Selecting Inductors for Buck Converters

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Introduction

This Application Note provides design information to help select an off-the-shelf inductor for any continuous-mode buck converter application.

The first part shows how the designer should estimate his requirements, specifically the required inductance.

The next part takes an off-the-shelf inductor and shows how to interpret the specs provided by the vendor in greater detail. A step-by-step procedure is provided.

Finally, all the previous steps are consolidated in a single design table, which answers the question: "How will the selected inductor actually perform in a specific application?"

The important point to note here is that though every inductor is designed assuming certain specific 'design conditions', that does not imply that these conditions cannot be varied. In fact every inductor can be satisfactorily used for many applications. But to be able to do this, the designer must know how to be able to accurately predict, or extrapolate, the performance of the inductor to a new set of conditions, which are his specific 'application conditions'. It will be shown that 'intuition' can be rather misleading. A detailed procedure is required and is presented in the form of the Design Table (*Table 2*) and the Selection Flow Chart (*Figure 2*).

Background: The Inductor Current Waveform

Refer to *Figure 1*, which shows the current through an inductor in continuous mode operation (bold line). Consider its main elements:

1. I_{DC}
   — is the geometrical center of the AC/ramp component
   — is the average value of the total inductor current waveform
   — is the current into the load, since the average current through the output capacitor, as for any capacitor in steady state, is zero
2. I_{PEAK} is $I_{DC}+\Delta I/2$, and it determines the peak energy in the core ($e = ½ * L * I^2$), which in turn is directly related to the peak field the core must withstand without saturating.
3. I_{TROUGH} is $I_{DC}-\Delta I/2$ and determines the constant residual level of current/energy in the inductor. Note that it depends on the load, even though it is not itself transferred to the load.
4. The AC component of the current is
   $I_{AC} = \Delta I = I_{PEAK} - I_{TROUGH}$
5. The DC component is the load current for the case shown in the figure.
   $I_{DC} = I_O$
   where I_O is the maximum rated load.
6. and 'r' is defined as the ratio of the AC to DC components (current ripple ratio) evaluated at maximum load, I_O. Note that by definition 'r' is a constant for a given converter/application (as it is calculated only at maximum load), and it is also defined only for continuous conduction mode.

$$r = \frac{\Delta I}{I_0}$$

A high inductance reduces ΔI and results in lower 'r' (and lower RMS current in the output capacitor), but may result in a very large and impractical inductor. So typically, for most buck regulators, 'r' is chosen to be in the range of 0.25–0.5 (at the maximum rated load). *See Appendix to this Application Note.* Once the inductance is selected, as we decrease the load on the converter (keeping input voltage constant), ΔI remains fixed but the DC level decreases and so the current ripple ratio increases. Ultimately, at the point of transition to discontinuous mode of operation, the DC level is ΔI/2 as shown in *Figure 1.* So

- The current ripple ratio at the point of transition to discontinuous mode is 2. Therefore, the upper limit for 'r' is also 2.
- The load at which this happens can be shown by simple geometry to be r/2 times I_O. So for example, if the inductance is chosen to be such that 'r' is 0.3 at a load of 2A, the transition to discontinuous mode of operation will occur at 0.15 times 2A, which is 300 mA.

Note: If the inductor is a 'swinging' inductor, its inductance normally increases as load current decreases and the point of transition to discontinuous mode may be significantly lower. We do not consider such inductors in this Application Note.

Estimating Requirements for the Application

There are two equivalent ways to go about calculating the required inductance and the designer should be aware of both.

BASIC METHOD TO CALCULATE L

From the general rule V = L*dI/dt we get during the ON time of the converter:

$$V_{IN} - V_{SW} - V_0 = L \times \frac{\Delta I}{D/f}$$

where V_{IN} is the applied DC input voltage, V_{SW} is the voltage across the switch when it is ON, D is the duty cycle and f is the switching frequency in Hz. Solving for ΔI we can write 'r' as:

$$r = \frac{(V_{IN} - V_{SW} - V_0) \times D}{L \times f \times I_0}$$

Now, for a buck regulator, we can show that the duty cycle is

Estimating Requirements for the Application (Continued)

$$D = \frac{V_O + V_D}{V_{IN} - V_{SW} + V_D}$$

where V_D is the forward drop across the catch diode (≅ 0.5V for a Schottky diode).

'r' can be finally written as:

$$r = \frac{(V_{IN} - V_{SW} - V_O) \times (V_O + V_D)}{(V_{IN} - V_{SW} + V_D) \times L \times f \times I_O}$$

and L is therefore

$$L = \frac{(V_{IN} - V_{SW} - V_O) \times (V_O + V_D)}{(V_{IN} - V_{SW} + V_D) \times r \times f \times I_O} \times 10^6 \ \mu H$$

where f is in Hz.

EXAMPLE 1

The input DC voltage is 24V into an LM2593HV buck converter. The output is 12V at a maximum load of 1A. We require an output voltage ripple of 30 mV peak-to-peak (±15 mV). We assume V_{SW} = 1.5V, V_D = 0.5V and f = 150,000 Hz.

