High Frequency Planar Magnetics for Power Conversion

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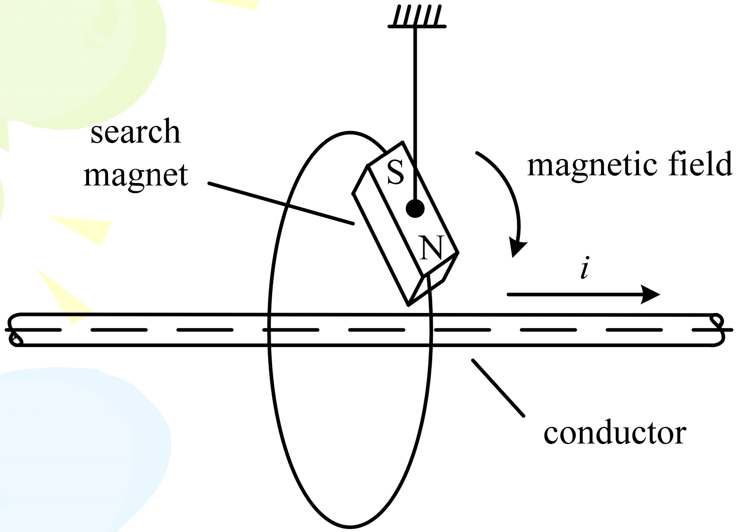
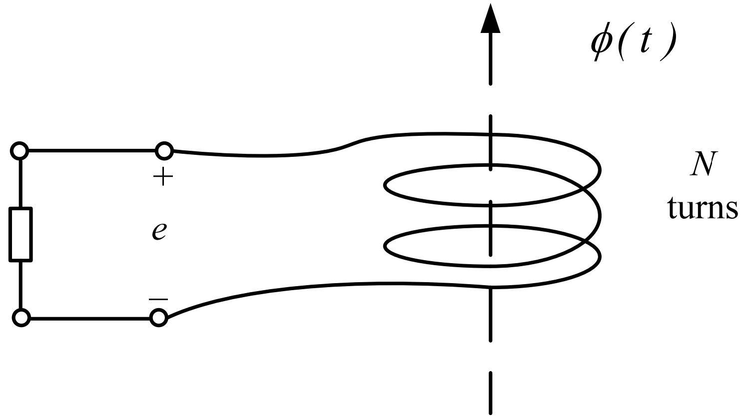
Outline

- Magnetics Basics Introduction (WGH)
  - ✓ Transformer Design
  - ✓ High Frequency Effects in the Winding
- Analytical Models for Planar Magnetics (WGH)
- Planar Magnetics – Fundamentals (ZO)
- Planar Magnetic Components Integration (ZO)

Magnetics Basics

Laws of Electromagnetism





Ampere's Law

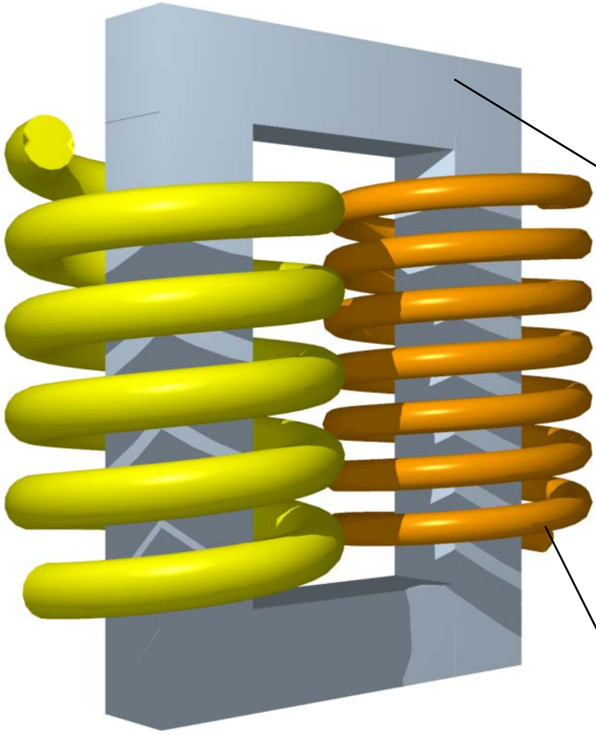
Faraday's Law

$$\sum H \cdot l = Ni$$

$$e = -N\frac{d\phi}{dt}$$

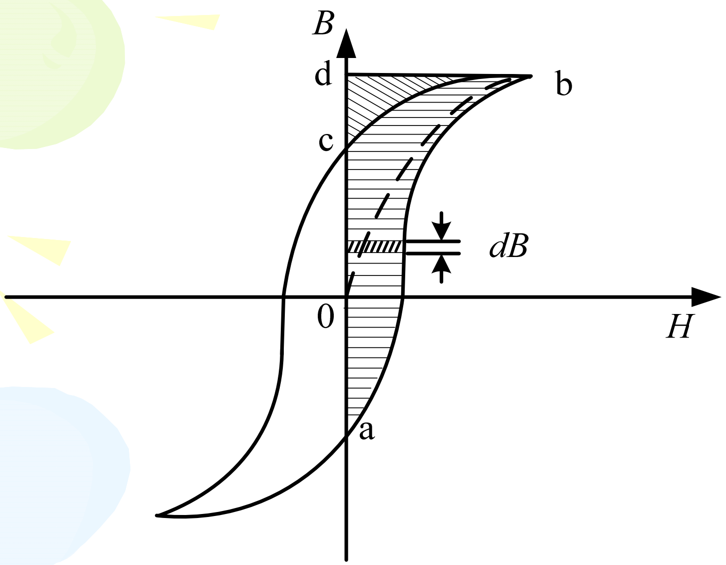
$$\mathbf{B} = \mu \mathbf{H} \qquad \mu = \mu_r \mu_0$$

Losses in Magnetic Components

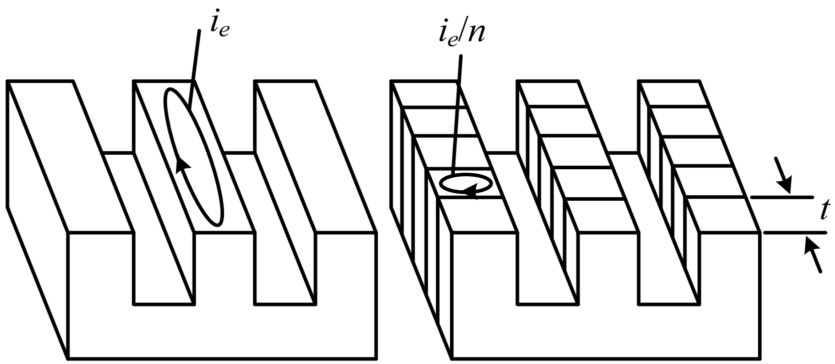


- Core losses
  - Hysteresis loss
  - Eddy current loss
- Copper losses
  - Skin effect loss
  - Proximity effect loss

Core Loss



Hysteresis loss in a ferromagnetic material

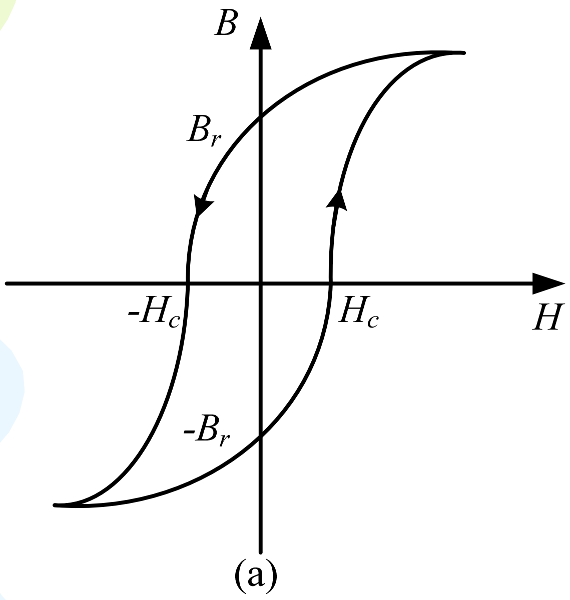


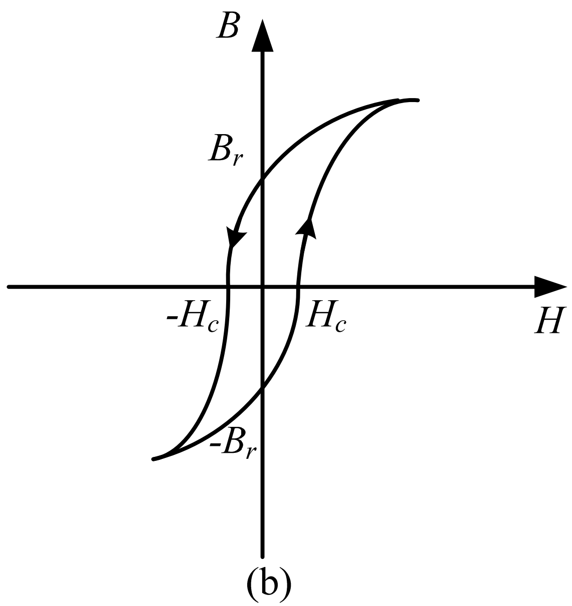
Eddy current loss in a ferromagnetic material

$$P_{fe} = K_c f^{\alpha} B_{max}^{\ \ \beta}$$

- Hysteresis loss is the area inside the B-H loop
- Eddy current loss is reduced by laminations

