Design guide for dual-stage PFC+LLC LED driver using ICL5102

Design guide

Revision 1.0

About this document

Scope and purpose

This document describes the procedure to design a dual-stage PFC+LLC offline LED driver using ICL5102 step by step based on the ICL5102 130 W Constant Current (CC) board.

Intended audience

This document is intended for anyone who needs to design an AC-DC offline LED driver with ICL5102 for dual-stage PFC+LLC control.

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Introduction

1 Introduction

The ICL5102 is a highly integrated multi-mode (Critical Conduction Mode – CrCM and Discontinuous Conduction Mode – DCM) Power Factor Correction (PFC) and resonant Half-Bridge (HB) combination controller. The integration of PFC and HB into a single controller enables reduction of external components and optimizes performance by harmonized operation of the two stages.

The two-stage approach divides the PFC responsibilities from the output current regulation functions. This ensures low variation in the output voltage and current and allows for low Total Harmonic Distortion (THD), high Power Factor (PF) and a greater ability to withstand AC-line perturbations. The multi-mode operation of the PFC converter provides excellent efficiency over the whole load range.

The resonant HB converter supports both LLC and LCC topologies with fixed or variable switching frequency control for highest efficiency, and Burst Mode (BM) enables low-standby power consumption.

1.1 Product highlights

- PFC controller with CrCM and DCM
- Resonant HB controller with fixed or variable switching frequency control
- Maximum 500 KHz HB switching frequency and soft-start frequency up to 1.3 MHz
- BM of HB controller with power limitation ensures low standby power of less than 300 mW in dim-to-off condition
- THD optimization ensures low harmonic distortion (THD less than 10 percent) down to 30 percent nominal load

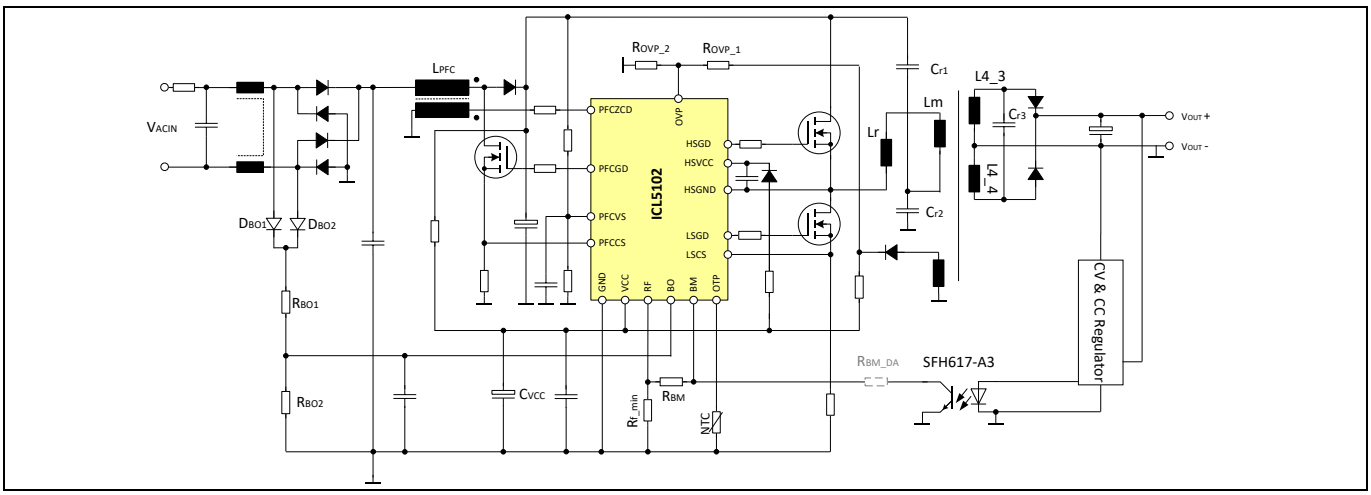
1.2 Design features

- Supports universal AC input voltage (90 to 305 V_{rms}) nominal
- Excellent system efficiency up to 94 percent
- Applicable power range up to 350 W

1.3 Target applications

- Offline AC-DC LED drivers for commercial and industrial lighting up to 350 W
- High-density AC-DC power supply

Figure 1 shows a typical LED driver application using ICL5102 with PFC+LCC topology.



Introduction

Figure 1 ICL5102 typical application with PFC+LCC topology

1.4 Pin configuration and description

ICL5102 pin assignments and basic pin descriptions are shown in **Figure 2** and **Table 1**.

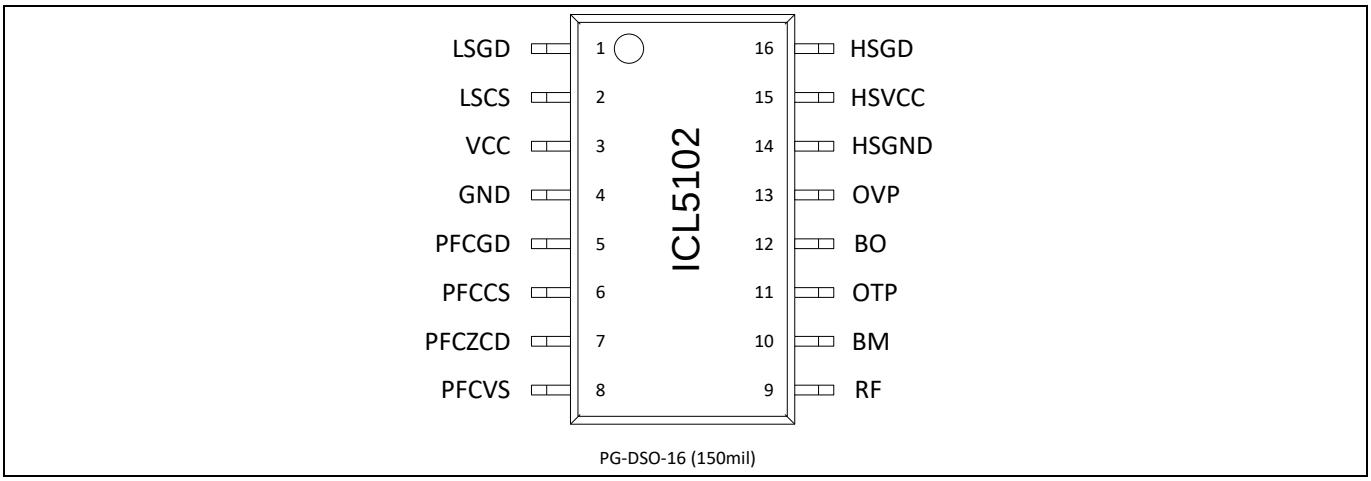


Figure 2 Pinning of ICL5102

Table 1 Pin definitions and functions

Name	Pin	Type	Function
LSGD	1	O	**HB Low-Side (LS) gate driver** Output for directly driving the HB LS MOSFET via a resistor
LSCS	2	I	**HB Current Sense (CS)** Connected to an external shunt resistor and the source of the HB LS MOSFET
V_{CC}	3	I	**Positive power supply** IC power supply
GND	4	-	**Ground** IC ground
PFCGD	5	O	**PFC gate driver** Output for directly driving the PFC MOSFET via a resistor
PFCCS	6	I	**PFC CS** Connected to an external shunt resistor and the source of the PFC MOSFET
PFCZCD	7	I	**PFC Zero-Crossing Detection (ZCD)** Connected to the PFC auxiliary winding via a resistor for PFC inductor current ZCD
PFCVS	8	I	**PFC bus voltage sense** Connected to a high-impedance resistor divider from the PFC controller output for bus voltage sensing
RF	9	I	**HB minimum switching frequency setting** Connected via an external resistor to GND for HB minimum switching frequency setting
BM	10	I	**BM enter/exit setting**

Introduction

Name	Pin	Type	Function
			Connected to an optocoupler and to the RF pin with an external resistor for BM enter/exit setting
OTP	11	I	**Over-Temperature Protection (OTP)** Connected to an external Negative Temperature Coefficient (NTC) thermistor for external OTP
BO	12	I	**Brown-in/out detection** Connected to the rectified input voltage via an external resistor for input brown-in/out detection
OVP	13	I	**Output Over-Voltage Protection (OVP)** Connected to the HB auxiliary winding via a resistor divider and diode for OVP of the secondary output voltage
HSGND	14	-	**High-Side (HS) ground** Ground for floating HS driver of HB
HSVCC	15	I	**HS V_{CC} power supply** Power supply of the HS floating driver of HB, supplied via bootstrap circuit
HSGD	16	O	**HS floating gate driver** Output for directly driving the HB floating HS MOSFET via a resistor

The ICL5102 pin connection schematic is shown in **Figure 3**.

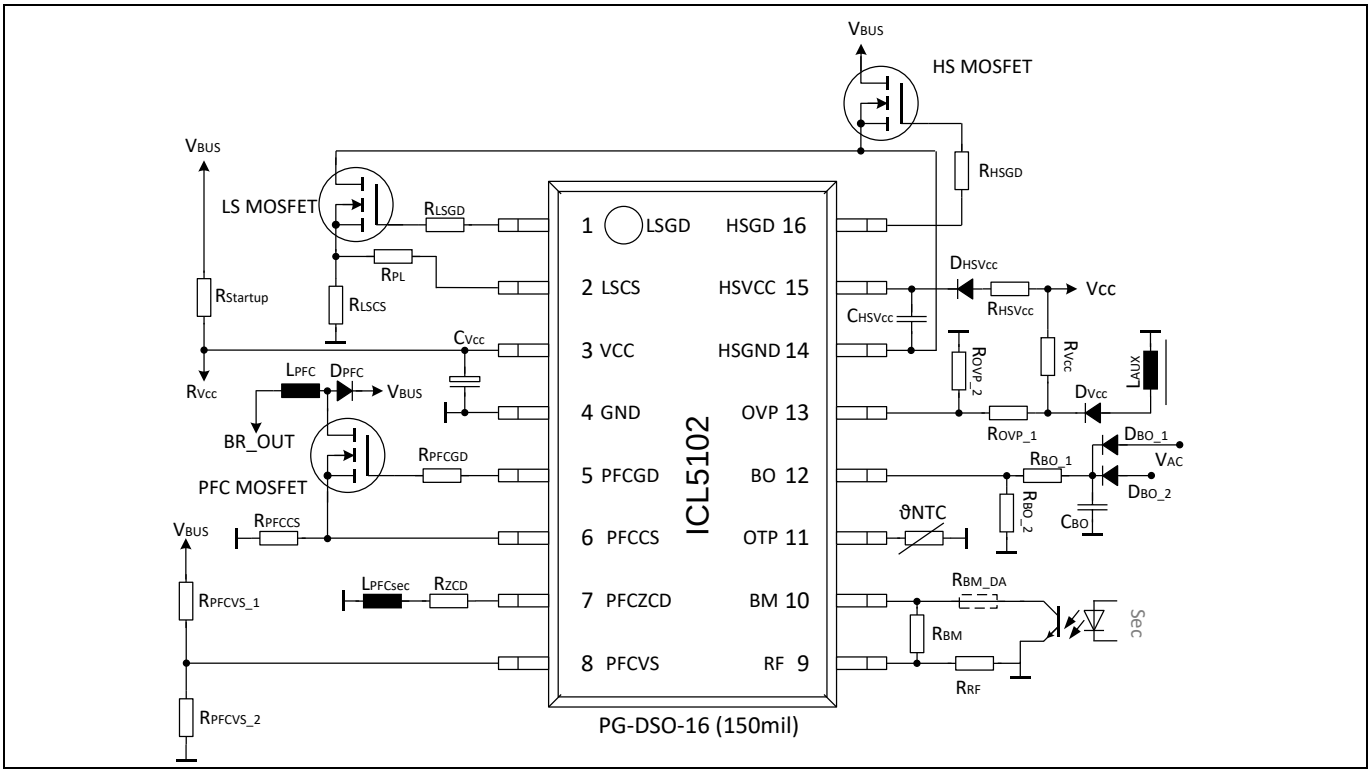


Figure 3 ICL5102 pin connection

2 Design specification

The design example used in this design guide is a 130 W CC mode reference design for direct driving of LED lighting applications based on the LLC topology. The customer can easily apply their own target specifications according to this example and obtain the design parameters by themselves.

2.1 System specification of a 130 W PFC+LLC reference design for LED lighting applications

The system specification of the ICL5102 130 W CC reference design for LED lighting applications is given in **Table 2**.

Table 2 System design specification for 130 W PFC+LLC reference design

Parameter	Symbol	Target value	Unit
General input/output specification			
Nominal input AC voltage	V_{IN_nom}	90 ~ 305	V_{RMS}
Line frequency	f_{AC_line}	47 ~ 63	Hz
Nominal output voltage	V_{OUT}	38 ~ 76	V_{DC}
Nominal output current	I_{OUT_nom}	1.75	A
Minimum output current	I_{OUT_min}	75	mA
Nominal output power	P_{O_nom}	130	W
Power factor	PF	More than 0.9	
THD	iTHD	Less than 10	%
Power efficiency	η	Less than 92	%
Maximum standby power at no load	$P_{O\_STB\_max}$	300	mW
Maximum time to light at 90 V AC	t_{T2L}	350	ms
Protections			
Brown-out threshold	V_{IN_BO}	71	V_{RMS}
Brown-in threshold	V_{IN_BI}	90	V_{RMS}
Maximum output over-voltage threshold	V_{OUT_OVP}	90	V_{DC}
OTP threshold	T_{OTP}	85	°C
Standard compliance			
Harmonics	–	EN61000-3-2 Class-C	–
EMI	–	EN55015	–
Safety	–	EN61347-2-13	–
Board dimensions			
Size	L x H x B	178 x 52 x 32	mm

Design specification

2.2 Schematic

The schematic of the ICL5102 130 W reference design for LED lighting applications is given in **Figure 4**.

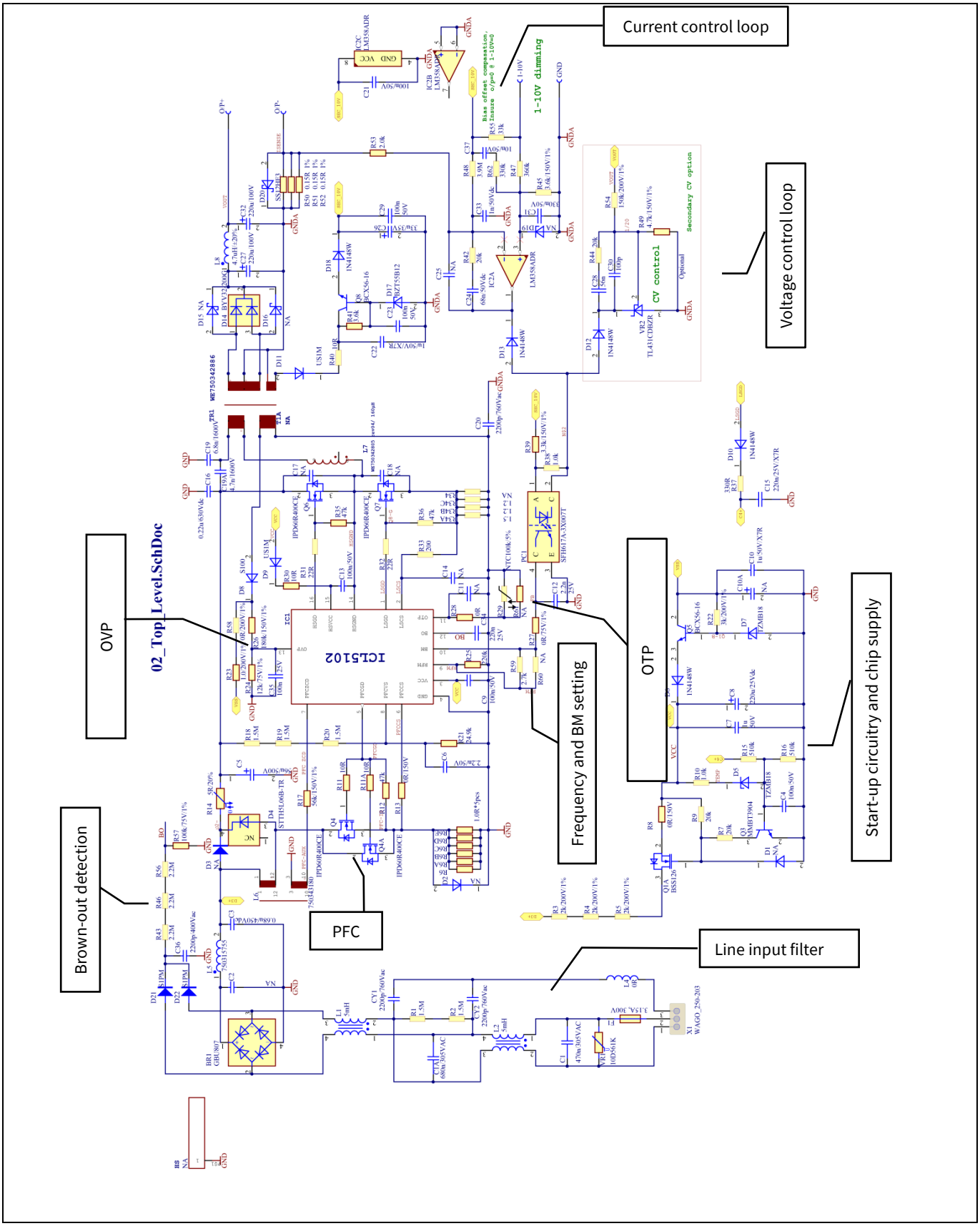


Figure 4 ICL5102 130 W reference design schematic

3 PFC boost converter design

PFC shapes the input current of the power supply to be in synchronization with the mains voltage, in order to maximize the real power drawn from the mains. In a perfect PFC circuit, the input current follows the input voltage as a pure resistor, without any input current harmonics. In the ICL5102 130 W reference design, PFC is implemented as a boost converter, which works in CrCM and DCM with constant on-time control to provide a constant high DC voltage for the HB control.

