ICE1PCS01/02

Boost Type CCM PFC Design with ICE1PCS01/02

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Abstract

Continuous conduction mode (CCM) PFC controllers, named ICE1PCS01/02, are developed based on a new control scheme. Compared to the conventional PFC solution, the new ICs does not need the direct sine-wave sensing reference signal from the AC mains. Average current control is implemented to achieve the unity power factor. In this application note, the design process for the boost PFC with ICE1PCS01/02 is presented and the design details for a 300W output power PFC with the universal input voltage range of 85~265VAC are included.

1 Introduction

The Pin layout of ICE1PCS01 and ICE1PCS02 is shown in Figure 1.

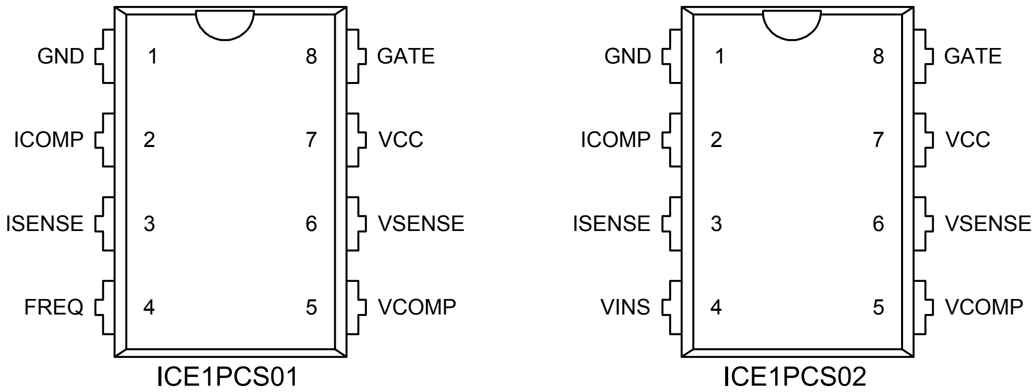

Figure 1 Pin Layout of ICE1PCS01 and ICE1PCS02

From the layout, it can be seen that most of Pins in ICE1PCS02 are the same as ICE1PCS01 except Pin 4. In ICE1PCS01, Pin 4 is to set the switching frequency. However, for ICE1PCS02, Pin 4 is for AC brown out detection and the switching frequency is fixed by internal oscillator at 65kHz. The typical application circuits of ICE1PCS01 and ICE1PCS02 are shown in Figure 2 and Figure 3 respectively.

