How to Design an Ultra-Wide Input Voltage Range Flyback

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Introduction

In modern technology, the flyback converter is one of the most widely used topologies for domestic and industrial applications that require an auxiliary power supply. Because flyback topology is so popular, designers often attempt to create a unified, single design for an ultra-wide input voltage (V_{IN}) range. This type of design allows flybacks to be used to their maximum functionality while avoiding long-term validation issues.

This article will describe how to design an ultra-wide input voltage range flyback using the MP023, a primary-side regulation (PSR) controller for low-power applications that provides an accurate constant voltage and constant current. The MP023 will be used for the practical example, and this article will provide the results of a 15W/5V flyback that accepts AC and DC voltages while operating across a wide input voltage range.

The MP023: A Primary-Side Regulation Flyback Controller

The MP023 is an offline, primary-side controller that provides optimal integrated regulation without the need for an optocoupler or secondary feedback circuit.

The MP023's variable off-time control allows its flyback converter to operate in discontinuous conduction mode (DCM). The MP023's current limit and maximum secondary duty cycle are configurable, which makes it simple to set the output current (I_{OUT}). Figure 1 shows the MP023's typical application circuit.

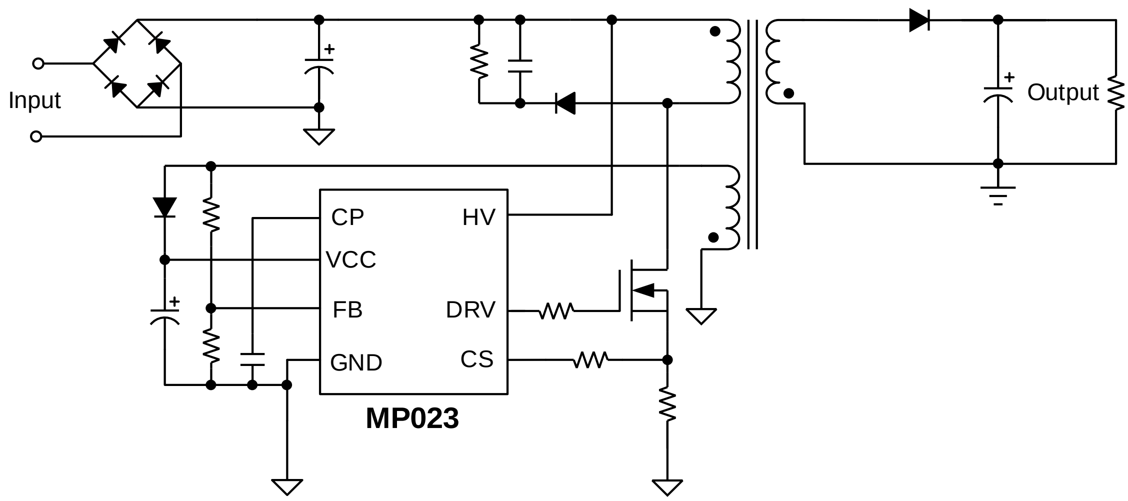

Figure 1: Typical Application of the MP023

The internal high-voltage start-up current source and power-saving technologies limit the no-load power consumption to below 30mW. Full protection features include V_{CC} under-voltage lockout (UVLO), overload protection (OLP), over-temperature protection (OTP), open-loop protection (OCkP), and over-voltage protection (OVP).

Designing a Flyback Converter Flowchart

There are many important design decisions and tradeoffs involved in designing an ultra-wide V_{IN} range flyback converter. The following sections will go through each step in the design process.

Figure 2 shows a flyback converter design flowchart.