The design specification of the PFC boost converter is given in **Table 3**.

Table 3 Design specification of the PFC boost converter

Parameter	Symbol	Value	Unit
Minimum AC input voltage	V_{IN_min}	90	V_{RMS}
Minimum AC input peak voltage	$V_{IN\_pk\_min}$	127	V_{RMS}
Maximum AC input voltage	V_{IN_max}	305	V_{RMS}
Maximum AC input peak voltage	$V_{IN\_pk\_max}$	431	V_{RMS}
AC-line frequency	f_{AC_line}	47 ~ 63	Hz
Maximum PFC output power	$P_{PFC\_out\_max}$	145	W
Minimum PFC bus voltage	V_{bus_min}	400	V
Nominal PFC bus voltage	V_{bus}	450	V
PFC OVP level 1	V_{bus_OVP1}		
Brown-out threshold	V_{IN_BO}	71	V_{RMS}
Brown-in threshold	V_{IN_BI}	90	V_{RMS}
Minimum PFC switching frequency	f_{PFC_min}	35	KHz
Estimated PFC efficiency	η_{PFC}	Less than or equal to 96	%

3.1 Main PFC boost inductor

As the key magnetic component of the PFC boost converter, the boost inductor has the main function of energy storage. Its inductance is given as the following formula:

$$L_{PFC} = \frac{V_{IN\_pk}^2 * (V_{bus} - V_{IN\_pk}) * \eta_{PFC}}{4 * V_{bus} * P_{PFC\_out} * f_{PFC}}$$

Where:

- L_{PFC} – inductance of the PFC boost inductor
- V_{IN_pk} – peak value of the input AC mains
- V_{bus} – bus voltage as the PFC output
- η_{PFC} – estimated power efficiency of the PFC boost converter
- P_{PFC_out} – output power of the PFC boost converter
- f_{PFC} – operation switching frequency of the PFC boost inductor

The maximum possible inductance should be calculated at both lowest (input brown-out threshold) and highest possible input voltage (input over-voltage threshold) with full load and minimum switching frequency.

PFC boost converter design

At V_{IN_min} = 71 V_{RMS}:

$$L_{PFC\_71} = \frac{(71 * \sqrt{2})^2 * (450 - 71 * \sqrt{2}) * 0.96}{4 * 450 * 145 * 35 * 10^3} \approx 370.3\mu H$$

At V_{IN_max} = 305 V_{RMS}:

$$L_{PFC\_305} = \frac{(305 * \sqrt{2})^2 * (450 - 305 * \sqrt{2}) * 0.96}{4 * 450 * 145 * 35 * 10^3} \approx 366.1\mu H$$

The suitable inductance must be less than the smaller one of both.

$$L_{PFC} < min(L_{PFC\_71}, L_{PFC\_305})$$

Other considerations regarding PFC choke inductance:

- The selected PFC inductance must be small enough to cover the maximum output power at the minimum input (e.g. to cover the brown-in/out feature).
- Bigger PFC inductance has the advantage at light load in comparison to smaller inductance due to longer on-time. This ensures smaller minimum output power in DCM when the LED load is small (e.g. 1 percent dimming) and avoids unwanted bus voltage ripples due to the limited minimum on-time of the IC controller.
- For the maximum output power, bigger PFC inductance has longer on-time and lower switching frequency. It must be guaranteed that these two parameters are still within the limit of ICL5102.
- Bigger PFC inductance leads to bigger choke size and more winding turns, which cause more winding loss. In contrast, smaller inductance has a smaller size and fewer winding turns but higher frequency, which could lead to more switching loss.

In the reference design, L_{PFC} = 0.36 mH is chosen to avoid magnetic saturation in all worst cases like start-up and load transient. After the PFC choke inductance is fixed, the choke relevant parameters can be calculated as follows with the assumption of the boundary conduction mode operation.

Maximum input current (RMS) happens at minimum AC input and maximum output power:

$$I_{in\_rms\_max} = \frac{P_{O\_PFC\_max}}{V_{IN\_BO} * \eta_{PFC}} = 2.12\ A$$

Maximum input peak current:

$$I_{in\_pk\_max} = \sqrt{2} * I_{in\_rms\_max} = 3.0A$$

Maximum inductor peak current:

$$I_{L,pk\_PFC\_max} = 2 * I_{in\_pk\_max} = 6.0\ A$$

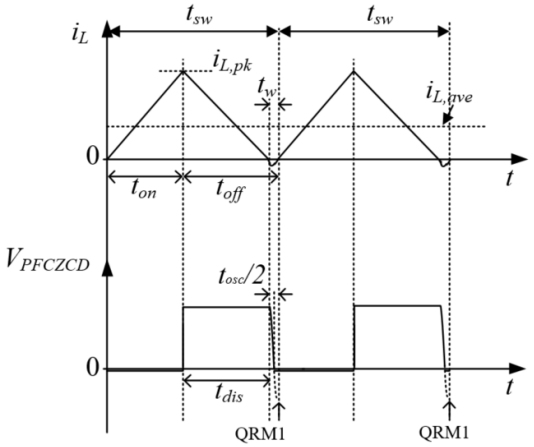


Figure 5 Boost inductor current waveform in a switching cycle

PFC boost converter design

According to **Figure 5**, other important parameters of the PFC boost converter can be calculated as follows.

Maximum on-time:

$$t_{on\_max} = \frac{L_{PFC} * I_{L,pk\_PFC\_max}}{\sqrt{2} * V_{IN\_BO}} = 21.5\ \mu s$$

If T_{osc} = 1.5 µs is assumed, off-time at minimum AC input and maximum output power:

$$t_{off} = \frac{L_{PFC} * I_{L,pk\_PFC\_max}}{V_{bus} - \sqrt{2} * V_{IN\_BO}} + 0.5 * T_{osc} = 6.92\ \mu s$$

The lowest frequency for maximum output power of PFC converter in QRM:

$$f_{PFC\_min} = \frac{1}{t_{on\_max} + t_{off}} = 35.2\ kHz$$

Maximum current (RMS) through the PFC inductor during on-time:

$$I_{L,PFC\_on\_rms\_max} = I_{L,pk\_PFC\_max} * \sqrt{\frac{1}{3} * t_{on\_max} * f_{PFC\_min}} = 3.0\ A$$

Maximum current (RMS) through the PFC inductor during off-time:

$$I_{L,PFC\_off\_rms\_max} = I_{L,pk\_PFC\_max} * \sqrt{\frac{1}{3} * t_{off} * f_{PFC\_min}} = 1.71\ A$$

Thus the maximum PFC inductor current (RMS):

$$I_{L\_rms\_max} = \sqrt{I^2_{L,PFC\_on\_rms\_max} + I^2_{L,PFC\_off\_rms\_max}} = 3.45\ A$$

To realize the ZCD of the inductor current for the CrCM switching, an additional auxiliary winding is introduced in the PFC inductor. The turns ratio of the main and auxiliary winding decides the amplitude of the oscillation at the PFCZCD pin. So that the comparator for the ZCD works correctly, the oscillation amplitude at the PFCZCD pin for ZCD must exceed $V_{PFCZCDTHRH\_max}$ = 1.6 V. The turns ratio is calculated as:

$$\frac{N_{p\_PFC}}{N_{a\_PFC}} < \frac{V_{bus} - V_{in\_pk\_max}}{1.6V} = 11.68$$

The important parameters of the PFC boost inductor are summarized in **Table 4**.

Table 4 PFC boost inductor design parameters

PFC boost converter			
Parameter	**Symbol**	**Value**	**Unit**
Main inductance of the PFC boost inductor	L_{PFC}	360	µH
Minimum switching frequency in QRM	f_{PFC_min}	35.2	kHz
Maximum inductor peak current	I_{L,pk_PFC_max}	6	A
Maximum input current (RMS)	$I_{in\_rms\_max}$	2.12	A
Maximum input peak current	$I_{in\_pk\_max}$	3	A
Maximum inductor current (RMS)	$I_{L\_rms\_max}$	3.45	A
Maximum on-time	t_{on_max}	21.5	µs
Turns ratio of primary to auxiliary winding	N_{p_PFC}/N_{a_PFC}	9:1	–

Based on the calculated specifications above, the inductor can be constructed according to different design requirements such as size, power efficiency and temperature by selecting different bobbin and core materials. In order to avoid core saturation and achieve an optimized core loss, the flux density B_{max} is recommended not to exceed 0.3.

PFC boost converter design

In the ICL5102 130 W CC reference design, the PFC boost inductor is constructed by Würth Electronik under part no. **750343180** as a design example. The specification sheet is given as shown in **Figure 6** and **Table 5**.

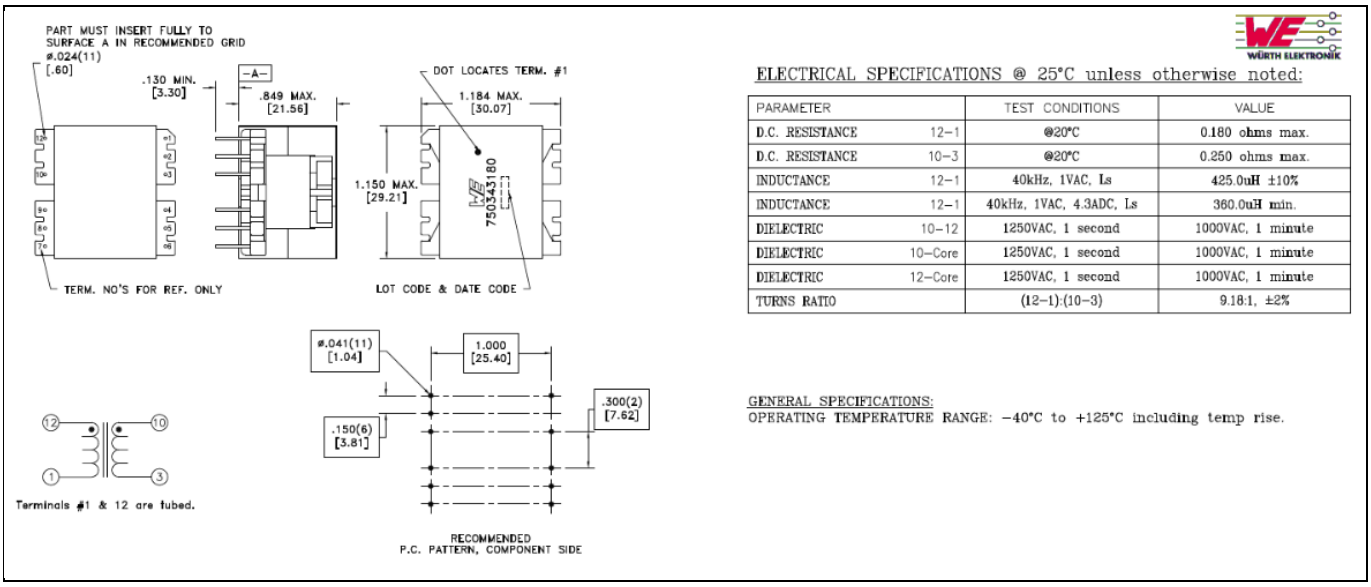


Figure 6 Würth PFC inductor 750343180

Table 5 Parameters of Würth PFC inductor 750343180

Parameter	Value	Unit
Inductance	360	µH
Bobbin	PQ2620	–
Core material	TP4A or DMR44	–
Turns ratio of primary to auxiliary winding	9.18	–
DC resistance primary winding	0.18	Ω
DC resistance auxiliary winding	0.25	Ω
Saturation current	7.2	A

3.2 Input bridge rectifier

The bridge rectifier usually has the highest semiconductor power loss in the LED driver. The following considerations must be taken into account for a proper selection.

- Maximum recurrent peak reverse voltage/DC blocking voltage:

Must be chosen to be higher than the peak of the maximum input voltage with at least 20 percent margin:

$$V_{BR\_RRM} > 1.2 * V_{IN\_pk\_max} = 518\,V$$

- Maximum rectified average forward current:

The maximum rectified average forward current is calculated as:

$$I_{BR\_avg\_max} = \frac{2\sqrt{2}}{\pi} * \frac{P_{O\_PFC\_max}}{V_{IN_{BO}} * \eta_{PFC}} = 1.91\,\mathrm{A}$$

PFC boost converter design

Using a bridge rectifier with a higher rated current can reduce the forward voltage drop, which reduces the total power dissipation at a small incremental cost.

- Peak surge forward current

The peak surge forward current must be chosen to cover the maximum surge current requirement.

- Forward voltage

This is directly related to the power efficiency; the forward voltage should be chosen to be as small as possible.

- Maximum power loss

The total maximum power loss is calculated using the maximum average input current flowing through two of the bridge rectifying diodes if the forward voltage is assumed as 1 V:

$$P_{BR\_loss\_max} = I_{BR\_avg\_max} * 2 * V_{BR\_F} = 3.83\ \text{W}$$

- Maximum diode temperature rise without heatsink

Assuming the thermal resistance of the bridge rectifier is R_{TH_JA} = 21°C/W (e.g. according to GBU80M), the maximum temperature rise of the diode without heatsink can be calculated as:

$$\Delta T_{BR\_max} = R_{TH\_JA} * P_{BR\_loss\_max} = 80.43\ °\text{C}$$

The important parameters for the bridge rectifier used in the 130 W reference design are summarized in **Table 6**.

Table 6 Bridge rectifier design parameters

Parameter	Symbol	Value	Unit
Maximum reverse voltage	V_{BR_RRM}	1000	V
Average rectified forward current	I_{BR_avg}	8	A
Forward voltage	V_{BR_F}	1	V

3.3 PFC boost diode

The selection of the boost diode is a major decision in the PFC boost converter design, and it is related to the converter efficiency. The following considerations should be taken into account:

- Reverse breakdown voltage

It must be chosen to be higher than the bus voltage with at least 20 percent margin:

$$V_{RRM\_D\_PFC} > 1.2 * V_{bus\_OVP1} = 594\ V$$

A 600 V diode is suitable here in the 130 W reference design.

- Average rectified forward current

It must be higher than the maximum average PFC diode current, calculated as follows:

$$I_{D\_PFC\_avg\_max} = \frac{\sqrt{3}}{2} * I_{L,PFC\_off\_rms\_max} = 1.05\ A$$

Using a diode with high current capability will benefit the power efficiency.

PFC boost converter design

- Forward voltage

This is directly related to the power efficiency. So the forward voltage should be chosen to be as small as possible.

- Reverse recovery time

As the PFC boost converter is controlled by ICL5102 in the QRM+DCM mode, the PFC boost diode current goes back to zero while the PFC MOSFET turns on. So there is no current commutation between the PFC diode and MOSFET and thus no switching loss by reverse recovery. It is not necessary to choose an ultra-fast diode.

- Power loss

The only power loss that should be considered is conduction loss. With a forward voltage of 0.5 V assumed, the diode conduction loss can be calculated as follows:

$$P_{loss\_D\_PFC} = I_{D\_PFC\_avg\_max} * V_{F\_D\_PFC} = 0.525\ \text{W}$$

- Thermal characteristics

With the thermal resistance of the diode $R_{D\_PFC\_TH\_JA}$ and ambient temperature T_A, the PFC diode temperature without heatsink is calculated as:

$$T_{D\_PFC} = P_{loss\_D\_PFC} * R_{D\_PFC\_TH\_JA} + T_A$$

The important parameters for the boost diode used in the 130 W reference design are summarized in **Table 7**.

