TI Designs: TIDA-01253
CISPR 25 Class 4 Automotive Dual Remote and Local USB Port Reference Design

Description

This automotive media port design allows customers to accelerate media port systems by taking advantage of a complete reference design comprising analog AEC-100-qualified integrated circuits (ICs). The reference design is specified for automotive media ports that require both local-port and remote-port support utilizing the USB BC1.2 specification. This design creates a robust, flexible solution which allows the system to charge both ports simultaneously while still accounting for voltage droop on a remote port.

Resources

TIDA-01253	Design Folder
LM74610-Q1	Product Folder
LMS3655-Q1	Product Folder
TPS254900-Q1	Product Folder
TPS2511	Product Folder
LM358	Product Folder
CSD16408Q5C	Product Folder

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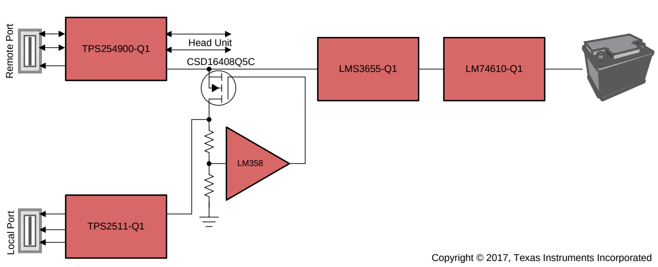
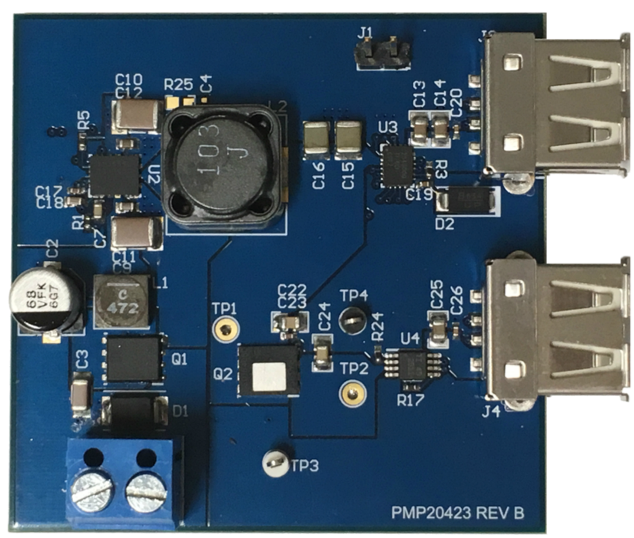
Features

- LM74610-Q1 Smart Diode Emulates Ideal Diode Rectifier and Protects Downstream Devices in Case of Reverse Polarity
- TPS254900 and TPS2511 Provide Remote Port and Local Port Negotiation Over D+/D– Lines, Respectively, to Enable Charging of Either Cable Compensated or Typical Local Port
- Components in System Provide Short-to-Battery Protection and Load Dump Protection Against Input Transients up to 40 V
- Automotive Grade LMS3635-Q1 Optimizes Power Supply Through Adjustable Switching Frequencies Outside of AM and FM Band Operation and Supports Both Ports by Providing up to 5.5-A Continuous Current Without Requiring External Diodes or MOSFETs

Applications

- Automotive Head Units with Remote Displays
- Remote Media Hub





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1 System Overview

1.1 System Description

This system is designed to allow a dual-USB type-A port from a head unit to provide cable compensation to a remote port while still operating within tight USB specifications on the local port. One port is a dedicated charging port (DCP) and the other is a charging downstream port (CDP) that allows the USB 2.0 data to pass to the head unit. A common challenge when trying to design for both a local and remote port using one input voltage is when the feedback loop for the remote port adjusts the voltage to compensate for the voltage drop across the long cable. This adjustment puts the local port voltage outside the required tolerances. To overcome this issue, use a very-low dropout regulator to maintain the local port regulation.

1.2 Key System Specifications

This design is specified to use a car battery input between 5.5 V and 36 V with a nominal voltage typically around 12 V to 14.5 V. Both USB type-A outputs ultimately support 5-V nominal; however, this design was created specifically such that the local port is within ±2.5% of 5 V and the remote port is within ±2.5% of 5 V after 2 m to 3 m of cable. The following Table 1 presents the key specifications for this design.

Table 1. Key System Specifications

PARAMETER	COMMENTS	MIN	TYP	MAX	UNIT
SYSTEM OPERATION					
V_{IN}	Input voltage of system	5.5	14.5	36	V
F_{sw}	Switching frequency of DC-DC	—	400	—	kHz
$R_{DS(ON)\text{-Remote}}$	ON-resistance of power switch (TPS254900)	—	50	—	mΩ
$R_{DS(ON)\text{-local}}$	ON-resistance of power switch (TPS2511)	—	70	120	mΩ
$I_{OS\text{-Remote}}$	Short-circuit current limit of TPS254900	2.42	2.59	2.760	A
$I_{OS\text{-Local}}$	Short-circuit current limit of TPS2511	2.11	2.3	2.5	A
OUTPUT VOLTAGES					
V_{Remote}	Remote port voltage	5.081	5.087	5.092	V
V_{Local}	Local port voltage	4.875	5.00	5.125	V
V_{BUS}	Output of LM73605 with cable compensation adjustment	5.0	5.1	5.6	V
OUTPUT CURRENT					
$I_{OUT\text{-Remote}}$	VBUS load current of remote port	—	2.4	—	A
$I_{OUT\text{-Local}}$	VBUS load current of local port	—	2.4	—	A

1.3 Block Diagram

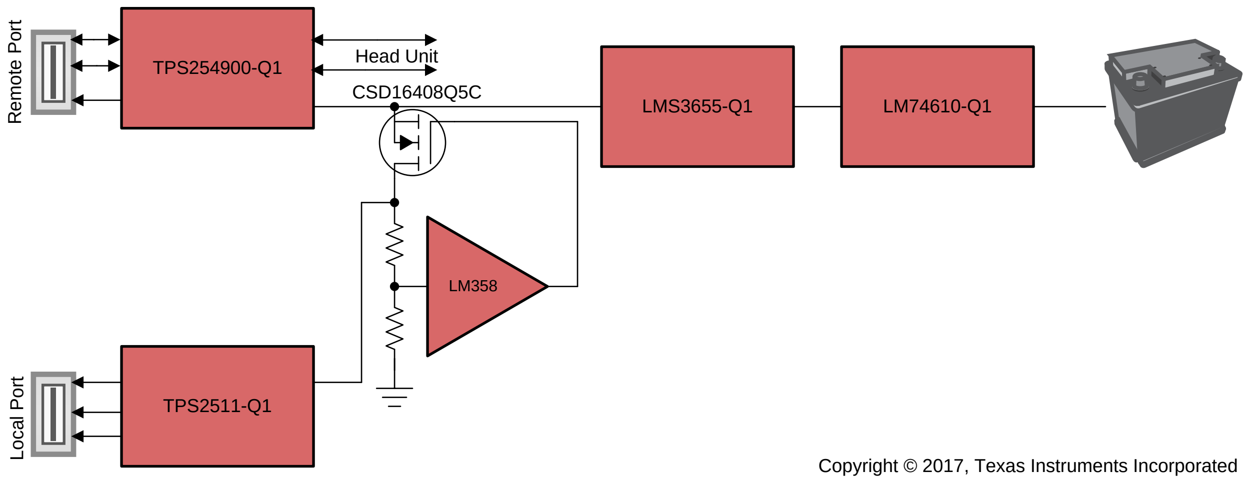

Figure 1. TIDA-01253 Block Diagram

1.4 *Highlighted Products*

1.4.1 LM74610-Q1 Zero I_Q Reverse-Polarity Protection Smart-Diode Controller

The LM74610-Q1 (Figure 2 and Figure 3) is a controller device that can be used with an N-channel MOSFET in a reverse-polarity protection circuit. The device has been designed to drive an external MOSFET to emulate an ideal diode rectifier when connected in series with a power source. A unique advantage of this scheme is that it is not referenced to ground and thus has zero I_Q.

The LM74610-Q1 controller provides a gate drive for an external N-channel MOSFET and a fast-response internal comparator to discharge the MOSFET gate in the event of reverse polarity. This fast pulldown feature limits the amount and duration of reverse current flow if the opposite polarity has been sensed. The device design also meets CISPR25 Class-5 electromagnetic interference (EMI) specifications and automotive ISO7637 transient requirements with a suitable transient voltage suppressor (TVS) diode.

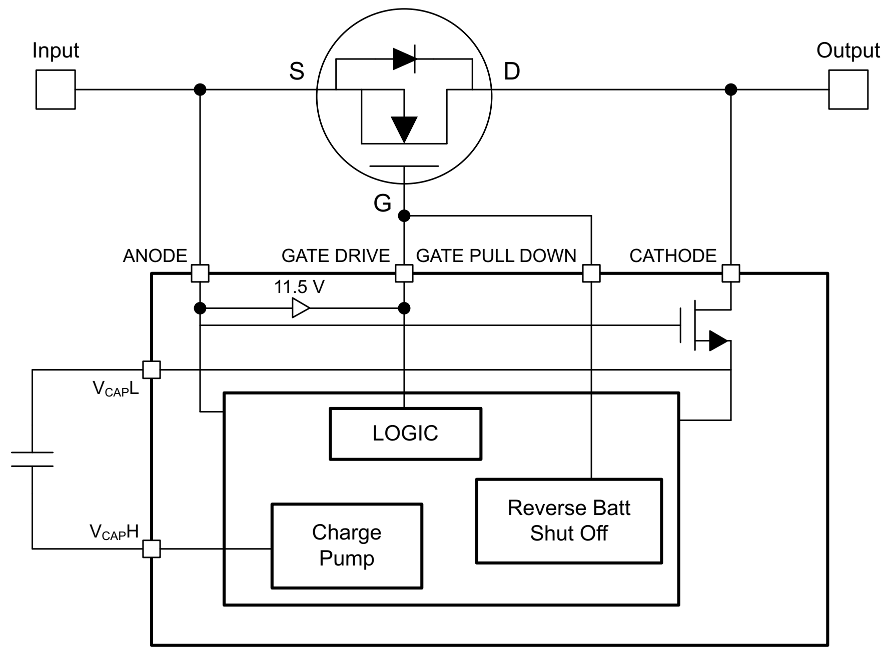

Figure 2. LM74610-Q1 Functional Block Diagram

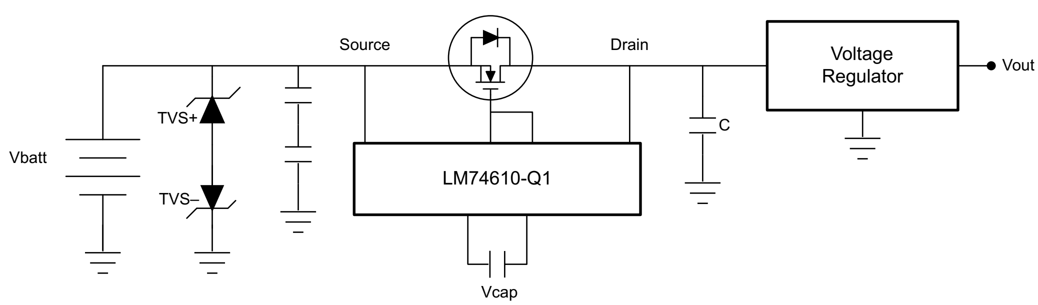

Figure 3. LM74610-Q1 Simplified Schematic

1.4.2 LMS3635-Q1 to 36-V, 5.5-A, 400-kHz Synchronous Step-Down Voltage Converter

The LMS3655 synchronous buck regulators are optimized for high-performance applications, providing an adjustable output of 1 V to 15 V. Seamless transition between PWM and PFM modes, along with a low quiescent current, ensures high efficiency and superior transient responses at all loads. Advanced high-speed circuitry allows the LMS3655 to regulate an input of 24 V to an output of 3.3 V at a fixed frequency of 400 kHz while also enabling a continuous load current of 5.5 A. An innovative frequency foldback architecture allows this device to regulate a 3.3-V output from an input voltage of only 3.5 V. The input voltage can range up to 36 V, with transient tolerance up to 42 V, easing input surge protection design. An open-drain reset output, with built-in filtering and delay, provides a true indication of system status. This feature negates the requirement for an additional supervisory component, saving cost and board space.

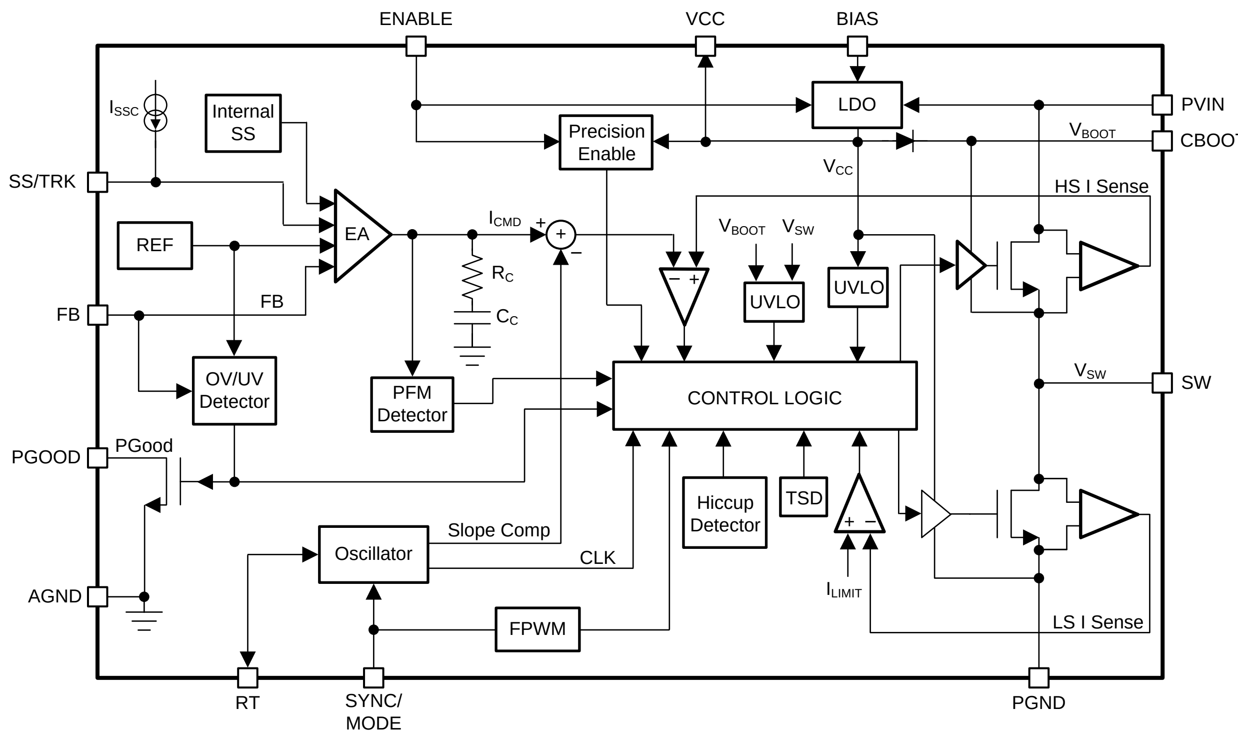

Figure 4. LMS3655-Q1 Functional Block Diagram

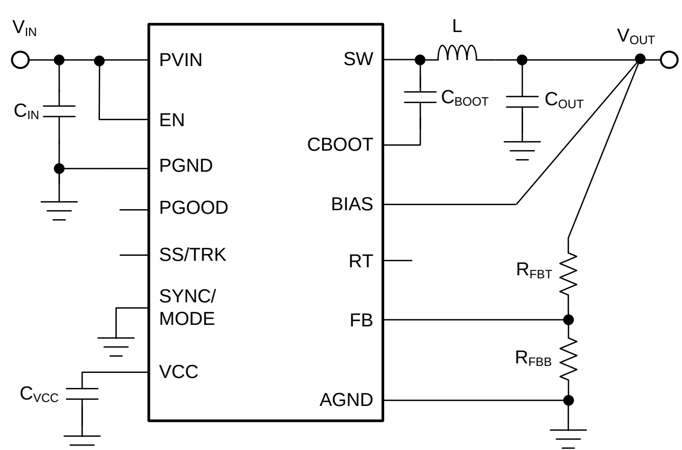

Figure 5. LMS3655-Q1 Simplified Schematic

1.4.3 TPS254900-Q1 Automotive Charging Port Controller and Power Switch With Short-to-Battery Protection

The TPS254900-Q1 (Figure 6 and Figure 7) is a USB charging port controller and power switch with short to-battery protection. The TPS254900-Q1 provides protection on the OUT, DM_IN, and DP_IN pins. These three pins withstand voltages up to 18 V. Upon a short-to-battery incident, an internal MOSFET quickly disconnects the pin. This feature provides protection to the power supply of the TPS254900-Q1, the upstream processor, or the upstream HUB.

The TPS254900-Q1 50-mΩ power switch has two selectable, programmable current limits that support port power management by providing a lower current limit that can be used when adjacent ports experience heavy loads. This feature is important in a system with multiple ports and the upstream supply is unable to provide full current to all ports simultaneously.

The TPS254900Q1 has a current-sense output that is able to control an upstream supply. This output control allows it to maintain 5 V at the USB port even during heavy charging currents. This feature is important in systems with long USB cables where significant voltage drops can occur while fast-charging portable devices. The unique IMON feature allows the system to register the load current by monitoring the IMON voltage and is used for dynamic port-power management.

To save space in the application, the TPS254900-Q1 also integrates IEC 61000-4-2 compliant protection from electrostatic discharge (ESD).

