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LPV Gain-scheduling Control for a Phase-shifted PWM Full-bridge Soft Switched Converter

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Abstract: This paper presents an output feedback LPV gain-scheduling control design for a class of PSPWM full-bridge power converters. The design starts with an averaged model of the converter with adequate complexity. One main contribution of this work is to demonstrate how to reformulate the averaged model into an LPV one. A gain-scheduling controller can then be synthesized taking advantage of the design framework of LPV-based H-infinity control. Forming the open-loop LPV system and synthesizing the controller, a repetitive control module and an anti-windup scheme are incorporated into the design. The proposed design is further justified with a numerical simulation.

Keywords: Anti-windup; LPV gain-scheduling; phase-shifted PWM full-bridge converter; repetitive control

1. INTRODUCTION

Conventional linear power converters are gradually replaced by switching power converters because of the advantages such as being compact, lightweight, high-efficiency and larger input voltage range. With the rapid development and prevalence of personal computers, mobile communication devices, and automotive electronics in recent years, necessity of stability and efficiency for converters is increased. Among the switching power converters, the phase-shifted pulse-width modulated (PSPWM) full bridge soft switched power converter (Mweene, Wright, & Schlecht, 1989; Redl, Sokal, & Balogh, 1991) and corresponding modifications (Brunoro & Vieira, 1999; Cho, Sabate, Hua, & Lee, 1996; Hua, Lee, & Jovanovic, 1993; Yungtaek Jang & Jovanovic, 2004; Yungtack Jang, Jovanovic, & Chang, 2003; Redl, Balogh, & Edwards, 1994) have become a widely used circuit topology due to various beneficial characteristics. The ZVS operation significantly reduces switching losses and stresses, and eliminates the need of primary snubbers. Hence, the circuit is capable of high switching frequency operation with improved power density and conversion efficiency.

The input and output voltages for switching power are often subject to fluctuations due to unstable power sources (such as batteries or utility lines) and varying load current, respectively. Most applications (in additional to laboratory power supplies) require constant output voltage or current of wide-ranging adjustability. Some even have need for variable output voltage. Therefore, feedback control has been incorporated into switching power converters to not only stabilize, but also improve the performance robustness of the output voltage.

Most of the closed-loop control designs study simple dc-dc (buck, boost, or buck/boost) converters with immediately obtained linear or nonlinear mathematical models. Design approaches ranging from linear to nonlinear control have been introduced in the literatures (Carbonell, Garcerá, & Hilario, 1999; Mariéthoz et al., 2010; Ng, Leung, & Tam, 1997; Oettmeier, Neely, Pekarek, DeCarlo, & Uthaichana, 2009; Carlos Olalla, Leyva, El Aroudi, & Garces, 2009; C Olalla, Leyva, El Aroudi, Garces, & Queinnec, 2010; Torres-Pinzon & Leyva, 2009). Aside from closed-loop stability, taking into account voltage/current disturbances, parametric uncertainty of components, and nonlinearities due to PWM switching has increasingly been deemed as an essential aspect of control design. Resorting to nonlinear control technique may directly tackle the nonlinearities. The drawback, in general, is the absence of a systematic design framework in unifying various design objectives, such as stability, rejection of disturbance, actuator saturation, and reduction of sensitivity to parametric uncertainty. By linearizing the averaged model of the converter around an operating point, linear robust control paradigm provides the aforementioned design framework, and the formulated problem can readily be solved using existing numerical tools. However, the design will only guarantee stability and performance for a small region around the operating point. Recent development in the field of linear matrix inequalities (LMI) has inspired a new development of linear robust control, which can now address certain category of nonlinear systems, i.e., those formulated as linear parameter varying (LPV) systems, with the so-called LPV gain scheduling control (Gahinet, P., & Apkarian, P. 1994; Becker, G., & Packard, A. 1994). Apart from inheriting the aforestated benefits of linear robust control design, utilization of LPV gain scheduling control can extend the operation range of a system.

