Design guide for adaptor with XDPS21071

How to design a 45 W USB-PD adaptor with Infineon’s latest ZVS digital controller XDPS21071

About this document

Scope and purpose

This is a design guide to help customers design a 45 W USB-PD adaptor with Infineon’s latest ZVS digital controller XDPS21071. It provides guidelines for power stage design, IC parameter settings, PCB layout and .dp V 1.0

Intended audience

This document is intended for power supply engineers who need to design adaptors with Infineon’s digital control ZVS IC, XDPS21071. It also provides insight into making a highly efficient transformer design with understanding of the loss mechanism.

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Abstract

1 Abstract

Infineon’s digital controller XDPS21071 is a fixed-frequency Discontinuous Conduction Mode (DCM) Zero Voltage Switching (ZVS) controller for a Flyback converter. This design guide provides information on the ZVS principle of the Flyback converter as well as related parameter settings. It gives a step-by-step design for power stage components with given specifications of a typical 45 W USB-PD adaptor design. The design equation can also be extended to other power ratings. Installation and usage of a Graphical User Interface (GUI) – .dp Vision – are covered, and customers will learn how to set IC parameters through the digital interface.
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Introduction

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Introduction

2.1

IC introduction

The XDPS21071[1] is a ZVS Flyback current mode controller with built-in HV start-up cell. The start-up cell makes the IC power supply much more efficient and flexible during no-load operation. The DSP in the controller is like the brain of the chip, making the controller much smarter than a conventional mixed-signal hardware chip. The DCM operation with ZVS function can be enabled at different input voltage levels, with frequency reduction mode and configurable burst mode power to get the best efficiency across line and load regulation. In addition a One-Time Programmable (OTP) unit is integrated to provide a wide set of programmable parameters to ease the design-in phase.

2.2

Pin configuration and description

The pin configuration is shown in Figure 1 and Table 1.

![Figure 1 Pin configuration](image)

Table 1 Pin definitions and functions

<table>
<thead>
<tr>
<th>Symbol</th>
<th>Pin</th>
<th>Type</th>
<th>Function</th>
</tr>
</thead>
<tbody>
<tr>
<td>ZCD</td>
<td>1</td>
<td>I</td>
<td>Zero Crossing Detection</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>The ZCD pin is connected to an auxiliary winding for zero crossing detection, positive pin voltage measurement and also to insert V_{cs} offset based on output voltage.</td>
</tr>
<tr>
<td>MFIO</td>
<td>2</td>
<td>I</td>
<td>Multi-Functional Input Output</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>The MFIO pin is connected to an optocoupler that provides an amplified error signal for PWM mode operation.</td>
</tr>
<tr>
<td>GPIO</td>
<td>3</td>
<td>IO</td>
<td>Digital General-Purpose Input Output</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>The GPIO pin provides a UART interface until brown-in. It is switched to weak pull-down mode and UART function is disabled during normal operation.</td>
</tr>
<tr>
<td>CS</td>
<td>4</td>
<td>I</td>
<td>Current Sense</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>The CS pin is connected via a resistor in series to an external shunt resistor and the source of the power MOSFET.</td>
</tr>
<tr>
<td>HV</td>
<td>5, 6, 7, 8</td>
<td>I</td>
<td>High Voltage input</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>The HV pin is connected to the rectified bulk voltage. An internally connected 600 V HV start-up cell is used for initial V_{cc} charge. Brown-in and brown-out detection are also provided.</td>
</tr>
<tr>
<td>GD1</td>
<td>9</td>
<td>I</td>
<td>FFR signal Gate Driver output</td>
</tr>
</tbody>
</table>
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<table>
<thead>
<tr>
<th>Symbol</th>
<th>Pin</th>
<th>Type</th>
<th>Function</th>
</tr>
</thead>
<tbody>
<tr>
<td>GD0</td>
<td>10</td>
<td>O</td>
<td>Gate Driver output</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>Output for directly driving the main power MOSFET.</td>
</tr>
<tr>
<td>VCC</td>
<td>11</td>
<td>I</td>
<td>Positive voltage supply</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>IC power supply.</td>
</tr>
<tr>
<td>GND</td>
<td>12</td>
<td>O</td>
<td>Power and signal ground</td>
</tr>
</tbody>
</table>

2.3 Product highlights

- Integrated 600 V start-up cell for fast start-up and direct bus voltage sensing
- Adaptive CS compensation for USB-PD
- FFR mode
- High-precision line and load regulation
- Primary-side output Over-Voltage Protection (OVP)
- Supports lowest no-load standby power
- One-pin UART interface for configuration

2.4 Simplified application diagram

![Simplified application diagram](image-url)
2.5 ZVS principles

Figure 3 shows the typical PWM sequences and related key waveforms for the ZVS Flyback.

After the primary MOSFET turns off at $t_0$, the Synchronous Rectifier (SR) MOSFET will turn on, delayed by a short blanking time. At $t_1$, the SR MOSFET turns off when the demagnetizing current ideally goes to zero, then the magnetizing inductance $L_p$ and $C_{eqv}$ will oscillate. The voltage of the primary MOSFET will oscillate from $V_{bulk}+V_{ref}$ to $V_{bulk}-V_{ref}$. If the auxiliary MOSFET is turned on at $t_2$, the resonant peak of the primary MOSFET will mean the magnetizing current is zero, then the $i_{mag}$ will build up as negative. During this controlled ZVS on-time, the $V_d$ of the primary MOSFET is clamped to $V_{bulk}+V_{ref}$. Once the peak current reaches $i_{zvs\_pk}$, the aux MOSFET is turned off, and because this current is stored in the magnetizing inductance and in the reverse direction, it will continue to flow in this direction and discharge the energy stored in $C_{eqv}$. This time duration in the IC is controlled by the $t_{ZVS\_dead}$ parameter, which is configurable. So at $t_4$, the drain voltage of the primary MOSFET reaches its minimum, and turns on the primary MOSFET, which reduces the turn-on losses significantly, which is almost ZVS. As seen in the diagram, the energy is proportional to $V_{bulk}$, and so is the ZVS on-time.

