

## Abstract

The [ISL6752/54EVAL1Z](#) is a new design based on the ISL6752EVAL1Z but with several design modifications to improve the efficiency from 90% to 95%. The control circuit has been moved off the main board onto a daughter card. Two different daughter cards are provided: one using the ISL6752 and the other using the ISL6754. Both control cards use the Intersil Zero Voltage Switching (ZVS) topology. The ISL6752 daughter card features pulse by pulse current limiting, and the ISL6754 daughter card features a patented method for average current limiting that results in a brick-wall current limit profile.

The ISL6752/54EVAL1Z Printed Circuit Board (PCB) layout has also been greatly improved over the ISL6752EVAL1Z. Even though the overall size of the board has been reduced, the copper losses have been reduced as well.

In addition to the ZVS function, this board also incorporates N-Channel FETs as secondary side rectifiers, also known as Synchronous Rectifiers (SR). Power dissipation of the secondary side rectifiers is reduced because the conduction losses of SRs are significantly less than the conduction losses of PN or Schottky diodes.

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## Scope

This application note covers implementation of Synchronous Rectifiers (SRs) and their associated drive circuits as used on the ISL6752/54EVAL1Z board. Implementation of the primary side ZVS controller, based on the ISL6752 daughter card, is described extensively in Intersil Application Note [AN1262](#), “Designing with the ISL6752, ISL6753 ZVS Full-Bridge Controllers.”

Also reviewed is the performance of this evaluation board. Oscillographs illustrate the performance of the power supply with load transients on the output. The ZVS switching of the bridge FETs is shown, and efficiency and load regulation are measured.

At the end of this application note, the schematics, Bill of Materials (BOM), and PCB layouts are included for reference.

TABLE 1. SPECIFICATIONS

Absolute Maximum Input Voltage	450VDC
Operating Input Voltage	350VDC to 450VDC
Maximum Input Current	2.5ADC
Rated Output Current	50ADC
Current Limit	60A ± 5%
Output Voltage	12V ± 5%
Efficiency at 100% (50A) Load	95%
Efficiency at 20% (10A) Load	92%



FIGURE 1. NEW ISL6752/54EVAL1Z

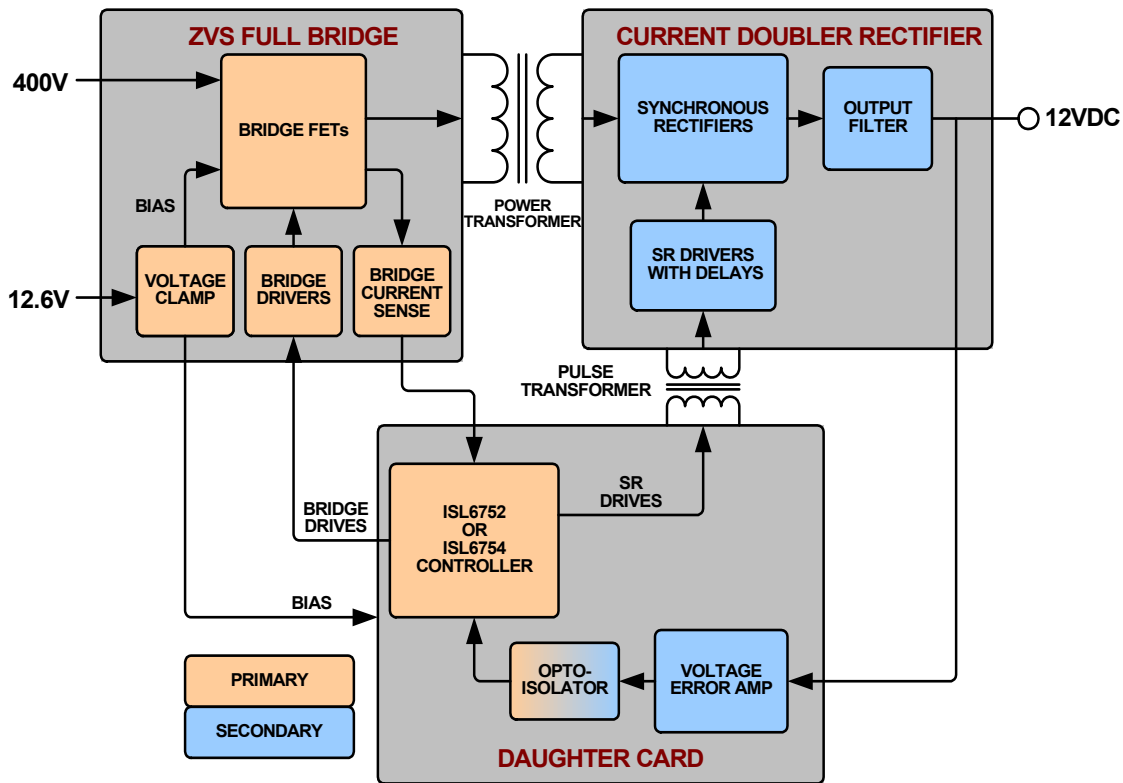


FIGURE 2. ISL6752/54EVAL1Z BLOCK DIAGRAM

## Block Diagram

The evaluation board is composed of several distinct circuit elements. The three main sections are the ZVS full bridge on the input, the current doubler rectifier on the output, and the controller daughter card. See [“Schematics- Main Board, ISL6752/54EVAL1Z” on page 25](#) and [“Schematics - Daughter Card” on page 27](#) for complete circuit details.

## Daughter Cards

The ISL6752 and ISL6754 control ICs are located on their respective daughter cards, as shown in [Figures 3 and 4](#). Both daughter cards have the control ICs on the primary side and the voltage error amplifier on the secondary side. Creepage spacing between the primary and secondary is maintained on the cards.

The ISL6752 and the ISL6754 control ICs are located on the primary side, eliminating the need for two AC line isolating gate drive transformers to drive the primary side bridge FETs. Instead, the low side FETs are driven directly by MOSFET drivers (on the main board), and the high side FETs are driven by a gate drive transformer that only requires operational insulation. Primary side control also simplifies design of the current sensing transformers because they also do not have to be AC line isolating.

A line isolation rated opto-coupler (D5 on the ISL6752DB or D2 on the ISL6754DB) passes the analog error signal generated by the error amplifier, U1, from the secondary to the primary. Opto D3 passes a digital signal from primary to secondary to turn off the SRS for diode emulation.

The only functional difference between the ISL6752 and ISL6754 daughter cards is how current limit is implemented. The ISL6752 uses pulse-by-pulse current limit, and the ISL6754 uses average current limit.

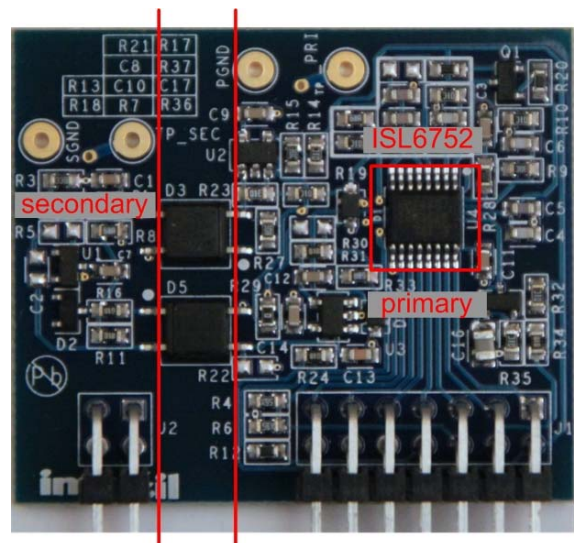


FIGURE 3. ISL6752 DAUGHTER CARD

Special test points located on the daughter cards aid in probing nodes on the daughter cards for evaluation. Test points PGND and TP\_PRI are located on the primary side, and SGND and TP\_SEC are located on the secondary. With these test points, the user can employ the spring-like probe accessories included with many scope probes. The scope signal pin is inserted in TP\_PRI (or TP\_SEC), and the short spring ground lead is inserted in PGND (or SGND).

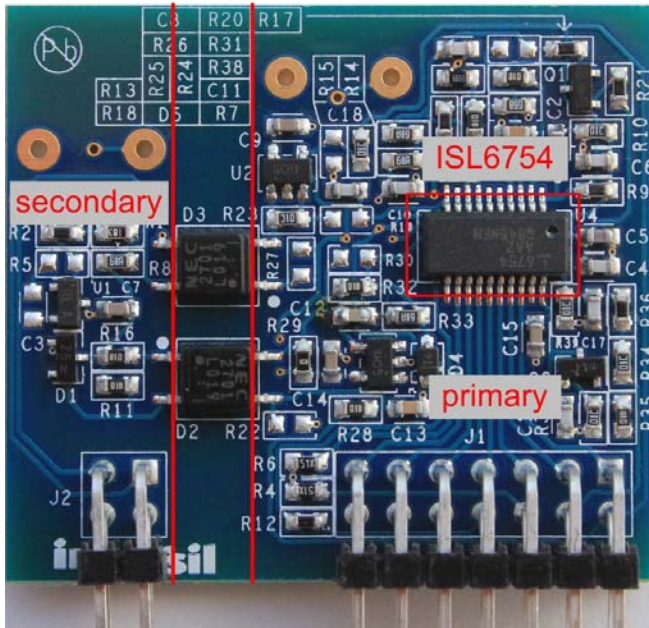


FIGURE 4. ISL6754 DAUGHTER CARD

To probe any node on the daughter card, solder a 30ga insulated wire between the desired node and the via that is associated with the TP\_PRI or TP\_SEC test point. This method not only simplifies probing of any node, but also implements the preferred technique of measuring small signals in the presence of high amplitude switching magnetic fields.

The ISL6752DBEVAL1Z and the ISL6754DBEVAL1Z daughter cards are also available as standalone evaluation kits.

## ZVS Full Bridge

The low side FETs, Q<sub>3</sub> and Q<sub>4</sub>, are driven directly by the ISL89160 MOSFET driver, U1 (Figure 5). The two high-side FETs, Q<sub>1</sub> and Q<sub>2</sub>, are indirectly driven by the ISL89160 driver, U2. A level translating gate drive transformer, T<sub>3</sub>, with complementary output windings, directly drives the high-side bridge FETs with a symmetrical square wave. The design of T<sub>3</sub> is simplified because it only needs 400V operational insulation, and it is always driven with a square wave, thus eliminating the problems associated with non-symmetrical drive waveforms.

Observe that the ISL89160 MOSFET drivers are located as close as possible to their respective bridge FETs to minimize the detrimental effects of parasitic inductance on the outputs of the drivers. Although the input signal lead lengths between the drivers and the daughter card are relatively long (about 5cm), they are shielded on top and bottom by ground planes to significantly reduce the noise injected on these lines. The

hysteresis of the ISL89160 inputs also reduces the possibility of noise corrupting the gate signals.

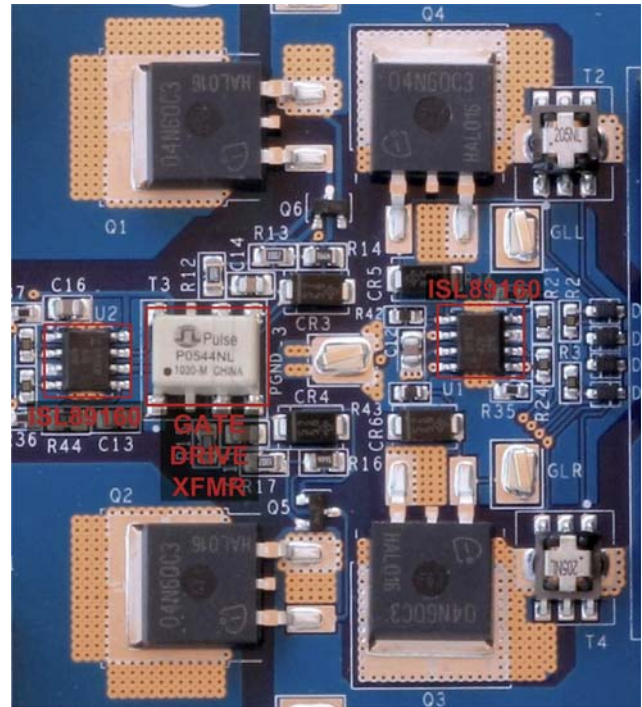


FIGURE 5. FULL BRIDGE

## High Voltage Protection

Because a bridge failure can cause catastrophic damage to the primary side control elements, a voltage crowbar, F1 and D<sub>3</sub>, and a voltage blocking diode, D<sub>4</sub>, are incorporated (Figure 6). D<sub>3</sub> clamps the bias voltage to a safe level. If 400V are applied to the V<sub>DD</sub> node, F1 opens shortly after D<sub>3</sub> conducts current. D<sub>4</sub> provides additional protection by blocking high voltage from being applied to the 13V lab supply. Note that a fully debugged power supply does not need these additional components. These parts are included on the evaluation board to minimize damage if the user accidentally introduces a fault while evaluating the circuits. The designer may want to keep F1 in the final design, to prevent a loud bang if the bridge does fail.

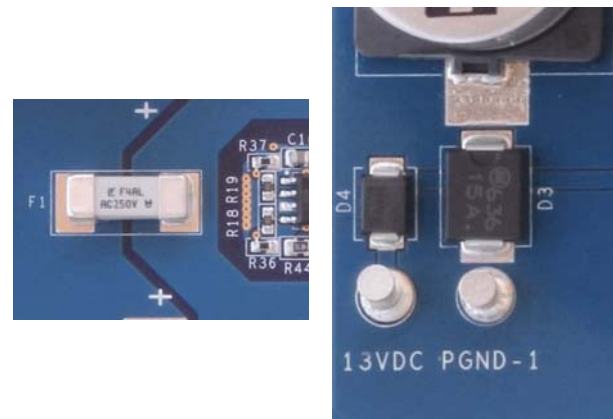


FIGURE 6. PROTECTION CIRCUITS

## Primary Side Current Sensing

The primary side bridge has two current sensing transformers, T<sub>2</sub> and T<sub>4</sub>, one on each leg on the drains of the low-side bridge FETs (Figure 7). Using two transformers allows each CT to reset during alternate half cycles. Alternate current sensing methods are reviewed in [“Current Sensing” on page 10](#).

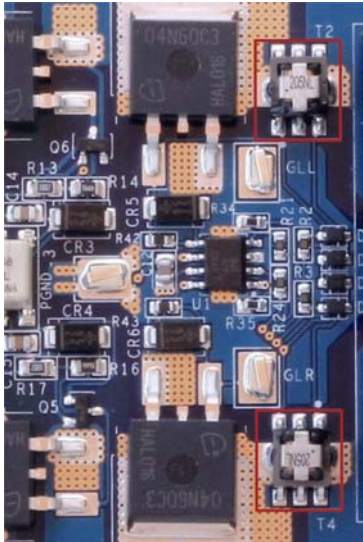


FIGURE 7. PRIMARY SIDE CURRENT SENSING

## Synchronous Rectifier Drive Circuit

Two banks of SRs are driven by the ISL89163 MOSFET driver, U4 (Figure 8). An RCD network on the inputs to this driver delays the turn-on of the SRs relative to the turn-off of the primary side bridge FETs.

The ISL89367, U108, can optionally be used to drive the SRs instead of the ISL89163. Review [“Schematics - Daughter Card” on page 27](#) to understand how to disconnect the ISL89163 and connect the ISL89367.

The pulse transformer, T<sub>6</sub>, crosses the isolation boundary to couple the control signals from the ISL6752/54 to the MOSFET drivers (Figure 9). Note that this transformer also provides the secondary side bias voltage for the MOSFET drivers.