Since, for loop stability reasons, we should not use any output capacitor of less than 100 mΩ, and since we do not wish to use an LC post filter, our ΔI must be

$$\Delta I = \frac{30\ mV}{100\ m\Omega} = 0.3A$$

So 'r' is

$$r = \frac{0.3}{1.0} = 0.3$$

The required inductance is

$$L = \frac{(24 - 1.5 - 12) \times (12 + 0.5)}{(24 - 1.5 + 0.5) \times 0.3 \times 150000 \times 1.0} \times 10^6 \ \mu H$$

L = 127 µH

The required energy handling capability is next calculated. Every cycle, the peak current is

$$I_{PEAK} = I_O + \frac{\Delta I}{2} = 1.0 + \frac{0.3}{2}\ A$$

$$I_{PEAK} = 1.15A$$

The required energy handling capability 'e' is

$$e = \frac{1}{2} \times L \times I_{PEAK}{}^2 \ \mu J$$

where L is in µH. So

$$e = \frac{1}{2} \times 127 \times 1.15^2 = 84 \ \mu J$$

Note: During a hard power-up (no soft start) or abnormal conditions like a short circuit on the output, the feedback loop is not effective in limiting the current to the value used above for calculating the energy handling capability. The current is actually going to hit the internal current limit of the device, I_{CLIM} in *Figure 1*, and this could be much higher than the steady state value calculated above. If the inductor has saturated, and if the input DC voltage is higher than 40V the current could slew up at a rate so high that the controller may not be able to limit the current at all, leading to destruction of the switch. Luckily, most off-the-shelf inductors are designed with large inherent air gaps and do not saturate very sharply even under overload conditions. **However we strongly recommend that at least when the input voltage is above 40V, the inductor should be sized to handle the worst case energy e_{CLIM}:**

$$e_{CLIM} = \frac{1}{2} \times L \times I_{CLIM}{}^2 \ \mu J$$

where L is in µH and I_{CLIM} is the internal limit of the regulator in amps.

VOLTSECONDS METHOD TO CALCULATE L

Talking in terms of voltseconds allows very general equations and curves to be generated. Here we talk of voltsµsecs or 'Et' which is simply the voltage across the winding of the inductor times the duration in µsecs for which it is applied.

Note:

- Current ramps up to the same peak value whether V (the applied voltage across inductor) is large but t (the time for which V is applied) is small, or whether V is small but t is large. So an infinite number of regulators with different combinations of input and output voltages but having the same voltseconds are actually the same regulator from the viewpoint of basic magnetics design. Et is what really counts. (The only exception to this is the Core Loss term since this depends directly on the absolute value of the frequency too, not just the Et).
- Also, Et can be calculated during the ON-time, (Vµsecs gained), or during the OFF-time (Vµsecs lost). Both will give the same result since there is no net change in Vµsecs per cycle in steady state.
- Also, remember that though Vµsecs is related to the energy in the core, it does not tell us the total energy. The Vµsecs gives information only about the AC component of 'r', i.e., ΔI. Combined with the DC component I_{DC}, it determines the peak current and energy of the inductor. So both I_O and Et are the variables on which our design procedure and tables are based upon. But a given application is completely defined by a certain I_O and Et (and frequency for the core loss term), and so these cannot be changed. Our only degree of freedom is L (or 'r') and we fix it according to the guidelines in the Appendix.

From the general equation V = L*dI/dt we can write that V*dt = L*dI. Here V*dt is the applied voltseconds. So by definition

$$Et = V\Delta t = L\Delta I \ V\mu secs$$

where L is in µH. 'r' can therefore be written as

$$r = \frac{Et}{L \times I_O}$$

Solving for L

$$L = \frac{Et}{r \times I_O} \ \mu H$$

which gives us an alternate and more general way of calculating L.

Estimating Requirements for the Application (Continued)

EXAMPLE 2

We repeat Example 1 from the viewpoint of Et.

The ON-time is

$$t_{ON} = \frac{D}{f} = \frac{(12 + 0.5) \times 10^6}{(24 - 1.5 + 0.5) \times 150000} \ \mu s$$

t_{ON} = 3.62 µsecs

So Et is

Et = $(V_{IN}–V_{SW}–V_O) \times t_{ON}$ = (24–1.5–12) x 3.62 Vµsecs

Et = 38.0 Vµsecs

L is therefore

$$L = \frac{Et}{r \times I_O} \ \mu H$$

$$L = \frac{38.0}{0.3 \times 1.0} \ \mu H$$

L = 127 µH

which gives us the same result as in Example 1 as expected.

SUMMARY OF REQUIREMENTS

- An inductance of 127 µH (or greater, based on maximum ‘r’ of 0.3)
- DC load of 1A (to ensure acceptable temperature rise, specify ΔT) **OR** steady state Energy handling capability of 84 µJ
- Peak load of 4.0A (to rule out core saturation if DC input voltage ≥40V) **OR** peak energy handling capability of 1016 µJ. (Max Current Limit of LM2593HV is 4.0A)
- Et of 38 Vµsecs
- Frequency 150 kHz

These can be communicated directly to a vendor for a custom-built design.

Characterizing an Off-the-Shelf Inductor

With reference to our design flow chart in *Figure 2*, the first pass selection is based upon inductance and DC current rating. We tentatively select a part from Pulse Engineering because its L and I_{DC} are close to our requirements, even though the rest does not seem to fit our application (see *Table 1* and bullets below). In particular the frequency for which the inductor was designed is 250 kHz, but our application is 150 kHz. We are intuitively lead to believe that since we are decreasing the frequency our core losses will go up, and so will the peak flux density. In fact the reverse happens in our case, and that is why it is important to follow the full procedure presented below. ‘Intuition’ can be very misleading.