ZVS pulse insertion is based on nano-DSP core and memory info. The IC knows the next switching cycle period and ZVS dead-time and ZVS pulse on-time, so the switching period minus these two parts will decide the ZVS pulse starting point, assuming the IC main gate turn-on time is also fixed. When the CS signal reaches the current command, the main gate off-point can also be decided.

![Figure 3 ZVS principles](image-url)
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Introduction
3 Flyback power stage design

3.1 45 W adaptor system specifications

Table 2 shows the simplified system specs for the nominal 45 W adaptor. Only the key specs are included here to dimension the power train components such as the bulk capacitor, transformer and MOSFETs.

Table 2 45 W adaptor system specifications

<table>
<thead>
<tr>
<th>Description</th>
<th>Symbol</th>
<th>Value</th>
<th>Unit</th>
<th>Test conditions</th>
</tr>
</thead>
<tbody>
<tr>
<td>Input AC voltage</td>
<td>V AC</td>
<td>90 to 264</td>
<td>V AC</td>
<td></td>
</tr>
<tr>
<td>Input AC frequency</td>
<td>f-line</td>
<td>50/60</td>
<td>Hz</td>
<td></td>
</tr>
<tr>
<td>Output voltage</td>
<td>V_out</td>
<td>5/9/12/15/20</td>
<td>V</td>
<td></td>
</tr>
<tr>
<td>Output voltage ripple</td>
<td>ΔV_out</td>
<td>150</td>
<td>mV</td>
<td>From 5 V to 20 V under steady-state load</td>
</tr>
<tr>
<td>Nominal output current</td>
<td>I_nom</td>
<td>3</td>
<td>A</td>
<td>3 A for voltage below 20 V</td>
</tr>
<tr>
<td></td>
<td></td>
<td>2.25</td>
<td>A</td>
<td>2.25 A for 20 V output</td>
</tr>
</tbody>
</table>

3.2 Bulk capacitor selection

Figure 4 shows the typical waveforms after the rectification bridge with a large bulk capacitor. T1 is the conduction period of the rectifier bridge diode. During this timeframe, input AC voltage will charge the bulk capacitor, and the rest of time is the discharging of the bulk capacitor. The difference between the energy stored in the bulk capacitor is equal to the output power times (T2 to T1).

The conducting period can be calculated using the following equation:

\[ T_1 = T_2 \times \frac{\pi - \sin^{-1}\frac{V_{\text{bulk min}}}{V_{pk}}}{\pi} \] (1)

Where T2 is the period of rectified sine waveform, i.e. T_{line}/2, V_{pk} is the AC peak input voltage, and V_{bulk min} is the minimum voltage of the bulk capacitor.

The energy discharged from the bulk capacitor during the rectifier off-period is:

\[ C_{\text{energy}} = 0.5 \times C_{\text{bulk}} (\frac{V_{pk}^2}{V_{pk}} - \frac{V_{\text{bulk min}}^2}{V_{\text{bulk min}}}) \] (2)

This energy should be equal to the input power times of the off-period, so we get the following equation:

\[ P_{in} \times (T_2 - T_1) = C_{\text{energy}} \] (3)

Using equations (1), (2), (1) and (3), with given line frequency and AC input voltage, either the value of the bulk capacitor or minimum bulk voltage can be calculated once one variable is fixed.

Here we choose C_{bulk} 100 µF as an example, with the given power requirement in Table 2, giving the minimum bulk voltage shown in Table 3.
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Flyback power stage design

![Figure 4](image)

**Figure 4**  Bulk capacitor voltage after rectification

<table>
<thead>
<tr>
<th>$V_{\text{in}}$ (V AC)</th>
<th>Freq. (Hz)</th>
<th>$V_{\text{out}}$ (DC)</th>
<th>$I_{\text{out}}$ (A)</th>
<th>$P_{\text{out}}$ (W)</th>
<th>$V_{\text{bulkmin}}$ (DC)</th>
</tr>
</thead>
<tbody>
<tr>
<td>90</td>
<td>47</td>
<td>20</td>
<td>2.25</td>
<td>45</td>
<td>85</td>
</tr>
<tr>
<td>90</td>
<td>47</td>
<td>15</td>
<td>3</td>
<td>45</td>
<td>85</td>
</tr>
</tbody>
</table>

Using 100 μF and considering -15 percent derating for the bulk cap value, the minimum $V_{\text{bulk}}$ 85 V appears at 90 V AC/47 Hz with 2.25 A loading. We assume 15 V/3 A has similar efficiency, so the bulk cap minimum voltage is the same, due to low reflection voltage under 15 V$_{\text{out}}$ once the transformer turns ratio is fixed, so 15 V/3 A is the worst condition for transformer design in terms of $B_{\text{max}}$ design.

### 3.3 Transformer design

Now, with the minimum bulk cap voltage information at different AC-line/frequency and output power, we will be able to design the transformer of the Flyback converter. To fully optimize performance, the system will be designed at boundary mode at low-line (90 V AC/47 Hz) at nominal 45 W. Since the gate-drive signal is only available when Zero Crossing (ZC) is detected, during over-load condition, on-time will be increased and also the off-time, so switching frequency will be reduced. The transformer design will need to guarantee that flux density $B_{\text{max}}$ is below saturation level while the wire size only caters for nominal power to handle the thermal issue.

Reference [2] shows how to design the inductance in critical mode operation. The relevant equations are (4) and (5).

\[
L_p \cdot I_{pk} \left( \frac{1}{V_{\text{bulk}}} + \frac{1}{\eta_{\text{reg}}} \right) + 0.5 \cdot T_r = \frac{1}{f_{\text{sw}}} \tag{4}
\]

\[
0.5 \cdot L_p \cdot I_{pk}^2 \cdot f_{\text{sw}} \cdot \eta = P_{\text{in}} \tag{5}
\]

Table 4 shows the calculated inductance will be 190 μH, and peak current is 1.87 A at 20 V output. Meanwhile peak current is 2.08 A at 15 V output for a constant power application, which leads to the worst case being 15 V/3 A load.

