

## Introduction

The MAXREFDES1256 is an isolated power supply capable of delivering 7V at up to 7A load current. High efficiencies are achieved by using the MAX17606 MOSFET driver on the secondary side for synchronous rectification. The input voltage range of the MAXREFDES1256 is 9V to 36V DC.

This design uses the MAX17597 peak-current-mode controller in the discontinuous-conduction mode (DCM) in a flyback topology at 100kHz.

The transformer provides a galvanic isolation of 1425VDC between the primary and secondary. Also, an optocoupler provides secondary side feedback to the controller for voltage regulation while maintaining isolation. This document also discusses a step-by-step procedure to implement this reference design and presents design equations to calculate all the design parameters. These equations show how to calculate the component values to modify this design for applications. The key features of the MAXREFDES1085 are:

- Input voltage range of 9V to 36V
- Synchronous rectification on the secondary side
- Isolated output of 7V at 49W
- Cycle-by-cycle current limit
- Resistor programmable UVLO/OVI threshold
- Galvanic isolation up to 1425V DC
- Proven PCB layout
- Fully assembled and tested

**Table 1. Design Specification**

PARAMETER	SYMBOL	MIN	MAX
Input Voltage	$V_{IN}$	9V	36V
Frequency	$f_{SW}$	100kHz	
Maximum Efficiency	$\eta$	89%	
Output Voltage	$V_{OUT}$	7V	
Output Voltage Ripple	$\Delta V_{OUT}$	1% of $V_{OUT}$ (max)	
Output Current	$I_{OUT}$	0	7A
Output Power	$P_{OUT}$	49W	

## Designed–Built–Tested

This document describes the hardware shown in [Figure 1](#). It provides a detailed systematic technical guide to designing a DC-DC discontinuous conduction mode (DCM) flyback using Maxim's MAX17597 current-mode controller. The power supply has been built and tested, details of which follow later in this document.

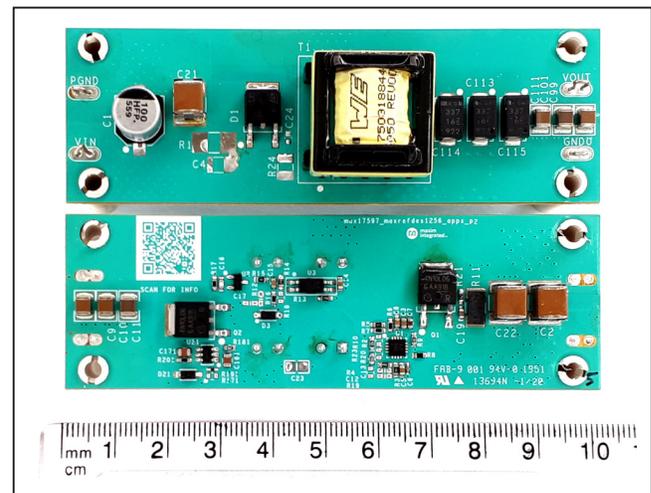


Figure 1. MAXREFDES1256 hardware.

## Generic Isolated Power Supply

Figure 2 shows a generic isolated power-supply block diagram. It consists of a power stage, an isolation transformer, rectifier, secondary-side error amplifier, and opto-coupler to provide a feedback for the primary side control. Different isolated power supplies are different depending upon how the transformer is being used in them.

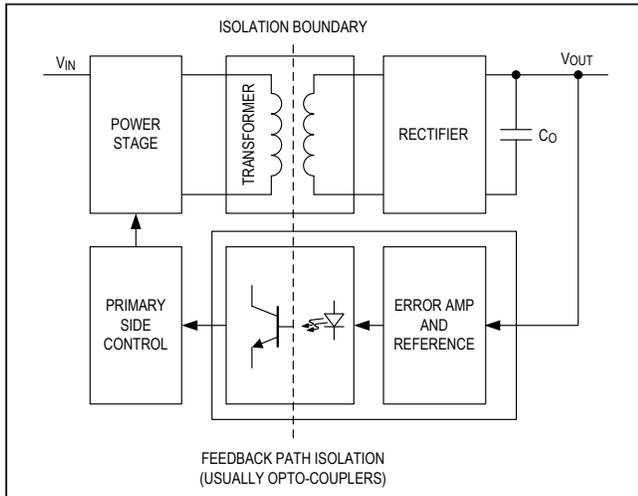


Figure 2. Generic isolated power supply.

## Flyback Principle

A transformer in a flyback configuration acts differently than its usual operation of transformation of energy from primary to secondary. During a transformer's usual operation, both primary and secondary windings conduct together at the same time to make the transfer of energy possible from primary to secondary. In a flyback configuration the primary and secondary windings do not conduct at the same time and the transformer acts more like a coupled inductor. Note that in this document we have used the following notations for the transformer turns ratio:

$$K = \frac{N_P}{N_S}$$

$$k = \frac{N_S}{N_P}$$

This means capital K for primary turns/secondary turns and small k for secondary turns/primary turns.

Figure 3 shows a simple flyback topology that consists of a transformer whose primary winding is connected to the drain of a switching MOSFET. The source of the MOSFET is connected to ground. The secondary winding is connected to the output capacitor through a rectifier diode. In this flyback configuration the current flows into the primary winding during the on time of the switching period and flows into the secondary winding during the off time of the switching period.

