# Applications Information

October 24, 2003

## *Component Selection for the Boost Converter Used in the A3935*

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This application note provides information to assist in the selection of components for designs using the Allegro MicroSystems A3935 three-phase MOSFET controller. In the first section, the general operation of the boost converter is described. A typical configuration is shown in figure 1. Then individual component selection considerations are discussed. A detailed overview of the selection of each component is given along with technical considerations. Finally, a worked example provides a clearer understanding of the selection of actual component values.

### The Modulated (PFM) Boost Converter

The output voltage of the boost converter is regulated by two control loops: the *inner loop*, which controls the current through the switch by hysteretic switching action, and the *outer loop*, which controls the ac ripple voltage on the output by means of hysteretic enabling/disabling of the current loop. During the switching action period, excess charge is dumped into the output capacitor, $C_{OUT}$. Consequently, the output voltage tends to rise until the hysteretic voltage level is reached (an example is shown in figure 4). At this point, the switching period terminates and the load is supplied only by $C_{OUT}$. Once the voltage drops by a predetermined value, the boost switching action occurs again.

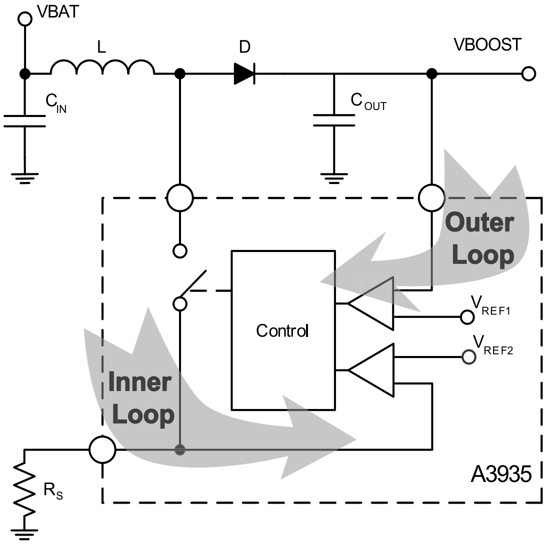

Figure 1. Boost regulator control loops

The key to successful operation is to ensure that, when the boost current is being controlled through the inner loop, enough charge (defined as electrical current multiplied by elapsed time) is transferred to both the load and $C_{OUT}$, so that when the outer loop determines the correct output voltage has been reached, $C_{OUT}$ can maintain the load current during the switcher dead time. (In this context, the term *dead time* refers to the internal process of the A3935. It is not to be confused with the more general use of the term when referring to the MOSFET drive, where dead time is a programmed interval required to prevent current shoot-through.)

### Sense Resistors

The sense resistor, $R_S$, is chosen to give a desired current trip point for the boost converter to terminate during current control switching. In most applications, the value of $R_S$ should be as low as possible, ensuring that the peak current limit of 300 mA is not exceeded. This approach ensures that the smallest inductance value can be selected. A ⅛W resistance value is more than adequate. The size of $R_S$ is also related to the amount of charge that can be transferred, as described in the following section on inductor selection.

### Inductors

Figure 2 illustrates the current through the inductor. At the beginning of the hysteretic current switching mode, the boost switch turns on and the current rises in the inductor until it reaches the programmed level, $I_{L(peak)}$, set by the sense resistor ($0.5V/R_S$). The switch on-time ($t_{on}$) is determined by the inductor value and the current trip point. The boost switch then turns off for a fixed period ($t_{off}$). The amplitude of the ac current ripple is determined by $t_{off}$ and the inductor value.

It is necessary to ensure that enough charge is passed during the hysteretic current switching period to supply the load continuously under worst case conditions (maximum load at minimum input volts), and to overcome power losses in the circuit. The reason that the worst case conditions occur at minimum line voltage is that the current ripple is at a maximum, giving a lower average inductor current. This is because the voltage across the inductor during the off-time is at a maximum.

Selecting the correct inductance value requires striking a balance between a value that is low enough to minimize size and cost, and one that is high enough to ensure that sufficient charge is transferred to $C_{OUT}$ and to the load under worst-case conditions (minimum input volts and maximum load). When selecting the inductor (L), as well as the inductance, parameters of importance are: maximum current rating, series resistance, and operating temperature.

