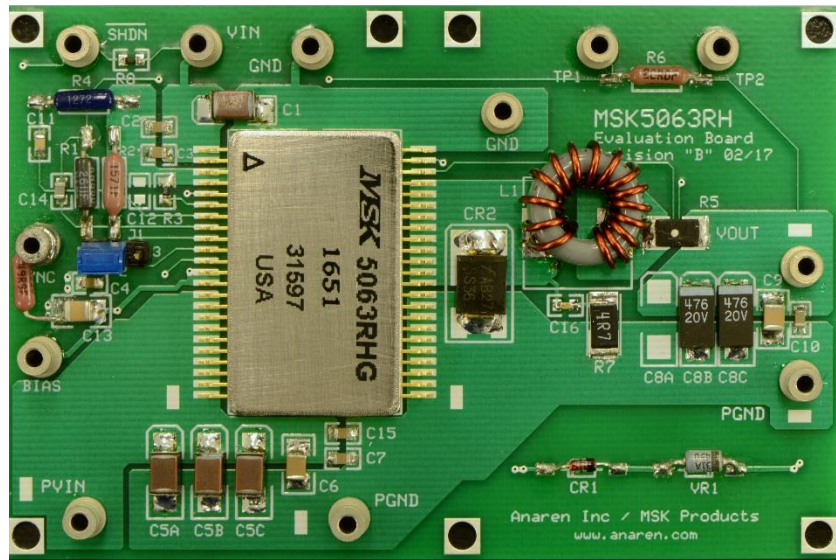


**MSK5063RH**  
**Evaluation Board User's Guide**  
 REV A, 5-12-17

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**Introduction**

The MSK5063RH is a Radiation Hardened wide input voltage range step-down synchronous switching regulator. The MSK5063RH integrates A RAD HARD Current Mode PWM controller, high and low side RAD HARD Power MOSFETs, bootstrap supply, and peripheral components in a hermetically sealed in a 46-pin flatpack. The wide input and output range and feature set make these regulators suitable for many medium to high power applications. The adjustable operating frequency gives designers flexibility to optimize performance while avoiding sensitive frequency bands. Synchronization with other switching regulators can be used to reduce EMI and component sizes.

The MSK5063RH Evaluation Board provides a platform from which to evaluate performance and experiment with design changes for your application. Excessive ripple, instability, and other costly technical problems in flight hardware can be avoided with evaluation early in the design cycle.

This application note is intended to be used in conjunction with the MSK5063RH data sheet and the LT3845 data sheet. Reference those documents for additional application information and specifications.

## Evaluation Board Setup

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Two generic configurations for the MSK5063RH Evaluation Board are detailed herein. Each version is tailored to different operating regimes. These are simply offered as a starting point for adaptation into a broad range of applications.

- The low output voltage configuration operates with  $4V \leq PVIN \leq 28V$ ; Common for POL applications
- The high output voltage configurations operates over  $18V \leq PVIN \leq 60V$ ; covering the MIL-STD-704F 28VDC bus

There are several possible schemes for supplying power to the MSK5063RH. The optimal approach for any given application will be depend on available bus voltages, capacities, and the required conversion ratio. VIN, BIAS, and PVIN input pins can be ganged together or supplied independently. Individual pin ratings must be observed and internal component power dissipation levels understood. Carefully review the pin functional descriptions and specifications provided in the MSK5063RH datasheet to ensure the startup and operating requirements are met.

Use the standard turret terminals to connect to your test equipment. Connect a power supply across the Vin and GND terminals. Connect a power supply to PVIN and PGND terminals. For the low output voltage configuration the BIAS terminal can be left unconnected or supplied from an external source. For the high output voltage configuration BIAS is back-fed from the regulated output. Other pin functions are treated separately later in this document.