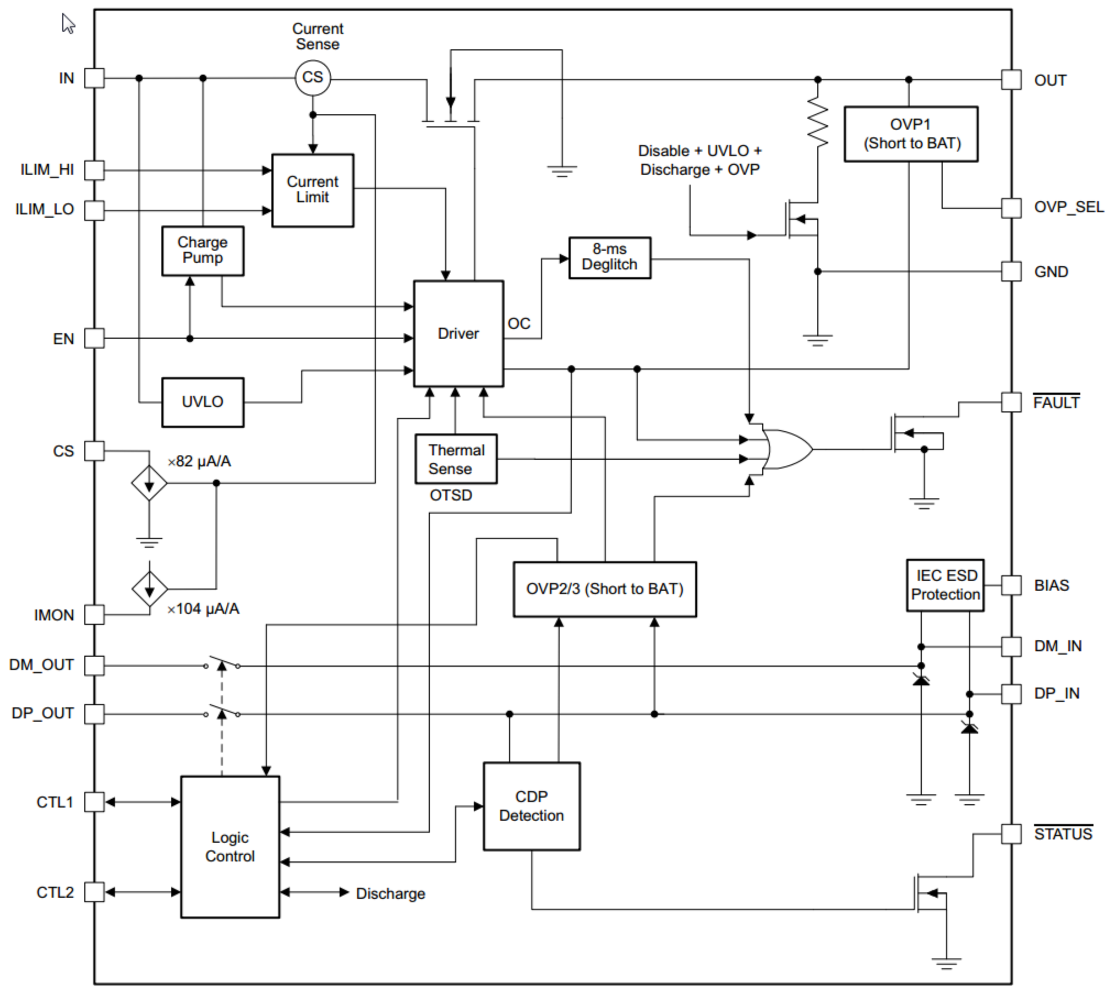

Figure 6. TPS254900-Q1 Functional Block Diagram

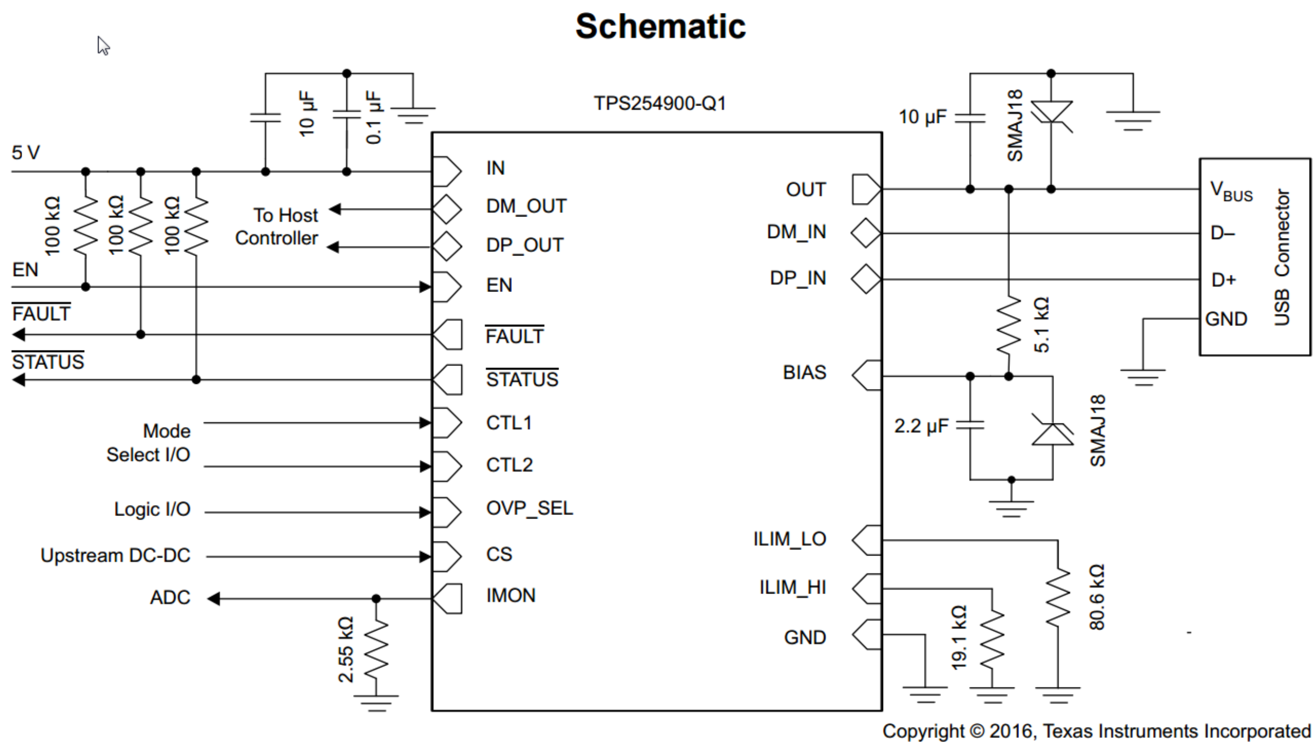

Figure 7. TPS254900-Q1 Simplified Schematic

1.4.4 TPS2511-Q1 Automotive Catalog, USB-Dedicated, Charging-Port Controller and Current-Limiting Power Switch

The following overview references various industry standards. TI always recommends to consult the latest standard to ensure the most recent and accurate information. Rechargeable portable equipment requires an external power source to charge its batteries. USB ports are convenient locations for charging because of an available 5-V power source. Universally-accepted standards are required to ensure host and client-side devices meet the power management requirements. Traditionally, USB host ports following the USB 2.0 specification must provide at least 500 mA to downstream client-side devices. Because multiple USB devices can be attached to a single USB port through a bus-powered hub, the client-side device is responsible for negotiating the power allotment from the host to guarantee the total current draw does not exceed 500 mA. The TPS2511-Q1 device provides 100 mA of current to each USB device (see Figure 8 and Figure 9). Each USB device can subsequently request more current, which is granted in steps of 100 mA up to 500 mA total. The host may grant or deny the request based on the available current. Additionally, the success of the USB technology makes the micro-USB connector a popular choice for wall adapter cables. This type of connection allows a portable device to charge from both a wall adapter and USB port with only one connector; however, this results in a common dilemma. With the gaining popularity of USB charging, the 500-mA minimum defined by the USB 2.0 specification (or 900 mA defined in the USB 3.0 specification) has become insufficient for many handsets, tablets, and personal media players (PMP), which have a higher-rated charging current. Wall adapters and car chargers can provide much more current than 500 mA or 900 mA to fast charge portable devices. Several new standards have been introduced which define protocol handshaking methods that allow host and client devices to acknowledge and draw additional current beyond the 500-mA (defined in the USB 2.0 specification) or 900-mA (defined in the USB 3.0 specification) minimum while using a single micro-USB input connector.

The TPS2511-Q1 supports three of the most common protocols:

- USB battery charging specification, revision 1.2 (BC1.2)
- Chinese Telecommunications Industry Standard YD/T 1591-2009
- Divider mode

These protocols utilize three types of charging ports which are defined to provide different charging current to client-side devices. These charging ports are defined as:

- Standard downstream port (SDP)
- Charging downstream port (CDP)
- Dedicated charging port (DCP)

The BC1.2 specification defines a charging port as a downstream-facing USB port that provides power for charging portable equipment.

The following Table 2 shows different port operating modes according to the BC1.2 specification.

Table 2. Operating Modes Table

PORT TYPE	SUPPORTS USB 2.0 COMMUNICATION	MAXIMUM ALLOWABLE CURRENT DRAWN BY PORTABLE EQUIPMENT
SDP (USB 2.0)	Yes	0.5 A
SDP (USB 3.0)	Yes	0.9 A
CDP	Yes	1.5 A
DCP	No	1.5 A

The BC1.2 specification defines the protocol necessary to allow portable equipment to determine what type of port it is connected to so that it can allot its maximum allowable current drawn. The handshaking process consists of two steps. During step one, the primary detection, the portable equipment outputs a nominal 0.6-V output on its D+ line and reads the voltage input on its D– line. The portable device concludes that it is connected to the charging port if the D– voltage is greater than the nominal data-detect voltage of 0.3 V and less than 0.80 V.

The second step, the secondary detection, is necessary for portable equipment to determine between a CDP and DCP. The portable device outputs a nominal 0.6-V output on its D– line and reads the voltage input on its D+ line. The portable device concludes it is connected to a CDP if the data line being read remains less than the nominal data detect voltage of 0.3 V. The portable device concludes it is connected to a DCP if the data line being read is greater than the nominal data-detect voltage of 0.3 V and less than 0.8 V.

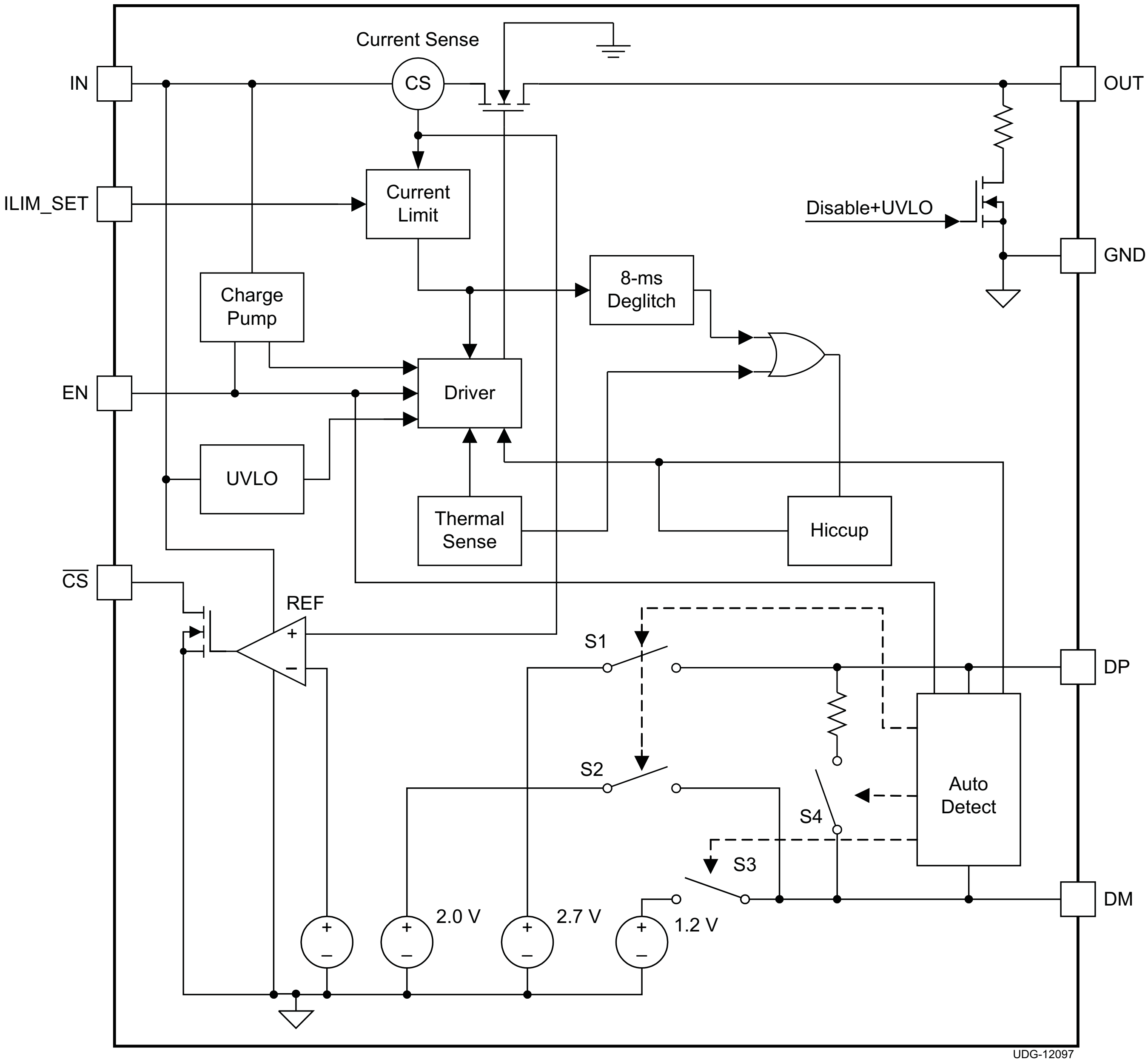

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Figure 8. TPS2511-Q1 Functional Block Diagram

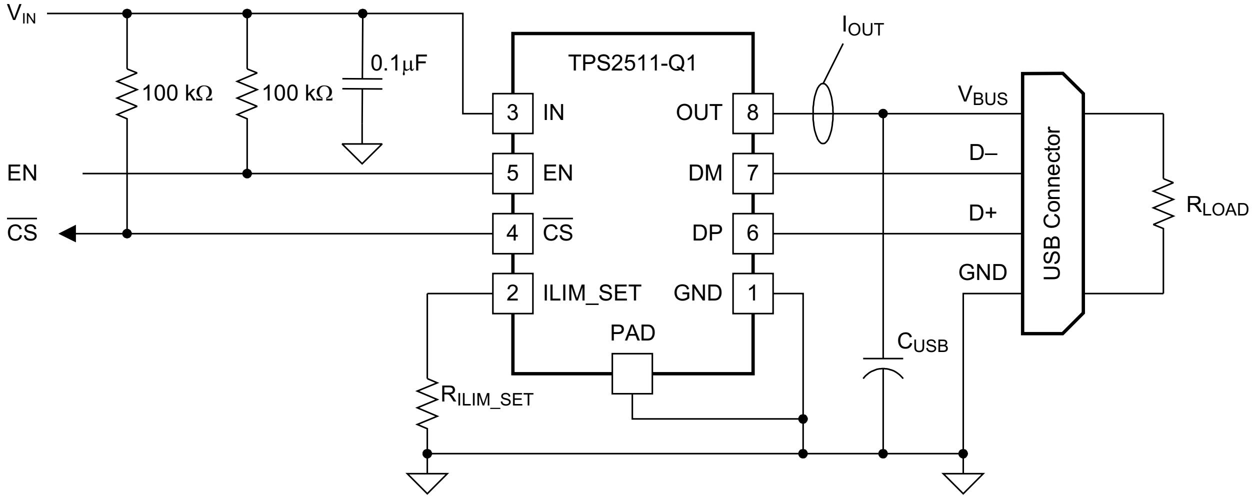


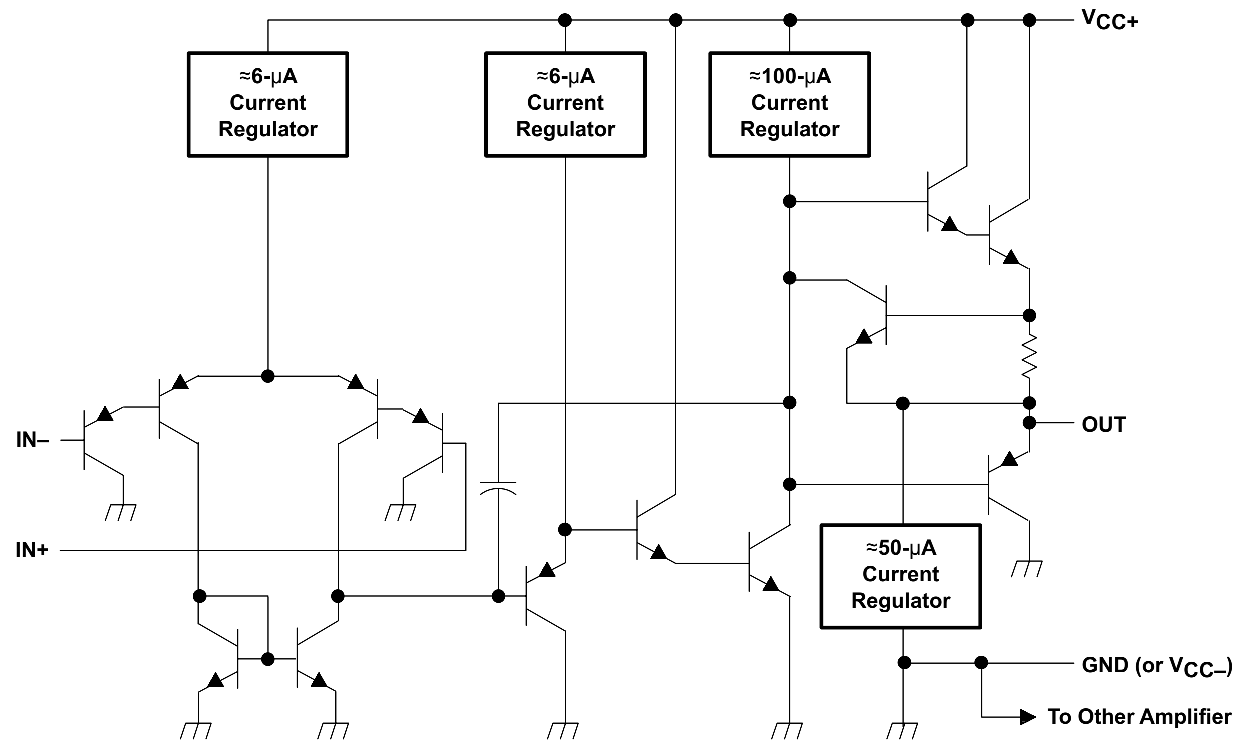
Figure 9. TPS2511-Q1 Simplified Schematic

1.4.5 LM358 Dual Operational Amplifiers

These devices consist of two independent, high-gain frequency-compensated operational amplifiers (op amps) designed to operate from a single supply over a wide range of voltages. Operation from split supplies is also possible if the difference between the two supplies is 3 V to 32 V (3 V to 26 V for the LM2904 device) and VCC is at least 1.5 V more positive than the input common-mode voltage. The low supply-current drain is independent of the magnitude of the supply voltage.