During the on-time when the primary switch is closed, a current,  $I_P$ , flows through the primary winding as shown in Figure 4.  $I_P$  can be written as follows:

$$I_P(t) = \frac{1}{L_P} \int_0^t V_{IN} dt = \frac{1}{L_P} V_{IN} t$$

The peak magnitude of the primary current can be written as follows:

$$I_{P-P} = \frac{1}{L_P} \int_0^{t_{ON}} V_{IN} dt = \frac{1}{L_P} V_{IN} t_{ON}$$

In the secondary winding, a negative voltage is induced due to the current flowing in to the primary. The rectifier diode is reverse-biased and no current is flowing in the secondary winding. The induced voltage in the primary can be written as:

$$V_S(t) = L_S \times \frac{dI_P(t)}{dt}$$

During the off-time when the primary switch opens as shown in Figure 5, the magnetic field in the primary winding collapses and the voltage at the winding reverses, while current keeps flowing in the same direction until the field fades away.

The secondary current  $I_S$  flows and the secondary and rectifier diode is forward-biased. Output voltage  $V_{OUT}$  is now available across the secondary coil if we ignore the forward voltage drop of the rectifier diode. The secondary winding voltage is now flown away to primary side as  $K \times V_{OUT}$ . This voltage is present across the switch until the current in the secondary winding decays to zero. Total voltage available across the switch during the off-time can be written as:

$$V_{SW} = V_{IN} + K \times V_{OUT}$$

This voltage also causes the breakdown of the magnetic flux in the primary winding (no current is flowing in the primary winding after this reset). Here we can see that unlike a usual transformer action where current flows in both the winding at the same time, in a flyback transformer the current flows into the primary winding during the on-time and into the secondary winding during the off-time. This is why we use the term "coupled storage inductor" for transformers used in flyback operation. It should be noted though that mechanically these transformers are like any transformer. Use in flyback operations makes transformers act differently as coupled inductors. The required duty cycle for a given input voltage and output voltage can be calculated from:

$$D = \frac{V_{OUT}}{V_{OUT} + V_{IN}}$$

where:

$$V_{OUT} = (V_{OUT} + V_F) \times \frac{N_P}{N_S}$$

Figure 6 shows a typical CCM mode flyback primary and secondary winding current, and Figure 7 shows a typical DCM mode flyback waveform.

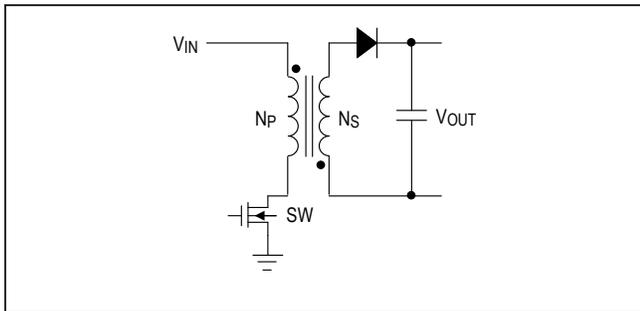


Figure 3. Simple flyback topology.

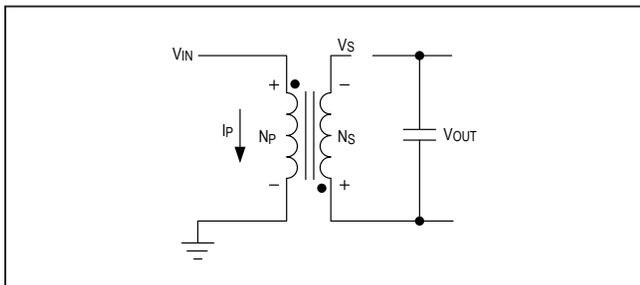


Figure 4. Flyback topology during on-time,  $t_{ON}$ .

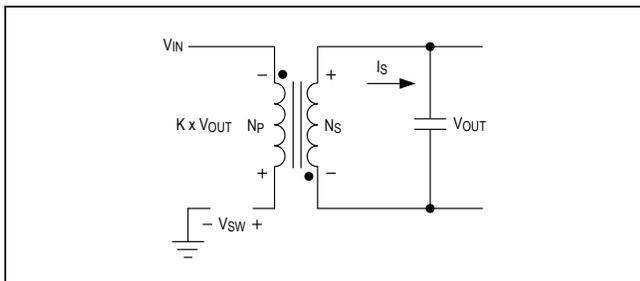


Figure 5. Flyback topology during off-time,  $t_{OFF}$ .