Connect measurement equipment using Kelvin Sensing techniques to ensure accurate load regulation and efficiency measurements. The ground wire from an oscilloscope probe acts as an antenna picking up radiated EM fields making waveforms such as output ripple voltage appear noisy. For improved results, the probe tip can be touched against the output turret with the bare ground shield pressed against the ground turret.

## MODE

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The MODE pin connection determines the regulator's behavior under light loads. Discontinuous Conduction Mode (DCM) disables the bottom MOSFET gate drive as the inductor current approaches zero thus causing the regulator to behave as a non-synchronous rectified buck regulator. DCM mode maximizes efficiency at low load current by reducing gate drive power and conduction losses. Continuous Conduction Mode (CCM) has the benefits of improved noise characteristics, faster response to large load steps, and does not require a minimum load to maintain tight regulation. Selecting either mode for the MSK5063RH is as simple as connecting MODE (pin 9) to VFB (pin 8) for DCM or BIAS (pins 12 & 13) for CCM. The MSK5063RH Evaluation board is furnished with a solderless jumper, J1, to allow quick evaluation of either configuration.

**Note:** Power down while changing J1.

## SHDN

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The Eval Board is factory equipped with a 100k $\Omega$  resistor pulling the  $\overline{\text{SHDN}}$  pin up to VIN. The  $\overline{\text{SHDN}}$  pin is internally clamped to  $\approx 6V$  and must be current limited to less than 1mA.

A simple Under Voltage Lockout (UVLO) circuit is implemented using just one additional resistors connected from the  $\overline{\text{SHDN}}$  pin to ground. UVLO hysteresis can be increased with a positive feedback resistor from the output to  $\overline{\text{SHDN}}$ .

**BIAS**

The BIAS pins are the output of the internal 8V linear regulator and is sourced from VIN. As with any linear regulator, efficiency and power dissipation must be considered carefully while operating at large input to output voltage differentials. Most of the internal control and gate drive circuitry is powered from the BIAS pin node. Internal losses can be reduced by driving this pin with an external higher voltage source.

The high output voltage version of the Evaluation Card diode OR's the output to BIAS pins. This approach works well for applications where the output voltage is greater than ≈8.7V. For output voltages greater than 14V, some means of down regulating before the BIAS pins is required to remain below the BIAS pin 14V absolute maximum rating. An appropriately sized Zener diode connected in series with the OR'ing diode is sufficient for this task. Adding a flyback winding to the output inductor as described below could be a more efficient option.

The low output voltage version of the Evaluation Board is configured to draw BIAS power either directly from VIN through the internal linear regulator, or back driven from an external supply attached to the BIAS turret. In low output voltage scenarios, a BIAS supply can be efficiently generated by adding a flyback winding to the output inductor and connecting one side to either ground or the main output and the other side through a blocking diode to the BIAS pins. Reference the simplified circuit diagram below.

