IEEE 1999 International Conference on Power Electronics and Drive Systems, PEDS'99, July 1999, Hong Kong.

Control of An Active Clamp Discontinuous Conduction Mode Flyback Converter

C. T. CHOI, C. K. LI and S. K. KOK

Department of Electronic & Information Engineering,
The Hong Kong Polytechnic University,
Kowloon, HONG KONG SAR, CHINA

Abstract – Design of an Active-clamp Discontinuous Conduction Mode Flyback Converter (ADCMFC) is presented with the emphasis in output efficiency. The control of such a converter using PI, PID and LQR control had been explored. Both the simulation results and a practical circuit had been used to verify the design and the performance of the converter. Further suggestion of using neuro-fuzzy and generic type of controller is also proposed.

I. Introduction

Active-clamp discontinuous conduction mode flyback converter (ADCMFC) is a kind of fixed frequency PWM controlled discontinuous conduction mode flyback converter. This kind of converter shows better performance than ordinary flyback converter while being operated at higher frequency, e.g. above 500kHz. The high efficiency and high frequency are achieved by the zero-voltage switching of the power MOSFET and zero-current switching of the rectifier diode [1]-[2]. The open-loop small-signal transfer function for CCM flyback converter can be obtained analytically [3]. However, the dynamics of such a converter can vary substantially under different loading conditions. Such change of model parameters has been studied [4]. The corresponding controller design is achieved in this paper to provide control law that can be adapted with the changing load.

II. Design Considerations of ADCMFC

To realize zero voltage switching (ZVS) of the ADCMF converter that is shown in Fig. 1, the magnetizing inductance cannot be in a large value. The transformer design is basically the same as the normal flyback converter. Therefore, the DC magnetizing current I_m relates the output current I_{out} as:

$$\begin{aligned} I_m &= I_{in} + N_s I_{out} / N_p \\ &= I_{in} / \delta \\ &= (N_s / N_p) I_{out} / (1-\delta), \end{aligned} \tag{1}$$

where I_{in} is the input current,
I_{out} is the output current and
δ is the duty ratio of Q1.

For the realization of ZVS of the switches, the magnetizing current must decrease under zero so that it can discharge the capacitor C_p .The magnetizing inductance L_m of the transformer meets the inequality,

$$L_m < \frac{V_{in}\delta}{2f(I_m + I_3)}, \tag{2}$$

where f is the switching frequency and I_3 is the reverse current for discharging the parasitic capacitance of the MOSFET and transformer. I_3 can be determined by

$$I_3 = (V_{in} + N_p V_{out} / N_s) / Z_r, \tag{3}$$

where Z_r is the characteristic impedance of the magnetizing inductance L_m and the parasitic capacitor C_p and defined by

$$Z_r = \sqrt{L_m / C_p}\ . \tag{4}$$

In the ADCMFC, ZVS is basically achieved by utilizing the magnetizing current, so the magnetizing inductance of the transformer is restricted. In this topology, for the realization of zero current switching (ZCS) of the secondary rectifier diode, the resonant frequency must be set to complete the half wave of the sinusoidal current. Therefore, capacitance of the clamp capacitor C_r and inductance of the leakage inductor L_r are limited by the inequality

$$T_{off} / 2 < \pi\sqrt{(C_r L_r)} < T_{off}, \tag{5}$$

where T_{off} is the ON period of Q2. The period of half wave $\pi\sqrt{(C_r L_r)}$ does not have to be within the ON period of Q2. The conduction period of the diode is less than the half of the resonant period at the low output current. Therefore, the resonant period $2\pi\sqrt{(C_r L_r)}$ can be set according to the ZCS at the maximum output current. It is appropriate that the resonant period must be maximized to reduce the RMS current for the conversion efficiency.

