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Augmenting ARMarkov-PFC predictive controller with PID-Type III to improve boost converter operation

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ABSTRACT

The goal of this research is to improve the dynamic performance of a boost converter in continuous voltage mode controlled by a Type III PID controller. To this end, a special predictive controller called 'ARMarkov-PFC' has been used. In fact, in order to improve the performance of the closed-loop system during startup and reference input tracking and also enhance the relative stability of the system, the proposed controller has been augmented to this controller. Accordingly, a new controller named ARMarkov-PFC Plus PID (APP) has been proposed. Simulation and experimental results reveal that the proposed controller enjoys adequate performance and stability.

1. Introduction

The boost converter is a highly-popular, versatile and increasing DC–DC switching converter; also it is usually an important part in many industrial applications like the renewable electrical power supplies, energy storage, and hybrid electric vehicles (Liu, Gao, Wang, & Wang, 2015; Mukherjee & Strickland, 2016).

This converter operates in two different modes: Continuous Conduction Mode (CCM) and Discontinuous Conduction Mode (DCM). Also, the boost converters are controlled by two different control structures: Voltage Mode Control (VMC) and Current Mode Control (CMC) (Arulselvi', Uma, & Chidambaram, 2004; Bag, Roy, Mukhopadhyay, Samanta, & Sheehan, 2013). When a boost converter operates in CCM and VMC structure, there is a Right Half Plane (RHP) zero in its duty cycle-to-output voltage transfer function (it is a non-minimum phase transfer function) resulting from small-signal analysis. The RHP zero in the transfer function causes a phase lag and jeopardizes the stability of the converter. Normally, to ensure the stability of these converters, their bandwidth is severely limited and their gain crossover frequency is tuned to a value smaller than the frequency corresponding to the RHP zero (Kittipeerachon & Bunlaksananusorn, 2004).

The boost converter in the VMC has a pair of complex poles as well. The positions of this pair of poles as well as the RHP zero depend on the duty cycle, input voltage, output voltage, load resistance and output capacitor. The existence of this complex poles along with the RHP zero, and the dependence of their positions on the working point have complicated the stabilization and control of this converter in quickly achieving an acceptable dynamic response and presented a special challenge (Kapat, Patra, & Banerjee, 2009; Yao, 2012; Zurbriggen & Ordonez, 2016).

One of the known methods of eliminating the effect of RHP zero is to use the boost converter in the DCM; Nevertheless, compared to the CCM, the DCM has several drawbacks of its own, including: (1) a higher dependency of the relationship between the converter's input and output on the amount of load, (2) a greater peak current and thus higher losses and lower efficiency for the same output power and (3) requiring an inductor with larger physical dimensions for the same output power (Sheehan, 2007; Yao, 2012).

Another way of simplifying a boost converter's dynamics and reducing its control complexity is to use it in the CMC scheme. In the CMC, it is easier to compensate and control the boost converter; because in this control mode, the transient response of the converter at lower frequencies acts as a single-pole system (Lynch, 2008; Maniktala, 2012). The biggest problem of operating a boost converter at the CMC scheme is the need for a current sensor with very quick dynamics and very low time delay. In case of using a digital scheme for controlling the boost converter, a fast A/D would be needed for taking several samples in every current cycle and converting them to digital; which is not easy to do and is very costly (Basso, 2008; Chen, Prodić, Erickson, & Maksimović, 2003).

In SMPSs, the structure and tuning of the controller should be chosen so that (1) the steady state error in the output voltage becomes zero, (2) an adequate relative stability achieves for the system (GM > 10 dB,

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PM > 45 deg) (Basso, 2008), (3) the circuit bandwidth is sufficiently large (ω_{0dB}), and (4) the circuit has a satisfactory transient response when eliminating the disturbances associated with load variations and input voltage (Kittipeerachon & Bunlaksananusorn, 2004; Olalla, Leyva, & El Aroudi, 2009) and also when following the reference input around working point and startup duration. As was mentioned, because of the existence of the RHP zero and the complexity of the boost converter's dynamics in the voltage mode control scheme and in the CCM, the abovementioned objectives are difficult to achieve. However, because of not needing a current sensor in the control loop, and due to the existing flaws in the DCM, this research is interested in investigating the boost converter operation in the CCM.

The most common traditional controllers like PID, lead–lag, etc. have been developed in the past to control the existing power converters and ensure desired converter performance for the different conditions (Ghosh, Banerjee, Sarkar, & Dutta, 2016; Liping, Hung, & Nelms, 2011). The simple models of these converters are obtained by using the linearization and signal averaging methods, and these models are then used for designing the controllers (Yang & Sen, 2001). For this reason, under problematic conditions, like severe changes in parameters and disturbances resulting from the load, and the changes occurring in the input voltage and at startup, a PID controller may not have a satisfactory performance (Liping et al., 2011). A robust LQR controller based on Linear Matrix Inequalities (LMI-LQR) has been used in Olalla, Queinnec, Leyva, and El Aroudi (2011) and Olalla, Leyva, El Aroudi, Garćes, and Queinnec (2010), for controlling a boost converter in the CCM. This type of controller has been employed to overcome the model uncertainty and the applied disturbances and to ensure the stability. The current feedback has been used to achieve the mentioned objectives; and this will create some problems in the practical implementation of the controller.