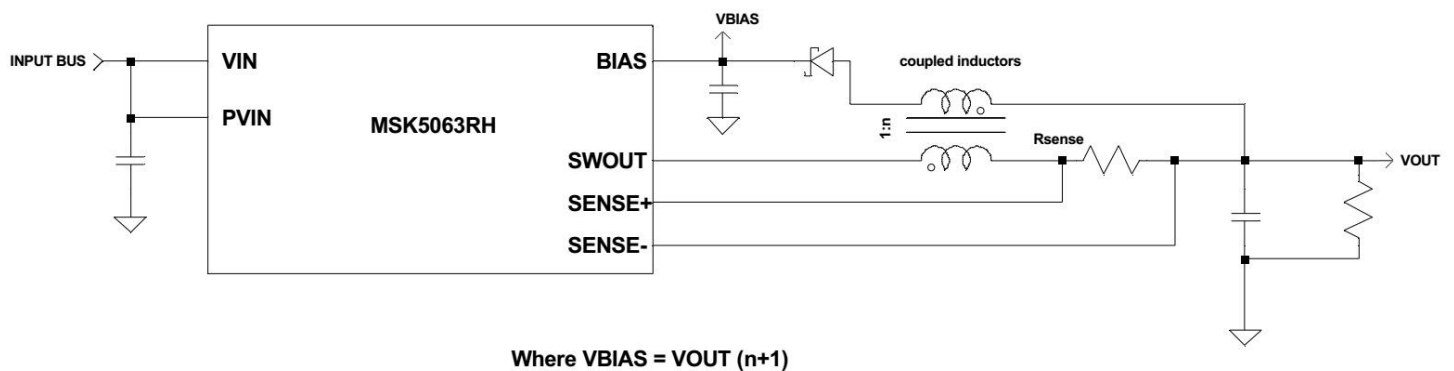


Figure 1 Simplified Flyback BIAS supply Schematic

**COMP**

The error amplifier in the MSK5063RH is an Operational Transconductance Amplifier (OTA). The COMP pin is the OTA's output. Type II compensation is implemented using R4, C11, and C12. C11 sets a pole at the origin, giving high DC gain for reduced regulation error. R4 and C11 form a zero, setting the mid-band gain. C12 sets the high frequency pole. C14 adds an additional zero if needed.

$$DC\ gain = \frac{R2}{R1 + R2} \times \frac{gm \times R4 \times C11}{C11 \times C12}$$

$$f_z = \frac{1}{2\pi \times R4 \times C11}$$

$$f_p = \frac{1}{2\pi \times R4 \times C11 \times C12}$$

## Switching Frequency Set Point

The Switching frequency of the MSK5063RH is programmable from 100kHz to 500kHz with a single resistor from the FSET pin to ground. Both versions of the Evaluation Board are configured for 300kHz operation,  $R_3 = 49.9\text{k}\Omega$ . Figure 2 plots the Switching Frequency vs  $R_{SET}$  relationship for quick reference. Choosing the switching frequency involves balancing efficiency, range of operation and components sizes. Higher switching frequencies are generally favored for the substantial savings in filter component board area at the expense of lower efficiency.

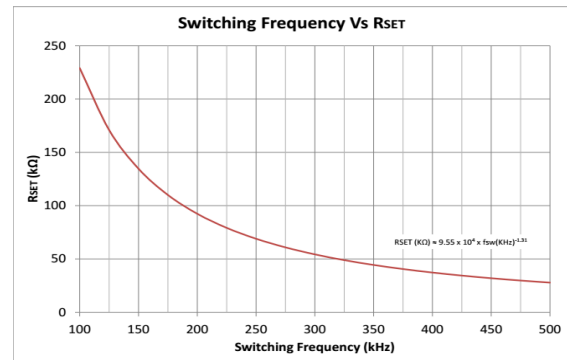


Figure 2

## Synchronization and Paralleling

The SYNC pin on the MSK5063RH evaluation board is pulled to GND through a 49.9kΩ resistor. Synchronization eliminates beat frequency issues that could otherwise occur when multiple switching regulators share a common power bus.

Multiple MSK5063RH modules can be paralleled by synchronizing and connecting the COMP pins. There are several advantages to paralleling multiple converters compared to a single larger switcher. Inductor size can be considerably reduced. For example, three 2A inductors store less energy ( $LI^2/2$ ) than one 6A inductor so are far smaller. Synchronizing three converters 120° out of phase with each other minimizes input and output ripple currents. This reduces the ripple rating, size and cost of filter capacitors.

## Output Voltage Programming

The output voltage is determined by the R1-R2 divider network connected between the Output, VFB pin and GND.

$$V_{out} = V_{FB} \times \left( \frac{R_1}{R_2} + 1 \right) \quad \longrightarrow \quad R_1 = R_2 \times \left( \frac{V_{out}}{V_{FB}} - 1 \right)$$

Given:  $V_{REF} = 1.231\text{V Typ.}$   
 (3.3V Eval Board):  $R_1 = 2.64\text{k}\Omega$ ,  $R_2 = 1.57\text{k}\Omega$   
 (15V Eval Board):  $R_1 = 17.56\text{k}\Omega$ ,  $R_2 = 1.57\text{k}\Omega$

The VFB pin is a high impedance node and is susceptible to noise pickup. Use minimal trace lengths to R1 and R2 and avoid routing near switching nodes.

