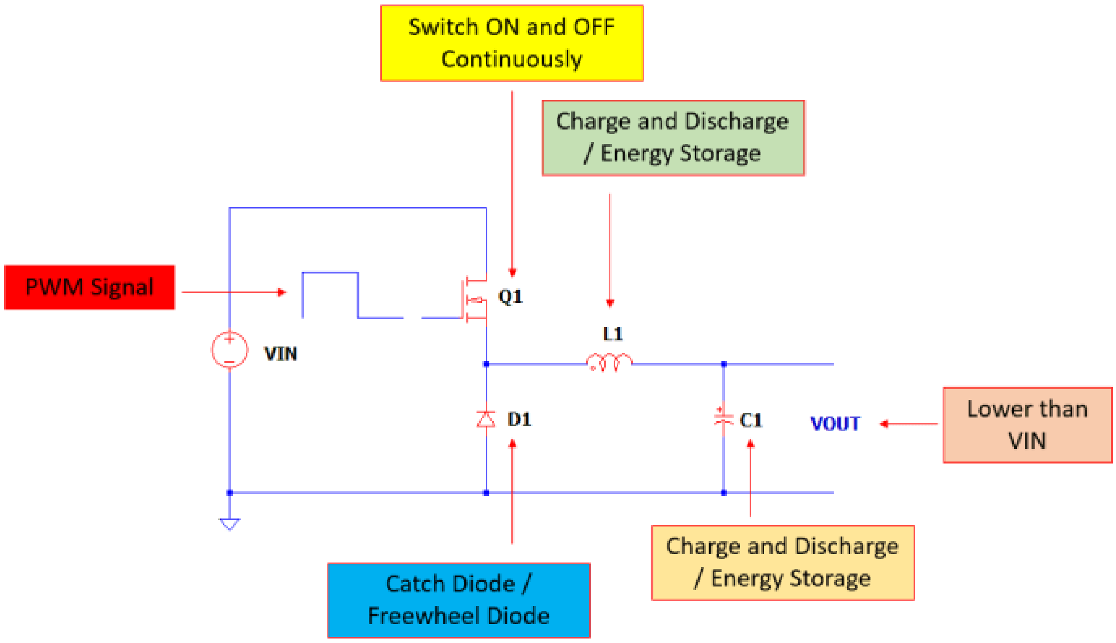


Buck Converter Design Tutorial

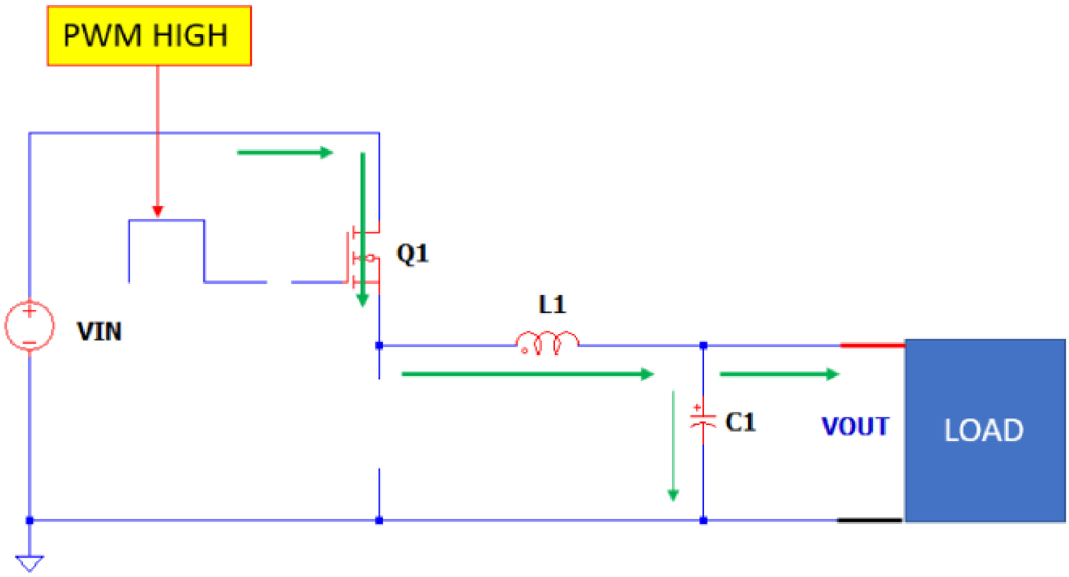
Buck Converter Basic Function

Buck converter operates by continuously turning ON and OFF a semiconductor switch like BJT, MOSFET or IGBT. The turning ON and OFF of the switch is determined by the duty cycle. The ideal duty cycle of a buck converter is simply

$$Duty\ Cycle = V_{Out} / V_{In}$$

Buck Converter Basic Operation – PWM is High

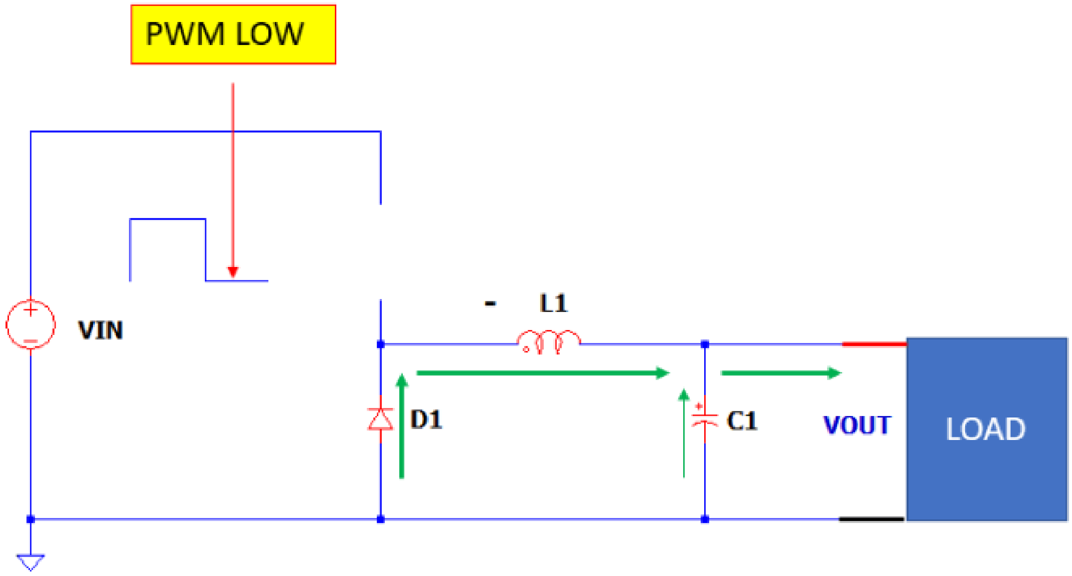
When the PWM is at high state, Q1 will conduct at saturation (very low voltage drop). D1 will be reversed biased and not part in the current loop. The current will flow from VIN, going to the channel of Q1, then charging L1 and a portion will charge C1 and finally the main current path will go to the load.



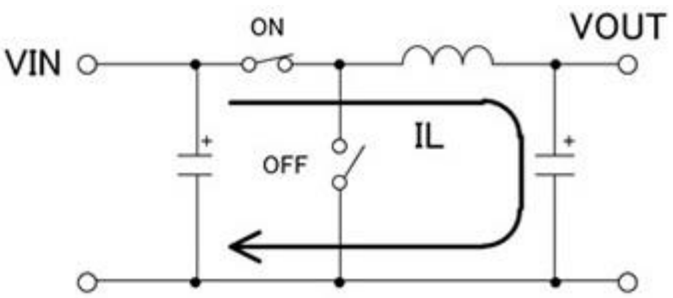
At this time, L1 will charge and the dot side will be at higher potential. L1's current will ramp up linearly.

Buck Converter Basic Operation – PWM is Low

When PWM is low, Q1 will turn off and not anymore part of the current loop. The dot side of the inductor L1 will become negative potential as the L1 will reversed polarity but maintaining the same direction of current. The current path will be from the D1, to L1 that is discharging at this time, then to the load. At this time also, C1 energy will help providing the need of the load.