Ferromagnetic Materials





(a) Hard magnetic materials

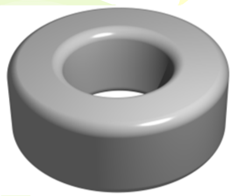
(b) Soft magnetic materials

Soft Magnetic Materials

The magnetic and operating properties of some soft magnetic materials

Materials	Ferrites	Nanocrystalline	Amorphous	Si Iron	Ni-Fe (Permalloy)	Powdered iron
Model	TDK P40	VIROPERM 500F	METGLAS 2605	AK Oriented M-4	MAGNETICS PERMALLOY 80	MICROMET -ALS 35µ
Permeability, μ_i	1500-4000	15000	10,000-150, 000	5,000-10,000	20,000	3-550
B_{peak}, T	0.45-0.81	1.2	1.56	2.0	0.82	0.6-1.3
ρ, µΩm	6.5×10^6	1.15	1.3	0.51	0.57	10^6
Curie temp. T_c, °C	215	600	399	746	460	665
P_{loss}	60 mW/ cm^3 at 0.1T/50kHz	588 mW/cm^3 at 0.3T/100kHz	72 m W/cm^3 at 0.2T/25kHz	2.295-30.6mW/cm^3 at 1.5T/50Hz	192.28mW/cm^3 at 0.2T/5kHz	126-315mW/cm^3 at 0.1T/10kHz

Core Shapes



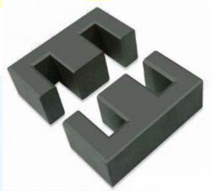
Toroid core

PQ core

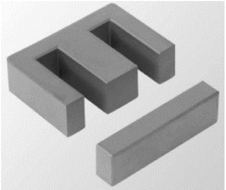
Pot core

RS/DS core

RM core



EE core

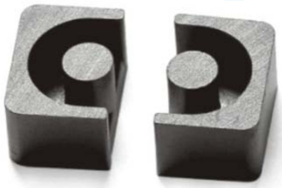


EI core

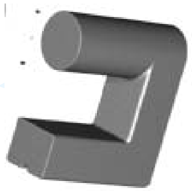
ER core

EFD core

ETD core



EP core



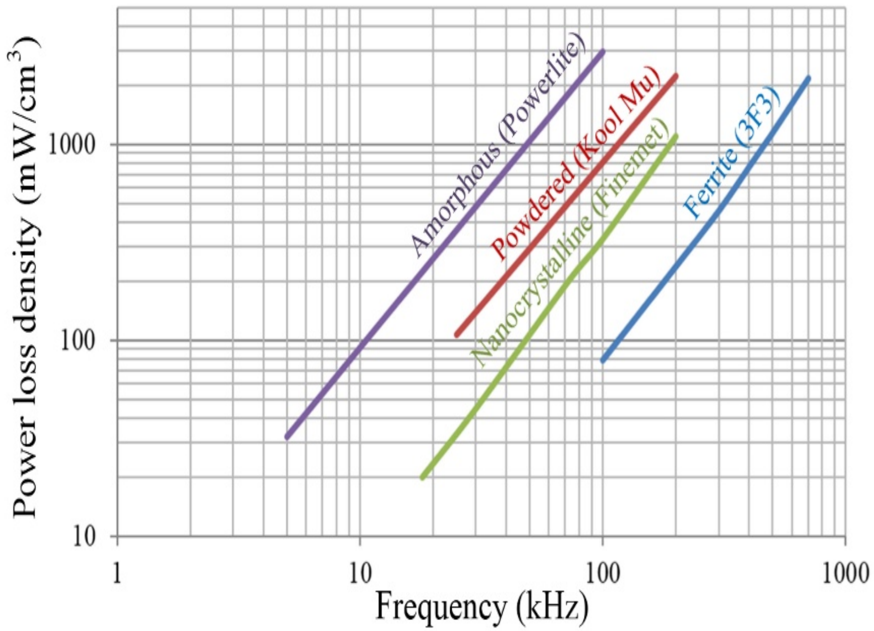
UR core

U core

C core

Planar core

Core Loss Density vs Frequency



Transformer Design

Basic Equations

Voltage equation

$$V_{rms} = \frac{k}{\tau/T} f\ N\ B_{max}\ A_m = K_v\ f\ N\ B_{max} A_m \quad \text{where} \quad k = \frac{V_{rms}}{<v>} \qquad K_v = \frac{k}{\frac{\tau}{T}} = \frac{k}{\tau f}$$

K_v =4.44 for a sinewave
=4.00 for a squarewave

Power equation

$$\sum VA = K_v f B_{max} J_o k_f k_u A_p$$

V_{rms}: the rms value of the applied voltage
$<v>$: the average value of the applied voltage
K_v: voltage waveform factor
k: the form factor
f: the frequency of the applied voltage
T: the period of the applied voltage
τ: the interval from the point where the flux density is zero to the point where it is at its maximum value
k_f: the core stacking factor
J_o: the current density in each winding

Transformer Losses

Winding losses

Total resistive losses $$P_{cu} = \sum RI^2 = \rho_w \sum_{i=1}^{n} \frac{N_i MLT (J_o A_{wi})^2}{A_{wi}}$$

$$k_u = \frac{\sum_{i=1}^{n} N_i A_{wi}}{W_a}$$ is window utilization factor

$$V_w = MLT \times W_a$$ is volume of the windings

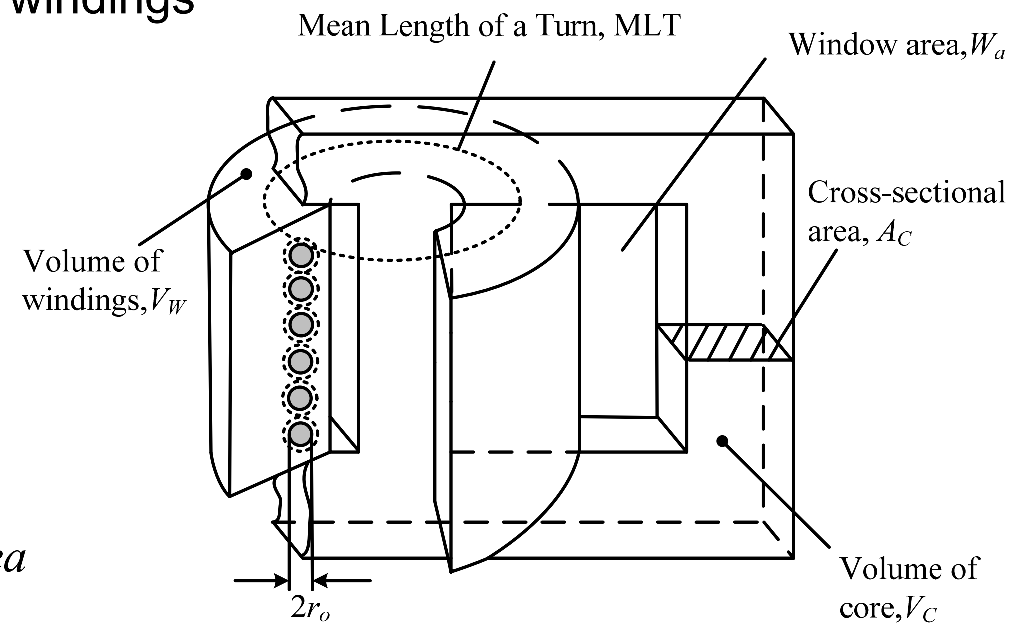
$$P_{cu} = \rho_w V_w k_u J_o^2$$

Core losses

$$P_{fe} = V_c K_c f^{\alpha} B_{max}^{\beta}$$

$$A_p = W_a \times A_c$$

$$Window\,area \times cross-sectional\,area$$



Typical layout of a transformer

Losses Optimization

Winding losses

$$P_{cu} = \rho_w V_w k_u \left[\frac{\sum \mathrm{VA}}{K_v f B_{\max} k_f k_u A_p} \right]^2 = \frac{a}{f^2 B_{\max}{}^2}$$