## Setting the Output Current Limit

R5 senses the inductor current. The maximum peak inductor current ( $I_{out\_max}$ ) is approximately  $100mV/R5$ . The maximum load current is the peak inductor current minus half of the peak-to-peak ripple current. Both versions of the Eval Board are equipped with a  $10m\Omega$  1W surface mount resistor.  $I_{out\_max}$  for either board is typically greater than 8A across all conditions.

$$R5 = \frac{100mV}{I_{out\_max} + (0.5 \times I_{ripple})}$$

## Inductor Selection

The critical parameters for inductor selection are its value, saturation current specification, Peak Current, and RMS current. A common starting point for peak to peak current ripple is 20% to 30% of the maximum load current. Use the following equations to determine the ripple current and RMS current:

$$\Delta I = \frac{V_{out} \times (V_{IN} - V_{out})}{L \times f_{sw} \times \Delta I} \quad I_{RMS} = \sqrt{I_{out}^2 + \frac{\Delta I^2}{12}}$$

The control IC has an internal fixed slope compensation ramp. The minimum inductance required to avoid sub-harmonic oscillation is given by:

$$L_{min} > V_{out} \times \frac{2D_{max} - 1}{D_{max}} \times \frac{8.33 \times R_{SENSE}}{f_{sw}}$$

Where:

D = Duty Cycle =  $V_{OUT}/V_{IN}$

$f_{sw}$  = Switching Frequency

$I_{out}$  = DC load Current

$\Delta I$  = Peak to Peak Ripple current

Provide sufficient derating to prevent saturation and thermal overstress when selecting the inductor.

## Input Capacitors

Buck switching regulators are constant power transfer devices and thus exhibit a negative input impedance. That is to say; as input voltage decreases, input current increases. This has important implication for input filtering and bypass capacitors. The power bus impedance must be maintained sufficiently below the input impedance of the converter in order to avoid control loop interaction or even creation of a negative resistance oscillator. The input bus impedance in many applications is inductive and will resonant with the regulator's input capacitors. Adequate damping must be implemented.

The primary factors in selecting the type of input capacitors for any buck switching regulators are the DC voltage rating, ripple voltage and ripple current ratings. Current is drawn from the input bus in trapezoidal pulses with fast rising edges. The input ripple current for a buck converter is high;  $\approx I_{OUT}/2$  at 50% duty cycle. Finding capacitors that economically satisfy all requirements with high reliability is a challenge. Tantalum capacitors in general are not well suited for high voltage and high ripple current applications. MIL-PRF-49470 series ceramic stack capacitors may be one viable option to consider.

The MSK5063RH Evaluation Board input capacitor selection was geared to work under a very broad range of conditions and is likely excessive for most real applications.

## Output Capacitor Selection

The output capacitors C7 and C8 are AVX TAZ series 47uF tantalum capacitors. See BOM for specific part number. These capacitors were selected to provide a design starting point using high reliability MIL-PFR-55365/4 qualified capacitors (CWR). The output capacitor value and its parasitic elements determine several important performance characteristics including; loop stability and bandwidth, load step response, output ripple voltage start, start-up surge current and output response due to Single Event Transients.

The output ripple voltage is determined by the inductor ripple current and output capacitor ESR. If lower output ripple voltage is required, increase the inductor value and/or reduce the ESR by choosing a different capacitor or placing more capacitors in parallel. For very low ripple, an additional LC filter added to the output may be a more suitable solution. Re-compensation of the loop may be required if the output capacitance is altered. The output contains very narrow voltage spikes caused by the parasitic inductance of C7 and C8. Ceramic capacitors C11 and C12 help absorb these spikes on the demo board. In application, distributed trace impedance and local bypass capacitors will help serve this function.

