Operation principles of a switched capacitor snubber circuit suggested for half-bridge DC–DC converters

M.T. Aydemir* and F. Evran

Department of Electrical and Electronics Engineering, Gazi University, Ankara, Turkey

(Received 19 October 2006; final version received 7 October 2008)

An auxiliary circuit to accomplish soft switching in half-bridge converters is presented. The complete circuit is analysed, the design process, simulation and experimental results are given. Results show that the topology can be used even at high power levels. There is no control complexity involved. Efficiency is better than the classical topology. Cost increase because of the auxiliary devices is very low since the devices conduct the current very briefly allowing the selection of low rated metal oxide semiconductor field effect transistors (MOSFETs).

Keywords: half-bridge; DC–DC converter; soft-switching; switching losses; switched capacitor

1. Introduction

DC–DC converters are commonly utilised in various devices. The type of the application determines the power requirement for the converter and, depending on the power level, topology selection is performed. In some applications power isolation is also desired.

Half-bridge (HB) converters are preferred for power levels up to a few kilowatts because of their simplicity and ease of control. Typically, hard-switching and pulse width modulation techniques are employed with HB converters. Hard switching causes high switching losses and electromagnetic interference (EMI) problems. There are various soft switching techniques suggested for buck, boost, buck-boost and full bridge DC–DC converters. However, the situation is not so for HB converters. Suggested soft switching topologies for the HB converter, which is the simplest isolated converter topology, can be classified under three headings: (a) Resonance converters, (b) Asymmetrical HB converters and (c) Auxiliary switch converters.

HB resonance converters can be divided into three groups as series, parallel and series-parallel converters. Steigerwald (1988) analyses and compares different resonance circuits for HB converters in great detail. An inductor in series with the primary winding of the transformer establishes a resonance circuit with bus dividing capacitors in the series resonance HB converter. In the parallel resonance HB converter, the series inductor makes the resonance circuit with the resonance capacitor across the primary winding. Both the bus dividing capacitors and primary winding parallel capacitor exist in the series-parallel HB converter to resonate with the series inductor. Power control is achieved by varying the switching frequency around the resonance frequency. As a result of zero-current or zero-voltage switching, switching losses theoretically become zero in resonance HB

*Corresponding author. Email: aydemirmt@gazi.edu.tr

ISSN 0020-7217 print/ISSN 1362-3060 online
© 2009 Taylor & Francis
DOI: 10.1080/00207210802447686
http://www.informaworld.com
converters. However, varying switching frequency complicates the control and magnetic design. Also, switch stresses are increased. Despite these negative sides, resonant converters are still used in certain applications like induction cooking (Sazak and Sami 2004).

The topology suggested by Kim, Huh and Cho (1991) utilises a center-tapped high frequency transformer. A capacitor connected in parallel with the secondary winding establishes resonance circuits again with the leakage inductance and parasitic capacitance of the transformer. This results in zero-voltage switching in the primary switches, and reduced reverse recovery losses in the secondary side rectifying diodes. Therefore it is possible to use this topology at high frequencies. However, it is limited to low power levels.

There are some topologies that are designed to take advantage of the leakage or magnetising inductance of the transformer (Farrington, Jovanovic and Lee 1993). In this case, it becomes unnecessary to add an external inductor. This concept is also applied to other types of converters. However, these converters require saturated inductors if a constant switching frequency is desired, and this increases the complexity of the converter.

Another soft switching circuit that uses the assistance of the transformer is suggested by Xiaoming and Barbi (2000). The topology is suggested to overcome the difficulties (control complexity, protection of the auxiliary switches, voltage variation at the center point) associated with the auxiliary resonance commutation pole inverter. It was shown that no additional protection for the auxiliary switches is needed and zero-voltage switching is achieved if the transformer turns ratio is below 0.5. However, this topology is also complex.

