

How to Design Flyback Converter With LM3481 Boost Controller

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ABSTRACT

The Texas Instruments LM3481 boost controller is a versatile low-side N-FET high-performance controller for switching regulators. The device has wide range of applications such as automotive start-stop application, one-cell or two-cell Li-ion battery-powered portable Bluetooth audio system, industrial system isolated supplies, and so forth, and supports topologies like boost, SEPIC, and flyback. This application note describes the design principles of the flyback converter with the circuit operation based on LM3481.

The report is broadly divided into three parts:

1. Principle of flyback converter and features of LM3481
2. Step-by-step calculation followed by simulation (on Simplis)
3. Test bench results

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1 LM3481 Description

The LM3481 is a versatile low-side N-FET controller which has a wide input supply voltage range of 2.97 V to 48 V and can easily configured in topologies like boost, SEPIC, and flyback. Current mode control provides superior bandwidth and transient response in addition to cycle-by-cycle current limiting. [Figure 1](#) shows the pin configuration of LM3481.

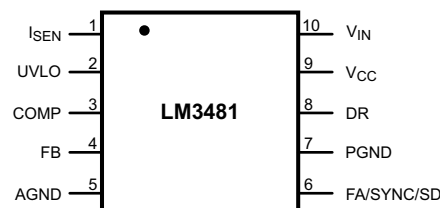


Figure 1. LM3481 Pin Configuration

2 Principle of Flyback Converter

The basic operation of the flyback is illustrated through circuit shown in [Figure 2](#). The flyback topology is essentially the buck-boost topology that is isolated by using a transformer as the storage inductor. The transformer not only provides isolation, but by varying the turn ratio, the output voltage can be adjusted.

The flyback converter operation is divided into two steps. In the first step, the switch Q is turned on and energy is stored (from the input) in the primary winding of the flyback transformer. On the secondary side, diode D is reverse biased and the load RL is supplied through energy stored in output capacitor Cout. In the second step, the switch Q is turned off and the energy stored in primary winding is transferred to secondary. The diode D is forward-biased and the energy is delivered to charge the output capacitor Cout as well as supply to the Load RL.

There are three basic modes of operations in flyback: that is, CCM (continuous conduction mode), DCM (discontinuous conduction mode) and CRM (critical conduction mode). The differentiation is based on energy-transfer of the primary winding of the flyback transformer. In CCM flyback, part of the energy stored in the flyback transformer remains in the transformer when the next ON period begins. Because the energy transfer is not complete, that is, all the stored energy is not transferred when switch Q is OFF, the primary current starts from a value greater than zero at the beginning of each cycle. The main waveforms are shown in [Figure 3](#).

In DCM flyback, all the stored energy from the secondary of the transformer is transferred to the output (to charge output capacitor as well as to supply the load) in the OFF time of switch Q. As a result input - or primary - current starts from zero at the beginning of each switching cycle.

The CRM flyback (also called transition mode (TM)), is just at the boundary between DCM and CCM, occurring when the stored energy just reaches zero at the end of the switching period.

In CCM mode (see [Figure 3](#)), the amount that the inductor current increases can be calculated by using a version of the familiar relationship:

$$V_L = L \frac{di_L}{dt}; \Delta i_L = V_L \frac{\Delta T}{L}$$

where

- V_L is the voltage across the Inductor (1)

The inductor current increase during the ON period can be approximated as:

$$\Delta I_{L(+)} = V_{IN} \times \left(\frac{D \times T_s}{L_{pri}} \right)$$

where

- L_{pri} is the primary inductance of the transformer
- D is the duty cycle of flyback converter
- T_s is the switching time of the flyback converter (2)

The inductor current decrease during the OFF state is given by:

$$\Delta I_{L(-)} = \frac{V_{OUT} \times N_{PS} \times (1-D) T}{L_{pri}} \quad (3)$$

In steady-state conditions, the current increase, $\Delta I_{L(+)}$, during the ON time and the current decrease during the OFF time, $\Delta I_{L(-)}$, must be equal. Otherwise, the inductor current would have a net increase or decrease from cycle-to-cycle which would not be a steady-state condition. Then the relationship of V_{IN} , V_{OUT} and duty cycle could be derived in [Equation 4](#).

$$V_{OUT} = \frac{V_{IN}}{N_{ps}} \times \frac{D}{1-D}$$

where

- N_{ps} is the turn ratio of primary windings to secondary windings (4)

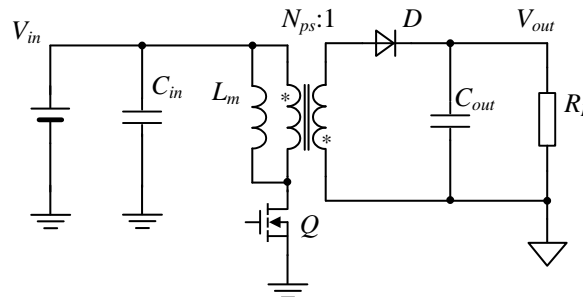


Figure 2. Topology of Flyback

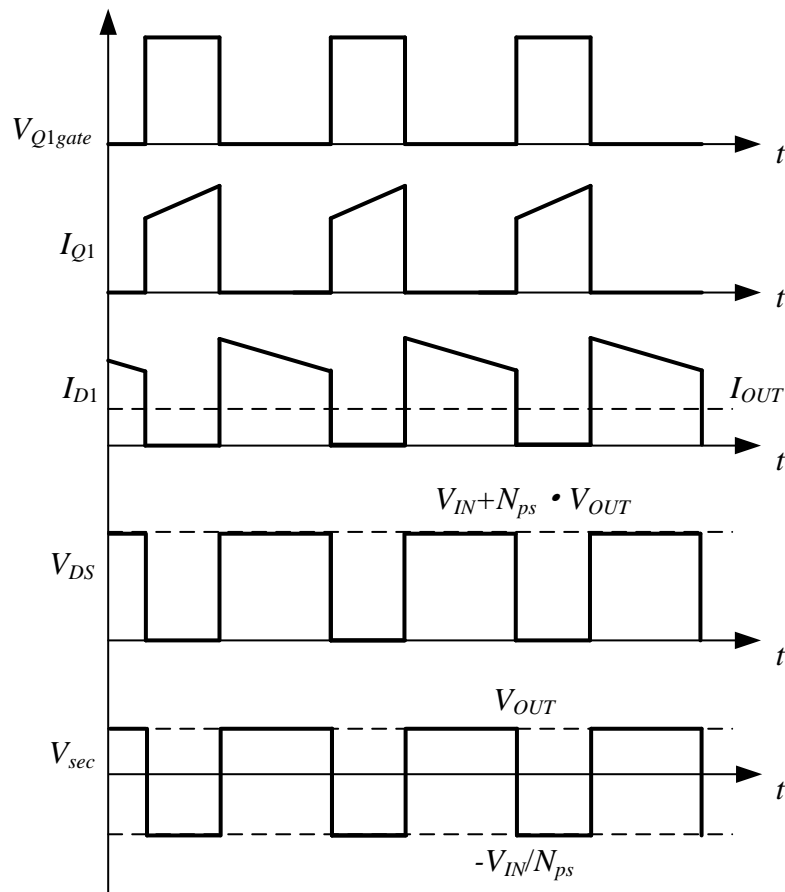


Figure 3. Flyback Main Waveform in CCM

3 Flyback Parameters Design

This section introduces flyback circuit design with details on transformer turns ratio, magnetizing inductance calculation, MOSFET/Diode voltage stress, flyback small signal modeling, and loop compensation.

