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TI Designs: PMP9774

Envelope-Tracking Power Supply Reference Design for Audio Power Amplifiers with TPS61088

TEXAS INSTRUMENTS

Designs Overview

This reference design delivers an envelope-tracking power supply circuit for audio power amplifier (PA) with TPS61088. By adding an audio envelope signal to the FB pin, the TPS61088's output voltage can change in accordance with the envelope of the audio signal. So the TPS61088 provides a dynamically changing supply voltage to the PA. Thus the PA can always keep high efficiency in the whole output power range.

Design Features

- Input voltage range: 2.7-4.2V
- Output voltage range: 5.5-11.75V
- The output voltage changes linearly with the amplitude of the audio signal
- The output voltage rise up from 5.5V to 11.75V within 1ms under a full range step up of the audio signal
- The PA can always operate at high-efficiency in the whole output power range

Design Resources

PMP9774	All Design files
TPS61088	Product Folder
OPA4377	Product Folder

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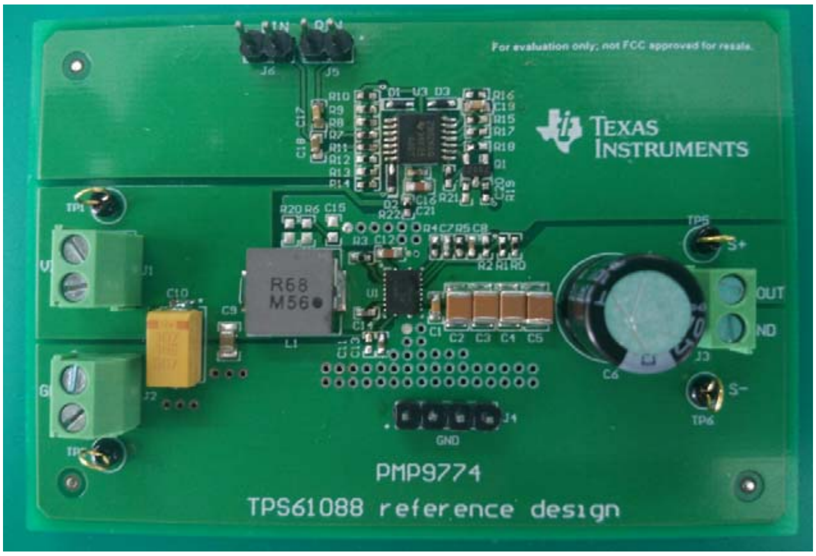

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1 Introduction

The traditional fixed power supply audio PA has been well established for many years. The shortage of this fixed power supply PA is that when the input audio signal's peak to average ratio is high, the conversion efficiency will be low. Because during the periods of the peaks, the PA requires a full voltage to be able to deliver the required power without running into distortion, but during the periods of lower signal amplitude, this full voltage is not required and keeping high power supply voltage means unnecessary power is dissipated.

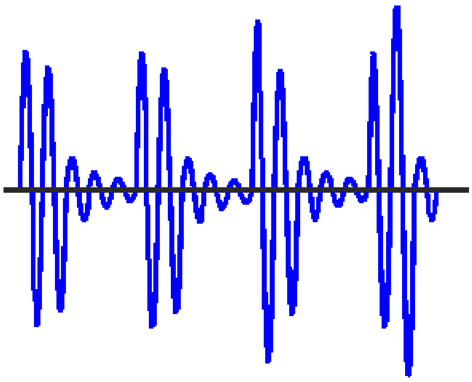

Figure 1. Audio Signal with High Peak to Average Ratio

This reference design delivers an envelope-tracking power supply circuit for the PA with TPS61088. The objective of envelope-tracking is to improve the efficiency of PA in the full range of the output power. By adding an audio envelope signal to the FB pin, the TPS61088's output voltage can change in accordance with the envelope of the audio signal. Thus the PA can always operate at high-efficiency in the whole output power range.

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2 Design Process

2.1 Specification

This reference design makes the output voltage change linearly with the amplitude of the input audio signal. The following table gives the detailed output voltages corresponding to different amplitude of the input audio signal.

Table 1. Performance Specification

Audio Signal _ peak(mV)	Output Voltage(V) (Typical)
0	5.50
50	6.54
100	7.58
150	8.63
200	9.67
250	10.71
300	11.75

