbrace{\begin{bmatrix} 0 & 0 & 0 & 1 \end{bmatrix}}_{C_m} \cdot \tilde{x} \quad (2.22)$$

where the states are:

$$\tilde{x}_1 = \tilde{i}_{L1} \quad \tilde{x}_2 = \tilde{i}_{L2} \quad \tilde{x}_3 = \tilde{V}_{ci} \quad \tilde{x}_4 = \tilde{V}_{dc}$$

The mathematical models obtained by considering power losses in the converter elements allow the design of the control strategy to regulate the output voltage with greater precision. With the equations obtained in the mathematical modeling, in the next chapter design equations for the power stage are proposed.

Chapter 3

Design of the power stage

The adequate selection of the elements of the converter is crucial, and when considering a system that works in a wide range of operation, the worst case scenario of the variables must be considered. In this way, compliance with the design criteria is guaranteed under any specified operating condition. This chapter reports the design procedure of the electrical elements of the charger/discharger converter. The design is performed based on the following criteria:

- Range of dc bus voltage
- Maximum voltage ripple accepted in the dc bus
- Maximum power exchange with the dc bus
- Range of the battery voltage
- Maximum current ripple accepted in the converter inductors
- Minimum efficiency accepted for the charger/discharger

From the steady-state equations (2.17)-(2.20), the averaged input current is calculated:

$$i_{int} = (i_{L1e} + i_{L2e}) \cdot d = \left(\frac{d}{1-d} \right) \cdot i_o \quad (3.1)$$

Then, the electrical efficiencies in discharge (η_d) and charge (η_c) conditions are calculated from equations (2.17) and (3.1), as follow:

$$\eta_d = 1 - \frac{i_o}{V_b} \cdot \left(R_{L1} \cdot \frac{d}{(1-d)} + R_{L2} \cdot \frac{(1-d)}{d} + R_{on} \cdot \frac{1}{d \cdot (1-d)} \right) \quad (3.2)$$

$$\eta_c = \frac{1}{1 - \frac{i_o}{V_b} \cdot \left(R_{L1} \cdot \frac{d}{(1-d)} + R_{L2} \cdot \frac{(1-d)}{d} + R_{on} \cdot \frac{1}{d \cdot (1-d)} \right)} \quad (3.3)$$

The previous equations relate the efficiency of the converter, the operation conditions and the parasitic resistances in both inductors and MOSFETs. From those equations, it is observed that increasing the bus current i_o , and operating with a bus voltage lower than the battery voltage, the converter efficiency is decreased. Therefore, the worst-case scenario for the converter efficiency is given in (3.4):

$$\eta_{dMin} = 1 - \frac{i_{oMax}}{V_{bMin}} \cdot \left(R_{L1} \cdot \frac{d}{(1-d)} + R_{L2} \cdot \frac{(1-d)}{d} + R_{on} \cdot \frac{1}{d \cdot (1-d)} \right) \quad (3.4)$$

Based on the previous equation, the maximum parasitic resistance allowed in the inductors is given in (3.5). Such an expression considers $R_{L1} = R_{L2}$, which is fair since both inductors support similar currents. In conclusion, the inductors must be selected with a maximum parasitic resistances R_{LMax} to ensure the desired efficiency.

$$R_{LMax} = \left((1 - \eta_{req}) \cdot \frac{V_{bMin}}{i_{oMax}} - R_{on} \cdot \frac{1}{d \cdot (1-d)} \right) \cdot \left(\frac{d \cdot (1-d)}{1 - 2d + 2d^2} \right) \quad (3.5)$$

The parasitic resistance of the MOSFETs R_{on} depend on the semiconductor characteristics. For this power interface, the maximum current and voltage that must be supported by both MOSFETs are given in (3.6) and (3.7), which are used to select the commercial MOSFETs for the charger/discharger, thus also helping to obtain the R_{on} value from the manufacturer datasheet.

$$i_{Mmax} = i_{L1max} + i_{L2max} \quad (3.6)$$

$$V_{Mmax} = V_{bmax} + V_{cimax} \quad (3.7)$$

The peak values of the ripples in the inductor currents and capacitor voltages are obtained from the differential equations (2.1)-(2.2) and (2.3)-(2.4), respectively. Such a calculation is performed by applying the volt-second balance in the inductors, and the charge balance in the capacitors [34]. Since the differential equations for i_{L1} , i_{L2} and V_{ci} are different in $u = 0$ and $u = 1$, those ripples correspond to triangular waveforms, thus the peak values are calculated from (2.1), (2.2) and (2.3) for the first section of the switching period, which has a duration of $d \cdot T_{sw}$: the peak ripple magnitude is calculated by multiplying the derivative of the state by $d \cdot T_{sw}$, and

dividing such a result by 2. On the other hand, the differential equations for V_{dc} in $u = 0$ and $u = 1$ are equal, hence the ripple has a second-order waveform: in this case the voltage ripple is calculated by integrating the capacitor ripple ($\Delta i_{L2} - i_o$) within the complete switching period $T_{sw} = 1/F$, where F is the switching frequency. The resulting ripple equations are:

$$\Delta i_{L1} = \frac{(V_b - (i_{L1e} + i_{L2e}) \cdot R_{on} - i_{L1e} \cdot R_{L1}) \cdot d}{2 \cdot L_1 \cdot F} \quad (3.8)$$

$$\Delta i_{L2} = \frac{(V_{cie} + V_b - V_{dce} - (i_{L1e} + i_{L2e}) \cdot R_{on} - i_{L2e} \cdot R_{L2}) \cdot d}{2 \cdot L_2 \cdot F} \quad (3.9)$$

$$\Delta V_{ci} = \frac{i_o \cdot d}{2 \cdot C_i \cdot F} \quad (3.10)$$

$$\Delta V_{dc} = \frac{\Delta i_{L2}}{8 \cdot F \cdot C_{dc}} = \frac{(V_{cie} + V_b - V_{dce} - (i_{L1e} + i_{L2e}) \cdot R_{on} - i_{L2e} \cdot R_{L2}) \cdot d}{16 \cdot C_{dc} \cdot L_2 \cdot F^2} \quad (3.11)$$

As in the efficiency criterion, the worst-case scenario is used to calculate the passive elements. From the ripple equations (3.8)-(3.11) it is observed that the parasitic resistances (R_{on} , R_{L1} and R_{L2}) reduce the ripple magnitude, thus the worst-case scenario corresponds to a loss-less device, which leads to the following minimum inductance and capacitance values:

$$L_{1Min} = \frac{V_{bMax} \cdot d}{2 \cdot \Delta i_{L1} \cdot F} \quad (3.12)$$

$$L_{2Min} = \frac{V_{bMax} \cdot d}{2 \cdot \Delta i_{L2} \cdot F} \quad (3.13)$$

$$C_{iMin} = \frac{i_o \cdot d}{2 \cdot \Delta V_{ci} \cdot F} \quad (3.14)$$

$$C_{dcMin} = \frac{V_{bMax} \cdot d}{16 \cdot \Delta i_{L2} \cdot L_2 \cdot F^2} \quad (3.15)$$

Finally, the elements selection must to include a safe margin to account for manufacturing tolerances. In this work it is recommended to adopt a 50 % safe margin, but any other margin can be used:

$$L_1 > 1.5 \cdot L_{1Min} \quad (3.16)$$

$$L_2 > 1.5 \cdot L_{2Min} \quad (3.17)$$

$$R_L < 0.9 \cdot R_{LMax} \quad (3.18)$$

It must be noted that a large inductor could also exhibit large parasitic resistances, thus degrading the efficiency. Therefore, it must be found a proper balance between inductance size and losses of the commercial devices. Then, using the selected inductances, the converter capacitances are selected using equations (3.14) and (3.15), as follow:

$$C_{dc} > 1.5 \cdot C_{dcMin} \quad (3.19)$$

$$C_i > 1.5 \cdot C_{iMin} \quad (3.20)$$

Finally, figure 3.1 summarizes the design process using a systematic flowchart.

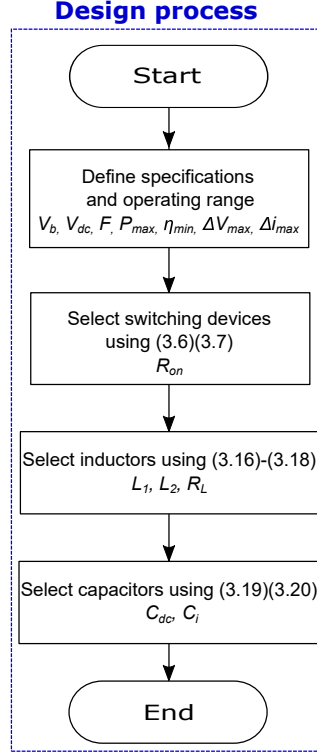


Figure 3.1: Flowchart of the design process

In short, considering the electrical efficiency of the charger/discharger was a wise option in the approach to the design procedure, since it guarantees the best use of energy in accordance with the criteria set. In the next chapter the control strategy for voltage regulation is proposed, where the mathematical modeling obtained in chapter 2 is considered, and it is evaluated in simulations for a case study applying the proposed design procedure.

Chapter 4

Adaptive LQG voltage controller

The charger/discharger must to regulate the bus voltage, which is needed to ensure the correct operation of the microgrid. Such a bus voltage regulation must be performed in any operation condition, i.e. charge, discharge and stand-by mode (null current exchange). Moreover, the controller must be stable in all the operation range, i.e. boosting and reducing voltage, hence the controller parameters must be adapted to the operation conditions. This process is performed by designing a Linear Quadratic Gaussian (LQG) controller with adaptive parameters. Therefore, in this chapter three control blocks are designed: a Linear Quadratic Integral (LQI) controller designed to regulate the bus voltage, an optimal observer to estimate the system states, and an adaptive law to adjust the LQG parameters depending on changes of the operation point. Three methods are proposed to calculate the adaptive parameters of the controller: two offline methods and another online method.

4.1 LQI controller design

The bus voltage regulation is described as an optimal control problem, thus it is needed to define a cost function based on a quadratic expression, which is solved using the LQR paradigm [42]. However, in order to ensure a null steady-state error, the integral of the error is included into the system, which results in the problem known as LQI [43]. The formalization of the previous problem is as follows: find the optimal duty cycle value $\tilde{d}^*(t)$ that minimizes expression (4.1).

$$J = \int_0^{t_f} \frac{1}{2} \cdot (\tilde{z}^T Q \tilde{z} + r \tilde{d}^2) \cdot dt \quad (4.1)$$

The previous expression represents the quadratic cost functional from the initial time to the final time t_f , where r is a positive scalar value and Q is 5×5 positive matrix (size imposed by the states $\tilde{x}$ defined in (2.16), plus the integral of the error):

$$Q = Q^T > 0 \quad r > 0$$

In addition, $\tilde{z}$ is the extended state vector resulting from the inclusion of the error integral, named $\tilde{x}_i$. Then, the state feedback law is imposed as given in (4.2), where K_x is a gain vector that multiplies the states $\tilde{x}$, and K_i is a scalar that multiplies the error integral.

$$\tilde{d} = -K \cdot \tilde{z} = - \begin{bmatrix} K_x & K_i \end{bmatrix} \cdot \begin{bmatrix} \tilde{x} \\ \tilde{x}_i \end{bmatrix} \quad (4.2)$$

The inclusion of the error integral produces the following extended system, where C_m has a negative sign due to the negative feedback of $\tilde{x}_4 = \tilde{V}_{dc}$ into the small-signal control system:

$$\dot{\tilde{z}} = \begin{bmatrix} \dot{\tilde{x}} \\ \dot{\tilde{x}_i} \end{bmatrix} = \underbrace{\begin{bmatrix} A_m & 0 \\ -C_m & 0 \end{bmatrix}}_{A_w} \cdot \underbrace{\begin{bmatrix} \tilde{x} \\ \tilde{x}_i \end{bmatrix}}_{\tilde{z}} + \underbrace{\begin{bmatrix} B_m \\ 0 \end{bmatrix}}_{B_w} \cdot \tilde{d} \quad (4.3)$$

Then, the optimization problem solution is calculated using the Hamiltonian matrix $H(\tilde{z}, \tilde{d}, P, t)$ given in (4.4), where the state equations are reported in (4.5), the costate equations are given in (4.6), and the steady-state conditions are given in (4.7).

$$H(\tilde{z}, \tilde{d}, P, t) = \frac{1}{2} \cdot (\tilde{z}^T \cdot Q \cdot \tilde{z} + r \cdot \tilde{d}^2) + P^T \cdot (A_w \cdot \tilde{z} + B_w \cdot \tilde{d}) \quad (4.4)$$

$$\dot{\tilde{z}}(t) = A_w \cdot \tilde{z} + B_w \cdot \tilde{d} \quad (4.5)$$

$$-\dot{P}(t) = Q \cdot \tilde{z} + A_w^T \cdot P \quad (4.6)$$

$$0 = r \cdot \tilde{d}^* + B_w^T \cdot P \quad (4.7)$$

In order to guarantee that the solution corresponds to the global optimal, the second order optimality condition given in (4.8) must be fulfilled:

$$\frac{\partial^2 H}{\partial d^2}(\tilde{z}^*, \tilde{d}^*, P^*, t) = 2 \cdot r > 0 \quad (4.8)$$

Since r is defined as a positive value then H is a convex function, thus $\tilde{d}^*$ corresponds to the minimum condition of H and it is the optimal solution around the operating point. The optimal value $\tilde{d}^*$ is obtained from equation (4.7), as follows:

$$\tilde{d}^* = -r^{-1} \cdot B_w^T \cdot P \quad (4.9)$$

$$\text{where } P \text{ is defined as } P = S_m \cdot \tilde{z} \quad (4.10)$$

In the previous expression S_m is a 5×5 matrix, which is required to calculate the stage-feedback vector gain K , and it will be discussed in Subsection 4.3. Replacing the state-space equations (4.5) into the costate equations (4.6), and considering that such an expression must be valid for all z conditions, the following restriction is obtained:

$$Q + A_w^T \cdot S_m + S_m \cdot A_w - S_m \cdot B_w \cdot r^{-1} \cdot B_w^T \cdot S_m + \dot{S}_m = 0 \quad (4.11)$$

The previous equation corresponds to the Riccati differential equation. Taking into account the control law defined in (4.9), and including the optimal duty cycle condition defined in (4.2), K is calculated as follows:

$$K = -r^{-1} \cdot B_w^T \cdot S_m \quad (4.12)$$

where S_m is the solution of differential equation (4.11). This problem is then formulated as an optimal regulator of infinite time, i.e. $t_f \rightarrow \infty$. In this case the Riccati differential equation is solved backwards starting from $t = t_f$, thus the initial condition is $S_m(t_f)$. The variable $S_m(t)$ has two components: a transient component and a steady-state one. Therefore, for $t \ll t_f$ the value of $S_m(t)$ corresponds to the steady-state solution of the Riccati equation S_{me} , which can be calculated making the derivative $\dot{S}_m(t)$ equal to zero in expression (4.11), which leads to:

$$S_{me} = \lim_{t \rightarrow \infty} S_m(t) \quad (4.13)$$

Based on the previous solution, the controller is implemented using the following time-invariant feedback law:

$$\tilde{d}^* = -r^{-1} \cdot B_w^T \cdot S_{me} \cdot \tilde{z} \quad (4.14)$$

4.2 State observer

One of the main drawbacks of a feedback state controller is the requirement of measuring all the states of the system, thus several sensing circuits are required for the experimental implementation, which increases the implementation cost and complexity. This drawback is addressed for the proposed solution by designing a state observer, which only requires to measure the dc bus voltage V_{dc} .

The structure of the linear observer is presented in figure 4.1, which corresponds to a full-state observer because all the states of the system are estimated. The inputs of the state observer are the measured voltage V_{dc} and the duty cycle d , which is known. The output of the observer is the vector of estimated states, which are the inductor currents and capacitor voltages, as they were defined in equation (2.16).

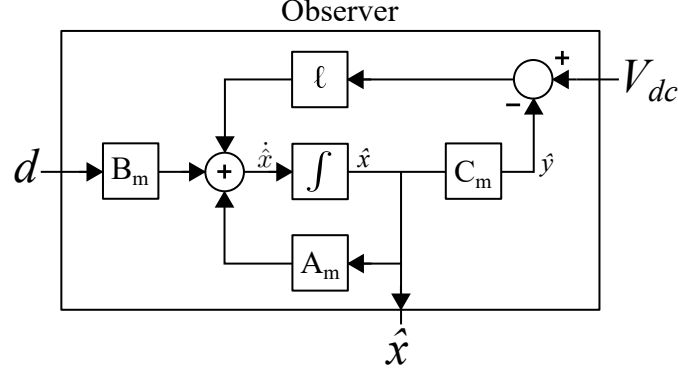


Figure 4.1: State observer

Considering the small-signal approximation of the system given in (2.21)-(2.22), and the feedback state law defined in (4.14) [44], the state observer is formalized as follow:

$$\dot{\hat{x}} = A_m \cdot \hat{x} + B_m \cdot d - \ell \cdot (y - \hat{y}) = (A_m + \ell \cdot C_m) \cdot \hat{x} + B_m \cdot d - \ell \cdot y \quad (4.15)$$

$$\hat{y} = C_m \cdot \hat{x} \quad (4.16)$$

where $\hat{y}$ corresponds to V_{dc} as shown in figure 4.1. Equation (4.15) shows that the observer gain vector ℓ defines the dynamic of the error estimation, such that the observer states converge to the states of the system. For determining ℓ , an optimization problem is defined to ensure the best estimation of the states are reached, and at the same time, the effect of the measurement noise in V_{dc} is minimized. Similarly to the calculation of feedback state vector K , the observer gain vector ℓ is determined in equation (4.18), which requires to solve the Riccati differential equation given in expression (4.17) [45]. In such an expression S_o corresponds to the optimal error covariance matrix of the state observer, and γ is a scalar used to adjust the convergence speed of the observer states [46].

$$A_m \cdot S_o + S_o \cdot A_m^T - S_o \cdot C_m^T \cdot \gamma^{-1} \cdot C_m \cdot S_o + B_m \cdot B_m^T + \dot{S}_o = 0 \quad (4.17)$$

$$\ell = S_o \cdot C_m^T \cdot \gamma \quad (4.18)$$

4.3 Adaptive parameters of the LQG controller

In order to guarantee a correct dynamic behavior of the charger/discharger in any operating condition, an adaptive process to ensure the stability of the LQG controller has been proposed. This means that the state feedback vector K and observer gain vector ℓ are dynamically calculated depending on the operating condition. Such an adaptation is performed by continuously calculating the A_m and B_m matrices, and continuously solving the Riccati differential equations.

