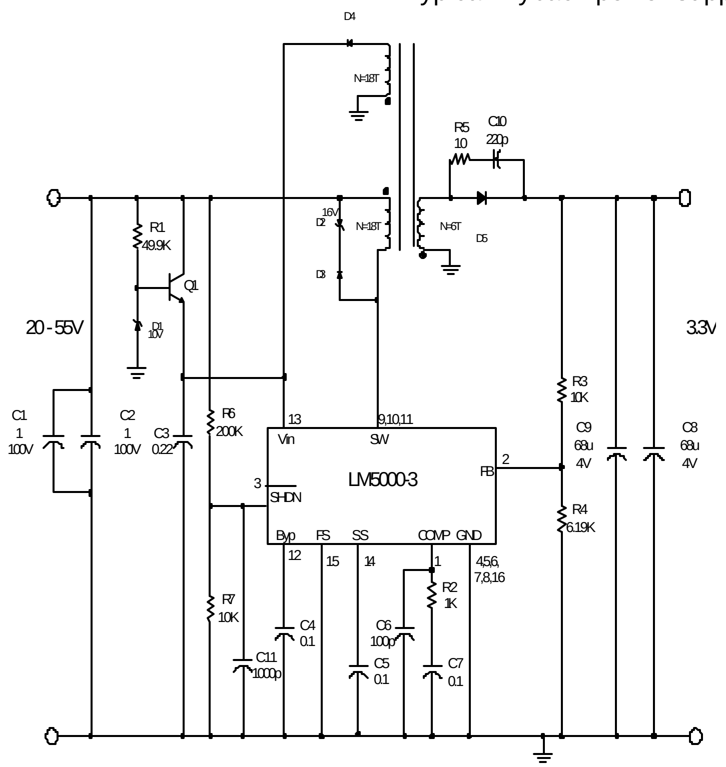
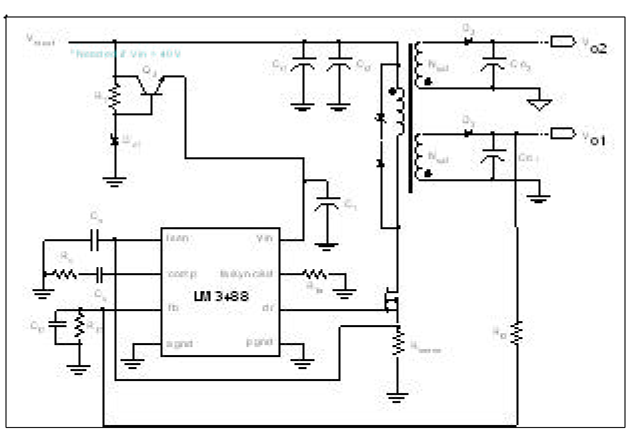
SWITCHING POWER SUPPLY DESIGN: CONTINUOUS MODE FLYBACK CONVERTER

Written by Michele Sclocchi
michele.sclocchi@nsc.com
Application Engineer
National Semiconductor

Typical Flyback power supply:



LM5000 FLYBACK CONVERTER



Notes:

Write down the power supply requirements on : $X_{XX} :=$

Get the results on: $Rsults_{XX} :=$

This Mathcad file helps the calculation of the external components of a typical continuous mode switching power supply.

Input voltage:

- Minimum input voltage: $Vi_{min} := 22 \cdot volt$ $\mu sec := 10^{-6} \cdot sec$

- Maximum input voltage: $Vi_{max} := 55 \cdot volt$

- Nominal input voltage: $Vi_{nom} := 36 \cdot volt$

Output:

- Nominal output voltage, maximum output ripple, minimum output current, maximum output current

$Vo1 := 3.3 \cdot volt$ $Vrp1 := 100 \cdot mV$ $Io1_{min} := 0.250 \cdot amp$ $Io1_{max} := 2. \cdot amp$

$Vo2 := 0 \cdot volt$ $Vrp2 := 120 \cdot mV$ $Io2_{min} := 0.000 \cdot amp$ $Io2_{max} := 0.000 \cdot amp$

$Vd_{fw} := 0.5 \cdot volt$ (diode's forward drop voltage)

$Po_{min} := (Vo1 + Vd_{fw}) \cdot Io1_{min} + (Vo2 + Vd_{fw}) \cdot Io2_{min}$ $Po_{min} = 0.95\,watt$

$Po_{max} := (Vo1 + Vd_{fw}) \cdot Io1_{max} + (Vo2 + Vd_{fw}) \cdot Io2_{max}$ $Po_{max} = 7.6\,watt$

- Switching Frequency: $fsw := 300 \cdot kHz$ $T := \frac{1}{fsw}$ $T = 3.33\,\mu sec$

- Transformer's Efficiency: $\eta := 0.90$ (Guessed value)

- Maximum drop voltage across the switching mosfet during the on time:

- On resistance of the Mosfet: $Rds_{on} := 0.180 \cdot ohm$

$$Vds_{on} := \frac{Po_{max}}{\eta \cdot Vi_{min}} \cdot Rds_{on} \qquad Vds_{on} = 0.07\,volt$$

1) Define Primary/secondary turns ratio: Nps1

Primary/secondary turns ratio can be selected as compromise between maximum voltage across the switching mosfet and desired max.-min. duty cycle.