3.1 Transformer Design

Flyback can be designed in CCM, DCM, or CRM mode based on the application requirements. CCM has smaller ripple as well as RMS current, lower MOSFET conduction and turnoff losses, better full load efficiency, and so forth. While DCM has no diode reverse recovery loss, lower inductance value which may result into smaller transformer, better no load efficiency, and so forth. For the DC-DC application, where the input voltage is low, CCM is an attractive option to minimize the primary RMS current particularly in higher power systems even though the system's dynamic behavior is considerably difficult to maintain. CCM is taken as an example in this report for a lower input voltage design application. The transformer turn ratio N_{ps} could be derived as shown in [Equation 5](#).

$$N_{PS} = \frac{V_{IN_min}}{V_{OUT}} \times \frac{D_{max}}{1 - D_{max}}$$

where

- D_{max} is the maximum duty cycle allowed in the design at lowest input voltage.

(5)

Suppose the efficiency of the flyback converter is η , the input power and output power relationship could be derived as shown in [Equation 6](#).

$$V_{IN} \times I_{LM} \times D \times \eta = P_O$$

where

- I_{LM} is the average current flowing in the transformer primary winding, which is also equals to the input average current (6)

Setting the magnetizing inductor current as:

$$\Delta I_L = K_L I_{LM}$$

where

- K_L is the fraction chosen (7)

According to [Equation 2](#), [Equation 4](#), [Equation 6](#), and [Equation 7](#), L_M function could be derived as shown in [Equation 8](#).

$$L_M(V_{IN}) = \frac{V_{IN}^2 \times \eta}{K_L \times P_O \times f_s} \times \frac{V_{OUT}^2 \times N_{ps}^2}{(V_{IN} + V_{OUT} \times N_{ps})^2}$$

where

- P_O is output power
- f_s is flyback converter frequency (8)

From this equation, it also could be observed that L_M has positive correlation with V_{IN} . That means the magnetizing inductor should be designed at highest input voltage to ensure the converter could run at CCM during all the input voltage range.

The peak current flowing in the primary winding can be calculated as:

$$I_{pk_L} = \left(1 + \frac{K_L}{2}\right) \times \frac{P_O \times (V_{IN} + V_{OUT} \times N_{ps})}{V_{IN} \times V_{OUT} \times \eta \times N_{ps}} \quad (9)$$

The calculated turns ratio, primary winding inductance and peak current can be used to determine the magnetic core of the transformer, number of turns on primary and secondary windings, as well as wire thickness.

3.2 MOSFET and Diode Selection

As shown in [Figure 3](#), the voltage across the MOSFET during off time is $V_{IN} + V_{OUT} \times N_{ps}$ and across the diode when it is reversed biased is $V_{OUT} + V_{IN} / N_{ps}$. The voltage rating of the MOSFET and the diode needs to be chosen with appropriate margin as both of them suffers from high voltage spike. In case of MOSFET, the primary leakage inductance resonates with output capacitance of MOSFET and similarly in case of diode, secondary leakage inductance resonates with diode capacitance and results into high voltage spikes. Considering the spike voltage, usually choose 1.5~2 times of the voltage stage as the voltage rating to ensure sufficient margin.

The voltage rating of MOSFET and Diode could be derived as shown in [Equation 10](#).

$$V_{ds_MOS} = (1.5 \sim 2) \times (V_{IN_MAX} + V_{OUT} \times N_{ps})$$

$$V_{R_diode} = (1.2 \sim 1.5) \times \left(V_{OUT} + \frac{V_{IN_MAX}}{N_{ps}} \right) \quad (10)$$

The current RMS rating of MOSFET and Diode could be derived as shown in [Equation 11](#).

$$\begin{aligned}
 I_{dsRMS} &= \sqrt{D \times \left(\left(\frac{V_{OUT} \times I_{OUT}}{V_{IN} \times D} \right)^2 + \frac{(K_L \times I_{LM})^2}{3} \right)} \\
 I_{R_diode} &= \sqrt{(1-D) \times \left(I_{OUT}^2 + \frac{(K_L \times I_{LM} \times N_{ps})^2}{3} \right)}
 \end{aligned} \tag{11}$$

The current rating of MOSFET and Diode should at least be I_{pk_LM} and I_{OUT} respectively.

3.3 CCM Flyback Small Signal Modeling

Figure 4 is LM3481 functional block diagram and Figure 5 is the Peak Current mode overall control implementation. The I_L (inductor current) disturbance includes two parts: V_{IN} and duty cycle, which is generated by the comparison of a current sampling RAMP and output of the EA (Error Amplifier). As the current waveform is in conjunction with external ramp, the modulator gain of the circuit F_m is:

$$F_m = \frac{1}{(S_n + S_e)T_s} \tag{12}$$

The meaning of PCM control implementation parameters are shown in Table 1.

Table 1. The Meaning of PCM Overall Control Implementation Parameters

Parameter	Description	Transfer Function	Function
T_s	Switching period	$H_{voi}(s)$	I_L to output voltage transfer function
S_n	On-time slope of the sensed-current waveform	$H_{EA}(s)$	Gain function of the error amplifier
S_e	External ramp	$H_{div}(s)$	Gain of the divided resistor network
R_i	Current sample resistor	$H_e(s)$	Sample-and-hold system gain function
$H_{di}(s)$	Duty cycle to I_L transfer function	$H_{vi}(s)$	Input voltage to I_L transfer function

Figure 6 is the average small signal model of flyback converter. According to the KCL and KVL principle, use Equation 13 and Equation 14:

$$D \times \hat{v}_{in}(s) + (V_{IN} + N_{ps} \times V_{OUT}) \times \hat{d}(s) = sL_m \times \hat{i}_{Lm}(s) + N_{ps} \times \hat{v}_{out}(s) \tag{13}$$

$$(1-D) \times N_{ps} \times \hat{i}_{Lm}(s) = \frac{V_{OUT}}{R_{Ld}} \hat{d}(s) + sC_{out} \times \hat{v}_{out}(s) + \frac{\hat{v}_{out}(s)}{R_L} \tag{14}$$

To separate the small signal disturbance, the gain function from duty cycle to inductor current could be calculated as shown in Equation 15.

$$H_{di}(s) = \frac{\hat{i}_{Lm}(s)}{\hat{d}(s)} = \frac{sC_{out} \times \frac{V_{IN}}{1-D} + \frac{1}{R_L} \times \left(\frac{V_{IN}}{1-D} + V_{IN} \times D \times N_{ps} \right)}{\left(\frac{V_{IN} \times D}{V_{OUT}} \right)^2 + \frac{L_m}{R_L} s + s^2 L_m C_{out}} \tag{15}$$

