# Design Of A Small, Efficient, Isolated Flyback Converter For 24-V Input Systems with Si9121

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## INTRODUCTION

The Si9121DY is an excellent choice for a non-isolated buck-boost converter IC capable of operating from a wide input–voltage range of –10 V to –60 V to produce a +3.3-V or +5.0-V output. The power IC contains both the controller and an internal switch to enable it to be used as a converter with the minimal amount of external components. This allows a small, cost-effective converter to be built. The design of this converter is described, in detail, in the Vishay Siliconix application note AN732, "Designing a High-Voltage Non-Isolated Buck-Boost Converter With the Si9121DY."

In some circumstances there is a requirement for a cost-effective, low-power, isolated converter with a minimum amount of external components and design expertise required. Although the Si9121 is designed for a non-isolated buck-boost converter, the internal MOSFET lends itself to use in isolated converters. This application note describes the design and construction of such a converter with a nominal 24-V input and a 3.3-V output with a continuous power rating of 1.3 W. The design uses the Siliconix Si9121DY power integrated circuit, described above as the controller, and implements its internal MOSFET as the primary switch. Using the internal MOSFET, coupled with a flyback transformer and a nominal amount of external parts, makes for a cost-effective power supply with a very small size envelope. The power IC includes an internal floating feedback-error amplifier, a fixed-frequency oscillator, an output-voltage-sensing resistor divider, and a depletion-mode MOSFET for start-up $V_{CC}$ regulation reducing the external component count to less than 15.

The IC features current-mode control to achieve good line transient response, and also includes undervoltage lockout, programmable soft-start, pulse-by-pulse current limit, hiccup mode with negligible power delivery and dissipation during continuous short-circuit conditions, automatic recovery from hiccup after fault removal, and overtemperature shut-down. The Si9121DY is available in a narrow-body SO-8 package and allows dissipation of up to 1.3 W.

A detailed study of the design of the flyback transformer is presented. An Excel spreadsheet design tool is also provided to enable readers to tailor the design to their specific requirements.

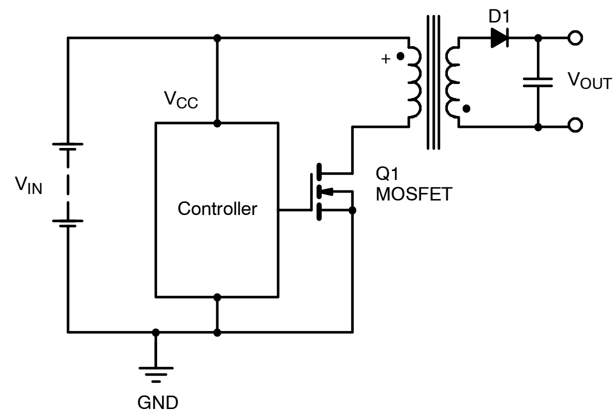


**FIGURE 1.** Block Diagram of Flyback Converter

## CIRCUIT TOPOLOGY AND DESCRIPTION

For this power level the most suitable isolated topology is the flyback converter. These converters can provide either single or multiple outputs. Flyback converters are more suitable than forward converters for relatively low power levels because of their lower circuit complexity as shown in Figure 1. Their relative simplicity results from the elimination of the output inductor and freewheel diode that would be present in the secondary stage of a forward converter. The energy acquired by the transformer during the on-time of the primary MOSFET is delivered to the output in the non-conducting period of the primary switch. During the conduction period of the primary MOSFET, the current flows into the positive terminal (+) of the primary winding through the switch to GND. The secondary winding is connected with reverse polarity, so there can be no current flow to the output, due to the blocking diode (D1). When the primary MOSFET ceases to conduct, the induced voltage is reversed by the collapse of the magnetic field, and the output capacitor is charged through the diode (D1). Since the circuit only requires one magnetic component—the transformer—flyback converters are simpler and cheaper than forward converters to design and build.

## TRANSFORMER FLYBACK INDUCTOR DESIGN

Because the major element of change from the non-isolated design described in application note AN732 is the inclusion of the transformer, it is the focus of this application note. For further circuit design information regarding the Si9121, please see AN732.

The area product is a method of assigning a power-handling ability to a magnetic core. It is the product of the window area $W_a$ of the available winding space and the cross-sectional area $A_c$ of the magnetic core. Multiplying the two together gives the area product or $A_p$ of the core. Two Vishay Siliconix application notes cover this topic in detail and approach the problem in two different ways. The AN713 application note uses an already-gapped core (delta core) and then finds the primary inductance, the primary turns, and then the inductance factor $A_L$, whereas the AN707 application note finds the primary inductance, the air gap, then $A_L$, and finally the primary turns. The method described and used in this paper combines aspects of both the AN707 and AN713.