- Nominal desired on Duty Cycle: $D_{nom} := 0.24$

$$Nps1 := \left(\frac{Vi_{nom} - Vds_{on}}{Vo1 + Vd_{fw}} \right) \cdot \frac{D_{nom}}{1 - D_{nom}} \qquad Nps1 = 3$$

The calculated turns ratio can be modified to optimise the windings

- Flyback voltage across the mutual inductance during the off time: Vfm

$Vfm := Nps1 \cdot (Vo1 + Vd_{fw})$ $Vfm = 11.35\,volt$

- Maximum voltage across the switching-mosfet:

$F_{spike} := 0.15$

$Vds_{max} := (F_{spike} + 1) \cdot (Vi_{max} + Vfm)$ $Vds_{max} = 76.3\,volt$

Safe factor (assume spikes of 20-30% of Vdc)

To reduce the maximum voltage across the switching mosfet reduce Nps turns ratio by reducing the desired on-duty cycle

- Slave output turns ratio:

$$Nps2 := \frac{Vfm}{Vo2 + Vd_{fw}} \qquad Nps2 = 22.7$$

2) Maximum and minimum duty cycle : Dmax and Dmin

To maintain the continuous mode of operation the dead time has to be equal zero (Ton+Toff = T), and to reset the core every cycle, the average voltage on the primary inductance must be equal zero: (Vi - Vds) * Ton = (Vo + Vd) * Nps * Toff, where Toff is equal to (T - Ton)

$$Ton_{max} := \frac{Vfm \cdot T}{(Vi_{min} - Vds_{on}) + Vfm} \qquad Ton_{max} = 1.14\,\mu sec$$

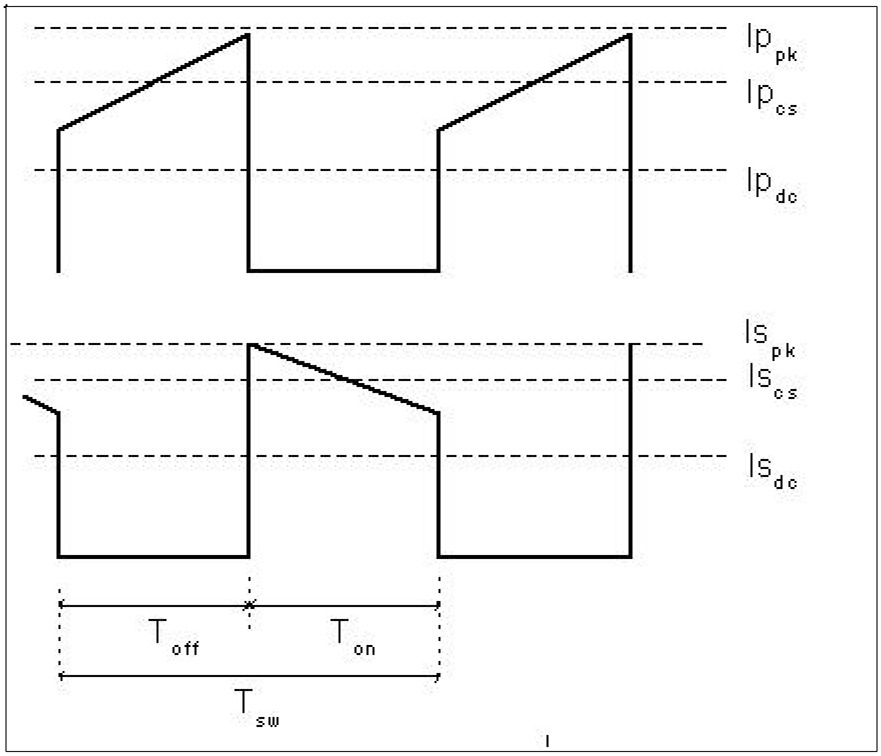
$$Ton_{min} := \frac{Vfm \cdot T}{(Vi_{max} - Vds_{on}) + Vfm}$$ $Ton_{min} = 0.57\,\mu sec$

Maximum duty cycle $D_{max} := \frac{Ton_{max}}{T}$ $D_{max} = 0.34$

Minimum duty cycle $D_{min} := \frac{Ton_{min}}{T}$ $D_{min} = 0.17$

3) Primary winding: Inductance, peak, AC, RMS current

In continuous mode the duty cycle changes with a change of input voltage. An increase of output current, will temporary increase the duty cycle until the average primary and secondary currents increase.



$$Ip_{cs} := \frac{Po_{max}}{(Vi_{min} - Vds_{on}) \cdot \eta \cdot D_{max}}$$ $Ip_{cs} = 1.13\,amp$

- Primary average current:

There are several criterias to select the primary and secondary inductances, following are explained two different solutions: the first one is to select the primary inductance in order to insure continuous mode of operation from full load to minimum load. (about 1/10-1/20 of the maximum load). **(3-a)**,

The second alternative criteria, is to calculate primary and secondary inductances by defining maximum secondary ripple current. **(3-b)**

3-a) Select primary inductance for continuous mode of operation at minimum load:

During the transition from discontinuous to continuous mode, the peak primary current it's about double the central average current Ipcs(min) .In order to maintain continuous mode at minimum load the maximul ramp amplitude has to be twice the minimum average current.

- Ramp amplitude:

$$\Delta Ip_a := \frac{2 \cdot Po_{min}}{\left(Vi_{min} - Vds_{on}\right) \cdot \eta \cdot D_{max}} \qquad \Delta Ip_a = 0.28\,amp$$

- Primary inductance: dIp= (Vi-Vds)*Ton/Lp

$$Lp_a := \frac{\left(Vi_{min} - Vds_{on}\right) \cdot Ton_{max}}{\Delta Ip_a} \qquad Lp_a = 88.29\,\mu H$$

3-b) Primary and secondary inductance for a maximum defined secondary peak to peak ripple current:

AC core losses, AC winding losses, and output ripple current are directly proportional to the current ramp amplitude of the primary and secondaries windings. Therefore in high current application, AC ripple currents could have a predominant role on the overall performance of the converter, a good compromise between transformer's size and AC currents can be obtained by selecting the most appropriate secondary ripple current:

- Desired secondary ripple current:

$\Delta Is\% := 30 \cdot \%$ (maximum value / average)

$$Is1_{cs} := \frac{Io1_{max}}{\left(1 - D_{max}\right)} \qquad Is1_{cs} = 3.03\,amp$$

- Ramp amplitude:

$$\Delta Is1_b := Is1_{cs} \cdot \Delta Is\% \qquad \Delta Is1_b = 0.91\,amp$$

- Secondary inductance :

$$Ls1_b := \frac{\left(Vo1 + Vd_{fw}\right) \cdot \left(T - Ton_{max}\right)}{\Delta Is1_b} \qquad Ls1_b = 9.17\,\mu H$$