Considering the practical crossover frequency is much higher than the corner frequency with PCM control, simplify Equation 15 to Equation 16:

$$H_{di}(s) = \frac{\hat{i}_{Lm}(s)}{\hat{d}(s)} \approx \frac{\frac{V_{IN}}{1-D} + V_{IN} \times D \times N_{ps}}{L_m \times s} \tag{16}$$

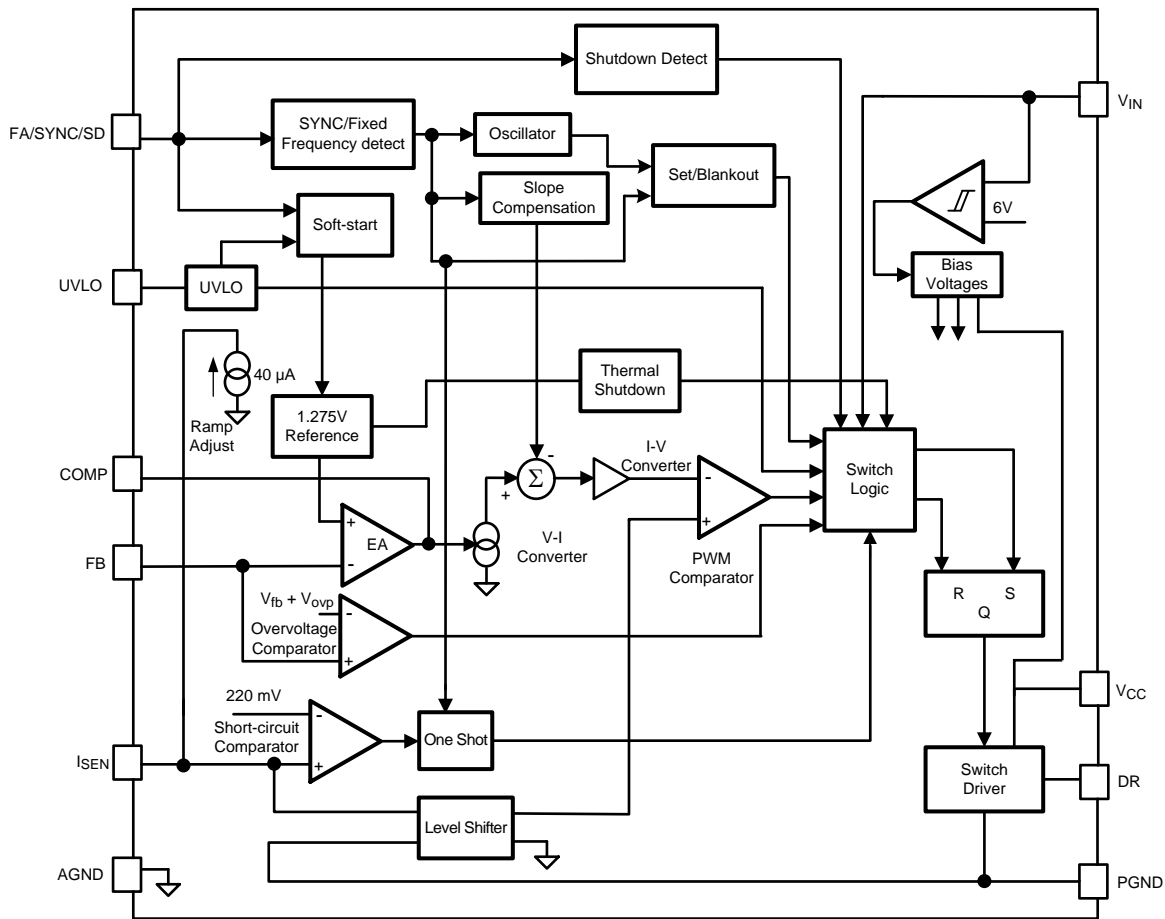


Figure 4. LM3481 Functional Block Diagram

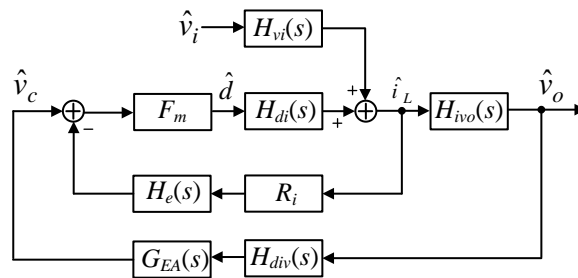


Figure 5. The Overall Control Implementation

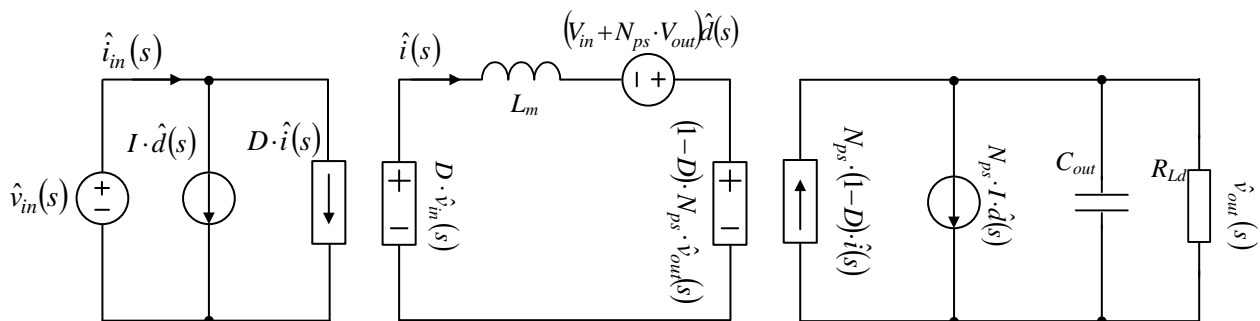


Figure 6. The Small Signal Model of CCM Flyback Converter

The gain function from magnetizing inductor current to output voltage could be calculated as shown in Equation 17.

$$H_{ivo}(s) = \frac{\hat{v}_o(s)}{\hat{i}_{Lm}(s)} = \frac{N_{ps} \times \left(1 - \frac{L_m \times V_{OUT}}{V_{IN} \times R_L} \times s\right)}{\frac{s \times C_{out}}{1-D} + \frac{1}{R_L} \times \left(\frac{1}{1-D} + D \times N_{ps}\right)} \quad (17)$$

The sample-and-hold system gain function is as shown in Equation 18.

$$H_e(s) = \frac{T_s \times S}{e^{T_s \times s} - 1} \approx 1 - \frac{s}{2/T_s} + \frac{s^2}{\left(\frac{\pi}{T_s}\right)^2} \quad (18)$$

Ignore the disturbance of V_{IN} , the control block for current regulation loop could be found in Figure 5 as below:

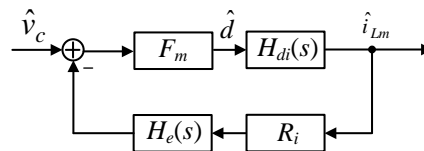


Figure 7. Control Block For Current Regulation Loop

The gain function from comp voltage to inductor current is shown in Equation 19.

$$H(s) = \frac{\hat{i}_{Lm}(s)}{\hat{v}_c(s)} = \frac{F \times H_{di}(s)}{1 + H_e(s) \times R_i \times F_m \times H_{di}(s)} \quad (19)$$

The gain function from comp voltage to output voltage is shown in Equation 20.

$$PS(s) = \frac{\hat{v}_o(s)}{\hat{v}_c(s)} = \frac{F_m \times H_{di}(s) \times H_{ivo}(s)}{1 + H_e(s) \times R_i \times F_m \times H_{di}(s)} \quad (20)$$

3.4 LM431 and Optocoupler Modeling

For the best output voltage accuracy, this design uses secondary-side regulation with a shunt regulator driving an optocoupler to control the COMP pin voltage of the LM3481. The FB pin of the LM3481 is grounded.

