

AN973: Design Guide for Si8281/83

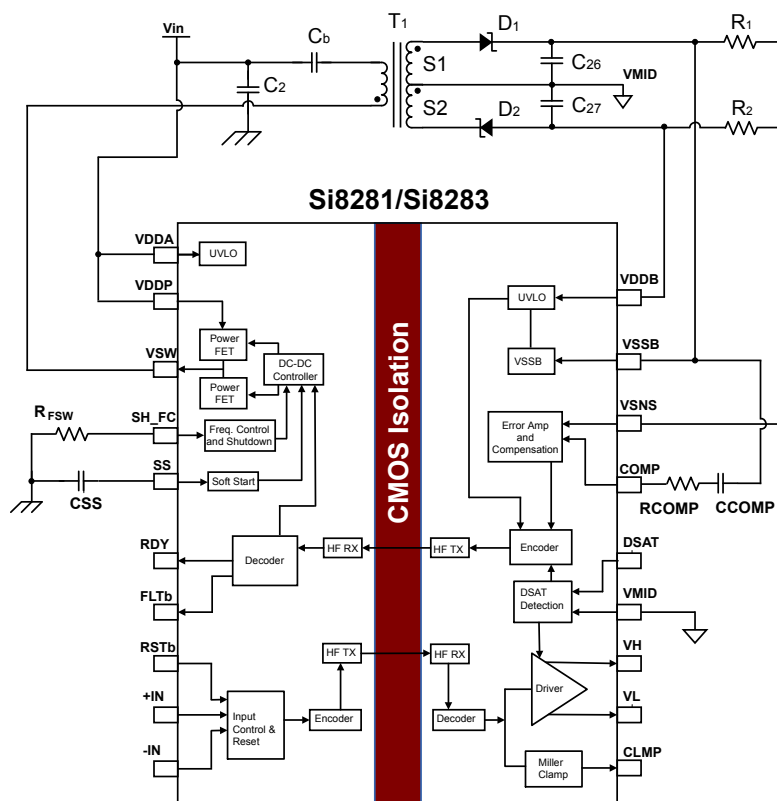
Isolated DC-DC with Internal Switch

The Si8281 and Si8283 products have an integrated isolated gate driver with an isolated dc-dc controller. The controller's internal switch can drive an external transformer directly to generate isolated voltages to power the driver's output. This integration allows the Si8281 and Si8283 products to operate with a single 3-5V input supply while generating isolated positive and negative voltages optimized for MOSFET and IGBT switching operations. This functionality eliminates the complexity of having a separated dc-dc circuit and performance limiting issues when a bootstrapping circuit is used in the driver high-side half-bridge configuration. This application note provides guidance for selecting external components necessary for the operation of the dc-dc controller.

The figure below shows the minimum external components required for the Si8281, Si8283 asymmetric half bridge fly back dc-dc circuit. They include a blocking capacitor C_b , input capacitor C_2 , fly back transformer T_1 , diodes D_1 and D_2 , output capacitors C_{26} and C_{27} , voltage sense resistors R_1 and R_2 , and compensation network components R_{COMP} and C_{COMP} . Note that R_{FSW} and the CSS are used only on the Si8283 product to select the dc-dc switching frequency and set maximum start-up duration.

KEY POINTS

- Theory of Operation
- Power transfer states & State Waveforms
- Blocking Cap and XFRM Power Split
- Transformer Design
- Compensation and Loop Stability
- Input and Output Caps
- Design Example



1. Introduction

The Si8281, Si8283 dc-dc controller is designed to operate with input voltages between 3.3-5.5V and generate isolated supply voltages for the integrated gate driver. The Si8281 operates at a fixed 250 kHz switching frequency while the Si8283 has programmable switching frequency ranging from 100 kHz to 1 MHz through the selection of the external RSW resistor and CSS capacitor. The controller operates in closed loop feedback across the isolation to achieve high level of output voltage accuracy and regulation for output power level up to 2 W with 5 V input and 1W with 3.3 V input. The inrush current is controlled to ramp up smoothly during start up. The Si8281 has a 40 ms soft-start period and the Si8283 soft-start period can be set with the value of the CSS.

The external transformer is driven by the controller's internal switches, and with the proper transformer N ratio, it produces the desired output voltage for a given input voltage level. It is possible to have two secondary windings to generate positive and negative voltages for the gate driver output stage. The controller provides one close loop regulation for the entire transformer secondary windings and the transformer splits the output voltages to generate VDDB and VSSB according to the secondary winding ratio.

2. Theory of Operation

The Figure below shows the Si8281-Si8283 dc-dc application circuit. The left side primary circuit has the internal SH-SL half-bridge power switches to drive the L-C series circuit comprised of the L_p transformer primary winding and the C_b capacitor between V_{in} and GND. The transformer secondary split windings are connected to the output diodes D1 and D2 to generate V_{DDB} and V_{SSB} on C26 and C27 bypass capacitors. The R1 and R2 form the closed loop feedback divider circuit, and RC and CC establish the compensation circuit.