- Primary inductance:

$$Lp_b := Ls1_b \cdot Nps1^2 \qquad Lp_b = 81.75\,\mu H$$

- Ramp amplitude:

$$\Delta Ip_b := \frac{\left(Vi_{min} - Vds_{on}\right) \cdot Ton_{max}}{Lp_b} \qquad \Delta Ip_b = 0.3\,amp$$

Select primary inductance (3-a) or (3-b):----> $Lp := Lp_b \qquad Lp = 81.75\,\mu H$

$$\Delta Ip := \frac{\left(Vi_{min} - Vds_{on}\right) \cdot Ton_{max}}{Lp} \qquad \Delta Ip = 0.3\,amp$$

- Primary average current:

$$Ip_{cs} := \frac{Po_{max}}{\left(Vi_{min} - Vds_{on}\right) \cdot \eta \cdot D_{max}} \qquad Ip_{cs} = 1.13\,amp$$

- Primary peak current:

$$Ip_{pk} := Ip_{cs} + \frac{\Delta Ip}{2} \qquad Ip_{pk} = 1.28\,amp$$

- Primary RMS current:

$$Ip_{rms} := \sqrt{D_{max} \cdot \left[Ip_{pk} \cdot \left(Ip_{cs} - \frac{\Delta Ip}{2} \right) + \frac{1}{3} \cdot \left[Ip_{pk} - \left(Ip_{cs} - \frac{\Delta Ip}{2} \right) \right]^2 \right]} \qquad Ip_{rms} = 0.66\,amp$$

- Primary DC current:

$$Ip_{dc} := \frac{Po_{max}}{\eta \cdot \left(Vi_{min} - Vds_{on}\right)} \qquad Ip_{dc} = 0.39\,amp$$

- Primary AC(rms) current:

$$Ip_{ac} := \sqrt{Ip_{rms}^2 - Ip_{dc}^2} \qquad Ip_{ac} = 0.54\,amp$$

$Edt := Vi_{min} \cdot Ton_{max}$ $\qquad Edt = 2.5 \times 10^{-5}\, volt \cdot sec$

4) Secondary winding: Inductance, peak, AC, RMS current

-Master output:

- Primary average current:

$$Is1_{cs} := \frac{Io1_{max}}{(1 - D_{max})} \qquad Is1_{cs} = 3.03\, amp$$

- Secondary inductance :

$$Ls1 := \frac{Lp}{Nps1^2} \qquad Ls1 = 9.17\, \mu H$$

- Ramp amplitude:

$$\Delta Is1 := \frac{(Vo1 + Vd_{fw}) \cdot (T - Ton_{max})}{Ls1} \qquad \Delta Is1 = 0.91\, amp$$

- Secondary peak current:

$$Is1_{pk} := Is1_{cs} + \frac{\Delta Is1}{2} \qquad Is1_{pk} = 3.49\, amp$$

- Secondary RMS current:

$$Is1_{rms} := \sqrt{(1 - D_{max}) \cdot \left[Is1_{pk} \cdot \left(Is1_{cs} - \frac{\Delta Is1}{2} \right) + \frac{1}{3} \cdot \left[Is1_{pk} - \left(Is1_{cs} - \frac{\Delta Is1}{2} \right) \right]^2 \right]}$$

$$Is1_{rms} = 2.47\, amp$$

- Secondary AC current:

$$Is1_{ac} := \sqrt{Is1_{rms}^2 - Io1_{max}^2} \qquad Is1_{ac} = 1.45\, amp$$

-First slave output:

- Primary average current:

$$Is2_{cs} := \frac{Io2_{max}}{(1 - D_{max})} \qquad Is2_{cs} = 0\, amp$$

- Secondary inductance :

$$Ls2 := \frac{Lp}{Nps2^2} \qquad Ls2 = 0.16\, \mu H$$

- Ramp amplitude:

$$\Delta Is2 := \frac{(Vo2 + Vd_{fw}) \cdot (T - Ton_{max})}{Ls2} \qquad \Delta Is2 = 6.92\, amp$$

- Secondary peak current:

$$Is2_{pk} := Is2_{cs} + \frac{\Delta Is2}{2} \qquad Is2_{pk} = 3.46\, amp$$

- Secondary RMS current:

$$Is2_{rms} := \sqrt{(1 - D_{max}) \cdot \left[Is2_{pk} \cdot \left(Is2_{cs} - \frac{\Delta Is2}{2} \right) + \frac{1}{3} \cdot \left[Is2_{pk} - \left(Is2_{cs} - \frac{\Delta Is2}{2} \right) \right]^2 \right]}$$

$$Is2_{rms} = 1.62\, amp$$

- Secondary AC current:

$$Is2_{ac} := \sqrt{Is2_{rms}^2 - Io2_{max}^2} \qquad Is2_{ac} = 1.62\, amp$$

5) Maximum Stress across the output diodes: Vdiode

-Maximum stress voltage on the cathode of diodes

$$Vdiode1_{max} := \frac{Vi_{max}}{Nps1} + Vo1 \qquad Vdiode1_{max} = 21.72\,volt$$

$$Vdiode2_{max} := \frac{Vi_{max}}{Nps2} + Vo2 \qquad Vdiode2_{max} = 2.42\,volt$$

Select a diode with Va-c>> Vdiode.max, and ultra-fast switching diode

$$Pdiode1_{max} := Is1_{rms} \cdot Vd_{fw} \cdot \left(1 - D_{max}\right) \qquad Pdiode1_{max} = 0.81\,watt$$

$$Pdiode2_{max} := Is2_{rms} \cdot Vd_{fw} \cdot \left(1 - D_{max}\right) \qquad Pdiode2_{max} = 0.53\,watt$$

$$Pdiode_{tot} := Pdiode1_{max} + Pdiode2_{max} \qquad Pdiode_{tot} = 1.35\,watt$$

6) Output ripple Specifications and Output Capacitors

To meet the output ripple specifications the output capacitors have to meet two criterias:
- satisfy the standard capacitance definition: **I=C*dV/dt** where t is the Toff time, V is 25% of the allowable output ripple.
- The Equivalent Series Resistance (ESR) of the capacitor has to provide less than 75% of the maximum output ripple. (Vripple=dI*ESR)

