CHAPTER 13bis SWITCH-MODE POWER SUPPLIES

16. THE DESIGN OF SWITCH-MODE POWER SUPPLIES

16.1 SMPS COMPONENTS

See p.13.30

16.2 DETERMINATION COMPONENTS OF A DC-DC CONVERTER

For the choice of the switching transistor and the free wheel diode we can use the waveforms from the study of the converter under consideration. This gives an easy way to determine the required current and voltage for the semiconductors. In addition we have to determine inductor and capacitor for buck, boost and buck-boost. Remark that a number of manufacturers offer a large choice of “off the self” inductors so it’s not worth the effort to construct homemade inductors. In the present section we restrict the study to determine the specifications of the choke (self inductance and current). For the design of the three basic types of converter we consider only continuous current through the coil. From this, the case of discontinuous current is easy to deduce.

16.2.1 Buck converter with continuous current (fig.13-4/13-5a textbook)

a. Choice of switching transistor

It has been assumed that we use a power mosfet.
The maximum transistor current is I_1 and the transistor has to resist a voltage V_i .

Fig. 5-24 shows an identical current propagation in which: $I_{RMS} = \delta \cdot \sqrt{\dfrac{I_1^2 + I_1 \cdot I_2 + I_2^2}{3}}$

The power dissipation through the mosfet is: $P = I_{RMS}^2 \, . \, R_{DS(ON)}$

b. Choice of free wheel diode

The maximum reverse voltage is V_i . Fig.13-5a shows that the maximum diode current is I_1 and the average current is to determine like in fig. 5-24: $I_{AV} = \dfrac{I_1 + I_2}{2}$

c. General formula for *L*

From (13-2): $L = V_i \dfrac{(1 - \delta)\,T}{\Delta i_L}$. The largest value of L is obviously for the case that

$\delta = \delta_{min}$ (with $V_i = V_{i\,max}$), so that: $L_{max} = \dfrac{V_o \cdot T \cdot (1 - \delta_{min})}{\Delta i_L}$ or $L = \dfrac{V_o \cdot t_{off\,max}}{\Delta i_L}$ (13-28)

d. Calculation of choke and filter capacitor

Given:
- input voltage: $V_i = 8$ to 15 V
- output voltage: $V_o = 5V \pm 0.1\%$
- maximum load current: $I_o = 2A$
- chopper frequency: 100kHz

Required: calculate the values of the choke and the filter capacitor

Sequence of operations	Numeric example
1. Determination of $t_{off\,max}$ $t_{off\,max} = (1 - \delta_{min})\,.\,T$	**1. $t_{off\,max}$ = ?** $t_{off\,max} = (1 - \frac{5}{15})\,.\,10^{-5} = 6.67\mu s$
2. Ripple current through the coil To minimize the peak current through the choke as well as the output ripple it is useful to assume $\Delta i_L \le 25\%\, I_o$. We obtain the required output ripple of 0.1% by filtering with C !	**2. Δi_L = ?** Suppose e.g. Δi_L = 20% of I_o $\Delta i_L = 0.2 \times 2 = 0.4A$
3. Self inductance of the choke $L = \frac{V_o\,.\,t_{off\,max}}{\Delta i_L}$	**3. L = ?** $L = \frac{V_o\,.\,t_{off\,max}}{\Delta i_L} = \frac{5 \times 6.67 \times 10^{-6}}{0.4} = 83.4\mu H$
4. Peak current through the coil The peak current through the coil is given by: $i_{L\,peak} = I_o + \frac{\Delta i_{L\,max}}{2}$ In case of a large Δi_L , it is important that we determine $i_{L\,peak}$ because the choke may not saturate with this peak current.	**4. $i_{L\,peak}$ = ?** $i_{L\,peak} = I_o + \frac{\Delta i_{L\,max}}{2} = 2 + \frac{0.4}{2} = 2.2A$
5. Determination of filter capacitor $C_{min} = \frac{\Delta i_L}{8\,.\,f\,.\,\Delta v_{C\,max}}$ Furthermore we must have $ESR \le \frac{\Delta v_{o\,max}}{\Delta i_L}$ If the ESR of the available capacitor is too high, it can be practical to switch two or more capacitors in parallel so that the resultant ESR decreases	**5. C = ?** $\Delta v_{C\,max} = \Delta v_{o\,max} = 0.1\,\%\,.\,5 = 5mV$ $C_{min} = \frac{\Delta i_L}{8\,.\,f\,.\,\Delta v_{C\,max}}$ $= \frac{0.4}{8 \times 10^5 \times 5 \times 10^{-3}} = 100\mu F$ $ESR = \frac{5 \times 10^{-3}}{0.4} = 0.0125\Omega$

14.2.2 Boost converter with continuous current (fig. 13-7/13-8a book)

a. Power mosfet

Identical with the choice discussed for the buck converter

b. Free wheel diode

Maximum reverse voltage: V_o

Diode current: see fig. 13-8a

c. General formula for *L*:

From (13-8) results: $L = \dfrac{\delta \,.\, V_i \,.T}{\Delta i_L}$

With $V_o = \dfrac{V_i}{1-\delta}$ it follows: $L = \dfrac{V_o \,.(\delta - \delta^2)\,.T}{\Delta i_L}$