Table 7 Boost diode design parameters

Parameter	Symbol	Value	Unit
Maximum reverse voltage	$V_{RRM\_D\_PFC}$	600	V
Average rectified forward current	$I_{F\_D\_PFC}$	4	A
Forward voltage	$V_{F\_D\_PFC}$	0.5	V

3.3.1 PFC power MOSFET

The selection of the PFC power MOSFET is based mainly on the consideration of the drain-source breakdown voltage and the power dissipation.

- Drain-source breakdown voltage

According to the operating bus voltage, the breakdown voltage should be chosen as:

$$V_{(BR)DSS} > 1.2 * V_{bus\_OVP1} = 594\ V$$

A 600 V MOSFET would be suitable; for applications with enhanced surge protection, higher breakdown voltage should be selected.

- Continuous drain current

The maximum continuous drain current is calculated as:

$$I_{D\_MOS\_PFC\_max} = 3 * I_{L,PFC\_on\_rms\_max} = 6.33\ A$$

- Conduction loss

PFC boost converter design

These losses are frequency independent and do not scale significantly with frequency. This is calculated as follows:

$$P_{con\_loss\_MOS\_PFC} = \left(\frac{I_{L,PFC\_on\_rms\_max}}{2}\right)^2 * R_{DS(ON)}$$

- Turn-on transition loss

As the converter works in QRM+DCM mode, the turn-on transition loss caused by the magnetizing current can be ignored because the current rises from zero when a switching cycle starts. But to discharge the parasitic capacitors like C_{oss} (MOSFET output capacitance) and C_{DS_total} (total parasitic drain-source equivalent capacitance) through the MOSFET channel can cause significant turn-on transition loss. These losses occur every switching cycle and are thus frequency dependent.

- E_{oss} and $^1/_2 \cdot C_{DS\_total} \cdot V^2$ loss

The energy stored in C_{DS_total} and C_{oss} at the time of turn-on must be dissipated in the MOSFET channel and CS resistor during the turn-on transition. These losses are fundamentally a function of the square of the voltage across it, and can be very significant during high-line conditions. They occur every switching cycle and are thus frequency dependent. To simplify the calculation, we assumed that the switching loss is approximately half of the conduction loss:

$$P_{sw\_loss\_MOS\_PFC} = \frac{1}{2} * P_{con\_loss\_MOS\_PFC}$$

- Gate driver loss

These losses also scale linearly with frequency, but are generally quite a small contribution to the overall losses (at switching frequencies of below 100 kHz) and depend almost exclusively on the MOSFET Q_g (total gate charge). The gate driver power is typically dissipated in the external gate resistor and the gate driver itself and thus does not need to be considered in the thermal calculation of the MOSFET.

In the 130 W reference design, two 600 V Infineon MOSFETs IPD60R400CE of CE family are used in parallel to minimize the power dissipation. With the $R_{DS(ON)}$ of 200 mΩ, the maximum total loss of each MOSFET is calculated as below:

$$P_{loss\_MOS\_PFC\_max} = 0.5 * (P_{sw\_loss\_MOS\_PFC} + P_{con\_loss\_MOS\_PFC}) = 0.75 * P_{con\_loss\_MOS\_PFC} = 0.96\,W$$

The maximum temperature rise of each MOSFET without heatsink is:

$$\Delta T_{MOS\_PFC\_max} = R_{TH\_JA} * P_{loss\_MOS\_PFC\_max} = 76.8\,°C$$

The important parameters for PFC MOSFETs are summarized in **Table 8**.

Table 8 PFC MOSFET design parameters

Parameter	Symbol	Value	Unit
Breakdown voltage	$V_{BR\_DSS\_PFC}$	650	V
MOSFET on-resistance	$R_{DS(ON)}$	400//400 = 200	mΩ
Maximum PFC MOSFET conduction loss	$P_{loss\_MOS\_PFC\_con}$	1.93	W

PFC boost converter design

3.3.2 PFC CS

To design the PFC CS shunt resistor, the following condition must be complied with:

$$I_{L,pk\_PFC\_max} * R_{CS_{PFC}} < V_{OCP1\_PFC\_min} = 0.95\ V$$

and:

$$R_{CS\_PFC} < \frac{0.95V}{I_{L,pk\_PFC\_max}} = 0.22\ \Omega$$

The value of the CS resistor is chosen to be 0.2 Ω with five resistors of 1 Ω connected in parallel. This splits the power dissipation and reduces the thermal stress. The maximum power loss of each shunt resistor is:

$$P_{loss\_shunt\_PFC\_max} = 0.2 * I^2_{L,PFC\_on\_rms\_max} * R_{CS\_PFC} = 0.2 * 2.11A^2 * 0.2 = 0.18\ W$$

This should be considered while selecting the proper shunt resistor type.

The important design parameters for bus voltage sensing are summarized in **Table 9**.

Table 9 PFC CS and ZCD design parameters

Parameter	Symbol	Value	Unit
PFC OCP1 maximum operating range	$V_{OCP1\_PFC\_min}$	0.95	V
PFC CS resistor	R_{CS_PFC}	0.2	Ω

3.3.3 PFC bus voltage sense

As shown in **Figure 7**, the bus voltage is measured at the PFCVS pin of ICL5102 through a resistor divider. This measurement is used as the input of the PFC output voltage regulator to generate the PWM control signal for the PFC MOSFET and offers protection functions for the PFC boost converter. It is strongly recommended to add a filter capacitor near the VS pin to filter the switching noise in order to get a precise and stable measurement result. The VS pin has a very low leakage current so the intolerance can be ignored.

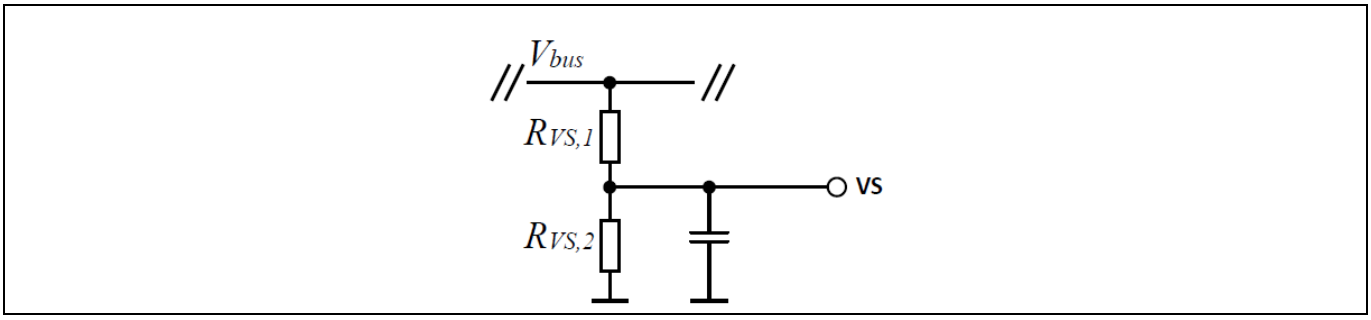


Figure 7 Bus voltage measurement

The calculation of the resistor divider is given as follows when the V_{bus} = 450 V is mapped to $V_{PFCVSREF}$ = 2.5 V.

$$\frac{R_{VS1\_PFC}}{R_{VS2\_PFC}} = \frac{V_{bus} - V_{PFCVSREF}}{V_{PFCVSREF}} = 179$$

To reduce the inaccuracy caused by the resistor divider, it is necessary to select the bus voltage sensing resistors with a tolerance of 1 percent or less. In the 130 W reference design, to reduce the voltage stress, the

PFC boost converter design

upper resistor R_{VS1_PFC} consists of three resistors each of 1.5 MΩ, and the lower resistor R_{VS1_PFC} is selected as 24.9 kΩ.

Note: In order to reduce the switching noise coupled in the bus voltage sense signal, a filter capacitor of 1 nF is strongly recommended to be placed directly near the PFCVS pin.

The important design parameters for bus voltage sensing are summarized in **Table 10**.

Table 10 Bus voltage sensing design parameters

Parameter	Symbol	Value	Unit
Nominal PFC boost converter output voltage	V_{bus}	450	V
ICL5102 PFCVS pin reference voltage	V_{REF}	2.5	V
Bus voltage sensing divider upper resistor	R_{VS1_PFC}	1.5 x 3	MΩ
Bus voltage sensing divider lower resistor	R_{VS2_PFC}	24.9	kΩ
Bus voltage sensing filter capacitor	C_{VS}	1	nF

3.3.4 PFC output capacitor

The PFC bus capacitor can be calculated with the following formula if the ESR of the capacitor is small enough to be neglected and the peak-to-peak voltage ripple is selected as 20 V. Please note that the tolerance of 20 percent of the capacitance needs to be taken into account as well.

$$C_{bus} = \frac{I_{out\_PFC\_max}}{2 * \pi * f_{line\_min} * V_{bus\_ripple\_pp}} * 1.2 = 46\ uF$$

With:

$$I_{out\_PFC\_max} = \frac{P_{O\_PFC\_max}}{V_{bus}} = 0.24\ A$$

Regarding the voltage rating with consideration for the OVP threshold, a 500 V capacitor is necessary. The ESR of the capacitor should be selected to be as small as possible, and the allowed maximum ripple current should have enough margin. In the 130 W reference design, one 500 V capacitor of 56 µF with low ESR is selected.

The important parameters for the bus capacitor selection are summarized in the following table.

Table 11 Bus capacitors design parameters

Parameter	Symbol	Value	Unit
Nominal PFC boost converter output voltage	V_{bus}	450	V
Maximum PFC boost converter output power	$P_{out\_PFC\_max}$	130	W
Bus voltage ripple (peak to peak)	$V_{bus\_ripple\_pp}$	20	V
AC input line frequency	f_{line}	45 ~ 66	Hz
PFC bus capacitor	C_{bus}	56	µF

3.3.5 Input voltage sensing and brown-in/out

The rectified input voltage is sensed through an external resistor divider which consists of R_{BO1} and R_{BO2} at the BO pin as shown in **Figure 8**. This input voltage sensing function enables the AC input brown-in/out to provide the under-voltage protection for the LED driver.

PFC boost converter design

It is strongly recommended to add a high-voltage capacitor C_{BO1} directly after the rectification diodes D_{BO1} and D_{BO1} so that the peak value of AC input is measured, which is not dependent on the load condition. Without the capacitor $C_{BO1,}$ the equalized AC input mean value is sensed at the BO pin, which shifts according to the load. Thus the brown-in/out threshold is different under different load conditions.

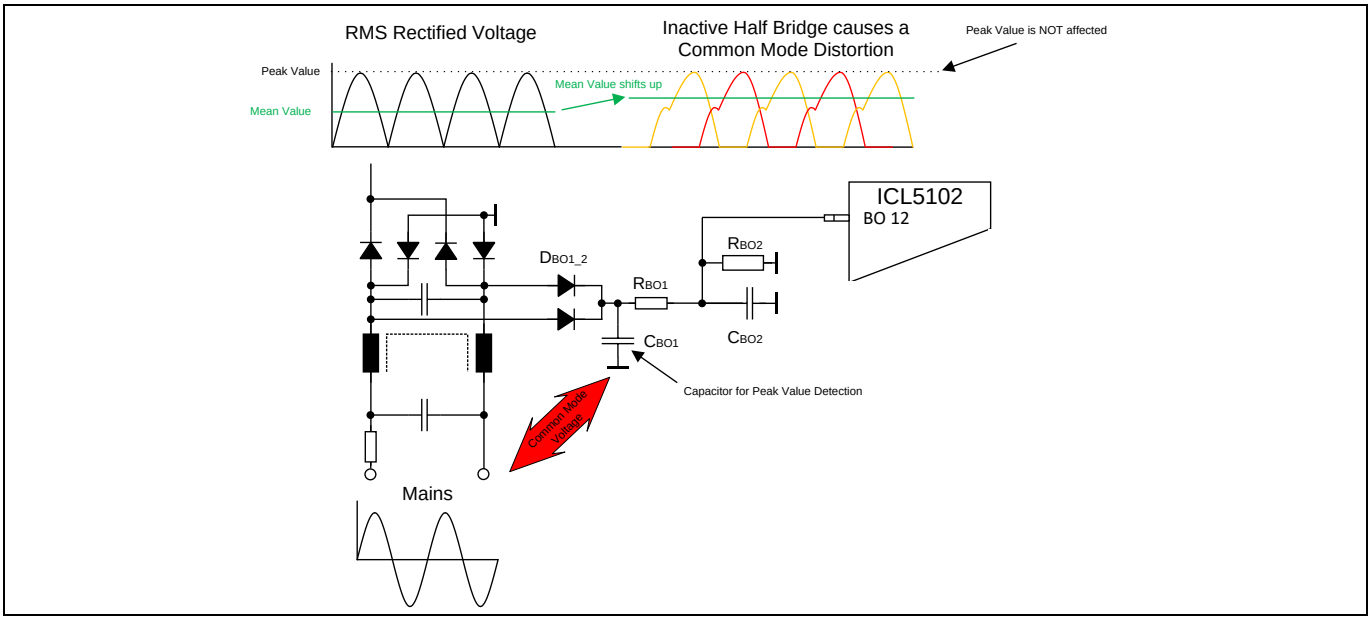


Figure 8 Input voltage sensing

At the initial start-up (AC plug-in), the voltage at the BO pin must typically exceed V_{BO} = 1.4 V to enable brown-in. During normal operation, brown-out is detected when the voltage at the BO pin drops below typically V_{BO} = 1.2 V for longer than 50 ms. ICL5102 will stop both PFC and HB switching and enter auto-restart. Normal system operation recovers if the voltage at the BO pin is above 1.4 V again.

The ratio of the divider is calculated as follows:

$$\frac{R_{BO1}}{R_{BO2}} = \frac{1.414 * V_{IN\_BO} - V_{D\_BO} - V_{BO\_out\_min}}{V_{BO\_out\_min}} = 86$$

With $V_{BO\_out\_min}$ = 1.14 V from ICL5102 datasheet Table 16, brown-in threshold V_{IN_BI} = 71 V and V_{D_BO} = 0.7 V.

In the ICL5102 130 W CC reference design, R_{BO1} = 6600 kΩ is selected. To reduce the voltage and power stress of the resistor, it is strongly recommended to split it into three 1206 resistors of 2200 kΩ each. To improve the accuracy of the measurement, resistors with tolerance of less than 1 percent should be selected. According to the divider ratio calculated previously, R_{BO2} = 76.8 kΩ is selected.

The brown-in voltage can be calculated after R_{BO1} and R_{BO2} are fixed:

$$V_{IN\_BI} = \frac{86 * V_{BO\_in\_max} + V_{D\_BO} + V_{BO\_in\_max}}{1.414} = 90\ V$$

The important design parameters for input voltage sensing are summarized in Table 12.

Table 12 Input voltage sensing design parameters

Parameter	Symbol	Value	Unit
Brown-out threshold	V_{IN_BO}	71	V_{RMS}
Brown-in threshold	V_{IN_BI}	90	V_{RMS}

PFC boost converter design

Parameter	Symbol	Value	Unit
BO pin divider upper resistor	R_{BO1}	6600	k Ω
BO pin divider lower resistor	R_{BO2}	76.8	k Ω

3.3.6 THD optimization

The input AC current becomes most distorted in the area when zero-crossing of AC input voltage occurs. In order to ensure the sinusoidal current waveform in this area, the ICL5102 extends the PFC on-time dynamically up to twice PFC maximum on-time according to the instantaneous value of the input voltage amplitude. The detection of AC input voltage zero-crossing is realized through the PFC auxiliary winding. When the voltage across the PFC auxiliary winding after the PFC MOSFET turns on reaches the minimum value, AC zero-crossing is detected. The concept of THD correction is shown in **Figure 9**.