**Table 4**  Calculated inductance and peak current

![Table 3](image)

**Table 3**  $V_{\text{bulk}}$ minimum voltage at different power conditions

<table>
<thead>
<tr>
<th>$V_{\text{in}}$ (V AC)</th>
<th>Freq. (Hz)</th>
<th>$V_{\text{out}}$ (DC)</th>
<th>$I_{\text{out}}$ (A)</th>
<th>$P_{\text{out}}$ (W)</th>
<th>$V_{\text{bulkmin}}$ (DC)</th>
</tr>
</thead>
<tbody>
<tr>
<td>90</td>
<td>47</td>
<td>20</td>
<td>2.25</td>
<td>45</td>
<td>85</td>
</tr>
<tr>
<td>90</td>
<td>47</td>
<td>15</td>
<td>3</td>
<td>45</td>
<td>85</td>
</tr>
</tbody>
</table>

Nominal
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Flyback power stage design

<table>
<thead>
<tr>
<th>F_{sw} (kHz)</th>
<th>T_{r} (µs)</th>
<th>V_{refl} (V DC)</th>
<th>V_{bulk} (V DC)</th>
<th>P_{out} (W)</th>
<th>η</th>
<th>L_{p} (µH)</th>
<th>I_{pk} (A)</th>
</tr>
</thead>
<tbody>
<tr>
<td>140</td>
<td>0.9</td>
<td>140</td>
<td>85</td>
<td>45</td>
<td>0.97</td>
<td>190</td>
<td>1.87</td>
</tr>
<tr>
<td>113</td>
<td>0.9</td>
<td>105</td>
<td>85</td>
<td>45</td>
<td>0.97</td>
<td>190</td>
<td>2.08</td>
</tr>
</tbody>
</table>

The next thing to do is decide the transformer turns based on the selected core shape/material, B_{max}. Core loss and copper loss are the main considerations here.

\[
B_{max} = \frac{L_{p} \cdot I_{pk}}{N \cdot A_{e}} \tag{6}
\]

Equation (6) shows the delta flux density or peak flux in DCM, where N is the primary turns of the transformer, and A_{e} is the effective flux area of the core.

Table 5 shows the peak current and maximum flux density under nominal power based on the selected EIQ25 core under a different output voltage. This indicates that when the system design at CRM is for 20 V/2.25 A, it will go to frequency reduction mode to stay in CRM at 15 V/3 A loading. If we design 15 V/3 A at CRM, then 20 V/2.25 A load will go deep into DCM. In the end, it is to balance efficiency of two design considerations to compare the thermal performance.

<table>
<thead>
<tr>
<th>V_{in} (AC)</th>
<th>P_{out} (W)</th>
<th>V_{out} (V)</th>
<th>V_{bulk} (V DC)</th>
<th>I_{pk} (A)</th>
<th>L_{p} (µH)</th>
<th>N_{p}</th>
<th>A_{e} (mm^2)</th>
<th>B_{max} (T)</th>
</tr>
</thead>
<tbody>
<tr>
<td>90</td>
<td>45</td>
<td>20</td>
<td>85</td>
<td>1.87</td>
<td>190</td>
<td>14</td>
<td>89</td>
<td>0.285</td>
</tr>
<tr>
<td>90</td>
<td>45</td>
<td>15</td>
<td>85</td>
<td>2.08</td>
<td>190</td>
<td>14</td>
<td>89</td>
<td>0.317</td>
</tr>
</tbody>
</table>

3.4 Output capacitor selection

One consideration when selecting the output capacitor value is based on the ripple requirement. The required ripple is 150 mV. The ripple has two parts: one is the current charge to the output capacitor; the other is dominated by the ESR of the capacitor. Figure 5 shows the current charge output capacitor for the Flyback converter.

Part 1 voltage ripple is calculated using the following equation:

\[
\Delta V_{o,1} = \frac{\Delta Q}{C} = \frac{(I_{pk} - I_{o})^2}{I_{pk} \cdot sec^2 \cdot C} \tag{7}
\]

Part 2 voltage ripple is calculated using the following equation:

\[
\Delta V_{o,2} = esr \cdot I_{ac} = esr \cdot \sqrt{i_{pk}^2 - i_{o}^2} \tag{8}
\]

Total output voltage ripple is the sum of \(\Delta V_{o,1} + \Delta V_{o,2}\).
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![Diagram](image)

**Figure 5**  Output voltage ripple

<table>
<thead>
<tr>
<th>Table 6</th>
<th>Output ripple calculation under 15 V/3 A</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\Delta V_{o,\text{spec}}$ (mV)</td>
<td>$I_{pk_sec}$ (A)</td>
</tr>
<tr>
<td>----------</td>
<td>-----------------</td>
</tr>
<tr>
<td>150</td>
<td>2.08 x 7</td>
</tr>
</tbody>
</table>

Table 6 shows the total ripple is $(13.3 + 118.3) = 131.6$ mV, which meets the requirements. The selected output capacitors are polymer capacitors, 330 µF/25 V and 680 µF/25 V.

### 3.5 MOSFET selection

#### 3.5.1 Primary MOSFET

The superjunction (SJ) MOSFET is the popular, HV discrete component used nowadays in AC-DC conversion. The latest CoolMOS™ P7 [3] technology shows the best $Q_g \times R_{ds(on)}$ Figure-of-Merit (FOM), and fast turn-off speed, which reduces the turn-off losses, so it is best suited to this design.

Based on the transformer design, the turns ratio is 7, so we can estimate the maximum voltage stress of the primary MOSFET based on the following equation. Derating is the most important consideration when selecting MOSFETs.

$$V_{ds} = V_{in} + V_{clamp}$$

So the maximum voltage rating during steady-state happens with maximum input RMS AC voltage, i.e. 264 V AC.

$$V_{ds} = 265 \times 1.414 + V_{clamp} = 90\text{ percent derating}; \text{ with a 700 V MOSFET, the } V_{clamp} \text{ should be kept below 255 V DC. This means properly dimensioning the snubber circuit is important, because the ZVS topology shown here will still have a turn-off spike caused by the energy in leakage inductance.}$$

The other selection consideration for the primary MOSFET is its $R_{ds(on)}$ value; usually conduction loss is the main concern at low-line, which is the worst case. A low $R_{ds(on)}$ MOSFET is preferred, but this means greater $C_{oss}$ and it also means more turn-on losses at high-line for conventional hard-switching or even QR Flyback. With ZVS technology, we now can choose a low $R_{ds(on)}$ MOSFET to optimize the conduction losses while maintaining low losses at high-line.
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As already known, different MOSFETs have different $R_{ds(on)}$ and $Q_{oss}$ combinations, and $Q_{oss}$ will need certain energy from the auxiliary winding to discharge it. Again, it is a trade-off between the losses gained and dissipated in term of system optimization.

