Magnetics in SMPS Basics

Roberto Scibilia
EU Reference Design Team
Freising

Texas Instruments

Magnetic structures

- Two main different classes:
  - Inductors: energy storage
  - Transformers: coupling and isolation
- Energy is stored in Non-Magnetic regions
  - air in gapped structures or distributed air gap structures (powdered iron cores, moly-permalloy)
- Magnetics is an easy path for flux (magnetic Bus Bars)
  - Link to other core sections (transformers)
    - Electrical insulation, magnetic coupling through flux
  - Link core to gap for storage (inductors)
- From the basic relation of SMPS

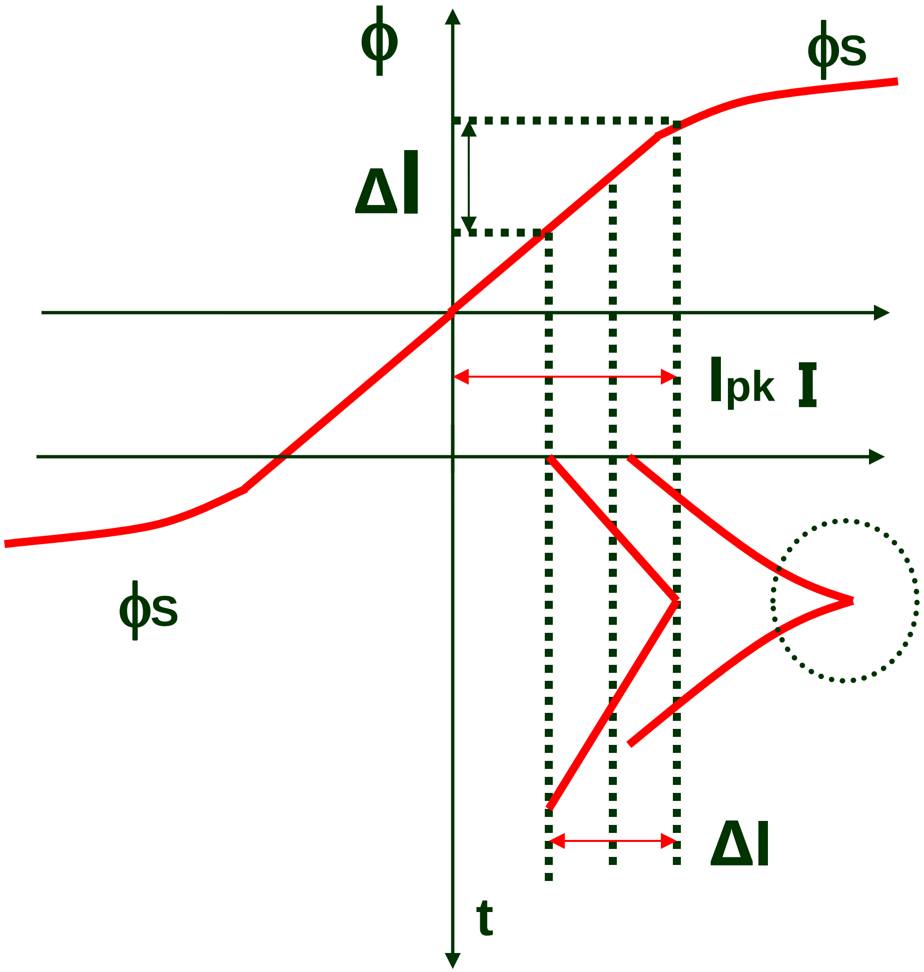
$$\frac{V}{N} \leq 4 B_S S f_S$$

by increase of f_{sw}, smaller cores (S, N) can be used with no saturation

Ferromagnetic materials

- **Spontaneous magnetization behaviour below a material-specific temperature (*Curie point*)**
- **Elementary atomic magnets are aligned in parallel within macroscopic regions, called *Weiss' domains*, normally oriented so that no magnetic effect is perceptible**
- **When a ferromagnetic body is placed in a magnetic field and the flux density B as a function of the magnetic field strength H is measured from H= 0 and B= 0, the *initial magnetization* curve is obtained.**
- **At higher field strength, whole domains overturn magnetically and finally the magnetic moments are moved into the direction of the field until *saturation* is obtained (all elementary magnets in the material are in the direction of the field)**
- **If H is now reduced again, the B curve is completely different. The relationship between H and B is shown in the *hysteresis loop***
- **Most used material at high switching frequency : ferrite (MnZn)**

Flux capability and Saturation



- Full utilization of flux capability

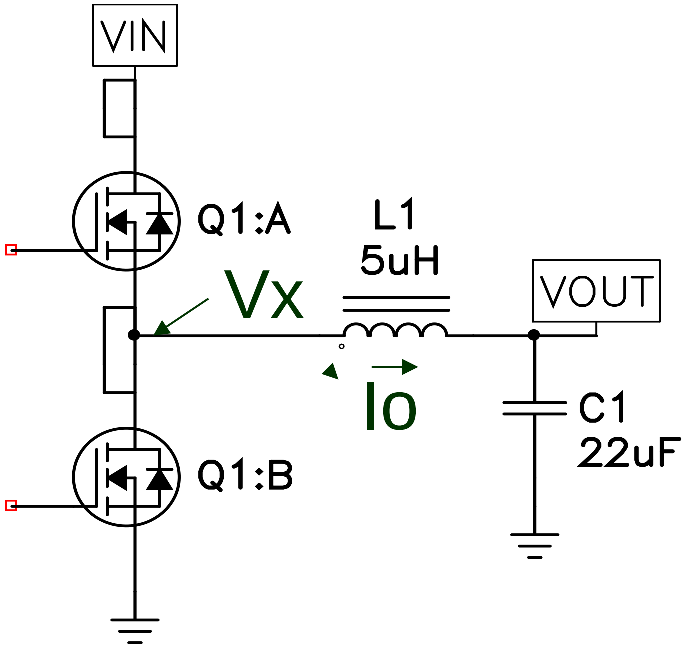
$$\lambda_{TOT} \leq \lambda_S \Rightarrow$$

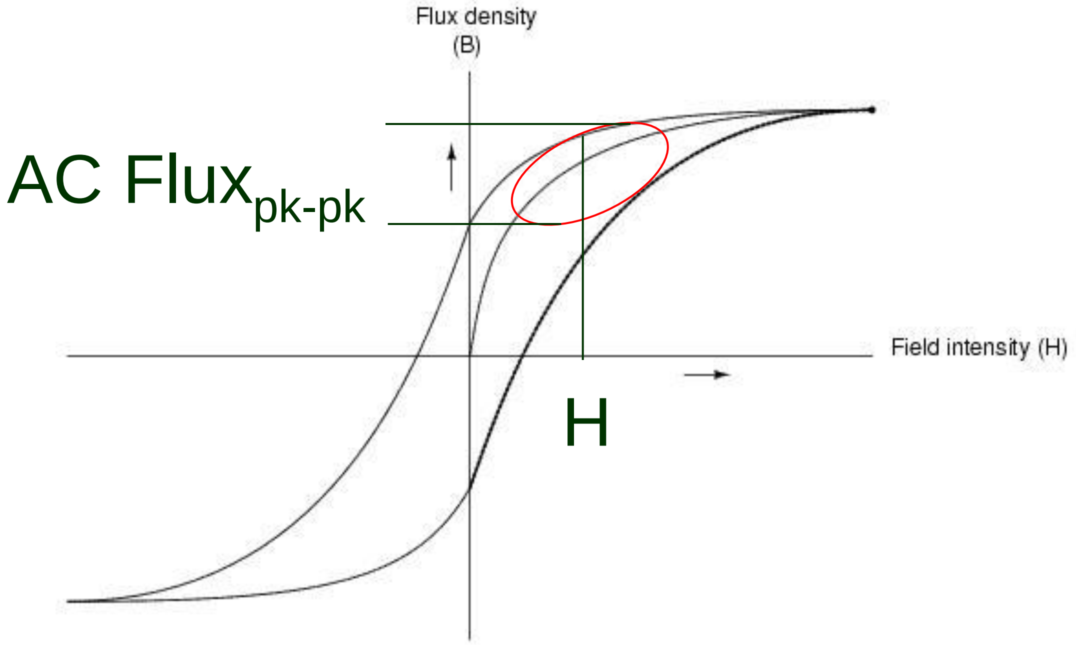
$$N \geq \frac{LI_{PK}}{B_s S}$$

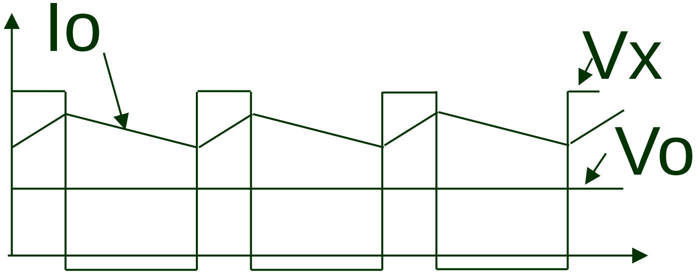
$$LI_{PK} \leq B_S SN$$

- DC current with superimposed triangular current ripple **ΔI**
- Current spikes indicate approach to **saturation**