With current mode control, a single pole is established by the load impedance and output capacitance. ESR of the bulk output capacitance forms a zero and is exploited to improve phase margin. Ceramic capacitors with equivalent bulk have comparatively negligible ESR. The impact on loop dynamics should be considered if using large values of ceramic capacitors. The Evaluation Boards are designed to provide good stability margins over a broad range of applications and conditions. There is no one size fits all solution, analysis and experimental verification will ensure performance and high reliability.

## Compensation and Measuring Loop Stability

The compensation components on the MSK5063RH evaluation board were selected to provide gain and phase margins sufficient for many High-Rel applications. Stability margins should be re-assessed after any component change.

The MSK5063RH Evaluation Board is designed to facilitate a common stability measurement technique. An isolation transformer is used to inject the AC output of a Network Analyzer into the control loop across R6 (terminals TP1 and TP2). Connect the inputs of the analyzer to TP1(loop input) and TP2(loop output). Use GND<sub>2,3</sub> for the ground reference for the network analyzer inputs. The analyzer will inject a frequency swept signal into the feedback loop, and plot the control loop gain and phase response. Figure 3 illustrates typical results for the 15V configuration at 1A and 8A. Phase Margin is greater than 71° and Gain Margin is greater than 23dB

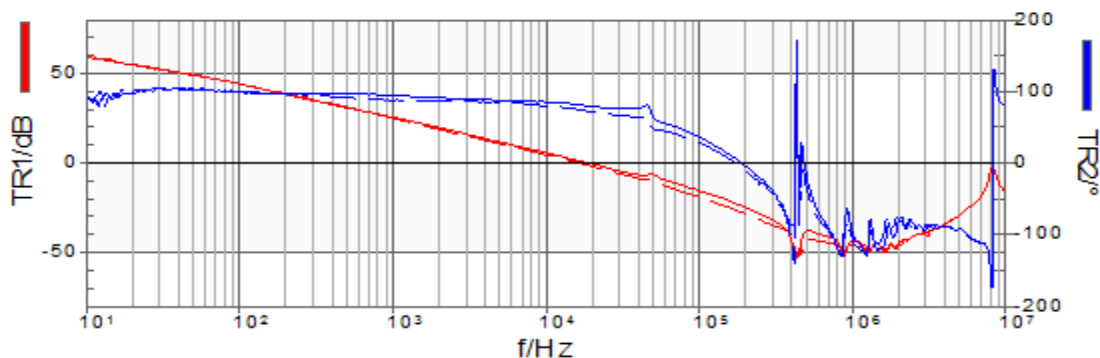


Figure 3, MSK5063RH Eval Board, 15Vout, 1A and 8A



## SWOUT Snubber

Parasitic inductance in the commutating loops and switch node capacitance form a series resonant circuit. The SWOUT node will ring after every transition at the resonant frequency given by:

$$f_{res} = \frac{1}{2\pi\sqrt{L_s C_p}}$$

Good layout practices and component selection can minimize but not eliminate these parasitic elements. Some applications may require dampening to control the amplitude and frequency of the switch node ring voltage. A dissipative RC snubber network is commonly used for this task, R7 + C16. The snubber component values can be optimized by first measuring the ring frequency and then iteratively adding capacitance to the switch node until the ring frequency is reduced by a factor of .707. This added capacitance is approximately equal to the switch node capacitance. Larger capacitors will reduce the ring amplitude and frequency at the expense of increased power dissipation.

