

## ISL81802EVAL1Z

### Evaluation Board

The ISL81802EVAL1Z dual-phase evaluation board (shown in [Figure 4](#)) features the [ISL81802](#), an 80V high voltage dual synchronous buck controller that offers external soft-start, independent enable functions, and integrates UV/OV/OC/OT protection. A programmable switching frequency ranging from 100kHz to 1MHz helps to optimize inductor size while the strong gate driver delivers up to 20A for the buck output.

### Key Features

- Wide input range: 18V to 80V
- High light-load efficiency in pulse skipping DEM operation
- Programmable soft-start
- Optional DEM/PWM operation
- Optional CC/HICCUP OCP protection
- Supports pre-bias output with soft-start
- PGOOD indicator
- OVP, OTP, and UVP protection
- Back biased from output to improve efficiency

### Specifications

The ISL81802EVAL1Z dual-phase evaluation board is designed for high current applications. The current rating of the ISL81802EVAL1Z is limited by the FETs and inductor selected. The ISL81802EVAL1Z electrical ratings are shown in [Table 1](#).

**Table 1. ISL81802EVAL1Z Electrical Ratings**

Parameter	Rating
Input Voltage	18V to 80V
Switching Frequency	200kHz
Output Voltage	12V
Output Current	20A
OCP Set Point	Minimum 22A at ambient room temperature

### Ordering Information

Part Number	Description
ISL81802EVAL1Z	High Voltage Dual Buck Controller Evaluation Board

### Related Literature

For a full list of related documents, visit our website:

- [ISL81802](#) device page

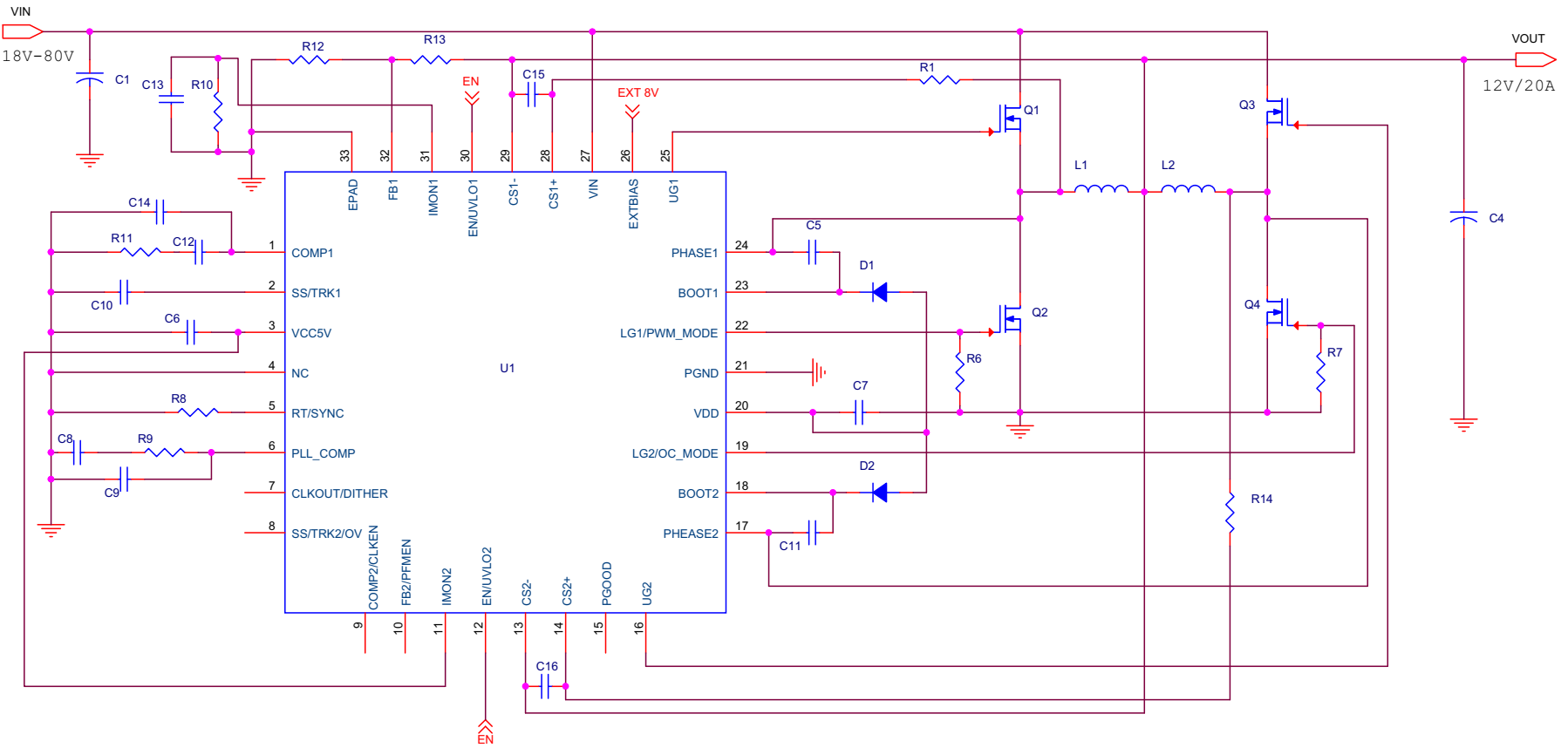


Figure 1. ISL81802EVAL1Z Block Diagram

## 1. Functional Description

The ISL81802EVAL1Z is the same test board used by Renesas application engineers and IC designers to evaluate the performance of the ISL81802 TQFN IC. The board provides an easy and complete evaluation of all the IC and board functions.

As shown in [Figure 3](#), 18V to 80V  $V_{IN}$  is supplied to J1 (+) and J2 (-). The regulated 12V output on J4 (+) and J5 (-) can supply up to 20A to the load. Due to the high power efficiency, the evaluation board can run at 20A continuously without airflow at ambient room temperature conditions.

Test points TP1 through TP23 provide easy access to the IC pin and external signal injection terminals.

As shown in [Table 2](#), connector J6 provides a selection of either Forced PWM mode (shorting Pin 1 and Pin 2) or DEM mode (shorting Pin 2 and Pin 3). Connector J7 provides a selection of either constant current limit (shorting Pin 1 and Pin 2) or HICCUP OCP (shorting Pin 2 and Pin 3). Connector J3 provides an option to disable the converter by shorting its Pin 1 and Pin 2.

### 1.1 Recommended Testing Equipment

The following materials are recommended for testing:

- 0V to 80V power supply with at least 30A source current capability
- Electronic loads capable of sinking current up to 30A
- Digital Multimeters (DMMs)
- 100MHz quad-trace oscilloscope

### 1.2 Operating Range

The input voltage range is from 18V to 80V for an output voltage of 12V. If the output voltage is set to a lower value, the minimum  $V_{IN}$  can be reset to a lower value by changing the ratio of  $R_1$  and  $R_5$ . The minimum EN threshold that  $V_{IN}$  can be set to is 4.5V.

The rated load current is 20A with the OCP point set at a minimum 22A at ambient room temperature conditions. The operating temperature range of this board is -40°C to +85°C.

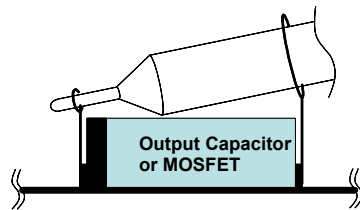
**Note:** Airflow is needed for higher temperature ambient conditions.

### 1.3 Quick Test Guide

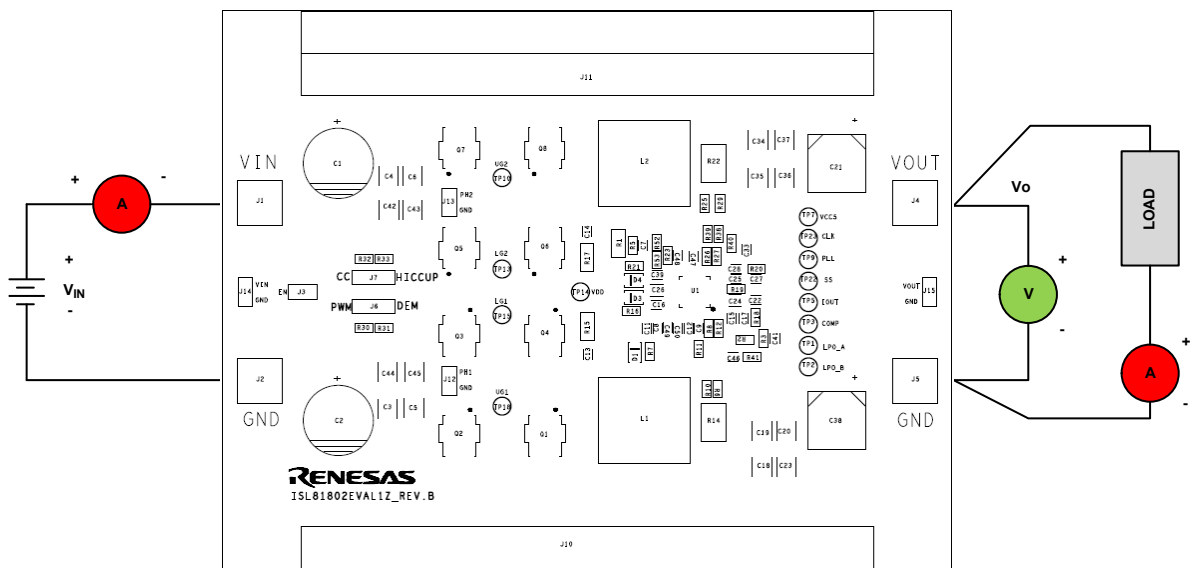
1. Jumper J6 provides the option to select PWM or DEM. Jumper J7 provides the option to select a constant current limit or HICCUP. See [Table 2](#) for the operating options. Ensure that the circuit is correctly connected to the supply and electronic loads before applying any power. See [Figure 3](#) for the proper setup.
2. Turn on the power supply.
3. Adjust the input voltage ( $V_{IN}$ ) within the specified range and observe the output voltage. The output voltage variation should be within 3%.
4. Adjust the load current within the specified range and observe the output voltage. The output voltage variation should be within 3%.
5. Use an oscilloscope to observe output voltage ripple and phase node ringing. For accurate measurement, see [Figure 2](#) for the proper test setup.