Buck Inductor

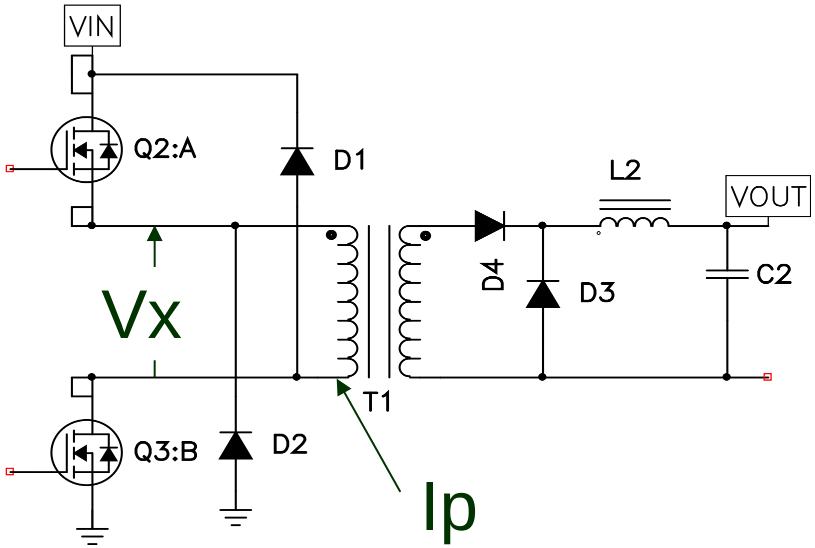


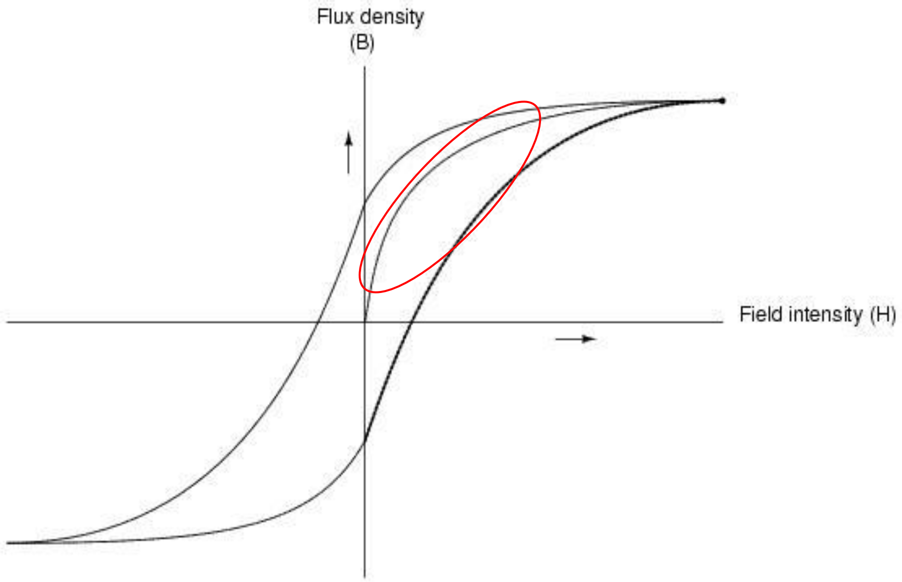


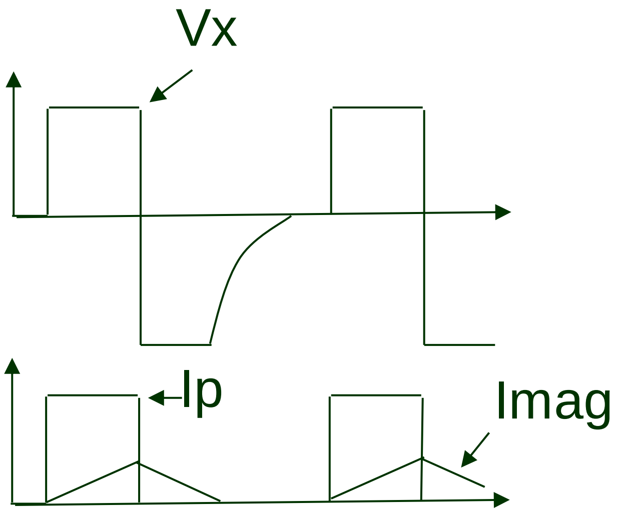


Quadrants	**1**
DC Magnetic Field (H)	**N * Io / lm**
AC Pk to Pk Current	**20% DC Typical**
AC Flux$_{pk\text{-}pk}$	**(1-D) * Vo / (Fs * Ac *N)**

2 Transistor Forward Transformer

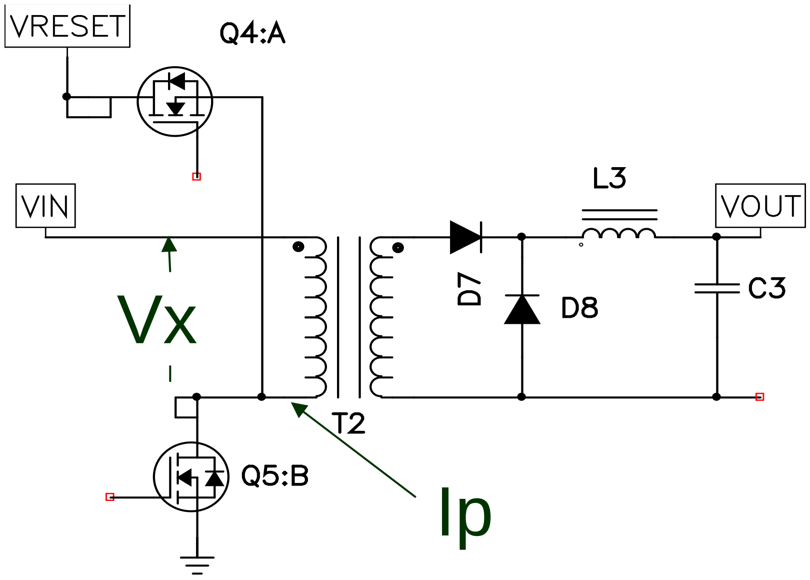


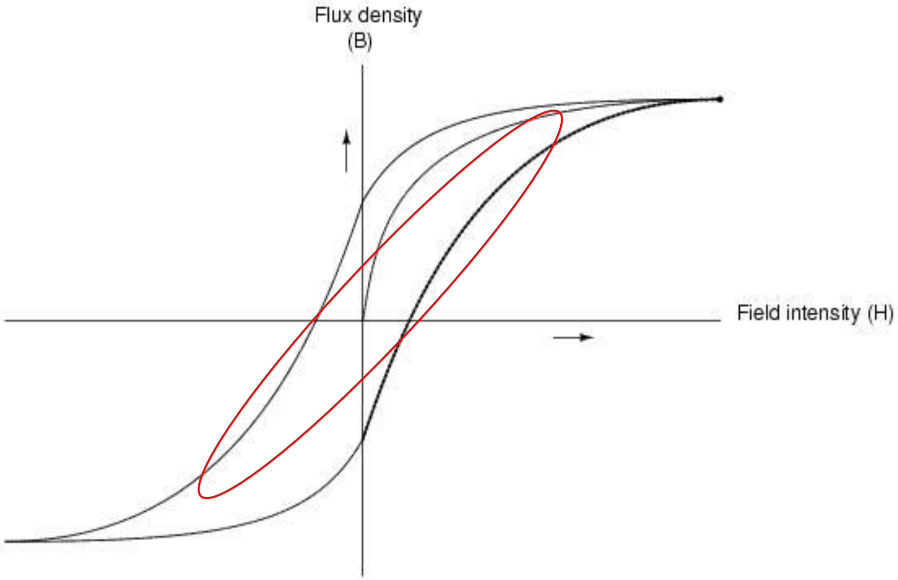


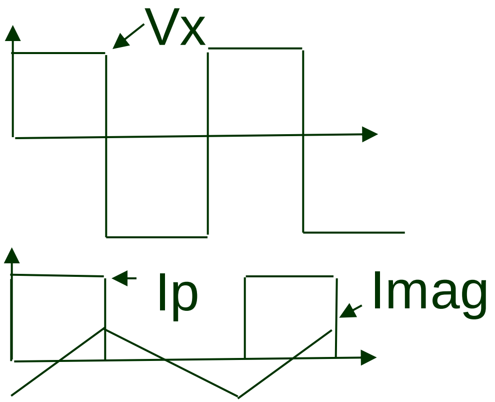


Quadrants	**1**
Magnetic Field (H)	**N * Imag / Im**
Primary AC Current	**$D^{1/2}$ * Ip**
AC Flux$_{pk-pk}$	**D * Vin / (Fs * Ac *N)**

Active Clamp Forward Transformer

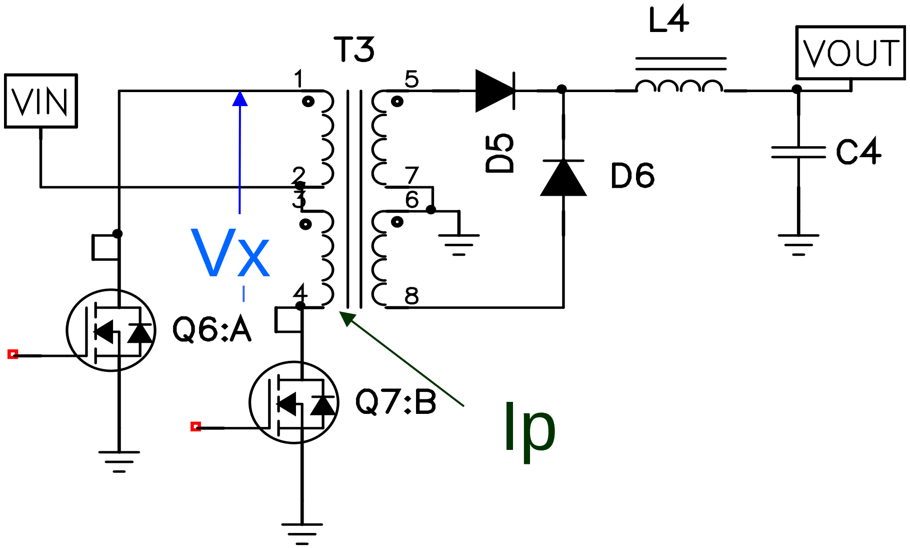


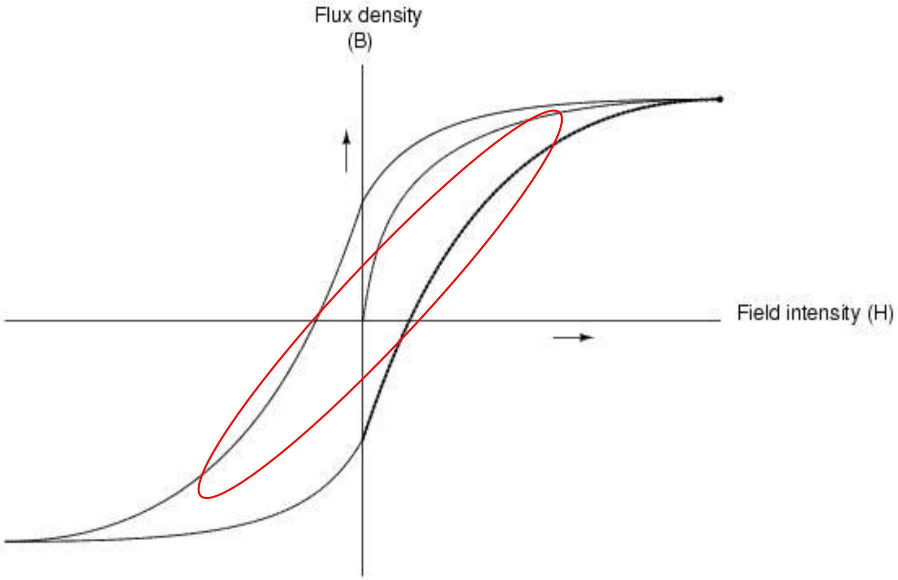


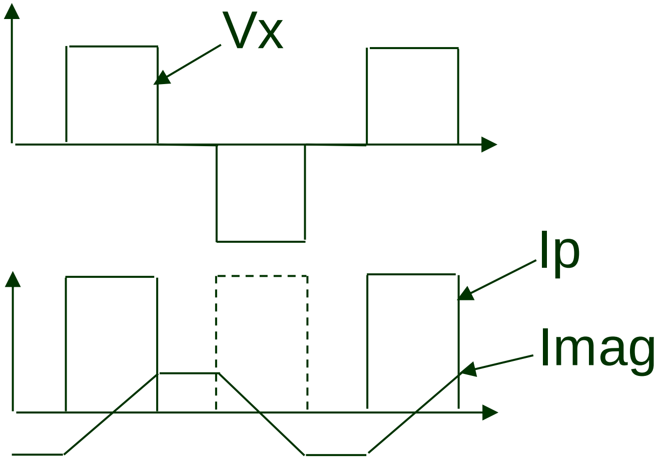


Quadrants	**2**
Magnetic Field (H)	**N * Imag / Im**
Primary AC Current	**$D^{1/2}$ * Ip**
AC Flux_{pk-pk}	**D * Vin / (Fs * Ac *N)**