The current rating chosen should be greater by some margin than the value set by $R_S$. However, keep in mind that the maximum peak current of the A3935 is 300 mA.

Manufacturer data sheets for inductors usually provide two current-level specifications. Of the following, use the parameter with the lower current value to determine the maximum allowable current level:

- Saturation level, where the inductance value typically drops by 10%. In most cases, this is the limiting current value.
- Temperature rise, where the inductor experiences a rise in temperature, in the order of 20°C to 50°C, when current reaches its full rated level.

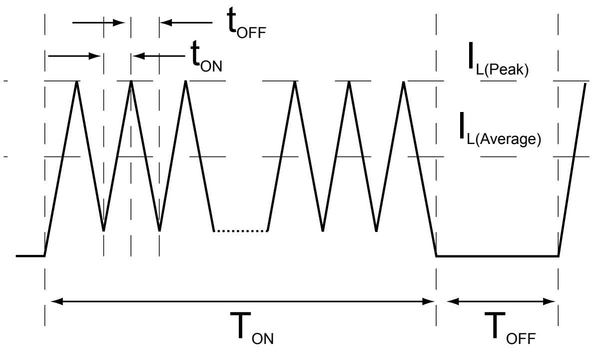

Figure 2. Current ripple through inductor L. $T_{ON}$ is switcher on-time, $T_{OFF}$ is switcher dead time.

When interpreting values for temperature, determine the method used by the manufacturer to define the maximum operating temperature. In many cases, manufacturers include this self-induced temperature rise, so the safe ambient temperature is correspondingly lower.

The following inductor series are recommended:

| Manufacturer | Series | Max. Ambient Temperature (ºC) | Package |
|---|---|---|---|
| TDK | SLF7045T | 105 | Surface Mount |
| Coilcraft | MSS7431H | 135 | Surface Mount |
| Coilcraft | DR0810 | 135 | Through-Hole |

## Capacitors

In this section, general considerations for capacitor selection are discussed. In later sections, the criteria for the output capacitor $C_{OUT}$ and the input capacitor $C_{IN}$ are examined.

**Capacitor Technologies**. The following considerations affect the choice of the type of capacitor:

- From the point of view of cost, performance, and acceptability, standard aluminum electrolytics are probably the optimal technology.
- Polymer aluminum electrolytics provide excellent performance in terms of esr, temperature stability, and size. However, they tend to be relatively expensive.
- Because of the capacitor output considerations, the use of ceramic capacitors may be restricted in applications with high loads. This is due to the cost penalty of a relatively large value capacitor. Note that the X7R, X5R, and X5S or better are the only dielectrics worth considering, especially in high-temperature applications.
- It is recommended to avoid the use of tantalums due to reliability issues.

**Temperature**. Electrolytics come in various temperature ratings: 85°C, 105°C, 125°C, and 150°C. Obviously, the maximum application ambient will influence the temperature chosen, but lifetime is also a consideration. It is worth noting that the lifetime doubles for every 10°C that a capacitor is derated with respect to its maximum rated temperature. For instance, a capacitor that is rated at 125°C and has a predicted life of 5,000 hours can have an expected life of 20,000 hours when operated in an ambient temperature of 105°C.

The following capacitor series are recommended:

| Manufacturer | Series | Max. Ambient Temperature (ºC) | Package |
|---|---|---|---|
| Nichicon | VR | 85 | Through-Hole |
| United Chemi-Con | SME | 85 | Through-Hole |
| Nichicon | BT | 125 | Through-Hole |
| United Chemi-Con | GXL | 125 | Through-Hole |
| Nichicon | UH/UB | 125 | Surface Mount |
| United Chemi-Con | MVH | 125 | Surface Mount |
| Nichicon | BX | 150 | Through-Hole |

## Output Capacitors ($C_{OUT}$)

Excess ouput voltage ripple can be caused by too much stored charge in the inductor L being dumped into the output capacitor $C_{OUT}$ on the final switching cycle. To prevent this, it is advisable to avoid a mismatch between L and $C_{OUT}$. It is recommended that the capacitor-to-inductor ratio, L µH / C µF, should be $<5$.