Core losses

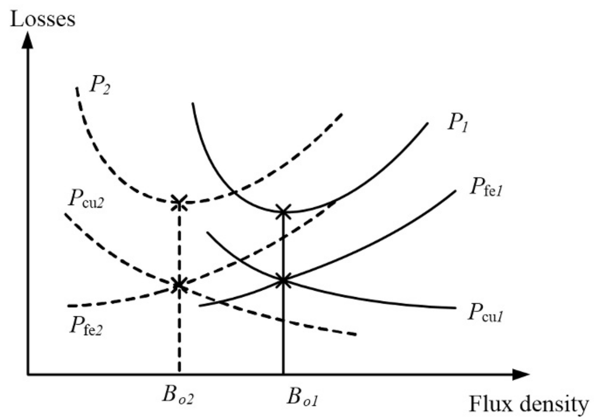
$$P_{fe} = V_c K_c f^{\alpha} B_{\max}{}^{\beta} = b f^{\alpha} B_{\max}{}^{\beta}$$

Total losses

$$P = \frac{a}{f^2 B_m{}^2} + b f^{\alpha} B_m{}^{\beta}$$

At a given operation frequency,

$$\frac{\partial P}{\partial B_{\max}} = -\frac{2a}{f^2 B_{\max}{}^3} + \beta b f^{\alpha} B_{\max}{}^{\beta-1} = 0$$

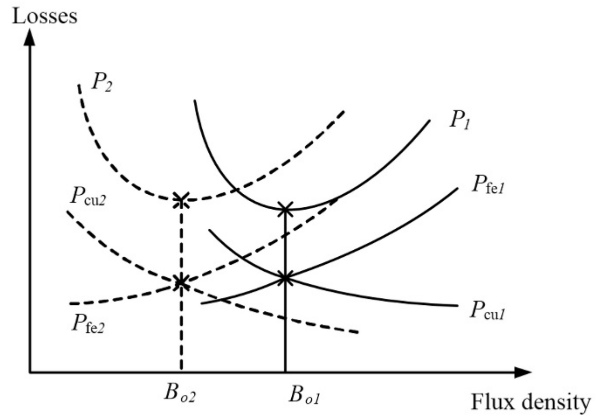


The minimum losses occur when

$$P_{cu} = \frac{\beta}{2} P_{fe}$$

Losses Optimization

$$P = \frac{a}{f^2 B_m{}^2} + b f^{\alpha} B_m{}^{\beta}$$



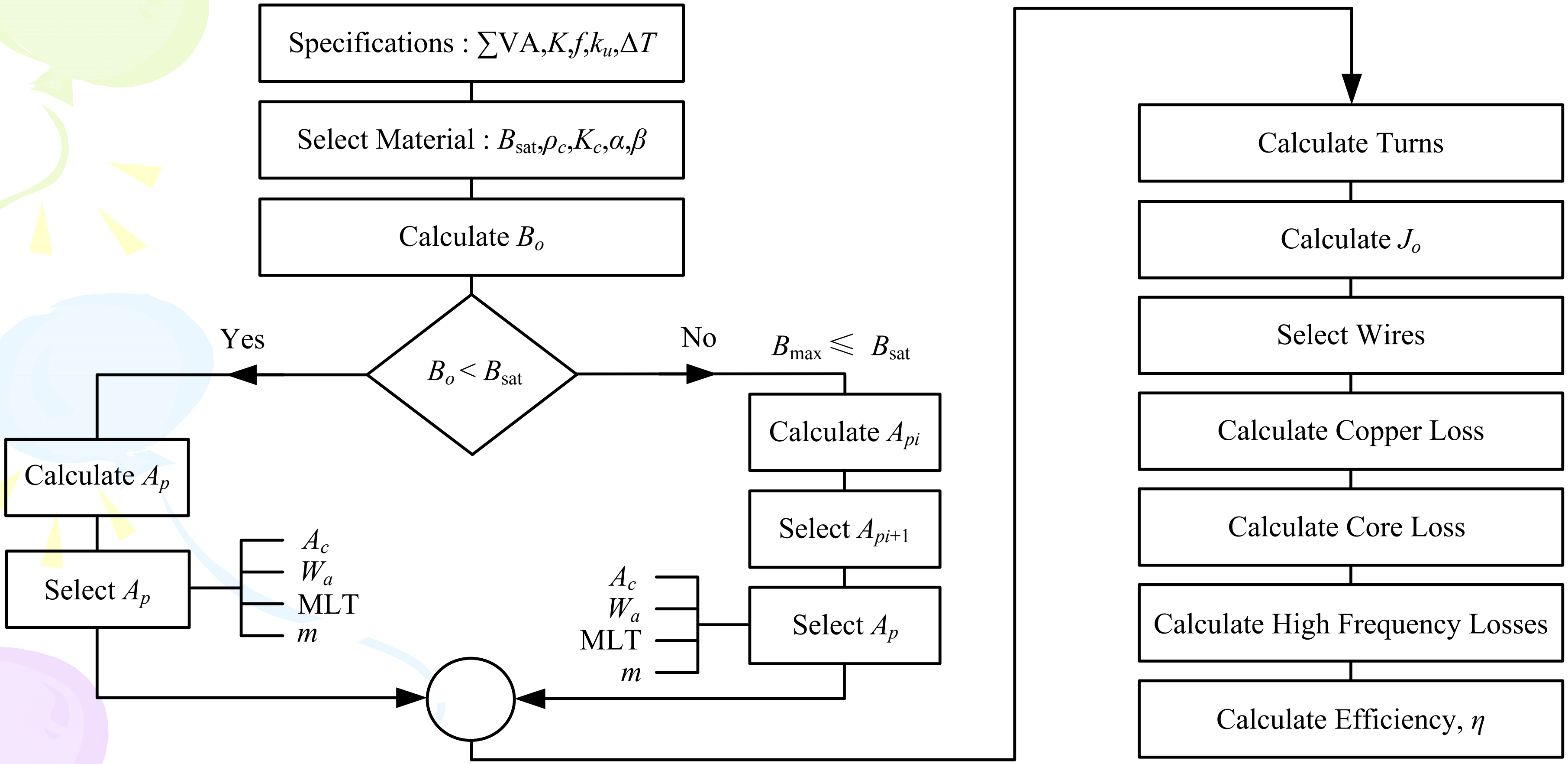
$$B_o^{\frac{7\beta-12}{12}} f^{\frac{7\alpha-12}{12}} = \frac{2^{7/12}\beta^{1/12}}{(\beta+2)^{2/3}} \frac{[hk_t \Delta T]^{2/3}}{[\rho k_w]^{1/12}[k_c K_c]^{7/12}} \left[\frac{K_v \sqrt{k_u}}{\sum \mathrm{VA}}\right]^{1/6}$$

$$A_p = \left[\frac{\beta+2}{\beta}\frac{1}{k_u \Delta T}\right]^{4/7} \left[\frac{\sum \mathrm{VA}}{K_v f B_o K_\theta}\right]^{8/7}$$

$$J_o = \sqrt{\frac{\beta}{\beta+2}\frac{hk_t}{\rho k_w}} \sqrt{\frac{\Delta T}{k_u}} \frac{1}{A_p^{1/8}} = K_\theta \sqrt{\frac{\beta}{\beta+2}} \sqrt{\frac{\Delta T}{k_u}} \frac{1}{A_p^{1/8}}$$

