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MAX16809 Evaluation Kit

General Description

The MAX16809 evaluation kit (EV kit) is a 16-channel, constant-current LED driver, capable of driving 40mA each to 16 LED strings with a total forward voltage of up to 32V. The MAX16809 EV kit is based on the MAX16809 device, which has 16 constant-current-sinking outputs with sink current settable using a single resistor and a high-performance, current-mode pulse-width-modulator (PWM) controller, for implementing a DC-DC converter that generates the supply voltage to drive the LED strings.

The MAX16809 EV kit operates at supply voltages between 9V to 16V and temperatures ranging from 0°C to +70°C. It features a PWM dimming control, adaptive control of the LED supply voltage, which depends upon the operating voltage of the LED strings, a built-in clock generator, and a low-current shutdown. The MAX16809 EV kit is a fully assembled and tested board.

Features

- ♦ **9V to 16V Supply Voltage Range**
- ♦ **40mA LED Current (Per Each LED String)**
- ♦ **Single-Resistor Current Adjust for 16 Channels**
- ♦ **Up to 32V LED String Voltage**
- ♦ **Boost Converter to Generate LED Supply Voltage**
- ♦ **Adaptive LED Supply Voltage Control Increases Efficiency**
- ♦ **PWM Dimming Control**
- ♦ **Output-Voltage-Spike Protection for Inductive-Output Lines**
- ♦ **Proven PCB Layout**

Ordering Information

PART	TEMP RANGE	IC PACKAGE
MAX16809EVKIT+	0°C to +70°C*	38 TQFN-EP†

+Denotes a lead-free and RoHS-compliant EV kit.

**This limited temperature range applies to the EV kit PCB only. The MAX16809 IC temperature range is -40°C to +125°C.*

†EP = Exposed paddle.

Component List

DESIGNATION	QTY	DESCRIPTION
C1–C4, C6, C7, C8, C10, C11, C19–C23, C25, C26	16	1nF ±10%, 50V X7R capacitors (0603) Murata GRM188R71H103KA01D KEMET C0603C103K5RACTU
C5, C24, C30, C31, C33	5	0.1µF ± 10%, 50V X7R capacitors (0603) Murata GRM188R71H104KA93D TDK C1608X7R1H104K
C9	1	1µF ±20%, 16V X7R capacitor (0805) Murata GRM21BR71C105KA01L TDK C2012X7R1C105K
C12, C13, C14	3	22µF ±20% 35V electrolytic capacitors Panasonic EEEFK1V220R
C15	1	1µF ±20%, 50V X7R capacitor (1210) KEMET C3225X7R1H105M Murata GRM32ER71H105KA01L
C16, C17, C18	3	22µF ±20%, 50V electrolytic capacitors Panasonic EEEFK1H220P
C27	1	560pF ±10%, 50V C0G capacitor (0603) KEMET C0603C561K5RACTU Murata GRM188R71H561KA01D
C28	1	10pF ±10%, 50V C0G capacitor (0603) TDK C1608C0G1H00DB Murata GRM1885C1H100JA01D
C29	1	220pF ±10%, 50V C0G capacitor (0603) KEMET C0603C221K5RACTU Murata GRM188R71H221KA01D
C32	1	100pF ±10%, 50V C0G capacitor (0603) KEMET C0603C101K5RACTU Murata GRM188R71H101KA01D

MAX16809 Evaluation Kit

Component List (continued)

DESIGNATION	QTY	DESCRIPTION
C34	1	10nF ±10%, 50V X7R capacitor (0603) KEMET C0603C103K5RACTU Murata GRM188R71H103KA01D
C35	0	Not installed, capacitor
D1	1	33V zener diode (SOD323) Diodes Inc. MMSZ5257BS-7
D2	1	1A, 40V Schottky diode (SMA) Diodes Inc. CMSH1-40M Central Semiconductor CMSH1-40M
D3–D10	8	15V dual zener diodes (SOT323) Diodes AZ23C15-7-F
D11	1	20mA switching diode (SOD323) Diodes Inc. 1N4148WS-7
D12	1	40V small-signal Schottky diode (SOD523) Diodes Inc. SDM03U40
GND, GND, PWM, $\overline{SHDN}$, VBIAS, VIN	6	Wire loops
J1	1	0.1in 20-pin header
L1	1	27µH, 3.2A inductor Coilcraft MSS1260-273ML
Q1	1	Switching transistor (SOT523) Diodes Inc. MMBT2222AT-7-F

DESIGNATION	QTY	DESCRIPTION
Q2	1	60V, 5.5A N-channel MOSFET Vishay Si4450DY
R1	1	430Ω ±1% resistor (0603)
R2	1	330kΩ ±1% resistor (0603)
R3	1	8.45kΩ ±1% resistor (0603)
R4, R6, R10	3	22kΩ ±1% resistors (0603)
R5	1	180kΩ ±1% resistor (0603)
R7	1	10.5kΩ ±1% resistor (0603)
R8	1	10Ω ±1% resistor (0603)
R9	1	1.2kΩ ±1% resistor (0603)
R10	1	22kΩ ±1% resistor (0603)
R11	1	50kΩ ±1% resistor (0603)
R12	1	75mΩ ±1% resistor (0603)
R13	0	Not installed, resistor
R14	1	2.2kΩ ±1% resistor (0603)
R15	1	10kΩ ±1% resistor (0603)
U1	1	MAX16809ATU+ (38-pin TQFN, 5mm x 7mm)
U2	1	Digital pnp transistor ROHM DTA114WKA
U3	1	Digital npn transistor ROHM DTC114WKA
U4	1	Dual inverter with hysteresis Texas Instruments SN74LVC2G14DCKR
—	1	PCB: MAX16809 Evaluation Kit+

Component Suppliers

SUPPLIER	PHONE	FAX	WEBSITE
Central Semiconductor Corp.	631-435-1110	631-435-3388	www.centralsemi.com
Coilcraft, Inc.	847-639-6400	847-639-1469	www.coilcraft.com
Diodes Inc.	805-446-4800	805-446-4850	www.diodes.com
KEMET Corp.	978-658-1663	978-658-1790	www.kemet.com
Murata Mfg. Co., Ltd.	770-436-1300	770-436-3030	www.murata.com
ROHM Co., Ltd.	858-625-3630	858-625-3670	www.rohm.com
TDK Corp.	847-390-4373	847-390-4428	www.component.tdk.com
Texas Instruments Inc.	—	—	www.ti.com
Vishay	402-563-6866	402-563-6296	www.vishay.com

Note: *Indicate that you are using the MAX16809 when contacting these component suppliers.*

Quick Start

Recommended Equipment

- One 16V, 5A adjustable power supply
- One 5V power supply
- 16 LED strings with a total forward voltage ≤ 32V
- One multimeter
- One PWM signal generator (optional)

Procedure

The MAX16809 EV kit is fully assembled and tested. Follow the steps below to verify operation. **Caution: Do not turn on the power supply until all connections are completed.**

1) Connect LED strings with operating voltage of approximately 32V between VLED (pins 1-4 of J1) and OUT0–OUT15 (pins 5-20 of J1). All 16 channels should have an LED string load connected of the same type.
2) Connect the DC power supply (16V, 5A) to VIN and GND.
3) Connect a DC power supply (0 to 5V) to VBIAS and GND.
4) Turn on the power supplies and apply 10V to VIN and 3V to 5V to VBIAS. Connect $\overline{\text{SHDN}}$ and PWM to 3V to 5V. All of the LEDs should turn on. Measure the current through any LED string, which should be 40mA ±7%.
5) Increase the supply voltage to 16V and the LED currents will be stable. Measure the current through any LED string, which should be 40mA ±7%.
6) Apply a PWM signal with amplitude of 3V to 5V and a frequency between 100Hz and 2kHz to the PWM input. The LED brightness should increase as the PWM duty cycle increases and viceversa.
7) Connect $\overline{\text{SHDN}}$ to GND and all LEDs should turn off.