Let us estimate the losses stored in the MOSFET junction capacitor at 230 V AC, considering the reflection voltage 140 V DC; the plateau of the primary MOSFET $V_{ds}$ will be $230 \times 1.414 + 140 = 465$ V DC, while from the IPD70R360P7S datasheet $C_{oss}$ curve, we find that from 465 V DC to 50 V DC, the $E_{oss}$ change is $(2.15 - 0.9) \mu$J, with 140 kHz switching frequency, which means power loss is $E_{oss} \times F_{sw} = 1.25 \times 140 = 175$ mW. We notice that from 50 V DC to zero, the $E_{oss}$ is around 0.9 µJ, which means 126 mW is needed to discharge from 50 V to zero. So when considering efficiency, the optimum turn-on point for the primary MOSFET is not really zero voltage; it should be around 50 V DC for CoolMOS™.

![Diagram 15: Typ. Coss stored energy](image)

Based on these considerations, IPD70R360P7S is chosen.

The conduction losses of the primary MOSFET at 15 V output and 3 A loading at 100°C junction temperature are:

$$P_{cond} = i_{rms}^2 \cdot R_{ds(on)} = \left( \sqrt{D/3} \cdot i_{pk} \right)^2 \cdot R_{ds(on)} = \left( \sqrt{0.53/3 \cdot 2.08} \right)^2 \cdot 0.522 = 0.399W$$

### 3.5.2 ZVS MOSFET

The transformer has one additional winding at the primary. This winding provides the energy to discharge the primary MOSFET’s equivalent $C_{oss}$ by controlling the on-time of the ZVS MOSFET. The designed transformer has $N_{pz} = N_p/N_{ZVS} = 14:1$ turns ratio, so similarly the $V_{ds}$ stress of this MOSFET can be calculated with $V_{\text{inmax}}/N_{pz} + V_{ZVS} = 374 V/14 + 10 V = 36.7 V$. Because this MOSFET only handles low energy, a small-signal MOSFET can be chosen. Here we choose BSL606SN.

### 3.5.3 SR MOSFET

The secondary-side SR MOSFET selected is the OptiMOS™ 5 series, considering the best $FOM_{gd}$ and $FOM_{oss}$.

Based on the turns ratio $N_{ps} = 14:2$, the MOSFET rating will be $374 V/7 + 20 = 73.4 V$. Due to the ZVS turns on the primary side, hardly any spike is seen at the output SR MOSFET in ZVS mode. So 100 V rating is chosen. Here we choose BSC0805LS.
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The RMS current in the secondary SR MOSFET is:

\[ i_{\text{rms}} = \sqrt{D_{\text{off}} \cdot i_{\text{pksec}}^3} = \sqrt{0.43^3 \cdot 2.08 \cdot 7} \approx 5.51 \text{A} \]

The conduction loss of the SR MOSFET is:

\[ i_{\text{rms}}^2 \cdot R_{\text{ds(on)}} = 5.51^2 \cdot 0.009 \text{ohm} = 0.273 \text{W} \]

3.5.4 SR IC selection

When the ZVS pulse is turned on, the SR \( V_{\text{ds}} \) will also drop below zero voltage. It will tend to mis-trigger the \( V_{\text{ds}} \) direct sensing SR IC, as shown in Figure 7. While using the voltage balance SR IC as shown in Figure 8, due to the SR arming function, there is no mis-triggering of the SR gate due to the ZVS pulse. For SR IC with direct sensing, the minimum on-time needs to be set to minimum, e.g. 50 ns.

**Figure 7** SR gate signal with \( V_{\text{ds}} \) direct sensing SR controller

**Figure 8** SR gate signal with voltage second balance SR controller
3.6 Snubber circuit design

When the primary MOSFET turns off, the energy stored in leakage inductance will cause significant ringing at the drain node. This ringing must be clamped to keep the MOSFET within safe operating range. A Zener clamp circuit as shown in Figure 9 is chosen, as it has high efficiency at light load since the Zener clamp will never conduct, so there is no residual resistor loss compared to an RCD snubber. This is the snubber circuit used in the 65 W demo board. In general, the energy that goes into the snubber circuit is leakage inductance stored energy. Depending on the D1 type, the losses will be different. Due to the reverse recovery characteristic, the current entering the clamp capacitor can return to the output side, and part of the energy is recovered to the input or output.

Figure 9 Snubber circuit

Figure 10 and 11 show the snubber current and clamp voltage waveforms when the MOSFET turns off.

In reality, there is reverse recovery of the snubber diode, as shown in Figure 10, which means not all the current into the snubber capacitors is lost; the longer the recovery, the lower the losses that can be achieved.

Table 7 shows the calculation results of energy entering the snubber tank during turn-off, which is around 1.16 W. Figure 10 also clearly shows that after the snubber diode current goes in the negative direction, the clamp voltage begins to decay.

This means not all the energy into the tank is lost, and the clamp voltage returns to the steady-state. Because the current returns to negative, the only way for this current path is to flow out of the dot terminal of the secondary-side transformer and partially to the output. So this energy is not fully dissipated.

Table 8 shows the losses in the Zener clamp, based on the measured waveforms. The power loss is calculated using 0.5 x I_z x V_{Zener,clamp} x duration x F_{sw}. It shows only 21 mW is dissipated in the Zener diode.