-Maximum outputs ripple:

$Vrp1 = 100\,mV \qquad Vrp2 = 120\,mV$

-Minimum output capacitance:

$$Co1 := \Delta Is1 \cdot \frac{\left(Ton_{max}\right)}{Vrp1 \cdot 0.25} \qquad Co1 = 41.39\,\mu F$$

-Maximum ESR value:

$$ESR1 := \frac{Vrp1 \cdot 0.75}{\Delta Is1} \qquad ESR1 = 0.08\,ohm$$

-Minimum output capacitance:

$$Co2 := \Delta Is2 \cdot \frac{\left(Ton_{max}\right)}{Vrp2 \cdot 0.25} \qquad Co2 = 262.14\,\mu F$$

$$ESR2 := \frac{0.75 \cdot Vrp2}{\Delta Is2}$$

-Maximum ESR value: $ESR2 = 0.01\,ohm$

7) Input capacitor:

The input capacitor has to meet the maximum ripple current rating Ip(rms) and the maximum input voltage ripple ESR value.

8) Switching Mosfet: Power Dissipation

The Mosfet is chosen based on maximum Stress voltage (section1), maximum peak input current (section 3), total power losses, maximum allowed operating temperature, and driver capability of the LM3488.

-The drain to source Breakdown of the mosfet (Vdss) has to be greater than:

$Vds_{max} = 76.3\,volt$

-Continuous Drain current of the mosfet (Id) has to be greater than:

$Ip_{pk} = 1.28\,amp$

- Maximum drive voltage:

The voltage on the drive pin of the LM3488, Vdr is equal to the input voltage when input voltage is less than 7.2V, and Vdr is equal to 7.2V when the input voltage is greater than 7.2V

$Vdr := 7.2 \cdot volt$

$Rdr_{on} := 7 \cdot ohm$

-Total Mosfet's losses and maximum junction temperature:

The goal in selecting a Mosfet is to minimize junction temperature rise by minimizing the power loss while being cost effective. Besides maximum voltage rating, and maximum current rating, the others three important parameters of a Mosfet are Rds(on), gate threshold voltage, and gate

capacitance.
The switching Mosfet has three types of losses, conduction loss and switching loss, and gate charge losses:
-**Conduction losses** are equal to: I^2*R losses, therefore the total resistance between the source and drain during the on state, Rds(on) has to be as low as possible.
-**Switching losses** are equal to: Switching-time*Vds*I*frequecy. The switching time, rise time and fall time is a function of the gate to drain Miller-charge of the Mosfet, Qgd, the internal resistance of the driver and the Threshold Voltage, Vgs(th) the minimum gate voltage which enables the current through drain source of the Mosfet.
-**Gate charge losses** are caused by charging up the gate capacitance and then dumping the charge to ground every cycle. The gate charge losses are equal to: frequency • Qg(tot) • Vdr
Unfortunately, the lowest on resistance devices tend to have higher gate capacitance.
Because this loss is frequency dependent, in very high current supplies with very large FETs, with large gate capacitance, a more optimal design may result from reducing operating frequency.
Switching losses are also effected by gate capacitance. If the gate driver has to charge a larger capacitance, then the time the Mosfet spends in the linear region increases and the losses increase. The faster the rise time, the lower the switching loss. Unfortunately this causes high frequency noise.

$n := 10^{-9}$

Mosfet:

$Rds_{on} := 0.200 \cdot ohm$ (Total resistance between the source and drain during the on state)

$Coss := 95 \cdot pF$ (Output capacitance)

$Qg_{tot} := 13 \cdot n \cdot coul$ (Total gate charge)

$Qgd_{miller} := 6.1 \cdot n \cdot coul$ (Gate drain Miller charge)

$Vgs_{th} := 2 \cdot volt$ (Threshold voltage)

- Conduction losses: Pcond

$Pcond := Rds_{on} \cdot Ip_{rms}^{2} \cdot D_{max}$ $\qquad Pcond = 0.03\,watt$

- Switching losses: Psw(max)

Turn On time:

$$t_{sw} := Qgd_{miller} \cdot \frac{Rdr_{on}}{Vdr - Vgs_{th}} \qquad t_{sw} = 8.21 \times 10^{-9}\,sec$$

$$Psw_{max} := \left(t_{sw} \cdot Vds_{max} \cdot Ip_{pk} \cdot fsw\right) + \frac{Coss \cdot Vds_{max}^{2} \cdot fsw}{2} \qquad Psw_{max} = 0.32\,watt$$

- Gate charge losses: Pgate

Average current required to drive the gate capacitor of the Mosfet:

$Igate_{awg} := fsw \cdot Qg_{tot}$ $\qquad Igate_{awg} = 3.9 \times 10^{-3}\,amp$

$Pgate := Igate_{awg} \cdot Vdr$ $\qquad Pgate = 0.03\,watt$

-Total losses: Ptot(max)

$Pmosfet_{tot} := Pcond + Psw_{max} + Pgate$ $\qquad Pmosfet_{tot} = 0.38\,watt$

-Maximum junction temperature and heat sink requirement:

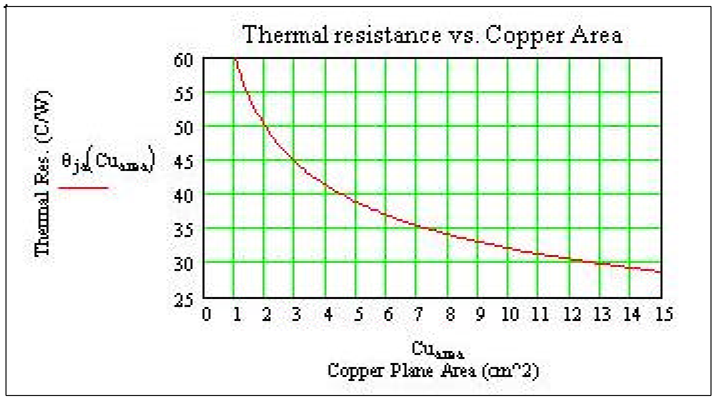
Maximum junction temperature desired: $Tj_{max} := 140$ Celsius

