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Robust digital control for interleave PFC boost converter

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Abstract: In recent years, improving of power factor and reducing harmonic distortion in electrical instruments are needed. In general, a current conduction mode boost converter is used for active PFC (Power Factor Correction). Especially, an interleave PFC boost converter is used in order to make a size compact, make an efficiency high and make noise low. In this paper, a robust digital controller for suppressing the change of step response characteristics and variation of output voltage at a load sudden change with high power factor and low harmonic is proposed. Experimental studies using a micro-processor for controller demonstrate that this type of digital controller is effective to improve power factor and to suppress output voltage variation.

Keywords: interleave PFC, boost converter, digital robust control, micro-processor

1 INTRODUCTION

In recent years, improving of power factor and reducing harmonic of power supply using nonlinear electrical instruments are needed. A passive filter and an active filter in AC lines are used for improving of the power factor and reducing the harmonic. Generally a current conduction mode boost converter is used for an active PFC (Power Factor Correction) in electrical instruments. Especially, an interleave PFC boost converter is used in order to make a size compact, make an efficiency high and make noise low. In the PFC boost converter, if a duty ratio, a load resistance and an input voltage are changed, the dynamic characteristics are varied greatly, that is, the PFC converter has non-linear characteristics. In many applications of the interleave PFC converters, loads cannot be specified in advance, i.e., their amplitudes are suddenly changed from the zero to the maximum rating. This is the prime reason of difficulty of controlling the PFC boost converter.

Usually, a conventional PI or analog IC controller designed to the approximated linear controlled object at one operating point is used for the PFC converter. In the nonlinear PFC boost converter system, those controllers are not enough for attaining good performance. In this paper, the robust controller for suppressing the change of step response characteristics and variation of output voltage at a load sudden change with high power factor and low harmonic is proposed. An approximate 2-degree-of-freedom (A2DOF) method [1] is applied to the interleave PFC boost converter with the load. The PFC converter is a nonlinear system and the models are changed at each operation point. The design and combining methods of two controllers which can cope with nonlinear system or changing of the models with one controller is proposed. One is an approximate 2-Dgree-of-freedom (A2DOF) controller for a current control system and another is a PI controller. These controller are actually implemented on a micro processor and is connected to the PFC converter. Experimental studies demonstrate that the digital controllers designed by proposed method satisfy the desired performances and are useful.

2 INERLEAVE PFC BOOST CONVERTER

2.1 State-space model of interleave boost converter

The interleave boost converter shown in Fig. 1 is manufactured.

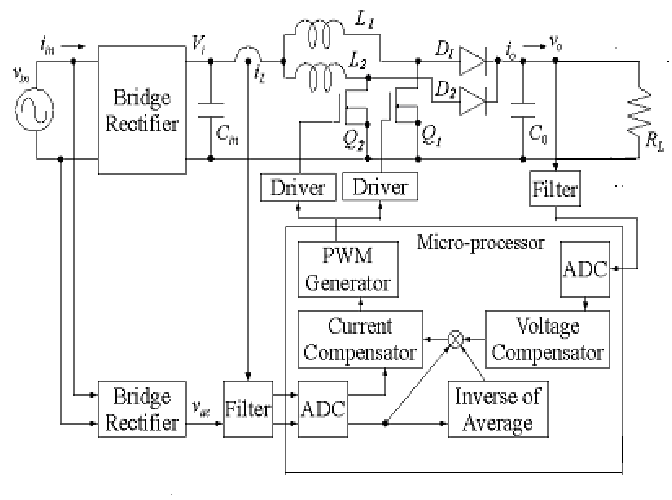

Fig.1 Interleave PFC boost converter

Fig.1, v_{in} is an input AC voltage, i_{in} is an input AC current, C_{in} is a smoothing capacitor, V_i is a rectifying and smoothing input voltage, Q_1 and Q_2 are MOSFETs or IGBTs, L_1 and L_2 are interleave boost inductances, D_1 and D_2 are interleave boost diodes, C_0 is an output capacitor, R_L is an output load resistance, i_L is the sum of inductor current, v_{ac} is an absolute value of the input AC voltage and v_o is an output voltage. The inductor currents i_L is controlled to follow the rectified input voltage v_{ac} for improved power factor, reduced harmonics and stable the output voltage.