The resistor value determines how quickly the ringing is damped. A good starting point for the resistor value is set equal to the snubber capacitor impedance at the ring frequency. Resistor values between 1Ω and 10Ω are common and can be adjusted as needed. The average power dissipation is equal to the energy stored and released from the snubber capacitance during a switch cycle, given by:

$$P_{snubber} \cong C_{snubber} \times PVIN^2 \times f_{Sw}$$

The majority of this power is dissipated in the snubber resistor, be sure to size it appropriately.

## Layout Considerations

Switch mode regulators generate very high di/dt and dv/dt signals in the switching loop components and traces. In many conditions there is significant energy contained in these high frequency switching paths.

Figure 4 illustrates a critical high frequency loop. The area enclosed by this loop should be kept as small to minimize parasitic inductance to EMI.

The lower the impedance of a trace, the less likely it is to be susceptible to near-field disturbances. Therefore, always choose the lowest impedance end of a branch for longer routes. For example consider the feedback divider top resistor.

Given a choice of locating the resistor close to the high impedance error amp input or the low impedance output, always place it close to the error amp and make the long run to the low impedance output voltage node.

Use ground planes and single point connections to keep a common reference. Make sure components are oriented such that the signal ground referenced nodes are placed away from regions of high current power ground return paths. The signal ground reference is the output decoupling capacitors C11 and C12 and GND<sub>2</sub>.