2.2 Reference Design Schematic

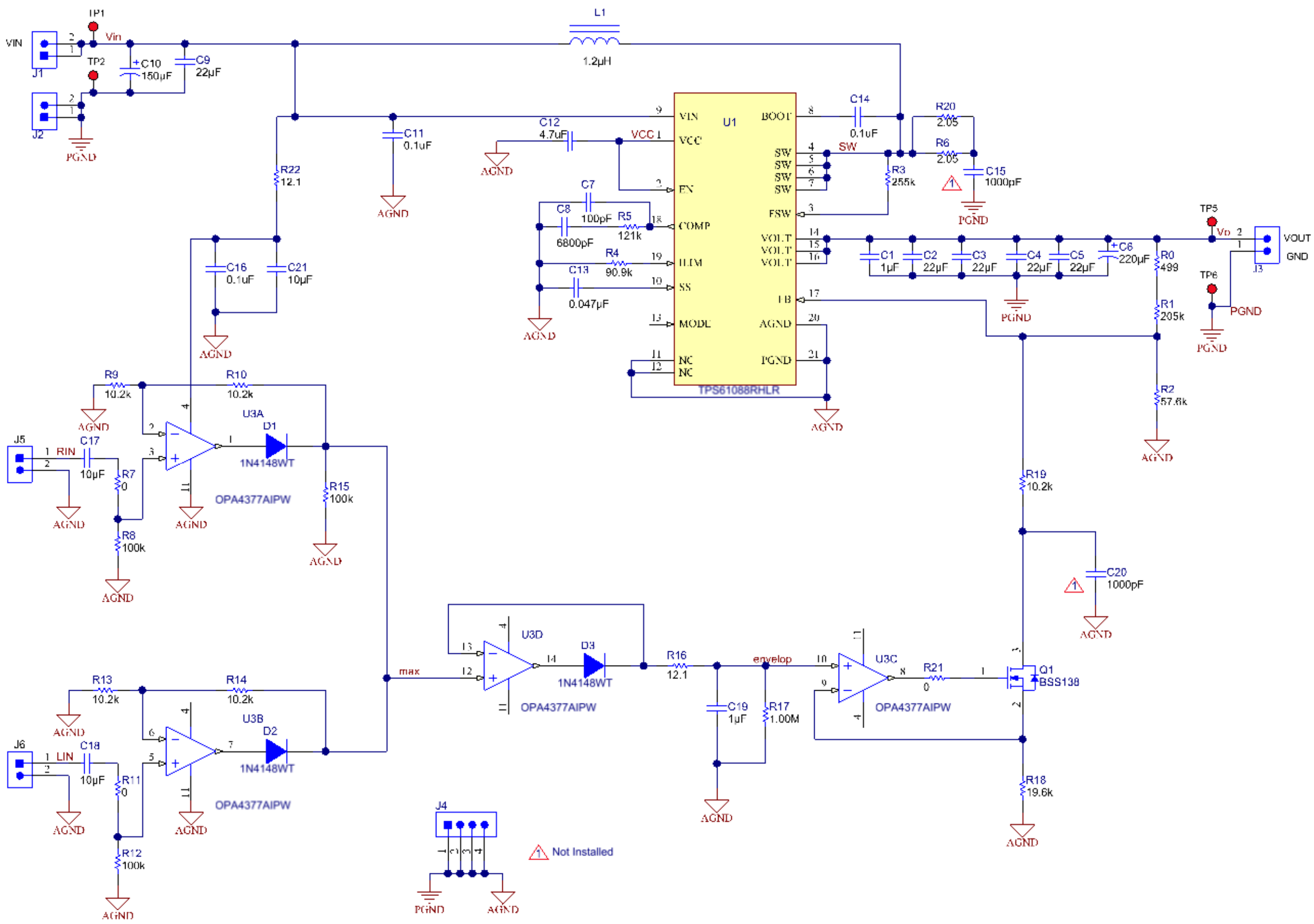

Figure 2. Schematic of the Reference Design

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2.3 Output Voltage Setting

In this reference design, during the theoretical calculation, we set the minimum output voltage $V_{o(min)}$ to 5.5V, set the maximum output voltage $V_{o(max)}$ to 11.75V.

Refer to the schematic in Figure.2, when the input audio signal is 0V, the output of the envelope-tracking circuit is 0V. So the current flowing through Q1 is 0A. In this case, the output voltage $V_{o(min)}$ is determined by R0, R1, R2 and the TPS61088's feedback regulation voltage V_{FB}. We put two resistors (R0 and R1) at the high side of the output voltage dividing network to make the resistors selection more flexible.

From the TPS61088's datasheet, we know that the FB pin's maximum leakage current is 100nA. So the current through the resistance divider should be higher than 20uA to ensure the output voltage precision. A standard low side resistor R2 of 57.6kΩ is selected. The high side resistors R0 and R1 can be calculated from the following equation:

$$\frac{V_{o(\min)} - V_{FB}}{R0 + R1} = \frac{V_{FB}}{R2} \tag{1}$$

Where

- V_{FB} is the TPS61088's feedback regulation voltage (V_{FB}=1.204V).

So the total value of R0 and R1 is 205.5kΩ. Here, we choose R0=499Ω, R1=205kΩ.