At some operating point, let v_o, i_L and μ, be V_s, I_s and μ_s, respectively. Then the linear approximate state equation of the boost converter using small perturbations $\Delta i_L = i_L - I_s$, $\Delta v_o = v_o - V_s$ and $\Delta\mu = \mu - \mu_s$ is as follows:

$$\begin{aligned} \dot{x}(t) &= A_c x(t) + B_c u(t) \\ y(t) &= C_c x(t) \end{aligned} \qquad (3)$$

where

$$A_c = \begin{bmatrix} -\dfrac{R_0}{L_0} & -\dfrac{1-\mu_s}{L_0} \\ \dfrac{1-\mu_s}{C_0} & -\dfrac{1}{R_L C_0} \end{bmatrix}, B_c = \begin{bmatrix} \dfrac{V_s}{L_0} \\ -\dfrac{I_s}{C_0} \end{bmatrix}$$

$$x(t) = \begin{bmatrix} \Delta i_L(t) \\ \Delta v_o(t) \end{bmatrix}, u(t) = \Delta\mu(t), y = \begin{bmatrix} y_i \\ y_v \end{bmatrix}, C_c = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$$

Here μ is duty ratio. When controlling the current i_L of the sum of each phase, R_0 (equivalent resistance of inductor) is $R_1R_2/(R_1+R_2)$ and L_0 is $L_1L_2/(L_1+L_2)$. And Δi_L, Δv_o, $\Delta\mu$ are small-signal variables. And y_i,= Δi_L is a small signal inductor current and $y_v = \Delta v_o$ is a small signal output voltage.

From this equation, matrix A and B of the interleave boost converter depends on duty ratio μ_s. Therefore, the converter response will be changed depending on the operating point and other parameter variations. The changes of the load R_L, the duty ratio μ_s, the output voltage V_s and the inductor current I_s in the controlled object are considered as parameter changes in eq. (1). Such parameter changes can be replaced with the equivalent disturbances inputted to the input and the output of the controlled object. Therefore, what is necessary is just to constitute the control systems whose pulse transfer functions from equivalent disturbances to the output y become as small as possible in their amplitudes, in order to robustize or suppress the influence of these parameter changes.

3 DIGITAL ROBUST CONTROLLERS

3.1 Discretization of controlled object

The continuous system of eq. (1) is transformed into the discrete system as follows:

$$\begin{aligned} x_d(k+1) &= A_d x_d(k) + B_d u(k) \\ y(k) &= C_d x_d(k) \end{aligned} \qquad (4)$$

3

where

$$A_d = \left[e^{A_c T}\right], B_d = \left[\int_0^T e^{A_c \tau} B_c d\tau\right], C_d = C_c$$

Here, in order to compensate the delay time by A/D conversion time and micro-processor operation time etc., one delay (state ξ_1) is introduced to input of the controlled object. Then the state-space equation is described as follows:

$$\begin{aligned} x_{dt}(k+1) &= A_{dt} x_{dt}(k) + B_{dt} v(k) \\ y(k) &= C_{dt} x_{dt}(k) \end{aligned} \qquad (3)$$

where

$$A_{dt} = \begin{bmatrix} e^{A_c T_s} & e^{A_c (T_s - L_d)} \int_0^{L_d} e^{A_c \tau} B_c d\tau \\ 0 & 0 \end{bmatrix} = \begin{bmatrix} a_{11} & a_{12} & a_{13} \\ a_{21} & a_{22} & a_{23} \\ 0 & 0 & 0 \end{bmatrix}$$

$$B_{dt} = \begin{bmatrix} \int_0^{T_s - L_d} e^{A_c \tau} B_c d\tau \\ 1 \end{bmatrix} = \begin{bmatrix} b_{11} \\ b_{21} \\ 1 \end{bmatrix} \quad x_{dt}(k) = \begin{bmatrix} x(k) \\ \xi_1(k) \end{bmatrix}$$

$$C_{dt} = \begin{bmatrix} C_c & 0 \end{bmatrix} = \begin{bmatrix} 1 & 0 & 0 \end{bmatrix}$$