Detailed Description

The MAX16809 EV kit is a 16-channel, constant-current LED driver capable of driving 40mA each to 16 LED strings, with a total forward voltage of up to 32V. The MAX16809 EV kit can drive a total of 160 white LEDs in 16 strings, with operating current up to 40mA. The MAX16809 EV kit can operate at input supply voltages between 9V and 16V.

The MAX16809 EV kit evaluates the MAX16809 IC, which has two major sections. The first section consists of 16 constant-current LED drivers capable of sinking up to 55mA when on and blocking up to 36V when off. The second section is a high-performance current-mode PWM controller that can control a DC-DC converter to generate the supply voltage for driving the LED strings. The MAX16809 EV kit uses the PWM controller to drive a boost converter, which takes a 9V to 16V input and generates a 33V LED supply voltage. To drive a constant current into an LED string, connect the LED string between the 33V output and any of the 16 constant-current-sink outputs. The resistor (R1) from the SET pin to ground programs the sink current of each output. The sink current of any output can be up to 55mA and the amplitude is the same value for all the outputs. The difference between the total forward voltage and the LED supply voltage drops between the constant-current-sink output and ground, and is dissipated as power in the device.

The LED supply voltage generated by the boost converter in the MAX16809 EV kit is adaptive. The LED string with the highest total forward voltage dominates the control loop, and the boost-converter voltage is adjusted so that the driver associated with that string receives just enough voltage required for current drive. All the other strings, with lower total forward voltages, will have excess supply voltage, which is then dropped in the associated driver. This feedback mechanism ensures that the linear-current-control circuit dissipates the minimum possible power. An on-board inverter (U4A) is configured to generate the clock input for the MAX16809. The constant-current output-driver circuits and U4 need a 3.3V to 5V input, which should be supplied externally. If 5V is not available, it can be generated using an emitter-follower buffer from the REF output of MAX16809.

Boost Converter

The boost converter that generates the 33V LED supply voltage operates at a switching frequency of 350kHz in continuous-conduction mode (CCM). The current-mode PWM controller in the MAX16809 drives the external MOSFET (Q2) to control the boost converter. The MOSFET is turned on at the beginning of every switching cycle and turned off when the current through the inductor (L1) reaches the peak value set by the error-amplifier-output voltage. Inductor current is sensed from the voltage across the ground-referenced current-sense resistor (parallel combination of R12 and R13). This current-sense information is passed on to the current-sense comparators in the MAX16809 through the CS pin.

During the on period of the MOSFET, the inductor stores energy from the input supply. When the switch is turned off, the inductor generates sufficient voltage in reverse direction to discharge the stored energy to VLED. This generated voltage forms a source, in series with the input supply voltage, and drives VLED through the rectifier diode (D2).

As the boost converter is operated in CCM, only part of the stored energy in the inductor is discharged to VLED. The advantages of CCM include reduced input and output filtering, reduced EMI due to lower peak currents, and higher converter efficiency. However, these advantages come at the cost of a right-half-plane zero in the converter-transfer function. Compensating this zero requires reducing the system bandwidth, which affects the converter-dynamic response. As the 16-channel, constant-current-sink outputs control the current through the LEDs, slower control of VLED does not affect the LED operation. Compensation of the feedback circuit is explained in the *Feedback Compensation* section.

An internal comparator turns off the gate pulse to the external MOSFET if the voltage at the CS pin exceeds 0.3V. The current through the inductor that produces 0.3V at the CS pin is the maximum inductor current possible (the actual current can be a little higher than this limit due to the 60ns propagation delay from the CS pin to the MOSFET drive output). This condition can happen when the feedback loop is broken, when the output capacitor charges during power-up, or when there is an overload at the output. This feature protects the MOSFET by limiting the maximum current passing through it during such conditions.

The RC filter, consisting of R9 and C10, removes the voltage spike across the current-sense resistors produced by the turn-on gate current of the MOSFET and the reverse-recovery current of D2. Without filtering, these current spikes can cause sense comparators to falsely trigger and turn off the gate pulse prematurely. The filter time constant should not be higher than required (the MAX16809 EV kit uses a 120ns time constant), as a higher time constant adds additional delay to the current-sense voltage, effectively increasing the current limit.

During normal operating conditions, the feedback loop controls the peak current. The error amplifier compares a scaled-down version of the LED supply voltage (VLED) with a highly accurate 2.5V reference. The error amplifier and compensation network then amplify the error signal, and the current comparator compares this signal to the sensed-current voltage to create a PWM drive output.

Power-Circuit Design

Initially, decide the input supply voltage range, output voltage VLED (the sum of the maximum LED total forward voltage and 1V bias voltage for the constant-current-sink output), and the output current I_{OUT} (the sum of all the LED string currents).

Calculate maximum duty cycle D_{MAX} using the following equation:

$$D_{MAX} = \frac{VLED + V_D - VIN_{MIN}}{VLED + V_D - V_{FET}}$$

where V_D is the forward drop of the rectifier diode D2 (~0.6V), VIN_{MIN} is the minimum input supply voltage (in this case, 9V), and V_{FET} is the average drain-to-source voltage of the MOSFET Q2 when it is on.

Select the switching frequency F_{SW} based on the space, noise, dynamic response, and efficiency constraints. Select the maximum peak-to-peak ripple on the inductor current IL_{PP}. For the MAX16809 EV kit, F_{SW} is 350kHz and IL_{PP} is ±30% of the average inductor current. Use the following equations to calculate the maximum average-inductor current IL_{AVG} and peak inductor current IL_{PEAK}:

$$IL_{AVG} = \frac{I_{OUT}}{1 - D_{MAX}}$$

Since IL_{PP} is ±30% of the average-inductor current IL_{AVG}:

$$IL_{PP} = IL_{AVG} \times 0.3 \times 2$$

$$IL_{PEAK} = IL_{AVG} + \frac{IL_{PP}}{2}$$

Calculate the minimum inductance value L_{MIN} with the inductor current ripple set to the maximum value:

$$L_{MIN} = \frac{(VIN_{MIN} - V_{FET}) \times D_{MAX}}{F_{SW} \times IL_{PP}}$$

Choose an inductor that has a minimum inductance greater than this calculated value.

Calculate the current-sense resistor (R12 in parallel with R13) using the equation below:

$$R_{CS} = \frac{0.3 \times 0.75}{IL_{PEAK}}$$

where 0.3V is the maximum current-sense signal voltage. The factor 0.75 is for compensating the reduction of maximum current-sense voltage due to the addition of slope compensation. Check this factor and adjust after the slope compensation is calculated. See the *Slope Compensation* section for more information.

The saturation current limit of the selected inductor (IL_{SAT}) should be greater than the value given by the equation below. Selecting an inductor with 10% higher IL_{SAT} rating is a good choice:

$$IL_{SAT} = 1.1 \times IL_{PEAK}$$

Calculate the output capacitor C_{OUT} (parallel combination of C16, C17, C18, and C24) using the following equation:

$$C_{OUT} = \frac{D_{MAX} \times I_{OUT}}{VLED_{PP} \times F_{SW}}$$

where $VLED_{PP}$ is the peak-to-peak ripple in the LED supply voltage. The value of the calculated output capacitance will be much lower than what is actually necessary for feedback loop compensation. See the *Feedback Compensation* section to calculate the output capacitance based on the compensation requirements.

Calculate the input capacitor C_{IN} (parallel combination of C12, C13, C14, and C5) using the following equation:

$$C_{IN} = \frac{IL_{PP}}{8 \times F_{SW} \times VIN_{PP}}$$

where VIN_{PP} is the peak-to-peak input ripple voltage. This equation assumes that input capacitors supply most of the input ripple current.

Selection of Power Semiconductors

The switching MOSFET (Q2) should have a voltage rating sufficient to withstand the maximum output voltage, together with the diode drop of D2, and any possible overshoot due to ringing caused by parasitic inductances and capacitances. Use a MOSFET with voltage rating higher than that calculated by the following equation:

$$V_{DS} = (VLED + V_D) \times 1.3$$

The factor of 1.3 provides a 30% safety margin.

