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Modified DCR Current Sensing Expands Inductor Choices For Peak-Current-Mode Buck Regulators

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Inductor dc-resistance (DCR) current sensing offers several benefits over the traditional current-sense resistor approach, including higher efficiency, lower cost and reduced solution size. In current-mode control, the current-sensing element is part of the control loop, and its characteristics directly impact loop stability and current-limit accuracy. The Texas Instruments (TI) LM25148, LM25149, LM25137-Q1, LM70880 and similar buck regulators incorporate output current sensing with internal slope compensation, which introduces specific considerations for inductor and current-sense resistor selection.

At first glance, internal slope compensation may seem to restrict your design choices. With careful design, however, DCR current sensing can provide significant flexibility, even when the inductor selected has a low inductance value and a relatively high DCR.

In this article, I'll present a method for designing and tuning a DCR current-sensing network to accurately replicate the inductor current in peak-current-mode buck converters. By matching the resistor-capacitor (RC) network's time constant (τ_{RC}) to the ratio of inductance and its DCR, you can align the sensed waveform with the actual inductor current for both dc scaling and ac ripple attenuation, enabling current-limit control and stable loop operation under different inductor values.

Temperature variation also impacts DCR sensing accuracy, as DCR typically increases and inductance can shift with core temperature.[1] Although not discussed in this article, you will need to factor these effects into current-limit margin planning from the outset, as they alter both the sensed voltage scaling and the matched τ_{RC}.

This article begins with a review of how DCR current sensing works. It then analyzes how the sensed waveform representing the current, V_{CS}, consists of ac and dc components, and how these components are affected by a change in the resistor value from that required for matching the inductor and R-C-network time constants. It's then shown how the scaling of the ac and dc components of the sensed waveform can be accomplished with a resistor divider applied to the R-C network.

This technique is the basis for applying inductor DCR current sensing in converter applications with current-mode control and fixed internal slope compensation, while allowing the designer greater latitude in the inductor selection. The inductor can then be optimized for size, efficiency or other parameters. In the latter sections of this article, I demonstrate how to apply this modification of the R-C network in a buck converter example.

DCR Current-Sensing Overview

As shown in Fig. 1, DCR current sensing works by developing a voltage across the inductor's winding resistance (R_{DCR}) and replicating this voltage across a parallel R_{CS}-C_{CS} network. The voltage developed across C_{CS} is then the sensed current signal.

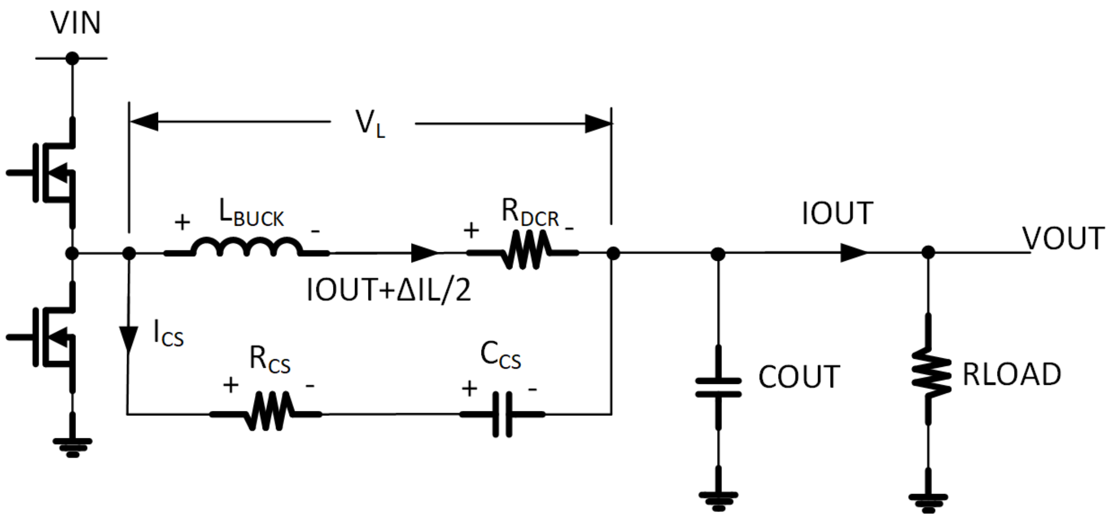

Fig. 1. DCR current-sensing circuit using R_{DCR} and an R_{CS}-C_{CS} network in a buck converter.