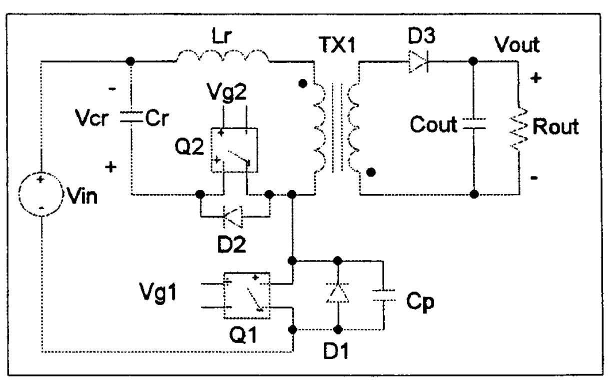


Fig. 1: Circuit of the ADCMFC

0-7803-5769-8/99/$10.00©1999IEEE

III. Experimental Setup

A practical converter was set up with the input voltage range of 75V-177V and output voltage of 24V, 1.2A Max and 29W. The switching frequency is 500kHz. The components and their values are:

Transformer TX1:
Core: TDK PQ26/20
Primary: 12 turns of enamel wire AWG#27
Secondary: 6 turns of enamel wire AWG#23
L_m: 20.42uH, L_r: 0.76uH

$C_r = 0.05\mu F$
$C_p = 340pF$
$C_{out} = 560\mu F, 35V$
$C_{in} = 470\mu F, 200V$
$Q_1, Q_2 = IRFBC30$ (International Rectifier)
$D_3 = STPR1010CT$ (SGS-Thomson)

The schematic diagram of the experiment is shown in Fig. 2. In this experiment, the efficiency of the converter was measured and compared with a conventional R-C-D clamp flyback converter. The results are tabulated in Table 1.

V_{in}	I_{out}	Active-clamp flyback	R-C-D clamp flyback
75V	1.2A	85.25%	78.35%
90V	1.2A	84.50%	78.06%
127V	1.2A	84.10%	77.25%
140V	1.2A	83.60%	76.70%
160V	1.2A	83.20%	73.70%
177V	1.2A	83.07%	71.45%

Table 1: Efficiency Comparison

IV. Controller Design for ADCMFC

If this converter is designed to operate in such a variation of input voltage and load condition, a complete model of the converter must be obtained before a controller can be designed. Here, the technique used by Banos and Gomez [5] is employed to identify the parameters of the model under the diversified operating conditions.

In the frequency response, magnitude and phase angle are two factors to determine the curve fitting function. The cost function for the identification routine is based on a square root of summation of the square of the magnitude error and the phase error. The cost function is

$$J = \sqrt{\sum |e|^2 + \sum \angle e^2} \quad (1)$$

where

$$|e| = W(\log|G_0(j\omega)| - \log|G_r(j\omega)|)$$

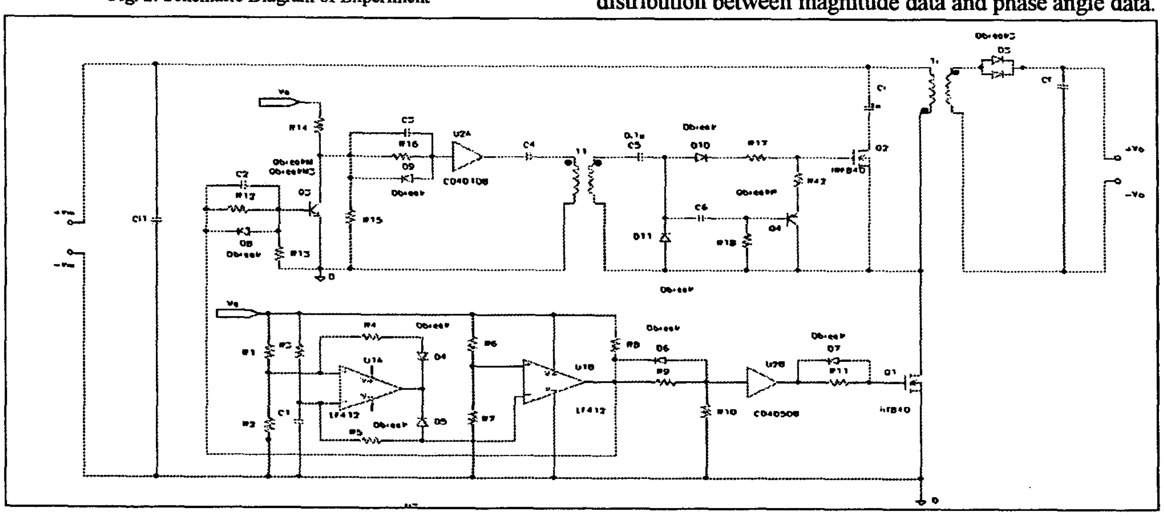
$$\angle e = \angle G_0(j\omega) - \angle G_r(j\omega)$$

