A NEW SENSORLESS CONTROL TECHNIQUE FOR PFC OPERATION OF AC-DC BUCK CONVERTER

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ABSTRACT

In this paper a new single phase ac-dc buck converter with a near unity power factor is presented. The principle of proposed control system is discussed. The good operation of circuit is verified by simulation carried out with PSCAD/EMTDC software.

I. INTRODUCTION

A buck type ac-dc single phase converter provides a dc output voltage less than the rms value of ac input voltage. Recently boost circuits are received more attention to achieve high power factor at input, because the power factor correction (PFC) operation of boost converters has a better performance than buck converters at lower cost [1-3].

Suggested methods usually add extra switches or inductors to the circuit [4]. Using extra voltage or current sensors is a usual strategy to achieve high power factor. Obviously all of these methods involve expense in power or control circuits [5].

In some of these suggested circuits, researchers use integrative elements or band pass filters to produce control signals from new modulation methods [6]. Clearly, these ways invite the complication of the control system. In this paper a buck-type single phase single switch ac-dc converter with PFC operation is proposed. This circuit can work in low switching frequency. Another improvement, obtained by this proposed control strategy, is that dc voltage and ac current sensors used in the conventional PFC converters are not required. In the proposed control system, by removing dc voltage sensor, ac line voltage and ac or dc current sensor, just a sinusoidal waveform is needed to synchronize the power circuit and the control circuit. The output dc voltage can be controlled by the command input signal $k = V_o / E_a$.

The topology of the proposed circuit and the principle of the sensorless control method are presented. The performance of the circuit and its control system in increasing power factor are shown by simulations.

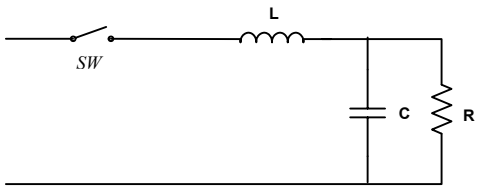


Figure 1. Conventional buck chopper

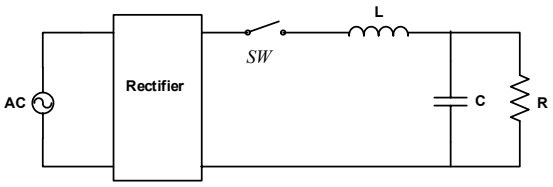


Figure 2. Ac-dc buck converter

Additionally, the optimization of the circuit parameters for decreasing the total harmonic distortion (THD) of the input current is discussed. Finally, the operation of the proposed circuit in several gain factors is examined.

II. TOPOLOGY AND PRINCIPLE OF CONTROL

The main configuration of this circuit is based on a conventional dc-dc buck chopper. A dc-dc buck circuit is shown in figure 1.

Considering the inductor voltage in a period of switching

$$\int_0^{T_s} v_L dt = \int_0^{t_{on}} v_L dt + \int_{t_{on}}^{T_s} v_L dt = 0 \qquad (1)$$

$$(V_{in} - V_o) t_{on} = V_o (T_s - t_{on}) \qquad (2)$$

$$\Rightarrow \frac{V_o}{V_{in}} = \frac{t_{on}}{T_s} = D \qquad (3)$$

Duty cycle D is the *ON* time of switch, t_{on}, to the cycle time of switching, T_s.

An ac-dc buck chopper is shown in figure 2. The duty cycle of switch *SW* (D) must be calculated to obtain an input current in phase with input voltage to achieve PFC operation. Before calculating D , we bring up a lemma.

Lemma: If a sinusoidal waveform, with frequency f and rms value A is multiplied by an *ON-OFF* waveform with frequency $4lf$ ($l = 1,2,3,...$) and duty cycle D a new waveform is obtained with rms value $A\sqrt{D/2\omega}$ (ω is the angular frequency of the sinusoidal waveform).

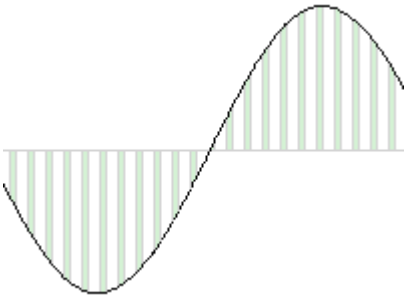


Figure 3. *Lemma*

The *lemma* is proofed in appendix. It must be considered that the rms value of the mentioned waveform and its absolute are the same, because in calculating rms the function is squared.

