EE-365 - W4
MAGNETICS
INDUCTOR DESIGN
TRANSFORMER DESIGN

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MAGNETIC COMPONENTS

Essential element of every power electronics converter...

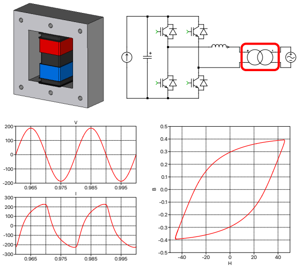


Figure 1 Line frequency transformer interfacing VSI and power grid

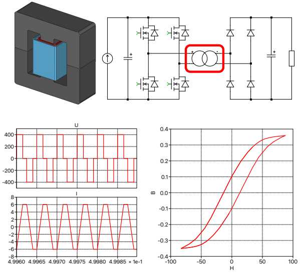


Figure 3 Isolation transformer in DCDC converter with phase-shift modulation

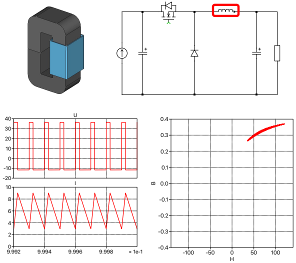


Figure 2 Filter inductor in Buck converter with DC-bias

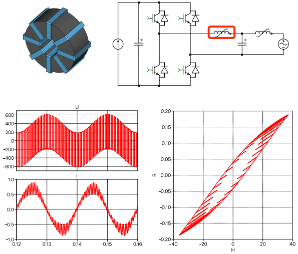
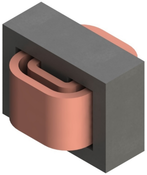
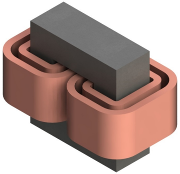


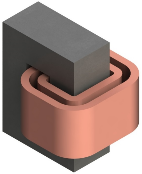
Figure 4 Filter inductor in VSI with both sinusoidal and PWM excitation

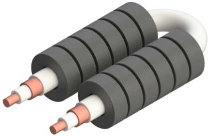
Construction Choices:

- Core-Winding Arrangement


Shell Type


Core Type

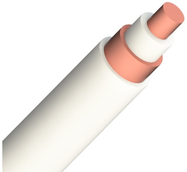

C-Type

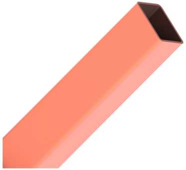

Coaxial Type

- Winding Types


Litz Wire


Foil


Coaxial


Hollow

Materials:

- Magnetic Materials
  - Silicon Steel
  - Amorphous
  - Nanocrystalline
  - Ferrites
- Windings
  - Copper
  - Aluminum
- Insulation
  - Air
  - Solid
  - Oil
- Cooling
  - Air natural/forced
  - Oil natural/forced
  - Water

MAGNETIC MATERIALS - SILICON STEEL

Ferromagnetic - Silicon Steel (SiFe)

- Iron based alloy of Silicon provided as isolated laminations
- Mostly used for line frequency transformers

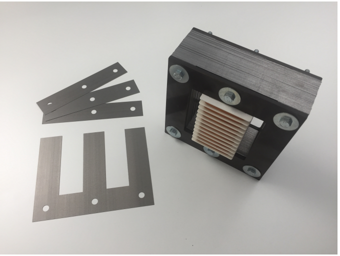
Advantages

- Wide initial permeability range
- High saturation flux density
- High Curie-temperature
- Relatively low cost
- Mechanically robust
- Various core shapes available (easy to form)

Disadvantages

- High hysteresis loss (irreversible magnetisation)
- High eddy current loss (high electric conductivity)
- Acoustic noise (magnetostriction)

Saturation B	Init. permeability	Core loss (10 kHz, 0.5T)	Conductivity
$0.8 \sim 2.2\,\mathrm{T}$	$0.6 \sim 100 \cdot 10^3$	$50 \sim 250$ W/kg	$2 \cdot 10^7 \sim 5 \cdot 10^7$ S/m



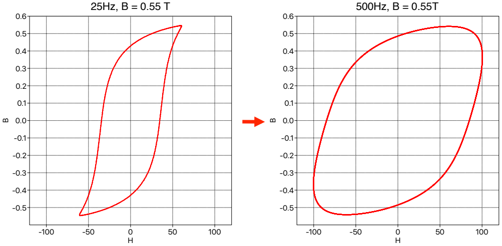


Figure 5 Example: Measured B-H curve of M330-35 laminate

MAGNETIC MATERIALS - AMORPHOUS ALLOY

Ferromagnetic - Amorphous Alloy

- Iron based alloy of Silicon as thin tape without crystal structure
- For both line frequency and switching frequency applications

Advantages

- High saturation flux density
- Low hysteresis loss
- Low eddy current loss (low electric conductivity)
- High Curie-temperature
- Mechanically robust

