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Low-Noise, High PSRR LDO for Powering Hi-Fi Audio Application

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LP5907	Product Folder
TPS72301	Product Folder
TPS65130	Product Folder
TPS62203EVM-211	Tool Folder

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- Very Quiet Power Supply for Audio Decoder and Power Amplifiers
- Small Solution Size and Cost
- Simple Design

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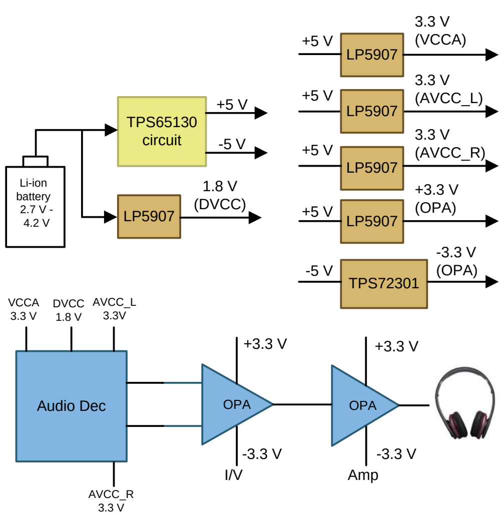
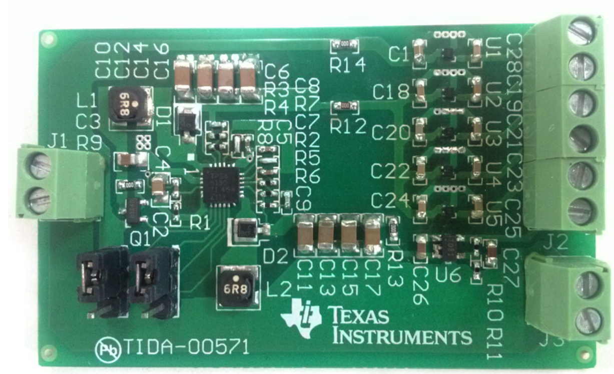
- Smartphone Hi-Fi Audio
- Battery-Powered Devices
- Portable Instruments





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1 System Description

This guide describes how LDO can be used as a quiet power for high-fidelity (hi-fi) audio in smartphone application. Smartphones usually use Li-ion battery as input voltage, which is in the range of 2.7 to 4.2 V. The supply voltage of audio decoder is 3.3 V. A DC-DC converter is used to convert battery voltage to a stable higher voltage, usually 5 V, which are with large ripple and noise. LDO is used to regulate 5 V to a quiet 3.3 V to power audio decoder. The DC-DC converter also provides ±5 V to supply the operational amplifier (OPA), but it is better to use LDO to reject ±5-V output ripple and noise, and provide quiet power for OPA.

The TIDA-00571 design provides all the design files and supporting documentation (schematic, Gerber, and test data), which can be used as a reference for power supplies for smartphone Hi-Fi audio that require low noise and excellent transient response. All files can be obtained from http://www.ti.com/tool/tida-00571.

2 Block Diagram

Figure 1 shows a high-level block diagram of the TIDA-00571 design. Note that the diagram shows an audio decoder and OPA; however, this design only covers power section while the audio decoder and OPA are not used nor populated in the design files.

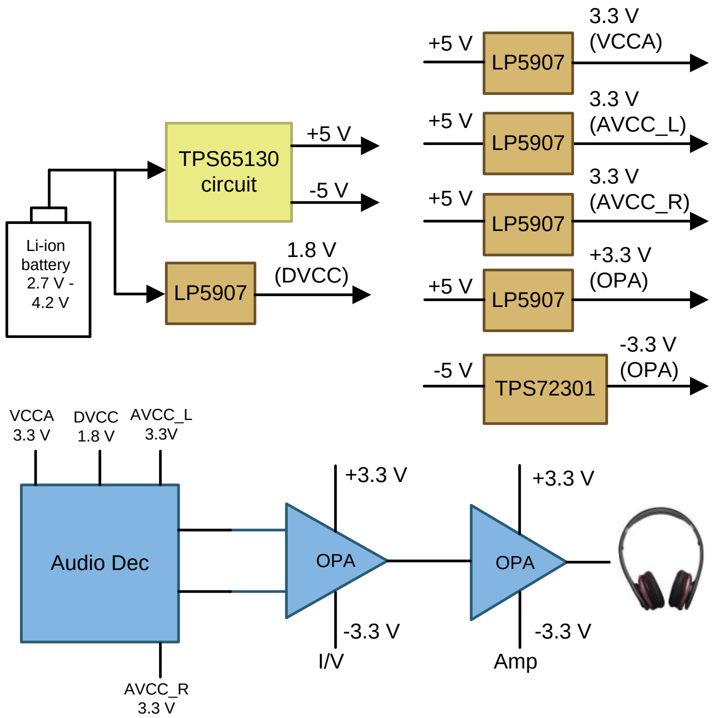

Figure 1. TIDA-00571 Block Diagram

3 Highlighted Products

The TIDA-00571 reference design features the following parts:

- LP5907: 250-mA, Ultra-Low Noise, High-PSRR Linear Regulators with Fixed Output Stable with Ceramic Output Capacitor
- TPS72301: 200-mA, Low Noise, High-PSRR Negative Output Linear Regulators with Adjustable Output Stable with Ceramic Output Capacitor
- TPS65130: Positive and Negative Output DC-DC Converter

3.1 LP5907

The LP5907 is a linear regulator capable of supplying a 250-mA output current with ultra-low noise and high PSRR ideal for RF and analog circuits.

With low noise and high PSRR, the LP5907 can suppress external noise and offers a very clean output, which is suitable for noise sensitive applications.

The LP5907 uses new, innovative design techniques to offer class-leading noise performance without a noise bypass capacitor, which reduces the BOM cost. In addition, the output tolerance of the LP5907 is ±2%, which can provide more accurate power for an audio decoder.

The LP5907 is designed to work with a 1-µF input and a 1-µF output ceramic capacitor. The device is available with fixed output voltages from 1.2 to 4.5 V in 25-mV steps.

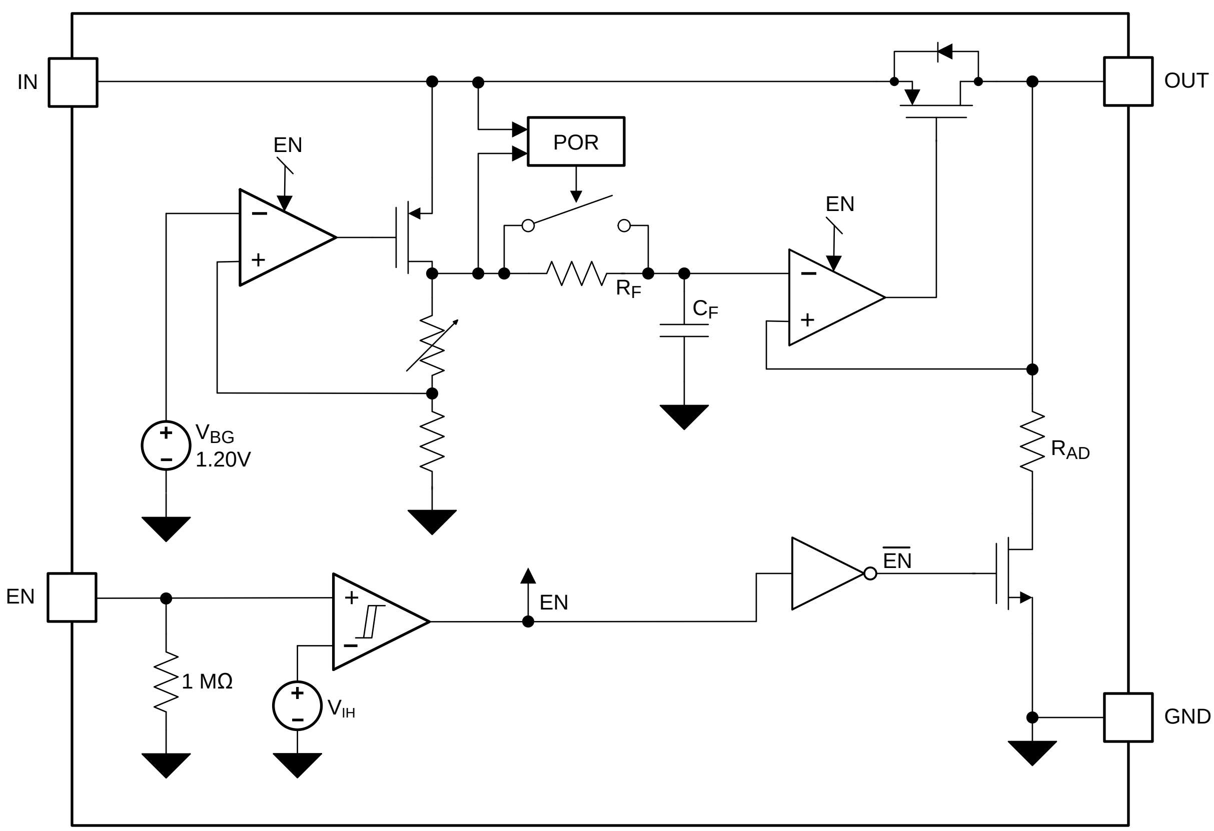

Figure 2. LP5907 Functional Block Diagram

3.1.1 Low Noise and High PSRR

- Very low noise
  - 6.5 μV_{RMS} at 1-mA load over BW = 10 Hz to 100 kHz
  - 10 μV_{RMS} at 250-mA load over BW = 10 Hz to 100 kHz
- High PSRR
  - 90 dB at 100 Hz
  - 82 dB at 1 kHz
  - 65 dB at 10 kHz