Note that the compensation shown around LM431 consists of a resistor R1 and a capacitor C1, which provide a zero in the compensation network. The C_{opto} represents the output capacitance of the optocoupler. R2, C2 usually are the compensation circuit connected to the IC comp pin.

The gain function from comp voltage to inductor current is shown in Equation 21.

$$H_{EA}(s) = \frac{\hat{v}_{comp}(s)}{\hat{v}_o(s)} = -\frac{K_{opto} R_{up}}{R_{opto}} \frac{(R_1 + R_{fb1}) \times C_1 \times s + 1}{R_{fb1} C_1 \times s} \frac{R_2 C_2 \times s + 1}{R_2 R_{up} C_2 C_{opto} \times s^2 + (C_2 R_{up} + C_{opto} R_{up} + R_2 C_2) \times s + 1} \quad (21)$$

Usually $C_2 \gg C_{opto}$, the $H_{EA}(s)$ could be simplified as shown in Equation 22.

$$H_{EA}(s) = \frac{\hat{v}_{comp}(s)}{\hat{v}_o(s)} = -\frac{K_{opto} R_{up}}{R_{opto}} \frac{(R_1 + R_{fb1}) \times C_1 \times s + 1}{R_{fb1} C_1 \times s} \times \frac{R_2 C_2 \times s + 1}{\left[(R_2 + R_{up}) C_2 \times s + 1\right]} \quad (22)$$

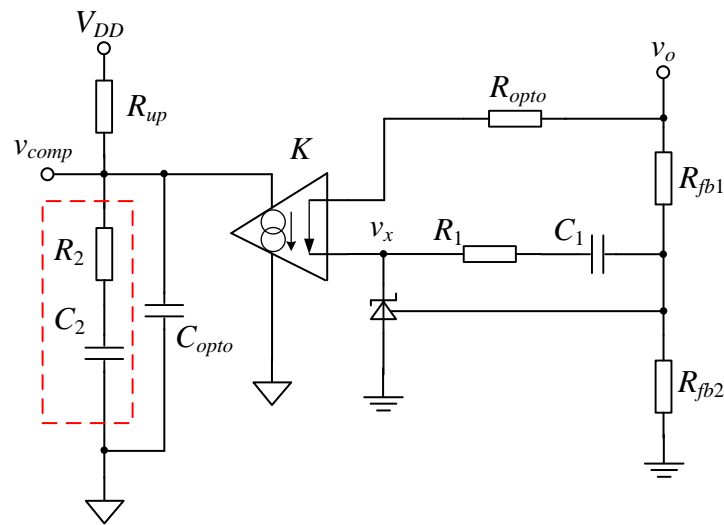


Figure 8. Typical LM431 Configuration With Output Bias and Optocoupler

4 LM3481 Flyback Parameters Design

Table 2 gives the performance specification of the example design presented. A CCM flyback was selected for this application.

Table 2. Design Specifications

PARAMETER	EXAMPLE VALUE
V _{in min}	5 VDC
V _{in max}	32 VDC
V _{out}	12 VDC
I _{out}	2 max
Target switching frequency	130 kHz

4.1 LM3481 Flyback Transformer Design

Maximum duty cycle of LM3481 is 0.85. Considering the tradeoff between V_{ds_MOS} and V_{R_diode}, set D_{max} = 0.7. The N_{ps} could be calculated as shown in Equation 23.

Setting f_s = 130 kHz, K_L = 0.2 and η = 0.85, the curve of L_m and V_{in} relationship could be derived as shown in Figure 9. Choose L_m = 12 μH.

$$N_{ps} = \frac{V_{IN_min}}{V_{OUT}} \times \frac{D_{max}}{1-D_{max}} = \frac{5 \times 0.7}{12 \times (1-0.7)} \approx 1 \quad (23)$$

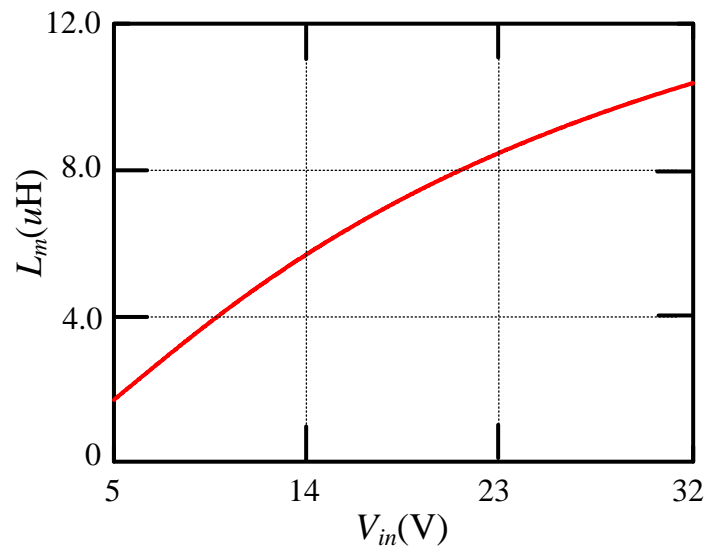


Figure 9. The Curve of Lm and Vin Relationship

Peak Primary Winding current can be calculated as shown below:

$$I_{pk_Lm} = \left(1 + \frac{K_L}{2}\right) \times \frac{P_o \times (V_{IN_min} + V_{OUT} \times N_{ps})}{V_{IN_min} \times V_{OUT} \times \eta \times N_{ps}} = 9.13 \text{ A} \quad (24)$$

Use 30% to 40% headroom on the Peak primary current to calculate the current sense resistor.

The limit on the ISEN pin of LM3481 is 100 mV so the current sense resistance can be chosen as 100 mV / 12 A = 8.33 mΩ. A current sense resistance of 6 mΩ was chosen for this application.

4.2 LM3481 Flyback MOSFET and Diode Selection

According to Equation 6, the voltage stress of MOSFET and Diode is:

$$\begin{aligned} V_{ds_MOS} &= V_{in_max} + V_o \times N_{ps} = 44 \text{ V} \\ V_{ds_diode} &= V_o + V_{in_max} / N_{ps} = 44 \text{ V} \end{aligned} \quad (25)$$

$$\begin{aligned} I_{dsRMS} &= \sqrt{D \times \left(\left(\frac{V_{OUT} \times I_{OUT}}{V_{IN} \times D} \right)^2 + \frac{(K_L \times I_{LM})^2}{3} \right)} = 10.86 \text{ A} \\ I_{R_diode} &= \sqrt{(1-D) \times \left(I_{OUT}^2 + \frac{(K_L \times I_{LM} \times N_{ps})^2}{3} \right)} = 9.31 \text{ A} \end{aligned} \quad (26)$$

To get good thermal performance, reasonable package for diode and MOSFET are needed to be chosen beside ensuring required voltage/current rating. Here MOSFET CSD19502Q5B (80V/100A, DNK0008A) and diode V10P-M3/86A (100V/10A, TO-227) are chosen.

