Real-Time Implementation of Input-State Linearization and Model Predictive Control for Robust Voltage Regulation of a DC-DC Boost Converter

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ABSTRACT The use of DC-DC step-up converters has significantly increased due to their implementation as power interfaces in microgrids (MGs), smart grids (SGs) and electrical vehicles. Step-up converters adapt the source voltage or current to the load specifications through an appropriate control algorithm, which is linear in most cases. However, linear algorithms mostly guarantee the system's stability and desired performances only around a relatively small neighborhood of the equilibrium point. Model predictive controllers (MPCs) have been proposed to improve the performance of the converter and broaden its operating region. However, MPCs have mostly been based on an approximated linear model of the converter, which contributes to a relatively narrow operating region. This work proposes an MPC algorithm based on an exactly linearized converter model. The converter model is linearized according to an exact input-state linearization control (ILC). To the best of our knowledge, this is the first work to present a real-time implementation of the ILC in the context of nonlinear DC-DC boost converter control. The objective of exact linearization is to continue using the same reduced-complexity linear MPC while extending the operation area of the system compared to classic linear control. Simulations and experimental results show that the static and dynamic performances of the proposed control are significantly better than those of the standard linear control.

INDEX TERMS Real-time implementation, model predictive control (MPC), nonlinear control (NLC), input-state linearization control (ILC).

I. INTRODUCTION
Step-up DC-DC converters are attracting increasing interest in academia and industry, especially with their use in converting DC renewable power in microgrids (MGs) and smart grids (SGs) [1]–[4] and in electric vehicles. The main function of step-up converters is to amplify the input source voltage or current to the level required by the load through an appropriate algorithm. Control algorithms for step-up converters can be classified into two categories: linear- and nonlinear-model-based. Linear algorithms are designed based on an approximated linear model of the converter derived around a specific equilibrium point. Polynomial and linear state space controllers are among the most well known in this category. These algorithms can satisfy acceptable dynamic and static performances while guaranteeing stability around the equilibrium point. However, mismatches in the system parameters or signal disturbances can cause the converter to operate too far from the specified equilibrium point. This situation can degrade the output performance or even cause system instability. Linear-model-based predictive controllers (LMPCs) have been proposed as a solution [5]–[7] to extend the operating region of this type of converter. LMPCs use a converter linear state space model to predict the future values of the state variables. The control signal is obtained by minimizing a cost function subject to constraints on the system state variables and inputs. An LMPC was first used in the process industry thirty years ago. Then, due to increased microprocessor performances, LMPCs were extended to the real-time control of power converters [6]. However, the use of the approximated linear model in an LMPC can reduce the operating area of the converter and limit its robustness. Other recent linear-model-based algorithms include dual-mode LQR-feedforward optimal control [8], internal model control [9] and nested reduced-order PI observers [10]. Those algorithms share the common objective...
of regulating the output voltage and mitigating the nonminimum phase zero (NMP) effect on the system performance. The algorithms improved the output dynamics and static performance when there were variations in the source voltage or the resistive load. However, the reduction in the inductor start-up inrush current was not always addressed. In [8], the output voltage continued showing permanent ripples of approximately 25%. In [9], for the reference tracking situation, the output voltage continued showing a transient overshoot. For the load-halving case, the output voltage continued showing either an undershoot close to that of the PID [9] or an overshoot of approximately 4.5% [10].

In contrast, nonlinear control algorithms are based on a system nonlinear model, which is considered to more realistically describe the converter’s behavior and can improve the system robustness in the case of disturbances or parameter mismatches [11], [12].

