UCC2897A Peak Current Mode Active-Clamp Forward Converter Small-Signal Modeling Design Consideration

ABSTRACT

UCC2897A is a peak current mode active-clamp controller. This paper discussed the modeling process and loop compensation for UCC2897A. An example has been implemented with the modeling and compensation.

Contents

1 UCC2897A Introduction: .............................................................................................................. 1
2 Peak Current Mode Small-Signal Circuitry ................................................................................. 2
3 Active Clamp Peak Current Modeling Analysis: ......................................................................... 4
4 A Design Example: ....................................................................................................................... 7
5 Conclusion: ................................................................................................................................. 10

FIGURES

Figure 1. UCC2897A Internal Block Diagram. ................................................................................ 2
Figure 2. Control Block for Peak Current Mode. ............................................................................. 3
Figure 3. Gain from Control to Inductor Current for Peak Current Mode .................................. 4
Figure 4. Peak Current Mode Active Clamp Forward Converter Topology. ................................. 4
Figure 5. The Peak current active clamp control block diagram. .................................................. 6
Figure 6. Small Signal Implementation from Control to Output. ..................................................... 7
Figure 7. 100-W Isolated Power Module with UCC2897A Device .................................................. 7
Figure 8. Overall Small Signal Circuitry Implementation ................................................................. 8
Figure 9. Simulation Results. .......................................................................................................... 9
Figure 10. Lab Test Results (V_{in} = 72, I_{o} = 30 A) ...................................................................... 10

1 UCC2897A Introduction:

The UCC2897A PWM controller simplifies implementation of the various active clamp/reset and synchronous rectifier switching power topologies. The UCC2897A is peak current-mode, fixed frequency, high performance pulse width modulator. It includes the logic and the drive capability for the P-channel auxiliary switch along with a simple method of programming the critical delays for proper active clamp operation, as showed in Figure.1.
2 Peak Current Mode Small-Signal Circuitry.

A peak current-mode converter, with continuous current mode, introduces a sampling-hold function, as shown in Equations 1 and 2. Figure 2 shows the control block for peak current-mode:

\[
H_e(s) = \frac{1 - e^{-s t_{SW}}}{s \times t_{SW}} \approx \frac{1}{s} \left( 1 + \frac{s}{2 t_{SW}} + \frac{\pi^2}{t_{SW}^2} \right); 
\]

(1)

\[
H_{vi}(S) = \frac{I_L(S)}{V_i(S)} \approx \frac{D}{S L_m} 
\]

(2)
Then, the gain from the error amplifier output to the inductor current can be got:

\[
\frac{\hat{i}_{L}}{\hat{e}} = H(S)
\]  \(3\)

\[
H(S) \approx \frac{1}{R_i} \frac{1}{1 + S\left[\frac{T_s L_m (S_n + S_f)}{V_i R_i} - \frac{T_s}{2}\right] + S^2 \frac{T_s^2}{\pi^2}}
\]  \(4\)

Define two components as below:

Resistor: \(R_a = \frac{2L_m}{t_{sw} \left[\frac{2(S_n + S_f)}{(S_n + S_f) - 1}\right]}\); and Capacitor: \(C_a = \frac{t_{sw}^2}{\pi^2 L_m}\)  \(5\)

Where:

- \(S_n\) is the rising rate.
- \(S_f\) is the falling rate.
- \(S_e\) is the slope compensation rate.

Hereby, the equivalent small signal model from control to inductor current can be obtained as shown in Figure 3.
3 Active Clamp Peak Current Modeling Analysis:

The control to current gain function can be described in Equation 6:

\[
\frac{\hat{u}_c}{R_s} \frac{1}{1 + \frac{S}{Q\omega_n} + \frac{S^2}{\omega_n^2}} = i_m^* + \frac{i_i}{n}
\]

(6)

With the active-clamping topology,
\[ S_n = \left( \frac{V_i}{L_m} + \frac{V_i - nV_o}{n^2 L_s} \right) R_s \]
\[ S_f = \left( \frac{DV_i}{(1-D)L_m} + \frac{V_o}{nL_s} \right) R_s \]

(7)

\( n \) is the transformer turn ratio between primary and secondary side windings.