### Determination of the Primary Inductance

The flyback converter in this circuit is designed to operate in discontinuous operation. To ensure this, the primary inductance $L_p$ needs to be limited to a maximum value. Therefore, the maximum primary inductance for discontinuous conduction at maximum load needs to be determined. The input power is defined as:

$$P_{IN} = \frac{P_{OUT}}{\eta} \quad (1)$$

where $\eta$ is the efficiency of the inductor.

If the flyback inductor/transformer is assumed to be 80% efficient, and the maximum output power $P_{OUT}$ is defined as 1.32 W (3.3 V @ 400 mA), then the input power will be 1.65 W.

The input power can also be defined as the product of the stored energy E in the magnetic field and the switching frequency $f_s$:

$$P_{IN} = Ef_s = \frac{L_p I_{pk}^2 f_s}{2} \quad (2)$$

This allows the required primary inductance to be determined, but it is also necessary to know the peak current $I_{pk}$ to be able to calculate this.

The peak current is defined by [1].

$$I_{pk} = \frac{V_{IN(min)} t_{ON(max)}}{L_p} \quad (3)$$

where $V_{IN(min)}$ is the minimum supply voltage and $t_{ON(max)}$ is the maximum on-time of the switch.

Therefore, to limit the primary inductance to ensure discontinuous operation, the maximum inductance is determined by:

$$L_{p(max)} \le \frac{V_{IN(min)} t_{ON(max)}}{I_{pk}} = \frac{V_{IN(min)}}{I_{pk}} \delta_{max} T_s \quad (4)$$

where $\delta_{max}$ is the maximum duty cycle and $T_s$ is the PWM switching period.

By combining equations 2 and 4 it is possible to determine the maximum primary inductance:

$$L_{p(max)} = \frac{V_{IN(min)}^2 \delta_{max}^2 T_s}{2P_{IN(max)}} \quad (5)$$

With a minimum input voltage of 10 V, a maximum duty cycle of 0.45, a switching frequency of 95 kHz, and a maximum input power, as determined above, of 1.65 W, the maximum inductance $L_{p(max)}$ from equation 5 is 64.6 μH. Substituting this value into equation 4 gives a peak primary current of $I_{pk}$ = 0.73 A.

### Core Selection

The area product method is used to determine the inductor core size [2] is defined as:

$$A_p = \left(\frac{2E \times 10^4}{B_m K_u K_j}\right)^{1.14} \quad (6)$$

where $A_p$ is the area product, $E = \frac{L_p I_{pk}^2}{2}$ and is the core energy-storage requirement, $B_m$ is the maximum flux density, $K_u$ is the window-utilization factor, and $K_j$ is the current-density coefficient.

However, this empirical equation [6] applies for the area product of single-winding inductors. To account for the secondary windings, which will handle the same energy as the primary, this equation needs to be multiplied by two. The result is:

$$A_p = 2\left(\frac{L_p I_{pk}^2 \times 10^4}{B_m K_u K_j}\right)^{1.14} \quad (7)$$

In this case, limiting the maximum flux density to 0.1500 tesla, $K_j$ to 433, and $K_u$ to 0.15 results in an area-product requirement of 0.0447 $cm^4$. From the manufacturer's databook,[3] the RM6 core fulfils this requirement with an area product of 0.0507 $cm^4$.

### Determination of the Number of Primary Turns Np

Once a particular core has been chosen, then the minimum air-gap $l_g$ requirement can be obtained from:

$$l_G = \frac{1.26 L_p I_{pk}^2}{A_c B_m^2} \quad (8)$$

[1] "Switching Power Supply Design," Pressman, McGraw Hill.
[2] "Magnetic Core Selection for Transformers and Inductors," McLyman.
[3] "Magnetic Products, Soft Ferrites, Data Handbook MA01, Philips."

where $A_c$ is the effective cross-section of the core.

In this case $L_{p(max)}$ = 64.6 μH, $I_{pk}$ = 0.73 A, $A_c$ = 0.32 cm$^2$, $B_m$ = 0.15 gives a minimum air gap of 60.2 μm. The closest "off-the-shelf" core gap for this RM6 core was 110 μm.

Once the minimum air gap has been determined, then it is possible to calculate the primary turns from the inductance value specified in the ferrite data for the particular core and air gap because:

$$N_p = 1000\sqrt{\frac{L_p}{A_l}} \tag{9}$$

The RM6 core, with the 110-μm air gap, had an inductance factor $A_l$ of 250 mH, and using the 64.6-μH inductance results in 16 primary turns.

**Determination of the Number of Secondary Turns $N_s$**

To guarantee discontinuous conduction mode at the maximum load current, it is necessary to limit the inductance of the secondary windings to some maximum value. The negative current slope during the flyback period is:

$$\frac{di}{dt} = \frac{I_s}{t_{fly}} = \frac{V_O + V_D}{L_s} \tag{10}$$

where $I_S$ is the peak current in the secondary winding, $t_{fly}$ is the flyback period, $V_O$ is the output voltage from the winding, $V_D$ is the voltage drop across the output diode, and $L_s$ is the secondary inductance.