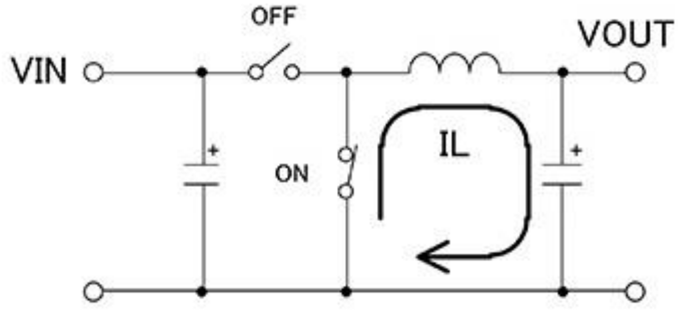


Below, a model of a basic step-down converter is used to explain the circuit operation.



$$IL = \frac{VIN - VOUT}{L} \times ton$$

- When the high-side switch (the transistor) turns on, a current IL flows in the inductor L, and energy is stored
- At this time, the low-side switch (the diode) is turned off
- The inductor current IL is expressed by the following equation （ton：ON-time）



$$IL = \frac{VOUT}{L} \times toff$$

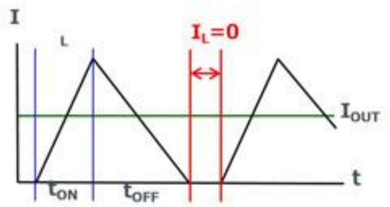
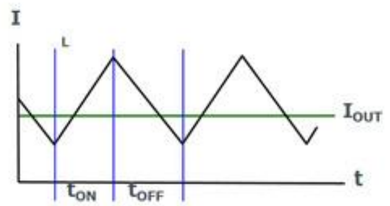
- When the high-side switch (the transistor) turns off, the energy stored in the inductor is output through the low-side switch (the diode)
- At this time, the high-side switch (the transistor) is OFF
- The inductor current IL is expressed by the following equation

  （toff： OFF time）

Discontinuous Mode and Continuous Mode

In switching operation, there are two modes, a discontinuous mode and a continuous mode. They are compared in the following table.

The "operation" item for comparison is the waveform of the inductor currents. In discontinuous mode, there is a period in which the inductor current IL is interrupted, hence the name, discontinuous mode. In contrast, in continuous mode there is no period in which the inductor current is zero.

Comparison item	Discontinuous mode	Continuous mode
Operation	 There is a zero-inductor current period between ON and OFF, so that the inductor current is not continuous.	 The inductor current flows continuously, which turns ON and OFF at the same frequency as the switching frequency.
Inductor	Inductance ↓, size ↓, cost ↓	Inductance ↑, size ↑, cost ↑
Rectifying Diode	Fast recovery type, cost ↓	Requires a faster recovery type, cost ↑
Switching Transistor	Allowable power ↑, size ↑, cost ↑	Allowable power ↓, size ↓, cost ↓
Efficiency	Switching loss ↓, efficiency ↑	Switching loss ↑, efficiency ↓

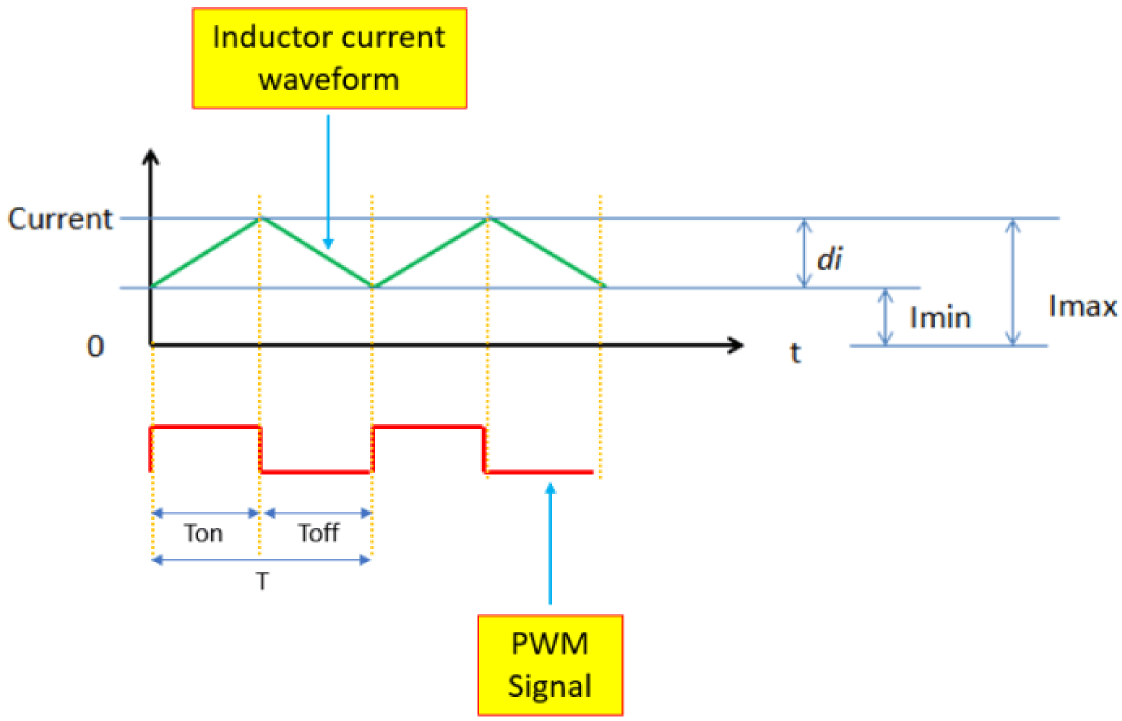
Comprehensive Buck Converter Design Tutorial

Basic function and operation of a buck converter has been tackled. So, here we go to our main topic which is buck converter design tutorial. Below is the outline of this buck converter design tutorial.

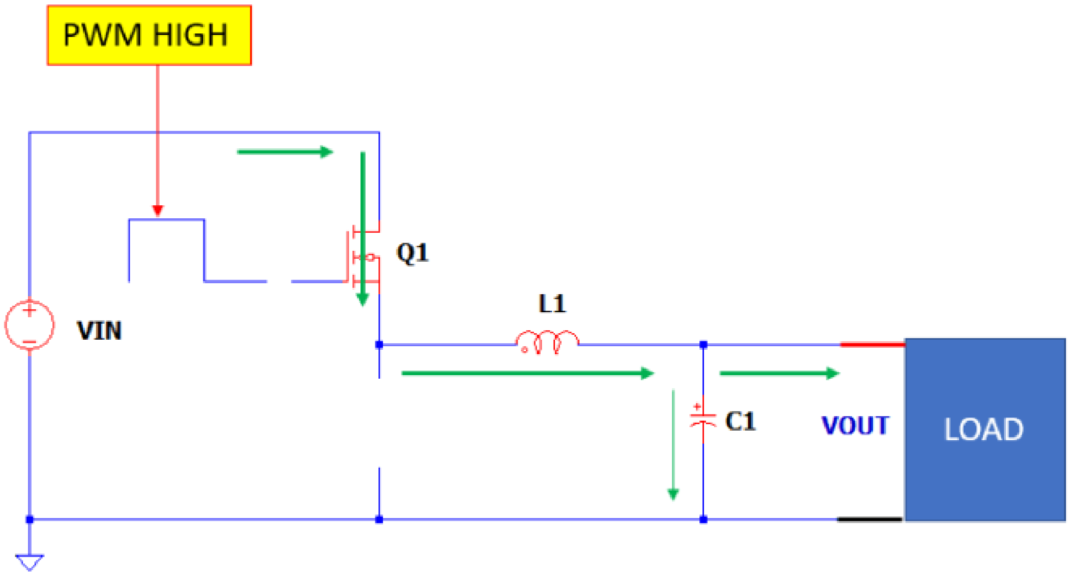
1. Inductor Ripple Current Derivation

To derive the inductor current equations, it is important to know its waveform. By the way, a buck converter is can be categorized as CCM, TM or DCM. CCM stands for continuous conduction mode while TM stands for transition mode or sometimes called as boundary mode. On the other hand, DCM stands for discontinuous conduction mode. CCM and TM are having the same analysis while DCM requires different one. For high power applications, it is unlikely to intentionally operate the buck converter at DCM mode. This will result to a very high losses and impractical.

However, there is a time that buck converter will enter DCM mode, and this is when the load is very light. So, the design point or component selection will be based on the heavy load and this is mostly at CCM. So, in this derivation, we will be considering a CCM operation. Below in green is the current waveform of the inductor operating at CCM. It rises linearly when the PWM signal is high. It then decreases linearly when the PWM signal is low.