The continuous drain-current rating of the selected MOSFET when the case temperature is at +70°C should be greater than that calculated by the following equation. The MOSFET must be mounted on a board, as per manufacturer specifications, to dissipate the heat:

$$ID_{RMS} = \left(\sqrt{\frac{IL_{AVG}^2}{D_{MAX}}}\right) \times 1.3$$

The MOSFET dissipates power due to both switching losses, as well as conduction losses. Use the following equation to calculate the conduction losses in the MOSFET:

$$P_{COND} = \frac{IL_{AVG}^2}{D_{MAX}} \times RDS_{ON}$$

where RDS_{ON} is the on-state drain-source resistance of the MOSFET with an assumed junction temperature of 100°C.

Use the following equation to calculate the switching losses in the MOSFET:

$$P_{SW} = \frac{IL_{AVG} \times VLED^2 \times C_{GD} \times F_{SW}}{2} \times \left(\frac{1}{I_{GON}} + \frac{1}{I_{GOFF}}\right)$$

where I_{GON} and I_{GOFF} are the gate currents of the MOSFET (with V_{GS} equal to the threshold voltage) when it is turned on and turned off, respectively, and C_{GD} is the gate-to-drain MOSFET capacitance. Choose a MOSFET that has a higher power rating than that calculated by the following equation when the MOSFET case temperature is at +70°C:

$$P_{TOT} = P_{COND} + P_{SW}$$

The MAX16809 EV kit uses a Schottky diode as the boost-converter rectifier (D2). A Schottky rectifier diode produces less forward drop and puts the least burden on the MOSFET during reverse recovery. If a diode with considerable reverse-recovery time is used, it should be considered in the MOSFET switching-loss calculation.

The Schottky diode selected should have a voltage rating 20% above the maximum boost-converter output voltage. The current rating of the diode should be greater than I_D in the following equation:

$$I_D = \left(\sqrt{\frac{IL_{AVG}^2}{1 - D_{MAX}}}\right) \times 1.2$$

Slope Compensation

When the boost converter operates in CCM with more than 50% duty cycle, subharmonic oscillations occur if slope compensation is not implemented. Subharmonic oscillations do not allow the PWM duty cycle to settle to a peak current value set by the voltage-feedback loop. The duty cycle oscillates back and forth about the required value, usually at half the switching frequency. Subharmonic oscillations die out if a sufficient negative slope is added to the inductor peak current. This means that for any peak current set by the feedback loop, the output pulse terminates sooner than normally expected. The minimum slope compensation that should be added to stabilize the current loop is half of the worst-case (max) falling slope of inductor current.

Adding a ramp to the current-sense signal, with positive slope in sync with the switching frequency, can produce the desired function. The greater the duty cycle, the greater the added voltage, and the greater the difference between the set current and the actual inductor current. In the MAX16809 EV kit, the oscillator ramp signal is buffered using Q1 and added to the current-sense signal with proper scaling to implement the slope compensation. Follow the steps below to calculate the component values for slope compensation.

Calculate the worst-case falling slope of the inductor current using the following equation:

$$IL_{SLOPE} = \frac{(VLED_{MAX} + V_D - VIN_{MIN})}{L_{MIN}}$$

From the inductor current falling slope, find its equivalent voltage slope across the current-sense resistor R_{CS} (R12 parallel with R13) using the following equation:

$$V_{SLOPE} = IL_{SLOPE} \times R_{CS}$$

The minimum voltage slope that should be added to the current-sense waveform is half of V_{SLOPE} for ensuring stability up to 100% duty cycle. As the maximum continuous duty cycle used is less than 100%, the minimum required compensation slope becomes:

$$VC_{SLOPE} = \frac{V_{SLOPE} \times (2D_{MAX} - 1) \times 1.1}{D_{MAX}}$$

The factor 1.1 provides a 10% margin. Resistors R9 and R10 determine the attenuation of the buffered voltage slope from the emitter of Q1. The forward drop of signal diode D11, together with the V_{BE} of Q1, almost cancel the 1.1V offset of the ramp waveform. Calculate the approximate slope of the oscillator ramp using the following equation:

$$VR_{SLOPE} = 1.7 \times F_{SW}$$

where 1.7V is the ramp amplitude and F_{SW} is the switching frequency.

Select the value of R9 such that the input bias current of the current-sense comparators does not add considerable error to the current-sense signal. The value of R10 for the slope compensation is given by the equation:

$$R10 = \left(\frac{VR_{SLOPE}}{VC_{SLOPE}} - 1\right) \times R9$$

LED Driver

The MAX16809 features a 16-channel, constant-current LED driver, with each channel capable of sinking up to 55mA of LED current. The LED strings are connected between VLED and the constant-current-sink outputs to drive regulated current through LED strings. The current through all 16 channels is controlled through resistor (R1) from the SET pin to ground. The MAX16809 EV kit sets the current through each string at 40mA and the maximum LED supply voltage to 33V. The MAX16809 EV kit drives LED strings with a total forward voltage of up to 32V.

A 4-wire serial interface with four inputs (DIN, CLK, LE, and $\overline{OE}$) individually control the constant-current outputs. In the MAX16809 EV kit, a 50kHz clock signal, generated by U4A, clocks 16 1s into the internal shift register by tying DIN and LE to 5V. The clock-generation circuit can be avoided if a microcontroller provides the function.

The output enable ($\overline{OE}$) can provide PWM dimming. An inverted PWM signal, generated by the inverter U4B, is necessary to drive the $\overline{OE}$ pin. When the PWM signal is low (LED drivers off), it also influences the feedback with the network formed by R6 and D12. See the *Adaptive LED Supply Voltage Control* section for more details.

If an inverted PWM signal is available, use the circuit shown in Figure1 to drive the $\overline{OE}$ input and feedback network.

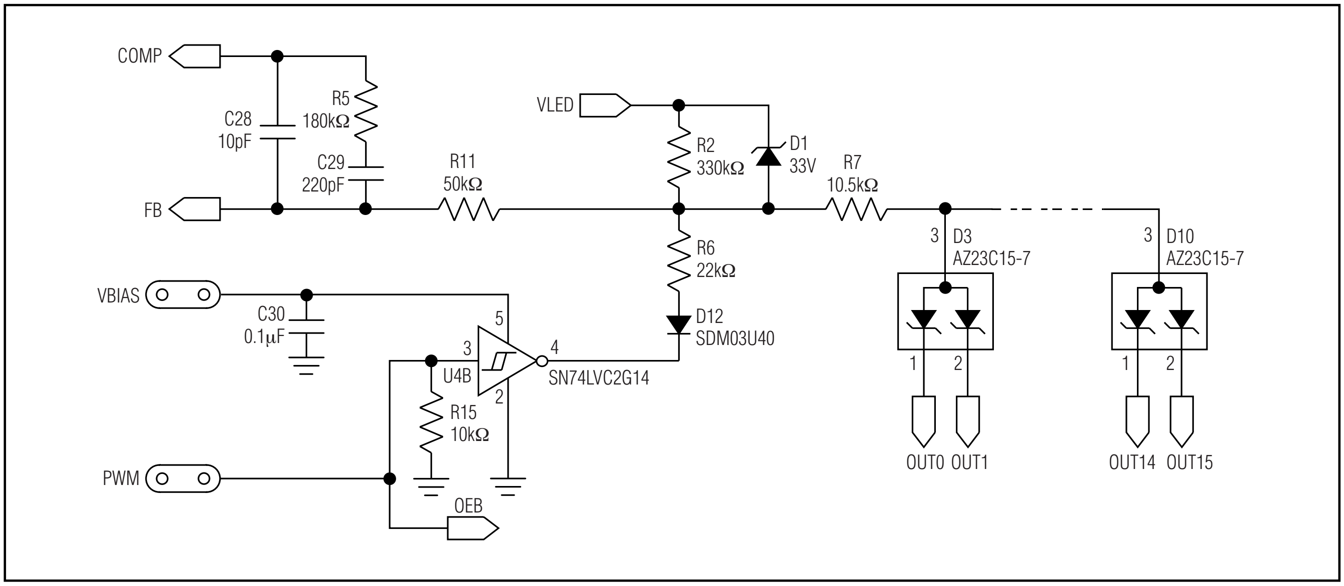

Figure 1. Inverting PWM Drive Circuit

Output Current Setting

The amplitude of the output sink currents for all 16 channels is set to the same value by the resistor (R1) from the SET pin to ground. The minimum allowed value of R_{SET} is 311Ω, which sets the output currents to 55mA. The maximum allowed value of R_{SET} is 5kΩ. The MAX16809 EV kit uses 430Ω for R_{SET}, which sets the output current to 40mA. To set a different output current, use the following equation:

$$R_{SET} = \frac{17100}{I_{OUT}}$$

where R_{SET} is the current-setting resistor (R1) value in ohms and I_{OUT} is the desired output current in milliamps.