Maximum ambient temperature: $Ta_{max} := 70$ Celsius

-Thermal resistance junction to ambient temperature:

$$\theta ja := \frac{Tj_{max} - Ta_{max}}{Pmosfet_{tot}} \qquad \theta ja = 183.35\,\frac{1}{watt} \quad \text{Celsius}$$

If the thermal resistance calculated is lower than that one specified on the Mosfet's data sheet a heat sink or higher copper area is needed.
For Example for a T0-263 (D2pak) package the Tja of the Mosfet versus copper plane area is:



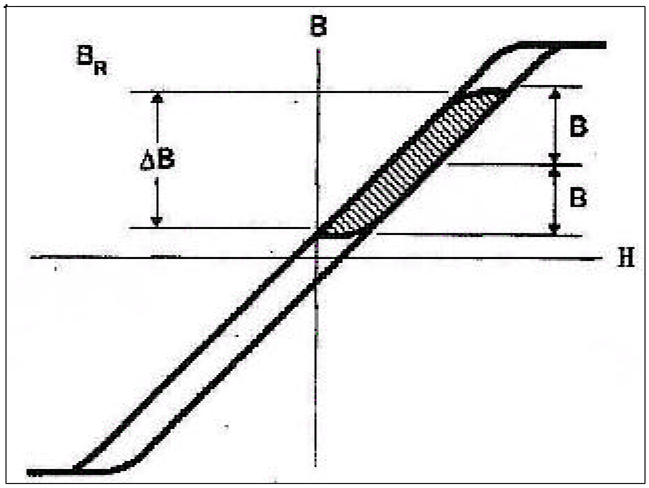
10) Current limit:

The LM3488 uses a current mode control scheme. The main advantages of current mode control are inherent cycle-by-cycle current limit for the switch, and simple control loop characteristics. Since the LM3488 has a maximum duty cycle of 100%, the current limit should be designed to have current limit just above the maximum primary peak current plus 20-30%

$$R_{sense} := \frac{160 \cdot mV}{Ip_{pk} \cdot 1.2} \qquad R_{sense} = 0.1\,\Omega$$

11) Transformer Design:

The inductor- transformer should be designed to minimize the leakage inductance, ac winding losses, and core losses.
In continuous mode of operation, the total amper-turns never goes to zero, therefore the transformer will have lower core losses, and high AC winding losses.
Unipolar pulses cause dc current to flow through the core winding, moving the flux in the core from Br towards saturation. When pulses goes to zero, the flux travels back to Br. The transformer in Flyback power supply acts as an energy storage device, therefore to do not saturate is used a air-gapped ferrite core or Molypermalloy Poweder cores with distributed air-gap.



The power handling capacity of the transformer core can be determined by its WaAc product area , where Wa is the available core window area, and Ac is the effective core cross-selectional area. The WaAc power output relationship is obtained with the Faraday's law:

E = 4 B Ac Nf 10^-8

Where:

E = applied voltage J = current density amp/cm^2
B = flux density in gauss K = winding factor
Ac = core are in cm^2 I = current (rms)

N = number of turns   Po = output power
f = frequency   Wa = window area in cm^2

-Select maximum current density of the windings:C   (280- 390 amp/cm^2, or 400-500 circular-mils/amp)

$J := 390 \cdot \frac{amp}{cm^2}$ $cir\_mil := 5.07 \cdot 10^{-6} \cdot cm^2$ $\frac{1}{J} = 505.74 \frac{cir\_mil}{amp}$

- winding factor: $K := 0.2$ (0.2-0.3 for flyback continuous mode)

-Select core material and maximum flux density:

It is assume that at high switching frequency (fsw>>25KHz) the limitation factor is the core losses, and temperature rise of the transformer

The type of ferrite material chosen will influence the core losses at the given operating conditions:

- F material has its lowerst losses at room temperature to 40°C.
- P material has lowerst losses at 70°C-80°C.
- R material has lowerst losses at 100°C-110°C.
- K material has lowerst losses at 40°C-60°C at elevated frequencies.

At high switching frequency it is necessary to adjust the flux density in order to limit core temperature rise: limiting core losses density to 100mW/cm^3 would keep the temperature rise at approximately 40°C.

Use the following formula to select the most appropriate maximum flux density:

-Maximum core losses density: $Pcored := 250$ mW/cm^3

for P material:

a = 0.158	b = 1.36	c = 2.86	for frequency f<100kHz
a = 0.0434	b = 1.63	c = 2.62	for frequency 100kHz<f<500kHz
a = 7.36*10^-7	b = 3.47	c = 2.54	for frequency f>500kHz

for K material:

a = 0.0530	b = 1.60	c = 3.15	for frequency f<500kHz
a = 0.00113	b = 2.19	c = 3.10	for frequency 500kHz<f<1 MHz
a = 1.77*10^-9	b = 4.13	c = 2.98	for frequency f>1MHz

$a1 := 0.0434$ $b1 := 1.63$ $c1 := 2.62$

$$B := \left[\frac{Pcored}{a1 \cdot \left(\frac{fsw}{kHz}\right)^{b1}}\right]^{\frac{1}{c1}} \cdot 10^3 \cdot gauss \qquad B = 783.75\, gauss$$

$\Delta B := B \cdot 2$ $\Delta B = 1.57 \times 10^3\, gauss$

-Topology constant: $Kt := \frac{0.00025}{1.97} \cdot 10^3$

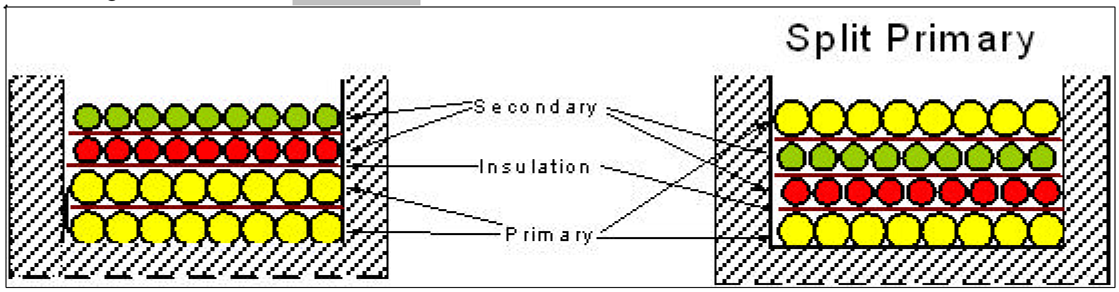
$$WaAc := \frac{Po_{max}}{Kt \cdot \Delta B \cdot fsw \cdot J} \qquad WaAc = 0.03\, cm^4$$