When PWM is high, the analysis will be:



The key equation to use is the voltage across an inductor that is

$$V_L = L \times di / dt$$

$$VL1 = L1 \cdot \frac{di}{dt}$$

$$VL1 = L1 \cdot \frac{di}{dt} \text{ solve}, di \rightarrow \frac{VL1 \cdot dt}{L1}$$

$$di\_Ton = \int_0^{Ton} \frac{VL1\_Ton}{L1} dt \text{ simplify} \rightarrow di\_Ton = \frac{Ton \cdot VL1\_Ton}{L1}$$

Finding VL1_Ton

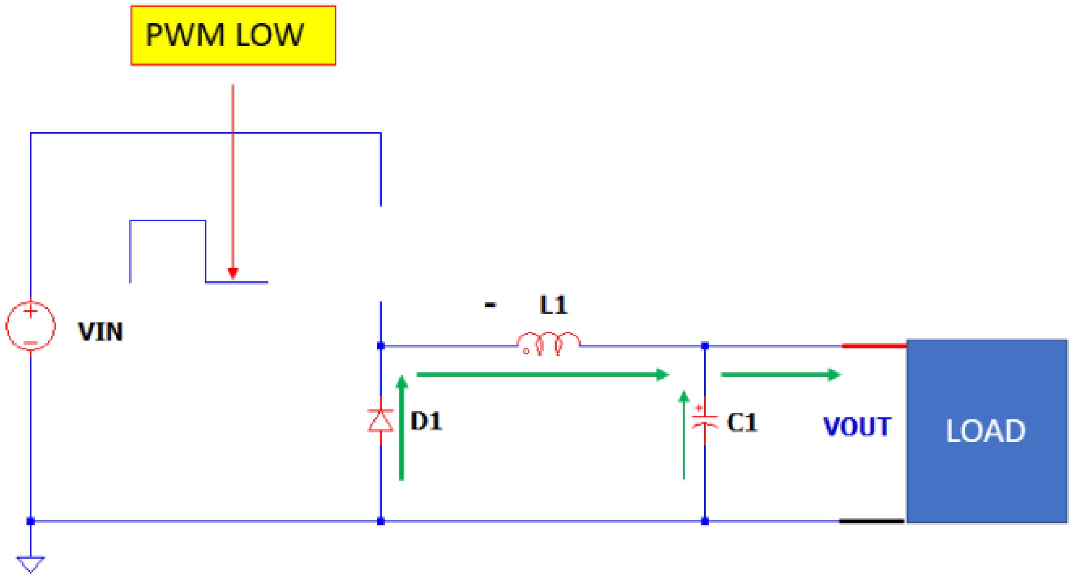
$$VIN - VQ1 - VL1\_Ton - VOUT = 0 \text{ solve}, VL1\_Ton \rightarrow VIN - VQ1 - VOUT$$

$$VL1\_Ton = VIN - VQ1 - VOUT$$

$$di\_Ton = \frac{Ton \cdot (VIN - VQ1 - VOUT)}{L1} \text{ substitute}, Ton = D \cdot T \rightarrow di\_Ton = -\frac{D \cdot T \cdot (VQ1 - VIN + VOUT)}{L1}$$

$$\boxed{di\_Ton = -\frac{D \cdot T \cdot (VQ1 - VIN + VOUT)}{L1}}$$

When PWM is low, the analysis will be:



$$di\_Toff = \int_{Ton}^{T} \frac{VL1\_Toff}{L1}\,dt \text{ simplify } \rightarrow di\_Toff = \frac{VL1\_Toff\cdot(T - Ton)}{L1}$$

$$di\_Toff = \frac{VL1\_Toff\cdot(T - Ton)}{L1} \text{ substitute}, Ton = D\cdot T \rightarrow di\_Toff = -\frac{T\cdot VL1\_Toff\cdot(D - 1)}{L1}$$

$$di\_Toff = -\frac{T\cdot VL1\_Toff\cdot(D - 1)}{L1}$$

Finding VL1_Toff

$$-VD1 + VL1\_Toff - VOUT = 0 \text{ solve}, VL1\_Toff \rightarrow VD1 + VOUT$$

$$VL1\_Toff = VD1 + VOUT$$

$$\boxed{di\_Toff = -\frac{T\cdot(VD1 + VOUT)\cdot(D - 1)}{L1}}$$

Both di_Ton and di_Toff will give the same result.

2. Duty cycle Derivation

If you examine the inductor current waveform, the rise and the fall are in equal magnitude. Therefore, both equations di_Ton and di_Toff above are can be equated and we derived final the duty cycle.

$$di\_Ton = di\_Toff$$

Substitute D with Dutycycle

$$-\frac{Dutycycle \cdot T \cdot (VQ1 - VIN + VOUT)}{L1} = -\frac{T \cdot (VD1 + VOUT) \cdot (Dutycycle - 1)}{L1}$$

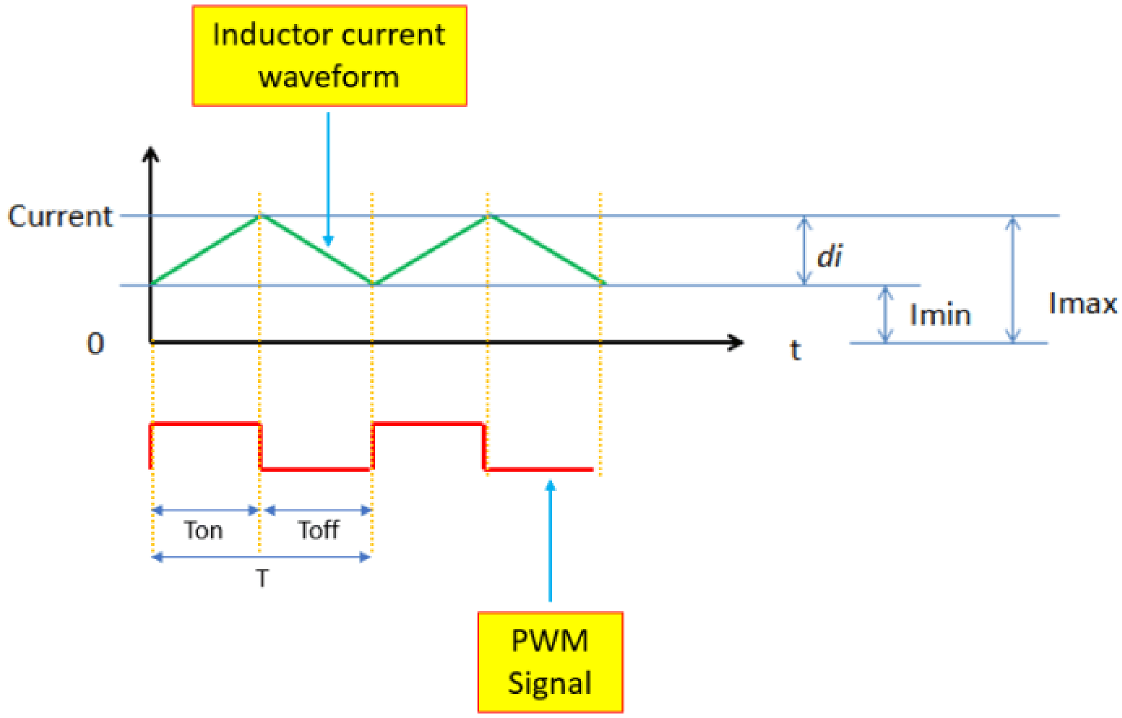
$$-Dutycycle \cdot (VQ1 - VIN + VOUT) = -(VD1 + VOUT) \cdot (Dutycycle - 1) \text{ solve}, Dutycycle \rightarrow \frac{VD1 + VOUT}{VD1 - VQ1 + VIN}$$

$$Dutycycle = \frac{VD1 + VOUT}{VD1 - VQ1 + VIN}$$ Dutycycle considering actual voltage drops

$$Dutycycle = \frac{VOUT}{VIN}$$ Dutycycle not considering the voltage drops. Ideal dutycycle.