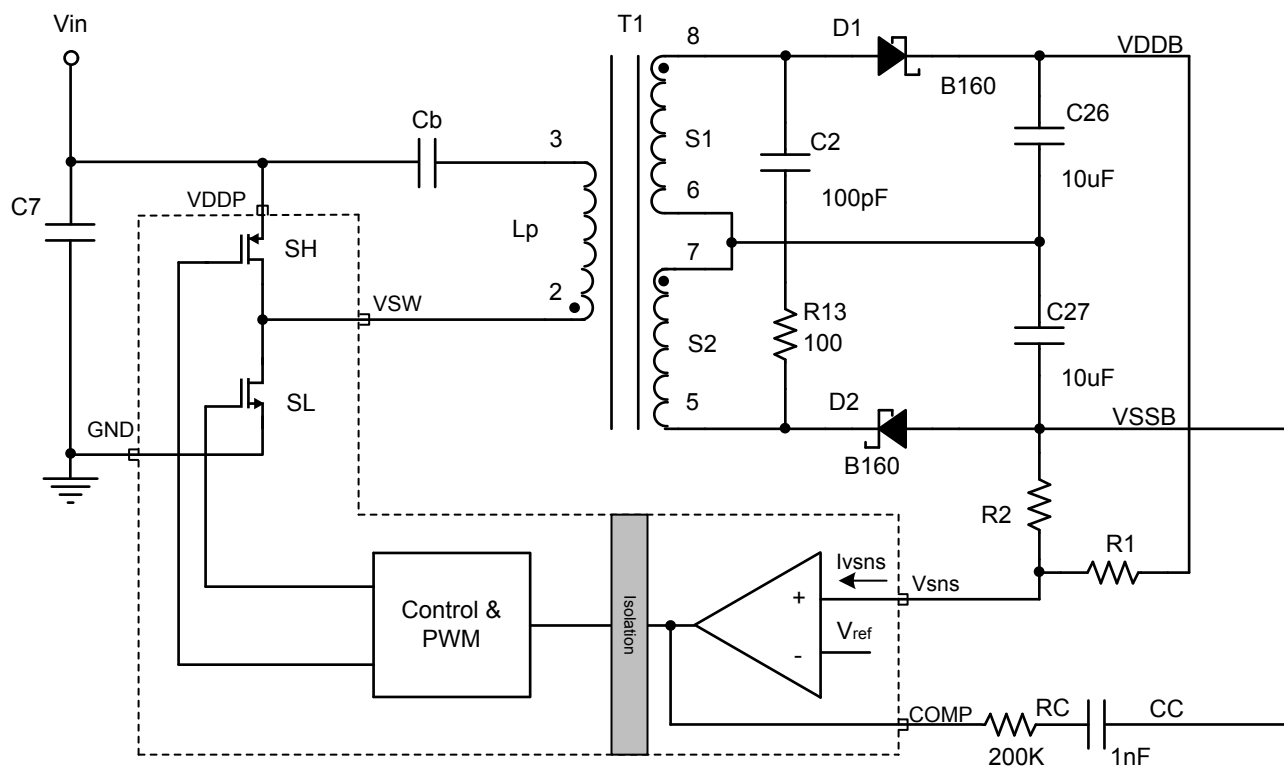


Figure 2.1. Si8281, Si8283 DC-DC Circuits

The Si8281, Si8283 dc-dc converter operates in two main power states. In the first state (t_0 – t_1), current from V_{in} charges the energy-storage elements C_b and L_p and this stored energy is delivered to the output during the second dc-dc state (t_1 – t_3). The sections below discuss the two converter operating states in detail.

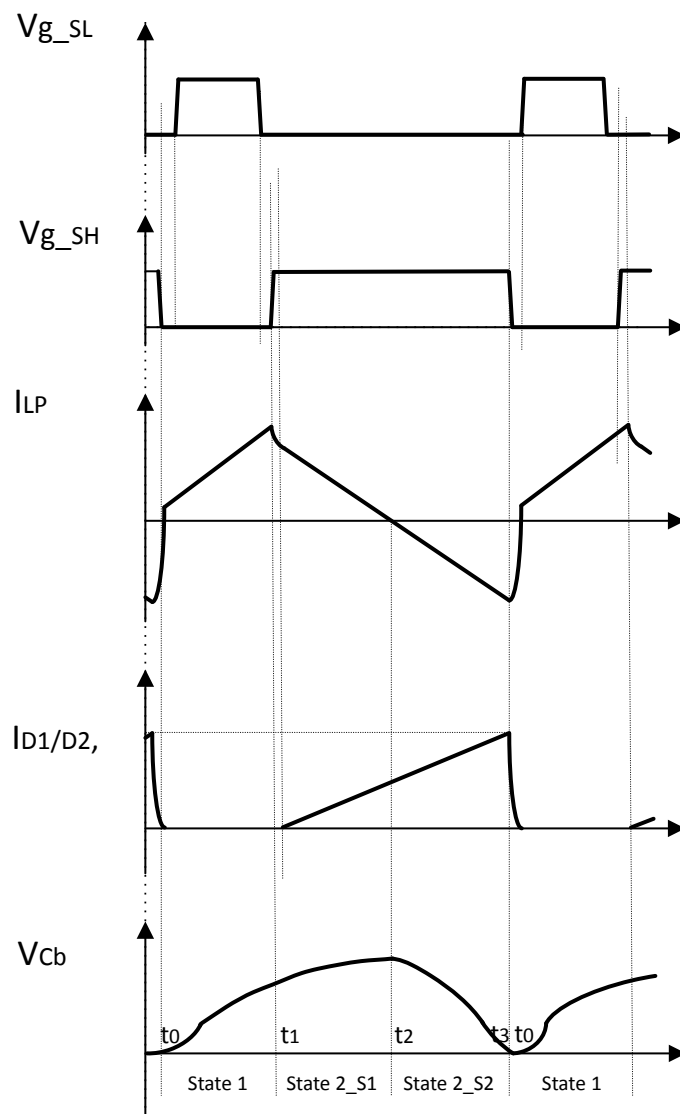


Figure 2.2. Si8281, Si8283 DC-DC State Waveforms

2.1 DC-DC State 1 (t_0-t_1): Storing Energy on Cb and Lp

The Figure below shows the dc-dc current flow during state 1. The SL switch conducts (SH is off) and the current flows starting from Vin through the Cb-Lp series circuit to ground. This current flow stores energy in Cb and Lp causing the voltage, and current to increase in Cb and Lp correspondingly. Notice that in this state, the voltages generated on the secondary windings are in reversed polarity with the diode D1 and D2 and consequently no current flows in the secondary circuits.