For accurate sensing, you must match the inductor's $\frac{L_{BUCK}}{R_{DCR}}$ τ_{RL} to the $R_{CS} \times C_{CS}$ τ_{RC}:

$$\tau_{RL} = \frac{L_{BUCK}}{R_{DCR}}, \tau_{RC} = R_{CS} \times C_{CS} \quad (1)$$

When the time constants are equal, the voltage across C_{CS} faithfully reproduces the inductor's current waveform.

τ_{RC} Derivation

During the on-time (D period) of the buck converter, the inductor's inductance and winding resistance R_{DCR} develop an inductor voltage (V_L). The design goal is for the voltage across C_{CS} to match the voltage across R_{DCR}.

Equation 2 expresses V_L during the D period as

$$V_L = (s \times L_{BUCK} + R_{DCR}) \times \left(IOUT + \frac{\Delta IL}{2}\right) \quad (2)$$

Assuming that the voltage across R_{CS} and C_{CS} equals V_L:

$$V_L = I_{CS} \times \left(R_{CS} + \frac{1}{s \times C_{CS}}\right) \quad (3)$$

Substituting equation 3 into equation 2 results in

$$I_{CS} \times \left(R_{CS} + \frac{1}{s \times C_{CS}}\right) = (s \times L_{BUCK} + R_{DCR}) \times \left(IOUT + \frac{\Delta IL}{2}\right) \quad (4)$$

Solving equation 4 for I_{CS} leads to

$$I_{CS} = \frac{(s \times L_{BUCK} + R_{DCR}) \times \left(IOUT + \frac{\Delta IL}{2}\right)}{\left(R_{CS} + \frac{1}{s \times C_{CS}}\right)} \quad (5)$$

Equation 6 calculates the voltage across C_{CS}:

$$V_{CS} = I_{CS} \times \frac{1}{s \times C_{CS}} \quad (6)$$

Substituting equation 6 into equation 5 results in

$$V_{CS} = \frac{(s \times L_{BUCK} + R_{DCR}) \times \left(IOUT + \frac{\Delta IL}{2}\right)}{\left(R_{CS} + \frac{1}{s \times C_{CS}}\right)} \times \frac{1}{s \times C_{CS}} \qquad (7)$$

Then, multiplying out equation 7 and simplifying yields

$$V_{CS} = \frac{\left(s \times \frac{L_{BUCK}}{R_{DCR}} + 1\right)}{(s \times C_{CS} \times R_{CS} + 1)} \times R_{DCR} \times \left(IOUT + \frac{\Delta IL}{2}\right) \qquad (8)$$

When the time constants match, you get

$$R_{CS} \times C_{CS} = \frac{L_{BUCK}}{R_{DCR}} \qquad (9)$$

The frequency-dependent term $s \times \frac{L_{BUCK}}{R_{DCR}}$ in the numerator of (8) is exactly canceled by the filter $R_{CS} \times C_{CS}$ term in the denominator, leaving

$$V_{CS} = R_{DCR} \times \left(IOUT + \frac{\Delta IL}{2}\right) \qquad (10)$$

Thus, with proper matching of time constants, the voltage across C_{CS} accurately reproduces the inductor's current waveform scaled by R_{DCR}.

It is possible to separate equation 10 into a dc component and an ac component. The dc component is the average value of the sensed voltage waveform:

$$V_{CSave} = R_{DCR} \times IOUT \qquad (11)$$

The ac component is the peak-to-peak variation of the sensed voltage:

$$V_{CSac} = R_{DCR} \times \frac{\Delta IL}{2} \qquad (12)$$

Fig. 2 is a graphical representation of both the dc and ac components of the current-sensed voltage.