Push-Pull Forward Transformer

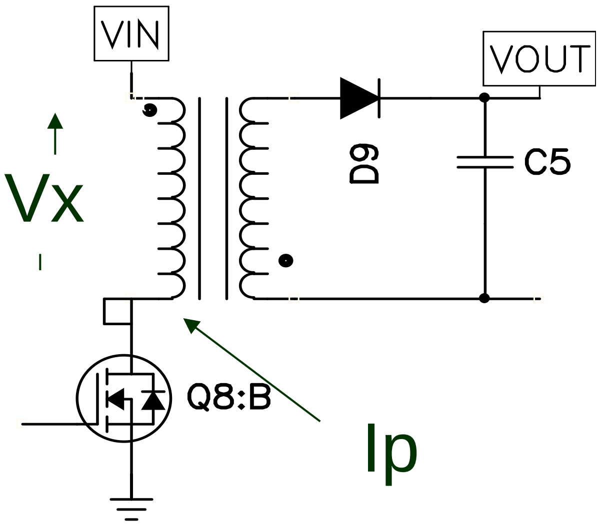


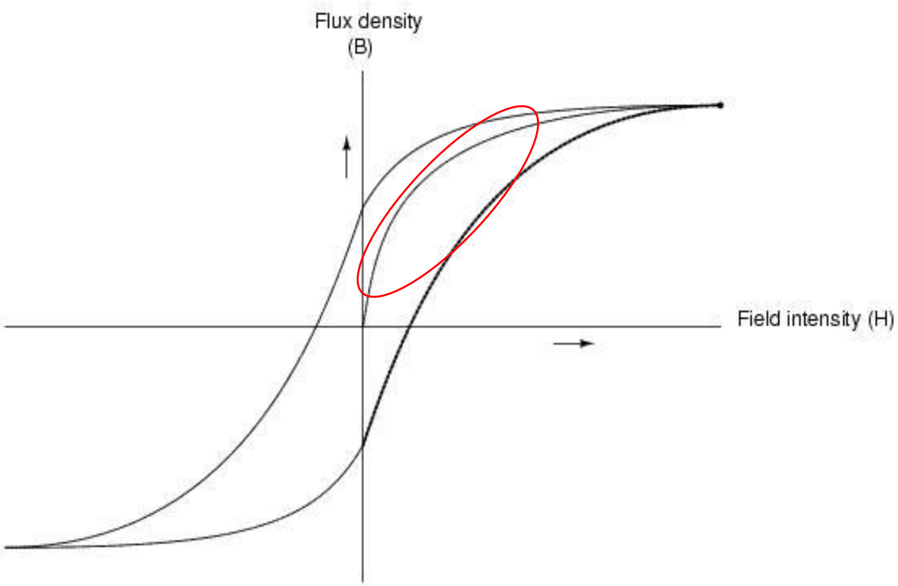


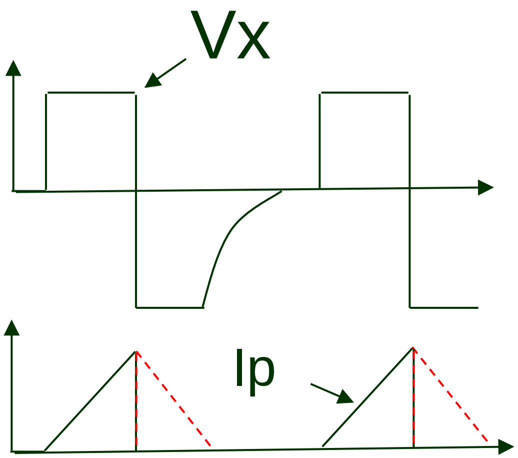


Quadrants	**2**
Magnetic Field (H)	**N * Imag / Im**
Primary AC Current	**$D^{1/2}$ * Ip**
AC $\text{Flux}_{pk\text{-}pk}$	**D * Vin / (Fs * Ac *N)**

Discontinous Flyback Transformer

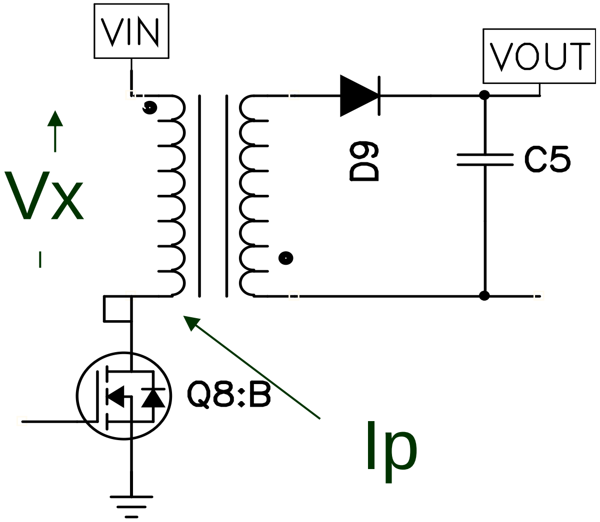


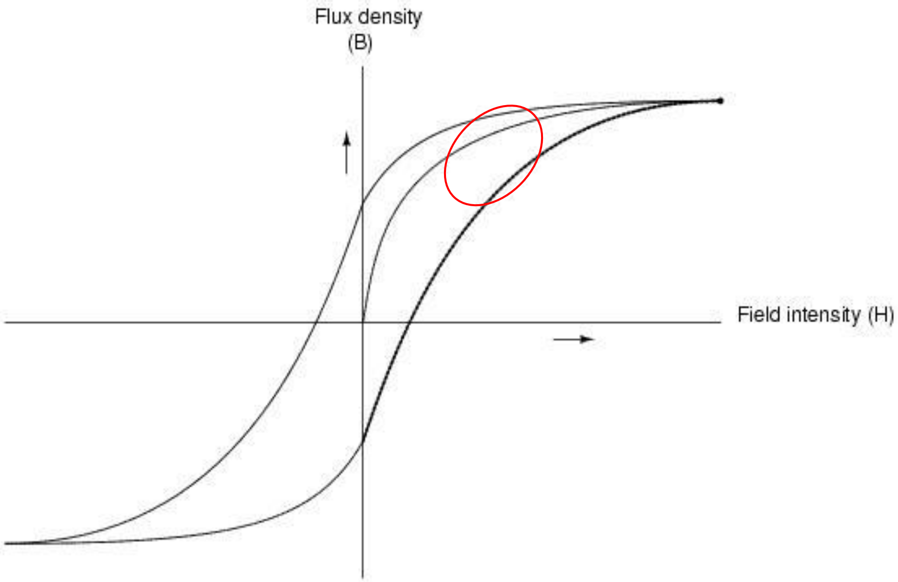


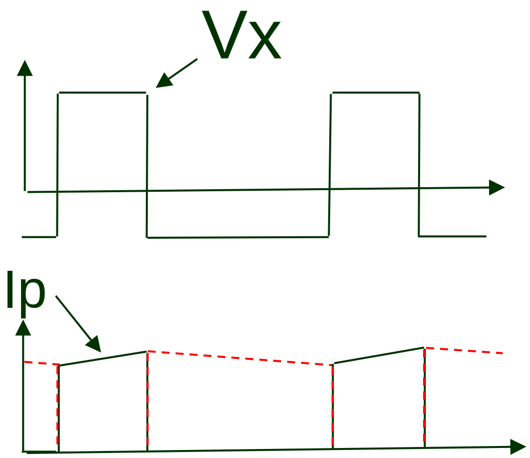


Output Power	$\frac{1}{2} * L * Ip^2$
Quadrants	**1**
Magnetic Field (H)	**N * Imag / lm**
Primary AC Current	$(D/3)^{1/2} * Ip$
AC Flux$_{pk-pk}$	**D * Vin / (Fs * Ac *N)**
AC Flux$_{pk-pk}$	**L * Ip / (Ac *N)**

Continous Flyback Transformer





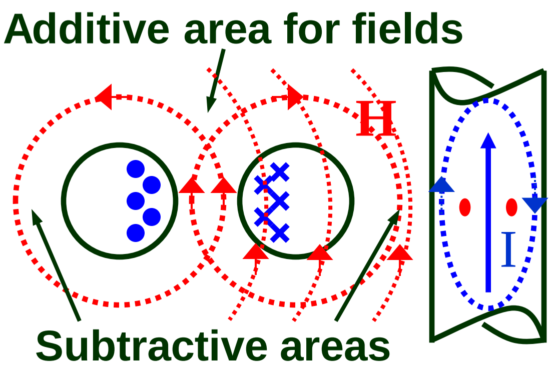


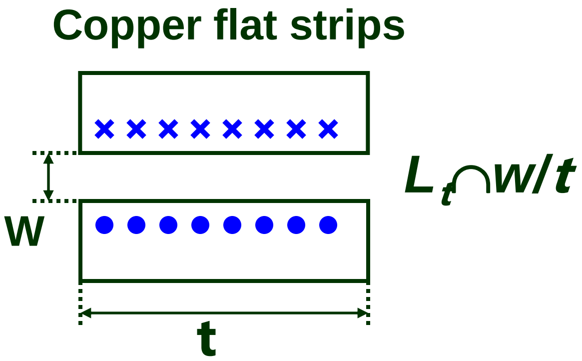
Quadrants	**1**
Magnetic Field (H)	**N * Imag / Im**
Primary AC Current	**$(D)^{1/2}$ * Ip**
AC Flux$_{pk-pk}$	**D * Vin / (Fs * Ac *N)**
AC Flux$_{pk-pk}$	**L * ΔIp / (Ac *N)**

Copper losses (Parasitics effects) (2/2)

- **PROXIMITY effect**
  - Circulating currents with opposite directions in adjacent conductors induces current concentrations in the adjacent sides of conductors
  - Again Eddy currents, induced by the current in an adjacent conductor
  - Field in between conductors store energy in parasitic parameter, named Leackage Inductance

$R_{ac}(f_{sw}) >> R_{dc}$

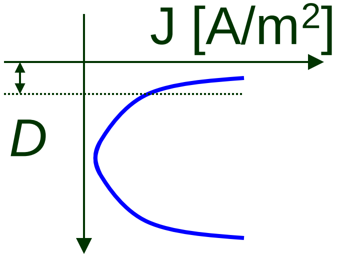




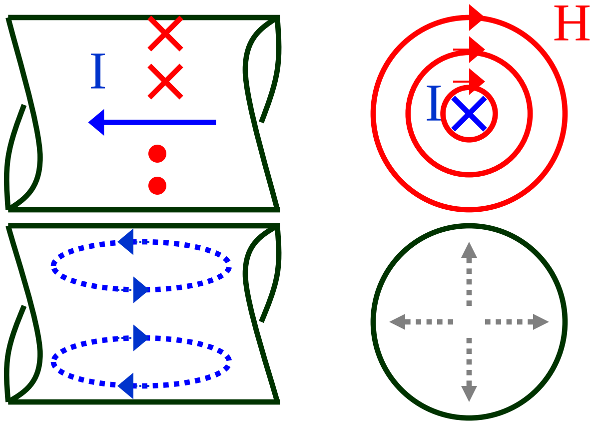
  - Resistance Factor, AC to DC Resistance of conductors Ratio is a function of *D*, of spacing and shape of the conductors in a layer and the number of layers in the winding portion
    - Phisically cutting in more portions (i.e interleaving) reduces eddy currents