In spite of its advantageous features, feedback control for soft switched PSPWM full-bridge converters is still confined to simple linear time-invariant design, e.g., proportional-integral (PI) or lead-lag compensators based on a linearized model

(Cho et al., 2010; Lim, Lim, & Chung, 2007; Liu, Meyer, & Liu, 2009; Tseng, K. H., & Chen, C. L., 2011). As pointed out by (Schutten & Torrey, 2003; VlatkoviC, Sabate, Ridley, Lee, & Cho, 1992), due to the increased number of topological stages and the PWM duty cycle being affected by input voltage, output voltage, and load current, the dynamics of a PSPWM full-bridge converter is much more sophisticated than that of a simple buck converter. A trade-off needs to be made regarding whether a simple model (e.g., linearized model) or a complex one (e.g., switched model) is to be established for the purpose of control design.

In this work, an advanced gain-scheduling control design is proposed for a PSPWM full-bridge zero-voltage-switching (ZVS) power converter, which can also cope with periodic disturbances and actuator saturation. First, an averaged model of the converter, with decent complexity and appropriate for subsequent control design, is formed. Next, a sequence of steps is presented, which details how to reformulate the averaged model into an LPV one. Note that the suggested steps do not involve approximation or linearization. Finally, an LPV gain-scheduling controller can then be synthesized utilizing corresponding design framework. Note that, in the process of forming the open-loop LPV system and synthesizing the feedback controller, the repetitive control module and the anti-windup scheme are integrated into the design. The feasibility and capability of the proposed design is demonstrated by numerical simulation.

This paper is organized as follows. Section 2 will detail how to formulate the averaged model of a PSPWM full-bridge power converter into an LPV model. Section 3 describes the procedure for synthesizing the output feedback LPV gain-scheduling controller. Numerical simulation and discussion are given in Section 4. Conclusion is made in Section 5.

2. PROBLEM FORMULATION

In this section, we demonstrate how to reformulate the averaged model of a PSPWM full-bridge ZVS converter (see Fig. 1) into an LPV model.

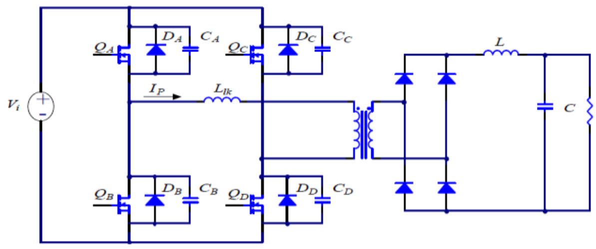

Fig. 1 Circuit topology of a PSPWM full-bridge converter

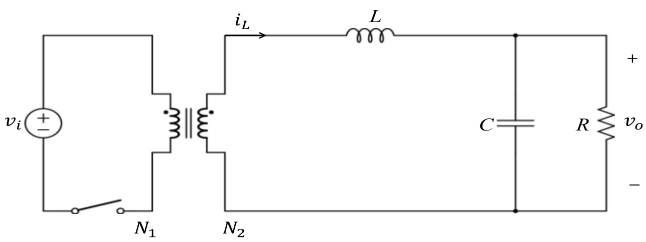

Fig. 2 The equivalent circuit of the converter

Consider the (simplified) equivalent circuit of the converter shown in Fig. 2. Based on Fig. 2, the averaged model of the converter is described by

$$\frac{d}{dt}\begin{bmatrix} i_L(t) \\ v_o(t) \end{bmatrix} = \begin{bmatrix} 0 & -\frac{1}{L} \\ \frac{1}{C} & -\frac{1}{RC} \end{bmatrix}\begin{bmatrix} i_L(t) \\ v_o(t) \end{bmatrix} + \begin{bmatrix} \frac{nd_{fb}}{L} \\ 0 \end{bmatrix} v_i(t) \quad (1)$$

where i_L is inductance current, v_i is input voltage, v_o is output voltage, d_{fb} is the effective duty cycle, $N_2/N_1 = n$ is the transformer turns ratio, L is inductance, C is capacitance, and R is resistance.