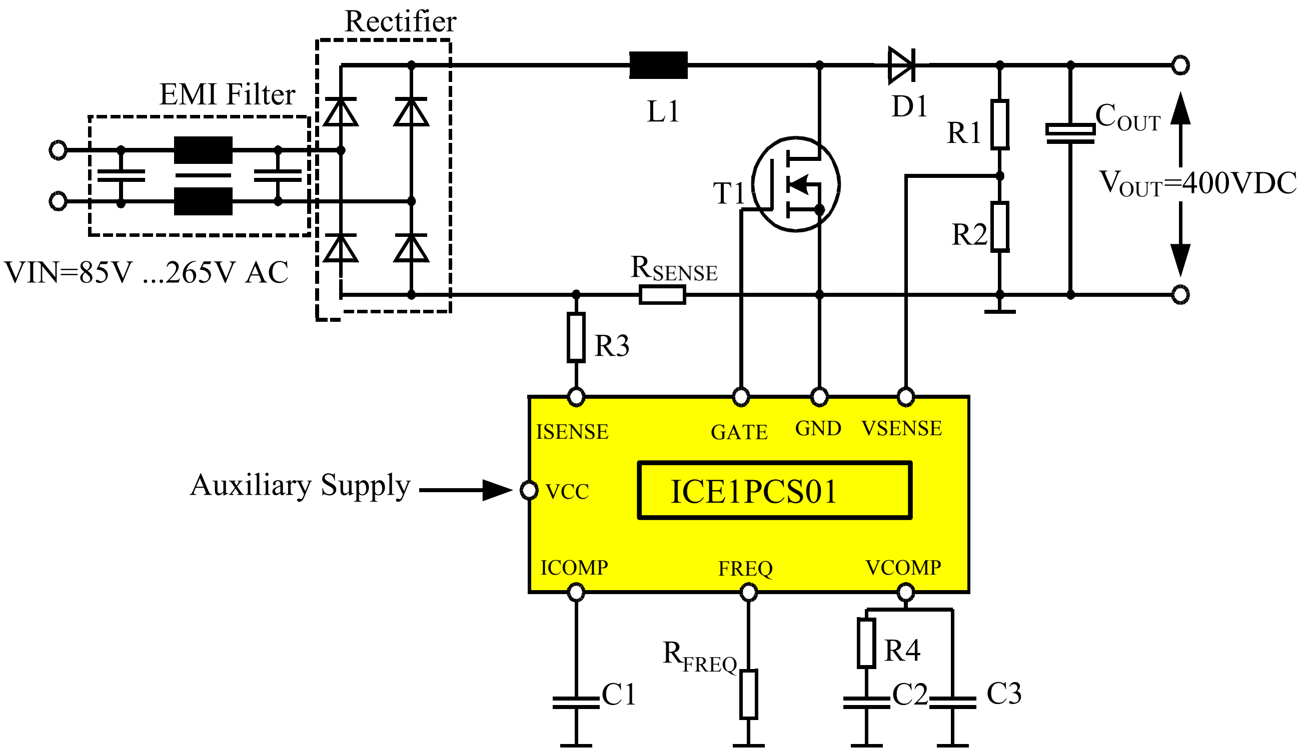

Figure 2 **Typical application circuit of ICE1PCS01**

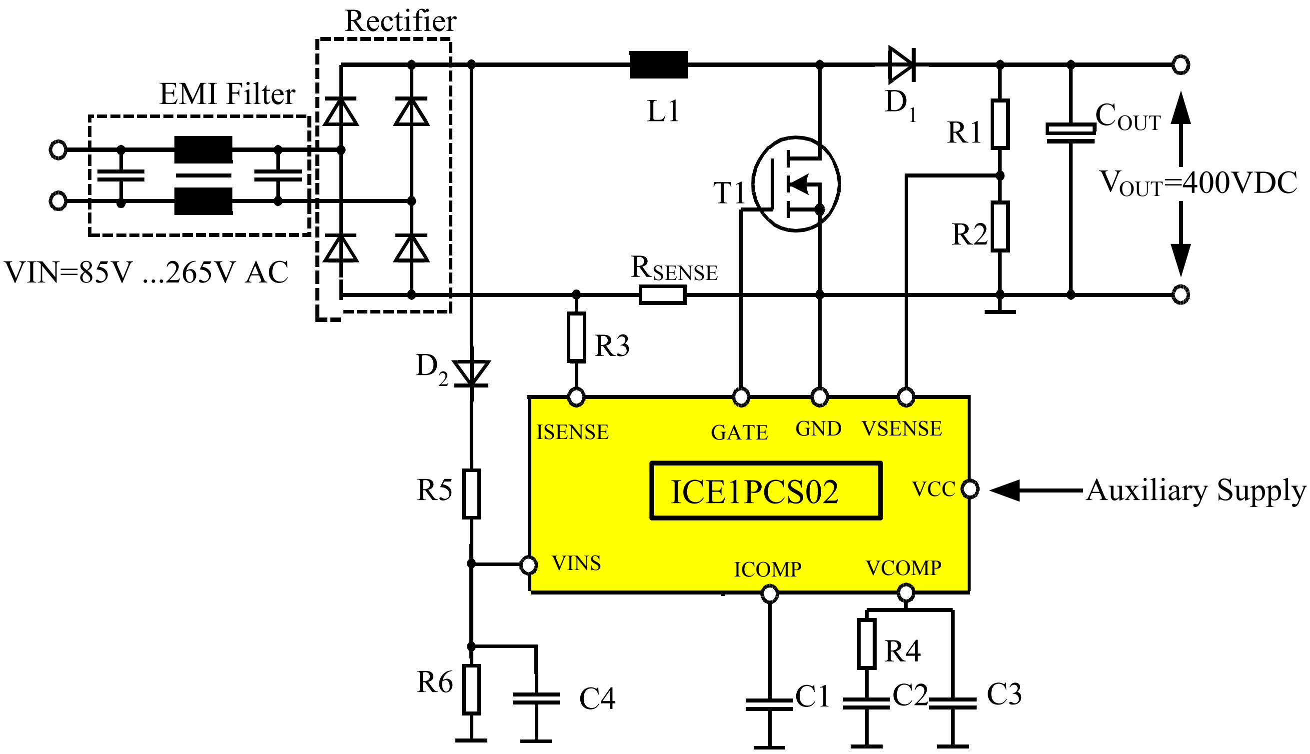

Figure 3 **Typical application circuit of ICE1PCS02**

2 Boost PFC design with ICE1PCS01/02

2.1 Target specification

The fundamental electrical data of the circuit are the input voltage range Vin, the output power Pout, the output voltage Vout, the operating switching frequency f_{SW} and the value of the high frequency ripple of the AC line current I_{ripple}. Table 1 shows the relevant values for the system calculated in this Application Note. The efficiency at rated output power Pout is estimated to 90 % over the complete input voltage range.

Input voltage	85VAC~265VAC
Input frequency	50Hz
Output voltage and current	390VDC, 0.76A
Output power	300W
Efficiency	>90% at full load
Switching Frequency	65kHz
Maximum Ambient temperature around PFC	70ºC

Table 1 Design parameter for the proposed design

2.2 Bridge rectifier

In order to obtain 300W output power at 85 V minimum AC input voltage, the maximum input RMS current is

$$I_{in\_RMS} = \frac{P_{out}}{V_{in\_\min} \cdot \eta} = \frac{300}{85 \cdot 90\%} = 3.92A$$

and the sinusoidal peak value of AC current is

$$I_{in\_pk} = \sqrt{2} \cdot I_{in\_RMS} = \sqrt{2} \cdot 3.92 = 5.54A$$

For these values a bridge rectifier with an average current capability of 6A or higher is a good choice. Please note here, that due to a power dissipation of approximately