In this category, linearizing control (LC) [13] enables the transformation of a nonlinear state space affine model into its exact linear real-time equivalent using specific transformation functions. The linearized system can then be regulated using a linear outer-loop algorithm. Thus, the closed-loop system shows improved robustness because the equivalent model more accurately represents the converter than an approximated linearization. Among the LC controllers, the input-output linearization derives a linearized system composed of two sets of state variables, which are called external and internal dynamics. The internal dynamics are unstable in the case of NMP systems such as buck, boost and buck-boost DC-DC converters. If the outer loop does not handle such dynamics, the system is closed-loop unstable. As a solution to this problem, researchers in [14] proposed a controller composed of an input-output linearization outer loop and a nonlinear stabilizing inner loop. The controller was designed using a Lyapunov scheme to stabilize both the external and the internal dynamics. Another solution to the NMP is the use of an input-state linearization control (ILC). This algorithm performs an exact linearization of the full nonlinear affine state space model. Thus, the internal dynamics that might cause instability are avoided. One of the drawbacks of this control is its relative complexity for real-time implementation. To the best of our knowledge, to date, there has been no prior experimental implementation of ILC for the control of a nonlinear boost converter. Nonlinear-model-based predictive controllers (NLMPCs) [15], [16] have also helped extend the operating area of the system. The closed-system stability was often assessed through Lyapunov methods [17], [18]. However, a permanent state offset remained when the NLMPC was used alone. This offset was corrected often using an outer observation loop or an integral action corrector. In [19], an output voltage-tracking controller for dynamic performance recovery was presented. To maintain the system performance after a disturbance, the controller used nonlinear observers that estimated the disturbances. Nonetheless, the inductor current ripples were close to those of the classic linear control, and the start-up current inrush was not addressed. The controller was tuned according to a trial-and-error procedure. A constrained near-time-optimal sliding-mode control was proposed in [20]. The algorithm used a switched affine analysis of the converter model. The objective of this control is to ensure the robustness of stability of the converter if a reference, load or input voltage changes. However, in the load-halving case, the output voltage continued showing a static error of approximately 1%. The inductor current ripples were above 100% in the start-up test. In [21], a discrete-time sliding-mode-based output voltage regulation was proposed. The control was composed of a current-based discrete-time sliding mode control inner loop and a discrete-time PI outer loop. The control ensured a fast response time and the benefits of digital implementation, such as programmability and noise immunity, and enabled a constant frequency-switching operation. However, the start-up inductor current continued exhibiting an inrush of approximately 2000%. The output voltage also presented a start-up overshoot. In addition, a voltage undershoot of approximately 21% was present in the load-halving case. In [22], a sensorless control using Lyapunov’s method was proposed. The inductor current sensor was replaced by a state observer, which resulted in a system free from the noise problem associated with sensors and the control cost reduction. Improved stability and output voltage tracking performances were achieved for large-signal perturbations. However, the dynamic performance was close to that of a PI, and the inductor current performance was not addressed.

In this work, a nonlinear control that both improves the output voltage regulation and limits the inductor current inrush during start-ups is proposed. To achieve this objective, a linear-model-based MPC outer loop is cascaded with an exact input-state linearizing control inner loop. We combine the classic linear MPC and input-state controllers to improve the conventional MPC robustness to the changes in the source voltage or output load and to offer an improved closed-loop reference tracking ability. The outer loop consists of a linear MPC to ensure the reference voltage tracking and reduce the inductor current start-up inrush by minimizing the inductor current, output voltage-related tracking errors and duty cycle values. The predictive component of the MPC will also help reduce the response time of the system.

The main contributions of this work are as follows:

- The first real-time experimental implementation of a complex and calculation-demanding ILC control in the context of a DC-DC boost converter with a limited-resource microcontroller.
- Improvement in the robustness and tracking ability of the classic linear MPC by cascading it with the internal ILC loop.
- A nonlinear control that both improves output voltage regulation and limits inductor current inrush during start-ups.

The remainder of this paper is organized as follows: in the second section, the affine nonlinear state space model of
the converter is described. In the third section, the input-state linearization algorithm is presented. In the fourth section, the MPC algorithm is outlined. Simulation results are presented in the fifth section to compare the performances of the proposed algorithm with those of the classic linear algorithms. Experimental results are presented in the sixth section to validate the simulation findings. A conclusion is provided in the seventh section to summarize the key points of this work.