When the input audio signal is at its maximum amplitude of 300 mV, the output voltage of the envelope-tracking circuit is 600mV. The current flowing through Q1 reaches its maximum value. The TPS61088's output voltage also reaches its maximum value. The maximum current flowing through Q1 ($I_{Q1(max)}$) can be calculated as:

$$I_{Q1(\max)} = \frac{(V_{o(\max)} - V_{FB})}{R0 + R1} - \frac{V_{FB}}{R2} = 30.42uA \tag{2}$$

The resistors R18 can be calculated by the following equation:

$$R18 = \frac{V_{envelop(\max)}}{I_{Q1(\max)}} \approx 19.6k\Omega \tag{3}$$

Here we choose R19=10kΩ.

Where

- $V_{envelop(max)}$= 600mV(two times of the maximum amplitude of the input audio signal).

2.4 Switching Frequency Setting

The TPS61088's switching frequency is set by the resistor R3 which is connected between the FSW pin and the SW pin. This resistor can be calculated by the following equation:

$$R3 = \frac{4 \cdot (\frac{1}{f_{sw}} - t_{DELAY} \cdot \frac{V_{o(\max)}}{V_{in}})}{C_{FREQ}} \approx 255k\Omega \tag{4}$$

Where

- f_{sw} is the desired switching frequency (f_{sw}=600kHz).
- t_{DELAY}=89 ns.
- C_{FREQ}=23 pF.

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- V_{in} is the input voltage.

2.5 Peak Current Limit Setting

The peak switch current limit is set by the external resistor R4 (Figure 2). We should make sure that the current limit point is higher than the required peak switch current at the lowest input voltage and highest output power condition. The current limit value under PFM mode can be calculated by the following equation:

$$I_{LIM} = \frac{1190000}{R4} = 13.1A \quad (5)$$

Where

- R4 is the resistance connected between the I_{LIM} pin and ground (R9=90.9KΩ).
- I_{LIM} is the peak switch current limit.

When the audio signal steps from 0V to 300mV, the TPS61088's output voltage will rise from 5.5V to 11.75V. In order to make the TPS61088's output voltage rise up quickly during this transition, we choose I_{LIM}=13.1A, which is 2.8A higher than the peak switch current (the peak switch current is 10.23A by equation 6). Higher current limit point means the output capacitor can be charged up more quickly, which also means the output voltage can rise up more quickly. But if we set the current limit value to an even higher point, the TPS61088 will have thermal issue.

If the MODE pin is short to ground, the current limit value is 1.6A lower than that of floating the MODE pin. We need to change R4 to 82.5k to ensure the current limit point is unchanged.

2.6 Inductor Selection

The inductor L1 is the most important component in the switching power supply design. Because it can affect the power supplier's steady state operation, transient behavior, loop stability, and the conversion efficiency.

One of the main parameter of the inductor is the saturation current. The saturation current of the selected inductor should be higher than the peak switching current at the maximum output power condition. Suppose the conversion efficiency ɧ=0.88 at the minimum input voltage condition, then the maximum input current is:

$$I_{in(\max)} = \frac{V_{o(\max)} \cdot V_{o(\max)}}{R_{load} \cdot V_{in(\min)} \cdot \eta} = 8.72A \quad (6)$$

Set the inductor L1's ripple current to about 35% of the average inductor current (input current). Then the peak switching current is:

$$I_{sw(peak)} = I_{in(\max)} + I_{in(\max)} \cdot \frac{0.35}{2} = 10.23A \quad (7)$$

As the current limit point (13.1A) is much higher than the peak switch current. The inductor L1's saturation current should be higher than the current limit point instead of the peak switch current.

Another main parameter of the inductor is the inductor value. The inductor value can be calculated by the following equation:

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$$L1=\left(\frac{V_{in(\min)}}{V_{o(\max)}}\right)^2\cdot\left(\frac{V_{o(\max)}-V_{in(\min)}}{\frac{V_{o(\max)}}{R_{load}}\cdot f_{sw}}\right)\cdot\left(\frac{\eta}{0.35}\right)\approx 1.2uH \quad (8)$$

Where

- R_{load} is the output load resistance (R_{load} = 6Ω).
- $V_{in(min)}$ is the minimum input voltage ($V_{in(min)}$ = 3V).

The inductor's ripple current can be calculated by the following equation:

$$\Delta I_{L1\_pp}=\frac{V_{in(\min)}\cdot(V_{o(\max)}-V_{in(\min)})}{L_1\cdot f_{sw}\cdot V_{o(\max)}}=3.2A \quad (9)$$

The inductor L1's DCR is also a key factor during the inductor selection. DCR should be as low as possible to minimize the power loss.

Finally, make sure the selected inductor type is fit for the high frequency application. At switching frequencies of 600 kHz, the inductor core loss, the proximity effect and the skin effect become very important.

2.7 Output Capacitor Selection

The output capacitors C2, C3, C4 and C5 can be calculated with the following equation:

$$C_{out}=\frac{V_{o(\max)}-V_{in(\min)}}{V_{o(\max)}\cdot f_{sw}}\cdot\frac{V_{o(\max)}}{R_{load}\cdot\Delta V_o} \quad (10)$$

Where

- ΔV_o is the required output voltage ripple.