$$P_{BR} = 2 \cdot V_F \cdot I_{in\_RMS} = 2 \cdot 1V \cdot 3.92A = 7.84W$$

the rectifier bridge should be connected to an appropriate heatsink. Assuming a maximum junction temperature T_{Jmax} of 125°C, a maximum ambient temperature T_{Amax} of 70°C, the thermal junction-to-case R_{thJC} of approximate 2.5 K/W and the thermal case to heatsink R_{thCHS} of approximate 1K/W, the heatsink must have a maximum thermal resistance of

$$R_{thHS\_BR} = \frac{T_{J\max} - T_{A\max}}{P_{BR}} - R_{thJC} - R_{thCHS} = \frac{125-70}{7.84} - 2.5 - 1 = 3.52K/W$$

2.3 Power MOSFET and Gate Drive Circuit

Due to the switch mode operation, the losses are only active during the on-time of the MOSFET. The duty cycle of the transistor in boost converters operating in CCM at minimum AC input RMS voltage is

$$D_{on} = 1 - \frac{V_{in\_\min}}{V_{out}} = 1 - \frac{85}{390} = 0.782$$

Since rms-values have the same effect on a system as DC-values, it is possible to calculate a characteristic duty cycle for the rms-value. Therefore, the on-state losses of the MOSFET in CCM-mode at a junction-temperature of 125°C are

$$P_{cond} = I_{in\_RMS}^{2} \cdot D_{on} \cdot R_{dson(125C)}$$

the MOSFET switching loss can be estimated as

$$P_{SW} = (E_{on} + E_{off}) \cdot f_{SW}$$

where, E_{on} and E_{off} are the switch-on and switch-off energy loss which data can be found in MOSFET datasheet, f_{SW} is the switching frequency.