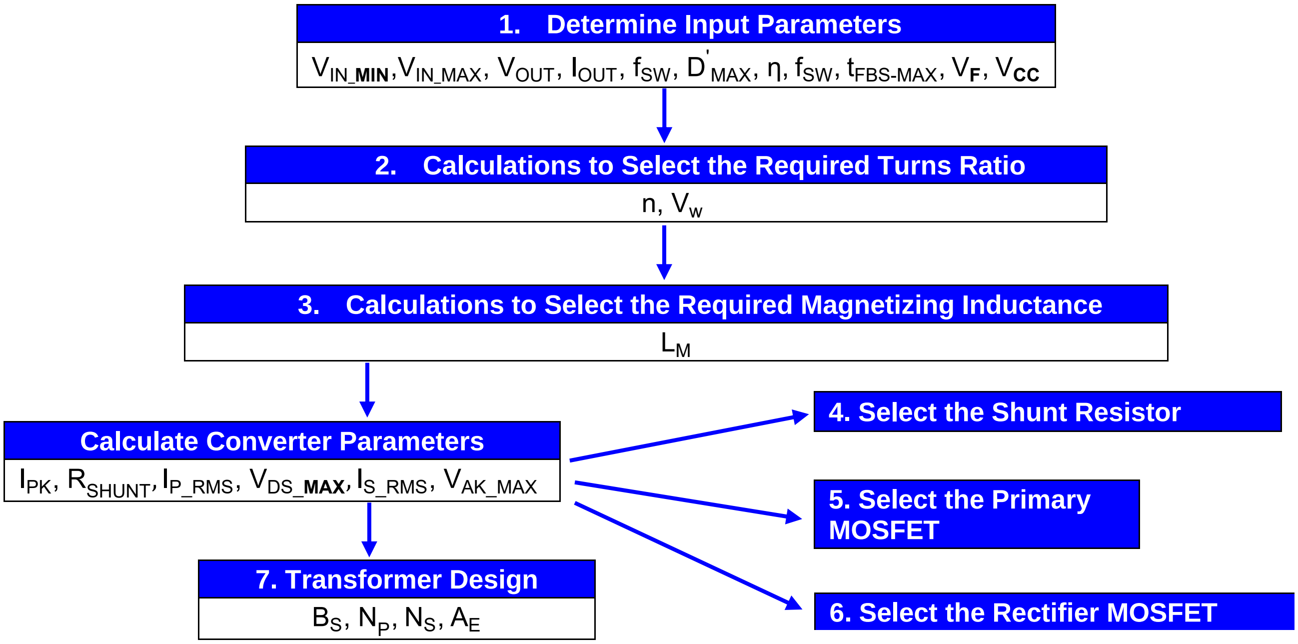

Figure 2: Control-Loop Design Flowchart

Flyback Converter Design Process and Calculations

Step 1: Design Inputs

Once the input parameters have been defined, it is time to design the overall converter. These parameters include the input voltage (V_{IN}), output voltage (V_{OUT}), output current (I_{OUT}), operation mode, switching frequency (f_{SW}), secondary duty cycle, estimated efficiency, feedback (FB) maximum sampling time, the secondary FET's forward voltage, and the IC supply voltage.

Table 1 shows a summary of the design inputs for the circuit discussed in this article. In this case, the input voltage ranges from $85V_{AC}$ to $576V_{AC}$, or from $90V_{DC}$ to $815V_{DC}$, by accepting both AC and DC input.

Table 1: Summary of Design Inputs

Design Input	Value
Minimum input voltage (V_{IN_MIN})	$85V_{AC}$ (or $90V_{DC}$)
Maximum input voltage (V_{IN_MAX})	$576V_{AC}$ (or $815V_{DC}$)
Output voltage (V_{OUT})	5V
Output current (I_{OUT})	3A
Operation mode	DCM
Switching frequency (f_{SW})	50kHz
Secondary duty cycle (D'_{MAX})	40%
Estimated efficiency (η)	85%
Rectifier MOSFET forward voltage (V_F)	0.1V
IC supply voltage	12V

The MP023 features output cable compensation in which, the duty cycle of the secondary side is limited to a certain value depending on the resistance or capacitance that is connected to the CP pin. According to the MP023's datasheet, connecting a 1μF capacitor to MP023's CP pin limits the secondary duty cycle to 40%.

To ensure that the results are realistic, the converter's estimated efficiency is defined to be relatively low (about 85%), as this is a common value for low-power flyback converters. For this application, I_{OUT} is defined to be 3A; a synchronous rectifier controller (e.g. the MP6908A) is used with a secondary MOSFET to improve efficiency and reduce thermal issues.