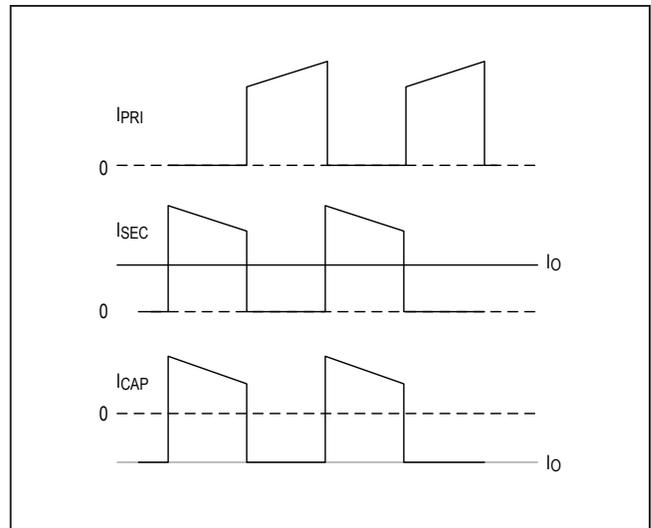


Figure 6. A typical CCM mode flyback primary and secondary winding current.

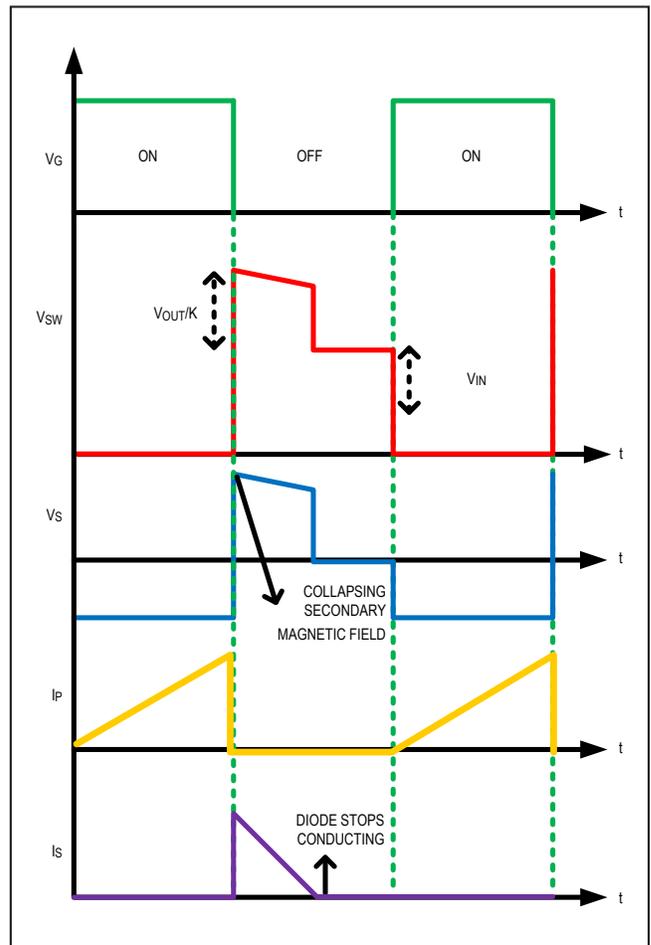


Figure 7. A typical DCM mode flyback topology waveform.

## Design Procedure for DC-DC Flyback Using MAX17597

Now that the basic principle of the DCM flyback is understood, a practical design can be illustrated. The design parameters are obtained by using expressions given in Maxim **Application Note 5504**. This document is primarily concerned with the power stage and the feedback loop design, and is intended to complement the information contained in the MAX17597 data sheet.

Flyback converters can be operated in discontinuous conduction mode (DCM) or continuous conduction mode (CCM). The component choices, stress level in power devices, and controller design vary depending on the operating mode of the converter. The design discussed in this document is a DCM design and expressions for calculating component values and ratings are presented to achieve the design goals.

### Step 1: Switching Frequency

For this design we selected a 100kHz switching frequency. The MAX17597 switching frequency is programmable between 100kHz and 1000kHz with a resistor  $R_{RT}$  connected between RT and SGND.  $R_{RT}$  is calculated as follows:

$$R_{RT} = \frac{10^{10}}{f_{SW}} \Omega$$

$$R_{RT} = \frac{10^{10}}{100k} = 100k\Omega$$

A standard 100k $\Omega$  resistor is selected for  $R_{RT}$ .

### Step 2: Transformer Magnetizing Inductance and Turns Ratio

In a DCM flyback converter, the energy stored in the primary inductance of the flyback transformer is delivered entirely to the output. The maximum primary-inductance value for which the converter remains in DCM at all operating conditions can be calculated as:

$$L_{PRI} \leq \frac{0.4 \times (V_{IN(MIN)} \times D_{MAX})^2}{(V_{OUT} + V_D) \times I_{OUT} \times f_{SW}}$$

where:

$$D_{MAX} = 0.9$$

$V_D = 0.1V$  since synchronous rectification is being used through the MAX17606 secondary MOSFET driver.

In this application, the DC input voltage varies from 9V DC to 36V DC. Substitute the values in the expression of  $L_{PRI}$  as follows:

$$L_{PRI} \leq \frac{0.4 \times (9 \times 0.9)^2}{(7 + 0.1) \times 7 \times 100k} = 5.28\mu H$$

For our design  $L_{PRI}$  is chosen as 1.3 $\mu H$ ,  $L_{PRI} = 1.3\mu H$ .