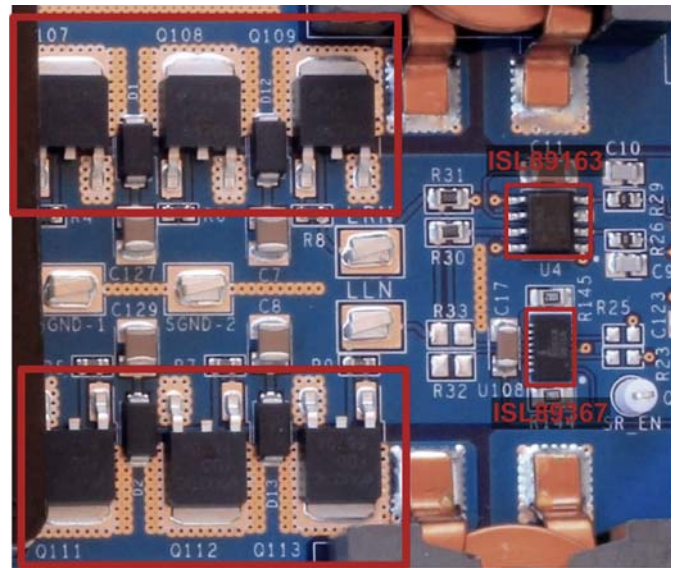


FIGURE 8. SRS AND DRIVERS



FIGURE 9. PULSE TRANSFORMER AND DRIVER

## Current Doubler Output

The current doubler output is composed of two banks of SRs, Q<sub>107</sub>...Q<sub>109</sub> and Q<sub>111</sub>...Q<sub>113</sub>; inductors L<sub>102</sub> and L<sub>103</sub>; and output filter capacitors, C<sub>133</sub>...C<sub>136</sub> as shown in [Figure 10 on page 7](#). One advantage of this topology is that the output current is shared by the two inductors, thus reducing conduction losses. Another advantage is that the secondary winding of the power transformer does not require a center tap.

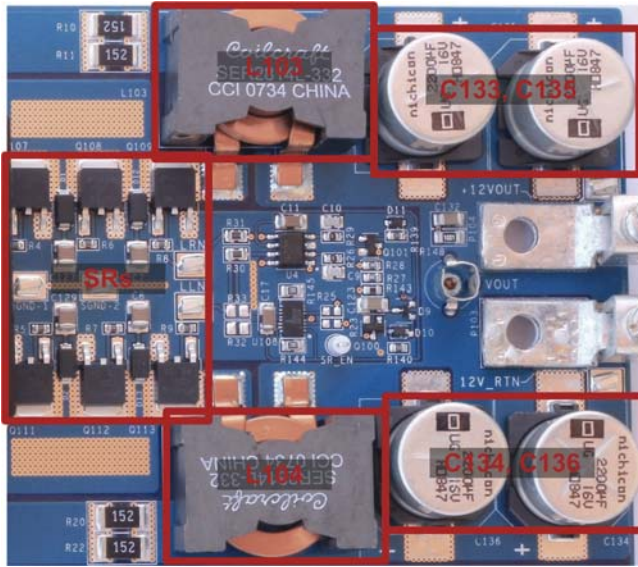


FIGURE 10. CURRENT DOUBLER OUTPUT

### Basic SR Principles

Replacing diodes with MOSFETs has two major advantages:

- Dramatically reduces conduction losses
- The applied duty cycle remains practically constant from no load to full load.

The disadvantages are:

- Additional complexity and cost
- Higher reverse recovery losses as compared to fast recovery diodes.
- When paralleling units for redundancy, provisions must be made to prevent current circulation among the paralleled units.

### SR Drive Timing Requirements

To emulate a diode, an SR must be driven ON when a diode would normally be conducting. Unlike a diode, if the SR is ON, the current through the SR can reverse if the voltage on the SR “cathode” becomes positive. The consequence is that if the SR is driven ON when the primary side is sourcing voltage to the secondary, the secondary side will be shorted by the SR.

Figure 11 illustrates the timing required to drive the SRs. Note that the rising edges of the two lower bridge FETs are delayed by the ISL6752/54 relative to the PWM signal. Likewise, the rising edges of the SRs gate signals are delayed by the ISL89163 relative to the falling edge of the PWM signal. These delays are necessary to prevent the overlap of drive signals that would result with high amplitude short circuit currents.

When an SR is turned off while current is flowing from source to drain, the current diverts from the FET channel to the internal body diode. Because the voltage drop across the body diode is higher than the channel, it is desirable to minimize dissipation by minimizing the duration of the current flow through the body diode.

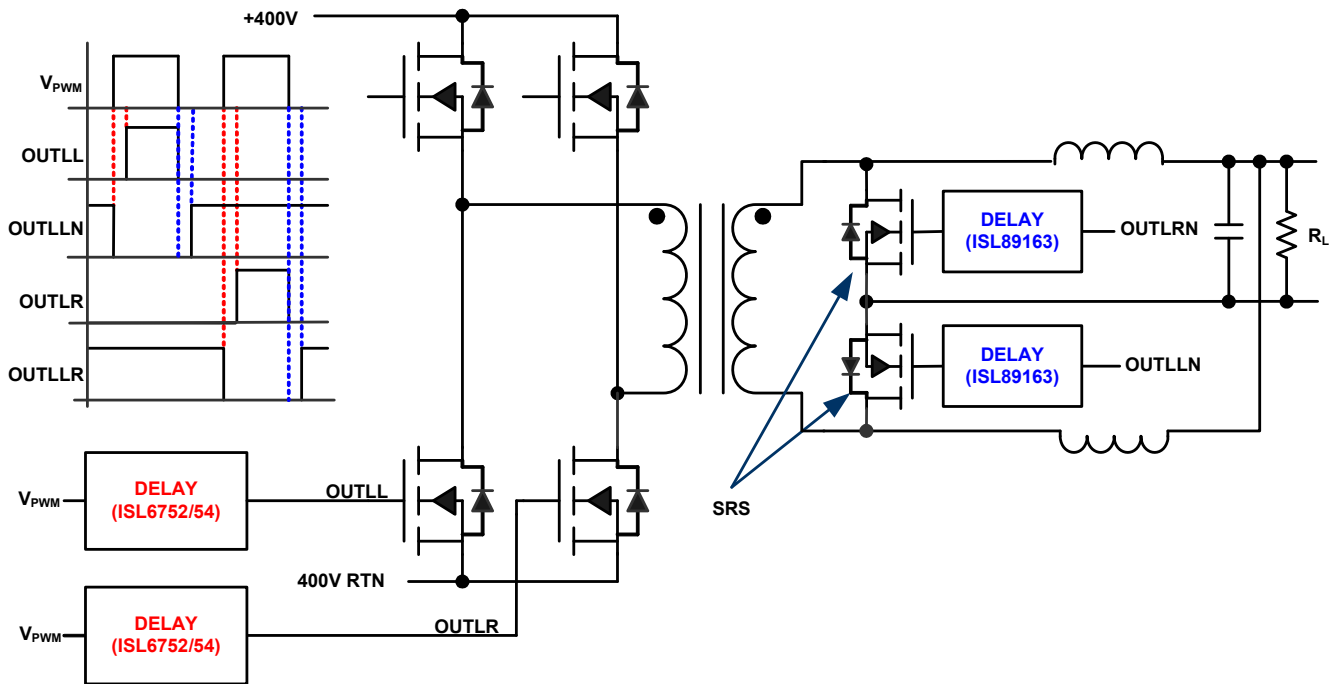


FIGURE 11. TIMING FOR SRS AND BRIDGE FETS

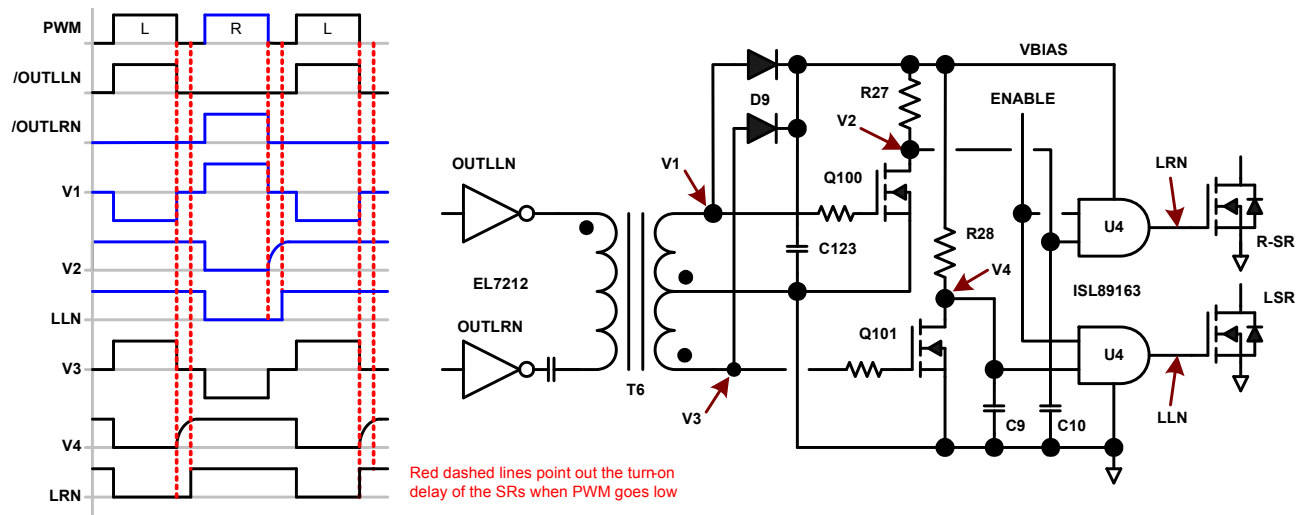


FIGURE 12. SIMPLIFIED SR DRIVE SCHEMATIC

## SR Drive and Bias

OUTLLN and OUTLRN in [Figure 12](#) are control signals from the ISL6752/54 that are used to drive the SRs. Because the ISL6752/54 is located on the primary side, a pulse transformer,  $T_6$ , is used to cross the isolation boundary. The simplified schematic of [Figure 12](#) illustrates the use of  $T_6$  to not only couple OUTLLN and OUTLRN to the secondary, but also to generate the bias for drivers on the secondary.

When /OUTLLN or /OUTLRN (outputs of EL7212) transitions to a logic high, it is necessary to turn off the associated SR quickly. For example, when /OUTLRN (blue) transitions high, V1 is high, and C10 is quickly discharged by Q100. U4 then drives R-SR off. In a similar manner, when /OUTLLN is high, U4 drives L-SR off.

When /OUTLLN or /OUTLRN transitions to a logic low, it is necessary to turn on the SRs after a time delay, to prevent the SRs from shorting the primary side bridge when it is sourcing current. For example, when /OUTLRN transitions to low, V1 is low and Q100 turns off, allowing C10 to be charged by R27. When the positive threshold of UR is exceeded, the output of U4 drives on R-SR. In a similar manner, the high to low transition of /OUTLLN results in the output of U4 driving on L-SR after a time delay.

Note that the cathodes of D9 are connected together to peak charge C123. Because C123 is large in value, after the initial charging, the voltage does not change significantly from cycle to cycle. An important aspect of generating the bias for U4 in this manner is that the thresholds for the logic transitions on the inputs of U4 are proportional to VBIAS, and the voltage to charge C9 and C10 is also VBIAS. Consequently, the delays generated by the RC networks are independent of the absolute value of VBIAS.

## Current Doubler

[Figure 13 on page 9](#) illustrates the current flow in the two inductors of the current doubler topology. Current flow in the circuit is correlated with the waveforms by color coding. The green waveform represents the sum of red and blue currents through  $R_{LOAD}$ . For circuit clarity, paralleled SRs and output capacitors of the ISL6752/54EVAL1Z board are not shown.

When using diodes (instead of SRs), if the average load current is less than half of the ramp current in the output inductors, the current in the inductors becomes discontinuous, and the duty cycle of the PWM is shortened to maintain the desired output voltage.

When using SRs, the inductor currents in  $L_1$  and  $L_2$  can become negative because current in SRs can flow bidirectionally; consequently, the duty cycle remains virtually unchanged. The benefit is that the load transient performance is the same for any load from zero up to current limit. Another advantage is that, for very light loads, the duty cycle is not reduced to very small duty cycles, pulse skipping does not occur, and the associated voltage jitter does not occur.

When designing the current doubler topology, the DC resistance of both halves must be equal. The PCB layout must be as symmetrical as possible, and the DCRs of the inductors should be reasonably equal. If not, the current between the two sides does not split equally. Because perfect physical PCB symmetry is not always possible, current sharing between inductors must be confirmed.

In [Figure 14 on page 9](#), inductor current waveforms are taken from the ISL6752/54EVAL1Z board. Current balance between the two inductors was achieved after one board revision. The inductor currents maintain the same waveform shape even at no load.



Another design consideration when using SRs is how to connect the outputs of multiple power supplies in parallel for redundancy or increased power capacity. A consequence of negative current flow in an SR (when a diode would otherwise be reverse biased and off) is that power can be transferred from the secondary to the primary if one of the paralleled outputs has a higher voltage. The voltage loop of the units with lower set point voltages attempts to pull down the voltage by sinking current from the

higher set point units. The primary side bridge capacitor is charged by the secondary side, eventually resulting in excessive voltage damage. This damage can be avoided by using OR-ing diodes (or FETs) on the paralleled outputs. Another solution is to turn off the SRs (Diode Emulation mode) when the current reverses in the SRs, but this eliminates some of the advantages of using SRs. Paralleling features are not implemented on the ISL6752/54EVAL1Z board.