HB converters have been considered for fuel cell applications by Aude, Dang Bang Viet, Yves, Paul and Jean (2006). In this study, a ZVS mode HB converter is compared for a 2.5 kW application. Results show that in spite of duty cycle limitations, the ZCS converter is superior to the ZVS one because of its simplicity and higher efficiency. The ZVS version described in the study has active switches both at the primary and the secondary sides of the circuit. The ZCS type, on the other hand, has only two active switches, both on the primary. However, the weak point of the ZCS type is the over voltage that occurs during turn-off instants.

Asymmetrical HB converters have been suggested to achieve zero-voltage switching (Imbertson and Mohan 1990, 1993, 1996, 1997; Yi-Hsin and Chern-Lin 2002; Mao, Qahouq, Luo and Batarseh 2004). The topology was first suggested by Imbertson and Mohan (1990). The physical structure is no different than that of an HB converter. The difference is in the control logic. Duty cycles of the switches are not equal, but varied so that the total will be 100%. This ideally presents zero-loss switching. However, the input–output relationship becomes non-linear, making it difficult to control it. This circuit can not be used for large input voltage variation range (Mao et al. 2004). Mao et al. (2004) suggest the use of current doublers on the secondary and an auxiliary switch in the primary to overcome this problem. However, this circuit is designed for low voltages and its performance at high output voltage levels is not known.

Aydemir, Bendre and Venkataramanan (2002) compare the hard switching and soft switching high power converters. They point out a simple HB converter, which is a modification of previous converter suggested by Barbi, Oliveira and Vieira (1989). In the original form, soft switching in an HB converter is achieved by using an auxiliary circuit between the mid-points of the bus capacitors and the mid-point of the switch leg. This topology has been overlooked in the literature. This study presents the modified converter. Simulation and experimental results show that the converter has better performance than the classical HB converter, and this is achieved with only a slight increase in cost because
the auxiliary switches are lower rated than the main devices, as they carry the load current for only short times. Since a preliminary version of this circuit is given by Aydemir and Evran (2008), some aspects are shortened here.

2. Circuit description
A soft switching HB (SSHb) converter diagram is shown in Figure 1. The snubber function is achieved through two capacitors between the DC bus. Two auxiliary switches with reverse blocking capacity are placed between the center point of the main devices and the capacitor leg. Main switches are turned on and off as they are in the classical HB converter. The auxiliary switches, on the other hand, are turned on slightly after the main switches and turned-off at a specific time after the turn-off of main switches. As they are turned on and off together at each switching of the main devices, their switching frequency is twice as high as that of the main switches. However, as will be shown later, the current handling is low and therefore, low rated switches can be used for this purpose. An external inductor is usually needed in series with the transformer primary to allow the current flow during the switching transient, as it is the case typically in soft switching converters.

Figure 2 shows all the waveforms related to the operation of the HB-switched-snubber converter. Circuit diagrams for each operation mode are shown in Figure 3.

3. Modal analysis
There are seven distinct modes of operation at each half cycle. Table 1 shows all the equations that are valid for the modes of operation. The last row gives the condition that defines the end of each mode. The variables used in the equations are defined in Table 2.

It is also necessary to define some important voltage and current values that are attained during the operation. At the end of the current fall period of Mode 2, switch voltage has the following level:

$$v_{cel}(t_f) = v_{cs1}(t_f) = V_f = \frac{I_{max}}{2C_S}$$

Figure 1. Circuit diagram of switched snubber HB.
Figure 2. Operational waveforms for switched snubber HB.

<table>
<thead>
<tr>
<th>Mode</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
<th>7</th>
<th>1'</th>
<th>2'</th>
<th>3'</th>
<th>4'</th>
<th>5'</th>
<th>6'</th>
</tr>
</thead>
<tbody>
<tr>
<td>Q1</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Q1s</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Q2s</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Q2</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>IL</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>I_Q1</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Vce1</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Vcs1</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Vp</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Vb</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>
To guarantee soft switching, it is a safe practice to choose a snubber capacitor value that will result in a $V_f$ that is lower than $V_s/2$. This condition yields a minimum capacitor value:

$$C_{\text{min}} = \frac{I_{\text{max}} \cdot I_f}{V_s}$$

Mode 4 is a resonance period. A resonance circuit is formed by $L_p$, $C_{s1}$ and $C_{s2}$. $Z_0$, $\omega_0$ and $C_r$ are the characteristic impedance, resonant frequency and resonant capacitor of the circuit, respectively.

