Application Note for Reduction of No-load Power Consumption

Prepared by Hommy Ding

Oct 09, 2011

ABSTRACT

This note presents a method of reducing no-load power consumption for flyback converters. Under normal operation, power loss of a flyback converter includes conduction loss, switching loss and control circuit loss. In no-load condition, the current in the circuit is very small which makes the conduction loss almost negligible. Switching loss and control circuit loss are major sources of power loss, and must be minimized to reduce no-load power consumption.

INDEX

NO-LOAD POWER LOSS ANALYSIS

The analysis presented in this note is based on a flyback converter with an MPS current mode controller. For a flyback converter working under no-load condition, power losses can be divided into six parts: (1) Xcap discharge loss, (2) Input capacitor loss, (3) RCD snubber loss, (4) Loss of switching components, (5) Transformer power loss, (6) Loss of control circuit.

1. Xcap

An Xcap is a kind of safety capacitor connected between L and N. It acts as a filter on the differential mode interference of the power supply. If the capacitance exceeds 0.1μF, when input is disconnected, the circuit automatically discharges the capacitor to avoid any potential electrical shock, the discharging time must not exceed 1s for pluggable power supplies. The relevant time constant is the product of the effective capacitance and the discharging resistance in the circuit. However, because determining the effective capacitance and resistance values precisely is difficult and the Xcap is usually the dominant capacitance, then we can estimate time constant as a function of the Xcap and the discharging resistors. So that the discharging time constant τ of a RC network is then:

$$\tau = R_x \cdot C_x \tag{1}$$

Where C_x is the Xcap, R_x is the discharging resistance. As the time constant τ can not exceed 1s, the discharging resistors must be smaller than $1/C_x$. The discharging resistors continuously dissipate power throughout operation. The power dissipated by the discharging resistors $P_{discharge}$ can be calculated as:

$$P_{discharge} = \frac{{V_{AC}}^2}{R_x} \tag{2}$$

Where V_{AC} is the rms value of the AC input voltage.

The power loss through discharging resistor contributes significantly to the no-load power loss especially in the high-input condition. To decrease the no-load power consumption, increase the discharging resistance, through in some instances the Xcap must decrease in order to increase the discharging resistance, which may deteriorate EMI performance. As a compromise, choose the appropriate Xcap and discharging resistors according to each application.

2. Input Capacitor

Power loss of electrical capacitor induced by the leakage current I_R can not be ignored when the capacitor voltage is very high. To decrease the no-load power consumption, lower the leakage current of the input capacitor as much as possible. The leakage current I_R can be calculated as:

$$I_R = K \cdot C_{in} \cdot V_{in} \tag{3}$$

Where K is the coefficient of leakage current, C_{in} is the capacitance, and V_{in} is the DC input voltage.

We can obtain the loss induced by the input capacitor $P_{Capacitor}$ as:

$$P_{Capacitor} = K \cdot C_{in} \cdot {V_{in}}^2 \tag{4}$$

For a flyback converter, C_{in} is defined by the power of the converter, and V_{in} is defined by the rms value of the AC input voltage. Therefore, the coefficient K dominates the loss induced by the capacitor. It is correlative with material purity of capacitor and use condition. At typical temperature, K is 0.01 for a general specific product, and it is 0.0001 for a premium product. The loss induced by the capacitor can be several to tens of mW. Choose a capacitor with a low leakage current to minimize the standby power loss.

3. RCD Snubber

In operation, the energy stored in the leakage inductance can not transfer to the output side of a flyback converter. This energy may result in a high voltage spike across the MOSFET and the rectifier diode, which can cause severe EMI noise and device failure.

The RCD snubber shown in Figure 1 suppresses the voltage spike to protect the component.