**Table 2. Operating Options**

Jumper	Position	Function
3	EN-GND	Disable output
	EN Floating	Enable output
6	Pin 1-2	PWM
	Pin 2-3	DEM
7	Pin 1-2	Constant current limit
	Pin 2-3	HICCUP



**Figure 2. Proper Probe Setup to Measure Output Ripple and Phase Node Ringing**



**Figure 3. Proper Test Setup**

## 2. PCB Layout Guidelines

Careful attention to Printed Circuit Board (PCB) layout requirements is necessary for the successful implementation of an ISL81802 based DC/DC converter. The ISL81802 switches at a high frequency; therefore, the switching times are short. At these switching frequencies, even the shortest trace has significant impedance and the peak gate drive current rises significantly in an extremely short time. The transition speed of the current from one device to another causes voltage spikes across the interconnecting impedances and parasitic circuit elements. These voltage spikes can degrade efficiency, generate EMI, and increase device voltage stress and ringing. Careful component selection and proper PCB layout minimize the magnitude of these voltage spikes.

Three sets of components are critical when using the ISL81802 DC/DC converter:

- Controller
- Switching power components
- Small-signal components

The switching power components are the most critical to the layout because they switch a large amount of energy, which tends to generate a large amount of noise. The critical small-signal components are those connected to sensitive nodes or those supplying critical bias currents. A multilayer PCB is recommended.

Complete the following steps to optimize the PCB layout.

1. Place the input capacitors, FETs, inductor, and output capacitor first. Isolate these power components on dedicated areas of the board with their ground terminals adjacent to one another. Place the input and output high frequency decoupling ceramic capacitors very close to the MOSFETs.
2. If signal components and the IC are placed separately from the power train, Renesas recommends using full ground planes in the internal layers with shared SGND and PGND to simplify the layout design. Otherwise, use separate ground planes for the power ground and the small signal ground. Connect the SGND and PGND together close to the IC. **Note: DO NOT** connect them together anywhere else.
3. Keep the loop formed by the input capacitor, the top FET, and the bottom FET as small as possible.
4. Ensure the current paths from the input capacitor to the FETs, the power inductor, and the output capacitor are as short as possible with maximum allowable trace widths.
5. Place the PWM controller IC close to the lower FETs. The low-side FETs gate drive connections should be short and wide. Place the IC over a quiet ground area. Avoid switching ground loop currents in this area.
6. Place the VDD bypass capacitor very close to the VDD pin of the IC and connect its ground end to the PGND pin. Connect the PGND pin to the ground plane by a via.  
**Note: DO NOT** connect the PGND pin directly to the SGND EPAD.
7. Place the gate drive components (BOOT diodes and BOOT capacitors) together near the controller IC.
8. Place the output capacitors as close to the load as possible. Use short, wide copper regions to connect output capacitors to load to avoid inductance and resistances.
9. Use copper filled polygons or wide, short traces to connect the junction of the upper FET, lower FET, and output inductor. Keep the PHASE nodes connection to the IC short. **Note: DO NOT** unnecessarily oversize the copper islands for the PHASE nodes. Because the phase nodes are subjected to very high dv/dt voltages, the stray capacitor formed between these islands and the surrounding circuitry tends to couple switching noise.
10. Route all high speed switching nodes away from the control circuitry.
11. Create a separate small analog ground plane near the IC. Connect the SGND pin to this plane. Connect all small signal grounding paths including feedback resistors, current monitoring resistors and capacitors, soft-starting capacitors, loop compensation capacitors and resistors, and EN pull-down resistors to this SGND plane.
12. Use a pair of traces with minimum loop for the input or output current sensing connection.
13. Ensure the feedback connection to the output capacitor is short and direct.