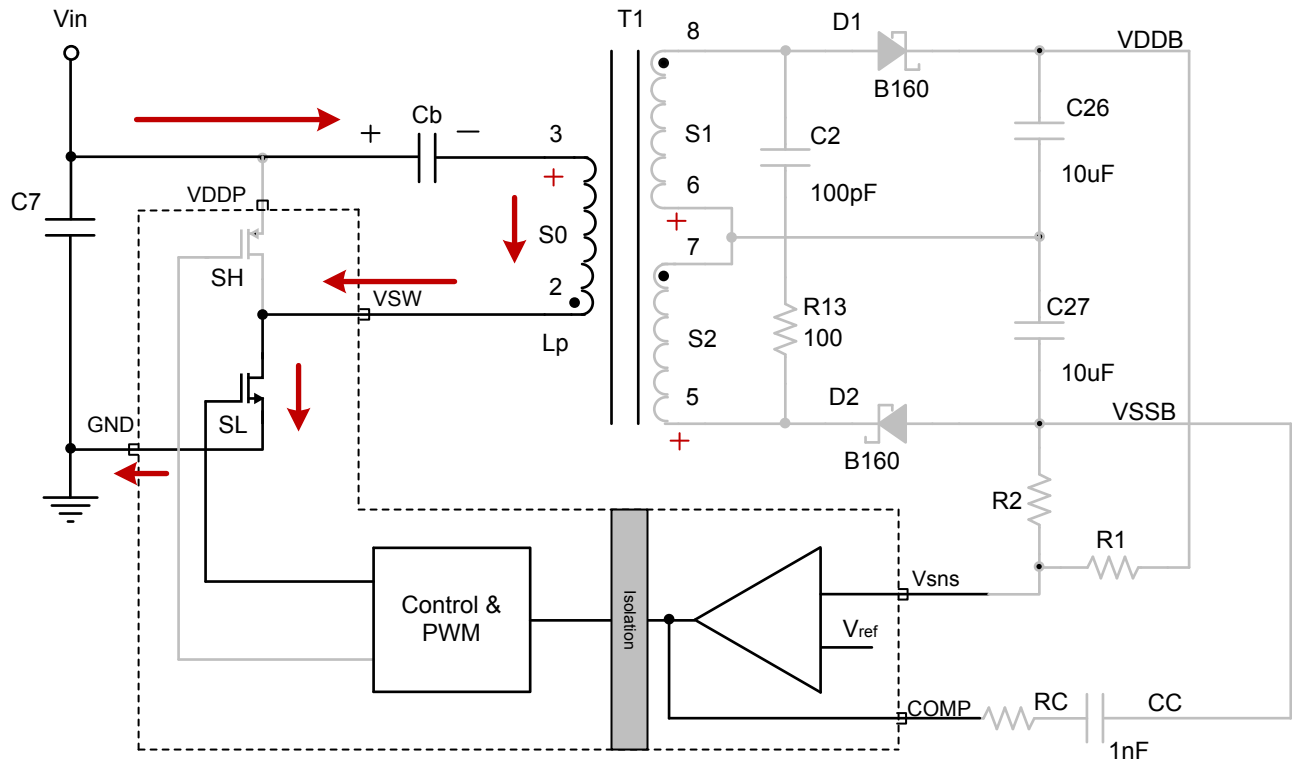


Figure 2.3. Si8281, Si8283 DC-DC State 1

As shown on the figure above, the SL/SH switches, Lp, and Cb form a basic synchronous buck circuit. Therefore, according to the buck converter topology, the voltage across Cb is directly proportional to the SL switch's on time D.

$$V_{Cb} = DV_{in}$$

Equation 1.

For optimum Si8281, Si8283 dc-dc operation it is strongly recommended to keep the duty cycle $D \leq 0.3$.

Assuming the voltage drop across the SL switching device is negligible, the inductor voltage is equal to Vin less V_{Cb}.

$$V_{Lp} = V_{in} - V_{Cb}$$

Equation 2.

Combine Equation 1 and Equation 2, we get the inductor voltage equation:

$$V_{Lp} = V_{in}(1 - D)$$

Equation 3.

The inductance of the primary LP inductor can be expressed as:

$$L_p = \frac{V_{LP}}{I_{LP}} t$$

Equation 4.

The SL switch on time duration Δt is the product of the duty cycle D and period of the switching frequency period $1/F_{sw}$:

$$t = \frac{D}{F_{SW}}$$

Equation 5.

By substituting Δt from [Equation 5](#) and V_{LP} from [Equation 3](#) into [Equation 4](#) and simplifying, we get the transformer primary inductor equation L_P .

$$L_P = \frac{V_{in} D (1 - D)}{F_{SW} I_{LP}}$$

Equation 6.

The current flow during state 1 increases voltage across C_b and the L_P current. Energy storage in these elements can be expressed as.

$$E_{Cb} = \frac{1}{2} C_b V_{Cb}^2$$

Equation 7.

$$E_{LP} = \frac{1}{2} L_P I_{LP}^2$$

Equation 8.

The energies expressed in [Equation 7](#) and [Equation 8](#) are the energies transferred from the input voltage's (V_{in}) charging of the storage components C_b and L_P during dc-dc state 1. These energies will be transferred to the output during the dc-dc state 2 as described in the following sections.

2.2 DC-DC State 2 (t_1 – t_3): Energy Transfer

The Energy Transfer state 2 begins when switch SL transitions to the off state and switch SH is turned on. The stored energies in L_p and C_b during state 1 is transferring to the output in two consecutive steps.

- The first step of power transfer (t_1 – t_2 of [Figure 2.2 Si8281, Si8283 DC-DC State Waveforms on page 4](#)) begins instantly after SL turns off and SH turns on and is illustrated in the figure below. The current on L_p continues to flow in the same direction (as of the end of the dc–dc state 1) but is now diverted to flow across SH switch towards the C_b capacitor. This current flow is energized by the T1 primary inductance and continues to charge the voltage across the C_b capacitor and induces voltages on the secondary windings S1, S2. By the transformer dot convention, the induced voltages have the same polarity as the D1, D2 diodes and generate current flow toward the RL load. The current flow in the T1 transformer primary and secondary windings facilitate the transferring of the L_p stored energy to the C_b capacitor and to the RL load. This first step of power transfer is the transformer primary inductor “fly back” phenomenon and ends when the ILP current in the T1 primary winding reached the zero level (see ILP waveform in [Figure 2.2 Si8281, Si8283 DC-DC State Waveforms on page 4](#)).