Applications include transducer amplifiers, DC amplification blocks, and all the conventional op-amp circuits that can now be implemented more easily in single-supply voltage systems. For example, these devices can be operated directly from the standard 5-V supply used in digital systems and can easily provide the required interface electronics without additional ±5-V supplies.

Figure 10 shows the functional block diagram of the LM358 and Figure 11 shows the simplified schematic.



COMPONENT COUNT	
Epi-FET	1
Diodes	2
Resistors	7
Transistors	51
Capacitors	2

Figure 10. LM358 Functional Block Diagram

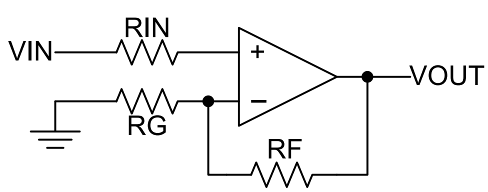


Figure 11. LM358 Simplified Schematic

1.4.6 CSD16408Q5C DualCool™ N-Channel NexFET™ Power MOSFET

The NexFET™ power MOSFET has been designed to minimize losses in power conversion applications with ultra-low Q_g and Q_{gd}. While utilizing the DualCool™ packaging, the devices are optimized for two-sided cooling and have low thermal resistance, all while being avalanche rated (see Figure 12).

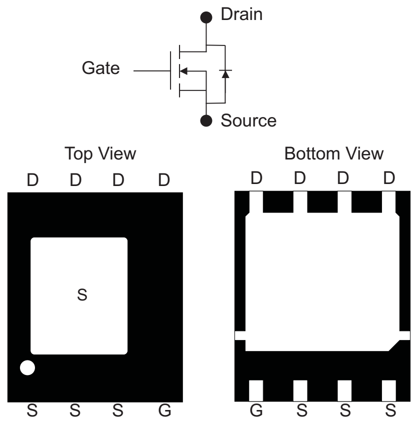


Figure 12. CSD16408Q5C Package and Symbol

2 System Design Theory

2.1 Introduction

The TIDA-01253 reference design has been designed to support two USB type-A ports: one port operating in CDP mode and the other operating in DCP mode as specified by USB BC1.2 specification (see Figure 13). The purpose of this design is to show dual-port functionality while using one power supply. The challenge, however, is when supporting a remote port, cable compensation is required due to the voltage drop across wires. This challenge requires additional voltage to compensate for the loss; however, the difficulty is providing an extra voltage for the remote port while also maintaining adequate voltage for the local port. The following subsections address this issue. As Figure 1 shows, the TIDA-01253 is a media interface subsystem that could be integrated in a head unit while supporting a remote port.

This TI Design has been specifically designed to meet automotive standards and uses parts that are already released or, at the time of this writing, planned to be released as automotive Q-100 grade-qualified. When designing a subsystem for automotive infotainment, satisfying relevant protection standards as required by automotive original-equipment manufacturers (OEMs) is important. In this TI Design, parts have been selected to support typical protection requirements such as reverse battery protection, current limiting, short-to-battery protection, as well as ESD protection. Additionally, power supplies typically require a spread spectrum to mitigate EMI and switching frequency outside of AM band operation while supporting wide input off-battery operation. In the circuit board layout (see Figure 1), care has been taken to improve thermal dissipation and EMI to ultimately pass CISPR-25 class-5 standards. The following sections explain each system subsection.

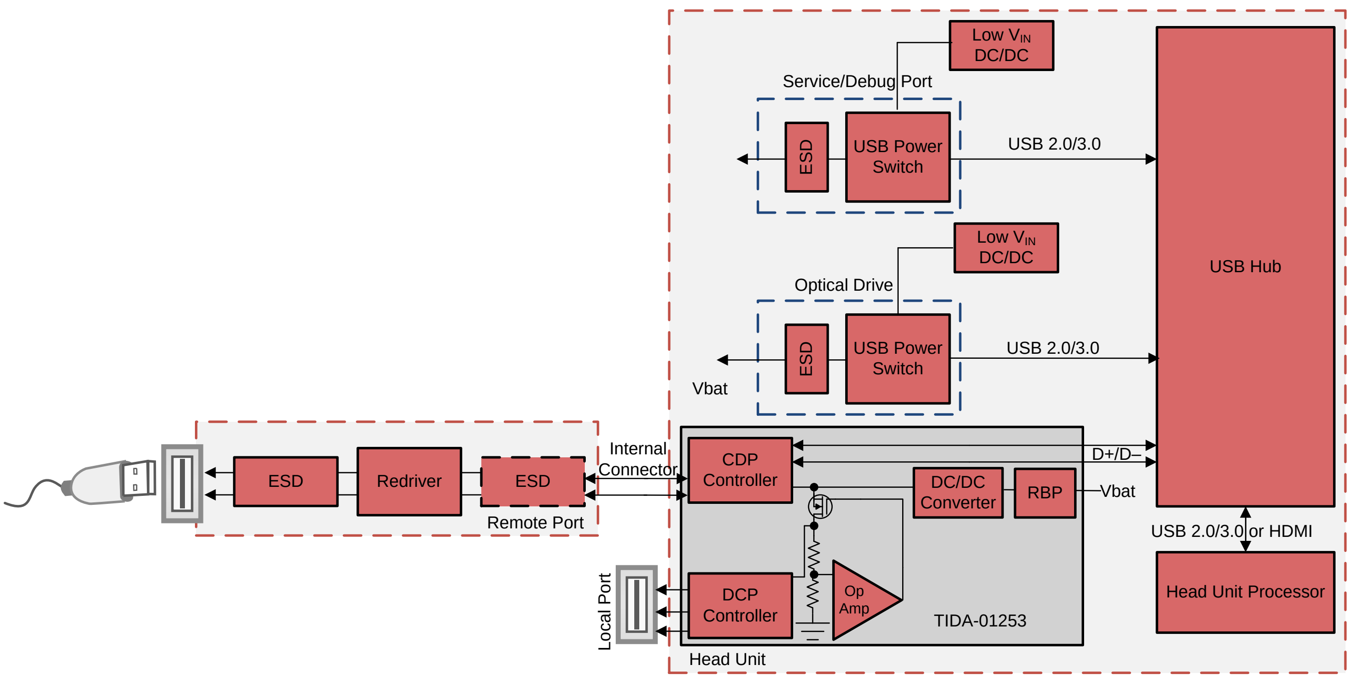


Figure 13. System Integration Concept

2.2 Power Supply

2.2.1 Reverse Battery Protection

The reverse battery protection that Figure 14 shows is a requirement for every electronic subsystem in a vehicle recognized by OEM standards as well as load dump protection standards ISO 16750-2. The LM74610 device is used to control the NFET to protect the load in a negative polarity condition. This device is used to emulate an ideal diode by using an NFET in series with the battery supply. This configuration has the advantage of a highly effective and efficient substitute for reverse battery protection to the traditional rectifier. The LM74610 has no power pin and no ground reference which means that it requires zero I_Q and reduces standby current drawn from the battery. Additionally, the voltage drop across the field-effect transistor (FET) is so inconsequential that it allows the wide V_{IN} buck regulator to operate at even lower battery input voltages. This feature is an advantage for scenarios such as cold crank when the battery voltage temporarily drops to as low as 3.5 V. A traditional diode solution usually has a 700-mV voltage drop, so the buck converter is incapable of maintaining the 5-V_{OUT} system voltage. With the smart diode solution, the buck still receives close to 5 V during this condition and can continue to operate.

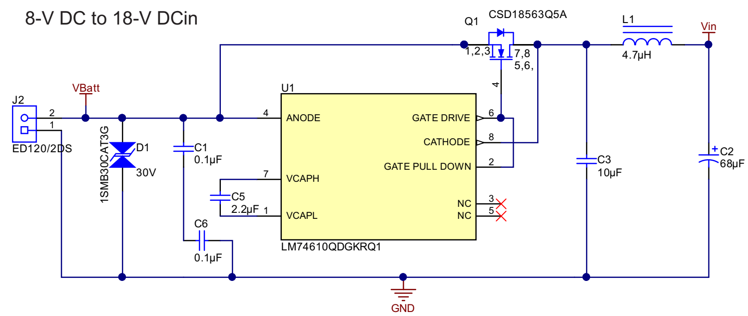


Figure 14. Schematic of Reverse Battery Protection

2.2.1.1 MOSFET

The LM74610-Q1 product folder has tools to help simulate using the WEBENCH® Design simulator tool. The data sheet also outlines MOSFET selection in the LM74610-Q1 data sheet [1]. Some important characteristics that the data sheet recommends to design around are:

- Continuous drain current (I_D)
- Maximum drain-to-source voltage ($V_{DS(MAX)}$)
- Gate-to-source threshold voltage ($V_{GS(TH)}$)
- Drain-to-source ON-resistance ($R_{DS(ON)}$)

The rating for I_D must exceed the load current, which is 6 A t ($I_D > I_{LOAD}$) in this system, and the body diode maximum current $I_S \geq I_D$. Also, the gate-to-source threshold voltage must be $V_{GS(TH)} \leq 3$ V. The NFET outlined in the following Table 3 has been selected for this application.

Table 3. MOSFET Parameters

AUTOMOTIVE N-CHANNEL 40-V (D-S) 175°C MOSFET (SQJ422EP)					
PARAMETER	COMMENTS	MIN	TYP	MAX	UNIT
I_D	Drain-to-source current	—	—	75	A
V_{DS}	Drain-to-source voltage	—	—	40	V
$V_{GS(TH)}$	Gate-to-Source voltage	1.5	2.5	2.5	V
$R_{DS(ON)}$	Drain-to-source ON resistance	—	4.3	—	mΩ

2.2.1.2 Input Capacitors

Due to the possibility of this subsystem PCB flexing, the likelihood of mechanical short increases. If a mechanical shortdoes occur, then a short transpires between the two terminals of the battery. To mitigate any chance of this short, the two input capacitors are placed in series but rotated 90º to one another in case the vibration in an automotive system causes one capacitor to crack. Some component manufacturers offer options for an integrated solution using automotive-grade multilayer ceramic capacitors (MLCC); but for this solution, discrete components have been used.

2.2.1.3 TVS Diodes

Although the transient voltage suppression (TVS) diodes used in this TI Design are not required by the LM74610-Q1 (as outlined in the LM74610-Q1 data sheet [1]), they are used to clamp the positive and negative voltage surges that may occur in the input. These transients are outlines in the ISO specification ISO 7637-2:2004 pulses 1 and 2a. An LC filter is also used downstream. This TI Design focuses on maintaining a fully functional system while transients may be present because this system must remain in operation during such transient conditions. Two specifications that must be considered are breakdown voltage and clamping voltage:

- Breakdown voltage is the voltage when the TVS diode goes into an avalanche similar to a Zener diode and is specified at a low-current value (typically 1 mA).
- Clamping voltage is the voltage which the TVS diode clamps to in high-current pulse situations.

Table 4 shows these parameters for the selected diodes.

Table 4. TVS Diode Parameters

600-W PEAK-POWER ZENER TRANSIENT VOLTAGE SUPPRESSOR BIDIRECTIONAL (1SMB30CAT3G)					
PARAMETER	**COMMENTS**	**MIN**	**TYP**	**MAX**	**UNIT**
V_{RWM}	Reverse standoff voltage	—	30	—	V
V_C	Clamping voltage	—	48.8	—	V
V_{BR}	Breakdown voltage	33.3	35.06	36.8	V

Choose the diode breakdown voltage such that the transients are clamped at the voltages that protect the MOSFET and the rest of the system. D1 clamps above the double battery (jump-start) and clamped load-dump voltages, but lower than the maximum operating voltage of the downstream devices. In this case, the diode starts to clamp around 30 V but has a maximum clamp voltage just below 48 V. The ideall voltage is somewhere around 35 V, which is why D1 has been chosen for its maximum clamping voltage. In regard to power levels for TVS diodes, the particular package used is SMBJ, which supports 600-W peak-power levels. This package is sufficient for ISO 7637-2 pulses and suppressed load-dump case (ISO-16750-2 pulse B). For unsuppressed load dumps (ISO-16750-2 pulse A), higher power TVS diodes such as SMCJ or SMDJ may be required. Refer to the LM74610-Q1 data sheet [1] for more information about designing the TVS diodes for this application

2.2.2 DC-DC Converter

For this design, the LMS3655 has been chosen for its optimization for automotive applications. In this particular application, this device was used for its 5-A current capability. However, the LMS3655 also meets many of the requirements for automotive applications such as:

- Wide V_{IN} operation with transient tolerance up to 42 V
- Wettable flanks
- Switching frequency outside AM band
- Low EMI and switch noise
- External frequency synchronization
- AEC-Q100 automotive qualified

Figure 15 shows the DC-DC schematic of the LMS3655.

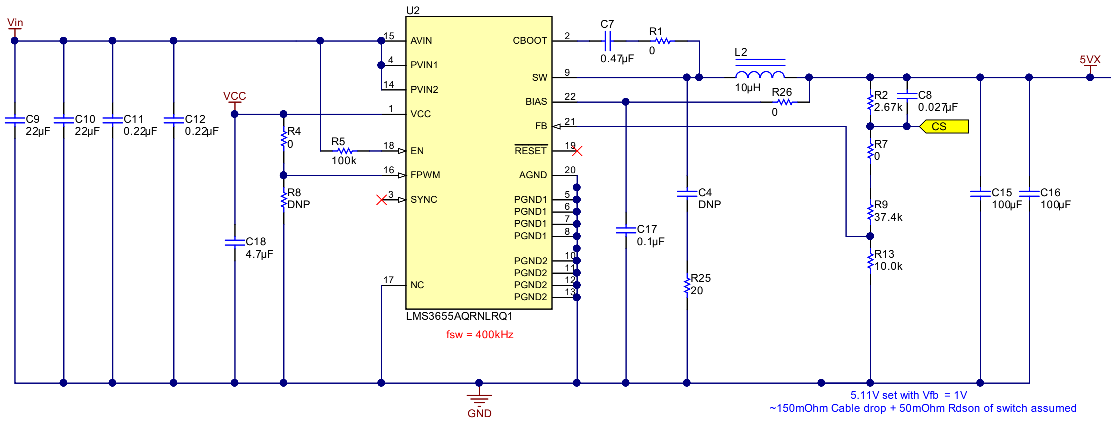


Figure 15. DC-DC Schematic

Table 5. Spreadsheet Parameters of Power Supply Design

PARAMETER	COMMENTS	MIN	TYP	MAX	UNIT
System Operation					
V_{IN}	Input voltage	5.5	14.5	36	V
F_{SW}	Switching frequency	—	400	—	kHz
Output Voltages					
V_{Remote}	Remote port voltage	5.081	5.087	5.092	V
V_{Local}	Local port voltage	4.875	5.00	5.125	V
V_{BUS}	Output of LM73605 with cable compensation adjustment	5.0	5.1	5.6	V
Output Current					
$I_{OUT-Remote}$	VBUS load current of remote port	—	2.4	—	A
$I_{OUT-Local}$	VBUS load current of local port	—	2.4	—	A
Component Values					
L_1	Inductor	—	10	—	µH
C_{IN}	Input capacitance	—	44.4	—	µF
C_{OUT}	Output capacitance	—	200	—	µF

2.2.2.1 Input Capacitors

Ceramic capacitors are typically used at the input of a power supply because they provide a low-impedance source to the regulator in addition to supplying ripple current and isolating switching noise from other circuits. Because the input of the regulator can encounter voltages up to 40 V, the rating of the capacitors must be able to accommodate these voltages (see Table 6). In addition, small high-frequency bypass capacitors connected directly between the VIN and GND pins are used to reduce noise spikes and aid in reducing conducted EMI. Additional high-frequency capacitors can be used to help manage conducted EMI or voltage spike issues that may be encountered. The following calculations help determine the right input capacitance to use.