The leakage inductance of the transformer should be targeted as low as possible. For this design, we achieved a 1.3% leakage inductance of 18nH,  $L_{LKG} = 18nH$ . A customized transformer 750318844 from Würth Elektronik is used in this design. This transformer also fulfills the specification of turns ratio and primary/secondary currents requirement of the design that is calculated step by step in this document. The transformer has dielectric isolation specification of 1425V DC.

### Step 3: Maximum Duty-Cycle Calculation with Selected $L_{PRI}$

Use the following expressions to calculate the maximum duty cycle of the converter for the selected frequency and magnetizing inductance:

$$D_{NEW} = \frac{\sqrt{2.5 \times L_{PRI} \times V_{OUT} \times I_{OUT} \times f_{SW}}}{V_{IN(MIN)}}$$

$$D_{NEW} = \frac{\sqrt{2.5 \times 1.3\mu \times 7 \times 7 \times 100k}}{9} = 0.443$$

Calculate the required transformer turns ratio (k) using the expressions as follows:

$$k = \frac{N_s}{N_p} = \frac{(V_{OUT} + V_D) \times (1 - D_{NEW})}{D_{NEW} \times V_{IN(MIN)}}$$

$$k = \frac{N_s}{N_p} = \frac{(7 + 0.1) \times (1 - 0.443)}{0.443 \times 9} = 0.99$$

For the present design, k is chosen as 1:1.

#### Step 4: Calculation of Peak/RMS Current

Primary and secondary RMS and primary peak currents calculations are needed to design the transformer in switched-mode power supplies. Also, primary peak current is used in setting the current limit. Use the following expressions to calculate the primary and secondary peak and RMS currents:

$$I_{PRIPEAK} = \frac{V_{IN(MIN)} \times D_{NEW}}{L_{PRI} \times f_{SW}} = \frac{9 \times 0.443}{1.3 \mu \times 100 \text{kHz}} = 30.70 \text{A}$$

$$I_{PRI RMS} = I_{PRIPEAK} \times \sqrt{\frac{D_{NEW}}{3}} = 30.7 \times \sqrt{\frac{0.443}{3}} = 11.8 \text{A}$$

$$I_{SECPEAK} = \frac{I_{PRIPEAK}}{k} = \frac{30.7}{0.99} = 30.99 \text{A}$$

$$I_{SECRMS} = \sqrt{\frac{2 \times I_{OUT} \times I_{PRIPEAK}}{3 \times k}} = 12.03 \text{A}$$

#### Step 5: Current-Limit Resistor Calculation

For the current-limit setting, the peak current can be calculated as follows:

$$I_{LIM} = 1.2 \times I_{PRIPEAK} = 1.2 \times 30.7 = 36.84 \text{A}$$

The device includes a robust overcurrent protection scheme that protects the device under overload and short-circuit conditions. A current-sense resistor, connected between the source of the MOSFET and PGND, sets the peak current limit. The current-limit comparator has a voltage trip level ( $V_{CS-PEAK}$ ) of 300mV. Use the following equation to calculate the value of  $R_{CS}$ :

$$R_{CS} = \frac{305 \text{m}}{I_{MOSFET}} = \frac{305 \text{m}}{36.84} = 8.28 \text{m}\Omega$$

where  $I_{MOSFET}$  is the peak current flowing through the MOSFET. A typical 8m $\Omega$  current-sense resistor is selected,  $R_{CS} = 8 \text{m}\Omega$ .

#### Step 6: MOSFET Selection

MOSFET selection criteria includes maximum drain voltage, peak/RMS current in the primary, and the maximum allowable power dissipation of the package without exceeding the junction temperature limits. The voltage seen by the MOSFET drain is the sum of the input voltage, the reflected secondary voltage on the transformer primary, and the leakage inductance spike. The MOSFET's absolute maximum  $V_{DS}$  rating must be higher than the worst-case drain voltage as follows:

$$V_{DS(MAX)} = V_{IN(MAX)} + \left( \frac{2.5 \times (V_{OUT} + V_D)}{k} \right)$$

$$V_{DS(MAX)} = 36 + \left( \frac{2.5 \times (7 + 0.1)}{0.99} \right) = 54 \text{V}$$

For this application we chose the Infineon IPD90N10S4L-06 100V, 90A n-channel MOSFET as the primary MOSFET.

#### Step 7: Snubber Selection

An RCD snubber reduces the maximum voltage stress on the MOSFET by clamping the voltage level. However, it also dissipates power and reduces efficiency. RCD snubbers may not always be required, however, it is always a good idea to leave place holders in the board for RCD and RC snubbers. Ideally, the external MOSFET experiences a drain-source voltage stress equal to the sum of the input voltage and reflected voltage across the primary winding during the off period of the MOSFET. In practice, parasitic inductors and capacitors in the circuit, such as leakage inductance of the flyback transformer, cause voltage overshoot and ringing in addition to the ideally expected voltage stress. Snubber circuits are used to limit the voltage overshoots to safe levels within the voltage rating of the external MOSFET. The snubber capacitor can be calculated using the following expression:

$$C_{SNUB} = \frac{2 \times L_{LK} \times I_{PRIPEAK}^2 \times k^2}{V_{OUT}^2}$$

$$C_{SNUB} = \frac{2 \times 25 \text{n} \times 30.7^2 \times 0.99^2}{5^2} = 942.95 \text{nF}$$

Considering the derating of the capacitor, we selected a 2.2 $\mu\text{F}$  capacitor,  $C_{SNUB} = 2.2 \mu\text{F}$ .