Disadvantages

- Relatively narrow initial permeability range
- Very high acoustic noise (magnetostriction)
- Limited core shapes available (difficult to form)
- Relatively expensive

Saturation B	Init. permeability	Core loss (10kHz, 0.5T)	Conductivity
$0.5 \sim 1.6\,\mathrm{T}$	$0.8 \cdot 10^3 \sim 50 \cdot 10^3$	$2 \sim 20$ W/kg	$< 5 \cdot 10^3$ S/m

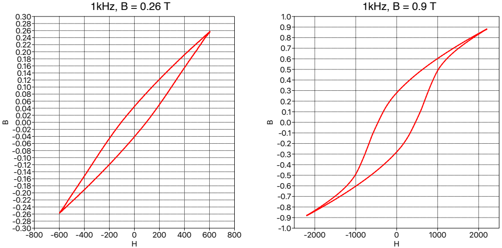


Figure 6 Example: Measured B-H curve of Metglas 2605SA

MAGNETIC MATERIALS - NANOCRYSTALLINE ALLOY

Ferromagnetic - Nanocrystalline Alloy

- Iron based alloy of silicon as thin tape with minor portion of crystal structure
- For both line frequency and switching frequency applications

Advantages

- Relatively narrow initial permeability range
- High saturation flux density
- Low hysteresis loss
- High Curie-temperature
- Low acoustic noise

Disadvantages

- Eddy current loss (compensated thanks to the thin tape)
- Mechanically fragile
- Limited core shapes available (difficult to form)
- Relatively expensive

Saturation B	Init. permeability	Core loss (10kHz, 0.5T)	Conductivity
$1 \sim 1.2\,\mathrm{T}$	$0.5 \cdot 10^3 \sim 100 \cdot 10^3$	< 50 W/kg	$3 \cdot 10^3 \sim 5 \cdot 10^4$ S/m

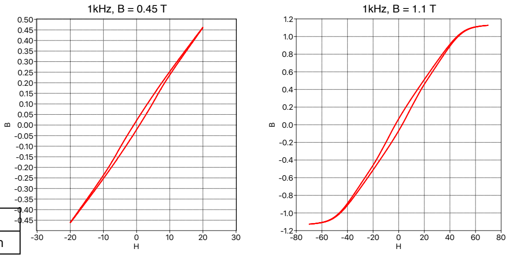
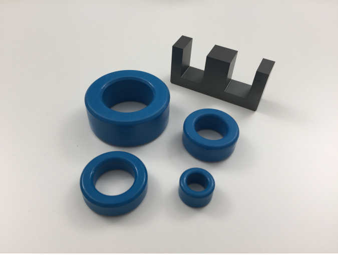


Figure 7 Example: Measured B-H curve of VITROPERM 500F

Ferrimagnetic - Ferrites

- Ceramic material made from powder of different oxides and carbons
- For both line frequency and switching frequency applications



Advantages

- Relatively narrow initial permeability range
- Low hysteresis loss
- Very low eddy current loss
- Low acoustic noise
- Relatively low cost
- Various core shapes available

Disadvantages

- Low saturation flux density
- Narrow range of initial permeability
- Magnetic properties deteriorate with temperature increase
- Mechanically fragile

Saturation B	Init. permeability	Core loss (10kHz, 0.5T)	Conductivity
$0.3 \sim 0.5$ T	$0.1 \cdot 10^3 \sim 20 \cdot 10^3$	$5 \sim 100$ W/kg	$< 1 \cdot 10^{-5}$ S/m

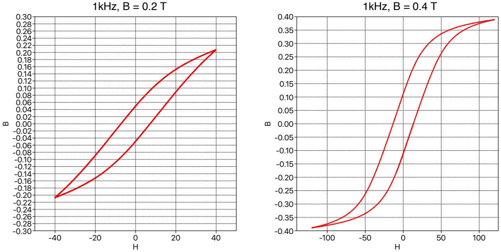


Figure 8 Example: Measured B-H curve of Ferrite N87

WINDING MATERIALS

Copper winding

- Flat wire - low frequency, easy to use
- Litz wire - high frequency, limited bending
- Foil - provide flat windings
- Hollow tubes - provide cooling efficiency
- Better conductor
- More expensive
- Better mechanical properties

Aluminium winding

- Flat wire
- Foil - skin effect differences compared to Copper
- Hollow tubes
- Difficult to interface with copper
- Offer some weight savings
- Cheaper
- Somewhat difficult mechanical manipulations

Copper Parameters

Electrical conductivity	$58.5 \cdot 10^6\ S/m$
Electrical resistivity	$1.7 \cdot 10^{-8}\ \Omega m$
Thermal conductivity	$401\ W/mK$
TEC (from 0° to 100° C)	$17 \cdot 10^{-6}\ K^{-1}$
Density	$8.9\ g/cm^3$
Melting point	$1083\ ^\circ C$

