Feedback Compensator Design for a Two-Switch Forward Converter

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Abstract—**Advantages of a two-switch forward converter over its single-switch counterpart are that it operates with reduced voltage stress on power semiconductor switches and does not require a transformer with a reset winding. The lower stress on the switching devices leads to increased reliability and, without the reset windings, the transformer production is simplified. In this paper, feedback compensator design for a two-switch forward converter is presented. A Proportional-plus-Integral (PI) compensator is chosen for its simplicity. Experimental results confirm that the properly designed PI compensator is capable of providing the satisfying output voltage regulation and transient response.**

Keywords-switched mode power supplies; forward converter; feedback compensation

I. INTRODUCTION

A two-switch forward converter [1] has many advantages over a conventional forward converter. It operates with reduced voltage stress on power semiconductor switches and does not require a transformer with a reset winding, simplifying the transformer design and production. Moreover, a turn-off voltage spike imposed on the power switches is less severed in this two-switch topology, as the circuit arrangement provides a path for the transformer's leakage inductance to release its stored energy to the input source. All these distinctive features make the two-switch forward converter attractive and replace the conventional forward converter in many applications. The desirable performance of any converter is good output voltage regulation and fast transient response to disturbances such as a step load or a step input voltage. These targets can be attained by feedback control of a converter's output voltage. A compensator is usually included in this feedback loop to adjust gain and phase of a control signal. This compensated control signal, when compared with a constant-frequency sawtooth signal, generates a pulse-width-modulated (PWM) gate signal that drives the converter towards maintaining the constant output voltage. Output voltage regulation and transient characteristics are largely dependent on the type of compensators used and how well it has been designed [2].

Existing literatures of a two-switch forward converter [1,3,4] had mainly focused on the circuit operation and design. This paper presents feedback compensator design for a two-switch forward converter. The design is illustrated using a PI compensator chosen for its simplicity. Experimental results confirm that the properly designed PI compensator is capable of providing the satisfying output voltage regulation and transient response.

II. TWO-SWITCH FORWARD CONVERTER

In Continuous Conduction Mode (CCM), the steady-state operation of the two-switch forward converter in Fig. 1 can be divided into three stages as follows:

When M_1 and M_2 are turned on (Fig. 2(a)), the input voltage, V_{in}, is applied across the primary winding, causing magnetic flux to build up in the transformer core, i.e. the magnetizing current, i_{Lm}, is increasing. The induced secondary voltage, v_2, forward biases the diode D_3. The secondary current flows through the inductor to the output capacitor (C) and load resistance (R). The primary current, i_{in}, is made up of two current components, i_{Lm}, and the inductor current reflected to the primary, i'_L.

When M_1 and M_2 are turned off, i.e. D_1 and D_2 are turned on (Fig. 2(b)), the negative V_{in} is applied across the primary winding, resetting magnetic flux in the transformer core and causing i_{Lm} to decrease. The polarity reversal of v_2 reverse biases D_3, while D_4 is forward biased to allow the continuous flow of i_L. The primary current, i_{in}, is equal, but opposite, to the demagnetizing current, i_{Lm}.

When both M_1 and M_2 and D_1 and D_2 are turned off (Fig. 2(c)), the transformer core has been completely demagnetized, i.e. i_{Lm} is equal to zero. On the secondary, i_L continues flowing through D_4

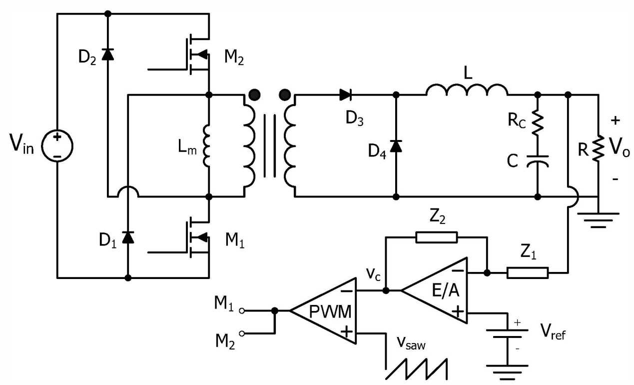
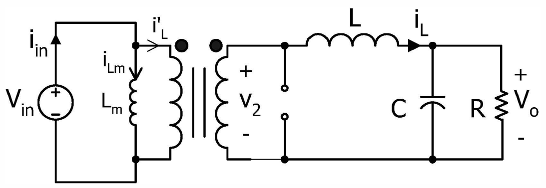
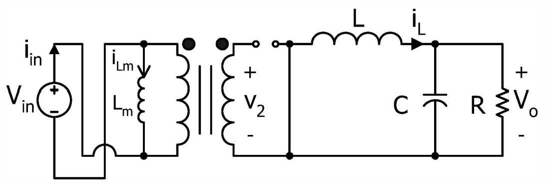