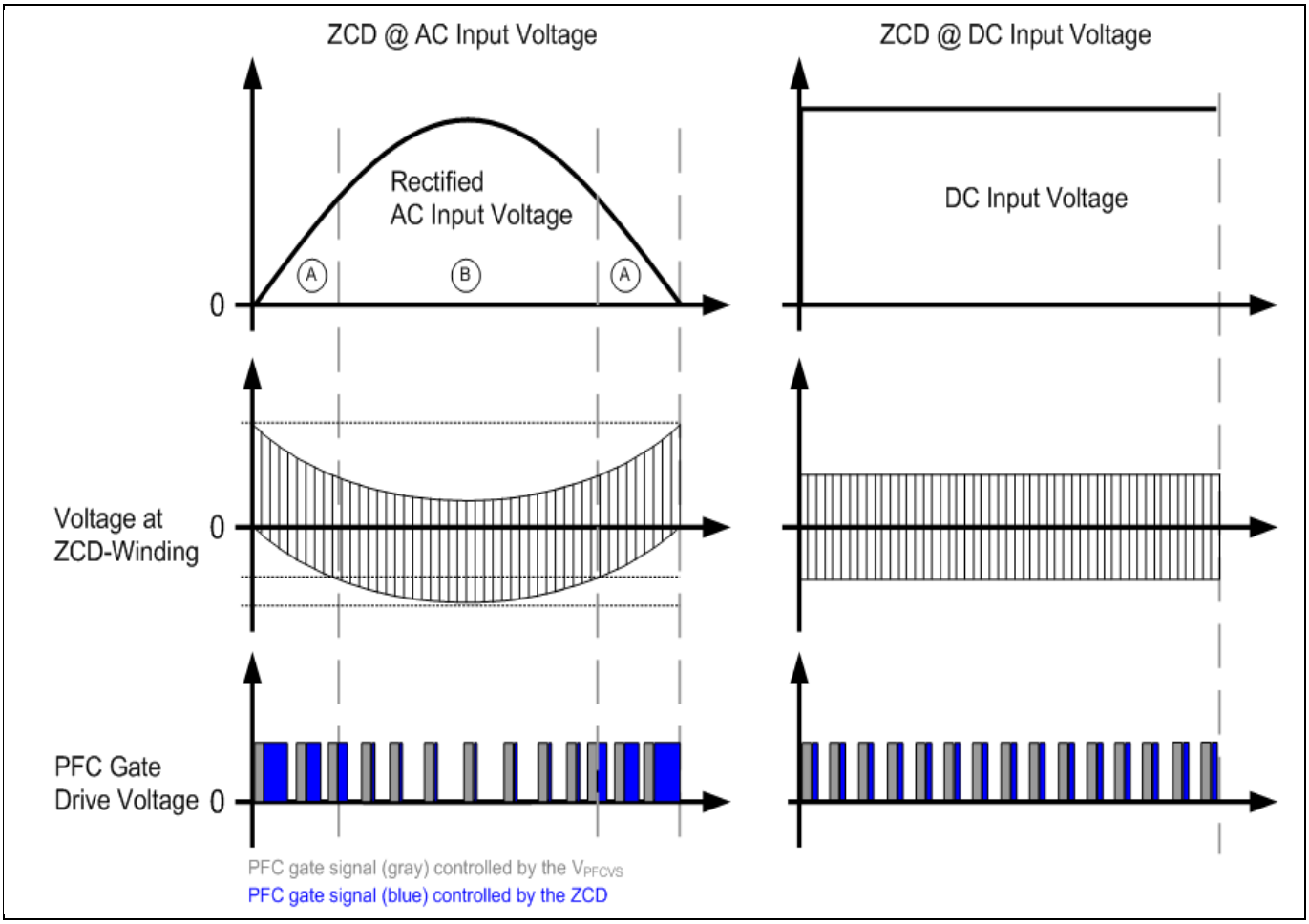


Figure 9 PFC THD correction

To adjust the on-time enlargement, an external resistor between the PFC choke auxiliary winding and PFCZCD pin is required, as shown in **Figure 10**. When the PFC MOSFET turns on, the input voltage is measured according to the current flowing out of the PFCZCD pin. A smaller current means AC input is in the area A (see **Figure 9**), which requires an enlargement of the PFC on-time. Generally, the smaller the current flowing out of the PFCZCD pin is, the bigger the PFC on-time enlargement. The value of the resistor should be selected so that the current is limited between I_{ZCD} = 500 µA ~ 1.2 mA.

PFC boost converter design

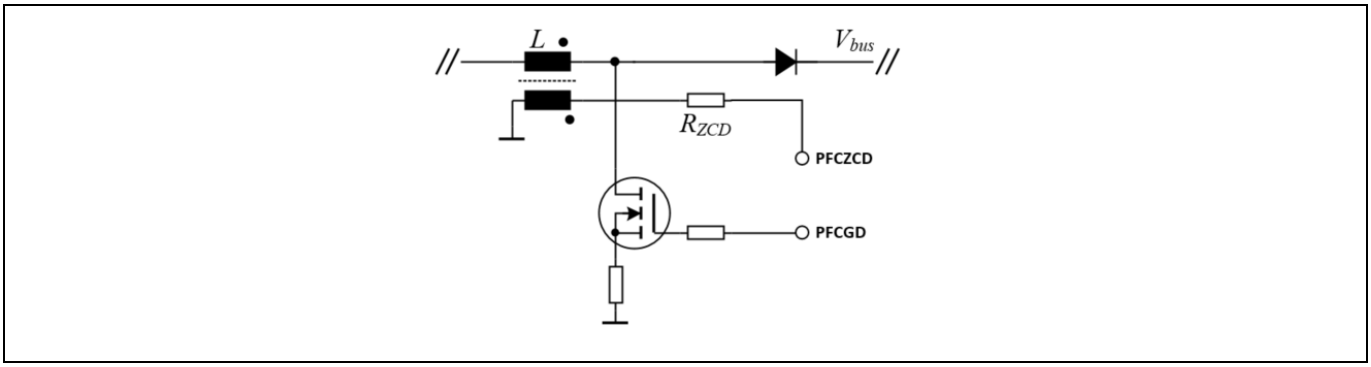


Figure 10 PFC THD optimization via an external resistor

The following formula shows the calculation of the connected ZCD resistor value with the assumption that I_{ZCD_max} = 1.2 mA flows out of the PFCZCD pin at maximum AC-line input:

$$R_{ZCD} = \frac{1.414 * V_{IN\_max}}{\frac{N_{p\_PFC}}{N_{a\_PFC}} * I_{ZCD\_max}} = 39.9\ \text{k}\Omega$$

The THD performance can be optimized by varying the resistor value around 40 kΩ.

Note: There is a trade-off between improvement in PF and reduction of THD. However, trying to achieve harmonic-free current may bring harmonics into the voltage and thus a lower PF. On the other hand, any attempt at PF unity by the usual methods will result in a non-sinusoidal current, which increases THD.

4 HB LLC resonant converter design

A resonant HB converter has many advantages, such as high efficiency, high power density and low EMI. ICL5102 provides independent control of the resonant HB converter (e.g. LLC or LCC topology) for Constant Voltage (CV) or CC output. It supports both fixed and variable switching frequency control.

The following sections provide the HB converter design guidelines using ICL5102 based on LLC topology with a secondary-side regulated CC control.

The design specification for the 130 W LLC CC reference design is given in the following table.

Table 13 LLC converter design specification

Parameter	Symbol	Value	Unit
Minimum DC input voltage	V_{bus_min}	400	V
Nominal DC input voltage	V_{bus}	450	V
Maximum DC input voltage	V_{bus_max}	490	V
Maximum LLC converter output power	P_{O_max}	130	W
Estimated LLC converter power efficiency	η_{FB}	Less than 96	%
LLC converter maximum output voltage	V_{out_max}	76	V
LLC converter maximum output voltage	V_{out_min}	38	V
LLC converter maximum output current	I_{out_max}	1750	mA
LLC converter output current	I_{out_min}	75	mA
LLC converter resonant frequency	f_r	100	kHz
Primary-side switching bridge topology	–	HB	–
Secondary-side rectifier topology	–	Full-Bridge (FB) rectifier	–

On the primary side, the HB is used as a switching bridge because ICL5102 is a HB controller. On the secondary side, a FB rectifier is preferred due to the high voltage and low current output characteristics. The design target is to achieve Zero Voltage Switching (ZVS) on the primary side and Zero Current Switching (ZCS) on the secondary side for the entire load range for optimized power efficiency and low-EMI performance.

4.1 LLC resonant tank design

Due to the non-linear nature of the resonant LLC topology, the following calculations of resonant tank are based on the First Harmonic Approximation (FHA). Furthermore, to simplify the design of the LLC transformer, an external resonant inductor is used.

4.1.1 Transformer turns ratio

According to the characteristics of the HB LLC converter with FB rectifier, the converter gain can be calculated as:

$$LLC\ Converter\ Gain = \frac{V_{out} + 2 * V_{F\_D}}{V_{bus}} = 0.5 * Resonant\ Tank\ Gain * Transformer\ Turns\ Ratio\ (N_s/N_p)$$

When the LLC converter operates at a frequency equal to the resonant frequency, the resonant tank gain is 1 when using an individual core for the series inductor. So the transformer turns ratio can be expressed as:

$$n = \frac{N_p}{N_s} = \frac{V_{bus}}{2 * (V_{out} + 2 * V_{F\_D})}$$

HB LLC resonant converter design

It is recommended to operate the LLC tank below resonant frequency (boost mode operation) for CC output. To ensure this for the widest load range, the minimum output voltage with maximum bus voltage as a working point is selected as LLC resonant operation to decide the transformer turns ratio:

$$n = \frac{N_p}{N_s} = \frac{V_{bus\_max}}{2 * (V_{out\_min} + 2 * V_{F\_D})} = 6.22$$

4.1.2 LLC resonant tank gain

After the transformer turns ratio is defined, the resonant tank voltage gain M_{LLC} can be summarized as:

- Unit gain M_{LLC} = 1 for maximum bus voltage and lowest output voltage
- Minimum gain M_{LLC_min} = M_{LLC} = 1
- Maximum gain M_{LLC_max} for minimum bus voltage and highest output voltage

$$M_{LLC\_max} = \frac{n * 2 * (V_{out\_max} + 2 * V_{F_D})}{V_{bus\_min}} = 2.4$$

4.1.3 Selecting the m value

The ratio of total primary inductance to resonant inductance is defined as the m value as follows:

$$m = \frac{L_r + L_m}{L_r}$$

Lower values of m can achieve higher boost gain and provide more flexible control and regulation, which is valuable in applications with wide input voltage range. Nevertheless, low values of m mean smaller magnetizing inductance L_m, and therefore higher magnetizing peak-to-peak current ripple, causing increased circulating energy and conduction losses.

For the 130 W LLC CC reference design, a reasonable initial value of m = 8 is first selected. Later, by few iterations, the m value can still be optimized as long as the maximum gain requirement for all load conditions can still be achieved.

4.1.4 Selecting the maximum quality factor value Q_{max}

Quality factor is dependent on the load and defined as:

$$Q = \frac{\sqrt{\frac{L_r}{C_r}}}{R_{ac}}$$

Heavy load conditions operate at high Q values while lighter loads have lower Q values. It is important to set a value for the Q_{max} associated with the maximum load point.

By using the ICL5102 LLC calculation tool, the analysis can be simplified. The peak gains are obtained and depicted in **Figure 11**, which shows how the peak gain (attainable maximum gain) varies with Q for different m values. It appears that higher peak gain can be obtained by reducing m or Q values. With a given resonant frequency and Q value, decreasing m means reducing the magnetizing inductance, which results in increased circulating current. There is a trade-off between the available gain range and conduction loss.

With the maximum gain M_{LLC_max} and defined m value, the maximum quality factor is achieved through the curve:

$$Q_{max} = 0.17$$

HB LLC resonant converter design

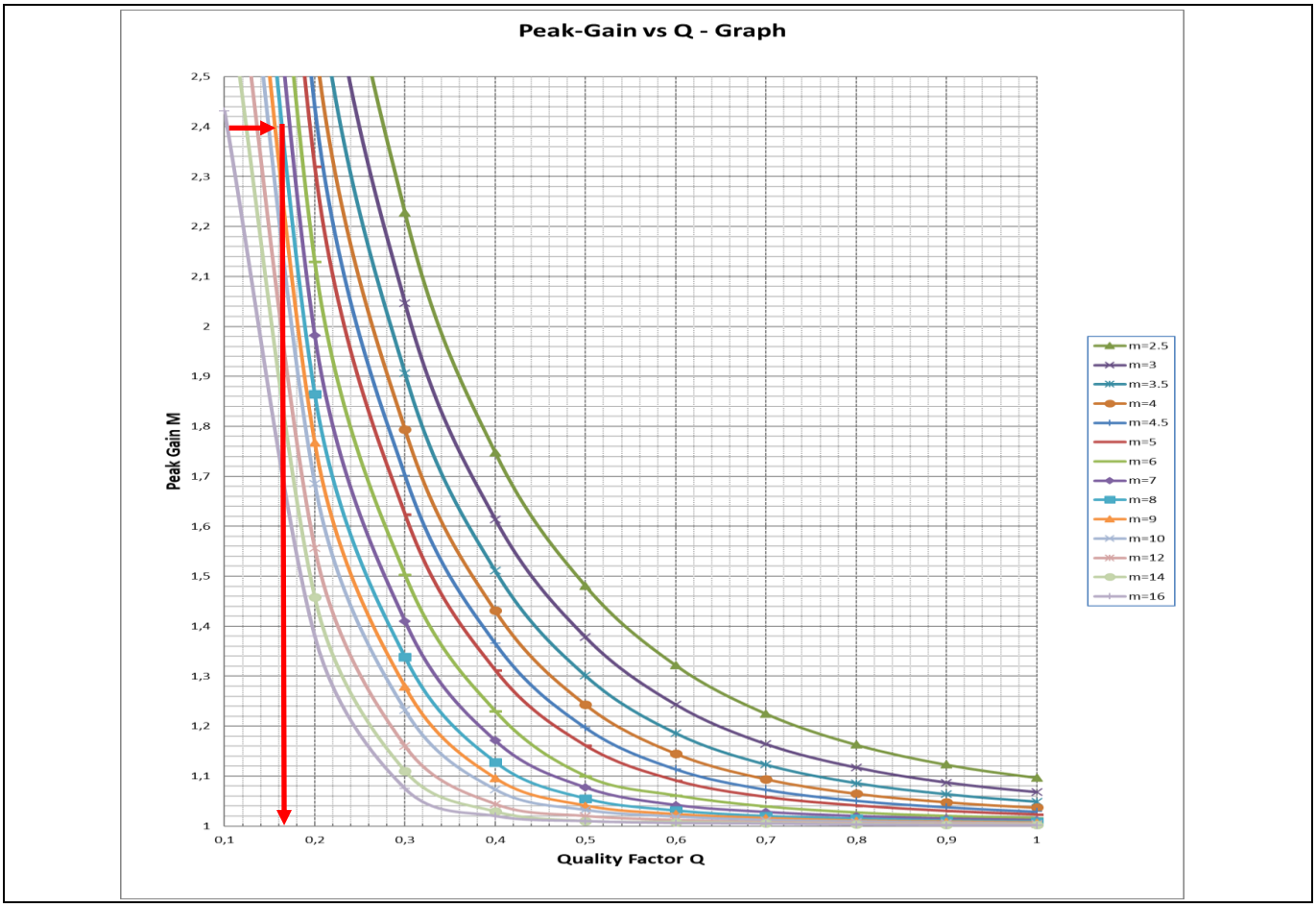


Figure 11 Peak gain vs. Q for different m values

4.1.5 Equivalent reflected resistance R_{ac}

FHA eliminates non-linearity and generates an equivalent resonant circuit, as shown in **Figure 12**. The input square wave voltage is replaced with first harmonic of the voltage square wave, while the secondary section with rectifier and load is replaced by equivalent load R_{ac}, which is calculated at full load as:

$$R_{ac} = \frac{8}{\pi^2} * n^2 * \frac{(V_{out\_max} + 2 * V_{F_D})}{I_{out\_max}} = 1388\,\Omega$$

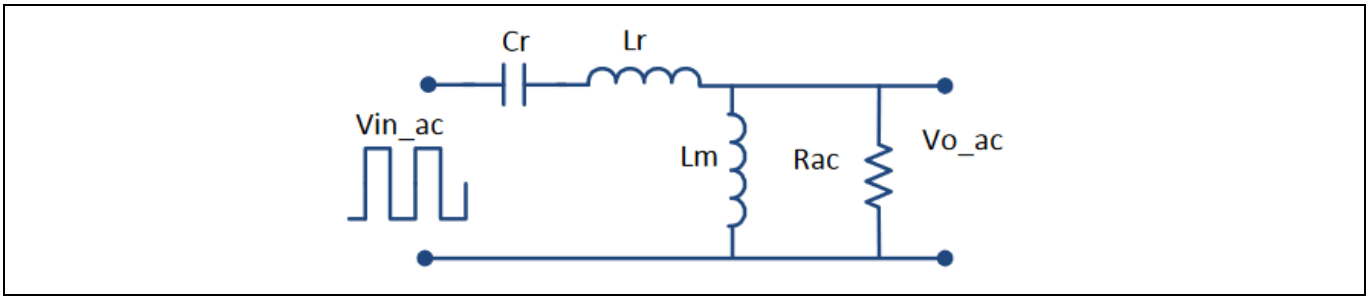


Figure 12 Equivalent resonant circuit

4.1.6 LLC resonant tank calculation

After the values of m, Q_{max} and R_{ac} are obtained, the value of the resonant capacitor can be calculated as:

HB LLC resonant converter design

$$C_r = \frac{1}{2 * \pi * Q_{max} * f_r * R_{ac}} = 6.74\ nF$$

To reduce the size of the resonant choke, a bigger resonant capacitor is selected for the 130 W reference design as:

$$C_r = 11.5\ nF$$

In order to reduce the start-up inrush current and the ripple current in normal operation through the resonant capacitor, one resonant capacitor can be split into two capacitors with standard values (one to the DC bus, the other to the ground). This helps to obtain the available capacitor values, too.

$$C_{r1} = 6.8\ nF$$

$$C_{r2} = 4.7\ nF$$

The inductance of the resonant choke is calculated as:

$$L_r = \frac{1}{4 * \pi^2 * C_r * f_r^2} = 220\ \mu H$$

The main inductance of the LLC transformer L_p is calculated as:

$$L_P = m * L_r = 1760\ \mu H$$

The magnetizing inductance L_m is calculated as:

$$L_m = L_P - L_r = 1540\ \mu H$$

The equivalent leakage inductance due to HB can be approximately estimated as:

$$L_{leak} = 0.5 * 8\% * L_m = 60\ \mu H$$

So the inductance of the separated resonant choke is calculated as:

$$L_{r\_ext} = L_r - L_{leak} = 160\ \mu H$$

4.1.7 Verification of the LLC resonant tank parameters

After all resonant tank parameters are decided, the working points of all corner cases must recalculated to ensure that the LLC resonant design target is still fulfilled.