Figure 3. Circuit diagrams for each mode of operation.
Table 1. Equations defining the modes.

<table>
<thead>
<tr>
<th>Mode</th>
<th>Equations</th>
<th>Explanation</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>$v_{ce1} = v_{cs1} = 0; v_{cs2} = v_{ce2} = V_s$</td>
<td>Ends when the upper switch is turned-off at an instant that is determined by the duty cycle controller.</td>
</tr>
<tr>
<td>2</td>
<td>$i_{sw}(t) = I_p^{\text{max}}(1 - \frac{t}{t_f})$&lt;br&gt;$v_{ce1} = v_{cs1} = \frac{I_p^{\text{max}}}{2C_s}t^2$</td>
<td>Ends at $t = t_f$</td>
</tr>
<tr>
<td>3</td>
<td>$i_{cs1} = \frac{I_p^{\text{max}}}{2}$&lt;br&gt;$v_{ce1}(t) = v_{cs1}(t) = V_f$&lt;br&gt;$v_{ce2}(t) = 0$</td>
<td>Ends when the switch voltages become equal to half the rail voltage.</td>
</tr>
<tr>
<td>4</td>
<td>$i_p(t) = I_p^{\text{max}} \cos(\omega_0 t)$&lt;br&gt;$v_{ce1}(t) = v_{cs1}(t) = \frac{V_s}{2} + Z_0 I_p^{\text{max}} \sin(\omega_0 t)$</td>
<td>Ends when the upper switch voltage reaches to the rail voltage.</td>
</tr>
<tr>
<td>5</td>
<td>$i_p(t) = I_p^{\text{max}} - \frac{V_f}{2C_s}t$</td>
<td>Ends when the primary current becomes zero.</td>
</tr>
<tr>
<td>6</td>
<td>$i_{cs1} = \frac{I_p^{\text{max}}}{2}$; $v_{ce1} = V_s$; $v_{cs2} = 0$</td>
<td>Ends when the next switch is turned on.</td>
</tr>
<tr>
<td>7</td>
<td>$i_p(t) = \frac{V_f}{2C_s}t$</td>
<td>Ends when the primary current catches up with the reflected load current.</td>
</tr>
</tbody>
</table>

$$Z_0 = \sqrt{\frac{L_p}{C_r}}; \quad \omega_0 = \frac{1}{\sqrt{L_p C_r}}; \quad C_r = 2C_s$$ (3)

When this mode ends, the upper switch voltage reaches $V_s$, the lower switch voltage reaches zero, and the parallel diode of the lower switch clamps on. The primary current at this instant is

$$i_p(t_f') = I_p^{\text{max}} \sqrt{1 - \left(\frac{V_s}{2Z_0 I_p^{\text{max}}}\right)^2}$$ (4)

In order for the upper switch voltage to reach the rail voltage before the primary current becomes zero, a certain condition needs to be met: This condition can be stated as

$$Z_0 I_p^{\text{max}} > \frac{V_s}{2}$$ (5)

This, in turn, requires that

$$C_s < C_s^{\text{max}} = 2\left(\frac{I_p^{\text{max}}}{V_s}\right)^2L_p$$ (6)

This condition can also be stated as

$$L_p > \left(\frac{V_s}{I_p^{\text{max}}}\right)^2 \frac{C_s}{2}$$ (7)

This equation shows why a larger primary inductance is required in this converter.
The length of Mode 5 is $t_{lf}$, which is the time the primary current becomes zero.

$$t_{lf} = \frac{2L_p I_{p1}}{V_s}$$

The auxiliary switches can be turned off any time in this mode. The required minimum delay value of the auxiliary gate is

$$t_{d\text{min}} = t_f + t'_f + t''_f$$

Mode 7 ends at $t = t_{lr}$ at which point the primary current reaches the reflected value of the load current.

$$t_{lr} = \frac{2L_p I_{p\text{min}}}{V_s}$$

where $I_{p\text{min}}$ is the reflected value of the minimum secondary current.