In figure 2 we suppose the ac line voltage and the desirable ac current are expressed as:

$$e_a = \sqrt{2}E_a \sin \omega t \tag{4}$$

$$i_a = \sqrt{2}I_a \sin \omega t \tag{5}$$

Where E_a and I_a are rms values of the ac line voltage and current respectively.

Voltage and current of inductor are expressed as:

$$v_L = \sqrt{2}E_a \sqrt{\frac{D}{2\omega}} \sin \omega t - V_o \tag{6}$$

$$i_L = \sqrt{2}I_a \sin \omega t \tag{7}$$

The differential equation shows the relation between voltage and current of L is:

$$v_L = L\frac{di_L}{dt} = L\sqrt{2}I_a \omega \cos \omega t \tag{8}$$

Neglecting the losses:

$$E_a I_a = \frac{V_o^2}{R} \tag{9}$$

The control variable k of the dc mean voltage V_o to ac rms voltage E_a is also defined by:

$$k = \frac{V_o}{E_a} \tag{10}$$

$$\Rightarrow D = (\sqrt{2}\frac{L\omega}{R}K^2\frac{\cos \omega t}{\sin \omega t} + \frac{K}{\sin \omega t})^2 \times \omega \tag{11}$$

The last equation gives the relation between duty cycle of *SW*, control variable k and circuit parameters. On the contrary, if *SW* is controlled by the switching duty cycle given by (11), there must be a sinusoidal current expressed by (5).

III. SIMULATION

Using PSCAD/EMTDC software we can simulate the operation of the proposed circuit. Power factor and control circuit are shown in figure 4 and figure 5 respectively. Circuit parameters during simulation are presented in table 1. By controlling the switch *SW*, with the duty cycle given by (11), the input current can be expected in phase with the input voltage if the frequency of the switching satisfies the condition of the *lemma*.

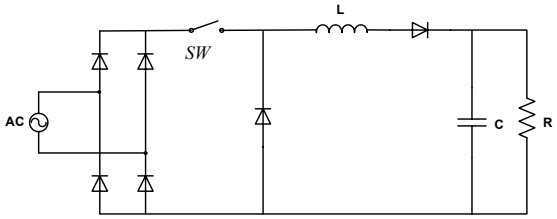


Figure 4. Configuration of proposed circuit

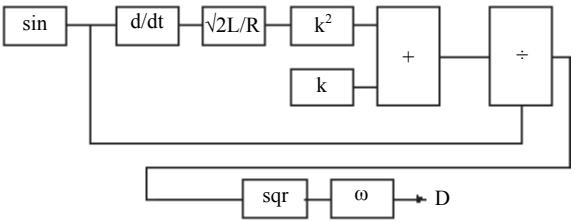
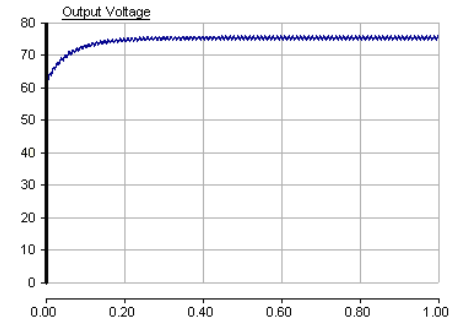


Figure 5. Proposed control system

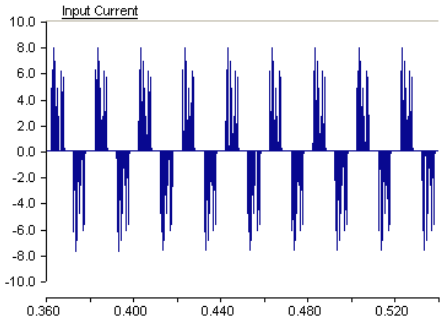
Table 1. Circuit parameters during simulation

Ac input frequency f	50 Hz
Input voltage rms E_a	100 volt
Switching frequency f_s	1200 Hz
L	1 mH
C	3 mF
R	100 Ω

The output voltage and input current for $k = 0.7$ are shown in figure 6. The phase difference of the input voltage and current is 1° in this case. Figure 7 shows the harmonics of the input current. As can be seen the largest harmonic after the main harmonic is the fifth.