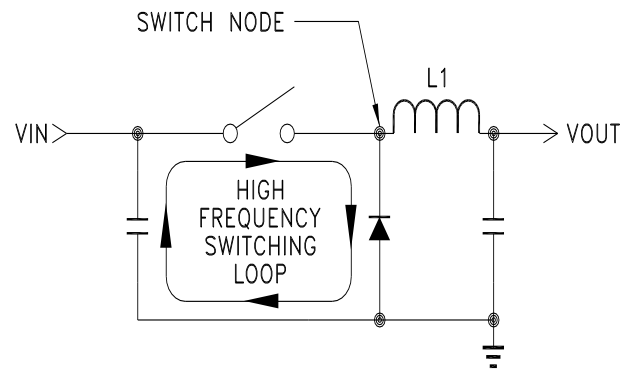


Figure 4

## Typical Performance

### 3.3V Configuration

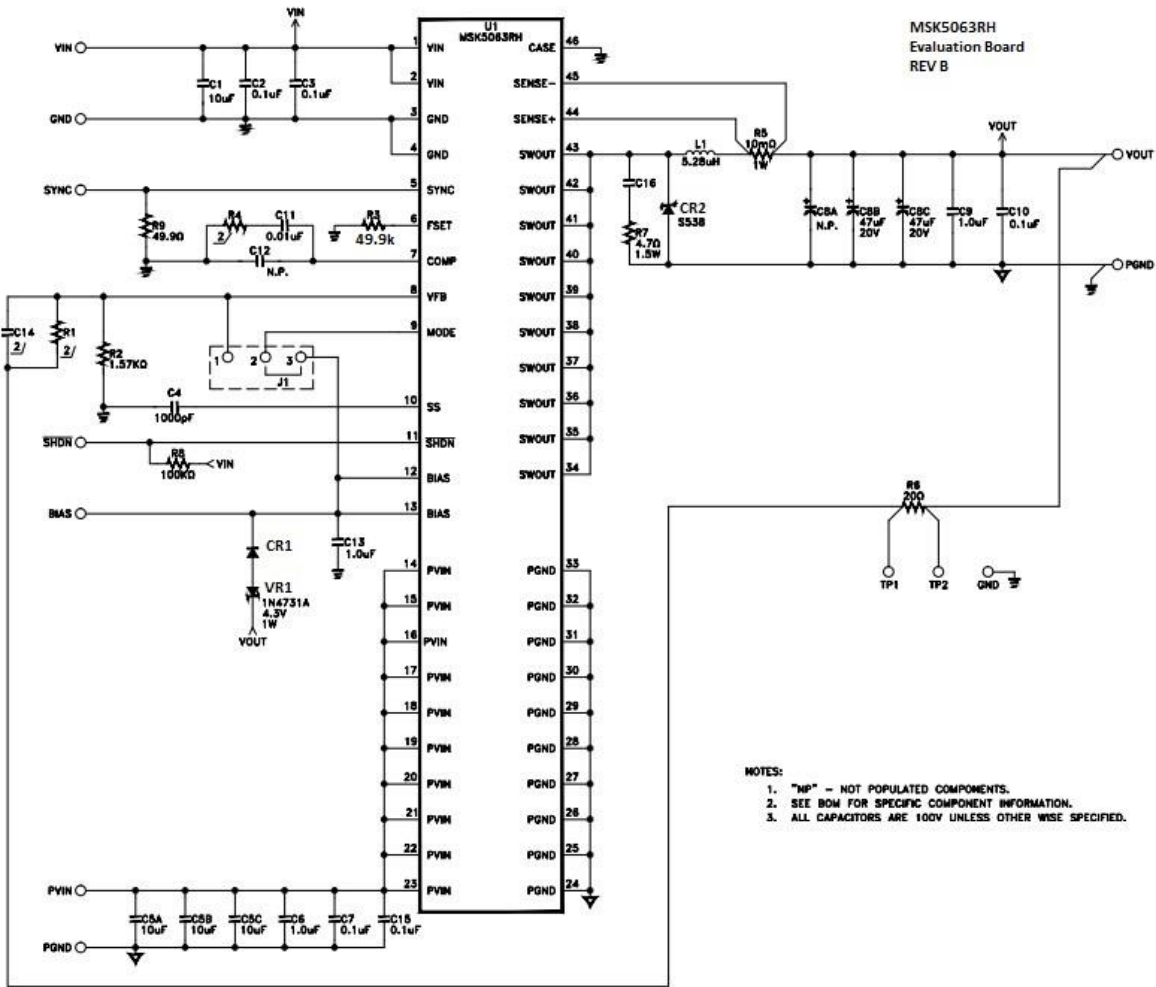
Parameter	Conditions	Units	Typical
Output Voltage	IL = 5A, PVIN = 5V	V	3.3
Switching Frequency	IL = 5A, PVIN = 5V	kHz	300
Output Ripple Voltage	IL = 5A, PVIN = 5V	% Vout	< 0.5
Line Regulation	IL = 5A, $4.5V \leq PVIN \leq 28V$	%	.05
Load Regulation	IL = 0A to 8A, PVIN = 5V	%	-.15
Efficiency	IL = 5A, PVIN = 5V	%	83
Current Limit	PVIN = 5V	A	10
Gain Margin	IL = 5A, PVIN = 5V	dB	20
Phase Margin	IL = 5A, PVIN = 5V	Deg	80

### 15V Configuration

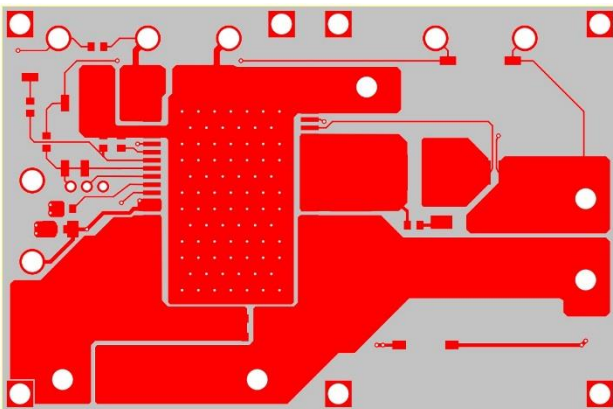
Parameter	Conditions	Units	Typical
Output Voltage	IL = 5A, PVIN = 28V	V	15
Switching Frequency	IL = 5A, PVIN n = 28V	kHz	300
Output Ripple Voltage	IL = 5A, PVIN = 28V	% Vout	< 0.5
Line Regulation	IL = 5A, $V_{in} = 18V \leq PVIN \leq 50V$	%	.05
Load Regulation	IL = 0A to 8A, PVIN = 28V	%	-.15
Efficiency	IL = 5A, PVIN = 28V	%	90
Current Limit	Vin = 28V	A	8.5
Gain Margin	IL = 5A, PVIN = 28V	dB	23
Phase Margin	IL = 5A, PVIN = 28V	Deg	75