## 2.1 ISL81802EVAL1Z Evaluation Board

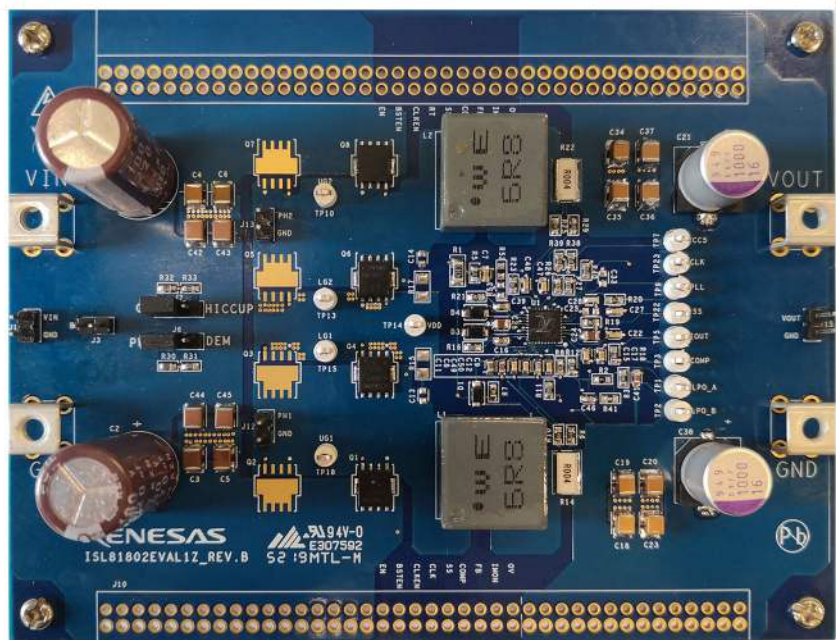


Figure 4. ISL81802EVAL1Z Evaluation Board, Top View

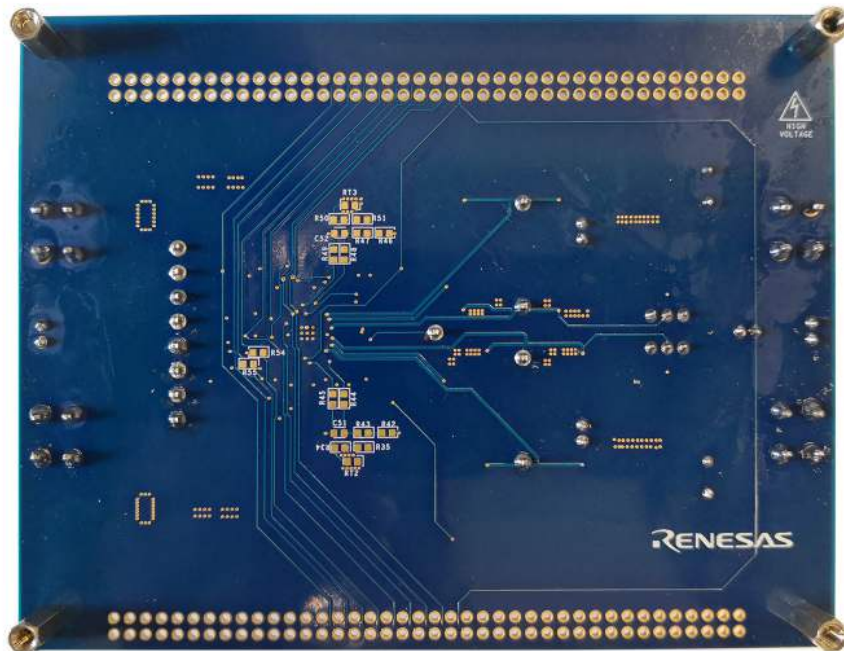


Figure 5. ISL81802EVAL1Z Evaluation Board, Bottom View

## 2.2 ISL81802EVAL1Z Circuit Schematic

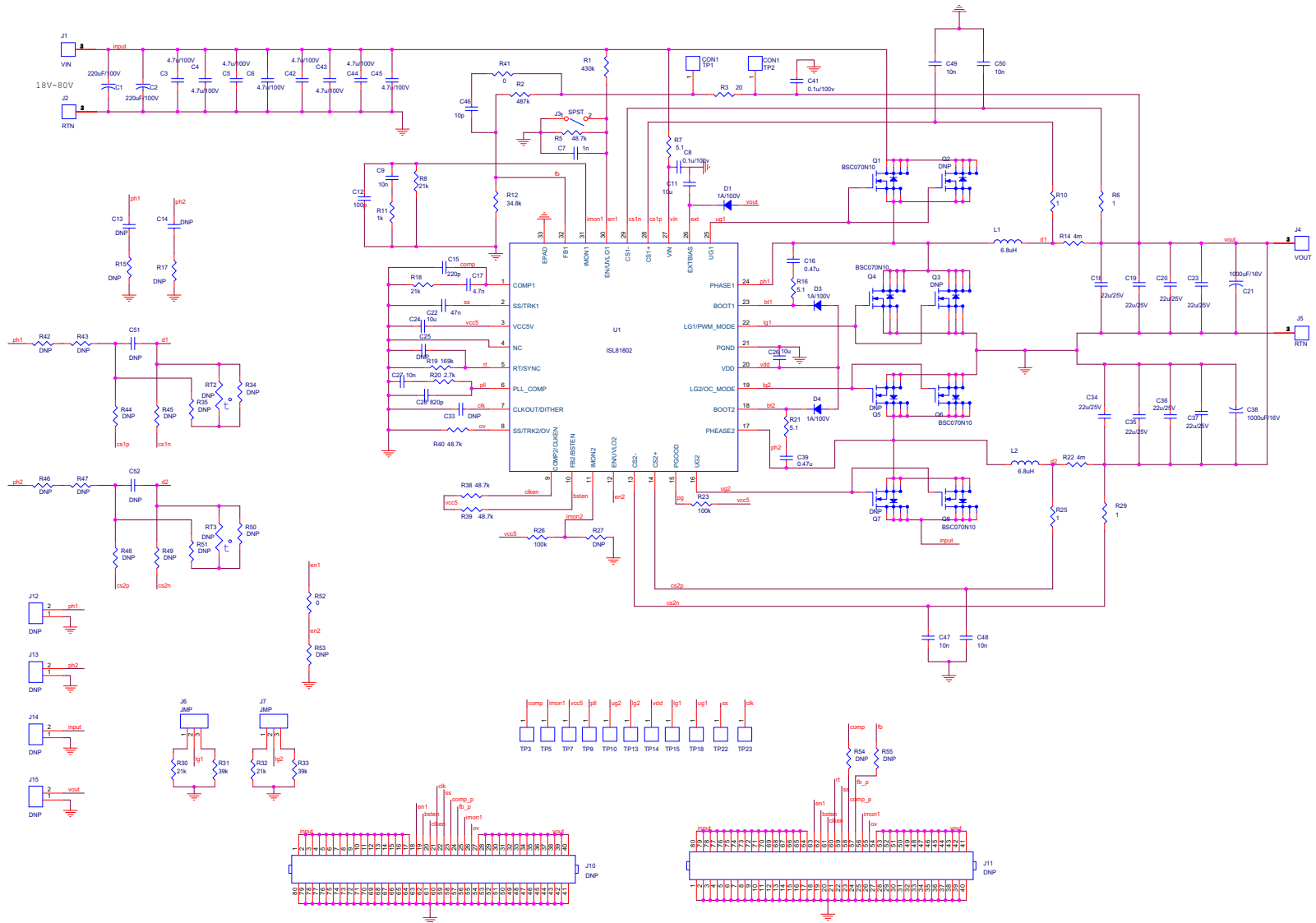


Figure 6. Schematic

## 2.3 Bill of Materials

Qty	Reference Designator	Description	Manufacturer	Manufacturer Part
1		PWB-PCB, ISL81802EVAL1Z, REVB, ROHS	Multilayer PCB Technology	ISL81802EVAL1ZREVBPCB
2	C1, C2	CAP, RADIAL, 12.5x26.5, 220µF, 100V, 20%, ALUM.ELEC., 5mm, ROHS	United Chemi-Con	EKZN101ELL221MK25S
8	C3, C4, C5, C6, C42, C43, C44, C45	CAP-AEC-Q200, SMD, 1210, 4.7µF, 100V, 10%, X7S, ROHS	TDK	CGA6M3X7S2A475K200AB
1	C7	CAP, SMD, 0603, 1000pF, 50V, 10%, X7R, ROHS	TDK	C1608X7R1H102K080AE
2	C8, C41	CAP, SMD, 0603, 0.1µF, 100V, 10%, X7R, ROHS	Vishay	GRJ188R72A104KE11D
6	C9, C27, C47, C48, C49, C50	CAP, SMD, 0603, 0.01µF, 100V, 5%, X7R, ROHS	Kemet	C0603C103J1RACTU
1	C11	CAP, SMD, 0603, 10µF, 16V, X5R, 10%, ROHS	Murata	GRM188R61C106KAALJ
1	C12	CAP, SMD, 0603, 100PF, 50V, X7R, 10%, ROHS	Kemet	C0603C101K5RACTU
1	C46	CAP, SMD, 0603, 10PF, 50V, X7R, 10%, ROHS	Kemet	C0603C100K5RACTU
1	C15	CAP, SMD, 0603, 220pF, 50V, 10%, X7R, ROHS	Murata	GRM188R71H221KA01D
2	C16, C39	CAP, SMD, 0603, 0.47µF, 25V, 10%, X7R, ROHS	Murata	GRM188R71E474KA12D
1	C17	CAP, SMD, 0603, 4700PF, 50V, X7R, ROHS	Murata	GCJ188R71H472KA01D
8	C18, C19, C20, C23, C34, C35, C36, C37	CAP, SMD, 1210, 22µF, 25V, X7R, ROHS	Murata	GRM32ER71E226KE15L
2	C21, C38	CAP-OSCON, SMD, 10mm, 1000µF, 16V, 20%, 12mΩ, ROHS	Panasonic	16SVPF1000M
1	C22	CAP, SMD, 0603, 0.047µF 25V X7R, ROHS	Kemet	C0603C473K3RACTU
2	C24, C26	CAP, SMD, 0805, 10µF, 16V, 10%, X7S, ROHS	Murata	GRM21BC71C106KE11L
1	C28	CAP, SMD, 0603, 820pF, 50V, 10%, X7R, ROHS	Kemet	C0603C821K5RACTU
0	C13, C14, C25, C51, C52, C33	CAP, SMD, 0603, DNP-PLACE HOLDER, ROHS		
3	D1, D3, D4	DIODE-RECTIFIER, SMD, 2P, S0D-123FL, 100V, 1A, ROHS	ON Semiconductor	MBR1H100SFT3G
4	J1, J2, J4, J5	HDWARE, TERMINAL, M4 METRIC SCREW, TH, 4P, SNAP-FIT, ROHS	Keystone	7795
1	J3	CONN-HEADER, 1x2, BRKAWY 1x36, 2.54mm, ROHS	BERG/FCI	68000-236HLF
2	J6, J7	CONN-HEADER, 1x3, BREAKAWY 1x36, 2.54mm, ROHS	BERG/FCI	68000-236HLF
0	J10, J11, J12, J13, J14, J15	2.54mm Headers, DNP-PLACE HOLDER, ROHS		
2	L1, L2	COIL-PWR INDUCTOR, SMD, 6.8µH, 20%, 15A, 4.1mΩ, ROHS	Würth	74439370068



Qty	Reference Designator	Description	Manufacturer	Manufacturer Part
1	U1	80V DUAL-BUCK PWM CONTROLLER, 32P, TQFN, 5x5, ROHS	Renesas Electronics America	ISL81802FRTZ
4	Q1, Q4, Q6, Q8	TRANSISTOR-MOS, N-CHANNEL, SMD, 8P, PPK SO-8, 100V, 80A, ROHS	Infineon	BSC070N10NS5ATMA1
0	Q2, Q3, Q5, Q7	DO NOT POPULATE OR PURCHASE		
1	R1	RES SMD 430kΩ 1% 1/4W 1206	Panasonic	ERJ-8ENF4303V
1	R2	RES SMD 487kΩ 1% 1/10W 0603	Yageo	RC0603FR-07487KL
4	R5, R38, R39, R40	RES SMD 48.7kΩ 1% 1/10W 0603	Yageo	RC0603FR-0748K7L
1	R3	RES SMD 20Ω 1% 1/10W 0603	Yageo	RC0603FR-0720RL
4	R6, R10, R25, R29	RES SMD 1Ω 1% 1/10W 0603	Panasonic	ERJ-3RQF1R0V
1	R7	RES SMD 5.1Ω 1% 1/10W 0603	Yageo	RC0603FR-075R1L
4	R8, R18, R30, R32	RES SMD 21kΩ 1% 1/10W 0603	Yageo	RC0603FR-0721KL
1	R11	RES SMD 1kΩ 1% 1/10W 0603	Yageo	RC0603FR-071KL
1	R12	RES SMD 34.8kΩ 1% 1/10W 0603	Yageo	RC0603FR-0734K8L
2	R14, R22	RES SMD 0.004Ω 3W 2512 WIDE	Susumu	KRL6432E-M-R004-F-T1
2	R16, R21	RES SMD 5.1Ω 1% 1/10W 0603	Yageo	RC0603FR-075R1L
1	R19	RES SMD 169kΩ 1% 1/10W 0603	Venkel	CR0603-10W-1693FT
1	R20	RES SMD 2.7kΩ 1% 1/10W 0603	Yageo	RC0603FR-072K7L
2	R23, R26	RES SMD 100kΩ 1% 1/10W 0603	Yageo	RC0603FR-07100KL
2	R31, R33	RES SMD 39kΩ 1% 1/10W 0603	Yageo	RC0603FR-0739KL
2	R41, R52	RES SMD 0Ω 1% 1/10W 0603	Yageo	RC0603FR-070RL
0	R34, R35, R42, R43, R44, R45, R46, R47, R48, R49, R50, R51, R53, R54, R55, R17, R15, R27, RT2, RT3,	RES, SMD, 0603, DNP-PLACE HOLDER, ROHS		
13	TP1, TP2, TP3, TP5, TP7, TP9, TP10, TP13, TP14, TP15, TP18, TP22, TP23	CONN-COMPACT TEST PT, VERTICAL, WHT, ROHS	Keystone	5007
3	J3, J6, J7	CONN-JUMPER, SHORTING, 2PIN, BLACK, GOLD, ROHS	Sullins	SPC02SYAN
4	Four corners	SCREW, 4-40x1/4in, PHILLIPS, PANHEAD, STAINLESS, ROHS	Keystone	2204
4	Four corners	STANDOFF, 4-40x3/4in, F/F, HEX, ALUMINUM, 0.25 OD, ROHS	Keystone	7795