II. SWITCHED BILINEAR MODEL OF THE STEP-UP CONVERTER

The switched bilinear model of the step-up converter in equation (1), namely, the CBM, is derived by using Fig. 1 according to the demonstration in [23]. This model is composed of two linear state space models corresponding to two switching intervals of the MOSFET. The model is formulated by applying Kirchhoff’s laws on each of the switching intervals. The state variables are the inductor current and capacitance voltage. The state space representation (1) is nonlinear because the temporal duty cycle \( d(t) \) multiplies the state variables. This representation also has an affine format, which makes it suitable for use with the ILC. This model aims to offer a more realistic and accurate representation of the converter characterisitcs.

\[
\dot{x}(t) = (d(t) * A_{c,1} + (1 - d(t)) * A_{c,2}) * x(t) + (d(t) * B_{c,1}) + (1 - d(t)) * B_{c,2} * v(t)
\]

which can be rewritten as

\[
\dot{x}(t) = A_{c,2} * x(t) + B_{c,2} * v(t) + ((A_{c,1} - A_{c,2}) * x(t) - (B_{c,1} - B_{c,2}) * v(t)) * d(t)
\]

such that

\[
x(t) = [i_L(t), v_c(t), v_D(t)]^T, \quad v(t) = [v_p(t), v_D(t)]^T
\]

and

\[
A_{c,1} = \begin{bmatrix} -(R_L + R_O) / L & 0 & 1 \\ 0 & -1 / RC & 0 \end{bmatrix}, \quad B_{c,1} = \begin{bmatrix} 1 \\ L \\ 0 \end{bmatrix}
\]

\[
A_{c,2} = \begin{bmatrix} -(R_L + R_D) / L & 0 & 1 \\ 0 & -1 / RC & 0 \end{bmatrix}, \quad B_{c,2} = \begin{bmatrix} 1 \\ L \\ 0 \end{bmatrix}
\]

\( R_L, \ R_O, \ R_D \): Resistances of the inductance, MOSFET passing state and Diode passing state, respectively.

An approximated linear transfer function \( v_c(s) / d(s) \) can be derived from (1) using a Taylor series expansion. For DC-DC step-up converters, \( v_c(s) / d(s) \) will have zeros that have strictly positive real parts, which makes the converter a nonminimum phase type. Thus, ILC is more suitable for our system than an input-output linearization [14]. The linearized system with ILC has identical dimensions to the original nonlinear system. Thus, the linearized system in this case does not contain internal dynamics that can cause instability. In the next section, the ILC algorithm will be developed for the bilinear converter.

III. INPUT-STATE LINEARIZING CONTROL

The ILC loop provides an exact real-time linear equivalent model [13]. In contrast to the classic approximated linearization tools such as Jacobian matrix and gradient methods, the ILC gives an exact linear equivalent for the nonlinear system. The equivalent model calculation with ILC is performed through state variable transformations. The linear equivalent model can then be controlled by one of the standard linear control strategies such as PI or state feedback. Fig. 2 illustrates a diagram of our complete control algorithm, where the specific equation numbers are provided for each step. In our case, (2) can be rewritten into the following format:

\[
\dot{x} = f(\bar{x}) + g(\bar{x}) * u(t)
\]

such that

\( f, \ g \): Continuously derivable functions on \( \mathbb{R}^n \),

\( n \): Number of state variables,

and

\[
f(\bar{x}) = A_{c,2} * x(t) + B_{c,2} * v(t) \quad g(\bar{x}) = (A_{c,1} - A_{c,2}) * x(t) - (B_{c,1} - B_{c,2}) * v(t) \quad u(t) = d(t)
\]

The converter bilinear model in (3) is input-state linearizable if and only if i) and ii) are satisfied:

i) \( \{g, \ ad_f g, \ldots, ad_f^{n-1} g\} \) are linearly independent,

ii) \( \{g, \ldots, ad_f^{n-2} g\} \) is involutive,

i.e., \( \forall i, j: \) If \( \{g, \ldots, ad_f^{n-2} g\} = \{f_1, \ldots, f_{n-2}\} \) then \( \text{rank}(f_1, \ldots, f_{n-2}) = \text{rank}(f_1, \ldots, f_{n-2}, ad_f f_j) \).