4.3.1 Offline calculation of the controller parameters

This section details the offline process to calculate the adaptive parameters of the controller for the defined range of operation. Two methods were proposed to adjust the controller parameters according to the operating point:

Lookup-Table method

This method consists of calculating the state feedback vector and the observer gain vector, solving Riccati's differential equation offline once for each operating point. Subsequently, the controller parameters are stored in a Lookup-table. The selection of a specific parameter from the Lookup-Table consists of searching for the closest value from the table (Nearest method), according to the bus reference voltage and the measured battery voltage.

Polynomial interpolation method

This method consists of defining a polynomial function that represents, with some precision, the variation of each element of the state feedback vector and of the observer gain vector. For this, it is decided to use the data calculated for the Lookup table and then a polynomial interpolation. The degree of the polynomial is defined in such a way that the root mean square error (RMSE) is significantly small. As a result, after calculating the value of the coefficients, functions that depend on V_b and V_{dc} are obtained, which are subsequently implemented in a control device.

4.3.2 Online calculation of controller parameters

The Riccati differential equation is solved online by the iterative procedure shown in figure 4.2, which provides $\frac{dS_m}{dt}$. This value is integrated to obtain the matrix S_m , which is used to determine the value of K from equation (4.12). The same procedure is used to solve the differential Riccati equation given in (4.17), which

provides S_o to determine the observer state gain using expression (4.18). Both S_m and S_o matrices can exhibit positive and negative values, hence those matrices are initialized with zeros to provide an homogeneous excursion of both positive and negative possibilities. This problem is defined as an infinite-time regulator, therefore the final S_m and S_o must be obtained in steady-state (derivates $\dot{S}_m$ and $\dot{S}_o$ equal to zero) for each operation point. The previous condition requires that the state observer calculations must be performed fast, which is needed to ensure that the estimated states are available prior to the calculation of the optimal value d . Finally, figure 4.3 reports the procedure used to calculate the online adaptive parameters of the LQG controller.

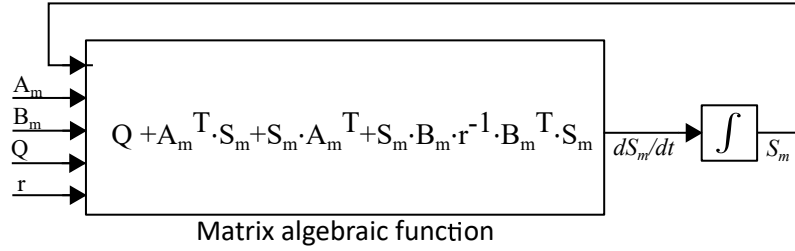


Figure 4.2: Block diagram used to calculate the numerical solution of the Riccati differential equation

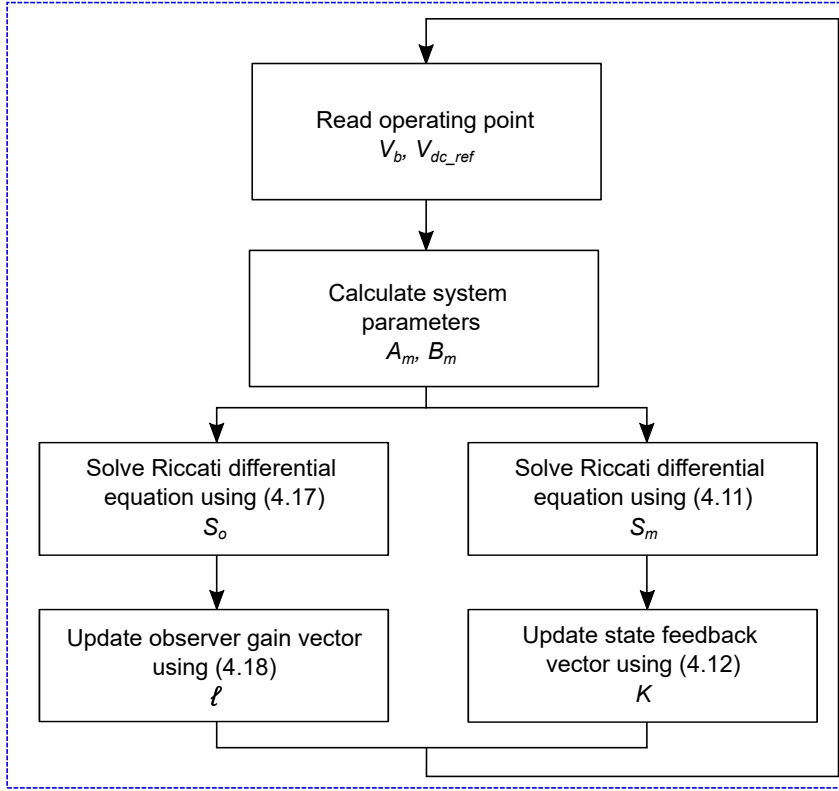


Figure 4.3: Flowchart to calculate the online adaptive parameters of the LQG controller

4.4 Structure of the adaptive LQG voltage controller

The structure of the complete control system requires the interaction of the state observer, the state feedback (LQI) controller, and the adaptive tuning of the parameters. Such a structure is observed in figure 4.4, where the LQI feedback is divided into the vector K_x and the scalar K_i , the last one processing the integral of the error. The adaptive tuning block produces the adaptive feedback vector K_x and the matrices used to process the state observer (A_m, B_m, ℓ) , which depend on the operating point defined by both the battery and the reference voltages. The real-time implementation of such a control system will be described in chapter 5.

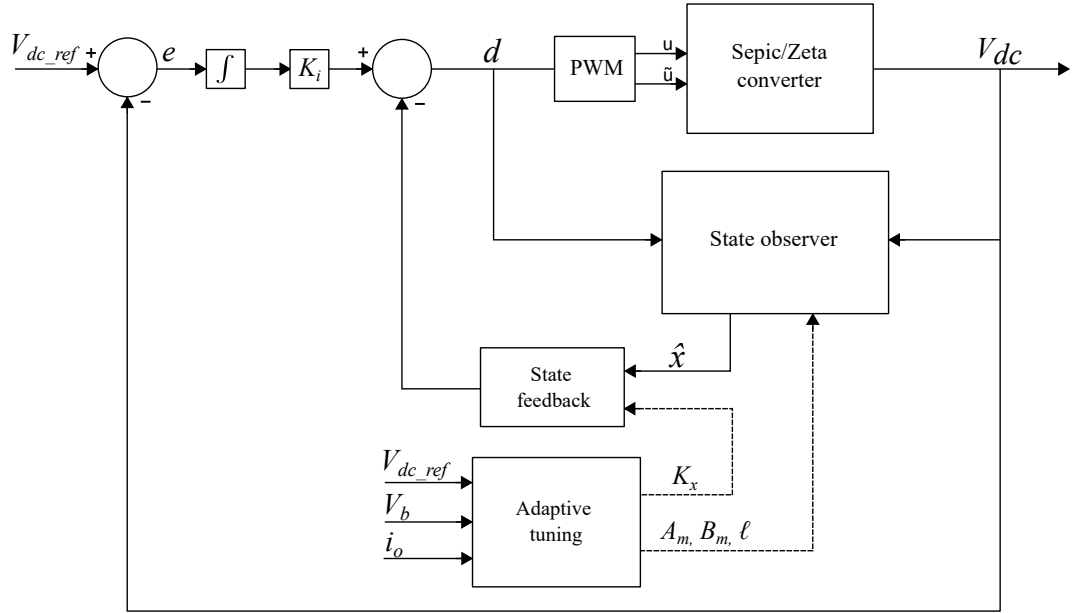


Figure 4.4: Structure of the adaptive LQG voltage controller

4.5 Design example and simulation results

This section illustrates the design process proposed in the flowchart of figure 3.1 based on the realistic requirements reported in Table 4.1. This application case is defined to provide a realistic design, which is experimentally verified in chapter 5. This proof-of-concept design considers a maximum battery power of $26W$, and the switching frequency F is chosen as 40 kHz due to the low requirements imposed to the PWM generator. The design example is developed in three phases: the calculation of the Sepic/Zeta converter parameters, the adaptive calculation of the controller parameters, and the verification of the design using circuital simulations.

Table 4.1: Parameters for the design example.

Parameter	Value
V_b	10–26 V
V_{dc}	10–26 V
F	40 kHz
I_{dc}	$[-1,1]$ A
n_{min}	90%
Δi_L	20%
ΔV_c	1%

Applying the design process of figure 3.1, the commercial devices of Table 4.2 are selected.

Table 4.2: Electronic devices selected for the Sepic/Zeta converter.

Parameter	Ref	Values
<i>Mosfets</i>	IRF3710	$R_{on} = 23 \text{ m}\Omega$, $V_{DS} = 100 \text{ V}$, $I_D = 57 \text{ A}$
L_1	2322-RC	$L = 680 \text{ }\mu\text{H}$, $DCR = 0.15 \text{ }\Omega$
L_2	2322-RC	$L = 680 \text{ }\mu\text{H}$, $DCR = 0.15 \text{ }\Omega$
C_i	UEP1J331MHD	$C = 330 \text{ }\mu\text{F}$, $V = 63 \text{ V}$
C_{dc}	UEP1J331MHD	$C = 330 \text{ }\mu\text{F}$, $V = 63 \text{ V}$

4.5.1 Calculation of controller parameters

First, concerning the state penalty matrix Q and the control penalty scalar r [47], those parameters are determined using an iterative process to ensure a performance of the controlled system with a maximum overshoot of 10% and a stabilization time lower than 10 *ms*. Moreover, K_i is defined as a constant because it does not depend on the operating point. The final values of the controller parameters are:

$$Q = \begin{bmatrix} 1 & 0 & 0 & 0 & 0 \\ 0 & 1 & 0 & 0 & 0 \\ 0 & 0 & 1 & 0 & 0 \\ 0 & 0 & 0 & 5 & 0 \\ 0 & 0 & 0 & 0 & 1 \end{bmatrix} \quad r = 1000 \quad K_i = 16$$

Second, the controller parameters K and ℓ are calculated offline with the Lookup-table and polynomial interpolation methods, and later with the online method.

Calculation of controller parameters with adaptive lookup-table

To generate the data for the Lookup table, it was decided to calculate the controller parameters by varying the battery voltage and the DC bus voltage in the operating range established in Table 4.1. For this, it was used the LQR function of the Control System toolbox package in Matlab, and the controller parameters were calculated by varying V_b and V_{dc} every 2 V. The established search algorithm is Nearest, which consists of returning the value of the table corresponding to the closest data according to the combination of the inputs (V_b , V_{dc}).

Calculation of controller parameters with adaptive polynomial interpolation

To establish the polynomial functions that represent the variation of the state feedback vector, a polynomial interpolation was performed with the data generated in the lookup table, for each element of the vector, as a function of the voltages V_b and V_{dc} . For this, the Curve Fitter tool of Matlab was used, which calculates the coefficients according to the polynomial degree selected for the inputs, using the Linear Least Squares method. For this case, a 4th order polynomial for V_b and 3rd for V_{dc} were selected, in such a way that a maximum root-mean-square error (RMSE) equal to 0.026 was obtained. This means that the resulting polynomial functions represent with sufficient accuracy the state feedback vector in the established design range.

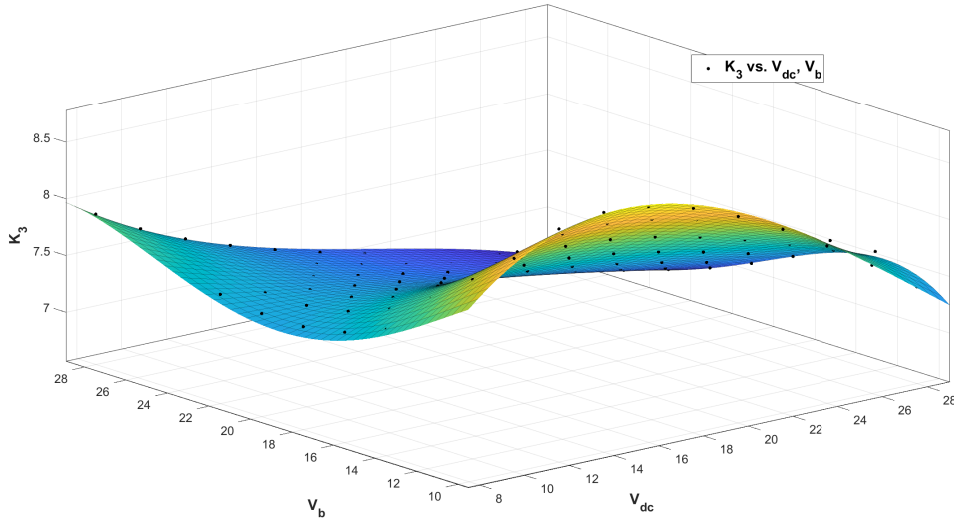


Figure 4.5: Fit plot of K_3

Figure 4.5 shows an example of the polynomial fit for the third element of the state feedback vector (K_3). It is evident that the resulting surface adequately represents the set of points generated for all the combinations of V_b and V_{dc} in the defined range.

The previous procedure is repeated to calculate the coefficients of the polynomial functions of the observer gain vector. The polynomial degree established for V_b and V_{dc} was 3, and as a result a maximum RMSE of 0.06 was obtained. Figure 4.6 shows the polynomial fit for the third element of the observer gain vector (ℓ_3) as an example, obtaining a surface that adequately represents the variation of this observer parameter.

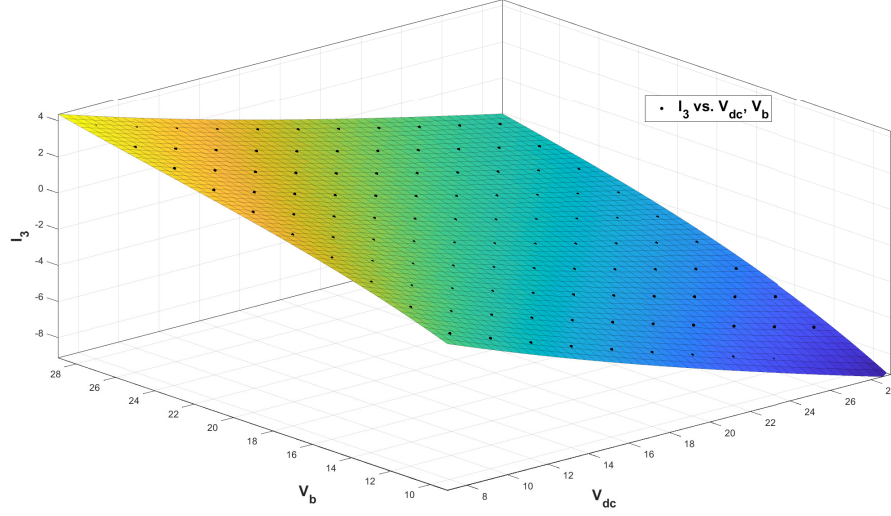


Figure 4.6: Fit plot of ℓ_3

Calculation of controller parameters with the online adaptive LQG

The controller parameters K and ℓ are calculated online following the procedure of figure 4.3. Due to the low complexity implementation requirements, the Euler numerical integration method is selected to solve $\frac{dS_m}{dt}$.

4.5.2 Simulations results

The designed battery charger/discharger is implemented in Simulink from Matlab [48] using the Simscape library and the Specialized Power Systems tools [49]. Such a library enables to model the power devices considering the parasitic resistances, voltage and inductance of the diode, and the on-resistance of the transistor. In addition, functions already built into Simulink are used to implement the proposed control scheme. The performance of the controlled system is tested for two goals: dc bus regulation and reference tracking, each one of them at different operating points.

Simulations results for bus voltage regulation with offline parameter adjustment

The first objective is to regulate the bus voltage V_{dc} under the presence of bus current disturbances (i_o), which occur due to changes of the power balance between the loads and sources of the microgrid. The microgrid bus current is defined to

impose different operating modes to the battery, i.e. charge, discharge and stand-by, all of them occurring in real situations. Moreover, some converter modes are also tested: buck (low bus voltage), boost (high bus voltage) and unitary gain.