Therefore, \( Q = \frac{1}{\pi \left( \frac{S_n + S_c}{S_n + S_f} - 0.5 \right)} \); \( \omega_n = \frac{\pi}{T_{sw}} \)

(8)

The clamping circuitry modeling:

\[ <i_{cm}> = (1-<d>)(1-<i_m>); <i_{cm}^\wedge> = I_{cm} + i_{cm}^\wedge ; <d> = D + d; <i_m> = I_m + i_m^\wedge \]

(9)

Where:

- \(<> \) means the cycle average function.
- \(^\wedge\) means the perturbation element.
- \( d \) is the duty cycle of the main switch;

Considering the magnetic operating in symmetrically, the average current \( I_c \) and \( I_m \) are zero.

\[ i_{cm}^\wedge = (1-D)i_m^\wedge \]

(10)

The Vds of MOSFET modeling:

\[ u_{ds}^\wedge = -\frac{D}{(1-D)} V_m^\wedge d + (1-D)u_{cm}^\wedge \]

(11)

And:

\[ u_{ds}^\wedge = -SL_m i_m^\wedge ; u_{cm}^\wedge = -\frac{1}{SC_m} i_{cm}^\wedge \]

(12)

\[ V_m^\wedge d = \left[ S \frac{(1-D)L_m}{D} + \frac{(1-D)^3}{SC_m} \right] i_m^\wedge \]

As result,

\[ L_c = \frac{(1-D)}{D} L_m; C_c = \frac{D}{(1-D)^3} C_m \]

Define:

(14)
Then: 

\[
\frac{v_m^\wedge}{n} \hat{d} = \frac{(SL_c + \frac{1}{SC_c}) \hat{i}_m}{n}; \quad (15)
\]

And: 

\[
\hat{i}_l = \frac{1}{(SL_s + \frac{R_L}{1 + SC_c R_L})} \frac{v_m^\wedge}{n} \hat{d} \quad (16)
\]

\[\text{Figure 5. The Peak current active clamp control block diagram.}\]

\[
G_{co}(S) = \frac{\hat{u}_o}{\hat{u}_c} = \frac{R_L (1 + S^2 L_c C_c)}{R_s (1 + \frac{S}{Q \omega_n} + \frac{S^2}{\omega_n^2}) [S^2 (L_c + n^2 L_s) C_o C_r R_L + S^2 (L_c + n^2 L_s) C_c + S(C_o + n^2 C_c) R_L + 1]}
\]

With \(C_o >> n^2 C_c\), this formula can be simplified as in Equation 18:

\[
G_{co}(S) = \frac{\hat{u}_o}{\hat{u}_c} \approx \frac{R_L (1 + S^2 L_c C_c)}{R_s (1 + \frac{S}{Q \omega_n} + \frac{S^2}{\omega_n^2}) [S^2 (L_c + n^2 L_s) C_c + 1] (SC_c R_L + 1)} \quad (18)
\]

The above Equation (18) introduced dual zeros and dual poles.

\[
f_c = \frac{1}{2 \pi \sqrt{L_c C_c}} = \frac{(1 - D)}{2 \pi \sqrt{L_m C_m}}; \quad f_p = \frac{1}{2 \pi \sqrt{(L_c + n^2 L_s) C_c}} = \frac{(1 - D)}{2 \pi \sqrt{(L_m + \frac{D}{1 - D} n^2 L_s) C_m}} \quad (19)
\]

Generally, to avoid the instability, the closed-loop crossover frequency “fc”, should be far less than half of the pole frequency “fp”. Here, “fp” should be the value with the maximum limited duty cycle “D”.

6
Figure 6 shows a small-signal circuitry implementation:

![Small Signal Implementation from Control to Output.]

**4 A Design Example:**
The design schematic (see Figure 7) and electric specification follow.

![100-W Isolated Power Module with UCC2897A Device]

T1: Current transformer specification is: turn ratio is 1:100;
T2: Transformer specification is: PA0810NL; turn ratio is 6:1; Lm is 345 µH.

L1 inductor specification is: PA0373; 2.1 µH;

\[V_{in} = 72V; V_{out} = 3.3V; I_{out} = 30A\]

Based on the schematic design in Figure.7, and the small signal analysis, we can get the overall simulation model as Figure.8:

**Figure 8. Overall Small Signal Circuitry Implementation**

Figure 9 shows the simulation results.
Figure 9. Simulation Results.

Figure 9 showed the cross-over frequency is 4.5 kHz, and phase margin is 111 degrees.

By lab test results, we can get the results as shown in Figure 10.

The test results showed cross-over frequency is 4.5 kHz and phase margin is 80 degrees.

Comparing the simulation and test results, they matched fully.
5 Conclusion:
The analysis shows the modeling and compensation is effective. The analysis revealed the zero and poles of the peak current mode control active-clamp forward converter, which is critical for the design of the active-clamp converter.

Reference:
1. Texas instruments, SLUS829D, UCC2897A datasheet.
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