Define $\vec{x}(t) = [i_L(t) \quad v_o(t)]^T$ and decompose each state variable into its steady-state and perturbation components:

$$\vec{x} = \vec{X} + \hat{\vec{x}}, v_i = V_i + \hat{v}_i, d_{fb} = D_{fb} + \hat{d}_{fb} \quad (2)$$

where $\vec{X} = [I_L \quad V_o]^T$, V_i, and D_{fb} are steady-state variables with $\hat{\vec{x}} = [i_L \quad v_o]^T$, $\hat{v}_i$, and $\hat{d}_{fb}$ corresponding perturbation terms. Since the steady-state variables should satisfy

$$\dot{\vec{X}} = \bar{A}\vec{X} + \bar{B}V_i = \begin{bmatrix} 0 & -\frac{1}{L} \\ \frac{1}{C} & -\frac{1}{RC} \end{bmatrix}\begin{bmatrix} I_L \\ V_o \end{bmatrix} + \begin{bmatrix} \frac{nD_{fb}}{L} \\ 0 \end{bmatrix} V_i = 0 \quad (3)$$

Equation (1) can be rewritten as

$$\dot{\hat{\vec{x}}} = \begin{bmatrix} 0 & -\frac{1}{L} \\ \frac{1}{C} & -\frac{1}{RC} \end{bmatrix}\hat{\vec{x}} + \begin{bmatrix} \frac{nD_{fb}}{L} \\ 0 \end{bmatrix}\hat{v}_i + \begin{bmatrix} \frac{nV_i}{L} \\ 0 \end{bmatrix}\hat{d}_{fb} + \begin{bmatrix} \frac{n}{L} \\ 0 \end{bmatrix}\hat{d}_{fb}\hat{v}_i \quad (4)$$

For a PSPWM full-bridge ZVS converter, because of leakage, there exists duty cycle loss ΔD on the secondary side, such that $D_{fb} = D_{pwm} - \Delta D$, where D_{pwm} is the duty cycle of the primary voltage. The duty cycle loss (VlatkoviC, Sabate, Ridley, Lee, & Cho, 1992)

$$\Delta D = 4nI_L L_{lk} fs - (1 - D_{pwm})V_o \frac{nL_{lk}}{L} \Big/ V_i \quad (5)$$

where L_{lk} is leakage inductance, and fs is switching frequency. Using (5), the effective duty cycle in steady state can be expressed by

$$D_{fb} = D_{pwm} - \frac{4nI_L L_{lk} fs - (1 - D_{pwm})V_o \frac{nL_{lk}}{L}}{V_i} \quad (6)$$

Denote $\hat{d}_{fb} = \hat{d}_b - (\hat{d}_{i_L} + \hat{d}_{v_i} + \hat{d}_{v_o})$, where $\hat{d}_{v_o}$, $\hat{d}_{i_L}$, $\hat{d}_{v_i}$ and $\hat{d}_b$ are duty cycle changes caused by the output voltage, filter inductor current, input voltage and duty cycle of the primary voltage. Each variation term of the duty cycle can be derived by taking partial derivative with respect to a corresponding variable (i_L , v_0 , or v_i):

$$\hat{d}_{fb} = \hat{d}_b - (\frac{\partial \Delta D}{\partial i_L}\Big|_{\substack{v_i=V_i\\ i_L=I_L\\ v_o=V_o}} \hat{i}_L + \frac{\partial \Delta D}{\partial v_o}\Big|_{\substack{v_i=V_i\\ i_L=I_L\\ v_o=V_o}} \hat{v}_o + \frac{\partial \Delta D}{\partial v_i}\hat{v}_i\Big|_{\substack{v_i=V_i\\ i_L=I_L\\ v_o=V_o}} \hat{v}_i)$$

$$= \hat{d}_b - \frac{4nL_{lk}f_s}{V_i}\hat{i}_L + \frac{(1-D_{pwm})nL_{lk}}{V_iL}\hat{v}_o + \frac{4nI_LL_{lk}f_s - (1-D_{pwm})V_o\frac{nL_{lk}}{L}}{V_i^2}\hat{v}_i \tag{7}$$