Function \( ad_f g \) is the ith-order Lie brackets of \( f \) and \( g \) and defined in (4). These brackets are mainly used to reduce the length of partial derivative formulas, as they represent an equivalent shorter expression of the original longer partial derivative formula.

\[
ad_f^0 g = g
\]

\[
ad_f^1 g = \dfrac{\partial g}{\partial x} * f(x) - \dfrac{\partial f}{\partial x} * g(x)
\]

\[
ad_f^2 g = ad_f(ad_f^{n-1} g)
\]
There is a linearizing control such that (2) can be expressed in the equivalent following LTI format:

\[
\begin{align*}
\dot{z}_1 &= z_2 \\
\dot{z}_2 &= z_3 \\
\vdots \\
\dot{z}_{n-1} &= z_n \\
\dot{z}_n &= v = L_f^L T(\bar{x}) + L_{L_1}^{-1} L_q T(\bar{x}) \ast u
\end{align*}
\]

(5)

\(L_f^n T(\bar{x})\) and \(L_{L_1}^{n-1} L_q T(\bar{x})\) are Lie derivatives of \(f\) and \(g\), respectively, and defined according to (6). For the Lie brackets, the objective of using these derivatives is mainly to reduce the length of the partial derivative formulas since these derivatives represent a shorter equivalent version for an identical original longer expression that contains partial derivative formulas.

\[
\begin{align*}
L_f^n T(\bar{x}) &= L_f(L_f^{n-1} T(\bar{x})) \\
L_f T(\bar{x}) &= \frac{\partial T}{\partial x} f(\bar{x}) \\
L_0^n T(\bar{x}) &= T(\bar{x}) \\
L_{L_1}^{n-1} L_q T(\bar{x}) &= L_{L_1}^{n-1} \left( \frac{\partial T}{\partial x} g(\bar{x}) \right)
\end{align*}
\]

(6)

Thus, the transformation function \(T(\bar{x})\) is calculated by solving the following system of partial differential equations when the Lie brackets are replaced using equation (4).

\[
\begin{align*}
\frac{\partial T}{\partial x} \ast ad_f^0 g &= 0 \\
\frac{\partial T}{\partial x} \ast ad_f^1 g &= 0 \\
\frac{\partial T}{\partial x} \ast ad_f^{n-1} g &= 0 \\
\frac{\partial T}{\partial x} \ast ad_f^{n-2} g &= 0 \\
\frac{\partial T}{\partial x} \ast ad_f^{n-1} g &\neq 0
\end{align*}
\]

(7)

The linearizing control signal \(u\) is then obtained from (5) and (7).

\[
u = (v - L_f^n T(\bar{x}))/L_f^{n-1} L_q T(\bar{x})
\]

(8)

**IV. MODEL PREDICTIVE CONTROL FORMULATION**

Our model predictive control (MPC) strategy aims to enhance the system response to significant perturbations through the prediction of future state variables and the minimization of the cost function, although other objectives can be implemented. The MPC outer loop provides the control input for the ILC control inner loop. Cascading both controls will provide a more accurate linear-model-based MPC for the converter. This improved accuracy aims to enhance the converter robustness in the case of significant changes in the input voltage, reference signal or output load.

The MPC prediction process uses the discrete mathematical model of the converter and real-time measures of the state variables [24]. The prediction objective is to calculate the control signal value on a future prediction horizon composed of N samples. The control signal is calculated to minimize a cost function \(g\). The main advantage of the MPC is to anticipate the future value of the control signal to ensure an optimal reference tracking performance of the state variables. This improves the system behavior in the case of significant perturbations. The optimal prediction of state variables enables a better use of the converter, which can increase its lifecycle. The constraints in the optimization process can also help protect the converter circuit from excessive voltage or current overshoots. MPC also has the advantage of being flexible and adaptive to most converter models: linear, nonlinear, deterministic and stochastic.

MPC can be implemented in three steps: First, the discrete state space model of the converter is formulated. Second, an optimization problem is defined through a cost function and respective constraints. Third, the limits of the cost function weights and constraints are tuned to achieve the desired behavior of the system.

In our case, the prediction state space model will be the converter linear equivalent obtained in the ILC stage. The equivalent model is then discretized through a zero-order hold on the input method with a sampling time of \(T\):

\[
\begin{align*}
z_1(k+1) &= z_1(k) + T \ast z_2(k) + \frac{T^2}{2} \ast v(k) \\
z_2(k+1) &= z_2(k) + T \ast v(k)
\end{align*}
\]

(9)

The optimization objective is to minimize a cost function \(C\). In our case, the objective of minimizing the voltage and current reference tracking errors and the control signal amplitude was selected. A quadratic format was selected for the cost function expression as follows:

\[
C(z_k, v_k) = \sum_{j=1}^{p} \sum_{i=1}^{n} \lambda_i \ast [z_i(k + j | k) - z_i^* (k + j | k)]^2 + \lambda_3 \ast [v(k + j | k) - v^* (k + j | k)]^2
\]