Initially, tests are carried out using the adaptive adjustment method based on the lookup table. Figures 4.7, 4.8 and 4.9 show the performance of the charger/discharger considering a 12 V battery. All the simulations start in stand-by mode, this means $i_{dc} = 0$ A. At 50 ms the bus current changes to 0.5 A, hence the battery enters in discharge mode ($i_{dc} > 0$ A), and the battery must to supply the power required by the bus. At 150 ms and 250 ms the bus current is increased to 1 A and decreased to 0.5 A, respectively. Then, at 350 ms the bus current changes to -0.5 A, thus the battery enters in charge mode ($i_{dc} < 0$ A), and the battery must to absorb the power excess present in the dc bus. An additional decrement of the bus current is applied at 450 ms, reaching a value of -1 A, and returning to -0.5 A at 550 ms. Finally, at 650 ms the battery reaches again the stand-by mode. The same test is performed for bus voltages equal to 10 V (figure 4.7, buck mode), 12 V (figure 4.8, unitary gain mode), and 16 V (figure 4.9, boost mode). The simulations using Adaptive LQG with Lookup-table, confirm the correct regulation of the bus voltage on any operation condition of the battery, thus the bus voltage is stable under any micro-grid condition. Furthermore, the controller adaptability is demonstrated since the change on the converter mode (buck, boost, unitary gain) is automatically compensated and the duty cycle, which is never saturated, confirms the correct operation of both the observer and adaptive stage

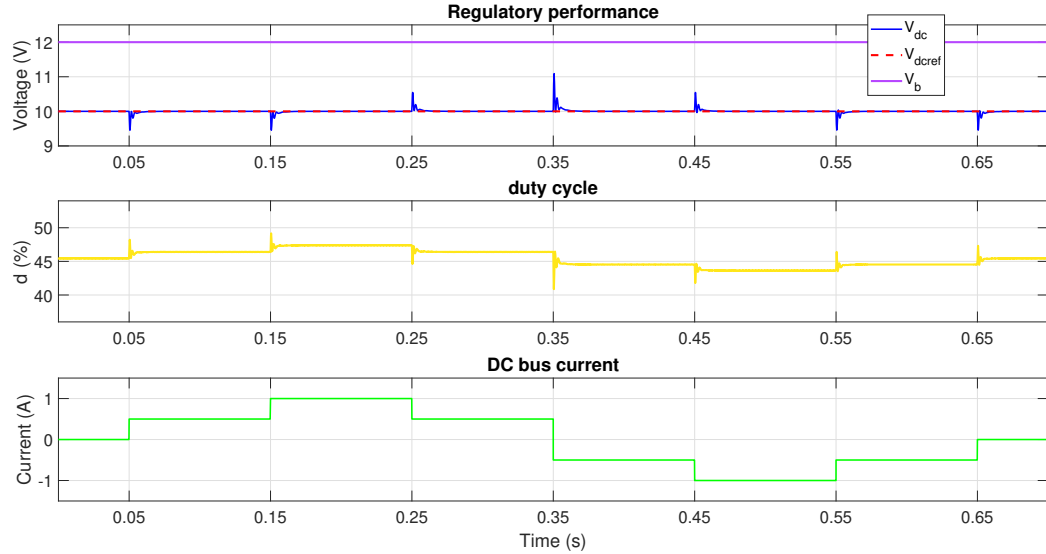


Figure 4.7: Performance with $V_b = 12$ V and reference voltage $V_{dcref} = 10$ V using Adaptive LQG with Lookup-table (buck mode)

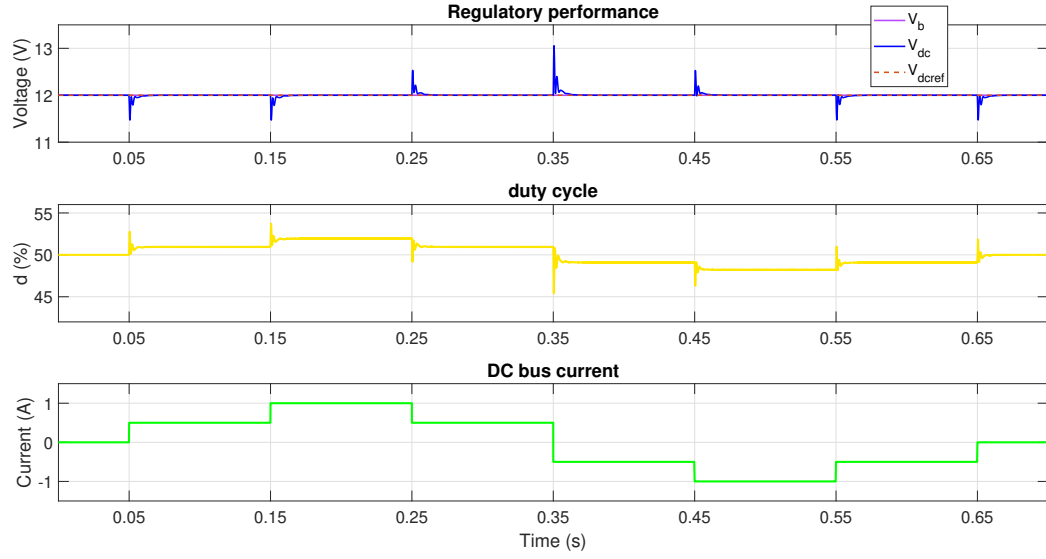


Figure 4.8: Performance with $V_b = 12$ V and reference voltage $V_{dcref} = 12$ V using Adaptive LQG with Lookup-table (unitary gain mode)

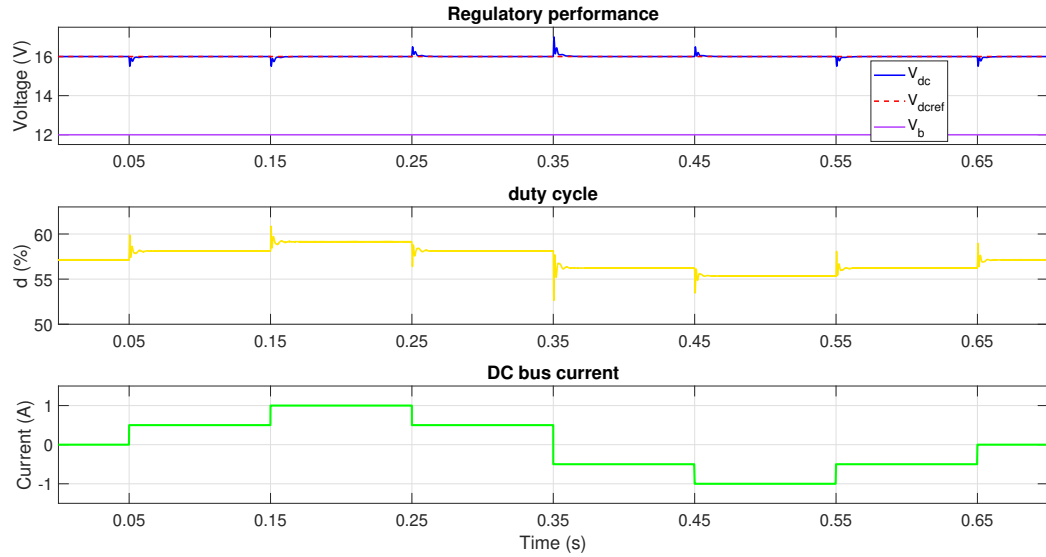


Figure 4.9: Performance with $V_b = 12$ V and reference voltage $V_{dcref} = 16$ V using Adaptive LQG with Lookup-table (boost mode)

The same previous test was conducted considering two batteries in series connection, i.e. $V_b = 24$ V, which enables to test the adaptability of the control system to changes on the battery voltage. The simulation results with such a new battery voltage are reported in figures 4.10 (buck mode), 4.11 (unitary gain mode) and 4.12

(boost mode). In this case the bus voltage is set to 20 V, 24 V and 26 V, respectively. As in the previous tests, the microgrid bus is successfully regulated for any operation condition, and without any modification to the proposed control system.

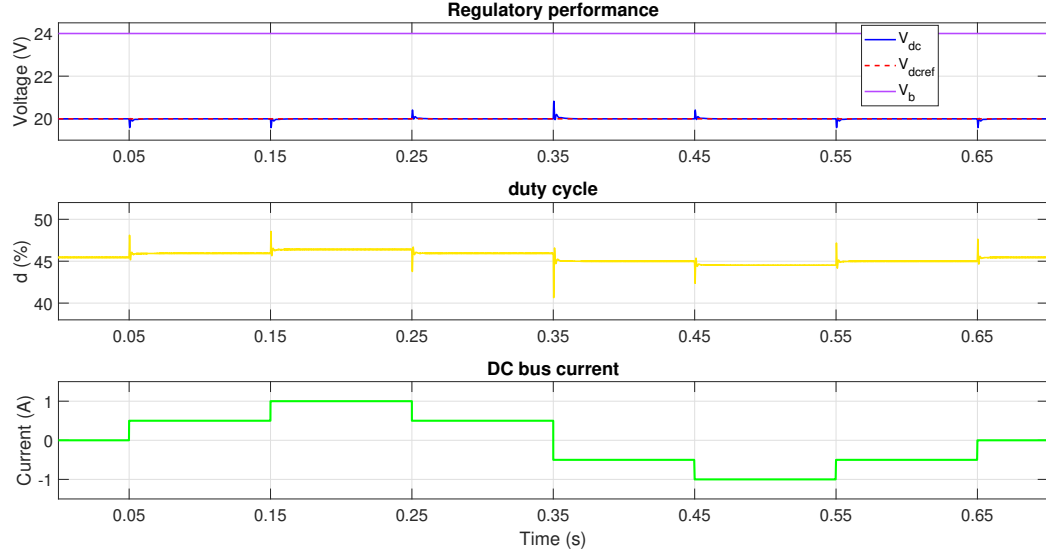


Figure 4.10: Performance with $V_b = 24$ V and reference voltage $V_{dcref} = 20$ V using Adaptive LQG with Lookup-table (buck mode)

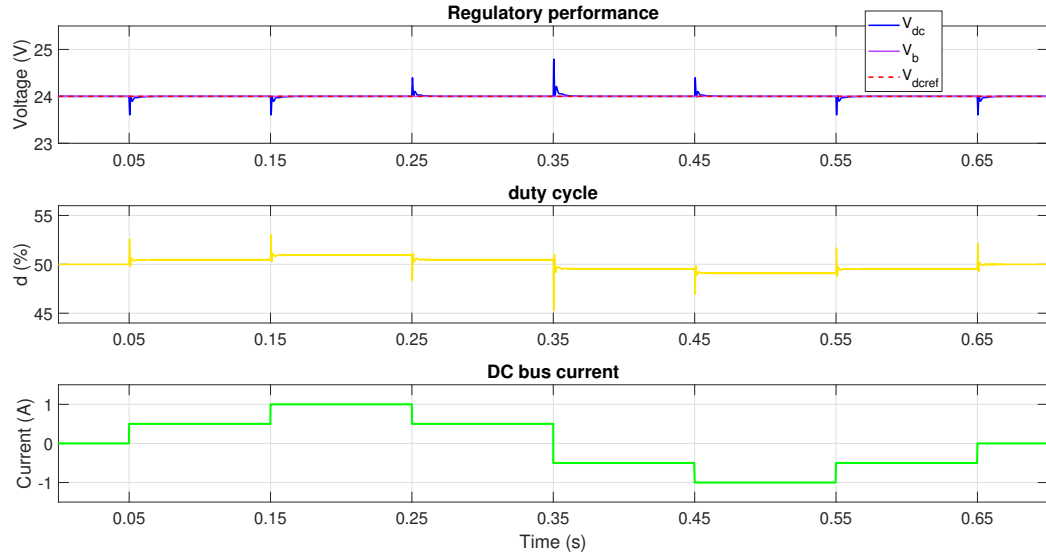


Figure 4.11: Performance with $V_b = 24$ V and reference voltage $V_{dcref} = 24$ V using Adaptive LQG with Lookup-table (unitary gain mode)

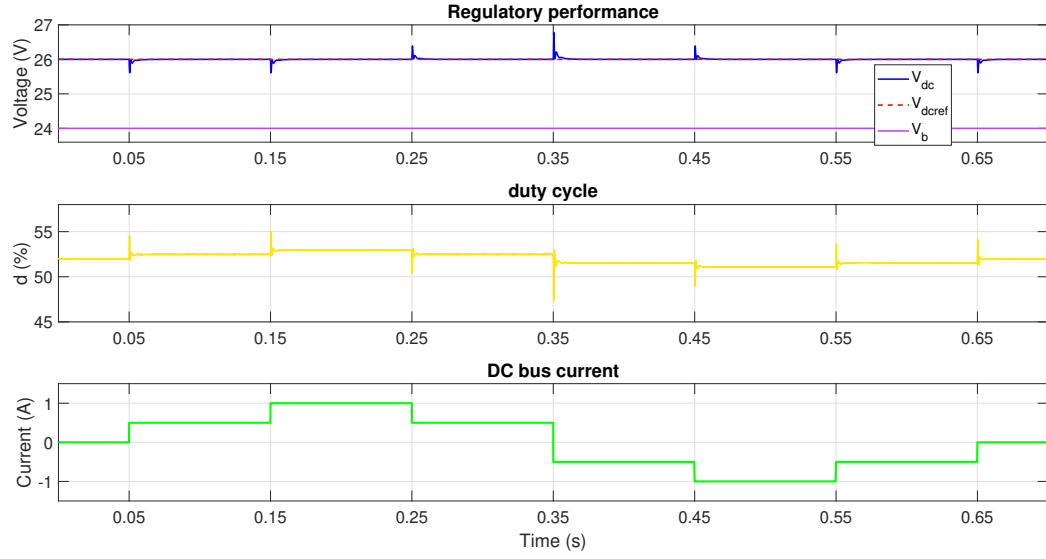


Figure 4.12: Performance with $V_b = 24\text{ V}$ and reference voltage $V_{dcref} = 26\text{ V}$ using Adaptive LQG with Lookup-table (boost mode)

Next, the tests are repeated for the adaptive parameter method based on the polynomial interpolation. The tests are carried out with one and two batteries connected in series, and testing the different modes of the converter. The simulation results are reported in the figures 4.13 (buck mode), 4.14 (unitary gain mode) and 4.15 (boost mode) that correspond to a single battery, and in 4.16 (buck mode), 4.17 (unitary gain mode) and 4.18 (boost mode) for two series-connected batteries.

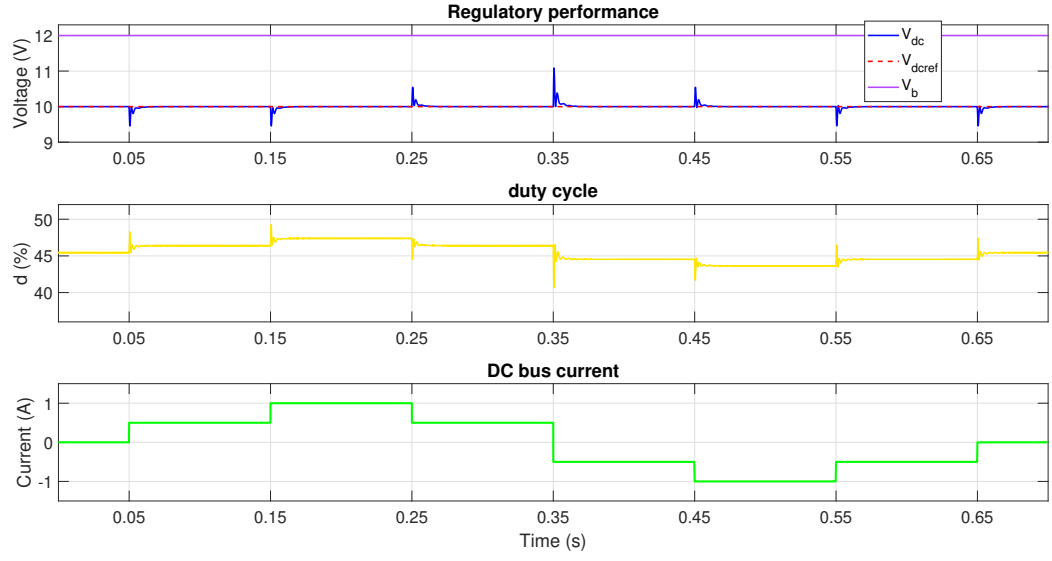


Figure 4.13: Performance with $V_b = 12$ V and reference voltage $V_{dcref} = 10$ V using Adaptive LQG with Polynomial interpolation (buck mode)

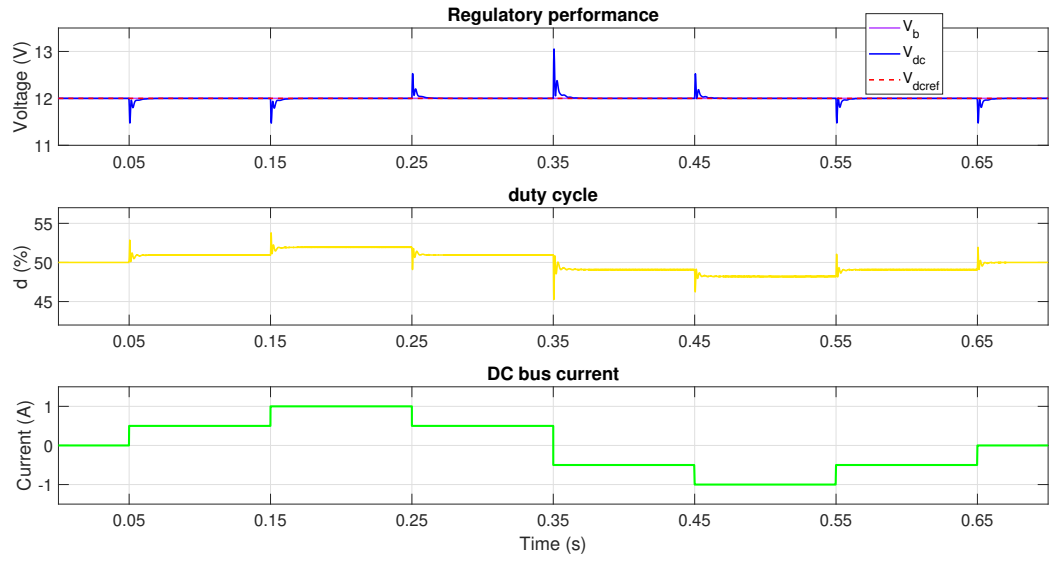


Figure 4.14: Performance with $V_b = 12$ V and reference voltage $V_{dcref} = 12$ V using Adaptive LQG with Polynomial interpolation (unitary gain mode)