Resonant frequency:

$$f_r = \frac{1}{2 * \pi * \sqrt{L_r * C_r}} = 100.1\ kHz$$

$$f_p = \frac{1}{2 * \pi * \sqrt{L_p * C_r}} = 35.3\ kHz$$

Quality factor:

$$Q = \frac{\sqrt{\frac{L_r}{C_r}}}{R_{ac}} = 0.1$$

Ratio of total primary inductance to resonant inductance:

$$m = \frac{L_r + L_m}{L_r} = 8$$

HB LLC resonant converter design

4.1.8 Achieving ZVS of the LLC converter

To ensure high power efficiency of the LLC converter, ZVS must be guaranteed for all input and load conditions. There are two main considerations to achieve ZVS.

- Operating regions of the LLC converter

As long as the converter switches at higher than resonant frequencies f_r, it always runs in ZVS mode. When the converter is switching between the resonant frequencies f_r and f_p, the load condition (Quality Factor, Q) determines whether the converter operates in ZVS mode or not. Under normal operating conditions, the LLC resonant converter operates at slightly higher than the resonant frequency f_r, which is optimal for high efficiency.

- MOSFET output capacitance

The resonant inductor current must be high enough to discharge the voltage of effective capacitance appearing in parallel with drain-sources of the power MOSFETs. A major part of this capacitance is the MOSFET output capacitance.

- Dead-time

Dead-time is required to completely discharge the effective capacitance appearing in parallel with drain-sources of the power MOSFETs to realize the ZVS operation in worst-case conditions applied to the converter.

The four most common waveforms and solutions for ZVS are shown in **Figure 13**, where modification of the circuit for optimal efficiency can be carried out according to the instructions given. **Figure 13** (a) illustrates the perfect ZVS diagram. In **Figure 13** (b), VDS cannot fully discharge to 0 V before the control signal V_{GS} is sent, thus it is recommended to reduce the magnetizing inductance value in the transformer, and slightly increase the dead-time or replace the power MOSFET with a higher $R_{DS(ON)}$. In **Figure 13** (c), the power MOSFET achieves ZVS, but the inductance current is insufficient to continually forward bias the body diode; thus it is recommended to slightly reduce dead-time. ZVS is achieved in **Figure 13** (d), but the excessive dead-time results in reduced efficiency; thus, it is recommended to reduce dead-time.

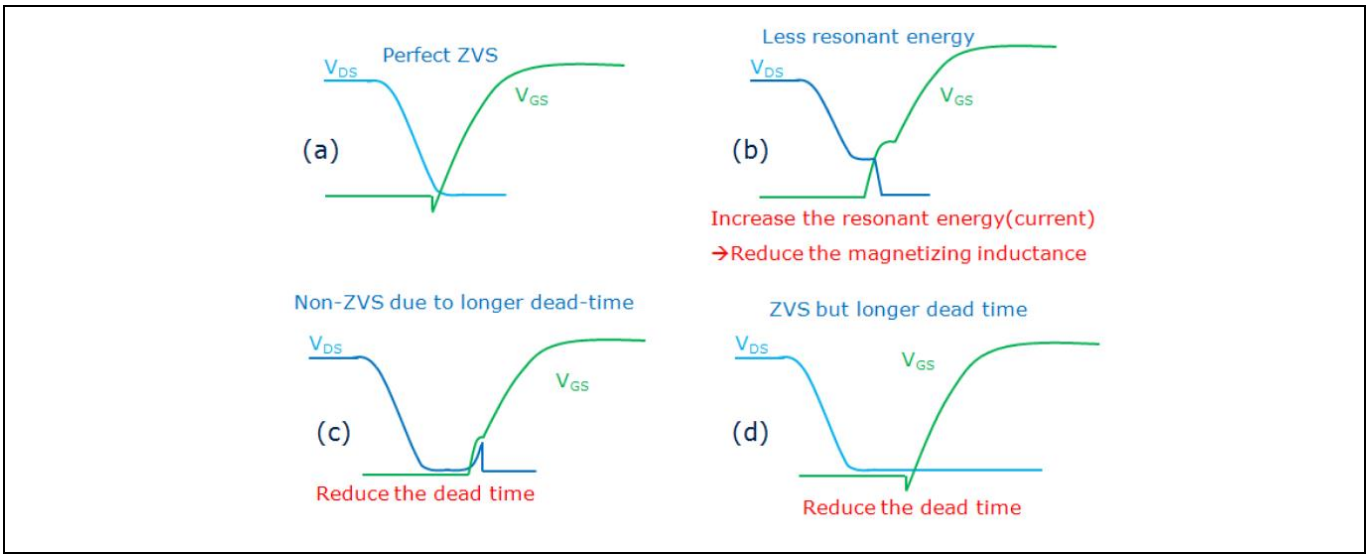


Figure 13 Waveforms for ZVS power MOSFET and solutions

4.2 LLC main transformer design

After the turns ratio and inductance of the main transformer have been confirmed, the design parameters of the transformer can be obtained.

HB LLC resonant converter design

With selected secondary turns:

$$N_s = 6\ turns$$

The primary turns can be calculated as:

$$N_p = n * N_s = 38.7\ \approx 40\ turns$$

In order to provide the V_{CC} power supply for ICL5102, an auxiliary winding is introduced. The turns ratio is calculated as if the V_{CC} = 15 V is assumed:

$$n_a = \frac{N_p}{N_a} = \frac{V_{bus}}{2 * (V_{cc} + V_{F\_D})} = 14.33$$

The auxiliary turns can be calculated as:

$$N_a = \frac{N_p}{n_a} = 2.79\ \approx 3\ turns$$

According to the following gain curves from different load conditions from the ICL5102 LLC design tool, the LLC converter minimum switching frequency at maximum gain can be obtained as:

$$f_{min} = 44.7\ kHz$$

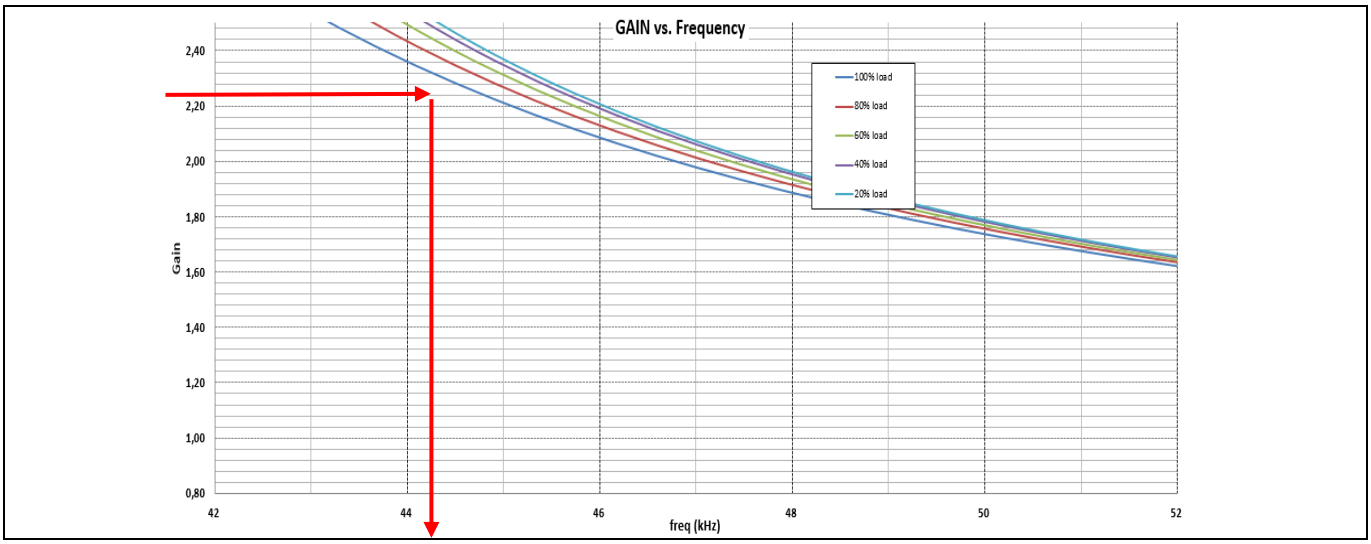


Figure 14 Gain vs. frequency for different load conditions

The RMS current on the primary side through the resonant tank can be calculated as:

$$I_{L\_p\_rms} = \frac{1}{\eta} * \sqrt{\left(\frac{\pi * I_{out\_max}}{2 * \sqrt{2} * n}\right)^2 + \left(\frac{n * (V_{out\_max} + 2 * V_{F\_D})}{4 * \sqrt{2} * f_r * (L_p - L_r)}\right)^2} = 0.73\ A$$

The primary peak current through the resonant tank can be calculated as:

$$I_{L\_p\_peak} = \sqrt{2} * I_{L\_p\_rms} = 1.03\ A$$

In the ICL5102 130 W CC reference design, the LLC external resonant inductor is constructed by Würth Electronik under part no. **750342805** as a design example. The specification sheet is given in **Table 14**.

Table 14 Parameters of Würth inductor 750342805

Parameter	Value	Unit
Inductance	160	µH

HB LLC resonant converter design

Parameter	Value	Unit
DC resistance	0.33	Ω
Saturation current	3.5	A

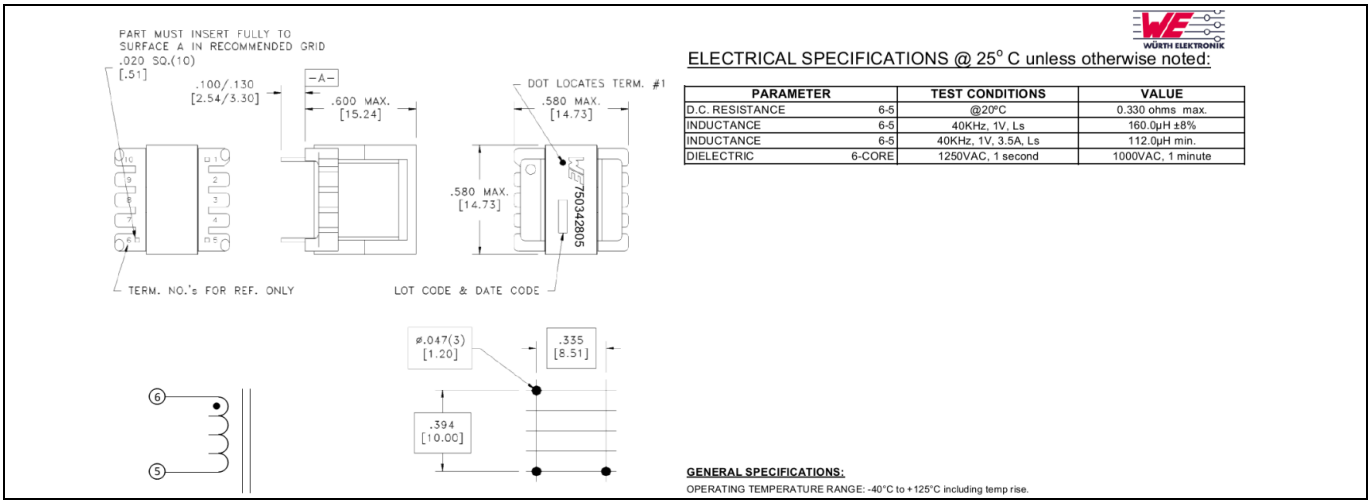


Figure 15 Würth inductor 750342805

The LLC main transformer is constructed by Würth Electronik under part no. **750342886** as a design example. The specification sheet is given in **Table 15**.

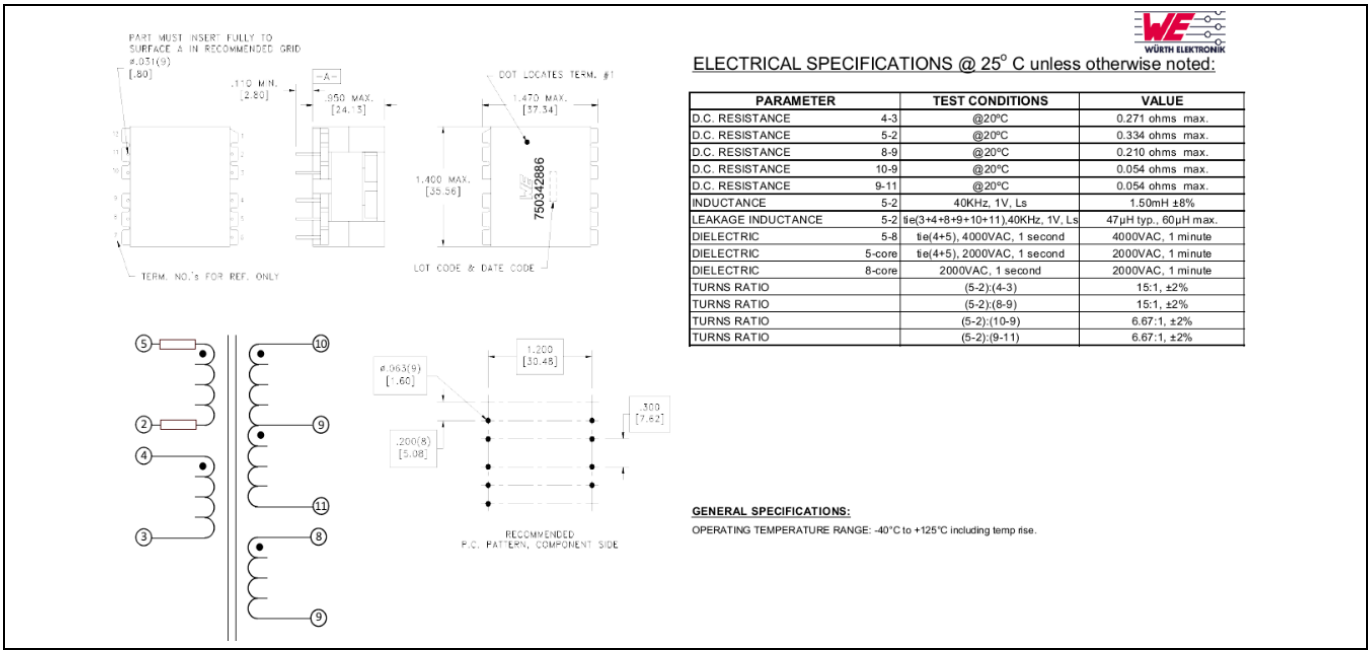


Figure 16 Würth inductor 750342886

Table 15 Parameters of Würth inductor 750342886

Parameter	Value	Unit
Main inductance	1500	µH
Leakage inductance	47	µH
Turns ratio N_p/N_s	6.67	–

HB LLC resonant converter design

Parameter	Value	Unit
Turns ratio N_p/N_a	15	–
DC resistance primary main winding	0.33	Ω
DC resistance secondary output winding	0.054	Ω
Saturation current	2	A

4.3 HB primary MOSFET selection

The selection of the HB primary power MOSFET is based mainly on the consideration of the drain-source breakdown voltage and the power dissipation.

- Drain-source breakdown voltage

According to the operating bus voltage, the breakdown voltage should be chosen as:

$$V_{(BR)DSS} > 1.2 * V_{bus\_OVP1} = 594\ V$$

A 600 V MOSFET would be suitable; for applications with enhanced surge protection, higher breakdown voltage should be selected.