3. Inductor RMS Current Derivation

We will start with the inductor RMS current is the total of the RMS of di and Imin in the below waveform.



$$I_{RMS\_inductor} = I_{RMS\_di} + Imin\_RMS$$

$$I_{RMS\_di} = \sqrt{\frac{1}{T} \cdot \int_0^T \left(\frac{t}{T} \cdot di\right)^2 dt} \text{ simplify } \rightarrow I_{RMS\_di} = \frac{\sqrt{3} \cdot \sqrt{di^2}}{3}$$

$$I_{RMS\_di} = \frac{\sqrt{3} \cdot \sqrt{di^2}}{3}$$

$$I_{RMS\_di} = \frac{di}{\sqrt{3}}$$

$$Imin\_RMS = \sqrt{\frac{1}{T} \cdot \int_0^T Imin^2 \, dt} \text{ solve}, Imin\_RMS \rightarrow \sqrt{Imin^2}$$

$$Imin\_RMS = Imin$$

$$I_{RMS\_inductor} = \frac{di}{\sqrt{3}} + Imin$$

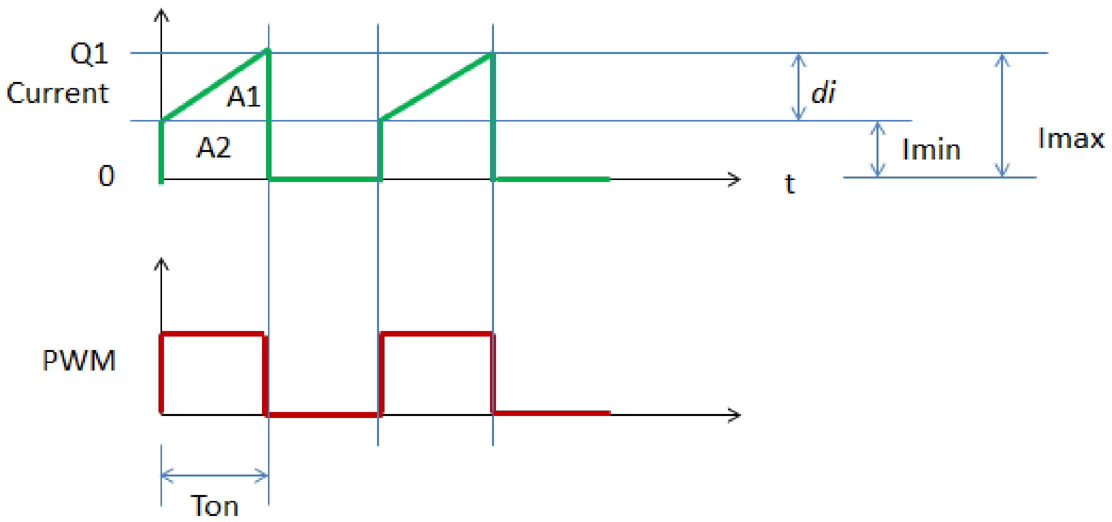
$$I_{RMS\_inductor} = \frac{di}{\sqrt{3}} + Imax - di$$

4. Inductor DC Current Derivation

The next buck converter inductor design formula will be for the DC current. But if you watch carefully at the buck converter schematic, the inductor is in series to the output load. Thus, the DC level of the inductor current is the same as the DC level of the load. This is the easiest derivation in this buck converter design tutorial.

5. Switch RMS Current Derivation

The switch on the buck converter is could be a BJT, MOSFET or IGBT. In this tutorial let us use MOSFET as it is the most popular one in low to medium-power applications. The current waveform of the MOSFET looks like below.



The RMS current of Q1 is the sum of the RMS of areas A1 and A2. A1 is a triangle while A2 is a rectangle.

RMS of Area A1

$$I_{RMS\_A1} = \sqrt{\frac{1}{T}\int_0^{Ton}\left(\frac{di\cdot t}{Ton}\right)^2 dt} \text{ simplify } \rightarrow I_{RMS\_A1} = \frac{\sqrt{\frac{3\cdot Ton\cdot di^2}{T}}}{3}$$

$$I_{RMS\_A1} = \frac{\sqrt{\frac{3\cdot Ton\cdot di^2}{T}}}{3} \text{ substitute}, Ton = D\cdot T \rightarrow I_{RMS\_A1} = \frac{\sqrt{3}\cdot\sqrt{D\cdot di^2}}{3}$$

$$I_{RMS\_A1} = \frac{\sqrt{3}\cdot\sqrt{D\cdot di^2}}{3}$$

$$I_{RMS\_A1} = di\cdot\sqrt{\frac{D}{3}}$$

RMS of Area A2

$$I_{RMS\_A2} = \sqrt{\frac{1}{T}\int_0^{Ton} Imin^2\, dt} \text{ simplify } \rightarrow I_{RMS\_A2} = \sqrt{\frac{Imin^2\cdot Ton}{T}}$$

$$I_{RMS\_A2} = \sqrt{\frac{Imin^2\cdot Ton}{T}} \text{ substitute}, Ton = D\cdot T \rightarrow I_{RMS\_A2} = \sqrt{D\cdot Imin^2}$$

$$I_{RMS\_A2} = \sqrt{D\cdot Imin^2}$$

$$I_{RMS\_A2} = Imin\cdot\sqrt{D}$$

So, the RMS of the switch current will be

$$I_{RMS\_Q1} = I_{RMS\_A1} + I_{RMS\_A2}$$

$$I_{RMS\_Q1} = di\cdot\sqrt{\frac{D}{3}} + (Imax - di)\cdot\sqrt{D}$$

Simplifying to get rid of Imax

$$I_{RMS\_Q1} = \sqrt{D}\cdot Imax - \sqrt{D}\cdot di + \frac{\sqrt{3}\cdot\sqrt{D}\cdot di}{3} \text{ solve}, Imax \rightarrow \frac{I_{RMS\_Q1} + \sqrt{D}\cdot di - \frac{\sqrt{3}\cdot\sqrt{D}\cdot di}{3}}{\sqrt{D}}$$

$$Imax = \frac{I_{RMS\_Q1} + \sqrt{D}\cdot di - \frac{\sqrt{3}\cdot\sqrt{D}\cdot di}{3}}{\sqrt{D}} \text{ simplify } \rightarrow Imax = di - \frac{\sqrt{3}\cdot di}{3} + \frac{I_{RMS\_Q1}}{\sqrt{D}}$$

$$Imax = di - \frac{\sqrt{3}\cdot di}{3} + \frac{I_{RMS\_Q1}}{\sqrt{D}}$$

$$I_{RMS\_Q1} = di\cdot\sqrt{\frac{D}{3}} + (Imax - di)\cdot\sqrt{D} \text{ simplify } \rightarrow I_{RMS\_Q1} = \sqrt{D}\cdot\left(Imax - di + \frac{\sqrt{3}\cdot di}{3}\right)$$

$$I_{RMS\_Q1} = di\cdot\sqrt{\frac{D}{3}} + \left(I_{Load} + \frac{di}{2} - di\right)\cdot\sqrt{D} \text{ simplify } \rightarrow I_{RMS\_Q1} = \sqrt{D}\cdot\left(I_{Load} - \frac{di}{2} + \frac{\sqrt{3}\cdot di}{3}\right)$$

$$I_{RMS\_Q1} = \sqrt{D}\cdot\left(I_{Load} - \frac{di}{2} + \frac{\sqrt{3}\cdot di}{3}\right)$$

6. Switch DC Current Derivation

RMS current of the MOSFET is always higher than the DC current and it is the value to use in computing the power dissipation to get the worst case. However, the DC level is may be needed for whatever reason a designer to come up. So, let us include it in this buck converter design tutorial.

The total DC level is also the sum of the DC level of A1 and A2 in the above waveform.

$$I_{DC\_A1} = \frac{1}{Tsw} \cdot \int_0^{Ton} \frac{di \cdot t}{Ton}\, dt \text{ simplify } \rightarrow I_{DC\_A1} = \frac{Ton \cdot di}{2 \cdot Tsw}$$

$$I_{DC\_A1} = \frac{Ton \cdot di}{2 \cdot Tsw} \text{ substitute}, Ton = D \cdot Tsw \rightarrow I_{DC\_A1} = \frac{D \cdot di}{2}$$

$$I_{DC\_A1} = \frac{D \cdot di}{2}$$

$$I_{DC\_A2} = \frac{1}{Tsw} \int_0^{Ton} Imin\, dt \text{ simplify } \rightarrow I_{DC\_A2} = \frac{Imin \cdot Ton}{Tsw}$$

$$I_{DC\_A2} = \frac{Imin \cdot Ton}{Tsw} \text{ substitute}, Ton = D \cdot Tsw \rightarrow I_{DC\_A2} = D \cdot Imin$$

$$I_{DC\_A2} = D \cdot Imin$$

$$I_{DC\_A2} = D \cdot (Imax - di)$$

$$I_{DC\_total} = \frac{D \cdot di}{2} + D \cdot (Imax - di)$$

Rewriting the equation to exclude Imax

$$I_{DC\_total} = I_{DC\_A1} + I_{DC\_A2}$$

$$I_{DC\_total} = \frac{D \cdot di}{2} + D \cdot (Imax - di) \text{ simplify } \rightarrow I_{DC\_total} = -\frac{D \cdot (di - 2 \cdot Imax)}{2}$$