The auxiliary switches can be turned on any time, after the main switch becomes completely on. Let us assume that the auxiliary switches are turned-on after the main switch voltage becomes zero, but before the current reaches $I_{p\text{min}}$. Before the turn-on of the auxiliary devices, snubber capacitors have the same voltage across them as their corresponding switches do. Therefore, ideally there is no power loss when the auxiliary switches are turned-on.

4. Design parameters and limitations
A prototype has been designed to show the operation principles of the converter. The following operation parameters have been used.

- Rated load: 10 A, 100 V
- Input voltage: 2 × 155 V ($V_s = 310$ V)
- Switching frequency: 20 kHz
- Transformer ratio: 1:1
- Main devices: IXGH32N60 IGBT
- Auxiliary devices: IRFP 460 MOSFET

The snubber capacitors used in the converter help to reduce the turn-off losses of the switches. However, as defined by Equation (6), there is a condition to be met in order for
the switch voltage to reach the rail voltage during turn-off. Therefore, one needs to be careful while choosing the primary inductance and snubber capacitance values.

Figure 4 shows how the snubber capacitor value varies as a function of the load current, with the primary inductance used as a parameter. The minimum value of the capacitor is calculated by using the fall time of the insulated gate bipolar transistor (IGBT) switch (in this case, 120 ns) and is shown with dashed lines in the figure.

Large capacitor values yield less turn-off voltages, while requiring large inductor values as well. If a soft switching region of 5–10 A is aimed, 10 nF and 8 μH can be chosen for the snubber capacitor and external inductor values in this application. On the other hand, the snubber capacitors and external inductor reduce the effective power transfer cycles.

Average value of the output voltage can be calculated by integrating the rectified secondary voltage. As seen in the last drawing of Figure 2, rectified voltage waveform pertains to non-zero values during the on time and fall time of the main switches, and while the snubber voltage changes between $V_f$ and $V_s/2$. The Voltage equation can be written as follows:

$$v_b = \begin{cases} \frac{V_s}{n} & t \leq dT \\ V_s - \frac{I_{p_{\text{max}}}}{2nC_s}t_f^2 & t \leq t_f \\ V_s - V_f \left(1 - \frac{t}{t_f'}\right) & t \leq t_f' \end{cases}$$

From (11), average value of the voltage can be calculated:

$$V_{\text{out}} = \frac{2}{n} \left(\frac{V_s}{2}dT + \frac{V_s}{2}dT_f - \frac{I_{p_{\text{max}}}}{6C_s}t_f^2 + \frac{1}{2} \left(\frac{V_s}{2} - V_f\right)t_f'\right)$$

Figure 4. Variation of snubber capacitance as a function of load current and total primary inductance.
5. Simulation results

Both the SSHB and hard-switched HB (HSHB) converters have been simulated by using the PSpice simulator. As the exact models of the devices used in the experimental work were not available, a comparison has been attempted by using an available device model (IXGH10N60) in both converters. Figure 6 shows the turn-off process of the SSHB converter. Because of the large external inductor, the primary current persists after the turn-off of the device and the snubber capacitor slowly charges to the upper rail voltage. Figure 7 shows the HSHB at the same conditions. As there is no snubber in the circuit, the device voltage jumps right at the beginning of the turn-off process.

Loss calculations can be made by using the data obtained from the simulations. However, these calculations are not meaningful if the device model is not very accurate. Yet, a comparison at same base can show if the method is effective. Table 3 shows the reduction percentages when SSHB is used instead of HSHB for various loads. Only the losses of the switches are included in the comparison. SSHB converter losses also include those of the auxiliary devices. The results show that the soft switching technique is more effective at higher currents, but still works even at very low load currents.