The rectifier or flyback duty cycle $\delta_r$ and is defined as:

$$\delta_r = \frac{t_{fly}}{T_s} \tag{11}$$

and the load current $I_O$ is related to the peak secondary current and duty cycle by:

$$I_S = \frac{2I_O}{\delta_r} \tag{12}$$

By combining equations 9, 10, and 11 and specifying a maximum duty cycle, the maximum secondary inductance can be specified as:

$$L_s \le \frac{\delta_r^2(V_O + V_D)T_s}{2I_O} \tag{13}$$

In this case, with a maximum duty cycle of 0.55, a maximum PWM switching frequency of 95 kHz, an output voltage of 3.3 V, an output current of 400 mA and a diode conduction voltage of 0.5 V gives a maximum secondary inductance of 15 μH.

$$N_s = 1000\sqrt{\frac{L_s}{A_l}} \tag{14}$$

By using equation 14, the number of secondary turns for NS1 is 7.7 turns, and therefore a value of 7 turns. The same number of turns should be used for NS2. Scaling NS3 for 12 V (plus 0.5 V) results in 23 turns.

$$N_{s3} = N_{s1}\frac{V_{O2} + V_D}{V_{O1} + V_D} \tag{15}$$

It should be noted that due to the original application of the Si9121DY, an additional winding, identical to the output winding, needs to be provided to give an isolated feedback signal to the controller. The second feedback winding, NS3, provides power to the Si9121DY during normal operation. This could be removed and the supply to the controller taken from $V_{IN}$.

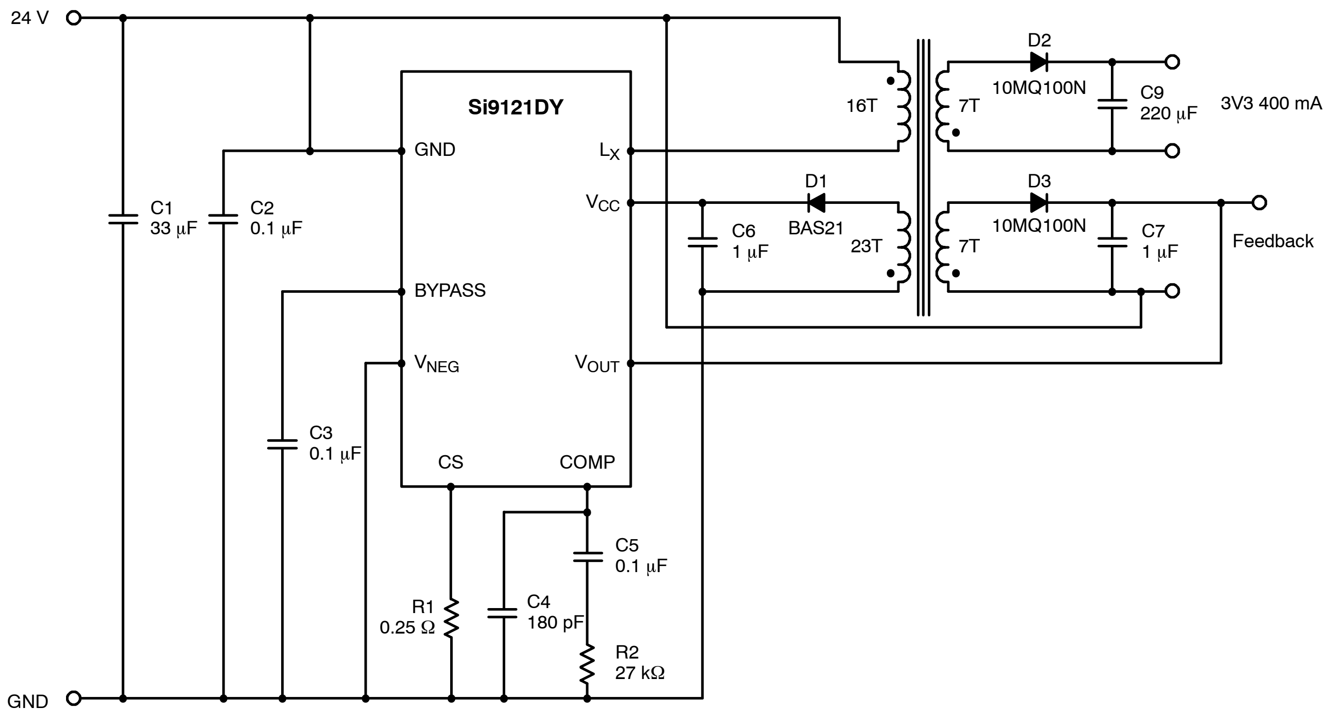


**FIGURE 2.** Circuit Schematic of Isolated Flyback Si9121

## WINDING CONSIDERATIONS

The skin effect can be present in copper conductors as well as in the magnetic core. If the skin depth $d_{skin}$ is less than the diameter of the copper wire, then eddy currents flow in the conductor. These eddy currents tend to flow in the opposite direction of the applied current in the center of the wire, and as such very little of the applied current flows through the center of the wire. The skin depth is dependent on frequency and is defined as [4]:

$$d_{skin} = \sqrt{\frac{2}{\varpi\mu\sigma}} = \sqrt{\frac{1}{\pi f_s \mu\sigma}} \tag{16}$$

where $\delta$ is the conductivity ($6.02 \times 10^7$) and $\mu$ is the magnetic permeability ($4\pi \times 10^{-7}$).

If the switching frequency is low, then the skin depth is greater than the radius of the wire, in which case significant eddy currents are not present and the entire cross-sectional area of the wire is being used. However, as the switching frequency is increased, the skin depth reduces and can cause a significant part of the copper conductor not to be utilized, causing the effective resistance of the conductor to be far larger than the dc resistance. Reducing the diameter of the wire can overcome this problem, and it has been shown[4] that if the diameter of the wire is less than twice the skin depth, the consequences of the skin effect are negligible.

In this case, a switching frequency of 95 kHz gives a skin depth of 0.21 mm, which means that the diameter of wire must be less than 0.42 mm. In this example, two lots of SWG31 were used for the primary, and three lots of SWG29 were used for the secondary.

Windings should be sequenced so that the highest-power secondary is closest to the primary. This keeps the lower-power secondary out of the highest field region and has the added benefit of minimizing the adverse effects of leakage inductance on cross-regulation.

## OUTPUT REGULATION WITH LOW PASS FILTER

A low-pass filter is included on the feedback winding to eliminate spurious voltage spikes, which give false information for the error amplifier. This filter consists of R7 and C7 as shown in Figure 3. The optimum value for R7 was determined to be 51 Ω. The filter removes the peak excursion seen at the beginning of the switching transient. This slows down the response of the system but gives the feedback a steady-state value rather than a peak value.

[4] “Power Electronics, Converters, Applications and Design,” Mohan, Undeland and Robbins,

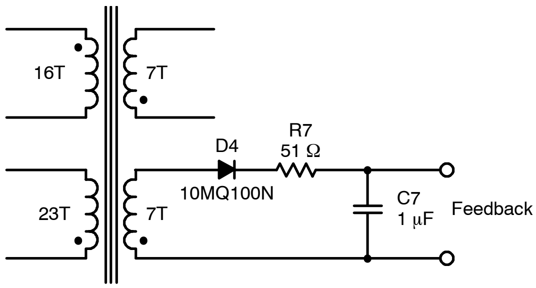


**FIGURE 3.** Inclusion of Low Pass Filter In Feedback Winding

## RESULTS

The output regulation achieved with the 51-Ω resistor in the feedback output is shown in Table 1. This design provides an accurate no-load output of 3.3-V but resulted in a 3.04-V output voltage the full-load current.

**TABLE 1**
**Output Regulation with Low Pass Filter In Feedback Winding**

| Current (mA) | Output Voltage (V) |
|---|---|
| 100 | 3.13 |
| 200 | 3.08 |
| 300 | 3.06 |
| 400 | 3.04 |

To increase the voltage output under full-load conditions, the no-load voltage output was increased to 3.5 V using a potential divider (Figure 4).

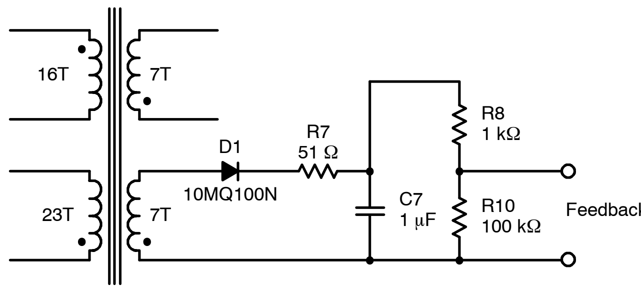


**FIGURE 4.** Inclusion of Potential Divider In Feedback Winding

This circuit results in a 3.3-V voltage output with a full-load-range regulation of +2% and –2.4% as shown in Table 2. The complete schematic is shown in Figure 11.

## OUTPUT REGULATION WITH A POTENTIAL DIVIDER

**TABLE 2**
**Output Regulation with Low Pass Filter And Increased Voltage Output**

| Current (mA) | Output Voltage (V) |
|---|---|
| 50 | 3.36 |
| 100 | 3.31 |
| 200 | 3.26 |
| 300 | 3.24 |
| 400 | 3.22 |

### Output Ripple and Transient Response

Figures 5 through 8 show the output ripple and transient response for the isolated Si9121DY converter at both 200-mA and 400-mA output currents for the complete schematic shown in Figure 11.