Considering the capacitance derating under certain DC bias, four 22-μF ceramic capacitors in parallel is fit for the ΔV_o=50mV application.

In some of the audio applications, electrolytic capacitors will be used as the decoupling capacitors between the power supply and the PA. So in order to match those applications, we add an extra 220uF electrolytic capacitor at the output side.

2.8 Compensation Circuit

The higher the cross over frequency f_c, the faster the loop response is. In order to make the output voltage rise up as quickly as possible when the audio signal step up from its minimum value to its maximum value, we need to choose a high cross over frequency f_c while keeping the loop stability at the same time. The cross over frequency at low output voltage condition is wider than that of the high output voltage condition. So we choose f_c=30kHz when $V_{o(min)}$=5.5V.

The COMP pin is the output of the internal trans-conductance error amplifier. The following equation can be used to calculate R5 (Rcomp) and C8 (Ccomp) (Figure 3).

$$R5=\frac{2\pi\cdot V_{o(\min)}\cdot R_{sense}\cdot f_c\cdot C_{out}}{(1-D)\cdot V_{FB}\cdot G_{EA}} \quad (11)$$

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$$C8 = \frac{R_{load} \cdot C_{out}}{2 \cdot R5} \quad (12)$$

Where:

- R_{sense} is the equivalent internal current sense resistor, which is 0.08 Ω.
- D is the switching duty cycle under V_{in}=4.2V and $V_{o(min)}$ = 5.5V.
- C_{out} is the output capacitance (effective C_{out} ≈250 uF).
- G_{EA} is the error amplifier's trans-conductance (G_{EA}=190 uA/V).

R5=121k and C8=6.8nF are used in this reference design.

The value of C7 can be calculated by the equation 13 :

$$C7 = \frac{R_{ESR} \cdot C_{out}}{R5} \quad (13)$$

We use four 22-µF ceramic capacitors and one 220uF electrolytic capacitor at the output side. The equivalent ESR of the output capacitor is very small, about 50mΩ. So we choose C7=100pF to compensate the ESR zero.

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3 PCB Layout

This reference design is implemented on a 2-layer PCB. All the components are placed on the top layer. Figure 3 shows the top layer and top silk screen. Figure 4 shows the layout of the bottom layer.

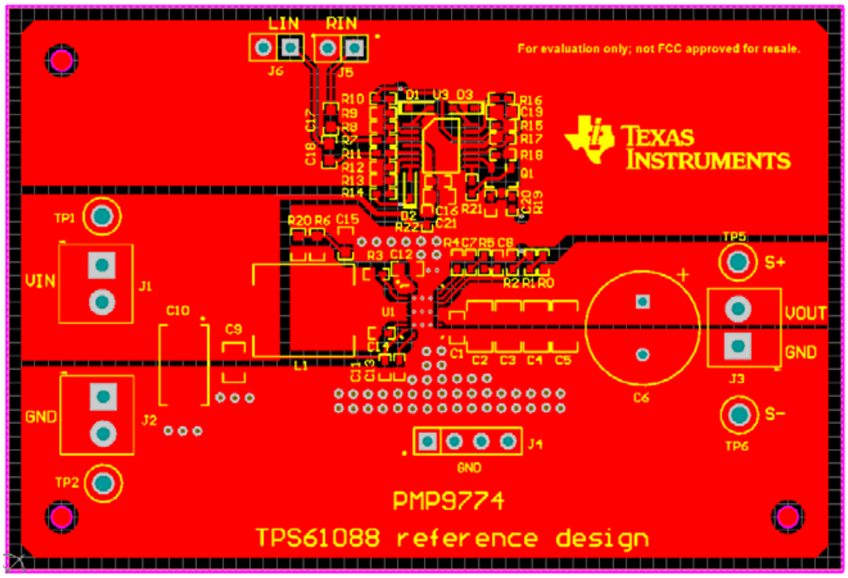

Figure 3 Top Layer and Top Silkscreen

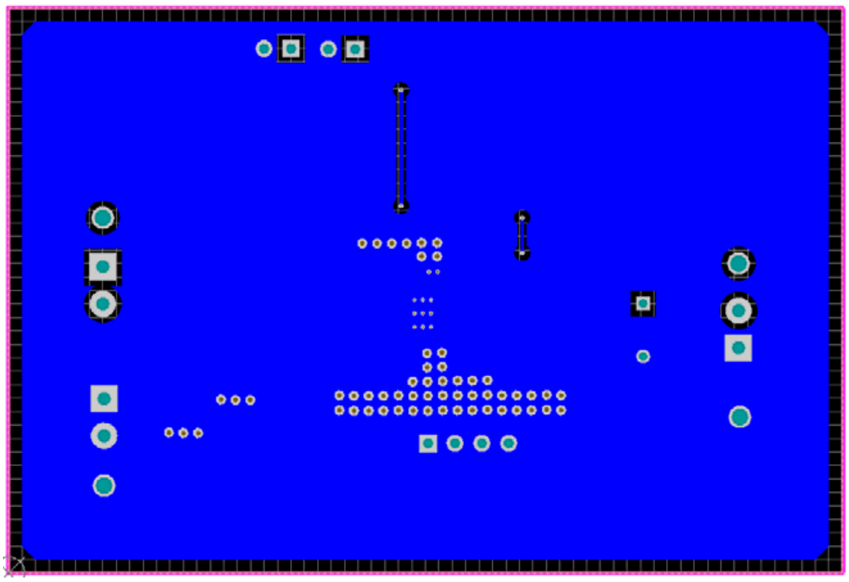

Figure 4 Bottom Layer

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4 Test Result

Figure 5 shows the output voltage versus the amplitude of the input audio signal.