Adaptive LED Supply Voltage Control

To reduce power dissipation in the IC, the MAX16809 EV kit features adaptive control of VLED based on the operating voltage of the LED strings. The constant-current-sink outputs can sink stable currents with output voltages as low as 0.8V. The voltage at each of the 16 outputs will be the difference between VLED and the total forward voltage of the LED string connected to that output. The MAX16809 EV kit implements a feedback mechanism to sense the voltage at each of the 16 constant-current-sink outputs. Using dual zener diodes (D3–D10), the MAX16809 EV kit selects the lowest driver voltage (with the greatest LED string voltage) to regulate. The boost-converter PWM then adjusts so that VLED is high enough for this sink output to settle to approximately 0.8V. All the other strings have sufficient voltage, as their total forward voltages are lower. The feedback mechanism ensures that the IC dissipates the minimum possible power. For adaptive control to function efficiently connect LED strings to all 16 channels and use an equal number of LEDs from the same bin in each string. If some of the 16 channels are not used, then the zener diodes (D3–D10) should be removed from the unused channels.

Use the equation below to calculate the value of R2 to get the required minimum voltage at the sink outputs:

$$R2 = \frac{(V_{FLED} + V_S - 2.5) \times R7}{2.5 - V_{DZ} - V_S}$$

where 2.5V is the feedback reference, V_{DZ} is the forward drop of the ORing diode (D3–D10), V_S = 0.5V is the required sink-output voltage, and V_{FLED} is the nominal total forward voltage of the LED strings. Select the value of R2 such that R7 is approximately 10kΩ.

The zener diodes (D3–D10) also provide output overvoltage protection. If an LED string gets partially or fully shorted, making the sink-output voltage go high, the 15V zener diode connected to that output conducts in reverse direction, and limits the VLED voltage. Under this condition, the other LED strings might not turn on.

When the outputs are off, the LED drivers are at high impedance and the feedback network now combines R6 and D12 to provide a path for the feedback current and to control VLED. Use the following equation to calculate the value of R6 to get the required LED supply voltage during PWM off time:

$$R6 = \frac{R2 \times (2.5 - 0.4)}{VLED_{OFF} - 2.5}$$

where 2.5V is the feedback-reference voltage, 0.4V is the total voltage dropped by D4 and PWM input, and $VLED_{OFF}$ is the desired LED supply voltage during PWM off time. $VLED_{OFF}$ should be set to the worst-case LED string voltage plus some additional headroom for the LED drivers (0.8V), as well as a reserve voltage (approximately 1V). The reserve voltage allows the MAX16809 to provide current for very short PWM dimming on-pulses. With pulses as low as 2µs, the VLED control loop is not able to react, and the output capacitors provide all the current. For longer PWM dimming pulses, the control loop reacts and the supply operates at the adaptive voltage level.

During an open LED condition, the 33V zener diode (D1) limits the maximum LED supply voltage to 35.5V. If VLED attempts to increase beyond this level, D1 conducts in reverse direction and pulls the FB pin high, which causes the boost regulator to cut back on the PWM signal and reduce the output voltage.

PWM Dimming

The PWM dimming controls the LED brightness by adjusting the duty cycle of the PWM input signal. A high voltage at the PWM input enables the output current; a low voltage turns off the output current. Connect a signal with peak amplitude of 3V to 5V and with frequency from 100Hz to 2kHz to the PWM input and vary the duty cycle to adjust the LED brightness. The LED brightness increases when the duty cycle increases and vice versa. If an inverted PWM signal is available, use that signal to implement PWM dimming, as shown in Figure 1.

Feedback Compensation

Like any other circuit with feedback, the boost converter that generates the supply voltage for the LED strings needs to be compensated for stable control of its output voltage. As the boost converter is operated in continuous-conduction mode, there exists a right-half-plane (RHP) zero in the power-circuit transfer function. This zero adds a 20dB/decade gain together with a 90-degree phase lag, which is difficult to compensate. The easiest way to avoid this zero is to roll off the loop gain to 0dB at a frequency less than half of the RHP zero frequency with a -20dB/decade slope. For a boost converter, the worst-case RHP zero frequency (F_{ZRHP}) is given by the following equation:

$$F_{ZRHP} = \frac{VLED(1 - D_{MAX})^2}{2\pi \times L \times I_O}$$

where D_{MAX} is the maximum duty cycle, L is the inductance of the inductor, and I_O is the output current, which is the sum of all the LED string currents.

The boost converter used in the MAX16809 EV kit is operated with current-mode control. There are two feedback loops within a current-mode-controlled converter: an inner loop that controls the inductor current and an outer loop that controls the output voltage. The amplified voltage error produced by the outer voltage loop is the input to the inner current loop that controls the peak inductor current.

The internal current loop converts the double-pole 2nd-order system, formed by the inductor and the output capacitor C_{OUT}, to a 1st-order system having a single pole consisting of the output filter capacitor and the output load. As the output load is a constant current (i.e., very high Thevenin impedance), this pole is located near the origin (0Hz). The phase lag created by the output pole for any frequency will be 90 degrees. Since the power-circuit DC gain is limited by other factors, the gain starts falling at -20dB/decade from a non-zero frequency before which the power-circuit gain stabilizes.

Total gain of the feedback loop at DC is given by the following equation:

$$G_{TOT} = G_P \times G_{EA} \times G_{FB}$$

where G_P is the power-circuit DC gain, and G_{EA} is the error-amplifier open-loop DC gain, typically 100dB. G_{FB} is the gain of the feedback network for adaptive control of the VLED, which is seen from VLED to the error-amplifier input (FB pin). The adaptive control senses the voltages at the 16 constant-current-sink outputs and adjusts the feedback to control these voltages to a minimum value (Figure 2). As the LEDs carry constant current, the voltage across the LEDs does not change with variations in VLED. Any change in VLED directly reflects to the constant-current-sink outputs and to the error-amplifier input, making G_{FB} equal to unity.

The DC gain of the power circuit is expressed as the change in the output voltage, with respect to the change in error-amplifier output voltage. As the boost converter in the MAX16809 EV kit drives a constant-current load, the power-circuit DC gain is calculated based on a constant-current load:

$$G_P = \frac{\Delta VLED}{\Delta EA_{OUT}}$$

Calculate the power-circuit DC gain using the following equation:

$$G_P = \frac{1}{\left(\frac{V_{IN}^2}{2 \times L \times F_{SW} \times VLED^2} + \frac{I_O}{V_{IN}}\right) \times R_{CS} \times 3}$$

where R_{CS} is the current-sense resistor, F_{SW} is the switching frequency, and the factor 3 is to account for the attenuation of error-amp output before it is fed to the current-sense comparator.

The power-circuit gain is lowest at the minimum input supply voltage and highest at the maximum input supply voltage. Any input supply voltage between 9V and 16V can be used for power-circuit gain calculation, as the final compensation values obtained are the same.