4.3 LM3481 Flyback CCM Flyback Transfer Function

Figure 10 shows the final application schematic of the flyback solution based on LM3481.

LM3481 has an internal V_{sL} (90 mV) slope compensation to avoid sub-harmonic oscillation when duty cycle is larger than 0.5. In 12Vin and 12-V, 2-A output condition, the modulator gain of the circuit F_m could be derived as shown in Equation 27.

$$S_n = \frac{V_{in}}{L_m} \times R_i = 6 \times 10^3; S_e = \frac{V_{sL} \times D}{T_s} = 5.85 \times 10^3; F_m = \frac{1}{(S_n + S_e) T_s} = 10.97 \quad (27)$$

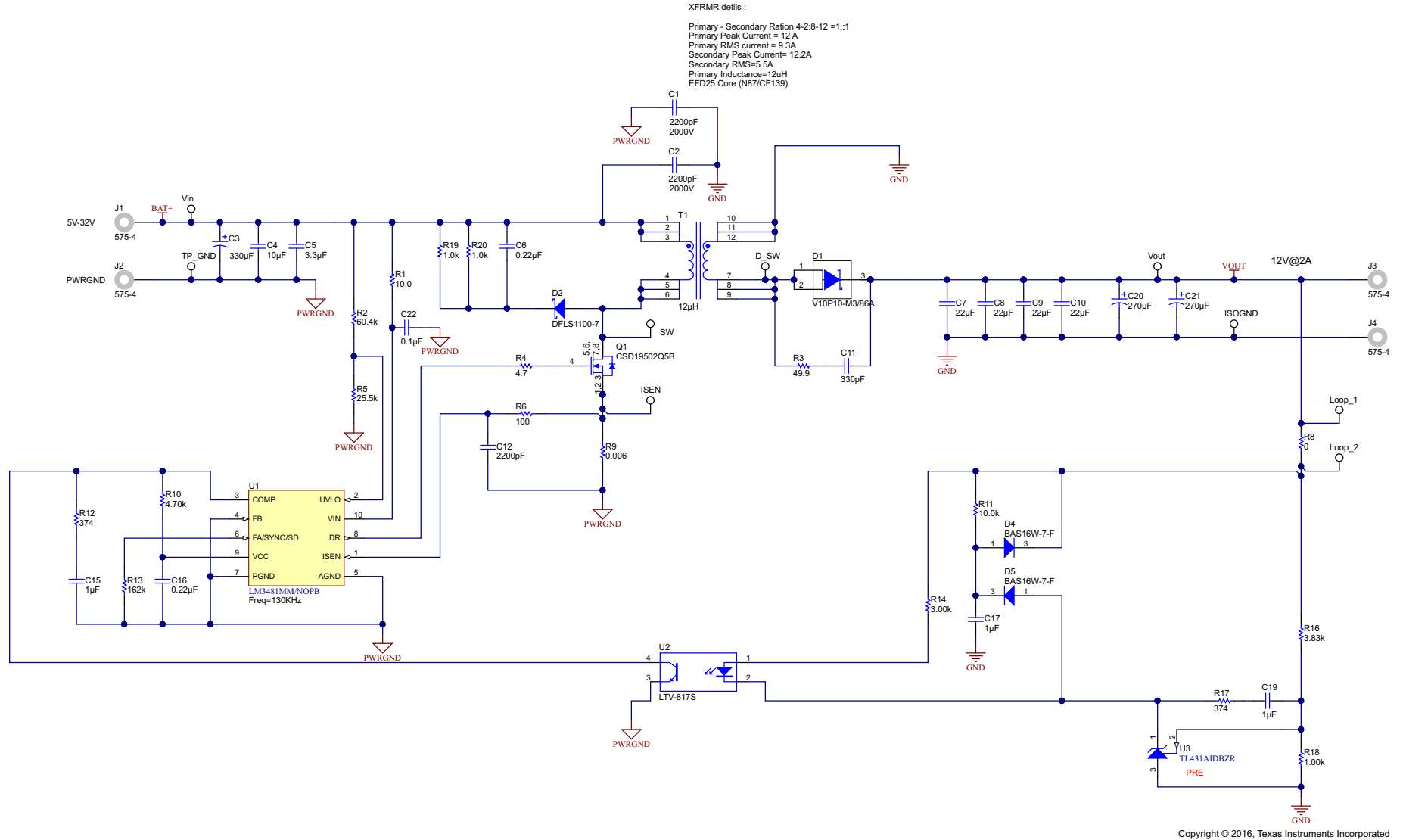


Figure 10. LM3481-Flyback Schematic

According to Equation 11 to Equation 16, $H_{di}(s)$, $H_{ivo}(s)$, $H_e(s)$ and $H(s)$ transfer function could be derived. Figure 11 shows the Mathcad calculated Bode plot of LM3481 flyback converter, and Figure 12 shows the simplis simulation schematic and results. It could be observed that the right half plane zero of flyback converter is at 33.2 kHz. The calculated results match the simulation results very well.