Table 7 Energy entering the snubber tank

<table>
<thead>
<tr>
<th>V_{clamp, bottom} (V)</th>
<th>V_{clamp, peak} (V)</th>
<th>C_{clamp} (nF)</th>
<th>F_{sw} (kHz)</th>
<th>P_{enter, snubber} (W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>123</td>
<td>151</td>
<td>2.2</td>
<td>137</td>
<td>1.16</td>
</tr>
</tbody>
</table>
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Flyback power stage design

Table 8  Power loss in Zener diode

<table>
<thead>
<tr>
<th>$I_{Zener}$ (mA)</th>
<th>$V_{Zener_clamp}$ (V)</th>
<th>Time (ns)</th>
<th>$F_{sw}$ (kHz)</th>
<th>$P_{Zener_dissipate}$ (mW)</th>
</tr>
</thead>
<tbody>
<tr>
<td>23</td>
<td>130</td>
<td>105</td>
<td>137</td>
<td>21</td>
</tr>
</tbody>
</table>

Figure 10  Snubber waveforms

Figure 11  Snubber waveforms
4 Control diagram

Figure 12 shows the control diagram of XDPS21071.

The secondary-side control feedback signal feeds into the MFIO pin through the optocoupler, and this signal is sampled after Leading-Edge Blanking (LEB) time every switching cycle and converted into a digital number by an 8-bit ADC, which takes around 1 µs. Based on this MFIO value, frequency and peak current will be picked up from the frequency law reference table. This value will be compared to the CS limit to see if maximum power is reached. If yes, after 5 ms blanking time, the IC will enter over load protection.

If not, the digital CS signal will be taken away from the Vcs offset, PDC correction and slope drop, and feed into the 8-bit DAC to convert back to an analog signal. It compares with the CS shunt resistor to form the classical current mode control.

The Vcs offset value is calculated based on output voltage, which means the maximum usable current command will be different for variable output voltage, and therefore it will have different output power. For calculation details please refer to 5.2.2.

The propagation delay correction is to compensate for propagation delay influenced by the slew rate of the CS signal. For example, \( P \text{ DC} = \frac{V_{\text{in}}}{L_p} \times T_{\text{delay}}, \) where \( V_{\text{bulk}} \) is the bulk capacitor voltage, \( L_p \) is main inductance, and \( T_{\text{delay}} \) includes both IC internal (~100 ns) and external MOSFET turn-off delay. The detailed calculation can be found in 5.6.

Also in the current command path is slope drop compensation; the slope drop starts from 0.4375 duty cycle, and after this point, IC current drops at a slew rate of 84 mV/µs. This is important for CCM operation condition but is not needed for DCM design.

![Control diagram](image-url)

**Figure 12** Overall control diagram
5 IC parameter settings

5.1 HV pin-related parameters and functions

The HV pin provides the start-up and bulk cap voltage sensing. The sensed voltage info will provide input power estimation and ZVS on-time adjustment. The brown-in/out function is also based on the HV pin, but on the current level.

The internal 600 V depletion start-up cell provides the charging current to the \( V_{CC} \) pin.

The charging current is calculated as
\[
    i_{hv}(t) = \frac{V_{bulk} - V_{CC}(t)}{R_{hv}}
\]
so it can be seen with a fixed HV pin external resistor, and the charging current will change proportionally with input AC line voltage.

When the \( V_{CC} \) is above 9 V, if the current flowing into the HV pin is greater than 1.156 mA, the IC starts switching. If the current is below 0.443 mA, the IC will stop switching after 1.06 ms.

So based on this current limit and the 100 kΩ HV pin resistor, we will see the brown-in voltage is
\[
    V_{brownin} = I_{hvbi} \times R_{hv} + I_{hvbi} \times 1.49 = 0.443 \text{ mA} \times (100 \text{ kΩ} + 0.99 \text{ kΩ}) = 44.7 \text{ V DC}.
\]

The start-up delay time needed is the time spent charging the \( V_{CC} \) capacitor before reaching the \( V_{CC} \) on threshold 20.5 V.

\[
    t_{delay} = C_{vcc} \times \frac{V_{cc,on}}{i_{hv} - i_{lk}}.
\]

\( i_{lk} \) is the HV pin leakage current before power-up, stated in the datasheet spec, typically ~30 µA.

\[
    R_{h} \text{ determines the current flowing out of the ZCD pin, based on designed } N_{aux} \text{ and } N_{s}. \text{ } R_{h} \text{ will be selected first based on the current limit of the ZCD pin. Once } R_{h} \text{ is decided based on the Over-Voltage Protection (OVP) requirement and } V_{ZCD} \text{ OVP threshold, } R_{i} \text{ can be calculated.}
\]

\[
    N_{aux} = 2, \quad N_{p} = 14, \quad V_{bulk} = 374 \text{ V}, \text{ from equation (10), } R_{i} \text{ is greater than or equal to } 13.4 \text{ kΩ, and here we choose } 39 \text{ kΩ.}
\]

The next thing is to decide the value of \( R_{i} \) based on equation (9) and the OVP requirement, \( N_{i} = 2, R_{h} = 39 \text{ kΩ, } V_{OVP} = 2.75 \text{ V, } V_{OVP} = 21.7 \text{ V, } R_{i} = 5.6 \text{ kΩ.}

To filter noise at the ZCD pin, a small capacitor such as 10 pF can be paralleled to the low-side resistor.
5.2.2 VCS offset based on VZCD

Figure 13 shows the VCS offset based on VZCD, while the ZCD voltage corresponds to the output voltage. The purpose of this compensation is to reduce the current command for different output voltages, so the maximum allowed power can be reduced. This feature is especially important for variable-output voltage applications when the system is dimensioned at full power at the highest voltage. So the power can be reduced at lower output voltages to meet Limited Power Supply (LPS) requirements.

ZCD pin voltage is from 1.2 V to 2.8 V, so the input range is 1.6 V, with the gain 1.5, the ADC input voltage will be 2.4 V and it is converted to the digital value internally with an eight-bit ADC.

The calculation procedure is to decide the zero point of VZCD above this threshold, there is no compensation.

Secondly, based on the targeted compensation value, calculate Kvcs_offset.

The output voltage:

\[
V_{ZCD_{\text{zeropoint}}} = \frac{R_l}{R_l + R_h} \times N_{aux} \times V_{o_{\text{zeropoint}}} \quad (11)
\]

Here we choose \( V_{o_{\text{zeropoint}}} = 13.5 \) V, with known \( R_l = 5.6 \) kΩ, \( R_h = 39 \) kΩ, \( N_{aux} = 2 \), \( N_s = 2 \), \( V_{ZCD_{\text{zeropoint}}} = 1.7 \) V.