(a)



(b)

Figure 6. Circuit waveforms for $k = 0.7$ (a) output dc voltage (b) input current

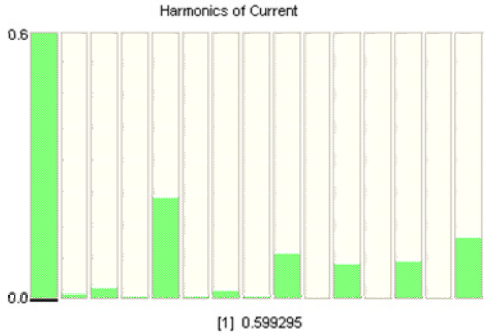
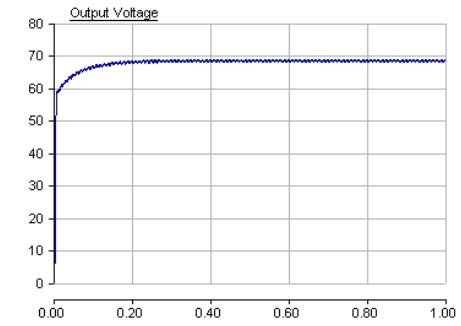
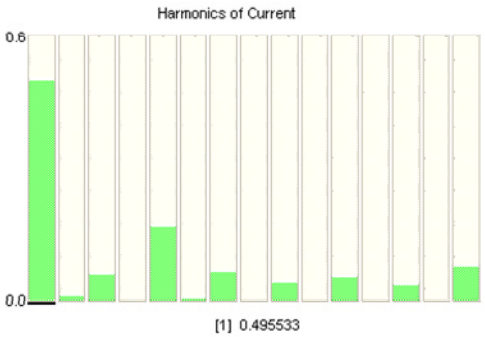


Figure 7. Harmonic spectrum of input current for $k = 0.7$

To show the good performance and low sensitivity of the proposed circuit in circuit parameters variation, the value of inductor is increased by 50%. The output dc voltage and the harmonics of the input current are shown in figure 8. As can be seen the output dc voltage is also proportional to control parameter k. The phase difference between input voltage and current is 4.3° and the power factor is 0.997 despite 50% increasing in inductance.



(a)



(b)

Figure 8. 50% increase in inductance (a) output dc voltage (b) harmonic spectrum of input current

Figure 9 shows the output dc voltage with $k = 0.7$ and 50% decreasing in capacitance C. In spite of this variation, the phase difference is 1°.

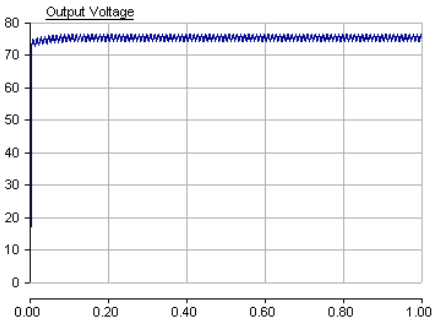
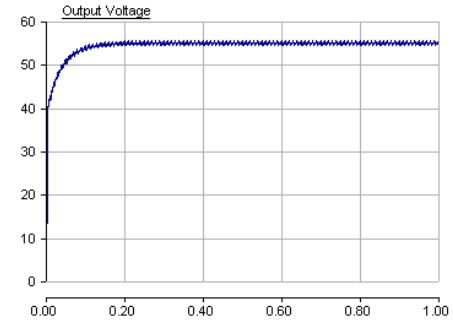


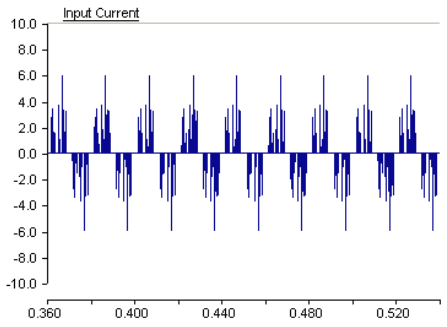
Figure 9. Output dc voltage with 50% decrease in capacitance

Figure 10 represents the circuit waveforms for $k = 0.5$, where the output dc voltage is proportional to control variable k, and the phase difference between input voltage and current is -2.8° that leads to 0.999 power factor in input.

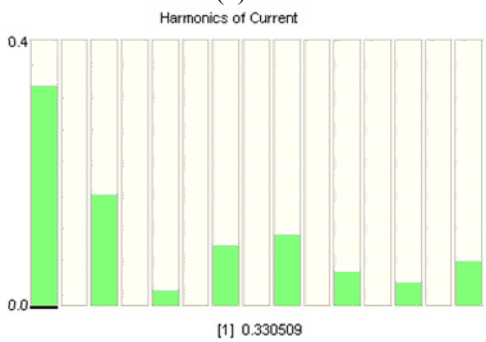
Figure 11 shows the output dc voltage when we change the desired voltage from 50 to 70 by controlling … at …. The quick response of the circuit, less than 5 cycles, is remarkable.