Calculate the frequency F_{P2}, at which the power-circuit gain starts falling, at -20dB/decade using the following equation:

$$F_{P2} = \frac{(1 - D_{MAX})}{2\pi \times C_{OUT} \times 3 \times R_{CS} \times G_P}$$

where C_{OUT} is the output filter capacitor, which is the parallel combination of C16, C17, C18, and C24. Adjust the output capacitance so that the product of F_{P2} and G_P is below F_{ZRHP} / 6. The value of output capacitance obtained this way will be much greater than the value obtained using the maximum output voltage ripple specification.

The compensation strategy is as follows. The gain-frequency response of the feedback loop should cross 0dB at or below half of the RHP zero frequency, with a slope of -20dB/decade for the feedback to be stable and have sufficient phase margin. The compensation network from COMP pin to FB pin of the MAX16809 (formed by R5, C28, C29, and R11) offers one dominant pole (P1), a zero (Z1), and a high-frequency pole (P3). There are two very low frequency poles and a zero in the loop before the crossover frequency. The function of the zero (Z1) is to compensate for the output pole and to reduce the slope of the loop gain from -40dB/decade to -20dB/decade, and also to reduce the phase lag by 90 degrees.

Choose the crossover frequency to be half of the worst-case RHP zero frequency:

$$F_C = \frac{F_{ZRHP}}{2}$$

Place the zero (Z1) at one-third of the crossover frequency, so that the phase margin starts improving from a sufficiently lower frequency:

$$F_{Z1} = \frac{F_C}{3}$$

Use the following equation to calculate the dominant pole location, so that the loop gain crosses 0dB at F_C:

$$F_{P1} = \frac{F_{ZRHP} \times F_{Z1}}{2 \times G_{TOT} \times F_{P2}}$$

Since the open-loop gain of the error amplifier can have variations, the dominant pole location can also vary from device to device. In the MAX16809 EV kit, the dominant pole location is decided by the error-amplifier gain, so the combined effect is a constant-gain-bandwidth product.

Select the value of R11 such that the input bias current of the error amplifier does not cause considerable drop across it. The effective AC impedance seen from the FB pin is the sum of R11 and R7. It is preferable to keep R7 much lower, compared to R11, to have better control on the AC impedance. Find C29 using the following equation:

$$C29 = \frac{1}{2\pi \times G_{EA} \times (R11 + R7) \times F_{P1}}$$

The location of the zero (Z1) decided by R5 and C29 is given by the following equation:

$$F_{Z1} = \frac{1}{2\pi \times R5 \times C29}$$

Place the high-frequency pole (P_3), formed by C28, C29, and R5, at half the switching frequency to provide further attenuation to any high-frequency signal propagating through the system. The location of the high-frequency pole (F_{P3}) is given by the following equation, and should be used to calculate the value of C28:

$$F_{P3} = \frac{1}{2\pi \times R5 \times \left(\frac{1}{C28} + \frac{1}{C29}\right)^{-1}}$$

The MAX16809 EV kit uses electrolytic capacitors at the output for filtering, so the zero produced by the ESR of the capacitors can be low enough to be within or near the crossover frequency. This zero should be compensated using an additional pole (P4) placed at the ESR zero location. The ESR zero frequency is calculated using the following equation:

$$F_{ZESR} = \frac{1}{2\pi \times ESR \times C_{OUT}}$$

Use the following equation to calculate the value of C35 to place the pole (P4) at the ESR zero frequency:

$$C35 = \frac{1}{2\pi \times F_{ZESR} \times R7}$$

If ceramic capacitors are used at the output for filtering, the frequency of zero produced by the ESR and the capacitance will be above the crossover frequency (0dB gain frequency) of the feedback loop and need not be considered in the compensation design.

Layout Considerations

LED driver circuits based on the MAX16809 device use a high-frequency switching converter to generate the supply voltage for LED strings. Proper care must be taken while laying out the circuit to ensure proper operation. The switching-converter part of the circuit has nodes with very fast voltage changes, producing high-frequency electric fields, and branches with fast current changes, producing high-frequency magnetic fields. As the circuit converts power, the amplitude of these fields will be high and can easily couple to sensitive parts of the circuit, creating undesirable effects. Follow the guidelines below to reduce noise as much as possible:

1) Connect the bypass capacitors from REF and VCC as close as possible to the device and connect the capacitor grounds to the analog ground plane using vias close to the capacitor terminals. Connect the AGND pin of the device to the analog ground plane using a via close to the pin. Lay the analog ground plane on the inner layer, preferably next to the top layer. Use the analog ground plane to cover the entire area under critical-signal components for the power converter.

2) Keep the oscillator timing capacitor and resistor very close to the RTCT pin and make the connection as short as possible. Connect the ground of the timing capacitor to the analog ground plane using a via close to the capacitor terminal. Make sure that no switching node is present near the RTCT node and keep the area of the copper connected to the pin small. Keep the REF connection to the timing resistor short and away from any switching node.

3) Have a power ground plane for the switching-converter power circuit under the power components (input filter capacitor, output filter capacitor, inductor, MOSFET, rectifier diode, and current-sense resistor). Connect all the ground connections to the power ground plane using vias close to the terminals.

4) There are two loops in the power circuit that carry high-frequency switching currents. One loop is when the MOSFET is on (from the input filter capacitor positive terminal, through the inductor, the MOSFET, and the current-sense resistor, to the input capacitor negative terminal). The other loop is when the MOSFET is off (from the input capacitor positive terminal, through the inductor, the rectifier diode, and the output filter capacitor, to the input capacitor negative terminal). Analyze these two loops and make the loop areas as small as possible. Wherever possible, have a return path on the power ground plane for the switching currents on the top-layer copper tracks, or through power components. This reduces the loop area considerably and provide a low inductance path for the switching currents. Reducing the loop area also reduces radiation during switching.

5) The gate-drive current of the MOSFET is another high-frequency switching current to consider. There are two major loops: one during the MOSFET turn-on edge and the second during the turn-off edge. The MOSFET turn-on loop is from the VCC bypass capacitor positive terminal, through the MOSFET driver in the device, the gate-drive resistor, the MOSFET gate to source (CGS and CGD), and the current-sense resistor to the VCC bypass capacitor negative terminal. There is no direct path for the current from the current-sense resistor to return to the VCC bypass capacitor through the ground plane, as the VCC bypass capacitor is connected to the analog ground plane and the current-sense resistor is connected to the power ground plane. The best solution is to connect the analog ground plane to the power ground plane directly under the MOSFET gate-drive track. This ensures that the turn-off current also has a return path on the ground plane.

6) The drain node of the MOSFET is a switching node. Keep this node area small to reduce radiation and capacitive coupling to other sensitive parts of the circuit. However, the track should be wide enough to carry the large switching currents.

7) Keep the node area and track length on the FB pin small to reduce any noise pickup.

8) Connect the power ground plane for the constant-current LED driver part of the circuit to the boost-converter output filter capacitor negative terminal.

Power Dissipation

The MAX16809 dissipates power during normal operating conditions. The heat transferred to the exposed pad from the die should be properly dissipated to the board to prevent the device from entering into thermal shutdown. The exposed pad land area on the top layer should be of the same size as that of the exposed pad. Thermal vias are used to carry the heat from the exposed pad to other layers of the board and spread it across the board area through copper planes. Thermal vias should have a 0.4mm hole size and should be placed at a distance of 1mm from center to center. For a four-layer board, these vias should be connected to the bottom ground plane and to one internal ground plane. Do not use thermal relief for the thermal vias; instead, use solid copper to get the minimum thermal resistance.

Use the following equation to calculate the total power dissipated in the MAX16809 device during normal operation:

$$P_D = \sum_{N=0-15} VS_N \times IN + I_B \times V_{IN}$$

where IN is the LED current in channel N, VS is the operating voltage of each of the LED driver outputs with respect to GND pins, I_B is the input bias current of MAX16809 including the average of MOSFET drive current, and V_{IN} is the input supply voltage. To dissipate 1W of power, the exposed pad of the device should be connected to a minimum of two square inches of copper ground plane with 70µ copper thickness.