Copper losses – Skin Effect

- **SKIN effect**
  - **Circulating current induces magnetic field inside wires and therefore other parasitic currents, resulting in a global effect of current concentrations in the external section of wire**
  - **Eddy currents:**
    - **induced in a conductor by the current in that conductor**
  - **D=Penetration depth, variable with switching frequency (here neglecting harmonics)**



$$D = \frac{7.5}{\sqrt{f_{sw}}}\left[cm\right]$$



Losses in Magnetics

- **CORE Losses**
  - **Hysteresis ($\frown f_{sw}$) + Eddy Currents ($\frown f_{sw}^{2}$)**

$$\frac{W}{V_e} = \int HdB \left[\frac{Joule}{m^3} \right]$$

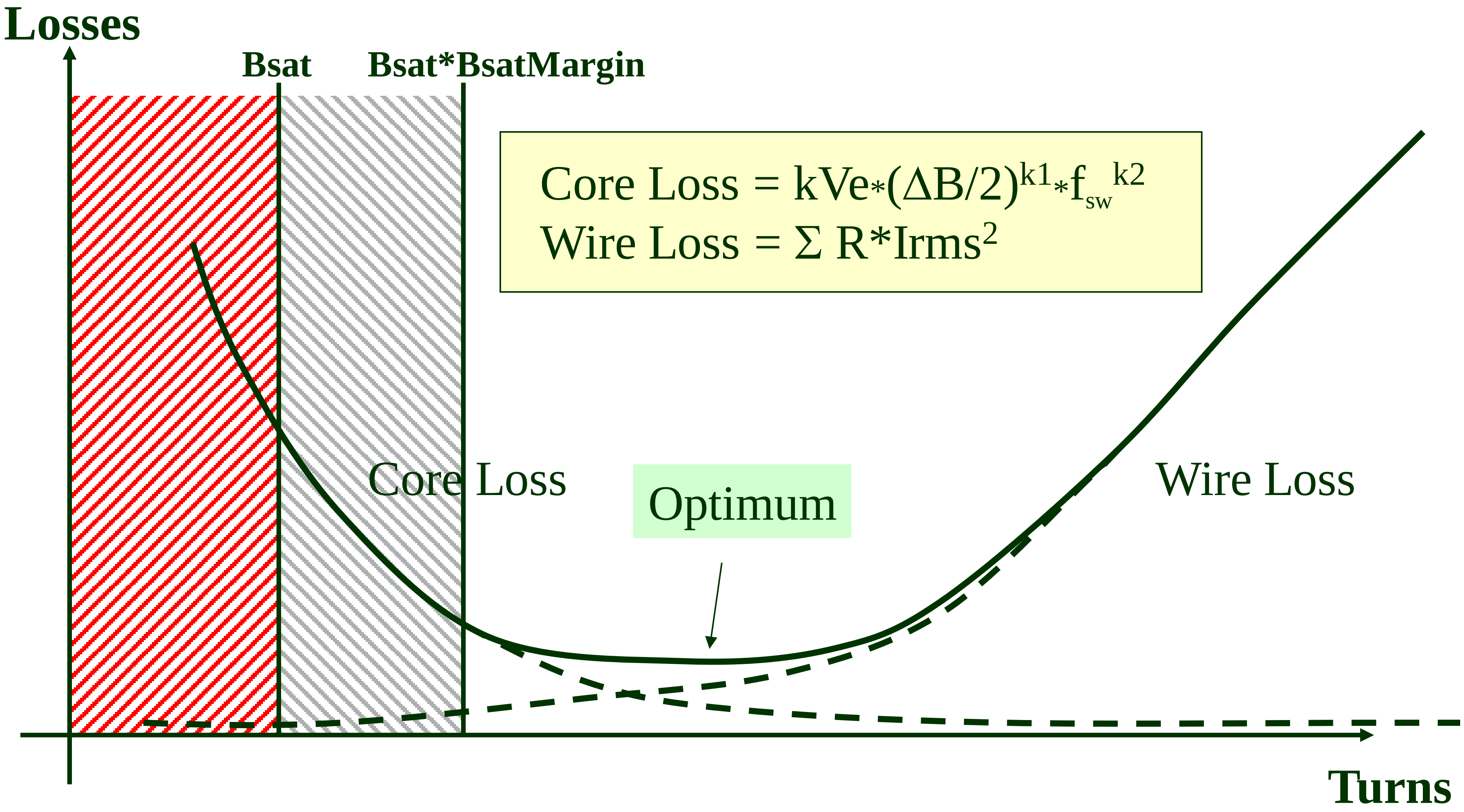
$$Core\ Losses = P_{fe} = kVe * (\Delta B/2)^{k1} * f_{sw}^{\ k2}$$

  - **K depends on the material and temperature; $2<K1<2.7$; $1.2<K2<1.7$**
  - **Temperature rise:** $\Delta T = T_x - T_{amb} = P_{fe} R_{\theta}$

- **COPPER Losses**
  - **Skin and Proximity effects**
  - **Resistance increases with frequency for equivalent reduction of wires section**

$$Copper\ Losses = P_{cu} = \Sigma R * Irms^2$$

Power Losses



$$\text{Core Loss} = kVe*(\Delta B/2)^{k1}*f_{sw}^{k2}$$

$$\text{Wire Loss} = \Sigma\, R*Irms^2$$

-To Lower Core Losses, Minimize Magnetic Field Variation i.e. Increase Turns

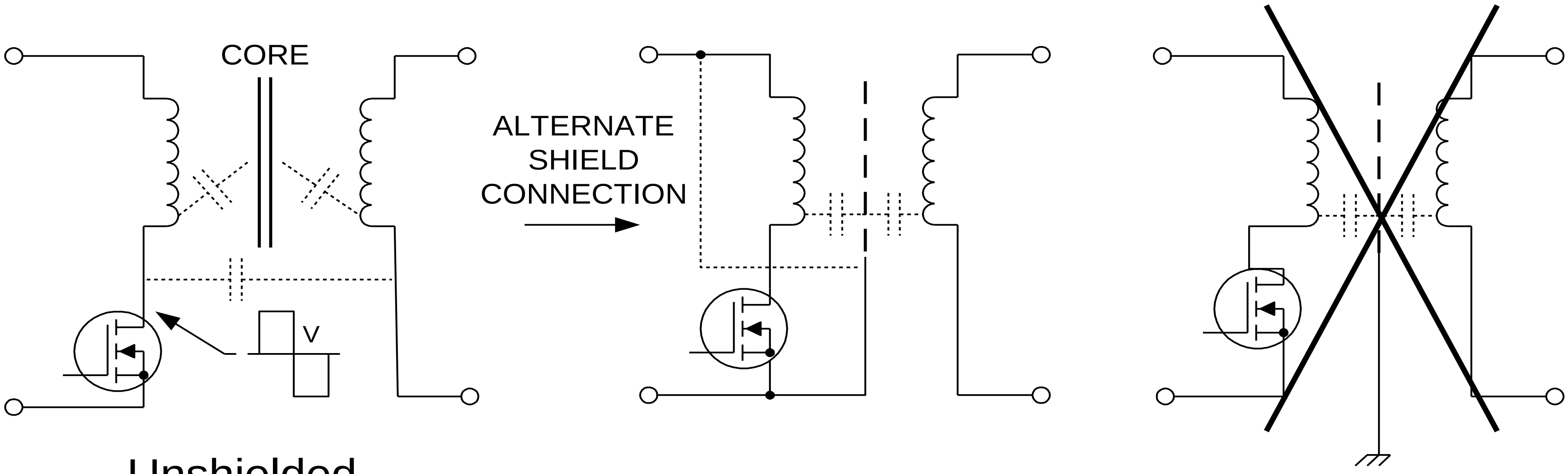
-But more Turns means Lower Diameter and Longer Winding i.e. Higher Resistance and Losses

LEAKAGE INDUCTANCE EFFECTS

- LEAKAGE INDUCTANCE IS THE MAIN CAUSE OF POOR REGULATION
  - LOAD REGULATION, CROSS-REGULATION
  - GREATER EFFECT THAN CIRCUIT RESISTANCES
- POWER LOSS IN SNUBBERS, CLAMPS
- REDUCED OUTPUT PULSE WIDTH
  - REQUIRES HIGHER PEAK VOLTAGE, CURREN

Electrostatic Shielding

Use of Primary Shield



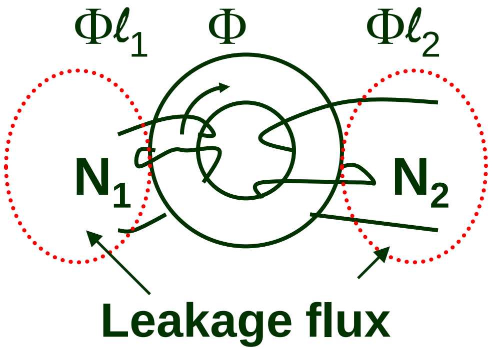
Unshielded Transformer

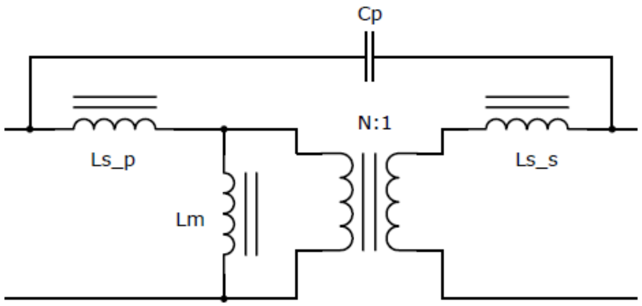
Correct

Incorrect!

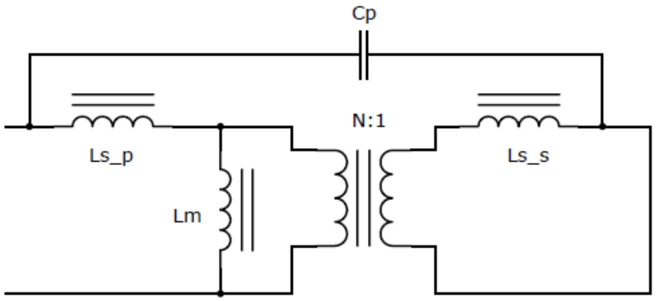
- Also called a Faraday shield
- Connect to V+ if turn-off is fastest, to return with faster turn-on

Modelling of Transformer

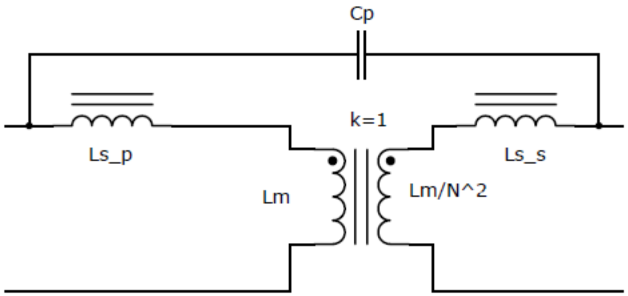




- Electrical Model
- Measuring the magnetising inductance
- Measuring parasitic capacitance