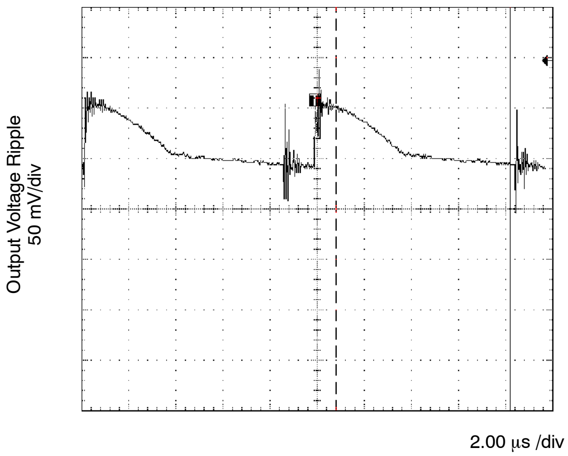


**FIGURE 5.** Voltage Output Ripple with 200-mA

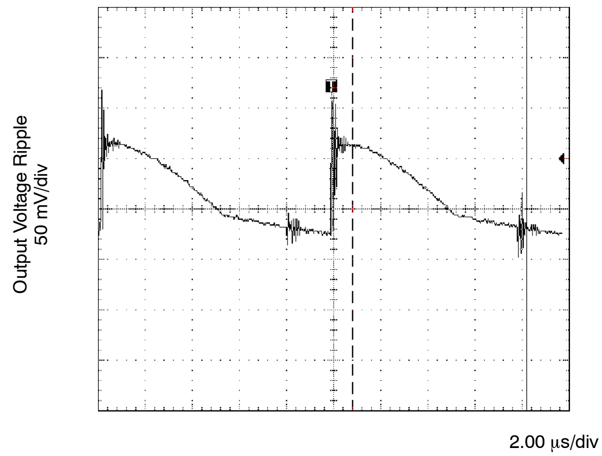


**FIGURE 6.** Voltage Output Ripple with 400-mA

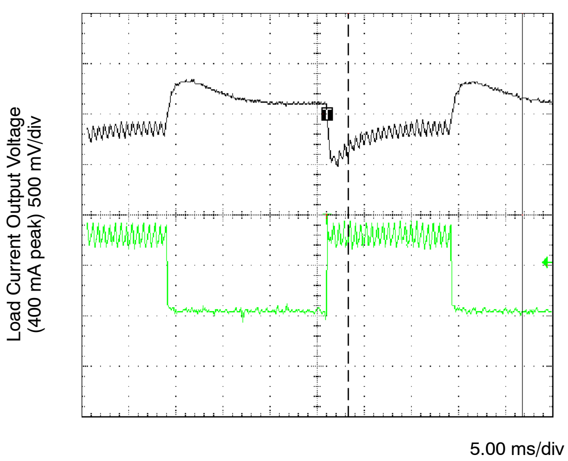


**FIGURE 7.** Transient Load Response with 400-mA Stepped Load

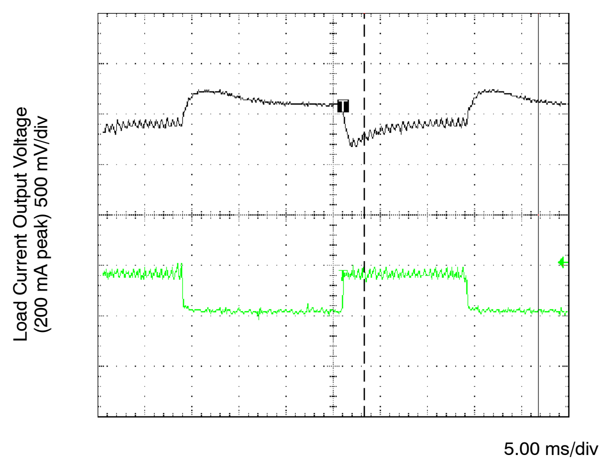


**FIGURE 8.** Transient Load Response with 200-mA Stepped Load

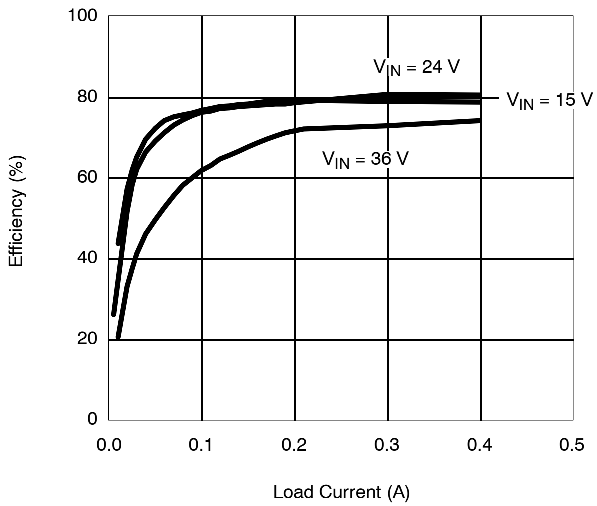


**FIGURE 9.** Efficiency vs. Output Current

**Input Filter**

It should be noted that MIL-STD-461 EMI testing was not performed for this supply. To meet CEO3 and CSO1 limits, some input filter redesign is required. [5]