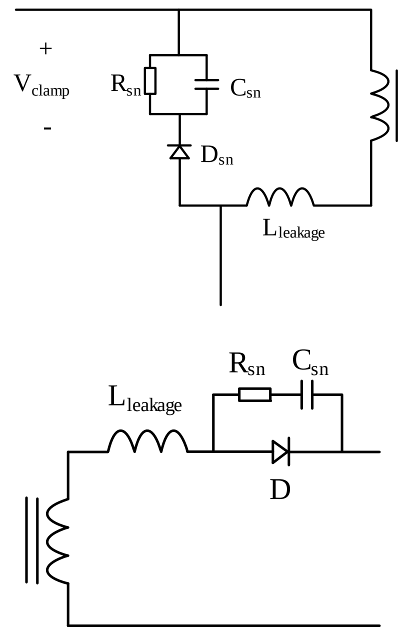

Figure 1: RCD Snubber on Primary and Secondary Side

The RCD snubber dissipates the energy of the leakage inductance and limits the voltage spikes. Accurate analysis of the RCD snubber power loss is affected by the leakage inductance, snubber diode and parasitic capacitance, but can be roughly estimated by assuming the energy stored in the leakage inductance is completely dissipated by the RCD snubber circuit in steady state.

The energy stored in the leakage inductance can be expressed as:

$$P_{leakage} = \frac{1}{2} \cdot L_{leakage} \cdot I_{pri\_peak}{}^{2} \cdot f \tag{5}$$

Where $L_{leakage}$ is the leakage inductance, I_{pri_peak} is the primary peak current, and f is the switching frequency.

When the converter works in no-load condition, the current sense voltage threshold V_{peak} can be obtained from the datasheet of the IC controller. So the primary peak current I_{pri_peak} is determined by the sense resistor R_{sense}. The peak current I_{pri_peak} under no-load condition is given as:

$$I_{pri\_peak} = \frac{V_{peak}}{R_{sense}} \tag{6}$$

Meanwhile, the converter enters burst mode when working in no-load condition. An equivalent switching frequency f_s can be used to substitute the switching frequency f and be calculated by (5).

$$f_s = \frac{N_{sw}}{N_{sw} \cdot t_{sw} + t_{burst}} \tag{7}$$

As shown in Figure 2, N_{sw} is the number of switchings in one burst time, t_{sw} is the switching period and t_{burst} is the burst time.

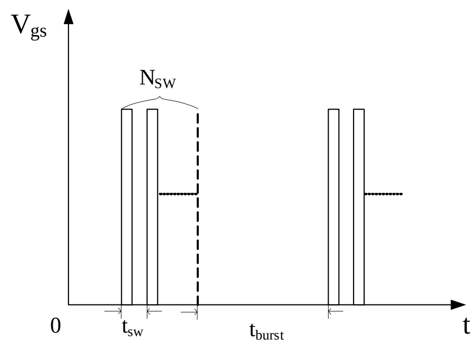

Figure 2: V_{gs} in Burst mode

By substituting (4) (5) into (3), we can obtain the energy stored in the leakage inductance as:

$$P_{leakage} = \frac{1}{2} \cdot L_{leakage} \cdot (\frac{V_{peak}}{R_{sense}})^2 \cdot \frac{N_{sw}}{N_{sw} \cdot t_{sw} + t_{burst}} \tag{8}$$

4. Switching Components

Generally, switching components include the MOSFET and diode in a flyback converter.

(1) MOSFET

Power loss of MOSFET can be divided into conduction loss, switching loss and gate driving loss. As mentioned above, the primary peak current in no-load condition can be calculated with the current sense voltage threshold V_{peak} to derive the conduction loss of MOSFET. Usually, the gate drive stage is integrated in the controller so the gate driving loss of MOSFET can be also included in the loss of the control circuit.

Figure 3 shows the flyback transformer magnetizing current at no-load condition; generally it is in DCM mode.

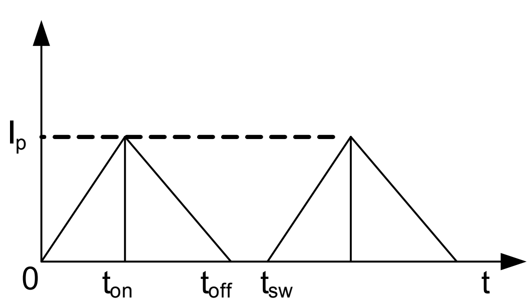

Figure 3: Magnetizing Current in Transformer

The primary side MOSFET on time t_{on} and secondary side diode on time t_{off} can be calculated as:

$$t_{on} = \frac{L_m \cdot I_{pri\_peak}}{V_{in}} \quad (9)$$

$$t_{off} = \frac{L_m \cdot I_{pri\_peak}}{N \cdot (V_{out} + V_F)} \quad (10)$$

Where L_m is the transformer magnetizing inductance, N is the transformer turn ratio (primary side to secondary side), V_{out} is the output voltage, and V_F is the forward voltage drop of secondary diode.