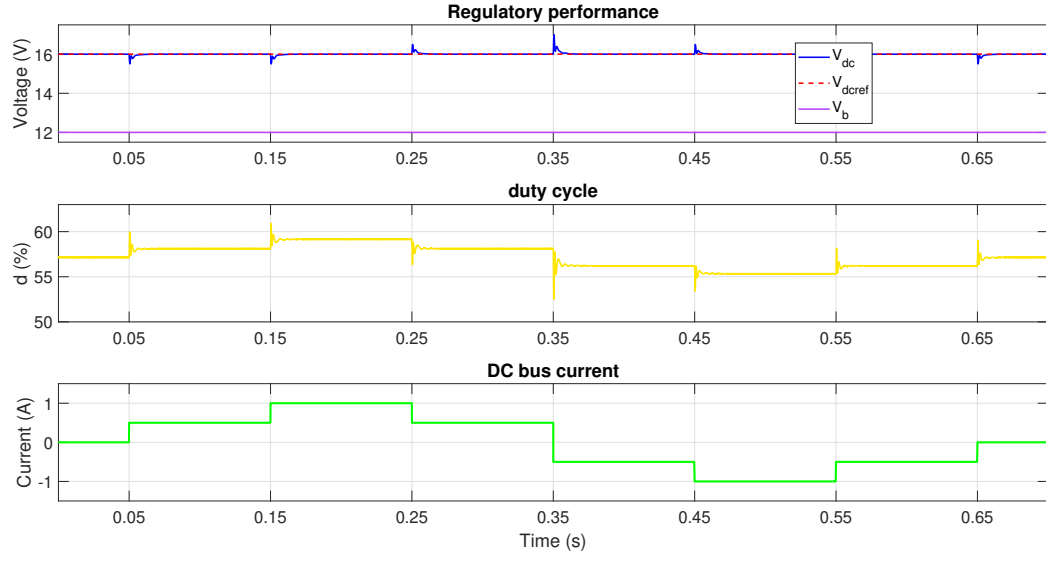


Figure 4.15: Performance with $V_b = 12$ V and reference voltage $V_{dc_{ref}} = 16$ V using Adaptive LQG with Polynomial interpolation (boost mode)

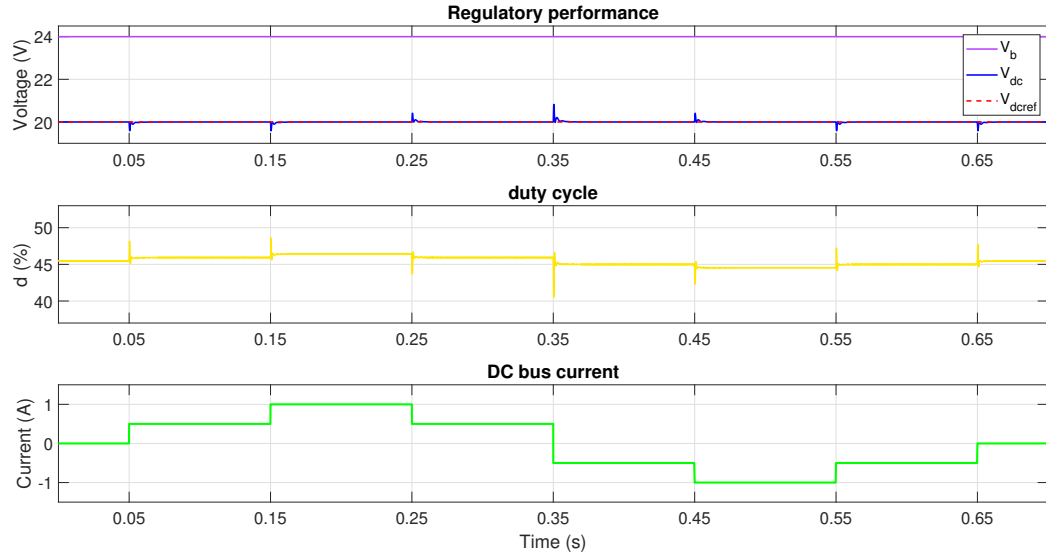


Figure 4.16: Performance with $V_b = 24$ V and reference voltage $V_{dc_{ref}} = 20$ V using Adaptive LQG with Polynomial interpolation (buck mode)

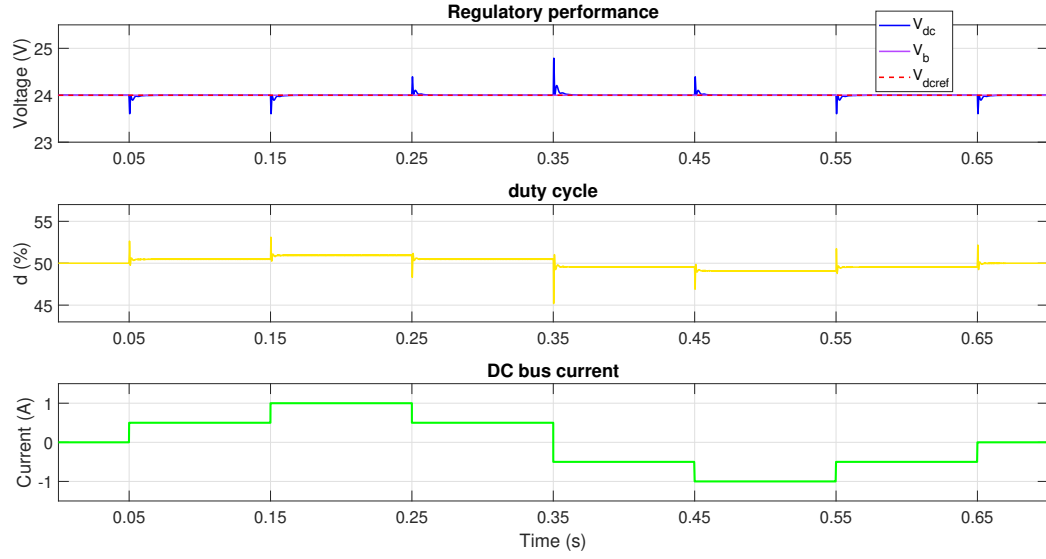


Figure 4.17: Performance with $V_b = 24$ V and reference voltage $V_{dcref} = 24$ V using Adaptive LQG with Polynomial interpolation (unitary gain mode)

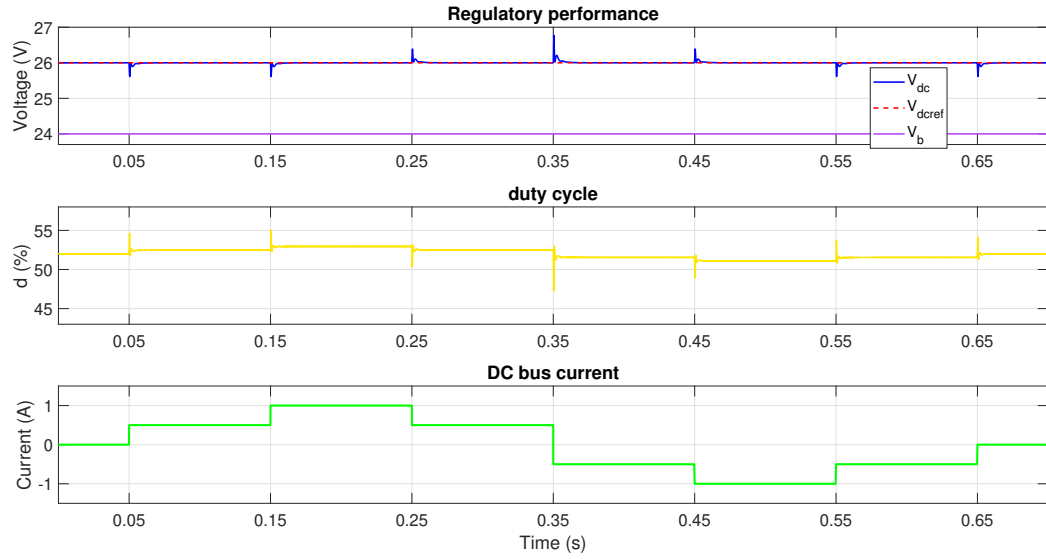


Figure 4.18: Performance with $V_b = 24$ V and reference voltage $V_{dcref} = 26$ V using Adaptive LQG with Polynomial interpolation (boost mode)

The tests of the adaptive controller with polynomial interpolation show adequate voltage regulation against variations in the DC bus current, and for the different modes of the converter. However, there are no significant differences in transient behavior in contrast to the Lookup table method. Nevertheless, the choice of the

method may be due to the limitations of the hardware to be implemented, since the Lookup table method consumes more memory, but the polynomial interpolation method consumes more processing.

Simulations results for bus voltage regulation with online parameter calculation

Exactly to the tests for the offline adaptive methods, the simulations start in stand-by mode, then the battery enters in discharge mode, next in charge mode and finally returns to stand-by mode. The test is performed for bus voltages equal to 10 V (figure 4.19, buck mode), 12 V (figure 4.20, unitary gain mode), and 16 V (figure 4.21, boost mode).

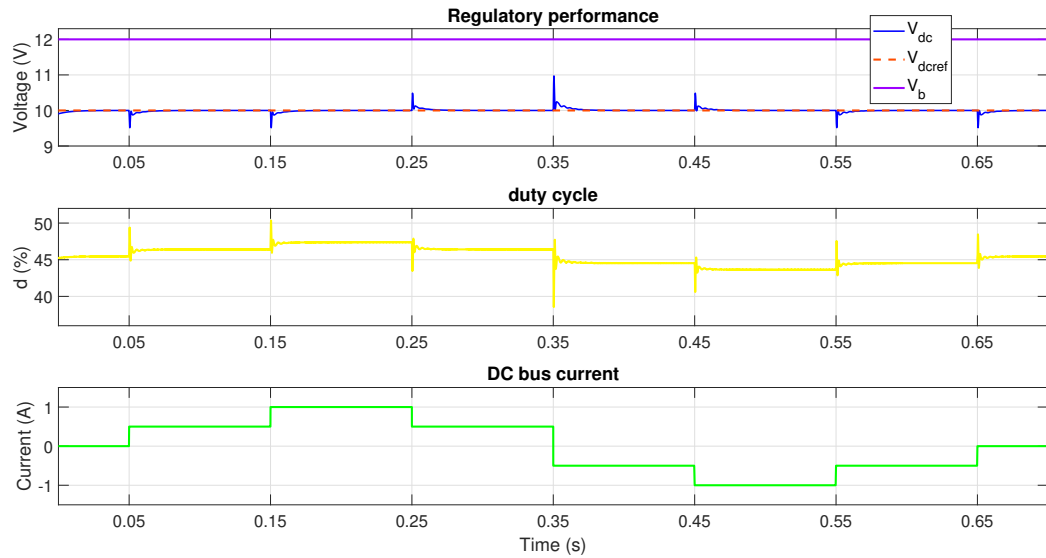


Figure 4.19: Performance with $V_b = 12$ V and reference voltage $V_{dcref} = 10$ V using online Adaptive LQG (buck mode)

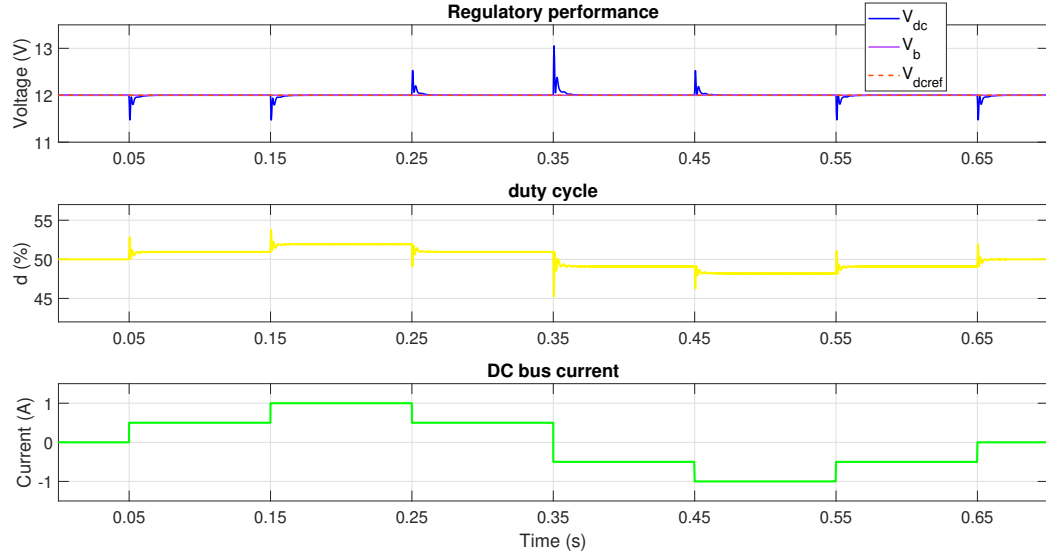


Figure 4.20: Performance with $V_b = 12\text{ V}$ and reference voltage $V_{dcref} = 12\text{ V}$ using online Adaptive LQG (unitary gain mode)

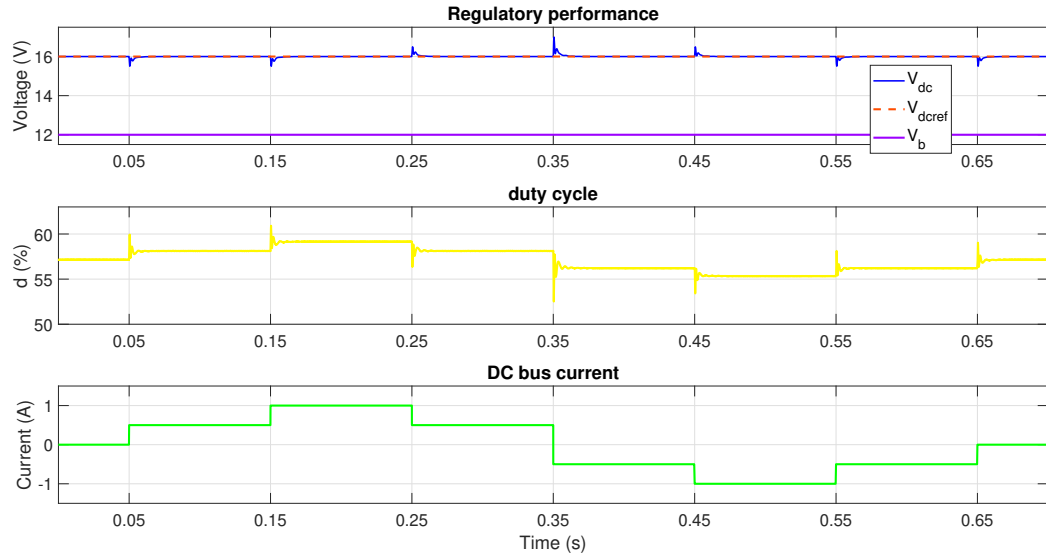


Figure 4.21: Performance with $V_b = 12\text{ V}$ and reference voltage $V_{dcref} = 16\text{ V}$ using online Adaptive LQG (boost mode)

Now, two batteries are connected in series and the tests are repeated to regulate the DC bus voltage at 20 V , 24 V and 26 V . In this way, the correct operation of the control system is verified with the online adjustment method against changes in the configuration of the batteries at the voltage level. Figures 4.22, 4.23 and 4.24

show the performance of the charger/discharger considering a 24 V battery.