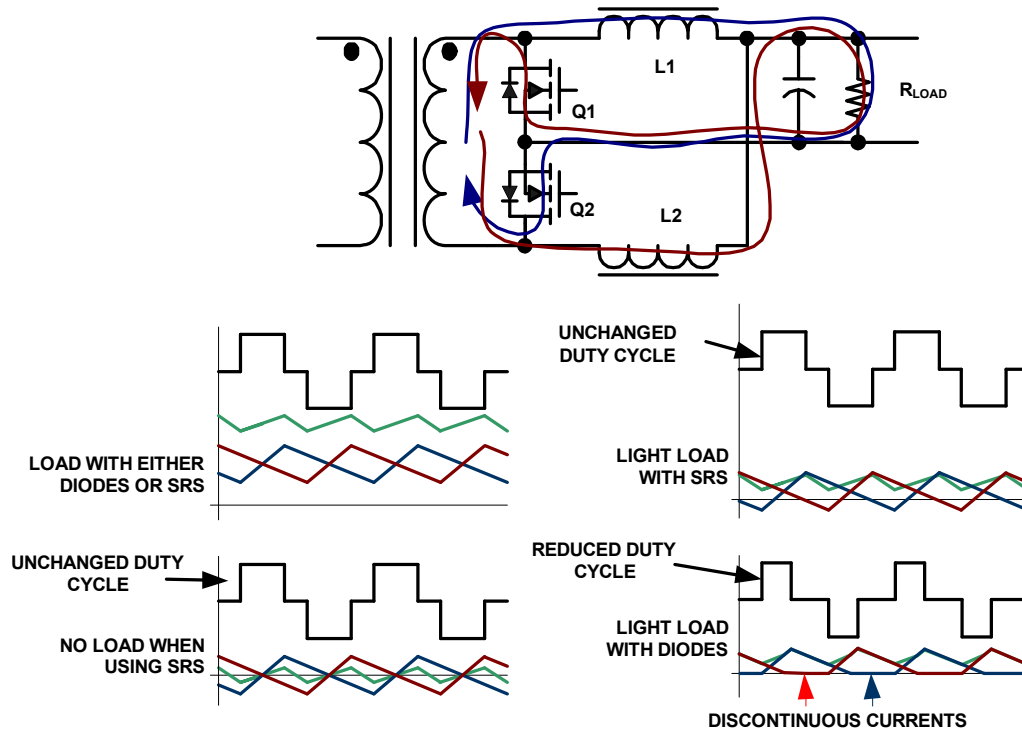


FIGURE 13. CURRENT FLOW IN TWO INDUCTORS OF CURRENT DOUBLER TOPOLOGY

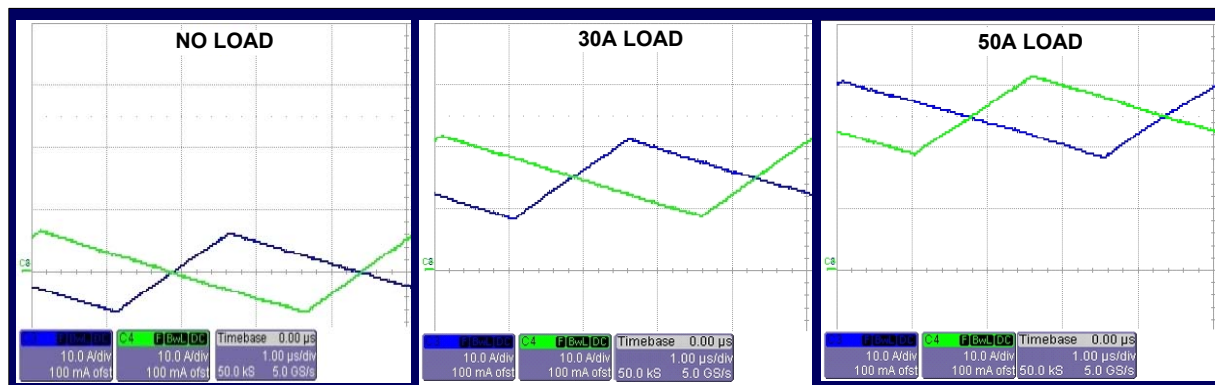


FIGURE 14. INDUCTOR CURRENT WAVEFORMS

## Current Sensing

Current flowing from the secondary to the primary can result in an unanticipated malfunction of the current sensing transformer circuit if reverse SR currents are not considered. Figure 15 shows a commonly used primary side current sensing circuit using one current sensing transformer (CT).

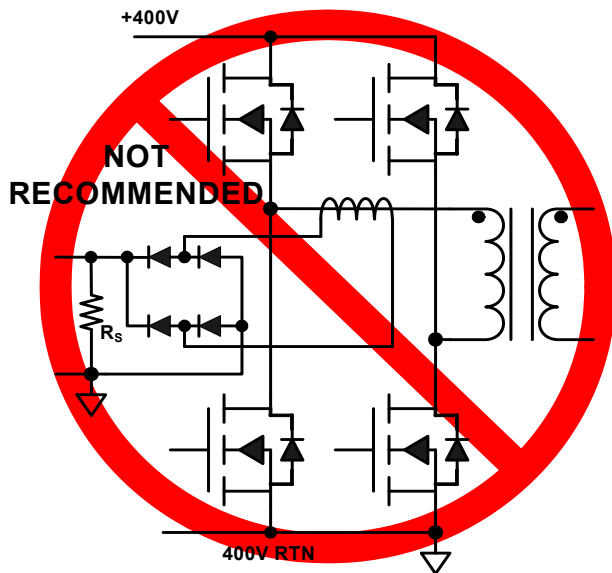


FIGURE 15. PRIMARY SIDE CURRENT SENSING CIRCUIT USING ONE CT

This circuit works well for peak current mode control if power is always flowing from primary to secondary, as is the case when diodes are used instead of SRs. Figure 16 illustrates the performance of the current sensing output when power always flows from primary to secondary.

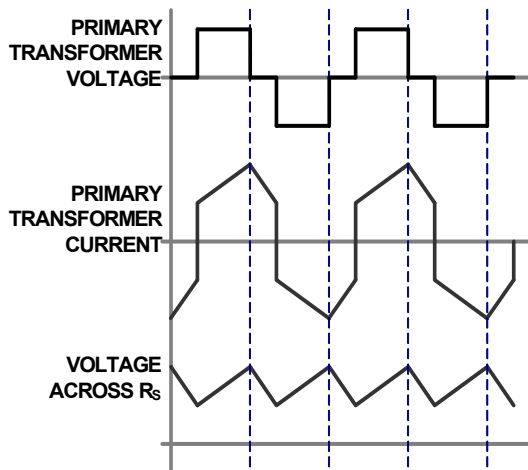


FIGURE 16. PERFORMANCE OF CURRENT SENSING OUTPUT

The voltage across  $R_s$  is as expected. The vertical dashed lines show when the power cycle is terminated at the required peak of the current.

Figure 17 illustrates what happens at no load to the sense voltage across  $R_s$ .

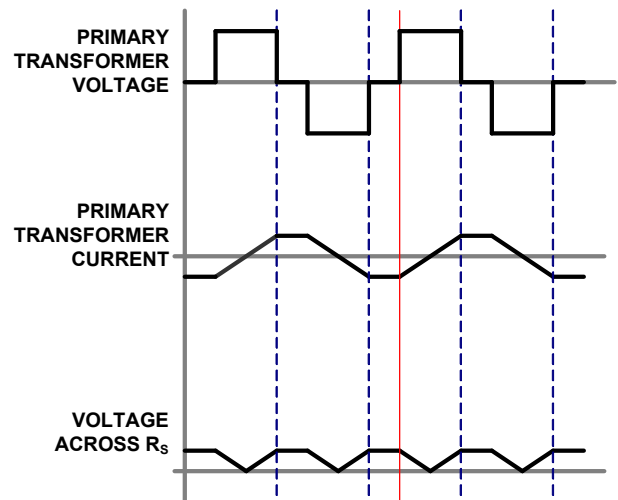


FIGURE 17. NO LOAD SENSE VOLTAGE ACROSS  $R_s$

Notice that the negative components of the primary transformer current are rectified, resulting in two peaks of current across  $R_s$  for each half cycle. Under steady state conditions, the rectified negative component may cause erratic performance because the cycle can terminate on the first peak (the inverted peak, as indicated by the vertical red line) instead of the required second peak. This condition can easily be corrected by having a small load across the output to ensure that the negative peak is always less than the positive.

A minimum load, however, does not correct a more serious problem that occurs when there is a large load step from a heavy load to no load. When the load current is interrupted, the output capacitor charges higher than the regulated voltage. As the regulation loop is starting to respond by slewing to a minimum duty cycle, the excessive voltage on the output capacitor starts to discharge back to the primary. This results in a large negative current at the beginning of the duty cycle, which causes the duty cycle to be terminated very early. The imbalance of the applied volt-seconds to the power transformer may saturate the power transformer and damage the power bridge.

Another scenario is that the current sensing transformer itself may saturate, which also damages the bridge. The control loop cannot maintain balanced alternate half cycles applied to the power transformer without valid current sense information.

There are three solutions to this problem.

1. Figure 18 illustrates the placements of two current sensing transformers, one on each drain leg of the bottom FETs.

In this configuration, only positive current flowing into the drains of the bottom FETs are sensed across  $R_s$ , solving the problem of rectified negative currents being impressed across  $R_s$ . An advantage of using two CTs is that there is a full half cycle available to reset the cores of the CTs. This is the solution used in the ISL6752/54EVAL1Z board.

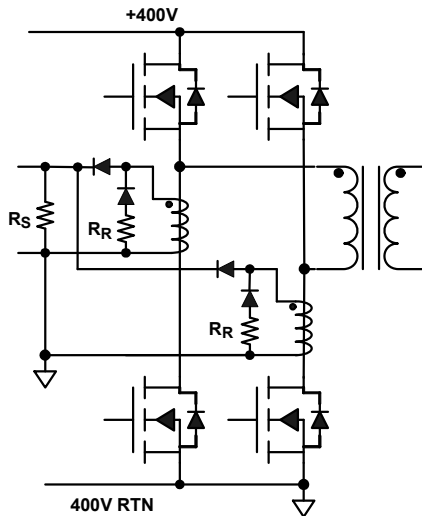


FIGURE 18. PLACEMENT OF TWO CURRENT SENSING TRANSFORMERS

2. Figure 19 shows a different current sensing implementation that also solves the problem shown in Figure 15. In this example, both drain currents of the bottom FETs are sensed by only one CT. There are some limitations that must be considered, however. The minimum time available to reset the core is the duration of the selected dead time between the two FETs on the same side of the bridge. To accommodate the resetting of the CT, this dead time can be made longer, but the consequences of reducing the maximum duty cycle available for output voltage regulation must be considered.

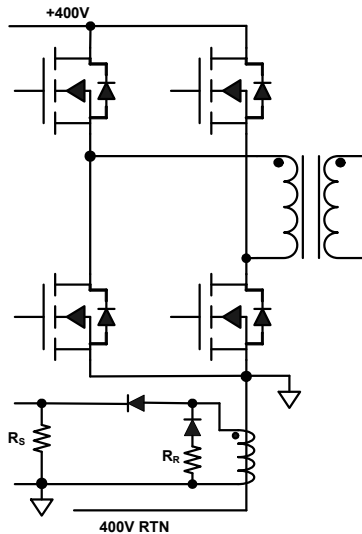


FIGURE 19. CURRENT SENSING TRANSFORMER IN THE COMMON SOURCE LEAD

If the dead time is kept short, then the peak voltage required for resetting the core is relatively large. For example, assume that the selected dead time is 2% of the duty cycle. The resulting worst-case reset voltage is shown approximately in Equation 1:

$$(0.98/0.02) \cdot V_{S\text{MAX}} = 49V \tag{EQ. 1}$$

In Equation 1,  $V_{S\text{MAX}}$  is 1V (the current limit voltage of the ISL6752); this is the ideal reset voltage. In practice, however, the parasitic capacitance of the output windings suppresses the peak voltage, and consequently, the reset time increases. If a custom current sensing transformer is designed, the effects of the parasitic capacitance can be minimized by increasing the space between turns. If a standard, off-the-shelf transformer is used, however, the output capacitance may be too large to allow long duty cycles. In this case, the two-transformer solution may be necessary.

Notice in Figure 19 that the 400V RTN is slightly more negative than signal ground. This configuration is recommended for applications that directly drive the bottom FETs with MOSFET drivers. If the 400V RTN and the MOSFET drivers are grounded, regenerative feedback will be present on the output of the MOSFET drivers because of the CT windings in the gate drive loop.

3. A variation on the current sense circuit in Figure 19 is to place the current sensing transformer in the common drain lead of the two high-side FETs, as shown in Figure 20.

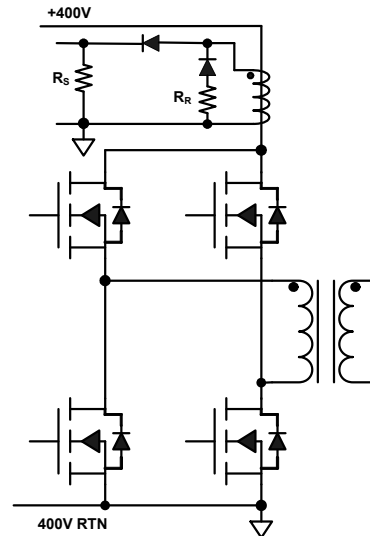


FIGURE 20. CURRENT SENSING TRANSFORMER IN THE COMMON DRAIN LEAD

The circuits shown in Figures 19 and 20 give exactly the same performance, but the problem associated with the gate drives (as explained in Figure 19) is avoided. The disadvantage of placing the CT at this location is that the CT must be designed with 400VDC operational insulation.

## Conclusion

This application note reviews the use of MOSFETs as synchronous rectifiers to replace conventional diodes. The advantages of improved power efficiency and load transient are reviewed along with implementation problems that must be solved.

The use of daughter cards for the ISL6752 and ISL6754 control ICs also allows comparison of cycle-by-cycle peak current limiting and average current limiting.

## References

- [1] Fred Greenfeld, Intersil Application Note [AN1246](#),  
“Techniques to Improve ZVS Full-bridge Performance”
- [2] Fred Greenfeld, Intersil Application Note [AN1262](#),  
“Designing with the ISL6752, ISL6753 ZVS Full-bridge  
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- [3] Richard Garcia, Intersil Application Note [AN1619](#),  
“Designing with ISL6752DBEVAL1Z and  
ISL6754DBEVAL1Z Control Cards”

## Evaluation Board Setup

The following sections cover the setup of the ISL6752/54EVAL1Z evaluation board. Also included are waveforms, performance parameters, PCB layout, and schematics.

## Setting Up

### Danger

- This evaluation unit should be used and operated only by persons experienced and knowledgeable in the design and operation of high voltage power conversion equipment.
- Use of this evaluation unit constitutes acceptance of all risk inherent in the operation of equipment having accessible hazardous voltage. Careless operation may result in serious injury or death.
- Use safety glasses or other suitable eye protection.
- A line isolated 400VDC supply is required.

### Lab Equipment Required

- DC bias power supply, 12.6VDC at 200mA minimum
- Adjustable 0VDC-400VDC regulated lab power supply, 2.5ADC minimum with current limit
- Fan to cool heatsinks
- Oscilloscope, digital preferred, with four channels, 20MHz minimum bandwidth
- Adjustable DC load (electronic or resistor), 70A at 12V, 100A at 0V min, >850W
- DC multimeter
- Infrared temperature probe (optional but highly recommended)

### Turn-On Procedure

1. Solder a wire between the DISABLE and PGND-1 lugs located on the lower left side of the main board. Optionally connect a switch between these two lugs.
2. Install either of the daughter control cards onto the main board.
3. Connect the DC load to the outputs of the evaluation board. Adjust the load to zero current.
4. With both supplies turned off, connect the DC bias supply to the +13V terminal and PGND.
5. Connect the 400V supply to +400V and 400V RTN.
6. Turn on the DC bias supply and adjust the current limit to 200mA. Adjust the voltage to +12.6 VDC. The lab supply current should be approximately 150mA.

### Caution

A voltage clamp, D3, is used to protect the primary side control circuit from catastrophic damage if the high voltage bridge fails. To prevent this clamp from conducting, do not adjust the bias supply above 13.5VDC.

7. Turn on the 400V supply and adjust the current limit to 2.5A. Adjust the voltage to 400VDC. **Do not exceed 450VDC.** The current should be approximately 45mA.
8. Turn on the fan and direct the air flow through the heatsinks mounted on the bottom of the board.
9. Using the test points that are adjacent to the output power lugs, measure the output voltage of 12V ±0.5VDC.

The output load and input voltage can now be safely adjusted.

Because there is no thermal shut-down circuit, it is important to maintain adequate airflow over the heatsinks, especially when applying large loads. It is recommended to measure the temperature of the power FETs (primary bridge and secondary SRs) to ensure that their temperatures do not exceed +85°C. It is usually necessary to have only a moderate airflow over the heatsinks, even under worst-case loads.

# Waveforms

## ZVS

In Figure 21, the drain-source voltage of the low-side FETs relative to the gate voltage is displayed to highlight the ZVS performance of the bridge. The load is at the rated 50A. Notice that full ZVS is not achieved because the minimum resonance voltage is about 25VDC. Also, the gate drive is turning on late (about 25ns), allowing the resonant voltage to start rising. Even

though the optimum zero voltage switching is not achieved, 98% of the switching losses are still recovered  $[(400^2 - 50^2) / 400^2 = 98\%]$ . This improvement over the ISL6752EVALZ was achieved by increasing the leakage inductance of the transformer and by using bridge FETs with less body capacitance.