The vendor also states that:

- The inductor is such that 380 mW dissipation corresponds to 50˚C rise in temperature.
- The core loss equation for the core is $6.11 \times 10^{-18} \times B^{2.7} \times f^{2.04}$ mW where f is in Hz and B is in Gauss.
- The inductor was designed for a frequency of 250 kHz.
- Et_{100} is the Vµsecs at which ‘B’ is 100 Gauss.

Note: For core loss equations it is conventional to use *half* the peak-to-peak flux swing. So, like most vendors, the ‘B’ above actually refers to ΔB/2. This must be kept in mind in the calculations that follow.

The step-by-step calculations are:

a) AC Component of Current:

This can be easily calculated from

Et = LΔI Vµsecs

where L is in µH.

So

$$\Delta I = \frac{Et}{L} = \frac{59.4}{137} = 0.434A$$

b) ‘r’:

So this inductor has been designed for the following ‘r’

$$r = \frac{\Delta I}{I_O} = \frac{0.434}{0.99}$$

r = 0.438

at a load current of 0.99A.

c) Peak Current:

$$I_{PEAK} = I_O + \frac{\Delta I}{2} = 0.99 + \frac{0.434}{2} \ A$$

I_{PEAK} = 1.21A

d) RMS Current:

$$I_{RMS} = \sqrt{I_O^2 + \frac{\Delta I^2}{12}} \ A$$

$$I_{RMS} = \sqrt{0.99^2 + \frac{0.434^2}{12}} \ A$$

I_{RMS} = 0.998A

e) Copper Loss:

This is

$P_{CU} = I_{RMS}^2 \times DCR$ mW

where DCR is in mΩ.

In most cases, to a close approximation, we can simply use I_{DC} instead of I_{RMS} in the above equation. Also sometimes, the vendor may have directly given the RMS current rating of the inductor.

$P_{CU} = 0.998^2 \times 387$ mW

P_{CU} = 385 mW

f) The AC Component of the B-Field:

This is proportional to the AC component of the inductor current.

Characterizing an Off-the-Shelf Inductor (Continued)

The vendor has provided the information that an Et of Et_{100} = 10.12 Vµsecs produces 100 Gauss (B). So since the inductor is designed for an Et = 59.4 Vµsecs, we get

$$B = \frac{Et}{Et_{100}} \times 100 = \frac{59.4}{10.12} \times 100 \text{ Gauss}$$

B = 587 Gauss

But this is half the peak-to-peak swing by convention. So

$\Delta B = 2 \bullet B$ = 1174 Gauss

CHECK: We can use the alternative form for ΔB as given in *Table 2*. We asked the vendor for more details than he had provided on the datasheet and we learned that the effective area of the core, A_e, is 0.0602 cm^2 and the number of turns is N = 84. So

$$\Delta B = \frac{100 \bullet Et}{N \bullet A_e} \text{ Gauss}$$

$$\Delta B = \frac{100 \bullet 59.4}{84 \bullet 0.0602} = 1175 \text{ Gauss}$$

which is what we expected.

g) The DC Component of the B-Field:

This is proportional to the DC component of the inductor current. In fact the instantaneous value of B can always be considered proportional to the instantaneous value of the current (for a given inductor).

The proportionality constant is known from f) above, i.e., a ΔI of 0.434A produces a ΔB of 1174 Gauss. So the DC component of the B-field must be

$$B_{DC} = \frac{\Delta B}{\Delta I} \times I_{DC} \text{ Gauss}$$

where

$I_{DC} = I_O$ = 0.99A

$$B_{DC} = \frac{1174}{0.434} \times 0.99 \text{ Gauss}$$

B_{DC} = 2678 Gauss

h) Peak B-Field:

Since B is proportional to I, we can write for the peak B-field

$$B_{PEAK} = B_{DC} + \frac{\Delta B}{2} \text{ Gauss}$$

$$B_{PEAK} = 2678 + \frac{1174}{2} \text{ Gauss}$$

B_{PEAK} = 3265 Gauss

i) Core Loss:

The vendor has stated that core loss (in mW) is 6.11 x 10^{-18} x $B^{2.7}$ x $f^{2.04}$ watts where f is in Hz and B is in Gauss.

$$P_{CORE} = 6.11 \times 10^{-18} \times \left[\frac{\Delta B}{2}\right]^{2.7} \times f^{2.04} \text{ mW}$$

P_{CORE} = 6.11 x 10^{-18} x $587^{2.7}$ x $250000^{2.04}$ mW

P_{CORE} = 18.7 mW

j) Total Inductor Loss:

$P = P_{CU} + P_{CORE}$ mW

P = 385 + 18.7 mW

P = 404 mW

k) Thermal Resistance of Inductor:

The vendor has stated that 380 mW dissipation corresponds to a 50˚C rise in temperature. So thermal resistance of the inductor is

$$R_{TH} = \frac{50}{\frac{380}{1000}} \text{ °C/W}$$

R_{TH} = 131.6˚C/W

l) Estimated Temperature Rise of Inductor:

$$\Delta T = R_{TH} \times \frac{P}{1000} \text{ °C}$$

$$\Delta T = 131.6 \times \frac{404}{1000} \text{ °C}$$

ΔT = 53˚C

Here the temperature rise 'ΔT' is the temperature of the core, 'T_{CORE}' minus the worst case ambient temperature '$T_{AMBIENT}$'. The 'ambient' is the local ambient around the inductor.

m) Energy Handling Capability of Core:

$$e = \frac{1}{2} \times L \times I_{PEAK}^2 \ \mu J$$

where L is in µH

$$e = \frac{1}{2} \times 137 \times 1.21^2 \ \mu J$$

e = 100 µJ

As before, we warn that the energy in the core during hard power-up or a short circuit on the outputs, may be significantly higher.