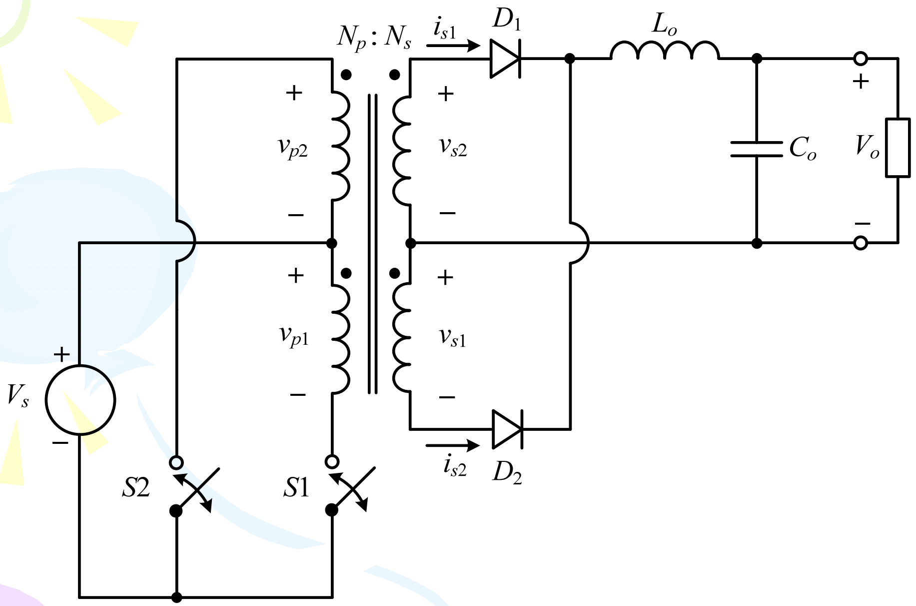
For *h*=10 W/m2 ºC, k_c=5.6, k_w=10, k_t=40, ρ=1.72 x 10-8 Ω-m, K_θ=48.224 x 10^3.

Design Methodology

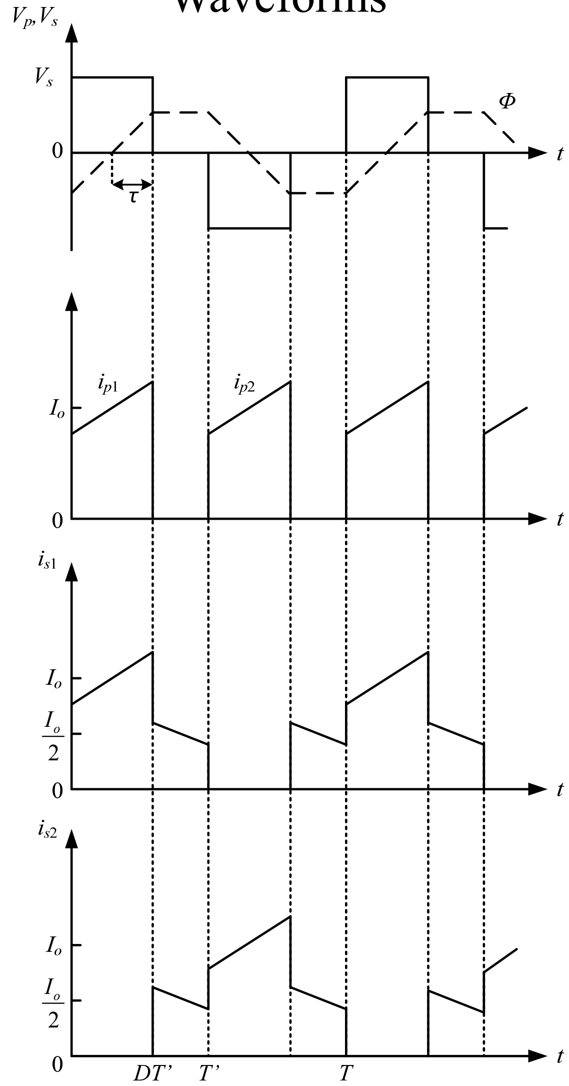


Push-pull Converter Transformer

Circuit



Waveforms



Push-pull Converter Transformer

Design specifications

Input	36 → 72 V
Output	24 V, 12.5 A
Frequency, f	50 kHz
Temperature Rise, ΔT	35 ºC
Ambient Temperature, T_a	45 ºC

Core data: EPCOS N67 Mn-Zn

K_c	9.12
α	1.24
β	2.0
B_{sat}	0.4 T

Core loss

$$P_{fe} = K_c f^{\alpha} B_m^{\beta}$$

Push-pull Converter Transformer

Calculations:

(3) *VA* ratings of the windings

$$P_o = (24+1)\times 12.5 = 312.5\ \mathrm{W}$$

$$\sum VA = \left(\frac{1}{k_{pp}}\left(\frac{P_o}{2}+\frac{P_o}{2}\right)+\frac{1}{k_{ps}}\left(\frac{P_o}{2}+\frac{P_o}{2}\right)\right) = \left(\sqrt{2}+\sqrt{\frac{1+D}{D}}\right)P_o$$

$$= \left(\sqrt{2}+\sqrt{\frac{1+0.67}{0.67}}\right)(312.5) = 935$$

(4) Optimum A_p

$$B_o = \frac{\left[(10)(40)(35)\right]^{2/3}}{2^{2/3}\left[(1.72\times 10^{-8})(10)(0.4)\right]^{1/12}\left[(5.6)(9.12)(50\ 000)^{1.24}\right]^{7/12}}$$

$$\cdot\left[\frac{(4.88)(50000)(1.0)(0.4)}{935}\right]^{1/6} = 0.126\,\mathrm{T}$$

The optimum flux density is less than B_{sat}

$$A_p = \left[\frac{\sqrt{2}(935)}{(4.88)(50000)(0.126)(1.0)(0.4)(48.2\times 10^3)\sqrt{(0.4)(35)}}\right]^{8/7}\times 10^8 = 2.693\,\mathrm{cm}^4$$

ETD44 Core Data

A_c	1.73 cm^2
W_a	2.78 cm^2
A_p	4.81 cm^4
V_c	17.70 cm^3
k_f	1.0
k_u	0.4
MLT	7.77 cm
ρ_{20}	1.72 μΩ-cm
α_{20}	0.00393

Push-pull Converter Transformer

Calculations:

(6) Wire size

$$J_o = K_t \sqrt{\frac{\Delta T}{2k_u}} \frac{1}{\sqrt[8]{A_p}} = (48.2\times10^3)\sqrt{\frac{35}{2(0.4)}} \frac{1}{\sqrt[8]{(4.81\times10^{-8})}} = 2.620\times10^6 \ \text{A/m}^2$$

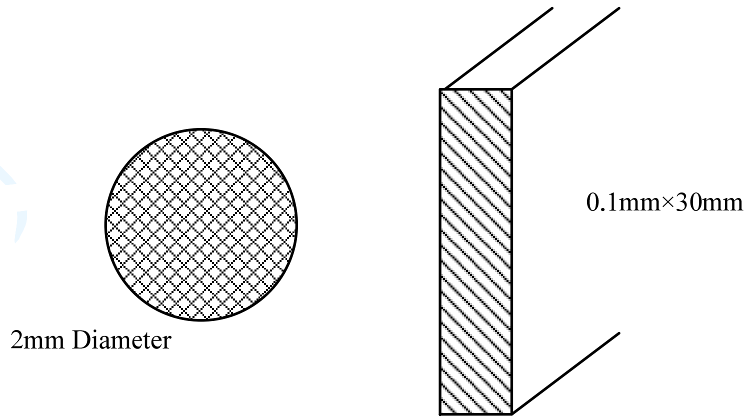
Primary windings:

$$I_p = \frac{P_o/2}{k_{pp}V_p} = \frac{312.5/2}{(0.707)(29.5)} = 7.5 \text{ A}$$