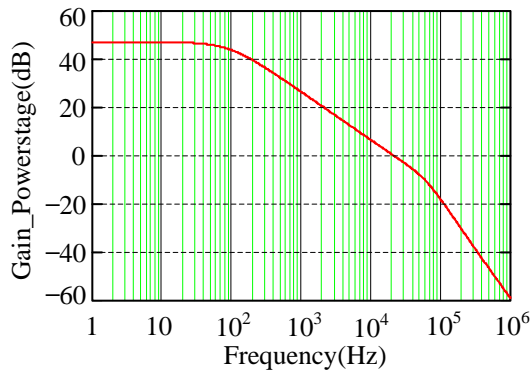


Figure 11. Calculated LM3481-Flyback Bode Plot – Gain-Frequency

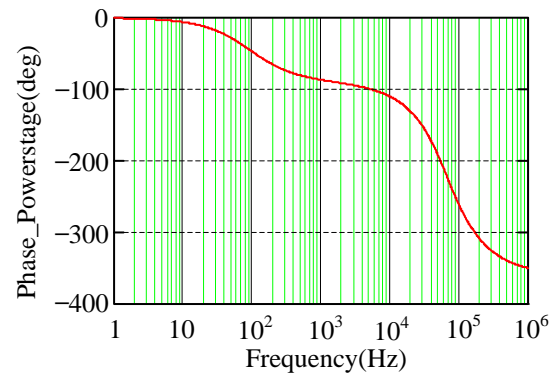


Figure 12. Calculated LM3481-Flyback Bode Plot – Phase-Frequency

LM3481 Flyback Isolated 12V in 12V@2A out

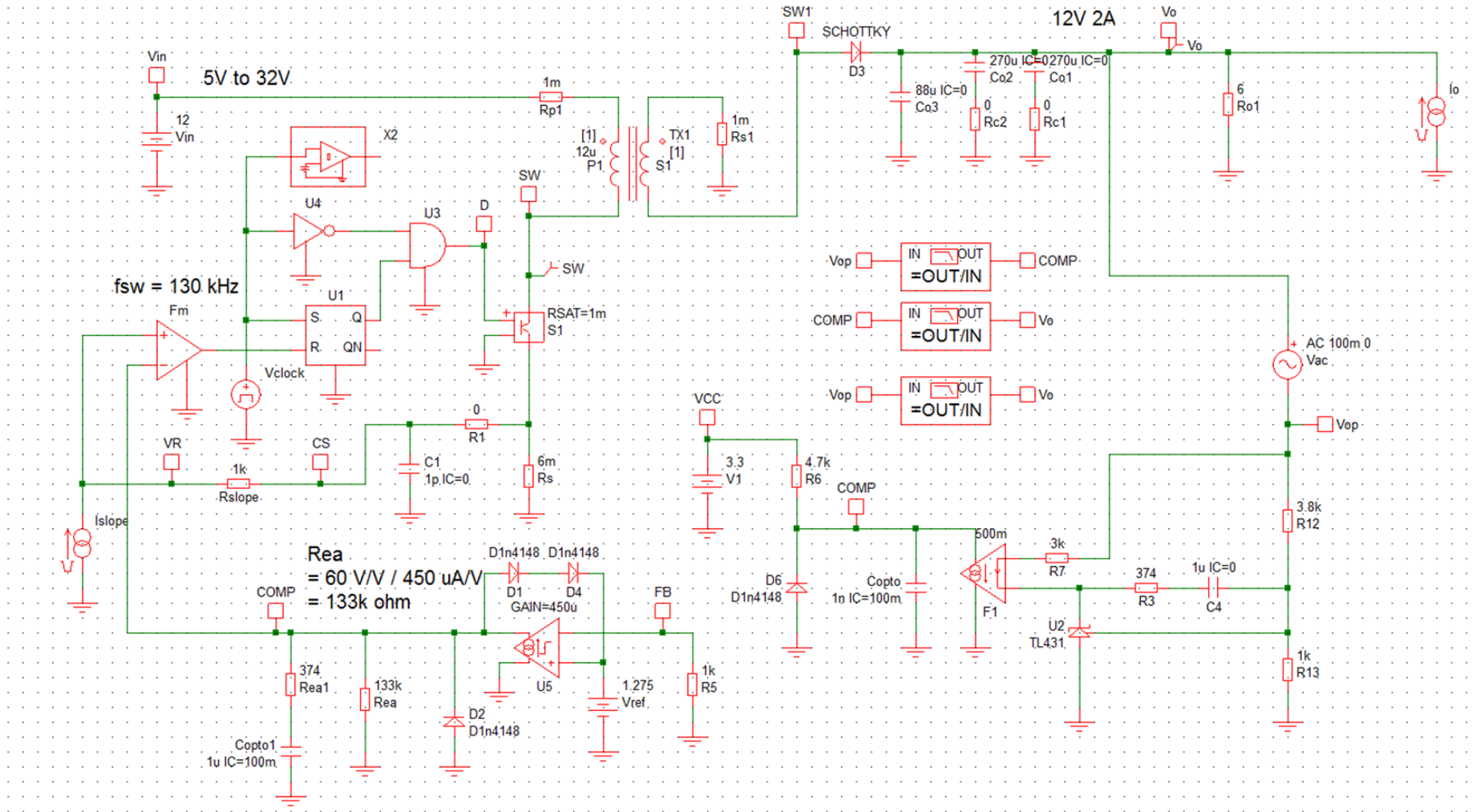
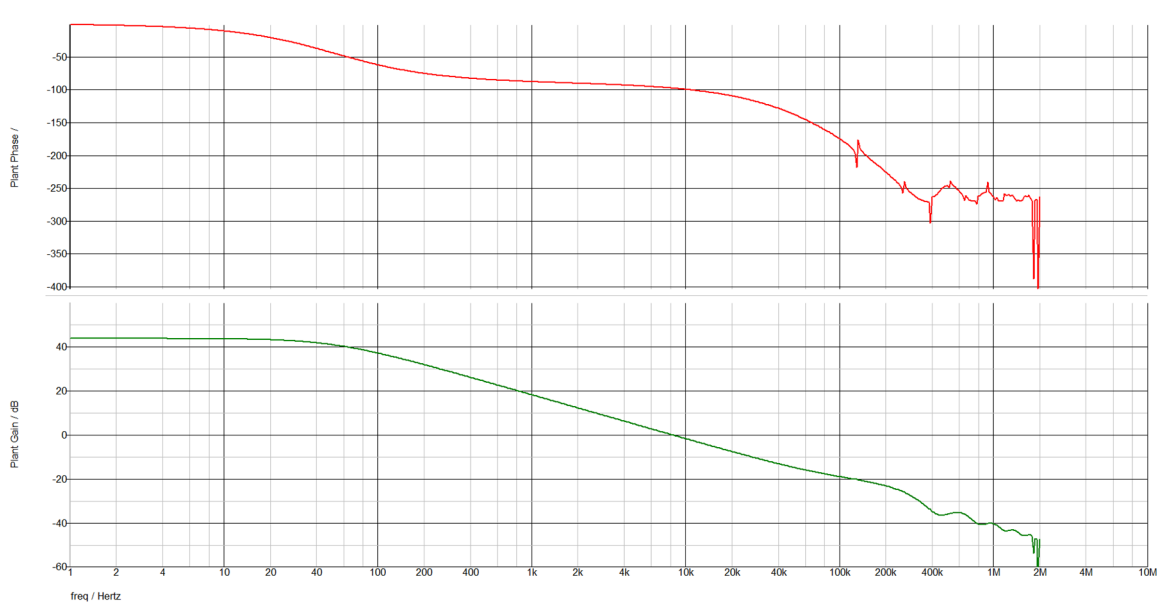


Figure 13. LM3481 Simplis Simulation Schematic


Figure 14. LM3481 Simplis Simulation Results

4.4 LM431 Compensation Circuit Design

Equation 18 provides a two-zero two pole compensation function. This section will introduce a step by step method to design the compensation circuit.

4.4.1 Step 1: Determine the Feedback Resister

Rfb1 and Rfb2 determine the output voltage. The reference voltage of LM431 and output voltage are $V_{ref} = 2.5 \text{ V}$ and $V_{out} = 12 \text{ V}$ respectively. Choose $R_{fb2} = 1 \text{ k}\Omega$, Rfb1 could be calculated as shown in Equation 28.

$$R_{fb1} = R_{fb2} \times \frac{V_o - V_{ref}}{V_{ref}} = 3.8 \text{ k}\Omega; \quad (28)$$

4.4.2 Step 2: Determine the Rup and Ropto, Kopto

As shown in Equation 22, the ratio of the pullup resistance R_{up} and R_{opto} has an impact on Mid gain of compensator. Setting the collector current of 1mA on the opto coupler gives the pull resistance of 4.7k (pullup voltage $V_{cc}/1 \text{ mA}$). Now the CTR (K_{opto} -current transfer ratio) is close to 1 for the chosen optocoupler. This results in the optocoupler LED current of 1 mA which is used to determine maximum allowed R_{opto} . Now $R_{opto} = (V_{out} - V_{ref} - V_{LED}) / 1 \text{ mA}$ where $V_{LED} = 1 \text{ V}$ and $V_{ref} = 2.5 \text{ V}$. This gives maximum value of R_{opto} as 8.5k. A 3K- Ω resistor is used for this application.