Aluminum Parameters

Electrical conductivity	$36.9 \cdot 10^6\ S/m$
Electrical resistivity	$2.7 \cdot 10^{-8}\ \Omega m$
Thermal conductivity	$237\ W/mK$
TEC (from 0° to 100° C)	$23.5 \cdot 10^{-6}\ K^{-1}$
Density	$2.7\ g/cm^3$
Melting point	$660\ ^\circ C$

INSULATING MATERIALS

Multiple influencing factors

- Operating voltage levels
- Over-voltage category
- Environment - IP class
- Temperature
- Moisture
- Cooling implications
- Ageing (self-healing?)
- Manufacturing complexity
- Partial Discharge
- BIL
- Cost

Dielectric properties

- Breakdown voltage (dielectric strength)
- Permittivity
- Conductivity
- Loss angle
- ...

Dielectric material	Dielectric strength (kV/mm)	Dielectric constant
Air	3	1
Oil	5 - 20	2 - 5
Mica tape	60 - 230	5 - 9
NOMEX 410	18 - 27	1.6 - 3.7
PTFE	60 - 170	2.1
Mylar	80 - 600	3.1
Paper	16	3.85
PE	35 - 50	2.3
XLPE	35 - 50	2.3
KAPTON	118 - 236	3.9

Figure 9 Variety of choices available...

DESIGN CONSTRAINTS

Electrical [1]

- Inductance
- $B < B_{sat}$
- Turns ratio
- Duty cycle
- Frequency
- $DCR < DCR_{max}$
- $J < J_{max}$
- Leakage inductance
- Self capacitance
- Self resonance
- Skin and Proximity effects
- EMI, EMC
- Shielding
- Efficiency
- Safety
- Isolation

[1] Source: George Slama, Würth Elektronik, APEC2019 Educational Seminar

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Mechanical

- $A_{wdg} > A_{wdg-min}$
- Size (L, W, H)
- Volume
- Surface area
- Weight
- Safety
- Creepage distances
- Clearance distances
- Insulation class
- Materials
- Environmental

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Thermal

- $T < T_{max}$
- $P_{wdq} < P_{wdg-max}$
- $P_{core} < P_{core-max}$
- Environmental

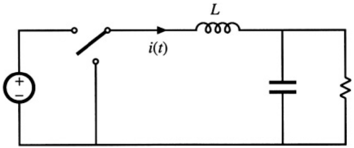
[1] Source: George Slama, Würth Elektronik, APEC2019 Educational Seminar

INDUCTOR DESIGN

Single winding structure design using K_g method...

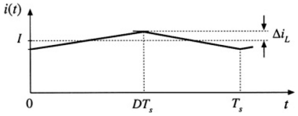
Typical DC inductor application (Buck converter)

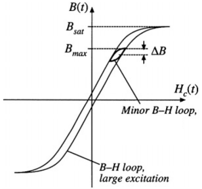
- It is a part of the LC output filter
- Designed to match the reference L



Inductor current and B-H loop

- Nonzero DC current component $I_{dc} > 0$
- Nonzero AC current components $\Delta I > 0$



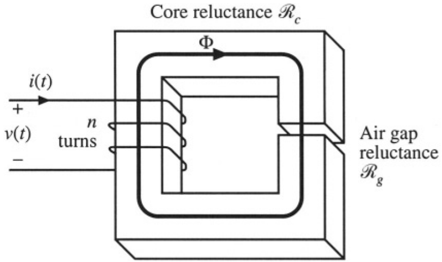


Assumptions

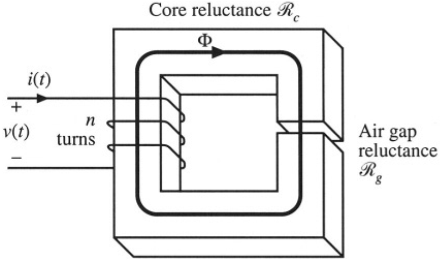
- Core and proximity losses are negligible
- Low frequency copper losses are dominant

Design constraints

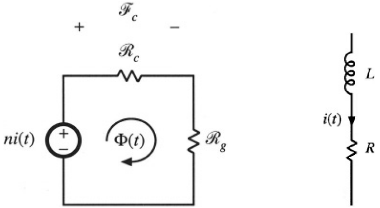
- Maximum flux density should not be exceeded
- Target inductance value should be matched
- Sufficient winding area of the core
- Winding resistance - current density - losses



Inductor geometry



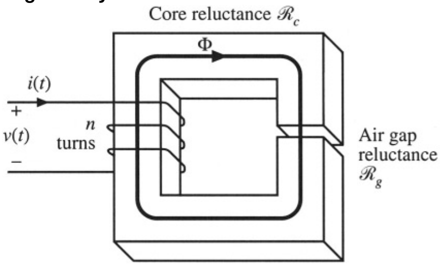
Equivalent reluctance and electric circuit