- Conduction loss

These losses are frequency independent and do not scale significantly with frequency. This is calculated as follows:

$$P_{con\_loss\_MOS\_HB} = \left(I_{L\_p\_rms}\right)^2 * R_{DS(ON)} = 0.21\ W$$

In the 130 W reference design, two 600 V Infineon MOSFETs IPD60R400CE of CE family are used for HS and LS. With the $R_{DS(ON)}$ of 290mΩ, the maximum total loss of each MOSFETs is calculated as below:

$$P_{loss\_MOS\_HB\_max} = 1.5 * P_{con\_loss\_MOS\_PFC} = 0.315\ W$$

The maximum temperature rise of each MOSFET without heatsink is:

$$\Delta T_{MOS\_PFC\_max} = R_{TH\_JA} * P_{loss\_MOS\_HB\_max} = 25.2\ °C$$

The important parameters for HB MOSFETs are summarized in **Table 16**.

Table 16 PFC MOSFET design parameters

Parameter	Symbol	Value	Unit
Breakdown voltage	$V_{BR\_DSS\_HB}$	650	V
MOSFET on-resistance	$R_{DS(ON)}$	400	mΩ
Maximum PFC MOSFET conduction loss	$P_{loss\_MOS\_HB\_con}$	0.315	W

4.4 HB low-side CS resistor selection

The LS CS resistor is connected to the ICL5102 LSCS pin and the source of the HB LS MOSFET in order to protect the HB converter from over-current. The capacitive mode protection and the self-adaptive dead-time are also implemented with the CS function at the LSCS pin.

HB LLC resonant converter design

According to the datasheet, there are two over-current protection levels. The LS CS resistor value can be calculated according to the over-current protection level 1 as following:

$$R_{LSCS} < \frac{V_{HB\_OCP1\_min}}{I_{L\_p\_peak}} = 0.71\ \Omega$$

In the ICL5102 130 W reference design, R_{LSCS} = 0.5 Ω is chosen. In order to achieve the precise resistor value and minimize the power dissipation, the CS resistor is split to three parallel connected resistors of 1.2 Ω // 1.2 Ω // 1.5 Ω.

To enable the power limitation function in BM, a resistor R_{PL} between the LSCS pin and the CS resistor is required, as shown in **Figure 17**.

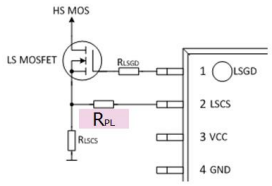


Figure 17 ICL5102 BM power limitation

In BM, during burst phase, the ICL5102 HB controller turns on with a switching frequency of f_{HB_BM} and steadily decreases it to f_{HB_max} to initialize a soft-on (see **ICL5102 datasheet**). After soft-on, the switching frequency continues to decrease to f_{HB_PL}, which is set by the resistor R_{PL}.

As shown in **Figure 18**, a reference threshold higher than 0 is set by the resistor R_{PL}. ICL5102 will integrate the area above and below this threshold according to the voltage at the LSCS pin. The HB switching frequency will adjust itself (by increasing and decreasing) until f_{HB_PL} is reached so that the integration of area B is the same as area A1 and area A2. By changing the R_{PL} resistor value, the switching frequency f_{HB_PL} is set and thus the transferred power in BM. The recommended value of the resistor R_{PL} is between 200 Ω and 1000 Ω.

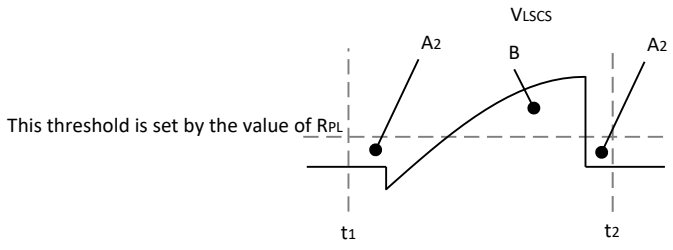


Figure 18 Threshold set by R_{PL} for limited power transfer in the BM

In order to filter the voltage spike at the LSCS pin so that the over-current protection is not wrongly triggered, an external filter consisting of R_{PL} and a small capacitor of some pF is recommended. The capacitor should be placed near the LSCS pin.

4.5 HB secondary rectifier diode selection

For the selection of the secondary rectifier diode,the following considerations should be taken into account:

- Reverse breakdown voltage

HB LLC resonant converter design

As the FB rectifier network is used on the secondary side, the diode voltage stress should be the same as the output voltage. However, due to the leakage inductance of the LLC transformer, the rectifier diodes could experience much higher voltage peak, thus a factor 2 margin should be taken into account:

$$V_{RRM\_D\_sec} > 2 * (V_{out\_max} + V_{F\_D\_sec}) = 155\ V$$

- Maximum RMS current

The maximum RMS current flowing through each rectifier diode is calculated as:

$$I_{D\_sec\_rms\_max} = \frac{\pi}{4} * I_{out\_max} = 1.37\ A$$

Using a diode with high current capability will benefit the power efficiency.

- Forward voltage

This is directly related to the power efficiency, so the forward voltage should be chosen to be as small as possible.

- Reverse recovery time

As hard commutation of the rectifier diode cannot always be prevented, it's better to choose an ultra-fast recovery diode.

- Power loss

The only power loss that should be considered is conduction loss. With a forward voltage of 0.5 V assumed, the diode conduction loss can be calculated as follows:

$$P_{loss\_D\_sec} = I_{D\_sec\_rms\_max} * V_{F\_D\_sec} = 0.7\ \text{W}$$

- Thermal characteristics

With the thermal resistance of the diode and ambient temperature T_A, the secondary rectifier diode temperature without heatsink is calculated as:

$$T_{D\_sec} = P_{loss\_D\_sec} * R_{D\_sec\_TH\_JA} + T_A$$

The important parameters for the boost diode used in the 130 W reference design are summarized in **Table 17**.

Table 17 Secondary rectifier diode design parameters

Parameter	Symbol	Value	Unit
Maximum reverse voltage	$V_{RRM\_D\_sec}$	200	V
Average rectified forward current	$I_{F\_D\_sec}$	18	A
Forward voltage	$V_{F\_D\_sec}$	0.7	V

4.6 HB secondary output capacitor

The ripple current flowing through the output capacitor is given as:

$$I_{cap\_sec\_rms\_max} = \sqrt{\left(\frac{\pi * I_{out\_max}}{2 * \sqrt{2}}\right)^2 - I_{out\_max}^2} = 0.84\ A$$

The output current ripple is defined as 2 percent of the maximum output current:

$$I_{out\_ripple} = 2\% * I_{out\_max} = 0.035\ A$$

HB LLC resonant converter design

The output voltage ripple is defined as 1 percent of the maximum output voltage:

$$V_{out\_ripple} = 1\% * V_{out\_max} = 0.76\,V$$

The equivalent serial resistance of the output capacitor is given as:

$$ESR = \frac{V_{out\_ripple}}{2 * I_{out\_ripple}} = 0.22\,\Omega$$

The value of the output capacitor is then given as:

$$C_{out} = \frac{1}{4 * \pi * f_{line} * \sqrt{\left(\frac{V_{out\_ripple}}{I_{out\_ripple}}\right)^2 - ESR^2}} = 80\,\mu F$$

Sometimes it is impossible to meet the ripple specification with a single output capacitor due to the high ESR of the electrolytic capacitor. Then, additional LC filter stages (post filter) or more parallel capacitors (to reduce ESR) can be used. When using the post filters, be careful not to make the corner frequency too low. Too low a corner frequency may make the system unstable or limit the control bandwidth. It is typical to set the corner frequency of the post filter at around 1/10 ~ 1/5 of the switching frequency.

The important parameters for the output capacitor used in the 130 W reference design are summarized in **Table 18**.

Table 18 Output capacitor design parameters

Parameter	Symbol	Value	Unit
Voltage rating	$V_{out\_C\_sec}$	100	V
Equivalent serial resistance	ESR	0.22	Ω
Output capacitor	C_{out}	220 // 220	µF

4.7 HB output regulation

In the ICL5102 130 W reference design, the HB LLC converter's output regulation includes three parts: constant current regulation, constant voltage regulation and output current dimming:

- Constant current regulation

In normal operation, the LLC converter provides a constant output current. The current regulation is realized through a PI regulator based on the operation amplifier LM358, as shown in **Figure 19**. The output current is measured through three parallel connected shunt resistors, R50, R51 and R52. These CS resistors must have high accuracy so that the current is regulated precisely. Output reference current is set through a divided voltage from either the auxiliary supply voltage or from the dimming voltage.

- Output current dimming

Based on the constant current regulation, a 1 to 10 V dimmer is implemented and provides the current reference signal for the output current regulation. A secondary auxiliary winding with a linear regulator is designed to provide power supply for the dimming circuit.

HB LLC resonant converter design

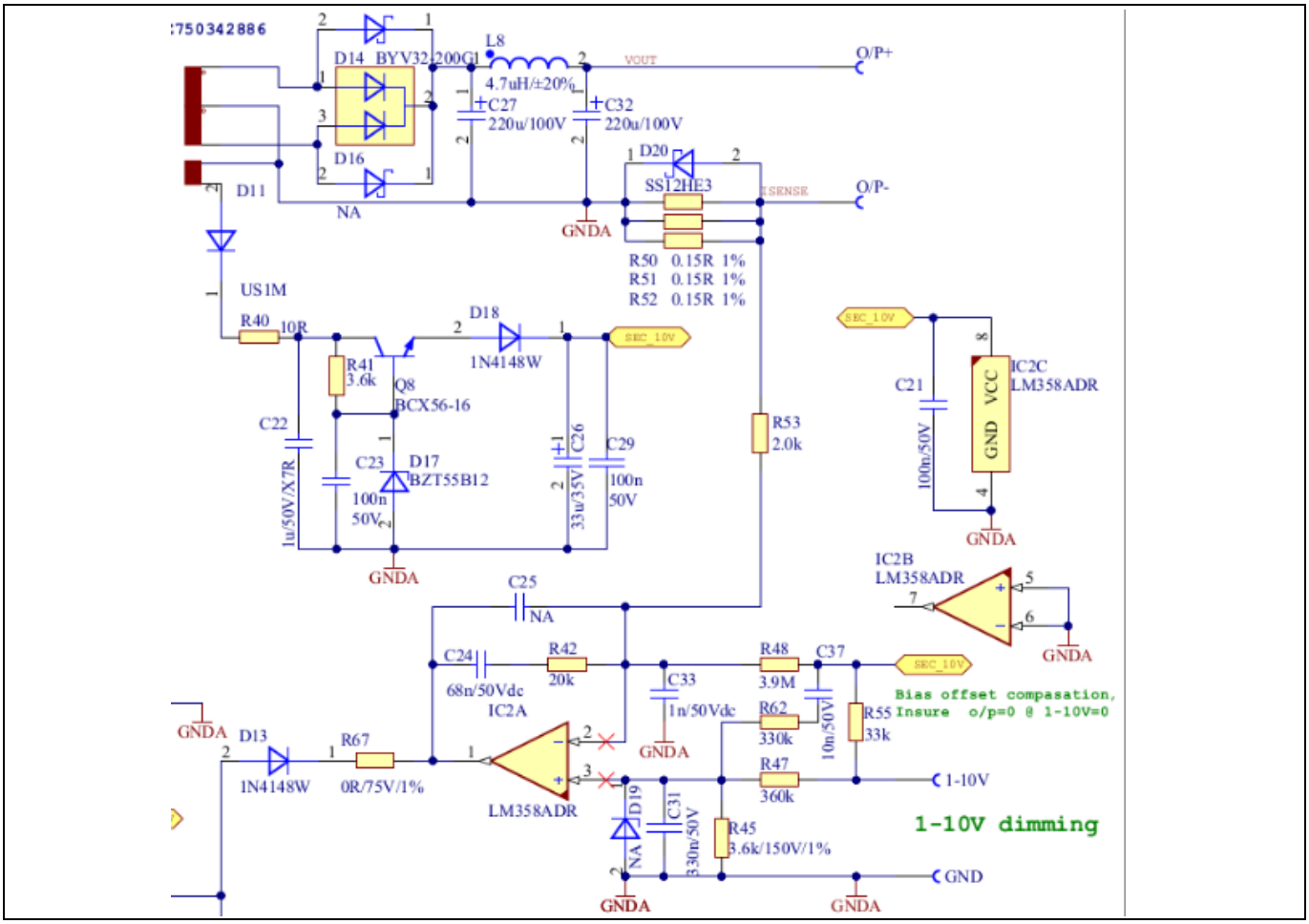


Figure 19 Secondary-side output current regulation

- Constant output voltage regulation

In addition to the current regulation, a constant output voltage regulation by using TL431 to prevent LEDs lighting in the dim-to-off condition is implemented so that the output voltage is not charged higher than the lowest possible LED forward voltage. This is shown in **Figure 20**.

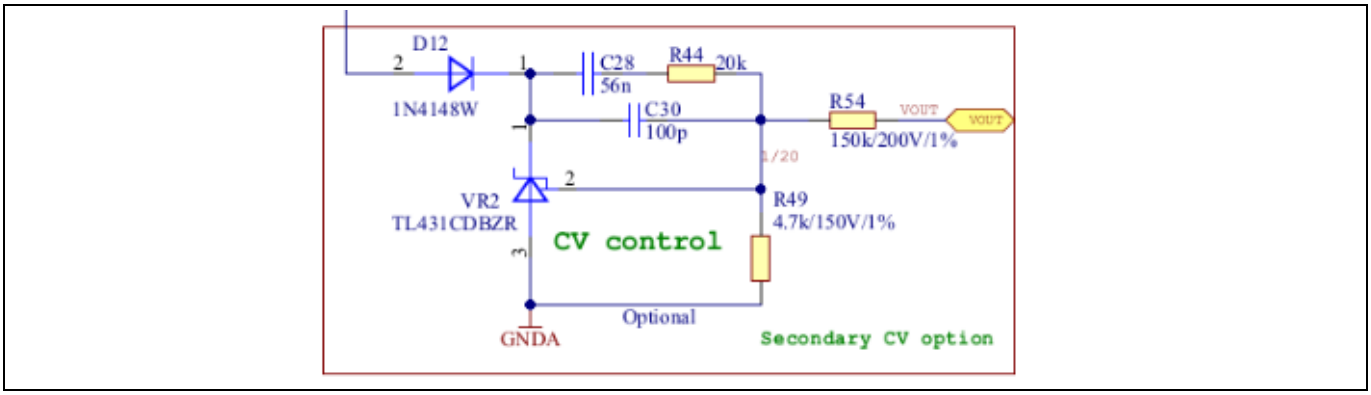


Figure 20 Secondary-side constant output voltage regulation

Once the dimmer brings the LED driver into the dim-to-off condition, the constant current regulator is then substituted by the constant output voltage regulator. ICL5102 will enter BM according to the output voltage regulation.

HB LLC resonant converter design

4.8 Frequency setting

In order to close the regulation loop, the feedback signal from the secondary-side current regulation circuit through the optocoupler must be connected to ICL5102.