Within the last two decades, numerous researchers have attempted to limit the destructive effects of the RHP zero in boost converters (Sable, Cho, & Ridley, 1991; Schoneman & Mitchell, 1988). A known method for reducing the effect of the RHP zero is to use a predictive control. The prediction-based algorithms can be useful in controlling the non-minimum phase systems; because they can predict the effect of the current trend of a process on its behavior in the future. The MPC is an attractive way of confronting the challenging issues in control. An MPC has been used in Bououden, Hazil, Filali, and Chadli (2014) for controlling a boost converter in the VMC structure (Holkar & Waghmare, 2010). An appropriate solution for overcoming the problem of burden and complex computations in the MPC and enabling its application in the online control of processes with fast dynamics is to employ a special type of MPC called the Predictive Functional Controller (PFC) (Mousavi Anzehaee & Haeri, 2011). It should be noted that the PFC used here is different from, and should not be mistaken with, the 'PFC' which is known in the power electronic systems as an abbreviation for Power Factor Correction.

The ARMarkov-PFC is a type of PFC developed for further reducing the computational burden and for practical implementation in quick processes (Kamrunnahar, Fisher, & Huang, 2002). In this predictive controller, the ARMarkov model architecture has been used to build the predictive model.

The objective of this research is to control a boost converter in the CCM and VMC structure and overcome its existing problems; so that the boost converter does not need a current feedback loop, has a satisfactory performance in reference input tracking at the startup duration and around working point, has an acceptable stability, is able to eliminate the load disturbances and input voltage changes and it also displays a good robustness against the changes of model parameters.

In order to achieve above-mentioned goals, at first, a model predictive controller named ARMarkov-PFC has been proposed and used to control the boost converter and its performance has been investigated. At last, this controller has been augmented to traditional PID Type III (well known as Type III) controller and a new efficient one named APP controller has been introduced.

In this research, by simulating in the PowerSim (of MATLAB®) environment, OrCad Pspise and by implementing in practice, the performance of the proposed controller has been compared, from different aspects, with the performance of Type III and LMI-LQR controllers.

In the sequel, the boost converter has been modeled in the CCM and VMC structure in Section 2. Also in this section, the designing of the controllers has been covered. The simulation and practical implementation results have been provided in Section 3; and finally, the conclusion of the research has been presented in Section 4.

2. Circuit diagram of boost converter and designing the controllers

This section deals with process modeling along with the design procedures for the two controllers used; namely the ARMarkov-PFC and Type III controllers.

Fig. 1 shows the electrical circuit of the boost converter with the parasitic elements in the capacitor (R_{esr}) and inductor (r_L). The values of the converter elements have been presented in Table 1. F_S is the switching frequency of the power supply.

The equations associated with the small-signal model of the boost converter in the CCM have been presented in Eq. (1) (Basso, 2008). In this converter, the mean change of the duty cycle (d) is used to control the mean change of the output voltage (v_o) as well. In these equations, $G_{v_o d}$ is the d to v_o transfer function.

$$G_{v_o d}(s) = \frac{v_o(s)}{d(s)} = G_{d0}\frac{\left(1+\frac{s}{\omega_{Z1}}\right)\left(1-\frac{s}{\omega_{RHPZ}}\right)}{1+\frac{s}{\omega_0 Q}+\frac{s^2}{\omega_0^2}} \tag{1}$$

where

$$Q = \frac{\omega_0}{\frac{r_L}{L}+\frac{1}{C(R_{Load}+R_{esr})}}, \qquad G_{d0} = \frac{V_{IN}}{(D')^2} = \frac{V_0^2}{V_{IN}},$$

$$\omega_{RHPZ} \approx \frac{R_{Load}}{L}\left(\frac{V_{IN}}{V_o}\right)^2, \qquad \omega_{z_1} = \frac{1}{CR_{esr}}$$

$$\omega_0 = \frac{1}{\sqrt{LC}}\sqrt{\frac{r_L+R_{Load}(D')^2}{R_{Load}}} \approx \frac{1}{\sqrt{LC}}\frac{V_{IN}}{V_o}, \qquad D' = 1-D$$

The $G_{v_o d}$ is a second-order transfer function with a rather large quality factor; which contains a zero in the left half plane and another zero in the right half (ω_{RHPZ}) of plane s.

2.1. Designing the controllers

PFC has been used in order to reduce the computation time needed for obtaining the control signal in the MPC family (Mousavi Anzehaee & Haeri, 2011). Thus, this controller is an appropriate one to control boost switching converters with very fast dynamics. Considering the criticality of computation time, a single-point PFC has been used here; meaning that this controller only has one coincident point. In fact, the process output coincides with its desired value only at one point of future sampling times. Moreover, to further reduce the computation time in the PFC, the ARMarkov model has been employed to build the predictive model and to compute the free response of the process.

2.1.1. ARMarkov-PFC

Fig. 2 shows the closed-loop system of the boost converter controlled by the ARMarkov-PFC. Now, in view of Fig. 2, the method of obtaining the necessary equations for the ARMarkov-PFC is explained. Based on the ARMarkov model configuration, the $\hat{y}(t+\mu)$, i.e. the predicted output at μ steps (sampling times) ahead of the single input-single output process can be written as follows:

$$\hat{y}(t+\mu) = g_1^{\mu}\Delta u(t+\mu-1) + g_2^{\mu}\Delta u(t+\mu-2) + \cdots + g_{\mu}^{\mu}\Delta u(t) + f(t+\mu) \tag{2}$$