Substituting (6) and (7) into (4), we may use the relationship $v_i = V_i + \hat{v}_i$ to combine terms related to the input voltage. Define $\hat{w} = \hat{v}_i$ as disturbance input and $\hat{u} = \hat{d}_b$ as the input of the actuator. We arrive at

$$\dot{\hat{x}}(t) = \begin{bmatrix} -R_d v_i(t)/V_iL & -1/L + (1-D_{pwm})n^2L_{lk}v_i(t)/V_iL^2 \\ 1/C & -1/RC \end{bmatrix}\hat{x}(t) + \begin{bmatrix} nv_i(t)/L \\ 0 \end{bmatrix}\hat{d}_b(t) + (\begin{bmatrix} n(D_{pwm}-\Delta D)/L \\ 0 \end{bmatrix} + \begin{bmatrix} +\dfrac{R_dI_L - (1-D_{pwm})V_on^2\frac{L_{lk}}{L}}{V_i^2L}v_i(t) \\ 0 \end{bmatrix})\hat{v}_i(t) \tag{8}$$

where $R_d = 4n^2L_{lk}fs$. Denote $sat(\hat{u})$ as the output of the actuator. We may define three varying parameters $\rho = \frac{sat(\hat{u})}{\hat{u}}$, $\varepsilon = v_i(t)$, and $\sigma = \varepsilon\rho$ such that $\varphi = (v_i(t), \frac{sat(\hat{u})}{\hat{u}}, v_i(t) \times \frac{sat(\hat{u})}{\hat{u}}) = (\varepsilon, \rho, \sigma)$. Choose $\hat{v}_o$ as the output of the system. Let $\hat{x}_1 = \hat{i}_L$, $\hat{y} = \hat{x}_2 = \hat{v}_o$, $\hat{w} = \hat{v}_i$, and $\hat{u} = \hat{d}_b$. Define

$$A = \begin{bmatrix} -\frac{R_d}{V_iL}\varepsilon(t) & -\frac{1}{L} + \frac{(1-D_{pwm})n^2L_{lk}}{V_iL^2}\varepsilon(t) \\ 1/C & -1/RC \end{bmatrix}, B_u = \begin{bmatrix} \frac{n}{L}\sigma(t) \\ 0 \end{bmatrix}$$

$$B_w = \begin{bmatrix} \frac{n}{L}(D_{pwm}-\Delta D) + \dfrac{R_dI_L - (1-D_{pwm})V_on^2\frac{L_{lk}}{L}}{V_i^2L}\varepsilon(t) \\ 0 \end{bmatrix}$$

An affine parametric varying system of (8) is represented by

$$\begin{bmatrix} \dot{\hat{x}}(t) \\ \hat{z}(t) \\ \hat{y}(t) \end{bmatrix} = \left[\begin{array}{c|cc} A(\varepsilon) & B_w(\varepsilon) & B_u(\sigma) \\ \hline C_z & D_{zw} & D_{zu} \\ C_y & D_{yw} & D_{yu} \end{array}\right] \begin{bmatrix} \hat{x}(t) \\ \hat{w}(t) \\ \hat{u}(t) \end{bmatrix} \tag{9}$$

where $\hat{x}(t)$ denote the state, $\dot{\hat{x}}(t)$ is the state derivative, $\hat{w}(t)$ is the disturbance, $\hat{z}(t)$ is the output related to system performance, $\hat{y}(t)$ is the output, and $\hat{u}(t)$ is the control input. Note that $C_z = \begin{bmatrix} 0 & 1 \end{bmatrix}$, $C_y = \begin{bmatrix} 0 & 1 \end{bmatrix}$, $D_{zw} = 1$, $D_{yw} = 1$, $D_{yu} = 0$ and $D_{zu} = 0$.

3. SYNTHESIS OF AN OUTPUT FEEDBACK LPV GAIN-SCHEDULING CONTROL

This section will present the procedure for synthesis of an output feedback LPV gain-scheduling controller. A repetitive control kernel and an anti-windup mechanism will be embedded into the controller to devote to rejection of periodic disturbances and abatement of actuator saturation.