The maximum of $(\delta - \delta^2)$ is found for $\delta = 0.5$ so that: $L_{max} = \dfrac{0.25 \text{ x } V_o \text{ x } T}{\Delta i_L}$ (13-29)

d. Determination of choke and filter capacitor

Given:
- input voltage: $V_i = 3$ to 5V
- output voltage: $V_o = 9\text{V} \pm 0.1\%$
- maximum load current: $I_o = 1\text{A}$
- chopper frequency: $f = 50\text{kHz}$

Required: calculate the values of *L* and *C*

Sequence of operations	**Numeric example**
1. Ripple current through the coil As discussed for the buck converter: $\Delta i_L \leq 25\% \,.\, I_o$	**1.** $\Delta i_L = ?$ $\Delta i_L = 20\% \,.\, I_o = 0.2 \text{ x } 1 = 200\text{mA}$
2. Determination of self inductance *L* $L = \dfrac{0.25 \,.\, V_o \,.\, T}{\Delta i_L}$	**2.** $L = ?$ $L = \dfrac{0.25 \text{ x } V_o \text{ x } T}{\Delta i_L} = \dfrac{0.25 \text{ x } 9 \text{ x } 2 \text{ x } 10^{-5}}{0.2} = 225\mu\text{H}$
3. Peak current through the coil $i_{L\,peak} = I_o + \dfrac{\Delta i_L}{2}$	**3.** $i_{L\,peak} = ?$ $i_{L\,peak} = 1 + \dfrac{0.2}{2} = 1.1\text{A}$
4. Determination of smoothing capacitor (13-8): $C_{min} = \dfrac{\Delta i_L}{8 \,.\, f \,.\, \Delta v_{C\,max}}$ $ESR \leq \dfrac{\Delta v_{o\,max}}{\Delta i_L}$	**4.** $C = ?$ $\Delta v_{C\,max} = \Delta v_{o\,max} = 0.1\,\% \text{ x } 9 = 9\text{mV}$ $C_{min} = \dfrac{\Delta i_L}{8 \,.\, f \,.\, \Delta v_{C\,max}} = \dfrac{0.2}{8 \text{ x } 50 \text{ x } 10^3 \text{ x } 9 \text{ x } 10^{-3}}$ $C_{min} = 55.6\mu\text{F}$ $ESR_{max} = \dfrac{9 \text{ x } 10^{-3}}{0.2} = 0.045\Omega$

Remark:

An output filter with small L and large C gives a good response on load surges.
For a good functioning TTL-circuit with a 5V-supply a dip of 250mV is not allowed for example in the case of a 25% load surge.

14.2.3 Buck-boost converter with continuous current (fig.13-9/13-10a)

a. Power mosfet

Maximum voltage: $V_i + V_o$

Determination of current: identical as for the buck converter

b. Diode

Maximum reverse voltage: $V_i + V_o$

Determination of current: identical to buck converter

c. General formula for the self inductance *L*

From (13-15) it follows: $L = \dfrac{V_o \,.\, (1 - \delta) \,.\, T}{\Delta i_L}$

From $V_o = V_i \dfrac{\delta}{1 - \delta}$ it follows that δ_{min} appears with $V_{i\,max}$.

The worst case for L is for δ_{min} (that is for $V_{i\,max}$) : $L = \dfrac{V_o \,.\, (1 - \delta_{min}) \,.\, T}{\Delta i_L}$

d. Determination of choke and filter capacitor

Given: input voltage: $V_i = 3$ to 15V
output voltage: $V_o = 9\text{V} \pm 0.1\%$
maximum load current: $I_o = 3\text{A}$
chopper frequency: $f = 100\text{kHz}$

Required: determine the values of L and C

Sequence of operations	**Numeric example**
1. Value of Δi_L Similar to the preceding numeric examples we suppose $\Delta i_L \leq 25\% \,.\, I_o$	**1.** $\Delta i_L = ?$ $\Delta i_L = 0.2 \text{ x } I_o = 0.2 \text{ x } 3 = 0.6\text{A}$
2. Determine δ_{min} $V_o = V_{i\,max} \cdot \frac{\delta_{min}}{(1 - \delta_{min})}$ $\delta_{min} = \frac{V_o}{V_{i\,max} + V_o}$	**2.** $\boldsymbol{\delta_{min}} = ?$ $\delta_{min} = \frac{9}{15+9} = 0.375$
3. Determine choke L $L = \frac{V_o \,.\, (1 - \delta_{min}) \,.\, T}{\Delta i_L}$	**3. L = ?** $L = \frac{9 \text{ x } (1 - 0.375) \text{ x } 10^{-5}}{0.6} = 93.75\mu\text{H}$
4. Peak current through choke $i_{L\,peak} = I_o + \frac{\Delta i_L}{2}$	**4. Peak current coil ?** $i_{L\,peak} = 3 + 0.3 = 3.3\text{A}$
5. Determination of smoothing capacitor (13-7): $C_{min} = \frac{\Delta i_L}{8 \,.\, f \,.\, \Delta v_{C\,max}}$ $ESR \leq \frac{\Delta v_{o\,max}}{\Delta i_L}$ Same remark concerning the *ESR* as for the buck converter	**5. Value of capacitor** $\Delta v_{C\,max} = \Delta v_{o\,max} = 0.1\% \,.\, V_o$ $= 0.1\% \text{ x } 9 = 9\text{mV}$ $C_{min} = \frac{\Delta i_L}{8.\, f \,.\, \Delta v_{C\,max}} = \frac{0.6}{8 \text{ x } 10^5 \text{ x } 9 \text{ x } 10^{-3}}$ $= 83.4\mu\text{F}$

14.3 Determination of the components for isolated converters

14.3.1 The isolated flyback converter

a. Generalities

We discuss the design of a flyback with discontinuous current. This operating mode is often mentioned as “total energy transfer”. The design of a flyback with continuous current (or incomplete energy transfer) is identical but the transistor current is now $(I_2 - I_1)$ instead of I_1.