In case of soft-start it should also be remembered that there are several ways to implement this feature, and not all lead to a reduction in switch or inductor current at start-up. The worst condition is *start-up with a short already present on the output.* The inductor waveforms should therefore be monitored on the bench during all conditions to check this out.

Also it will be seen that all inductors of a 'family', i.e., using the same core will typically have the same rated energy capability. So if this core is found to be inadequate, normally the only way out is to move to a physically larger core/inductor. Other options include the use of improved and more expensive core materials.

Characterizing an Off-the-Shelf Inductor (Continued)

SUMMARY OF INDUCTOR PARAMETERS

- The inductor is designed for about 50˚C rise in temperature over ambient at a load of 1A.
- The copper losses (385 mW) predominate (as is usual for such inductors/core materials) and the core losses are relatively small.
- The peak flux density is about 3200 Gauss, which occurs at a peak instantaneous current of 1.2A.
- The rated energy handling capability of the core is 100 µJ.

Note: Most vendors do not explicitly provide the material used, though an astute designer can figure this out by looking at the exponents of B and f in the core loss equation provided, or of course simply by asking the vendor. In this case we know that the material is ferrite and can typically handle a peak flux density of over 3000–4000 Gauss before it starts to saturate. (Caution: not all ferrite grades are similar in this regard and also that the saturation flux density B_{SAT} falls as the core heats up.)

Evaluating the Inductor for the Actual Application

Above we have the limits of the inductor operating under its design conditions. We will now *extrapolate* its performance to our specific application conditions. Unprimed parameters are the original 'design values', and the corresponding primed parameters are the extrapolated 'application values'.

The following are the design conditions (these may be allowed to change):

- I_{DC}
- Et
- f
- $T_{AMBIENT}$

The 'Application Conditions' are:

- I'_{DC}
- Et'
- f'
- $T'_{AMBIENT}$

In going from the 'Design Conditions' to the 'Application Conditions' the following are considered constant

- L
- DCR
- Rth
- The core loss equation

And, finally, to 'approve' the inductor for the given application we need to certify

- 'r' is acceptable (choice of L).
- B_{PEAK} OK.
- $I_{PEAK} < I_{CLIM}$.
- ΔT OK (evaluate $P_{CU} + P_{CORE}$).
- $B_{CLIM} < B_{SAT}$ (if DC input voltage is ≥40V).

We assume the vendor has provided all the following inputs:

- Et (Vµsecs)
- Et_{100} (Vµsecs per 100 Gauss)
- L (µH)
- I_{DC} (Amps)
- DCR (mΩ)
- f (Hz)
- The form for core losses (mW) as $a*B^{b}*f^{c}$, where B is in Gauss, f in Hz. Note that B is half the peak-to-peak flux swing.
- Thermal resistance of inductor in free air (˚C/W)

If any of these are unknown, the vendor should be contacted. *Table 2* condenses the step-by-step procedure given earlier and also shows how to 'extrapolate' the performance of the inductor.

EXAMPLE 3

This shows the complete selection procedure. Refer to *Table 2* and *Figure 2*. We have seen that the 'Design Conditions' of the inductor are:

- Et = 59.4 Vµsecs
- f = 250,000 Hz
- I_{DC} = 0.99A

Our 'Application Conditions' are

- Et' = 38 Vµsecs
- f' = 150,000 Hz
- I'_{DC} = 1A

(We assume that $T_{AMBIENT}$ is unchanged so we can ignore it above).

We need to verify that using the inductor in the given application:

a) current ripple ratio 'r' is close to desired

b) peak flux density/current are within bounds

c) temperature rise is acceptable

Using *Table 2*:

a) 'r':

Design Value:

$$r = \frac{Et}{L \bullet I_{DC}}$$

$$r = \frac{59.4}{137 \bullet 0.99}$$

r = 0.438

Extrapolated to our Application:

$$r' = r \bullet \left[\frac{Et' \bullet I_{DC}}{Et \bullet I'_{DC}}\right]$$

$$r' = 0.438 \bullet \left[\frac{38 \bullet 0.99}{59.4 \bullet 1}\right]$$

r' = 0.277

We expected 'r" to be slightly lower than 0.3 since the chosen inductor has a higher inductance than we required (137 µH instead of 127 µH). This is acceptable however as the output voltage ripple will be less than demanded.

b) Peak Flux Density

Design Value:

Evaluating the Inductor for the Actual Application (Continued)

$$B_{PEAK} = \frac{200}{Et_{100}} \cdot \left[(I_{DC} \cdot L) + \frac{Et}{2} \right] \text{Gauss}$$

$$B_{PEAK} = \frac{200}{10.12} \cdot \left[(0.99 \cdot 137) + \frac{59.4}{2} \right] \text{Gauss}$$

B_{PEAK} = 3267 Gauss

Extrapolated to our Application:

$$B'_{PEAK} = B_{PEAK} \cdot \left[\frac{2 \cdot L \cdot I'_{DC} + Et'}{2 \cdot L \cdot I_{DC} + Et} \right] \text{Gauss}$$

$$B'_{PEAK} = 3267 \cdot \left[\frac{2 \cdot 137 \cdot 1 + 38}{2 \cdot 137 \cdot 0.99 + 59.4} \right] \text{Gauss}$$

B'_{PEAK} = 3084 Gauss

which is less than B_{PEAK} and therefore acceptable.

c) Peak Current

To ensure that the regulator will deliver rated load, we need to ensure that the peak current is less than the internal current limit of the Switcher IC.