3.1 Synthesis of Output Feedback Gain-Scheduling

For the LPV system represented in (9), suppose an LPV output feedback dynamic controller is to be designed from $\hat{y}$ to $\hat{u}$, i.e.,

$$\begin{bmatrix} \dot{\hat{x}}_k(t) \\ \hat{u}(t) \end{bmatrix} = \begin{bmatrix} A_k(\varphi) & B_k(\varphi) \\ C_k(\varphi) & D_k(\varphi) \end{bmatrix} \begin{bmatrix} \hat{x}_k(t) \\ y(t) \end{bmatrix} \tag{10}$$

Equation (10) is a full-order design in the sense that $\hat{x} \in R^n$ implies $\hat{x}_k \in R^n$. The controller matrices in (10) are function of parameter φ , which indicates how gain-scheduling is achieved.

Define $\hat{x}_{cl} = [\hat{x} \quad \hat{x}_k]^T$. The closed-loop system with (9) and (10) can be expressed as

$$\begin{bmatrix} \dot{\hat{x}}_{cl}(t) \\ \hat{z}(t) \end{bmatrix} = \begin{bmatrix} A_{cl}(\varphi) & B_{cl}(\varphi) \\ C_{cl}(\varphi) & D_{cl}(\varphi) \end{bmatrix} \begin{bmatrix} \hat{x}_{cl}(t) \\ \hat{w}(t) \end{bmatrix} \tag{11}$$

Denote the LPV closed-loop system as P_{cl} . In the following, we summarize the main theoretic results for synthesizing an LPV H_∞ controller, which takes the form of (10).

The quadratic LPV γ -performance problem: The LPV closed-loop system P_{cl} is exponentially stable and the induced L_2 norm of the system is less than a scalar $\gamma > 0$, i.e.,

$$\left\| C_{cl}(\varphi)(sI - A_{cl}(\varphi))^{-1}B_{cl}(\varphi) + D_{cl}(\varphi) \right\|_\infty = \sup_{\|\hat{w}\|_2 \neq 0} \|\hat{z}\|_2 / \|\hat{w}\|_2 < \gamma$$

for all φ belonging to a parameter variation set Φ , if there exist symmetric matrices (R, S) such that the following matrix inequalities hold for all $\varphi \in \Phi$ (Becker, G., & Packard, A. 1994; Gahinet, P., & Apkarian, P. 1994; Becker, G., & Packard, A. 1994):

$$\begin{bmatrix} N_R(\varphi) & 0 \\ 0 & I \end{bmatrix}^T \left[\begin{array}{cc|c} A(\varphi)R + RA(\varphi)^T & RC_z^T & B_w(\varphi) \\ C_zR & -\gamma I & D_{zw} \\ \hline B_w^T(\varphi) & D_{zw} & -\gamma I \end{array}\right] \times \begin{bmatrix} N_R(\varphi) & 0 \\ 0 & I \end{bmatrix} < 0 \tag{12}$$

$$\begin{bmatrix} N_S(\varphi) & 0 \\ 0 & I \end{bmatrix}^T \left[\begin{array}{cc|c} A(\varphi)S + SA(\varphi)^T & SB_w(\varphi) & C_z^T \\ B_w^T(\varphi)S & -\gamma I & D_{zw}^T \\ \hline C_z & D_{zw} & -\gamma I \end{array}\right] \times \begin{bmatrix} N_S(\varphi) & 0 \\ 0 & I \end{bmatrix} < 0 \qquad (13)$$