With regard to the incomplete energy transfer we find for the total energy transfer:

- the transistor peak current will be larger (I_p)
- a relative small self inductance is sufficient

(energy $= \frac{L_p \cdot I_p^2}{2}$ so that, with a larger I_p, a smaller L_p will suffice)

- the heat loss ($I_p^2 \cdot R_p$) increases
- the ripple on de the input capacitor (mains rectifier) increases

The flyback with continuous current requires a larger transformer (demand a larger L_p !), that’s the reason that mostly for a discontinuous current is chosen. When the transistor operates in continuous mode (switch on with current I_2) the switch on losses will be larger than for total energy transfer (switch on with zero current).

b. Numerical example

Suppose an off-the-line supply with a mains input between 180 and 260V. De required output is 5V – 8A.

With a permitted mains fluctuation of + 6% and −10% we find for an off-the-line supply (via bridge rectifier and filter capacitor) as boundaries for V_i :

$V_{i\,max} = 260 \text{ x } \sqrt{2} \text{ x } 1.06 = 390\text{V}$
$V_{i\,min} = 180 \text{ x } \sqrt{2} \text{ x } 0.9 - 20 = 210\text{V}$

The 20V is a practical value who takes into account the ripple voltage on the input rectifier.

Initially we suppose a maximum loaded SMPS (5V – 8A) and we take into account the changes in the input voltage. The maximum duty ratio δ_{max} coincides with $V_{i\,min}$,while we need δ_{min} to become the same output power in case of $V_{i\,max}$.

If afterwards the load decreases, the controller will adjust the duty ratio (to decrease) so that the output voltage remains constant ($V_o = 5\text{V}$).

Given:	• input voltage:	$V_i = 210$ to 390V
	• output voltage:	$V_o = 5\text{V} \pm 0.1\%$
	• maximum output current:	$I_o = 8\text{A}$
	• chopper frequency:	$f_c = 100\text{kHz}$

Required: determine the components of an isolated flyback converter

Sequence of operations	**Numerical example**
1. Peak current transformer primary (13-22): $I_c = \frac{2 . P_o}{\eta . V_i . \delta_{max}}$ We know that we must have $\delta_{max} \leq 0.5$. A practical value is for example 0.45. The efficiency of such a converter is roughly about 80%. The duty ratio δ_{max} must coincide with the minimum input voltage $V_{i\,min}$, so that: $I_{c\,max} = \frac{2 . P_o}{\eta . V_{i\,min} . \delta_{max}}$	**1. Peak current transformer primary** $I_{C\,max} = \frac{2 . P_o}{\eta . V_{i\,min} . \delta_{max}} = \frac{2 \text{ x } 5 \text{ x } 6}{0.8 \text{ x } 210 \text{ x } 0.45}$ $\approx 0.8\text{A}$ $I_{C\,max}$ = peak current transistor = peak current in transformer primary (A) P_o = output power (W) η = efficiency (for example 80%) $V_{i\,min}$ = minimum input voltage (V) δ_{max} = maximum duty ratio (≤ 0.5)
2. Choice of the transistor - blocking voltage: $2 . V_{i\,max}$ - peak current: $I_{C\,max} = \frac{2 . P_o}{\eta . V_{i\,min} . \delta_{max}}$ - average current: $\frac{I_{C\,max}}{2} . \delta_{max}$ - $I_{RMS} = \frac{I_{max} . \sqrt{\delta}}{\sqrt{3}}$ - with a power mosfet: $P_D = I^2_{RMS} . R_{DS(ON)}$	**2. Transistor specifications ?** - maximum blocking voltage: $2 . V_{i\,max} = 2 \text{ x } 390 = 780\text{V}$ - peak current: 0.8A - average current: $\frac{0.8}{2} \text{ x } 0.45 = 0.18\text{A}$ - power dissipation (mosfet): (fig. 5-22!) $I_{RMS} = \frac{I_{max} . \sqrt{\delta}}{\sqrt{3}} = \frac{0.8 \text{ x } \sqrt{0.45}}{\sqrt{3}} = 0.3\text{A}$ $P_D = I^2_{RMS} . R_{DS(ON)}$
3. Determination of the self inductance of the transformer (13-20): $I_1 = \frac{V_i}{L_1} . \delta . T$ With $V_{i\,min}$, the duty ratio must be δ_{max} The peak current through the primary is to write as: $I_p = \frac{V_{i\,min}}{L_p} . \delta_{max} . T$, from which: $L_p = \frac{V_{i\,min} . \delta_{max} . T}{I_p}$	**3. L_p = ?** $L_p = \frac{V_{i\,min} . \delta_{max} . T}{I_p} = \frac{210 \text{ x } 0.45 \text{ x } 10^{-5}}{0.8} =$ $= 1.18\text{mH}$

4. Determination of δ_{max}

It is common practice to take an off-time $t_3 = (1 - \delta) \cdot T$ (fig. 13-13), than is:

$V_o = V_i \cdot \frac{\delta}{1-\delta}$, from which:

$V_o = V_{i\,min} \cdot \frac{\delta_{max}}{1-\delta_{max}}$ and also:

$V_o = V_{i\,max} \cdot \frac{\delta_{min}}{1-\delta_{min}}$

With $\frac{V_{i\,max}}{V_{i\,min}} = K$ we find finally:

$$\delta_{min} = \frac{\delta_{max}}{(1-\delta_{max}) \cdot K + \delta_{max}}$$

4. δ_{min} = ?

$$K = \frac{V_{i\,max}}{V_{i\,min}} = \frac{390}{210} = 1.857$$

$$\delta_{min} = \frac{\delta_{max}}{(1-\delta_{max}) \cdot K + \delta_{max}}$$

$$= \frac{0.45}{(1 - 0.45) \text{ x } 1.857 + 0.45} = 0.306$$

The control range for δ is: $0.306 \leq \delta \leq 0.45$. This control range compensates for the input voltage variations. If the load decreases ($I_o < 8$A) then the control will reduce δ so that V_o remains constant at 5V.

5. Choice of transformer core

Important core parameters are among other things the power capacity, the available winding area and the maximum frequency for low core loss operation. Power and areas are related with the following expression:

$$A_e A_w = \frac{0.682 \times P_o \times 10^5}{f \times B_{max} \times J}$$

A_e = effective cross sectional core (cm²)
A_w = bobbin winding area (cm²)
P_o = power capacity of the core (W)
f = chopper frequency (Hz)
B_{max} = maximum allowable flux density (T)
(1T = 1 Wb/m² = 10^4 gauss)
J = current density in the coil (A/m²)

With a chosen frequency f (e.g. 100 kHz) and a specific current density (e.g. 5A/mm²) we can determine the value of $A_e A_w$ for a supposed B_{max} (e.g. 0.3 Tesla). In the catalogue of the manufacturer we search for a core with a 50% larger $A_e A_w$ product. The increase of 50% is necessary to take into account the bobbin winding area enlarged with the insulation and air around the conductors of the coils. Next we have to look if the supposed B_{max} is allowed for the chosen core.
Instead of this procedure we can use nomographs put at our disposal by core manufacturers.
We made the choice for ferrite cores from the US firm “Coilcraft” (see table 13-2 on page 13.16).

5. Determination of transformer core

$P_o = 5 \times 8 = 40$W
We decide to take the EFD25 ferrite core from Coilcraft as he is able to handle 50W.
We find $A_e = 0.59$ cm² and $A_w = 0.4175$ cm² .

Remark:
American wire manufacturers give the current density frequently in circular mils per ampere (c.m./A), see table 13.3 on p.13BIS.17.

1 circular mil = surface area of a circle with a diameter of 0.001 inch
1 c.m. = 5×10^{-4} mm²
1000 c.m. = 0.5 mm²
1000 c.m./A = 0.5 mm²/A or 2A/mm²
For supply transformers a current density of 2.5A/mm² is common practice. Due to the small number of windings in the case of a ferrite core transformer for SMPS a larger current density of e.g. 400 c.m. (= 5A/mm²) can appear.

6. Determination of the transformer windings

a. Primary coil

From the general equation $e = N \cdot \frac{d\phi}{dt}$ it follows:

$$N_p = \frac{(E - T) \,.\, 10^4}{B \,.\, A_w} \text{ with:}$$

E - T = volt-seconds product of the core

(= flux $= L_p \,.\, I_p$!)

B = maximum allowed induction (Tesla)

(1T = 10^4 gauss)

A_e = effective cross sectional core

(cm² from there 10^4)

Coilcraft gives an (E-T) nomograph (fig. 13-27) for different E-cores and taking into account a B = 0.3 Tesla (3000 gauss) induction.

From fig. 13-27 (p. 13BIS.15) follows the number of turns (to prevent the core from saturation and to ensure low core loss)

b. Secondary coil

$$V_s = V_o + V_D \text{ and } V_s = \frac{V_i}{n} \cdot \frac{\delta \,.\, \eta}{1 - \delta}$$

Herein $n = \frac{N_p}{N_s}$

The worst case is minimum input voltage ($V_{i\,min}$) together with maximum output load ($\delta = \delta_{\max}$).

$$\frac{N_p}{N_s} = n = \frac{V_{i\,min} \cdot \delta_{max} \cdot \eta}{(V_o + V_D) \,.\, (1 - \delta_{max})}$$

$$N_s = N_p \cdot \frac{(V_o + V_D) \,.\, (1 - \delta_{max})}{\eta \,.\, V_{i\,min} \cdot \delta_{max}}$$

6. Number of primary and secondary (winding) turns

a. N_p **= ?**

(**E -T**) = $V_{i\,min} \cdot \delta_{max} \cdot T$ = 210 x 0.45 x 10^{-5}

= 945Vμs

We find (**E -T**) also from:

(E - T) = $L_p \,.\, I_p$ = 1.18 x 10^{-3} x 0.8 = 945Vμs

With a core EFD25 we find (fig. 13-27, p.13BIS.15): N_p = 54 turns .

The RMS current in the primary is (fig. 5-22 textbook):

$$I_{RMS} = I \,.\, \frac{\sqrt{\delta}}{\sqrt{3}} = 0.8 \text{ x } \frac{\sqrt{0.45}}{\sqrt{3}} = 0.3\text{A}$$

Table 13-3 (p. 13BIS.17) shows that for 1000 c.m./A an AWG25 is necessary. By analogy with the numerical example on p. 5.35 (textbook) we find for an AWG25: u = 0.33868 x 0.515 x$\sqrt{100}$ = 1.744.