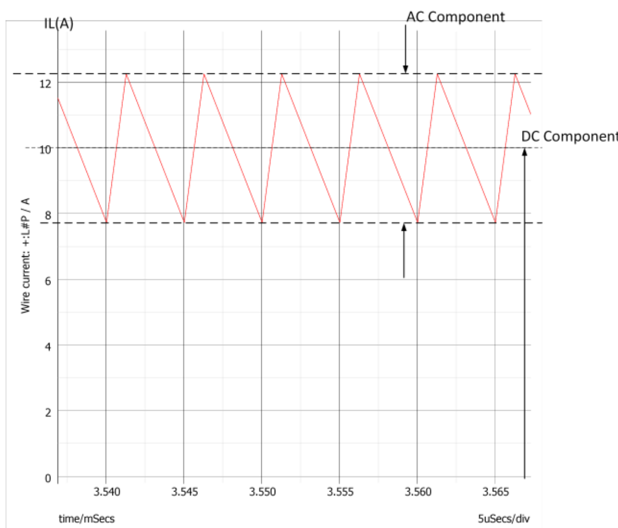

Fig. 2. Dc (average) and ac (peak-to-peak) components of the inductor current corresponding to the sensed voltage signal across C_{CS}.

You can adjust the ac component by making the $R_{CS} \times C_{CS}$ τ_{RC} different from the $\frac{L_{BUCK}}{R_{DCR}}$ τ_{RL} per the following equations:

If $R_{CS} \times C_{CS} > \frac{L_{BUCK}}{R_{DCR}}$,

$$V_{CSac} < R_{DCR} \times \frac{\Delta IL}{2} \quad (13)$$

If $R_{CS} \times C_{CS} < \frac{L_{BUCK}}{R_{DCR}}$,

$$V_{CSac} > R_{DCR} \times \frac{\Delta IL}{2} \quad (14)$$

For example, consider a 10-V input, 5-V output buck converter switching at 500-kHz. The τ_{RC} matches the τ_{RL} time constant, as shown in Fig. 3.

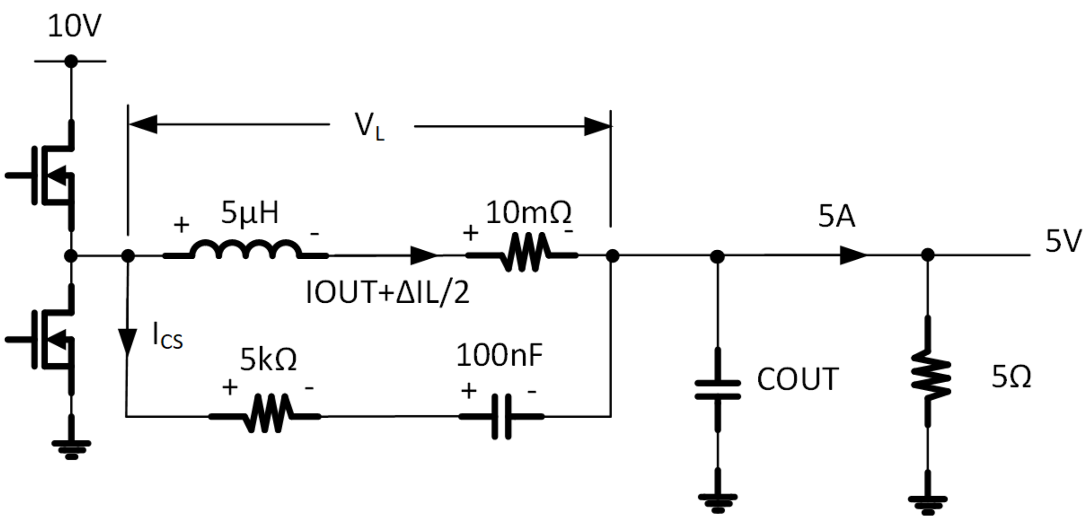

Fig. 3. Example buck converter showing a DCR current-sensing implementation. A 10-V to 5-V, 5-A buck converter operating at 500 kHz with L_{BUCK} = 5 µH, R_{DCR} = 10 mΩ, and RC network values of R_{CS} = 5kΩ and C_{CS} = 100 nF match the $\frac{L_{BUCK}}{R_{DCR}}$ time constant.

Using equation 11, the dc component is $V_{CSac} = 50\ \text{mV}$.