It was found after testing the board that the maximum  $V_{DS}$  voltage was measured to be 73V at maximum load and maximum input voltage. Since the MOSFET is rated for 100V, this design can be used without a RCD Snubber to achieve higher efficiencies.

The power that must be dissipated in the snubber resistor is calculated using the following expressions:

$$P_{\text{SNUB}} = 0.833 \times L_{\text{LKG}} \times I_{\text{PRIPEAK}}^2 \times f_{\text{SW}}$$

$$P_{\text{SNUB}} = 0.833 \times 0.25 \times 30.7^2 \times 100\text{k} = 1.962\text{W}$$

The snubber resistor is calculated based on the below expression:

$$R_{\text{SNUB}} = \frac{6.25 \times V_{\text{OUT}}^2}{P_{\text{SNUB}} \times k^2} = \frac{6.25 \times 5^2}{1.962 \times 0.99^2} = 159\Omega$$

A standard 150Ω, 3W resistor is selected,  $R_{\text{SNUB}} = 150\Omega$ .

The voltage rating of the snubber diode is:

$$V_{\text{DSNUB}} = V_{\text{IN(MAX)}} + \left(2.5 \times \frac{V_{\text{OUT}}}{k}\right)$$

$$V_{\text{DSNUB}} = 36 + \left(2.5 \times \frac{7}{0.99}\right) = 53.67\text{V}$$

We selected the STMicroelectronics STTH4R02B 200V, 4A diode as the snubber diode.

### Step 8: Secondary MOSFET Selection

The maximum operating drain-source voltage rating of the secondary MOSFET must be higher than the sum of the output voltage and the reflected input voltage. We use the following expression to calculate the secondary MOSFET voltage rating:

$$V_{\text{SEC}} = 1.25 \times (k \times V_{\text{IN(MAX)}} + V_{\text{OUT}})$$

$$V_{\text{SEC}} = 0.99 \times (0.43 \times 36 + 7) = 53.3\text{V}$$

We selected Infineon's IPD90N10S4L-06 100V MOSFET as the secondary side for synchronous rectification using MAX17606.

### Step 9: Feedback Resistor ( $R_U$ , $R_B$ ) Selection

A standard 10kΩ resistor is selected,  $R_B = 10\text{k}\Omega$ .

$$R_U = \left(\frac{V_{\text{OUT}}}{V_{\text{REF}}} - 1\right) \times R_B$$

where  $V_{\text{REF}}$  is the reference set by the secondary-side controller ( $V_{\text{REF}} = 2.5\text{V}$  for TL431AQDBZRQ1 is used in this design).

$$R_U = \left(\frac{7}{2.5} - 1\right) \times 10\text{k} = 18\text{k}\Omega$$

A standard 18kΩ resistor is selected,  $R_U = 18\text{k}\Omega$ .

### Step 10: Soft-Start Capacitor

The soft-start period for the devices can be programmed by selecting the value of the capacitor  $C_{\text{SS}}$  connected from the SS pin to SGND. Capacitor  $C_{\text{SS}}$  can be calculated as:

$$C_{\text{SS}} = 8.264 \times t_{\text{SS}}$$

where  $t_{\text{SS}}$  is expressed in ms and the resultant value of  $C_{\text{SS}}$  is in nF.

$$C_{\text{SS}} = 8.264 \times t_{\text{SS}} = 8.264 \times 12 = 99.17\text{nF}$$

A standard 100nF is selected as the soft-start capacitor,  $C_{\text{SS}} = 100\text{nF}$ .

### Step 11: Input Capacitor Selection

For DC-DC applications, X7R ceramic capacitors are recommended due to their stability over the operating temperature range. The effective series resistance (ESR) and effective series inductance (ESL) of a ceramic capacitor are relatively low, so the ripple voltage is dominated by the capacitive component. For the flyback converter, the input capacitor supplies the current when the main switch is on. We used the following equation to calculate the input capacitor for a specified peak-to-peak input switching ripple:

$$C_{\text{IN}} = \frac{D_{\text{NEW}} \times I_{\text{PRIPEAK}} \left[1 - (0.5 \times D_{\text{NEW}})\right]^2}{2 \times f_{\text{SW}} \times V_{\text{IN\_RIP}}}$$

### Step 12: Output Capacitor Selection

X7R ceramic output capacitors are preferred in industrial applications due to their stability over temperature. The output capacitor is usually sized to support a step load of a certain percentage of the rated output current so that the output voltage deviation is contained to 3% of the rated output voltage. The output capacitance can be calculated by using the below expressions:

$$t_{\text{RESPONSE}} \cong \left(\frac{0.33}{f_{\text{C}}} + \frac{1}{f_{\text{SW}}}\right)$$

$$C_{\text{OUT}} = \frac{I_{\text{STEP}} \times t_{\text{RESPONSE}}}{\Delta V_{\text{OUT}}}$$

where  $I_{\text{STEP}}$  is the load step,  $t_{\text{RESPONSE}}$  is the response time of the controller,  $\Delta V_{\text{OUT}}$  is the allowable output voltage deviation, and  $f_{\text{C}}$  is the target closed-loop crossover frequency. In our application, we selected  $f_{\text{C}} = 5\text{kHz}$ .