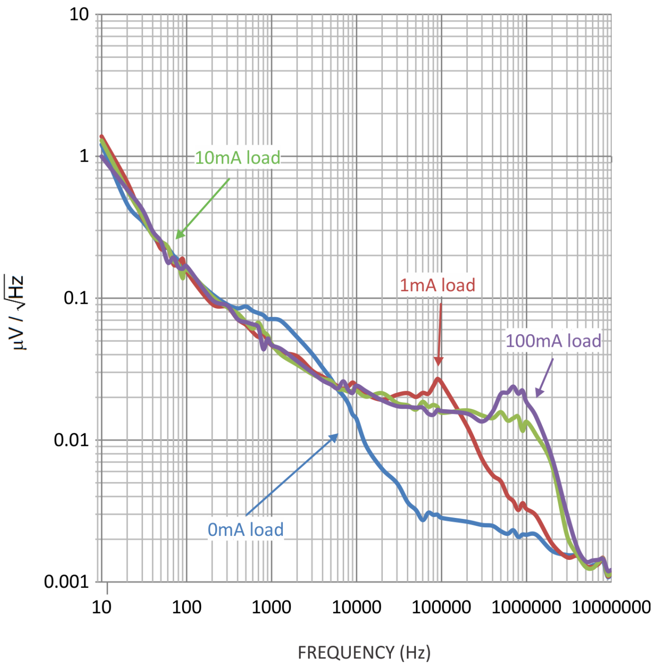

Figure 3. LP5907 Noise Density Test

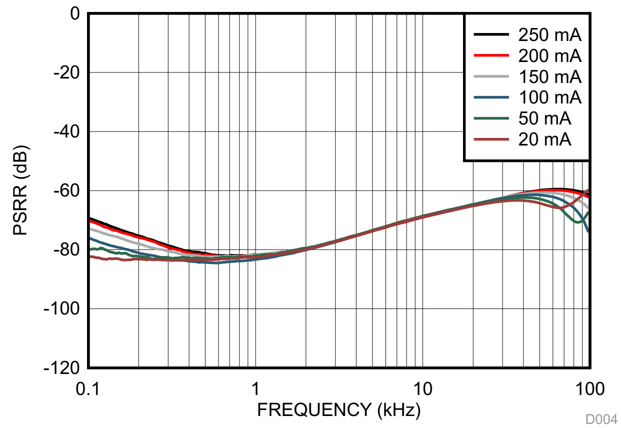

Figure 4. PSRR Loads Averaged 100 Hz to 100 kHz

3.1.2 LP5907 Solution versus Standard Buck Converter

Table 1. LP5907 Solution versus Standard Buck Converter

FEATURE	BENEFIT
Low noise and high PSRR	Quiet power to achieve high fidelity
Small solution size and cost	Competitive advantage in high-volume mobile applications

3.2 TPS72301

The TPS72301 is a negative output linear regulator capable of supplying 200-mA output current. The device offers an ideal combination of features to support low-noise applications. This device operates with input voltages from –10 to –2.7 V, and support outputs from –10 to –1.2 V. This regulator is stable with small, low-cost ceramic capacitors, and includes enable (EN) functions. Thermal short-circuit and overcurrent protections are provided by internal detection and shutdown logic. High PSRR (65 dB at 1 kHz) and low noise (60 μV_{RMS}) make the TPS72301 suitable for low-noise applications.

The TPS72301 uses a precision voltage reference to achieve a 2% overall accuracy over load, line, and temperature variations and is available in a small SOT23-5 package.

3.3 TPS65130

The TPS65130 is dual-output DC-DC converter generating a positive output voltage up to 15 V and a negative output voltage down to –15 V with output currents in a 200-mA range in typical applications, depending on input voltage to output voltage ratio. With a total efficiency up to 85%, the device is ideal for portable battery powered equipment. The input voltage range of 2.7 to 5.5 V allows the TPS65130 to be directly powered from a Li-Ion battery, from 3-cell NiMH/NiCd or alkaline batteries. The TPS65130 comes in a small 4×4-mm QFN-24 package. Together with a minimum switching frequency of 1.25 MHz, it enables designing small power supply applications since it requires only a few small external components.

The converter operates with a fixed frequency PWM control topology, and, if Power Save Mode is enabled, it uses a pulse-skipping mode at light load currents. The converter operates with only 500-µA device quiescent currents. Independent enable pins allow power-up and power-down sequencing for both outputs. The device has an internal current limit overvoltage protection and a thermal shutdown for highest reliability under fault conditions.

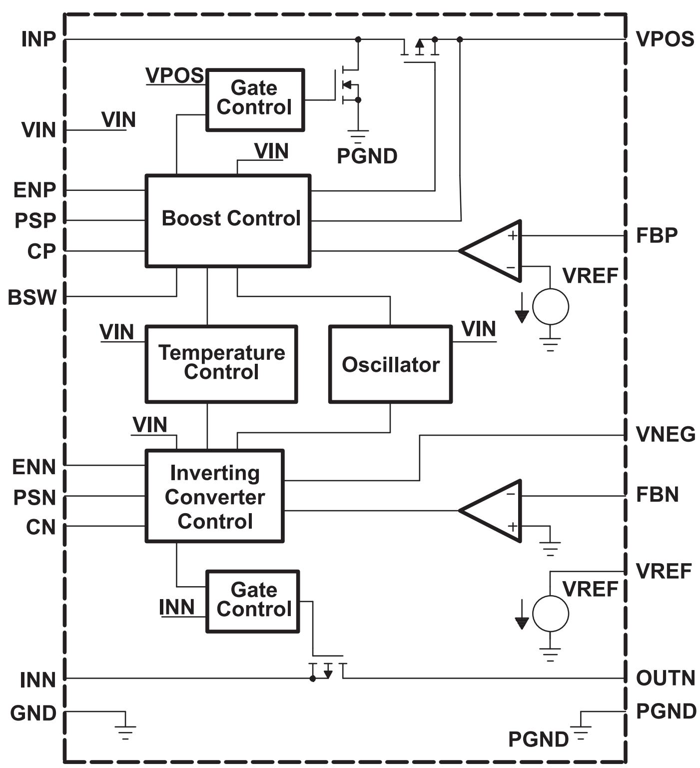

Figure 5. TPS65130 Functional Block Diagram

4 System Design Considerations and Component Selection

As shown in Figure 1, the TIDA-00571 only covers the power section while the audio decoder and OPA are not used nor populated in this design. Normally, the audio decoder needs three types of analog power and a digital power supply. Table 2 shows an example power supply rails for ES9018K2M, which is most popular in smartphone Hi-Fi audio applications. DVCC can be either 1.8 V or 3.3 V. This guide has 1.8 V selected.

Table 2. ES9018K2M Power Supply

POWER SUPPLY		VOLTAGE	CURRENT NOMINAL	CURRENT STANDBY
Digital power supply voltage	DVCC	1.8 V ± 5% 3.3 V ± 5%	13.0 mA 14.2 mA	0 mA 0 mA
Analog core supply voltage	VCCA	3.3 V ± 5%	0.8 mA	0 mA
Analog power supply voltage	AVCC_L AVCC_R	3.3 V ± 5%	3.0 mA	0 mA

Considering the Li-ion battery voltage range (usually 2.7 to 4.2 V) when the battery is in deep discharge, it cannot provide continuous 3.3 V to power audio decode directly. In hi-fi audio, it is necessary to keep the power supply ripple and noise from interacting with the audio decoder output to achieve high fidelity. The problem is resolved by boosting battery voltage to a stable higher voltage, usually 5 V, and then stepped down to 3.3 V. The buck converter and LDO can be used to step down the voltage. Comparing the buck converter, LDO with low-noise, and high PSRR can provide a much cleaner output, and the LDO does not need many external components, which simplifies the design.

The power amplifier follows audio decoder, which amplifies decoded sound and delivers that to headphones. Normally, headphones used for portable applications typically have low impedances (16 or 32 Ω) and therefore do not require high voltages to produce loud sounds. For example, to deliver 10 mW to 32-Ω headphones only requires an output voltage of .566 V_{RMS}, typical line-level audio (4 dBu, 1.228 V_{RMS}) is used as the maximum output voltage. The power amplifier needs positive and negative power supply, the positive supply can come from boost output, however for negative supply, it can be generated by buck-boost converter. The TPS65130 integrates boost and buck-boost controllers into a small package. The output of positives and negatives of the TPS65130 can be adjusted by external divided resistors. The TPS65130 can output ±5 V to supply power amplifier directly; however, for a better output in hi-fi, the LDO can be used to step down boost and buck-boost outputs to generate clean power supply for the amplifier.

To deliver 10 mW to a 16-Ω load, the amplifier must output:

$$I_{OUT} = \sqrt{\frac{P_{OUT}}{R_L}} = \sqrt{\frac{10\ mW}{16}} = 25\ mA_{RMS} \tag{1}$$

The design considerations on this section apply to the given parameters. If the design requires other parameters than those stated in this document, review the ratings on the datasheet of the mentioned devices, consider using an alternative part from Section 3, or perform an easy parametric search at www.ti.com/ldo.

4.1 Capacitor Selection for LP5907

Input and output capacitors are necessary for loop stability and eliminate high-frequency noise. These capacitors must be selected correctly for good performance.

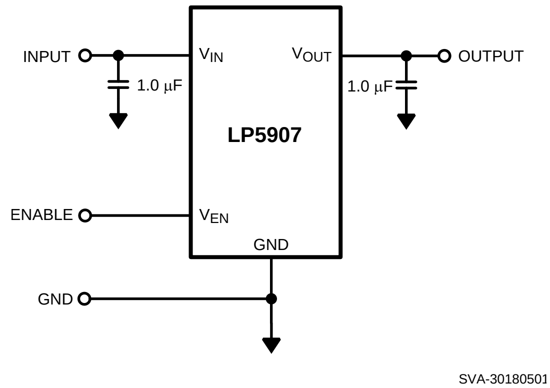

Figure 6. LP5907 Typical Application Circuit

4.1.1 Input Capacitor

An input capacitor is required for stability. The input capacitor should be at least equal to, or greater than, the output capacitor for good load transient performance. A capacitor of at least 1 µF has to be connected between the LP5907 input pin and ground for stable operation over a full-load current range. Basically, it is fine to have more output capacitance than input as long as the input is at least 1 µF. The input capacitor must be located at a distance of no more than 1 cm from the input pin and returned to a clean analog ground. Any good quality ceramic, tantalum, or film capacitor may be used at the input.

To ensure a stable operation, employ good PCB practices to minimize ground impedance and keep input inductance low. If these conditions cannot be met, or if long leads are to be used to power source to the LP5907, then increase the input capacitor to at least 10 µF. Also, tantalum capacitors can suffer catastrophic failures due to surge current when connected to a low-impedance source of power. If a tantalum capacitor is used at the input, it must be verified by the manufacturer to have a surge current rating sufficient for the application. Consider the initial tolerance, applied voltage de-rating, and temperature coefficient when selecting the input capacitor to ensure the actual capacitance is never less than 0.7 µF over the entire operating range.

4.1.2 Output Capacitor

The LP5907 is designed specifically to work with a very small ceramic output capacitor, typically 1 μF. A ceramic capacitor (dielectric types X5R or X7R) in the 1- to 10-μF range, and with ESR between 5 to 500 mΩ, is suitable in the LP5907 application circuit. For this device, the output capacitor should be connected between the OUT pin and a good connection back to the GND pin. It may also be possible to use tantalum or film capacitors at the device output, V_{OUT}, but these are not as attractive due to its size and cost (see Section 4.1.3).

The output capacitor must meet the requirement for the minimum value of capacitance and have an ESR value that is within the range 5- to 500-mΩ range for stability. Like the input capacitor, consider the initial tolerance, applied voltage de-rating, and temperature coefficient when selecting the input capacitor to ensure the actual capacitance is never less than 0.7 μF over the entire operating range.