**Possible Alternative Circuit Configurations**

With a negative input voltage, it is possible to utilize the feedback winding as an additional non-isolated supply, as shown in Figure 12. This is a suggested circuit configuration, with the operation, regulation, response, and efficiency of the circuit not investigated in this application note.

## CONCLUSIONS

This application note shows the basic design of a small isolated flyback converter. Efficiencies of over 80% can be achieved at full load, and regulation of ±2.5% (ignoring no-load conditions) is achieved as shown in Figure 10. Figure 11 shows the circuit schematic for the final design. An Excel design tool for the design of the flyback transformer is demonstrated and attached in the appendix.

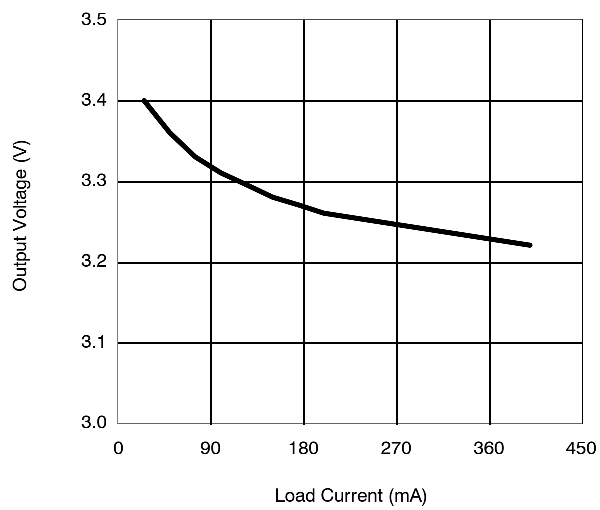


**FIGURE 10.** Output Regulation For A 100-mA to 400-mA Load

Application notes and data sheets are available on the internet at www.vishay.com. The applications notes of interest are:

- AN707 – Designing DC/DC Converters with the Si9110 Switch Mode Controller
- AN713 – A 1 watt Flyback Converter Using the Si9100
- AN732 – Designing a High Voltage Non-Isolated Buck-Boost Converter with the Si9121DY

[5] Middlebrook, R. D., "Input Filter Considerations In Design and Application of Switching Regulators," IEEE Industry Applications Society Annual Meeting, Oct. 11-14, 1976.