The time constants match:

$$5\ \text{k}\Omega \times 100\ \text{nF} = 0.5\ \text{ms and } 5\ \mu\text{H}/10\ \text{m}\Omega = 0.5\ \text{ms}.$$

The inductor ripple current is

$$\Delta IL = \frac{VIN - VOUT}{L_{BUCK}} \times \frac{\frac{VOUT}{VIN}}{Fsw} = \frac{10\ V - 5\ V}{10\ \mu H} \times \frac{\frac{5\ V}{10\ V}}{500\ kHz} = 1\ A\ \text{pk to pk}$$

Therefore, using equation 12,

$$V_{CSac} = 10\ m\Omega \times \frac{1\ A}{2} = 5\ mV$$

Adding the dc component from equation 11 (50 mV), the peak sensed voltage is

$$V_{CS} = 50\ mV + 5\ mV = 55\ mV$$

Example With Modified R_{CS}

If you increase R_{CS} to 10 kΩ, the τ_{RC} becomes

$$10\ k\Omega \times 100\ nF = 1\ ms$$

The inductor τ_{RL} remains

$$\frac{5\ \mu H}{10\ m\Omega} = 0.5\ ms$$

From equation 10, you know that the ac component is still 5 mV. The voltage across R_{DCR} remains unchanged. But because τ_{RC} is now double, equation 8 shows that the denominator term increases, reducing the multiplication factor (which was 1 in the matched example) to 0.5 in this unmatched example. The increased denominator in effect halves the ac component to 2.5 mV, making the new peak sensed voltage

$$V_{CS} = R_{DCR} \times \left(IOUT + \frac{\Delta IL}{2} \times 0.5\right) = 50\ mV + 2.5\ \ mV = 52.5\ mV$$

V_{CSave} Attenuation Technique

Adding a resistor-divider to the sensed network (as shown in Fig. 4) reduces the average sense voltage (V_{CSave}).

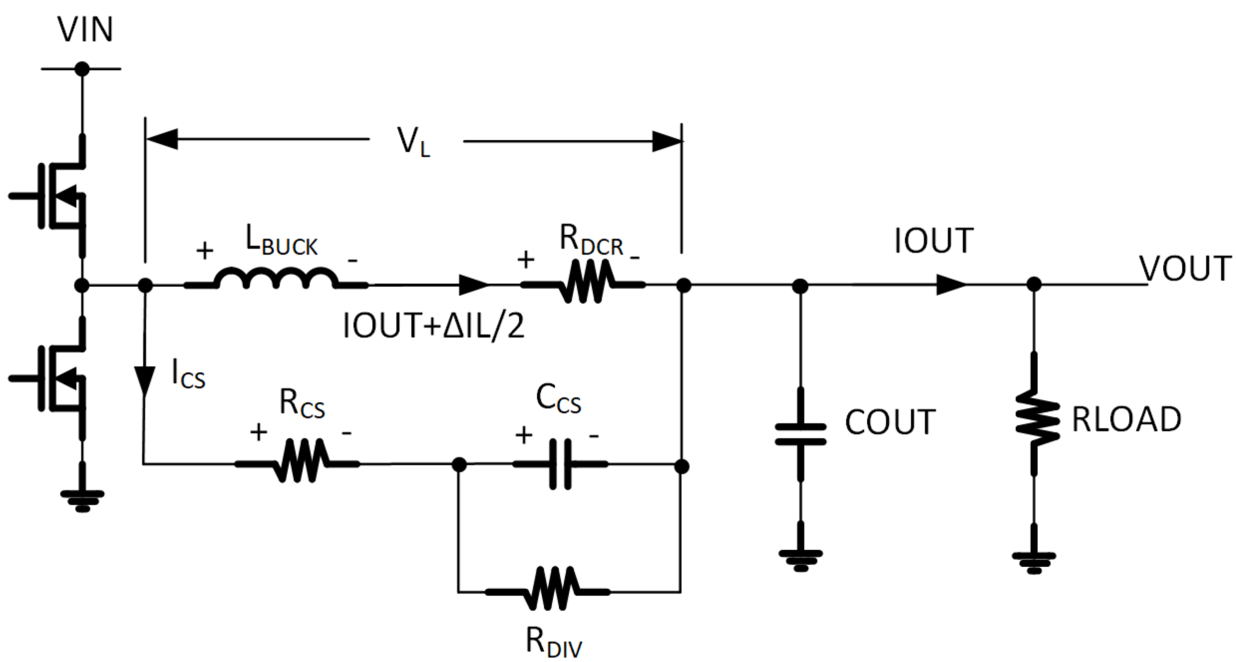

Fig. 4. DCR current sensing with a resistor-divider to attenuate the average sensed current.