For 300W design, if SPP20N60C3 is used, the conduction loss is

$$P_{cond} = 3.92^{2} \cdot 0.782 \cdot 0.42 = 5.05W$$

assuming the switching current is about 6A and gate drive resistance Rg=3.6Ω, then the switching loss is

$$P_{SW} = (0.007mWs + 0.015mWs) * 65kHz = 1.43W$$

the total loss is

$$P_{MOS\_total} = P_{cond} + P_{SW} = 6.48W$$

the required heatsink for the MOSFET is

$$R_{thHS\_MOS} = \frac{T_{J\max} - T_{A\max}}{P_{MOS\_total}} - R_{thJC\_MOS} - R_{thCHS} = \frac{125-70}{6.48} - 0.6 - 1 = 6.89K/W$$

the gate drive resistance is used to drive MOSFET as fast as possible but also keep dv/dt within EMI specification. In this 300W example, 3.6Ω gate resistor is chosen for SPP20N60C3 MOSFET.

Beside gate drive resistance, one 10kΩ resistor is also commonly connected between MOSFET gate and source to discharge gate capacitor.

2.4 Boost Diode

The boost diode D1 has a big influence on the system's performance due to the reverse recovery behaviour. So the Ultra-fast diode with very low t_{rr} and Q_{rr} is necessary to reduce the switching loss. The new diode technology of silicon carbide (SiC) Schottky shows its outstanding performance with almost no reverse recovery behaviour. The switching loss due to the boost diode can be ignored with choosing SiC Schottky diode. Only conduction loss is calculated as below.

$$P_{diode} = V_{F} \cdot I_{in\_RMS} \cdot (1 - D_{on}) = 2V \cdot 3.92A \cdot (1 - 0.782) = 1.71W$$

For a rule of thumb, the SiC diodes provide a output power Pout of a CCM-PFC-system of 100 W to120 W per rated ampere of the diode. This means for example, that the SDT04S60 from Infineon Technologies which is rated for a forward current IF = 4 A is capable for a system of Pout = 4*100 W = 400 W system in minimum. Therefore, this diode would be suitable for the proposed design.

The required heatsink for boost diode is

$$R_{thHS\_diode} = \frac{T_{J\max} - T_{A\max}}{P_{diode}} - R_{thJC\_diode} - R_{thCHS} = \frac{125-70}{1.71} - 4.1 - 1 = 27.06K/W$$

The SiC boost diodes often have a poor surge current capability, so that they may break down. Therefore a so called bypass diode is necessary such as the diode D3 as Figure 4. For the proposed system, 1N5408 is suitable.

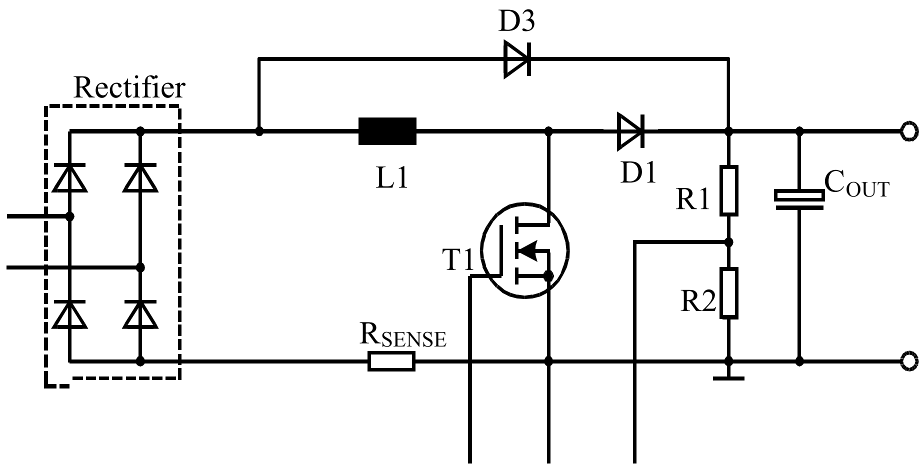

Figure 4 inrush current bypass diode

2.5 Boost inductor

The peak current that the inductor must carry is the peak line current at the lowest input voltage plus the high frequency ripple current. The high frequency ripple current peak to peak, I_{HF}, can be related to maximum input power and minmum input voltage as equation below.

$$I_{HF} = k \cdot \sqrt{2} \cdot \frac{P_{in\_max}}{V_{in\_min}}$$

Where, k must be kept reasonably small, and is usually optimized in the range of 15% to 25% for cost effective design based on the current magnetic component status. If k is too high, the larger AC input filter is required to filter out this ripple noise. If k is too low, the value of the inductance is too large and leads to big size of the magnetic core.