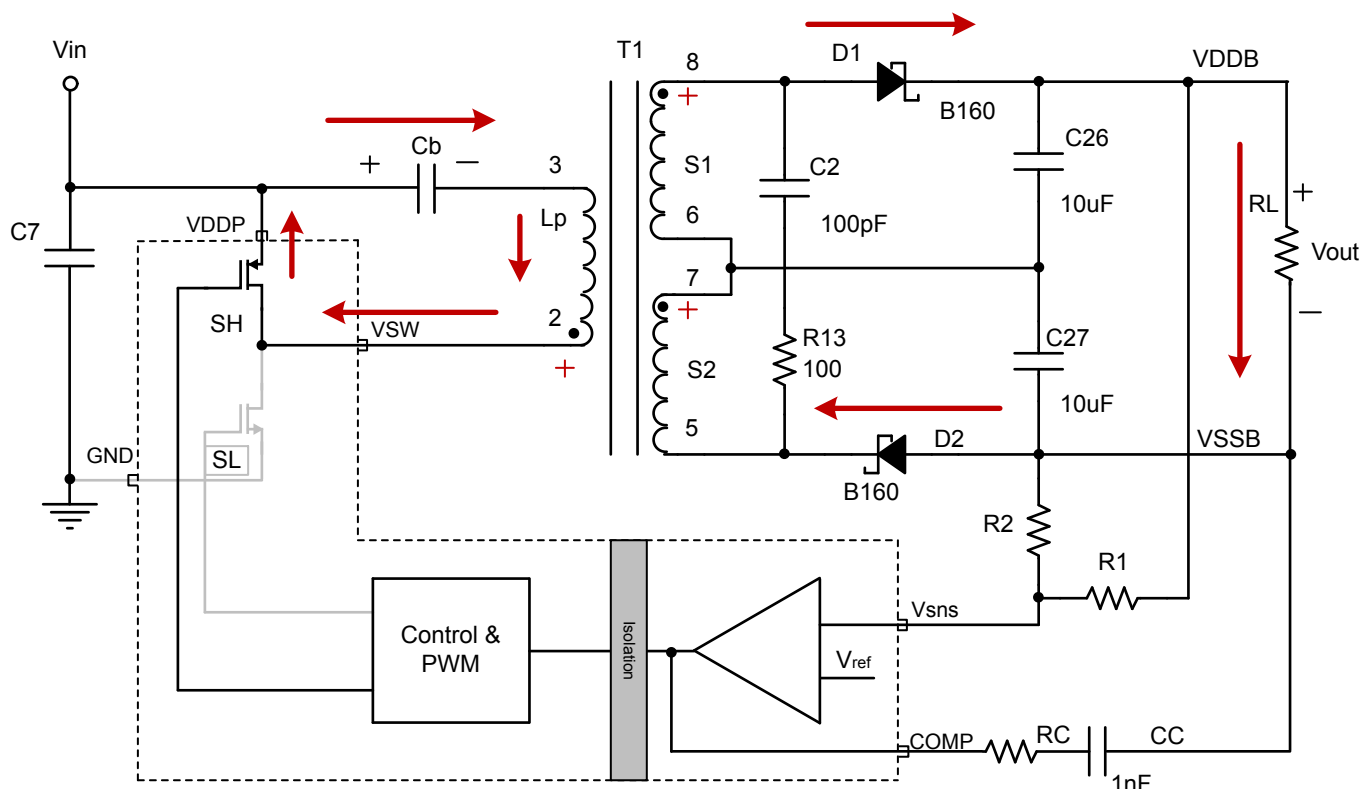


Figure 2.4. Si8281, Si8283 DC-DC State 2, Step 1

- Step 2 of the power transfer begins after the L_p stored energy is exhausted (t_2 – t_3 of [Figure 2.2 Si8281, Si8283 DC-DC State Waveforms on page 4](#)). At this point, the current flow on the T1 primary winding is reversed, and is now powered by the C_b accumulated charge. The figure below shows the current flow during the second power transfer

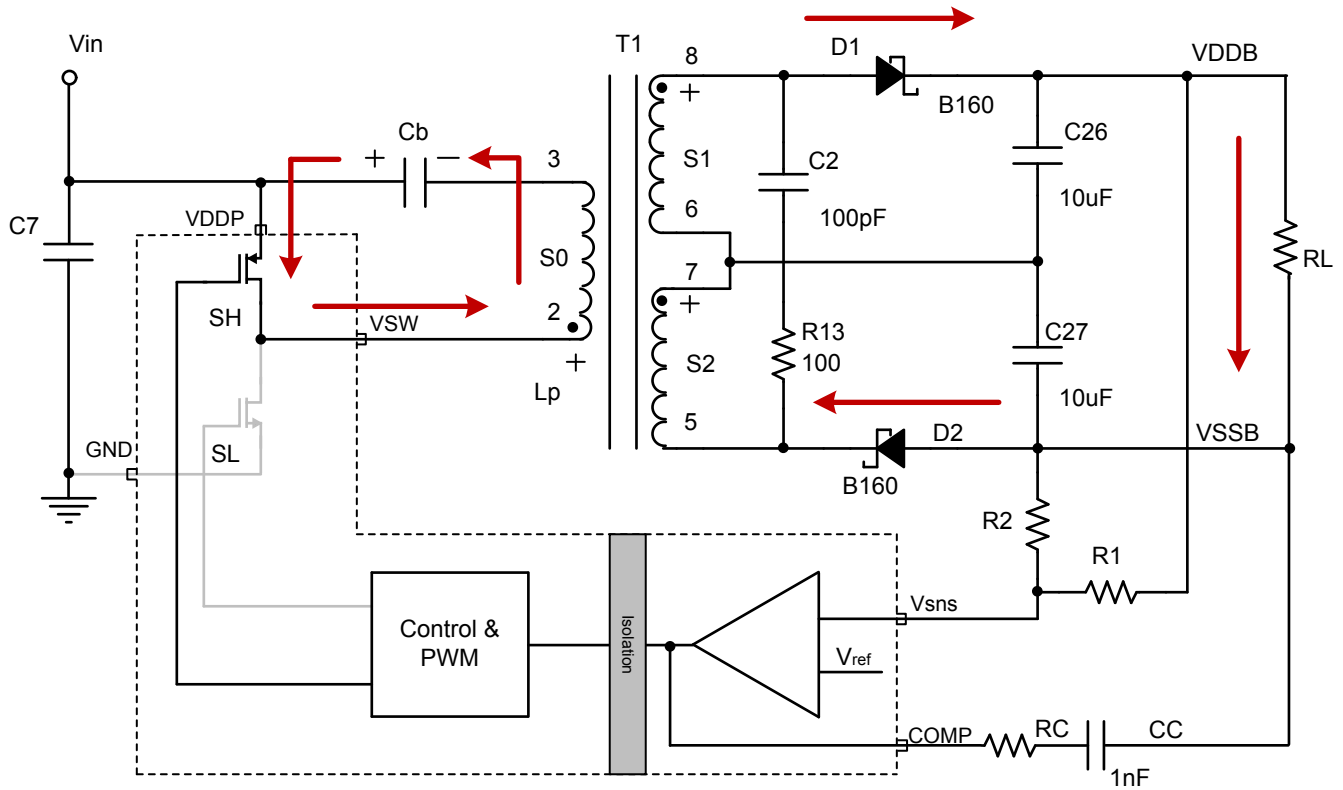


Figure 2.5. Si8281, Si8283 DC-DC State 2, Step 2

In Step 2, the current flow begins at the Cb positive terminal and flows through the SH switch and the transformer primary's winding then back to the Cb negative terminal. This current flow induces voltage on the secondary windings with same polarity as the diodes D1 and D2 and generates current flow to the RL load. This operation is similar to a forward dc-dc converter and therefore has the following primary to secondary voltage transfer equation.

$$V_{DDB} + V_{SSB} = V_{Cb} \times (N_1 + N_2)$$

Where N1 and N2 are transformer ratios: $N_1 = S1/S0$, and $N_2 = S2/S0$

Combining the above equation with the buck voltage transfer equation 1, we get the voltage transfer equation from Vin to output voltages VDDB and VSSB.

$$V_{DDB} + V_{SSB} = DV_{in}(N_1 + N_2)$$

Equation 9.

By dividing the secondary voltage according to the corresponding windings, one can calculate the voltage transfer equations for VDDB, and VSSB as shown below.

$$V_{DDB} = DV_{in}N_1$$

$$\text{therefore: } N_1 = \frac{V_{DDB}}{DV_{in}}$$

Equation 10.

$$V_{SSB} = -DV_{in}N_2$$

$$\text{therefore: } N_2 = \frac{V_{SSB}}{DV_{in}}$$

Equation 11.

The Step 2 power transfer ends when the charge in the Cb capacitor is exhausted or the switching cycle ends, whichever comes first. The converter then starts a new switching cycle begins with dc-dc state 1 and the energy transfer scheme starts anew.