Table 6. Input Capacitor Selection

NOMINAL CAPACITANCE	VOLTAGE RATING	TYPE	PART NUMBER
2x22 µF	25 V	X7R	C4532X7R1E226M250KC
0.22 µF	50 V	X5R	GRM155R61H104ME14D

NOTE: Following the recommendations from the DC-DC regulator is important; however, following the correlation of the input filter is also important to pass CISPR-25 EMI testing. The following calculations describe how the LC filter was calculated with respect to the system.

As Figure 16 shows, the following parameters must be identified to calculate the ideal component selection.

- Board I/Os: Input voltage (V_{IN}), output voltage (V_{OUT}), and output current (I_{OUT})
- Estimated efficiency
- Switching frequency (F_{SW})
- Filter inductor (L_f)
- Input capacitance (C_{IN}) and damping capacitor (C_d) equivalent series resistance (ESR)
- Maximum noise (V_{MAX})

With these parameters defined, the following calculations are conducted in Equation 1:

$$\text{Target attenuation} = 20 \times \text{LOG}(I_{OUT} / \pi^2 \times F_{SW} \times 1000 \times C_{IN} \times 1e - 6) \times \text{SIN} \times (\pi \times D) / 1e - 6) - V_{MAX} \quad (1)$$

where,

- Duty cycle D = V_{OUT} / V_{IN}.

After calculating the target attenuation, the filter capacitor (C_f) can be calculated. Two options are available for calculating the C_f and the greater of the two options in Equation 2 and Equation 3 provides the value to select.

$$C_{fA} = C_{IN} \times 1e - 6 / (C_{IN} \times 1e - 6 \times L_f \times 1e - 6 \times (2 \times \pi \times F_{SW} \times 1000 / 10)^2 - 1) / 1e - 6 \quad (2)$$

$$C_{fB} = 1 / (L_f \times 1e - 6) \times (10\text{\^{}}(TA / 40) / (2 \times \pi \times F_{SW} \times 1000))^2 / 1e - 6 \quad (3)$$

The two equations calculate to C_{fA} = 3.8 µF and C_{fB} = 27.7 µF, respectively. The 4.7-µF value has been chosen for this design.

For C_d, the damping capacitor is recommended to be $C_d \geq 4 \times C_{IN}$, where C_{IN} is the input capacitance to the DC-DC and thus determined by the input ripple required for the switcher. This equation produces a rather large C_d of 108 µF for this particular design. A standard high-frequency capacitor was chosen but note that adjusting these values based on the table in the following Figure 16 can help change the impedance at the switching frequency and thus troubleshoot any EMI issues that may arise.

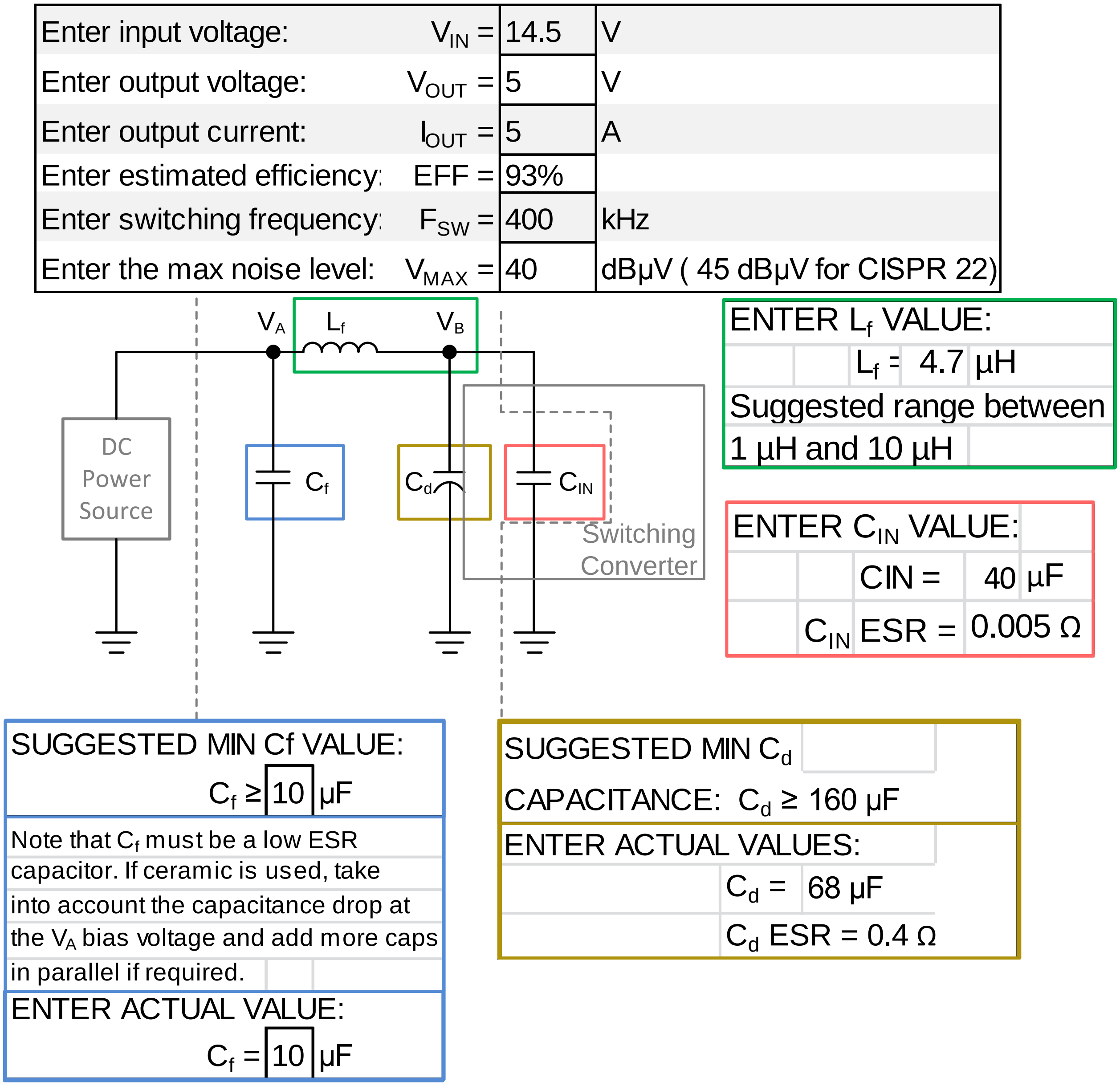


Figure 16. LC Filter Calculation

2.2.2.2 DC-DC Inductor Selection

There are two necessary criteria for selecting an output inductor:

- Inductance
- Saturation current rating

Inductance is usually calculated based on the desired peak-to-peak ripple current, ΔIL, that flows in the inductor along with the load current. Two tradeoffs exist for this selection:

- Higher inductance means lower ripple current and lower ripple voltage at the output; however, this usually leads to a higher inductor cost.
- Lower inductance means a smaller package size and thus a more cost-effective device.

An estimate for peak-to-peak ripple current through the inductor should start between 0.2 to 0.4 of the maximum output current. From this range, the minimum inductance can be calculated in Equation 4 from this value:

$$L_{min} = (V_{IN(Max)} - V_{OUT}) / (f_{sw} \times \Delta IL)) \times D \quad (4)$$

where,

- $V_{IN(MAX)}$ = 14.5,
- V_{OUT} = 5.11,
- D = 0.352,
- $F_{SW\ =\ 400\ kHz}$,
- ΔIL = 0.4

For this design, L_{MIN} = 4.128 µH and a standard 10 µH has been chosen for cost optimization.

The second criterion is inductor saturation current rating. The LMS3655 has an accurately-programmed valley current limit. During an instantaneous short, the peak inductor current can be very high due to a momentary increase in duty cycle. Because this current is limited by the high-side switch current limit, TI advises to select an inductor with a saturation current close to the high-side current limit, which is approximately 8 A; therefore, a saturation current can be chosen to match this value.

2.2.2.3 Output Voltage and Capacitor Selection

The output capacitor is responsible for filtering the output voltage and supplying load current during transients. Capacitor selection depends on application conditions as well as ripple and transient requirements. Best performance is achieved by using ceramic capacitors or combinations or ceramic and polymer capacitors. For high-output voltage conditions, such as 12 V and above, finding ceramic caps that are rated for an appropriate voltage can be challenging. In such cases, a low ESR SP™ or POSCAP™ type capacitor should be chosen. For those high V_{OUT} conditions, a low-value ceramic capacitor can be introduced in parallel so as to reduce the output ripple and noise spikes, while a higher-value electrolytic or polymer cap provides large bulk capacitance to supply transients. For a given input and output requirement, the following inequality in Equation 5 provides an approximation for a required absolute minimum output cap:

$$Cout > 1 / (fsw \times \Delta Vo / Iout \times r) \times ((r\text{^}2 / 12 \times (1+D')) + (D \times (1+r))) \quad (5)$$

where,

- R = Ripple ratio of the inductor ripple current (ΔIL / Iout),
- ΔV_O = Target output voltage undershoot,
- D = 1 – duty cycle,
- F_{SW} = Switching frequency,
- I_{OUT} = Load current.

Using low ESR output capacitors in low-voltage applications is critical because this limits potential output voltage overshoots when the input voltage falls below the device normal operating range. To optimize the transient behavior, a feed-forward capacitor can be added in parallel with the upper feedback resistor. For this design, 2×100-µF,10-V, X5R ceramic capacitors are used in parallel.

For further recommendations regarding component selection, refer to the LMS3655 data sheet [2].

2.3 Remote Port Regulation

The TPS254900-Q1 has been chosen for its optimization with automotive applications. The TPS254900 meets many of the requirements for automotive applications such as:

- Support for VBUS, D+, D– short-to-battery protection
- Cable compensation

The operation of the charging port circuit depends on the TPS254900 presented in Figure 17 for providing the electrical signatures on D+ and D– to support BC1.2 CDP-compliant charging.

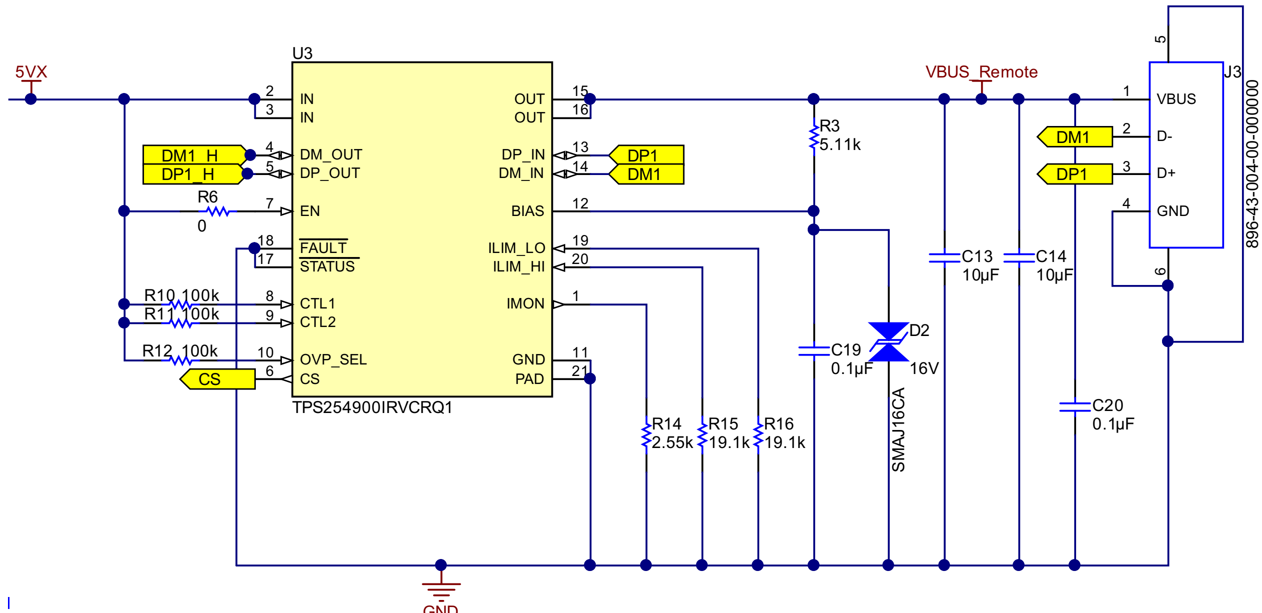


Figure 17. TPS254900-Q1 Remote CDP Port Controller

The internal switch of the TPS254900 is connected between the 5-V power source and VBUS. A CDP port is a USB port that follows USB BC1.2 requirements and supplies a minimum of 1.5 A per port. The port provides power and meets USB2.0 requirements for device enumeration. USB2.0 communication is supported and the host controller must be active to allow charging. What separates a CDP from an SDP is the host-charge handshaking logic that identifies this port as a CDP. A CDP is identifiable by a compliant BC1.2 client device and allows for additional current draw by the client device. As observable in the previous Figure 17, CTL1 and CTL2 are connected through R6 and R7 for the configuration of SDP, SDP1, or CDP mode. See Table 7 for the configurations.

Table 7. CTL1 and CTL2 Configurations

CTL1	CTL2	MODE	FAULT REPORT	NOTES
0	0	Client	OFF	Power Switch is disabled, only analog switch is on
0	1	SDP	ON	Standard SDP
1	0	SDP1	ON	No OUT discharge between CDP and SDP1 for port power management (PPM)
1	1	CDP	ON	No OUT discharge when changing between 10 and 11

The TPS254900-Q1 has been chosen in particular for its USB2.0 data support, integrated short-to-battery protection, and because of its cable compensation feature through the CS pin. This particular feature can be used as shown in Figure 15.

When a load draws current through a long or thin wire, an IR drop occurs which reduces the voltage delivered to the load. In the vehicle from the voltage regulator 5-V output to the VPD_IN (input voltage of portable device), the total resistance of power switch $RDS_{(ON)}$ and cable resistance causes an IR drop at the PD input. So the charging current of most portable devices is less than their expected maximum charging current. The TPS254900-Q1 device detects the load currents and sets a proportional sink current that can be used to adjust output voltage of the up-stream regulator to compensate the IR drop in the charging path. The gain GCS of sink current proportional to load current is 75 µA/A.

To design the feedback network on the regulator, the following calculations can be used:

- Choose RG according to the voltage regulator guidelines
- Calculate RFA as specified in Equation 6:

$$RFA = (RDS_{(ON)} + RWIRE) / GCS \quad (6)$$

Where:

- RWIRE is typically 80 mΩ/m to 120 mΩ/m and up to 3-m long
- GCS = 75 µA/A

Calculate RFB as specified in Equation 7:

$$RFB = VOUT / (VFB / RG) – RG – RFA \quad (7)$$

For this design, the voltage was set to 5.11 V with V_{FB} = 1 V assuming a 150-mΩ cable drop and 50-mΩ $RDS_{(ON)}$ of the TPS254900 switch.

2.4 Local Port Controller

A unique challenge when designing for both a local port and remote port is the issue that the DC-DC regulator must be set at a higher voltage (5.11 V in this case) so that the cable compensation can compensate for the voltage droop across a long wire. The issue arises when the local port must be within a tight set of requirements, usually ±2.5%. This set of restrictions becomes tricky to design without using two different regulators; so, to design for this, a very small regulation differential between input and output must be maintained. Usually this type of design restriction can be solved with a low dropout regulator (LDO); however, because this is an automotive USB port, two more challenges arise: The dropout must not only be 50 mV to 100 mV, but the LDO must also be automotive-qualified and the LDO must support up to 2.4 A. For this design, a custom LDO (see Figure 18) has been created using a low $RDS_{(ON)}$ MOSFET and op amp. This discrete LDO regulates the voltage for the local port controller, which is shown in Figure 19.