$$A_w = I_p / J_o = 2.863 \text{ mm}^2$$

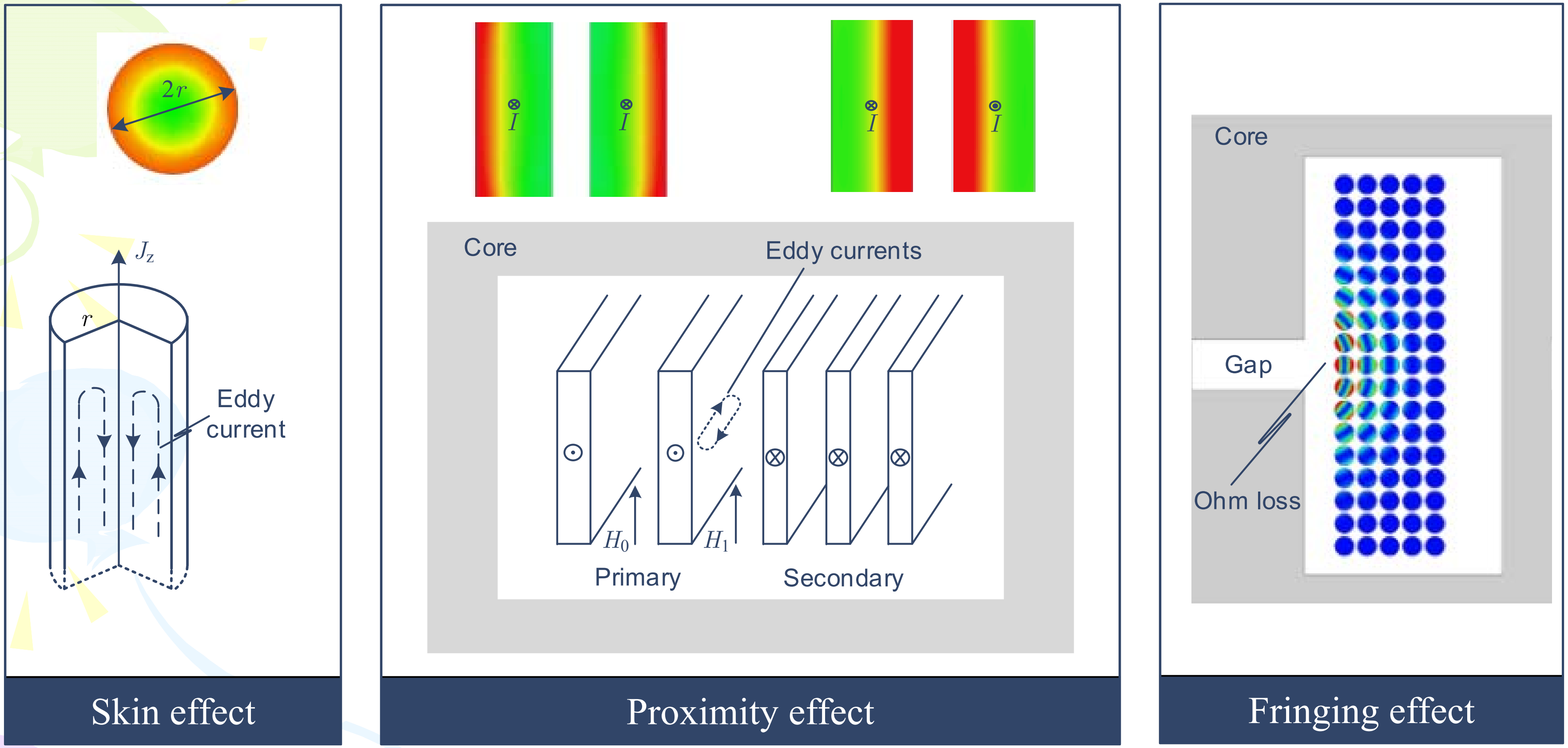
Skin depth at 50 kHz = 0.295 mm

Standard 0.1×30 mm copper foil with a dc resistance of 5.8 mΩ/m @ 20ºC meets this requirement or a 2 mm diameter wire.



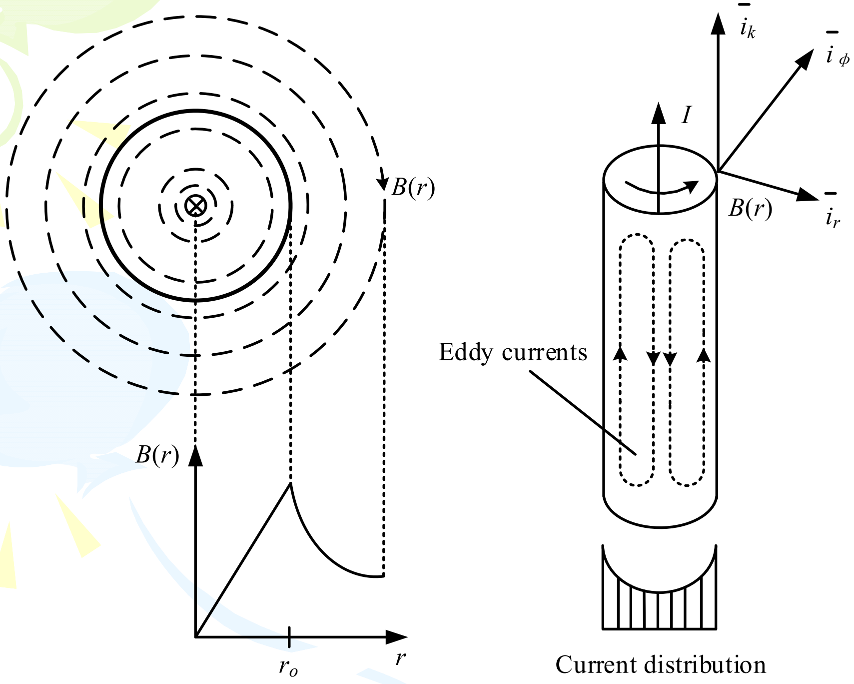
High Frequency Effects in the Windings

Design Issues for High Frequency

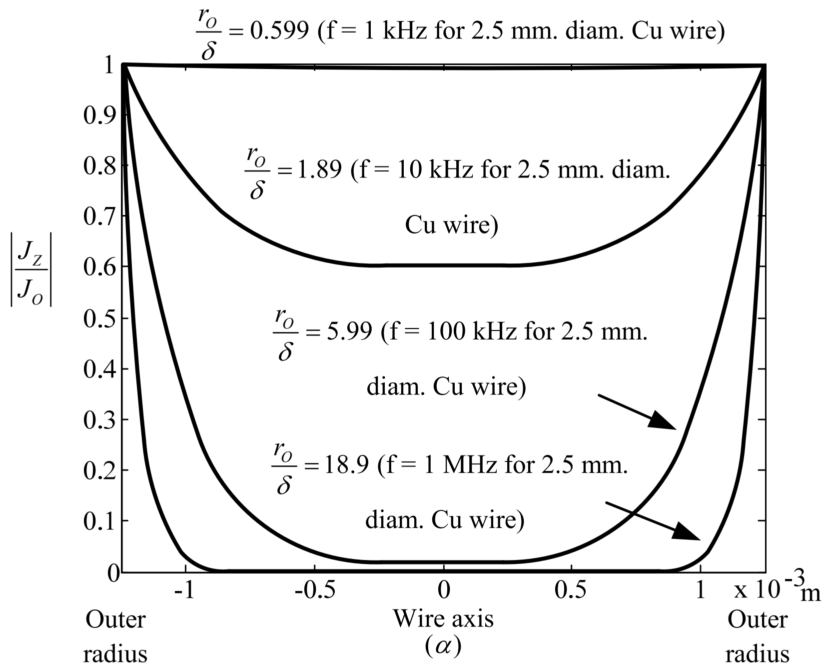


- High frequency winding loss
- Core loss: Steinmetz equation, iGSE.
- Parasitic parameters: leakage inductance, stray capacitance