The primary side current can be calculated as:

$$I_{pri}(t) = \frac{V_{in}}{L_m} \cdot t, 0 \le t \le t_{on} \quad (11)$$

From (7) (9), the primary side rms and average current can be obtained as:

$$I_{pri\_rms} = \sqrt{\frac{1}{t_{SW}} \int_0^{t_{on}} I_{pri}(t)^2 dt} \quad (12)$$

$$I_{pri\_avg} = \frac{1}{t_{SW}} \int_0^{t_{on}} I_{pri}(t) dt \quad (13)$$

The conduction loss of MOSFET $P_{MOSFET\_conduction}$ in no-load condition is given as:

$$P_{MOSFET\_conduction} = I_{pri\_rms}{}^2 \cdot R_{ds(on)} \quad (14)$$

Where $R_{ds(on)}$ is the on state resistance of MOSFET.

The converter works in DCM mode in no-load condition. That means that the switch turns on in zero-current condition. So the turn on power loss of MOSFET is dominated by the equivalent primary-side parasitic capacitance C_{oss} which includes the MOSFET junction capacitance, transformer parasitic capacitance, diode junction capacitance etc. In no load condition, the converter usually operates in deep DCM mode, and the MOSFET drain-source voltage is V_{ds}. The power loss during MOSFET turn-on is:

$$P_{MOSFET\_Coss} = \frac{1}{2} \cdot C_{oss} \cdot V_{ds}{}^2 \cdot f_s \quad (15)$$

Figure 4 shows the voltage and current waveforms of MOSFET during turn-off.

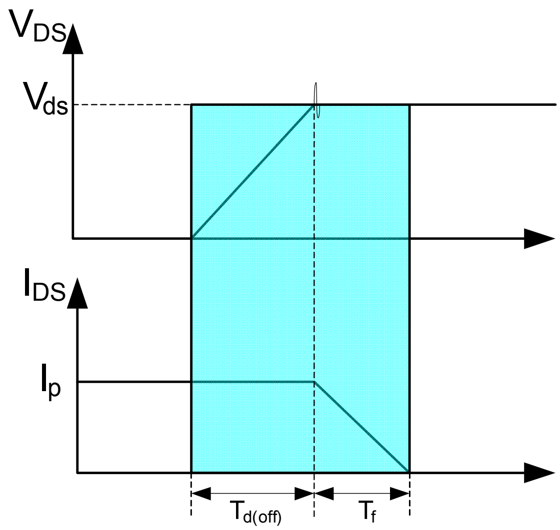

Figure 4: Voltage and Current Waveforms of MOSFET during Turning off

The switching off loss of MOSFET $P_{MOSFET\_switching}$ is:

$$P_{MOSFET\_switching} = \frac{1}{2} \cdot V_{ds} \cdot I_p \cdot (t_{d(off)} + t_f) \cdot f_s \quad (16)$$

We can find turn-off delay time $t_{d(off)}$ and fall time t_f in the datasheet of MOSFET with a given gate resistance. f_s is the equivalent switching frequency given in (7).

(2) Diode

For low-output-voltage application, use a Schottky diode to reduce the conduction loss and avoid potential diode reverse-recovery problem. The diode works in zero-current condition. The diode switch-off loss can be ignored and the snubber capacitance C_{diode} dominates the switch-on loss. So the power loss during diode turn-on is:

$$P_{diode\_swintch} = \frac{1}{2} \cdot C_{diode} \cdot \left(\frac{V_{in}}{N} + V_{out} \right)^2 \cdot f_s \quad (17)$$

The conduction loss of diode is induced by the forward voltage drop V_F of diode. V_F always changes with the forward current. For simplicity, use a constant value to calculate the power loss. In no-load condition, V_F is the voltage of diode under the secondary peak current I_{sec_peak}.