4.4.3 Step 3: Calculate the value of R1, R2

Before designing the RC network, usually need to determine the crossover frequency first. In this case we set $f_c = 3 \text{ kHz}$. Figure 11 shows that the current loop DC gain at 3 kHz is about 16.2 dB. That means, the compensation circuit need to provide a -16.2-dB gain at 3-kHz frequency to get a 3-kHz crossover frequency. Meanwhile the gain-frequency must pass across 0 db with -20 dB/dec to ensure enough phase margins.

The Mid-frequency gain of $H_{EA}(s)$ is $K_{opto} \times R_{up} \times R1 / (R_{opto} \times R_{fb1})$, so R1 could be calculated as shown in Equation 29.

$$R_1 = \frac{10^{(-16.2/20)} \times R_{fb1} \times R_{opto}}{K_{opto} \times R_{up}} = 374 \Omega; \quad (29)$$

To reduce the variables and simplify the design, setting $R2=R1=374 \Omega$. Then the compensation transfer function could be considered as one-pole one-zero system.

4.4.4 Step 4: Calculate the Value of C1, C2

The RC network (R1C1 and R2C2) is used to provide a zero to boost the phase and are placed below the crossover frequency. In this application zero were set at $f_{zero} = 400 \text{ Hz}$. C1 and C2 could be calculated as shown in Equation 29.

$$C_1 = C_2 = \frac{1}{2\pi \times R_1} = 1 \mu\text{F} \tag{30}$$

Figure 15 shows the Mathcad calculated Bode plot of compensation circuit, and Figure 17 shows the simplis simulation schematic and results. It could be observed that the calculated results and simulation results matched well.

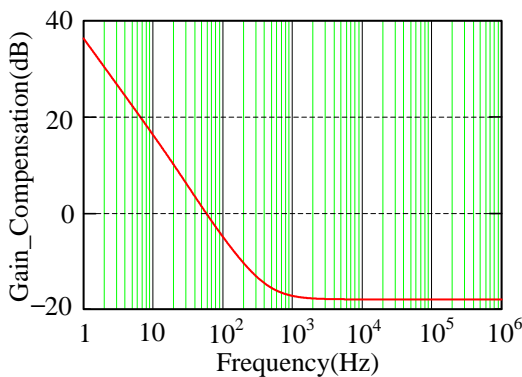


Figure 15. Compensation Circuit Bode Plot Calculated Results – Gain-Frequency

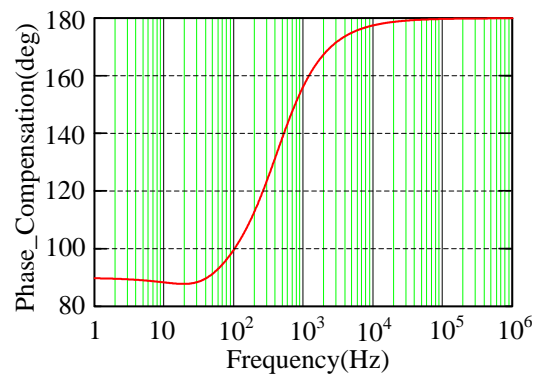


Figure 16. Compensation Circuit Bode Plot Calculated Results – Phase-Frequency

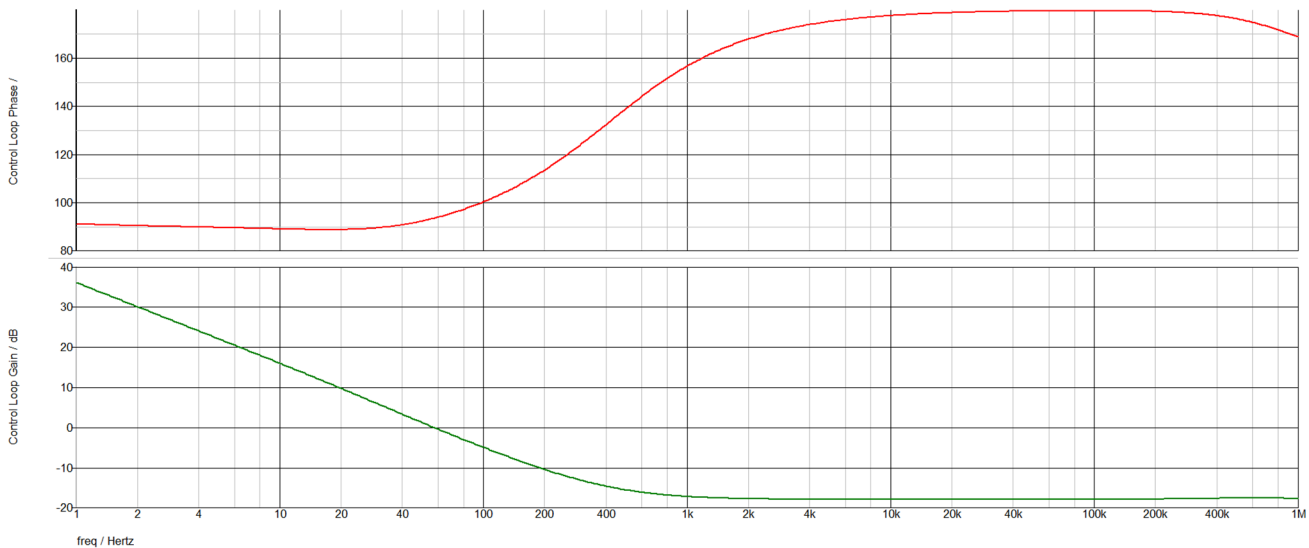


Figure 17. Compensation Simplis Simulation Results

4.5 LM3481 Flyback Loop Response Performance

Figure 18 and Figure 20 show the calculated and simulation results of LM3481 flyback after compensation. It could be observed that the crossover frequency F_c is exactly at 3 kHz and there is about 77° of phase margin. The calculated results and simulation results matched well.