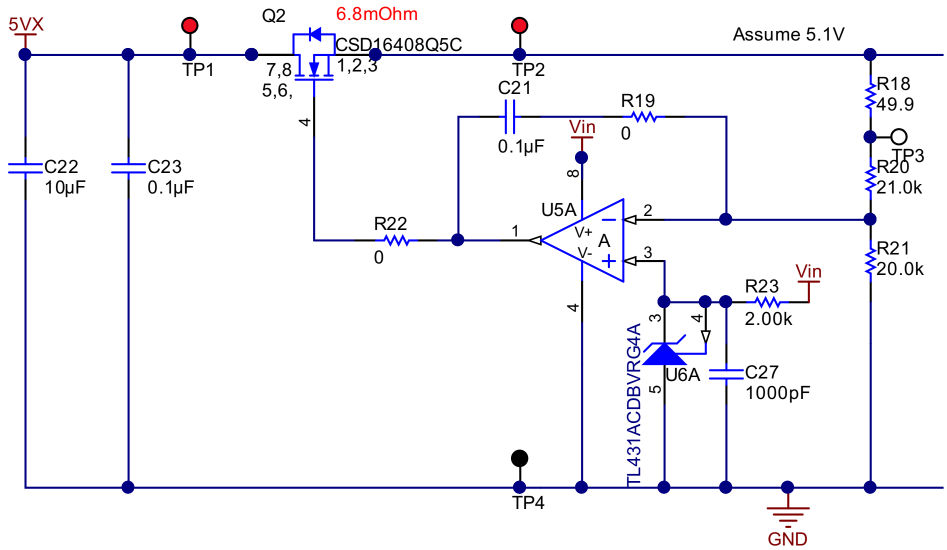


Figure 18. Custom LDO With Low Dropout

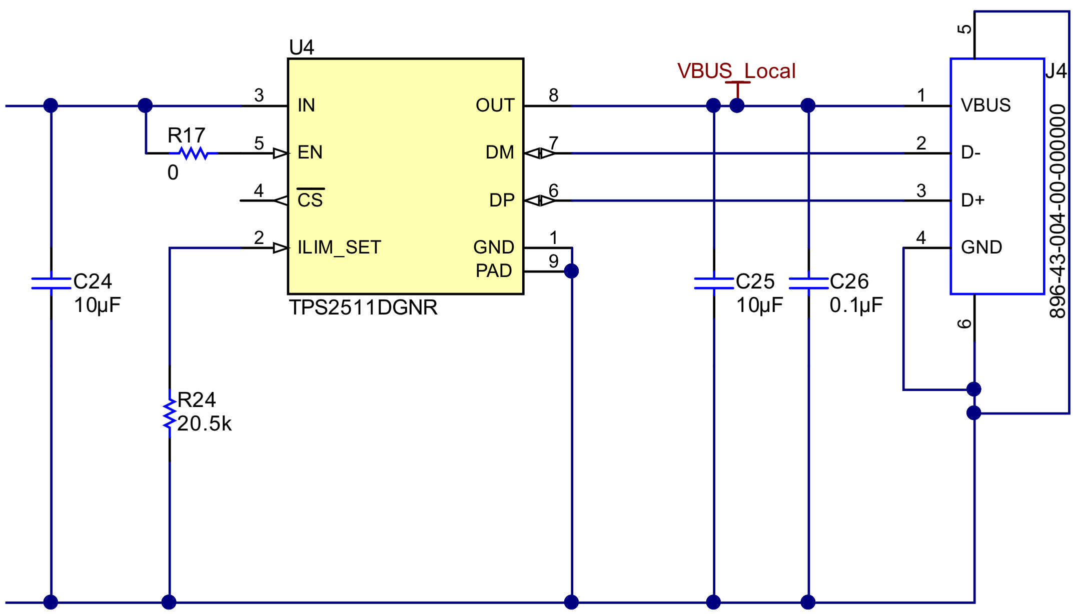


Figure 19. TPS2511 Local Port DCP Controller

The following Table 8 shows the voltages at the extremes of a no-load and full-load on either or both of the ports.

Table 8. Load Current Truth Table

REMOTE LOAD	LOCAL LOAD	VBUS	V_REMOTE	V_LOCAL
0	0	5.11 V	4.98 V	5.05 V
0	1	5.11 V	4.98 V	5.11 – (76.8 mΩ × 2.4 A) = 4.92 V
1	0	5.4 V	≈5.28 V	5.4 V
1	1	5.4 V	≈5.28 V	5.4 V – (76.8 mΩ × 2.4 A) = 5.22 V

3 Getting Started Hardware

This board is designed to passthrough data on the remote port in CDP mode and support dedicated charging (DCP) on the local port. Although there is not a USB output port to connect a USB host, D+ and D– can be pulled high or low to enable this detection.

To get started, simply supply 12 V to the J4 pin and plug in a USB type-A connector to either or both USB ports (see Figure 20).

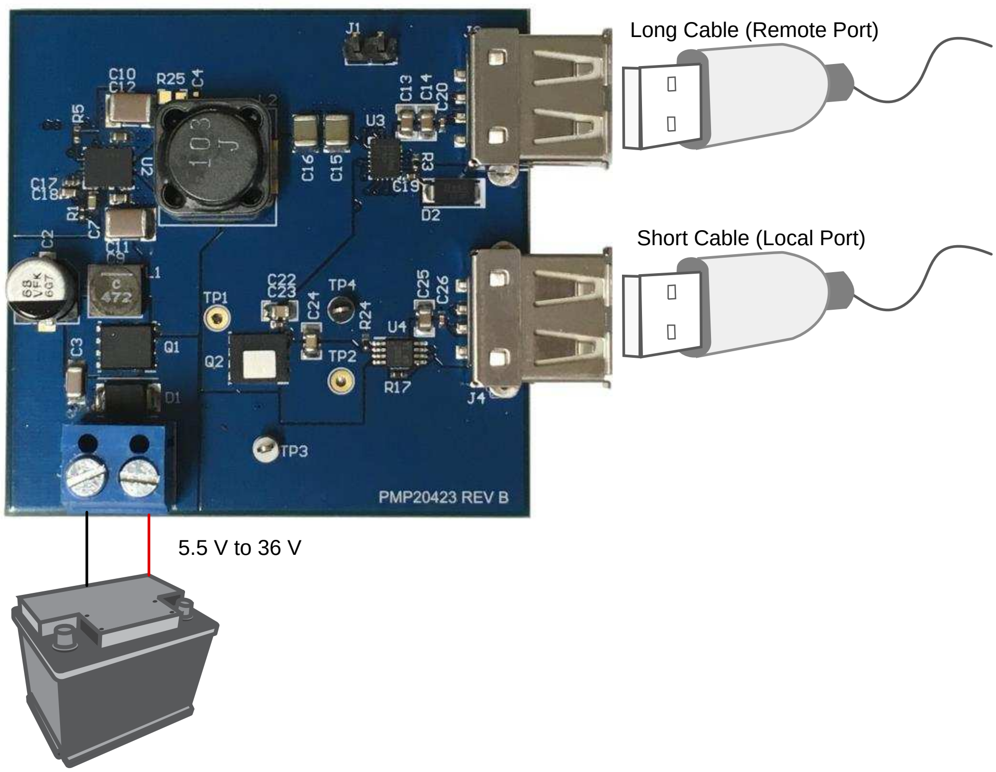


Figure 20. Getting Started Hardware

4 Testing and Results

The four-layer, 2-oz copper layer board was tested with a 1-m USB cable for the remote port. The local port was a point of load connection to an electronic load.

4.1 Start-Up

The following Figure 21 shows the start-up of the LMS3655 device at the TPS254900 output (remote port, pink), the start-up of the linear regulator at the TPS2511 output (local port, green), and V_{IN} (gold) where V_{IN} = 12 V with both ports at a 0-A load.

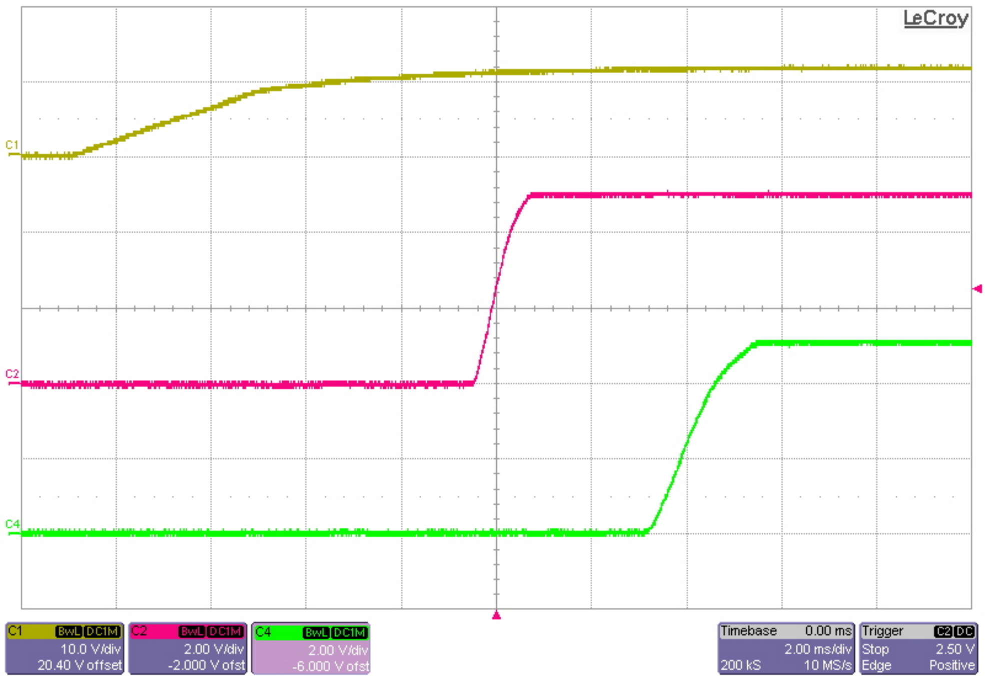


Figure 21. Startup Voltage at V_{IN} = 12 V, I_{OUT} = 0 A

4.2 Output Voltage

The converter voltage is one of the main concerns in this design. By measuring the output voltage based on the truth table from Table 8, the designer can determine if both ports can support tight requirements even using a single DC-DC. The following Table 9 shows the measurements. As the preceding Table 8 shows, 0 indicates no load and 1 indicates a full load at 2.4 A.

Table 9. Truth Table Measurements

REMOTE LOAD	LOCAL LOAD	VBUS	REMOTE PORT VOLTAGE	LOCAL PORT VOLTAGE
0	0	5.089 V	5.089 V	5.089 V
0	1	5.086 V	5.086 V	4.876 V
1	0	5.617 V	5.081 V	5.094 V
1	1	5.610 V	5.092 V	4.890 V

4.2.1 Output Ripple

The waveform in the following Figure 22 was captured at the remote port with V_{IN} = 12 V and the remote port loaded to 2.4 A while the local port was loaded to 0 A.

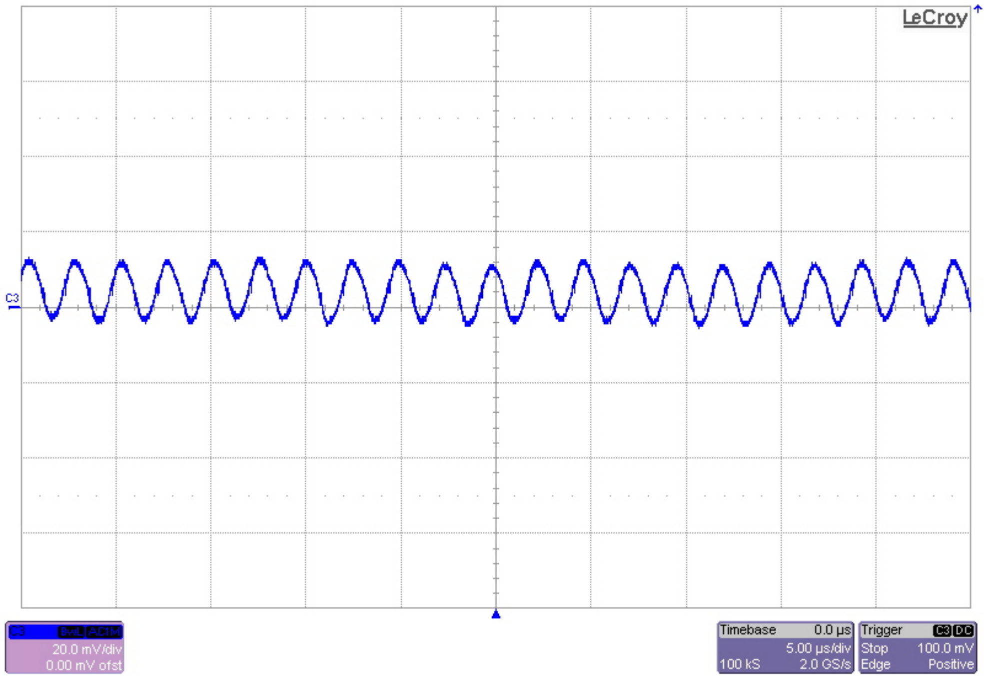


Figure 22. Output Ripple V_{IN} = 12 V, Remote Port = 2.4 A, Local Port = 0 A

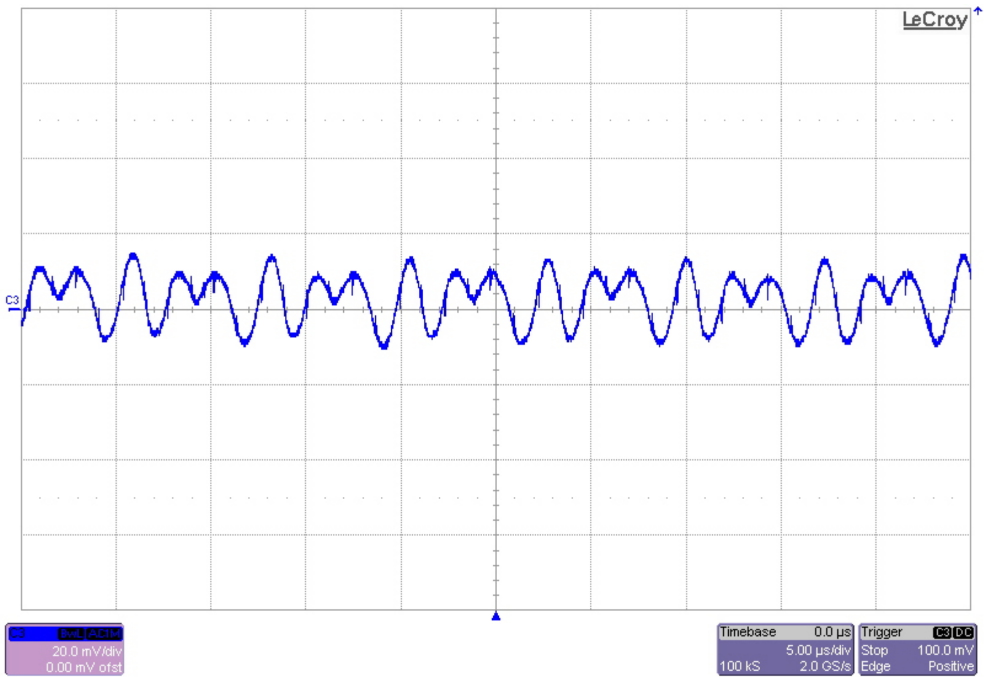


Figure 23. Output Ripple V_{IN} = 12 V, Remote Port = 2.4 A, Local Port = 2.4 A

4.2.2 Load Transients

The capture in the following Figure 24 was taken at the remote port, with the remote load current being stepped between 0 A and 2.4 A with V_{IN} = 12 V. The offset phenomena shown is due to the cable compensation offset voltage, which jumps to approximately 500 mV at 2.4 A. The local port was loaded at 2.4 A.

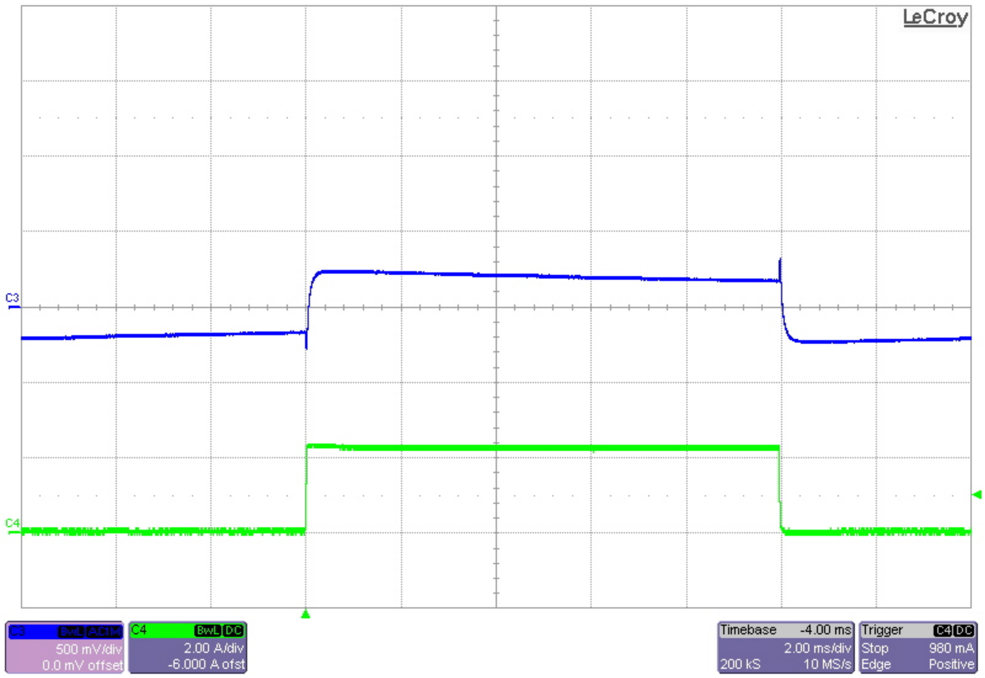


Figure 24. Transient V_{IN} = 12 V, Remote Load = 0 A to 2.4 A, Local Load = 2.4 A (Static)

The following Figure 25 was taken at the local port, with the local port stepping from 0 A to 2.4 A and remote port fully loaded to a static 2.4 A.