- Measuring leakage inductance



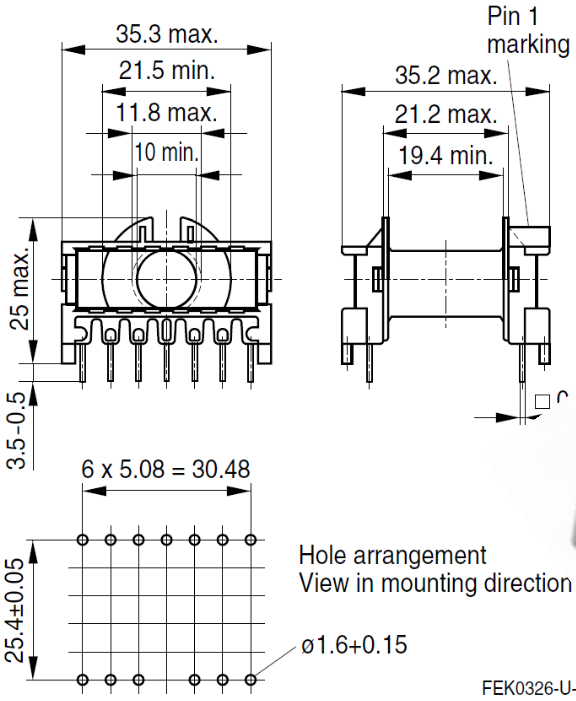
- SPICE Model

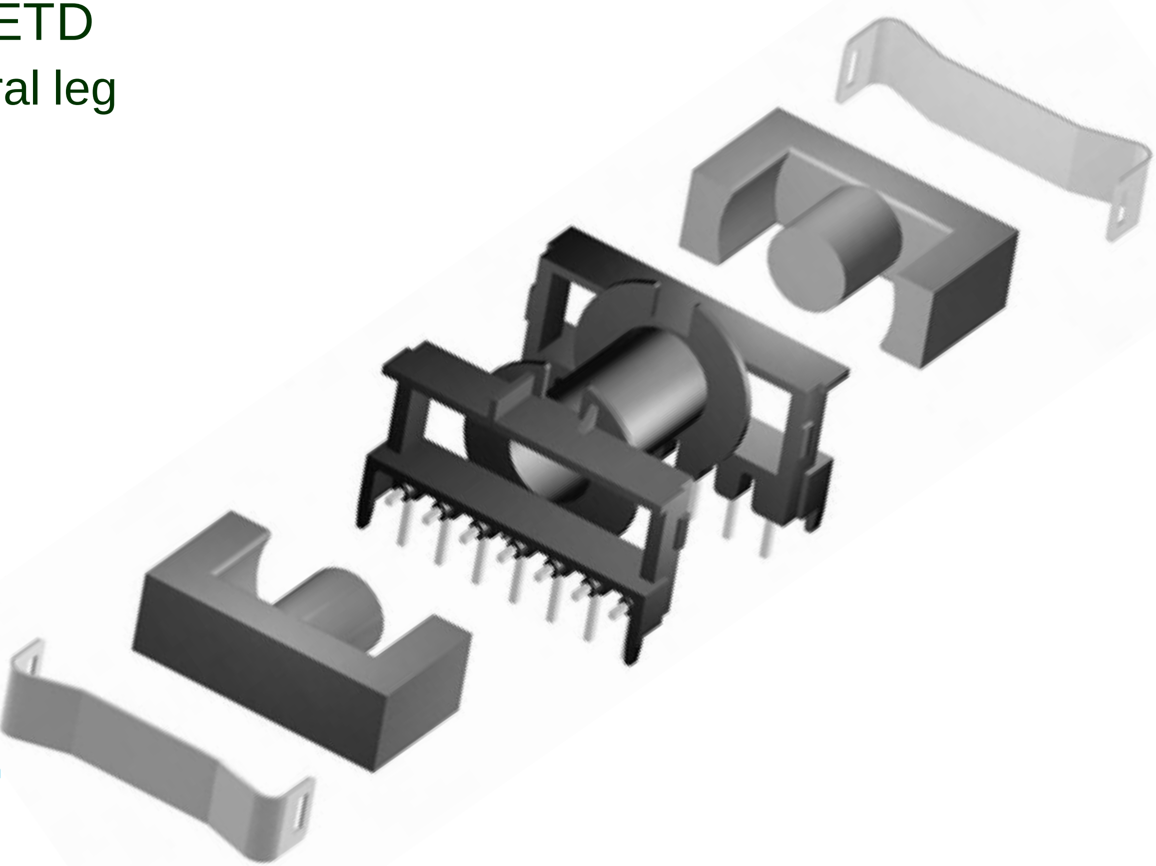
Example of Inductor Design

- We want to build an inductor with the following characteristics:
- L = 185uH
- DC current = 4A
- Peak to peak ripple = 1.4A (35% pk-pk)
- Switching frequency 100KHz
- ETD platform

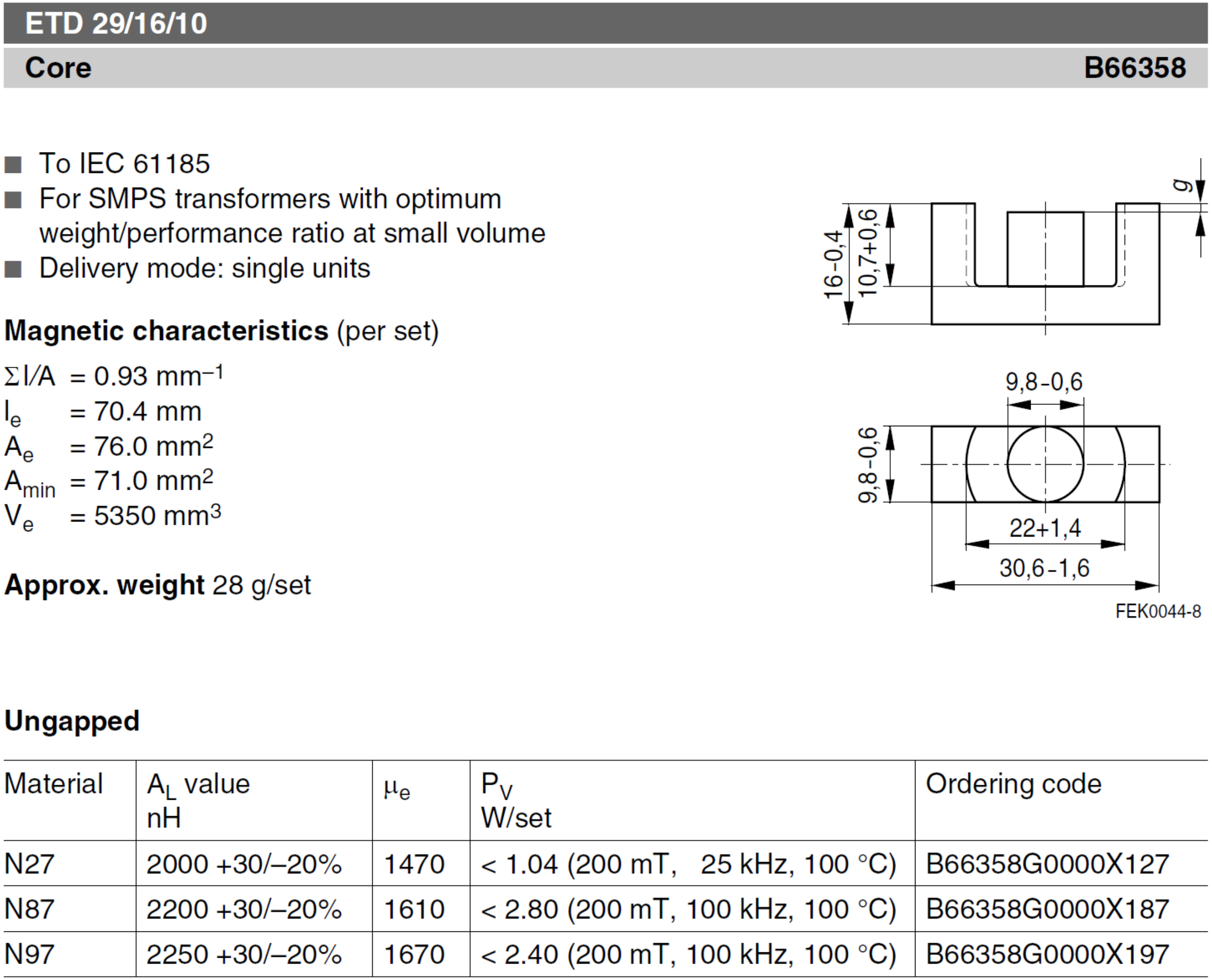
Inductor Design
Core Geometry Approach (Example)

- Mechanical structure ETD
  - core gapped in central leg





Inductor Design
Core Geometry Approach (ETD core)



ETD 29/16/10

Core **B66358**

- To IEC 61185
- For SMPS transformers with optimum weight/performance ratio at small volume
- Delivery mode: single units

Magnetic characteristics (per set)

$\Sigma l/A = 0.93\ \text{mm}^{-1}$
$l_e = 70.4\ \text{mm}$
$A_e = 76.0\ \text{mm}^2$
$A_{min} = 71.0\ \text{mm}^2$
$V_e = 5350\ \text{mm}^3$

Approx. weight 28 g/set

Ungapped

Material	A_L value nH	μ_e	P_V W/set	Ordering code
N27	2000 +30/–20%	1470	< 1.04 (200 mT, 25 kHz, 100 °C)	B66358G0000X127
N87	2200 +30/–20%	1610	< 2.80 (200 mT, 100 kHz, 100 °C)	B66358G0000X187
N97	2250 +30/–20%	1670	< 2.40 (200 mT, 100 kHz, 100 °C)	B66358G0000X197

Example of Inductor Design

- In order to prevent saturation, you have to know (almost) only one formula:

$$N \geq \frac{LI_{PK}}{B_s A_e}$$

Where:

L = inductance

N = number of turns

I_{pk} = DC + ripple current through the inductor

B_s = peak flux (keep it < 300mT)

A_e = effective core area

Here we have L=185uH, I_{pk} = 4.7A, B_s = 300mT, A_e = 76.2 mm²

Which means:

$$N \geq 38 turns$$

Inductor Design: basic formulas

But since: $L = A_L * N^2$ (with N=38, L=185uH)