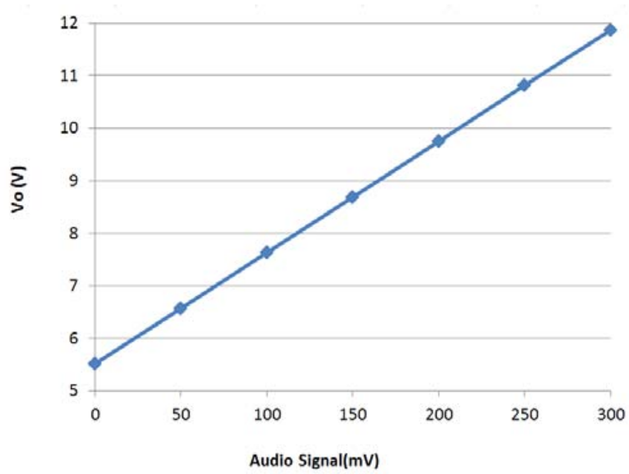

Figure 5 V_o VS. Audio Signal

Figure 6.1, Figure 6.2 , Figure 6.3 and Figure 6.4 show the transition of the output voltage rise up from $V_{o(min)}$ to $V_{o(max)}$ when the input audio signal steps up from 0mV to 300mV (V_{in}=3.6V, R_{load}=6 ohm).