1.744 < 6 → table 5.8 (p. 5.33 textbook) gives K = 1.046. The ohmic resistance $R_{AC} = K \,.\, R_{DC}$ increases with 4.6%. The Joule losses in the winding increases also with 4.6%

To conduct the same current we can for example connect two AWG28 wires in parallel. Calculation for an AWG28 gives that R_{AC} = 1.026 x R_{DC} (Joule power loss increases only with 2.6% instead of 4.6%!). We need now 108 turns.

A wire diameter of 0.37 mm gives a surface area of 0.1 mm².

The bobbin winding area is: 108 x 0.1 = 10.8 mm²

b. N_s **= ?**

$$N_s = N_p \cdot \frac{(V_o + V_D) \,.\, (1 - \delta_{max})}{\eta \,.\, V_{i\,min} \cdot \delta_{max}}$$

$$N_s = 54 \text{ x } \frac{(5 + 0.5) \text{ x } (1 - 0.45)}{0.8 \text{ x } 210 \text{ x } 0.45} = 2.16 \rightarrow N_s = 3 \,!$$

RMS current: $\frac{54}{3}$ x 0.3 = 5.4A

For example six AWG20 wires in parallel give a total of 18 turns of 0.892 mm (surface area 11.2 mm²).

Surface area (primary + secondary):

0.22 cm² + 50% (additional insulation + air between conductors) : 0.33 cm²

0.33 cm² < 0.4175 cm² (EFD25) : OK!

7. Determination of air gap Due to the biasing we can have a saturated core. To prevent from this we can take a larger cross sectional area or a core with air gap. For a compact SMPS-design, we prefer the second solution: $l_g = 4 . \pi . A_e . \frac{N_p^2}{L_p}$ (13-29) Let $\frac{L_p}{N_p^2} = A_L$. When we know A_e of the core, it will be easy to determine de length of the air gap. Fig. 13-29 (p. 13BIS.16) shows a nomograph of a few Coilcraft-cores. With A_L and the core known, we can at once find the length of the air gap.	**7. Length air gap ?** $A_L = \frac{L_p}{N_p^2} = \frac{1.18 \text{ x } 10^{-3}}{54^2} = 404$ nH/turn^2 Fig. 13-29 (p. 13BIS.16): $l_g = 0.16$ mm
8. Output diode flyback Peak current: $i_{s\,peak} = n . i_{p\,peak}$ Average current: $I_{AV} = I_o$ Maximum reverse voltage: $V_{RRM} \geq \frac{v_{p\,peak}}{n}$	**8. Choice of the output diode** $i_{s\,peak} = 0.8 \text{ x } \frac{54}{3} = 14.4\text{A}$ $I_{AV} = 8\text{A}$ $V_{RRM} \geq \frac{390}{54/3} = 21.66\text{V}$
9. Determination of filter capacitor (13-7): $C_{min} = \frac{\Delta i_L}{8 . f . \Delta v_{C\,max}}$ ESR $\leq \frac{\Delta v_{C\,max}}{\Delta i_L}$	**9. Choice smoothing capacitor** $\Delta v_{o\,max} = 0.1\% \text{ x } 5 = 5\text{mV}$ $C_{min} = \frac{14.4}{8 \text{ x } 10^5 \text{ x } 5 \text{ x } 10^{-3}} = 3600\mu\text{F}$ ESR $\leq \frac{5 \text{ x } 10^{-3}}{14.4} = 0.34\text{ m}\Omega$ To attain these capacitor and ESR values, we have to switch several capacitors in parallel.

14.3.2 The isolated forward converter

We start from the same AC-supply (180 to 260V) as with the flyback. The DC input voltage for the converter is 210 to 390V (see p. 13BIS.6).

Given: . input voltage: $V_i = 210$ to 390V
. output voltage: $V_o = 12V \pm 0.1\%$
. maximum load current: $I_o = 2A$
. chopper frequency: $f_c = 100kHz$

Required: make a choice of transformer, choke and smoothing capacitor.

Sequence of operations	Numerical example
1. Choice of transformer As with design of the flyback, we select a transformer from Coilcraft	**1. Choice transformer** $P_o = 12 \text{ x } 2 = 24W$ We select an EFD20 (p. 13BIS.15) since this transformer can handle up to 30W
2. Determination of the primary current $\eta . W_i = W_o$ $\eta . \frac{I_{p\,max} + I_{p\,min}}{2} . \delta_{max} . T . V_{i\,min} = V_o . I_o . T$ With $\eta = 80\%$, $\delta_{max} = 0.4$; $P_o = V_o . I_o$ and $\Delta i_L = I_{p\,max} - I_{p\,min} = 20\% . I_{p\,max}$ we find: $I_{p\,max} = \frac{3.47 \text{ x } P_o}{V_{i\,min}}$	**2. Determination of the primary current** $I_{p\,max} = \frac{3.47 \text{ x } P_o}{V_{i\,min}}$ $I_{p\,max} = \frac{3.47 \text{ x } 24}{210} = 0.397A$ $I_{p\,min} = 0.8 \text{ x } I_{p\,max} = 0.32A$ We calculate under 3c below that $I_{\mu\,peak} = 0.0257A$ so that the primary current fluctuates between 0.32A and 0.4227A From fig. 5-24 (textbook): $I_{p\,RMS} = 0.149A$ With a current density of 2.5A/mm² the necessary wire surface area is 0.06 mm² or 119.2 c.m. We switch two AWG32-wires (64 c.m. = 0.032mm²) in parallel