$$I_{\text{STEP}} = 0.5 \times I_{\text{OUT}} = 0.5 \times 7 = 3.5\text{A} \text{ (50\% of } I_{\text{OUT}})$$

$$\Delta V_{\text{OUT}} = 0.03 \times 7 = 210\text{mV} \text{ (3\% of } V_{\text{OUT}}, \text{ typ)}$$

$$t_{\text{RESPONSE}} \cong \left(\frac{0.33}{5\text{k}} + \frac{1}{100\text{k}}\right) = 76\mu\text{s}$$

$$C_{\text{OUT}} = \frac{I_{\text{STEP}} \times t_{\text{RESPONSE}}}{\Delta V_{\text{OUT}}} = \frac{3.5 \times 76\mu}{0.150} = 1267\mu\text{F}$$

Due to the DC-bias characteristics, 6 x 100µF 10V MLCC capacitors are selected for this design.

Capacitor values change with temperature and applied voltage. Refer to the capacitor data sheets to select capacitors that guarantee the required output capacitance across the operating range. For design calculations, use the worst-case derated value of capacitance, based on temperature range and applied voltage. In our case the worst-case derated value of capacitors is 1200µF.

For the flyback converter, the output capacitor supplies the load current when the main switch is on, and therefore the output voltage ripple is a function of load current and duty cycle. Use the following expression to estimate the output capacitor ripple:

$$\Delta V_{COUT} = \frac{I_{OUT} \times [I_{PRIPEAK} - (K \times I_{OUT})]^2}{I_{PRIPEAK}^2 \times f_{SW} \times C_{OUT}}$$

$$\Delta V_{COUT} = \frac{7 \times [30.7 - (0.99 \times 7)]^2}{30.7^2 \times 100\text{kHz} \times 1200\mu} = 34.9\text{mV}$$

### Step 13: Loop Compensation

Optocoupler feedback is used in isolated flyback converter designs for precise control of isolated output voltage. Figure 8 shows the overall scheme of the optocoupler feedback.

Use  $R_{FB} = 470\Omega$  (typ), for an optocoupler transistor current of 1mA. Select  $R1 = 49.9\text{k}\Omega$  and  $R2 = 22\text{k}\Omega$  (typical values) to use the full range of available COMP voltage. U3 is a low-voltage adjustable shunt regulator with a 2.5V reference voltage. In this design we selected a Texas Instruments TL431AQDBZRQ1 2.5V shunt regulator.

Calculate  $R_{LED}$  using the below expression:

$$R_{LED} = 400 \times CTR \times (V_{OUT} - 2.7)$$

$$R_{LED} = 400 \times 1 \times (7 - 2.7) = 1.72\text{k}\Omega$$

A standard 1.74kΩ resistor is selected,  $R_{LED} = 1.74\text{k}\Omega$ .

The bandwidth of typical optocouplers limits the achievable closed-loop bandwidth of opto-isolated converters. Considering this limitation, the closed-loop crossover frequency can be chosen at the nominal input voltage by selecting  $f_C = 5\text{kHz}$ . Closed-loop compensation values are designed based on the open-loop gain at the desired crossover frequency,  $f_C$ . The open-loop at  $f_C$  is calculated using the following expressions.

$$f_P = \frac{I_{OUT}}{\pi \times V_{OUT} \times C_{OUT}} = \frac{7}{\pi \times 7 \times 1200\mu} = 209.3\text{Hz}$$

$$G_{PLANT} = \frac{f_P}{f_C} \times \sqrt{\frac{L_{PRI} \times f_{SW} \times V_{OUT}}{8 \times I_{OUT}}} \times \frac{V_{IN}}{V_{IN} \times R_{CS} + 50 \times 10^3 \times L_{PRI}}$$

$$G_{PLANT} = \frac{209.3}{5\text{kHz}} \times \sqrt{\frac{1.3\mu \times 100\text{k} \times 7}{8 \times 7}} \times \frac{36}{36 \times 91\text{m} + 50 \times 10^3 \times 1.3\mu}$$