The digital number \( V_{ZCD_{\text{dig,zeropoint}}} = (V_{ZCD_{\text{zeropoint}}} - 1.2) \times 1.5/2.4 \times 255 = 79 \); this parameter can be set in the .dp Vision GUI.

Next is to decide the target compensation value, e.g., 80 mV at \( V_{ZCD} = 1.2 \) V, so the compensated digital value of \( V_{cs_{\text{dig}}} = 80 \) mV/0.6 V x 255 = 34, then \( K_{vcs\_offset} = V_{cs_{\text{dig}}}/(0-V_{ZCD_{\text{zeropoint}}}) \times 65535 = 28240 \), and this slew rate is also a configurable parameter in the .dp Vision GUI.

![Figure 13 VCS offset compensation based on VZCD voltage](image)

5.3 Gate-driver related settings and functions

The two gate drivers are double-loop controlled, with external current loop and internal voltage loop. The default sourcing current is 0.15 A and 10 V clamp. The sink current limit is 0.5 A. Unlike the conventional voltage source-based driver, only one gate resistor is needed to limit the sink current. In general 10 Ω below the resistor is used; it is tuned based on switch-off losses versus turn-off speed.
5.4 MFIO pin-related parameters

The MFIO pin provides the feedback information for the IC. Based on the voltage measured at the MFIO pin, the IC will check the reference table to get the desired frequency and current limit command. The maximum frequency is configurable, which enables the customer to fine-tune system efficiency based on limited available core size/shape.
5.5 V<sub>CC</sub> pin-related parameters

V<sub>CC</sub> provides the operating voltage of the IC and also has Under-Voltage Lockout (UVLO) protection when V<sub>CC</sub> drops below 7.2 V. Brown-in is activated when V<sub>CC</sub> is above 9 V.

V<sub>CC</sub> also has a V<sub>CC</sub> OVP function.
5.6 CS pin-related parameters

The CS pin’s main functions are current mode control and OCP. For the current command, it is influenced by propagation delay compensation, $V_{CS}$ offset and slope-drop compensation, and these parameters can be tuned according to different system dimensions.

The current command is given by equation (16):

$$\text{CS}_\text{OCP1LV} = (I_{pk}) - PDC\_Correction - V_{cs\_offset} - \text{slopedrop}$$ (16)

$$PDC\_Correction = \text{INT}(2^{-16} \times PDC\_FACTOR \times V_{bulk} + PDC\_OFFSET)$$ (17)

Where $PDC\_FACTOR$ and $PDC\_OFFSET$ are two configurable parameters using .dp Vision – propagation delay-related parameters.

$$PDC\_FACTOR = \frac{R_{CS} \cdot t_{pd}}{L_p \cdot V_{cslsb}} \cdot (2^{16} - 1)$$ (18)

$$PDC\_Correction = \frac{R_{CS} \cdot t_{pd}}{L_p \cdot V_{cslsb}} \cdot (2^{16} - 1) = \frac{0.25\text{ohm} \cdot 352\text{ns}}{190\mu\text{H} \cdot 2.34\text{mV}} \cdot 65535 = 13000 = 32\text{CBh}$$

$PDC\_Correction = 13000 \times 85\text{V}/65535 = 17$ bits, each bit is 2.34 mV, so PDC-induced $V_{CS}$ reduction is 2.34 mV x 17 = 40 mV.

Where $R_{CS}$ is the current sense resistor, $t_{pd}$ is the propagation delay, which includes internal comparator delay and external turn-off delay, $L_p$ is the main inductance value, and $V_{cslsb}$ is the DAC LSB (least significant bit) of the CS signal.

Slope compensation-induced drop can be calculated as follows:

$$slopeedrop = (D_{on} - 0.4375) \times T_{sw} \times 84\text{mV}/\mu\text{s}$$ (19)

With $D_{on} = n \times V_{out}/(V_{bulk} + n \times V_{out})$, $n = 6$, $V_{out} = 20\text{V}$, $V_{bulk} = 85\text{V}$, $T_{sw} = 7.14\mu\text{s}$, slope compensation-induced $V_{CS}$ drop is $(0.5853 - 0.4375) \times 7.14 \times 84 = 89\text{mV}$.

As max $V_{CS}$ at low-line is 0.6 V, so $R_{CS} = (600\text{mV} - 40\text{mV} - 89\text{mV})/I_{pk} = 0.471\text{V}/2.08\text{A} = 0.23\text{Ω}$. As we see in the calculation procedure, the PDC calculation is influenced by $R_{CS}$, which is unknown before we get the final $R_{CS}$ value, and also the propagation delay needs to be calibrated. And peak current calculations are also influenced by the efficiency assumption, etc., so $R_{CS}$ would be tuned slightly to tailor it to a real design based on real test results. In the 45 W reference board, $R_{CS} = 0.25\text{Ω}$ is chosen.
6  Burst mode operation

To maintain efficiency at light load, the IC needs to keep the system working in burst mode. XPDS21071 will stop the core engine and reduce power consumption depending on feedback voltage, which reflects load signal. And the entry level of $V_{\text{MFIO}}$ is different based on the ZCD voltage. Table 9 shows the different burst mode entry levels for different output voltages. Different MFIO voltage means different peak current based on frequency.

So the power for entry into burst mode will change accordingly for three different output voltages.

Based on $N_{\text{aux}} = 2$, $N_{s} = 2$, ZCD divider $R_{h} = 39 \, \text{k} \Omega$, $R_{i} = 5.6 \, \text{k} \Omega$, $V_{\text{out}} = (R_{h} + R_{i})/R_{i} \times V_{ZCD} \times N_{s}/N_{\text{aux}}$, $V_{ZCD} = 2.152 \, V$ means $V_{\text{out}} = 17 \, V$, $V_{ZCD} = 1.723 \, V$ means $V_{\text{out}} = 13.7 \, V$.

One reason to have different power is that efficiency is different for different output voltages. Meanwhile, during burst mode operation, the burst mode settings are same for variable output voltages, so adjusting burst mode entry is necessary.