Design Value:

$$I_{PEAK} = I_{DC} + \frac{Et}{2 \cdot L} \text{ A}$$

$$I_{PEAK} = 0.99 + \frac{59.4}{2 \cdot 137} \text{ A}$$

I_{PEAK} = 1.21A

This corresponds to a B-field of 3267 Gauss as calculated above.

Extrapolated to our Application:

$$I'_{PEAK} = I_{PEAK} \cdot \left[\frac{(2 \cdot L \cdot I'_{DC}) + Et'}{(2 \cdot L \cdot I_{DC}) + Et} \right] \text{A}$$

$$I'_{PEAK} = 1.21 \cdot \left[\frac{(2 \cdot 137 \cdot 1.0) + 38}{(2 \cdot 137 \cdot 0.99) + 59.4} \right] \text{A}$$

I'_{PEAK} = 1.14A

This corresponds to a B-field of 3084 Gauss as calculated above and is less than I_{CLIM}. (Min Current Limit of LM2593HV is 2.3A).

d) Temperature Rise:

Design Values:

$$P_{CU} = DCR \cdot (I_{DC}^2 + \frac{Et^2}{12 \cdot L^2}) \text{ mW}$$

$$P_{CU} = 387 \cdot (0.99^2 + \frac{59.4^2}{12 \cdot 137^2}) \text{ mW}$$

P_{CU} = 385 mW

$$P_{CORE} = a \cdot [\frac{Et}{Et_{100}} \cdot 100]^b \cdot f^c \text{ mW}$$

where the vendor has provided that a = $6.11*10^{-18}$, b = 2.7 and c = 2.04. So

$$P_{CORE} = 6.11 \cdot 10^{-18} \cdot [\frac{59.4}{10.12} \cdot 100]^{2.7} \cdot f^{2.04} \text{ mW}$$

P_{CORE} = 18.7 mW

So

$$\Delta T = Rth \cdot \frac{P_{CU} + P_{CORE}}{1000} \, ^\circ C$$

$$\Delta T = \frac{50}{0.380} \cdot \frac{385 + 18.7}{1000} \, ^\circ C$$

ΔT = 53˚C

because the vendor has stated that 380 mW dissipation in the inductor causes 50˚C rise in temperature.

Extrapolated to our Application:

$$P'_{CU} = P_{CU} \cdot \frac{(12 \cdot I'^2_{DC} \cdot L^2) + Et'^2}{(12 \cdot I_{DC}^2 \cdot L^2) + Et^2} \text{ mW}$$

$$P'_{CU} = 385 \cdot \frac{(12 \cdot 1^2 \cdot 137^2) + 38^2}{(12 \cdot 0.99^2 \cdot 137^2) + 59.4^2} \text{ mW}$$

P'_{CU} = 389 mW

$$P'_{CORE} = P_{CORE} \cdot \left[\left(\frac{Et'}{Et}\right)^b \cdot \left(\frac{f'}{f}\right)^c \right] \text{mW}$$

$$P'_{CORE} = 18.7 \cdot \left[\left(\frac{38}{59.4}\right)^{2.7} \cdot \left(\frac{150000}{250000}\right)^{2.04} \right] \text{mW}$$

P'_{CORE} = 2 mW

So,

$$\Delta T' = \Delta T \cdot \left[\frac{P'_{CU} + P'_{CORE}}{P_{CU} + P_{CORE}} \right] {}^\circ C$$

$$\Delta T' = 53 \cdot \left[\frac{389 + 2}{385 + 18.7} \right] {}^\circ C$$

ΔT' = 51˚C

which is considered to be acceptable in this application.

Conclusions

By the detailed selection procedure above, we can expect the selected inductor to work well for the given lower frequency application example. As mentioned earlier, we would have 'intuitively' thought that since the inductance and current rating is about what we need, if we lowered the frequency from 250 kHz to 150 kHz, the peak current and field would increase. But they actually decrease as we can now see. The reason being, that the inductor was designed for a higher Et in mind (59.4 Vµsecs vs. our 38 Vµsecs). As stated earlier, Et in effect, defines the regulator configuration itself, so we did not just lower the frequency, we actually went to an entirely different input-output voltage combination to what the inductor had been originally designed for. As we can now guess, the original inductor had been probably designed for a much higher applied voltage to what we subjected it to. But this was not obvious at first sight. The full procedure as given in *Figure 2* and *Table 2* is therefore necessary to avoid such 'errors of intuition'.

The data sheets of National's Simple Switchers also generally include simple nomograms and these are useful in most cases, but limit the selection to certain previously specified or custom built inductors and are also based on certain assumptions. In particular, there are many factors to consider when fixing a certain current ripple ratio 'r', which happens to be the key input in the process of selection an inductor. Nomograms are easy to use but assume a certain 'r' which may not be ideal for all purposes. In fact in the example discussed above, we did in fact select an inductance higher than what the nomograms may have recommended, because of output voltage ripple considerations.

In general, this Application Note should help in selecting a more optimum and readily available off-the-shelf inductor.

Appendix: Optimizing the Size of the Inductor

The size of the inductor is related to the energy handling capability required. The energy handling capability is $½*L*I_{PEAK}^2$. For a given application, if we reduce inductance, it seems that this would increase ΔI and thereby I_{PEAK}, which would cause the energy requirement to increase since it depends on square of current. However, a detailed calculation shows agin that reality is counterintuitive. The energy handling requirement is actually substantially *reduced* if the inductance is decreased. In terms of 'r', we can in fact write the energy handling capability as

$$e = \frac{I_O \bullet Et}{8} \bullet \left[r \bullet \left(\frac{2}{r} + 1 \right)^2 \right] \mu J$$

where 'r' is $\Delta I/I_O$ and Et is in Vµsecs.