Skin Effect Factor

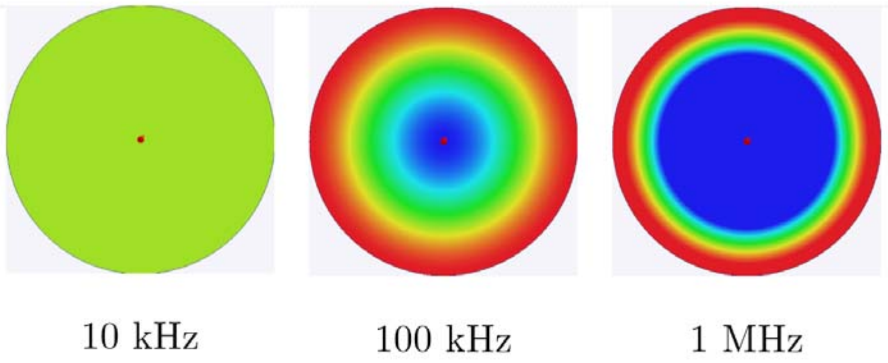


Eddy currents in a circular conductor

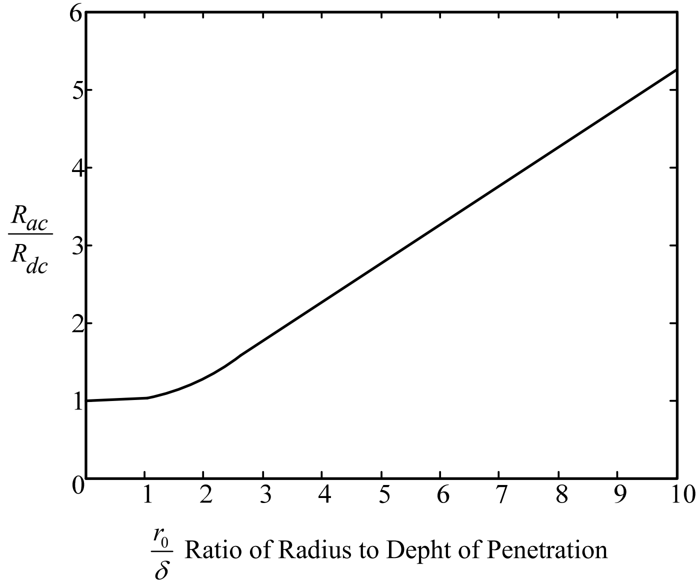


Current distribution in a circular conductor

Skin Effect



Current distribution in a circular conductor



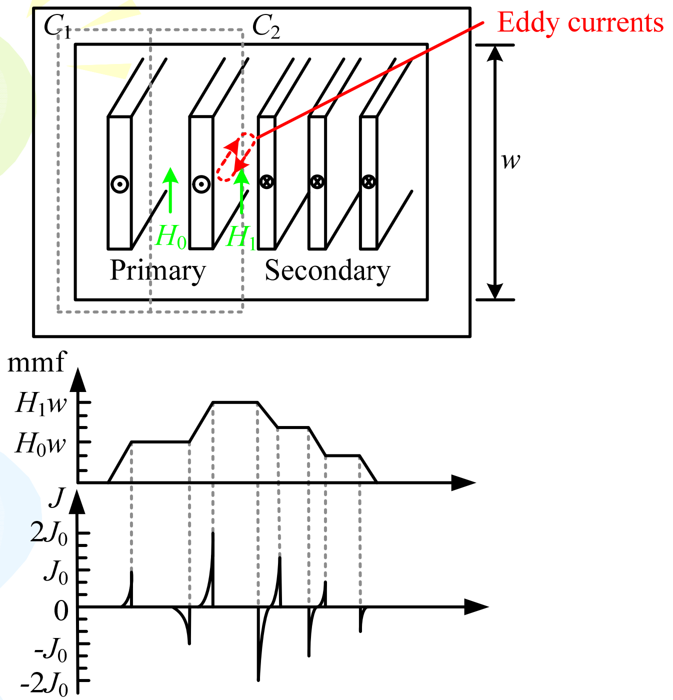
R_{ac}/R_{dc} due to the skin effect

$$\frac{R_{ac}}{R_{dc}} = 1 + \frac{(r_o/\delta)^4}{48 + 0.8(r_o/\delta)^4} \qquad r_o/\delta < 2$$

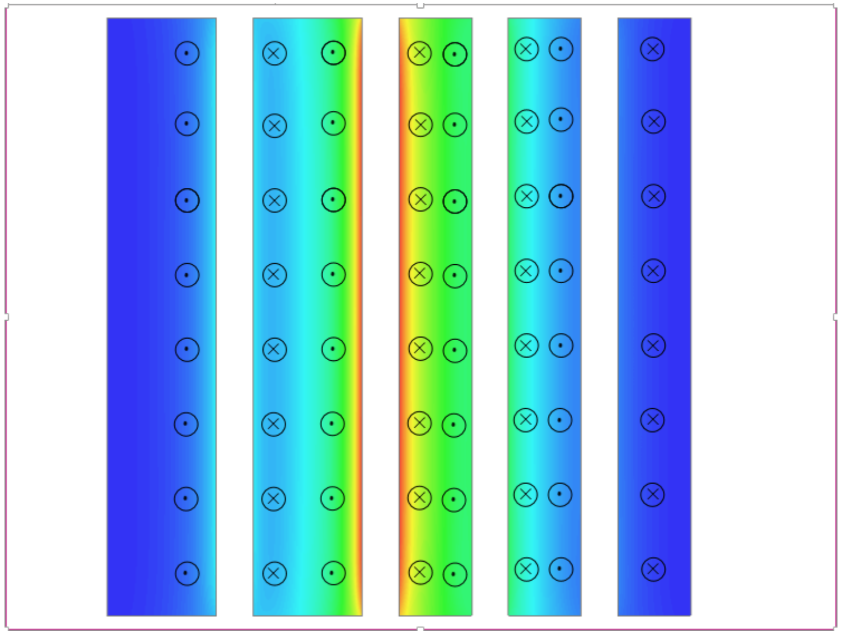
$$\delta = \frac{1}{\sqrt{\pi f \mu \sigma}}$$

- The ac resistance is proportional to the square root of frequency at very high frequencies.

Proximity Effect

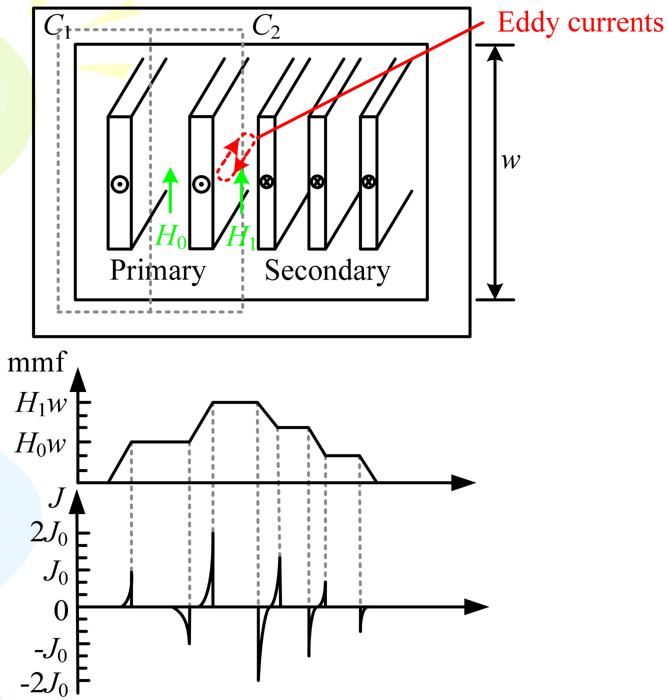


Transformer cross-section with current density distribution

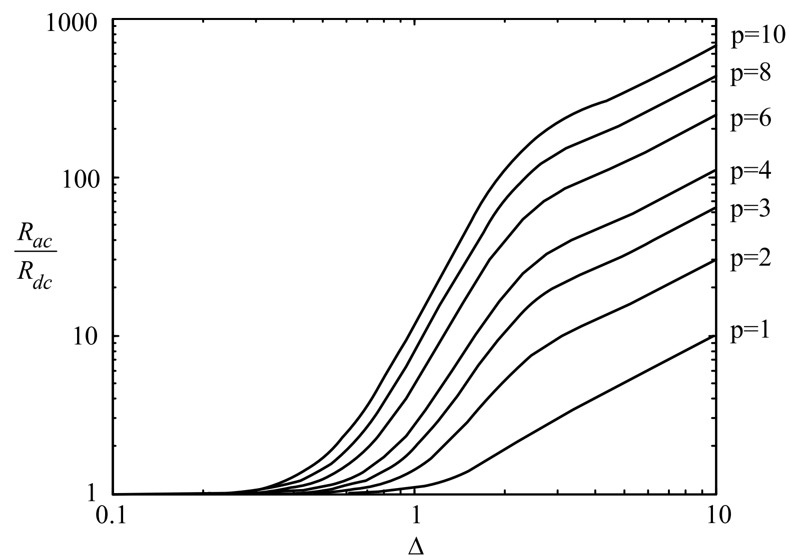


Proximity effect factor for sinusoidal excitation

Proximity Effect



Transformer cross-section with current density distribution



R_{ac}/R_{dc} due to the proximity effect

$$\frac{R_{ac}}{R_{dc}} = \phi_{prox}(\Delta) = 1 + \frac{5p^2 - 1}{45}\Delta^4 \text{ where } \Delta = \frac{d}{\delta}$$

$$\delta = \frac{1}{\sqrt{\pi f \mu \sigma}}$$

- As the number of layers *p* increase there is a substantial increase in the ac resistance for a given layer thickness *d* and frequency *f*.