4.1.3 Capacitor Characteristics

The LP5907 is designed to work with ceramic capacitors on the input and output to take advantage of the benefits they offer. For capacitance values in the 1- to 10-μF range, ceramic capacitors are the smallest, least expensive, and have the lowest ESR values, making them best for eliminating high-frequency noise. The ESR of a typical 1-μF ceramic capacitor is in the range of 20 to 40 mΩ, which easily meets the ESR requirement for stability for the LP5907.

A better choice for temperature coefficient in a ceramic capacitor is X7R. This type of capacitor is the most stable and holds the capacitance within ±15% over the temperature range. Tantalum capacitors are less desirable than ceramic for use as output capacitors because they are more expensive when comparing equivalent capacitance and voltage ratings in the 1- to 10-μF range.

Another important consideration is that tantalum capacitors have higher ESR values than equivalent size ceramics. This means that while it may be possible to find a tantalum capacitor with an ESR value within the stable range, it would have to be larger in capacitance (which means bigger and more costly) than a ceramic capacitor with the same ESR value. Also note that the ESR of a typical tantalum increases about 2:1 as the temperature goes from 25°C down to –40°C, so some guard band must be allowed.

4.2 Component Selection for TPS72301

The TPS72301 is an adjustable regulator. The output voltage is typically set by an external resistor divider at the adjustable pin. Use the equations in Figure 7 to determine the values for the resistor divider.

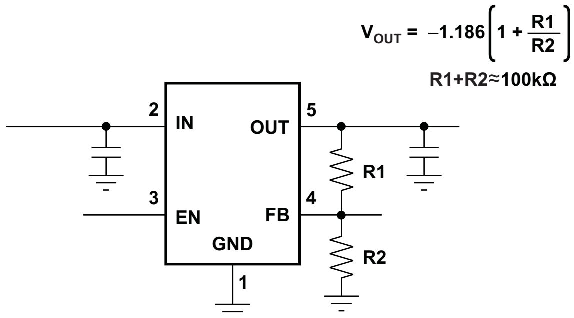

Figure 7. TPS72301 Typical Configuration

The TPS72301 datasheet specifies that VFB is typically –1.186 from the FB pin to ground. For V_{OUT} = –3.3 V, a 63.4 kΩ resistor is selected for R1, and a 35.7 kΩ resistor is selected for R2. Appropriate input and output capacitors should be used for the intended application. The TPS72301 requires a 2.2-µF ceramic output capacitor to be used for stable operation.

4.3 Component Selection for TPS65130

The design considerations on this section apply to the given parameters. If the design requires other parameters than the stated in this guide, review the ratings and specs on the datasheets of the mentioned devices.

Table 3. Design Parameters

DESIGN PARAMETERS	VALUE
Input voltage	2.7 to 4.2 V
Output voltage (positive)	5 V
Maximum Output current (positive)	200 mA
Output voltage (negative)	-5 V
Maximum Output current (negative)	-200 mA

4.3.1 Typical Application Schematic

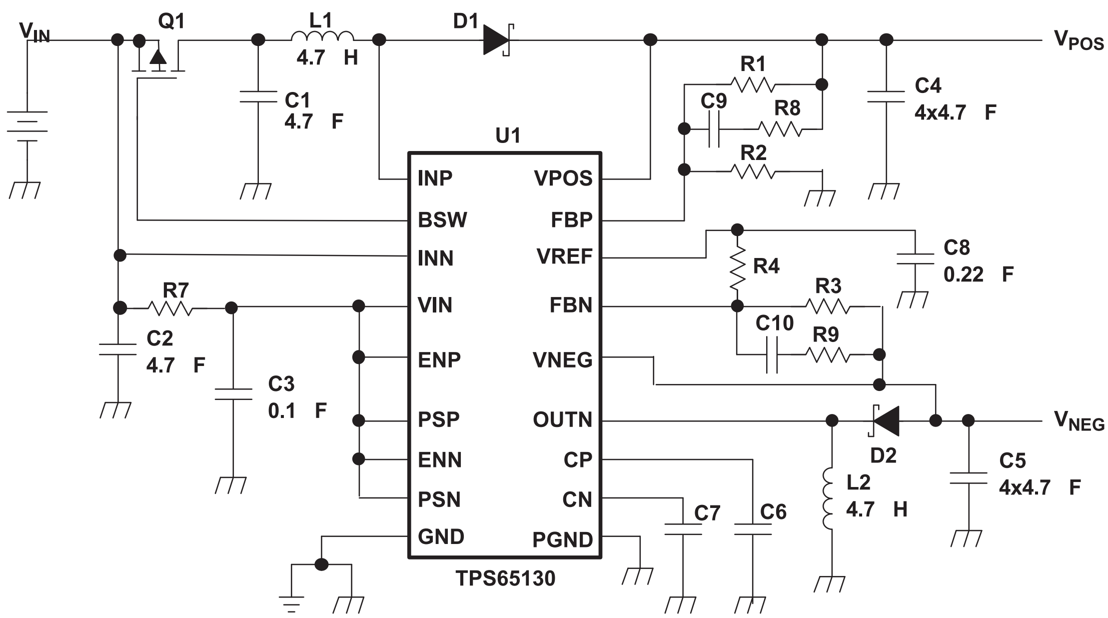

Figure 8. TPS65130 Application Schematic

4.3.2 Setting the Output Voltage

The output voltage of the TPS65130 boost converter stage can be adjusted with an external resistor divider connected to the FBP pin. The typical value of the voltage at the FBP pin is the reference voltage, which is 1.213 V. The output voltage at the boost converter is 5 V. To achieve an appropriate accuracy, the current through the feedback divider should be about 100 times higher than the current into the FBP pin. A typical current into the FBP pin is 0.05 µA, and the voltage across R2 is 1.213 V. Based on those values, the recommended value for R2 should be lower than 200 kΩ to set the divider current at 5 µA or higher. Depending on the needed output voltage (V_{POS}), the value of the resistor R1 can then be calculated using Equation 2:

$$R1 = R2 \times \left(\frac{V_{POS}}{V_{REF}} - 1 \right) \tag{2}$$

For a 5-V output, if a resistor of 180 kΩ has been chosen for R2, a 560-kΩ resistor is needed to program the desired output voltage.

The output voltage of the TPS65130 inverting converter stage can also be adjusted with an external resistor divider. It must be connected to the FBN pin. In difference to the feedback divider at the boost converter, the reference point of the feedback divider is not GND but V_{REF}. Therefore, the typical value of the voltage at the FBN pin is 0 V. The output voltage at the inverting converter is –5 V. Considerations for the feedback divider current are similar to the considerations at the boost converter. For the same reasons, the feedback divider current should be in the range of 5 µA or higher. The voltage across R4 is 1.213 V. Based on those values, the recommended value for R4 should be lower than 200 kΩ to set the divider current at the required value. The value of the resistor R3, depending on the needed output voltage (V_{NEG}), can be calculated using Equation 3:

$$R3 = R4 \times \left(\frac{V_{REF} - V_{NEG}}{V_{REF}} - 1 \right) \tag{3}$$

For a –5-V output, if a resistor of 162 kΩ has been chosen for R4, a 665-kΩ resistor is needed to program the desired output voltage.

4.3.3 Inductor Selection

An inductive converter normally requires two main passive components for storing energy during the conversion. In selecting the right inductor, it is recommended to keep the possible peak inductor current below the current-limit threshold of the power switch in the chosen configuration. For example, the current-limit threshold of the switch for the boost converter and for the inverting converter is nominally 800 mA at TPS65130. The highest peak current through the switches and the inductor depend on the output load, the input voltage (V_{IN}), and the output voltages (V_{POS}, V_{NEG}). Estimation of the peak inductor current in the boost converter can be done using Equation 4. Equation 5 shows the corresponding formula for the inverting converter.

$$I_{LP} = \frac{V_{POS}}{V_{IN} \times 0.64} \times I_{OUT\_P} \tag{4}$$

$$I_{LN} = \frac{V_{IN} - V_{NEG}}{V_{IN} \times 0.64} \times I_{OUT\_N} \tag{5}$$

The second parameter for choosing the inductor is the desired current ripple in the inductor. Normally, it is advisable to work with a ripple of less than 20% of the average inductor current. A smaller ripple reduces the losses in the inductor, as well as output voltage ripple and EMI. But in the same way, output voltage regulation gets slower, causing higher voltage changes at fast load changes. In addition, a larger inductor usually increases the total system cost. Keeping those parameters in mind, the possible inductor value can be calculated using Equation 6 for the boost converter and Equation 7 for the inverting converter.

$$L_P = \frac{V_{IN} \times (V_{POS} - V_{IN})}{\Delta I_{LP} \times f_S \times V_{POS}} \tag{6}$$

$$L_N = \frac{V_{IN} \times V_{NEG}}{\Delta I_{LN} \times f_S \times \left(V_{NEG} - V_{IN}\right)} \tag{7}$$

Parameter f is the switching frequency and ΔI_L is the ripple current in the inductor, that is, 25% x I_L. V_{IN} is the input voltage. So, the calculated inductance value for the boost inductor is 6.9 µH and for the inverting converter inductor is 7.3 µH. With these calculated values and the calculated currents, it is possible to choose a suitable inductor. In typical applications, a 6.8-µH inductor is recommended. The device has been optimized to work with inductance values between 3.3 and 6.8 µH. Nevertheless, operation with higher inductance values may be possible in some applications. Detailed stability analysis is then recommended.

4.3.4 Input and Output Capacitors

At least a 4.7-μF input capacitor is recommended for the input of the boost converter and for the input of the inverting converter (INN) to improve transient behavior of the regulators and EMI behavior of the total power supply circuit. A ceramic capacitor or a tantalum capacitor with a smaller ceramic capacitor (100 nF) in parallel, placed close to the input pins, is recommended.

One of the major parameters necessary to define the capacitance value of the output capacitor is the maximum allowed output voltage ripple of the converter. This ripple is determined by the capacitance and the ESR. Due to its low ESR, V_{RIPPLE_ESR} could be neglected for ceramic capacitors. It is possible to calculate the minimum capacitance needed for the defined ripple by using Equation 8 for the boost converter output capacitor and Equation 9 for the inverting converter output capacitor.

$$C_{min\_P} = \frac{I_{OUT\_P} \times \left(V_{POS} - V_{IN}\right)}{f_S \times \Delta V_P \times V_{POS}} \tag{8}$$

$$C_{min\_N} = \frac{I_{OUT\_N} \times V_{NEG}}{f_S \times \Delta V_N \times \left(V_{POS} - V_{IN}\right)} \tag{9}$$

Parameter f is the switching frequency and ΔV is the maximum allowed ripple. With a chosen ripple voltage in the range of 5 mV, a minimum capacitance of 13 μF is needed. Note that capacitor degradation increases the ripple greatly.