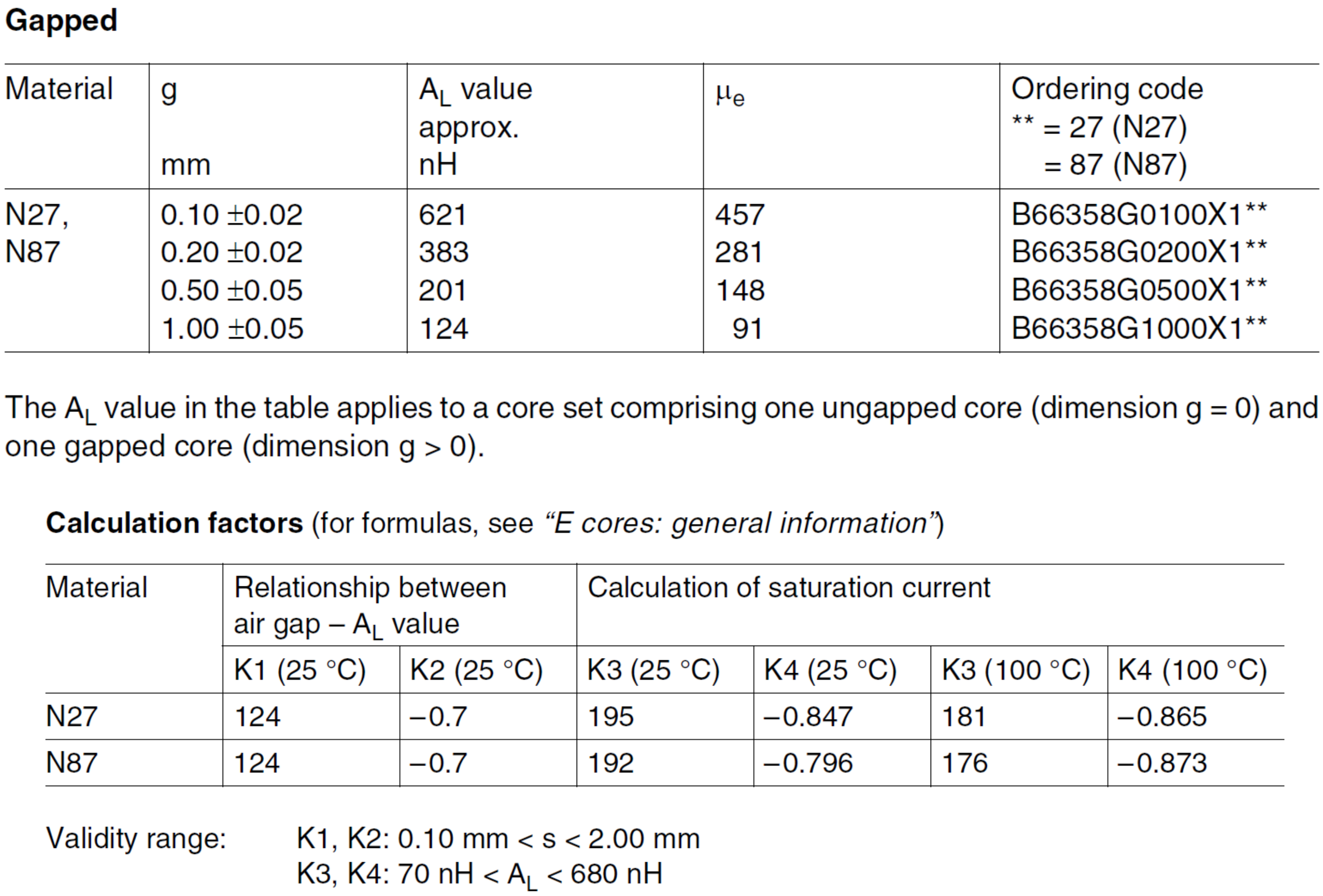
$$A_L = \frac{L}{N^2} = 128 nH / turns^2$$

Where: A_L = inductance / square turns

So, we chose as first shot $A_L = 128$

Now let's read on the datahseet which gap should we use.

Inductor Design: material details



Gapped

Material	g mm	A_L value approx. nH	μ_e	Ordering code ** = 27 (N27) = 87 (N87)
N27, N87	0.10 ±0.02	621	457	B66358G0100X1**
	0.20 ±0.02	383	281	B66358G0200X1**
	0.50 ±0.05	201	148	B66358G0500X1**
	1.00 ±0.05	124	91	B66358G1000X1**

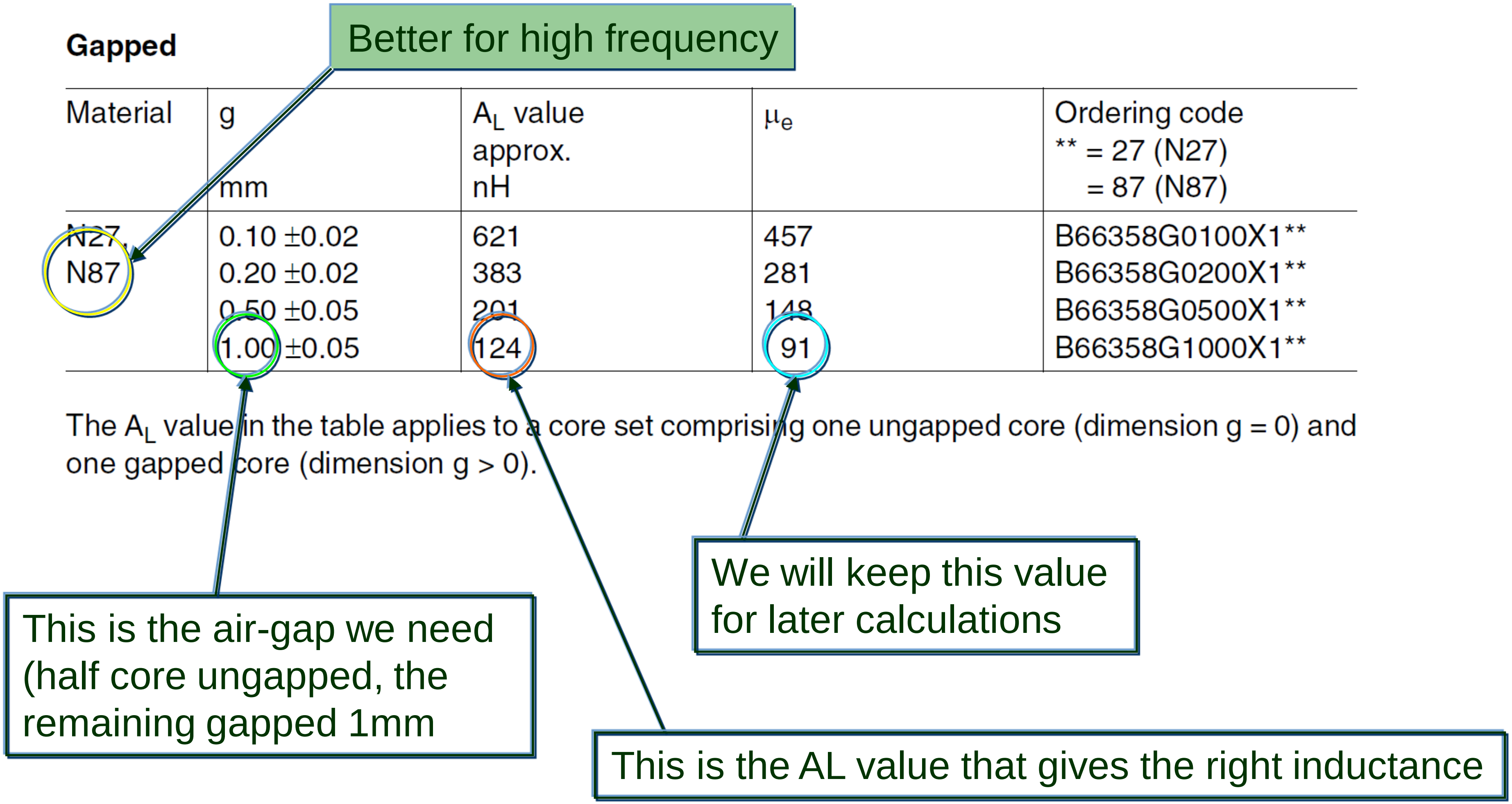
The A_L value in the table applies to a core set comprising one ungapped core (dimension g = 0) and one gapped core (dimension g > 0).

Calculation factors (for formulas, see *"E cores: general information"*)

Material	Relationship between air gap – A_L value		Calculation of saturation current			
	K1 (25 °C)	K2 (25 °C)	K3 (25 °C)	K4 (25 °C)	K3 (100 °C)	K4 (100 °C)
N27	124	–0.7	195	–0.847	181	–0.865
N87	124	–0.7	192	–0.796	176	–0.873

Validity range:
K1, K2: 0.10 mm < s < 2.00 mm
K3, K4: 70 nH < A_L < 680 nH

Inductor Design: core choice



Better for high frequency

Gapped

Material	g mm	A_L value approx. nH	μ_e	Ordering code ** = 27 (N27) = 87 (N87)
N27, N87	0.10 ±0.02	621	457	B66358G0100X1**
	0.20 ±0.02	383	281	B66358G0200X1**
	0.50 ±0.05	201	148	B66358G0500X1**
	1.00 ±0.05	124	91	B66358G1000X1**

The A_L value in the table applies to a core set comprising one ungapped core (dimension g = 0) and one gapped core (dimension g > 0).

This is the air-gap we need (half core ungapped, the remaining gapped 1mm

This is the AL value that gives the right inductance

We will keep this value for later calculations

Inductor Design

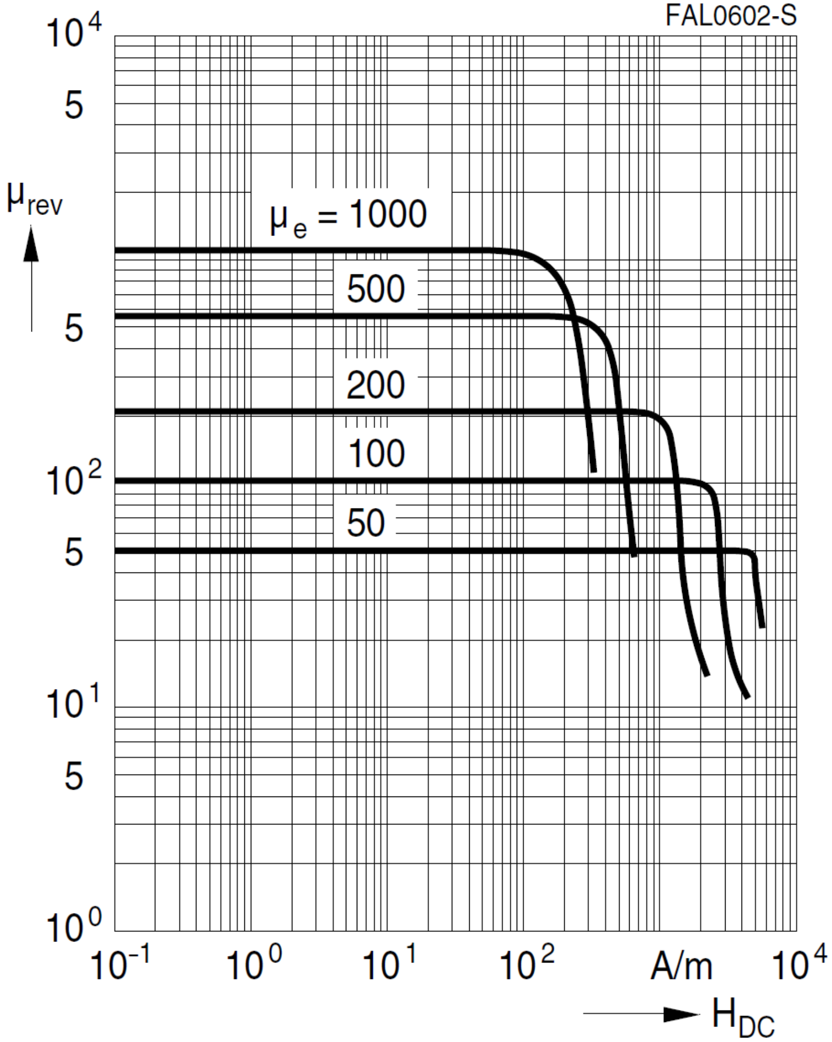
- So, we chose as first shot:
- AL = 128
- Gap = 1mm
- $\mu_e = 91$

Now let’s check if we are close to saturation (we should not, theoretically)

$$H = \frac{N \cdot I_{pk}}{Le} = 2537\ A \cdot Turns/m$$

•L_e = effective length of the magnetic path

DC magnetic bias
of P, RM, PM and E cores
($\hat{B} \leq 0.25$ mT, f = 10 kHz, T = 100 °C)