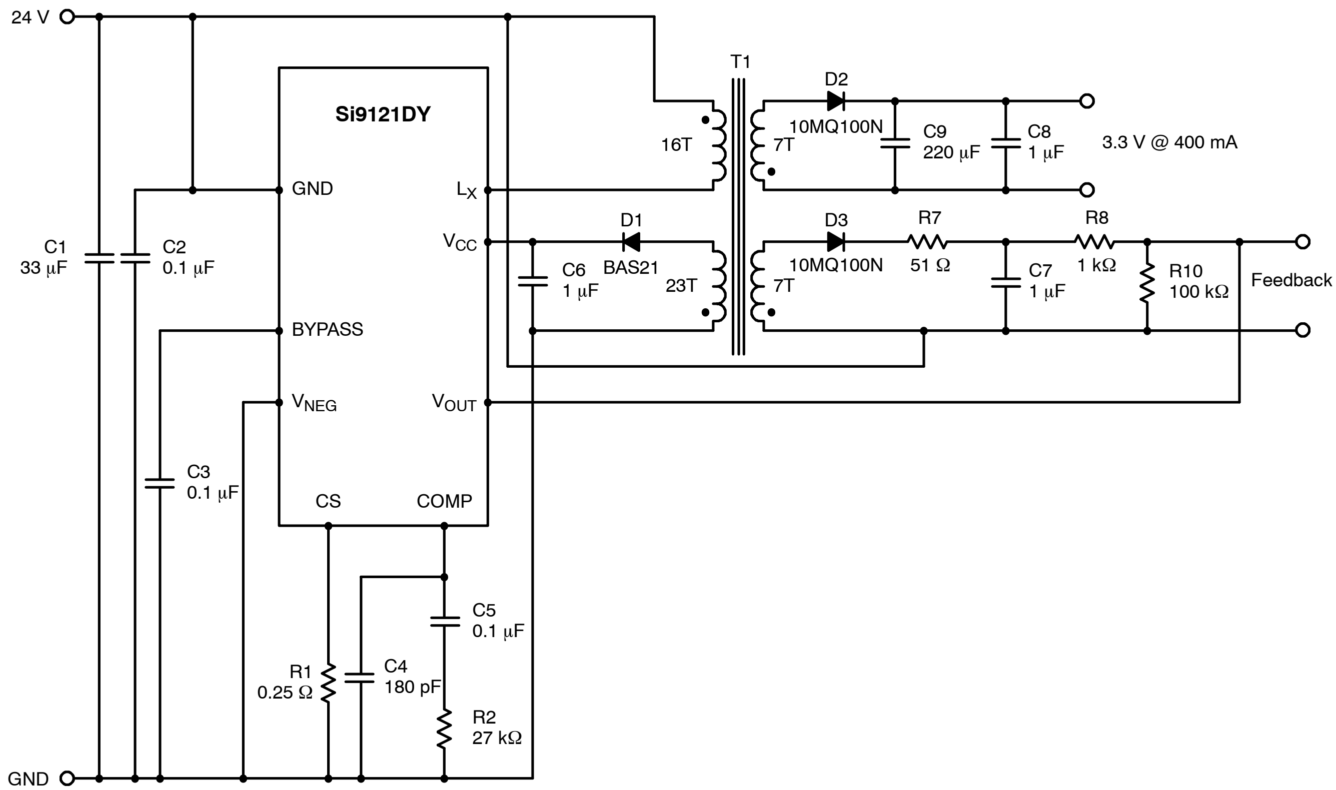


**FIGURE 11.** Circuit Schematic For Final Design

**TABLE 3: BILL-OF-MATERIALS FOR FINAL DESIGN (Figure 11)**

| Item | Qty | Designator | Part Type | Description | Footprint | Vendor Part# | Manufacturer |
|---|---|---|---|---|---|---|---|
| | | | | | | | |
| 1 | 1 | R1 | 0.25 Ω | RES,1%, 1/2 W, PWER Metal Strip | WSL-2010 | WSL-2010 | Vishay Dale |
| 2 | 1 | R2 | 27 kΩ | RES,1%, 1/8 W | 0805 | CRW08057502RF | Vishay Dale |
| 3 | 1 | R7 | 51 Ω | RES,1%, 1/8 W | 0805 | CRW080551R0F | Vishay Dale |
| 4 | 1 | R8 | 1 kΩ | RES,1%, 1/8 W | 0805 | CRW08051001R0F | Vishay Dale |
| 5 | 1 | R10 | 100 kΩ | RES,1%, 1/8 W | 0805 | CRW08051003R0F | Vishay Dale |
| | | | | | | | |
| 6 | 1 | C1 | 33 µF | CAP, ELEC, 80 V, PF | RB.14/.32 | UPF1K330MPH6 | Nichicon |
| 7 | 3 | C2, C3, C5 | 0.1 µF | CAP, CER | 0805 | VJ0805Y104KXXA | Vishay Vitramon |
| 8 | 1 | C4 | 180 pF | CAP, CER | 0805 | VJ0805Y181KXXA | Vishay Vitramon |
| 9 | 3 | C6, C7, C8 | 1 µF | CAP, CER, 50 V | 0805 | VJ0805Y105KXAAT | Vishay Vitramon |
| 10 | 1 | C9 | 220 µF | CAP, TAN, 10 V | 594D_D | 594D227X0010D2T | Vishay Sprague |
| | | | | | | | |
| 11 | 1 | D1 | BAS21-GS08 | Schottky Diode, 200 V | SMA | BAS21-GS08 | Vishay Telefunken |
| 12 | 2 | D2, D3 | 10MQ100N | Schottky Diode, 35 V | SMA | 10MQ100N | I. R. |
| | | | | | | | |
| 13 | 1 | T1 | RM6R-3B7-A250 | Transformer Core | | RM6R-3B7-A250 | Philips |
| | | | | | | | |
| 14 | 1 | U1 | Si9121DY-3 | Power IC | SO-8 | SI9121DY-3 | Vishay Siliconix |