4.3.5 Feedback Divider

To speed up the control loop, feedforward capacitors are recommended in the feedback divider, parallel to R1 (boost converter) and R3 (inverting converter). Equation 10 shows how to calculate the appropriate value for the boost converter, and Equation 11 for the inverting converter.

$$C9 = \frac{6.8\ \mu s}{R1} \tag{10}$$

$$C10 = \frac{6.8\ \mu s}{R2} \tag{11}$$

To avoid coupling noise into the control loop from the feedforward capacitors, the feedforward effect can be bandwidth-limited by adding a series resistor. Any value between 10 and 100 kΩ is suitable. The higher the resistance, the lower the noise coupled into the control loop system.

4.3.6 Compensation Capacitors

The control loops of both converters are completely compensated internally. The complex internal input voltage, output voltage, and input-current feedforward system has built-in error correction, which requires external capacitors. A 10-nF capacitor at CP of the boost converter and a 4.7-nF capacitor at CN of the inverting converter are recommended.

5 Test Setup and Results

NOTE: The TIDA-00571 EVM is not available for purchase; however, reference design files can be downloaded at http://www.ti.com/tool/tida-00571.

5.1 *Buck Converter versus LDO Output*

Since TPS65130 boosts the battery voltage to 5 V, buck converter or LDO can be used to step down the boost output to supply the audio decoder. Two test setups were made to quantify the benefits of using LDO: The first setup used a buck converter to step down the TPS65130 boost output to generate 3.3 V and 1.8 V for audio decoder with the required load. The second setup used LDO to step down the TPS65130 boost output to generate the same output voltage and load. Then the results from those two scenarios were compared to determine which setup yields a cleaner output. A low-output voltage ripple and noise means that the output will be better.

The test equipment is as in Table 4.

Table 4. Test Equipment

EQUIPMENT	DEVICE NUMBER	FUNCTION
Voltage supply	Agilent E3631A	Constant voltage supply 3.6-V DC For uniformity, consistent power supply was used instead of a battery
Digital multimeter	Agilent 34401A	Used to measure buck or LDO output ripple voltage
Spectrum analyzer	Agilent 4395A	Used to measure buck or LDO output noise

5.1.1 Buck Converter Setup and Output Results

The typical buck converter TPS6220x was taken to step down the TPS65130 boost output. This design guide uses the TPS62203EVM-211, which can provide 3.3 V. Table 5 lists the specifications:

Table 5. TPS62200/TPS62203EVM-211 Specification

SPECIFICATION	TEST CONDITIONS	MIN	TYP	MAX	UNIT
Input voltage range	TPS62203EVM-211	3.3		6	V
Output voltage range	TPS62203EVM-211		3.3		V
Output current		0		300	mA

Figure 9 represents the test setup for using the TPS62203EVM-211 to step down the TPS65130 boost output. The TPS65130 circuit design is calculated in Section 4.1, the load is as listed in Table 2.

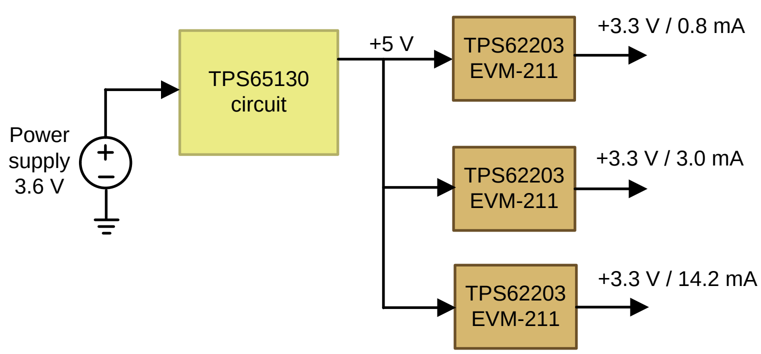

Figure 9. Test Setup — Buck Converter Output

The buck converter output was measured in Table 6.

Table 6. Buck Converter Output - Test Results

LOAD OF BUCK CONVERTER	OUTPUT VOLTAGE [V]	OUTPUT NOISE (10 Hz to 100 kHz) [μV_{RMS}]
3.3 V/0.8 mA	3.33	7036
3.3 V/3 mA	3.33	6943
3.3 V/14.2 mA	3.33	6451

5.1.2 LDO Output Results

Figure 10 represents the test setup for using LDO to step down TPS65130 boost output. The LDO circuit design is as in Section 4.2, the load is the same as list in Table 2.

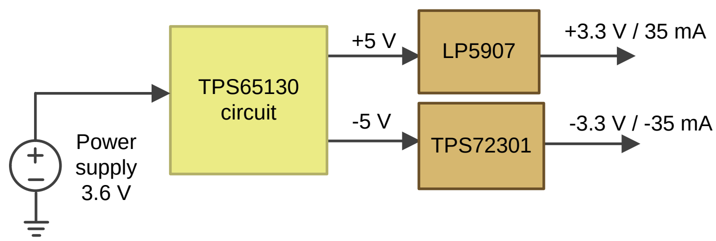

Figure 10. Test Setup — LDO Output

The LDO output was measured in Table 7.

Table 7. LDO Output - Test Results

LOAD OF LDO	OUTPUT VOLTAGE [V]	OUTPUT NOISE (10 Hz to 100 kHz) [μV_{RMS}]
3.3 V/0.8 mA	3.30	9.33
3.3 V/3 mA	3.30	8.64
3.3 V/14.2 mA	3.30	12

5.1.3 Comparison

By comparing Table 6 and Table 7 and their output noise spectral density as below, the advantages of using LDO over buck converter are as follows:

- V_{IN} = 5 V, V_O = 3.3 V, I_O = 0.8 mA

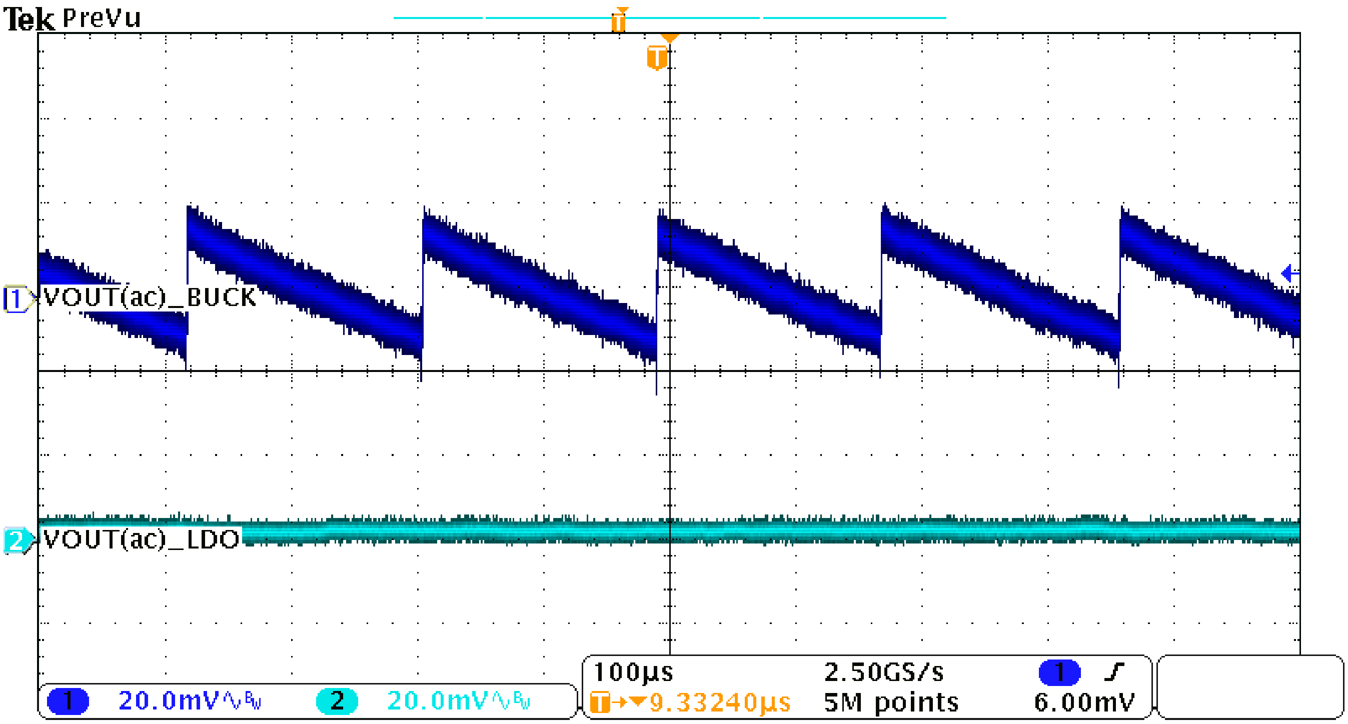

Figure 11. Ripple Comparison at 3.3 V/0.8 mA

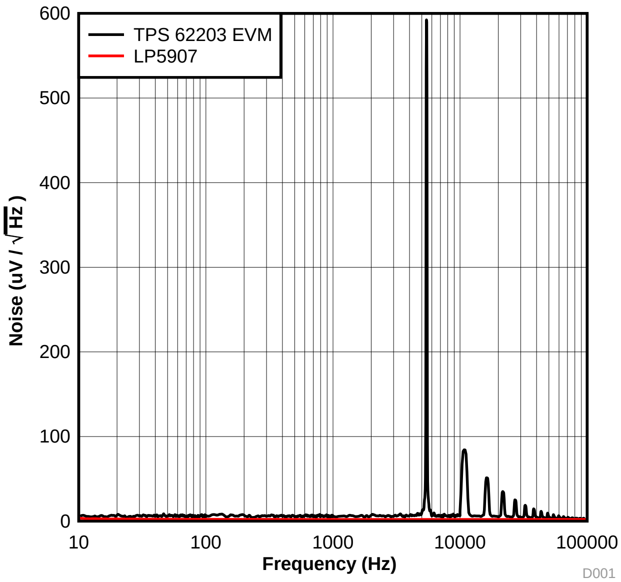

Figure 12. Noise Comparison at 3.3 V/0.8 mA

- V_{IN} = 5 V, V_O = 3.3 V, I_O = 3 mA

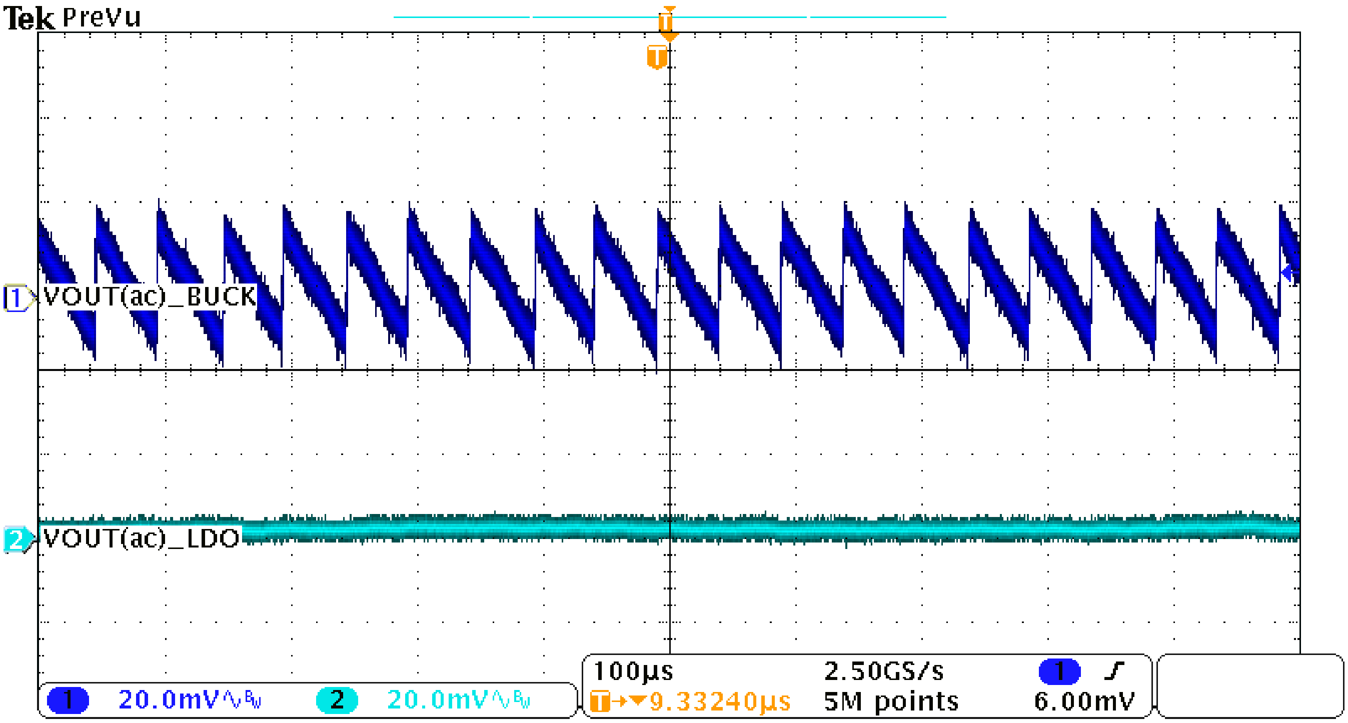

Figure 13. Ripple Comparison at 3.3 V/3 mA

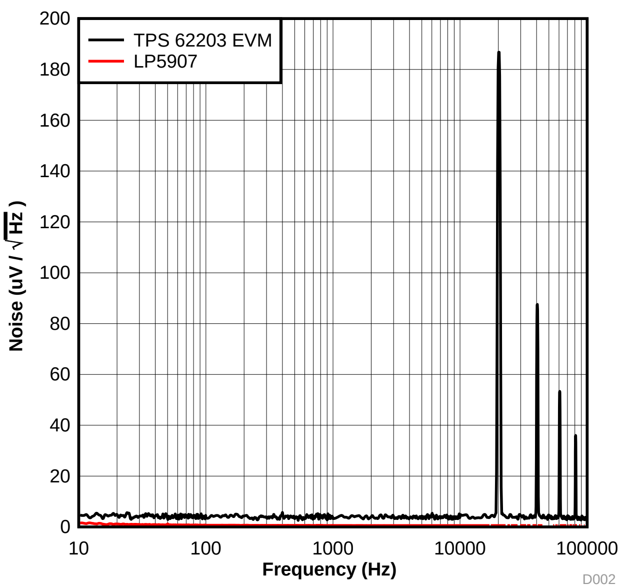

Figure 14. Noise Comparison at 3.3 V/3 mA

- V_{IN} = 5 V, V_O = 3.3 V, I_O = 14.2 mA

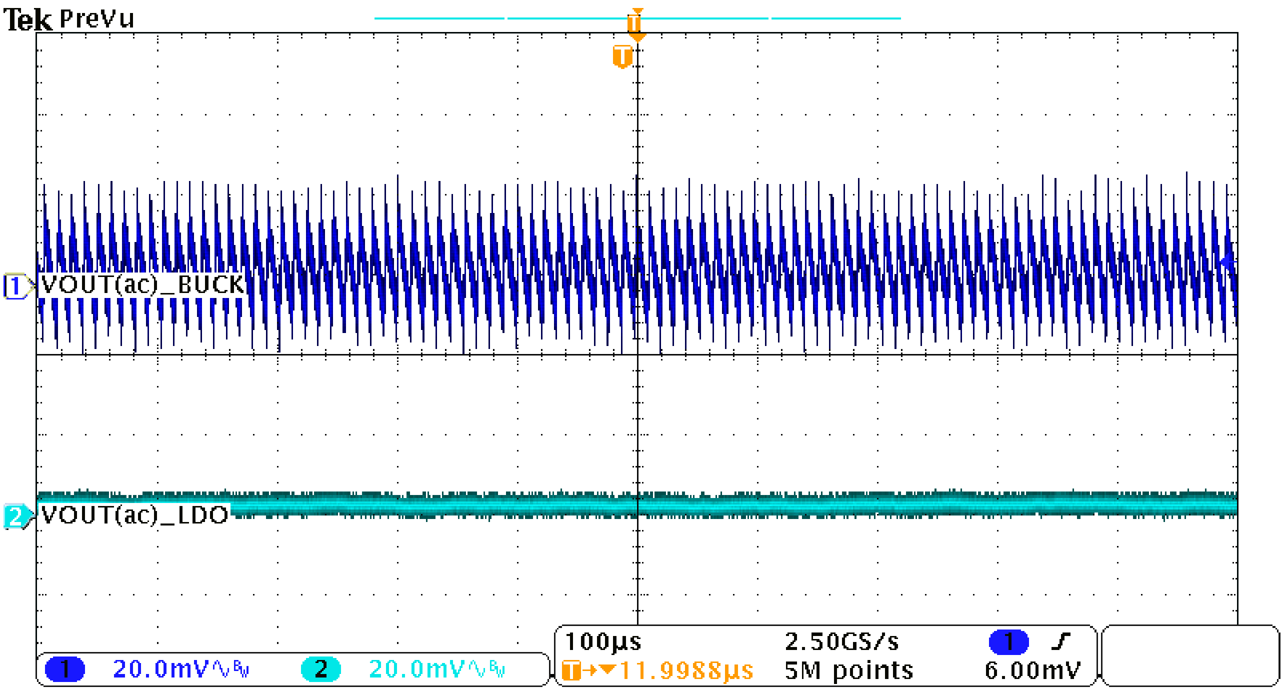

Figure 15. Ripple Comparison at 3.3 V/14.2 mA

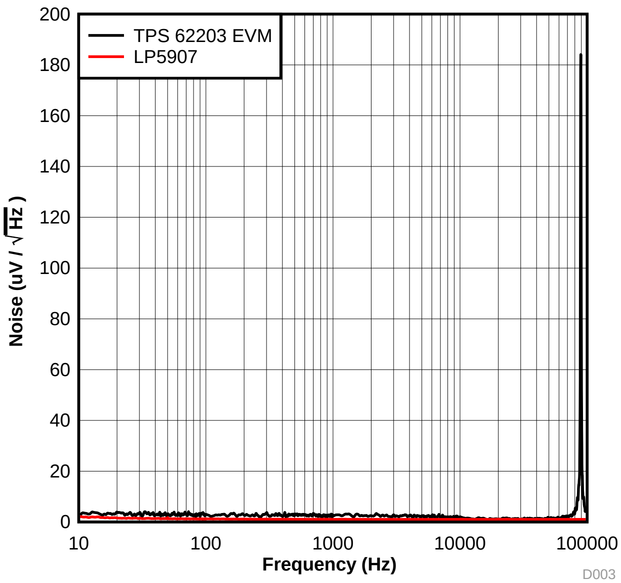

Figure 16. Noise Comparison at 3.3 V/14.2 mA

- Due to switching, buck represents a larger ripple than LDO. When the load is light, buck converters work during power save mode; its switching frequency is reduced, which represents big noise spurs in noise curves. Especially at a 0.8- or 3-mA very light load, the spurs occur in audio bands from 20 Hz to 20 kHz, which harms audio output. By comparing Table 6 and Table 7 and the ripple and noise curves, the LP5907 output ripple and noise is much lower than the buck converter, which almost eliminates the power supply impact to audio output.
- Meanwhile, the LP5907 solution size is smaller than the buck converter. The LP5907 needs two small capacitors; however, the buck converter normally needs two capacitors and a inductor, which means added cost and size.

5.2 TPS65130 Output versus LDO Output

The TPS65130 output is ±5 V, which can supply power amplifier directly; however, LDO can also be used after the TPS65130 output to generate quiet power for amplifier. Two test setups were made to quantify the benefits of using LDO. The first setup used TPS65130 boost and buck-boost output directly to supply amplifier.

The second setup used LDO after the TPS65130 boost and buck-boost output to supply amplifier, in this setup, the supply voltage for amplifier is lower than the first setup, but it is enough for amplifier. The loading for the two setups is as the assumption of delivering 10 mW to a 16-Ω load, the maximum output current is

$$I_{OUT} = \sqrt{\frac{P_{OUT}}{R_L}} \times \sqrt{2} = \sqrt{\frac{10\ \text{mW}}{16}} \times \sqrt{2} = 35\ \text{mA} \quad (12)$$

The test equipment is as the same as in Table 4.

5.2.1 TPS65130 Output Results

Figure 17 represents the test setup for using the TPS65130 output to supply amplifier. The TPS65130 circuit design is calculated in Section 4.1.

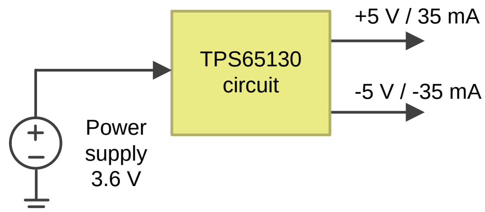

Figure 17. TPS65130 Output for Amplifier

The TPS65130 output was measured in Table 8.

Table 8. TPS65130 Output Test Results

TPS65130 OUTPUT	OUTPUT VOLTAGE [V]	OUTPUT NOISE (10 Hz to 100 kHz) [μV_{RMS}]
5 V/35 mA	4.94	9585
–5 V/–35 mA	–5.0	199

5.2.2 LDO Output Results

Figure 18 represents the test setup for using LDO after the TPS65130 to supply amplifier. The LDO output voltage ripple was measured on Table 8. The TPS65130 circuit design is calculated in Section 4.1, the LDO circuit design is as in Section 4.2, and Section 4.3.