During normal operation, the ICL5102 HB controller uses CCO to determine the switching frequency. The switching frequency is determined by current I_{RF} that flows out of the RF pin. The RF pin maintains a constant voltage of V_{RF} = 2.5 V. This voltage, together with the voltage at pin V_{BM}, resistors R_{BM} and R_{RF}, and the optocoupler, defines the current flowing out of the RF pin as shown in the following formula and **Figure 21**:

$$I_{RF} = I_1 + I_2 = I_{BM} + I_{OP} + \frac{V_{RF}}{R_{RF}}$$

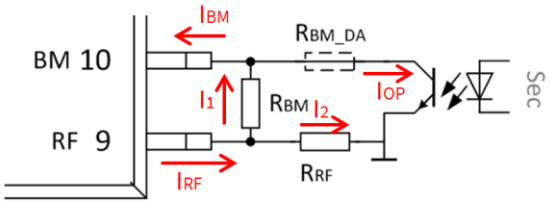


Figure 21 ICL5102 RF pin current definition

In order to determine the two resistors R_{RF} and R_{BM}, the HB LLC converter minimum and maximum switching frequency are needed.

$$f_{min} = 41.6\ kHz$$

$$f_{max} = 260\ kHz$$

The CCO of the ICL5102 HB controller is defined linearly with the constant slew rate C_{FC} as shown in **Figure 22**.

$$C_{FC} = 400\ KHz/mA$$

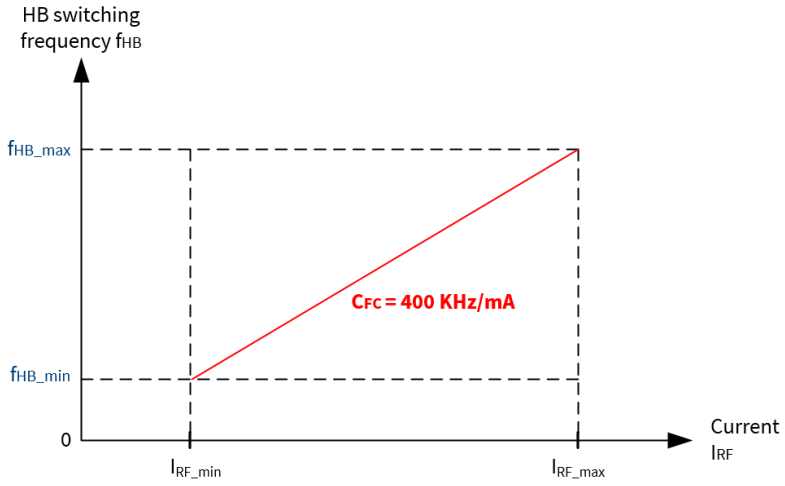


Figure 22 CCO of ICL5102 in normal operation

The calculation of the two resistors R_{RF} and R_{BM} is given below:

HB LLC resonant converter design

$$R_{RF} = C_{FC} * \frac{15V}{7 * f_{min} - f_{max}} = 212.76\ k\Omega$$

$$R_{BM} = C_{FC} * \frac{1.5V}{f_{max} - f_{min}} = 2.71\ k\Omega$$

Both resistor values are selected so that the ICL5102 130 W reference design does not enter BM in order to prevent the LED blinking. BM is only for standby in the dim-to-off operation.

4.9 Primary output OVP

ICL5102 provides another possibility for fast OVP on the primary side by sensing the winding voltage. This function is realized by connecting the OVP pin to the auxiliary winding of the HB transformer, as shown in **Figure 23**.

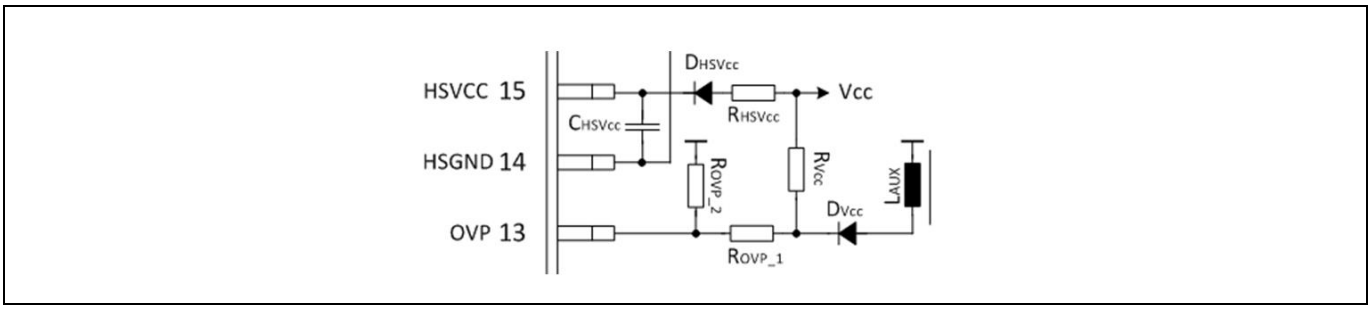


Figure 23 HB output OVP detection circuit

To decide the R_{OVP_1} and R_{OVP_2}, the following formula can be used:

$$\frac{R_{OVP\_2}}{R_{OVP\_1}} = \frac{V_{OVP\_ref}}{V_{out\_OVP} * \frac{n}{n_a} - V_{F\_D\_Vcc} - V_{OVP\_ref}}$$

If the primary output voltage level is defined as 90 V, the divider ratio can be obtained as:

$$\frac{R_{OVP\_2}}{R_{OVP\_1}} = 15$$

5 Power supply for ICL5102

The power supply for the controller ICL5102 is provided by the capacitors connected to the V_{CC} pin. It is strongly recommended to use both an electrolytic capacitor and a ceramic capacitor parallel connected as V_{CC} capacitors. Due to its high capacitance, an electrolytic capacitor is suitable as charge storage but has a bad AC coupling behavior. A ceramic capacitor on the other hand has an excellent AC decoupling effect, but has a capacitance derating strongly dependent on voltage and temperature.

The V_{CC} capacitors can be charged either by high-voltage start-up circuit, by PFC auxiliary winding or by HB auxiliary winding according to different requirements and conditions. This is described in the following sections.

5.1 Power-up

At start-up, after the AC input is applied, the V_{CC} capacitors must be charged by the high-voltage start-up circuit before the V_{CC} reaches the on-threshold. The start-up circuit can be designed as illustrated in **Figure 24**.

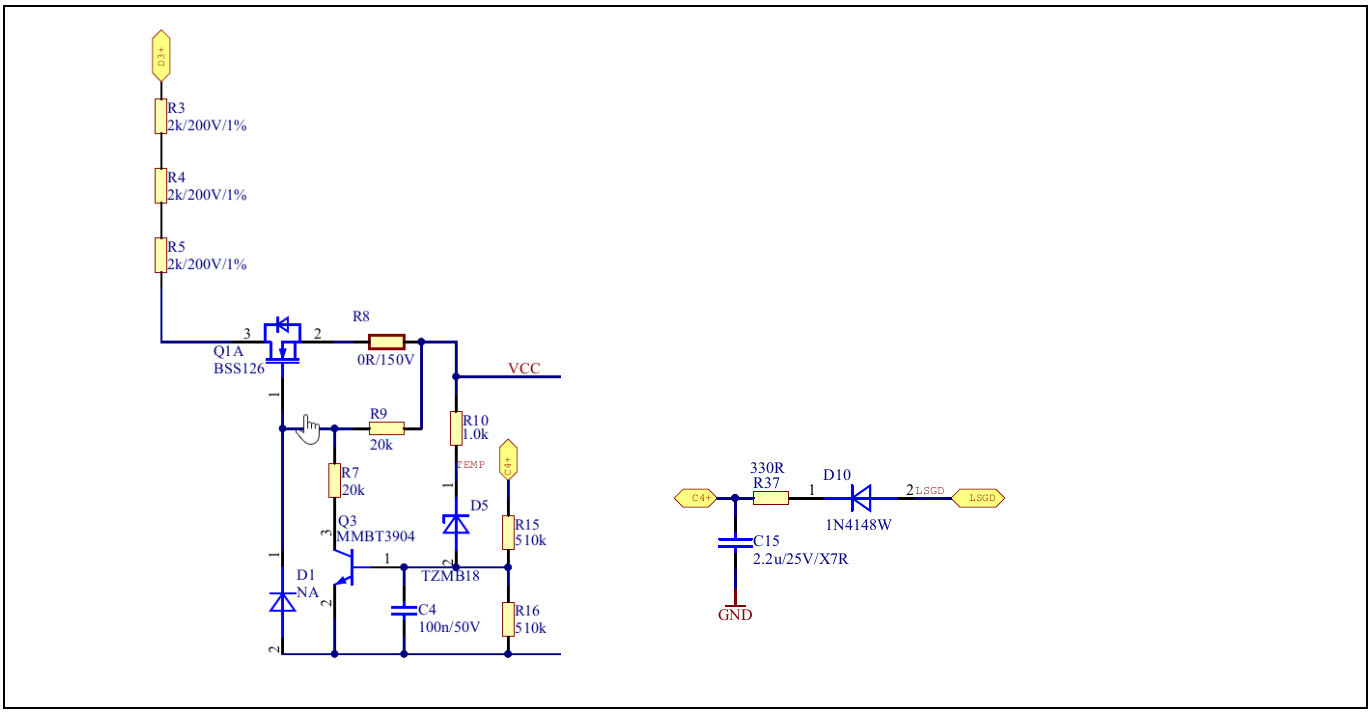


Figure 24 V_{CC} power supply from the start-up circuit

The start-up circuit is connected to the rectified AC input after the bridge rectifier via the current limitation resistors R3, R4 and R5. Depletion MOSFET BSS126 is used so that the V_{CC} charge function happens automatically as long as the V_{CC} is under a certain threshold dependent on the customer's design.

The charge current limited by the resistors R3, R4 and R5 is dependent on the time-to-light from the LED driver system requirement. Assuming the required time-to-light is 500 ms, the V_{CC} capacitor is 220 µF, and PFC and HB start-up time are each 50 ms, the minimum charge current can be calculated as:

$$I_{Vcc\_charge\_min} = C_{Vcc} * \frac{V_{Vcc\_on\_max}}{t_{time\_to\_light} - t_{PFC\_start} - t_{HB\_start}} = 12.2\ mA$$

To ensure that this charge current can be provided at low-line input so that the time-to-light is under 500 ms, the current limitation resistor value is calculated as:

Power supply for ICL5102

$$R_{HV} = \frac{\sqrt{2} * V_{in\_min\_rms}}{I_{Vcc\_charge\_min}} = 10\ \text{k}\Omega$$

And the maximum charge current at high-line input is:

$$I_{Vcc\_charge\_max} = \frac{\sqrt{2} * V_{in\_max\_rms}}{R_{HV}} = 43\ mA$$

It is strongly recommended to split the R_{HV} to three serial connected resistors as R3, R4 and R5 to minimize the power and voltage stress of each current limitation resistor.

5.2 Power supply during operation

After the V_{CC} on-threshold is reached, ICL5102 will start the first PFC stage to boost the bus voltage to V_{bus} = 450 V and then the HB stage begins to work. If there is no V_{CC} power supply from the PFC auxiliary winding, the V_{CC} high-voltage start-up circuit should stop working at the latest when the HB begins to work and the HB auxiliary winding will take over the charging of the V_{CC} capacitors. The start-up circuit can be disabled either by a certain V_{CC} voltage level or by the switching of the HB LS gate, as shown in **Figure 24**.

It is recommended to regulate the V_{CC} voltage not higher than 16 V, in order to reduce the power consumed by ICL5102 and avoid over-temperature. A reference circuit using a linear regulator for the V_{CC} power supply circuit is shown in **Figure 25**.

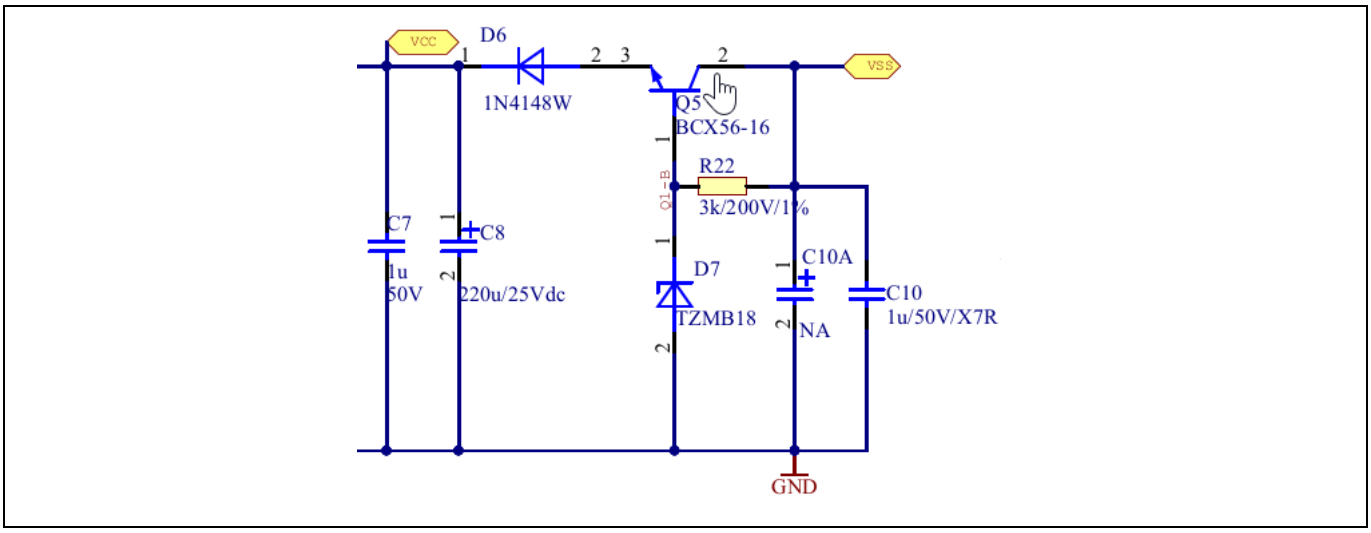


Figure 25 Linear regulator for V_{CC} power supply

5.3 Power supply for high-side driver

Due to the floating ground of the HS gate driver of ICL5102, the power supply for HS is separated and realized through the bootstrap circuit. This is shown in **Figure 26**.

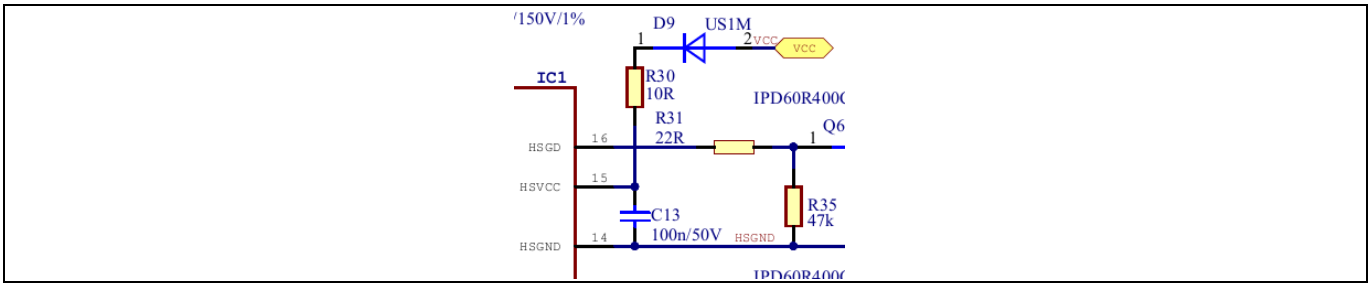


Figure 26 Bootstrap circuit for high-side power supply

Power supply for ICL5102

To optimize HB switching behavior during start-up and BM operation, the HS V_{CC} capacitor is targeted to reach its turn-on threshold within the first LS gate pulse. Thus, its capacitance can be calculated as:

$$C_{HSVcc} < -\frac{0.5 * T_{HB\_switching} + t_{dead\_time}}{ln(1 - \frac{V_{HSVcc\_on}}{V_{cc} - V_{D\_BS}}) * R_{HSVcc}}$$

Assuming the HB soft-start frequency begins with 300 kHz with 500 ns dead-time, HS V_{CC} on-threshold is maximum 11V, V_{CC} is 15 V, bootstrap diode forward voltage is 0.7 V, current limitation resistor is 10 Ω, the capacitor value is:

$$C_{HSVcc} < 120nF$$

In the ICL5102 130 W reference design, a 100 nF capacitor is selected.

The diode in the bootstrap circuit must be a 600 V high-voltage one according to the bus voltage level. The reverse recovery speed must be ultra-fast.

5.4 Other considerations for ICL5102 power supply

In the dim-to-off condition, ICL5102 is working in BM. To ensure the power supply for ICL5102 so that the V_{CC} does not fall below the UVLO threshold, other auxiliary winding for the power supply such as PFC or external resonant inductor auxiliary windings could be taken into account.

6 OTP

ICL5102 provides OTP through an external connected NTC temperature sensor at the OTP pin. The source current out of the OTP pin is I_{OTP} = 100 µA. The current generates a voltage drop on the connected NTC sensor. Once the voltage at the OTP pin decreases below V_{OTP_off} = 625 mV longer than the blanking time $t_{OTP\_blanking}$ = 620 µs in normal operation, both PFC and HB stages stop switching and ICL5102HV will enter auto-restart. PFC and HB operations recover after the voltage at the OTP pin is higher than V_{OTP_start} = 703 mV for longer than $t_{OTP\_blanking}$. This is shown in **Figure 27**.