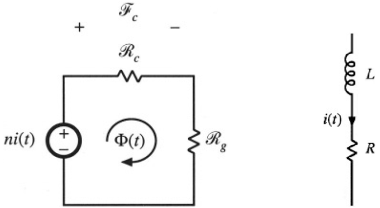
Design constraints

- Based on the geometry equivalent reluctance circuit is derived
- It is assumed that the core and the air gap have the same cross sections (fringe flux is neglected)
- The core and air gap reluctance can be calculated as:
$$\mathfrak{R}_c = \frac{l_c}{\mu_c A_c} \quad \text{and} \quad \mathfrak{R}_g = \frac{l_g}{\mu_0 A_c}$$
- Solution of the reluctance circuit yields:
$$ni = \Phi(\mathfrak{R}_c + \mathfrak{R}_g)$$
- Where it can be assumed that $\mathfrak{R}_c \ll \mathfrak{R}_g$ due to $\mu_0 \ll \mu_r$
$$ni = \Phi\mathfrak{R}_g$$
- Maximum operation flux density B_{max} must be less than the saturation flux density of the given core material B_{sat}
- B_{max} should be achieved at the maximum current point I_{max}
$$nI_{max} = B_{max}A_c\mathfrak{R}_g$$
- The B_{max} constraint can be expressed as:
$$nI_{max} = B_{max}\frac{l_g}{\mu_0}$$

Inductor geometry



Equivalent reluctance and electric circuit



Reference magnetizing inductance derivation

- The reference inductance value must be obtained
- The inductance is equal to

$$L = \frac{n^2}{\mathfrak{R}_g} = \frac{\mu_0 A_c n^2}{l_g}$$

Reference resistance derivation

- The reference resistance value must be obtained
- The resistance is equal to

$$R = \rho \frac{l_b}{A_w}$$

- A_w is the bare conductor cross section
- l_b is the total length of the winding wire

$$l_b = n(MLT)$$

- (MLT) is the mean-length-per-turn of the winding
- R can be expressed as

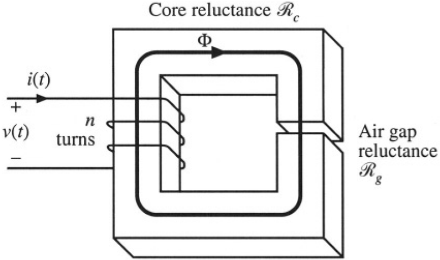
$$R = \rho \frac{n(MLT)}{A_w}$$

- R must not be too high due to thermal considerations
- Current density should typically be bounded (cooling!)

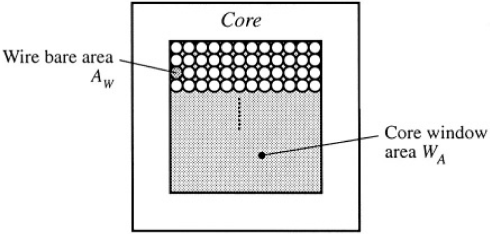
$$J_{rms} = \frac{I_{rms}}{A_w} \leq [3-6] \frac{A}{mm^2}$$

INDUCTOR DESIGN - WINDING AREA CONSTRAINT

Inductor geometry



Core window area utilization



Core window area sizing

- The winding must fit through the core window
- The total bare wire cross section can be expressed as:

$$nA_w$$

- Conductors are not perfectly packed, e.g. round wire
- Additional insulation material/spacing may be needed
- The relation between the core window area W_A and total bare wire cross section can be expressed as:

$$W_A K_u \geq nA_w$$

- K_u is the window utilization factor, or fill factor
- K_u is in $[0.7 - 0.55]$ range for round wire
- Additional multiplicative coefficient in range $[0.95 - 0.65]$ for insulated wire

INDUCTOR DESIGN - THE CORE GEOMETRICAL CONSTANT

We have identified four constraints:

- Core should not saturate

$$nI_{max} = B_{max}\frac{l_g}{\mu_0}$$

- Target inductance should be achieved

$$L = \frac{n^2}{\mathfrak{R}_g} = \frac{\mu_0 A_c n^2}{l_g}$$

- Winding should fit into available core window area

$$W_A K_u \geq n A_w$$

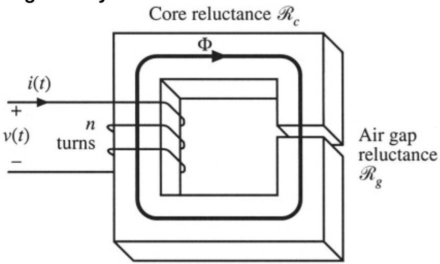
- Winding resistance should be less or equal than desired (Losses!)