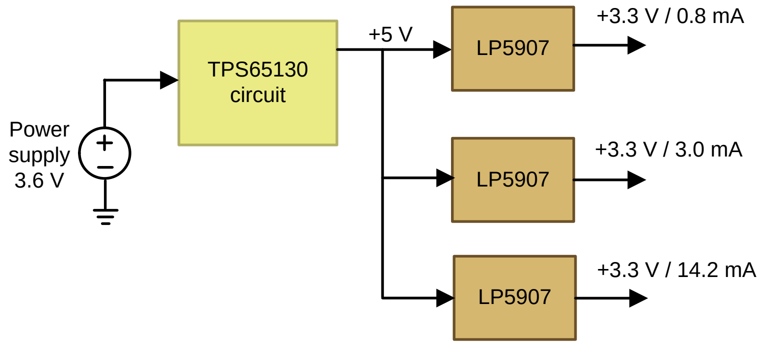

Figure 18. LDO Output for Amplifier

The LDO output was measured in Table 9.

Table 9. LDO Output After TPS65130 Test Results

LDO	OUTPUT VOLTAGE [V]	OUTPUT NOISE (10 Hz to 100 kHz) [μV_{RMS}]
LP5907 (3.3 V/35 mA)	3.33	17.8
TPS72301 (-3.3 V/-35 mA)	-3.31	163

5.2.3 Comparison

By comparing Table 8 and Table 9, and their output noise spectral density as below, the advantages of using LDO over directly using TPS65130 output are as follows:

- TPS65130 boost output and LP5907

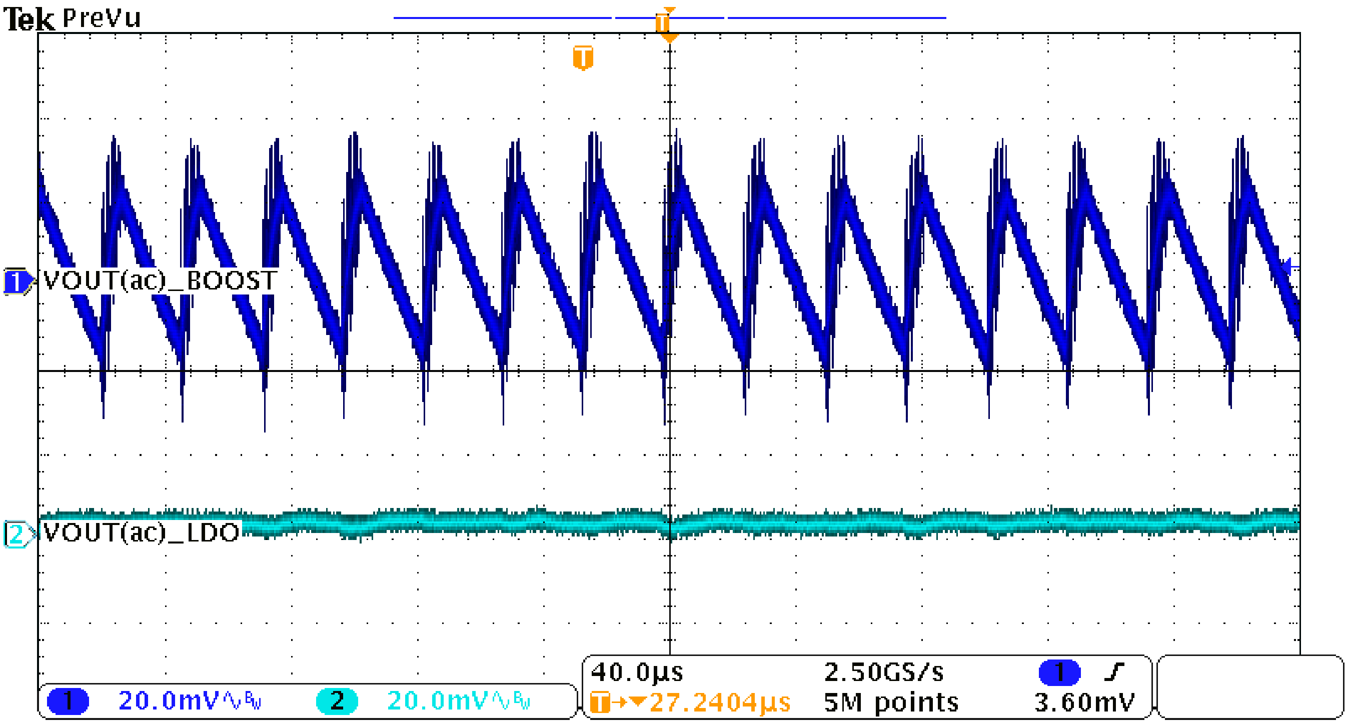

Figure 19. Ripple Comparison Between LP5907 and TPS65130 Boost Output

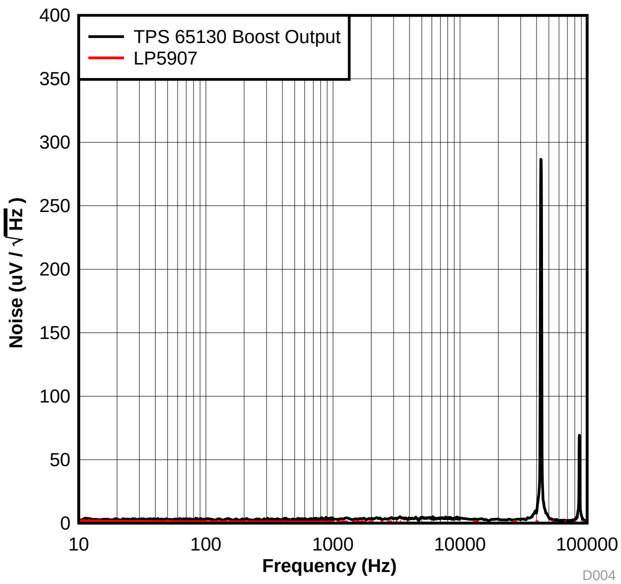

Figure 20. Noise Comparison Between LP5907 and TPS65130 Boost Output

- TPS65130 buck-boost output and TPS72301

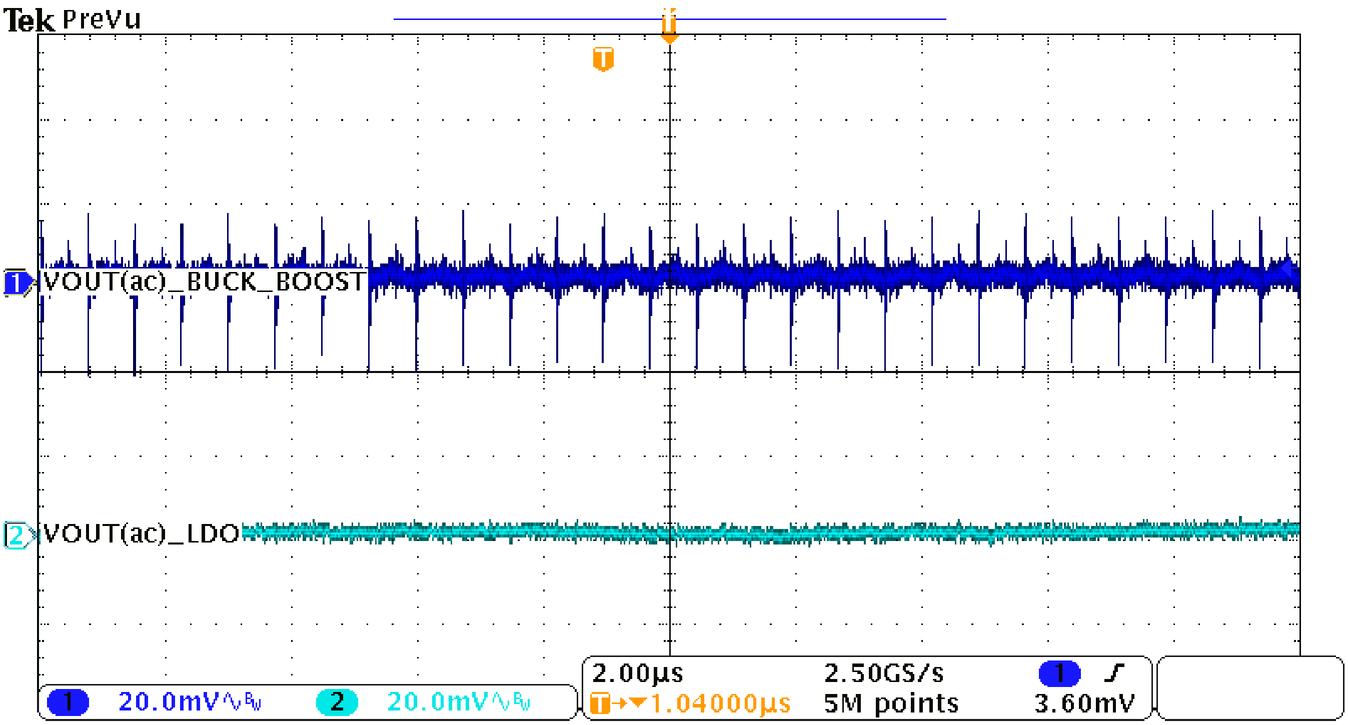

Figure 21. Ripple Comparison Between TPS72301 and TPS65130 Buck-Boost Output

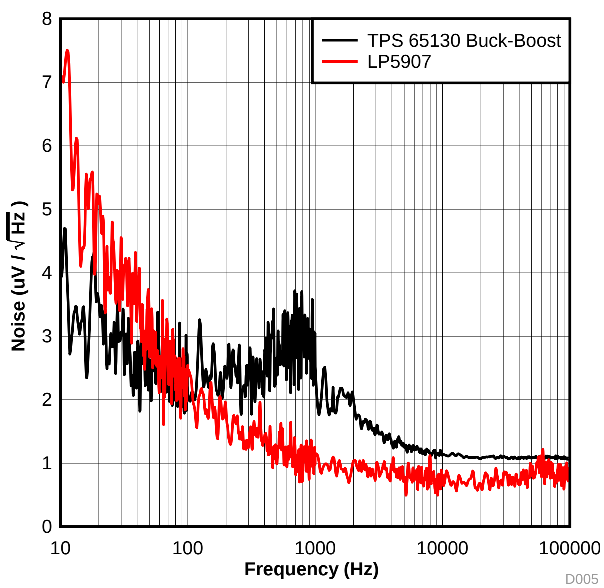

Figure 22. Noise Comparison Between TPS72301 and TPS65130 Buck-Boost Output

- Due to LDO and high PSRR, the output ripple of the TPS65130 can be almost eliminated by adding an LDO after the TPS65130 output, and the LDO output noise is much smaller than the TPS65130.
- By comparing Table 8, Table 9, and their output ripple and noise spectral densities, the LDO has a big advantage over the TPS65130 output to power amplifier. The quiet output of LDO significantly improves the final audio output noise +THD to achieve high fidelity.