$$\begin{bmatrix} R & I \\ I & S \end{bmatrix} \geq 0 \qquad (14)$$

where $N_R(\varphi)$, $N_S(\varphi)$ are orthonormal bases of the null space of $[B_u^T(\varphi) \quad D_{zu}^T]$, $[C_y^T \quad D_{yw}^T]$ respectively, and R, $S \in R^{n \times n}$ are symmetric matrices. For a polytopic LPV system satisfying the following two assumptions: (i) There is no direct transmission from $\hat{u}$ to $\hat{y}$, i.e., $D_{yu} = 0$. (ii) The matrices B_u, C_y, D_{zu}, and D_{yw} are constant matrices (independent of the varying parameters). It can be shown that (12) and (13) hold if and only if they hold for the matrices corresponding to the vertices of the parameter polytope. In other words, only the LMIs corresponding to the vertices of the parameter polytope need to be formed for solving matrices *R* and *S*. For the formerly formulated LPV system, it can be seen that most matrices satisfy both assumptions except that $B_u(\varphi)$ is a parameter dependent matrix. The parameter dependency of the $B_u(\varphi)$ matrix can be removed by filtering the input channel, as will be discussed in next section.

3.2 Repetitive and Anti-windup Control

The proposed control configuration is depicted in Fig. 3. The performance weighting W_1 is for the measured output error, which should supply suitable gain (e.g., >1) in the frequency region where non-periodic disturbances locate. It is also selected as a stable low-order filter, e.g.,

$$W_1 = k + e\rho / (s + \omega_b \rho + b)$$

where *e, b,* ω_b and *k* are constants. These parameters will be specified in next section. The open-loop LPV system $G(\varphi)$ is expressed by (9).

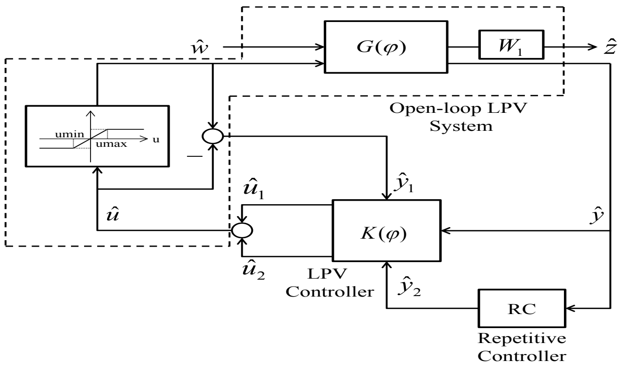

Fig. 3 Control structure with repetitive and anti-windup control

In this study, we will consider a low-order and attenuated repetitive controller, which takes the form of

$$RC(s) = \frac{1}{s/\omega_r + 1} \prod_{i=1}^{k} \frac{s^2 + 2\zeta_i \omega_{ni} s + \omega_{ni}^2}{s^2 + 2\xi_i \omega_{ni} s + \omega_{ni}^2}$$

where *k* is the number of periodic frequencies to be rejected, ω_{ni} is the i^{th} disturbance frequency, and ζ_i and ξ_i are two damping ratios that satisfy $0 < \xi_i < \zeta_i < 1$. We can adjust the gain of *RC*(s) at those periodic frequencies by varying the values of ζ_i and ξ_i. A low-pass filter of roll-off frequency ω_r is also added, which serves as similar role as the q-filter used in a digital repetitive controller. Suppose that *RC*(s) has the following state space realization

$$\dot{\hat{x}}_{rc} = A_{rc}\hat{x}_{rc} + B_{rc}\hat{y}$$
$$\hat{y}_2 = C_{rc}\hat{x}_{rc}$$

As shown in Fig. 3, the repetitive controller is augmented to the 'to-be-designed' LPV controller. Furthermore, consider the anti-windup scheme (Wu, F., Grigoriadis, K. M., & Packard, A., 2000) that feeds the difference between the actuator input and output back to the controller, as shown in Fig. 3. This corresponds to create a new input for the controller, i.e.

$$\hat{y}_1 = (\rho - 1)\hat{u}$$

This controller now has three inputs $(\hat{y}, \hat{y}_1, \hat{y}_2)$ and two outputs $(\hat{u}_1, \hat{u}_2)$. If the control $\hat{u}$ does not saturate (i.e., $\rho = 1$), then $\hat{y}_1 = 0$ and this additional input is deactivated. If the control $\hat{u}$ saturates (i.e., $\rho < 1$), then $\hat{y}_1 \neq 0$, which provides additional degree of freedom for manipulating the control $\hat{u}$.