$$R = \rho\frac{n(MLT)}{A_w}$$

- Core geometry related parameters: A_c, W_A, MLT
- Design specifications: $I_{max}, B_{max}, \mu_0, L, K_u, R, \rho$
- Uknown variables: n, l_g, A_w
- Eliminating unknown variables leads to the following inequality:

$$\frac{A_c^2 W_A}{(MLT)} \geq \frac{\rho L^2 I_{max}^2}{B_{max}^2 R K_u}$$

Inductor geometry



Merged constraints

- Combining the presented constraints leads to

$$\frac{A_c^2 W_A}{(MLT)} \geq \frac{\rho L^2 I_{max}^2}{B_{max}^2 R K_u}$$

- The **right side** of this inequality are specifications or other known quantities
- The **left side** of the inequality is a function of the core geometry alone

The core geometrical constant - K_g

- It is necessary to choose a core whose geometry satisfies the given inequality
- The Core Geometrical Constant is defined as

$$K_g = \frac{A_c^2 W_A}{(MLT)}$$

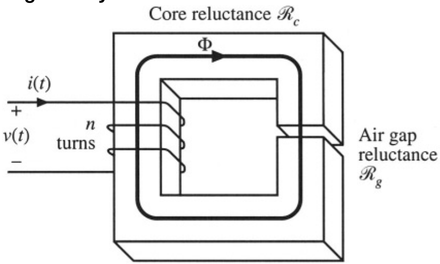
- It describes the effective electrical size of magnetic cores in applications where copper loss and maximum flux density are specified
- It can be computed for any core based on its geometry
- The core selection boils down to selecting a core with high enough K_g

$$K_g \geq \frac{\rho L^2 I_{max}^2}{B_{max}^2 R K_u}$$

Conclusions

- Different electric parameters affect the core size
- Different geometries can satisfy the K_g constraint, showing the trade-off between utilized conductor (W_A) and core material (A_c)

Inductor geometry



Specified units and constants:

- Electrical: $\rho\,[\Omega - cm]$, $I_{max}\,[A]$, $L\,[H]$, $R\,[\Omega]$, $K_u\,[p.u.]$, $R\,[\Omega]$, $B_{max}\,[T]$
- Geometrical: $A_c\,[cm^2]$, $W_A\,[cm^2]$, $(MLT)\,[cm]$
- Constants: $\rho = 1.724 \cdot 10^{-6}\Omega - cm$, $\mu_0 = 4\pi \cdot 10^{-7}\,H/m$

Step-by-step design:

- Determine the core size:

$$K_g \geq \frac{\rho L^2 I_{max}^2}{B_{max}^2 R K_u} \cdot 10^8\,[cm^5]$$

- Determine the air gap length:

$$l_g = \frac{\mu_0 L I_{max}^2}{B_{max}^2 A_c} \cdot 10^4\,[m]$$

- Determine the number of turns:

$$n = \frac{L I_{max}}{B_{max} A_c} \cdot 10^4$$

- Determine the wire size:

$$A_w \leq \frac{K_u W_A}{n}\,[cm^2]$$

Control check:

- Check current density:

$$J_{rms} = \frac{I_{rms}}{A_w} \leq 3\frac{A}{mm^2}$$

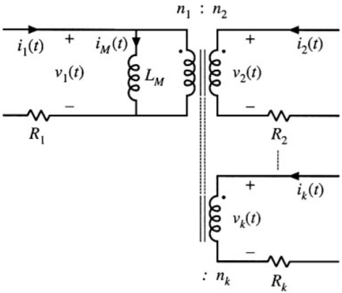
- Resistance can be computed for verification:

$$R = \rho\frac{n(MLT)}{A_w}$$

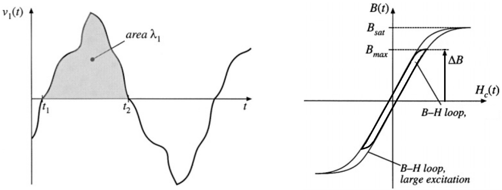
TRANSFORMER DESIGN

Multiple windings structure...

A k-winding transformer



Inductor Current and BH Loop



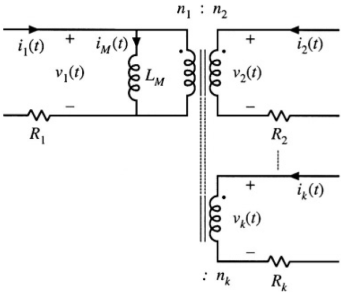
Assumptions

- High frequency Skin losses are negligible
- Proximity losses are negligible
- Both Winding and Core losses are considered
- B_{max} is limited by core losses rather than saturation due to large BH loop