For a given application, Et is fixed as is I_O, so the term in square brackets gives 'e' the shape shown in *Figure 3*. We can see that the energy handling requirement (size of inductor) decreases as 'r' increases (L decreasing). The best value is the 'knee' and so it is a good idea to target an 'r' of 30%–40%. No great improvement in the size of the inductor will take place by increasing 'r' much more than this, but the RMS current in the output cap, and also the RMS current in the input capacitor (especially for large duty cycles), will increase substantially. The absolute value of the RMS ripple current in the input capacitor is much higher than in the output capacitor, and the designer should watch out for the cost penalty on the input capacitor too! Refer to *Table 3* for the complete set of optimization equations expressed as a function of 'r'.

While optimizing, the following points need to be considered:

- For a given application, having defined input and output voltages and load current, Et is fixed as are D and I_O. So the only degree of freedom is in selecting the 'r'. The equations in *Table 3* are therefore written in terms of 'r'.
- *Table 3* provides the general equations required for optimization but also provides the values at an 'r' of 0.3 in the adjacent column as a benchmark. This is also equivalent to the 'flat top approximation' often used for quick estimates.
- *Figure 3* plots the variation of each parameter, normalized to the benchmark values (i.e., set to unity at an 'r' of 0.3).
- Note that when calculating dissipation in the switch, one must consider whether the switch is a bipolar transistor or a FET. If it is a FET, we need to apply I^2*R where I is the RMS switch current and R is the R_{DS} of the FET. If it is bipolar, we need to use V*I where V is the saturation voltage across the switch, and I is the average switch current. That is why both have been provided in *Table 3*. Also note that for a bipolar switch, the dissipation is seems almost independent of 'r'. In practice, the saturation voltage drop depends on the instantaneous value of current, so dissipation does increase slightly with 'r'.
- Referring to *Figure 3* we can see that the RMS inductor current hardly changes over a very wide range of 'r'. That is why, earlier in this Application Note, it was mentioned that for the purpose of evaluating copper losses we may use I_O^2 instead of I_{RMS}^2.
- The core losses also increase substantially with increasing 'r'. It can be shown that even if we keep the same core size, *the flux density B_{AC} will go up as* $r^{½}$. Going to a smaller core could increase this further.
- The RMS capacitor currents in the input and output are the main components to consider because they can increase rapidly with 'r'. So for example, if we increase 'r' to 0.6 (from 0.3), the energy handling requirement of the inductor falls by about 35% but the dissipation in the output capacitors (if ESR is unchanged) will increase by 400%! Alternatively stated, we now need to select an output capacitor with twice the ripple current rating.
- It must also be kept in mind that there is an output voltage ripple ΔV = ESR x ΔI associated with the current ripple. Now, the ESR of the output capacitor cannot usually be decreased below 100 mΩ–200 mΩ (with voltage mode control) for loop stability reasons. So for high loads (and high 'r'), the dissipation in the output capacitor will necessarily be high since we cannot reduce ESR further. This may call for physically large sized output capacitors to handle the dissipation. In addition, the output voltage ripple will be high too, and since we cannot reduce this by reducing the ESR, we will need to add on a post LC filter. So, for high load currents, it may become necessary to decrease 'r' substantially. This in turn will lead to a large inductor with slow transient response ability.

Appendix: Optimizing the Size of the Inductor (Continued)

- We have implied that the physical size of the inductor is related to its energy handling capability only. This in turn suggests that we are talking of inductor designs that are core-saturation limited. While this is usually true if the core material is a ferrite, it may not be true of some powdered iron inductors for example. The size of these may be limited not by core saturation but by core losses, which depend on flux swing, or ΔI, not I (or 'e'). So, while *Figure 3* is still valid, the criterion of 'best choice' may change. It may be necessary to choose or restrict 'r' to much smaller values than the 'knee'.

This completes the information required to optimize not only the inductor but the buck regulator itself. The key factor affecting cost/size of almost all the components is the current ripple ratio 'r' and this needs to be carefully optimized as discussed above.

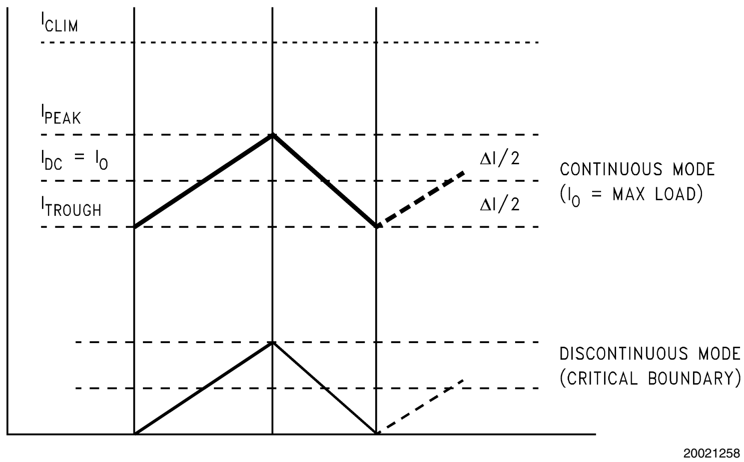

FIGURE 1. Inductor Current Waveform

TABLE 1. Specifications of Available Inductor

Part Number	Reference Values			Control Values	Calculation Data
	I_{DC} (Amps)	L_{DC} (µH)	Et (Vµsecs)	DCR (nom) mΩ	Et_{100} (Vµsecs)
P0150	0.99	137	59.4	387	10.12