In Figure 22, resonant switching with 50% load still saves 84% of the switching losses. Other techniques can be used to improve ZVS performance. For more information, see Application Note [AN1246](#), "Techniques to Improve ZVS Full-bridge Performance".

## ZVS Waveforms



FIGURE 21. RESONANT SWITCHING WITH 100% (50A) LOAD

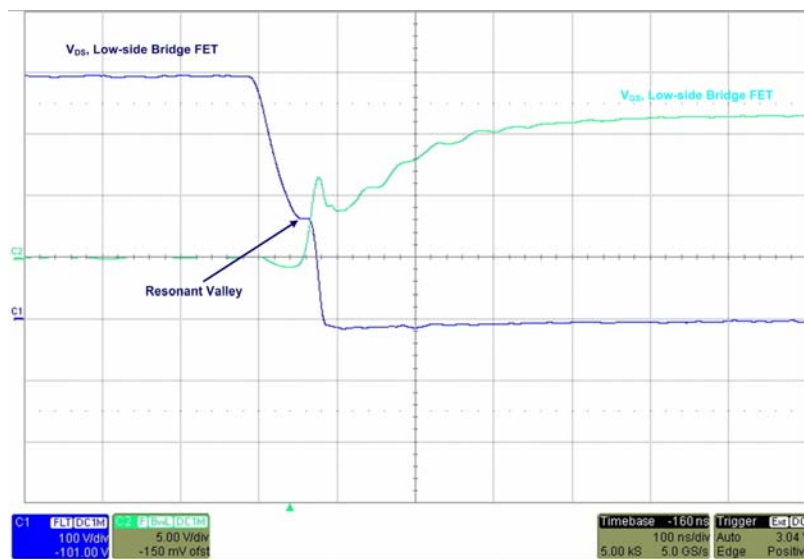


FIGURE 22. RESONANT SWITCHING WITH 50% (25A) LOAD

Load Transients Waveforms



FIGURE 23. STEP LOAD: 0A TO 12.5A (12.5A DELTA)



FIGURE 24. STEP LOAD: 25A TO 37.5A (12.5A DELTA)

Load Transients Waveforms (Continued)

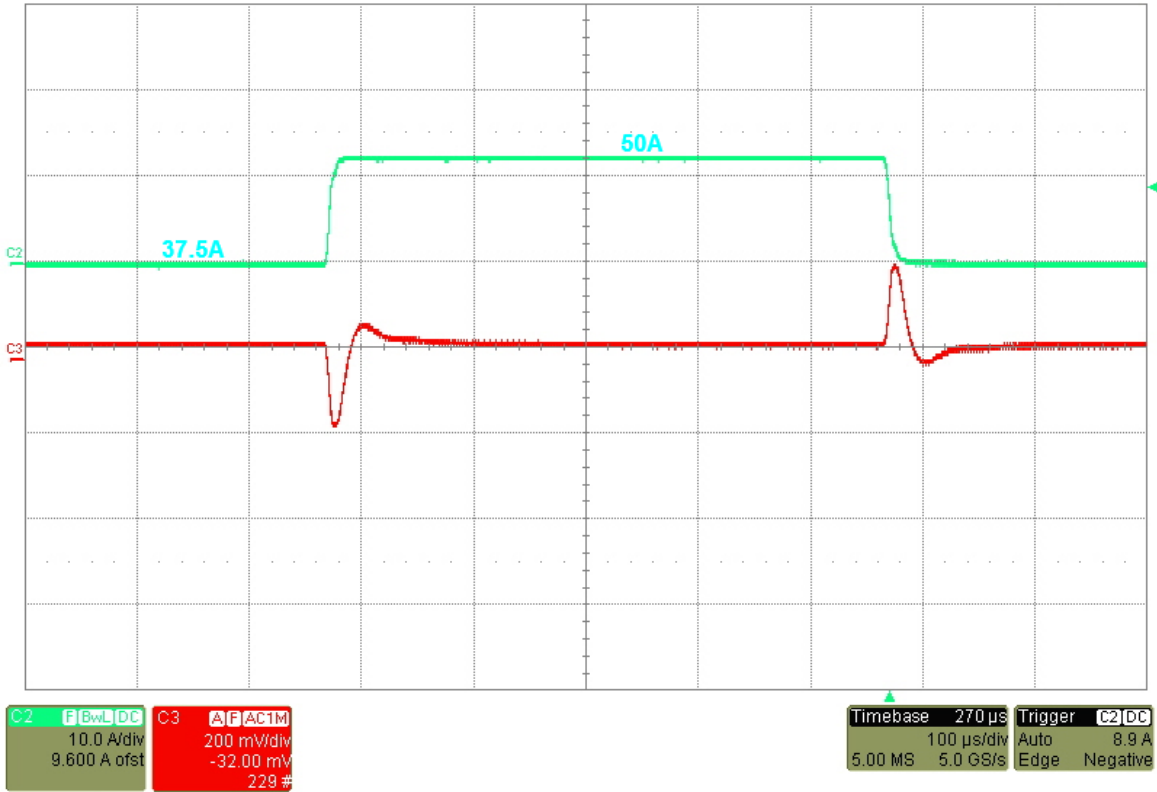


FIGURE 25. STEP LOAD: 37.5A TO 50A

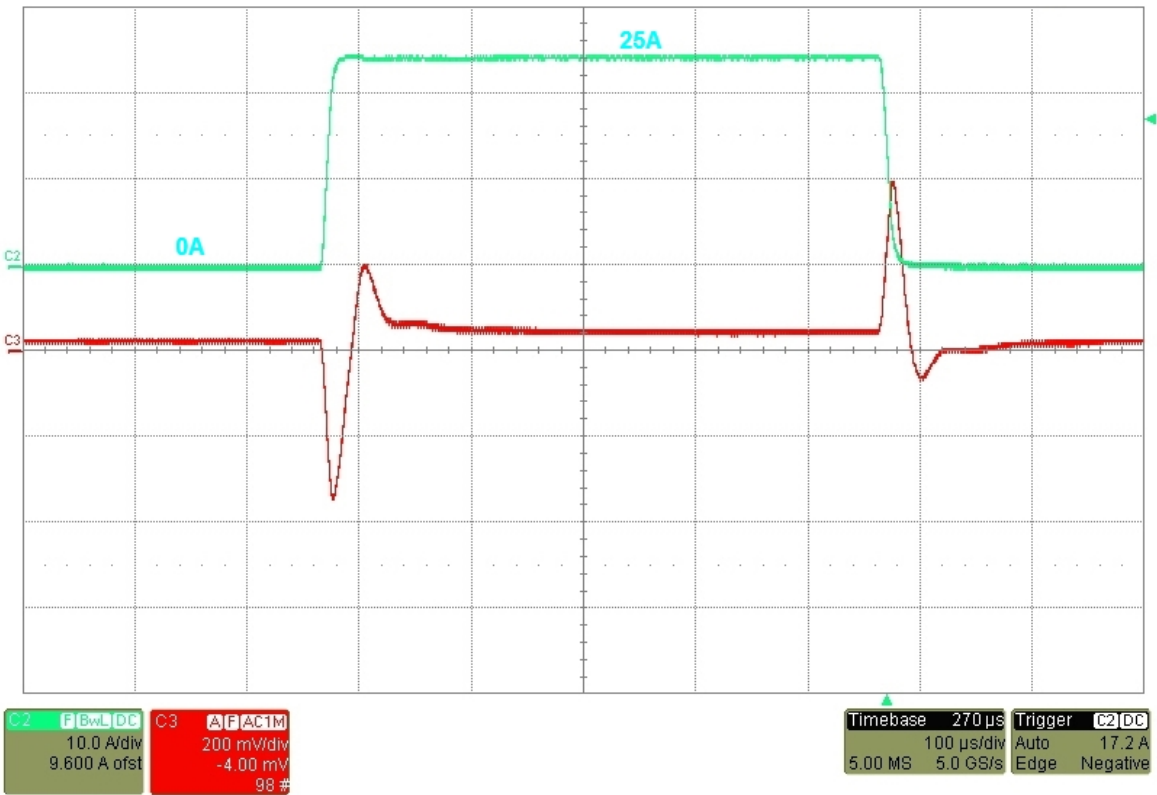


FIGURE 26. STEP LOAD: 0A TO 25A

Load Transients Waveforms (Continued)

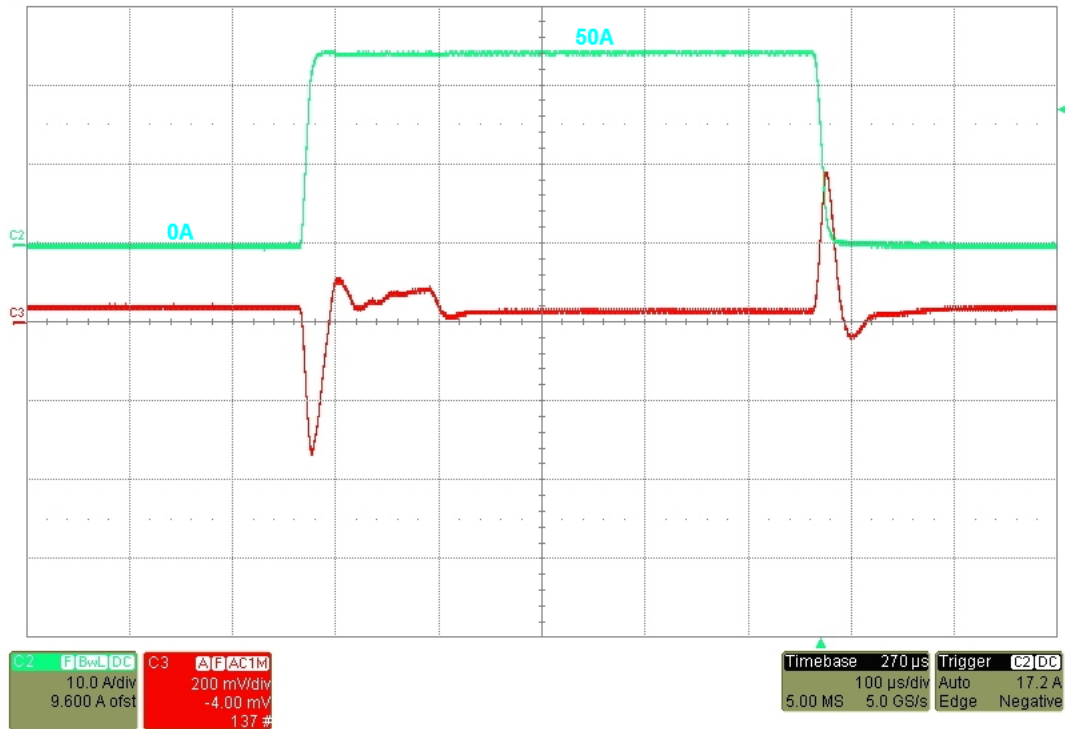
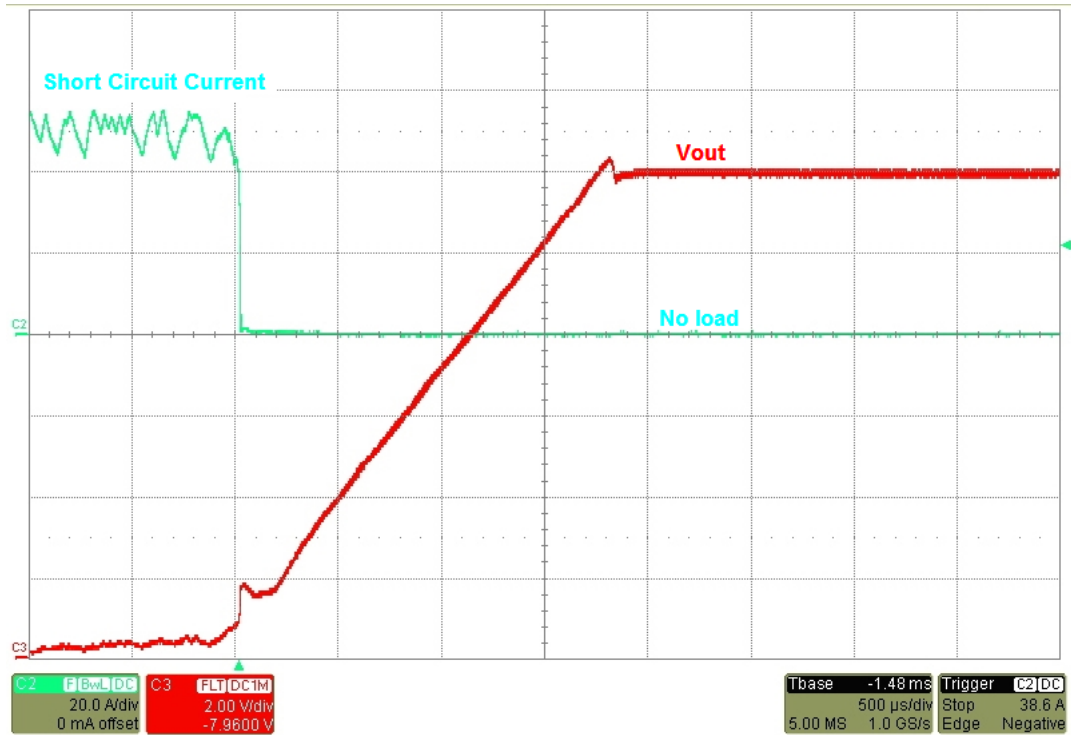


FIGURE 27. STEP LOAD: 0A TO 50A



NOTE:

$V_{OUT}$  recovers after a short circuit is removed when using the ISL6754DBEVAL controller.

After the short is removed,  $V_{OUT}$  increases linearly because the output capacitance is being charged with a constant current (~55A).