The resistor R_{DIV} forms a divider with R_{CS} to scale down the dc level of the sensed current signal. Adding R_{DIV} has three effects:

- Dc attenuation. The average sensed voltage will become lower according to the resistor-divider ratio.
- Ac attenuation. The ac component of the sensed voltage will also become lower when using the same resistor-divider ratio, decreasing the amplitude of the ripple voltage.
- τ_{RC} change. The effective resistor of the RC filter becomes the parallel combination of R_{CS} and R_{DIV}, which changes the τ_{RC} and therefore the ac gain of the sensing network.

Equation 15 expresses this behavior:

$$V_{CS} = \frac{\left(s\times\frac{L_{BUCK}}{R_{DCR}}+1\right)}{\left(s\times C_{CS}\times\frac{R_{CS}\times R_{DIV}}{R_{CS}+R_{DIV}}+1\right)}\times R_{DCR}\times\left(IOUT+\frac{\Delta IL}{2}\right) \qquad (15)$$

A Crafty Approach To DCR Current Sensing

Current-mode controllers that use DCR current sensing typically incorporate internal slope compensation. Stable performance requires strict adherence to the specified inductor value and R_{DCR}. Failure to meet these specifications can result in instability, either from too much slope compensation (overdamped) or too little (underdamped).

A fixed internal slope compensation can limit your inductor choice. By adjusting both the dc and ac components of the sensed signal, however, DCR current sensing allows you to compensate for nonideal inductors and their R_{DCR} values and to use inductors optimized for size, cost or efficiency—without sacrificing loop stability.

Application Example

TI's LM5148[2] is a peak-current-mode controller with internal slope compensation. When selecting the inductor and current-sense resistor, you must ensure that the values set the correct current limit, and that the internal slope compensation is not too high or too low.

Too small an inductor results in a large ripple current vs. the slope compensation, creating possible subharmonic oscillation at duty cycles ≥50%. Too large an inductor results in a small ripple current vs. the slope compensation, creating a risk of loop instability from excessive slope compensation.

The example conditions are:

- V_{IN} = 48 V
- V_{OUT} = 12 V
- Maximum continuous I_{OUT} = 10 A
- Switch frequency (F_{sw}) = 200k Hz
- Current-limit target = 12 A peak
- Recommended L_{BUCK} = 10 µH
- Recommended external R_s = 4 mΩ

The peak current is then

$$I_{PEAK} = IOUT + \frac{\frac{VIN - VOUT}{L_{BUCK}}\times\frac{\frac{VOUT}{VIN}}{Fsw}}{2} = 10\ A + \frac{\frac{48\ V - 12\ V}{10\ \mu H}\times\frac{\frac{12\ V}{48\ V}}{200\ kHz}}{2} = 12.25\ A\ pk$$

Checking Slope Compensation

To optimize the slope compensation, verify that the downslope is equal to the slope compensating voltage (V_{SLOPE}) gained up by the current-sense amplifier, resulting in

$$V_{DOWNSLOPE} = \frac{VOUT}{L_{BUCK}} \times \frac{1}{Fsw} = \frac{12\ V}{10\ \mu H} \times \frac{1}{200\ kHz} = 6\ A$$

The LM5148 data sheet electrical table lists the V_{SLOPE} at 200 kHz as 50 mV/µs. The time period is 5 µs, so

$$V_{SLOPE} = 5\ \mu s \times \frac{50\ mV}{\mu s} = 300\ mV$$

The downslope at the modulator stage is then

$$V_{DOWNSLOPEactual} = \frac{VOUT}{L_{BUCK}} \times \frac{1}{Fsw} \times R_S \times A_{CS} = \frac{12\ V}{10\ \mu H} \times \frac{1}{200\ kHz} \times 4\ m\Omega \times 10 = 240\ mV$$

Since this is close to the target V_{SLOPE}, the loop will be stable without subharmonic oscillations over the operating range.