For example, we choose k 22%. Then,

$$I_{HF} = 22\% \cdot \cdot \sqrt{2} \cdot \frac{P_{in\_max}}{V_{in\_min}} = 1.2A$$

The peak current passing through inductor is

$$I_{L\_pk} = I_{in\_peak} + \frac{I_{HF}}{2} = 5.54 + \frac{1.2}{2} = 6.14A$$

The boost inductance must be

$$L_{boost} \geq \frac{D \cdot (1-D) \cdot V_{out}}{I_{HF} \cdot f_{SW}}$$

D=0.5 will generate the maximum value for the above equation.

$$L_{boost} \geq \frac{0.5 \cdot (1-0.5) \cdot 390V}{1.2A \cdot 65kHz} = 1.25mH$$

The magnetic core of the boost choke can be either magnetic powder or ferrite material.

(1) sendust powder toroid core
The required effective magnetic volume of the core, V_e, is

$$V_e \geq \mu_r \mu_0 L_{boost} (\frac{I_{L\_pk}}{B_{\max}})^2 = 125 \cdot 1.257e-6 \cdot 1.25mH (\frac{6.14A}{0.8T})^2 = 11.6e-6m^3 = 11.6cm^3$$

where, μ_r is the relative permeability which is fixed by core manufacturer; μ_0 is magnetic field constant which is equal to 1.257e-6; B_{max} is the maximum magnetic flux density for the selected magnetic material (for sendust, B_{max} is up to 0.8T.) Select a core with similar Ve value from the magnetic core datasheet. For example, the core type CS468125 from Chang Sung Corporation is suitable for this case. The parameters of CS468125 are V_e=15.584cm^3, A_e=1.34cm^2, C=11.63cm, μ_r=125. The turn number of the boost choke winding is

$$N_{toroid\_boost} = \sqrt{\frac{L_{boost} \cdot C}{\mu_r \mu_0 A_e}}$$

where, C is the magnetic path length and A_e is the effective magnetic cross section area.

The copper loss of the winding wire can be calculated on I_{in_RMS}.

$$P_{L\_boost} = I_{in\_RMS}^2 \cdot R_{L\_boost}$$

Selecting the proper wire type to fullfil the loss and thermal requirement for the choke.

(2) ferrite core
To make sure the ferrite core will not go into saturation, the turn number of the boost choke winding with ferrite core is

$$N_{ferrite\_boost} \geq \frac{I_{L\_pk} \cdot L_{boost}}{B_{\max} \cdot A_{\min}}$$

where, B_{max} is up to 0.3T according to ferrite material specification; A_{min} is the minimum magnetic cross section area.

The winding wire copper loss calculation is the same as in the above section of sendust powder toroid core.

2.6 AC line current filter

As decribed in section 2.5, there is high frequency ripple current peak to peak I_{HF} passing through boost choke. This ripple will also go into AC line power network. The current filter is necessary to reduce the amplitude of high frequency current component. The filtering circuit consists of a capacitor and an inductor as shown in Figure 5.

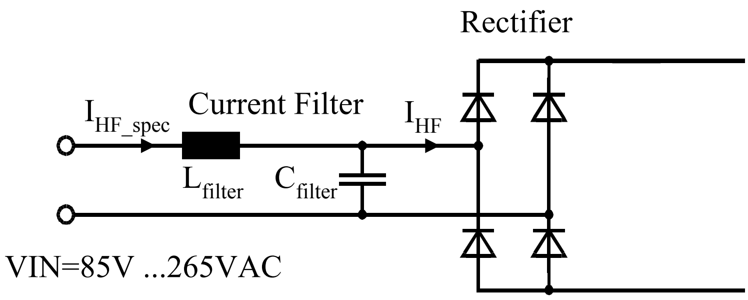

Figure 5 AC line current filter

The required L_{filter} is

$$L_{filter} \geq \frac{\frac{I_{HF}}{I_{HF\_spec}} + 1}{(2\pi f_{SW})^2 C_{filter}}$$

normally there is one EMI X2 capacitor which can act as C_{filter}. In this example, if we define I_{HF_spec} as 0.2A peak to peak and asumming X2 capacitance 0.47μF, then

$$L_{filter} \geq \frac{\frac{1.2A}{0.2A} + 1}{(2\pi \cdot 65kHz)^2 \cdot 0.47\mu F} = 89\mu H$$