6. Experimental results

A prototype of the SSHB converter has been built and tested in the laboratory. Following is a list of the parameters used in the prototype.

- Input voltage: $2 \times 155$ V dc (Rectified from 220 V mains)
- Output voltage: 100 V
Transformer turns ratio: 1:1
External primary inductance: 8 mH
Snubber capacitors: 10 nF each
Main switches: IXGH32N60 IGBTs
Snubber switches: IRFP 460
Delay time between gates: 1 ms (fixed)
Switching frequency: 20 kHz.

Table 3. Converter losses for HSHB and SSHB.

<table>
<thead>
<tr>
<th>Load (A)</th>
<th>Loss reduction (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>27</td>
</tr>
<tr>
<td>8</td>
<td>21</td>
</tr>
<tr>
<td>6</td>
<td>13</td>
</tr>
<tr>
<td>4</td>
<td>11</td>
</tr>
<tr>
<td>2</td>
<td>14</td>
</tr>
</tbody>
</table>

Figure 6. Turn-off of main device in SSHB converter: Collector-emitter voltage and snubber capacitor voltage (upper plot); Collector current and primary current (lower plot).

Figure 7. Turn-off of main device in HSHB converter: Collector-emitter voltage (upper plot); Collector current (lower plot).
Figure 8 shows the general operation of the converter at full load (10 A). The persistence of the primary current after turn-off, and voltage variation of the switch are clearly seen in the figure. The snubber voltage clamps at the rail voltage when the current totally diminishes. The switch voltage on the other hand resonates as a result of the interaction of the parasitic capacitances and the leakage inductance. The soft turn-off of the device is shown in detail in Figure 9.

Employing a large external inductor in the primary also helps to reduce the turn-on losses of the switch. The turn-on of the device is shown in Figure 10. Figure 11 shows the current waveforms of the auxiliary switch along with the primary current. For most of the time, no current flows through these devices. Therefore, their current ratings need not be too high.

![Figure 8](image1.png)

Figure 8. Device voltage (100 V/div), Snubber voltage (100 V/div), collector current (5 A/div) and primary current (2 V/div, 4 A/V, output of a current transformer) during turn-off (SSHB; 10 A load current).

![Figure 9](image2.png)

Figure 9. Turn-off period in SSHB, 10 A load current.
The converter has also been run in hard-switched mode for comparison purposes. Turn-on and -off waveforms of the main switches in the hard-switched operation with the rated load are shown in Figures 12 and 13.

Accurate measurement of power losses of the switches has not been possible because of probe sensitivity levels. Therefore, input and output power losses have been measured and efficiencies of the topologies have been calculated. Input power was measured by using a Fluke® 43B power quality analyser at the AC input terminal. The output power was measured across the adjustable resistive load. Efficiencies of the converters are shown in Table 4. For all current levels the soft switched topology is superior to the hard switching converter. These figures include all the losses of the converters (switch, transformer, input- and output-rectifier).

Figure 10. Turn-on period in SSHB, 10 A load current.

Figure 11. Auxiliary device current (upper trace; 5 A/div) and primary current.
7. Conclusion

The auxiliary switched capacitor snubber SSHB converter has been shown to be effective in reducing the losses and increasing the efficiency. The circuit has two low rated auxiliary switches, and adds almost no complexity to typical HB converters.

Table 4. Percent efficiencies of the two topologies for different load currents.

<table>
<thead>
<tr>
<th>Load current (A)</th>
<th>SSHB (%)</th>
<th>HSHB (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>85</td>
<td>84</td>
</tr>
<tr>
<td>4</td>
<td>82</td>
<td>74</td>
</tr>
<tr>
<td>6</td>
<td>80</td>
<td>75</td>
</tr>
<tr>
<td>8</td>
<td>79</td>
<td>77</td>
</tr>
<tr>
<td>10</td>
<td>78</td>
<td>76</td>
</tr>
</tbody>
</table>
References