Appendix: Optimizing the Size of the Inductor (Continued)

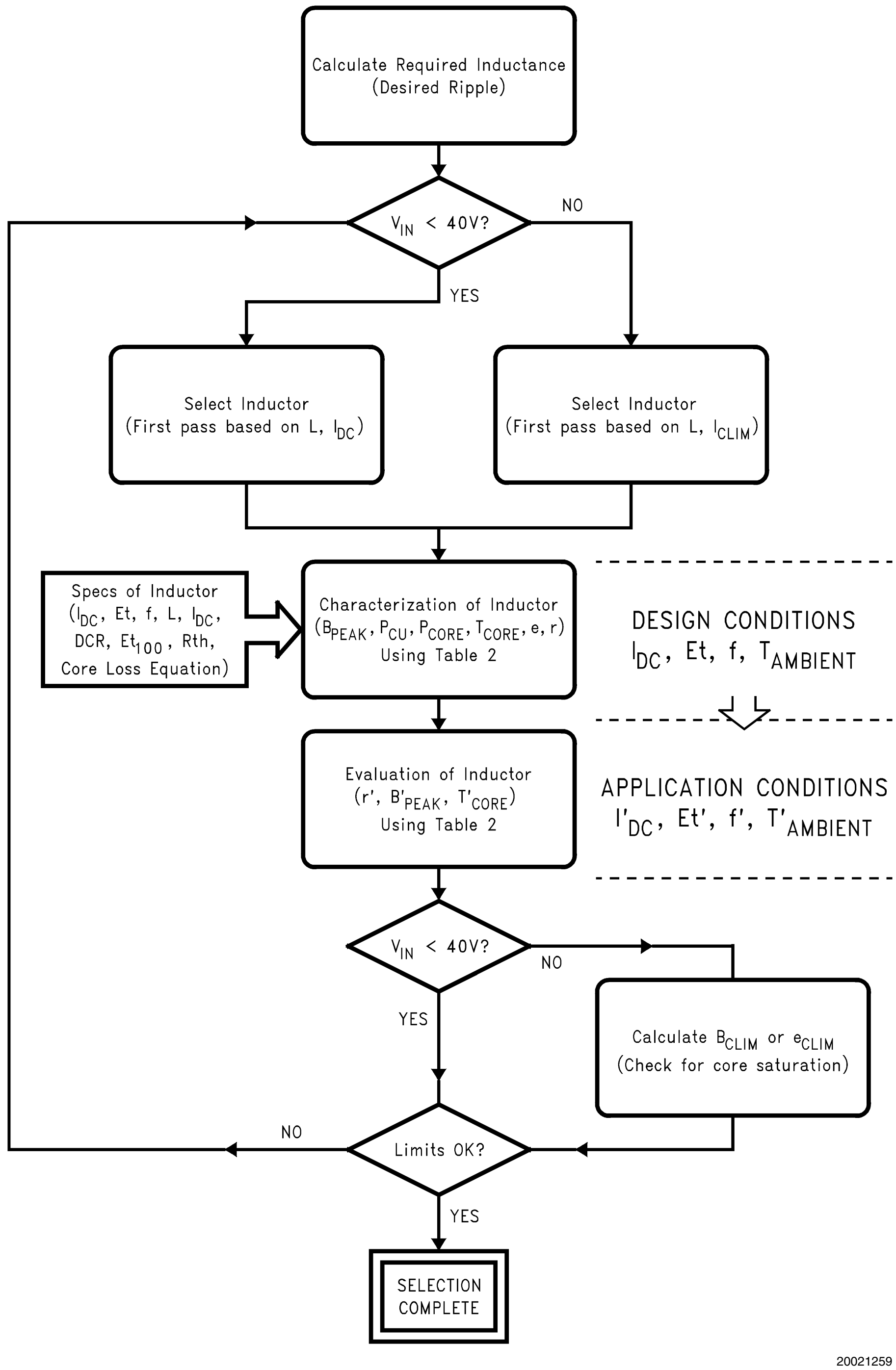

FIGURE 2. Design Flow Chart for Selection of Inductor

Appendix: Optimizing the Size of the Inductor (Continued)

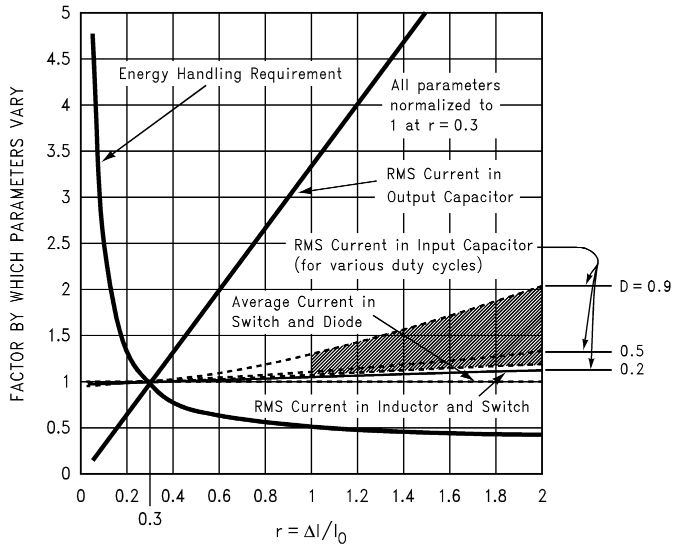

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FIGURE 3. Optimization Chart for Setting 'r'