Tuning For Design Constraints

What if you selected a different inductor and removed the external sense resistor, choosing to employ DCR current sensing? If $L_{BUCK} = 2.2\ \mu H$ and $R_{DCR} = 8\ m\Omega$, assuming that τ_{RC} matches τ_{RL}, the downslope at the modulator would be excessive, both because of the higher inductor ripple current from the smaller inductance value and from the doubled R_{DCR}, which also increases the dc sense voltage for a given current.

These effects cause both the dc and ac components of the sensed signal to be too large, resulting in an early current-limit engagement. In addition, the ripple voltage presented back to the modulator would be much larger compared to the slope compensating ramp, potentially resulting in loop instability because the loop is underdamped.

Reducing the dc level by a factor of two using R_{DIV} allows the current limit to engage at the intended load current to which the RC network is tuned, thus producing the correct downslope for stability. Additionally, increasing the ac levels will counter the effects of the attenuating signal caused by the addition of R_{DIV}.

Consider a scenario with matched time constants. A C_{CS} of 100 nF and the new inductor would result in an R_{CS} value of

$$Rcs = \frac{L_{BUCK}}{R_{DCR} \times C_{CS}} = \frac{2.2\ \mu H}{8m\Omega \times 100\ nF} = 2.75\ k\Omega$$

Since R_{CS} and R_{DIV} are effectively in parallel, for τ_{RC} calculations you must select their values so that the equivalent resistance is 2.75 kΩ, which requires both R_{CS} and R_{DIV} to be 5.5 kΩ.

With the dc level attenuated and the time constants matching, you must now adjust the ac levels to match the required ripple current in order to realize adequate slope compensation.

The ripple current-sense voltage for the nominal inductor (10 µH, 4 mΩ) is

$$\Delta IL \times R_{DCR} = \frac{VIN - VOUT}{L_{BUCK}} \times \frac{\frac{VOUT}{VIN}}{Fsw} \times R_{DCR} = \frac{48\ V - 12\ V}{10\ \mu H} \times \frac{\frac{12\ V}{48\ V}}{200\ kHz} \times 4\ m\Omega = 18\ mV\ pk\ to\ pk$$

With the new inductor (2.2 µH, 8 mΩ), the ripple current-sense voltage becomes

$$\Delta IL \times R_{DCR} = \frac{VIN - VOUT}{L_{BUCK}} \times \frac{\frac{VOUT}{VIN}}{Fsw} \times R_{DCR} = \frac{48\ V - 12\ V}{2.2\ \mu H} \times \frac{\frac{12\ V}{48\ V}}{200\ kHz} \times 8\ m\Omega = 164\ mV\ pk\ to\ pk$$

The new ripple voltage sensed is almost 10 times larger than the target, so you must attenuate the ac component using time-constant mismatching, increasing the τ_{RC} by a factor of 10 to slow the sensed voltage waveform and reduce the sensed ac ripple voltage.

The adjusted R_{CS} is

$$R_{CS} = \frac{L_{BUCK}}{R_{DCR} \times C_{CS}} \times 10 = \frac{2.2\ \mu H}{8\ m\Omega \times 100\ nF} \times 10 = 27.5\ k\Omega$$

Impact Of New Inductor On Stability

To find your starting point R_{COMP} and C_{COMP} values use the LM5148 calculator spreadsheet. These values are partly based on the recommended buck inductor and current sense resistor values. If you switch to a different inductor with a different DCR you will get a modified ripple current and thus a modified peak current sensed across its DCR. By applying the mismatched τ_{RC} you sense a different voltage level.

In this example, we would measure ~160 mV across the R_{DC} and we are scaling that down by the τ_{RC} to ~50 mV, let's say a factor of 4 reduction. In essence the ratio of which we reduced the sensed voltage (peak current across DCR divided by the scaled voltage across C_{CS}) is the same ratio by which you adjust your RCOMP.

So, you start with the recommended RCOMP and scale it down by that factor. R_{COMP} in this example is reduced with compensator capacitors C_{COMP} scaled inversely to preserve zero/pole locations. It's a straightforward rule of thumb to ensure stable operation when you deviate from the original inductor values.