3. Number of turns transformer bobbins

a. Primary coil

We use again the nomograph from Coilcraft (fig. 13-27).

b. Secondary coil

$N_s = N_p \cdot \frac{V_s}{V_p}$; $V_s = \frac{V_o}{\delta_{max}}$; $i_{s\,peak} = n \cdot i_{p\,peak}$

with $n = \frac{N_p}{N_s}$

Fig. 5-24 (textbook):

$$I_{RMS} = \delta \cdot \sqrt{\frac{I_1^2 + I_1 \cdot I_2 + I_2^2}{3}}$$

For the secondary current, we have to take $(1 - \delta)$ instead of δ !

c. Degaussing coil

This coil must have as much turns as the primary and must be strong coupled with the primary to create a very small leakage inductance. Both coils are mostly bifilar wound. The surface area of the wire is determined by the current through the bobbin:

$$i_{\mu\,peak} = \frac{\delta_{max} \cdot T \cdot V_{i\,min}}{L_o}$$

$$L_o = N_p^2 \cdot \frac{4 \text{ x } \pi \text{ x } 10^{-7} \text{ x } \mu_r \text{ x } A_e}{l_e}$$

Fig. 5-22 (textbook): $I_{\mu\,RMS} = \frac{i_{\mu\,peak} \cdot \sqrt{\delta}}{\sqrt{3}}$

3. Number of turns transformer bobbins

a. Primary coil

$$(\text{E - T}) = V_{i\,min} \cdot \delta_{max} \cdot T$$
$$= 210 \text{ x } 0.4 \text{ x } 10^{-5} = 840\text{V}\mu\text{s}$$

From fig. 13-27 (p. 13BIS.15) it follows:
$N_p = 88$ turns

b. Secondary coil

$V_s = \frac{12}{0.4} = 30\text{V}$; $N_s = N_p \cdot \frac{V_s}{V_p} = 88 \text{ x } \frac{30}{210} = 12.57$

We take 13 turns.

$i_{s\,peak} = 0.397 \text{ x } \frac{88}{13} = 2.69\text{A}$

Fig. 5-24 (textbook) : $I_{s\,RMS} = 1.46\text{A}$

With a current density of 2.5A/mm² we need a wire of 0.584 mm² (= 1168 c.m.).
We chose for three AWG24 wires in parallel.
AWG24 (404 c.m. = 0.2 mm²)

c. Degaussing coil

Suppose $\mu_r = 5000$ and with A_e and l_e from the EFD20 it follows:

$$L_o = 88^2 \text{ x } \frac{4 \text{ x } \pi \text{ x } 10^{-7} \text{ x } 5000 \text{ x } 0.31 \text{ x } 10^{-4}}{0.0461}$$
$$= 32.72\text{mH}$$

$$i_{\mu\,peak} = \frac{0.4 \text{ x } 10^{-5} \text{ x } 210}{32.72 \text{ x } 10^{-3}} = 0.0257\text{A}$$

$$i_{\mu\,RMS} = \frac{0.0257 \text{ x } \sqrt{0.4}}{\sqrt{3}} = 0.0093\text{A}$$

With a current density of 2.5A/mm² we need wires of 0.00372 mm² (= 7.44 c.m.).
We take the finest wires possible from table 13.2 (AWG35: 31.4 c.m. = 0.0157 mm²) with 88 turns (equal number as for the primary!).

4. Total area of the bobbins After the wire sizes have been determined, it is necessary to check fit, to see if the available winding area (chosen core) will accommodate the copper calculated in the previous steps. !	**4. Total area of the bobbins** **a. Primary coil:** 2 x 88 wires AWG32 (64 c.m. = 0.032 mm^2) Bobbin area: 2 x 88 x 0.032 = 5.63 mm^2 **b. Degaussing coil:** 88 wires AWG35 (31.4 c.m. = 0.0157 mm^2) Bobbin area: 88 x 0.0157 = 1.38 mm^2 **c. Secondary coil:** 3 x 13 wires AWG24 (404 c.m. = 0.2 mm^2) Bobbin area: 3 x 13 x 0.2 = 7.8 mm^2 **d. Total area:** 5.63 + 1.38 + 7.8 = 14.81 mm^2 + 50% (insulation, air between wires): 0.222 cm^2. This is smaller than the allowed area A_w for the EFD20 core. OK!
5. Choke *L* By analogy to the buck converter it follows: $L_{max} = \frac{V_o \,.\, T \;.\, (1-\delta_{min})}{\Delta_{i\,L}}$; $\frac{\delta_{min}}{\delta_{max}} = \frac{V_{i\,min}}{V_{i\,max}}$; *so that:* $\delta_{min} = \delta_{max} \cdot \frac{V_{i\,min}}{V_{i\,max}}$	**5. Choke *L*** $\delta_{\text{min}} = 0.4 \text{ x } \frac{210}{390} = 0.215$ $0.215 \leq \delta \leq 0.4$ These are the limits of the duty ratio to take in account the fluctuation of the input voltage between 390 and 210V while the converter is maximum loaded (I_o = 2A). With a smaller load, the control chain will decrease the value of δ (who corresponds with V_i) until V_o = 12V ! $\Delta i_L = 20\% \,.\, i_{s\,peak} = 0.2 \text{ x } 2.69 = 0.538\text{A}$ $L = \frac{12 \text{ x } 10^{-5} \text{ x } (1-0.215)}{0.538} = 175\mu\text{H}$
6. Determination of the smoothing capacitor $C_{min} = \frac{\Delta\, i_L}{8 \,.\, f \,.\, \Delta\, v_{C\,max}}$ ESR $\leq \frac{\Delta v_{o\,max}}{\Delta i_L}$	**6. Smoothing capacitor** $\Delta i_L = 0.538\text{A}$ $\Delta v_{C\,max} = 0.1\% \text{ x } 12 = 12\text{mV}$ $C_{min} = \frac{0.538}{8 \text{ x } 10^5 \text{ x } 12 \text{ x } 10^{-3}} = 56\mu\text{F}$ ESR $\leq \frac{12 \text{ x } 10^{-3}}{0.538} = 0.022\Omega$