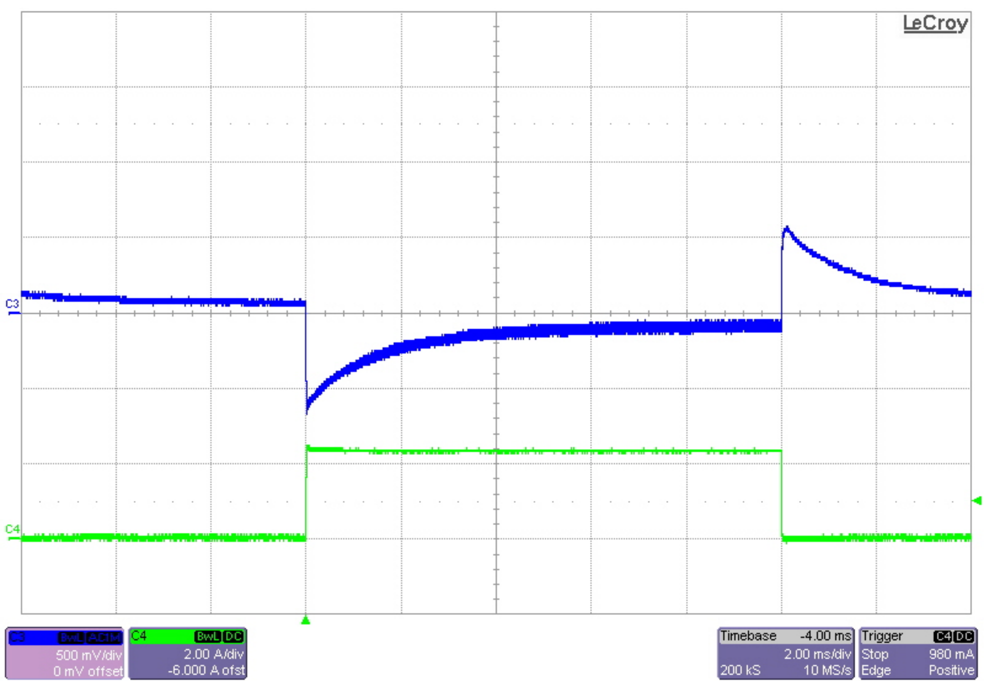


Figure 25. Transient V_{IN} = 12 V, Remote Load = 2.4 A (Static), Local Load = 0 A to 2.4 A

The capture in the following Figure 26 was taken at the local port. The local port is unloaded during this time at 0 A, while the remote port is being stepped from 0 A to 2.4 A.

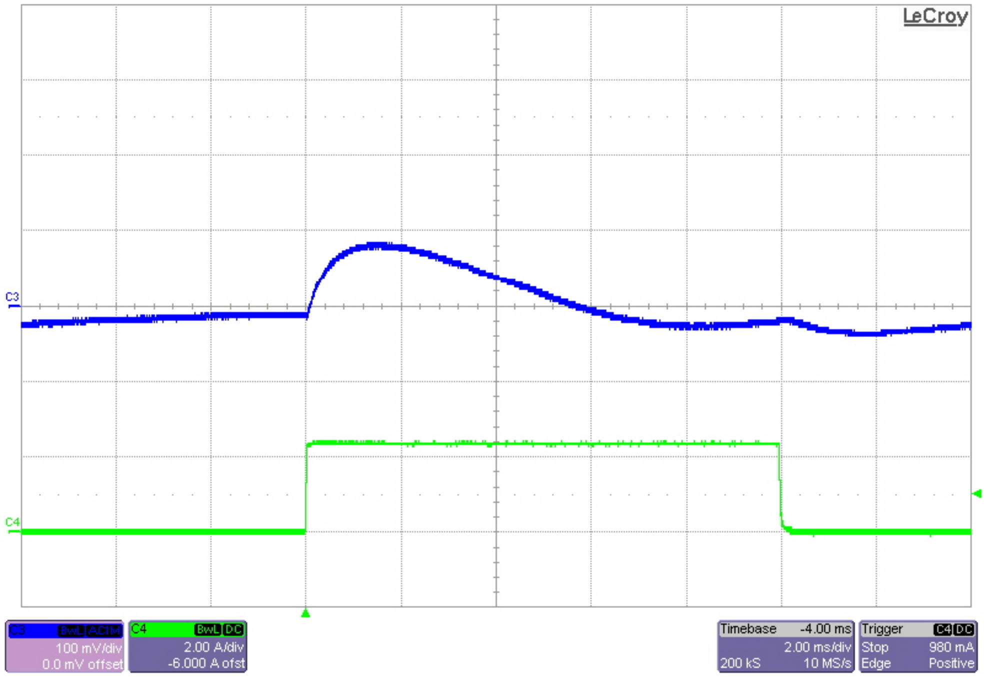


Figure 26. Transient V_{IN} = 12 V, Remote Load = 0 A to 2.4 A, Local Load = 0 (Static)

4.3 Switch Node

The following Figure 27 shows the L1 switch node. This image was taken at V_{IN} = 12 V with both ports loaded at 2.4 A, each.

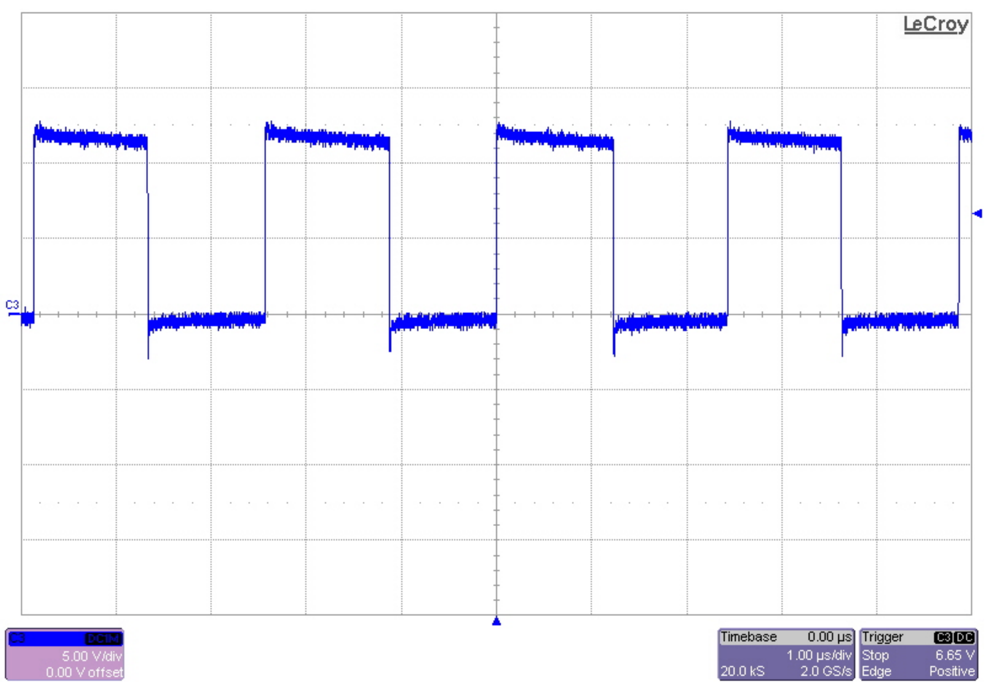


Figure 27. Switch Node V_{IN} = 12 V, Remote Load = 2.4 A, Local Load = 2.4 A

4.4 Loop Response

The following Figure 28 shows the loop response for the DC-DC LMS3655.

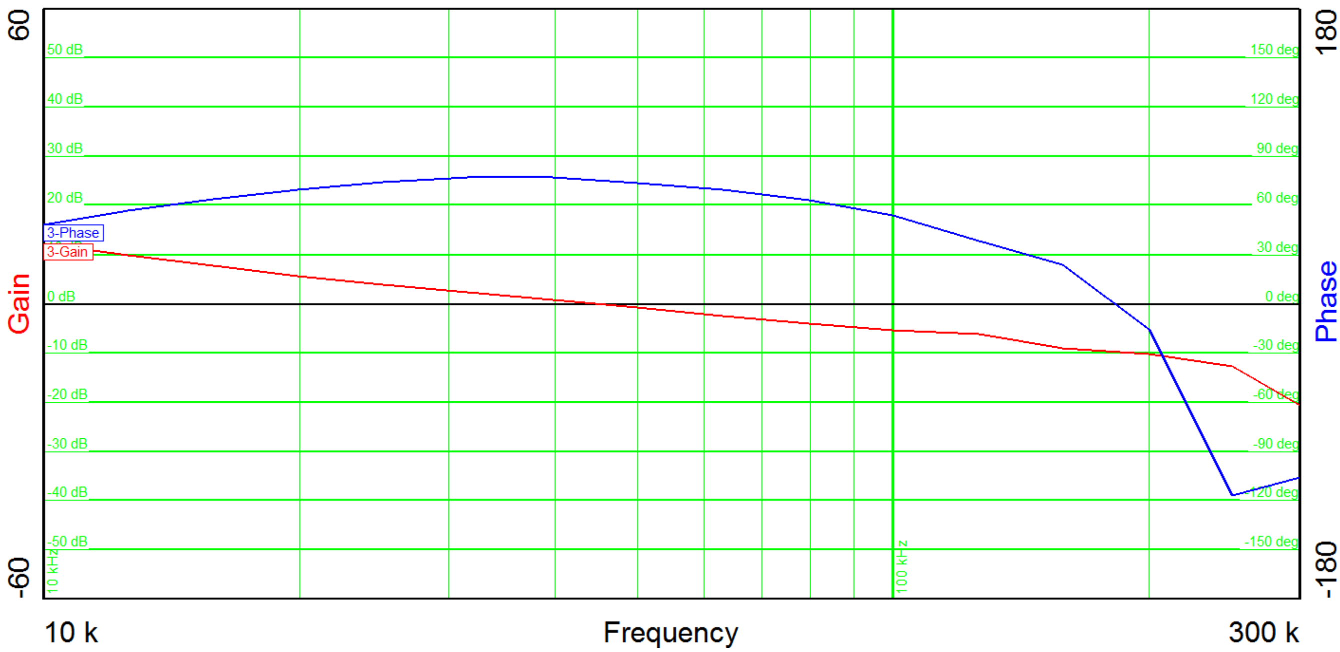


Figure 28. Loop Response, V_{IN} = 12 V, Full Load

4.5 Thermal Images

The following Figure 29 and Figure 30 are the thermal images of the board taken with V_{IN} = 12 V, a load of 2.4 A on each port, and no airflow. The board thermal was taken after 15 min of full loading.

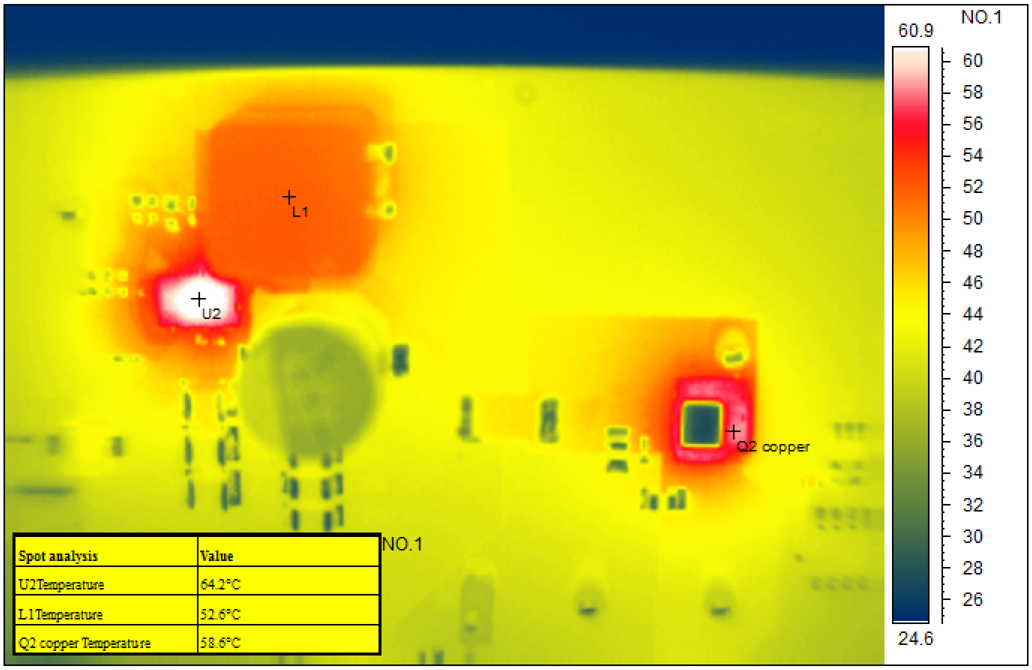


Figure 29. Thermal Image at Full Load—Front Side

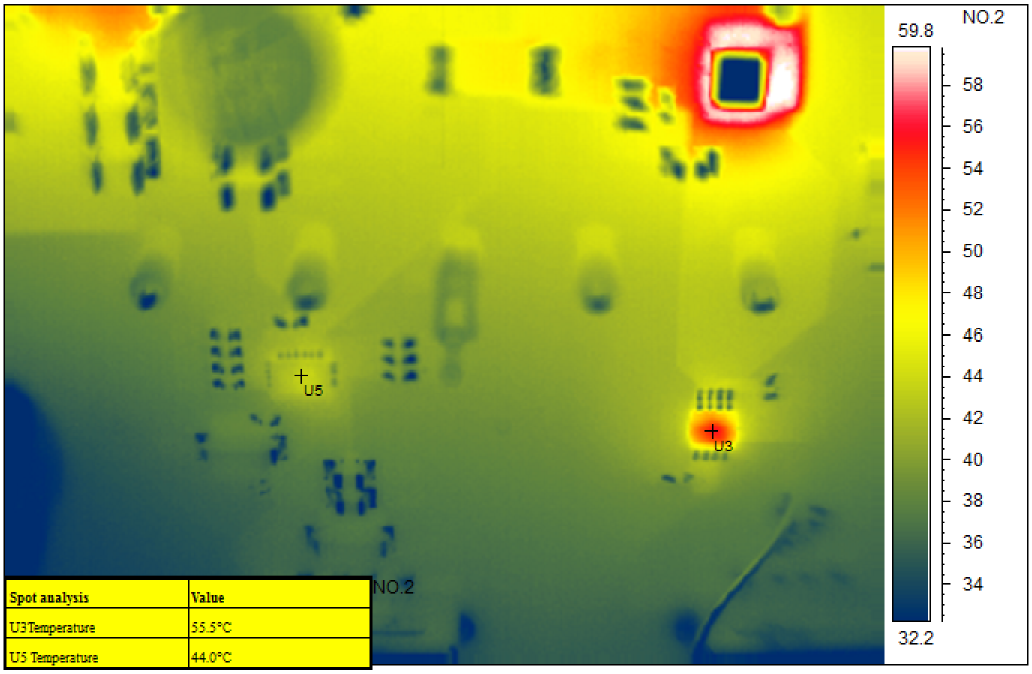


Figure 30. Thermal Image at Full Load—Back Side

4.6 CISPR 25 Class 4 Testing

This reference design was built and tested with the goal to pass CISPR 25 standards and currently meets Class 4 of the CISPR classification standards. This CISPR EMI testing is performed using a third-party facility and follows this international standard "to protect onboard receivers from disturbances produced by conducted and radiated emissions arising in a vehicle," [1] which means that this particular design has been tested to make sure that it does not interfere with other equipment in the vehicle. As outlined in the standard, radiated disturbances do not disrupt the broadcast and mobile service or band. For this particular reference design, the broadcast standards were tested at peak, quasi-peak, and average ratings. This section outlines the individual tests for radiated and conducted emissions. For both types of tests, a car battery or a 14.5-V power supply was used in conjunction with short cables to test at the optimized performance. Additionally, resistive loads were used to emulate similar types of loads that USB devices may draw.

4.6.1 Radiated Emissions

The absorber-lined shielded-enclosure (ALSE) procedure was used for testing radiated emissions. This procedure involves a chamber that has RF absorber material along the walls and ceiling of the enclosure. Table 10 and Table 11 show the broadcast lists of examples for average, quasi-peak, and peak limits for radiated disturbances for this particular ALSE method.

(1) CISPR 25 Ed. 3.0 b: 2008, pg. 7

Table 10. ALSE Average Levels for CISPR 25 Radiated Emissions

SERVICE OR BAND	FREQUENCY (MHz)	LEVELS IN dB (µV/m)				
		CLASS 1	CLASS 2	CLASS 3	CLASS 4	CLASS 5
BROADCAST		AVG	AVG	AVG	AVG	AVG
LW	0, 15 - 0, 30	66	56	46	36	26
MW	0, 53 - 1, 8	52	44	36	28	20
SW	5, 9 - 6, 2	44	38	32	26	20
FM	76 - 108	42	36	30	24	18
TV band I	41 - 88	42	36	30	24	18

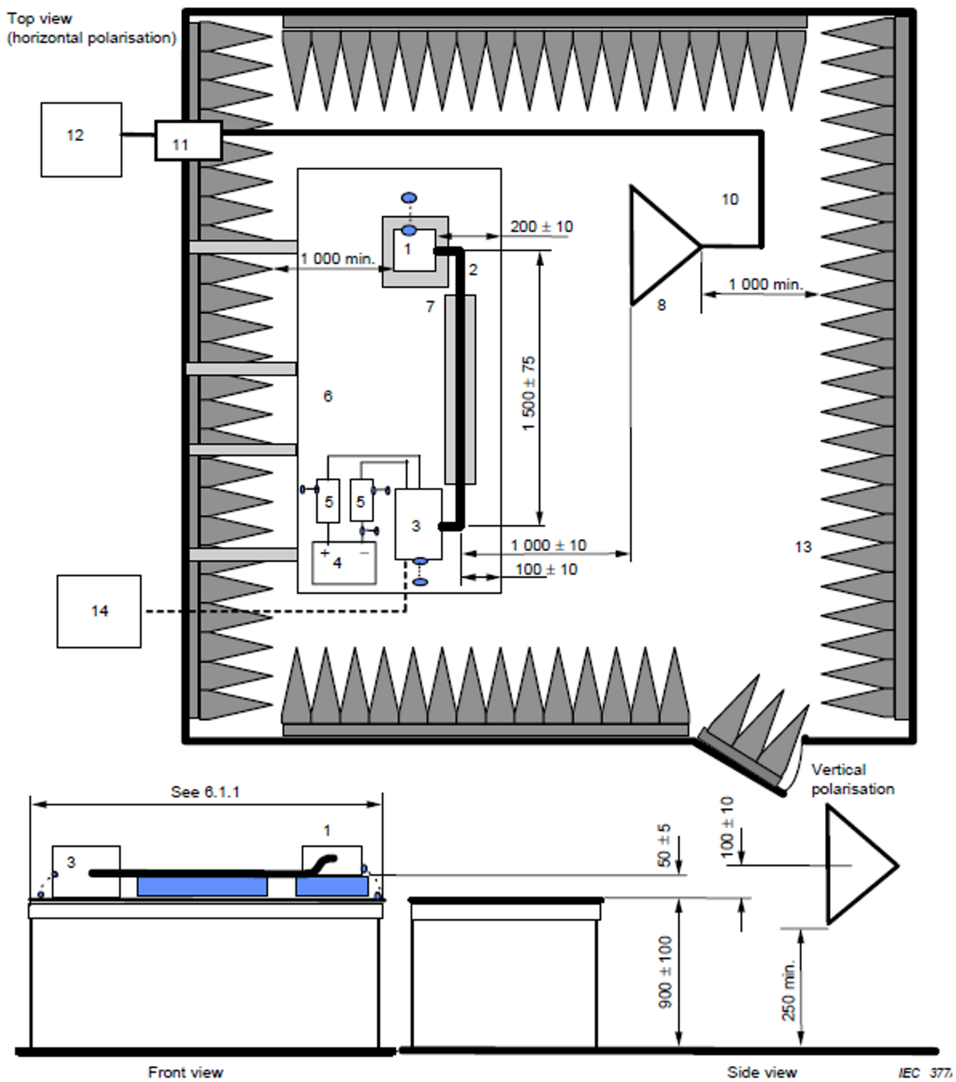
Table 10. ALSE Average Levels for CISPR 25 Radiated Emissions (continued)