2.3 Sizing the Energy-Storage Elements, C_b and L_p

In the steady state, the total energy charging the primary side component C_b and L_p is the same as the energy transferred to the secondary output in every cycle. For the relationship between the output power and the required storage energy on C_b and L_p , it has been shown by Tso-Min Chen and Chern-Lin Chen in an IEEE publication (see [5. References](#)) that:

$$E_{Cb} = P_o T_S D$$

Equation 12.

$$E_{LP} = P_o T_S (1 - D)$$

Equation 13.

[Equation 12](#) and [Equation 13](#) indicate the splitting of the storage energy between the C_b capacitor and the L_p primary inductor to support a given PO power output is strictly dependant on the switching duty cycle D .

The subsequent sections in this application note combine these two equations with the input energy-storage [Equation 7](#) and [Equation 8](#) to derive equations for the blocking capacitor C_b and transformer primary inductance L_p .

C_b Equation

Energy charging C_b from the input is the same as energy C_b transfers to the output: Combining [Equation 7](#) and [Equation 12](#):

$$\begin{aligned} \frac{1}{2} C_b V_{Cb}^2 &= P_o T_S D \\ C_b &= \frac{2 P_o T_S D}{V_{Cb}^2} = \frac{2 P_o D}{F_{SW} V_{Cb}^2} \end{aligned}$$

Equation 14.

L_p Equation

Input energy charging L_p is the same as the energy L_p transfers to the output: Combining [Equation 8](#) and [Equation 13](#) gives:

$$\begin{aligned} \frac{1}{2} L_p I_{LP}^2 &= P_o T_S (1 - D) \\ L_p &= \frac{2 P_o (1 - D)}{F_{sw} I_{LP}^2} \end{aligned}$$

Equation 15.

Combining [Equation 15](#) and [Equation 6](#) and simplifying gives:

$$\begin{aligned} \frac{V_{in} D (1 - D)}{F_{SW} I_{LP}} &= \frac{2 P_o (1 - D)}{F_{SW} I_{LP}^2} \\ I_{LP} &= \frac{2 P_o}{V_{in} D} \end{aligned}$$

Equation 16.

Substituting I_{LP} from [Equation 16](#) to [Equation 15](#) and simplifying gives:

$$L_p = \frac{V_{in}^2 D^2 (1 - D)}{2 P_o F_{SW}}$$

Equation 17.

[Equation 14](#) and [Equation 17](#) provide calculations for sizing the energy-storage element blocking capacitor C_b , and transformer primary inductance L_p base from the input voltage, output power, and dc-dc switching conditions.

2.4 Sizing the Input Capacitor

The input capacitor, C₇, provides current to C_b and L_p during the first state (t₀–t₁) of the dc-dc switching cycle (see current flow in the figure below). The I_{C7} current is governed by Equation 16 and has a discontinuous triangular wave shape as shown below.

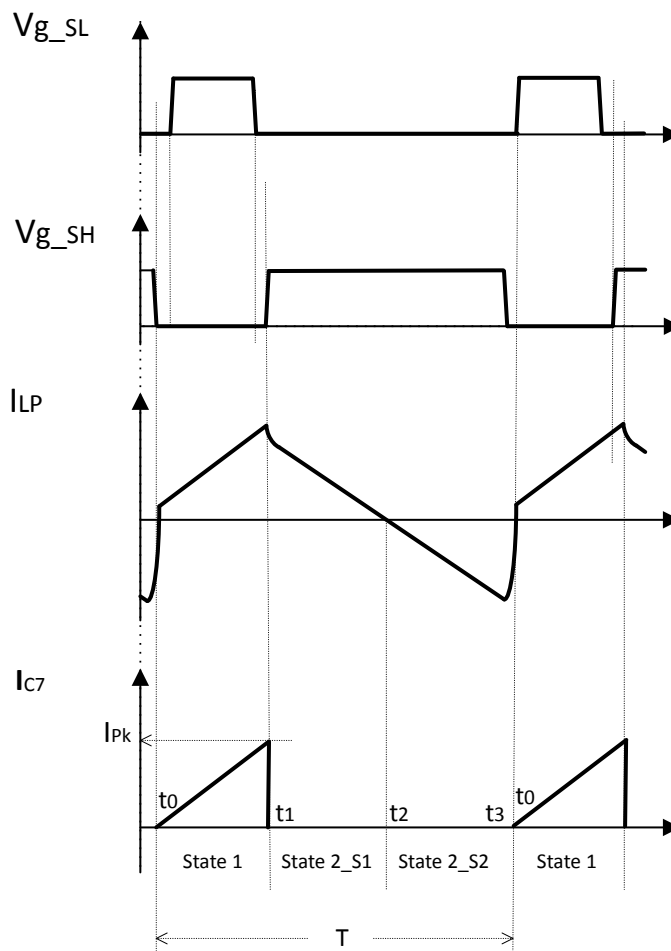


Figure 2.6. Input Capacitor C7 Current Waveform

The standard RMS current equation for a discontinuous triangular waveform is given in Equation 18.

$$I_{C7_rms} = I_{PK} \sqrt{\frac{t1}{3T}}$$

Equation 18.

Substitute I_{pk} from Equation 16 and t₁ = D/F_{SW} to Equation 18 and simplifying gives:

$$I_{C7_rms} = \frac{2P_o}{DV_{in}} \sqrt{\frac{D}{3}}$$

Equation 19.

From which, the equation for C₇ based on the RMS current and voltage is:

$$C_7 = \frac{I_{C7_rms}}{V_{C7_rms}} = \frac{2P_o}{DV_{in}V_{C7_rms}} \sqrt{\frac{D}{3}}$$

Equation 20.

2.5 Sizing the Output Capacitor

On the transformer's secondary windings, current flows through D1 and D2 only during the $(1-D)T_{SW}$ portion (off-time) of the steady state cycle. During the DT_{SW} portion (on-time) of the cycle, I_{LOAD} is sourced solely by the output capacitor C26–27 (C26 in series C27). Therefore, output peak-to-peak voltage ripple on C26–27 can be calculated by:

$$V_{PK-PK_RIPPLE} = \frac{I_{LOAD}DT_{SW}}{C_{26-27}}$$

Which can be rearranged to yield the equation to calculate the output VDD-B-VSSB capacitors:

$$C_{26-27} = \frac{DI_{LOAD}}{F_{SW}V_{PK-PK_RIPPLE}} = \frac{DI_{LOAD}}{\sqrt{2}F_{SW}V_{RMS}}$$

Equation 21.

2.6 Selecting the D1 and D2 Diodes

As illustrated in the figure below, the current flowing through the diode during $(1-D)T_{SW}$ must provide I_{LOAD} during that portion of the cycle plus the charge current for C10 to supply the load during D.

