Front-end power supply with power factor correction

Conventional PWM forward converter

- $V_g = 270-380V$
- $V_{ds}(t)$
- $max \ V_{ds} = 2V_g + \text{ringing, } = 760V + \text{ringing}$
- $P \approx 200W$
- $D \leq \frac{1}{2}$
- On-state transistor current $= \frac{P}{V_g D}$
- $I$
Magnetizing current must operate in DCM!

- Peak transistor voltage occurs while transformer is being reset.
- We could reset the transformer with less peak voltage if we used the discontinuous interval.
The active clamp forward converter

Q1 and Q2 are driven as in the half-bridge circuit:

Q1 gate drive

\[ \text{on} \quad \text{off} \]

\[ DT_3 \quad T_3 \]

Q2 gate drive

\[ \text{off} \quad \text{on} \]

\[ D' T_3 \quad T \]

- Better utilization of transformer and transistors
- Zero-voltage switching, related to QSW and ZVT (somewhat)
- Not limited to \( D \frac{T}{2} \)
Approximate analysis: hard-switched case, ignore resonant transitions and dead time

Transformer primary voltage, magnetizing current:

\[ v_{pri}(t) \rightarrow V_s \]

\[ \rightarrow D T_s \rightarrow \]

\[ \leftarrow D T_s \rightarrow \]

\[ -V_b \]

\[ t \]

\[ i_M(t) \uparrow \]

\[ \frac{V_s}{L_M} + \Delta i_m \]

\[ \frac{-V_b}{L_M} \]

\[ 0 \]

\[ t \]

Voh-sec balance: \( D V_s - D' V_b = 0 \)

\[ V_b = \frac{D}{D'} V_b \]

Charge balance on \( C_b \):

\( \langle i_b(t) \rangle \uparrow \)

\[ \text{equal areas} \]

\[ -\Delta i_M \text{ must } = -(\text{peak}) \]

so that \( \langle i_b \rangle = 0 \)

\( \langle i_M \rangle = 0 \) implies that \( \langle i_b \rangle = 0 \)
$V_b$ can be viewed as a flyback converter output, in which the flyback converter consists of $L_m$ (buck-boost inductor), $Q_1$ (transistor), and $Q_2/D_2$ (output diode, bidirectional switch). By use of a current bidirectional switch, there is no discontinuous conduction mode, and $L_m$ operates in CCM.

The peak transistor voltage is

$$\text{max } V_{ds} = V_g + V_b = V_g \left(1 + \frac{D}{D'}\right) = \frac{V_g}{D'}$$

which is less than the conventional value of $2V_g$ when $D > \frac{1}{2}$. To appreciate the benefit of this, let's consider a design example:

270V $\leq V_g \leq$ 350V

200W load

Conventional case: peak $V_{ds} = 2V_g + \text{ringing}$

= 700V + \text{ringing}
Let \( \max D = 0.5 \) \((\text{at } V_g = 270 \text{ V})\)

then \( \min D \) \((\text{at } V_g = 350 \text{ V})\) is 
\[
\frac{(0.5)(270)}{(350)} = 0.3857
\]

the on-state transistor current, neglecting ripple, is given by

\[
<i_g> = D \times I = D \times i_{d-on}
\]

with \( P = 200 \text{ W} = V_g <i_g> = D \times V_g \times i_{d-on} \)

so \( i_{d-on} = \frac{P}{D \times V_g} = \frac{200 \text{ W}}{(0.5)(270 \text{ V})} \approx 1.5 \text{ A} \)

active clamp case:

\(1\) suppose we choose the same turns ratio. Then \( D \) is the same: 0.3857 < 0.5

and \( i_{d-on} \) is the same. But the peak \( V_{ds} \) is \( \frac{V_g}{D} \)

<table>
<thead>
<tr>
<th>( V_g )</th>
<th>( D )</th>
<th>peak ( V_{ds} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>270</td>
<td>0.5</td>
<td>540V</td>
</tr>
<tr>
<td>350</td>
<td>0.3857</td>
<td>570V</td>
</tr>
</tbody>
</table>

which is considerably lower than 700V.
(2) Suppose we operate at a higher duty cycle, say \( D = 0.5 \) at \( V_g = 350\,V \).

Then peak \( V_{ds} = \frac{350}{1-\frac{1}{2}} = 700\,V \) (same as conventional).

but we can use a lower turns ratio that leads to lower reflected current in \( Q_1 \):

\[
I_{d-on} = \frac{P}{D \cdot V_g} = \frac{200\,W}{(\frac{1}{2}) \cdot (350\,V)} = 1.15\,A
\]