6 Design Files

6.1 Schematics

To download the schematics, see the design files at TIDA-00571.

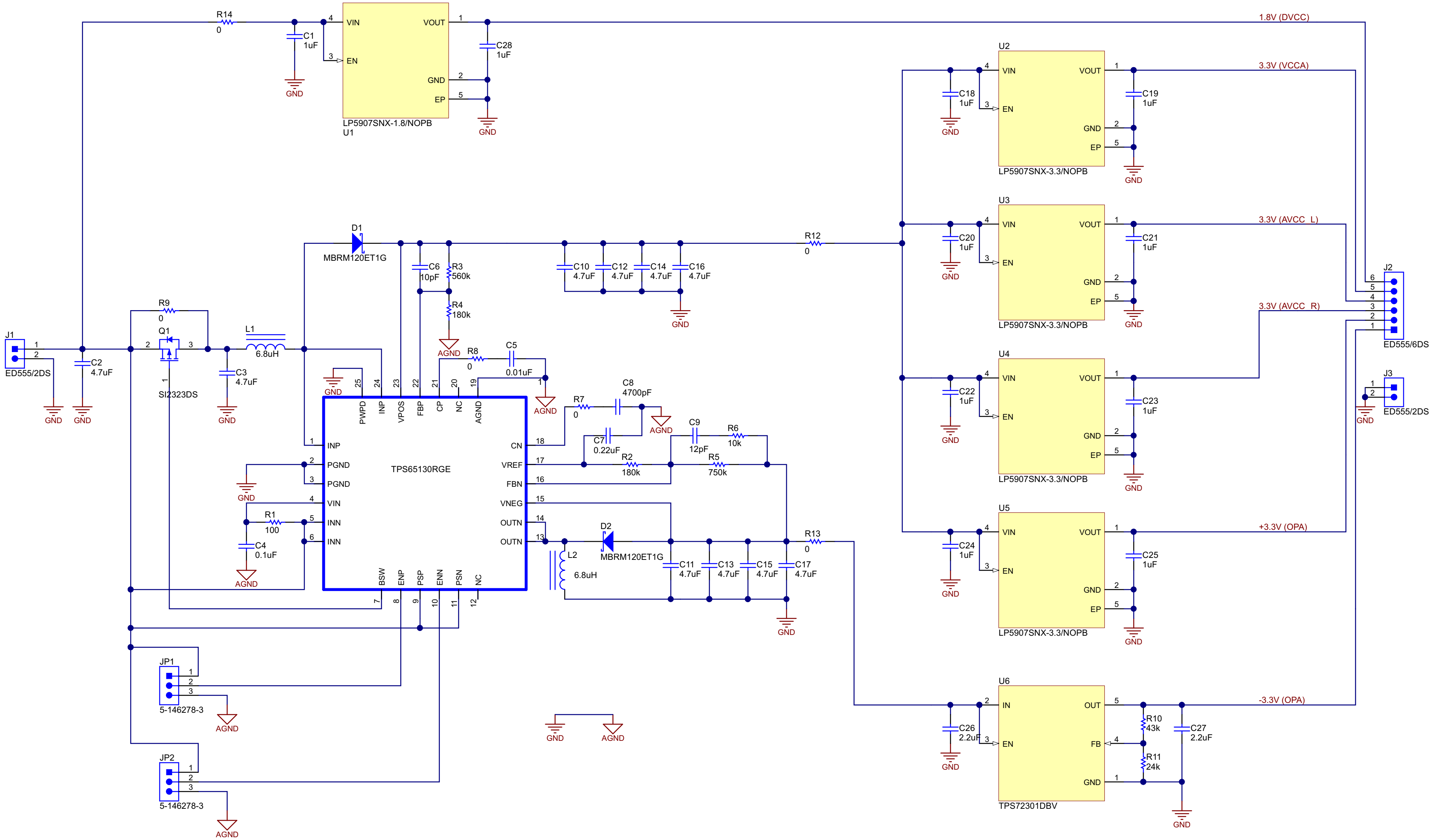

Figure 23. TIDA-00571 Schematic

6.2 Bill of Materials

To download the bill of materials (BOM), see the design files at TIDA-00571.

Table 10. BOM

ITEM #	DESIGNATOR	QTY	VALUE	PART NUMBER	MANUFACTURER	DESCRIPTION	PACKAGE REFERENCE
1	!PCB	1		TIDA-00571	Any	Printed Circuit Board	
2	1	1	TPS65130RGE	TPS65130RGE	TI	IC, Positive and Negative Output DC-DC Converter	QFN24
3	C1, C18, C19, C20, C21, C22, C23, C24, C25, C28	10	1uF	GRM188R71A105KA61D	MuRata	CAP, CERM, 1 uF, 10 V, +/- 10%, X7R, 0603	603
4	C2, C3	2	4.7uF	GRM21BR60J475KA11L	MuRata	CAP, CERM, 4.7 uF, 6.3 V, +/- 10%, X5R, 0805	805
5	C4	1	0.1uF	GRM155R71A104KA01D	MuRata	CAP, CERM, 0.1 uF, 10 V, +/- 10%, X7R, 0402	402
6	C5	1	0.01uF	GRM155R71C103KA01D	MuRata	CAP, CERM, 0.01 uF, 16 V, +/- 10%, X7R, 0402	402
7	C6	1	10pF	GRM1555C1H100JA01D	MuRata	CAP, CERM, 10 pF, 50 V, +/- 5%, C0G/NP0, 0402	402
8	C7	1	0.22uF	GRM155R60J224KE01D	MuRata	CAP, CERM, 0.22 uF, 6.3 V, +/- 10%, X5R, 0402	402
9	C8	1	4700pF	GRM155R71H472KA01D	MuRata	CAP, CERM, 4700 pF, 50 V, +/- 10%, X7R, 0402	402
10	C9	1	12pF	GRM1555C1H120JA01D	MuRata	CAP, CERM, 12 pF, 50 V, +/- 5%, C0G/NP0, 0402	402
11	C10, C11, C12, C13, C14, C15, C16, C17	8	4.7uF	GRM31CR71E475KA88L	MuRata	CAP, CERM, 4.7 uF, 25 V, +/- 10%, X7R, 1206	1206
12	C26, C27	2	2.2uF	GRM188R71A225KE15D	MuRata	CAP, CERM, 2.2 uF, 10 V, +/- 10%, X7R, 0603	603
13	D1, D2	2	20V	MBRM120ET1G	ON Semiconductor	Diode, Schottky, 20 V, 1 A, POWERMITE	POWERMITE
14	J1, J3	2		ED555/2DS	On-Shore Technology	Terminal Block, 6A, 3.5mm Pitch, 2-Pos, TH	7.0x8.2x6.5mm
15	J2	1		ED555/6DS	On-Shore Technology	Terminal Block, 6A, 3.5mm Pitch, 6-Pos, TH	20.5x8.2x6.5mm
16	JP1, JP2	2		5-146278-3	TE Connectivity	Header, 100mil, 3x1, Tin, TH	Header, 3x1, 100mil, TH
17	L1, L2	2	6.8uH	VLCF4018T-6R8NR94-2	TDK	Inductor, Shielded, Ferrite, 6.8 uH, 0.94 A, 0.131 ohm, SMD	4.0x1.8x4.0mm
18	LBL1	1		THT-14-423-10	Brady	Thermal Transfer Printable Labels, 0.650" W x 0.200" H - 10,000 per roll	PCB Label 0.650"H x 0.200"W

Table 10. BOM (continued)

ITEM #	DESIGNATOR	QTY	VALUE	PART NUMBER	MANUFACTURER	DESCRIPTION	PACKAGE REFERENCE
19	Q1	1	-20V	SI2323DS	Vishay-Siliconix	MOSFET, P-CH, -20 V, -3.7 A, SOT-23	SOT-23
20	R1	1	100	CRCW0402100RFKED	Vishay-Dale	RES, 100, 1%, 0.063 W, 0402	402
21	R2, R4	2	180k	CRCW0402180KJNED	Vishay-Dale	RES, 180 k, 5%, 0.063 W, 0402	402
22	R3	1	560k	CRCW0402560KJNED	Vishay-Dale	RES, 560 k, 5%, 0.063 W, 0402	402
23	R5	1	750k	CRCW0402750KFKED	Vishay-Dale	RES, 750 k, 1%, 0.063 W, 0402	402
24	R6	1	10k	CRCW040210K0JNED	Vishay-Dale	RES, 10 k, 5%, 0.063 W, 0402	402
25	R7, R8	2	0	CRCW04020000Z0ED	Vishay-Dale	RES, 0, 5%, 0.063 W, 0402	402
26	R9, R12, R13, R14	4	0	RC0603JR-070RL	Yageo America	RES, 0, 5%, 0.1 W, 0603	603
27	R10	1	43k	CRCW040243K0JNED	Vishay-Dale	RES, 43 k, 5%, 0.063 W, 0402	402
28	R11	1	24k	CRCW040224K0JNED	Vishay-Dale	RES, 24 k, 5%, 0.063 W, 0402	402
29	U1	1		LP5907SNX-1.8/NOPB	Texas Instruments	Ultra Low-Noise, 250-mA Linear Regulator for RF and Analog Circuits, DQN0004A	DQN0004A
30	U2, U3, U4, U5	4		LP5907SNX-3.3/NOPB	Texas Instruments	Ultra Low-Noise, 250-mA Linear Regulator for RF and Analog Circuits, DQN0004A	DQN0004A
31	U6	1		TPS72301DBV	Texas Instruments	200mA Low-Noise, High-PSRR Negative Output Low-Dropout Linear Regulators, DBV0005A	DBV0005A
32	!PCB	1		TIDA-00571	Any	Printed Circuit Board	

6.3 Layer Plots

To download the layer plots, see the design files at TIDA-00571.