FIGURE 28. SHORT CIRCUIT RELEASE WITH ISL6754



Output Ripple and Noise Waveforms

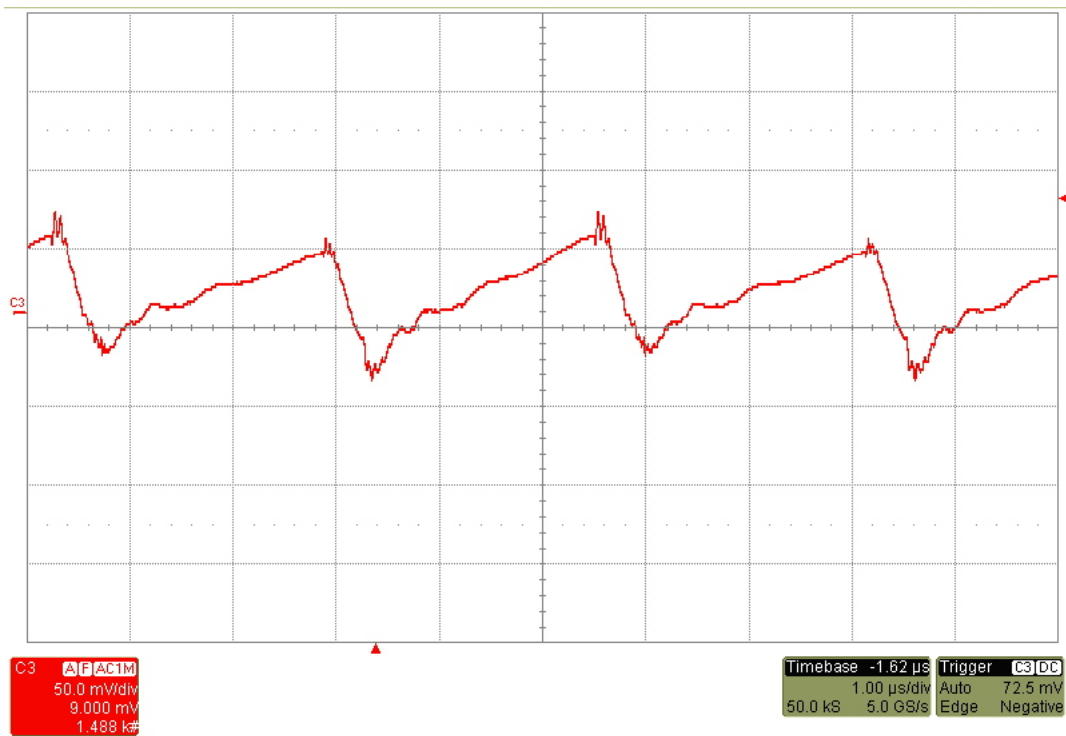


FIGURE 29. OUTPUT RIPPLE, 50A LOAD, 40MHZ BANDWIDTH

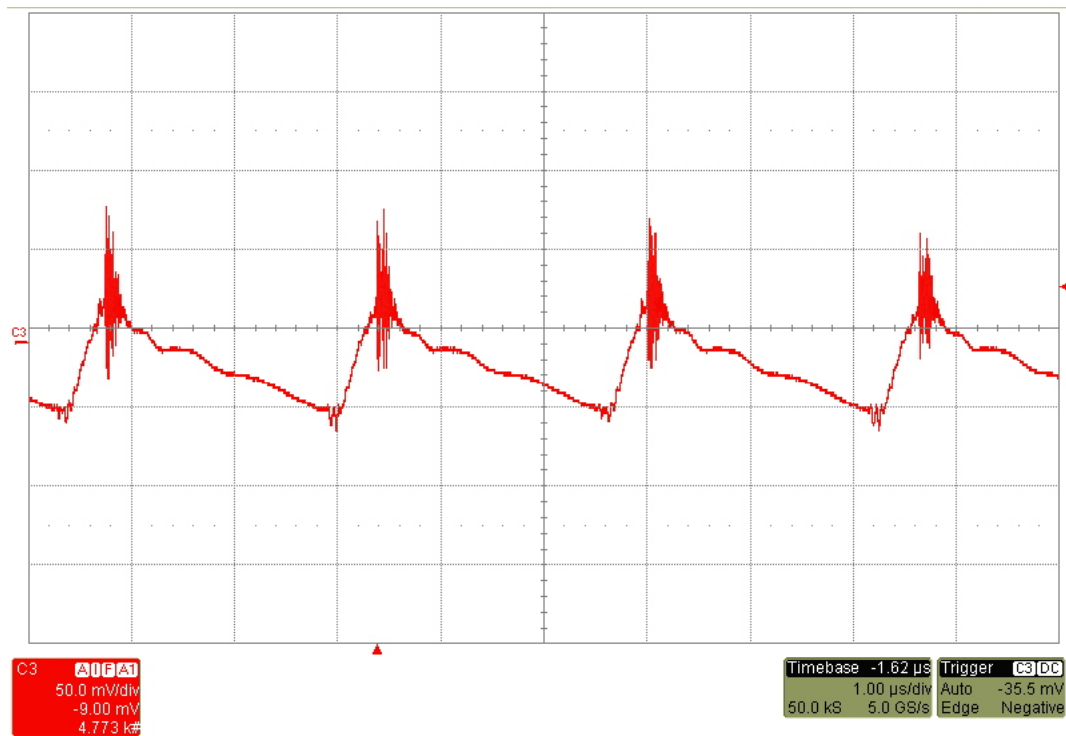


FIGURE 30. OUTPUT RIPPLE, 50A LOAD, 145MHZ BANDWIDTH

Transformer Current, Primary Winding Waveforms

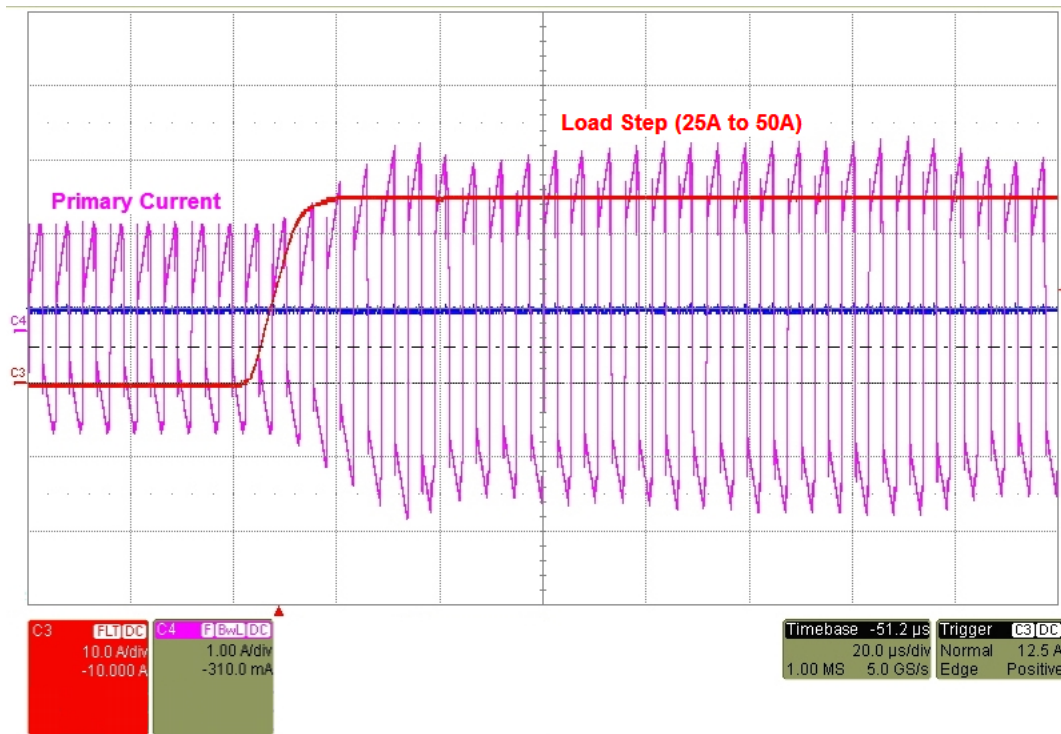


FIGURE 31. PRIMARY TRANSFORMER CURRENT VS OUTPUT LOAD TRANSIENT (25A TO 50A)

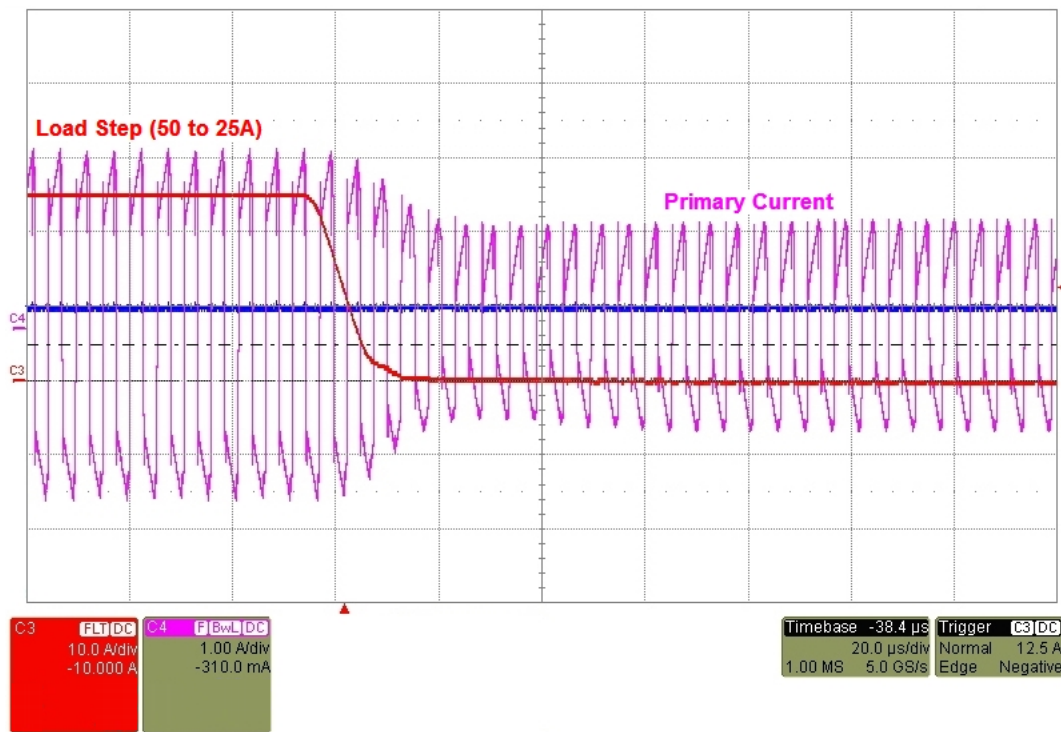


FIGURE 32. PRIMARY TRANSFORMER CURRENT VS OUTPUT LOAD TRANSIENT (50A TO 25A)

## Performance Curves

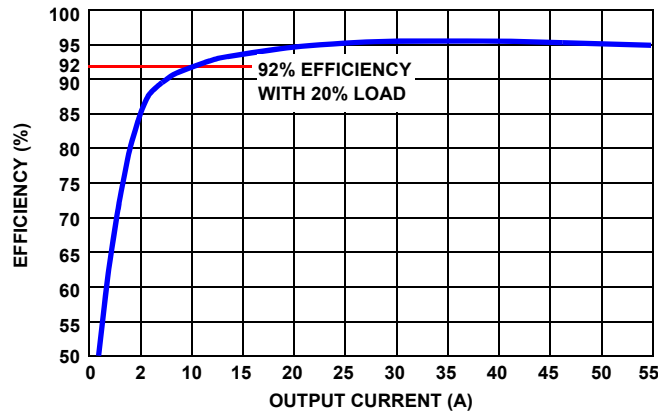


FIGURE 33. POWER EFFICIENCY VS LOAD (ISL6752 OR ISL6754)

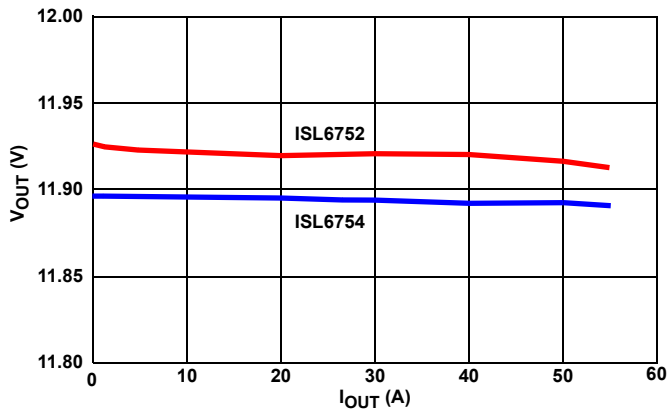


FIGURE 34. LOAD REGULATION

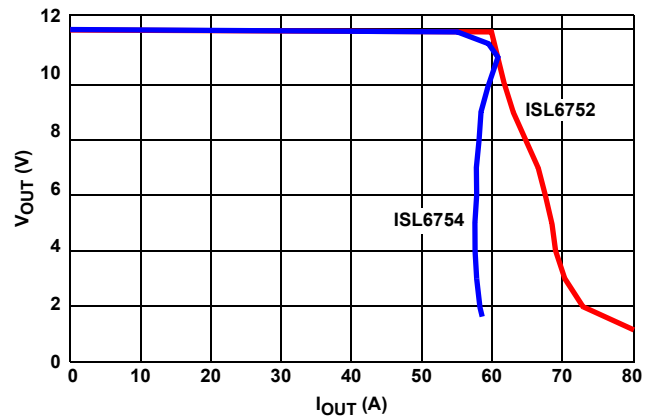


FIGURE 35. PULSE BY PULSE VS AVERAGE CURRENT LIMIT

# PCB Layout

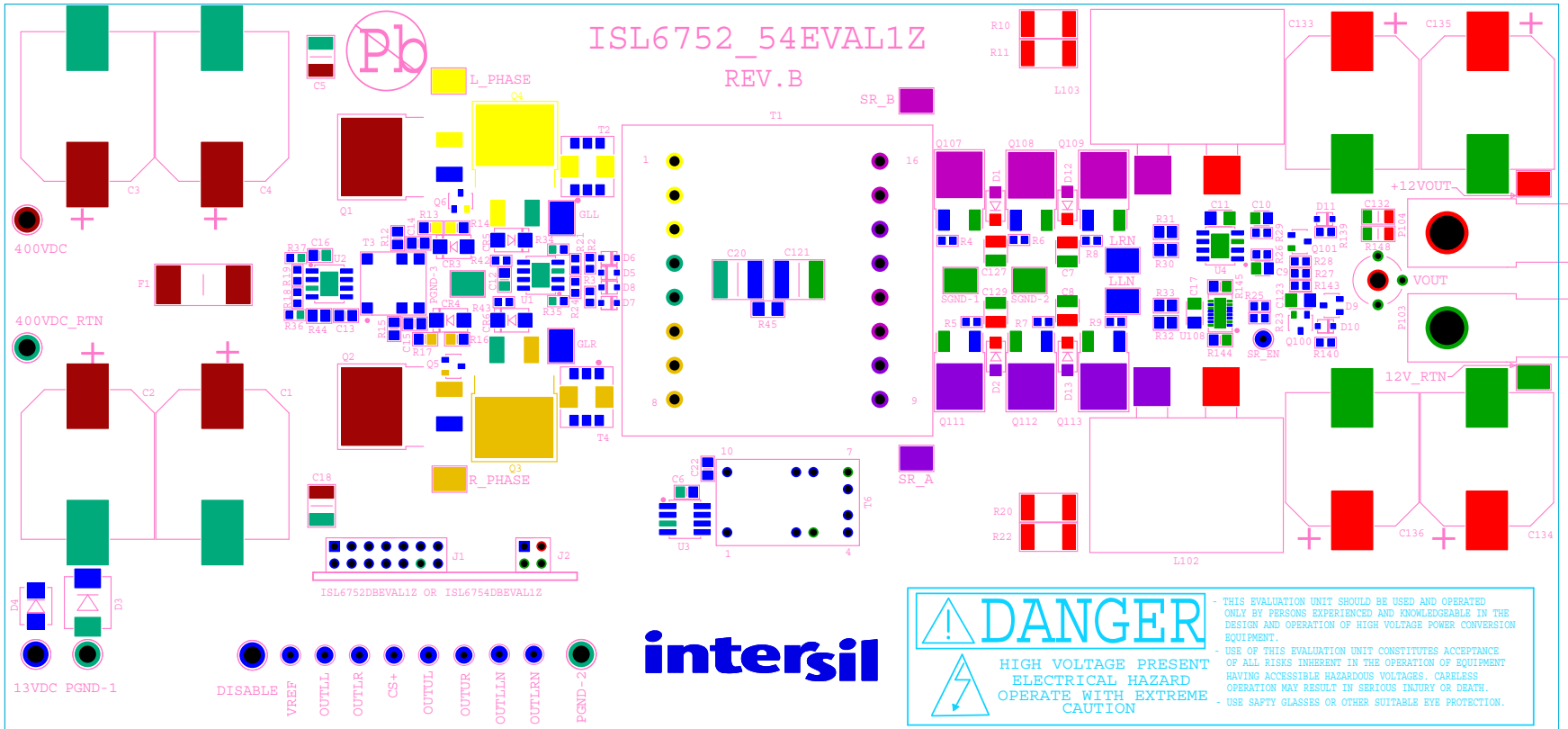
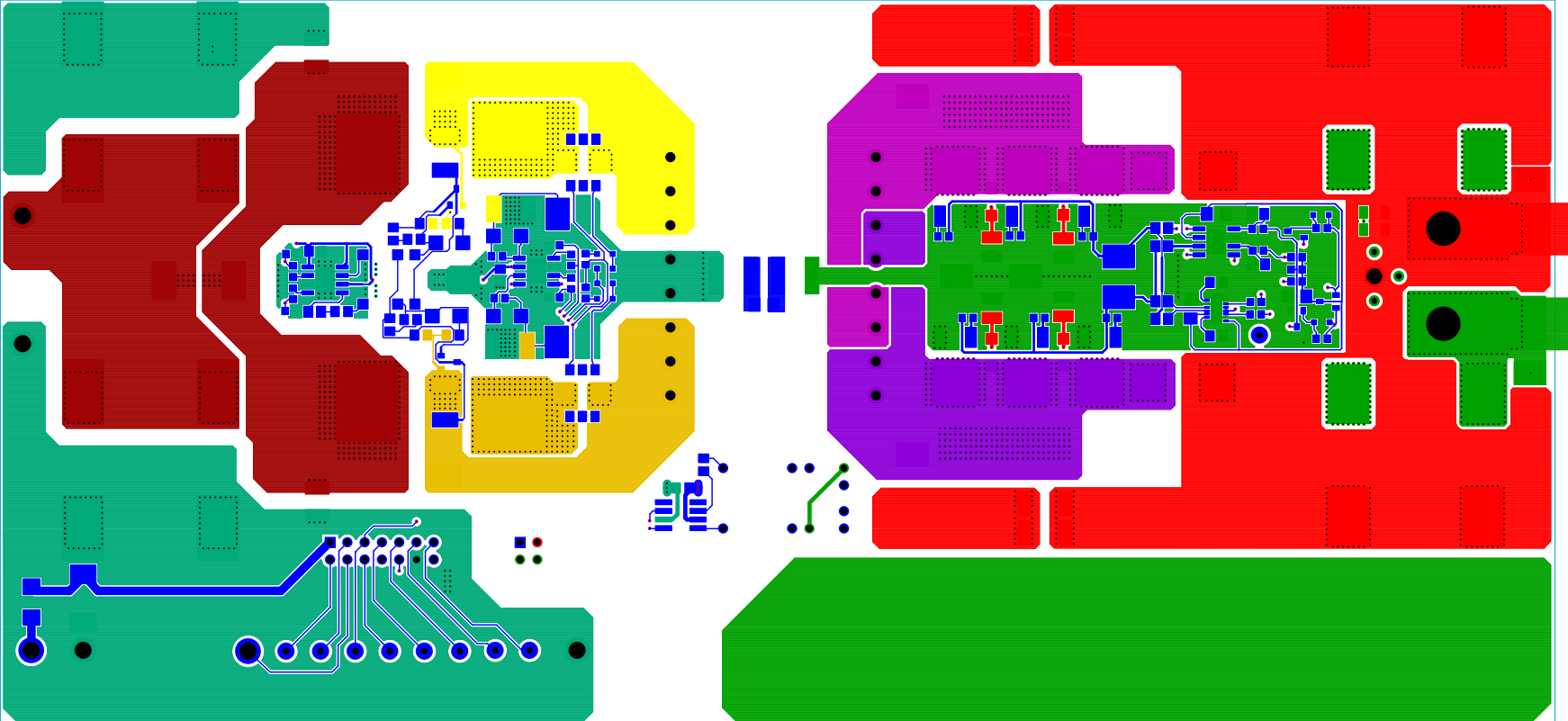