The leakage inductance of EMI common mode choke can be used for current filter. If the leakage inductance is large enough, no need to add the additional differential mode inductor for filtering. Otherwise, a current filter choke is necessary. The calculation method for the current filter choke is the same as for boost choke.

2.7 Boost Output Bulk Capacitance

The bulk capacitance has to fullfil two requirements, output double line frequency ripple and holdup time.

(1) output double line frequency ripple limit.

The inherent PFC always presents 2*f_L ripple. The amplitude of ripple voltage is dependant on output current and bulk capacitance as below.

$$C_{out} \geq \frac{I_{out}}{\pi \cdot 2 * f_L \cdot V_{out\_ripple\_pp}}$$

where, I_{out} is the PFC output current, $V_{out\_ripple\_pp}$ is the output voltage ripple (peak to peak), and f_L is the AC line frequency.

Please note that ICE1PCS01/02 has enhance dynamic block which is active when Vout exceed ±5% of regulated level. The enchanc dynamic block should be designed to work only during load or line change. During steady state with constant load, the enhance dynamic block should not be triggered, otherwise THD will be deteriorated. That means the target $V_{out\_ripple\_pp}$ must be lower than 10% of V_{out}. For this example, Vout=390VDC, then $V_{out\_ripple\_pp}$ must be lower than 39V. if we define $V_{out\_ripple\_pp}$=8V, then

$$C_{out} \geq \frac{I_{out}}{\pi \cdot 2 * f_L \cdot V_{out\_ripple\_pp}} = 306\mu F$$

(2) holdup time requirement

After the PFC stage, there is commonly a PWM stage to provide isolated DC output for end user. Some applications, especially computing, have the holdup time requirement. It means that PWM stage should be able to provide the isolated output even if AC input voltage become zero for a short holdup time. The common specification for this holdup time is 20ms. If minimum input voltage for PWM stage is defined as 250VDC, then the bulk capacitance will be

$$C_{out} \geq \frac{2 \cdot P_{out} \cdot t_{holdup}}{V_{out}^2 - V_{out\_min}^2} = \frac{2 \cdot 300W \cdot 20ms}{390^2 - 250^2} = 134\mu F$$

the final C_{out} capacitance should be higher value calculated from the above two requirement.

2.8 Current Sense Resistor

The current sense resistance is calculated based on the IC soft over current control threshold and peak current carried by boost choke.

When the Isense signal reaches the soft over control threshold, IC will reduce the internal control voltage and accordingly the duty cycle is reduced in the following cycles. Finally the boost choke current is limited. According to IC datasheet, soft over current control threshold is -0.66V maximum. So the current sense resistor should be

$$R_{sense} \leq \frac{0.66V}{I_{L\_pk}} = \frac{0.66V}{6.14A} = 0.11\Omega$$

According to Figure 2 and Figure 3, the transistor current as well as the diode current is sensed with R_{sense}. That means that also the inrush current is sensed there leading to a large negative voltage drop at R_{sense}, because the inrush current is in the range of about 150 A to 200 A. It is therefore necessary to limit the current into Pin 2 (ISENSE) to 1 mA, which is realized with resistor R3. A value of R3 = 220Ω is sufficient for this resistor.

2.9 Output voltage sensing divider

The output voltage is set with the voltage divider represented by R_1 and R_2 in Figure 2 and Figure 3. First, choose the value of the lower resistor R_2. Then the value of the upper resistor R_1 is

$$R_1 = \frac{V_{out} - V_{ref}}{V_{ref}} \cdot R_2$$

where, Vref is IC internal reference voltage for voltage sensing, 5V typical.