## 2.4 Board Layout

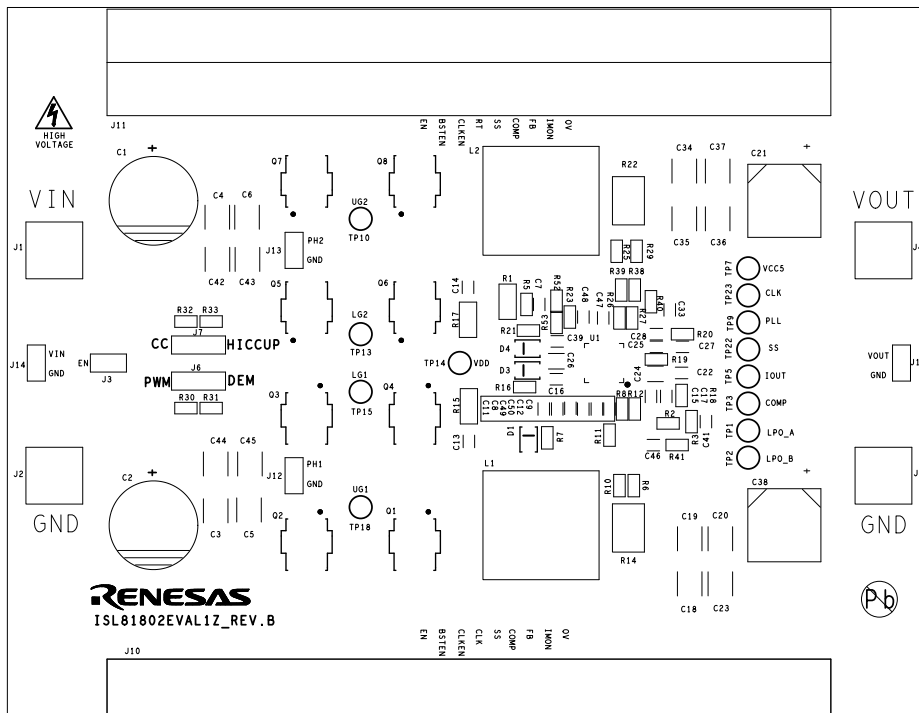


Figure 7. Silkscreen Top

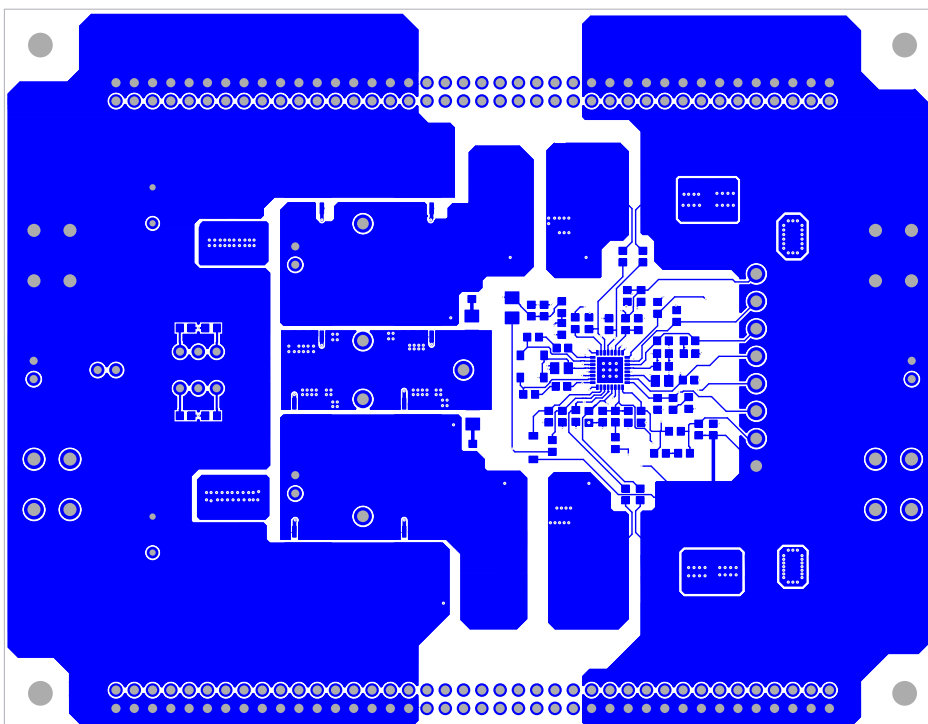


Figure 8. Top Layer

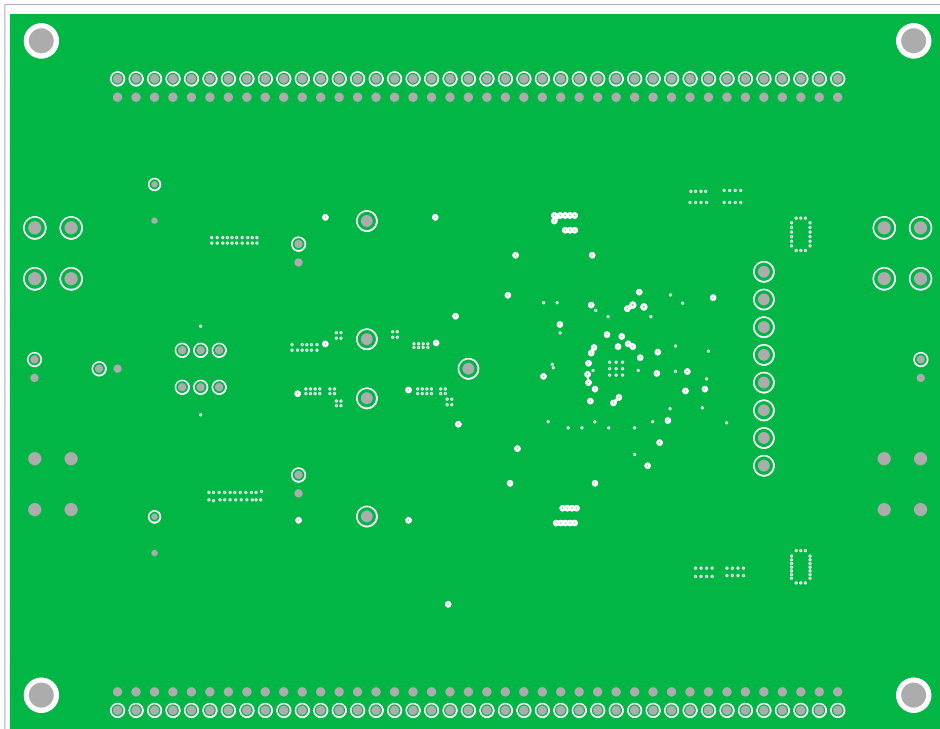


Figure 9. Second Layer (Solid Ground)