Porosity Factor

A round conductor of diameter *D* is equivalent to a square conductor of side length

$$d = \sqrt{\frac{\pi}{4}}D$$

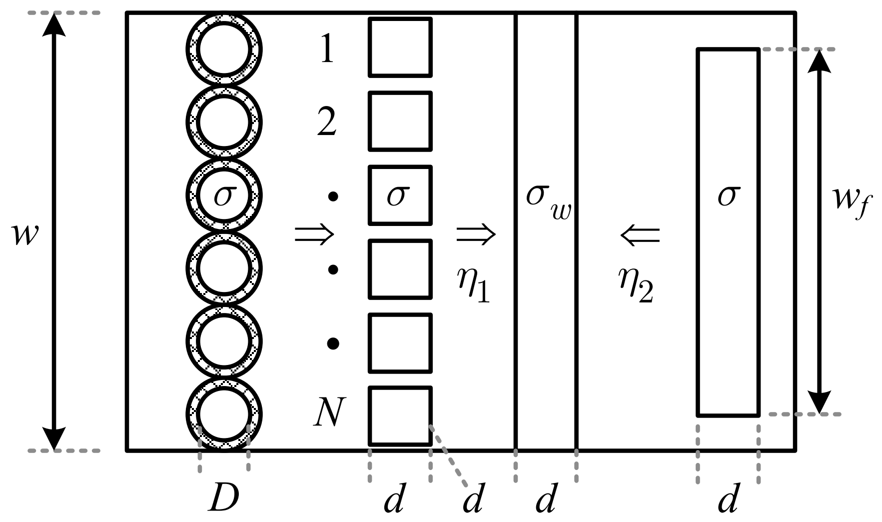
The porosity factor

$$\eta = \frac{Nd}{w}$$

The effective conductivity

$$\sigma_w = \eta\sigma$$

$$\delta = \frac{1}{\sqrt{\pi f \mu \sigma_w}}$$



Porosity factor for foils and round conductors

Proximity Effect: arbitrary waveform

An arbitrary periodic current waveform may be represented by its Fourier series

$$i(t) = I_{dc} + \sum_{n=1}^{\infty} \hat{I}_n \cos(n\omega t + \varphi_n)$$

The total power loss due to all the harmonics

$$P = R_{dc} I_{dc}^2 + \sum_{n=1}^{\infty} R_{dc} \phi_{prox}(\Delta_n) I_{n,rms}^2$$

so

$$\delta = \frac{1}{\sqrt{\pi f \mu \sigma}} \qquad \Delta_n = \frac{d}{\delta_n} = \sqrt{n}\frac{d}{\delta_o} = \sqrt{n}\Delta$$

$$\frac{R_{eff}}{R_{dc}} = \frac{I_{dc}^2 + \sum_{n=1}^{\infty} \phi_{prox}(\Delta_n) I_{n,rms}^2}{I_{rms}^2} \qquad \phi_{prox}(\Delta_n) = 1 + \frac{5p^2 - 1}{45}\Delta_n^4 = 1 + \frac{5p^2-1}{45} n^2 \Delta^4$$

$$= \frac{I_{dc}^2 + \sum_{n=1}^{\infty} I_{n,rms}^2 + \frac{5p^2-1}{45}\Delta^4 \sum_{n=1}^{\infty} n^2 I_{n,rms}^2}{I_{rms}^2}$$

Proximity Effect

$$i(t) = I_{dc} + \sum_{n=1}^{\infty} \hat{I}_n \cos(n\omega t + \varphi_n)$$

$$\frac{di(t)}{dt} = I' = -\omega \sum_{n=1}^{\infty} n\hat{I}_n \sin(n\omega t + \varphi_n)$$

$$I_{rms}^2 = I_{dc}^2 + \sum_{n=1}^{\infty} I_{n,rms}^2$$

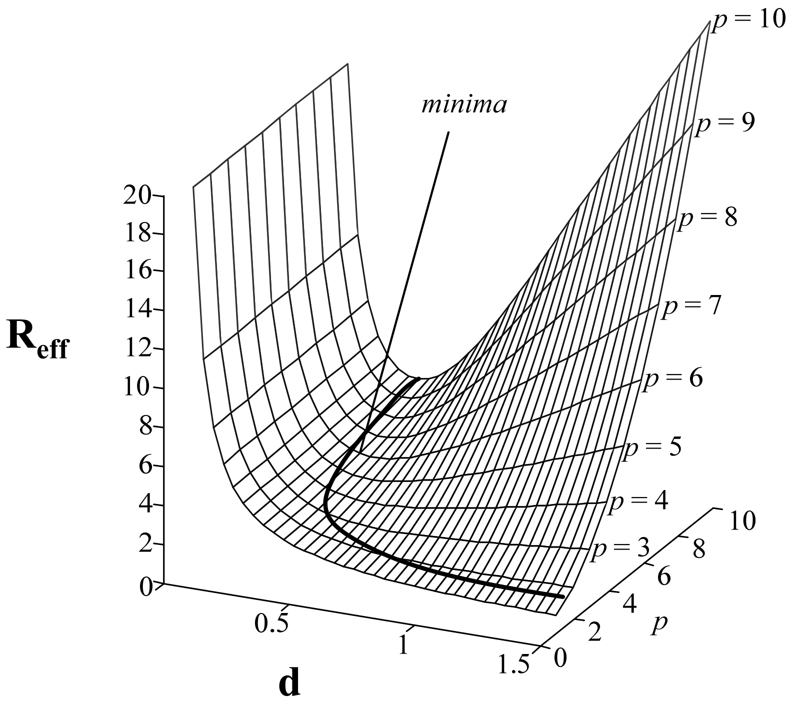
$$I'^2_{rms} = \omega^2 \sum_{n=1}^{\infty} n^2 I_{n,rms}^2$$

$$\frac{R_{eff}}{R_{dc}} = \frac{I_{dc}^2 + \sum_{n=1}^{\infty} I_{n,rms}^2 + \frac{5p^2-1}{45}\Delta^4 \sum_{n=1}^{\infty} n^2 I_{n,rms}^2}{I_{rms}^2}$$

$$= \frac{I_{rms}^2 + \frac{5p^2-1}{45}\Delta^4 \left[\frac{I'_{rms}}{\omega}\right]^2}{I_{rms}^2}$$

$$= 1 + \frac{5p^2-1}{45}\Delta^4 \left[\frac{I'_{rms}}{\omega I_{rms}}\right]^2$$

$$\Delta = \frac{d}{\delta_o}$$



Optimum Thickness

The optimum value of *Δ*

$$\Delta_{opt} = \sqrt{\frac{15}{5p^2 - 1}} \sqrt{\frac{\omega I_{\text{rms}}}{\left[\frac{di}{dt}\right]_{\text{rms}}}}$$

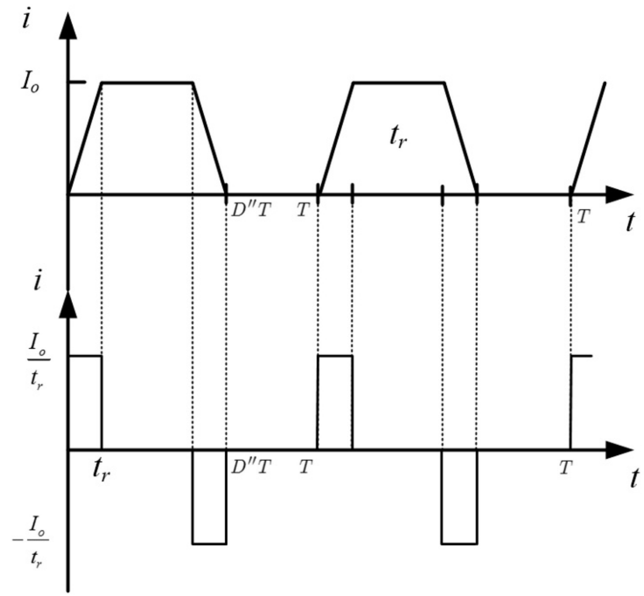
Finally

$$\frac{R_{eff}}{R_{dc}} = 1 + \frac{1}{3}\left(\frac{\Delta}{\Delta_{\text{opt}}}\right)^4$$

$$\left(\frac{R_{eff}}{R_{dc}}\right)_{\text{opt}} = \frac{4}{3}$$

Optimum Winding Thickness: Pushpull

$$\Delta_{opt} = \sqrt{\frac{15}{5p^2 - 1}}\sqrt{\frac{\omega I_{\mathrm{rms}}}{I'_{\mathrm{rms}}}}$$