$$I_{DC\_total} = -\frac{D \cdot (di - 2 \cdot Imax)}{2}$$

$$I_{DC\_total} = \frac{D \cdot (2 \cdot Imax - di)}{2}$$

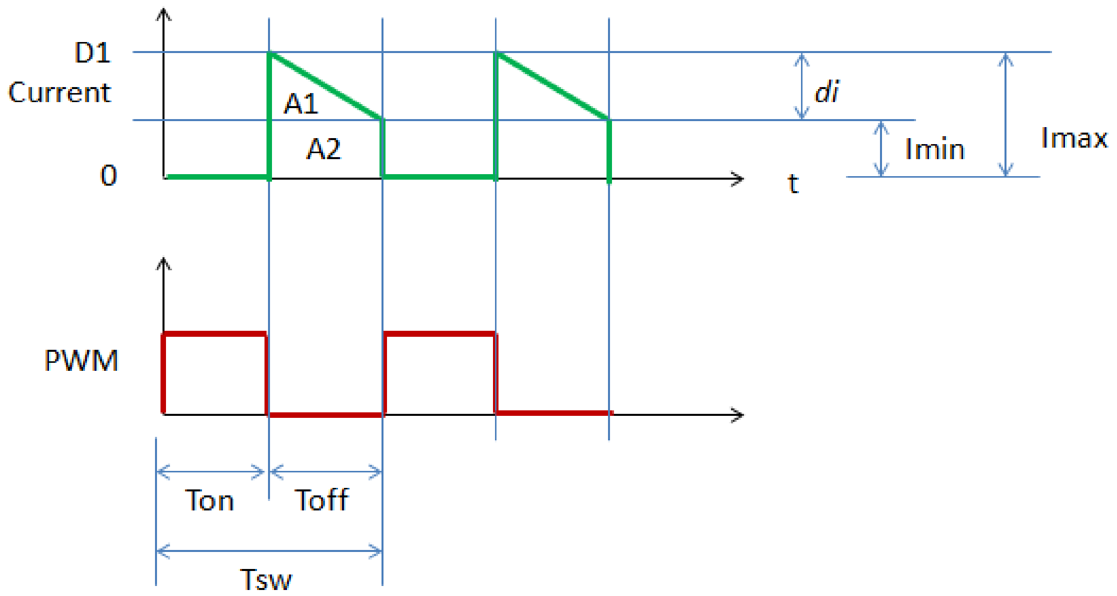
$$Imax = I_{load} + di - \frac{D \cdot di}{2}$$ from inductor current derivation

$$I_{DC\_total} = -\frac{D \cdot \left[di - 2 \cdot \left(I_{load} + di - \frac{D \cdot di}{2} \right) \right]}{2} \text{ simplify } \rightarrow I_{DC\_total} = \frac{D \cdot \left(2 \cdot I_{load} + di - D \cdot di \right)}{2}$$

$$I_{DC\_total} = \frac{D \cdot \left(2 \cdot I_{load} + di - D \cdot di \right)}{2}$$

7. Diode RMS Current Derivation

Referring to below waveform, we can calculate the RMS current of the diode. The diode will conduct only when the MOSFET is not conducting.



$$I_{RMS\_A1} = \sqrt{\frac{1}{T}\cdot\int_{Ton}^{T}\left(\frac{t - Ton}{Toff}\right)^2 \cdot di^2\, dt} \text{ simplify } \rightarrow I_{RMS\_A1} = \frac{\sqrt{\frac{3\cdot di^2\cdot(T - Ton)^3}{T\cdot Toff^2}}}{3}$$

$$I_{RMS\_A1} = \frac{\sqrt{\frac{3\cdot di^2\cdot(T - Ton)^3}{T\cdot Toff^2}}}{3} \text{ substitute}, Toff = T - Ton \rightarrow I_{RMS\_A1} = \frac{\sqrt{3}\cdot\sqrt{\frac{di^2\cdot(T - Ton)}{T}}}{3}$$

$$I_{RMS\_A1} = \frac{\sqrt{3}\cdot\sqrt{\frac{di^2\cdot(T - Ton)}{T}}}{3} \text{ substitute}, Ton = D\cdot T \rightarrow I_{RMS\_A1} = \frac{\sqrt{3}\cdot\sqrt{-di^2\cdot(D - 1)}}{3}$$

$$I_{RMS\_A1} = \frac{\sqrt{3}\cdot\sqrt{-di^2\cdot(D - 1)}}{3}$$

$$I_{RMS\_A1} = di\cdot\sqrt{\frac{1 - D}{3}}$$

$$I_{RMS\_A2} \equiv \sqrt{\frac{1}{Tsw}\cdot\int_{Ton}^{Tsw} Imin^2\,dt} \text{ simplify } \rightarrow I_{RMS\_A2} \equiv \sqrt{-\frac{Imin^2\cdot(Ton - Tsw)}{Tsw}}$$

$$I_{RMS\_A2} \equiv \sqrt{-\frac{Imin^2\cdot(Ton - Tsw)}{Tsw}} \text{ substitute}, Ton \equiv D\cdot Tsw \rightarrow I_{RMS\_A2} \equiv \sqrt{-Imin^2\cdot(D-1)}$$

$$I_{RMS\_A2} \equiv Imin\cdot\sqrt{(1-D)}$$

$$I_{RMS\_D1} \equiv I_{RMS\_A1} + I_{RMS\_A2}$$

$$I_{RMS\_D1} \equiv di\cdot\sqrt{\frac{1-D}{3}} + (Imax - di)\cdot\sqrt{1-D}$$

$$Imax\_D1 \equiv I_{RMS\_D1} + di\cdot\left(1 - \sqrt{\frac{1-D}{3}}\right)$$

$$I_{RMS\_D1} \equiv di\cdot\sqrt{\frac{1-D}{3}} + (Imax - di)\cdot\sqrt{1-D} \text{ simplify } \rightarrow I_{RMS\_D1} \equiv \sqrt{1-D}\cdot\left(Imax - di + \frac{\sqrt{3}\cdot di}{3}\right)$$

$$I_{RMS\_D1} \equiv \sqrt{1-D}\cdot\left(Imax - di + \frac{\sqrt{3}\cdot di}{3}\right)$$

$$I_{RMS\_D1} \equiv \sqrt{1-D}\cdot\left[Imax - di\cdot\left(1 - \frac{\sqrt{3}}{3}\right)\right]$$

$$Imax \equiv I\_load + \frac{di}{2}$$

$$I_{RMS\_D1} \equiv \sqrt{1-D}\cdot\left[I\_load + \frac{di}{2} - di\cdot\left(1 - \frac{\sqrt{3}}{3}\right)\right] \text{ simplify } \rightarrow I_{RMS\_D1} \equiv \sqrt{1-D}\cdot\left(I\_load - \frac{di}{2} + \frac{\sqrt{3}\cdot di}{3}\right)$$

$$I_{RMS\_D1} \equiv \sqrt{1-D}\cdot\left(I\_load - \frac{di}{2} + \frac{\sqrt{3}\cdot di}{3}\right)$$

8. Diode DC Current Derivation

We will still use above waveform in the determination of the DC current of the diode.

$$I_{DC\_A1} = \frac{1}{T}\cdot\int_{Ton}^{T}\left(\frac{t-Ton}{Toff}\right)\cdot di\,dt \text{ simplify } \rightarrow I_{DC\_A1} = \frac{di\cdot(T-Ton)^2}{2\cdot T\cdot Toff}$$

$$I_{DC\_A1} = \frac{di\cdot(T-Ton)^2}{2\cdot T\cdot Toff} \text{ substitute}, Toff = T - Ton \rightarrow I_{DC\_A1} = \frac{T\cdot di - Ton\cdot di}{2\cdot T}$$

$$I_{DC\_A1} = \frac{T\cdot di - Ton\cdot di}{2\cdot T} \text{ substitute}, Ton = D\cdot T \rightarrow I_{DC\_A1} = -\frac{di\cdot(D-1)}{2}$$

$$I_{DC\_A1} = -\frac{di\cdot(D-1)}{2}$$

$$I_{DC\_A1} = \frac{di\cdot(1-D)}{2}$$

$$I_{DC\_A2} = \frac{1}{Tsw}\cdot\int_{Ton}^{Tsw} Imin\,dt \text{ simplify } \rightarrow I_{DC\_A2} = -\frac{Imin\cdot(Ton-Tsw)}{Tsw}$$

$$I_{DC\_A2} = -\frac{Imin\cdot(Ton-Tsw)}{Tsw} \text{ substitute}, Ton = D\cdot Tsw \rightarrow I_{DC\_A2} = -Imin\cdot(D-1)$$