With the primary peak current I_{pri_peak}, the secondary peak current I_{sec_peak} can be calculated as:

$$I_{sec\_peak} = N \cdot I_{pri\_peak} \quad (18)$$

Figure 5 shows the diode current waveform.

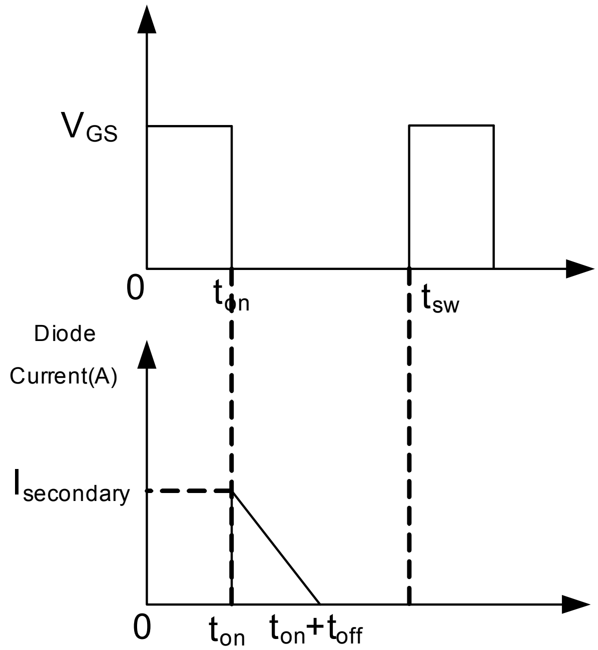

Figure 5: Current Waveform of Diode

The average value of secondary current can be written as follow:

$$I_{sec\_avg} = \frac{\frac{1}{2} \cdot I_{sec\_peak} \cdot t_{off}}{t_s} \tag{19}$$

The conduction loss of diode in no-load condition can be obtained as:

$$P_{Diode\_conduction} = V_F \cdot I_{sec\_avg} \tag{20}$$

5. Transformer

Transformer power loss can be divided into copper loss P_{copper} and core loss P_{core}. The copper loss is caused by the winding resistance. The resistance contains DC impedance and AC impedance. So the copper loss also contains DC loss and AC loss. The DC loss can be obtained as:

$$P_{copper\_DC} = I_{pri\_rms}^2 \cdot R_{pri\_winding} + I_{sec\_rms}^2 \cdot R_{sec\_winding} \tag{21}$$

Where $R_{pri\_winding}$ is the resistance of primary winding, $R_{sec\_winding}$ is the resistance of secondary winding, which are given as:

$$R_{pri\_winding} = \rho_{copper} \cdot \frac{N_{pri} \cdot L_{bobbin}}{N_{pri\_strand} \cdot S_{pri\_copper}} \tag{22}$$

$$R_{sec\_winding} = \rho_{copper} \cdot \frac{N_{sec} \cdot L_{bobbin}}{N_{sec\_section} \cdot S_{sec\_copper}} \tag{23}$$

Where

- ρ_{copper} is the conductivity of copper,
- N_{pri} is the number of primary turns,
- N_{sec} is the number of secondary turns,
- L_{bobbin} is the mean length of the turn,
- N_{pri_strand} is the strands of primary winding wire,
- N_{sec_strand} is the strands of secondary winding wire,
- S_{pri_copper} is the cross section area of single primary winding wire,
- S_{sec_copper} is the cross section area of single secondary winding wire.

The analysis of AC loss is difficult to calculate because of the difficulty in calculating the AC impedance and AC current accurately. Based on the AC transformer winding resistance calculation model of Dowell, the AC power loss can be calculated as:

$$P_{copper\_AC} = I_{pri\_ACrms}{}^2 \cdot \alpha R_{pri\_winding} + I_{sec\_ACrms}{}^2 \cdot \alpha R_{sec\_winding} \tag{24}$$

Where α is the empirical factor to estimate the AC resistance due to the calculation model. It is about 1.5 to 2 in this note.