- Select a core with Area Product larger than : ---> $WaAc = 0.03\, cm^4$

Core selected:

- Manufacture: Ferroxcube.com
- Material: P
- Shape: E core
- Part number: RM6S/I3-F3
- Core Area: Ae $Ae := 0.37 \cdot cm^2$
- Bobbin area: Wa $Wa := 0.14 \cdot cm^2$
- Core volume: Ve $Ve := 1.09 \cdot cm^3$
- Window length lw $lw := 0.67 \cdot cm$

- Area product: Used ---------------------------> $Ae \cdot Wa = 0.05\,cm^4$
- Inductance per 1000 turns without airgap :
- first length of turn: $Lt := 24 \cdot cm$



- Primary inductance: Primary turns

$$Np_c := \frac{Lp \cdot Ip_{pk}}{\Delta B \cdot Ae} \qquad Np_c = 18.07$$

The number of turns has to be rounded to the higher or lower integer value: $Np := 18$

$$\frac{Np \cdot Ae \cdot \Delta B}{Ip_{pk}} = 81.45\,\mu H$$

- Secondary inductance: Secondary turns

$$Ns1_c := \left(\frac{Np}{Nps1}\right) \qquad Ns1_c = 6.03 \quad \dashrightarrow \quad Ns1 := 6$$

$$Ns2_c := \left(\frac{Np}{Nps2}\right) \qquad Ns2_c = 0.79 \quad \dashrightarrow \quad Ns2 := 0$$

-Air-gap length

The air-gap length is proportional to the effective gap section area (Ag).
Ag is equal to the core section times the finging coefficient, that take in consideration the finging flux in the air-gap. Since Ag depends on the air-gap length itself, the air-gap length (Lg) can be calculated with few iterations of a loop cycle.

$$\mu_o := 4 \cdot \pi \cdot 10^{-7} \cdot \frac{henry}{m}$$

$$Lg := \begin{array}{|l} Ag \leftarrow \dfrac{Ae}{cm^2} \\ \text{for } \ i \in 0..4 \\ \quad \begin{array}{|l} lgap \leftarrow \mu_o \cdot \dfrac{cm}{henry} \cdot Np^2 \cdot \left(\dfrac{Ag}{\dfrac{Lp}{henry}}\right) \\ Ag \leftarrow \dfrac{Ae}{cm^2} \cdot \left(1 + \dfrac{lgap}{\sqrt{\dfrac{Ae}{cm^2}}} \cdot \log\left(\dfrac{2 \cdot \dfrac{lw}{cm}}{lgap}\right)\right) \end{array} \\ (lgap) \cdot cm \end{array}$$

(Air-gap length)

$$Lg = 7.68 \times 10^{-3}\,in \qquad Lg = 0.2\,mm$$

- Primary and secondary wire size:

Maximum current density: $J = 390 \frac{amp}{cm^2}$

Primary rms current: $Ip_{rms} = 0.66\,amp$

Primary:

by wire area:

$Wp_{cu} := \frac{Ip_{rms}}{J}$ $\quad Wp_{cu} = 1.7\,10^{-3} \cdot cm^2$

or by wire size:

$AWGp := -4.2 \cdot \ln\left(\frac{Wp_{cu}}{cm^2}\right)$ $\quad AWGp = 26.79$

(Approximated AWG wire size, for more precision refer to wire size table)

Primary Wire selected:

Wire size $AWG_{Lp} := 28$

Bare area (copper plus insulation) $Wa_{Lp} := 1.05 \cdot 10^{-3} \cdot cm^2$

Copper area: $Wcu_{Lp} := 0.8 \cdot 10^{-3} \cdot cm^2$

Diameter $Dcu_{Lp} := 0.0366 \cdot cm$

Number of strands: $Nst_{Lp} := 2$

- Number of primary turns per layer:

$Ntl_{Lp} := floor\left(\frac{lw}{Dcu_{Lp}}\right)$ $\quad Ntl_{Lp} = 18$

- Number of primary layers:

$Nly_{Lp} := ceil\left(\frac{Np \cdot Nst_{Lp}}{Ntl_{Lp}}\right)$ $\quad Nly_{Lp} = 2$

Secondary: Master

by wire area:

$Ws1_{cu} := \frac{Is1_{rms}}{J}$ $\quad Ws1_{cu} = 6.34\,10^{-3} \cdot cm^2$

or by wire size:

$AWGs1 := -4.2 \cdot \ln\left(\frac{Ws1_{cu}}{cm^2}\right)$ $\quad AWGs1 = 21.26$

Secondary Wire selected:

Wire size $AWG_{Ls1} := 25$

Bare area (copper plus insulation) $Wa_{Ls1} := 2 \cdot 10^{-3} \cdot cm^2$

Copper area: $Wcu_{Ls1} := 2.514 \cdot 10^{-3} \cdot cm^2$

Diameter $Dcu_{Ls1} := 0.0505 \cdot cm$

Number of strands: $Nst_{Ls1} := 4$

- Number of secondary turns per layer:

$Ntl_{Ls1} := floor\left(\frac{lw}{Dcu_{Ls1}}\right)$ $\quad Ntl_{Ls1} = 13$

- Number of secondary layers:

$$Nly_{Ls1} := \mathrm{ceil}\left(\frac{Ns1 \cdot Nst_{Ls1}}{Ntl_{Ls1}}\right) \qquad Nly_{Ls1} = 2$$