$$I_{DC\_A2} = Imin\cdot(1-D)$$

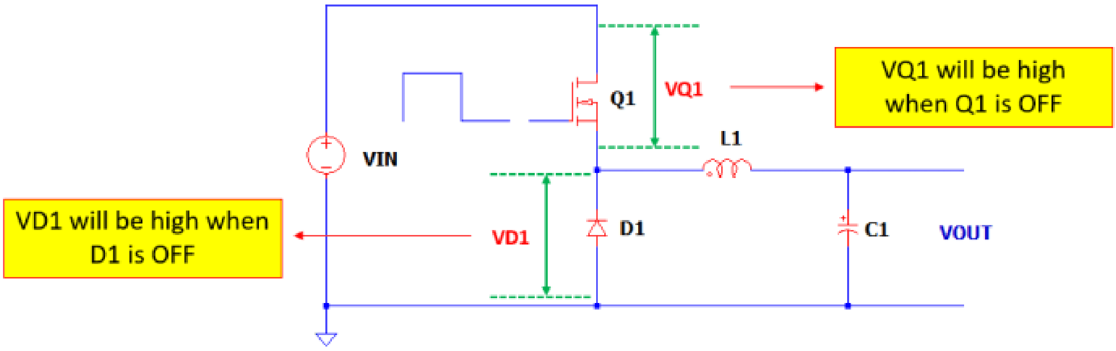
$$I_{DC\_A2} = (Imax - di)\cdot(1-D)$$

$$I_{DC\_D1} = \frac{di\cdot(1-D)}{2} + (Imax - di)\cdot(1-D)$$

$$I_{DC\_D1} = \frac{di\cdot(1-D)}{2} + \left(I\_load + \frac{di}{2} - di\right)\cdot(1-D) \text{ simplify } \rightarrow I_{DC\_D1} = -I\_load\cdot(D-1)$$

$$I_{DC\_D1} = I\_load\cdot(1 - Dutycycle)$$

9. Switch and Diode Voltage Derivation



$$VQ1\ max = VIN\ max + VSpike$$

Vspike is due to the parasitic inductance and it can be assumed to be 40-70% of VIN.

$$VD1\ max = VIN\ max + Vspike$$

Vspike is due to the parasitic inductance and it can be assumed to be 50-120% of VIN.

10. Switch Power Losses Derivation

The switch power losses are composed of two factors. The first is conduction loss and the second is switching loss. Conduction loss is due to the fixed voltage drop on the switch while the switching loss is due to the switching action of the switch. In this tutorial, we emphasize using a MOSFET. So, the below equations are valid for the MOSFET.

Conduction Loss

$$Pconduction = I_{RMS}^{2} \cdot RDSon$$

Switching Loss

$$Ploss\_gatecharge = \frac{1}{2} Qgtotal \cdot Vdrive \cdot Fsw$$

$$Ploss\_Coss = \frac{1}{2} \cdot Coss \cdot Vmax\_FET^{2} \cdot Fsw$$

$$Ploss\_trise\_tfall = \frac{1}{2} \cdot (trise + tfall) \cdot I_{RMS} \cdot Vdrive \cdot Fsw$$

Total MOSFET Power Loss

$$Ploss\_total = Pconduction + Ploss\_gatecharge + Ploss\_Coss + Ploss\_trise\_tfall$$

Where;

RDSon - drain to source on state resistance of the MOSFET
Qgtotal - total gate charge of the MOSFET
Vdrive - voltage applied to the MOSFET gate
Fsw - switching frequency
Coss - output capacitance of the MOSFET
Vmax_FET - drain voltage of the MOSFET when open base
trise - rise time of the MOSFET
tfall - fall time of the MOSFET

11. Switch Power Stress and Thermal Considerations

Power stress of the switch is just actual power dissipation divided by power capability.

$$P_{stress} = P_{dissipation\ actual} / P_{dissipation\ capability}$$

Power dissipation capability is can be derived from the datasheet information.

For without heatsink (the switch is not mounted on a heatsink):

$$P_{dissipation\ capability} = (T_{jmax} – T_{amax}) / R_{thjc}$$

Where;

Tjmax – maximum junction temperature of the device

Tamax – maximum ambient temperature of operation

Rthjc – thermal resistance from junction to case

In case needed to compute for the device actual junction temperature, it can be done as below:

$$T_{jactual} = (P_{dissipation\ capability} \times R_{thjc}) + T_{amax}$$

For with heatsink (the switch is mounted on a heatsink):

$$P_{dissipation\ capability} = (T_{jmax} – T_{cmax}) / (R_{thjc} + R_{thchs} + R_{thhsa})$$

Where;

Tjmax – maximum junction temperature of the device

Tcmax – maximum allowed case temperature

Rthjc – thermal resistance from junction to case

Rthchs – thermal resistance from case to heatsink. This is the thermal resistance of the material that bond the heatsink and the case.

Rthhsa – thermal resistance from heatsink to air. This actually the thermal resistance of the heatsink used.

The actual device junction temperature is can be computed as:

Tjactual = [Pdissipation capability X (Rthjc + Rthchs +Rthhsa)] + Tcmax

12. Diode Power Losses Derivation

$$Ploss\ diode = Irms \times VF$$

Where;

VF - diode forward voltage

Irms - RMS current to the diode

13. Diode Power Stress and Thermal Considerations

Power stress of the diode is just actual power dissipation divided by power capability.

Pstress = Pdissipation actual / Pdissipation capability

Power dissipation capability is can be derived from the datasheet information.

For without heatsink (the diode is not mounted on a heatsink):

Pdissipation capability = (Tjmax – Tamax) / Rthjc

Where;

Tjmax – maximum junction temperature of the device

Tamax – maximum ambient temperature of operation

Rthjc – thermal resistance from junction to case

In case needed to compute for the device actual junction temperature, it can be done as below:

$$Tjactual = (Pdissipation\ capability \times Rthjc) + Tamax$$

For with heatsink (the diode is mounted on a heatsink):

$$Pdissipation\ capability = (Tjmax - Tcmax) / (Rthjc + Rthchs + Rthhsa)$$

Where;

Tjmax – maximum junction temperature of the device

Tcmax – maximum allowed case temperature

Rthjc – thermal resistance from junction to case

Rthchs – thermal resistance from case to heatsink. This is the thermal resistance of the material that bond the heatsink and the case.

Rthhsa – thermal resistance from heatsink to air. This actually the thermal resistance of the heatsink used.

The actual device junction temperature is can be computed as:

$$Tjactual = [Pdissipation\ capability \times (Rthjc + Rthchs + Rthhsa)] + Tcmax$$

14. Inductor Power Losses Derivation

The power loss of the inductor is composed of two parts: DC and AC losses. In low switching frequency and low power, AC loss is small and thus simply not included in the calculation. But for very high switching frequency, you can assume a switching loss almost the same to the DC loss. DC losses is also sometimes referred to copper loss while switching loss is also called core loss.

DC loss

$$Ploss\_DC = Irms^2 \cdot DCR$$

Where;

Ploss_DC - loss of the inductor due to DC resistance
Irms - RMS current to the inductor
DCR - DC resistance of the inductor

15. Output Capacitor Selection

Below output capacitance (C1) calculation is generic. However, specific controllers may have their own equation to derive the value of the output capacitance as this has something to do with the loop compensation. Considering no effect of ESR, equation below is can be used to determine the size of output capacitor.

$$C1 = di / (Fsw \times Vripple)$$

For electrolytic capacitors the ESR is huge, so it needs to consider it in the analysis. The calculated capacitance above should have an ESR of not higher than below equation.

$$ESR = Vripple / di$$

Where;

ESR – equivalent series resistance

di – inductor ripple current

Fsw – switching frequency

Vripple – allowable output ripple voltage

Ripple Current

The selected output capacitor should have a ripple current rating of higher than the result of below equation.

$$irip\_cap = \sqrt{Irms\_inductor^2 - I\_load^2}$$

Where;

Irms_inductor – inductor RMS current
I_load – load current

16. Buck Converter Efficiency Equation Derivation

Buck converter efficiency is can be computed using below equation.

Efficiency = (Pout / Pin) X 100%

Pout = Iout X Vout

Pin = Pout + Ploss total

Efficiency = [Iout X Vout / (Pout + Ploss total)] X 100%

Where;

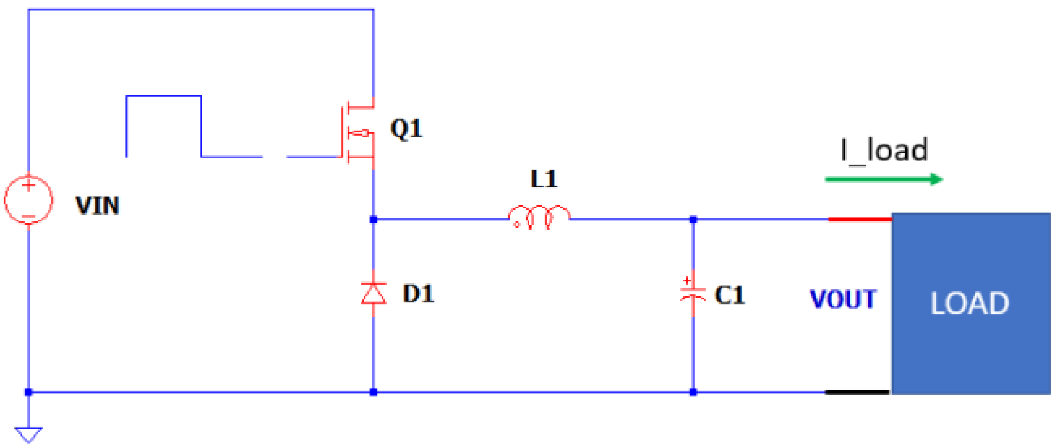
Iout – load current

Vout – output voltage

Pout – total power losses

17. Buck Converter Design Tutorial – Sample Design with Parts Selection

We are done with all the necessary equations. Let us apply this buck converter design tutorial to actual design scenario.