Simulation Results

The simulation sets both R_{CS} and R_{DIV} at 27.5 kΩ, which preserves the correct dc scaling while ensuring that the parallel combination is large enough to achieve the 10 times ac attenuation number.

The simulation of piecewise linear systems (SIMPLIS) plotted results (Fig. 5) show:
- The theoretical current sensed across R_{DCR} for the recommended inductor.
- The sensed voltage using the tuned R_{CS}, R_{DIV} and C_{CS} values with a modified inductor parameter.
- Current though the inductor for both the modified and recommended inductors.

As shown, the tuned network matches the sensed waveforms closely to the ideal case, achieving both the correct dc levels and the reduced ac amplitudes required for proper slope compensation.

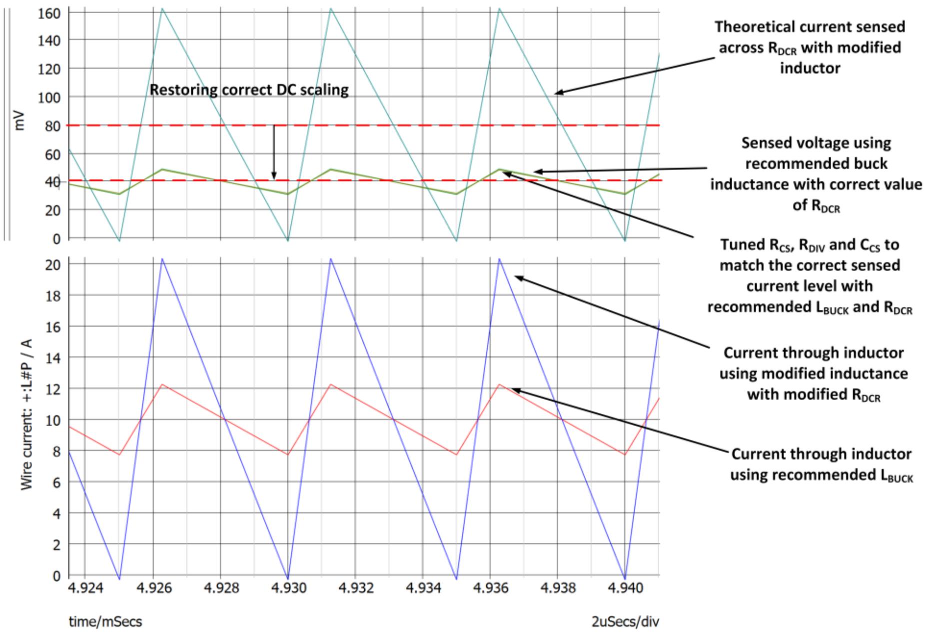

Fig. 5. Simulated comparison of theoretical sensed current waveforms for recommended inductor and modified inductor values, showing tuned R_{CS}, R_{DIV} and C_{CS} restoring correct dc scaling and ac content.

Conclusion

DCR current sensing with TI peak-current-mode buck regulators offers an efficient, low-cost alternative to discrete current-sense resistors while maintaining precise control and protection. By carefully tuning the R_{CS}-C_{CS} network, adjusting both dc scaling and ac ripple content, you can compensate for nonideal inductor parameters such as low inductance values and a higher DCR without compromising stability.

Modern buck controllers use internal slope compensation that can limit choices in component selection to meet a given specification. The method presented in this article reopens the door to more flexible component choices. This flexibility allows for optimization of efficiency, size, cost or thermal performance. At the same time, DCR current sensing ensures reliable operation across the full load range, making it a practical and adaptable solution for a variety of power designs.

Reference

1. "Inductor DCR Current Sensing with Temperature Compensation: An Accurate, Lossless Approach for POL Regulators," Tim Hegarty, How2Power, November, 2013.
2. LM5148 3.5-V to 80-V, current mode synchronous buck controller product page.

About The Author

David Baba serves as a senior applications engineering manager at Texas Instruments where he has over 28 years of experience in dc-dc power-supply design. He holds a bachelor of engineering (BEng (hons)) degree from the University of Surrey in the United Kingdom.

For more on current sensing, see How2Power's Design Guide, locate the "Design Area" category and select "Test and Measurement". Also keyword search "current sense" or "current sensing".