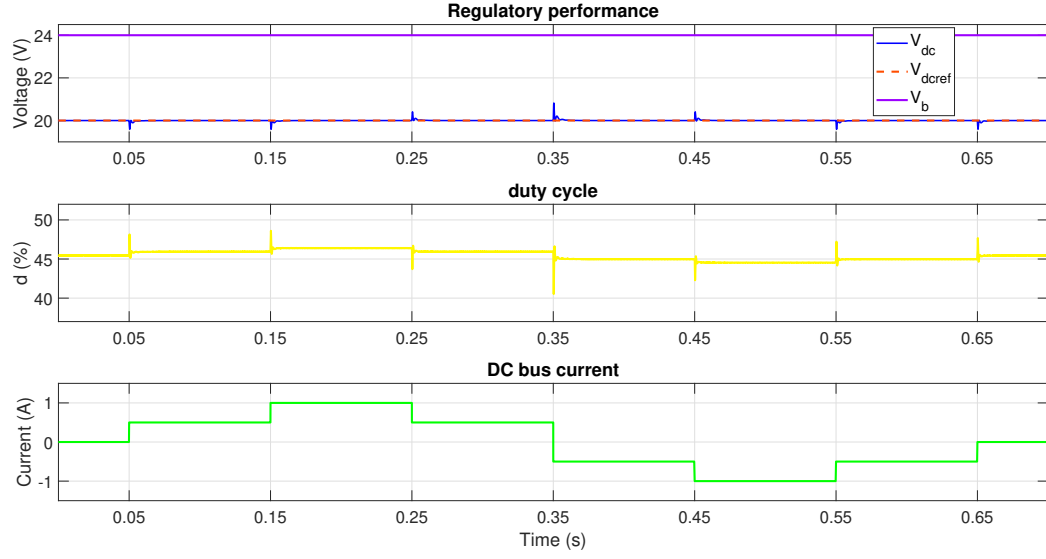


Figure 4.22: Performance with $V_b = 24$ V and reference voltage $V_{dcref} = 20$ V using online Adaptive LQG (buck mode)

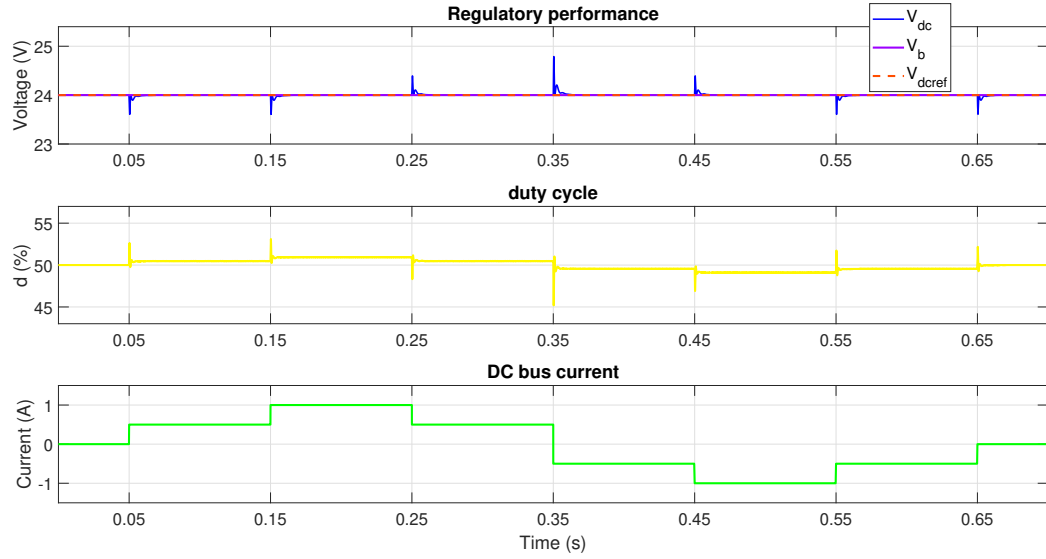


Figure 4.23: Performance with $V_b = 24$ V and reference voltage $V_{dcref} = 24$ V using online Adaptive LQG (unitary gain mode)

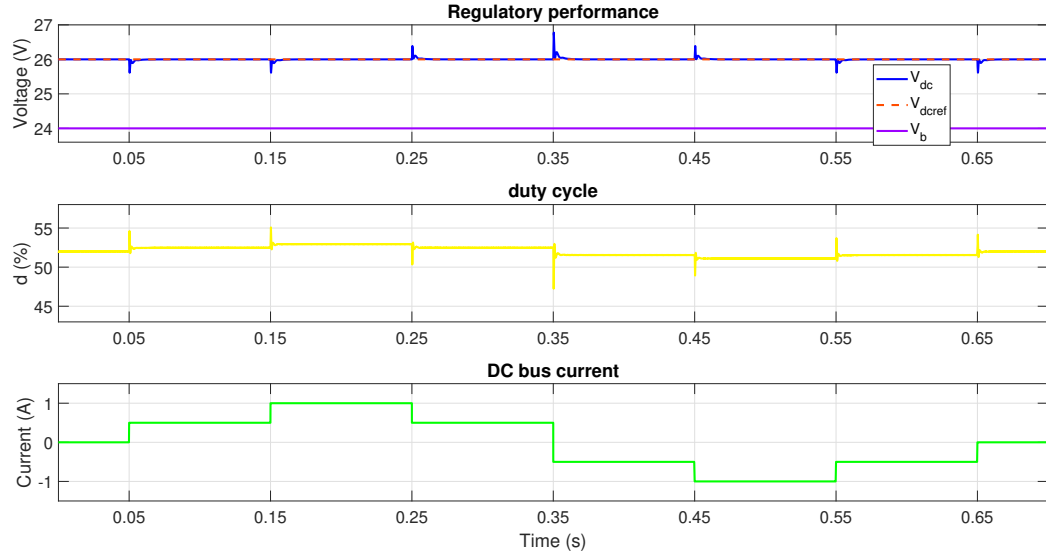


Figure 4.24: Performance with $V_b = 24$ V and reference voltage $V_{dcref} = 26$ V using online Adaptive LQG (boost mode)

Table 4.3 summarizes the performance of the charger/discharger under the different operating conditions and modes considered in this section. The results point out that the maximum overshoot is 9.7 %, and the settling time is under 10 ms in all cases using the online adaptive LQG; while the maximum overshoot is 10.9 %, and the settling time is under 10 ms using the offline adaptive LQG methods, hence the design requirements are fulfilled. Although there is lower overshoot in the online Adaptive LQG method with the tests at $V_b = 12$ V, compared to the other offline methods, there is also a longer settling time. This happens because of slight inaccuracies in the controller parameters with the offline adaptive methods, thus an error in the settling time occurs.

Table 4.3: Summary of the dynamic performance for bus voltage regulation.

Test Conditions		Overshoot (%)			Settling Time (ms)		
V_b (V)	V_{dc} (V)	Lookup table	Polynomial interp.	LQG online	Lookup table	Polynomial interp.	LQG online
12	10	10.9	10.85	9.7	3.3	3.65	3.83
	12	8.8	8.76	8.75	3.2	3.45	3.46
	16	6.25	6.21	6.19	2.9	3	3.02
24	20	4.13	4.1	4.1	0.75	0.75	0.75
	24	3.3	3.29	3.29	0.671	0.672	0.67
	26	3	2.98	3	0.63	0.624	0.63

The simulations using the three proposed methods to adapt the parameters of the controller confirm the correct regulation of the bus voltage on any operation condition of the battery, thus the bus voltage is stable under any microgrid condition. Moreover, the controller adaptability is demonstrated since the change on the converter mode (buck, boost, unitary gain) is automatically compensated, hence it could be used to interface any commercial battery with any bus voltage level. Finally, the simulations also show the duty cycle, which is never saturated, thus confirming the correct operation of both the observer and adaptive stage.

Simulations for reference tracking with the offline parameter adjustment

Two simulation cases were conducted to evaluate the controller performance for reference tracking. In these tests the bus voltage reference changes at 60 V/s around the battery voltage, which also forces the charger/discharger to enter in boost, buck and unitary gain modes. Figure 4.25 shows the simulation results considering a battery voltage equal to 12 V , where the reference profile changes between $V_{dc} = 16 \text{ V}$ and $V_{dc} = 10 \text{ V}$. In addition, the bus current changes from 1 A to -1 A , which sets the battery in both charge and discharge modes. Therefore, the battery charger/discharger must to follow the reference profile both charging and discharging the battery. The tests show an adequate tracking of the references in all the conditions; however, unexpected transients occur in the reference changes. This occurs due to the sudden changes in the controller parameters when the operating point changes significantly.

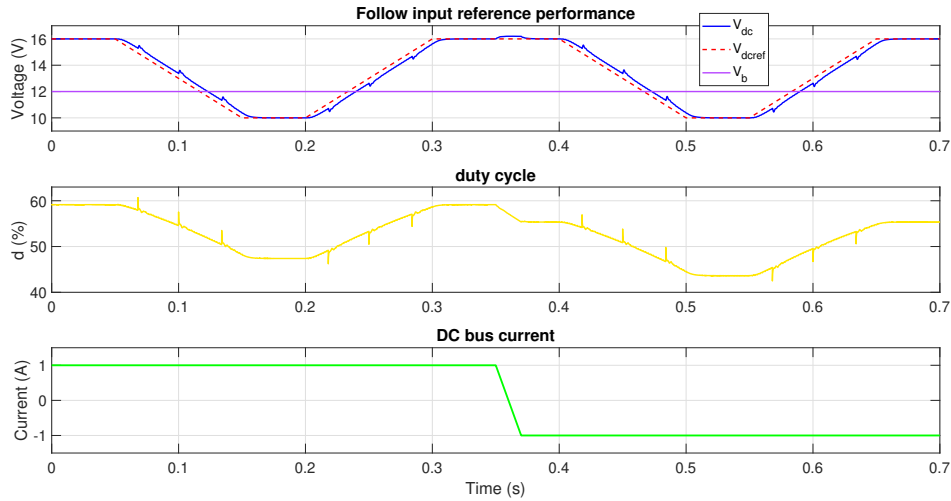


Figure 4.25: Reference tracking performance at $V_b = 12 \text{ V}$ using Adaptive LQG with Lookup-table

The second test considers two batteries in series connection, i.e. $V_b = 24 \text{ V}$.

Those simulation results are reported in figure 4.26. In this case, for evaluating the buck and boost modes of the converter, the reference changes between $V_{dc} = 20\text{ V}$ and $V_{dc} = 26\text{ V}$, and the bus current sets the battery in both charge and discharge modes. As in the previous case, the controller ensures a satisfactory tracking of the reference at any condition, but there are unexpected transients in the ramp zones, the same as in the test with a single battery.

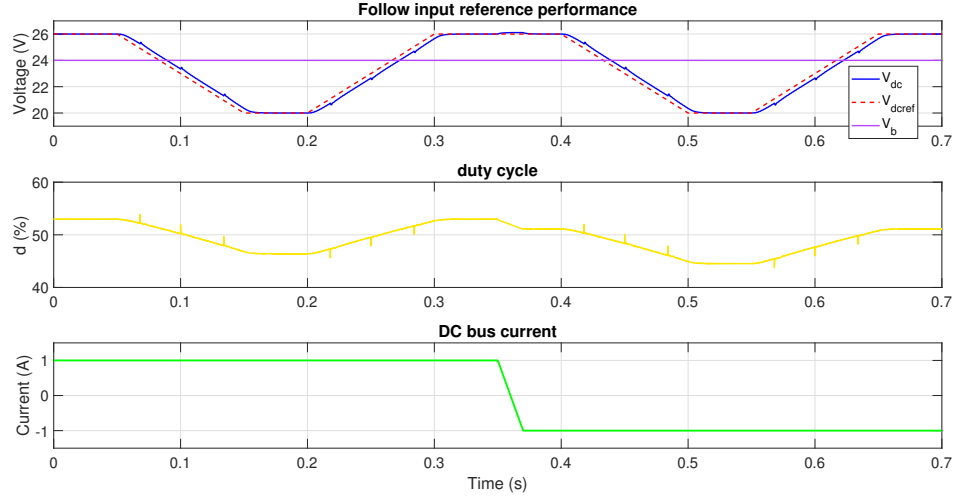


Figure 4.26: Reference tracking performance at $V_b = 24\text{V}$ using Adaptive LQG with Lookup-table

The previous tracking test scenario is repeated for the adaptive method with the polynomial interpolation. Figures 4.27 and 4.28 show the results for one and two batteries, respectively, where the controller ensures adequate tracking performance under the different conditions of the converter. In addition, there are no unexpected transients in the ramp zones since the adaptive adjustment of the parameters correspond to non-discontinuous functions, unlike the Lookup table method.

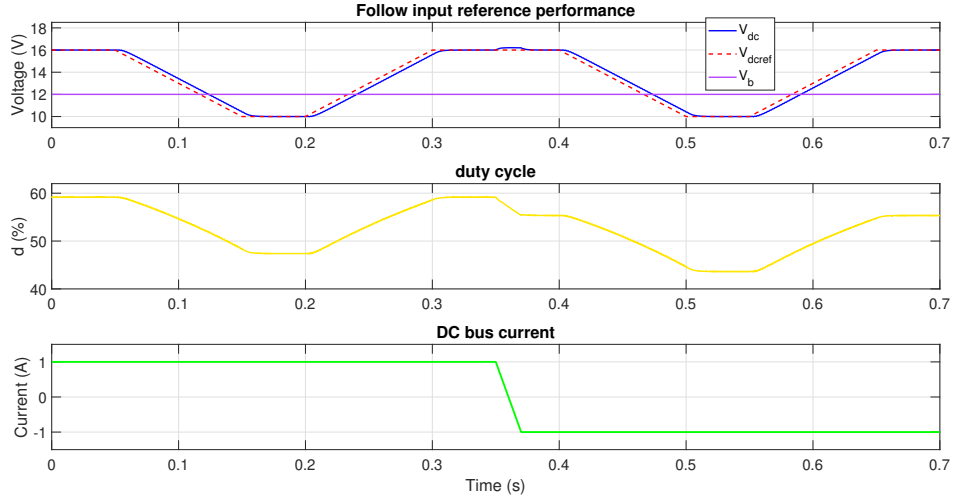


Figure 4.27: Reference tracking performance at $V_b = 12V$ using Adaptive LQG with Polynomial interpolation

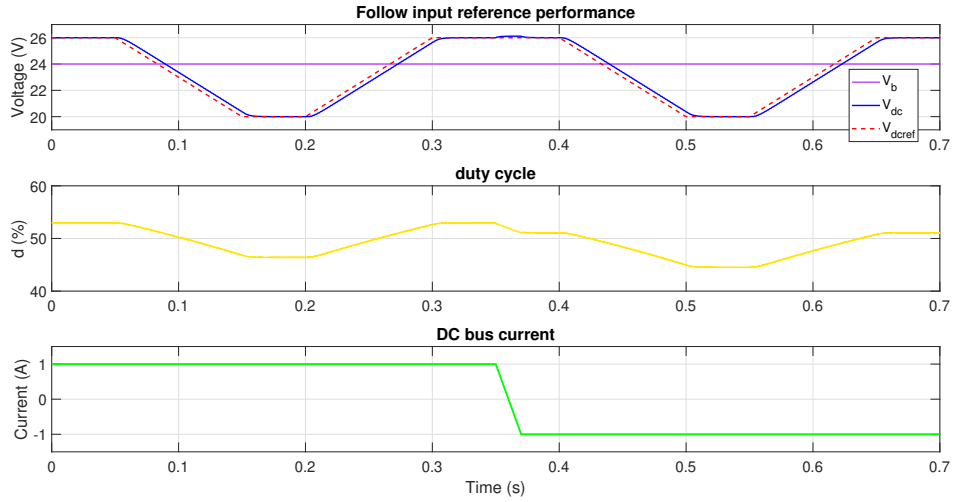


Figure 4.28: Reference tracking performance at $V_b = 24V$ using Adaptive LQG with Polynomial interpolation

Simulations for reference tracking with online parameter adjustment

In the same way as in the offline adaptive methods, the same conditions are evaluated for the online adaptive LQG method. The results are shown in figures 4.29 and 4.30, which test the charger/discharger conditions and Buck, Boost and unity gain modes of the converter. As in the previous tests, the controller ensures the correct operation

of the system, which in this case corresponds to a satisfactory reference tracking in any operation condition.

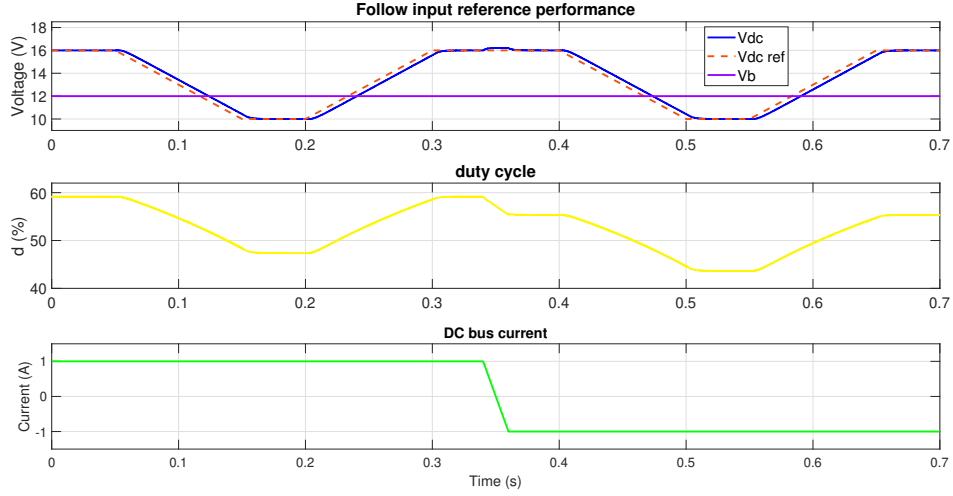


Figure 4.29: Reference tracking performance at $V_b = 12V$

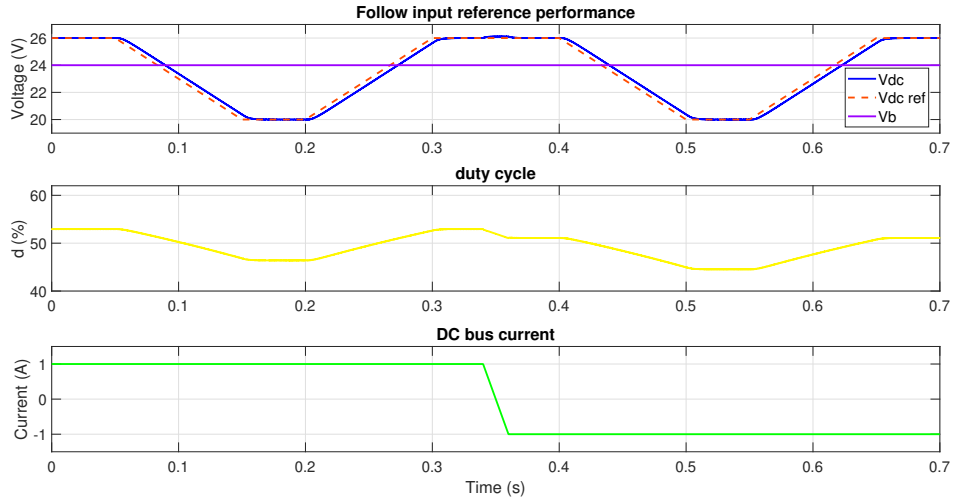


Figure 4.30: Reference tracking performance at $V_b = 24V$

The simulation results using Adaptive LQG reported in figures 4.25 and 4.26 with Lookup table, figures 4.27 and 4.28 with polynomial interpolation and figures 4.29 and 4.30 with the online method, demonstrate the correct operation of the proposed solutions under charge and discharges modes and reference tracking. The previous results show a delay of 7 ms between the bus voltage change and the reference signal, which is shorter than the settling time defined for the controller

design (10 ms). Moreover, the simulations also confirm that the control signal d is always far from the saturation limits. Therefore, the proposed adaptive controller ensures the global stability of the Sepic/Zeta converter in a large operation range, and achieving the desired performance for any commercial battery and microgrid requirements.