$$G_{PLANT} = 0.042 \times 0.1275 \times 101.98 = 0.546$$

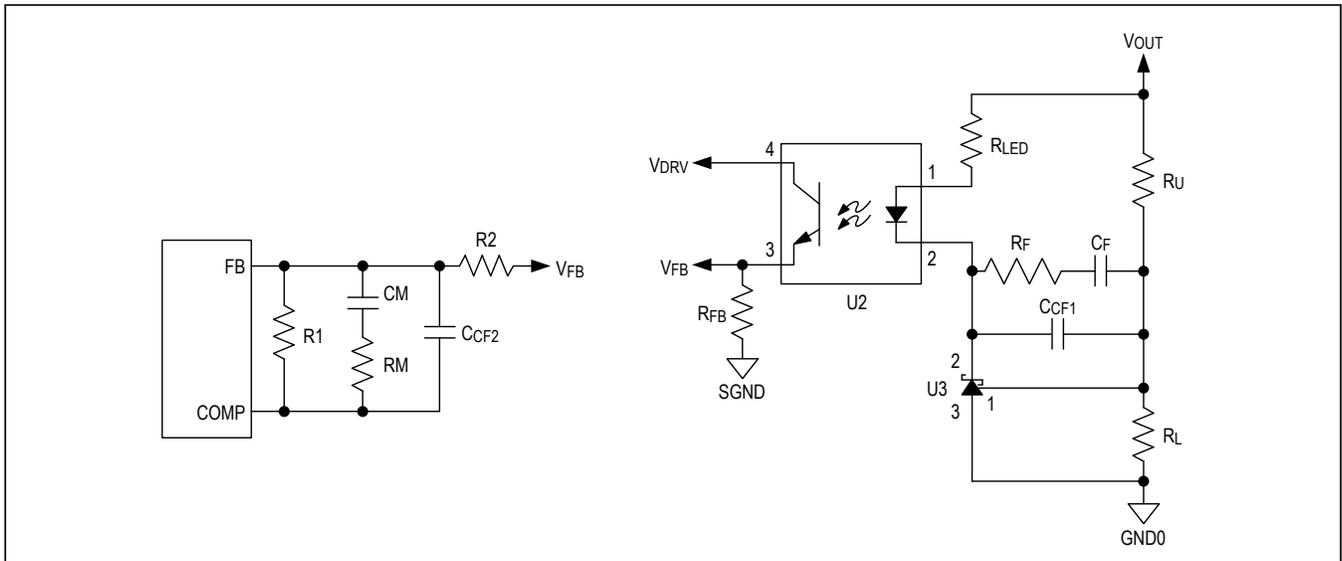


Figure 8. A typical opto-coupler-based feedback compensation.

Three controller configurations are suggested in **Application Note 5504** based on open-loop gain and the  $R_{LED}$  value. For typical designs, the current transfer ratio (CTR) of the optocoupler designs can be assumed to be unity. It is known that the comparator and gate-driver delays associated with the input voltage variations affect the optocoupler CTR. Depending on the optocoupler selected, variations in CTR causes wide variations in bandwidth of the closed-loop system across the input-voltage operating range. It is recommended to select an optocoupler with less CTR variations across the operating range. Checking the condition as stated in **Application Note 5504**:

$$G_{PLANT} \times CTR \times \frac{R_{FB}}{R_{LED}} \times \frac{R_1}{R_2} =$$

$$0.546 \times 1 \times \frac{470}{1.74k} \times \frac{49.9k}{22k} = 0.334$$

As stated in **Application Note 5504**, as  $0.334 < 0.8$ , configuration 1 is selected. **Figure 9** shows a typical schematic of configuration 1. The  $R_F$  value can be calculated from the expression below:

$$R_F = \left[ \frac{R_{LED} \times R_2}{G_{PLANT} \times CTR \times R_{FB} \times R_1} - 1 \right] \times R_U$$

$$R_F = \left[ \frac{1.74k \times 22k}{0.546 \times 1 \times 470 \times 49.9k} - 1 \right] \times 18k = 35.8k\Omega$$

A typical 35.7k value is selected as  $R_F$ ,  $R_F = 35.7k\Omega$ .

The  $C_F$  value can be calculated as follows:

$$C_F = \frac{1}{2\pi \times (R_U + R_F) \times f_P}$$

$$\frac{1}{2\pi \times (18k + 35.7k) \times 209.3} = 14.16nF$$

A standard 15nF capacitor is selected,  $C_F = 15nF$ .

The  $C_{CF1}$  value can be calculated as follows:

$$C_{CF1} = \frac{1}{\pi \times R_F \times f_{SW}}$$

$$= \frac{1}{\pi \times 35.7k\Omega \times 100kHz} = 89.2pF$$

A standard 100pF capacitor is selected,  $C_{CF1} = 100pF$ .

#### Step 14: EN/UVLO and OVI Setting

The device's EN/UVLO pin serves as an enable/disable input, as well as an accurate programmable input UVLO pin. The device does not commence startup operation unless the EN/UVLO pin voltage exceeds 1.21V. The device turns off if the EN/UVLO pin voltage falls below 1.15V. A resistor-divider from the input DC bus to ground can be used to divide down and apply a fraction of the input DC voltage ( $V_{DC}$ ) to the EN/UVLO pin. The values of the resistor-divider can be selected so the EN/UVLO pin voltage exceeds the 1.23V turn-on threshold at the desired input DC bus voltage. The same resistor-divider can be modified with an additional resistor ( $R_{OVI}$ ) to implement input overvoltage protection in addition to the EN/UVLO functionality as shown in **Figure 10**.