G_r are the measured data, G_0 are the optimized data and W is the weighting factor. The optimization method selected for the cost function is Nelder-Mead simplex algorithm, which is implemented in the optimization toolbox of the MATLAB under the function fmins(). The operation of this algorithm is to adjust the magnitudes and phase angles of the optimized transfer function in order to get the minimum value of cost function J. The optimization will be stopped when the tolerance of J reaches 1E-3.

The curve fitting optimization of the transfer function is applied up to 100kHz because the switching frequency of ADCMFC is 500kHz. The best control speed is only up to one-fifth of the operating frequency.

The weighting factor W can be used to balance the weighting of errors between the magnitude data and the phase angle data. The selection of W depends not only on the least cost function but also on the balance of the two errors. A simulation test has been done to select this weighting factor W so that it has almost equal proportion of the error distribution between magnitude data and phase angle data.

Fig. 2: Schematic Diagram of Experiment



Before the optimization starts, it is necessary to determine the number of zeros and poles first. A simulation test has also been used. The cost function J is the least if 3 zeros and 4 poles are used. In fact, the error improvement is at the high frequency side only. In real application, the controller response for this ADCMFC is less than 100kHz and thus high frequency side improvement may not be significant. Therefore, the selection of 1 zero and 2 poles is determined because it provides sufficient accuracy and simpler transfer function format.

The frequency response can be approximated by a second order linear transfer function of the following form

$$G(S) = K_a \frac{s+z}{(s-p_1)(s-p_2)}. \qquad (6)$$

Notice that there are one right-half-plane zero and two left-half-plane poles.

Then, the controllers can be designed accordingly. The most common types of PI, PID and LQR controllers are used for the evaluation of the converter performance.

The proportional term in the PI and PID controller is to increase the transient response of the system and to reduce the magnitude of any steady state errors. The integral term is to remove any steady state position error. However, this kind of controller will introduce extra phase lag to the system. Therefore, the corresponding parameters must be chosen very carefully. The corresponding transfer function of the PI and PID controller is given as

$$PI(s) = (K_p s + K_v)/s \qquad (7)$$

and

$$PID(s) = K_g + K_d s + K_i / s. \qquad (8)$$

The specifications of the closed loop controlled converter are

- The roll-off bandwidth is around 80kHz,
- The phase margin is greater than 30°,
- Overshoot is less than 25%.

The reason to have the system operating in an under-damped condition is to achieve rapid transient response. This is critical in this ADCMFC converter control because it is running at very high operating frequency 500kHz.

Optimization technique [4]-[5] is used to obtain the controller parameters accordingly. The cost function for the optimization is

$$J = \sqrt{\sum (1-y_i)^2}\, DF + DT \qquad (9)$$

where y_i is the discrete time step response of the system, DT the time domain requirement error and DF the frequency domain requirement error.

For Linear Quadratic Regulator (LQR) design [6], state space feedback is used and similar closed-loop performance requirement is assumed. However, in the LQR design, the DF term in the cost function J is ignored.

V. Results

From extensive computer simulation and practical experiment, the overshoot of the closed-loop optimal PI and PID controller is tabulated in Table 2.

V_{in} (V)	I_{out} (A)	PI Overshoot(%)	PID Overshoot(%)
75	1.20	13.46	15.29
	0.60	19.41	20.18
	0.20	12.93	16.44
90	1.20	15.13	17.12
	0.60	21.09	20.86
	0.20	14.95	15.56
127	1.20	19.54	21.93
	0.60	23.23	24.12
	0.20	14.74	16.26
140	1.20	23.15	25.00
	0.60	24.18	24.83
	0.20	15.12	18.74
160	1.20	24.30	24.05
	0.60	24.96	24.99
	0.20	15.77	18.79
177	1.20	24.32	24.56
	0.60	24.04	25.00
	0.20	16.88	18.69

Table 2: Percentage of Overshoots for PI and PID Control under Various Operating Conditions.