Step 2: Calculations to Select the Required Turns Ratio

Calculate the maximum turns ratio (*n*) according to the specified V_{IN_MIN} to provide a sufficient I_{OUT} due to the maximum on-duty limitation on the secondary side. The maximum turns ratio can be calculated with Equation (1):

$$n < \frac{\left(1 - D'_{MAX}\right) \times V_{DC}}{(V_{OUT} + V_F) \times D'_{MAX}} \rightarrow n < \frac{(1 - 0.4) \times 90}{(5 + 0.1) \times 0.4} \rightarrow n < 26.47 \qquad (1)$$

After calculating the maximum turns ratio to deliver the maximum power with V_{IN_MIN}, select *n*. The maximum turns ratio is a compromise between the secondary RMS current and the maximum reverse voltage of the secondary MOSFET.

Since this use case utilizes synchronous rectification, the secondary MOSFET'a reverse voltage is important, as low-voltage MOSFETs are cost-effective and easier to obtain. For this design, a maximum turns ratio of 15 was selected; this selection will be validated in Step 5.

Next, calculate the output-reflected voltage (V_W) that the primary winding will experience during the second half of the switching cycle. V_W can be estimated with Equation (2):

$$V_w = n \times (V_{OUT} + V_F) = 15 \times (5 + 0.1) = 76.5V \qquad (2)$$

V_W is important when calculating the primary MOSFET's maximum reverse voltage.

Step 3: Calculations to Select the Required Magnetizing Inductance

Since the compensator's passive components are internal to the controller, the MP023 can sample the auxiliary voltage that it supplies to close the loop-gain system (the flyback converter and integrated compensator). The FB maximum sampling time defines the time during which the controller samples the auxiliary voltage for regulation (see Figure 3).

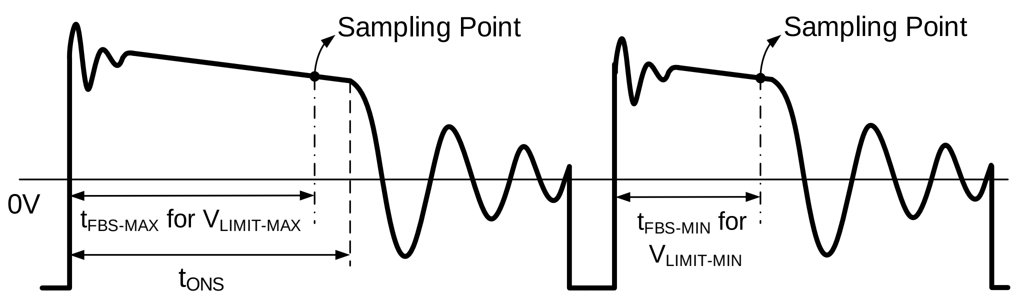

Figure 3: FB Voltage Sampling Point

According the MP023, the secondary MOSFET's minimum on time (t_{S_ON}) must meet the requirements in Equation (3):

$$t_{S\_ON} = I_{PK} \times \frac{N_S \times L_M}{(V_{OUT} + V_F) \times N_P} > t_{FBS\_MAX} + t_{FBS\_SD} \qquad (3)$$

Where t_{FBS_MAX} is the FB maximum sampling time, and t_{FBS_SD} is the FB sampling duration.

To calculate the magnetizing inductance and its peak current value, consider the mode. In this scenario, as the MP023 works in DCM, the output power (P) can be estimated with Equation (4):

$$P = \eta \times \frac{1}{2} \times L_M \times I_{PK}^2 \times f_{SW} \quad (4)$$

When considering Equation (3) and Equation (4), the minimum magnetizing inductance can be calculated with Equation (5):

$$L_{M\_MIN} > \left(\frac{(t_{FBS\_MAX} + t_{FBS\_SD}) \times N_{PS} \times (V_{OUT} + V_F) \times \sqrt{f_{SW}}}{\sqrt{2 \times P_{OUT}}}\right)^2 \quad (5)$$

Equation (5) can be simplified with Equation (6):

$$L_{M\_MIN} > \left(\frac{(3.5 \times 10^{-6} + 330 \times 10^{-9}) \times 15 \times (5 + 0.1) \times \sqrt{50 \times 10^3}}{\sqrt{2 \times 5 \times 3}}\right)^2 \rightarrow L_M > 143.1\mu H \quad (6)$$