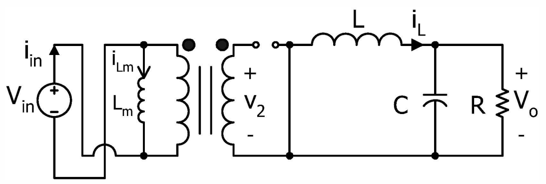
Figure 1. Two-switch forward converter.



(a)



(b)



(c)

Figure 2. Two-switch forward converter: (a) when M_1 and M_2 are on, (b) when D_1 and D_2 are on, and (c) when M_1-M_2 and D_1-D_2 are off.

From the above operation, the magnetic core is demagnetized by the conduction of D_1 and D_2 which applied the negative voltage to the primary. This flux reset scheme not only eliminates a need of the tertiary winding, but also reduces voltage stress on the power MOSFET switches (M_1 and M_2) to half of that in the conventional forward converter. The circuit on the secondary operates in a similar manner to a buck converter. Therefore, the output voltage is DC with a small switching ripple. In CCM, the relationship between the input and output voltages is given by:

$$\frac{V_o}{V_{in}} = nD \tag{1}$$

where n is a secondary-to-primary turn ratio and D is a duty cycle of M_1 and M_2.

As seen in Fig. 1, the output voltage is regulated to a desirable value by feedback control. The control signal, V_c, is generated by amplifying the difference between the reference voltage, V_{ref}, and sensed output voltage. It is then compared with the sawtooth signal, V_{saw}, at the PWM comparator to produce the duty-cycle signal that drives M_1 and M_2 towards maintaining the constant output voltage. The feedback compensator represented by Z_1 and Z_2 dictates gain and phase of the control signal. The properly designed compensator can result in a closed-loop converter having stable operation, good voltage regulation, and fast transient response.

III. Overview of Compensator Design

A control block diagram of the two-switch forward converter in Fig. 1 can be drawn as shown in Fig. 3 [2].

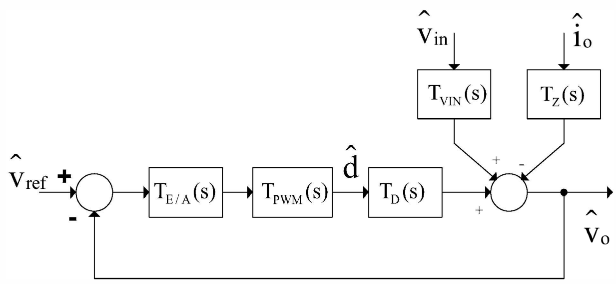


Figure 3. Two-switch forward converter control block diagram.

An output voltage variation ($\hat{v}_o$) is dependent on variations in a duty cycle ($\hat{d}$), input voltage ($\hat{v}_{in}$), and load current ($\hat{i}_o$), which can be expressed as

$$\hat{v}_o = T_D(s)\hat{d} + T_{VIN}(s)\hat{v}_{in} + T_Z(s)\hat{i}_o \tag{2}$$

where $T_D(s)$, $T_{VIN}(s)$, and $T_Z(s)$ represent a small-signal duty cycle-to-output voltage transfer function, a small-signal input voltage-to-output voltage transfer function, and an output impedance transfer function respectively. The variables $\hat{v}_{in}$ and $\hat{i}_o$ are treated as being disturbances to the system and can be ignored (i.e. set to zero) when designing the compensator. As a result, an open loop transfer function of the converter, T(s), is defined as:

$$T(s) = T_{E/A}(s)T_{PWM}(s)T_D(s) \tag{3}$$

where $T_{E/A}(s)$ and $T_{PWM}(s)$ are the transfer functions of the compensator and PWM comparator respectively. It can be shown that by designing $T_{E/A}(s)$ to make T(s) sufficiently large in magnitude, output voltage regulation of the converter can be improved and, at the same time, the effects of the input voltage and load current variations on the output voltage are reduced. Nonetheless, the large T(s) tends to cause damped oscillation in the output response and, if too large, it can cause instability. Hence, the main objective of the $T_{E/A}(s)$ design is to reach compromise between good output performance and stability.

The compensator can be designed by the frequency response method [5], which essentially involves placement of poles and zeros of $T_{E/A}(s)$, in such a way that the resulting open loop transfer function, T(s), exhibits the high DC gain, rolls off with a slope of –20dB/decade, and crosses the unity gain (0dB) line at reasonably high frequencies. The characteristics of T(s) are qualitatively related to time-domain performances of the converter. The high DC gain indicates good output voltage regulation, the large crossover frequency (or bandwidth) is akin to the fast output response, and the slope of –20dB/decade at the crossover frequency ensures the system will have an adequate phase margin, an indicator of good stability.

IV. CONVERTER TRANSFER FUNCTIONS

The two-switch forward converter is the topology derived from a buck converter. It hence possesses the same form of duty cycle-to-output voltage transfer function, $T_D(s)$, as the buck converter, with the gain lowered by the transformer's turn ratio.

$$T_D(s) = \frac{\hat{v}_o(s)}{\hat{d}(s)} = \frac{nV_{in}(1+\frac{s}{\omega_z})}{1+\left(\frac{s}{Q\omega_0}\right)+\left(\frac{s}{\omega_0}\right)^2} \quad (4)$$

where $\omega_z = \frac{1}{R_c C}$, $\omega_0 = \frac{1}{\sqrt{LC}}$, $Q = \frac{1}{\omega_0}\left(\frac{R}{RR_c C + L}\right)$. The transfer function is composed of double poles at ω_0 and a zero at ω_z.

The PWM comparator transfer function, $T_{PWM}(s)$, is expressed by [2]:

$$T_{PWM}(s) = \frac{\hat{d}(s)}{\hat{v}_c(s)} = \frac{1}{V_S} \quad (5)$$

where V_S is an amplitude of the sawtooth signal. Multiplying (4) and (5) yields the transfer function $T_D(s)T_{PWM}(s)$:

$$T_D(s)T_{PWM}(s) = \frac{\frac{nV_{in}}{V_s}(1+\frac{s}{\omega_z})}{1+\left(\frac{s}{Q\omega_0}\right)+\left(\frac{s}{\omega_0}\right)^2} \quad (6)$$

The PI compensator in Fig. 4 is chosen to compensate the feedback loop and its transfer function is expressed by

$$T_{E/A}(s) = \frac{K_{PI}(1+\frac{s}{\omega_{z,\,PI}})}{s} \quad (7)$$

where $K_{PI} = 1/C_1R_1$, $\omega_{z,\,PI} = 1/R_2C_1$. $T_{E/A}(s)$ has a pole at origin and a zero at $\omega_{z,\,PI}$. In the design process, $\omega_{z,\,PI}$ must be placed below ω_0 of $T_D(s)T_{PWM}(s)$; otherwise a phase lag due to the pole at origin of the PI compensator and that due to the double poles at ω_0 of $T_D(s)T_{PWM}(s)$ will combine and potentially cause instability. Hence, the crossover frequency of the open-loop transfer function in (3) must be less than ω_0 of $T_D(s)T_{PWM}(s)$. K_{PI} controls the gain of the PI compensator. Large K_{PI} will increase a speed of the output voltage response but, if too large, the system will become unstable.

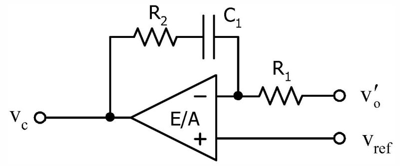


Figure 4. PI compensator.