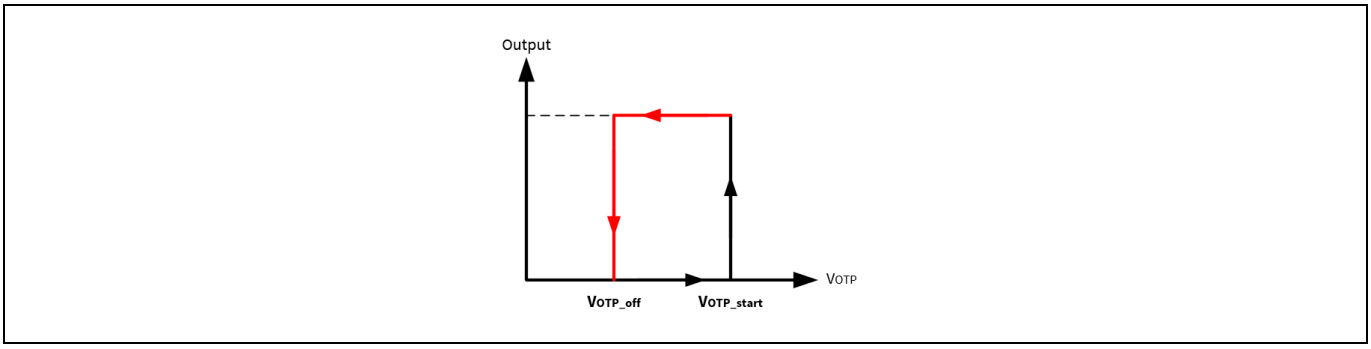


Figure 27 External OTP

In order to trigger the OTP threshold, the connected NTC should have a resistor value:

$$R_{OTP\_off\_min} = \frac{V_{OTP\_off\_min}}{I_{OTP\_max}} = 5.6\ \text{k}\Omega$$

In order to recover from the OTP, the connected NTC should have a resistor value:

$$R_{OTP\_start\_max} = \frac{V_{OTP\_start\_max}}{I_{OTP\_min}} = 7.8\ \text{k}\Omega$$

With the calculated values, the NTC temperature sensors of EPCOS 8507 series are compared as in the example in **Figure 28**.

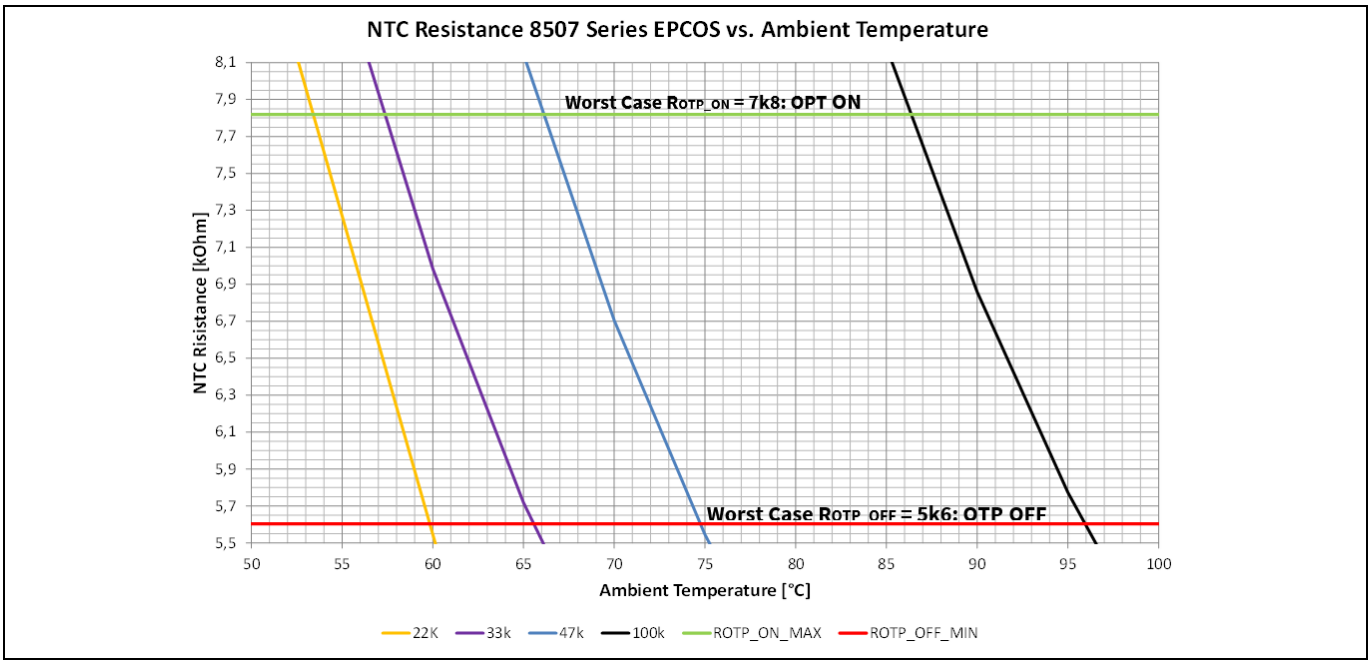


Figure 28 Characteristics of EPCPS NTC 8507 series

OTP

From the characteristics, 100 kΩ NTC is suitable and selected for the ICL5102 130 W reference design.

Additionally, to modify the resistance curve according to different requirements, serial or parallel resistors can be added to the NTC. **Figure 29** shows examples of how to modify the characteristics of NTC.

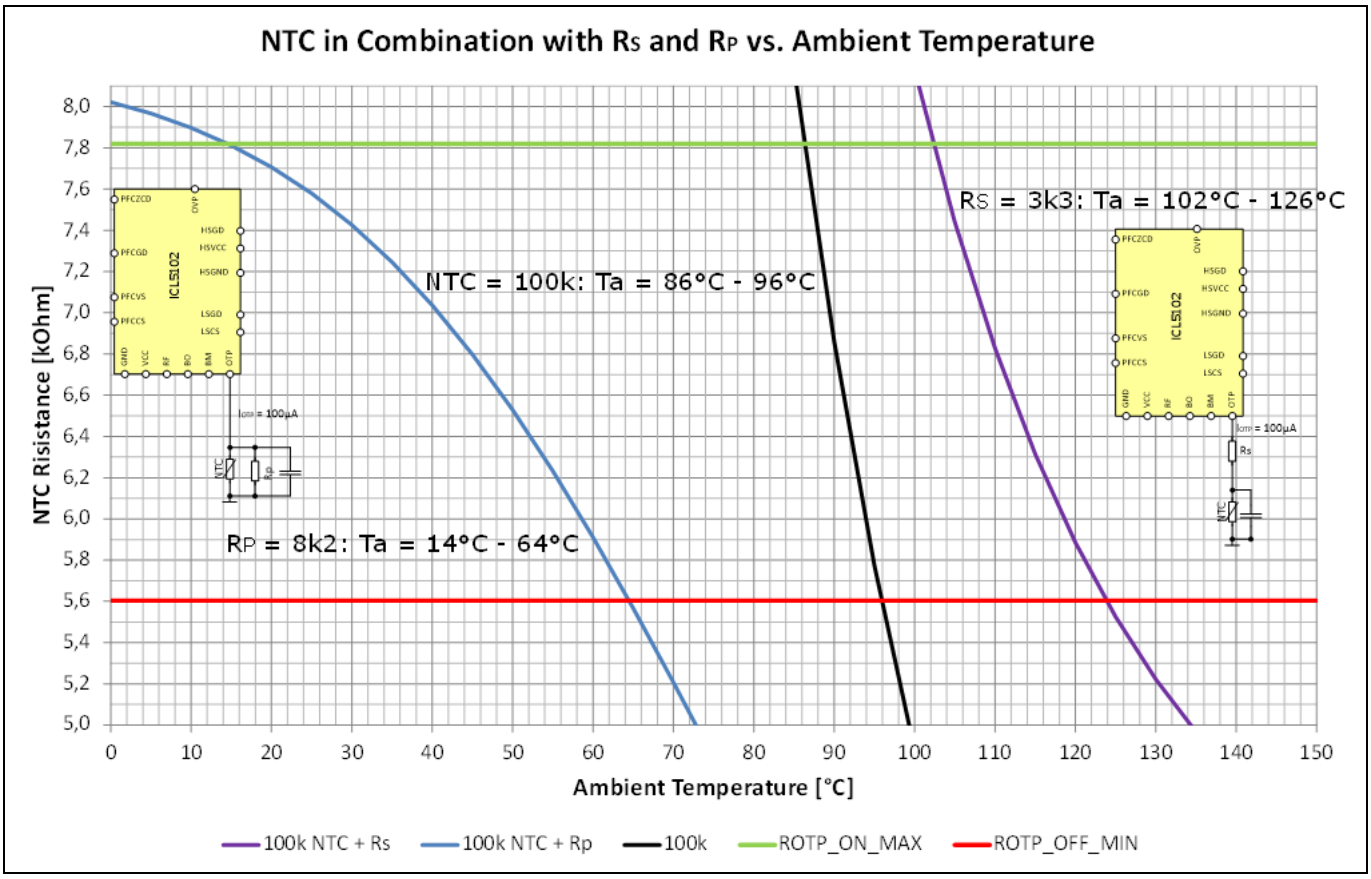
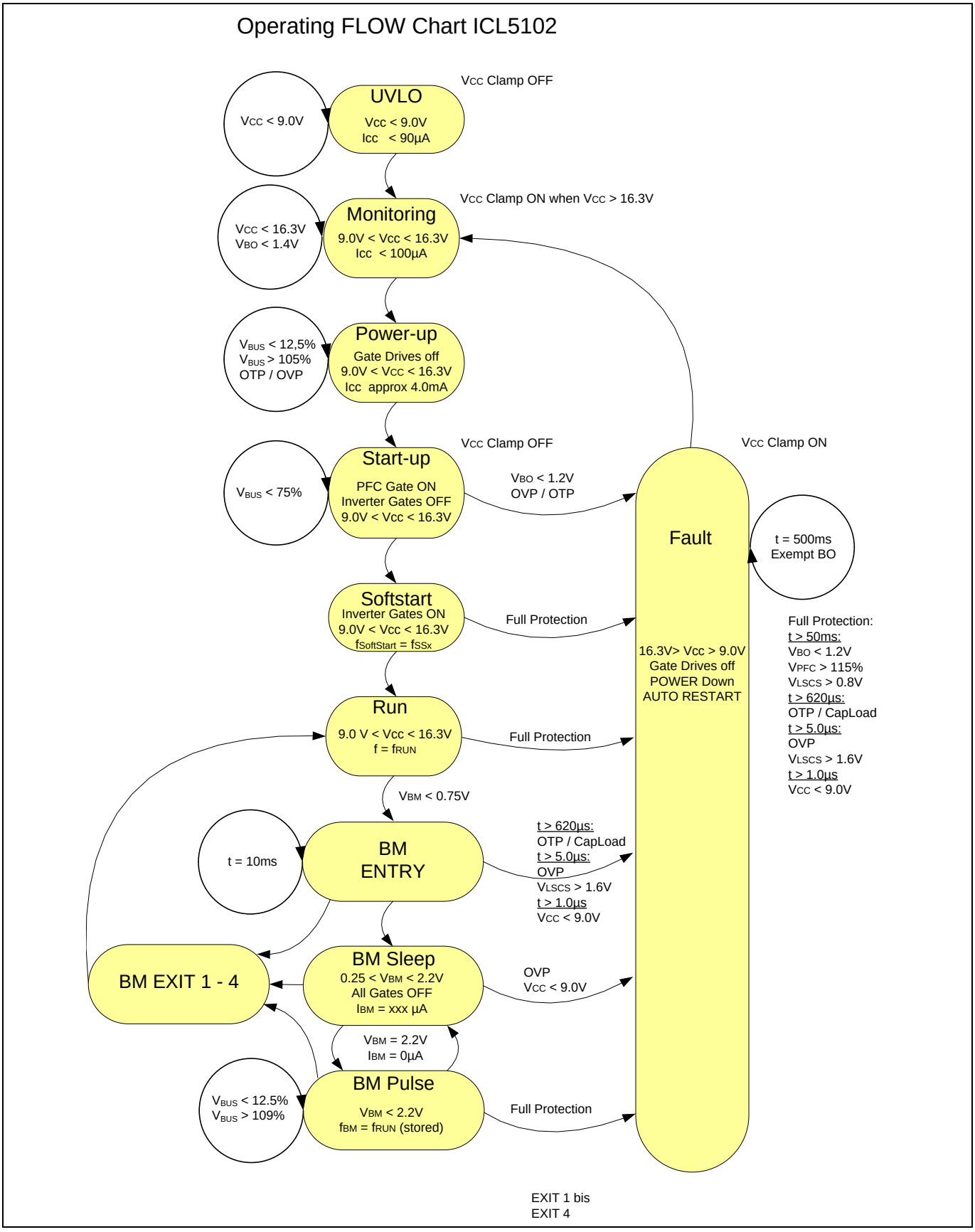


Figure 29 Modified characteristics of EPCPS NTC 8507 series

It is recommended to place a good-quality ceramic capacitor close to the OTP pin to prevent mis-triggered OTP protection from noise.

To disable the external OTP, a resistor of 20 kΩ can be added at the OTP pin instead of the NTC to keep the voltage higher than V_{OTP_start} = 703 mV.

7 ICL5102 operation flow chart



ICL5102 operation flow chart

Figure 30 ICL5102 operation flow chart

8 Protection features

All ICL5102 protection features are summarized in the following fault matrix.

Table 19 ICL5102 fault matrix

Description of fault	Characteristics of fault			Operating mode Detection is active							Consequence
	Definition of fault	Action	Minimum duration of effect	Monitoring	Power-up 130 µs	Start-up	Soft-start	Run mode	Burst pulse	Burst sleep	Reaction
Supply voltage V_{CC} less than 16.0 V before power-up	Below start-up threshold	W	1 µs	X							Prevents power-up
Supply voltage V_{CC} less than 9.0 V after power-up	Below UVLO threshold	A	1 µs	X	X	X	X	X	X	X	Power-down Auto-restart
Brown-out detection V_{BO} less than 1.2 V	BO	A	50 ms			X	X	X	X		Power-down Auto-restart when V_{BO} more than 1.4V
Brown-in control V_{BO} less than 1.4 V	BI	W	1 µs	X							Prevents power-up
Over-temperature detection V_{OTP1} less than 703 mV	OTP	W	620 µs		X						Prevents power-up
Over-temperature detection V_{OTP2} less than 625 mV	OTP	A	620 µs			X	X	X	X		Power-down Auto-restart when V_{OTP} more than 703mV
Bus voltage less than 12.5 percent of rated level	Open-loop detection	W	1 µs		X						Keep all gate drives off, restart when V_{BUS} more than 12.5 percent
Bus voltage less than 12.5 percent of rated level	Open-loop detection	W	1 µs			X	X	X	X		Stops PFC FET Restart when V_{BUS} more than 12.5 percent
Bus voltage less than 75 percent of rated level	PFC under-voltage	W	1 µs			X					Prevents start-up until V_{BUS} more than 75 percent Keep HB gate drives off
Bus voltage more than 105 percent of rated level	PFC over-voltage	W	5 µs		X						Keep all gate drives off Auto-restart after V_{BUS} less than 105 percent
Bus voltage more than 109 percent of rated level	PFC over-voltage	W	5 µs			X	X	X	X		Stops PFC FET Restart when V_{BUS} less than 105 percent
Bus voltage more than 115 percent of rated level	Inverter over-voltage	A	50 ms				X	X	X		Power-down Auto-restart

Protection features

Description of fault	Characteristics of fault			Operating mode Detection is active							Consequence
Output over voltage V_{OVP} more than 2.5 V	OVP	W	5 µs		X						Prevents power-up
Output over voltage V_{OVP} more than 2.5 V	OVP	A	5 µs			X	X	X	X	X	Stops ALL FETs Restart when V_{OVP} less than 2.5V
Capacitive load operation below resonance	Capacitive load	A	620 µs				X	X	X		Power-down Auto-restart
Capacitive load control	Capacitive load control	N	Half-cycle				X	X			Increases HB frequency
N = handled during normal operation	W = Wait while condition is present										A = Auto-restart
	Definition of fault	Action	Minimum duration of effect	Monitoring	Power-up 130 µs	Start-up	Soft-start	Run mode	Burst pulse	Burst sleep	Reaction
Voltage at PFCCS pin V_{PFCCS} more than 1.0 V	PFC over-current control	N	200 ns			X	X	X	X		Stops on-time of PFC FET immediately
Voltage at LSCS pin V_{LSCS} more than 0.8 V	Over-current control	N	Half-cycle				X	X	X		Increases HB frequency
Voltage at LSCS pin V_{LSCS} more than 0.8 V	Over-current shut-down	A	50 ms				X	X			Power-down Auto-restart
Voltage at LSCS pin V_{LSCS} more than 1.6 V	Inverter over-current	A	500 ns				X	X	X		Power-down Auto-restart
N = handled during normal operation	W = Wait while condition is present										A = Auto-restart

Protection features

[1] ICL5102 datasheet

[2] ICL5102 130 W reference board engineering report

[3] ICL5102 LLC design tool

[4] Primary-side MOSFET selection for LLC topology, application note, Infineon Technologies

Revision history

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1.0	2019-02-26	First release

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