The open-loop LPV system with repetitive and anti-windup control can be shown to have the following state-space representation

$$\begin{bmatrix} \dot{\hat{x}}(t) \\ \dot{\hat{x}}_{rc}(t) \\ \hat{z}(t) \\ \hat{y}(t) \\ \hat{y}_1(t) \\ \hat{y}_2(t) \end{bmatrix} = \left[\begin{array}{cc|ccc} A(\varepsilon) & 0 & B_w(\varepsilon) & B_u(\sigma) & B_u(\sigma) \\ B_{rc}C_y & A_{rc} & B_{rc}D_{yw} & 0 & 0 \\ \hline C_z & 0 & D_{zw} & D_{zu} & D_{zu} \\ C_y & 0 & D_{yw} & 0 & 0 \\ 0 & 0 & 0 & \rho - 1 & \rho - 1 \\ 0 & C_{rc} & 0 & 0 & 0 \end{array}\right] \begin{bmatrix} \hat{x}(t) \\ \hat{x}_{rc}(t) \\ \hat{w}(t) \\ \hat{u}_1(t) \\ \hat{u}_2(t) \end{bmatrix} \qquad (15)$$

We can further eliminate the parameter dependency of the input and output matrices by considering the dynamic responses of the sensor and actuator. Basically each input or output channel of the open-loop system (15) is passed through a low-pass filter

$$\begin{aligned} H(s) &= C_o(sI - A_o)^{-1}B_o \\ F(s) &= C_i(sI - A_i)^{-1}B_i \end{aligned} \qquad (16)$$

respectively, before connecting to the parameter dependent controller. The bandwidth of the low-pass filters depends on the sensor and actuator dynamics. For negligible senor or actuator dynamics, the bandwidth can be assigned to be much larger than that of the open-loop system to ensure least interference. The overall open-loop LPV system with parameter-free input-output matrices can be found after some matrix manipulations, which is depicted in Fig. 4.

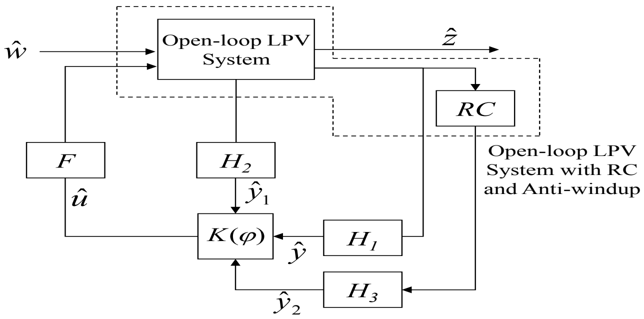

Fig. 4 Control structure with repetitive, anti-windup control, and isolated filters

The results summarized in previous section can then be used to synthesize an LPV gain-scheduling controller.

4. SIMULATION RESULTS

The open-loop LPV model is formulated preiously. The nominal system parameters listed in Table 1 are specified in accordance with a typical PSPWM full-bridge ZVS power converter in our laboratory. The polytopic parameter variation set Φ of four vertices is specified as

$$\Phi = \{(\varepsilon,\rho,\sigma): 150 \le \varepsilon \le 170, 0.1 \le \rho \le 1, \sigma = \varepsilon\rho\}$$

Based on the above simulation setup, with the varying range of ρ specified, the parameters of the weighting filter W_1 can be properly determined to reflect the different performance requirement for the unsaturated (ρ=1) and saturated (ρ<1) system. The low-pass filters *H*(s) and *F*(s) are selected as