For the LQR design, there is no significant overshoot as expected. The phase margin for the three different designs is tabulated in Table 3.

V_{in} (V)	I_{out} (A)	PI	PID	LQR
		PM (deg.)	PM (deg.)	PM (deg.)
75	1.20	64.19	61.05	90.14
	0.60	56.98	55.84	90.58
	0.20	65.04	59.16	90.60
90	1.20	61.96	58.75	90.46
	0.60	54.67	55.00	90.70
	0.20	61.65	60.63	90.81
127	1.20	56.83	53.40	90.76
	0.60	52.66	51.43	91.17
	0.20	61.95	59.50	91.01
140	1.20	52.98	50.49	91.64
	0.60	52.14	51.25	92.01
	0.20	61.02	55.49	91.19
160	1.20	52.00	52.36	91.86
	0.60	52.44	52.39	92.63
	0.20	60.51	55.88	91.51
177	1.20	52.07	51.74	92.42
	0.60	54.06	52.70	93.02
	0.20	59.37	56.60	91.64

Table 3: Phase Margin of the Three Different Designs

From the investigation carried out, we have learnt that when selecting the high frequency magnetic material for the power transformer, the switching loss of the power transformer would become the major thermal problem in high switching frequency. The other problem is that a high current rating device must be used for the output rectifier diode. The investigation also proves that the ADCMFC can perform much better in terms of efficiency and output ripple than that of the traditional R-C-D clamp flyback design. Optimal PI, PID and LQR controller parameters can be obtained by the optimization technique. For LQR design, the overshoot can be totally eliminated. However, it cannot guarantee specified roll-off frequency because the prescribed cost function does not include any frequency response constraint. For PI and PID controller, the roll-off frequency is 80kHz, which is faithfully preserved. The corresponding overshoots are also within the design limit. The overshoot value is around 15% for V_{in} at 75-90V and it reaches 25% for V_{in} at 160-188 V. The phase margin also falls into the region of around 65° at low V_{in} but it degrades to 50° at high V_{in}.

VI. Conclusion

From the study here, the observation is that there is little difference between the performance of the closed loop system under PI and PID. LQR is not suitable due to its uncontrolled frequency characteristics. With the design methodology suggested in this paper, engineers can design appropriate controllers more systematically without too much "trial and error" approach. Moreover, neuro-fuzzy and generic type of controller can be used to adaptively control the optimal operating points of this ADCMFC.

References

[1] K. Yoshida, T. Ishii, and N. Nagagata, "Zero Voltage Switching Approach for Flyback Converter," Proceedings of 14th International Telecommunications Energy Conference, 1992, INTELEC '92., pp. 324-329.

[2] R. Watson, F. C. Lee, and G. C. Hua, "Utilization of an Active-Clamp Circuit to Achieve Soft Switching in Flyback Converters," IEEE Transactions on Power Electronics, vol. 11, No. 1, pp. 162-169, January 1996.

[3] M. K. Kazimierczuk and S. T. Nguyen, "Small-Signal Analysis of Open-Loop PWM Flyback DC-DC Converter for CCM," Proceedings of the IEEE 1995 National Aerospace and Electronics Conference, 1995, NAECON 1995, vol. 1, pp. 69-76.

[4] S. K. Kok, "Modeling and Control of an Active Clamp Discontinuous Conduction Mode Flyback Converter (ADCMFC)", MSc dissertation, The Hong Kong Polytechnic University, June 1996.

[5] A. Banos and F. Gomez, "Parametric Identification of Transfer Functions from Frequency Response Data", Computing & Control Engineering Journal, pp.137-144, June, 1995.

[6] F. H. F. Leung, P. K. Tam, C. K. Li, "The Control of Switching DC-DC Converters - A General LQR Problem," IEEE Transactions on Industrial Electronics, Vol. 38, No. 1, pp. 65-71, February 1991.