(10)

such that

\[p: \text{prediction horizon, which is set to 1,} \]

\[\lambda_i: \text{weighting factors, which are tuned to ensure zero static error and a minimal response time,} \]

\[z_i^*, v^*: \text{values of } z_i \text{ and } v \text{ when } i_{L_2} = i_{L_2, ref} \text{ and } v_c = v_{c, ref}. \]

The solution of the optimization problem is the control input \(v_k\) that minimizes \(C\) at the instant \(k + 1\). Duty cycle \(d(k)\) is then calculated using (8) such that

\[
d(k) = (v(k) - L_f^n T(\bar{x}(k)))/L_f^{n-1} L_q T(\bar{x}(k))
\]

(11)

**V. SIMULATION RESULTS**

These simulation results show how the proposed algorithm improves the system response in three cases: 1) converter start-up; 2) output reference voltage tracking; and 3) output load reduction. The system response is compared with that of a linear PID control. PID parameters were tuned using a pole placement strategy to minimize the overshoot and response time and eliminate the static error. The system nominal configuration is described in Table 1. The PWM frequency is 45.45 kHz. The simulations were performed using SimPowerSystems on Matlab2016b with a sample time of T=22 µs. The computer has an Intel Core i7 3.6 GHz processor and 8 GB of RAM.
The start-up of the converter (Fig. 4) aims to compare the dynamic behaviors of our proposed control and a linear PID. As shown in Fig. 5, the PID start-up exhibits a voltage start-up overshoot of approximately 10 V and an inrush current above 5 A. Voltage and current undershoots are also observed at approximately t=0.02 s. The voltage curves show that the proposed control has a faster response time and no starting overshoot or static error. The current curve with the new control shows an overshoot of 4.5 A and a nominal current of 0.55 A.

B. STEP-UP CHANGE IN THE OUTPUT REFERENCE VOLTAGE

The step-up change in the reference voltage of the converter (Fig. 6) aims to assess the new control reference tracking ability. At t=0.1 s, the reference voltage is changed from 20 V to 30 V. In this case, the new control shows a good tracking ability with no overshoot or static error and a response time of 8 ms. As shown in Fig. 7, the PID control after t=0.1 s presents a slower response time of 30 ms. The PID voltage curve also presents an overshoot at the reference changing time and voltage and current undershoots after t=0.1 s.

C. STEP CHANGE IN THE LOAD

The step change in the load (Fig. 8) aims to confirm the enhanced performance of our controller in the case of a load reduction. The resistive load will be reduced to half its value at t=0.1 s. The reference voltage is 20 V. For the new control, an undershoot of 0.1 V occurs. The voltage returns to its nominal value at t=0.101 s. As shown in Fig. 9, for the PID control, a voltage undershoot of 0.71 V occurs at t=0.1 s. The converter voltage returns to its permanent value at t=0.11 s. In addition, a current overshoot of approximately 0.5 A occurs at t=0.1 s.

D. STEP-UP CHANGE IN THE SOURCE VOLTAGE

The step-up change in the source voltage (Fig. 10) aims to confirm the enhanced performance of our controller in the occurrence of a step-up increase in the input source voltage. The input voltage will be increased by 50% at t=0.1 s. The output voltage shows a reference tracking error of 0.25 V.
VI. EXPERIMENTAL RESULTS

These experimental results aim to validate the improvement in system performances shown by the simulations. The nominal configuration of the system (Fig. 12) is identical to that in the simulation, and the experimental components are listed in Table 1. In this study, switching losses were neglected because we used a TP65H035WS gallium nitride MOSFET and a GB01SLT12-220 silicon carbide Schottky power diode, which caused an overall start-up switching loss of 0.61%. The conduction losses were reduced by selecting the inductance, GAN and diode with minimal resistances $R_L$, $R_{ON}$ and $R_D$, respectively. The algorithm was embedded into a TI F28379D microcontroller with a 32-bit MCU working under a 200-MHz clock.