The working voltage of $C_{OUT}$ should be at least 25 V, although in systems where load dumps can occur, the ratings of both $C_{IN}$ and $C_{OUT}$ should be rated for at least 50 V.

As the converter operates in switcher mode, excess charge is transferred into $C_{OUT}$ until the output voltage, $V_{BOOST}$, rises to the upper hysteretic threshold. As charge is dumped into $C_{OUT}$, there is a natural increase in the output voltage, determined by the capacitance itself:

$$V_{BOOST} = I \times (t / C).$$

In addition, a further increase in output voltage is caused by the current passing through the electrolytic esr. Both of these effects influence when the switcher on-time terminates (see figure 2).

The output voltage ripple is essentially a low frequency sawtooth waveform with switching ripple superimposed on it. The size of the switching ripple is largely dictated by the esr of the capacitor and the load current/ charge current. See figure 4, which illustrates the switching ripple in the bottom trace (annotated as esr ripple), as well as the low frequency sawtooth (caused by switcher on/ switcher dead time).

The maximum current passing through $C_{OUT}$ should be well-derated with respect to the maximum current rating of $C_{OUT}$. Throughout $T_{ON}$ (switcher on-time; see figure 2), the current carried by $C_{OUT}$ fluctuates: during each $t_{ON}$ interval, when the switch is closed, $C_{OUT}$ carries the full output current, and during each $t_{OFF}$ interval, when the switch is open, $C_{OUT}$ carries charging current. Also, during $T_{OFF}$ (switcher dead time), $C_{OUT}$ carries the full output current.

**Current Ripple Rating**. When selecting a suitable current ripple rating, a point worth noting is that ripple can be defined at both 120 Hz and 100 kHz ratings. For a given capacitor, the two values can be quite different, because ripple current is determined by the esr, and esr varies with frequency. Manufacturers usually provide tables of current ripple factor values to allow appropriate selection for different operating frequencies.

Noteworthy also is that the capacitor experiences two different frequencies: during the hysteretic current period (the higher-frequency voltage ripple caused by the switcher), and during the hysteretic voltage period (the lower-frequency switching ripple caused when the switcher toggles between switching mode and dead time). To make the appropriate selection, it is necessary to determine the current for both cases, at the corresponding frequencies.

**Esr**. Esr is directly related to ripple rating. However, the specifications for less expensive capacitors often omit the value for either the esr or the impedance. This is probably because of an element of specmanship between manufacturers. When only one is given, it is usually the esr. As mentioned before, the esr has a bearing on the ripple voltage period. Because of the specification problem mentioned above, for lower-cost capacitors, predicting the operating frequency may be difficult.

Esr varies considerably with both temperature and frequency. The esr has a negative temperature coefficient, and typically diminishes by approximately 50% when the temperature rises from 25°C to 105°C. Therefore, at higher temperatures the esr tends to reduce while the hysteretic voltage period tends to increase, resulting in more switching cycles. Esr diminishes rapidly when the frequency changes in the range 100 Hz to 10 kHz, but the rate of change tends to flatten for frequency changes in the range 10 kHz to 100 kHz.

In the majority of applications, frequencies of above 10 kHz should be realizable (at least in terms of switching frequency), so esr variation as a function of frequency should be minimal. Esr also tends to be quoted as a maximum, with the actual value being considerably less.

In the interests of reducing costs, having a reasonably large esr is not necessarily an issue. It means that the number of switching cycles could be reduced, as the hysteretic voltage is reached earlier. This effects the voltage ripple frequency and may have a bearing on EMI filtering. Note that the capacitor has no effect on the actual boost switching frequency, which instead is largely determined by the value of: the inductor, the fixed off-time duration, and the sense resistor $R_S$.