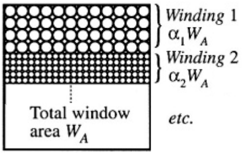
Design constraints

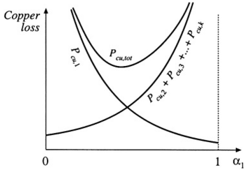
- Optimal core window area allocation
- Core losses
- Flux density
- Copper losses
- Total losses
- Optimal flux density

A k-winding transformer



Winding allocation





Problem statement

- A portion of the core window area needs to be allocated for each winding
- Increasing the allocated area of 1-th ($R_1 \searrow$) winding decreases the available area for the rest ($R_i \nearrow$)
- Overall winding losses ($P_{cu,tot}$) should be minimized

Optimal core window area allocation

- There exists an optimal winding space allocation for each winding α_i that minimizes $P_{cu,tot}$
- It can be expressed as a function of the apparent power of each winding

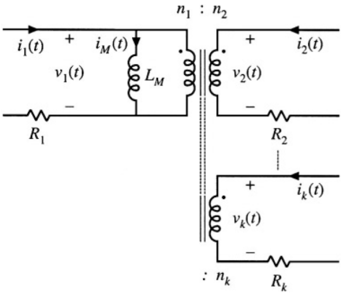
$$\alpha_i = \frac{V_i I_i}{\sum_{j=1}^{k} V_j I_j}$$

- Each winding will get a portion of core window area proportional to its apparent power

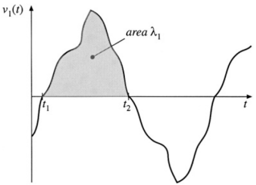
$$A_{wi} = \frac{\alpha_i K_u W_A}{n_i}$$

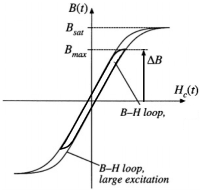
TRANSFORMER DESIGN - CORE LOSSES & FLUX DENSITY

A k-winding transformer



Volt-second and B-H loop





Core losses

- Core losses cannot be neglected due to large BH loop
- For sinusoidal excitation they can be expressed with Steinmetz equation

$$P_{fe} = K_{fe}(\Delta B)^{\beta} A_c l_m$$

- l_m is the core mean magnetic path length, hence $A_c l_m$ is the core volume
- K_{fe} and β is the Steinmetz parameters defined for the chosen core material
- Core losses P_{fe} increase with the increase of max flux density ΔB

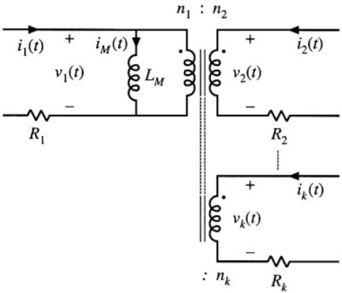
Flux density

- For a given applied voltage waveform with λ_1 volt-seconds, the maximum flux density can be expressed as

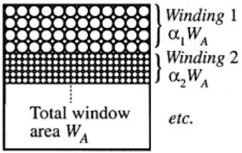
$$\Delta B = \frac{\lambda_1}{2 n_1 A_c}$$

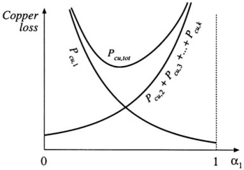
- ΔB and core losses can be decreased by increasing n_1
- However increase of n_1 increases copper losses, since it results in more turns of smaller wire
- Therefore there exists a ΔB that optimizes the total losses

A k-winding transformer



Winding allocation (e.g. primary and secondary)





Total copper losses:

- When the core window area is allocated to the various windings according to their relative apparent powers (optimal allocation), the total copper loss is then

$$P_{cu} = \frac{\rho(MLT)n_1^2 I_{tot}^2}{W_A K_u}$$

- I_{tot} is the sum of the rms winding currents, referred to winding 1.

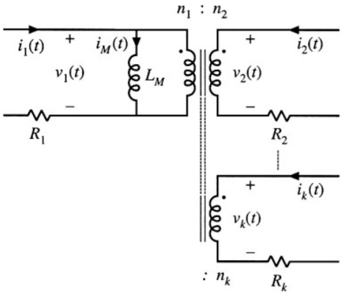
$$I_{tot} = \sum_{j=1}^{k} \frac{n_j}{n_1} I_j$$

- Expression for flux density can be used to eliminate n_1 resulting in

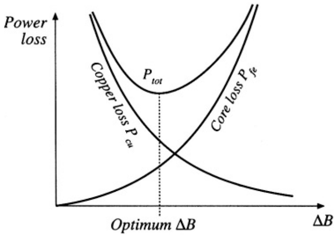
$$P_{cu} = \left(\frac{\rho\lambda_1^2 I_{tot}^2}{4K_u}\right)\left(\frac{(MLT)}{W_A A_c^2}\right)\left(\frac{1}{\Delta B}\right)^2$$