$$H(s) = F(s) = \frac{1}{s/(2\pi \times 5 \times 10^4) + 1}$$

to reflect negligible sensor and actuator dynamics. The low-order attenuated repetitive controller is specified such that the periodic disturbances aimed for rejection are tentatively set at 10000, 15000 and 18000 Hz. Fig. 5 shows the magnitude curves of *H(s)*, *F(s)*, *RC(s)* and $W_1(s)$. A feasible LPV gain-scheduling controller is acquired using MATLAB Robust Control Toolbox (Balas, G., Chiang, R., Packard, A., & Safonov, M. 2005) according to the above parameter settings, and reaches γ=1.717958. Structured singular value (μ) is employed to evaluate the nominal performance of the control system. The sensitivity plots for the four nominal closed -loop systems lying at the vertices of the polytope are shown in Fig. 6. A numerical simulation is also performed under the environment of Simulink to justify the design. The input voltage perturbation *w* is a combination of three sinusoids of amplitude equal to one and frequencies set to 1 kHz, 1.5 kHz, and 1.8 kHz, respectively. Fig. 7 shows that the controller successfully stabilizes the output voltage.

5. CONCLUSION

Current research on feedback control of dc-dc power converters mostly focuses on systems with simple circuit topology (buck, boost, or buck/boost). In particular, control for soft switched PSPWM full-bridge converters is still limited to linearized design with PI or lead-lag compensators. This motivates the work presented in this paper. Although the parametric uncertainty is not considered in this work, it can be readily incorporated into the proposed design framework. This along with experimental verification will be pursued in our future study.

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Table 1 Nominal parameters of the converter

Nominal Parameters	Unit
n=0.5 (transformer turns ratio)	
$V_i = 160$	volt
$V_o = 50$	volt
$I_L = 10$	Ampere
$D_{fb} = 62.5$	%
$D_{pwm} = 87.36$	%
$R = 5$	Ω
$C = 940$	μF
$L = 300$	μH
$L_{lk} = 20$	μH
$f_s = 100$	kHz
$R_d = 4n^2 L_{lk} f_s = 2$	Ω

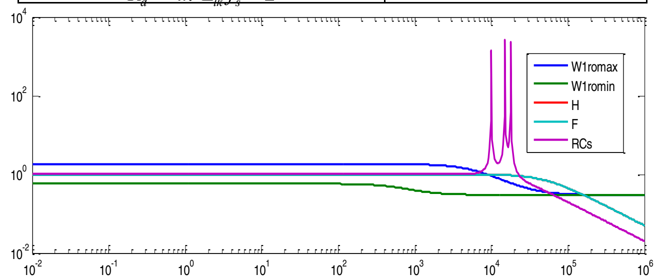

Fig. 5 Magnitude plots of *H(s)*, *F(s)*, *RC(s)*, *W1_romax* and *W1_romin*

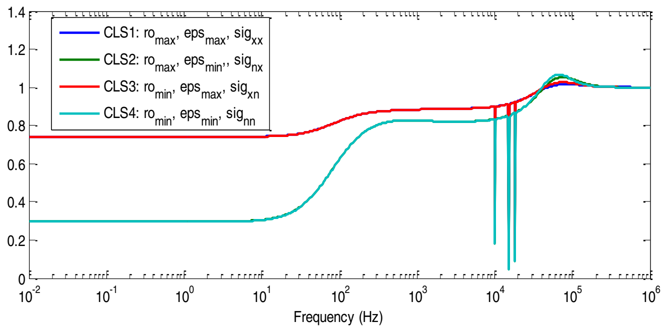

Fig. 6 Sensitivity plots corresponding of the closed-loop systems at the four vertices

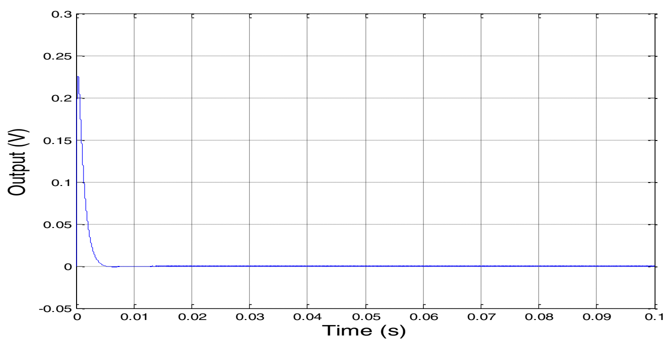

Fig. 7 The response of the output (perturbed) voltage