## Input Capacitors ($C_{IN}$)

The process of selecting the input capacitor, $C_{IN}$, is similar to that for the output capacitor. The main purpose of $C_{IN}$ is to provide a low-impedance path for the inductor ripple current, in order to minimize the so-called residue ac ripple current that appears on the VBAT supply line from the source voltage. In effect, the residue current is conducted EMI. This means that $C_{IN}$ should have a relatively low impedance with respect to the source impedance (including resistance of PCB traces, connectors, etc.).

Typically, where a low-impedance source is used, $C_{IN}$ should carry the ripple current portion of the inductor. Where other types of sources are used, $C_{IN}$ may supply more of the dc component of the inductor current as well. In general, the lower the source impedance, the less current the capacitor has to carry. In extremely noise-sensitive applications, it may be necessary to increase the high-frequency impedance of the source supply by incorporating a differential-mode inductor in series with VBAT.

The operating voltage should be at least 50 V. However, a lower value could be used in applications where the supply voltage is guaranteed to be less.

## Diodes

From a cost perspective, a 1N4148 high-speed signal diode is optimal. As the device operates at high speed, the body capacitance is relatively low. Hence, the noise produced when the current runs discontinuously is minimal. The higher voltage drop is tolerable given the magnitude of the output voltage.

## R-C Snubbers

At the end of the switching period, when the current through the boost diode drops to zero, the parasitic capacitances in the body diode, boost switch ($C_{oss}$), and so forth, resonate with the boost inductor. This resonance can be a major source of E-field emissions. This type of noise, despite being low energy, can be a major source of E-field radiation and in many instances may fail EMI emission standards. To remove this ringing effect, a lossless R-C damping filter can be applied between the anode of the diode and ground.

Additional selection considerations rely on results derived during the component selection example. These are discussed in the section Choosing R-C Snubber Values, after the worked example below.

## Component Selection Example

This section provides a sample set of calculations for determining values for components to be used in the boost section of the A3935.

### Assumptions

- VBAT ≥ 7 V (worst case)
- VBOOST output = 15.6 V
- Boost supply load = 40 mA
- Off time = 5 µs (typical)
- Maximum ambient temperature = 70°C
- Minimized component costs

### Calculations

1. Determine average inductor current, $I_{L(average)}$.

The first step is to determine the switching duty cycle, *D*. It can be found using the boost transfer function:

$$\frac{V_{BOOST}}{V_{BAT}} = \frac{1}{1 - D}$$

Because of the 1V drop in the 1N4148, the actual duty cycle compensates for this drop, so the output voltage appears to the control circuit to be 16.6 V.

$$D = 1 - \frac{V_{BAT}}{V_{BOOST}} = 1 - \frac{7}{16.6} = 0.58$$

Note that voltage drops in the switch and inductor are ignored.

The corresponding input current can now be found, again through the boost transfer function:

$$I_{IN} = I_{OUT} \times \frac{1}{1-D} = 40\text{ mA} \times \frac{1}{1-0.58} = 95\text{ mA}$$

The above input current assumes 100% efficiency. To establish the actual current, the typical efficiency of the converter can be taken from the graph shown in figure 3 for loads between 10 mA and 40 mA.

At VBAT = 7 V, the efficiency is approximately 84%. The input current then is 95 mA / 0.84 = 113 mA.

Now if the switching period was continuous, and there was no switcher dead time, the average inductor current would also be 113 mA.

A hysteretic duty cycle for the output voltage has to be introduced to ensure control is maintained. Introduce a duty cycle of say 0.7, to allow for tolerance effects: drop in inductance value, maximum fixed off-time, etc.

This means the average inductor current (during the switching period) is now 113 mA / 0.7 = 161 mA.

Choose a sense resistor ($R_S$) with a value of 1.8 Ω. This sets the maximum peak current to 0.5 V / 1.8 Ω = 278 mA.