14.4 Switchmode transformer ferrite E-core series EP and EFD (Coilcraft)

The EP and EFD-series cores are designed to about 50W or less. EP cores are suitable for lower power and offer the smallest PCB area. EFD cores offer the lowest profile (height). The recommended core sizes are shown in table 13-2.

TABLE 13-2 CORE DATA

	EP7	**EP10**	**EP13**	**EFD15**	**EFD17**	**EFD20**	**EFD25**
Power capacity @ 100kHz	10W	12W	20W	20W	25W	30W	50W
A_e (cm²) (effective cross sectional area)	0.10	0.11	0.20	0.14	0.21	0.31	0.59
l_e (cm) (mean magnetic path lenght)	1.57	1.92	2.47	3.29	3.88	4.61	5.65
A_w (cm²) (bobbin winding area)	0.045	0.122	0.141	0.173	0.198	0.286	0.4175
Required board space (mm)	13.2 x 10.9	15.2 x 12.7	17.8 x 13.5	22 x 17.2	24.1 x 17.4	30 x 20.6	32.7 x 26.8
Maximum height (mm)	9.0	11.0	12.3	8.5	10.0	11.4	14.0
Average length per turn (cm)	1.79	2.15	2.38	2.60	3.15	3.90	4.64

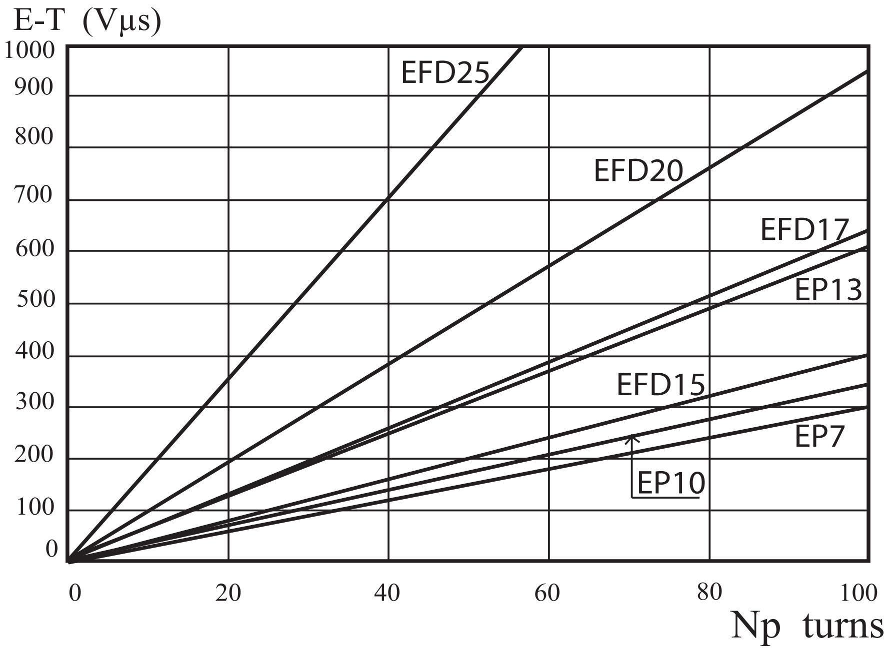

Fig. 13-27: Volt time product vs primary turns for EP and EFD cores (Coilcraft)