The rms AC current value of primary and secondary side can be can be estimated by:

$$I_{pri\_ACrms} = \sqrt{I_{pri\_rms}{}^2 - I_{pri\_avg}{}^2} \tag{25}$$

$$I_{sec\_ACrms} = \sqrt{I_{sec\_rms}{}^2 - I_{sec\_avg}{}^2} \tag{26}$$

The core loss can be calculated with an empirical formula as below:

$$P_{Core} = C_m \cdot (f_s)^x \cdot (B_{max})^y \cdot (C_{t0} - C_{t1} \cdot T_{Core} + C_{t2} \cdot T_{Core}{}^2) \cdot V_e \tag{27}$$

Where C_m, x, y, C_{t0}, C_{t1}, and C_{t2} are coefficients related to the material of core, B_{max} is the maximum magnetic flux, T_{core} is the temperature of core.

6. Control Circuit

(1) IC Controller

The power loss of IC controller contains internal IC consumption and the loss induced by the start up circuit.

For internal IC consumption, the power loss can be calculated as:

$$P_{IC} = V_{CC} \cdot I_{CC} \tag{28}$$

Where V_{cc} is the supply voltage of the IC controller, I_{cc} is the operation current in no-load condition. To decrease the no-load power consumption, select the lowest-possible voltage. This voltage must be higher than the lowest operating voltage of the IC controller. Select the voltage V_{CC} using the turn-ratio between the secondary and auxiliary winding for a given output voltage.

The IC controller needs a circuit to start up. Some ICs have an HV pin, some have startup resistors. Both the circuits will consume the power.

For the ICs with an HV pin, we can find the leakage current $I_{leakage}$ from HV pin in the datasheet. The loss induced by the leakage current can be estimated as:

$$P_{Leakage\_current} = V_{in} \cdot I_{Leakage} \tag{29}$$

Where V_{in} is the DC input voltage. Minimize the leakage current on the HV pin to decrease this power loss. However, it is mainly determined by the chip process.

For the IC with startup resistors, estimate the loss induced by the startup resistors as:

$$P_{startup} = \frac{\left(V_{in} - V_{CC}\right)^2}{R_{startup}} \tag{30}$$

To decrease this power loss, use large startup resistors. However, large startup resistors slow the startup speed and may even result in startup failure.

(2) Feedback Circuit

For a flyback converter with isolated output, typically adopt an optocoupler and three-terminal programmable shunt regulator like the TL431 to achieve output voltage feedback. For example, the HFC0300 feedback circuit shown in Figure 6 consumes some power for normal operation. To minimize the no-load power consumption, minimize the power loss of the feedback circuit. For an isolated flyback converter, the output voltage usually powers the optocoupler and the regulator. If the converter has multiple outputs, choose the lower voltage as the supply voltage.

Alternatively, choose a regulator with a lower operating current and an optocoupler with a high current transfer ratio (CTR).

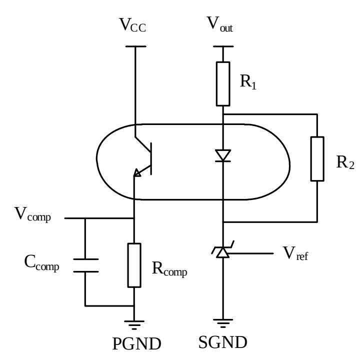

Figure 6: Feedback Circuit

For the HFC0300, V_{comp} is about 3.1V when the converter is working in burst mode condition. The primary and secondary optocoupler current can be estimated as:

$$I_{sec\_photocoupler} = \frac{V_{comp}}{R_{comp}} \tag{31}$$

$$I_{pri\_photocoupler} = \frac{I_{sec\_photocoupler}}{CTR} \tag{32}$$

Where CTR is the current transfer ratio of the optocoupler.

The power loss of feedback circuit contains two parts as shown below:

$$P_{feedback} = V_{CC} \cdot I_{sec\_photocoupler} + V_{out} \cdot I_{pri\_photocoupler} \tag{33}$$

Using a regulator with a low operating current and high CTR optocoupler, the primary and secondary side current of the feedback circuit drops and reduces the power loss. If the voltage supply for the feedback circuit is very high (typically range 12V- 24V), we can save 10mW-20mW power loss by choosing a regulator with a lower operating current and an optocoupler with a higher CTR.