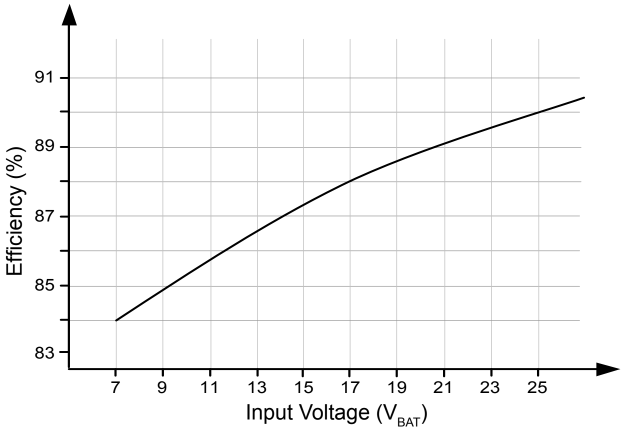

Figure 3. Converter efficiency. Assumes loads between 10 mA and 40 mA.

With the average current and the peak current calculated, the peak-to-peak current can be determined:

$$2 \times (278\text{ mA} - 161\text{ mA}) = 234\text{ mA}.$$

2. Select the inductor.

With a fixed off-time (from the data sheet), ripple current, and voltage across the inductor (difference between the output and input voltage) known, then:

$$L = \frac{\Delta V \times t_{off}}{\Delta I} = \frac{(16.5\text{ V} - 7\text{ V}) \times 5\ \mu\text{s}}{234\text{ mA}} = 203\ \mu\text{H}$$

Choose L= 220µH. In this example, a TDK SLF7045 part was chosen. The saturation current level is quoted as 400 mA (based on a 10% drop in inductance).

The current level for a 20°C temperature rise is quoted as 430 mA. Maximum peak current in this application is 278 mA, so there is plenty of margin.

The switching frequency can be determined either by using the duty cycle and the off-time or working with the current slope. Using the current slope, note that during switch on-time, VBAT appears across the inductor. Therefore:

$$t_{on} = \frac{L \times \Delta I}{\Delta V} = \frac{220\ \mu\text{H} \times 234\text{ mA}}{7} = 7.4\ \mu\text{s}$$

The overall period = 5 µs + 7.4 µs = 12.4 µs.

3. Choose the output capacitor value.

To prevent the output voltage rising too much due to excess charge stored in the inductor on the final switching cycle, the capacitor should not be greater than L / 5. With a 220 µH inductor chosen, the capacitor should be greater than 44 µF. For this example a 100 µF is chosen, however, a lower value could be selected.

Note that choosing a larger capacitor and/or a smaller esr results in a lower hysteretic voltage frequency. The voltage hysteresis trip point is determined by both the charge into the capacitor and the voltage drop caused by the esr. Therefore, a larger capacitor requires more charge to produce the same voltage and a lower esr produces a smaller voltage drop. The waveforms shown in figure 4 demonstrate this effect. It is worth noting that with a variation of esr due to tolerance and temperature effects, it is very difficult to predict the output voltage ripple frequency. Although the output voltage ripple may vary, the ripple voltage duty cycle remains constant. This is because the variation in esr affects both the switcher period and the switcher dead time.

In this example, a low-cost Nichicon VR series capacitor is selected, with a working voltage of 25 V. Relative to most other electrolytic ranges, this has one of the lowest ripple current ratings and low cost. The ripple current rating is 190 mA rms at 85°C, 120 Hz. From the waveforms shown in figures 4 and 5, it can be seen that the capacitor has to cope with two frequency components. There is the low-frequency component of the output voltage ripple at 1750 Hz, and the switching-frequency component at 1 / 12.4 µs = 80 kHz.

From the current ripple factor table, the current rating at 1 kHz = 1.34 × 190 mA, and the current rating at 10 kHz (the maximum frequency quoted) is 1.5 × 190 mA. The capacitor carries full output current during the switcher dead time (at 1750 Hz). During the switcher period, the capacitor alternatively carries the charging current (approximately 120 mA) and the output current (40 mA). Without working out the actual rms currents, clearly there is ample margin in the capacitor selected.

4. Choose the input capacitor value.

To minimize conducted EMC emissions, ensure the impedance of the input capacitor is lower than the VBAT source impedance so that the ripple current is supplied by the capacitor. This may necessitate the use of a low-impedance capacitor.

From the point of view of ripple current rating, the capacitor only has to supply the ripple current set by the inductor. In this case, the ac current is 234 mA, therefore, the rms current, $I_{rms} = (234 \text{ mA}) / 3^{1/2} = 135$ mA.