- Given expression is grouped into three terms: specifications, core geometry and ΔB influence

A k-winding transformer



Optimal flux density



Optimal flux density derivation:

- Both P_{fe} and P_{cu} are expressed as explicit functions of ΔB

$$P_{fe} = K_{fe}(\Delta B)^{\beta} A_c l_m$$

$$P_{cu} = \left(\frac{\rho \lambda_1^2 I_{tot}^2}{4K_u}\right)\left(\frac{(MLT)}{W_A A_c^2}\right)\left(\frac{1}{\Delta B}\right)^2$$

- It is possible to find the optimal ΔB that minimizes the total losses P_{tot}

$$P_{tot} = P_{fe} + P_{cu}$$

- Optimal ΔB yields

$$\frac{dP_{tot}}{d(\Delta B)} = 0$$

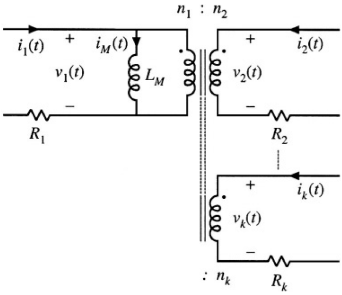
- Solution of this equation providing the optimal ΔB is given as

$$\Delta B = \left[\frac{\rho \lambda_1^2 I_{tot}^2}{2K_u} \frac{(MLT)}{W_A A_c^3 l_m} \frac{1}{\beta K_{fe}}\right]^{\left(\frac{1}{\beta+2}\right)}$$

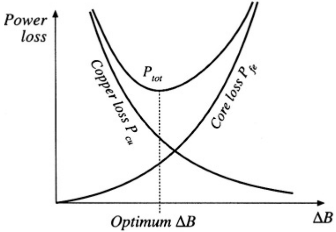
- It is important to check if this value exceeds the B_{sat} of the given material

TRANSFORMER DESIGN - THE CORE GEOMETRICAL CONSTANT

A k-winding transformer



Optimal flux density



Merged constraints:

- Combining the previously derived equations leads to

$$\frac{W_A A_c^{(2(\beta-1)/\beta)}}{(MLT) l_m^{(2/\beta)}} \left[\left(\frac{\beta}{2}\right)^{-\left(\frac{\beta}{\beta+2}\right)} + \left(\frac{\beta}{2}\right)^{\left(\frac{2}{\beta+2}\right)} \right]^{-\left(\frac{\beta+2}{\beta}\right)} = \frac{\rho \lambda_1^2 I_{tot}^2 K_{fe}^{(2/\beta)}}{4 K_u P_{tot}^{((\beta+2)/\beta)}}$$

- The terms on the left side depend on the core geometry
- The terms on the right side depend on application specifications and the desired core material characteristics
- The core geometrical constant K_{gfe} is defined as:

$$K_{gfe} = \frac{W_A A_c^{(2(\beta-1)/\beta)}}{(MLT) l_m^{(2/\beta)}} \left[\left(\frac{\beta}{2}\right)^{-\left(\frac{\beta}{\beta+2}\right)} + \left(\frac{\beta}{2}\right)^{\left(\frac{2}{\beta+2}\right)} \right]^{-\left(\frac{\beta+2}{\beta}\right)}$$

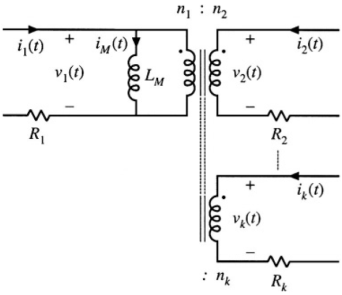
- The core selection boils down to selecting a core with high enough K_{gfe}

$$K_{gfe} \geq \frac{\rho \lambda_1^2 I_{tot}^2 K_{fe}^{(2/\beta)}}{4 K_u P_{tot}^{((\beta+2)/\beta)}}$$

Conclusions

- K_{gfe} is a measure of the magnetic size of a core for applications in which core loss is significant
- It depends on selected core material characteristics

A k-winding transformer



Specified units and constants:

- Electrical: $\rho\,[\Omega - cm]$, $I_{tot}\,[A]$, $\frac{n_j}{n_1}\,[p.u.]$, $\lambda_1\,[V - sec]$, $P_{tot}\,[W]$, $K_u\,[p.u.]$, $\beta\,[]$, $K_{fe}\,[]$, $\Delta B\,[T]$
- Geometrical: $A_c\,[cm^2]$, $W_A\,[cm^2]$, $(MLT)\,[cm]$, $l_m\,[cm]$, $A_{wi}\,[cm^2]$
- Constants: $\rho = 1.724 \cdot 10^{-6}\Omega - cm$, $\mu_0 = 4\pi \cdot 10^{-7}\,H/m$

Step-by-step design:

- Determine the core size:

$$K_{gfe} \geq \frac{\rho \lambda_1^2 I_{tot}^2 K_{fe}^{(2/\beta)}}{4 K_u P_{tot}^{((\beta+2)/\beta)}} \cdot 10^8\,[cm^5]$$

- Evaluate peak ac flux density

$$\Delta B = \left[\frac{\rho \lambda_1^2 I_{tot}^2}{2K_u} \frac{(MLT)}{W_A A_c^3 l_m} \frac{1}{\beta K_{fe}}\right]^{\left(\frac{1}{\beta+2}\right)}$$

- Evaluate primary number of turns

$$n_1 = \frac{\lambda_1}{2\Delta B A_c} \cdot 10^4$$

- Choose numbers of turns for other windings based on desired transformation ratios

$$\frac{n_j}{n_1}\,[p.u.]$$

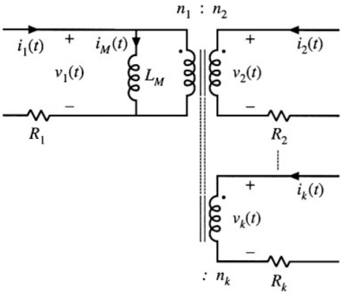
- Evaluate fraction of window area allocated to each winding

$$\alpha_j = \frac{n_j I_j}{n_1 I_{tot}}$$

- Evaluate the wire size

$$A_{wj} \leqslant \frac{\alpha_j K_u W_A}{n_1}$$

A k-winding transformer



Specified units and constants:

- Electrical: $\rho\,[\Omega - cm]$, $I_{tot}\,[A]$, $\frac{n_j}{n_1}\,[p.u.]$, $\lambda_1\,[V - sec]$, $P_{tot}\,[W]$, $K_u\,[p.u.]$, $\beta\,[]$, $K_{fe}\,[]$, $\Delta B\,[T]$
- Geometrical: $A_c\,[cm^2]$, $W_A\,[cm^2]$, $(MLT)\,[cm]$, $l_m\,[cm]$, $A_{wi}\,[cm^2]$
- Constants: $\rho = 1.724 \cdot 10^{-6}\Omega - cm$, $\mu_0 = 4\pi \cdot 10^{-7}H/m$

Control check:

- Optimal ΔB does not saturate the core:
$$\Delta B \leqslant B_{sat}$$
- Magnetizing inductance, referred to primary winding:
$$L_m = \frac{\mu n_1^2 A_c}{l_m}$$
- Peak AC magnetizing current, referred to primary winding:
$$i_{M,pk} = \frac{\lambda_1}{2L_m}$$
- Winding resistances:
$$R_j = \frac{\rho n_j (MLT)}{A_{wj}}$$

EXAMPLES OF HIGH-POWER MEDIUM FREQUENCY TRANSFORMERS

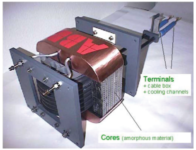


ABB: 350kW, 10kHz

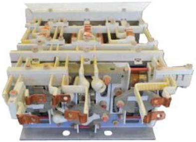
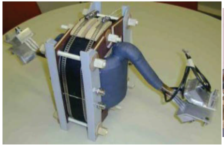
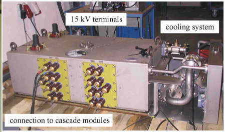


ABB: 3x150kW, 1.8kHz

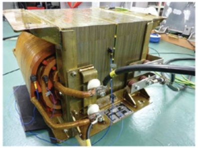


BOMBARDIER: 350kW, 8kHz

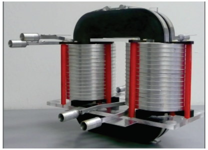


ALSTOM: 1500kW, 5kHz

IKERLAN: 400kW, 6kHz

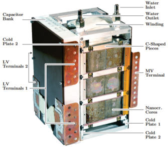


IKERLAN: 400kW, 600Hz

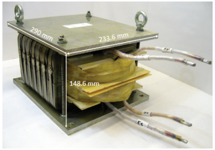


FAU-EN: 450kW, 5.6kHz

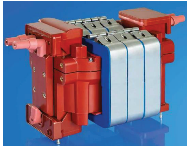
CHALMERS: 50kW, 5kHz



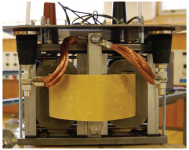
ETHZ: 166kW, 20kHz



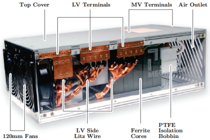
EPFL: 300kW, 2kHz



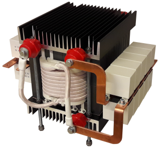
STS: 450kW, 8kHz



KTH: 170kW, 4kHz



ETHZ: 166kW, 20kHz



EPFL: 100kW, 10kHz

?

ACME: ???kW, ???kHz