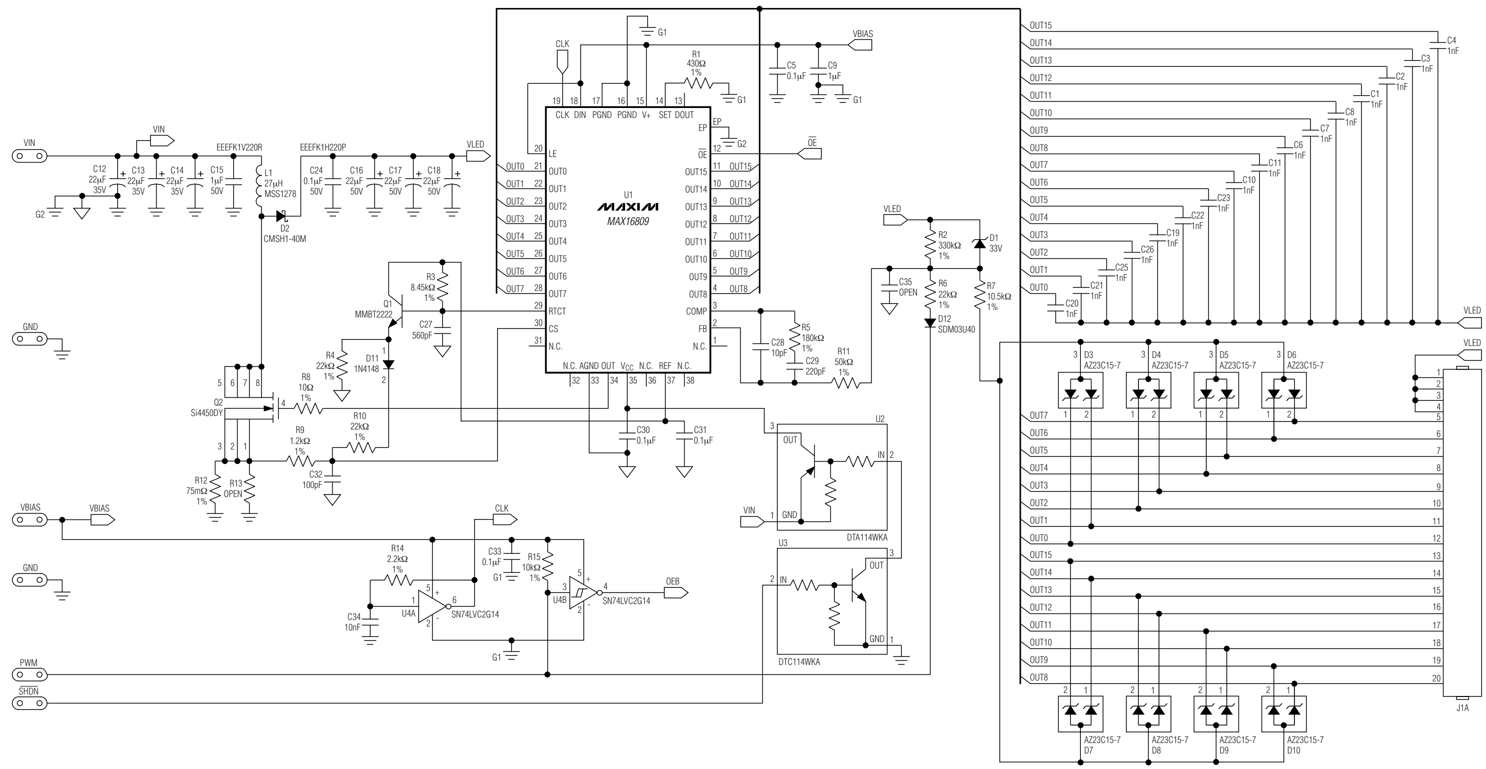

Figure 2. MAX16809 EV Kit Schematic

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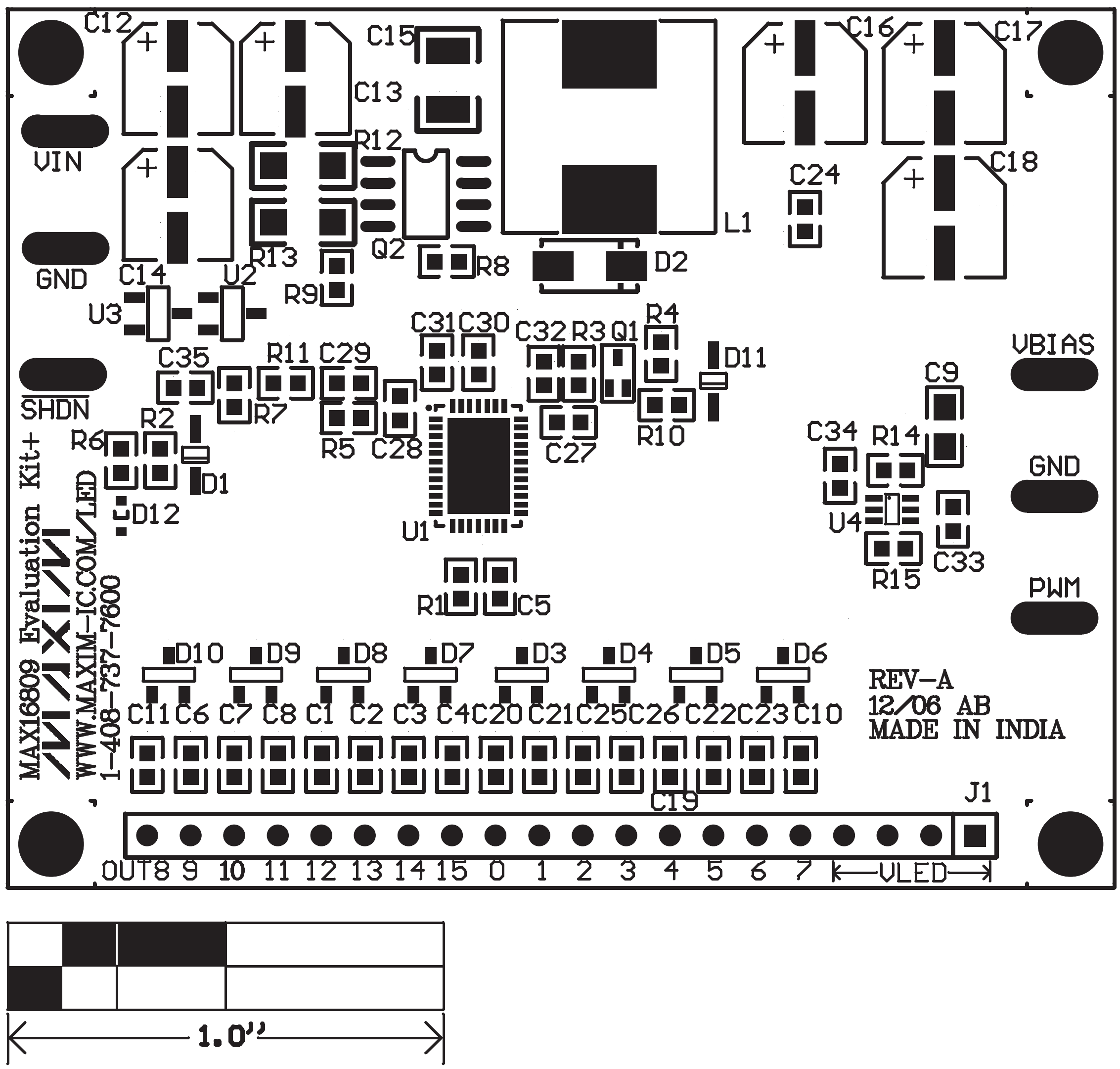