FIGURE 36. SILKSCREEN TOP

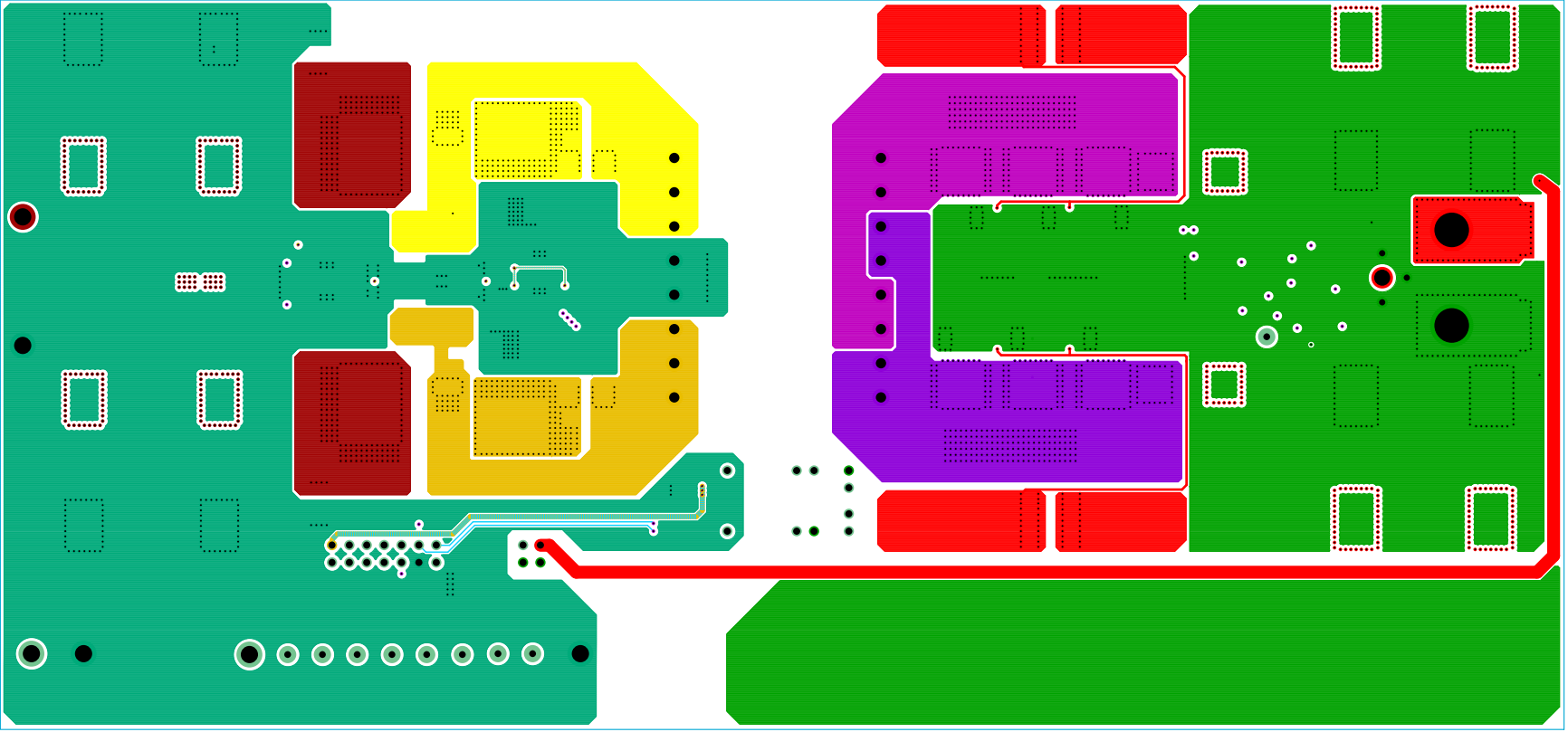
# PCB Layout (Continued)



LAYER 1

FIGURE 37. LAYER 1

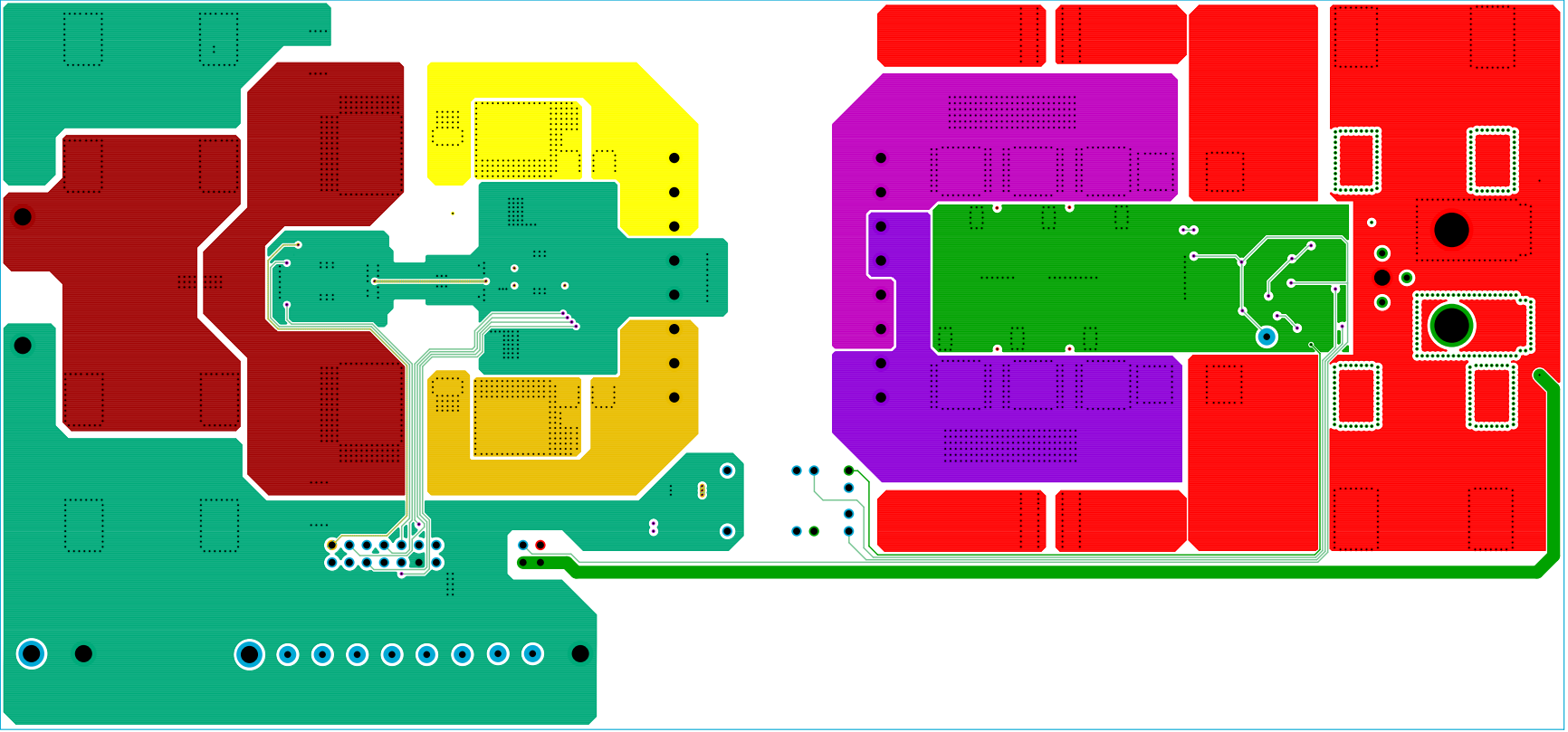
# PCB Layout (Continued)



LAYER 2

FIGURE 38. LAYER 2

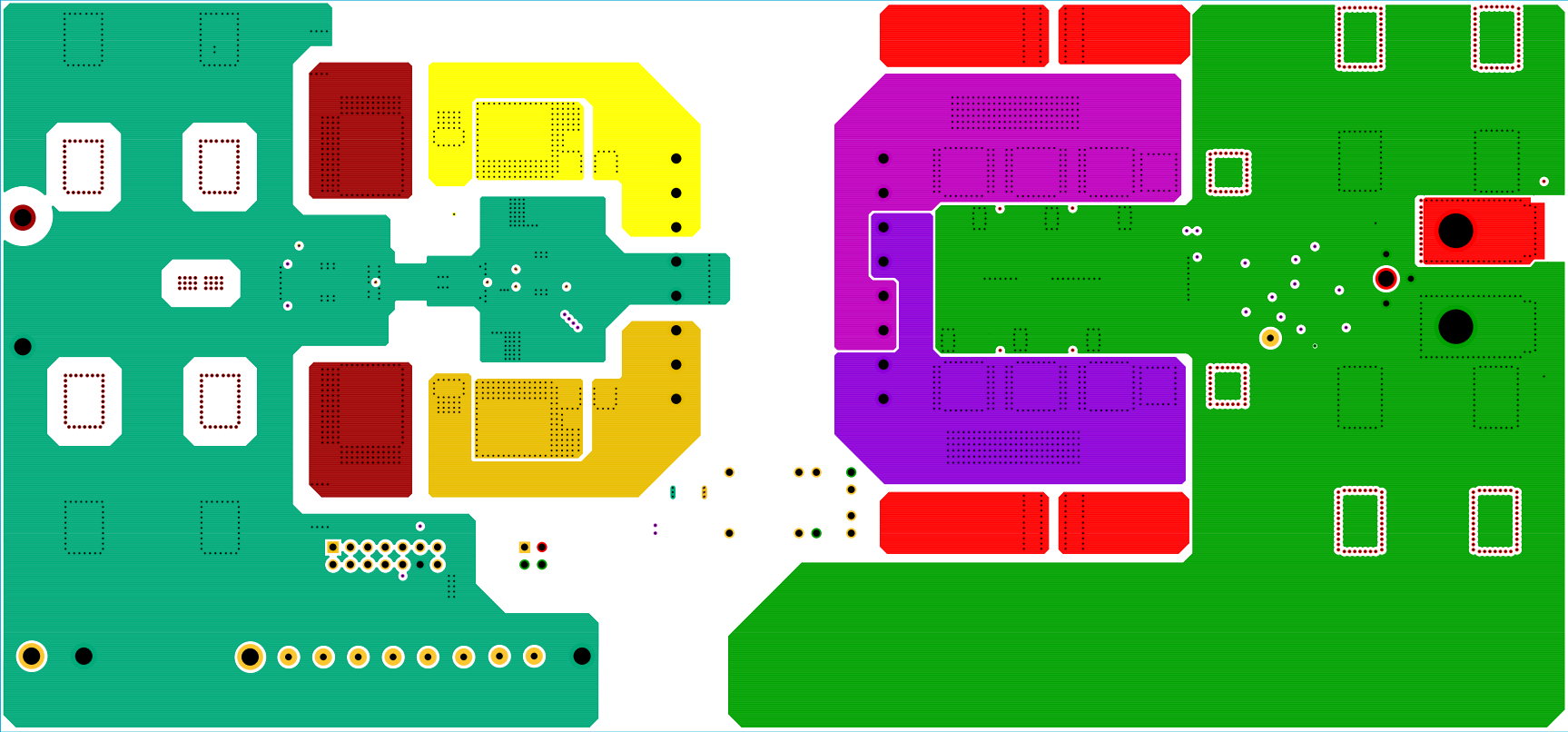
# PCB Layout (Continued)