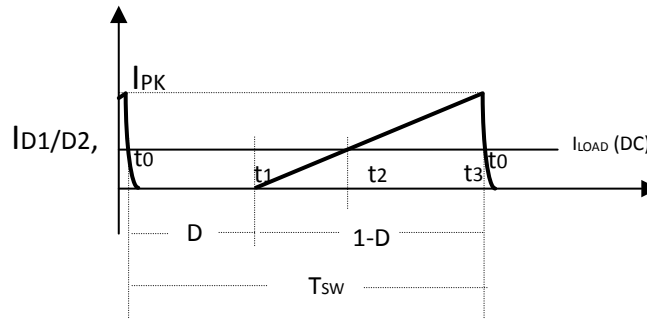


Figure 2.7. Diode Current and I_{LOAD}

During this time, the average I_{LOAD} current must equal the diode current in the triangular waveform (areas under the curve are equal).

$$I_{LOAD}T_{SW} = \frac{I_{PK}(1-D)T_{SW}}{2}$$

Simplifying:

$$I_{PK} = \frac{2I_{LOAD}}{(1-D)}$$

Equation 22.

Also, note that the diode average current is same as output current.

$$I_{D_AV} = I_{LOAD} = \frac{P_{out}}{V_{out}} = \frac{P_{out}}{V_{DDB} + V_{SSB}}$$

Equation 23.

Moreover, the Diode voltage rating must handle the output voltage plus the reversed voltage generated by the transformer during D period:

$$V_{DIODE_MAX} = V_{OUTX} + (V_{in} - DV_{in})N$$

Equation 24.

When operating the converter at high ambient temperatures ($>70^{\circ}\text{C}$), low-leakage rectifying diodes are needed in the power supply output. The following table describes a diode that was tested in the circuit and will fulfill these requirements:

Table 2.1. DC-DC Supply Output Rectifier Diode for High Working Temperatures

Mfr P/N	Mfr / Website	Vf(max)	Leakage Current (Reverse Current, Max)
MBR0580S1	Diodes Inc. www.diodes.com	800 mV @ 0.5 A, 25°C	280 μA typ @ $V_R = 80\text{ V}$, $T_A = +125^{\circ}\text{C}$, Short duration pulse test used to minimize self-heating effect.

The other low-leakage rectifying diodes has been tested was PMEG6020ELR. Since PMEG6020ELR only has 60 V V_R reverse voltage rating, PMEG6020ELR can be only used when the application allows.

2.7 Determining the VSNS Voltage Divider and R1 and R2 Values

Figure 2.1 Si8281, Si8283 DC-DC Circuits on page 3 shows the error amplifier with the negative (–) input connected to the internal Vref = 1.05 V, and the VDDB – VSSB voltage divider R1, R2 circuit connected to the positive (+) input. There is a small leakage current I_{VSNS} into the V_{SNS} pin and the equation for the voltage divider can be expressed by:

$$V_{DDB} - V_{SSB} = V_{ref} \left(\frac{R_1}{R_2} + 1 \right) + (R_1 \times I_{VSNS})$$

From data sheet, Vref = 1.05 V and $I_{VSNS} = \pm 500$ nA. For simplicity, the $(R_1 \times I_{VSNS})$ can be ignored if $R_1 \leq 200$ k Ω ($< \pm 0.1$ V error). The above equation becomes:

$$V_{DDB} - V_{SSB} = 1.05 \left(\frac{R_1}{R_2} + 1 \right)$$

Solving for R2 yields:

$$R_2 = \frac{1.05 \times R_1}{V_{DDB} - V_{SSB} - 1.05}$$

Equation 25.

2.8 Programming Soft Start and Setting the Switching Frequency (Si8283 Only)

The Si8281 soft start period and dc-dc switching frequency are fixed at 40 ms and 250 kHz. The Si8283 soft start and switching frequency are programmable through the external CSS and Rsw. The equations for selecting CSS and Rsw are presented in sections 2.12.7 and 2.12.8 of the Si8281_82_83_84 data sheet.

2.9 Calculating Values for the Compensation Network

The compensation network is comprised of RC and CC connecting between COMP pin and VSSB. RC is fixed to 200 k Ω to match the internal compensation RINT. That leaves the C11 as the main parameter to place the compensation zero in relationship to the cross over frequency to ensure loop stability. Below is the equation for estimating the Si8281, Si8283 dc-dc cross over frequency:

To achieve good phase margin, it is suggested to place the zero between 1/4th to 1/10th of the estimated crossover frequency. The zero placement in the equation below was chosen to lead the crossover frequency by a factor of 6:

$$f_C = \frac{200 \times 10^3 \times 3}{R_1 \times (N1 + N2)} \times \frac{1}{2C_{26} - 27}$$

Equation 26.

$$C_C = \frac{6}{2f_c \times R_c}$$

Equation 27.

2.10 Calculating Values for the RC Snubber Network

During the internal MOSFETs on-off switching transitions, the transformer leakage inductance and the output diode's parasitic capacitors undergo LC resonance ringing and generate high voltage spikes across the internal Si828x switching MOSFETs. To minimize this resonance ringing, a simple RC snubber network (R13+C2) is placed across the transformer secondary winding (see Figure 2.1.). This snubber RC network provides damping effect to the transformer's leakage inductance and diode parasitic capacitor resonance and minimizes voltage ringing during switching. Below are the equations to calculate the values of the RC snubber circuit.

The value of the capacitor C2 is selected to control the resonance ringing while not dissipate too much power on R13 to minimize its effect on the dc/dc efficiency. A general rule of thumb is to set the snubber power dissipation at 0.5% of the total output power dissipation (2 W).

$$P_{R13} = 0.5 \% \times 2 = C_2 V^2 f_{sw} = C_2 (V_{DDB} + |V_{SSB}|)^2 f_{sw}$$

Simplify and solve for C2:

$$C_2 = \frac{0.01}{(V_{DDB} + |V_{SSB}|)^2 f_{sw}}$$

Equation 28.

The RC timing of the R13-C2 need to be sufficiently small to ensure that the charge on C2 is dissipated before the next switching cycle:

$$R_{13} C_2 = \frac{1}{40} T_{sw} = \frac{1}{40 f_{sw}}$$

Solving for R13:

$$R_{13} = \frac{1}{40 C_2 f_{sw}}$$

Equation 29.

3. Design Example

Consider the following design requirements for the Si8281 device:

Table 3.1. Design Requirements

Parameter	Value
Input Voltage	5 V
Input Voltage Ripple	≤ 150 mV (RMS)
VDDDB	17 V
VSSB	-7 V
Output Voltage Ripple	≤ 120 mV (RMS)
Maximum power output, PO	2 W
Si8281 fixed switching frequency, Fsw	250 kHz

3.1 Transformer Design

The design of the primary inductance for the Si8281 asymmetric half bridge fly back transformer is similar to the design of the fly back transformer. The transformer primary inductance must be able to support the desired output power at the given input voltage and switching duty cycle D. However, the design of the secondary windings are similar to the design of the forward converter since the transformers' turns ratios determine the output voltages.