(a)



(b)



(c)

Figure 10. Circuit waveforms for $k = 0.5$ (a) output dc voltage (b) input current (c) Harmonic spectrum of input current

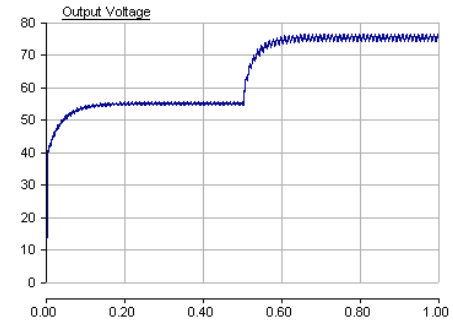


Figure 11. Gain variation from $k = 0.5$ to $k = 0.7$ at $t = 0.5\,\text{sec}$

IV. CIRCUIT PARAMETERS OPTIMIZATION

The represented parameters in table 1 are the usual values near the experimental values. To find the best values and to reach the less input current THD, some of these parameters of the circuit can be optimized. In order to minimize input current THD, golden section optimization method is used to find the best values of L, C and R. These parameters are shown in table 2.

Table 2. Optimized circuit parameters

L	0.9 mH
C	2.3 mF
R	91.64 Ω

By adjusting the parameters to that are shown in table 2 the performance of the circuit is improved. For example in $k = 0.8$ and with the parameters of the table 1, the phase difference between input voltage and current is 2.01° but with the parameters of table 2 the phase difference decrease to 1.75°.

V. OPERATION IN SEVERAL GAINS

In this section we examine the operation of the proposed circuit in several gains. For this examination we change the gain factor between $k = 0.40$ and $k = 0.98$, then the specifications of the circuit for each gain are recorded. The statistical results of these simulations are represented in table 3 to show the range of variation for each of these specifications.

Table 3. Statistical results from multiple simulations

	Minimum	Maximum	Mean
Gain factor	0.40	0.98	0.69
Phase difference between input voltage and current	0.80	6.28	2.67
Input power factor	0.994	0.999	0.998
Output dc voltage regulation in percent	0.00098	9.86	5.97
Input current THD	33.36	77.44	45.33
Dc component of input current per its rms value	0	0.27	0.08

The represented statistics in table 3 verify the good performance of the proposed circuit in power factor correction and output dc voltage regulation. Another advantage of this circuit is the low value of the dc component of the input current. This advantage can improve the experimental behaviour of the circuit.

Because of the configuration of the buck circuit with a switch in input current path, flowing of the discontinuous input current is inevitable. In spite of this fact, the multiple simulation results show that near 70% of simulations have a THD less than 45%.

VI. CONCLUSION

A new sensorless single phase single switch buck ac-dc converter with PFC operation is presented. Unlike conventional PFC converters, no dc voltage and ac current sensors are used.

Simulations are carried out using PSCAD/EMTDC. Because of presence of a switch in input current path, flowing of a sinusoidal current is impossible. But the amplitudes of the input current harmonics are negligible till fifth harmonic. The circuit parameters are optimized to obtain the least input current THD.

The proposed circuit has a low sensitivity against the variation of the circuit parameters. The output dc voltage can be controlled with a less than 10% error. The smallness of the dc component of the input current is another advantage.

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APPENDIX

The proof of the lemma:

$$y_{rms} = [\frac{1}{2\pi}\int_0^{2\pi} y^2(x)dx]^{1/2}$$

$$= [\frac{1}{2\pi}(\int_0^{DT} y^2(x)dx + \int_T^{T+DT} y^2(x)dx + ...)]^{1/2}$$

$$= [\frac{1}{2\pi}[A^2(\frac{DT}{2} + \frac{\sin 2\omega 0 - \sin 2\omega DT}{4\omega} + \frac{DT}{2} + \frac{\sin 2\omega T - \sin 2\omega(T+DT)}{4\omega} + ...)]]^{1/2}$$

If $\frac{2\pi}{\omega} = KT$ then:

$$y_{rms} = [\frac{A^2}{2\pi}[\frac{KDT}{2} + \frac{\sin \omega DT}{2\omega} \times (\cos DT\omega \sum_{i=1}^{K} \cos\frac{4\pi}{K}(i-1) - \sin DT\omega \sum_{i=1}^{K} \sin\frac{4\pi}{K}(i-1))]]^{1/2}$$

According to the switching frequency condition in the *lemma*:

$$T = \frac{1}{4lf} \quad \Rightarrow \quad y_{rms} = \sqrt{\frac{A^2 KDT}{4\pi}}$$

$$KT = \frac{2\pi}{\omega} \quad \Rightarrow \quad y_{rms} = A\sqrt{\frac{D}{2\omega}}$$