LAYER 3

FIGURE 39. LAYER 3

# PCB Layout (Continued)



LAYER 4

FIGURE 40. LAYER 4



# Schematics- Main Board, ISL6752/54EVAL1Z

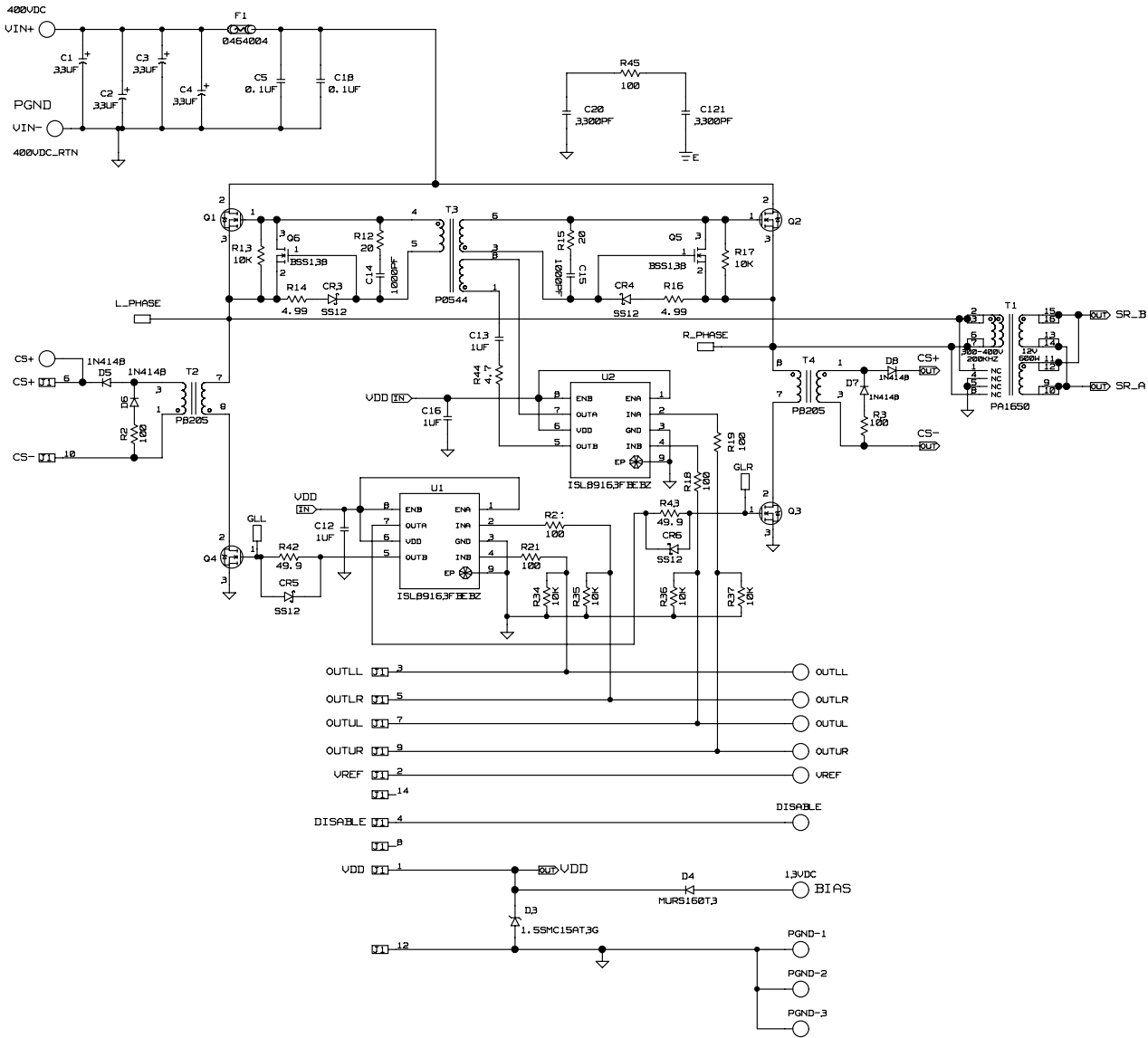


FIGURE 41. MAIN BOARD, PRIMARY, ISL6752\_54EVAL1Z

**Schematics- Main Board, ISL6752/54EVAL1Z (Continued)**

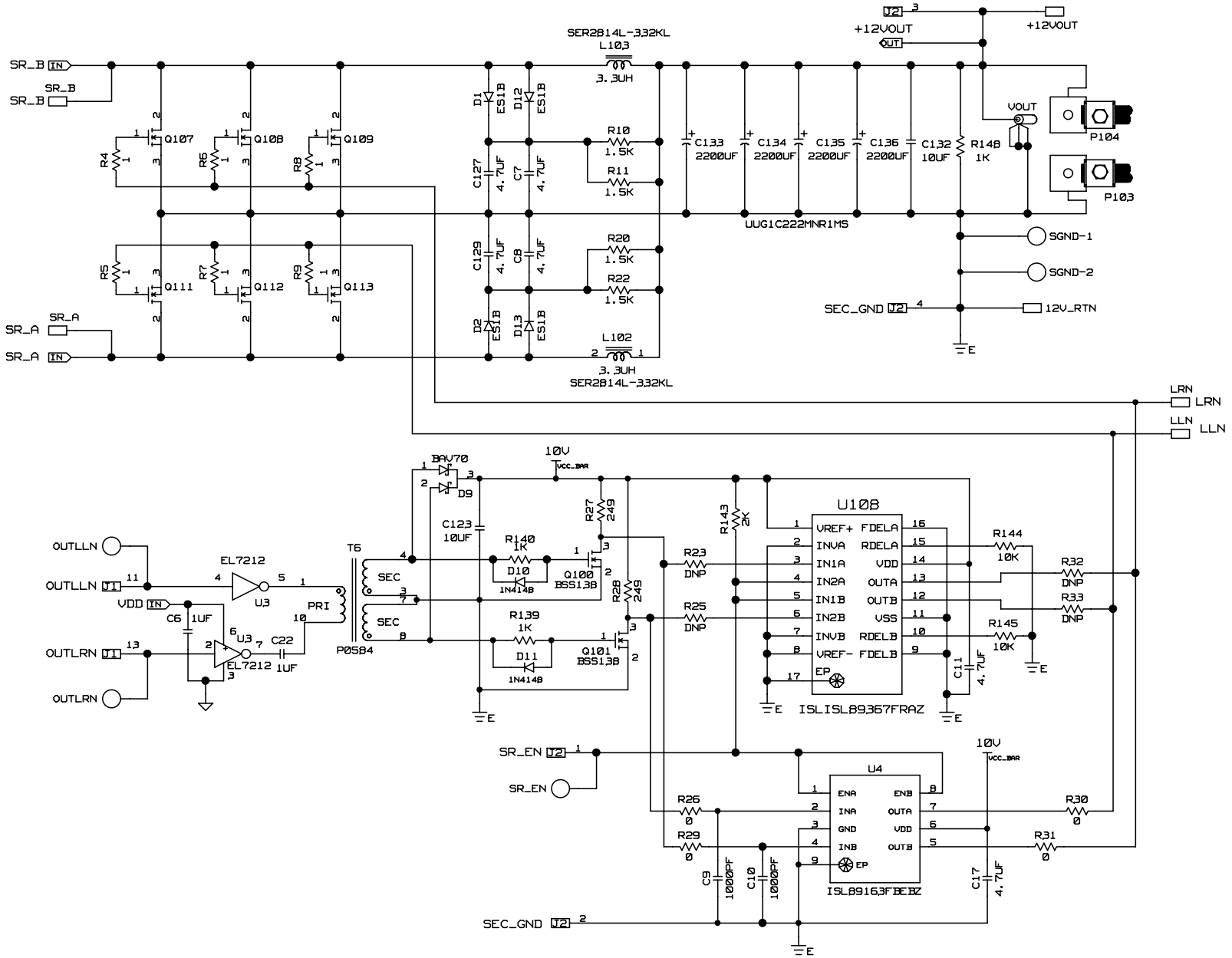


FIGURE 42. MAIN BOARD, SECONDARY, ISL6752\_54EVAL1Z

# Schematics - Daughter Card

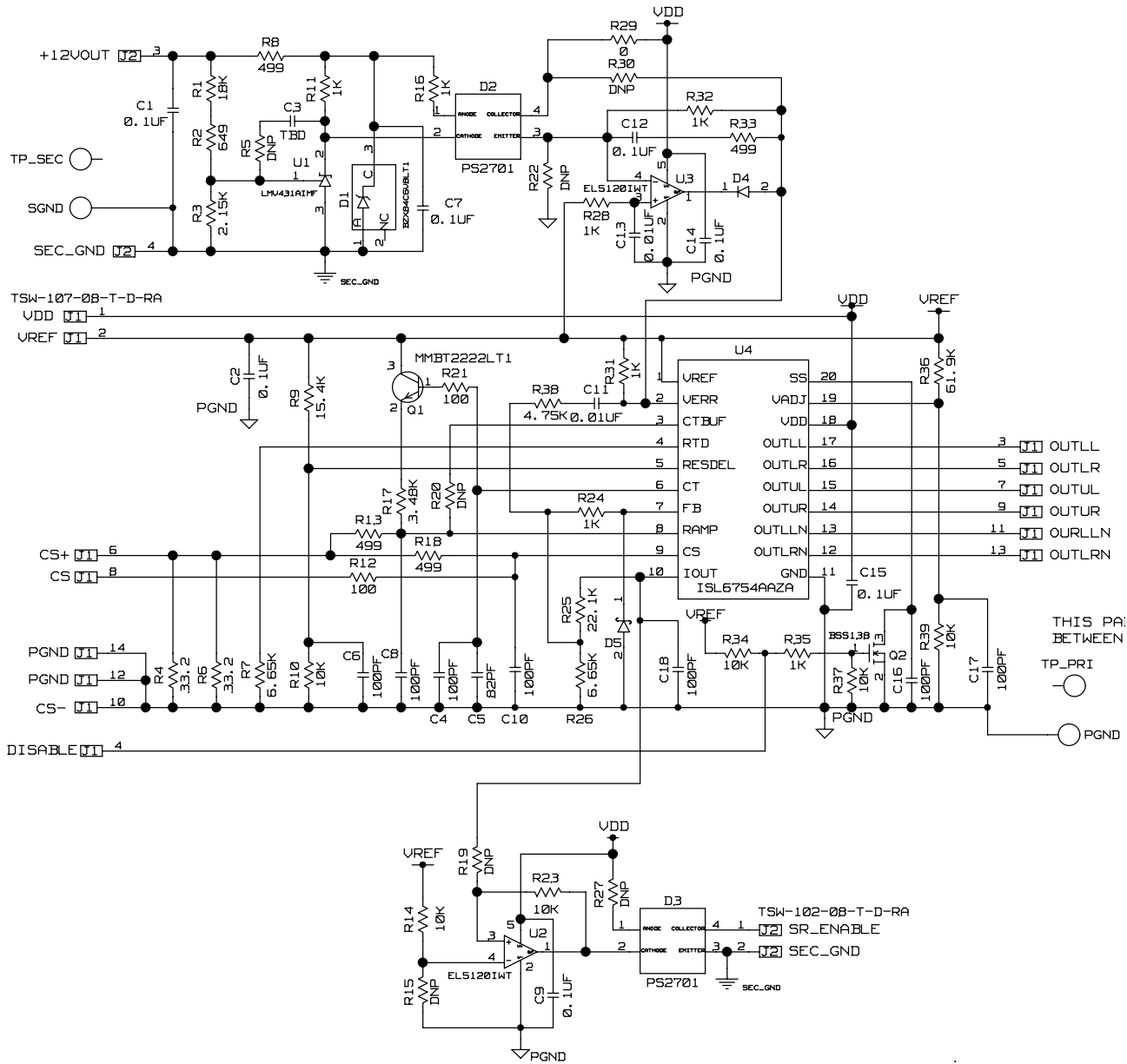


FIGURE 43. DAUGHTER CARD, ISL6754BEVAL

Schematics - Daughter Card (Continued)