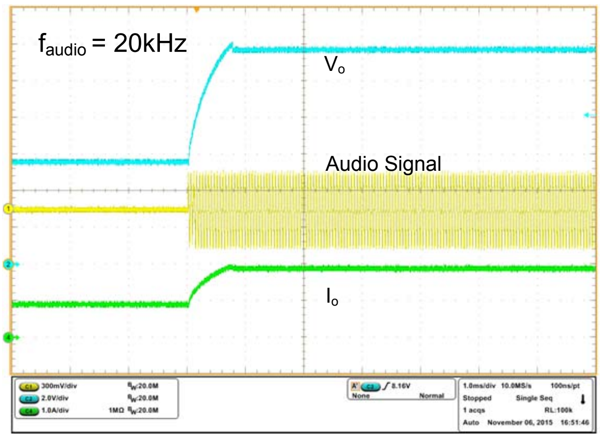

Figure 6.1 Output Voltage Transition at f_{audio}=20kHz

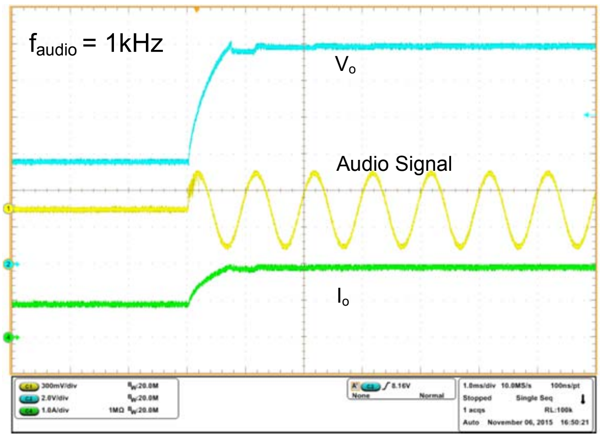

Figure 6.2 Output Voltage Transition at f_{audio}=1kHz

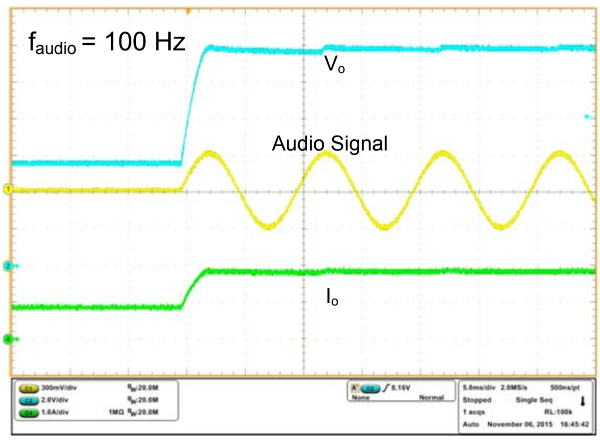

Figure 6.3 Output Voltage Transition at f_{audio}=100Hz

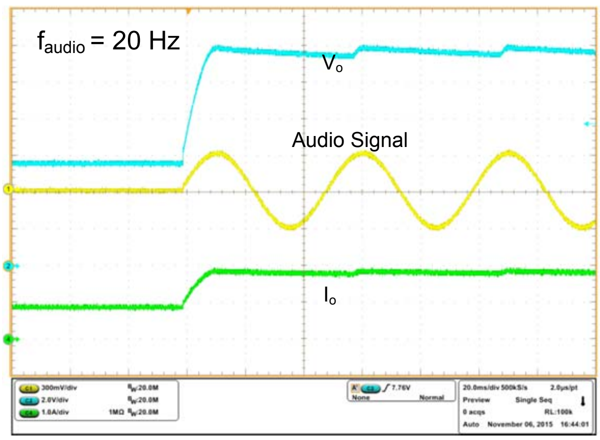

Figure 6.4 Output Voltage Transition at f_{audio}=20Hz

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Figure 7.1 and Figure 7.2 shows the inductor L1's current, TPS61088's SW pin voltage and the output voltage ripple when the peak amplitude of the input audio signal is 0V and 300mV respectively (f_{audio}=1 KHz).

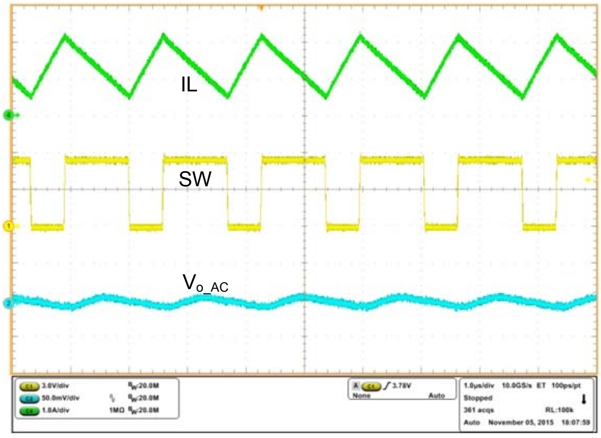

Figure 7.1 Switching Waveforms at V_{audio}=0V

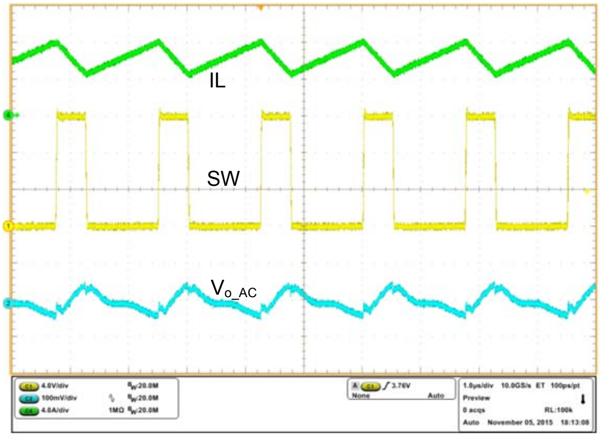

Figure 7.2 Switching Waveforms at V_{audio}=300mV

Figure 8.1 and Figure 8.2 show the start-up waveform of the inductor current, SW voltage and the output voltage when the peak amplitude of the input audio signal is 0V and 300mV respectively (f_{audio}=1 KHz).

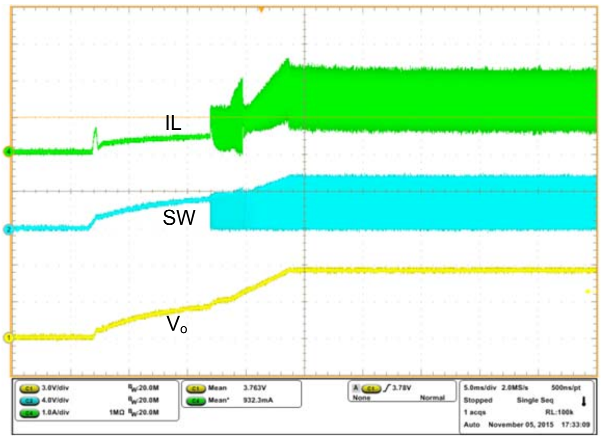

Figure 8.1 Startup Waveforms at V_{audio}=0V

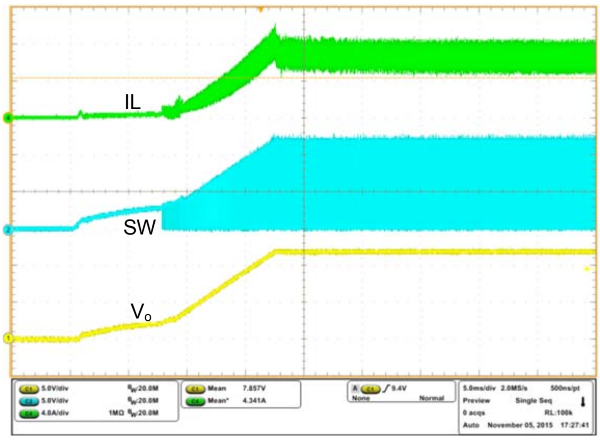

Figure 8.2 Startup Waveforms at V_{audio}=300mV

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Figure 9.1 and Figure 9.2 show the load transient response of the output voltage.