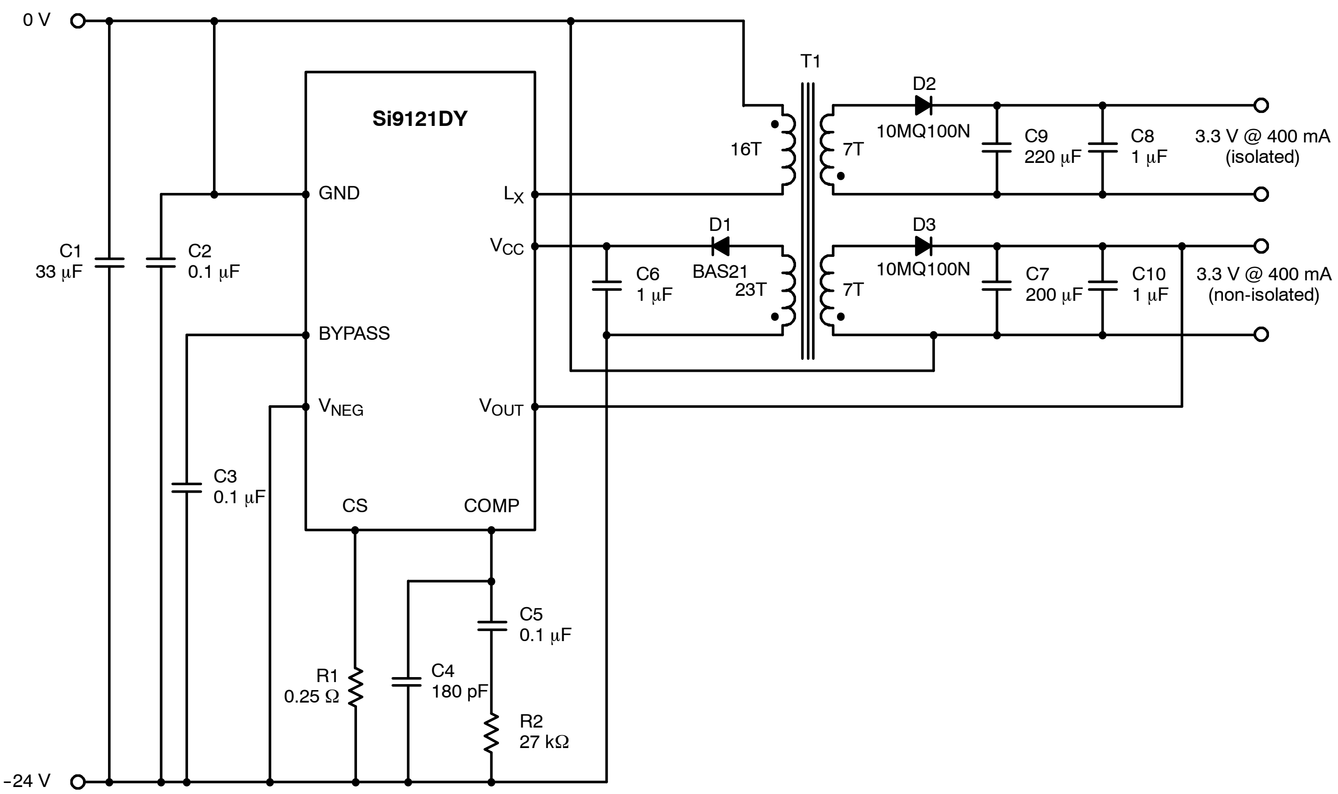


**FIGURE 12.** Suggested Circuit Schematic For Negative Input Voltage

| **TABLE 4: BILL-OF-MATERIALS FOR NEGATIVE INPUT VOLTAGE (Figure 12)** | | | | | | | |
|---|---|---|---|---|---|---|---|
| **Item** | **Qty** | **Designator** | **Part Type** | **Description** | **Footprint** | **Vendor Part#** | **Manufacturer** |
| | | | | | | | |
| 1 | 1 | R1 | 0.25 Ω | RES,1%, 1/2 W, PWER Metal Strip | WSL-2010 | WSL-2010 | Vishay Dale |
| 2 | 1 | R2 | 27 kΩ | RES,1%, 1/8 W | 0805 | CRW08057502RF | Vishay Dale |
| | | | | | | | |
| 3 | 1 | C1 | 33 µF | CAP, ELEC, 80 V, PF | RB.14/.32 | UPF1K330MPH6 | Nichicon |
| 4 | 3 | C2, C3, C5 | 0.1 µF | CAP, CER | 0805 | VJ0805Y104KXXA | Vishay Vitramon |
| 5 | 1 | C4 | 180 pF | CAP, CER | 0805 | VJ0805Y181KXXA | Vishay Vitramon |
| 6 | 3 | C6, C8, C10 | 1 µF | CAP, CER, 50 V | 0805 | VJ0805Y105KXAAT | Vishay Vitramon |
| 7 | 2 | C7, C9 | 220 µF | CAP, TAN, 10 V | 594D_D | 594D227X0010D2T | Vishay Sprague |
| | | | | | | | |
| 8 | 1 | D1 | BAS21-GS08 | Schottky Diode, 200 V | SMA | BAS21-GS08 | Vishay Telefunken |
| 9 | 1 | D2, D3 | 10MQ100N | Schottky Diode, 35 V | SMA | 10MQ100N | I. R. |
| | | | | | | | |
| 10 | 1 | T1 | RM6R-3B7-A250 | Transformer Core | | RM6R-3B7-A250 | Philips |
| | | | | | | | |
| 11 | 1 | U1 | Si9121DY-3 | Power IC | SO-8 | SI9121DY-3 | Vishay Siliconix |