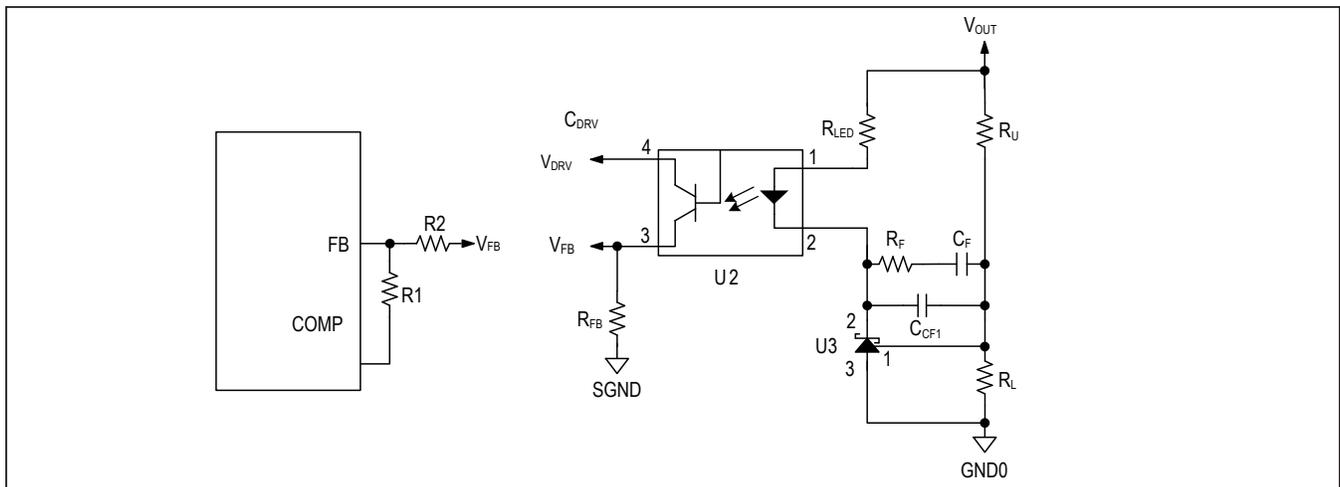


Figure 9. Opto-coupler feedback compensation configuration 1 schematic.

When voltage at the OVI pin exceeds 1.21V, the device stops switching and resumes switching operations only if voltage at the OVI pin falls below 1.15V. For given values of startup DC input voltage ( $V_{START}$ ) and input overvoltage-protection voltage ( $V_{OVI}$ ), the resistor values for the divider can be calculated as follows:

Select  $R_{OVI} = 10k\Omega$ .

$$R_{EN} = R_{OVI} \times \left( \frac{V_{OVI}}{V_{START}} - 1 \right)$$

where  $V_{OVI} = \text{maximum allowed overvoltage} = 37V$ .

$$R_{EN} = 10k \times \left( \frac{37}{8} - 1 \right) = 36.25k\Omega$$

A standard 36.5k $\Omega$  resistor is selected,  $R_{EN} = 36.5k\Omega$ .

The same resistor-divider can be modified to implement input overvoltage protection. When the voltage at the OVI pin exceeds 1.215V (typ), the device stops switching. The device resumes switching operations only if the voltage at the OVI pin falls below 1.1V (typ).

$$R_{EN\_TOP} = [R_{OVI} + R_{EN}] \times \left[ \frac{V_{START}}{1.21} - 1 \right]$$

$$R_{EN\_TOP} = [10k + 36.5k] \times \left[ \frac{8}{1.21} - 1 \right] = 260.9k\Omega$$

A standard 261k $\Omega$  resistor is selected for  $R_{EN\_TOP}$ .

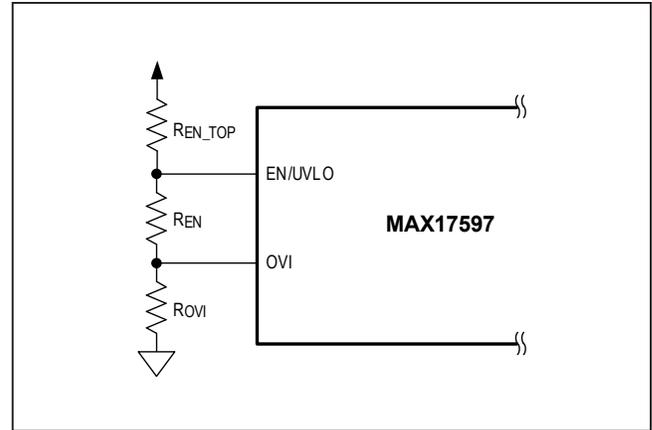


Figure 10. Programming EN/UVLO and OVI.

## Design Resources

Download the complete set of [Design Resources](#) including the schematics, bill of materials, PCB layout, and test files.

## Revision History

REVISION NUMBER	REVISION DATE	DESCRIPTION	PAGES CHANGED
0	7/20	Initial release	—

---

**Maxim Integrated**  
www.maximintegrated.com

*Maxim Integrated cannot assume responsibility for use of any circuitry other than circuitry entirely embodied in a Maxim Integrated product. No circuit patent licenses are implied. Maxim Integrated reserves the right to change the circuitry and specifications without notice at any time. The parametric values (min and max limits) shown in the Electrical Characteristics table are guaranteed. Other parametric values quoted in this data sheet are provided for guidance.*

© 2020 Maxim Integrated Products, Inc. All rights reserved. Maxim Integrated and the Maxim Integrated logo are trademarks of Maxim Integrated Products, Inc., in the United States and other jurisdictions throughout the world. All other marks are the property of their respective owners.