Given Values:

Below are the minimum given to start the calculations.

Vin := 24V	Input voltage
Vout := 12V	Output voltage
I_load := 10A	Load current
VQ1 := 0.1V	Estimated on state voltage drop of the MOSFET Q1
VF := 0.7V	Forward voltage drop of the diode. this can be replaced by MOSFET on state voltage if a synchronous buck converter is used.
Fsw := 300kHz	Switching frequency
Vout_ripple := 240mV	Output ripple voltage

Designer's Call

%inductor_ripple := 10%	The design engineer will set the amount of inductor ripple current to select the inductance. Usual value is ranging from 10%-50% of the load current. Too low ripple current needs bigger inductor. High ripple current requires smaller inductor size but sacrifices efficiency and may require expensive MOSFETs and diode.

By setting the %inductor_ripple to 100% means the converter operation is in the transition mode or boundary mode.

By setting the %inductor_ripple to 100% means the converter operation is in the transition mode or boundary mode. But in this sample design we will set to 10% only that means a CCM operation.

Dutycycle Calculation

$$\text{Dutycycle} := \frac{\text{Vout} + \text{VF}}{\text{Vin} - \text{VQ1} + \text{VF}}$$

$$\text{Dutycycle} = 51.626\cdot\%$$

Where;

Vout - output voltage
VF - diode voltage drop or MOSFET on state voltage for synchronous buck converter
Vin - input voltage
VQ1 -MOSFET on state voltage drop

Inductance Calculation

For very detailed explanations on how the inductor of a buck converter derived, read the tutorial http://electronicsbeliever.com/sizing-the-inductor-of-buck-converter-and-setting-its-operation/

$$\text{L1} := -\frac{\text{Dutycycle}\cdot\frac{1}{\text{Fsw}}\cdot(\text{VQ1} - \text{Vin} + \text{Vout})}{\text{\%inductor\_ripple}\cdot\text{I\_load}}$$

$$\text{L1} = 20.478\cdot\mu\text{H}$$

This is the theoritical value of the inductance based from the **%inductor_ripple** assumed. Choose a standard value near to this.

$$\text{L1\_selected} := 22\mu\text{H}$$

This is the selected inductance and this will be used to continue the calculations.

Inductor Ripple Current Derivation

$$\text{di} := -\frac{\text{Dutycycle}\cdot\frac{1}{\text{Fsw}}\cdot(\text{VQ1} - \text{Vin} + \text{Vout})}{\text{L1\_selected}}$$

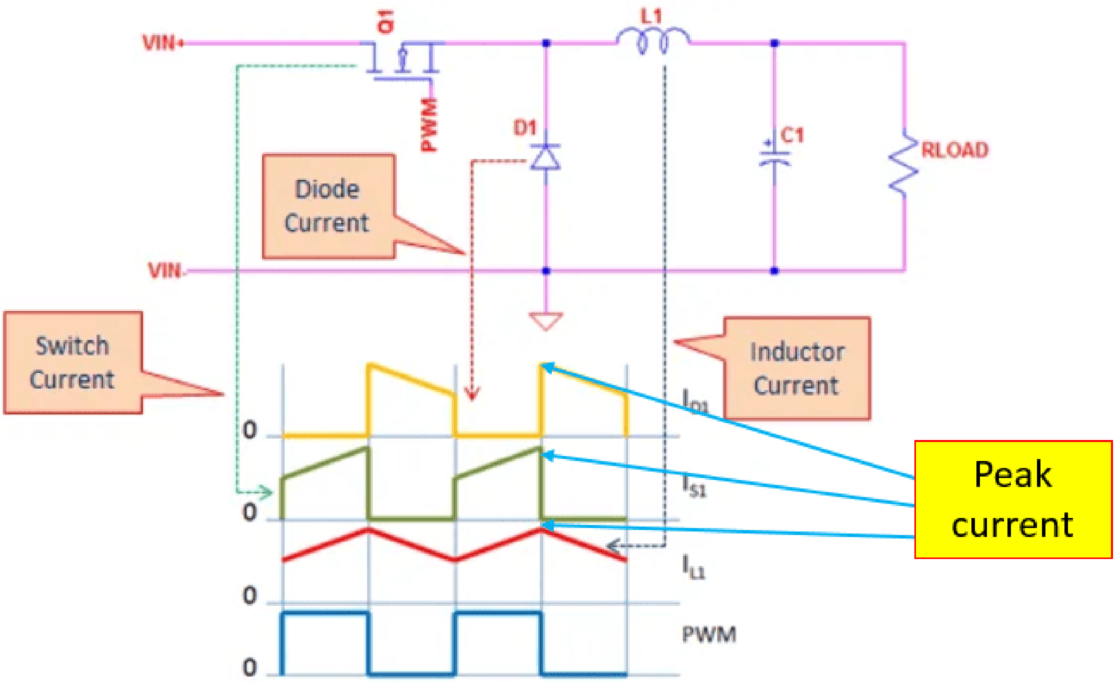
$$\text{di} = 0.931\cdot\text{A}$$

Where;

di - inductor ripple current
Fsw - switching frequency
VQ1 - MOSFET on state voltage drop
Vin - input voltage
Vout - output voltage
L1_selected - selected inductance value

Peak Current Calculation

MOSFET Q1, Diode D1 and the inductor L1 will have the same peak current.



$$Imax := I\_load + \frac{di}{2}$$

$Imax = 10.465\,A$

This is the level of the peak currents seen by the diode, MOSFET and inductor considering no effect of noise.

Inductor RMS Current

(Note: the DC value of the inductor current is the same to the load current.)

$$Irms\_inductor := \frac{di}{\sqrt{3}} + Imax - di$$

$Irms\_inductor = 10.072\,A$

Design Note 1: *Choose an inductor that has the value of L1_selected, with a RMS current rating higher than Irms_inductor and a saturation current rating higher than Imax.*

Inductor Power Losses

$DCR := 0.05\Omega$ DC resistance of the inductor

$$Ploss\_inductor := DCR \cdot Irms\_inductor^2$$

$Ploss\_inductor = 5.072\,W$ Inductor power loss

MOSFET Q1 RMS and DC Current

$$Irms\_Q1 := di \cdot \sqrt{\frac{Dutycycle}{3}} + (Imax - di) \cdot \sqrt{Dutycycle}$$

$Irms\_Q1 = 7.237\,A$

$$Idc\_Q1 := \frac{Dutycycle \cdot di}{2} + Dutycycle \cdot (Imax - di)$$

$Idc\_Q1 = 5.163\,A$

Design Note 2: *Select a MOSFET with an RMS current or DC current higher than Irms_Q1. The peak current rating must be higher than Imax. The selected MOSFET should have a voltage rating higher than the maximum input voltage. The rule of thumb is to select a voltage rating of twice the maximum input voltage. For instance, a MOSFET of 30V rating is can be used to a maximum input voltage of 12V.*

MOSFET Q1 Power Losses

To know the power losses, below information must be known:

$RDSon := 0.0094\Omega$	this is the typical RDSon of the MOSFET at 25'C centigrade.
$RDSon\_norm := 1.5$	This is the normalized RDSon of the MOSFET at junction temperature of interest. For instance, for high power applications, MOSFET junction temperature will be 100'C usually, so get the normalized value at 100'C.

In case the MOSFET selected specifies a RDSon value at desired junction temperature of interest (say at 100'C), just put 1 to **RDSon_norn** above for the template to work correctly.

$Qgtotal := 110nC$	This is the total gate charge specification of the MOSFET. Use may use the max value for worst case or the typical value for typical result.
$Coss := 420pF$	COSS specification of the MOSFET. Use may use the max value for worst case or the typical value for typical result.
$trise := 79ns$	rise time of the MOSFET. Use may use the max value for worst case or the typical value for typical result.
$tfall := 45ns$	fall time of the MOSFET. Use may use the max value for worst case or the typical value for typical result.
$Vdrive := 12V$	Voltage applied to the gate of the MOSFET
$Vmax\_FET := 24V$	Drain voltage of the MOSFET. Idealy, just equal to the maximum input voltage.