V. COMPENSATOR DESIGN

The two-switch forward converter has the following circuit values: L = 66μH, C = 300μF, n = 0.05, R_c = 0.083Ω, and R = 0.5 to 5Ω. The switching frequency is 100kHz. The DC input voltage, V_{in}, in Fig. 1 is obtained by rectifying the AC main supply of 220V±10%; hence V_{in} is ranging from 280V to 342V. The output voltage is fed back and regulated to 5V by the control IC UC3825 [6]. The compensator's circuit components, comprising R_1, R_2, and C_1, is connected around the error amplifier (E/A), which is accessible through pins 1 (an inverting input), 2 (a non-inverting input), and 3 (an output) of the IC. The PWM comparator and sawtooth signal generator are both internal to the IC. The amplitude of the sawtooth signal, V_s, is 1.8V.

In designing the compensator, the converter is assumed to be operating under the minimum input voltage and maximum output current condition, i.e. V_{in} = 280V and R = 0.5Ω. Substituting all the relevant parameters into (6), the DC gain, ω_0, and ω_z of the transfer function $T_D(s)T_{PWM}(s)$ are found to be -8.9dB, 7,107rad/sec, and 40,210rad/sec respectively. The PI compensator in (7) is designed for two different values of gain, i.e. K_{PI} = 1000 and K_{PI} = 2000. In both cases, the zero $\omega_{z,\,PI}$ of the PI compensator is placed at the frequency $0.4\omega_0$ or 2850rad/sec. For K_{PI} = 1000, component values of the compensator in Fig. 4 are: R_1 = 10kΩ, R_2 = 3.3kΩ, C_1 = 0.1μF, and for K_{PI} = 2000 the component values are: R_1 = 5kΩ, R_2 = 3.3kΩ, C_1 = 0.1μF.

Bode plots of the converter's open-loop transfer function are shown in Fig. 5. The compensator with the gain K_{PI} = 2000 results in the open-loop transfer function having higher gain and bandwidth (700rad/s for K_{PI} = 2000 versus 400rad/s for K_{PI} = 1000) Thus, it is expected to have better voltage regulation and faster transient response than when K_{PI} =1000.

VI. EXPERIMENTAL VERIFICATION

The two-switch forward converter was prototyped with the designed PI compensator being used in the feedback loop. Its output voltage was measured under a varying input voltage and load current. The result is presented in Tables 1 and 2 for K_{PI} = 1000 and K_{PI} = 2000 respectively. As seen in the tables, the output voltage is regulated to around 5V despite the changing input voltage and load current. At a given input voltage, when the load current increases from minimum to maximum, the output voltage drops only 0.013V for K_{PI} = 1000 and 0.012V for K_{PI} = 2000. The reason for good output voltage regulation can be explained through the closed-loop small-signal input voltage-to-output voltage transfer function (or audio-susceptibility) and output current-to-output voltage transfer function (or output impedance) derived from Fig. 3 and are given in (8) and (9) respectively.

$$\frac{\hat{v}_o}{\hat{v}_{in}} = \frac{T_{VIN}(s)}{1+T(s)} \tag{8}$$

$$\frac{\hat{v}_o}{\hat{i}_o} = \frac{T_Z(s)}{1+T(s)} \tag{9}$$

At steady state (s=0), the gain of T(s) is theoretically infinite due to the pole at origin of the PI compensator. Hence, the term 1+T(s) in dominators of (8) and (9) is very large. As a result, the output voltage variation becomes almost independent of the input voltage and load current variations.

The output voltage transient responses under a step load are given in Fig. 6 and 7. In both cases, the output voltage exhibits a damped oscillatory characteristic which is typical for the PI compensator. It was earlier predicted from Fig. 5 that the compensator with K_{PI} =2000 would yield the faster response as it yielded the larger system bandwidth. This prediction can be verified by observing the settling time of the responses in Fig. 6 and 7. The settling time is approximately 1.4ms for K_{PI} = 1000 (Fig. 6), and 1ms for K_{PI} = 2000 (Fig. 7).