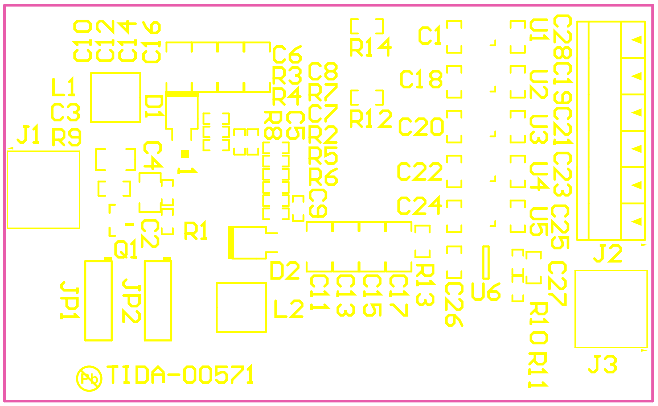

Figure 24. Top Silkscreen

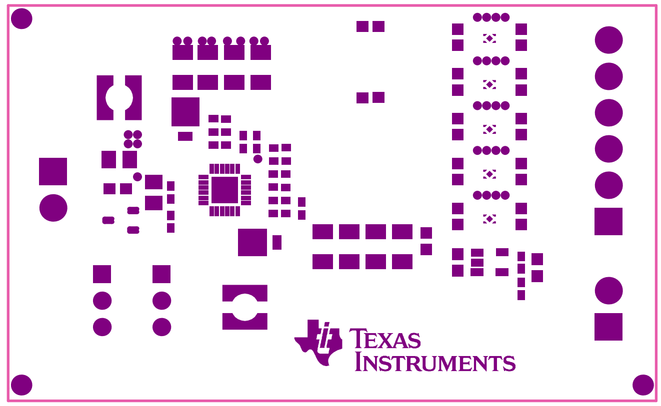

Figure 25. Top Solder Mask

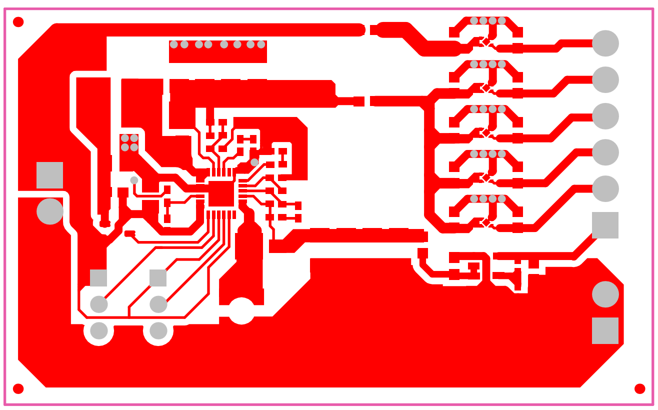

Figure 26. Top Layer

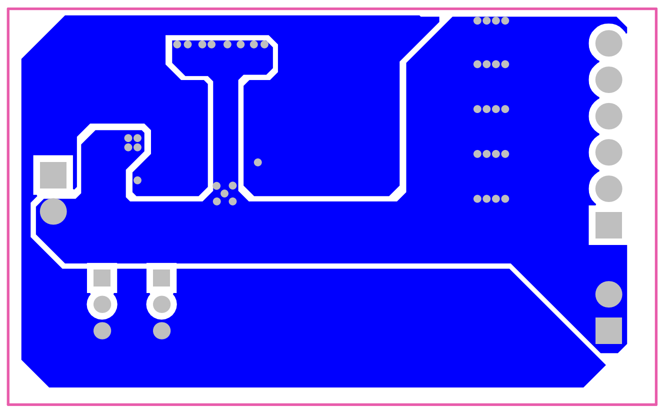

Figure 27. Bottom Layer

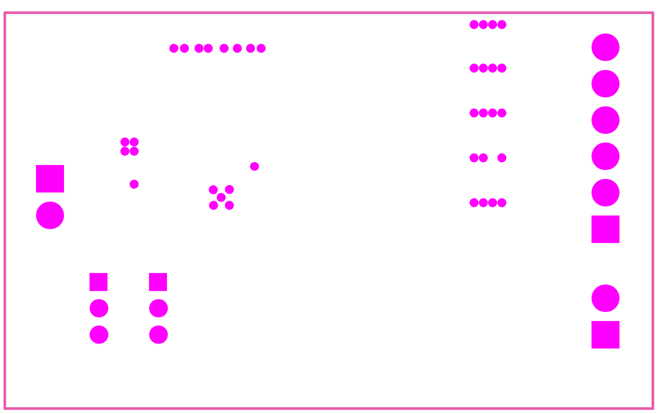

Figure 28. Bottom Solder Mask

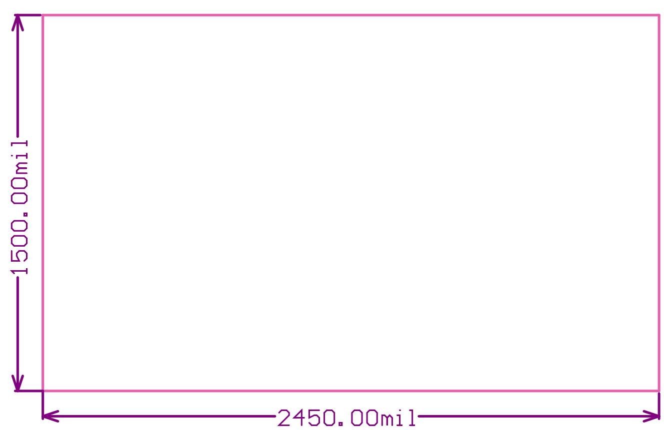

Figure 29. Mechanical Dimensions

6.4 Altium Project

To download the Altium project files, see the design files at TIDA-00571.

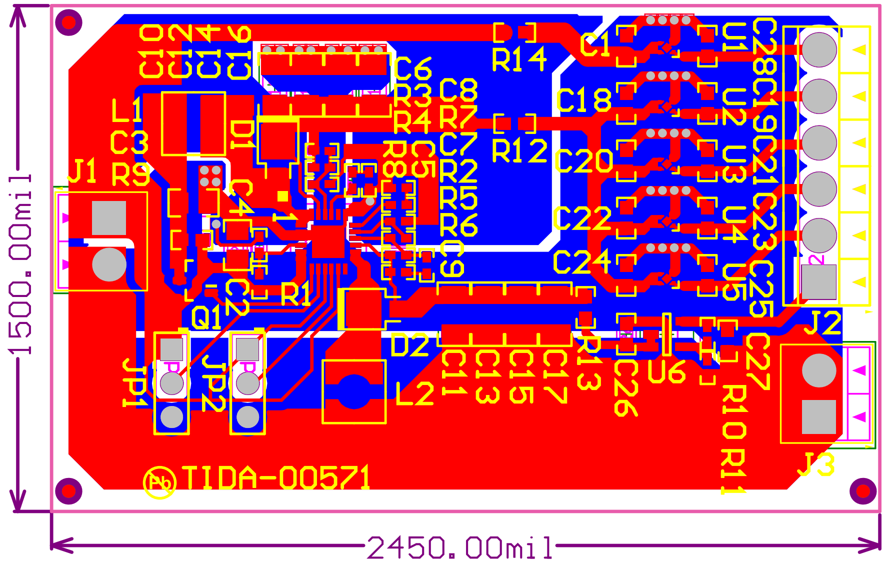

Figure 30. Multilayer Composite Print

6.5 Gerber Files

To download the Gerber files, see the design files at TIDA-00571.

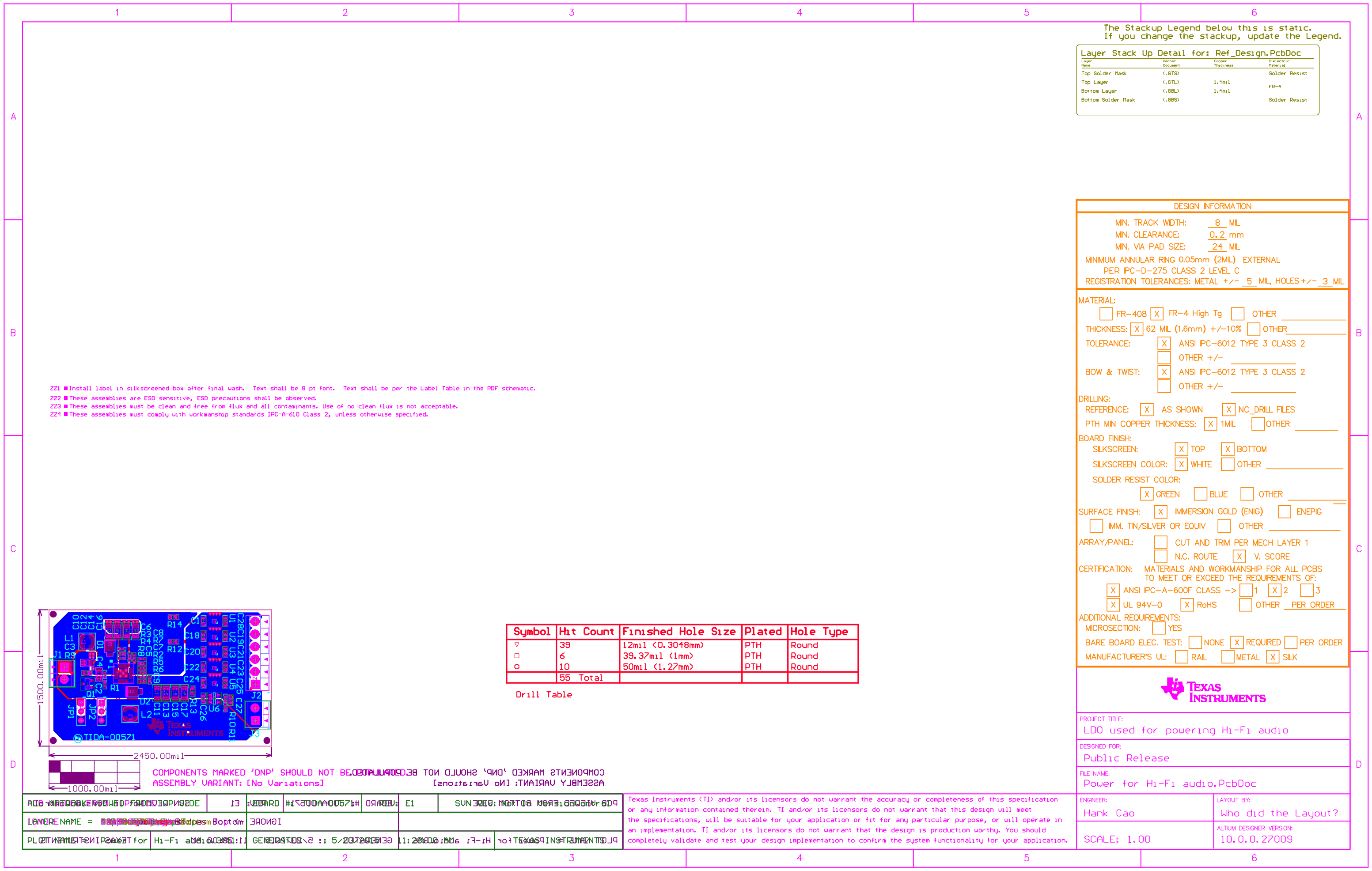

Figure 31. Fabrication Drawing

6.6 Assembly Drawings

To download the assembly drawings, see the design files at TIDA-00571.

7 About the Author

HANK CAO is a system engineer at Texas Instruments on the mobile power devices RF power group at Texas Instruments. Hank joined TI in 2012. Hank earned his Master of Power Electronics from Nanjing University of Aeronautics and Astronautics.

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