Conduction Loss

$$Pconduction\_Q1 := Irms\_Q1^2 \cdot RDSon \cdot RDSon\_norm$$

$Pconduction\_Q1 = 0.738\,W$ loss due to on state resistance

Switching Loss

$$Ploss\_gatecharge := \frac{1}{2} Qgtotal \cdot Vdrive \cdot Fsw$$

$Ploss\_gatecharge = 0.198\,W$ loss due to gate charge

$$Ploss\_Coss := \frac{1}{2} \cdot Coss \cdot Vmax\_FET^2 \cdot Fsw$$

$Ploss\_Coss = 0.036\,W$ loss due to COSS

$$Ploss\_trise\_tfall := \frac{1}{2} \cdot (trise + tfall) \cdot Irms\_Q1 \cdot Vdrive \cdot Fsw$$

$Ploss\_trise\_tfall = 1.615\,W$ loss due to rise and fall times

Total Power Loss of Q1

$$Ploss\_total\_Q1 := Pconduction\_Q1 + Ploss\_gatecharge + Ploss\_Coss + Ploss\_trise\_tfall$$

$Ploss\_total\_Q1 = 2.588\,W$ total losses of Q1

MOSFET Q1 Power Capability Without Heatsink

To know if the selected MOSFET Q1 is able to handle the **Ploss_total_Q1** above, the following information should be known.

$Tj\_max := 175\Delta°C$ Maximum junction temperature of the MOSFET

$Tamb := 50\Delta°C$ Maximum ambient or surrounding temperature of the MOSFET

$Tc\_max := 100\Delta°C$ Maximum case temperature allowed

$Rthja := 60\,\frac{\Delta°C}{W}$ Thermal resistance from junction to ambient of the selected MOSFET. This is the one to use if the MOSFET is not attached to a heat sink and no air cooling.

$Rthjc := 10\,\frac{\Delta°C}{W}$ Thermal resistance from junction to case. This is the one to use when the MOSFET is intended to attach on a heat sink.

Compute for the MOSFET Q1 Power Capability

Without heat sink and natural cooling

$$Pcapability\_Q1\_without\_heatsink := \frac{Tj\_max - Tamb}{Rthja}$$

$Pcapability\_Q1\_without\_heatsink = 2.083\,W$ This is the maximum power capability of the MOSFET. Operation above this will damage the MOSFET.

Diode D1 RMS and DC Current

$$Irms\_diode := \sqrt{1 - Dutycycle}\cdot\left(Imax - di + \frac{\sqrt{3}\cdot di}{3}\right)$$

$Irms\_diode = 7.005\ A$

$$Idc\_diode := \frac{di\cdot(1 - Dutycycle)}{2} + (Imax - di)\cdot(1 - Dutycycle)$$

$Idc\_diode = 4.837\ A$

Design Note 3: *The selected diode should have current continuous current rating higher than* ***Irms_diode. The*** *peak current rating must be higher than* ***Imax.*** *The peak inverse voltage rating of the diode must be higher than the maximum input voltage. A diode of 50V is suitable to an input voltage of up to 24V for instance.*

Diode D1 Power Loss

$VF = 0.7\ V$

$Ploss\_D1 := VF\cdot Irms\_diode$

$Ploss\_D1 = 4.904\ W$

Where;

VF - diode forward voltage

Irms_diode - RMS current to the diode. This is the same RMS current to Q2 if a synchronous buck converter is used as Figure 2 above.

Diode D1 Power Capability without Heatsink

$$Pcapability\_D1\_without\_heatsink := \frac{Tj\_max\_D1 - Tamb}{Rthja\_D1}$$

$Pcapability\_D1\_without\_heatsink = 2.083\ W$

This is the maximum power capability of the diode D1. Operation above this will damage the diode.

$$PowerStress\_D1\_without\_heatsink := \frac{Ploss\_D1}{Pcapability\_Q1\_without\_heatsink}$$

$PowerStress\_D1\_without\_heatsink = 235.375\cdot\%$

Power stress of D1. For high reliability, this must be lower than 80%.

Diode D1 Power Capability with Heatsink

For with heat sink, additional informations must be known.

$Rthchs\_D1 := 0.1 \frac{\Delta °C}{W}$

This is the thermal resistance from case to heat sink. Basically this is the thermal resistance of the bonding of the diode body and the heat sink. If the bonding is very good, this is negligible.

$Rthhsa\_D1 := 1 \frac{\Delta °C}{W}$

This is actually the thermal resistance of the heat sink used. Get this value from the heat sink datasheet.

$$Pcapability\_D1\_with\_heatsink := \frac{Tj\_max\_D1 - Tc\_max}{Rthjc\_D1 + Rthchs\_D1 + Rthhsa\_D1}$$

$Pcapability\_D1\_with\_heatsink = 6.757\,W$

This is the power capability of the diode D1 with heat sink

$$PowerStress\_D1\_with\_heatsink := \frac{Ploss\_D1}{Pcapability\_D1\_with\_heatsink}$$

$PowerStress\_D1\_with\_heatsink = 72.574 \cdot \%$

This is the power stress of the diode D1 with heat sink. Do not exceed 80% for higher reliability.

Output Capacitor C1 Selection

$$C1 := \frac{di}{Fsw \cdot Vout\_ripple}$$

$C1 = 12.928 \cdot \mu F$

this is the minimum capacitance to use just to meet the required output ripple voltage

Select a standard value capacitor higher than the computed.

$C1\_selected := 22 \mu F$

The selected capacitor should have ESR of not higher than

$$ESR := \frac{Vout\_ripple}{di}$$

$ESR = 0.258\,\Omega$ using ESR higher than this will not meet the required output ripple voltage

Where;

ESR - equivalent series resistance
di - inductor ripple current
Fsw - switching frequency
Vripple - allowable output ripple voltage

The selected output capacitor should have a ripple current rating higher than

$$irip\_cap := \sqrt{Irms\_inductor^2 - I\_load^2}$$

$irip\_cap = 1.202\,A$ select a capacitor with a ripple current rating higher than this value.

Where;

irip_cap - computed RMS ripple current on C1
Irms_inductor - inductor RMS current
I_load - load current

The voltage rating must be higher than the output voltage with enough margin.

Buck Converter Efficiency Calculation

Finally, the buck converter efficiency is

$$Ploss\_total := Ploss\_inductor + Ploss\_total\_Q1 + Ploss\_D1$$

$Ploss\_total = 12.564\,W$

$$Efficiency := \frac{I\_load \cdot Vout}{I\_load \cdot Vout + Ploss\_total}$$

$Efficiency = 90.522 \cdot \%$ This is the buck converter efficiency

Where;

Iout - load current
Vout - output voltage
Ploss_total - total power losses

Operation Mode Checking

A buck converter is can be a CCM, DCM or transition mode. In CCM, the current of the inductor will not touch zero. On the other hand, the current on DCM will go below zero while the current on the transition mode is just exactly at zero modes.

This section tells if the buck converter is operating in CCM or DCM. If you declare a %inductor_ripple on the upper portion of this template less than 100%, the operation of the buck converter is surely CCM.

$$Imin := Imax - di$$

$$OperationMode := \begin{cases} \text{"CCM"} & \text{if } Imin > 0 \\ \text{"Boundary"} & \text{if } Imin = 0 \\ \text{""DCM"} & \text{otherwise} \end{cases}$$

OperationMode = "CCM"

Summary Table

The table shows all the important equations required for the designing of BUCK converter.

S.No	Parameters	Equation
1	Maximum switch current	$\langle I_{switch\_max} \rangle = \langle I_{omax} \rangle . D$
2	Maximum switch voltage	$V_{switch-max} = V_{dcmax}$
3	Peak current rating of inductor	$I_{Lmax} = V_o [\frac{1}{R_{max}} + \frac{(1 - D_{max})}{2Lf}]$
4	Critical Inductance	$L_c = \frac{(1 - D_{max}) R_{max}}{2f}$
5	Peak inverse voltage of diode	$V_{PIV} = V_{PRM} = V_{dcmax}$
6	Current rating of diode	$I_{Fmax} > I_{omax} (1-D_{min})$
7	Minimum capacitance	$C = \frac{(1 - D_{min})}{8Lf^2 \Delta V_o / V_o}$
8	RMS current rating of capacitor	$I_{crms} = \frac{(1 - D_{min}).V_o}{2\sqrt{3}L}$
9	Voltage rating of capacitor	$V_{cmax} = V_o + \Delta V_o/2$