A. START-UP OF THE CONVERTER

The start-up of the converter (Fig. 13) aims to validate the converter dynamics and static performances. In this test, the output reference voltage is 20 V. The converter is turned on at $t=10$ ms. The test shows a zero static error and a response time of 8 ms. The overshoot of the inductor current is approximately 1.6 A.

B. STEP-UP CHANGE IN THE REFERENCE VOLTAGE

The step-up change in the reference voltage of the converter (Fig. 14) aims to validate the real-time reference tracking ability of the controller. At $t=0.1$ s, the reference voltage is changed from 20 V to 30 V. In this case, the new control shows no static error and a response time of 8 ms. The inductor current overshoot is approximately 3.2 A.
TABLE 2. Qualitative comparison of the boost nonlinear control algorithms.

| Control | Response time reduction | Reference voltage tracking | Voltage ripples elimination | Load halving response improvement | Current overshoot reduction |
|---------|------------------------|----------------------------|-----------------------------|-----------------------------------|---------------------------|
| [8]     | ✓                      | ✓                          | ×                           | ×                                 | ×                         |
| [9]     | ✓                      | ✓                          | ×                           | ×                                 | ×                         |
| [10]    | ✓                      | ✓                          | ✓                           | ×                                 | ×                         |
| [19]    | ✓                      | ✓                          | ✓                           | ×                                 | ×                         |
| [21]    | ✓                      | ✓                          | ✓                           | ×                                 | ×                         |
| [22]    | ✓                      | ✓                          | ✓                           | ×                                 | ×                         |
| Proposed control | ✓          | ✓                          | ✓                           | ×                                 | ✓                         |

FIGURE 13. Start-up of the voltage control with the new control: (a) Output voltage; (b) Inductor current.

FIGURE 14. Step-up change in the reference voltage with the new control: (a) Output voltage; (b) Inductor current.

FIGURE 15. Step change in the load with the new control: (a) Output voltage; (b) Inductor current.

C. STEP CHANGE IN THE LOAD

The step change in the load (Fig. 15) aims to confirm the enhanced performance of our controller when the load decreases. The resistive load will be reduced to half its value at t=0.1 s. The current then doubles from its nominal value with no overshoot. The voltage static error remains at approximately 0.5 V.

D. QUALITATIVE COMPARISON

Table 2 compares the proposed control features with those of recent algorithms in the literature. The response to load halving is considered to improve if the ratio of the voltage undershoots of the classic and proposed controls is greater than 4.

VII. CONCLUSION

This paper presents a cascaded MPC and ILC of the output voltage of nonlinear step-up converters. Our first goal is to use a fully nonlinear converter model to improve the output voltage dynamics and static response in the case of reference voltage, input voltage or load variations. The second objective is to limit the inductor current inrush during start-ups and transients. To the best of our knowledge, this is the first implementation of ILC control in the context of the real-time control of a nonlinear DC-DC boost converter. Three simulation and experimental validation tests were performed to verify the success of each goal: 1) start-up of the converter; 2) step-up change in the reference voltage; and 3) halving of the output load. The performances of the proposed controller were compared with those of a linear PID. Both tests show that the new algorithm offers better response time and reference voltage tracking features while limiting the inductor current inrush during transients and start-ups. The output load of our system belongs to a resistive type, however the proposed system can potentially be used everywhere a DC to DC boost converter is needed such as in: PV powered systems, regulated DC power supplies, battery charging systems, as well as in the boosting mode for Plug in Hybrid Electrical Vehicles (PHEV) regenerative breaking applications [25]. In those cases, the robust algorithm will help maintain the accuracy and rapid response time of the output signal. The reduction of the inductance current overshoot will help with energy savings and extend the appliance lifetime, which makes The proposed system it suitable for use in small power heating or lighting applications, where accurate outputs and rapid responses are required. In these cases, the robust algorithm will help maintain the accuracy and produce a fast response time of the output signal. A decrease in the inductor current overshoot will improve the energy-saving potential and extend the appliance lifetime. Future enhanced real-time implementations of the ILC can allow the use of standard linear state feedback control as the outer loop for the ILC. Standard pole placement techniques can then be applied to...
more systematically tune the nonlinear controller according to the performances specified by customers. This process can facilitate the introduction of this type of nonlinear control into an industrial or academic context.

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