4.5.3 Simulations for state observer and electrical efficiency

Figure 4.31 reports the performance of the state observer under a step change on the reference at $V_b = 24\text{ V}$. In this case the duty cycle d changes a 10 % around the steady state value (0.464), and the estimated states converge to conditions close to the real values. The results show small steady-state errors, which do not significantly affect the controller performance as observed in both the simulations and experimental results of the complete system.

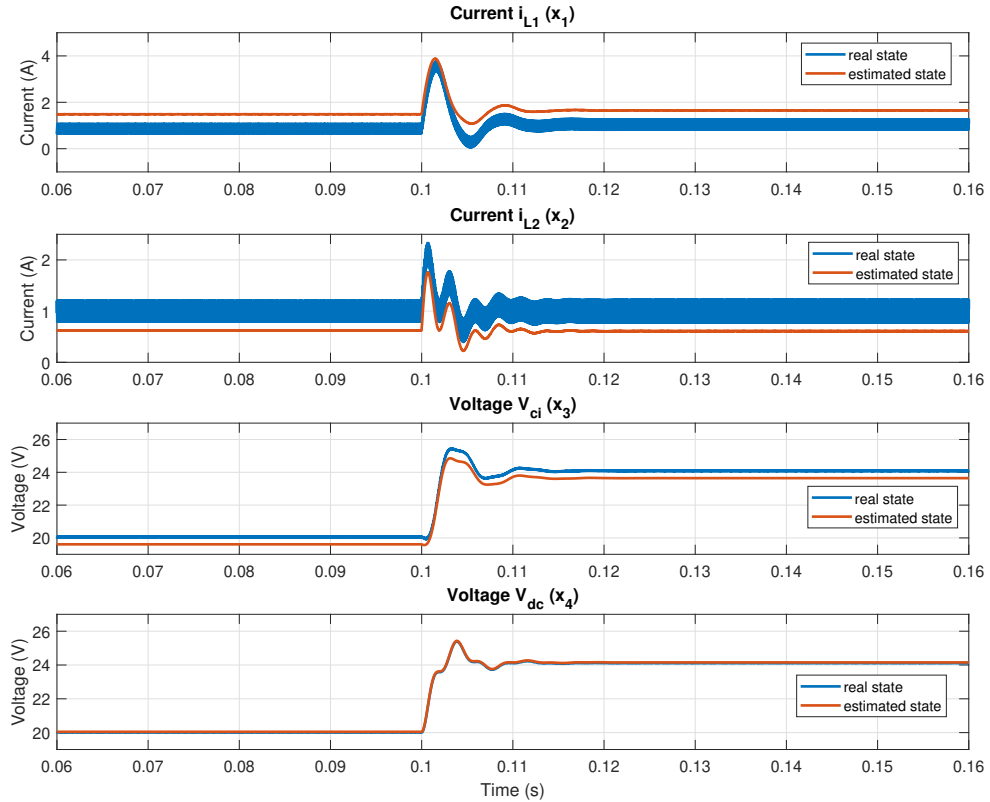


Figure 4.31: State observer performance under step changes at $V_b = 24\text{ V}$

Figure 4.32 shows the plot of electrical efficiency according to the variation of the bus voltage (V_{dc}) and the battery voltage (V_b). This is done according to the elements obtained in the converter design procedure. It is evident that the minimum

efficiency obtained corresponds to a value around 94 %. In this way, the minimum efficiency requirement for the entire defined operating range is achieved.

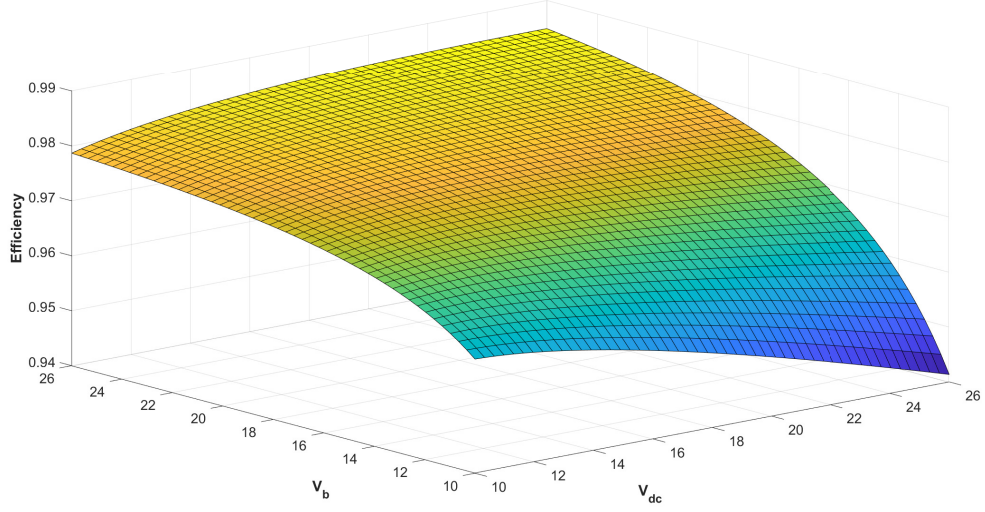


Figure 4.32: Variation of electrical efficiency in the defined voltage range

The next chapter carry out the validation of the proposed solution using an experimental prototype.

Chapter 5

Experimental validation

The results obtained in the simulations are adequate, however, they do not include all the effects present in reality. Therefore, to validate the design procedure and the proposed control strategies, the battery charger/discharger designed for the case study in the previous chapter is experimentally implemented.

5.1 Experimental implementation

Figure 5.1 shows the scheme of the test bench designed to validate the battery charger/discharger, which have the following blocks:

- Power stage: consists of a 12V battery, the bidirectional Sepic/Zeta converter and the DC bus. A connection of the commercial electronic load (BK 8514) and the commercial source (BK XLN6024), using rectifier diodes in parallel to avoid reverse currents, is used to emulate a controllable dc bus.
- Sensing hardware: LV25P voltage sensors are used to measure both the bus and battery voltages. The outputs of those sensors are isolated and scaled to the levels required by the control card.
- Control Card: the Texas Instruments Delfino TMS320F28379D control card is used to implement the adaptive controller and state observer. This control card has a 32-bit double core processor of the C2000 series at 200 MHz, four Analog-to-Digital Converters (ADC), three 12-bit buffered Digital-to-Analog (DAC) outputs, 24 Pulse Width Modulator (PWM) channels, input/output digital signals, among other features. In this experimental validation, a PWM module is used to generate the gate signals of the MOSFETs M_1 and M_2 , also introducing the dead-time required between both MOSFETs according to the datasheet. The control signal d is extracted from the control card,

using a DAC channel, for monitoring purposes, and the V_{dc} and V_b values are acquired using two ADC channels. In terms of software, the proposed adaptive controllers was implemented using the Embedded Coder Support Package for Texas Instruments C2000 Processor available in Matlab, which allows to generate C code from a Simulink model.

- MOSFET drivers: two opto-coupled HCPL 3120 drivers have been used to generate the isolated gate voltages for each MOSFET.

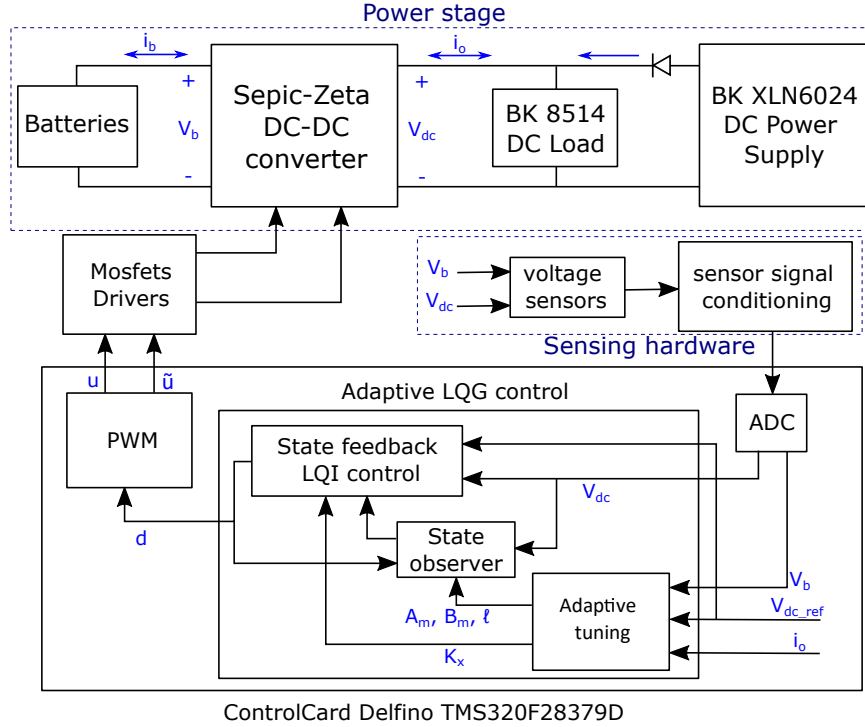


Figure 5.1: Test bench scheme of the battery charger/discharger

The experimental prototype is shown in figure 5.2, where the power stage, sensors and control card are observed. Figure 5.3 shows the complete experimental setup, which includes both the battery and the emulated microgrid bus. Such an experimental prototype is subjected to same test conditions previously defined for the simulations, including multiple battery modes and converter operation conditions.

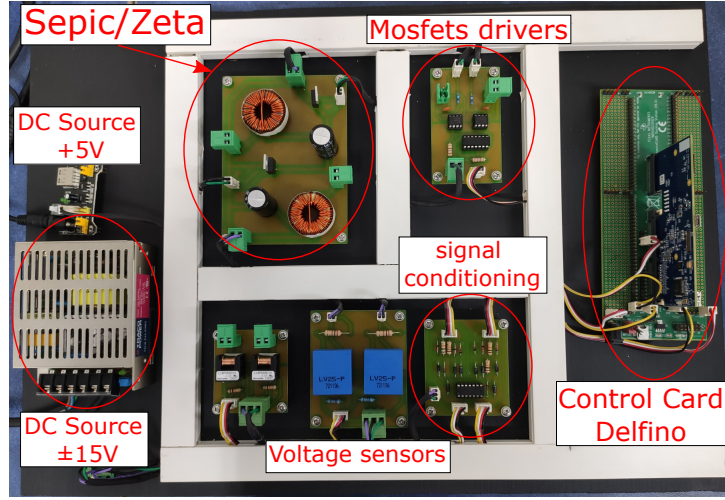


Figure 5.2: Charger/discharger prototype

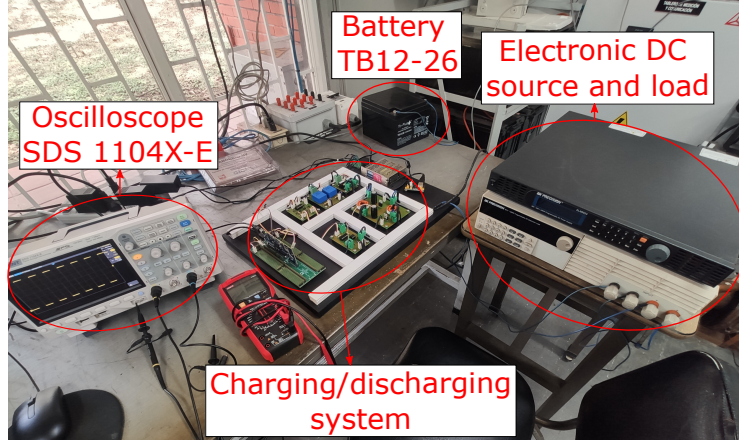


Figure 5.3: Experimental setup

5.2 Experimental results for bus voltage regulation with offline parameter adjustment

For a correct comparison, the same test conditions applied in the simulations are executed for the experimental validation. The disturbance profile includes upshifts, downshifts, shifts between charge, discharge and standby modes. Initially the proposed offline adaptive parameter methods were tested. Figures 5.4a) and 5.4b) show the performance of the charger/discharger using the adaptive LQG with Lookup table and the adaptive LQG with polynomial interpolation, where the bus voltage is regulated at 10 V (buck mode) for charging, discharging and stand-by mode. These

tests demonstrate the correct operation in terms of regulation of the offline parameter adjustment methods, and no appreciable differences are observed between the two methods.

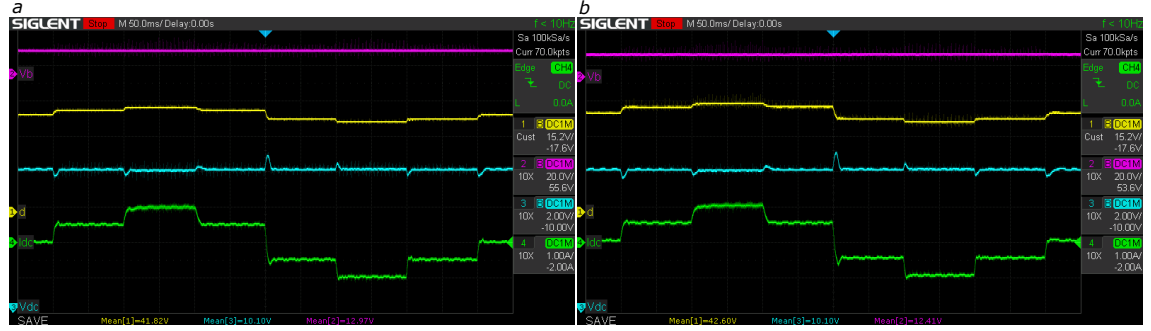


Figure 5.4: Bus regulation with $V_b = 12\text{ V}$ and reference $V_{dcref} = 10\text{ V}$ (Buck mode) using Adaptive LQG a) with Lookup-table, b) with polynomial interpolation. Magenta: battery voltage V_b , yellow: duty cycle d , cyan: bus voltage V_{dc} , green: bus current i_o

The previous test is repeated for voltage regulation at 12 V (unitary gain mode) and 16 V (Boost mode), with both offline methods. Those experimental results are shown in the figures 5.5 and 5.6, where an adequate performance is appreciated under different conditions of the DC bus current and different modes of the converter.

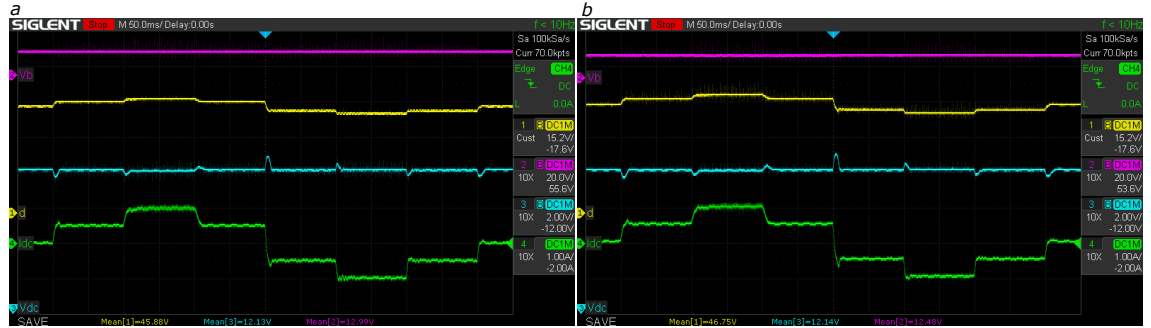


Figure 5.5: Bus regulation with $V_b = 12\text{ V}$ and reference $V_{dcref} = 12\text{ V}$ (unitary gain mode) using Adaptive LQG a) with Lookup-table, b) with polynomial interpolation. Magenta: battery voltage V_b , yellow: duty cycle d , cyan: bus voltage V_{dc} , green: bus current i_o

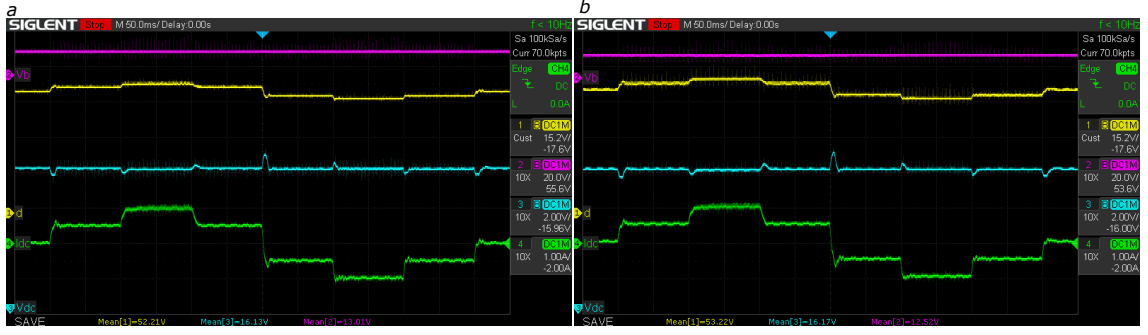


Figure 5.6: Bus regulation with $V_b = 12\text{ V}$ and reference $V_{dc_{ref}} = 16\text{ V}$ (Boost mode) using Adaptive LQG a) with Lookup-table, b) with polynomial interpolation. Magenta: battery voltage V_b , yellow: duty cycle d , cyan: bus voltage V_{dc} , green: bus current i_o

The next test considers the use of two series-connected batteries ($V_b = 24\text{ V}$). For this case, DC bus voltage references of 20 V (Buck mode), 24 V (unitary gain mode) and 26 V (Boost mode) are tested. Figures 5.7, 5.8 and 5.9, show the experimental results of the charger/discharger with both offline adaptive parameters methods.