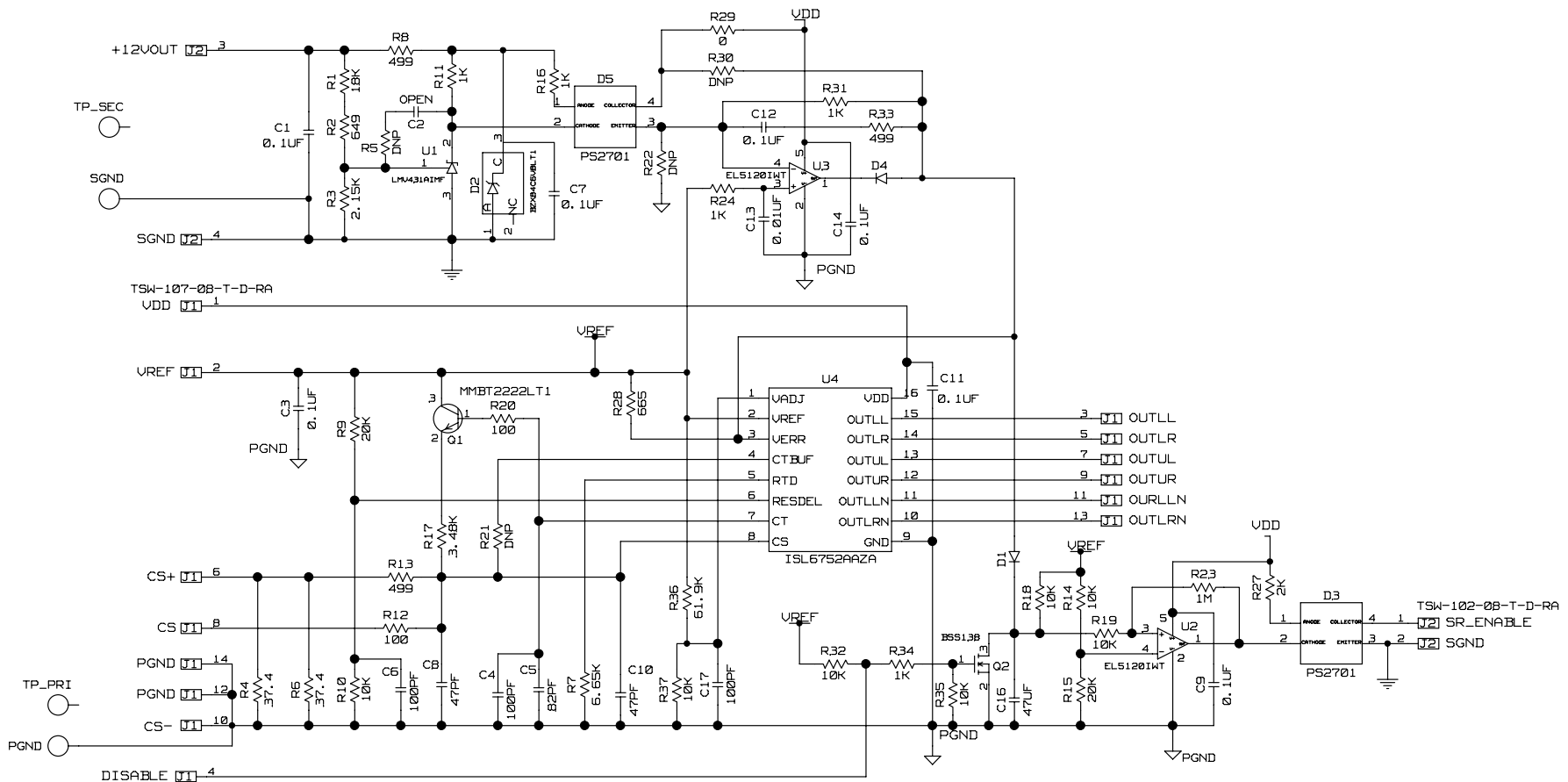


FIGURE 44. DAUGHTER CARD, ISL6752DBEVAL1Z

**Bill of Materials**

PART NUMBER	QTY	UNITS	REFERENCE DESIGNATOR	DESCRIPTION	MANUFACTURER	MANUFACTURER PART
ISL6752/54EVAL1ZREVBPCB	1	ea		PWB-PCB, ISL6752_54EVAL1Z, REV B, ROHS	IMAGINEERING INC	ISL6752/54EVAL1ZREVBPC B
C3216X7R1C475K-T	2	ea	C11, C17	CAPACITOR, SMD, 1206, 4.7µF, 16V, 10%, X7R, ROHS	TDK	C3216X7R1C475K
C4532X7R2J104K-T	2	ea	C5, C18	CAP, SMD, 1812, 0.1µF, 630V, 10%, X7R, ROHS	TDK	C4532X7R2J104K
GA355QR7GF332KW01L-T	2	ea	C20, C121	CAP, SMD, 2220, 3300pF, 250V, 10%, X7R, ROHS	MURATA	GA355QR7GF332KW01L
H1046-00102-100V10-T	2	ea	C14, C15	CAP, SMD, 0805, 1000pF, 100V, 10%, X7R, ROHS	VENKEL	C0805X7R101-102KNE
H1046-00102-50V5-T	2	ea	C9, C10	CAP, SMD, 0805, 1000pF, 50V, 5%, NPO, ROHS	PANASONIC	ECU-V1H102JCX
H1046-00105-25V10-T	5	ea	C6, C12, C13, C16, C22	CAP, SMD, 0805, 1.0µF, 25V, 10%, X5R, ROHS	AVX	08053C105KAT2A
H1065-00106-25V10-T	1	ea	C123, C132	CAP, SMD, 1206, 10µF, 25V, 10%, X5R, ROHS	VENKEL	C1206X5R250-106KNE
H1082-00475-50V10-T	4	ea	C7, C8, C127, C129	CAP, SMD, 1210, 4.7µF, 50V, 10%, X7R, ROHS	MURATA	GRM32ER71H475KA88L
SER2814L-332KL	2	ea	L102, L103	COIL-PWR INDUCTOR, SMD, 3.3µH, 10%, 48A, 1.2mΩ, ROHS	COILCRAFT	SER2814L-332KL
UUG1C222MNL1ZD	4	ea	C133-C136	CAP, SMD, 16X16.5, 2200µF, 16V, 20%, AL.EL., ROHS	NICHICON	UUG1C222MNL1ZD
UUG2W330MNL1MS	4	ea	C1-C4	CAP, SMD, 18X21.5, 33µF, 450V, 20%, ALUM.ELEC, ROHS	NICHICON	UUG2W330MNL1MS
131-4353-00	1	ea	VOUT	CONN-SCOPE PROBE TEST PT, COMPACT, PCBMNT, ROHS	TEKTRONIX	131-4353-00
1514-2	6	ea	a) 13VDC, 400VDC, PGND-1, PGND-2, DISABLE	CONN-TURRET, TERMINAL POST, TH, ROHS	KEystone	1514-2
1514-2	0	ea	b) 400VDC_RTN	CONN-TURRET, TERMINAL POST, TH, ROHS	KEystone	1514-2
5002	9	ea	a) CS+, VREF, OUTLL, OUTLR, OUTUL, OUTUR	CONN-MINI TEST POINT, VERTICAL, WHITE, ROHS	KEystone	5002
5002	0	ea	b) SR_EN, OUTLLN, OUTLRN	CONN-MINI TEST POINT, VERTICAL, WHITE, ROHS	KEystone	5002
5016-T	13	ea	a) GLL, GLR, LLN, LRN, SR_A, SR_B, PGND-3, 2, +12VOUT, 12V RTN	CONN-COMPACT TEST POINT, SMD, ROHS	KEystone	5016
5016-T	0	ea	b) SGND-1, SGND-	CONN-COMPACT TEST POINT, SMD, ROHS	KEystone	5016
5016-T	0	ea	c) L_PHASE, R_PHASE	CONN-COMPACT TEST POINT, SMD, ROHS	KEystone	5016

**Bill of Materials (Continued)**

PART NUMBER	QTY	UNITS	REFERENCE DESIGNATOR	DESCRIPTION	MANUFACTURER	MANUFACTURER PART
SLW-102-01-S-D	1	ea	J2	CONN-SOCKET STRIP, LP, 2X2, 2.54mm, GOLD/TIN, ROHS	SAMTEC	SLW-102-01-S-D
SLW-107-01-S-D	1	ea	J1	CONN-SOCKET STRIP, LP, 2X7, 2.54mm, GOLD/TIN, ROHS	SAMTEC	SLW-107-01-S-D
1N4148WS-7-F-T	6	ea	D5-D8, D10, D11	DIODE-RECTIFIER, SMD, SOD-323, 2P, 75V, 150mA, ROHS	DIODES INC.	1N4148WS-7-F
BAV70-T	1	ea	D9	DIODE-DUAL, SWITCHING, SMD, SOT23, 100V, 125mA, ROHS	NXP SEMICONDUCTORS	BAV70, 215
ES1B-13-F-T	4	ea	D1, D2, D12, D13	DIODE-RECTIFIER, SMD, 2P, SMA, 100V, 1A, ROHS	DIODES INC.	ES1B-13-F
MURS160T3G-T	1	ea	D4	DIODE-RECTIFIER, SMD, SMB, 2P, 1A, 600V, ROHS	ON SEMICONDUCTOR	MURS160T3G
SS12T3G-T	4	ea	CR3-CR6	DIODE-SCHOTTKY RECTIFIER, SMD, SMA, 2P, 20V, 1A, ROHS	ON SEMICONDUCTOR	SS12T3G (T3= T&R)
0464004.DR	1	ea	F1	FUSE, SMD, 12X4.5, 4A, 250V, NANO2 FAST ACTING, ROHS	LITTELFUSE	0464004.DR
1.5SMC15AT3G-T	1	ea	D3	TVS, UNIDIRECTION, SMD, SMC, 1500W, 12.8V (peak rev) ROHS	ON SEMICONDUCTOR	1.5SMC15AT3G
EL7212CSZ	1	ea	U3	IC-DUAL CHANNEL MOSFET DRIVER, 8P, SOIC, ROHS	INTERSIL	EL7212CSZ
ISL89163FBEBZ	3	ea	U1, U2, U4	IC-6A PWR MOSFET DRIVER, 8P, EPSOIC, ROHS	INTERSIL	ISL89163FBEBZ
ISL89367FRTAZ	1	ea	U108	IC-6A MOSFET DRIVER, 12V INPUT, 16P, DFN, 3X5, ROHS	INTERSIL	ISL89367FRTAZ
BSS138LT1G-T	4	ea	Q5, Q6, Q100, Q101	TRANSIST-MOS, N-CHANNEL, SMD, 3P, SOT23, 50V, 200mA, ROHS	ON SEMICONDUCTOR	BSS138LT1G
FDD6670A	6	ea	Q107-Q109, Q111-Q113	TRANSISTOR-MOS, N-CHANNEL, SMD, DPAK, 30V, 66A, ROHS	FAIRCHILD	FDD6670A
SPB04N60C3	4	ea	Q1-Q4	TRANSIST-MOS, N-CHANNEL, 3P, PG-TO263-8, 650V, 4.5A, ROHS	INFINEON TECHNOLOGY	SPB04N60C3
H2511-00010-1/10W1-T	6	ea	R4-R9	RES, SMD, 0603, 1 $\Omega$ , 1/10W, 1%, TF, ROHS	PANASONIC	ERJ-3RQF1R0V
H2511-00R00-1/10W-T	2	ea	R26, R29	RES, SMD, 0603, 0 $\Omega$ , 1/10W, TF, ROHS	VENKEL	CR0603-10W-000T
H2511-01000-1/10W1-T	6	ea	R2, R3, R18, R19, R21, R24	RES, SMD, 0603, 100 $\Omega$ , 1/10W, 1%, TF, ROHS	VENKEL	CR0603-10W-1000FT

**Bill of Materials (Continued)**

PART NUMBER	QTY	UNITS	REFERENCE DESIGNATOR	DESCRIPTION	MANUFACTURER	MANUFACTURER PART
H2511-01001-1/10W1-T	2	ea	R139, R140	RES, SMD, 0603, 1k, 1/10W, 1%, TF, ROHS	PANASONIC	ERJ-3EKF1001V
H2511-01002-1/10W1-T	4	ea	R34-R37	RES, SMD, 0603, 10k, 1/10W, 1%, TF, ROHS	KOA	RK73H1JT1002F
H2511-02001-1/10W1-T	1	ea	R143	RES, SMD, 0603, 2k, 1/10W, 1%, TF, ROHS	KOA	RK73H1JT2001F
H2511-02490-1/10W1-T	2	ea	R27, R28	RES, SMD, 0603, 249Ω, 1/10W, 1%, TF, ROHS	VENKEL	CR0603-10W-2490FT
H2511-049R9-1/10W1-T	2	ea	R42, R43	RES, SMD, 0603, 49.9Ω, 1/10W, 1%, TF, ROHS	VENKEL	CR0603-10W-49R9FT
H2511-DNP	0	ea	R23, R25	RES, SMD, 0603, DNP PLACEHOLDER, ROHS		
H2512-00200-1/8W1-T	2	ea	R12, R15	RES, SMD, 0805, 20Ω, 1/8W, 1%, TF, ROHS	KOA	RK73H2AT20R0F
H2512-004R7-1/8W1-T	1	ea	R44	RES, SMD, 0805, 4.7Ω, 1/8W, 1%, TF, ROHS	YAGEO	RC0805FR-074R7L
H2512-00R00-1/8W-T	2	ea	R30, R31	R30, R31 RES, SMD, 0805, 0Ω, 1/8W, TF, ROHS	YAGEO	RC0805JR-070RL
H2512-01002-1/8W1-T	4	ea	R13, R17, R144, R145	RES, SMD, 0805, 10k, 1/8W, 1%, TF, ROHS	VENKEL	CR0805-8W-1002FT(PbFREE)
H2512-04R99-1/8W1-T	2	ea	R14, R16	RES, SMD, 0805, 4.99Ω, 1/8W, 1%, TF, ROHS	YAGEO	RC0805FR-074R99L
H2512-DNP	0	ea	R32, R33	RES, SMD, 0805, DNP-PLACEHOLDER, ROHS		
H2513-01000-1/4W1-T	1	ea	R45	RES, SMD, 1206, 100Ω, 1/4W, 1%, TF, ROHS	STACKPOLE/FCI	RMC1/8 100R 1% T/R
H2513-01001-1/4W1-T	1	ea	R148	RES, SMD, 1206, 1.00k, 1/1%, 4W, TF, ROHS	YAGEO	RC1206FR-071KL
H2515-01501-1W5-T	4	ea	R10, R11, R20, R22	RES, SMD, 2512, 1.5k, 1W, 5%, TF, ROHS	VISHAY/DALE	CRCW25121K50JNEG
G108092LF	1	ea	T1	TRANSFORMER-PWR, CUSTOM, 3.2μH, 20%, TH, 16P, 46X45, ROHS	GCI TECHNOLOGIES	G108092LF
P0544NL	1	ea	T3	TRANSFORMER-GATE DRIVE, SMD, 8P, 9X8.6, 8P, 3950μH, ROHS	PULSE	P0544NL
P0584NL	1	ea	T6	TRANSFORMER-GATE DRIVE, TH, 6P, 450μH, 1:1:1RATIO, ROHS	PULSE	P0584NL
P8205NL-T	2	ea	T2, T4	TRANSFORMER-CURRENT SENSE, SMD, 8P, 500μH, 10A, ROHS	PULSE	P8205NL
61215-3B-02500G	2	ea	HS1, HS2	(SEE ASSEMBLY INSTRUCTIONS) HEATSINK-BLACK ANODIZED, 2.5, CUSTOM CUT, ROHS	AAVID/FCI	61215-3B-02500G

**Bill of Materials** (Continued)

PART NUMBER	QTY	UNITS	REFERENCE DESIGNATOR	DESCRIPTION	MANUFACTURER	MANUFACTURER PART
ISL6752DBEVAL1ZFG	1	ea	BAG & SHIP W/BOARD	PWBFG, ISL6752DBEVAL1Z, ROHS	INTERSIL	ISL6752DBEVAL1ZFG
ISL6754DBEVAL1ZFG	1	ea	BAG & SHIP W/BOARD	PWBFG, ISL6754DBEVAL1Z, ROHS	INTERSIL	ISL6754DBEVAL1ZFG
KPA8CTP	2	ea	P103, P104	HDWARE, MTG, CABLE TERMINAL, 6-14AWG, LUG & SCREW, ROHS	BERG/FCI	KPA8CTP
10X12-STATIC-BAG	1	ea	Place assy in bag	BAG, STATIC, 10X12, ZIP	INTERSIL	212403-015
BP100-0.005-00-1112-NA	2	ea	HS1, HS2 (SEE ASSEMBLY INSTRUCTIONS)	INSULATION-BOND-PLY 5mil ADHESIVE TAPE, ROHS	BERGQUIST	BP100-0.005-00-1112-NA
LABEL-SERIAL NUMBER	1	ea		LABEL-FOR SERIAL NUMBER AND BOM REV #	INTERSIL	LABEL-SERIAL NUMBER



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(Rev.4.0-1 November 2017)



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**Renesas Electronics America Inc.**  
1001 Murphy Ranch Road, Milpitas, CA 95035, U.S.A.  
Tel: +1-408-432-8888, Fax: +1-408-434-5351

**Renesas Electronics Canada Limited**  
9251 Yonge Street, Suite 8309 Richmond Hill, Ontario Canada L4C 9T3  
Tel: +1-905-237-2004

**Renesas Electronics Europe Limited**  
Dukes Meadow, Millboard Road, Bourne End, Buckinghamshire, SL8 5FH, U.K  
Tel: +44-1628-651-700, Fax: +44-1628-651-804

**Renesas Electronics Europe GmbH**  
Arcadiastrasse 10, 40472 Düsseldorf, Germany  
Tel: +49-211-6503-0, Fax: +49-211-6503-1327

**Renesas Electronics (China) Co., Ltd.**  
Room 1709 Quantum Plaza, No.27 ZhichunLu, Haidian District, Beijing, 100191 P. R. China  
Tel: +86-10-8235-1155, Fax: +86-10-8235-7679

**Renesas Electronics (Shanghai) Co., Ltd.**  
Unit 301, Tower A, Central Towers, 555 Langao Road, Putuo District, Shanghai, 200333 P. R. China  
Tel: +86-21-2226-0888, Fax: +86-21-2226-0999

**Renesas Electronics Hong Kong Limited**  
Unit 1601-1611, 16/F., Tower 2, Grand Century Place, 193 Prince Edward Road West, Mongkok, Kowloon, Hong Kong  
Tel: +852-2265-6688, Fax: +852-2886-9022

**Renesas Electronics Taiwan Co., Ltd.**  
13F, No. 363, Fu Shing North Road, Taipei 10543, Taiwan  
Tel: +886-2-8175-9600, Fax: +886-2-8175-9670

**Renesas Electronics Singapore Pte. Ltd.**  
80 Bendemeer Road, Unit #06-02 Hyflux Innovation Centre, Singapore 339949  
Tel: +65-6213-0200, Fax: +65-6213-0300

**Renesas Electronics Malaysia Sdn.Bhd.**  
Unit 1207, Block B, Menara Amcorp, Amcorp Trade Centre, No. 18, Jln Persiaran Barat, 46050 Petaling Jaya, Selangor Darul Ehsan, Malaysia  
Tel: +60-3-7955-9390, Fax: +60-3-7955-9510

**Renesas Electronics India Pvt. Ltd.**  
No.777C, 100 Feet Road, HAL 2nd Stage, Indiranagar, Bangalore 560 038, India  
Tel: +91-80-67208700, Fax: +91-80-67208777

**Renesas Electronics Korea Co., Ltd.**  
17F, KAMCO Yangjae Tower, 262, Gangnam-daero, Gangnam-gu, Seoul, 06265 Korea  
Tel: +82-2-558-3737, Fax: +82-2-558-5338