3.2 A2DOF digital current controller

The transfer function from the reference input r_i' to the output y_i is specified as follows:

$$W_{r_i'y_i}(z) = \frac{(1+H_1)}{(z+H_1)} \frac{(1+H_2)}{(z+H_2)} \frac{(1+H_3)}{(z+H_3)} \times \frac{(z-n_{1i})}{(1-n_{1i})} \frac{(z-n_{2i})}{(1-n_{2i})} \qquad (4)$$

Here H_i, i=1,…,3 are the specified arbitrary parameters, n_{1i} and n_{2i} are the zeros of the discrete-time controlled object. This target characteristic W_{ryi} is realizable by constituting the model matching system shown in Fig.5 using the following state feedback to the controlled object (5).

$$v = -Fx_{dt} - G_i r_i' \qquad (5)$$

Here $F = [f_1 \quad f_2 \quad f_3]$ and G_i are selected suitably. In Fig. 5, q_v and q_{yi} are the equivalent disturbances with which the parameter changes of the controlled object are replaced.

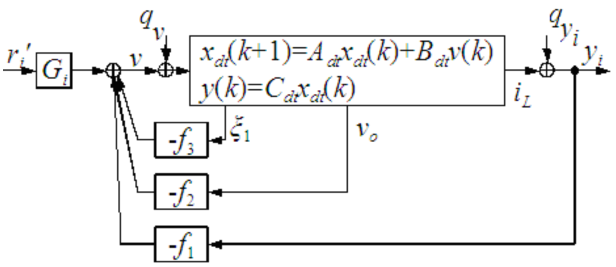

Fig.2 Model matching system using state feedback

It shall be specified that the relation of H_1 and H_3 become $|H_1| >> |H_3|$ and $n_{1i} \approx H_2$. Then $W_{ri'yi}$ can be approximated to the following first-order discrete-time model:

$$W_{r_i'y_i}(z) \approx W_{mi}(z) = \frac{1+H_1}{z+H_1} \qquad (5)$$

The transfer function $W_{Qyi}(z)$ between the equivalent disturbance $Q_i = [q_v \ q_{yi}]^T$ to y_i of the system in Fig.2 is defined as

$$W_{Qyi}(z) = \left[W_{qvyi}(z) \quad W_{qyyi}(z)\right] \quad (10) \qquad (6)$$

The system added the inverse system and the filter to the system of Fig.2 is constituted as shown in Fig.3.

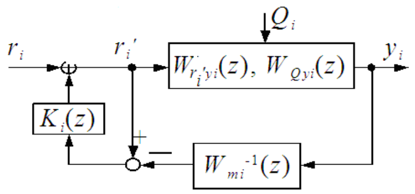

Fig. 3 System Reconstituted with Inverse System and F ilter

In Fig. 3, the transfer function $K_i(z)$ is as follows:

$$K_i(z) = \frac{k_{zi}}{z-1+k_{zi}} \qquad (7)$$

The transfer functions between r_i -y_i, q_{ui} -y_i and q_{yi} -y_i of the system in Fig.6 are given by

$$y_i = \frac{1+H_1}{z+H_1}\frac{z-1+k_{zi}}{z-1+k_{zi}W_{si}(z)}W_{si}(z)r_i \qquad (8)$$

$$y_i = \frac{z-1+k_{zi}}{z-1+k_{zi}}\frac{z-1+k_{zi}}{z-1+k_{zi}W_{si}(z)}W_{Qyi}(z)Q_i \qquad (9)$$

where

$$W_{si}(z) = \frac{(1+H_3)(z-n_{1i})}{(z+H_3)(1-n_{1i})}$$

Here, if $W_{si}(z) \approx 1$, then eq. (8) and eq. (9) are approximated, respectively as follows:

$$y_i = \frac{1+H_1}{z+H_1}r_i \qquad (10)$$

$$y_i = \frac{z-1}{z-1+k_{zi}}W_{Qyi}(z)Q_i \qquad (11)$$

From eq. (10), (11), it turns out that the characteristics from r to y can be specified with H_1 and the characteristics from Q_i to y_i can be independently specified with k_{zi}. That is, the system in Fig. 3 is an A2DOF system, and its sensitivity against disturbances becomes lower with the increase of k_{zi}. If equivalent conversion of the controller in Fig.3, we obtain Fig. 4. Then, substituting a system of Fig. 2 to Fig. 4, A2DOF digital integral type control system will be obtained as shown in Fig. 5. In Fig. 5, the parameters of the controller are as follows:

$$k_1 = -f_1 - \frac{Gk_{zi}}{1+H_1}, \quad k_2 = -f_2 \qquad (12)$$
$$k_3 = -f_3, \quad k_{ii} = G_i k_{zi}, \quad k_{ri} = G_i$$

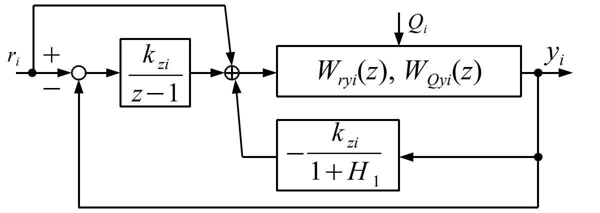

Fig. 4 Equivalent Conversion of the Robust Digital

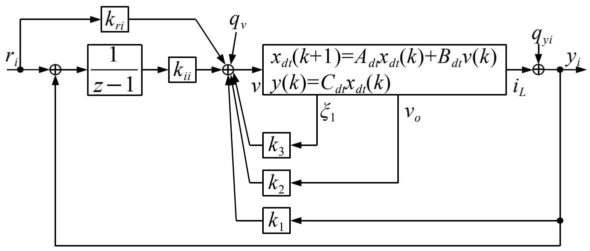

Fig. 5 Approximate 2DOF Digital Integral Type Curren t Control System

4 DIGITAL VOLTAGE CONTROLLER

4.1 Addition of u_v and v_{ac} to r_i

Add the multiplier in front of the reference input r_i of the current control system. Let the inputs of the multiplier be v_{ac} and u_v as shown in Fig. 6. v_{ac} is the absolute value of the input voltage v_{in} and u_v is a new input. This addition is for making the inductor current i_L follow the AC voltage v_{ac}.

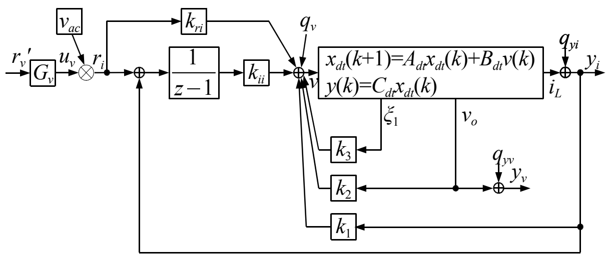

Fig. 6 Current Control System Added Multiplier

Next, the digital PI voltage controller is added to the input of Fig. 6. Then the digital robust control system including the A2DOF current controller and the PI voltage controller is obtained as shown in Fig, 7.

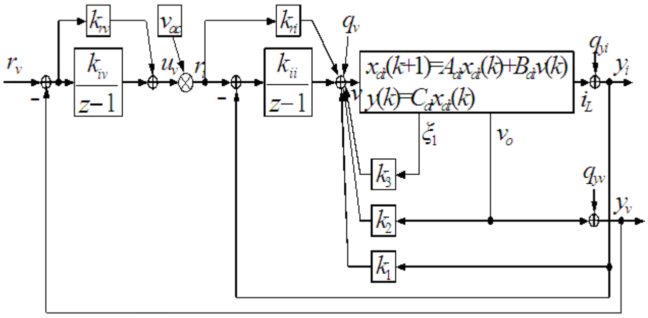

Fig. 7 Digital Robust Control System Including the A2DOF Current Controller and the PI Voltage Controller

4 Expermental Studies

All experimental setup system is manufactured. A micro-controller (RX) from Renesas Electronics is used for the digital controller. The digital PI current, and voltage controllers were implemented on 1 micro-processor.