If R_2=10kΩ,

$$R_1 = \frac{390-5}{5} \cdot 10k\Omega = 770k\Omega$$

It is recommended to take resistor values with a tolerance of 1% for R_1 and R_2. Due to the voltage stress of R1, it is recommended to split this value into few resistors in series.

2.10 Frequency setting (only for ICE1PCS01)

The frequency of the ICE1PCS01 is adjustable in the range of 50 kHz up to 200 kHz. The external resistor R_{FREQ} according to Figure 2 programs a current which controls the oscillator. The given points of the resistor-frequency-characteristic are (250 kHz / 18 k.), (125 kHz / 33 k.) and (50 kHz / 82k.) and Figure 6 shows the curve through those points.

Figure 6 Resistor-frequency characteristic

2.11 AC Brown-out Shutdown (only for ICE1PCS02)

Brown-out occurs when the input voltage VAC falls below the minimum input voltage of the design (i.e. 85V for universal input voltage range) and the VCC has not entered into the VCCUVLO level yet. For a system without input brown out protection (IBOP), the boost converter will increasingly draw a higher current from the mains at a given output power which may exceed the maximum design values of the input current and lead to over heat of MOSFET and boost diode. ICE1PCS02 provides a new IBOP feature whereby it senses directly the input voltage for Input Brown-Out condition via an external resistor/capacitor/diode network as shown in Figure 7. This network provides a filtered value of VIN which turns the IC on when the voltage at Pin 4 (VINS) is more than 1.5V. The IC enters into the standby mode and gate is off when VINS goes below 0.8V. The hysteresis prevents the system to oscillate between normal and standby mode. Note also that input voltage needs to at least 16% of the rated VOUT in order to overcome open loop protection and powerup the system (referred to application note of ICE1PCS01).

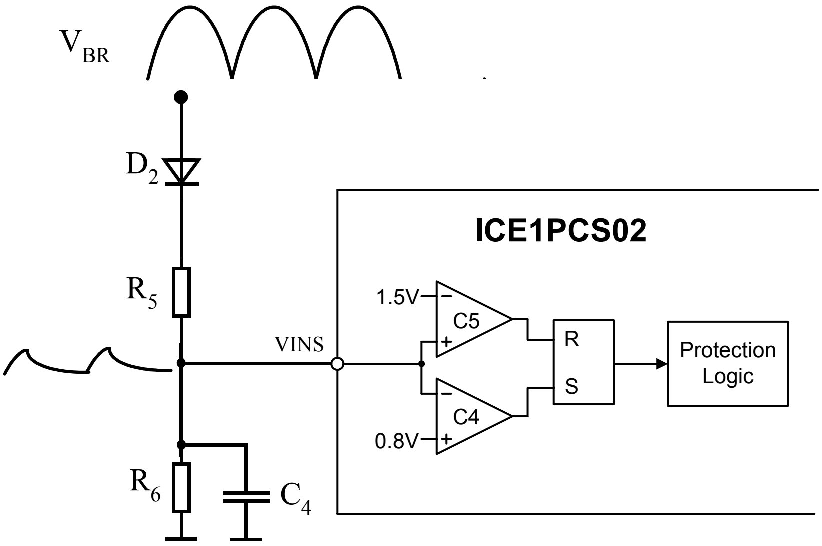

Figure 7 Block diagram of voltage loop

Because of the high input impedence of comparator of C4 and C5, R5 can be high ohmic resistance to reduce the loss. From the datasheet, the bias current on VINS Pin is 1µA maximum. In order to have the design consistence, the current passing through R5 and R6 has to be much higher than this bias current, for example 7µA. Then R6 is:

$$R_6 = \frac{0.8V}{7\mu A} = 114k\Omega$$

R6 is selected 120KΩ. R5 is selcted by

$$R_5 = \frac{\sqrt{2} \cdot V_{AC\_on} - 1.5V}{1.5V} \cdot R_6$$

where, V_{AC_on} is the minimum AC input voltage (RMS) to start PFC, for example 70VAC.

$$R_5 = \frac{\sqrt{2} \cdot 70V - 1.5V}{1.5V} \cdot 120k\Omega = 7.8M\Omega$$

Due to the voltage stress of R_5, it is recommended to split this value into few resistors in series.