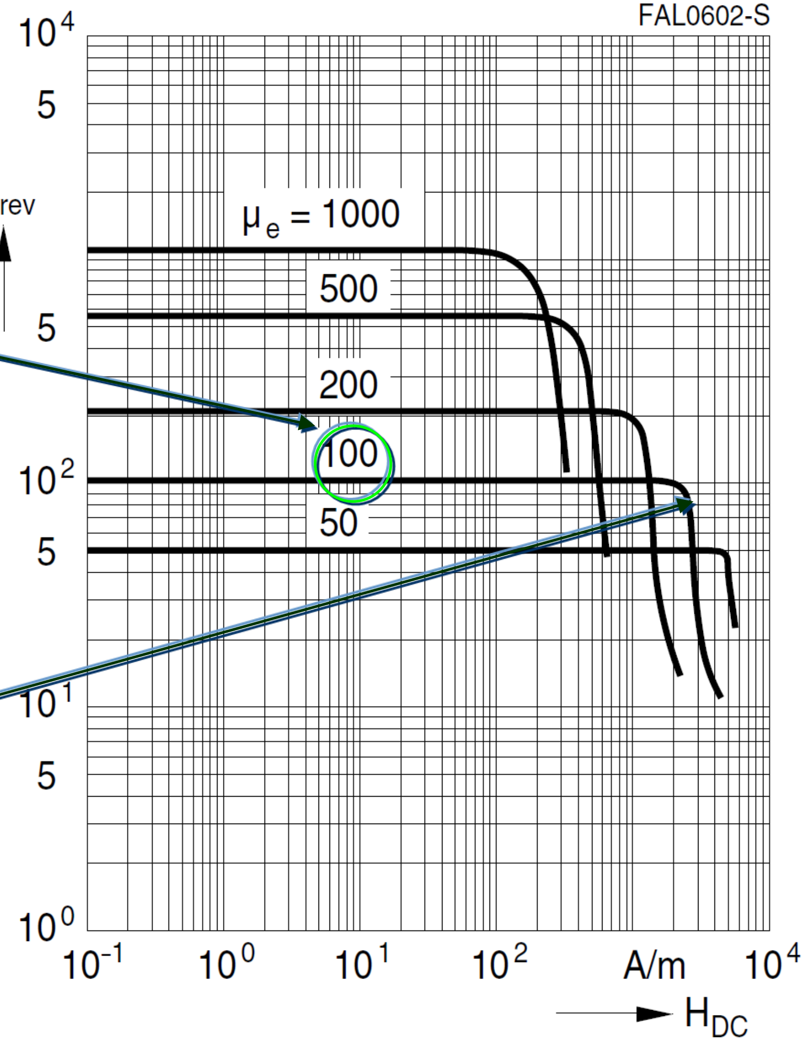


Inductor Design

- So, we chose as first shot:
- AL = 128
- Gap = 1mm
- $\mu_e = 91$

It looks like we are close to saturation, so we should take a higher number of turns and increase the gap.

$$H = \frac{N \cdot I_{pk}}{Le} = \boxed{2537\,A} \cdot Turns/m$$



DC magnetic bias
of P, RM, PM and E cores
($\hat{B} \leq 0.25$ mT, f = 10 kHz, T = 100 °C)

Inductor Design: core losses

Now let’s calculate the core losses:

The peak to peak ripple is 1.4A, so 700mA peak value (we have to consider the peak value for core losses calculations)

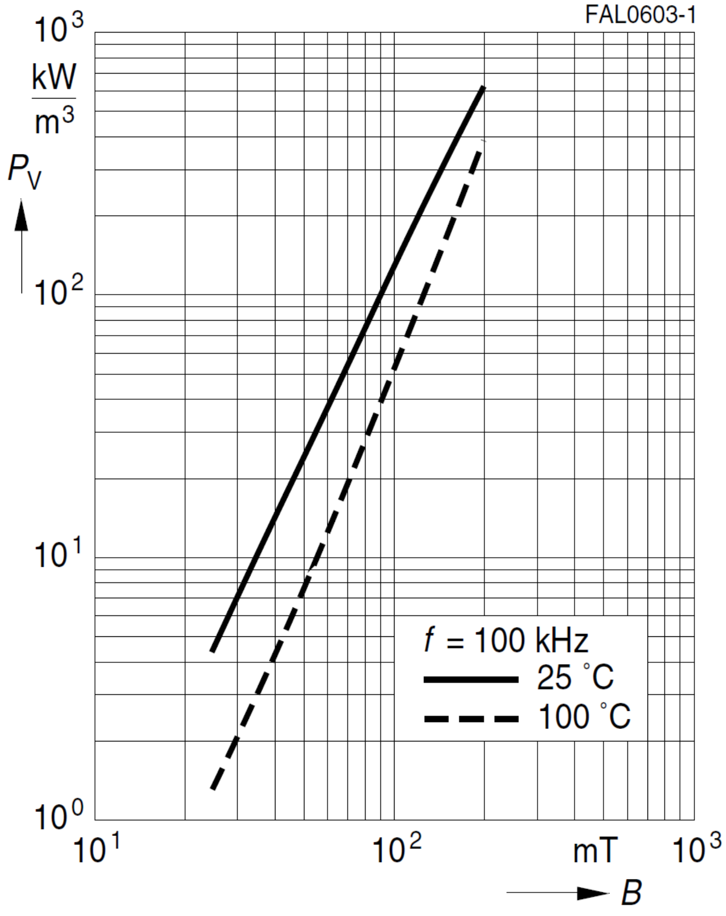
$$\Delta H = \frac{N \cdot \Delta I_{pk}}{Le} = 377.8 \qquad \Delta B = \mu_0 \cdot \mu_r(H) \cdot \Delta H = 43.2mT$$

Where:

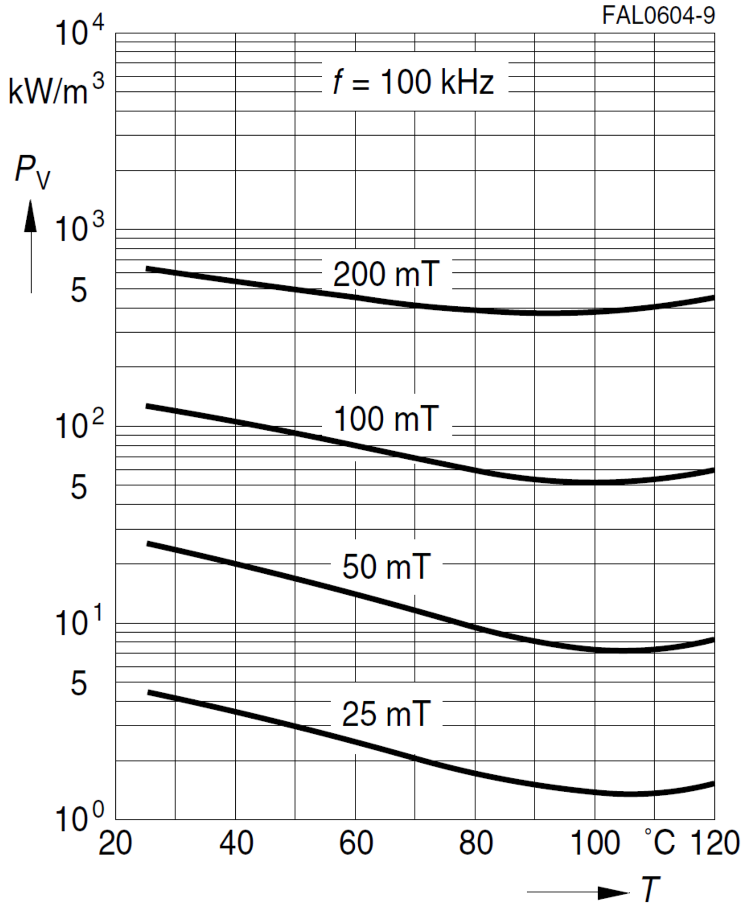
- μ_0 = 1.257 E-6 = magnetic field constant
- μ_r = μe = 91 (if we are not in saturation)

Inductor Design: core losses

Relative core losses
versus AC field flux density
(measured on R34 toroids)

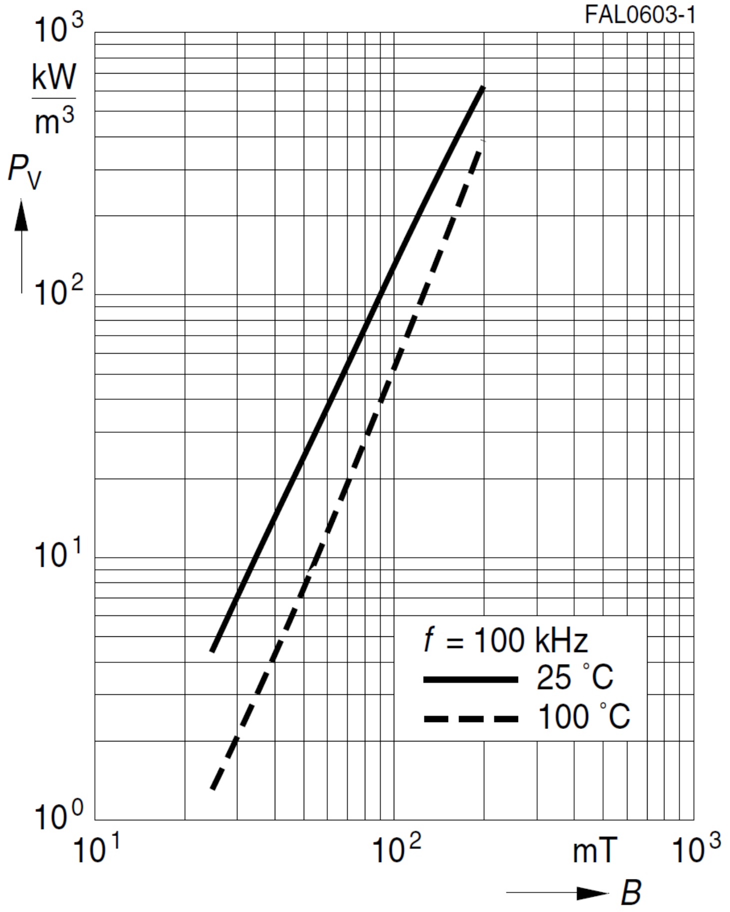


Relative core losses
versus temperature
(measured on R34 toroids)



Inductor Design

Relative core losses
versus AC field flux density
(measured on R34 toroids)