The design parameters of the A2DOF current control system have been determined as

$$H_1=-0.999866 \quad H_2=-0.6 \quad H_3=0.1 \quad k_{zi}=0.5$$

And the parameters of the PI voltage controller have been determined as

$$k_{rv}=8 \qquad k_{iv}=0.01$$

The experiment results are shown in Fig. 8, 9, 10. The experiment result of the steady state at load RL=500Ω by proposed method are shown in Fig. 8. The input current waveform and the phase are the almost same as the input voltage at each load and PFC of the converter at load RL=500Ω are 0.991 and 0.985, respectively. The experiment result of the steady state at load RL=500Ω by usual phase lead-lag method are shown in Fig. 10. PFC of the converter at load RL=500Ω are 0.985. The experiment result of load sudden change from 1kΩ to 500Ω is shown in Fig. 6. In Fig, 6, the output voltage variation in sudden load change is less than 3V (0.78%). It turns out that the digital robust controller proposed is effective practically.

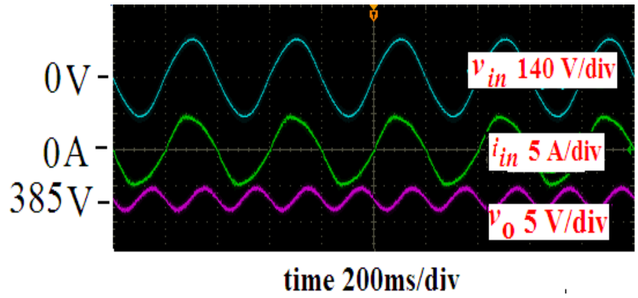

Fig.8 Experimental Results of Steady State Waveform, at load RL=500Ω by proposed method

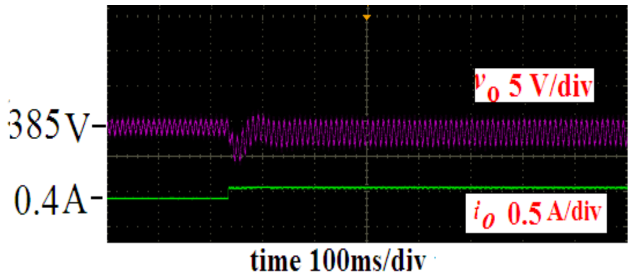

Fig.9 Experimental Results of Sudden Load Change from 1kΩ to 500Ω controlling the current for every phase

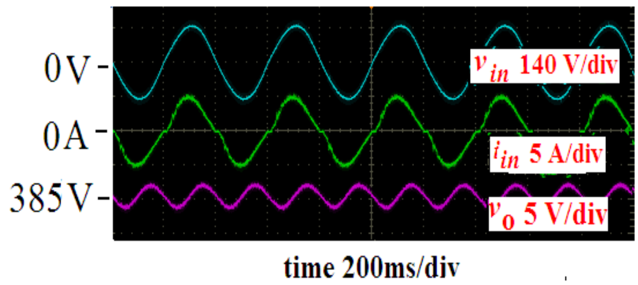

Fig.10 Experimental Results of Steady State Wavefor ms at load RL=500Ω by usual phase lead-lag method

4 CONCLUSION

. In this paper, the concept of controller of non-linear interleave PFC boost converter to attain good robustness was given. It was shown from experiments that the proposed A2DOF digital current controller can attain better performance.

REFERENCES

[1] K. Higuchi, K. Nakano, T. Kajikawa, E. Takegami, S. Tomioka, K. Watanabe, "A New Design of Robust Digital Controller for DC-DC Converters", IFAC 16th Triennial World Congress, (CD-ROM), 2005