It also is possible to use the Nichicon VR series capacitor rated at 100 µF. For 12 V battery systems, use a 25 V working voltage.

Note that if load dumps can occur in the application, both input and output capacitors should be rated accordingly.

## Results

In figure 4 it can be seen that the average current through the inductor is approximately 160 mA and the output voltage duty cycle is roughly 0.7, which ties up well with the calculations.

The bottom trace shows the ripple caused by the switching currents through the esr of the output capacitor superimposed on the voltage ramping up on the output capacitor due to the additional charge pumped into the capacitor on each switching cycle. The combination of the esr ripple and the charge being dumped into the output capacitor causes the hysteretic

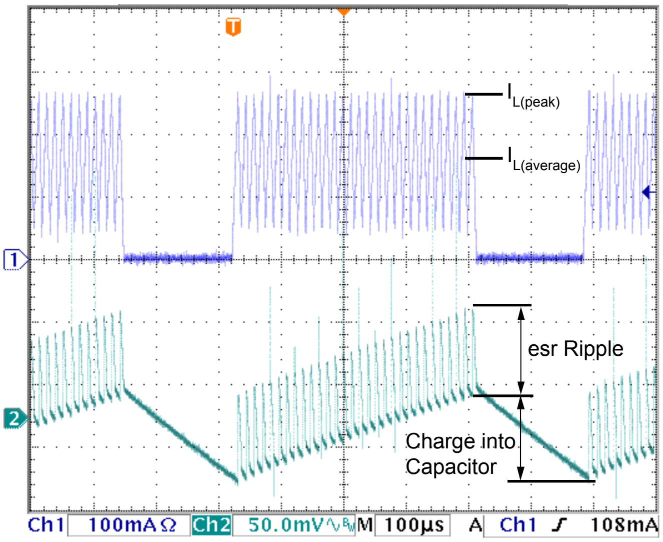

Figure 4. Current through the inductor (L). The lower trace illustrates the output voltage ripple.

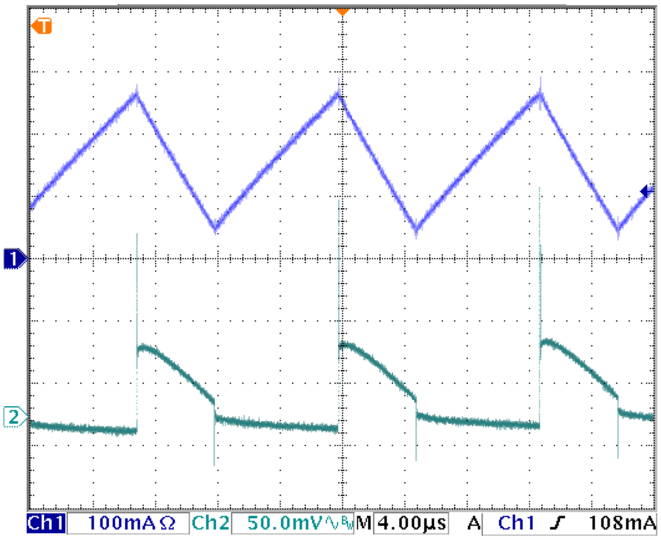

Figure 5. Current through the inductor (L). Expanded view.

switching action to terminate and reinitiate again. The hysteresis on the output voltage is approximately 120 mV, which aligns with the typical figure in the specification.

Figure 5 is an expanded view of the same waveforms as in figure 4. During the switch on-time, energy is stored in the inductor. At the same time, the output capacitor supplies the 40 mA load current. When the switch turns off, some of the energy stored in the inductor is transferred to both the load and the output capacitor. Because the esr dominates the output impedance, it can be seen that the output voltage follows the shape of the current waveform during the charging phase. After each charging sequence, it can be seen that the output volts increase slightly.

## Choosing R-C Snubber Values

As mentioned previously, at the end of the switching period, when the current through the boost diode drops to zero, the parasitic capacitance, dominated by the body diode capacitance and the output capacitance ($C_{oss}$) of the boost switch, resonates with the boost inductor. In this particular example, the waveform in figure 6 was recorded at worst case conditions, VBAT = 7 V, and $I_{load}$ = 40 mA.