After calculating the minimum magnetizing inductance required for the application, calculate its maximum value, which is limited by the fixed maximum secondary duty cycle. L_{M_MAX} can be calculated with Equation (7):

$$L_{M\_MAX} < \left(\frac{\left(D'_{MAX} \times \frac{1}{f_{SW}}\right) \times N_{PS} \times (V_{OUT} + V_F) \times \sqrt{f_{SW}}}{\sqrt{2 \times P_{OUT}}}\right)^2 \quad (7)$$

It can be simplified with Equation (8):

$$L_{M\_MAX} < \left(\frac{\left(0.4 \times \frac{1}{50 \times 10^3}\right) \times 15 \times (5 + 0.1) \times \sqrt{50 \times 10^3}}{\sqrt{2 \times 5 \times 3}}\right)^2 \rightarrow L_M < 624.24\mu H \quad (8)$$

Therefore, the magnetizing inductance must be between 143.1µH and 624.24µH. However, L_M is a tradeoff between the RMS currents and the size of the transformer. It is recommended to use a transformer between 60% and 80% of the maximum calculated value to achieve its full power without limiting the secondary duty cycle. For this example, a magnetizing inductance of 400µH is used.

Once the transformer value is chosen, calculate the peak current with Equation (9):

$$I_{PK} = \sqrt{\frac{2 \times P_{OUT}}{\eta \times L_M \times f_{SW}}} = \sqrt{\frac{2 \times 5 \times 3}{0.85 \times 400 \times 10^{-6} \times 50 \times 10^3}} = 1.328A \quad (9)$$

Because this application is designed to have an ultra-wide V_{IN}, it is important to ensure that the minimum on time at a high V_{IN} exceeds the leading-edge blanking time. The blanking time is the time during the first switching cycle, when the controller’s internal comparator is turned off to avoid activating short-circuit protection (SCP) due to shoot-though.

The minimum on time (t_{ON}) can be estimated with Equation (10):

$$t_{ON} = I_{PK} \times \frac{L_M}{V_{IN}} = 1.328 \times \frac{400 \times 10^{-6}}{815} = 652ns > 380ns \quad (10)$$

According to this calculation, the selected magnetizing inductance is suitable for the application.

Step 4: Shunt Resistor Calculations

Once the peak current value has been calculated, design the shunt resistor to correctly close the peak current-controlled loop.

According to the MP023's datasheet, the worst-case minimum voltage limit to sense the current is 0.464V. The shunt resistance (R_{SHUNT}) can be calculated with Equation (11):

$$R_{SHUNT} = \frac{V_{CS}}{I_{PK}} = \frac{0.464}{1.328} = 0.35\Omega \tag{11}$$

The designer must choose a shunt resistor that can withstand its own power dissipation. The primary RMS current can be estimated with Equation (12):

$$I_{P\_RMS} = I_{PK} \times \sqrt{\frac{D}{3}} = I_{PK} \times \sqrt{\frac{\frac{I_{PK} \times L_M \times f_{SW}}{V_{IN}}}{3}} = 1.328 \times \sqrt{\frac{\frac{1.328 \times 400 \times 10^{-6} \times 50 \times 10^{3}}{90}}{3}} = 0.417A \tag{12}$$

In this use case, the power dissipation is about 61mW.

Step 5: Primary MOSFET Calculations

In step 5, select the appropriate primary MOSFET for the application. As the maximum peak and RMS currents are calculated, calculate the maximum voltage that the MOSFET must withstand using Equation (13):

$$V_{DS\_MAX} = V_{IN\_MAX} + V_w + 20\% \times Margin = (815+76.5)x1.2 = 1070\ V \tag{13}$$

For this use case, a primary MOSFET with a maximum reverse voltage of 1200V is required.

Step 6: Rectifier MOSFET Calculations

Similar to the primary MOSFET calculations, the synchronous rectifier's maximum reverse voltage can be estimated with Equation (14):

$$V_{DS\_MAX} = V_{OUT} + \frac{V_{IN\_MAX}}{n} + 40\% \times Margin = \left(5+\frac{815}{15}\right) \times 1.4 = 83V \tag{14}$$

Thus, a rectifier MOSFET with a maximum reverse voltage of 120V to 150V is needed.