C_4 is used to modulate the ripple at the VINS pin. The timing diagram of VINS pin when IC enters brown-out shutdown is shown in Figure 8.

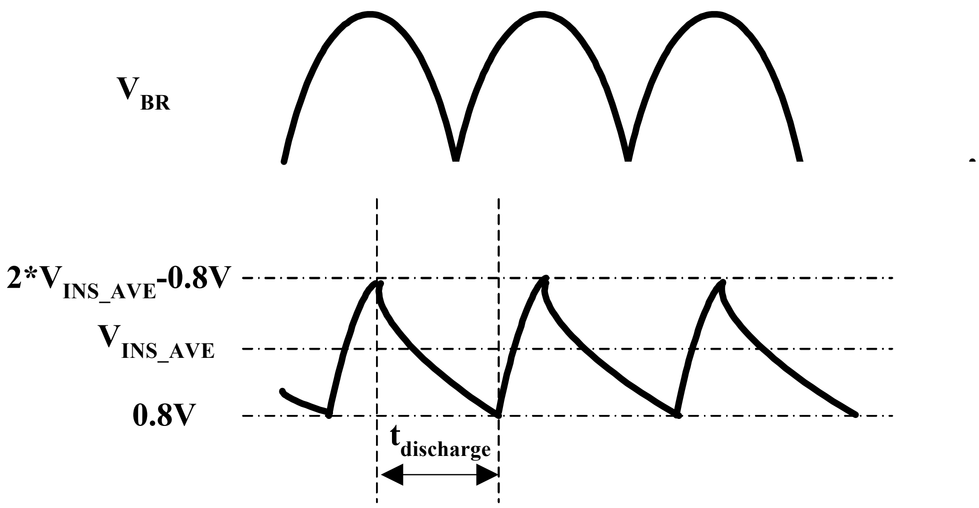

Figure 8 Timing diagram of VINS Pin when IC enters brown-out shutdown

If the bottom level of the ripple voltage touches 0.8V, PFC is in standby mode and gate is off. The ripple voltage defines PFC brown out off threshold of AC input voltage (RMS), V_{AC_off}. C_4 can be obtained from the following equation. Assuming $V_{INS\_AVE} = \frac{R_6}{R_5 + R_6} \cdot V_{AC\_off}$, where, V_{AC_off} is the maximum AC input voltage (RMS) to switch off PFC, for example 65VAC.

$$(2 \cdot \frac{R_6}{R_5 + R_6} \cdot V_{AC\_off} - 0.8) \cdot e^{-\frac{t_{discharge}}{R_6 C_4}} = 0.8V$$

assuming $t_{discharge}$ is equal to half cycle time of line frequency, ie. $t_{discharge} = \frac{1}{2 f_L}$, then

$$C_4 = \left(2 f_L R_6 \ln \frac{2 \cdot \frac{R_6}{R_5 + R_6} V_{AC\_off} - 0.8V}{0.8V} \right)^{-1}$$

$$C_4 = \left(2 \cdot 50Hz \cdot 120k\Omega \ln \frac{2 \cdot \frac{120k\Omega}{7.8M\Omega + 120k\Omega} 65V - 0.8V}{0.8V} \right)^{-1} = 219nF$$

2.12 IC supply

Because of the low Vcc turn-on/off hysteresis, ICE1PCS01/02 is not supposed to be supplied by self supply method with auxiliary winding on boost choke. The ICs can only be supplied by external DC applying to Vcc. One Electrolyte capacitor 47uF and one ceramic capacitor 100nF is recommended to connect between Vcc and GND to filter out the noise.

2.13 Voltage loop and current loop compensation

Please refer to Reference [5] for detail.

References

[1] Infineon Technologies: ICE1PCS01 - Standalone Power Factor Correction Controller in Continuous Conduction Mode; Preliminary datasheet; Infineon Technologies; Munich; Germany; Sept. 2002.

[2] Luo Junyang, Jeoh Meng Kiat, Huang Heng Cheong, A New Continuous Conduction Mode PFC IC With Average Current Mode Control, PEDS 2003; pp. 1110-1114, Nov 2003.

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