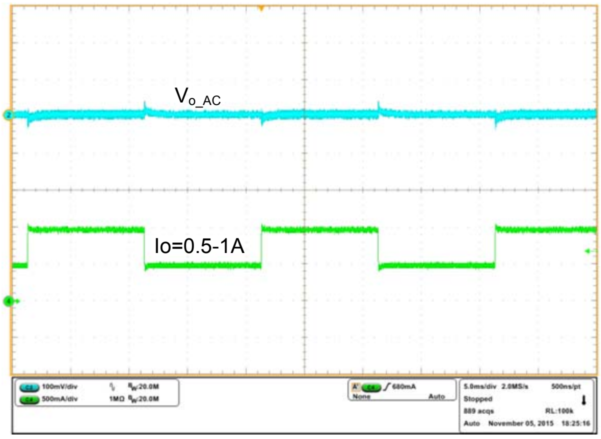

Figure 9.1 Load Transient (V_{audio}=0V)

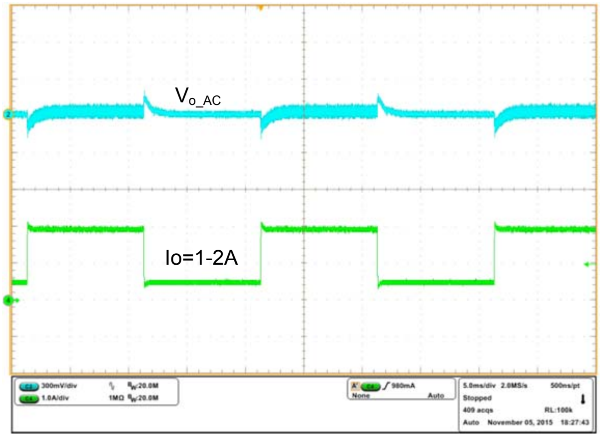

Figure 9.2 Load Transient (V_{audio}=300mV)

Figure 10.1 shows the loop bode plot when the input audio signal is 0V ($V_{o(min)}$=5.6V).

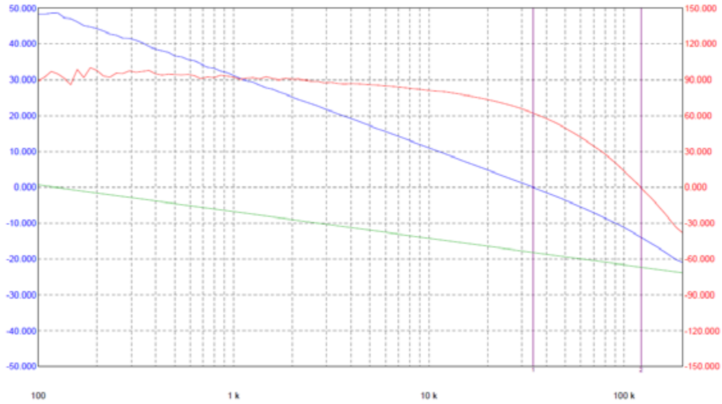

Figure 10.1 Loop Bode Plot (V_{in}=3.6V, R_{load}=6 ohm)

Figure 10.2 shows the loop bode plot when the peak amplitude of the input audio signal is 300mV ($V_{o(max)}$=11.8V).

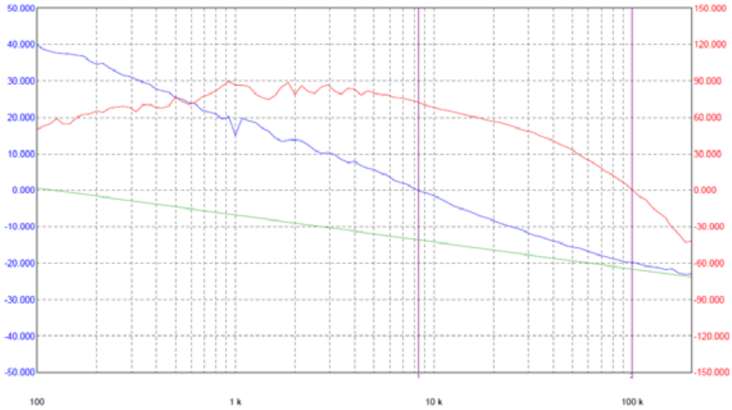

Figure 10.2 Loop Bode Plot (V_{in}=3.6V, R_{load}=6 ohm)

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5 Further Discussion

In this reference design, the peak-amplitude detection circuit and the envelope-tracking circuit is realized by OPA4377. This is a rail-to rail output (the output swing from rail is smaller than 40mV in the full temperature range), good small signal response and low cost operational amplifier.