The secondary RMS current is also important to select the optimal rectifier MOSFET. The secondary RMS current (I_{S_RMS}) can be calculated with Equation (15):

$$I_{S\_RMS} = I_{PK\_S} \times \sqrt{\frac{D'}{3}} = I_{PK} \times n\sqrt{\frac{D'}{3}} = 1.328 \times 15 \times \sqrt{\frac{0.4}{3}} = 7.27A \tag{15}$$

With this calculation in mind, a rectifier MOSFET with a low on resistance ($R_{DS(ON)}$) is needed for this application.

Step 7: Transformer Design

Step 7 involves the transformer. There are many design decisions involved in choosing a transformer, such as the core material and core shape. For this output power level and input voltage, the EF20 (E20/10/6) is suitable in terms of size and effective area.

The primary number of turns (N_P) for this transformer can be estimated with Equation (16):

$$N_P = \frac{L_M \times I_{PK}}{B_{MAX} \times A_E} = \frac{400 \times 10^{-6} \times 1.328}{0.275 \times 32.1 \times 10^{-6}} \approx 60 \tag{16}$$

Because f_{SW} is 50kHz, there are some core materials (such as the N27 or N97) that can be used for up to 0.3T of maximum magnetic flux density. To achieve the selected turns ratio with the lowest number of primary turns, a value of 0.275T is selected.

Once N_P has been calculated, the secondary turns (N_S) can be calculated with Equation (17):

$$N_S = \frac{N_P}{n} = \frac{60}{15} \approx 4 \tag{17}$$

After selecting the IC's supply voltage (V_{CC}), the auxiliary number of turns (N_{AUX}) can be estimated with Equation (18):

$$N_{AUX} = (V_{CC} + 0.6)x\frac{N_S}{V_{OUT}} = (12 + 0.6)\,x\frac{4}{5} \approx 10 \tag{18}$$

These calculations result in a transformer made with the following turns ratio: N_P:N_S:N_{AUX} = 60:4:10.

Final Design

Figure 5 shows the circuit's final design once the values for the vital components have been calculated.