TABLE 2. Complete Design Table for Evaluating the Inductor for a Given Application

Design Parameters	Design Conditions I_{DC}, Et, f, $T_{AMBIENT}$	Application Conditions $I'_{DC} = I_O$, Et', f', $T'_{AMBIENT}$
AC Component of Current Amps	$\Delta I = \frac{Et}{L}$	$\Delta I' = \Delta I \cdot \left[\frac{Et'}{Et}\right]$
Current Ripple Ratio 'r' ($\Delta I/I_{DC}$)	$r = \frac{Et}{L \cdot I_{DC}}$	$r' = r \cdot \left[\frac{Et' \cdot I_{DC}}{Et \cdot I'_{DC}}\right]$
Peak Current in Inductor Amps	$I_{PEAK} = I_{DC} + \frac{Et}{2 \cdot L}$	$I'_{PEAK} = I_{PEAK} \cdot \left[\frac{(2 \cdot L \cdot I'_{DC}) + Et'}{(2 \cdot L \cdot I_{DC}) + Et}\right]$
RMS Current in Inductor Amps	$I_{RMS} = \sqrt{I_{DC}^2 + \frac{Et^2}{12 \cdot L^2}}$	$I'_{RMS} = I_{RMS} \cdot \left[\frac{(12 \cdot I'^2_{DC} \cdot L^2) + Et'^2}{(12 \cdot I_{DC}^2 \cdot L^2) + Et^2}\right]^{1/2}$
AC Flux Density Gauss	$\Delta B = \frac{Et}{Et_{100}} \cdot 200 = \frac{100 \cdot Et}{N \cdot A_e}$	$\Delta B' = \Delta B \cdot \left[\frac{Et'}{Et}\right]$
Peak Flux Density Gauss	$B_{PEAK} = \frac{200}{Et_{100}} \cdot \left[(I_{DC} \cdot L) + \frac{Et}{2}\right]$	$B'_{PEAK} = B_{PEAK} \cdot \left[\frac{2 \cdot L \cdot I'_{DC} + Et'}{2 \cdot L \cdot I_{DC} + Et}\right]$
Copper Losses mW	$P_{CU} = DCR \cdot (I_{DC}^2 + \frac{Et^2}{12 \cdot L^2})$	$P'_{CU} = P_{CU} \cdot \frac{(12 \cdot I'^2_{DC} \cdot L^2) + Et'^2}{(12 \cdot I_{DC}^2 \cdot L^2) + Et^2}$
Core Losses mW	$P_{CORE} = a \cdot [\frac{Et}{Et_{100}} \cdot 100]^b \cdot f^c$	$P'_{CORE} = P_{CORE} \cdot \left[\left(\frac{Et'}{Et}\right)^b \cdot \left(\frac{f'}{f}\right)^c\right]$
Energy in Core µJ	$e = \frac{1}{2} \cdot L \cdot [I_{DC} + \frac{Et}{2 \cdot L}]^2$	$e' = e \cdot \left[\frac{(2 \cdot L \cdot I'_{DC}) + Et'}{(2 \cdot L \cdot I_{DC}) + Et}\right]^2$

Appendix: Optimizing the Size of the Inductor (Continued)

TABLE 2. Complete Design Table for Evaluating the Inductor for a Given Application (Continued)

Design Parameters	Design Conditions I_{DC}, Et, f, $T_{AMBIENT}$	Application Conditions $I'_{DC} = I_O$, Et', f', $T'_{AMBIENT}$
Temperature Rise (ΔT) ˚C	$\Delta T = Rth \bullet \frac{P_{CU} + P_{CORE}}{1000}$	$\Delta T' = \Delta T \bullet \left[\frac{P'_{CU} + P'_{CORE}}{P_{CU} + P_{CORE}}\right]$

Et in Vµsecs, DCR in mΩ, L in µH, f in Hz, Effective Area Ae in cm^2, N is number of turns

TABLE 3. Optimization Table for Fixing Current Ripple Ratio 'r'

Parameters	As a Function of 'r'	For 'r' = 0.3 (to a first approximation)
Energy Handling Capability µJ	$\frac{I_O \bullet Et}{8} \bullet \left[r \bullet \left(\frac{2}{r} + 1\right)^2\right]$	$2.2 \bullet I_O \bullet Et$
RMS Current in Output Cap Amps	$I_O \bullet \frac{r}{\sqrt{12}}$	$0.09 \bullet I_O$
RMS Current in Input Cap Amps	$I_O \bullet \sqrt{D \bullet \left[1 - D + \frac{r^2}{12}\right]}$	$I_O \bullet \sqrt{D \bullet (1 - D)}$
RMS Current in Inductor Amps	$I_O \bullet \sqrt{1 + \frac{r^2}{12}}$	I_O
RMS Current in Switch Amps	$I_O \bullet \sqrt{D \bullet \left[1 + \frac{r^2}{12}\right]}$	$I_O \bullet \sqrt{D}$
Average Current in Switch Amps	$I_O \bullet D$	$I_O \bullet D$
Average Current in Diode Amps	$I_O \bullet (1–D)$	$I_O \bullet (1–D)$

$r = \Delta I/I_O$, Et in Vµsecs

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