Relative core losses
versus temperature
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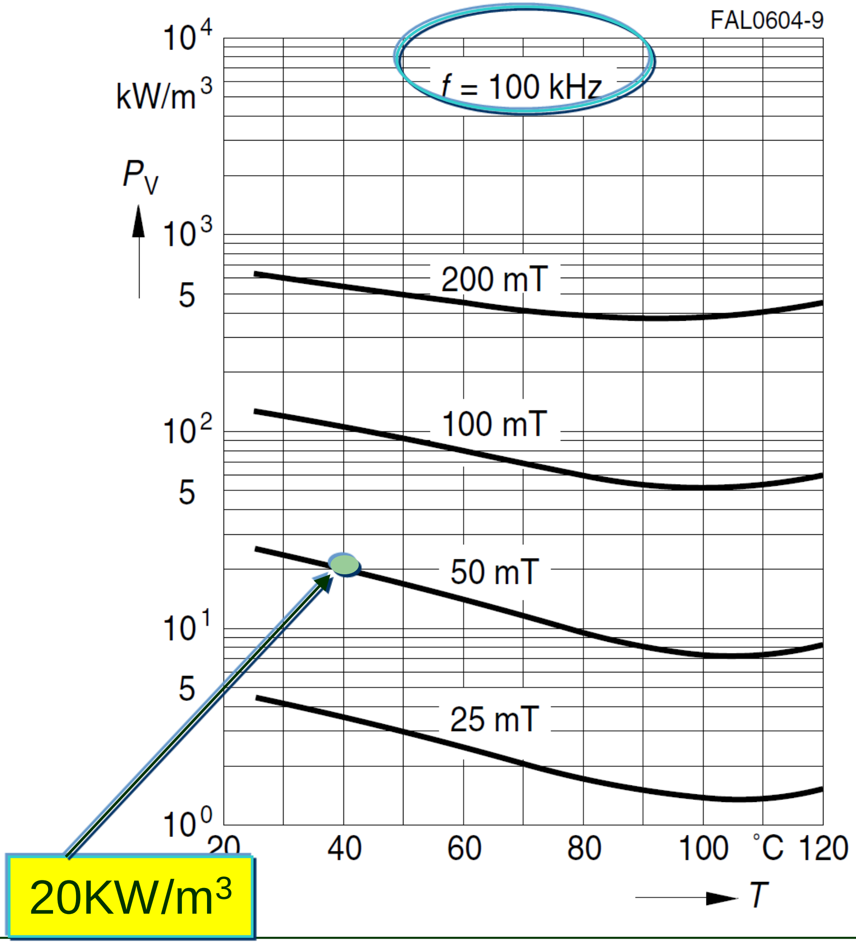


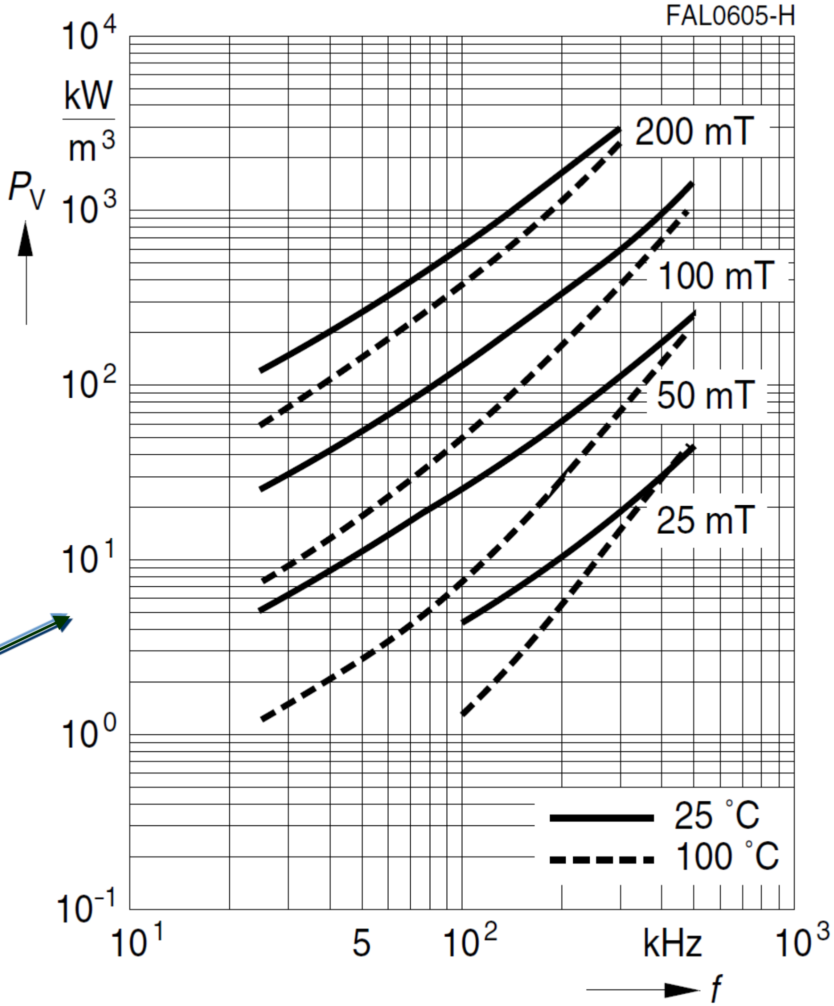
20KW/m³

Inductor Design: core losses

Core losses calculation result:

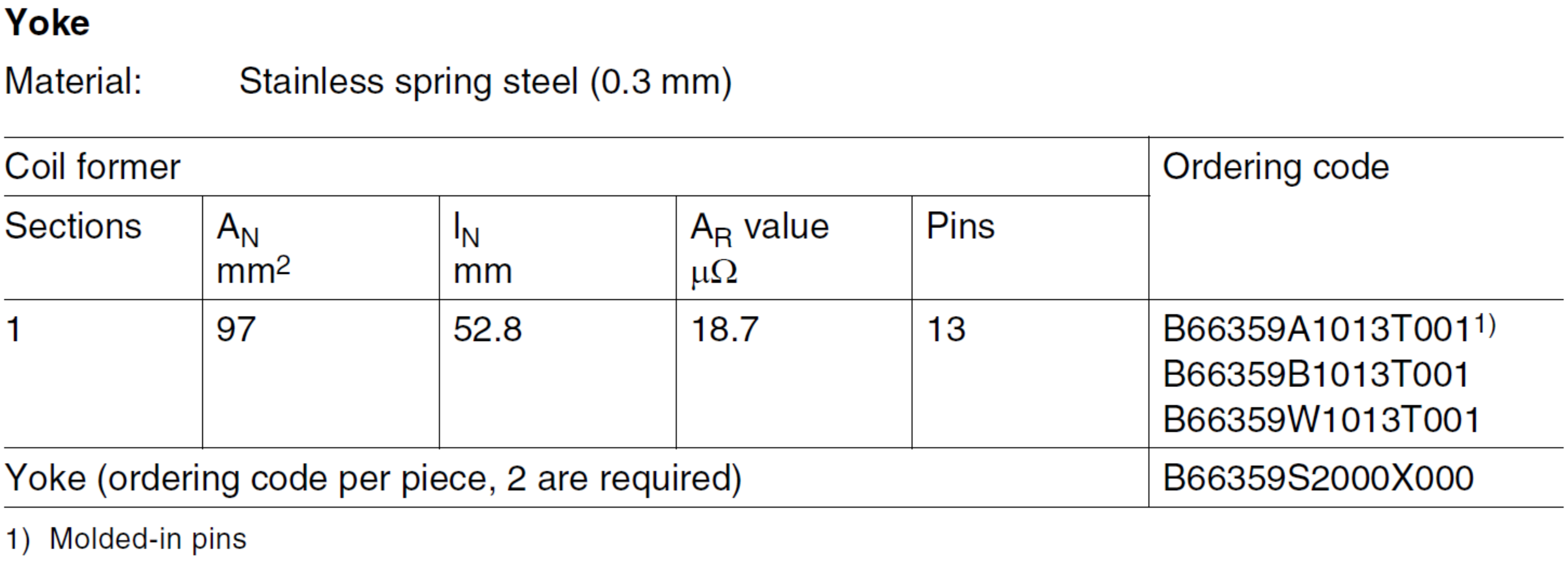
- The Volume of the core is $V = 5350\ mm^3$
- The losses per m^3 are $Pv = 20KW/m^3$
- So the core losses are Pv * V = 107mW

If your switching frequency is different from 100KHz, use this graph!



Inductor Design: winding

- The current is composed of a DC value + a small AC, so we will avoid LITZ wire.
- The window available for the winding is $A_N = 97mm^2$:
- The average length of one turn is L_N = 52.8mm



Yoke

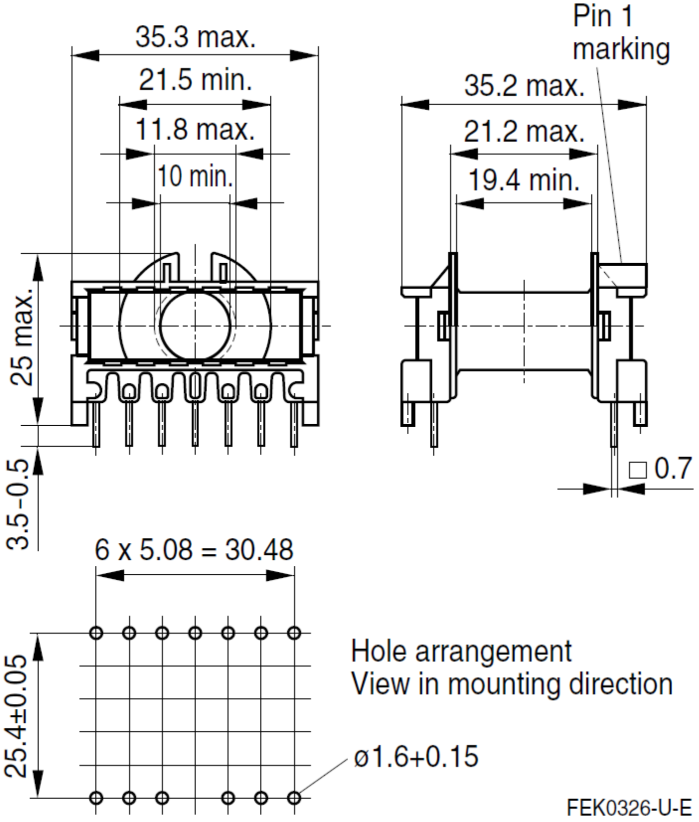
Material: Stainless spring steel (0.3 mm)

Coil former					Ordering code
Sections	A_N mm^2	l_N mm	A_R value µΩ	Pins	
1	97	52.8	18.7	13	B66359A1013T001 1) B66359B1013T001 B66359W1013T001
Yoke (ordering code per piece, 2 are required)					B66359S2000X000

1) Molded-in pins

Inductor Design: winding

- Let's suppose to take only half of the winding, so ~ 50mm^2
- We have 38 turns; the maximum cross section we can achieve is 50mm^2 / 38 = A_S = 1.31 mm^2 (= 1.3mm diameter)



- The width of the coil former is 19.4mm
- Result: in theory we can fit 15 turns for each layer, in reality we will fit 13 turns
- We will need at the end 3 layers.
- Final isolation: 2 MYLAR 0.05mm layers
- The available height of the coil former is (21.5mm – 11.8mm)/2 = 4.85mm…it will fit!

Inductor Design: winding

- Now we can calculate the DC resistance with:

$$R_{DC} = \frac{\rho \cdot N \cdot L_N}{A_S}$$

Where:

ρ = resistivity of copper @ 25C = 0.017

A_S = Wire section in mm²

- This results in a RDC = 26 milli-Ohm
- The DC losses will be: $P_{DC} = R_{DC} * I_{RMS}^2 = 0.42W$

Since: $$I_{RMS} = \sqrt{I_{DC}{}^2 + I_{pk}{}^2 / 3} = 4.02\,A_{RMS}$$

We didn’t calculate on purpose the AC losses since the ripple part is small, but will become more important as the inductor will be used for DCM or TM: use LITZ or multi-strand.