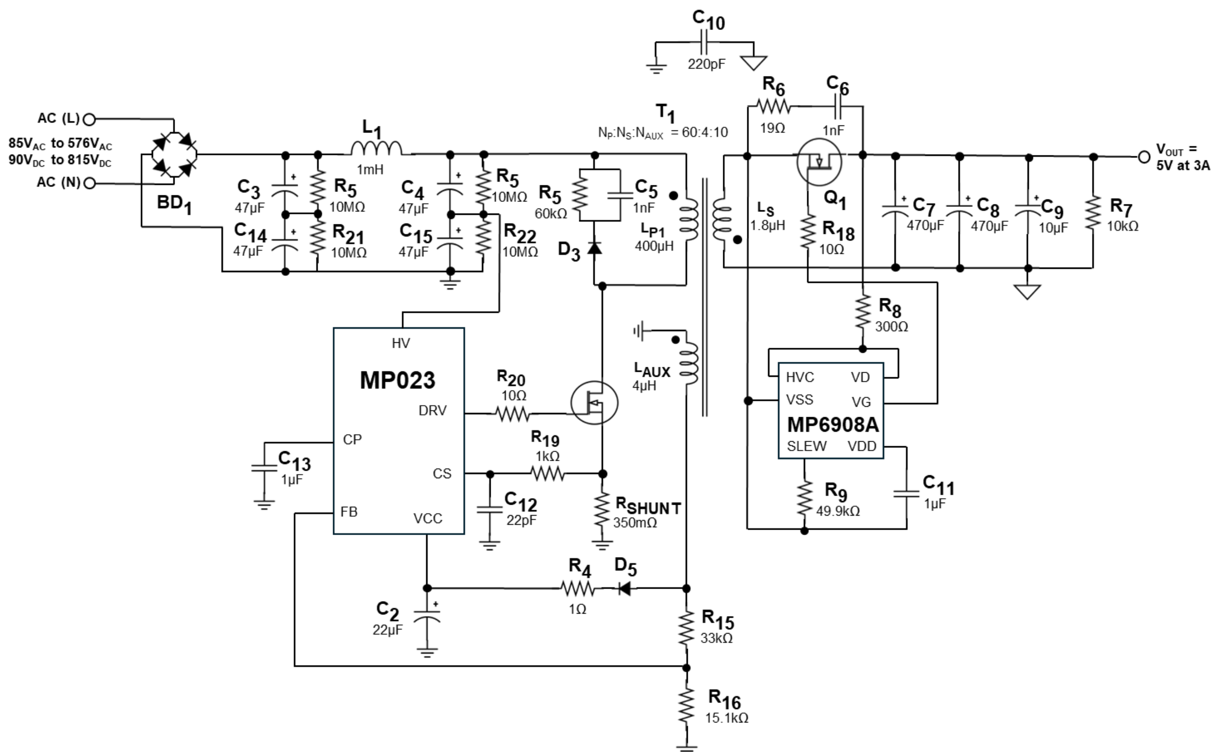

Figure 4: Final Design Circuit Schematic

Experimental Results

To correctly corroborate all the above calculations, a prototype of the ultra-wide input voltage range flyback was manufactured (see Figure 5).

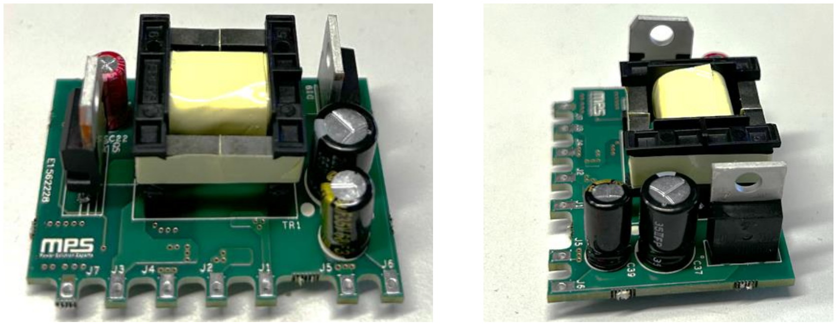

Figure 5: Prototype of the Ultra-Wide Input Voltage Range Flyback (PCB without the Input Filter)

This prototype was mounted without the input filter to make it a flexible PCB, which could be inserted into another PCB with different input filtering components.

Figure 6 shows the result of the converter validation at the minimum voltage. The blue trace denotes the drain-to-source voltage of the primary MOSFET (V_{DS}), while the pink trace denotes the primary current sensed through the shunt resistor.

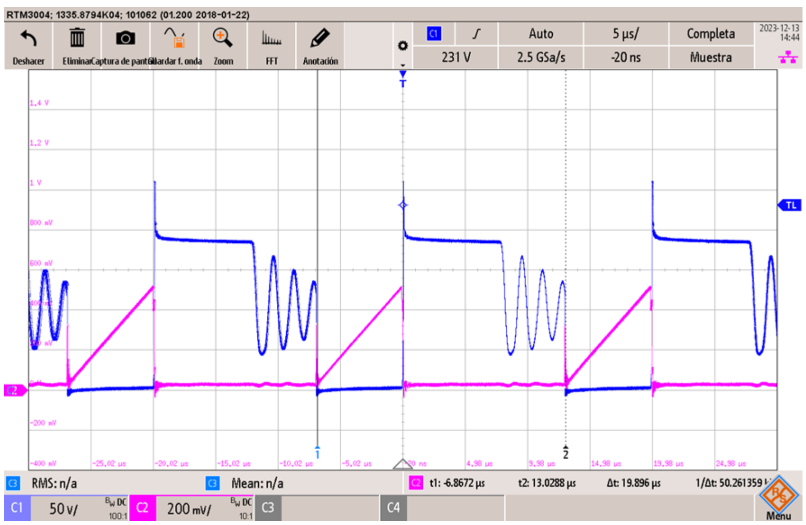

Figure 6: Converter Validation at the Minimum Input Voltage

Figure 7 shows the result of the converter validation at the maximum voltage. The blue trace denotes the drain-to-source voltage of the primary MOSFET (V_{DS}).