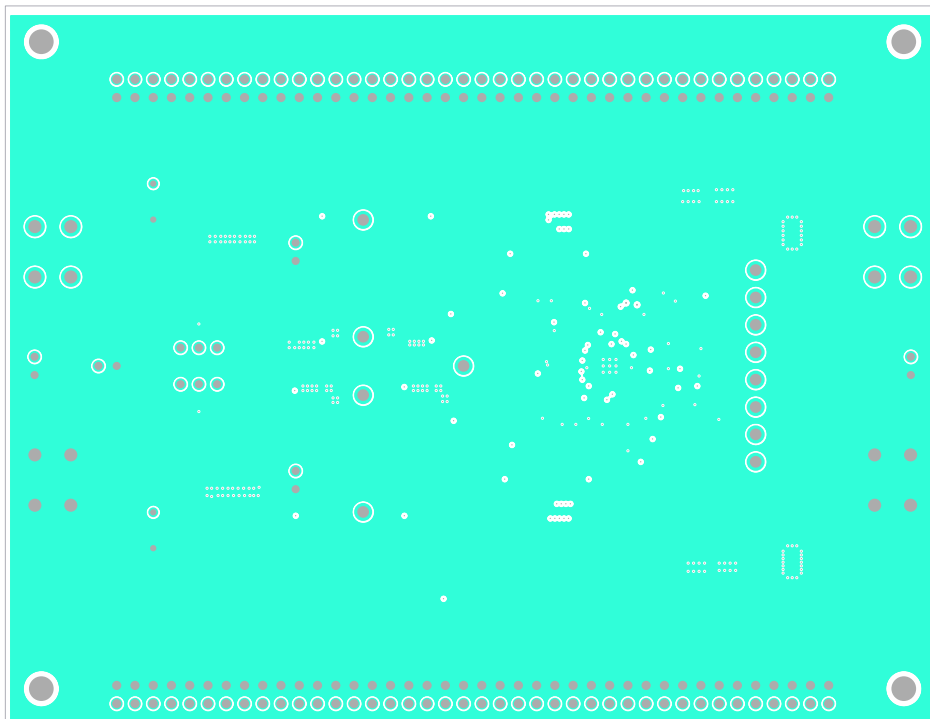


Figure 10. Third Layer

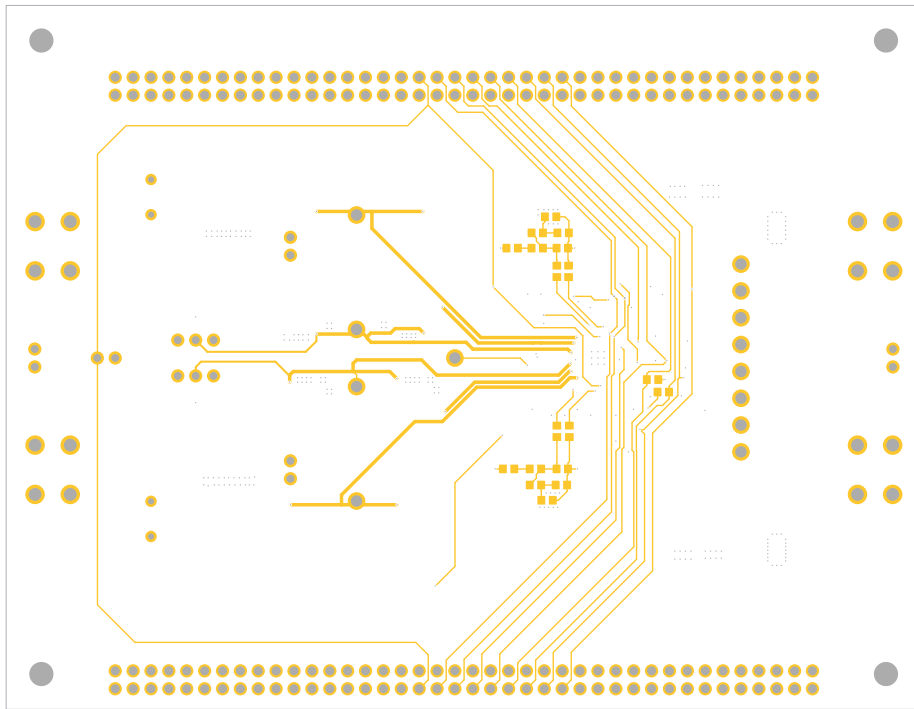


Figure 11. Bottom Layer

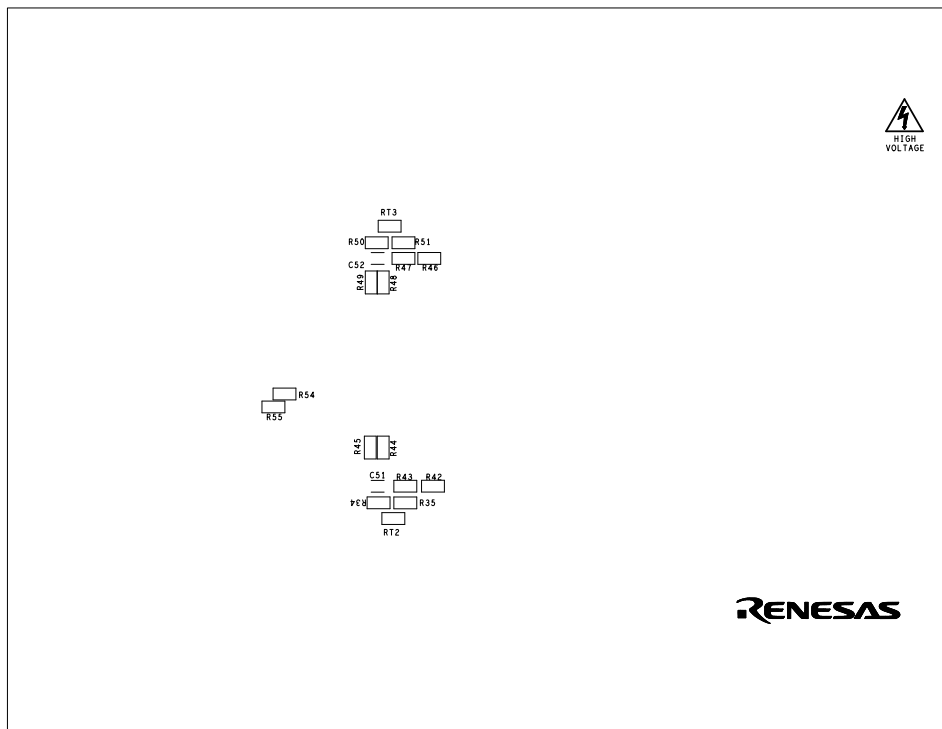


Figure 12. Silkscreen Bottom

## 3. Design Example

### 3.1 Design Requirements

Parameter	Rating
Input Voltage	18V to 80V
Switching Frequency	200kHz
Output Voltage	12V
Output Current	20A
OCP Set Point	22A
Output Mode	Dual phase
PWM Mode	Forced PWM
OCP Mode	Constant current

### 3.2 Frequency Setting

The default switching frequency of the PWM controller is determined by the resistor  $R_T$  ( $R_{19}$ ). It adjusts the default switching frequency from 100kHz to 1MHz. The  $R_T$  value for  $f_{SW} = 200\text{kHz}$  is calculated using [Equation 1](#).

$$(EQ. 1) \quad R_T = \left( \frac{34.7}{f_{SW}} - 4.78 \right) = \frac{34.7}{0.2} - 4.78 = 168.72\text{k}\Omega$$

where  $f_{SW}$  is the switching frequency in MHz. Select a standard value resistor  $R_T = 169\text{k}\Omega$ .

### 3.3 Output Voltage Setting

The output voltage can be set from 0.8V up to a level determined by the feedback voltage divider. A resistive divider from the output to ground sets the output voltage. Connect the center point of the divider to the FB\_OUT pin. With  $V_{OUT} = 12\text{V}$  and  $R_{FBO1}$  ( $R_2$ ) = 487k, the  $R_{FBO2}$  ( $R_{12}$ ) value is calculated using [Equation 2](#).

$$(EQ. 2) \quad R_{FBO2} = \frac{0.8\text{V} \times R_{FBO1}}{V_{OUT} - 0.8\text{V}} = \frac{0.8\text{V} \times 487\text{k}\Omega}{12\text{V} - 0.8\text{V}} = 34.78\text{k}\Omega$$

where  $R_{FBO1}$  ( $R_2$ ) is the top resistor of the feedback divider network and  $R_{FBO2}$  ( $R_{12}$ ) is the bottom resistor connected from FB\_OUT to ground. To avoid an unstable state during hiccup, the value of  $R_{FBO1}$  and  $R_{FBO2}$  in parallel should be no less than 30k. Select a standard value resistor  $R_{FBO2} = 34.8\text{k}\Omega$ .

### 3.4 UVLO Setting

The ISL81802 has input UVLO protection. When the voltage on the EN/UVLO pin reaches 1.8V, the PWM modulator is enabled. Accurate UVLO feature can be implemented by feeding the  $V_{IN}$  into the EN/UVLO pin using a voltage divider,  $R_{UV1}$  ( $R_1$ ) and  $R_{UV2}$  ( $R_5$ ). The  $V_{IN}$  UVP rising threshold is calculated using [Equation 3](#).

$$(EQ. 3) \quad V_{UVRISE} = \frac{V_{UVLO\_THR}(R_{UV1} + R_{UV2}) - I_{LEAK}R_{UV1}R_{UV2}}{R_{UV2}} = \frac{1.8\text{V}(430\text{k}\Omega + 48.7\text{k}\Omega) - 2.8\mu\text{A}(430\text{k}\Omega)(48.7\text{k}\Omega)}{48.7\text{k}\Omega} = 16.49\text{V}$$

The  $V_{IN}$  UVP falling threshold is calculated using [Equation 4](#).

$$(EQ. 4) \quad V_{UVFALL} = \frac{V_{UVLO\_THR}(R_{UV1} + R_{UV2}) - I_{UVLO\_HYST}R_{UV1}R_{UV2}}{R_{UV2}} = \frac{1.8\text{V}(430\text{k}\Omega + 48.7\text{k}\Omega) - 6.8\mu\text{A}(430\text{k}\Omega)(48.7\text{k}\Omega)}{48.7\text{k}\Omega} = 14.77\text{V}$$

where  $V_{UVLO\_THR}$  is the 1.8V UVLO rising threshold and  $I_{UVLO\_HYST}$  is the 6.8μA UVLO hysteresis current.

### 3.5 Soft-Start Capacitor

The soft-start time for dual-phase is set by the value of the soft-start capacitor  $C_{SS}$  ( $C_{22}$ ) connected from SS/TRK1 to GND. The soft-start time with  $C_{SS} = 47\text{nF}$  is calculated using [Equation 5](#).

$$(EQ. 5) \quad t_{SS} = 0.8V \left( \frac{C_{SS}}{4\mu A} \right) = 0.8V \times \left( \frac{47\text{nF}}{4\mu A} \right) = 9.4\text{ms}$$

When the soft-start time set by external  $C_{SS}$  or tracking is less than 1.7ms, an internal soft-start circuit of 1.7ms takes over the soft-start.

### 3.6 MOSFET Considerations

The MOSFETs are selected based on  $r_{DS(ON)}$ , gate supply requirements, and thermal management considerations. The maximum operation voltage of the MOSFETs in Buck mode is decided by the maximum  $V_{IN}$  voltage.

The power loss of the upper and lower MOSFETs for each phase is calculated using [Equation 6](#) and [Equation 7](#). The equations assume linear voltage current transitions and do not model power loss due to the reverse recovery for the body diode of the lower MOSFET.

$$(EQ. 6) \quad P_{UPPERMAX} = \frac{(I_{OUT})^2 (r_{DS(ON)}) (V_{OUT})}{V_{INMAX}} + \frac{(I_{OUT}) (V_{INMAX}) (t_{SW}) (f_{SW})}{2}$$

$$= \frac{(10A)^2 (6\text{m}\Omega) (12V)}{80V} + \frac{(10A)(80V) \left( \frac{6\text{nC}}{\left( \frac{8V-4.9V}{3.3\Omega} \right)} + \frac{6\text{nC}}{\left( \frac{4.9V}{3.3\Omega} \right)} \right) (200\text{kHz})}{2} = 0.09\text{W} + 0.834\text{W} = 0.843\text{W}$$

$$(EQ. 7) \quad P_{LOWERMAX} = \frac{(I_{OUT})^2 (r_{DS(ON)}) (V_{INMAX} - V_{OUT})}{V_{INMAX}}$$

$$= \frac{(10A)^2 (6\text{m}\Omega) (80V - 12V)}{80V} = 0.51\text{W}$$

Ensure that all MOSFETs are within their maximum junction temperature at high ambient temperature by calculating the temperature rise according to package thermal resistance specifications.

### 3.7 Inductor Selection

The inductor value determines the ripple current of the converter. The ripple voltage is a function of the ripple current and the output capacitor(s) ESR. Assume the ripple current ratio is 80% of the inductor average current at the maximum input voltage and the full output load condition. The inductor value for each phase is calculated using [Equation 8](#).

$$(EQ. 8) \quad L_{INMIN} = \frac{(V_{INMAX} - V_{OUT}) (V_{OUT})}{(f_{SW}) (0.8 \times I_{OUTMAX}) (V_{INMAX})} = \frac{(80V - 12V) (12V)}{(200\text{kHz}) (0.8 \times 10A) (80V)} = 6.375\mu\text{H}$$

The recommended inductor value is 6.8 $\mu\text{H}$ . Then the ripple current and peak current are calculated using [Equation 9](#), [Equation 10](#), and [Equation 11](#).

$$(EQ. 9) \quad \Delta I_{LMAX} = \frac{(V_{INMAX} - V_{OUT}) (V_{OUT})}{(f_{SW}) (L) (V_{INMAX})} = \frac{(80V - 12V) (12V)}{(200\text{kHz}) (6.8\mu\text{H}) (80V)} = 7.5\text{A}$$

$$(EQ. 10) \quad I_{LRMS} = \sqrt{(I_{OUTMAX})^2 + \frac{(\Delta I_{LMAX})^2}{12}} = \sqrt{(10A)^2 + \frac{(7.5A)^2}{12}} = 10.23\text{A}$$

$$(EQ. 11) \quad I_{LPEAKMAX} = \frac{I_{OUTOCP}}{2} + \frac{\Delta I_{LMAX}}{2} = \frac{22A}{2} + \frac{7.5A}{2} = 14.75A$$

The saturation current of the inductor should be larger than 14.75A. The heat rating current of the inductor should be larger than 10.23A.