EXAMPLE

In order to show the validity of no-load consumption analysis, a flyback converter controlled by the HFC0300 was built and tested. The AC input is 90 V_{rms} to 264V_{rms}; the outputs are 5V/3A and 24V/1.5A respectively. The circuit of the converter is shown in Figure 7.

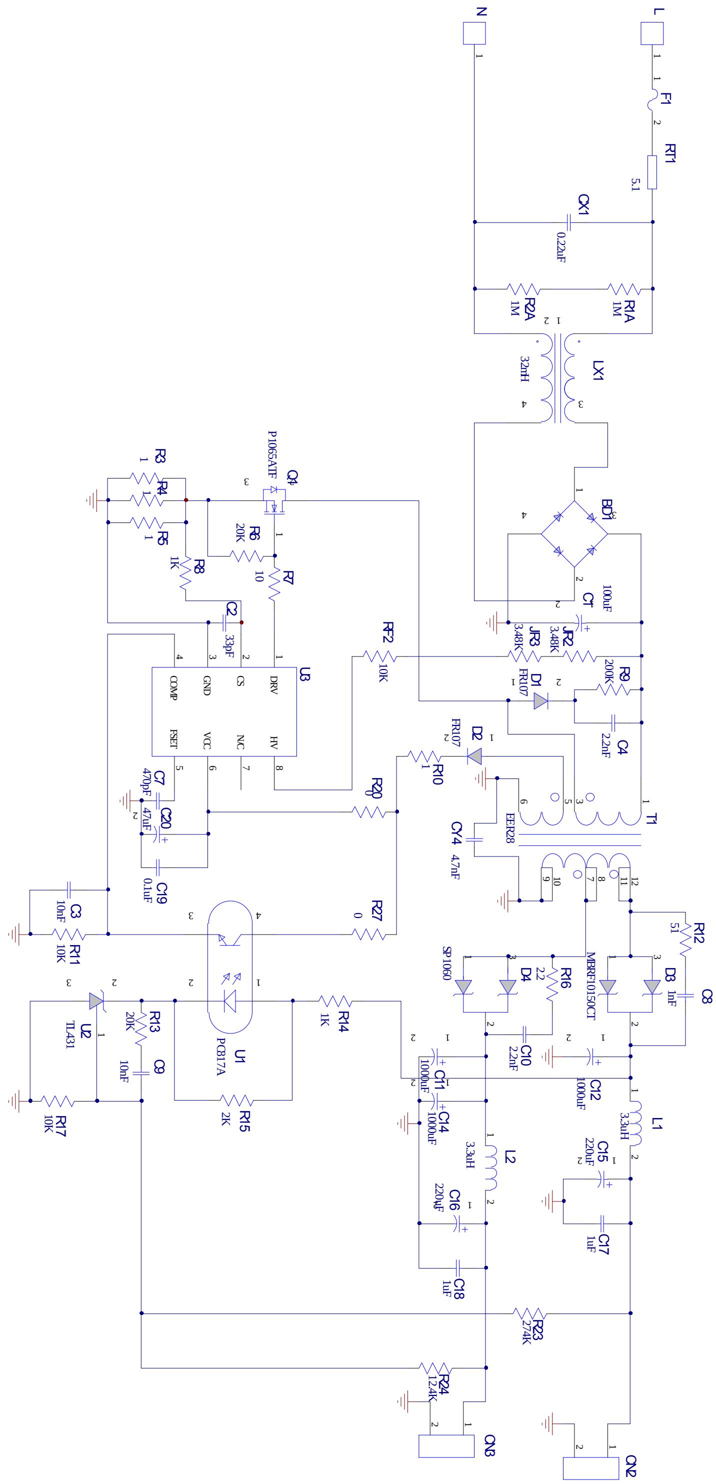

Figure 7: Circuit of HFC0300 Controlled Flyback Converter

The measured no-load power consumption was 91.5mW at 264V_{rms} AC input. To decrease the standby power loss, some measure is given as: 1) Increase the discharging resistors to 4MΩ. 2) Optimize the transformer design. 3) Choose 5V output voltage as the supply voltage of feedback circuit and use a regulator of 1.24V reference. The measured no-load power consumption decreased to 51.7mW, and the waveform of burst mode at no-load condition is as shown in Figure 8.