Figure 3. MAX16809 EV Kit Component Placement Guide—Component Side

MAX16809 Evaluation Kit

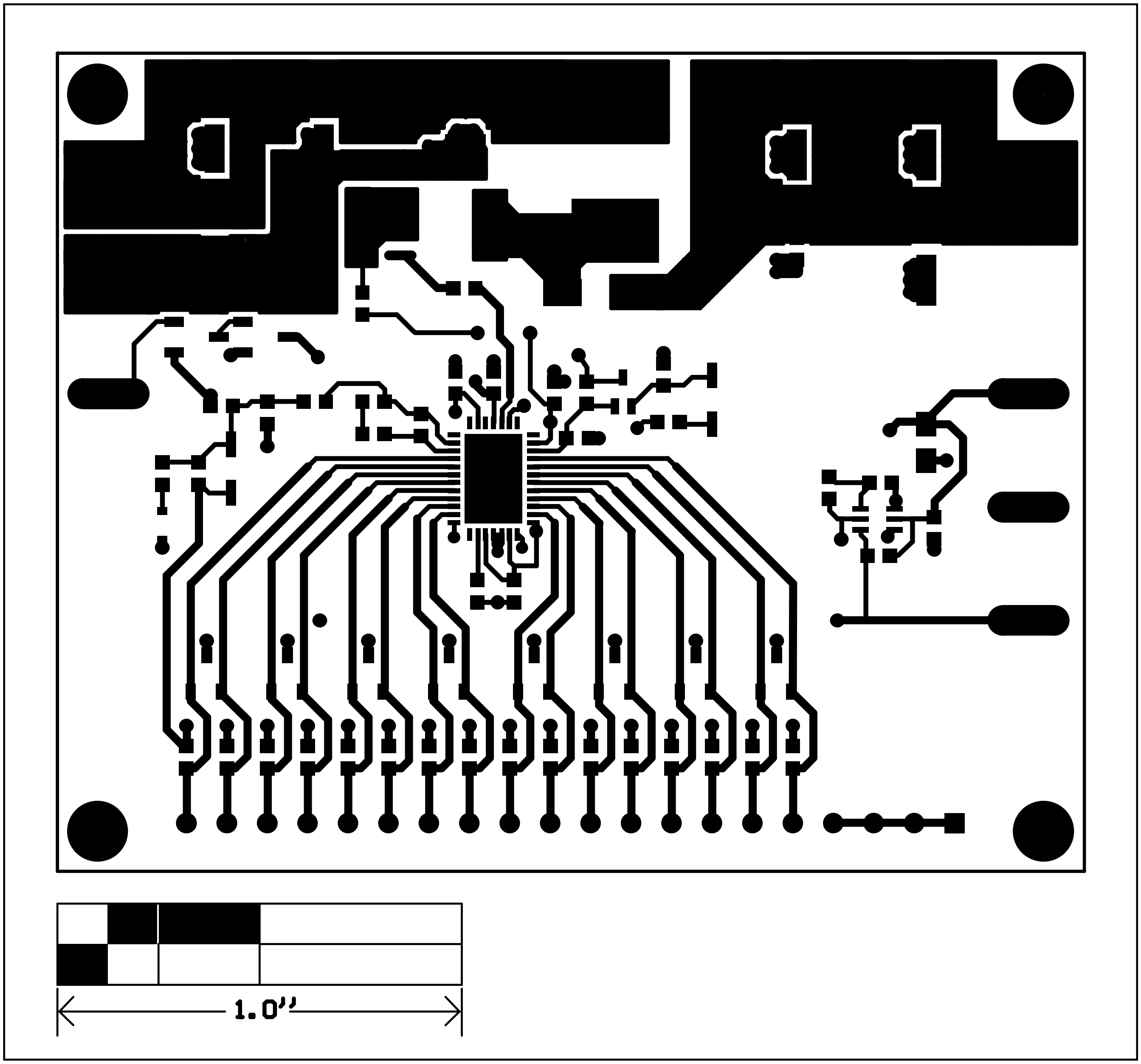

Figure 4. MAX16809 EV Kit PCB Layout—Component Side

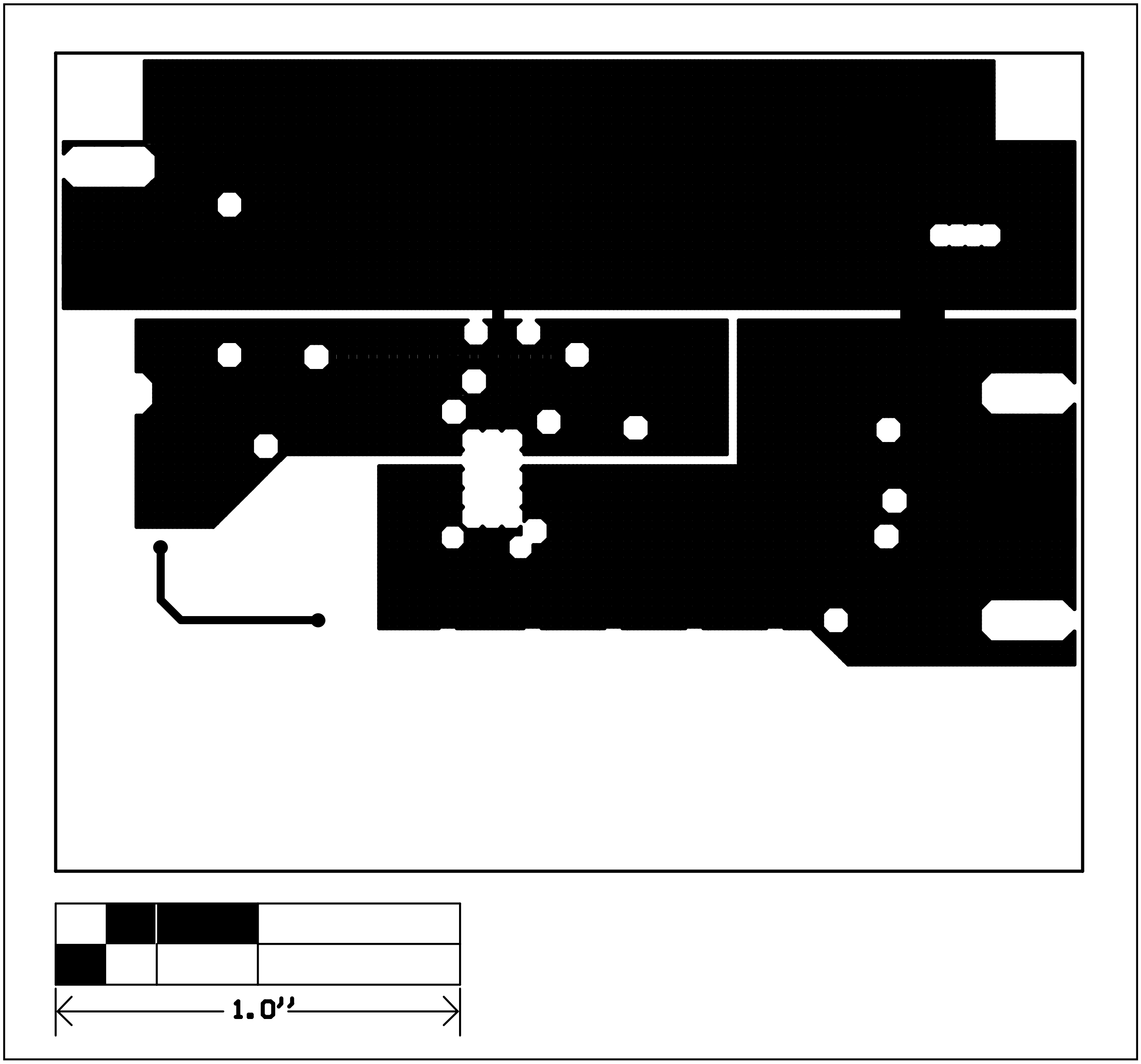