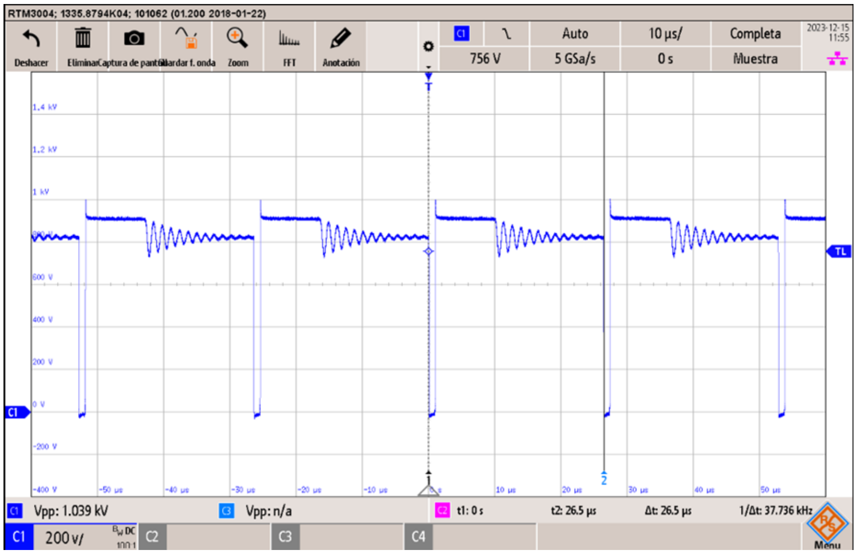

Figure 7: Converter Validation at the Maximum Input Voltage

Figure 8 shows the efficiency results for this design at different input voltages.

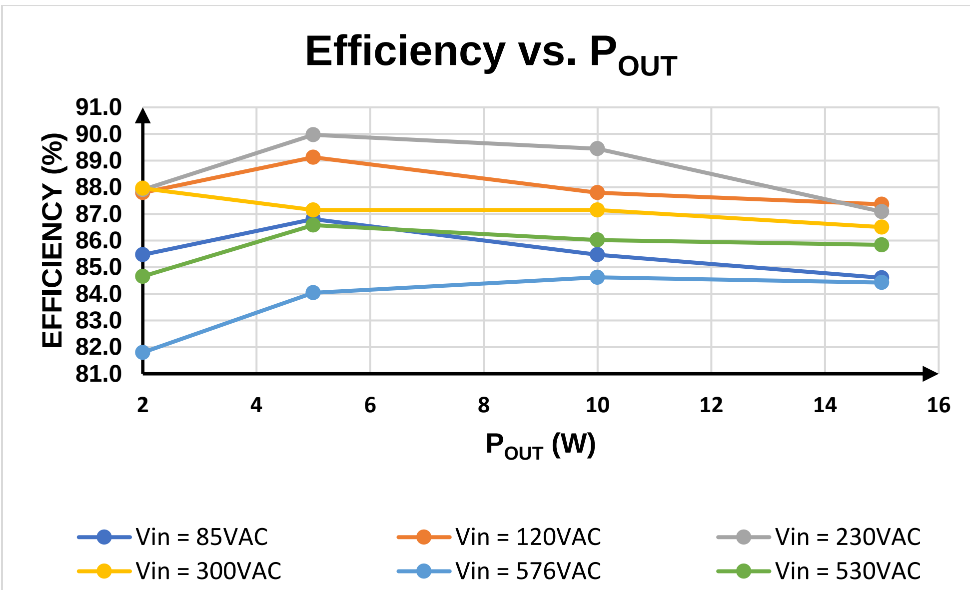

Figure 8: Efficiency Results

Figure 8 shows that the efficiency of the converter is quite high due to the use of synchronous rectification in the secondary side. In addition, using a primary MOSFET with a relatively low gate-charge capacitance reduces the switching losses at a high V_{IN}.

Conclusion

The use of a flyback with a wide V_{IN} range is useful in a large number of industrial applications that require a three-phase input. This article provided a simple series of steps to follow to optimize a flyback design while using the MP023. These steps included calculating the required turn ratio, magnetizing inductance, and shunt resistance, as well as selecting key parameters to optimize the design of the primary and secondary MOSFETs. Design validation results were provided to demonstrate the consistency and feasibility of the equations described in this article.

In addition to the MP023, MPS provides a number of flyback converters with primary-side regulation. Explore MPS's robust portfolio to find a solution that meets your design needs.