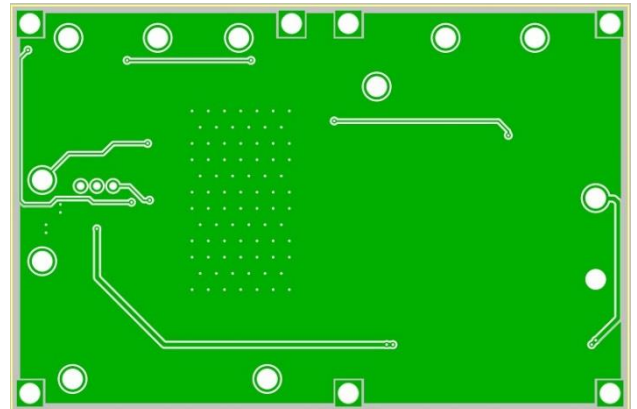
**MSK5063RH Evaluation Board Schematic**



**MSK5063RH Evaluation Board PCB Artwork**



Top Side



Bottom Side

**MSK5063RH Evaluation Board Bill Of Materials**

Ref Des	Description	Manufacturer	Part Number
U1	Rad Hard High Voltage Synchronous Switching Regulator Controller	Anaren, Inc. – MSK Products	MSK5063RHG
CR1	1N4148		
VR1	1N4731A, Zener diode, 4.3V, 1W		
CR2	Diode, Schottky, 3A 80V	Fairchild	SS38
C2, 3, 10, 15	0.1uF, 10%, 100V, X7R, 8050	AVX	08053C104K
C4	1000pF, 10%, 100V, X7R, 8050	AVX	08053C102K
C1, 5A, 5B, 5C	10uF, 20%, 100V, X7S	TDK	CKG45NX7S2A106M
C6, 9, 13, 15	1.0uF, 10%, 100V, X7R, 1210	AVX	12103C105K
C8A, 12	Not Populated		
C8B, 8C	47uF, 10%, 20V, Tantalum, Case Size H	AVX	TAZH476K020L
C11	0.01uF, 10%, 100V, X7R, 8050	AVX	08053C103K
C16	470pF, 10%, 100V, X7R, 6030	AVX	06031C741K
C14	<b>3.3V</b> 6800pF, 1%, 100V, NPO, 8050	AVX	08051A6821JAT
	<b>15V</b> Not Populated		
L1	5.28uH, MPP toroid, bifilar 14 turns, 20AWG Cu	Magnetics Inc	Core: 55050-A2
R1	<b>3.3V</b> 2.6kΩ, 1%, 1/8W, RNC		
	<b>15V</b> 17.5KΩ, 1%, 1/8W, RNC		
R2	1.57KΩ, 1%, 1/8W, RNC		
R3	49.9k, 1% 1/8W, 8050		CRCW060330K1FKEA
R4	<b>3.3V</b> 12.5kΩ, 1% 1/8W, RNC		
	<b>15V</b> 25kΩ, 1% 1/8W, RNC		
R5	0.010Ω, 1W, 0.1%, 2512	Vishay Dale	Y14870R01000B9R
R6	Resistor 20Ω, 1% 1/8W, RNC		
R7	4.7Ω, 5%, 1.5W	Vishay Dale	CRCW25124R70JNEGHP
R8	100kΩ	Vishay Dale	CRCW0603100KFKEA
R9	49.9Ω, 1%, 1/8W, RNC		

**Notes:**

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