Figure 5. MAX16809 EV Kit PCB Layout—Inner Layer 1

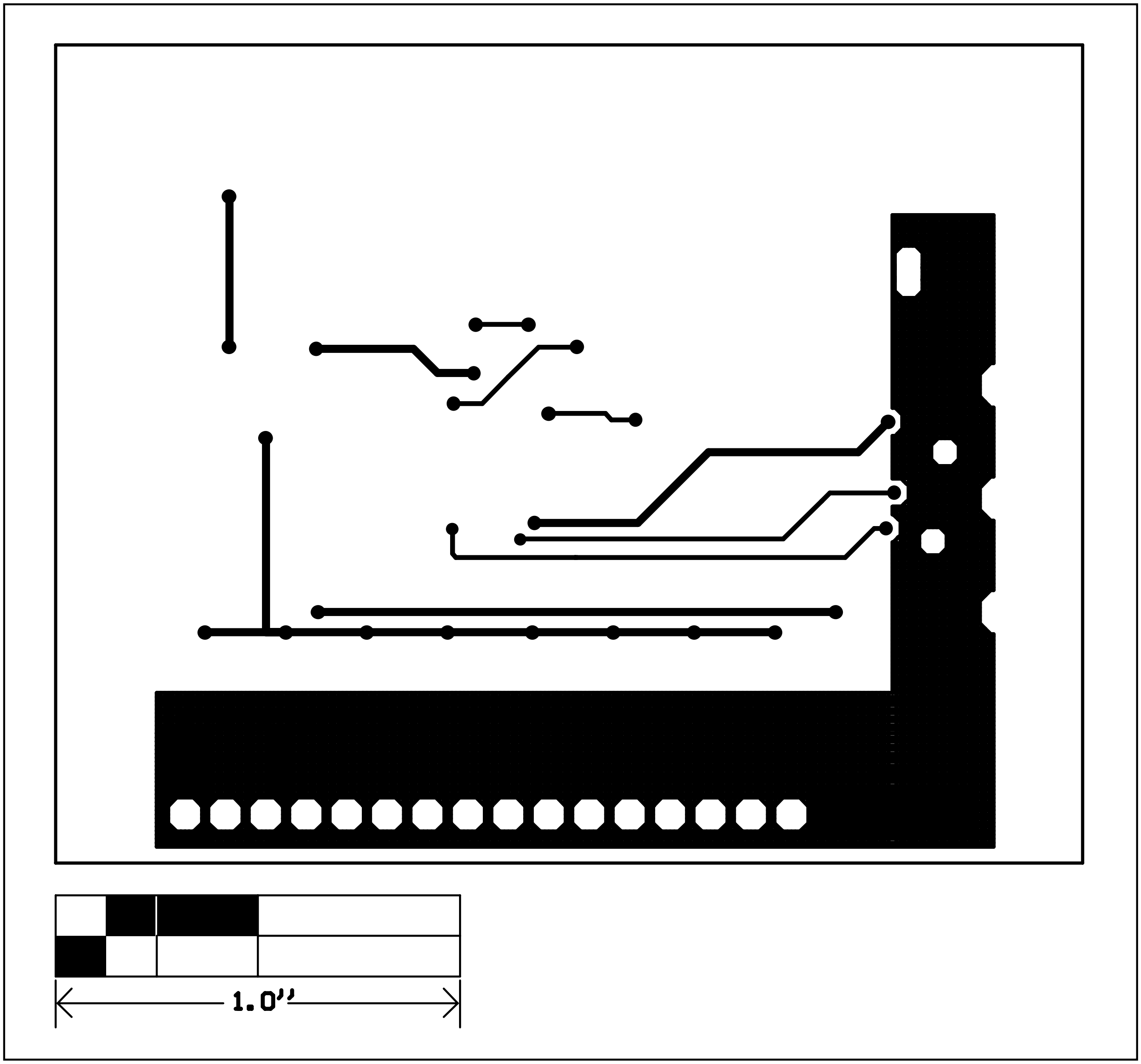

Figure 6. MAX16809 EV Kit PCB Layout—Inner Layer 2

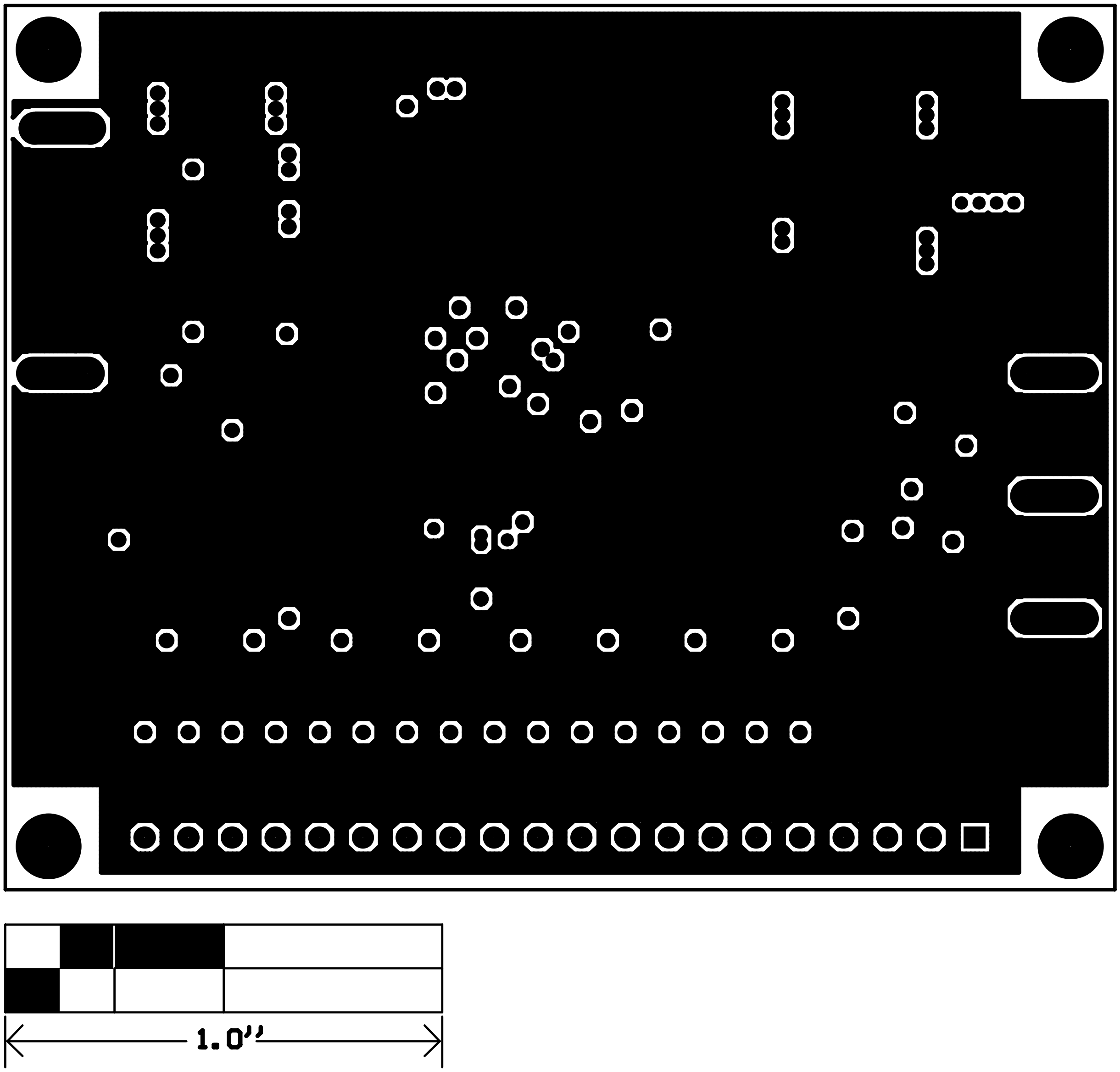

Figure 7. MAX16809 EV Kit PCB Layout—Solder Side