**Table 9  Burst mode entry level per $V_{ZCD}$**

<table>
<thead>
<tr>
<th>Burst mode entry threshold</th>
<th>Symbol</th>
<th>Min.</th>
<th>Typ.</th>
<th>Max.</th>
<th>Unit</th>
<th>Conditions</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$V_{\text{MFIOBMEN1}}$</td>
<td>0.57</td>
<td>0.607</td>
<td>0.644</td>
<td>V</td>
<td>$V_{ZCD}$ less than 1.723 V</td>
</tr>
<tr>
<td></td>
<td>$V_{\text{MFIOBMEN2}}$</td>
<td>0.419</td>
<td>0.455</td>
<td>0.491</td>
<td>V</td>
<td>1.723 V less than or equal to $V_{ZCD}$ less than 2.512 V</td>
</tr>
<tr>
<td></td>
<td>$V_{\text{MFIOBMEN3}}$</td>
<td>0.372</td>
<td>0.408</td>
<td>0.443</td>
<td>V</td>
<td>$V_{ZCD}$ greater than or equal to 2.152 V</td>
</tr>
</tbody>
</table>

Burst mode operation power is defined as follows:

$$P_{\text{bst}} = 0.5 \times L_{p} \times I_{pk}^2 \times f_{\text{bst}}$$  \hspace{1cm} (20)

Based on default $V_{CS}$ in burst mode 0.128 V, the peak current is 0.128 V/0.25 Ω = 0.512 A, $f_{\text{bst}}$ = 50 kHz. With known inductance, the burst mode power can be calculated.
7  Tips on PCB layout

In principle, there are two key switching loops, which need to be as small as possible. One is from bulk cap V+ to transformer winding to Q30 to Rsense to bulk cap V-. The other is the secondary-side rectification loop from the secondary transformer winding to the output filter capacitor and Q100 SR MOSFET.

For XDPS21071 ground connection, star ground to bulk cap negative V- is necessary. The ZCD winding’s ground should connect to filtering capacitor C50 first, then all other pins’ grounds should tie to this IC’s ground first before connecting to C50’s ground.

![System layout recommendation](image-url)

Figure 15  System layout recommendation
8 Usage of .dp Vision

8.1 Installation of .dp Vision

Customers can install the user interface .dp Vision on their own computer, following the installation instructions.

1) Install the latest .dp Vision_2.0.9.4 from the folder (double-click “dpVision_2.0.9.4.msi”).
2) Connect the IFX IDP21071 sample to the dpIFGen2 interface board using VCC, UART and GND pins.

![dpIFGen2 interface board (side view)](image)

3) Connect the dpIFGen2 interface board using the USB cable provided to the laptop USB port and open the .dp Vision GUI by double-clicking on “dpVision.exe” or the icon shown below.

![.dp Vision icon](image)

4) Open “XDPS21071_with_assistant_trials_rev1.07.csv” using menu → File → Open → browse to folder → \.....\dpVision\Parameters.
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**Figure 18  Power device on/off button and device status**

5) Press the “Power Device On/Off” button → “Device Status” should turn green.

**Figure 19  Power device on/off button and device status**

6) Press the “Test Configuration Set” button and the parameter values are loaded in the IDP2105 sample and the application firmware starts running.

7) Change the parameters under the “Value” field.
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<table>
<thead>
<tr>
<th>Design parameters</th>
<th>Name</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td><strong>Hardware Configuration</strong></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td><strong>Control Feature</strong></td>
<td>k_PDC</td>
<td>13000</td>
<td></td>
</tr>
<tr>
<td></td>
<td>k_PDC_OFFSET</td>
<td>0</td>
<td></td>
</tr>
<tr>
<td></td>
<td>t_CSLEB</td>
<td>269</td>
<td>ns</td>
</tr>
<tr>
<td></td>
<td>t_ZVSdead</td>
<td>220</td>
<td>ns</td>
</tr>
<tr>
<td></td>
<td>k_ZVSon</td>
<td>3200</td>
<td></td>
</tr>
<tr>
<td></td>
<td>t_ZVSon_OFFSET</td>
<td>30</td>
<td>ns</td>
</tr>
<tr>
<td></td>
<td>I_G00_Drive</td>
<td>31</td>
<td>mA</td>
</tr>
<tr>
<td></td>
<td>V_BULK_ZVS_ON</td>
<td>250</td>
<td>VDC</td>
</tr>
<tr>
<td><strong>Protection</strong></td>
<td>T_JOTP</td>
<td>130</td>
<td>degreeC</td>
</tr>
<tr>
<td></td>
<td>t_OCP2</td>
<td>600</td>
<td>ns</td>
</tr>
<tr>
<td></td>
<td>t_PeakPower</td>
<td>30</td>
<td>ms</td>
</tr>
<tr>
<td></td>
<td>EN_PeakPower</td>
<td>Enabled</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Response_OWP</td>
<td>Auto-Restart</td>
<td></td>
</tr>
<tr>
<td><strong>Burst Configuration</strong></td>
<td>Vcs_Burst</td>
<td>0.128</td>
<td>V</td>
</tr>
<tr>
<td></td>
<td>Freq_Burst</td>
<td>50.0</td>
<td>kHz</td>
</tr>
<tr>
<td></td>
<td>v_Burst_Pause</td>
<td>1.35</td>
<td>V</td>
</tr>
<tr>
<td></td>
<td>v_Burst.Exit</td>
<td>2.00</td>
<td>V</td>
</tr>
<tr>
<td></td>
<td>t_ReEntry_Burst</td>
<td>5</td>
<td>ms</td>
</tr>
<tr>
<td><strong>Frequency Law</strong></td>
<td>Freq_Law_Max</td>
<td>140.0</td>
<td>kHz</td>
</tr>
<tr>
<td></td>
<td>MFI0_Point_C</td>
<td>1.00</td>
<td>V</td>
</tr>
<tr>
<td></td>
<td>MFI0_Point_B</td>
<td>1.80</td>
<td>V</td>
</tr>
<tr>
<td></td>
<td>Vcs_at_Point_BC</td>
<td>0.45</td>
<td>V</td>
</tr>
<tr>
<td></td>
<td>Freq_cap_at_special_condition</td>
<td>105.0</td>
<td>kHz</td>
</tr>
<tr>
<td><strong>Adaptive Vcs offset</strong></td>
<td>k_Vcs_offset</td>
<td>28000</td>
<td></td>
</tr>
<tr>
<td></td>
<td>Vcs_offset_Vzcdzeropoint</td>
<td>79</td>
<td></td>
</tr>
<tr>
<td></td>
<td>EN_Vcs_offset</td>
<td>Enabled</td>
<td></td>
</tr>
</tbody>
</table>