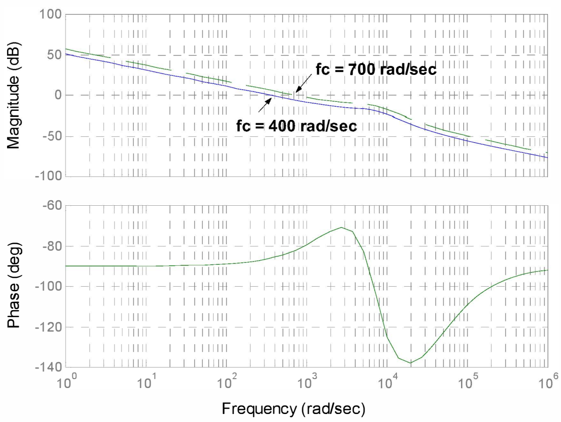


Figure 5. Bode plots of the open loop transfer function for K_{PI} = 1000 (solid line) and K_{PI} = 2000 (dashed line).

TABLE I. MEASURED OUTPUT VOLTAGE WITH K_{PI} = 1000

I_o	2A	4A	6A	8A	10A
V_{in} = 280V_{dc}	5.010	5.007	5.004	5.002	4.999
V_{in} = 311V_{dc}	5.010	5.007	5.005	5.002	4.999
V_{in} = 342V_{dc}	5.010	5.007	5.004	5.002	4.999

TABLE II. MEASURED OUTPUT VOLTAGE WITH K_{PI} = 2000

I_o	2A	4A	6A	8A	10A
V_{in} = 280V_{dc}	5.005	5.002	4.999	4.997	4.995
V_{in} = 311V_{dc}	5.005	5.002	4.999	4.997	4.995
V_{in} = 342V_{dc}	5.005	5.002	4.999	4.997	4.995

VII. CONCLUSION

In this paper, feedback compensator design for a two-switch forward converter has been presented. The PI compensator was selected because of its simplicity. The compensator design involved selection of the gain, $K_{PI,}$ and zero, $\omega_{z,\,PI}$, to get high open-loop gain and bandwidth possible, at the same time not breaching stability criterion. It was experimentally shown that the properly designed PI compensator is capable of providing the satisfying output voltage regulation and transient response.

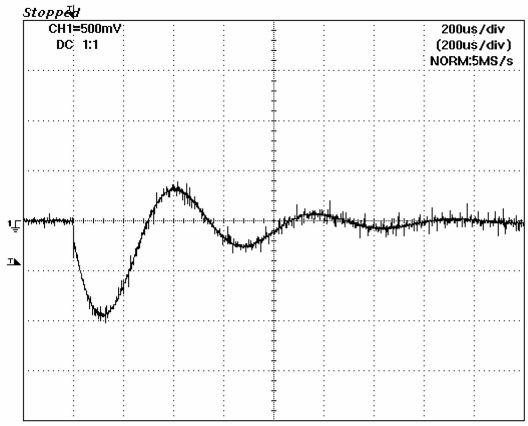


Figure 6. Output voltage response of the converter (K_{PI} = 1000), when the load current is stepped from 1A to 5A (X-scale: 200µs/div, Y-scale: 0.5V/div)

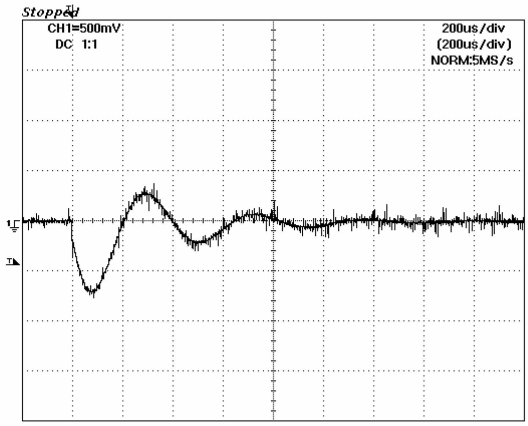


Figure 7. Output voltage response of the converter (K_{PI} = 2000), when the load current is stepped from 1A to 5A (X-scale: 200µs/div, Y-scale: 0.5V/div)

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