Secondary: Slave

by wire area:

$$Ws2_{cu} := \frac{Is2_{rms}}{J} \qquad Ws2_{cu} = 4.16 \cdot 10^{-3} \cdot cm^2$$

or by wire size:

$$AWGs2 := -4.2 \cdot \ln\left(\frac{Ws2_{cu}}{cm^2}\right) \qquad AWGs2 = 23.03$$

Secondary Wire selected:

Wire size	$AWG_{Ls2} := 26$
Bare area (copper plus insulation)	$Wa_{Ls2} := 1.63 \cdot 10^{-3} \cdot cm^2$
Copper area:	$Wcu_{Ls2} := 1.28 \cdot 10^{-3} \cdot cm^2$
Diameter	$Dcu_{Ls2} := 0.0452 \cdot cm$
Number of strands:	$Nst_{Ls2} := 1$

- Number of secondary turns per layer:

$$Ntl_{Ls2} := \mathrm{floor}\left(\frac{lw}{Dcu_{Ls2}}\right) \qquad Ntl_{Ls2} = 14$$

- Number of secondary layers:

$$Nly_{Ls2} := \mathrm{ceil}\left(\frac{Ns2 \cdot Nst_{Ls2}}{Ntl_{Ls2}}\right) \qquad Nly_{Ls2} = 0$$

- Copper area:

$$Wcu_{tot} := \left(Dcu_{Lp} \cdot Nly_{Lp} + Dcu_{Ls1} \cdot Nly_{Ls1} + Dcu_{Ls2} \cdot Nly_{Ls2}\right) \cdot 1.15 \cdot lw$$

$$Wcu_{tot} = 0.13\,cm^2$$

- Window utilizzation:

$$Wu := \frac{Wcu_{tot}}{Wa} \qquad Wu = 95.87\,\%$$

Important: if Window utilisation is greater than 90%, (Copper area>> than bobbin area) a core with larger window area, or smaller wire sizes must be selected.

- Core losses:

$$Pcore := Ve \cdot \left[\left(\frac{B}{10^3 \cdot gauss}\right)^{c1} \cdot a1 \cdot \left(\frac{fsw}{kHz}\right)^{b1}\right] \cdot \frac{10^{-3} \cdot watt}{cm^3} \qquad Pcore = 0.27\,watt$$

- Winding copper losses:

There are two effects, which can cause the winding losses to be significantly greater than (I^2*Rcu): skin and proximity effects.

Skin effect causes current in a wire to flow only in a thin skin of the wire.

Skin depth: distance below the surface where the current density has fallen to 1/e of its value at the surface: (Sd)

$$Sd := \frac{6.61}{\sqrt{\frac{fsw}{Hz}}} \cdot cm \qquad Sd = 0.01\,cm \qquad Lt = 24\,cm \qquad Nly_{Lp} = 2$$

To minimize the AC copper losses in a transformer, if the wire diameter is greater than two times the skin depth, a multy strands winding or litz wires should be considered

If $Dcu_{Lp} = 0.04\,cm$ is greater than $Sd \cdot 2 = 0.02\,cm$

Primary winding length:

$$Ldf_{Lp} := \begin{array}{|l} L1 \leftarrow Lt \\ \text{for } i \in 1 .. (Nly_{Lp} - 1) \\ \quad L1 \leftarrow L1 + 4 \cdot Dcu_{Lp} \\ L1 \end{array}$$

$$Lcu_{Lp} := \begin{array}{|l} L1 \leftarrow Lt \\ L \leftarrow 0 \cdot cm \\ \text{for } i \in 1 .. (Nly_{Lp} - 1) \\ \quad \begin{array}{|l} L \leftarrow L + L1 \cdot Ntl_{Lp} \\ L1 \leftarrow L1 + 4 \cdot Dcu_{Lp} \end{array} \\ \left[L + L1 \cdot \left[Np - (Nly_{Lp} - 1) \cdot Ntl_{Lp} \right] \right] \end{array}$$

$Np = 18$ $\quad Lcu_{Lp} = 432\,cm$ $\quad Ldf_{Lp} = 24.15\,cm$ $\quad 7.15 \cdot Np = 128.7$

Copper resistivity: (20C) $\quad \rho_{20} := 1.724 \cdot 10^{-6} \cdot ohm \cdot cm$

-Maximum temperature of the winding: $Tmax_{cu} := 80$

$$\rho := \rho_{20} \cdot \left[1 + 0.0042 \cdot (Tmax_{cu} - 20) \right]$$

$$Rdc_{Lp} := \rho \cdot \frac{Lcu_{Lp}}{Wcu_{Lp} \cdot Nst_{Lp}} \qquad Rdc_{Lp} = 0.58\,ohm$$

$$Rac_{Lp} := \frac{Rdc_{Lp} \cdot \left(\frac{Dcu_{Lp}}{2 \cdot Sd} \right)^2}{\left(\frac{Dcu_{Lp}}{2 \cdot Sd} \right)^2 - \left(\frac{Dcu_{Lp}}{2 \cdot Sd} - 1 \right)^2} \qquad Rac_{Lp} = 0.66\,ohm$$

$$\frac{Rac_{Lp}}{Rdc_{Lp}} = 1.13$$

$$Pcu_{Lp} := Rdc_{Lp} \cdot Ip_{dc}^{\,2} + Rac_{Lp} \cdot Ip_{ac}^{\,2} \qquad Pcu_{Lp} = 0.28\,watt$$