Figure 20  Value field

8) Changing the parameters will activate the “Save Configuration Set” button. Save the configuration and again press the “Power on/off” button and click on “Test Configuration Set”, which will load the new configuration set into the chip’s RAM area. Click on “Functions/Burn Configuration Set”, which will burn the new configuration set into OTP memory.

Figure 21  “Save Configuration Set” button

![Image of dp Vision interface]

**Figure 22** “Help” button

### 8.2 Parameter setting with .dp Vision

Table 8 lists all the parameters configurable by customers through .dp Vision. The setting of the major parameters has been explained in detail in this design guide. The description of each parameter together with an explanatory image will be displayed with a click on each parameter name. In addition, a user manual can be obtained via the “Help” button.

#### Table 10 Configurable parameters

<table>
<thead>
<tr>
<th>Parameters</th>
<th>Parameter symbol</th>
<th>Default</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>Propagation delay compensation for peak current control</td>
<td>PDC_FACTOR</td>
<td>13000d</td>
<td>Propagation delay compensation factor</td>
</tr>
<tr>
<td></td>
<td>PDC_OFFSET</td>
<td>0d</td>
<td>Propagation delay compensation offset</td>
</tr>
<tr>
<td>Leading-Edge Blanking (LEB)</td>
<td>$t_{CSLEB}$</td>
<td>269 ns</td>
<td>Blanking filter at CS pin to avoid erroneous turn-off of GD0 due to leading-edge spike at GD0 turn-on</td>
</tr>
<tr>
<td>ZVS dead-time</td>
<td>$t_{ZVSdead}$</td>
<td>220 ns</td>
<td>Dead-time between end of ZVS pulse at GD1 and start of GD0</td>
</tr>
<tr>
<td>ZVS pulse length factor</td>
<td>$k_{ZVSonfactor}$</td>
<td>3200</td>
<td>ZVS pulse length factor</td>
</tr>
<tr>
<td>Gate driver capability</td>
<td>$I_{GD0_drive}$</td>
<td>31 mA</td>
<td>Sourcing current of gate driver 0</td>
</tr>
<tr>
<td></td>
<td>$T_{JOTP}$</td>
<td>130°C</td>
<td>Internal over-temperature detection level</td>
</tr>
<tr>
<td></td>
<td>$t_{ocp2}$</td>
<td>600 ns</td>
<td>Blanking time for OCP2 of $V_{CS}$ signal</td>
</tr>
<tr>
<td></td>
<td>$t_{peakpower}$</td>
<td>30 ms</td>
<td>Blanking time for over-load protection</td>
</tr>
<tr>
<td>Protections</td>
<td>En_OLP</td>
<td>Enabled</td>
<td>To enable or disable over-load protection</td>
</tr>
<tr>
<td></td>
<td>$Response_OVP$</td>
<td>Auto-restart</td>
<td>Protection mode for OVP, configurable for AR or latch</td>
</tr>
<tr>
<td>Burst mode parameters</td>
<td>$Vcs_bst$</td>
<td>0.128 V</td>
<td>Burst mode current limit</td>
</tr>
<tr>
<td></td>
<td>$Freq_bst$</td>
<td>50.0 kHz</td>
<td>Burst mode frequency</td>
</tr>
<tr>
<td></td>
<td>$V_{bst_pause}$</td>
<td>1.35 V</td>
<td>Pause threshold at MFIO pin during on-phase in burst mode operation</td>
</tr>
</tbody>
</table>
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## Usage of .dp Vision

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{\text{bst_exit}}$</td>
<td>2.00 V</td>
<td>Burst mode exit voltage at MFIO pin</td>
</tr>
<tr>
<td>$T_{\text{reentry_bst}}$</td>
<td>5 ms</td>
<td>Minimum time to re-enter burst mode</td>
</tr>
</tbody>
</table>

### Frequency law settings

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$F_{\text{sw_A}}$</td>
<td>140 kHz</td>
<td>Frequency settings for point A</td>
</tr>
<tr>
<td>$V_{\text{MFIO_C}}$</td>
<td>1.00 V</td>
<td>MFIO pin corner point C voltage</td>
</tr>
<tr>
<td>$V_{\text{MFIO_B}}$</td>
<td>1.80 V</td>
<td>MFIO pin corner point B voltage</td>
</tr>
<tr>
<td>$V_{\text{CS_BC}}$</td>
<td>0.45 V</td>
<td>CS limit between point B and C</td>
</tr>
<tr>
<td>$f_{\text{clamp}}$</td>
<td>105 kHz</td>
<td>Frequency clamp when $V_{\text{in}}$ is greater than 200 V, $V_{\text{ZCD}}$ is less than 1.28 V</td>
</tr>
</tbody>
</table>

### Adaptive $V_{CS\_offset}$

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>$K_{V_{CS_offset}}$</td>
<td>28000</td>
<td>Gradient for compensation curve</td>
</tr>
<tr>
<td>$V_{\text{CS_offset}, V_{\text{zcdzeropoint}}}$</td>
<td>79</td>
<td>ZCD voltage level (digital value) without $V_{CS_offset}$</td>
</tr>
<tr>
<td>$En_{V_{CS_offset}}$</td>
<td>Enabled</td>
<td>To enable or disable $V_{CS_offset}$ compensation</td>
</tr>
</tbody>
</table>


9 References

[1] XDPS21071 2.0 datasheet


Design guide for adaptor with XDPS21071
How to design a 45 W USB-PD adaptor with Infineon’s latest ZVS digital controller

Revision history

<table>
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<th>Date of release</th>
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