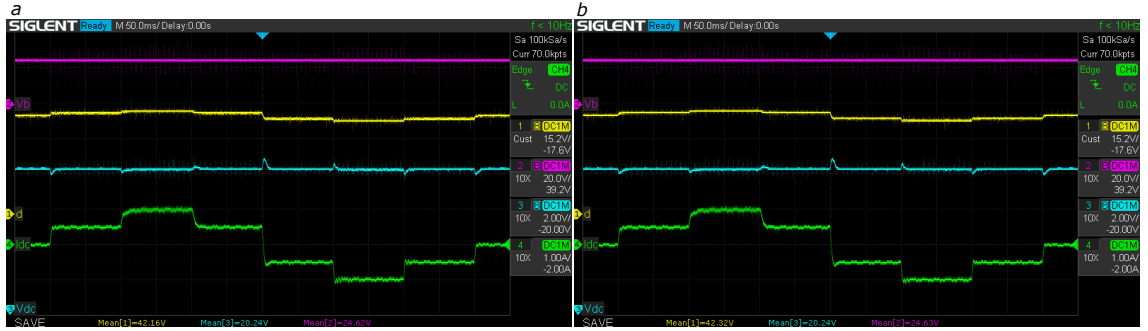


Figure 5.7: Bus regulation with $V_b = 24\text{ V}$ and reference $V_{dc_{ref}} = 20\text{ V}$ (Buck mode) using Adaptive LQG a) with Lookup-table, b) with polynomial interpolation. Magenta: battery voltage V_b , yellow: duty cycle d , cyan: bus voltage V_{dc} , green: bus current i_o



Figure 5.8: Bus regulation with $V_b = 24\text{ V}$ and reference $V_{dc_{ref}} = 24\text{ V}$ (unitary gain mode) using Adaptive LQG a) with Lookup-table, b) with polynomial interpolation. Magenta: battery voltage V_b , yellow: duty cycle d , cyan: bus voltage V_{dc} , green: bus current i_o

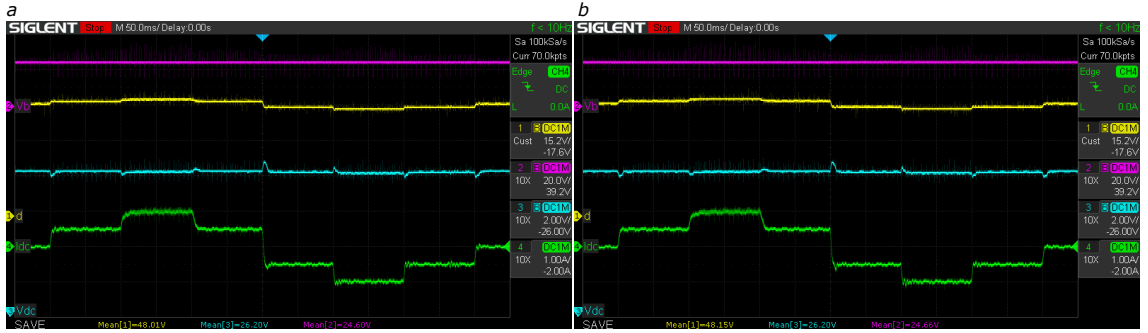


Figure 5.9: Bus regulation with $V_b = 24\text{ V}$ and reference $V_{dc_{ref}} = 26\text{ V}$ (Boost mode) using Adaptive LQG a) with Lookup-table, b) with polynomial interpolation. Magenta: battery voltage V_b , yellow: duty cycle d , cyan: bus voltage V_{dc} , green: bus current i_o

The experimental results of the DC bus voltage regulation, using the offline controller parameter adjustment, show the correct performance of the system under charge/discharge conditions and in the different modes of the converter (Buck, Boost and unitary gain). In addition, the variation of the battery voltage allowed to evaluate the stability of the controller at different operating points.

5.3 Experimental results for bus voltage regulation with online parameter adjustment

The experimental validation of the proposed control system using the online adaptive LQG also considers the comparison with a classical solution, i.e. a proportional-integral (PI) control law, which illustrates the adaptability of the new solution. The

PI controller was designed at the operating point ($V_b = 12\text{ V}$ and $V_{dc} = 16\text{ V}$) using the pole placement technique. Figure 5.10a) shows the performance of the charger/discharger under the action of the PI controller, where the bus voltage is regulated at 10 V (buck mode) for charging, discharging and stand-by mode. The experiment considers bus currents of -1.0 A , -0.5 A , 0.0 A , 0.5 A and 1.0 A with step changes, which is the same bus current profile considered in the simulations. The experiment shows that the PI controller is unable to regulate the bus voltage for the deep charging condition of -1.0 A , which is caused by the large deviation from the operating point used to design the PI controller. In contrast, figure 5.10b) shows the experimental performance of the charger/discharger under the action of the proposed adaptive controller, where the bus voltage is successfully regulated for all the operation conditions. Moreover, the adaptive LQI controller provides a much stable bus voltage, also exhibiting smaller overshoots and shorter settling times.

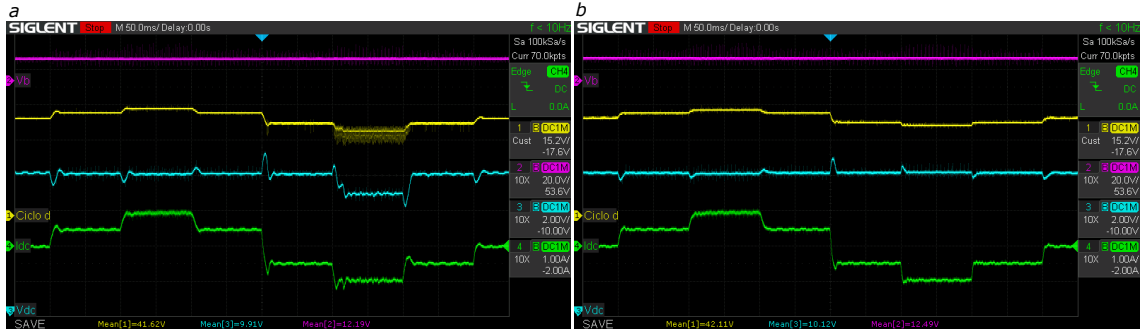


Figure 5.10: Bus regulation with $V_b = 12\text{ V}$ and reference $V_{dc_{ref}} = 10\text{ V}$ (Buck mode) using a) PI, b) Adaptive LQG controller. Magenta: battery voltage V_b , yellow: duty cycle d , cyan: bus voltage V_{dc} , green: bus current i_o

The same previous test was performed for unitary gain (12 V) and boost (16 V) modes to validate the controller capability to adapt to different bus voltages without any modification. The results of those experiments are reported in figures 5.11 and 5.12, which put into evidence the satisfactory performance of the proposed solution (figures b). In addition, those results also show the improved performance of the adaptive LQI controller over the classical PI controller in all operation conditions: the new solution provides smaller overshoots and shorter settling times.

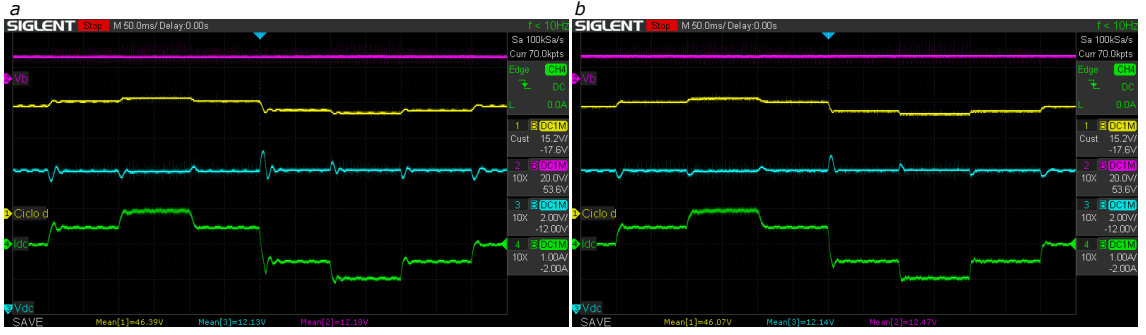


Figure 5.11: Bus regulation with $V_b = 12\text{ V}$ and reference $V_{dc_ref} = 12\text{ V}$ (unitary gain mode) using a) PI, b) Adaptive LQG controller. Magenta: battery voltage V_b , yellow: duty cycle d , cyan: bus voltage V_{dc} , green: bus current i_o

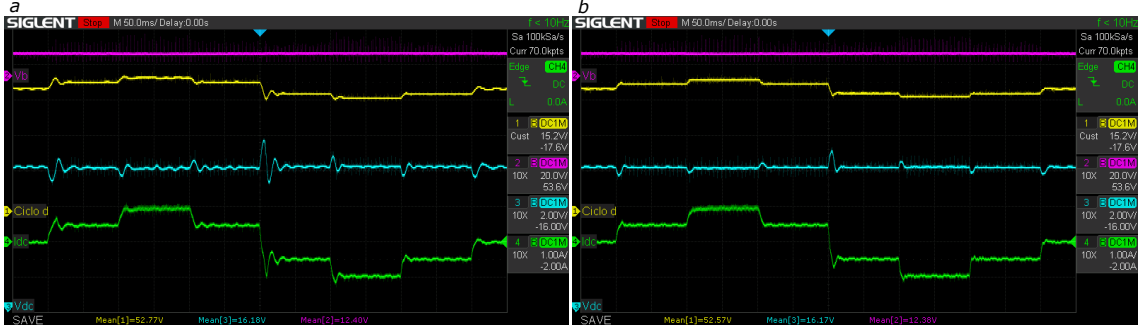


Figure 5.12: Bus regulation with $V_b = 12\text{ V}$ and reference $V_{dc_ref} = 16\text{ V}$ (boost mode) using a) PI, b) Adaptive LQG controller. Magenta: battery voltage V_b , yellow: duty cycle d , cyan: bus voltage V_{dc} , green: bus current i_o

The next set of experiments are aimed to validate the capability of adapting to changes on the battery voltage without any modification. Those experiments consider two batteries in series connection, thus increasing V_b to 24 V . The experiments consider the same bus current profile used in the previous tests, which forces the battery pack to operate in charge, discharge and stand-by modes. Figures 5.13, 5.14 and 5.15 report the experimental results for bus voltages equal to 20 V (buck mode), 24 V (unitary gain mode) and 26 V (boost mode), respectively. The satisfactory performance of the proposed controller confirms the adaptability of the new solution to changes on both the battery and the bus voltages, even under different operation modes of the battery. In addition, the experimental behavior of the proposed battery charger/discharger for a single battery (12 V) or a two-battery pack (24 V) put into evidence the improved performance over the classical PI solution. Moreover, all the experiments confirmed the correct duty cycle operation, which is never saturated.

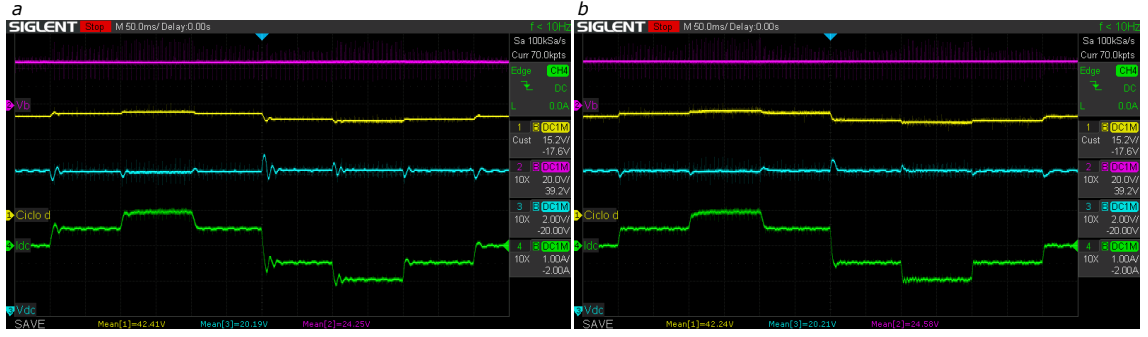


Figure 5.13: Bus regulation with $V_b = 24\text{ V}$ and reference $V_{dc_{ref}} = 20\text{ V}$ (buck mode) using a) PI, b) Adaptive LQG controller. Magenta: battery voltage V_b , yellow: duty cycle d , cyan: bus voltage V_{dc} , green: bus current i_o

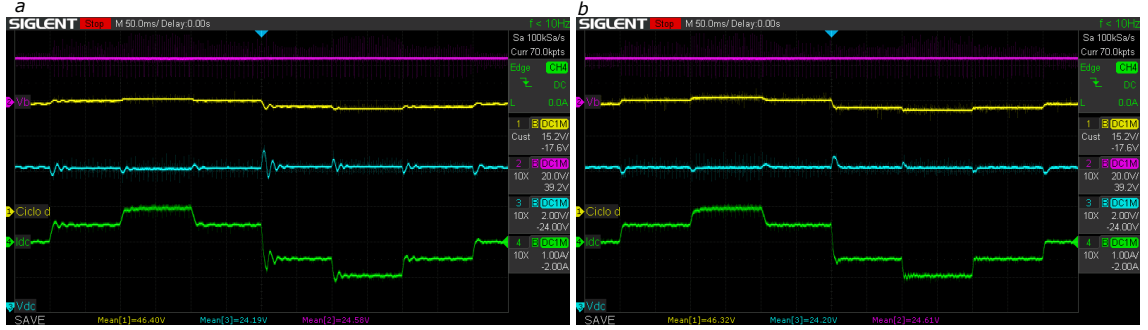


Figure 5.14: Bus regulation with $V_b = 24\text{ V}$ and reference $V_{dc_{ref}} = 24\text{ V}$ (unitary gain mode) using a) PI, b) Adaptive LQG controller. Magenta: battery voltage V_b , yellow: duty cycle d , cyan: bus voltage V_{dc} , green: bus current i_o

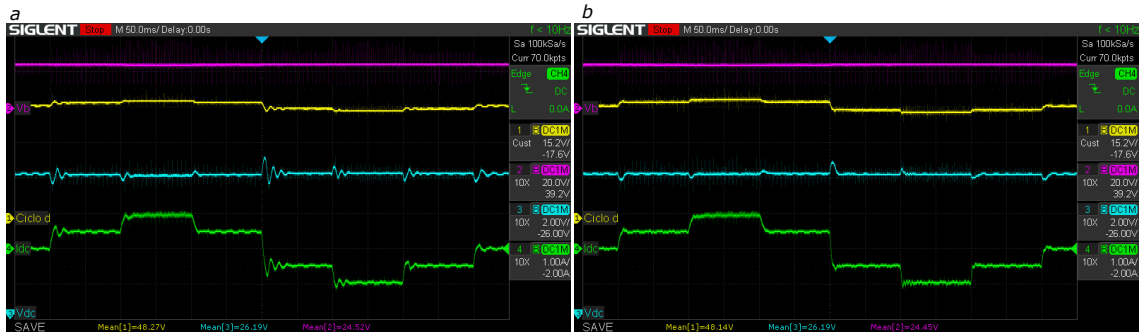


Figure 5.15: Bus regulation with $V_b = 24\text{ V}$ and reference $V_{dc_{ref}} = 26\text{ V}$ (boost mode) using a) PI, b) Adaptive LQG controller. Magenta: battery voltage V_b , yellow: duty cycle d , cyan: bus voltage V_{dc} , green: bus current i_o

Table 5.1 summarizes the experimental performance of the proposed solution and the classical PI controller. The data confirm that the adaptive LQI controller

always provides a maximum overshoot lower than 10 % and settling times shorter than 10 *ms*, thus fulfilling the design criteria. On the contrary, the PI controller exhibits poor performance in the majority of the tests, where a maximum overshoot of 12 % is observed, and longer settling times are also provided. Moreover, the proposed controller regulates the bus voltage in all the operation conditions, which is not ensured by the PI controller: for example, the PI controller was unable to regulate the bus voltage in charge mode with a bus current of 1 *A* (see Figure 5.10 a)), where the bus voltage has an steady-state error higher than 10 %.

Table 5.1: Dynamic performance of experiments for bus voltage regulation.

Test		Overshoot (%)		Settling Time (ms)	
V_b (V)	V_{dc} (V)	PI	Adaptive LQG	PI	Adaptive LQG
12	10	12	8.75	13.2	9
	12	9	7	18	8.8
	16	9.49	5.85	24.2	8.7
24	20	4.67	3.16	10.55	7.22
	24	4.09	2.63	11.33	7.46
	26	3.63	2.63	7.1	7.17

Despite the simulation and experimental tests were performed for the same conditions, the experimental performance of adaptive LQI controller exhibits some differences as it is observed in Tables 4.3 and 5.1. Those differences are caused by non-modeled parasitic effects on the wires, the internal resistances of the capacitors, parasitic inductances and capacitances on the MOSFETs, among other sources. However, both simulated and experimental implementations of the proposed solution fulfill the designed criteria, which confirms the viability of the solution for commercial applications.

5.4 Experimental results for reference tracking with the offline methods

The final set of experiments was conducted to verify the performance of the proposed solution in tracking a changing reference. These new experiments consider a particular battery mode (charge, discharge, stand-by) and apply a change on the bus voltage reference to force the converter into buck and boost modes. The battery charger/dischargers starts operating with a bus voltage of 10 *V* (buck mode), changing to 16 *V* (boost mode), and returning to buck mode again. The tracking experimental results for the adaptive adjustment of the controller parameters with

Lookup table and polynomial interpolation methods are shown in the figure 5.16, where a single battery is used ($V_b = 12\text{ V}$). The proposed offline adaptive controllers guarantee a correct tracking of the references under different operating points in the different modes of the converter (Buck and Boost).