SERVICE OR BAND	FREQUENCY (MHz)	LEVELS IN dB (µV/m)				
		CLASS 1	CLASS 2	CLASS 3	CLASS 4	CLASS 5
BROADCAST		AVG	AVG	AVG	AVG	AVG
TV band II	174- 230	46	40	34	28	22
DAB III	171 - 245	50	34	28	22	16
TV band IV/V	468 - 944	55	49	43	37	31
DTTV	470 - 770	59	53	47	41	35
DAB L band	1447 - 1494	42	36	30	24	18
SDARS	2320 - 2345	48	42	36	30	24

Table 11. ALSE Peak and Quasi-Peak Levels for CISPR 25 Radiated Emissions

SERVICE OR BAND	FREQUENCY (MHz)	LEVELS IN dB (µV/m)									
BROADCAST		CLASS 1		CLASS 2		CLASS 3		CLASS 4		CLASS 5	
		PEAK	QUASI	PEAK	QUASI	PEAK	QUASI	PEAK	QUASI	PEAK	QUASI
LW	0,15-0,30	86	73	76	63	66	53	56	43	46	33
MW	0,53-1,8	72	59	64	51	56	43	48	35	40	27
SW	5,9-6,2	64	51	58	45	52	39	46	33	40	27
FM	76-108	62	49	56	43	50	37	44	31	38	25
TV Band I	41-88	52	—	46	—	40	—	34	—	28	—
TV Band III	174-230	56	—	50	—	44	—	38	—	32	—
DAB III	171-245	50	—	44	—	38	—	32	—	26	—
TV Band IV	468-944	65	—	59	—	53	—	47	—	41	—
DTTV	470-770	69	—	63	—	57	—	51	—	45	—
DAB L Band	1447-1494	52	—	46	—	40	—	34	—	28	—
SDARS	2320-2345	58	—	52	—	46	—	40	—	34	—

For the ALSE method of radiated emissions testing, four different antennas are used to measure the full range of testing and three antennas were tested in a vertical polarization and horizontal polarization. Additionally, ambient readings were taken before each test. Figure 31 shows a diagram of the first test using a rod antenna (only one polarization option). This test was conducted for the frequency range between 150 kHz to 30 MHz. This band is particularly important to test because it captures the range for the AM and FM radio frequencies.



Key

1 EUT (grounded locally if required in test plan)

2 Test harness

3 Load simulator (placement and ground connection according to 6.4.2.5)

4 Power supply (location optional)

5 Artificial network (AN)

6 Ground plane (bonded to shielded enclosure)

7 Low relative permittivity support ($\varepsilon_r \leq 1.4$)

8 Horn antenna

10 High-quality coaxial cable e.g. double-shielded (50 Ω)

11 Bulkhead connector

12 Measuring instrument

13 RF absorber material

14 Stimulation and monitoring system

Figure 31. Radiated Emissions Test Setup

The following figures show the ambient reading and the radiated emissions of the design. The first peak at 400 kHz is easy to identify and corresponds to the switching frequency of the LMS3655-Q1 device. Subsequent peaks are also observable at the fundamentals for the switching frequency at approximately 1.2 MHz, 2.0 MHz, and so on. This reference design has passed all peak and quasi-peak limits as indicated in the CISPR 25 Class 4 documentation. For the 530-kHz to 1.8-MHz CISPR 25 Class 5 specification, the quasi-peak indicates a limit at 27 dB and this design remains at 30.7 dB. For average measurements, the documentation dictates a limit of 20 dB while this design sits at 25.8 dB. For further improvements to the design (to pass the CISPR 25 Class 5 standard), see *LMS3635-Q1 3.5-A, LMS3655-Q1 5.5-A, 36-V Synchronous, 400-kHz Step-Down Converter* .

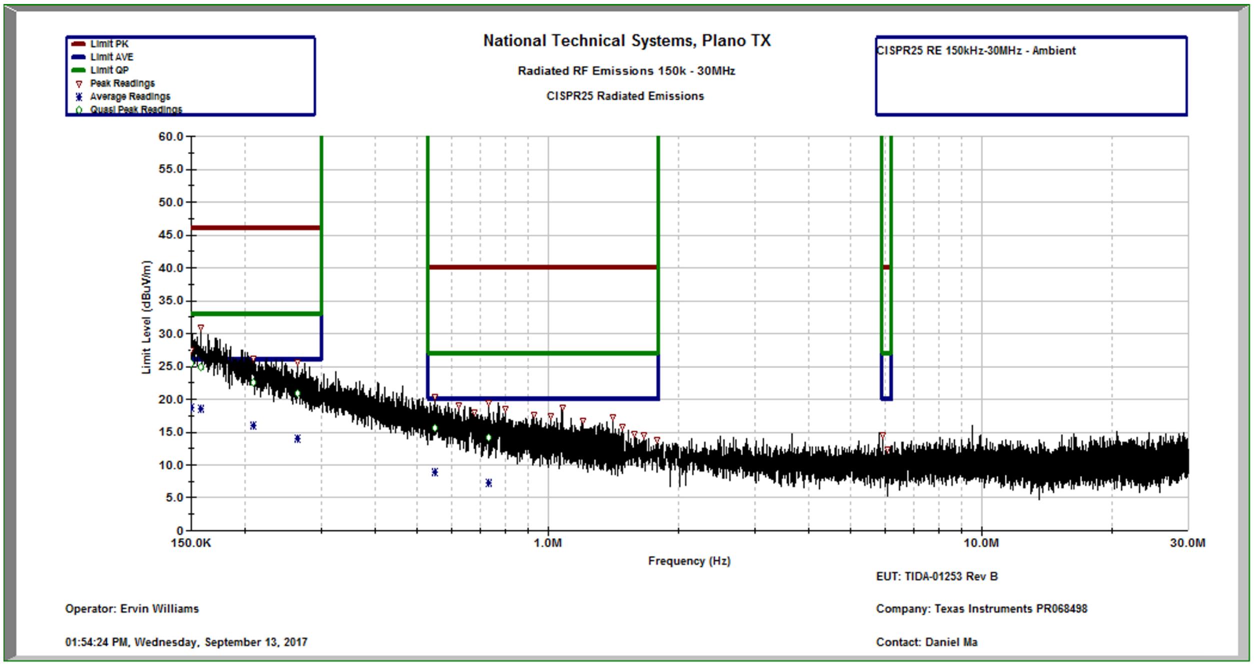


Figure 32. Radiated Ambient 150 kHz to 30 MHz

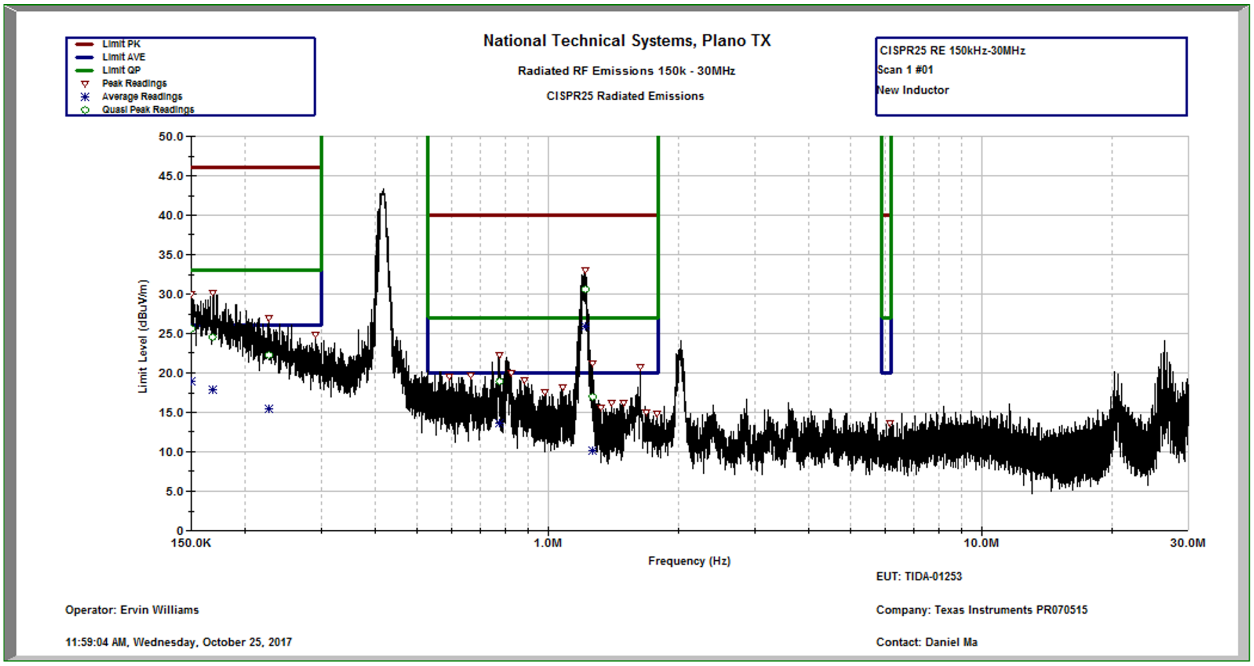


Figure 33. Radiated Emissions 150 kHz to 30 MHz

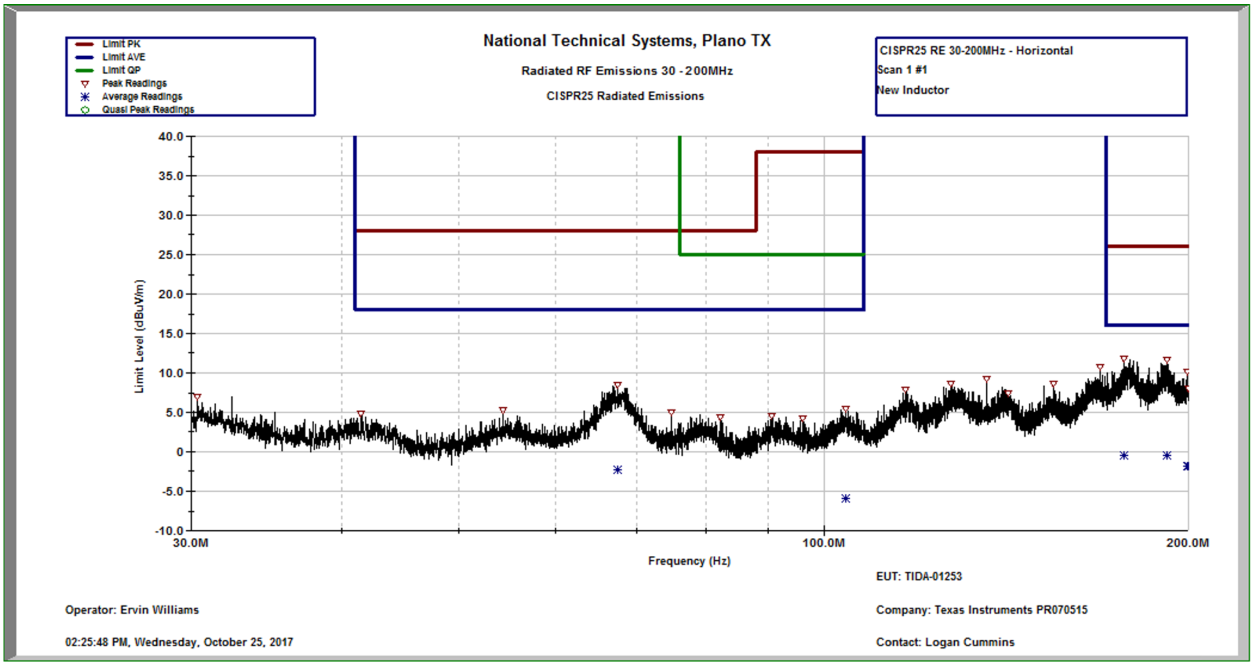


Figure 34. Radiated Emissions 30 MHz to 200 MHz

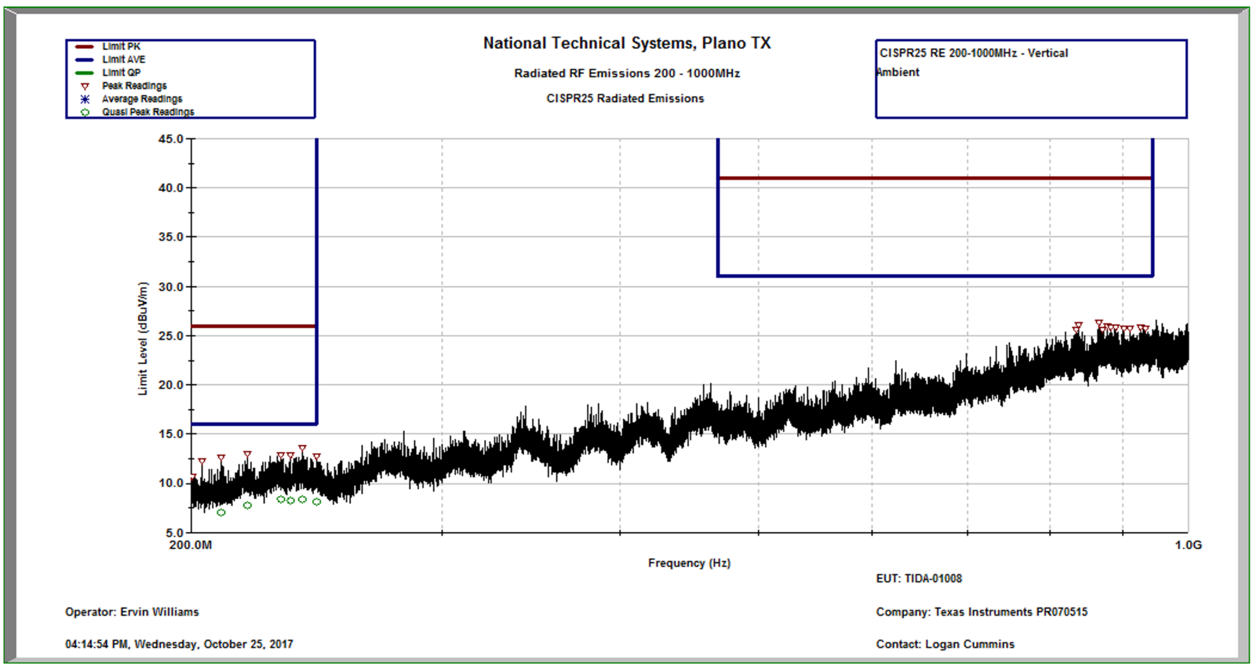


Figure 35. Radiated Emissions 200 MHz to 1 GHz

4.6.2 Conducted Emissions

A similar test setup is required for conducted emissions testing; however, no antennas are used. There are several variations in testing that are dependent upon the system under test. For this particular reference design, the voltage method was used and the following Table 12 and Table 13 indicate the average, quasi-peak, and peak limits for these tests.