From Equation 15 and selecting D = 0.25 (D ≤ 0.3 condition from equation 1):

$$L_P = \frac{V_{in}^2 D^2 (1-D)}{2P_o F_{sw}} = \frac{52 \times 0.32 \times (1-0.25)}{2 \times 2 \times 250,000} = 1.5 \mu H$$

Equation 31.

From Equation 16, calculate peak current on the primary winding:

$$I_{pk} = \frac{2P_o}{V_{in} D} = \frac{2 \times 2}{5 \times 0.25} = 3.2 A$$

Note: The transformer primary winding saturation current must be greater than 3.2 A.

Equation 32.

From Equation 10 and Equation 11, calculate the transformer turns ratio N1, N2:

$$N_1 = \frac{V_{DDB}}{DV_{in}} = \frac{17}{0.25 \times 5} = 13.6 = > 15$$

$$N_2 = \frac{V_{SSB}}{DV_{in}} = \frac{7}{0.25 \times 5} = 5.6 = > 5$$

Note: Rounding N1 = 15 and N2 = 5 for manufacturability

Equation 33.

Table 3.2. Transformer Parameters Summary

Input Voltage	Turns Ratio	Primary Inductance	Saturation Current	Leakage Inductance
3.0–5.5 V	P:S1:S2-1:15:5	1.5 μH	>3.2 A	>125 ns

The above transformer design has the same characteristics as the UMEC UTB02269s transformers listed in the Si828x data sheet and is used in the Si8281 customer evaluation board. Moreover, note that the value of the transformer leakage inductance is dependant on the construction quality of the transformer and should be kept below 125 nH while the winding resistance (both primary and secondary) should be kept as low as possible for optimum dc-dc efficiency.

3.2 Blocking Capacitor Cb Selection

The blocking capacitor should be sized sufficiently large to transfer the required power to the output. Using Equation 14 and allowing a voltage variation ΔV_{Cb} = 0.7 V across the blocking cap dictates that:

$$C_b = \frac{2P_o D}{F_{sw} V_{Cb}^2} = \frac{2 \times 2 \times 0.25}{250000 \times 0.7^2} = 8.2 \mu F$$

Equation 34.

For this design, round up the capacitance value to Cb = 10 μF and use a ceramic capacitor with a voltage rating of 25 V to ensure low ESR rating

3.3 Input Capacitor Selection

In most applications, VIN also supplies the VDDA pin that powers the dc-dc controller and Si828x left side silicon circuitry. It is recommended to minimize voltage ripple at VDDA = 0.15 V_{RMS}. Solving Equation 20:

$$C_7 = \frac{2P_o}{DV_{in}V_{C7_rms}}\sqrt{\frac{D}{3}} = \frac{2 \times 2}{0.25 \times 5 \times 0.15}\sqrt{\frac{0.25}{3}} = 6.5\mu F$$

For this design, round up the input capacitor value to 10uF and use a 25V rated ceramic capacitor for low ESR rating.

3.4 Output Capacitor Selection

The output capacitors C26 and C27 (C26 + C27 series value) can be set to the same value and can be calculated based on Equation 21.

$$C_{26-27} = \frac{DI_{LOAD}}{\sqrt{2}F_{SW}V_{RMS}} = \frac{D(P_o/(V_{DDB} + V_{SSB}))}{\sqrt{2}F_{SW}V_{RMS}}$$

$$C_{26-27} = \frac{0.25 \times (2/(17+7))}{\sqrt{2} \times 250,000 \times 0.15} = 5\mu F$$

For this design, rounding the series output capacitance value to 5 μF makes the value for C26 and C27 right at 10 μF each. For minimized high frequency ripple, it is recommended to use ceramic capacitors with voltage rating of 25 V for both C₂₆ and C₂₇.

$$C_{26} = C_{27} = 10\mu F$$

3.5 D1 and D2 Diode Selection

Equations 23 and 24 define the parameters for selecting the D1 and D2 diodes.

Calculate diode average current from Equation 23:

$$I_{D_AV} = \frac{P_{out}}{V_{DDB} + V_{SSB}} = \frac{2}{17+7} = 85mA$$

Then calculate peak current from Equation 22:

$$I_{PK} = \frac{2I_{LOAD}}{(1-D)} = \frac{2 \times 85}{(1-0.25)} = 225mA$$

From Equation 24 calculate the voltage rating for D1:

$$V_{D1_MAX} = V_{OUTX} + (V_{in} - DV_{in})N1 = 17 + (5 - 0.25 \times 5) \times 15 = 73 V$$

From Equation 24 calculate the voltage rating for D2:

$$V_{D2_MAX} = V_{OUTX} + (V_{in} - DV_{in})N2 = 7 + (5 - 0.25 \times 5) \times 5 = 26 V$$

For simplicity, the 100 V 1A ES1B diode is selected for both D1 and D2. This diode gives ample headroom for this application. For high working temperature applications, the low-leakage diode is required. Please refer to 2.6 Selecting the D1 and D2 Diodes and Table 2.1 DC-DC Supply Output Rectifier Diode for High Working Temperatures on page 12 for more details.

3.6 VSNS Voltage Divider Calculation

The values of the R1 and R2 determine the VDDB-VSSB voltage level while the transformer winding ratio between the two secondary windings determine the split voltage between VDDB and VSSB. From Equation 25 and selecting R1 = 178 k to keep the voltage error due to the ±500 nA leakage current into the VSNS pin small: Verr = 178 k x ±500 nA = 0.089 V (< 0.1 V). The equation below can be used to calculate R2

$$R_1 = 178k$$

$$R_2 = \frac{1.05 \times R1}{V_{DDB} + V_{SSB} - 1.05} = \frac{1.05 \times 178}{17 - (-7) - 1.05} = 8.14k = 8.25k$$

For this design, round up the R2 value to the nearest 1% resistor of 8.25 kΩ.

3.7 Compensation Network Calculation

Calculate the cross over frequency using Equation 31:

$$f_c = \frac{200 \times 10^3 \times 3}{R_1 \times (N1 + N2)} \times \frac{1}{2C_{26-27}}$$

$$f_c = \frac{200 \times 10^3 \times 3 \times 288}{178 \times 10^3 \times (15 + 5)} \times \frac{1}{2 \times 288 \times 10 \times 10^{-6}} = 2.7 \text{ kHz}$$

Calculate the compensation capacitor CC using Equation 32:

$$C_C = \frac{6}{2f_c \times R_c} = \frac{6}{2 \times 2.7 \times 10^3 \times 200 \times 10^3} = 1.8 \text{ nF}$$

3.8 Calculating Snubber Network C2, R

From Equation 28, calculate C2 value as shown below:

$$C_2 = \frac{0.01}{(V_{DDB} + |V_{SSB}|)^2 f_{sw}} = \frac{0.01}{24^2 \times 250000} = 70 = 100 \text{ pF}$$

For this application, round up C2 value to 100 pF for simple BOM selection.