The maximum DC power dissipation in the inductor is calculated using [Equation 12](#).

$$(EQ. 12) \quad P_{LMAX} = (I_{OUT})^2(DCR) = (10A)^2 \times (4.1m\Omega) = 0.41W$$

### 3.8 Output Capacitor Selection

The minimum capacitor value required to provide the full, rising step, transient load current during the response time of the inductor is shown in [Equation 13](#).

$$(EQ. 13) \quad C_{OUTMIN} = \frac{L(I_{TRAN})^2}{2(V_{INMIN} - V_{OUT})(\Delta V_{OUT})} = \frac{6.8\mu F \times (10A - 0A)^2}{2(18V - 12V)(12V \times \frac{1.5}{100})} = 314.8\mu F$$

where  $C_{OUTMIN}$  is the minimum output capacitor(s) required,  $I_{TRAN}$  is the transient load current step, and  $\Delta V_{OUT}$  is the drop in output voltage allowed during the load transient. Choose a capacitor no less than 314.8 $\mu$ F for each phase. 1000 $\mu$ F electrolytic capacitor and 88 $\mu$ F MLCC in total are used for each phase on this board.

The output voltage ripple is due to the inductor ripple current and the ESR of the output capacitors as defined by [Equation 14](#).

$$(EQ. 14) \quad V_{RIPPLE} = \Delta I_{LMAX} \times ESR = 7.5A \times 5m\Omega = 37.5mV$$

### 3.9 Input Capacitor Selection

The important parameters for the input capacitors are the voltage rating and the RMS current rating. The capacitor voltage rating should be at least 1.25 times greater than the maximum input voltage and 1.5 times is a conservative guideline. The AC RMS input current varies with the load given in [Equation 15](#).

$$(EQ. 15) \quad I_{RMS} = \sqrt{\left(D - \frac{\text{floor}(2 \times D)}{2}\right) \times \left(\frac{1 + \text{floor}(2 \times D)}{2} - D\right) \times I_{OUT}}$$

where  $\text{floor}(2 \times D)$  equals 0 when  $D < 0.5$  and equals 1 when  $D \geq 0.5$ .

The maximum RMS current for two phases in total supplied by the input capacitance occurs at  $D = 0.25$  and  $D = 0.75$ , and only  $D = 0.25$  is within the range of this application. Therefore, the maximum AC RMS current is shown in [Equation 16](#).

$$(EQ. 16) \quad I_{RMSMAX} = \sqrt{D \times \left(\frac{1}{2} - D\right)} \times I_{OUTMAX} = \sqrt{0.25 \times \left(\frac{1}{2} - 0.25\right)} \times 10A = 2.5A$$

Renesas recommends using a mix of input bypass capacitors to control the voltage ripple across the MOSFETs. Use ceramic capacitors for the high frequency decoupling and bulk capacitors to supply the RMS current. Two 220 $\mu$ F electrolytic capacitors with 2.2A rating current and eight 4.7 $\mu$ F ceramic capacitors are used to share the 2.5A RMS input current on this board.

### 3.10 First Level Peak Current Limit and Sense Resistor Selection

The inductor peak current is sensed by the sense resistor  $R_S$  ( $R_{14}$ ). When the voltage drop on  $R_S$  reaches the set point  $V_{OCSET-CS}$  typical 85mV, it triggers the pulse-by-pulse peak current limit. With the current limit set point  $I_{OCPP1} = 2 \times I_{OUTMAX} = 20A$  for each phase, the value of the sense resistor is calculated using [Equation 17](#).

$$(EQ. 17) \quad R_S = \frac{V_{OCSET-CS}}{I_{OCPP1}} = \frac{85mV}{20A} = 4.25m\Omega$$

Select a standard value resistor  $R_S = 4m\Omega$ . Then the actual peak current limit is calculated using [Equation 18](#).

$$(EQ. 18) \quad I_{OCPP1} = \frac{V_{OCSET-CS}}{R_S} = \frac{85mV}{4m\Omega} = 21.25A$$

The maximum power dissipation in  $R_S$  is calculated by [Equation 19](#).

$$(EQ. 19) \quad P_{RSMAX} = (I_{OUT})^2 R_S = (10A)^2 (4m\Omega) = 0.4W$$

Therefore, a sense resistor with 1W power rating is sufficient for this application.

### 3.11 Second Level Hiccup Peak Current Protection

In the output dead short condition especially at high  $V_{IN}$ , the inductor current runs away with the minimum on PWM duty. The ISL81802 integrates a second level hiccup type of peak current protection. The second level peak current protection set point  $I_{OCPP2}$  is calculated using [Equation 20](#).

$$(EQ. 20) \quad I_{OCPP2} = \frac{V_{OCSET-CS-HIC}}{R_S} = \frac{115mV}{4m\Omega} = 28.75A$$

### 3.12 Output Average Overcurrent Protection and $R_{IM}$ Selection

The ISL81802 provides either constant current or hiccup type of overcurrent protection for output average current. The OCP mode is set by a resistor connected between the LG2/OC\_MODE pin and ground. With output constant current/hiccup set point  $I_{OUTOCP} = 22A$  for two phases in total, the current monitoring resistor  $R_{IM}$  ( $R_8$ ) is calculated using [Equation 21](#).

$$(EQ. 21) \quad R_{IM} = \frac{1.2}{I_{OUTOCP} \times R_S \times G_{mCS} + 2 \times I_{CSOFFSET}} = \frac{1.2V}{22A \times 4m\Omega \times 195\mu S + 2 \times 20\mu A} = 20.99k\Omega$$

where  $I_{CSOFFSET}$  is the output current sense op amp internal offset current, typical 20 $\mu$ A. Select a standard value resistor  $R_{IM} = 21k\Omega$ .

### 3.13 Output Mode Selection

When the IMON2 pin voltage is higher than 3V, the IC is set for one output dual-phase application, and the original IMON2 current monitor function pin is disconnected from the IMON2 pin and internally connected to the IMON1 pin. The IMON2 pin is connected to VCC5 using  $R_{26}$  for dual-phase setting on this board.

### 3.14 PWM Mode Selection

You can set the ISL81802 to either forced PWM mode or DE mode. The mode is set by a resistor  $R_{PWMMODE}$  ( $R_{30}$  or  $R_{31}$ ) connected between the LG1/PWM\_MODE pin and GND. The boundary resistor value for  $R_{PWMMODE}$  is calculated using [Equation 22](#).

$$(EQ. 22) \quad R_{PWMMODE} = \frac{0.3V}{10\mu A} = 30k\Omega$$



A resistor less than 30kΩ sets the converter to forced PWM mode, while a resistor higher than 30kΩ sets the converter to DE mode. Considering the tolerance in all temperature ranges, Renesas recommends using 21kΩ to set Forced PWM mode and 39kΩ to set DE mode.

### 3.15 Overcurrent Protection Mode Selection

The ISL81802 is set to either a constant current or hiccup type of overcurrent protection for output average current by selecting a different value of the resistor  $R_{OCMODE}$  ( $R_{32}$  or  $R_{33}$ ) connected between LG2/OC\_MODE and GND. The boundary resistor value for  $R_{OCMODE}$  is calculated using [Equation 23](#).

$$(EQ. 23) \quad R_{OCMODE} = \frac{0.3V}{10\mu A} = 30k\Omega$$

A resistor less than 30kΩ sets the converter to constant current mode, while a resistor higher than 30kΩ sets the converter to Hiccup mode. Considering the tolerance in all temperature ranges, Renesas recommends using 21kΩ to set constant current and 39kΩ to set the Hiccup mode.

### 3.16 Phase Lock Loop (PLL)

The PLL of the ISL81802 ensures the wide range of accurate clock frequency and phase setting. It also makes the internal clock easily synchronized to an external clock with the frequency either lower or higher than the internal setting. The external compensation network of  $R_{PLL}$  ( $R_{20}$ ),  $C_{PLL1}$  ( $C_{27}$ ), and  $C_{PLL2}$  ( $C_{28}$ ) is needed to connect to the PLL\_COMP pin to ensure PLL stable operation. Renesas recommends choosing 2.7kΩ for  $R_{PLL}$ , 10nF for  $C_{PLL1}$ , and 820pF for  $C_{PLL2}$ .

### 3.17 Feedback Loop Compensation

Due to the current loop feedback, the modulator has a single pole response with -20dB slope at a frequency determined by the load using [Equation 24](#).

$$(EQ. 24) \quad F_{PO} = \frac{1}{2\pi \cdot R_O \cdot C_O} = \frac{1}{2\pi \cdot \frac{12V}{10A} \cdot 1088\mu F} = 122Hz$$

where  $R_O$  is load resistance and  $C_O$  is total load capacitance for each phase. For this type of modulator, a Type 2 compensation circuit is usually sufficient.

[Figure 13](#) shows a Type 2 amplifier and its response, along with the responses of the current mode modulator and the converter. The Type 2 amplifier, in addition to the pole at origin, has a zero-pole pair that causes a flat gain region at frequencies between the zero and the pole. The  $R_{COMP}$ ,  $C_{COMP1}$ , and  $C_{COMP2}$  network connected on the Gm regulator output COMP pin is needed to compensate the loop for stable operation. The loop stability can be affected by many different factors such as  $V_{IN}$ ,  $V_{OUT}$ , load current, switching frequency, inductor value, output capacitance, and the compensation network on the COMP pin.

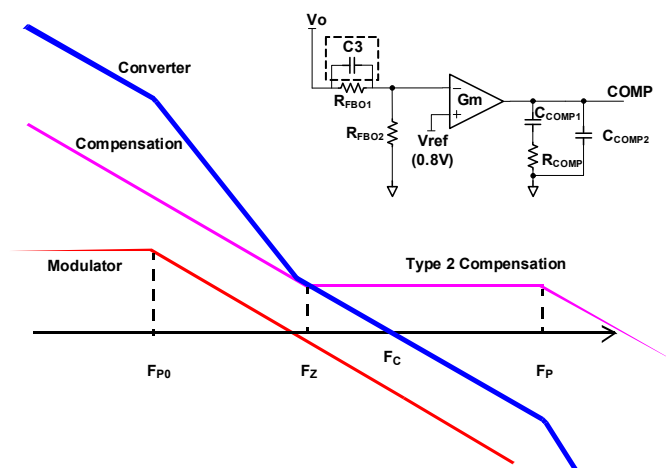


Figure 13. Feedback Loop Compensation

High amplifier zero frequency gain and modulator gain are chosen to satisfy most typical applications. The crossover frequency appears at the point where the modulator attenuation equals the amplifier high-frequency gain. The crossover frequency  $F_C$  is usually about 1/10 to 1/30 of switching frequency. To fulfill the various applications, large value capacitors are used on the output side. Therefore, a reasonable target crossover frequency  $F_C$  in this application is 4kHz.