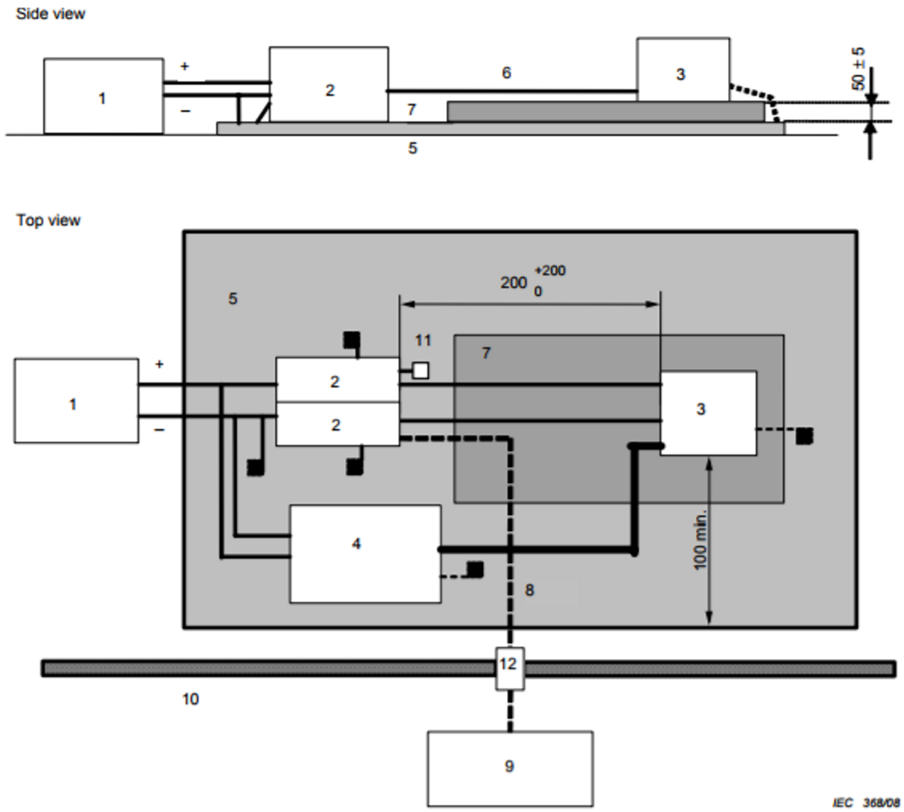
Table 12. ALSE Average Levels for CISPR 25 Conducted Emissions

SERVICE OR BAND	FREQUENCY (MHz)	LEVELS IN dB (µV)				
		CLASS 1	CLASS 2	CLASS 3	CLASS 4	CLASS 5
BROADCAST		AVG	AVG	AVG	AVG	AVG
LW	0, 15 - 0, 30	90	80	70	60	50
MW	0, 53 - 1, 8	66	58	50	42	34
SW	5, 9 - 6, 2	57	51	45	39	33
FM	76 - 108	42	36	30	24	18
TV band I	41 - 88	48	42	36	30	24
TV band II	174- 230	Conducted emission --Voltage method Not applicable				
DAB III	171 - 245					
TV band IV/V	468 - 944					
DTTV	470 - 770					
DAB L band	1447 - 1494					
SDARS	2320 - 2345					

Table 13. ALSE Peak and Quasi-Peak Levels for CISPR 25 Conducted Emissions

SERVICE OR BAND	FREQUENCY (MHz)	LEVELS IN dB (µV/m)									
BROADCAST		CLASS 1		CLASS 2		CLASS 3		CLASS 4		CLASS 5	
		PEAK	QUASI	PEAK	QUASI	PEAK	QUASI	PEAK	QUASI	PEAK	QUASI
LW	0, 15 - 0, 30	110	97	100	87	90	77	80	67	70	57
MW	0, 53 - 1, 8	86	73	78	65	70	57	62	49	54	41
SW	5, 9 - 6, 2	77	64	71	58	65	52	59	46	53	40
FM	76 - 108	62	49	56	43	50	37	44	31	38	25
TV Band I	41 - 88	58	—	52	—	46	—	40	—	34	—
TV Band III	174 - 230	Conducted Emission --Voltage Method Not Applicable									
DAB III	171 - 245										
TV Band IV	468 - 944										
DTTV	470 - 770										
DAB L Band	1447 - 1494										
SDARS	2320 - 2345										

For the voltage method of conducted emissions testing, both the positive terminal and negative terminal of the line impedance stabilization networks (LISNs) were used to measure the full range of testing. Additionally, ambient readings were taken before each test. Figure 36 shows a diagram of the conducted test. This test was performed for the
150-kHz to 108-MHz frequency range. This band is particularly important to test because it captures the range for the AM and FM radio frequencies.



Key

1 Power supply (may be placed on the ground plane)
2 Artificial network
3 EUT (housing grounded if required in test plan)
4 Load simulator (metallic casing grounded if required in test plan)
5 Ground plane
6 Power supply lines
7 Low relative permittivity support ($\varepsilon_r \leq$ 1,4)
8 High-quality coaxial cable e.g. double-shielded (50 Ω)
9 Measuring instrument
10 Shielded enclosure
11 50 Ω load
12 Bulkhead connector

NOTE The EUT housing ground lead, when required in the test plan, should not be longer than 150 mm.

Figure 36. Conducted Emissions Test Setup

The following figures show the ambient reading, low-side reading, and return reading, respectively. A jump is present in the ambient reading due to the change in bandwidth to 30 MHz.

For the conducted emissions measurements, the 400 kHz is easy to identify and corresponds to the switching frequency of the LMS3655-Q1 device. On the return side, subsequent peaks are observable at the fundamentals for the switching frequency at approximately 1.2 MHz, 2 MHz, and so on. This reference design has passed all peak and quasi-peak limits as indicated in the CISPR 25 Class 5 documentation for ranges 150 kHz to 108 MHz.

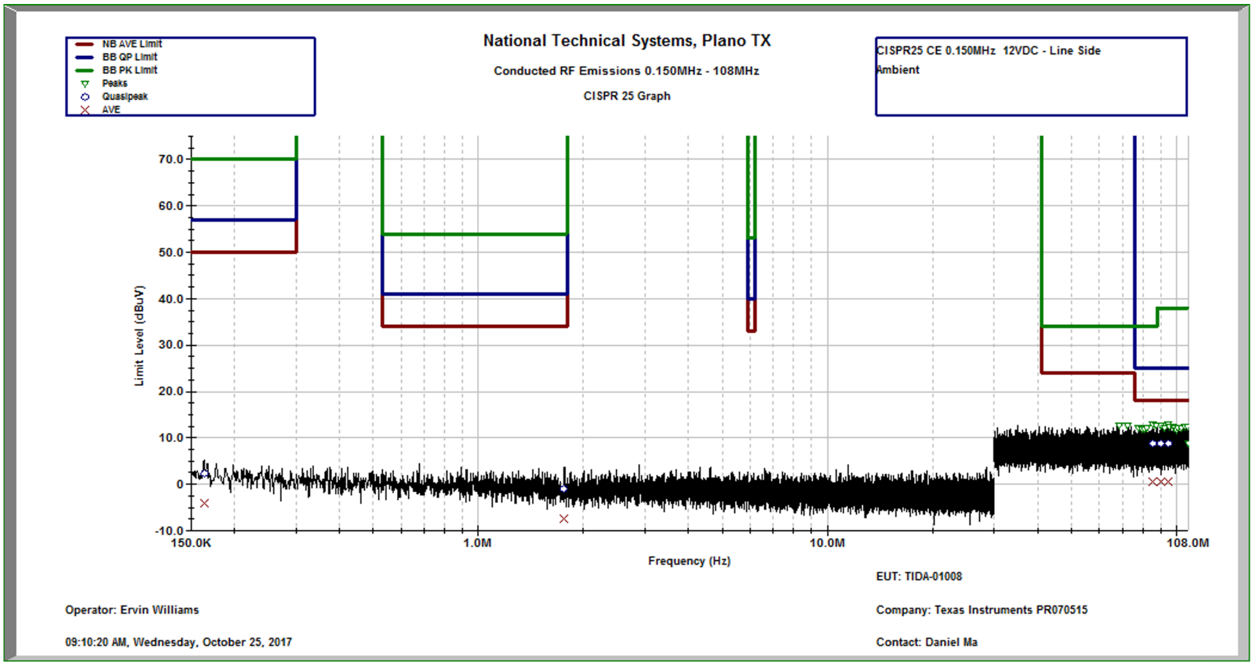


Figure 37. Conducted Emissions Ambient 150 kHz to 108 MHz

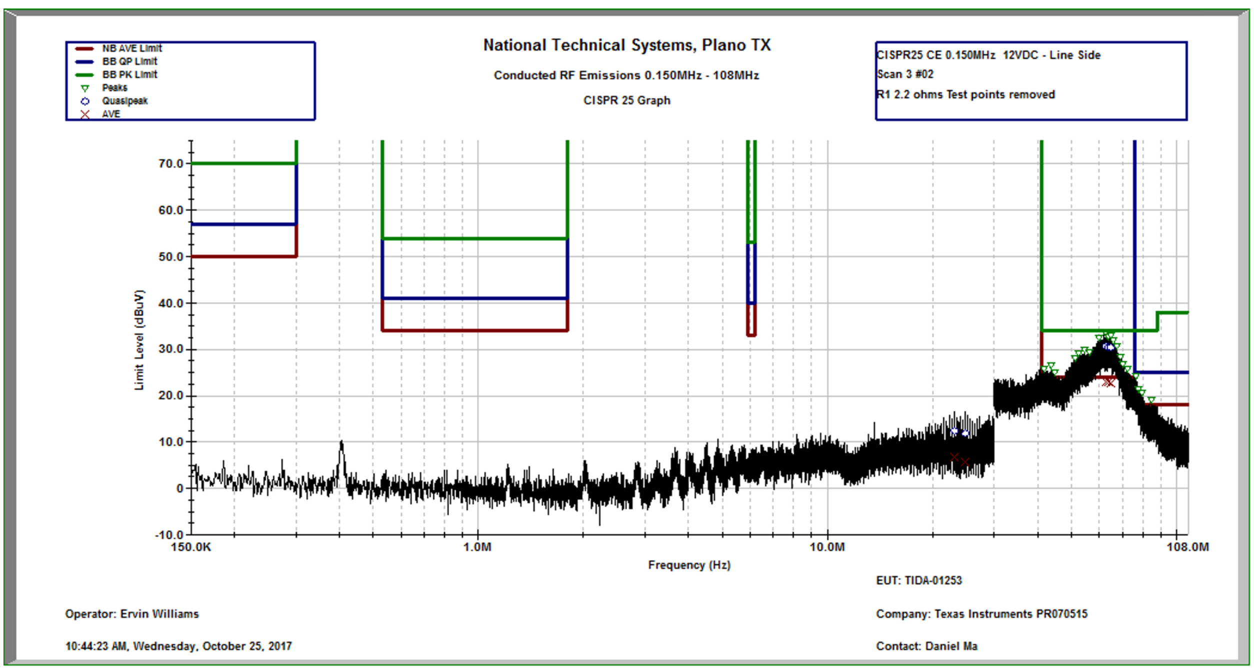


Figure 38. Conducted Emissions Line 150 kHz to 108 MHz

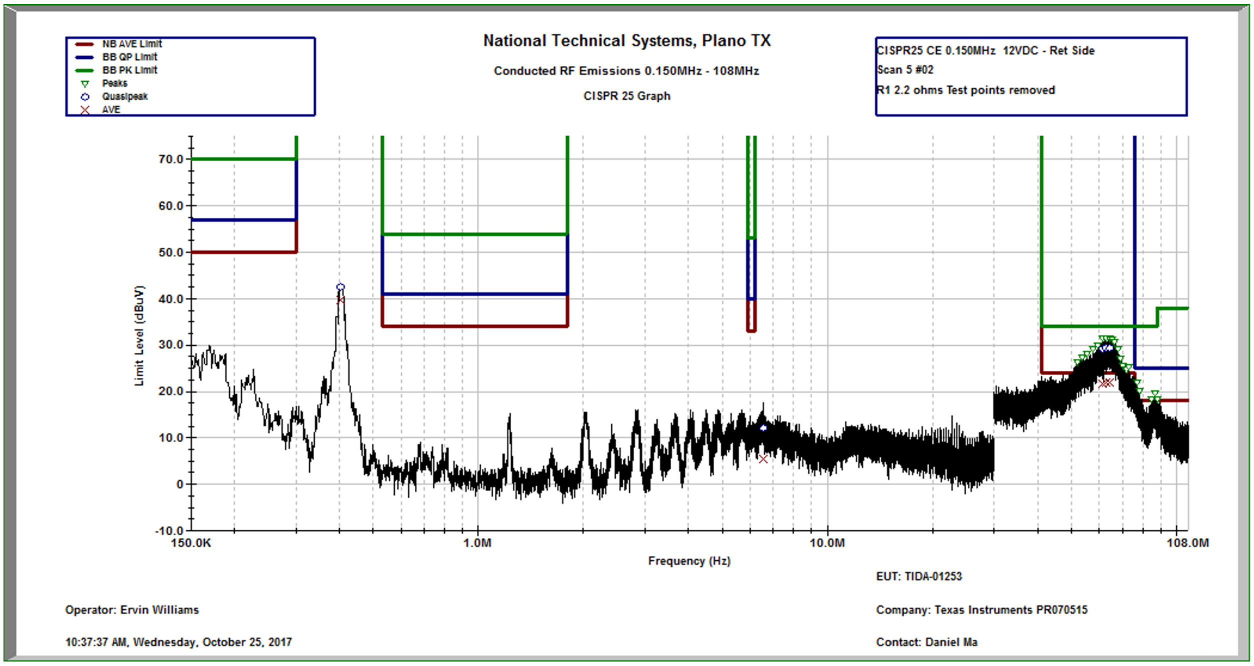


Figure 39. Conducted Emissions Return 150 kHz to 108 MHz

5 Design Files

5.1 Schematics

To download the schematics, see the design files at TIDA-01253.

5.2 Bill of Materials

To download the bill of materials (BOM), see the design files at TIDA-01253.

5.3 PCB Layout Recommendations

The TIDA-01253 PCB is a four-layer board:

- The top layer has the majority of components
- The second layer is a GND plane
- The third layer is signals
- The fourth layer is also a GND plane

Follow the recommendation from the corresponding data sheet diligently for most of the components, while making a few minor deviations to optimize for EMI constraints. Fully fill the top and bottom layer with copper for the ideal thermal performance. For a layout guide, follow the recommendations for individual components in the following subsections and their respective data sheets. The following subsection walks through some recommendations for the board layout.

5.3.1 DC-DC Converter

The more challenging layout of this design is the DC-DC converter. The following recommendations have been considered to maximize performance and reduce EMI:

- Compact layout and feedback resistors
- Ground plane and shape routing
- Thermal design

5.3.1.1 Compact Layout and Feedback Resistors

EMI is reduced by keeping switching current loops small. Minimizing the size of high di/dt areas helps to reduce radiated EMI.

1. The switching current loops for the switch node and other switching signals are kept small.
2. The input capacitors, Cin, are placed close to the input and use a common ground with the output capacitors.
3. The feedback, soft-start, and enable components are routed to the AGND terminal of the device at a single point of ground, which prevents any switched currents or load currents from flowing in the analog ground traces. Poor grounding can result in degraded load regulation or erratic output-voltage ripple behavior.
4. The size of the FB node was minimized because the feedback node is a significant source of noise. Keeping feedback resistors away from the switch node helps to minimize the coupling of this noise. All of the feedback resistors R10, R14, and R5 are placed close to the FB terminal and the feedforward capacitor C7 is placed directly in parallel with R5.

5.3.1.2 Ground Plane and Shape Routing

The input plane and output plane are kept as wide as possible to reduce any voltage drops on the input or output of the converter to maximize efficiency.

5.3.1.3 Thermal Design

An array of heat-sinking vias are connected from the exposed pad to the fourth-layer planes. For best results, use a minimum via diameter of 12 mil with thermal vias spaced 46.8-mil apart. As indicated in Section 4, 2 oz of copper per layer is used to ensure enough copper area is available for heat-sinking to maintain the junction temperature below 125ºC.

5.3.2 TPS254900 Layout Recommendation

To follow the recommendations for this particular part, refer to the data sheet [3]. The following instructions highlight some of the slight variances from the data sheet recommendation that pertain specifically to this reference design. Many of the surrounding components are simply configuration passives and it is important to create a close, compact placement between these components and the TPS254900 device. One of the most important considerations for the layout of this component are the DP_IN, DN_IN, DP_OUT, and DN_OUT signals. USB-IF standards for the USB2.0, universal serial bus 2.0 specification, and data require that these lines must be 90-Ω differential pairs. Additionally, trace length matching is a concern, so adding trace length is beneficial to maintaining signal integrity. A free online Impedance calculator from Mantaro Product Development Services [4] can be used to calculate the differential micro strip impedance, trace width, separation, and so on. For the DP_IN/DN_IN signal pair coming from the USB type-A connector, the signals are differentially routed on the top layer and signal layer 2. A GND plane is situated directly below the DP_IN/DN_IN layer to keep the signals from coupling with other traces. The same GND layer consideration should be taken into consideration for DP_OUT/DN_OUT, which is routed to J1.

5.3.3 LM74610 Layout Recommendation

For the reverse battery protection, the input capacitors are placed on the front of the board and the ideal recommendation is to rotate them 90º from each other due to possible flexing of the board. Take care to keep these components close to the input connectors of the board. The input to the LM74610 passes through the TVS diodes to pin 4 and the anode of Q1. To make routing easier, Q1 is placed below the LM74610 with C12 so that the gate drive and gate pull down of Q1 and VCAPH/VCAPL can be routed on signal layer 2 underneath the part. The output and cathode of Q1 goes directly to the pi filter. For additional EMI protection, a common-mode choke can be used with the pi filter to reduce high-frequency EMI.

5.4 Altium Project

To download the Altium project files, see the design files at TIDA-01253.

5.5 Gerber Files

To download the Gerber files, see the design files at TIDA-01253.

5.6 Assembly Drawings

To download the assembly drawings, see the design files at TIDA-01253.

6 Software Files

To download the software files, see the design files at TIDA-01253.

7 Related Documentation

1. Texas Instruments, *LM74610-Q1 Zero IQ Reverse Polarity Protection Smart Diode Controller*
2. Texas Instruments, *LMS3635-Q1 3.5-A, LMS3655-Q1 5.5-A, 36-V Synchronous, 400-kHz Step-Down Converter*
3. Texas Instruments, *TPS254900-Q1 Automotive USB Host Charger With Short-to-V_{BATT} Protection*
4. Mantaro.com, *Impedance Calculators*

7.1 Trademarks

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8 About the Author

HOPE BOVENZI is a Systems Engineer at Texas Instruments. Hope earned her bachelor of science in electrical engineering from the University of California at Davis in 2012. As a member of the Automotive Systems Engineering team at Texas Instruments, she is responsible for developing reference design solutions for the Automotive Infotainment and Cluster segment and has a background in power design.

9 Special Thanks

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