Some low-cost amplifiers, like LM324 and LM2902, are not fit for this application. Because when the input audio signal's frequency is above 5kHz, the output waveform of the peak-amplitude detection circuit is getting distortion.

Another low cost amplifier, LMV324 is also an option for this application. But its output swing low voltage is 180mV, which is pretty high. In order to make LMV324 fit for this application, we do some modification on the schematic. A pre-biased voltage is added to the output of the envelope detection circuit.

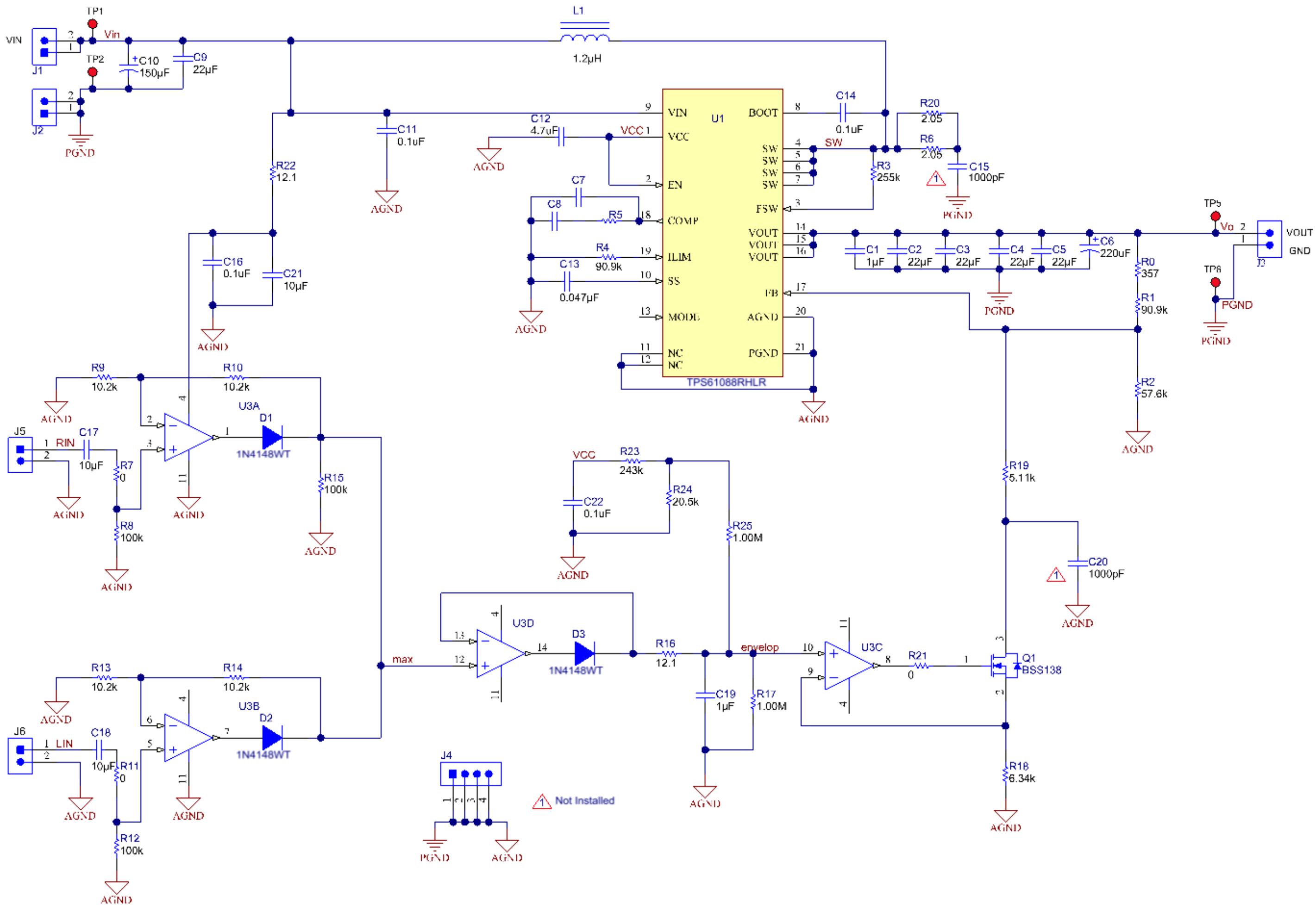

Figure 11. Schematic of the Reference Design with LMV324

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With the new schematic (Fig.11), when the output voltage of the envelope detection circuit is below 200mV (the input audio signal is lower than 100mV), the output voltage of TPS61088 is clamped at 6V. When the input audio signal is higher than 100mV, the output voltage of TPS61088 will increase linearly with the increment of the audio signal.

Table 2. Performance Specification with LMV324

Audio Signal _ peak(mV)	Output Voltage(V) (Typical)
0-100	6.00
150	7.45
200	8.90
250	10.35
300	11.80

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