The compensation zero  $F_Z$  is usually placed between  $F_{PO}$  and  $F_C$ . Setting  $F_Z$  to 1.6kHz, with  $C_{COMP1}$  ( $C_{17}$ ) = 4.7nF, the  $R_{COMP}$  ( $R_{18}$ ) is calculated using [Equation 25](#).

$$(EQ. 25) \quad R_{COMP} = \frac{1}{2\pi \cdot F_Z \cdot C_{COMP1}} = \frac{1}{2\pi \cdot 1.6\text{kHz} \cdot 4.7\text{nF}} = 21.17\text{k}\Omega$$

Select a standard value resistor  $R_{COMP} = 21\text{k}\Omega$ . A larger  $C_{COMP1}$  makes the loop more stable by giving a larger phase margin, but the loop bandwidth is lower. Lower  $R_{COMP}$  improves stability but slows the loop response.

A high-frequency pole  $F_P$  is placed by a capacitor  $C_{COMP2}$  ( $C_{15}$ ) in parallel with  $R_{COMP}$  and  $C_{COMP1}$ . Set the frequency of this pole at about 7 to 10 times of crossover frequency  $F_C$  to provide attenuation of switching ripple and noise on COMP, while avoiding excessive phase loss at the crossover frequency. For a target  $F_P = 35\text{kHz}$ , the  $C_{COMP2}$  is calculated using [Equation 26](#).

$$(EQ. 26) \quad C_{COMP2} = \frac{1}{2\pi \cdot R_{COMP1} \cdot F_P} = \frac{1}{2\pi \cdot 21\text{k}\Omega \cdot 35\text{kHz}} = 216.6\text{pF}$$

Select a standard value capacitor  $C_{COMP2} = 220\text{pF}$ .

Some phase boost can be achieved by connecting capacitor  $C_3$  in parallel with the upper resistor  $R_{FBO1}$  of the divider. These values provide a good starting point for the compensation design, and the final compensation network should be optimized with bench test.

### 3.18 Parallel Connection

The ISL81802EVAL1Z evaluation board can operate in parallel, in a daisy chain setup. [Figure 14](#) shows the wiring of two units in parallel and [Figure 15](#) shows three units in parallel.

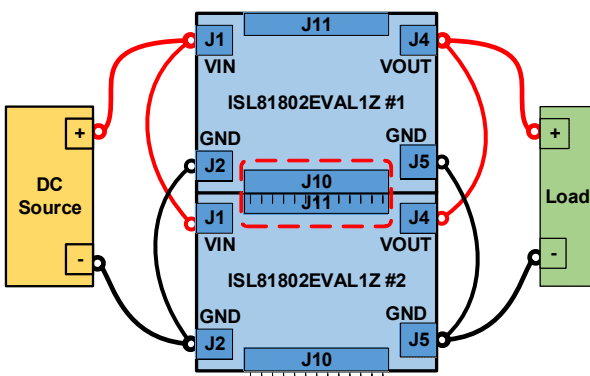


Figure 14. Setup for Two Units in Parallel

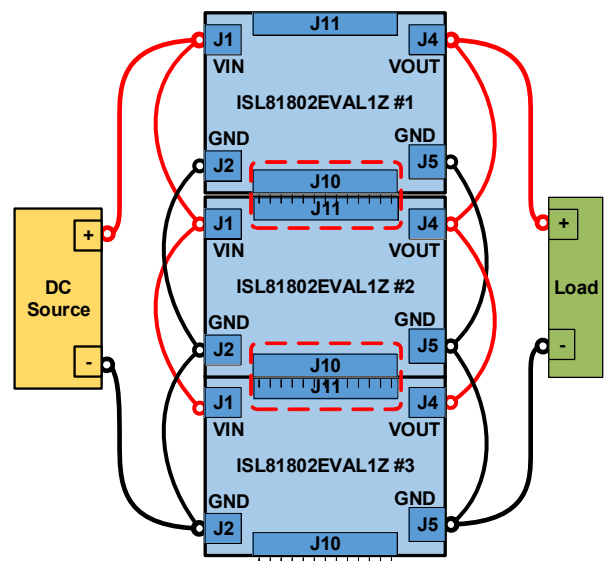


Figure 15. Setup for Three Units in Parallel

[Table 3](#) shows the CLKOUT/DITHER phase settings with different EN/UVLO2 pin connection and IMON2 pin voltage.

**Table 3. CLKOUT and Channel 2 Phase Shift vs EN/UVLO2 and IMON2 Voltage**

CLKOUT Phase Shift (°)	Channel 2 Phase Shift (°)	IMON2 Voltage (V)	EN/UVLO2
90	180	0 to 4.3	Tie to EN/UVLO1
60	180	4.7 to 5	Tie to EN/UVLO1
240	120	3 to 5	Tie to SGND

**Notes:**

1. CLKOUT Phase Shift: CLKOUT rising edge delay after UG1 rising edge.
2. Channel 2 Phase Shift: UG2 rising edge delay after UG1 rising edge.

On the ISL81802EVAL1Z board, the IMON2 pin is tied to 5V and EN/UVLO2 is tied to EN/UVLO1, which leads to a default 60° CLKOUT Phase Shift.

### 4. Typical Performance Curves

$V_{IN} = 48V$ ,  $T_A = 25^{\circ}C$ , unless otherwise noted.

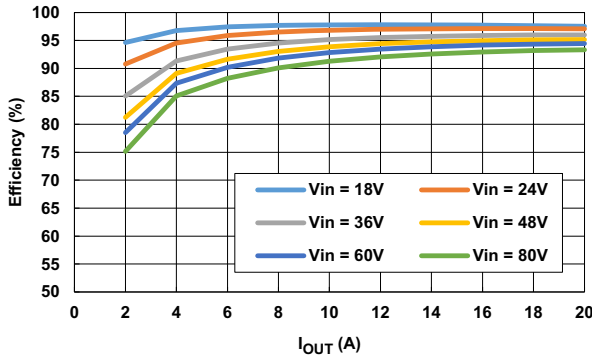


Figure 16. Efficiency, CCM

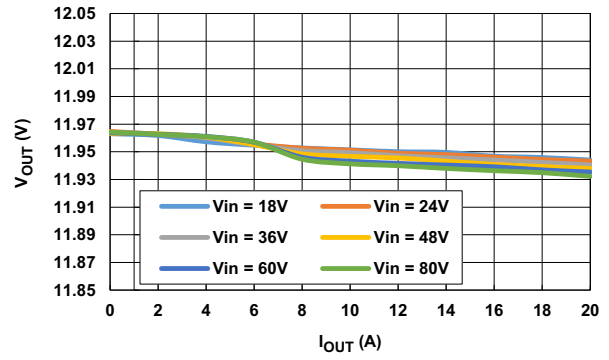


Figure 17. Load Regulation, CCM

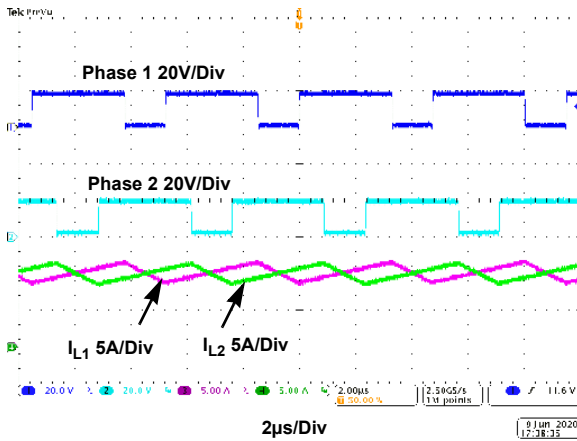


Figure 18. Phase 1, Phase 2,  $I_{L1}$ ,  $I_{L2}$ ,  $V_{IN} = 18V$ ,  $I_{OUT} = 20A$

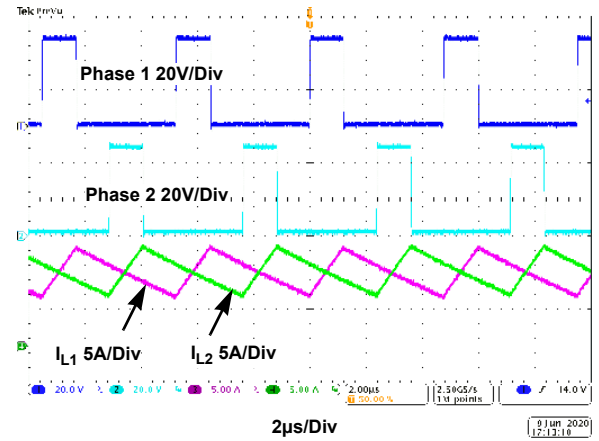


Figure 19. Phase 1, Phase 2,  $I_{L1}$ ,  $I_{L2}$ ,  $V_{IN} = 48V$ ,  $I_{OUT} = 20A$

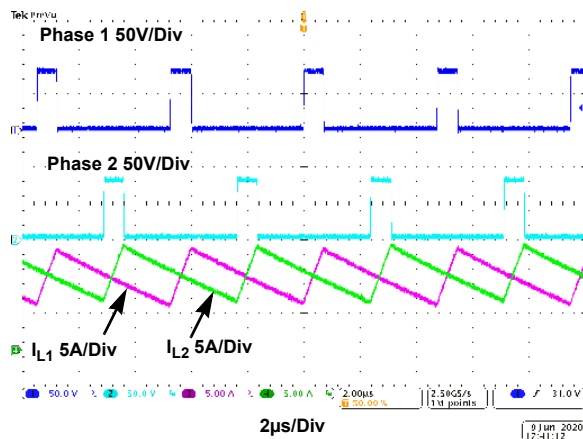


Figure 20. Phase 1, Phase 2,  $I_{L1}$ ,  $I_{L2}$ ,  $V_{IN} = 80V$ ,  $I_{OUT} = 20A$