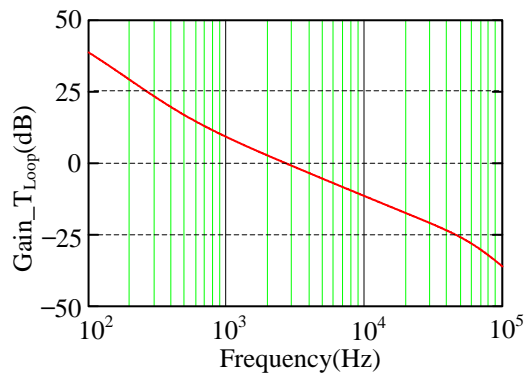


Figure 18. LM3481 Flyback Bode Plot Calculated Results – Gain-Frequency

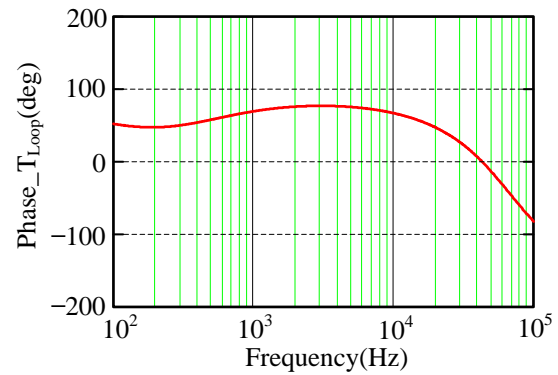


Figure 19. LM3481 Flyback Bode Plot Calculated Results – Phase-Frequency

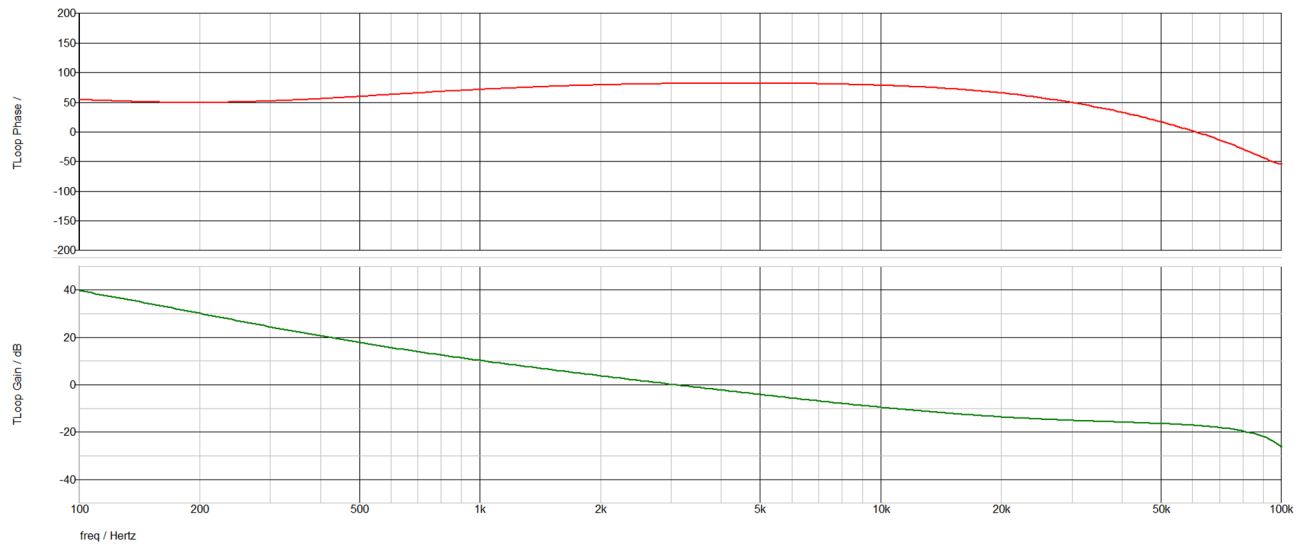


Figure 20. LM3481 Flyback Bode Plot Simulation Results

5 Bench Verification

Figure 21 and Figure 22 show the output voltage ripple and rectifier diode switching waveform (anode to ISO GND) for the LM3481-FlybackEVM. The output current is the rated full load of 2 A and at of $V_{IN} = 5.5$ V and 28 V respectively. It can be observed that the design works in CCM mode over wide input range.

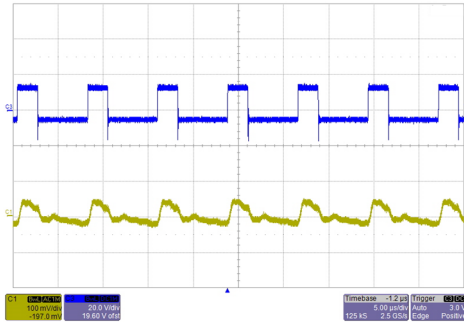


Figure 21. Maximum Output Ripple at Different Input Voltage – $V_{IN} = 5.5\text{ V}$

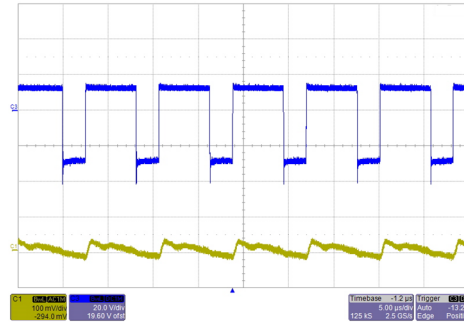


Figure 22. Maximum Output Ripple at Different Input Voltage – $V_{IN} = 28\text{ V}$

Figure 23 shows the LM3481-Flyback load transients performance. The current step is from 50% to 100% of maximum rated load at $V_{IN} = 12\text{ V}$ with 70-mA/µs slew rate. Figure 24 shows the loop-response characteristics. Gain and phase plots are shown for V_{IN} voltage of 12 V with load current of 2 A, which is verified with the previous theory analysis.

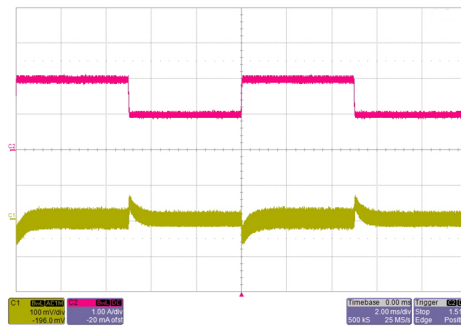


Figure 23. Load Transient Response at 12 VIN

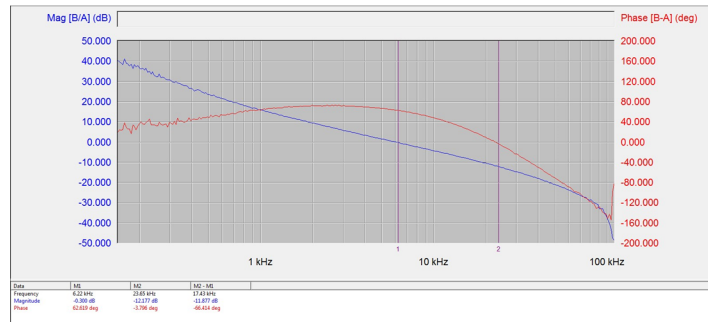


Figure 24. LM3481 Flyback EVM Bode Plot Tested Results

6 Conclusion

The LM3481 device is a versatile low-side N-FET high-performance controller which can support Wide Vin of 2.97 V to 48 V and can easily be configured in flyback topology in various isolated as well as non-isolated single or multiple output applications.

7 References

1. [LM3481 High-Efficiency Controller for Boost, SEPIC, and Flyback DC-DC Converters](#) (SNVS346)
2. [LM3481 Flyback-EVM User Guide](#) (SNVU528)
3. [How to Configure LMZ30604 Power Module with Ceramic Capacitor](#) (SNVA732)
4. [Under the Hood of Flyback SMPS Designs](#) (SLUP261)

Revision History

NOTE: Page numbers for previous revisions may differ from page numbers in the current version.

Changes from Original (September 2016) to A Revision	Page
• Changed cross reference.....	9

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