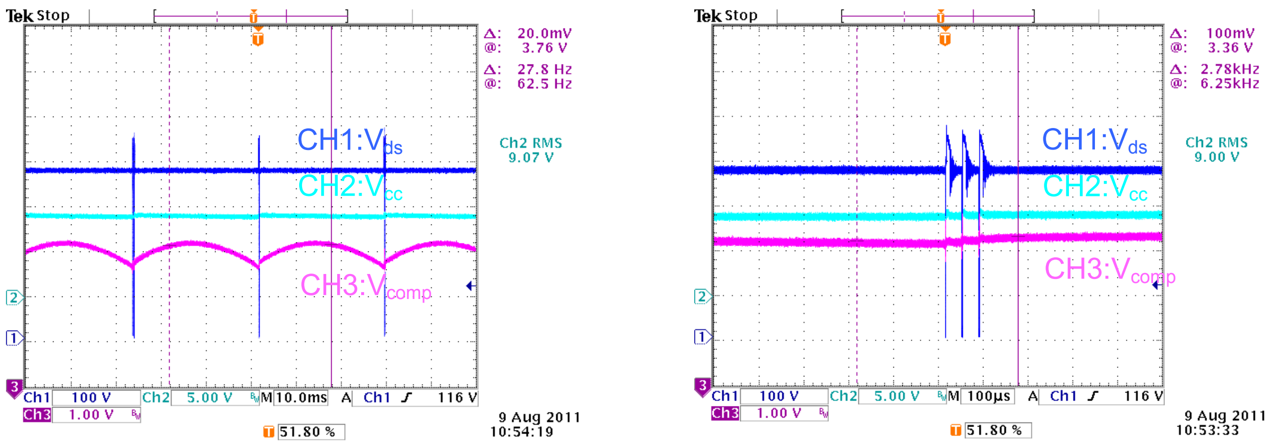

Figure 8: Waveform of Burst Mode at No-load

We can obtain the equivalent switch frequency f_s from the waveform. As per the analysis of no-load power consumption previously discussed, we can calculate the no-load power loss of each part and summarized them in Table 1.

Table 1: No-Load Power Loss Breakdown

Input Voltage		264VAC	
f_s		107Hz	
No-load power loss breakdown			
Discharging resistor	17.42mW	**Input capacitor**	7.27mW
RCD snubber	0.15mW	**MOSFET**	3.46mW
Diode	1.76mW	**Transformer**	0.12mW
IC(HFC0300)	15.34mw	**Feedback circuit**	1.07mW

Thus, we can find the loss through the discharging resistor and IC are the major part of total no-load power loss. However, with the increased equivalent frequency, the power loss of MOSFET, diode and transformer increase significantly and dominates.

APPLICATION SUGGESTIONS

Follow the suggestions below to decrease the no-load power consumption by decreasing the loss on each component.

Discharging Resistors

Small discharging resistors will cause higher power loss. Choose a suitable Xcap to optimize the no-load loss and EMI.

Electrical Capacitor

Choose an appropriate capacitor with relatively low leakage current, balanced against increased cost.

Switch Component

- Choose a MOSFET with low $R_{ds(on)}$, high switching speed, and low output capacitance.
- Use an application-appropriate gate drive resistor for the MOSFET to balance efficiency against EMI.
- Use a Schottky diode with a low forward voltage drop

Transformer

- Choose an appropriate winding size and use multiple strands of wire
- Use a transformer with a sandwich winding structure to decrease the leakage inductance
- Choose a low loss core material

Control Circuit

- Optimize the IC losses by decreasing the loss on the startup circuit and the operation current in no-load condition.
- Use a regulator with a low operating current and an optocoupler with a high-CTR
- Design an appropriate feedback circuit to decrease the equivalent frequency as much as possible.