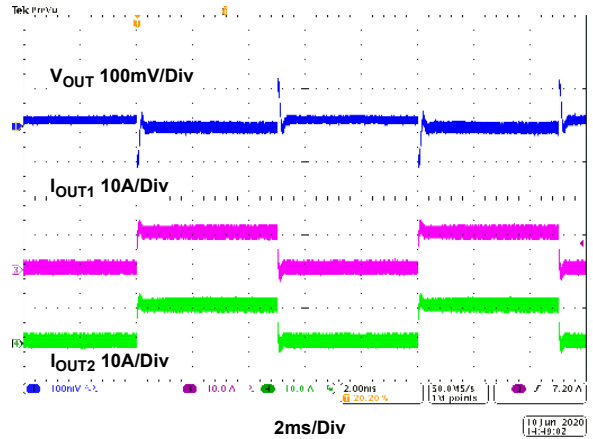


Figure 21. Load Transient,  $V_{IN} = 18V$ ,  $I_{OUT} = 0A$  to  $20A$ ,  $2.5A/\mu s$ , CCM

$V_{IN} = 48V$ ,  $T_A = 25^{\circ}C$ , unless otherwise noted. (Continued)

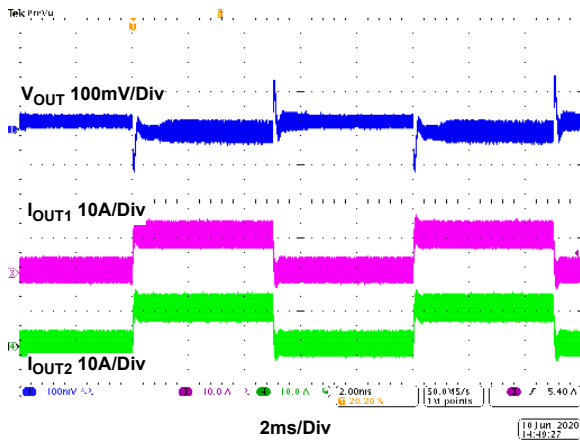


Figure 22. Load Transient,  $V_{IN} = 48V$ ,  $I_{OUT} = 0A$  to  $20A$ ,  $2.5A/\mu s$ , CCM

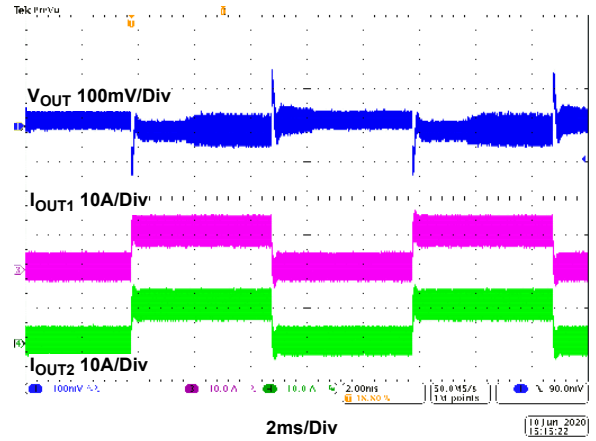


Figure 23. Load Transient,  $V_{IN} = 80V$ ,  $I_{OUT} = 0A$  to  $20A$ ,  $2.5A/\mu s$ , CCM

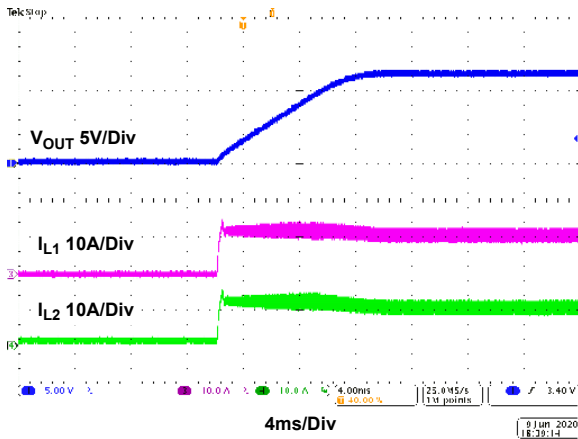


Figure 24. Start-Up Waveform,  $V_{IN} = 18V$ ,  $I_{OUT} = 20A$ , CCM

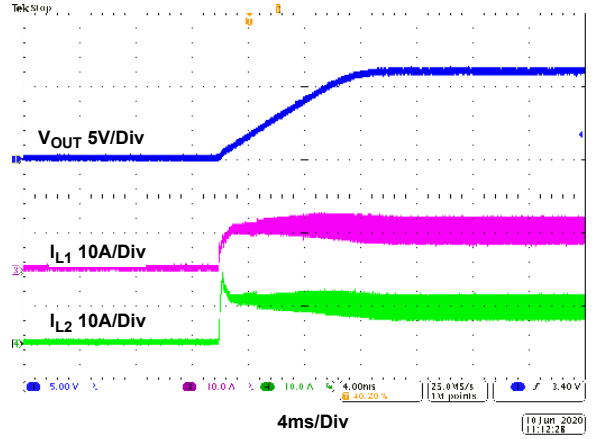


Figure 25. Start-Up Waveform,  $V_{IN} = 48V$ ,  $I_{OUT} = 20A$ , CCM

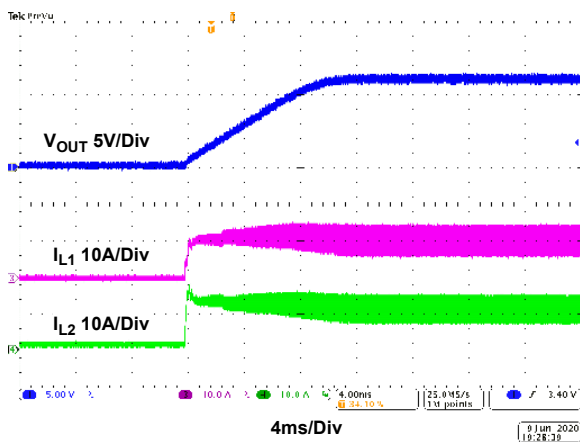


Figure 26. Start-Up Waveform,  $V_{IN} = 80V$ ,  $I_{OUT} = 20A$ , CCM

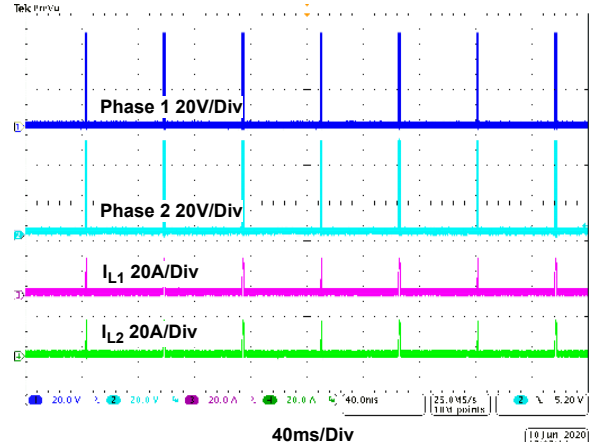


Figure 27. Short-Circuit Waveform

$V_{IN} = 48V$ ,  $T_A = 25^\circ C$ , unless otherwise noted. (Continued)

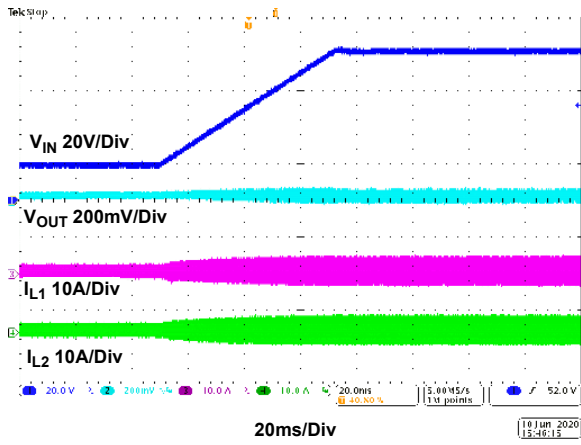


Figure 28. Line Transient,  $V_{IN} = 18V$  to  $80V$ ,  $1V/ms$ ,  $I_{OUT} = 0A$

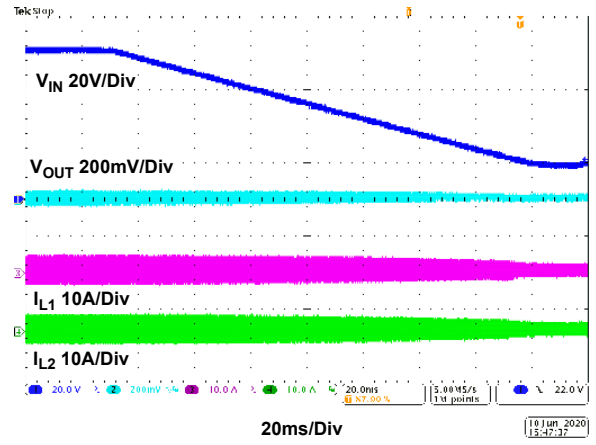


Figure 29. Line Transient,  $V_{IN} = 80V$  to  $18V$ ,  $1V/ms$ ,  $I_{OUT} = 0A$

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## 5. Revision History

Rev.	Date	Description
1.00	Aug.24.20	Initial release

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### Corporate Headquarters

TOYOSU FORESIA, 3-2-24 Toyosu,  
Koto-ku, Tokyo 135-0061, Japan  
[www.renesas.com](http://www.renesas.com)

### Contact Information

For further information on a product, technology, the most up-to-date version of a document, or your nearest sales office, please visit:  
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