In figure 6, the lower trace shows the current through the inductor-diode loop decaying to 0 A. As the current reaches zero, the voltage across the switch resonates. This is shown in the upper trace.

Effectively, a tuned circuit exists between the boost inductor and the parasitic capacitance which causes this voltage resonance. The simplest way of suppressing this noise is to add a lossless R-C damping filter. The resistor effectively adds in extra damping resistance at a particular frequency determined by the filter capacitor, without adding much in the way of power loss.

It is necessary to determine the characteristic impedance of the L-C circuit. From the equation describing the cut-off frequency of an L-C filter, the parasitic capacitance can be found:

$$f_1 = \frac{1}{2\pi\sqrt{L \times C}}$$

Therefore, from figure 6:

$$f_1 = 1.4\ \text{MHz} = \frac{1}{2\pi\sqrt{220\mu\text{H} \times C}}$$

where C = 58 pF.

Introducing a snubber capacitor of 150 pF provides a reasonable cut-off point. This value, along with the parasitic value of 58 pF, introduces a resonance point of 744 kHz.

The characteristic resistance, $R_0$ , is then:

$$R_0 = \sqrt{\frac{L}{C}} = \sqrt{\frac{220\ \mu\text{H}}{208\ \text{pF}}} = 1\ \text{k}\Omega$$

where the R-C snubber values are:

R = 1 kΩ, ⅛ W

C = 150 pF, 50 V (X7R ceramic)

The waveforms that result after addition of the snubber are shown in figure 7. Clearly the resonance has been reduced considerably. Although it is possible to provide more optimal damping, there is always a trade-off with power loss. In this particular example, only a few additional milliwatts were noted.

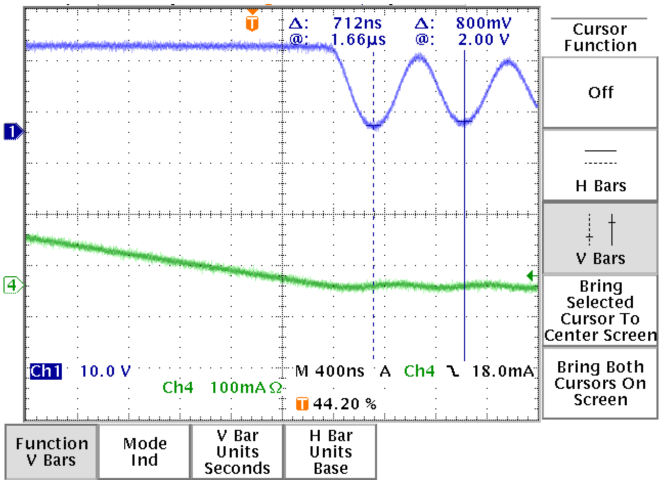

Figure 6. Current at worst case conditions

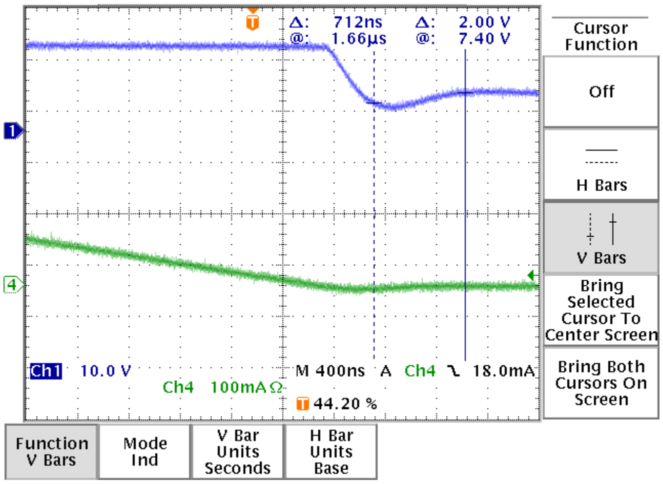

Figure 7. Current after addition of R-C snubber