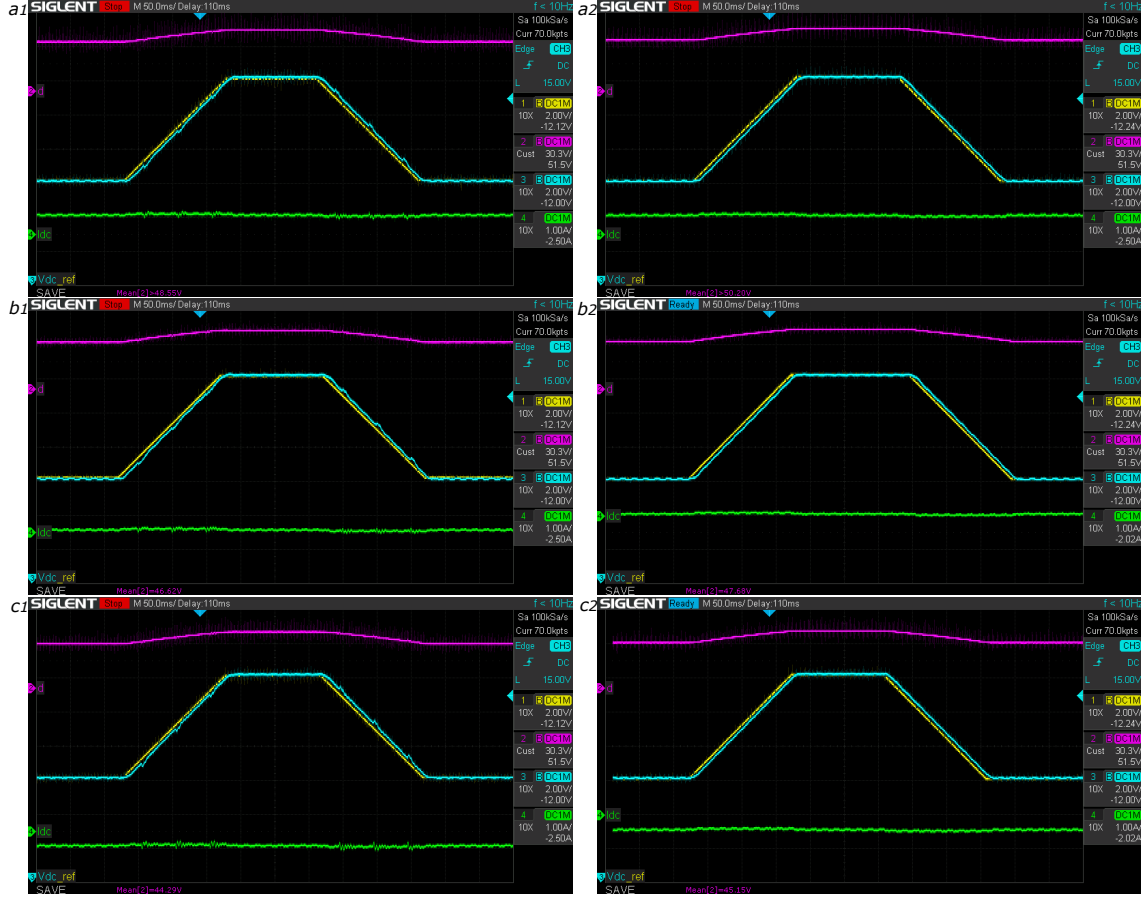


Figure 5.16: Reference tracking performance at $V_b = 12\text{ V}$ using Adaptive LQG with Lookup-table: a1) discharge mode, b1) stand-by mode, c1) charge mode, and with polynomial interpolation: a2) discharge mode, b2) stand-by mode, c2) charge mode. Magenta: duty cycle d , yellow: bus voltage reference V_{dcref} , cyan: bus voltage V_{dc} , green: bus current i_o

Also, the tracking tests are evaluated with two batteries connected in series ($V_b = 24\text{ V}$). For this case the bus voltage reference starts at 20 V (Buck mode), later changes to 26 V (Boost mode), and finally returns to 20 V . Figure 5.17 shows the experimental behavior of the charger/discharger with the proposed offline adaptive methods. The same performance as the tests with a battery are observed, those results show the correct performance of the controller evaluating different operating points.

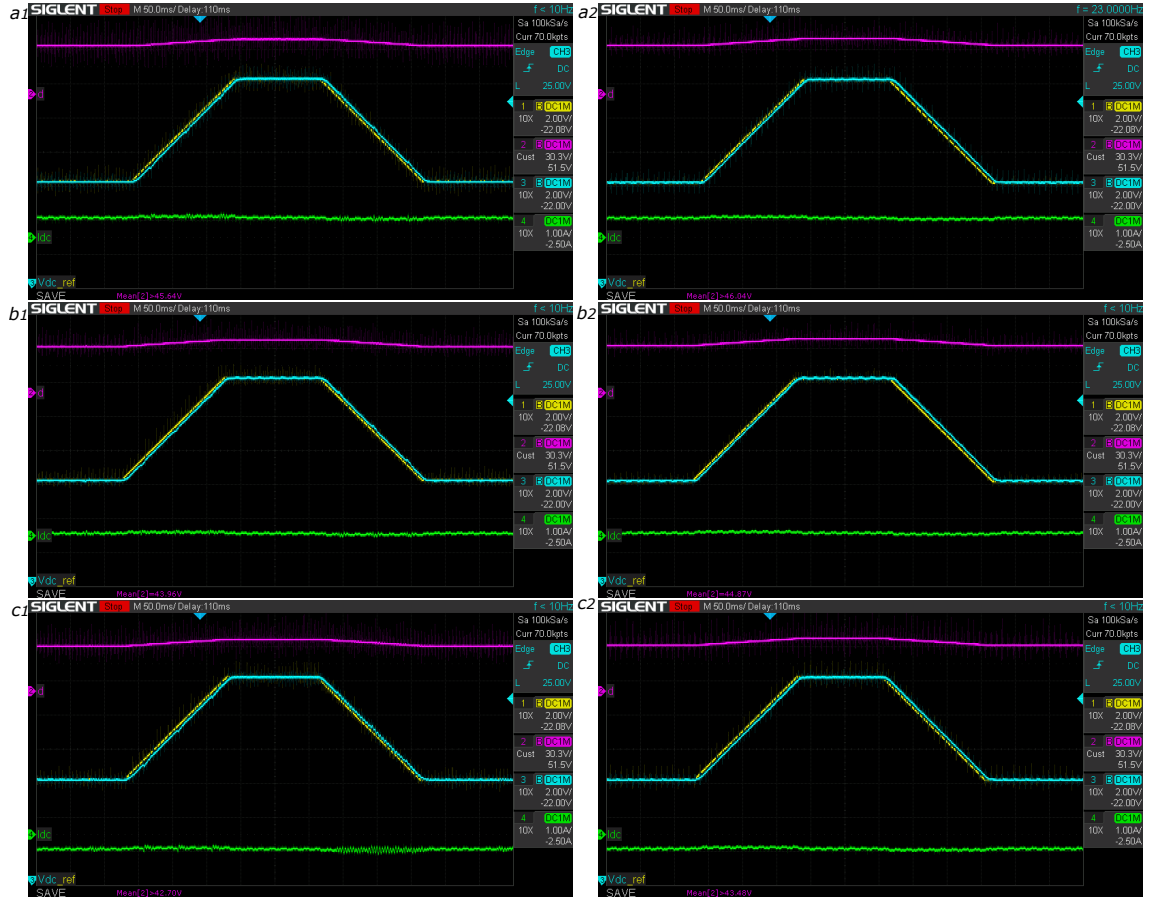


Figure 5.17: Reference tracking performance at $V_b = 24\text{ V}$ using Adaptive LQG with Lookup-table: a1) discharge mode, b1) stand-by mode, c1) charge mode, and with polynomial interpolation: a2) discharge mode, b2) stand-by mode, c2) charge mode. Magenta: duty cycle d , yellow: bus voltage reference V_{dcref} , cyan: bus voltage V_{dc} , green: bus current i_o

5.5 Experimental results for reference tracking with online adjustment

The same conditions of the previous tests are evaluated with the Adaptive LQG online. First, it is evaluated with a single battery $V_b = 12\text{ V}$ with changes of bus reference between 12 V (Buck mode) and 16 V (Boost mode), resulting in the experimental behavior shown in figure 5.18. Subsequently, it is evaluated with two batteries connected in series $V_b = 24\text{ V}$, giving rise to the experimental results shown in figure 5.19. As in the previous experimental tests, the proposed adaptive controller ensures a correct regulation of the bus voltage under any operation condition. Furthermore, the duty cycle is never saturated, which ensures the system

controllability

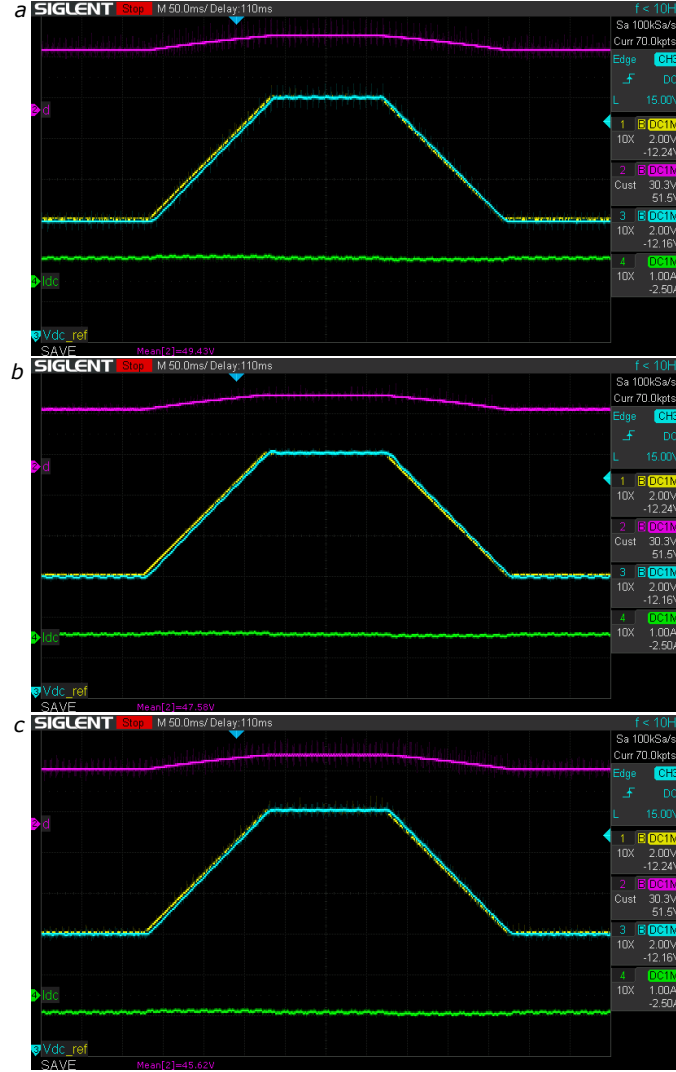


Figure 5.18: Reference tracking performance at $V_b = 12$ V: a) discharge mode, b) stand-by mode, c) charge mode. Magenta: duty cycle d , yellow: bus voltage reference V_{dcref} , cyan: bus voltage V_{dc} , green: bus current i_o

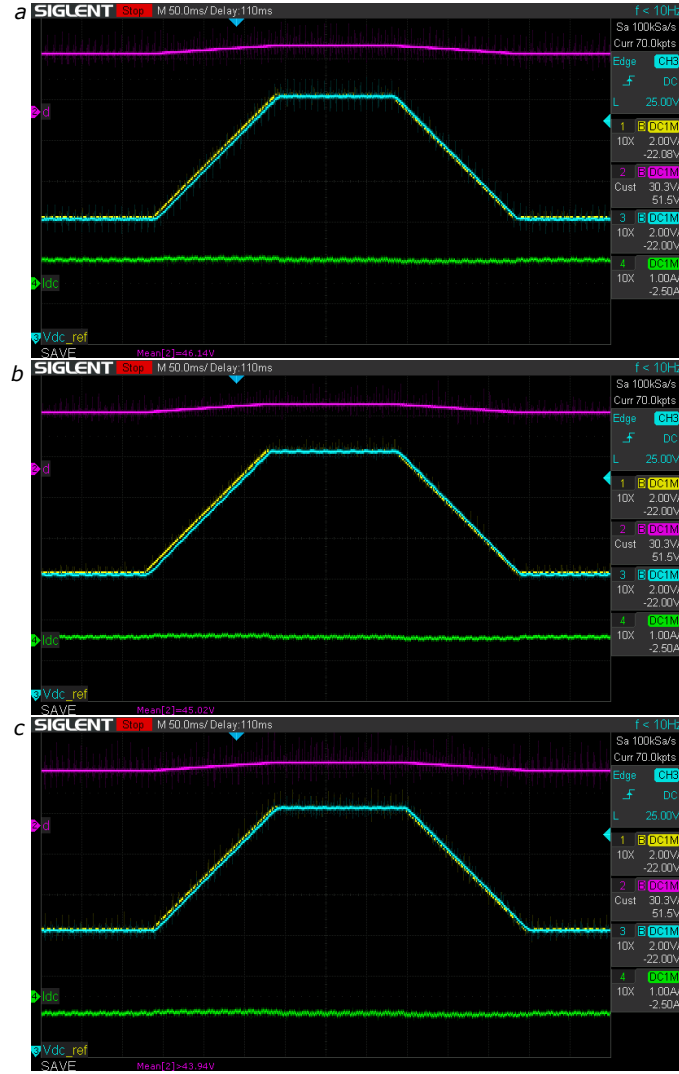


Figure 5.19: Reference tracking performance at $V_b = 24\text{ V}$: a) discharge mode, b) stand-by mode, c) charge mode. Magenta: duty cycle d , yellow: bus voltage reference V_{dcref} , cyan: bus voltage V_{dc} , green: bus current i_o

Finally, the experimental tests presented in this chapter have demonstrated the following conditions:

- The proposed adaptive LQG offline controllers are suitable for adjusting the controller parameters and guaranteeing a correct performance under any condition of the battery (charge, discharge, standby) and the converter (buck, boost, unitary gain), within the designed operating range.
- The proposed adaptive LQG online controller is able to adjust the parameters to ensure global stability and the desired performance under any operation

condition of the battery (charge, discharge, stand-by) and the converter (buck, boost, unitary gain).

- The experiments and simulation fulfill the design criteria, hence the design procedure proposed in this document is correct. Moreover, this also confirms the validity of the mathematical model developed for control purposes.
- It is possible to implement the proposed adaptive LQG controllers using commercial devices, which ensures the commercial viability of the proposed solution.
- The proposed controllers provide a better performance in comparison with classical solutions like PI controllers, which are not able to ensure global stability.

Chapter 6

Conclusions and future work

6.1 Conclusions

In this research work a battery charger/discharger able to regulate the bus voltage under different operation conditions and perturbations was presented. The charger/discharger is based on the bidirectional Sepic/Zeta topology to enable the interaction of microgrid buses and battery packs with any voltage. The proposed mathematical model and design process include the main parasitic losses, thus providing high accuracy and effectiveness.

The designed LQG adaptive control includes an state observer to avoid many sensors, and it was designed to ensure zero steady-state error and system stability over a large operating range. Moreover, the adaptive controllers guarantee the desired performance of the charger/discharger in different battery and converter modes, thus the proposed solutions are able to operate under different gain relations among bus voltage and battery voltage, hence it is suitable to produce a general battery interface for any commercial storage and microgrid devices.

Due to the performance of the adaptive parameter method with polynomial interpolation, it is a suitable alternative that does not consume significant computational resources, this compared to the online adaptive parameter method which requires solving the Riccati's differential equation constantly. However, it does not guarantee a correct performance and overall stability outside the designed operating range, in contrast to the online tuning method that calculates the optimal controller for any operating point.

The core of the designed system is a set of matrix algebraic operations and integrators, which are used to implement the proposed adaptive controllers. This solution was demonstrated to be suitable for implementation in many processing platforms such as DSPs, FPGAs, microcontrollers, among others. Such a condition was verified using an experimental implementation, which implemented the power

stage using commercial hardware, and the control system was implemented using a commercial microprocessor. The experiments conducted verified the accuracy of the design procedure, where the main design criteria were fulfilled (maximum overshoot and settling time). Moreover, the experiments also confirmed the advantages over classical linear controllers: both global stability and desired performance are achieved in any operation condition (charge, discharge, stand-by modes of the battery; and buck, boost and unitary gain modes of the converter). Therefore, the proposed solution is suitable to interface any battery pack with any microgrid. In fact, it is useful to replace series-connected battery packs with single batteries, since the boosting factor of the Sepic/Zeta converter will remove the need of a battery balancing system commonly required in series connection. In addition, such a modification will allow to hot-swapping any battery from the microgrid, thus reducing the maintenance stops.

In addition, the delay in reference tracking can be reduced by modifying the dynamic criteria adopted for the controller design, in particular reducing the settling time of the bus voltage. However, it must be taken into account that reducing the settling time also could increase the maximum overshoot, which could lead to undesired (or even dangerous) overvoltage conditions in the dc bus. Therefore, the complete dynamic performance must be analyzed.

6.2 Future work

In this implementation, the adaptive tuning was performed considering the worst case for the output current i_o , thus it is introduced as a constant. However, i_o could be measured, thus not inserted as a constant (worst-case) value. This is a future improvement which will require to add a current sensor, thus increasing the system complexity.

A more exhaustive analysis about the efficiency of the Sepic/Zeta converter can be carried out. More variables of power losses in the circuit can be considered, in such a way that they can contribute to the design and improve the efficiency of the converter. Consequently, the electrical efficiency results can be analyzed in a comparative way with other solutions.

Adaptive adjustment methods present a similar dynamic performance. Therefore, a comparative analysis of the adaptive methods in terms of the computational power required and the implementation costs can be performed.

Finally, an additional future work concerns the inclusion of the battery charging/discharging profile into the control system, which could be operated in agreement with secondary control loops of the microgrid.

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