Secondary winding length:

$$Ldf_{Ls1} := \begin{array}{|l} L1 \leftarrow Ldf_{Lp} \\ \text{for } i \in 1 .. (Nly_{Ls2} - 1) \\ \quad L1 \leftarrow L1 + 4 \cdot Dcu_{Ls1} \\ L1 \end{array}$$

$$Lcu_{Ls1} := \left| \begin{array}{l} L1 \leftarrow Ldf_{Lp} \\ L \leftarrow 0 \cdot cm \\ \text{for } i \in 1 .. (Nly_{Ls1} - 1) \\ \quad \left| \begin{array}{l} L \leftarrow L + L1 \cdot Ntl_{Ls1} \\ L1 \leftarrow L1 + 4 \cdot Dcu_{Ls1} \end{array} \right. \\ L \leftarrow 0 \text{ if } Nly_{Ls1} \leftarrow 1 \\ \left[L + L1 \cdot \left[Ns1 - (Nly_{Ls1} - 1) \cdot Ntl_{Ls1} \right] \right] \end{array} \right.$$

$$Lcu_{Ls1} = 146.09\,cm$$

$$Rdc_{Ls1} := \rho \cdot \frac{Lcu_{Ls1}}{Wcu_{Ls1} \cdot Nst_{Ls1}} \qquad Rdc_{Ls1} = 0.03\,ohm$$

$$Rac_{Ls1} := \frac{Rdc_{Ls1} \cdot \left(\frac{Dcu_{Ls1}}{2 \cdot Sd} \right)^2}{\left(\frac{Dcu_{Ls1}}{2 \cdot Sd} \right)^2 - \left(\frac{Dcu_{Ls1}}{2 \cdot Sd} - 1 \right)^2} \qquad Rac_{Ls1} = 0.04\,ohm$$

$$\frac{Rac_{Ls1}}{Rdc_{Ls1}} = 1.37$$

$$Pcu_{Ls1} := Rdc_{Ls1} \cdot Io1_{max}^{\;2} + Rac_{Ls1} \cdot Is1_{ac}^{\;2} \qquad Pcu_{Ls1} = 0.22\,watt$$

$$Lcu_{Ls2} := \left| \begin{array}{l} L1 \leftarrow Ldf_{Ls1} \\ L \leftarrow 0 \cdot cm \\ \text{for } i \in 1 .. (Nly_{Ls2} - 1) \\ \quad \left| \begin{array}{l} L \leftarrow L + L1 \cdot Ntl_{Ls2} \\ L1 \leftarrow L1 + 4 \cdot Dcu_{Ls2} \end{array} \right. \\ L \leftarrow 0 \text{ if } Nly_{Ls2} \leftarrow 1 \\ \left[L + L1 \cdot \left[Ns2 - (Nly_{Ls2} - 1) \cdot Ntl_{Ls2} \right] \right] \end{array} \right.$$

$$Lcu_{Ls2} = 0\,cm$$

$$Rdc_{Ls2} := \rho \cdot \frac{Lcu_{Ls2}}{Wcu_{Ls2} \cdot Nst_{Ls2}} \qquad Rdc_{Ls2} = 0\,ohm$$

$$Rac_{Ls2} := \frac{Rdc_{Ls2} \cdot \left(\frac{Dcu_{Ls2}}{2 \cdot Sd} \right)^2}{\left(\frac{Dcu_{Ls2}}{2 \cdot Sd} \right)^2 - \left(\frac{Dcu_{Ls2}}{2 \cdot Sd} - 1 \right)^2} \qquad Rac_{Ls2} = 0\,ohm$$

$$\frac{Rac_{Ls2}}{Rdc_{Ls2}} = 0$$

$$Pcu_{Ls2} := Rdc_{Ls2} \cdot Io2_{max}^{\;2} + Rac_{Ls2} \cdot Is2_{ac}^{\;2} \qquad Pcu_{Ls2} = 0\,watt$$

$$Pcu_{tot} := Pcu_{Lp} + Pcu_{Ls1} + Pcu_{Ls2} \qquad Pcu_{tot} = 0.49\,watt$$

$Pcore = 0.27\,watt$

-Total transformer's losses:

$Ptrans_{tot} := Pcu_{tot} + Pcore$ $\quad$ $Ptrans_{tot} = 0.77\,watt$

-Transformer's efficiency:

$$\eta_{Tra} := \frac{Po_{max}}{Po_{max} + Ptrans_{tot}} \qquad \eta_{Tra} = 90.84\,\%$$

12) Total Power Supply Efficiency:

$Ptrans_{tot} = 0.77\,watt$ $\quad$ $Pdiode_{tot} = 1.35\,watt$ $\quad$ $Pmosfet_{tot} = 0.38\,watt$

$Pout := Vo1 \cdot Io1_{max} + Vo2 \cdot Io2_{max}$

$$\eta_{tot} := \frac{Pout}{Pout + Ptrans_{tot} + Pdiode_{tot} + Pmosfet_{tot}} \qquad \eta_{tot} = 72.55\,\%$$

13) Select the proper switching frequency:

The operating frequency of the power supply should be selected to obtain the best balance between switching losses, total transformer losses, size and cost of magnetic components and output capacitors.

High switching frequency reduces the output capacitor value and the inductance of the primary and secondary windings, and therefore the total size of the transformer.

In the same manner, higher switching frequency increases the transformer losses and the switching losses of the switching transistor. High losses reduce the overall efficiency of the power supply, and increase the size of the heat-sink required to dissipate the heat.

Notes:

Wire table:

AWG Wire Size	Bare Area cm ^2 10^-3	Area cm ^2 10^-3	Diameter cm
10	52,61	55,9	0,267
11	41,68	44,5	0,238
12	33,08	35,64	0,213
13	26,36	28,36	0,19
14	20,82	22,95	0,171
15	16,51	18,37	0,153
16	13,07	14,73	0,137
17	10,39	11,68	0,122
18	8,23	9,32	0,109
19	6,53	7,54	0,098
20	5,19	6,06	0,0879
21	4,12	4,84	0,0785
22	3,24	3,86	0,0701
23	2,59	3,13	0,0632
24	2,05	2,514	0,0566
25	1,62	2	0,0505
26	1,28	1,603	0,0452
27	1,02	1,313	0,0409
28	0,8	1,05	0,0366
29	0,647	0,854	0,033
30	0,506	0,678	0,0294

References:

1. Rudolf P. Severns, Gordon E. Bloom "Modern DC to DC switchmode power converter circuits".
2. Magnetics application notes.
3. Colonel Wm. T. McLyman "Transformer and Inductor Design Handbook"

4. Pressman "Switching Power Supply Design"
5. R. Martinelli, C. Hymowitz, Intusoft "Designing a 12.5W 50kHz Flyback Transformer"