From Equation 29, calculate R13 value as shown below:

$$R_{13} = \frac{1}{40C_2 f_{sw}} = \frac{1}{40 \times 100 \times 10^{-12} \times 250000} = 100$$

Equation 28 also gives us the power dissipation calculation for R13.

$$P_{R13} = 0.5 \% \times 2 = 0.01 \text{ W}$$

Select SMD resistor package size of 0402 or larger to handle the 0.01 W power dissipation

3.9 Design Summary

The following table lists component selections that meet the design requirements.

Table 3.3. Design Summary

Part Reference	Description	Manufacturer	Manufacturer Part Number
Cb, C26, C27	CAP, 10 μ F, 50 V, \pm 10%, X7R, 1210	Venkel	C1210X7R500-106M
CC	CAP, 1.8 nF, 50 V, \pm 10%, X7R, 0603	Venkel	C0603X7R160-182K
C2	CAP, 100 pF, 50 V, \pm 20%, X7R, 0603	Venkel	C0603C0G500-101M
D1, D2	DIO, FAST, 100 V, 1.0A, DO-214AC	Diodes	ES1B
RC	RES, 200 k Ω , 1/10 W, \pm 1%, ThickFilm, 0603	Venkel	CR0603-10W-2003F
R1	RES, 178 k Ω , 1/10 W, \pm 1%, ThickFilm, 0603	Venkel	CR0603-10W-1783F
R2	RES, 8.25 k Ω , 1/10 W, \pm 1%, ThickFilm, 0603	Venkel	CR0603-10W-8252F
R13	RES, 100 Ω , 1/10 W, \pm 1%, ThickFilm, 0603	Venkel	CR0603-10W-1000F
T1	TRANSFORMER, POWER, FLYBACK, 1.5 μ H PRIMARY, 125nH LEAKAGE, 1:15:5, 2 TAP, SMT	UMEC	UTB02269s

4. Conclusion

This application note provides the user with a comprehensive theory of operation of the Si8281 Si8283 dc-dc converter and a method to calculate the parameters of the external components. The user should consult application note AN1009 as well as the overall driver system design to determine the power design requirements as shown in [Table 3.1 Design Requirements on page 15](#). This application note also provides a design example with simple step-by-step calculations to determine the parameters of the custom transformer as well as the external components. [Table 3.3 Design Summary on page 18](#) provides a typical BOM design that can satisfy most applications.

5. References

Tso-Min Chen and Chern-Lin Chen, *Characterization of Asymmetrical Half Bridge Flyback Converter*, 2002 IEEE 33rd Annual Power Electronics Specialists Conference, Volume 4 Conference Proceedings, pgs. 921-926 (2002)

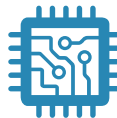


**Connecting Everyone
and Everything,
All the Time**



Portfolio

www.skyworksinc.com



Quality

www.skyworksinc.com/quality



Support & Resources

www.skyworksinc.com/support

Copyright © 2021 Skyworks Solutions, Inc. All Rights Reserved.

Information in this document is provided in connection with Skyworks Solutions, Inc. ("Skyworks") products or services. These materials, including the information contained herein, are provided by Skyworks as a service to its customers and may be used for informational purposes only by the customer. Skyworks assumes no responsibility for errors or omissions in these materials or the information contained herein. Skyworks may change its documentation, products, services, specifications or product descriptions at any time, without notice. Skyworks makes no commitment to update the materials or information and shall have no responsibility whatsoever for conflicts, incompatibilities, or other difficulties arising from any future changes.

No license, whether express, implied, by estoppel or otherwise, is granted to any intellectual property rights by this document. Skyworks assumes no liability for any materials, products or information provided hereunder, including the sale, distribution, reproduction or use of Skyworks products, information or materials, except as may be provided in Skyworks' Terms and Conditions of Sale.

THE MATERIALS, PRODUCTS AND INFORMATION ARE PROVIDED "AS IS" WITHOUT WARRANTY OF ANY KIND, WHETHER EXPRESS, IMPLIED, STATUTORY, OR OTHERWISE, INCLUDING FITNESS FOR A PARTICULAR PURPOSE OR USE, MERCHANTABILITY, PERFORMANCE, QUALITY OR NON-INFRINGEMENT OF ANY INTELLECTUAL PROPERTY RIGHT; ALL SUCH WARRANTIES ARE HEREBY EXPRESSLY DISCLAIMED. SKYWORKS DOES NOT WARRANT THE ACCURACY OR COMPLETENESS OF THE INFORMATION, TEXT, GRAPHICS OR OTHER ITEMS CONTAINED WITHIN THESE MATERIALS. SKYWORKS SHALL NOT BE LIABLE FOR ANY DAMAGES, INCLUDING BUT NOT LIMITED TO ANY SPECIAL, INDIRECT, INCIDENTAL, STATUTORY, OR CONSEQUENTIAL DAMAGES, INCLUDING WITHOUT LIMITATION, LOST REVENUES OR LOST PROFITS THAT MAY RESULT FROM THE USE OF THE MATERIALS OR INFORMATION, WHETHER OR NOT THE RECIPIENT OF MATERIALS HAS BEEN ADVISED OF THE POSSIBILITY OF SUCH DAMAGE.

Skyworks products are not intended for use in medical, lifesaving or life-sustaining applications, or other equipment in which the failure of the Skyworks products could lead to personal injury, death, physical or environmental damage. Skyworks customers using or selling Skyworks products for use in such applications do so at their own risk and agree to fully indemnify Skyworks for any damages resulting from such improper use or sale.

Customers are responsible for their products and applications using Skyworks products, which may deviate from published specifications as a result of design defects, errors, or operation of products outside of published parameters or design specifications. Customers should include design and operating safeguards to minimize these and other risks. Skyworks assumes no liability for applications assistance, customer product design, or damage to any equipment resulting from the use of Skyworks products outside of Skyworks' published specifications or parameters.

Skyworks, the Skyworks symbol, Sky5®, SkyOne®, SkyBlue™, Skyworks Green™, Clockbuilder®, DSPLL®, ISOModem®, ProSLIC®, and SiPHY® are trademarks or registered trademarks of Skyworks Solutions, Inc. or its subsidiaries in the United States and other countries. Third-party brands and names are for identification purposes only and are the property of their respective owners. Additional information, including relevant terms and conditions, posted at www.skyworksinc.com, are incorporated by reference.

Skyworks Solutions, Inc. | Nasdaq: SWKS | sales@skyworksinc.com | www.skyworksinc.com

USA: 781-376-3000 | Asia: 886-2-2735 0399 | Europe: 33 (0)1 43548540 | [in](#) [f](#) [t](#) [y](#)