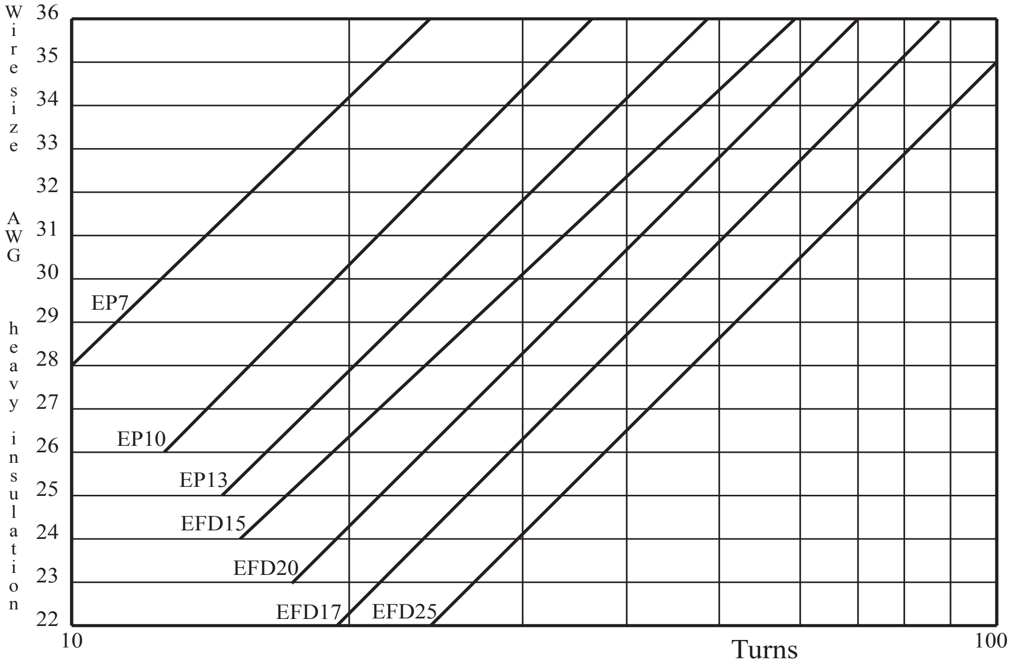

Fig. 13-28: Turns per layer versus wire size

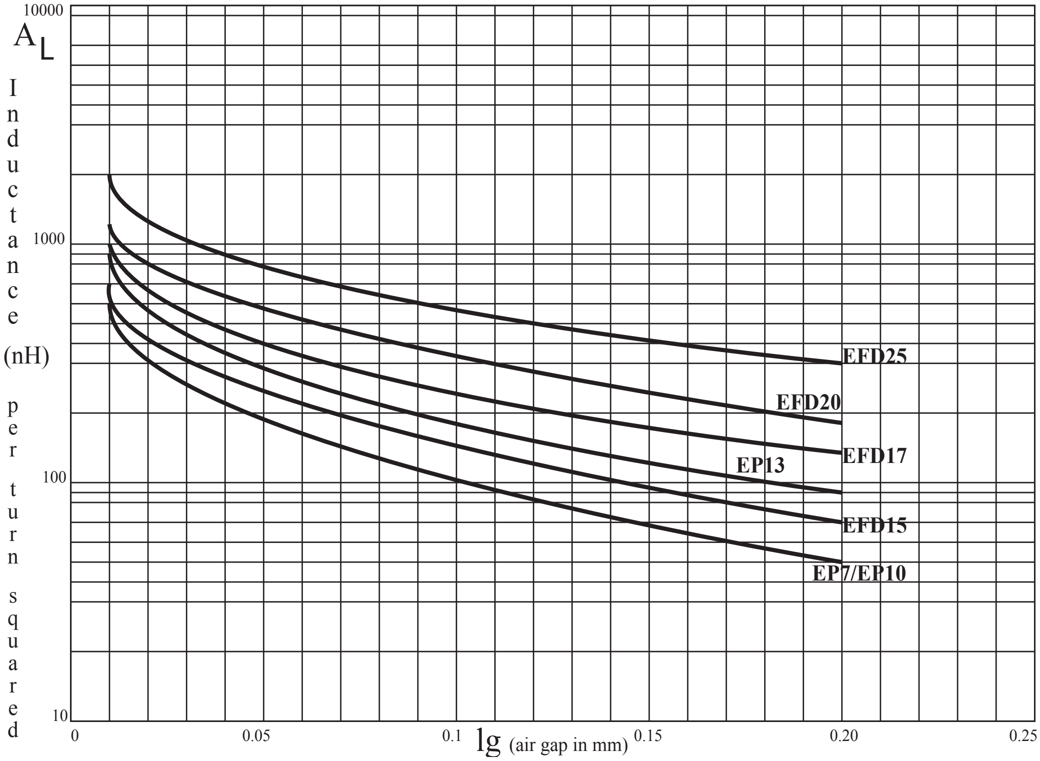

Fig. 13-29: Inductance factor A_L versus air gap l_g

14.4 Wire size (AWG = American Wire Gauge)

TABLE 13-3 Heavy insulated magnet wire specifications

AWG	Diameter over insulation (inches)		Nominal circular mil area	Resistance per 1000 ft	Current capacity in milliamperes based on 1000 c.m./A	AWG
	Min.	Max.				
8	0.130	0.133	16510	0.6281	16510	8
9	0.116	0.119	13090	0.7925	13090	9
10	0.104	0.106	10.380	0.9985	10380	10
11	0.0928	0.0948	8230	1.261	8226	11
12	0.0829	0.0847	6530	1.588	6529	12
13	0.0741	0.0757	5180	2.001	5184	13
14	0.0667	0.0682	4110	2.524	4109	14
15	0.0595	0.0609	3260	3.181	3260	15
16	0.0532	0.0545	2580	4.020	2581	16
17	0.0476	0.0488	2050	5.054	2052	17
18	0.0425	0.0437	1620	6.386	1624	18
19	0.0380	0.0391	1290	8.046	1289	19
20	0.0340	0.0351	1020	10.13	1024	20
21	0.0302	0.0314	812	12.77	812.3	21
22	0.0271	0.0281	640	16.20	640.1	22
23	0.0244	0.0253	511	20.30	510.8	23
24	0.0218	0.0227	404	25.67	404	24
25	0.0195	0.0203	320	32.37	320.4	25
26	0.0174	0.0182	253	41.02	252.8	26
27	0.0157	0.0164	202	51.44	201.6	27
28	0.0141	0.0147	159	65.31	158.8	28
29	0.0127	0.0133	128	81.21	127.7	29
30	0.0113	0.0119	100	103.7	100	30
31	0.0101	0.0108	79.2	130.9	79.21	31
32	0.0091	0.0098	64	162	64	32
33	0.0081	0.0088	50.4	205.7	50.41	33
34	0.0072	0.0078	39.7	261.3	39.69	34
35	0.0064	0.0070	31.4	330.7	31.36	35