Skin depth

$$\delta_0 = \frac{66}{\sqrt{f}} = \frac{66}{\sqrt{(50\times 10^3)}} = 0.295\,mm$$

Optimum layer *Δ*

$$\Delta_{\mathrm{opt}} = \sqrt[4]{\frac{\left[D - \frac{8t_r}{3T}\right]\pi^2 \frac{t_r}{T}}{(5p^2-1)15}} = \sqrt[4]{\frac{\left[0.67 - \frac{(8)(0.025)}{3}\right]\pi^2 (0.025)}{[(5)(6)^2 - 1]/15}} = 0.3342$$

Optimum layer thickness

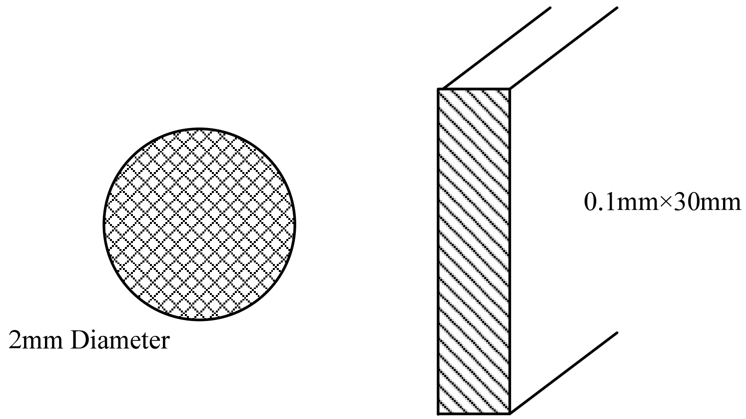
$$d_{\mathrm{opt}} = \Delta_{\mathrm{opt}}\delta_o = (0.3342)(0.295) = 0.1\,\mathrm{mm}$$

Optimum Winding Thickness: Pushpull

Effective ac resistance: foil $\dfrac{R_{\text{eff}}}{R_{\text{dc}}} = \dfrac{4}{3} = 1.3$

AC resistance of round conductor

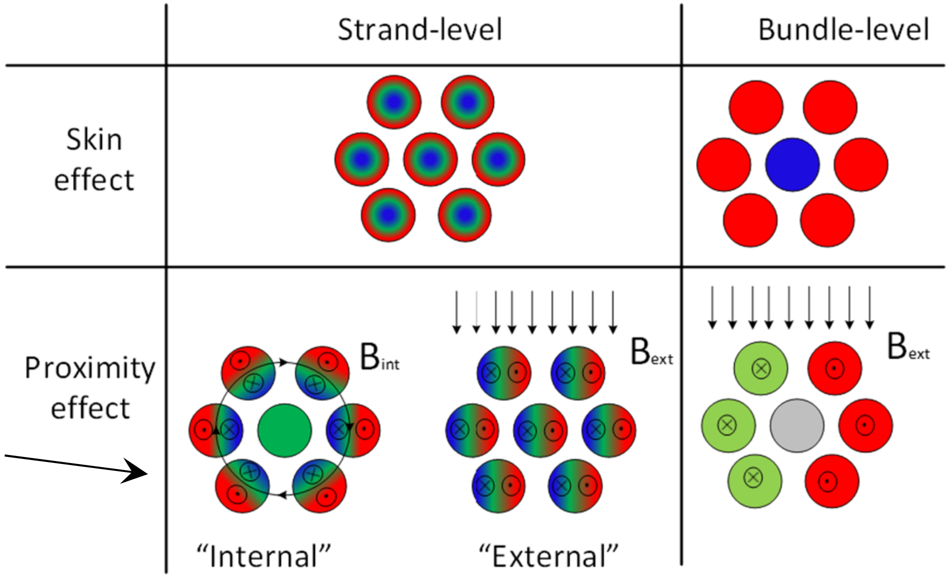
$$\frac{R_{\text{ac}}}{R_{\text{dc}}} = 0.25 + (0.5)\left(\frac{r_o}{\delta_0}\right) = 0.25 + (0.5)\left(\frac{1.0}{0.295}\right) = 1.95$$



Round versus foil conductor

Could replace solid wire with stranded Litz wire

Litz Wire



Avoid with radial and angular transposition

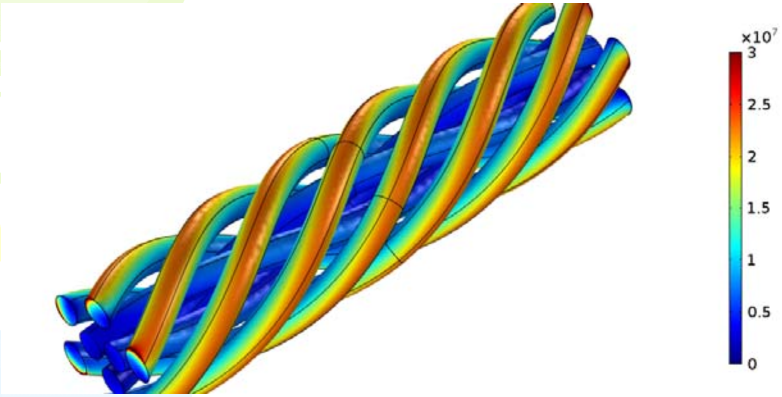
Negligible

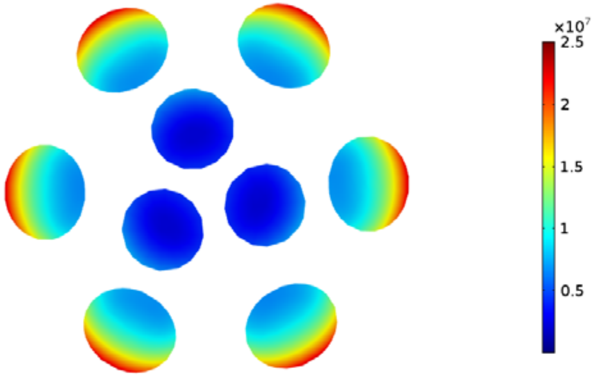
Avoid with radial transposition, simple twisting

- Litz wire reduces the window utilisation factor, core may be 30% larger for same temperature rise
- Use strands with diameter less than δ/4
- Proximity effect occurs at strand level when wire is twisted
- Twisting cancels proximity effect at bundle level

Sullivan C. R., Zhang R. Y., "Analytical Model for Effects of Twisting on Litz-wire Losses", IEEE 15th Workshop on Control and Modelling for Power Electronics, (COMPEL), pp. 1-10. 2014

Litz Wire: skin effect

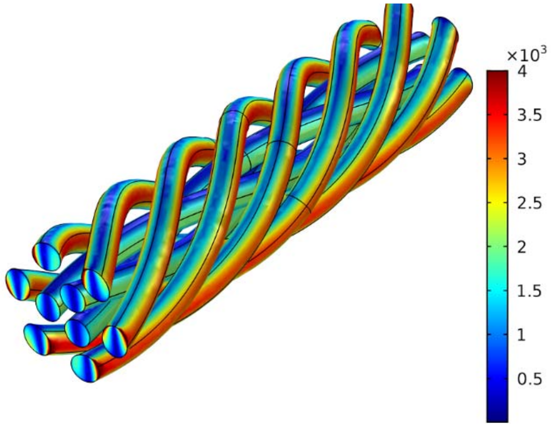


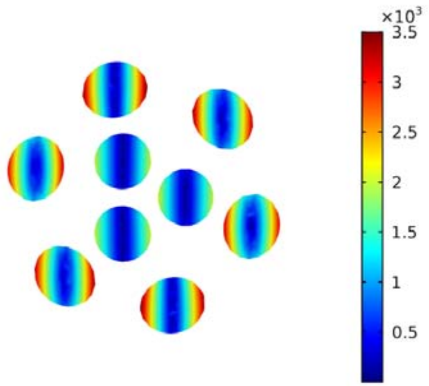


- Skin effect acts like a solid conductor at the bundle level
- Use strands with diameter less than δ/4

Acero J., Lope I., Burdio J.M., Carretero C., Alonso R., “Loss Analysis of Multistranded Twisted Wires by Usinf 3D-FEA Simulation”, IEEE 15th Workshop on Control and Modelling for Power Electronics, (COMPEL), pp. 1-6, 2014

Litz Wire: proximity effect

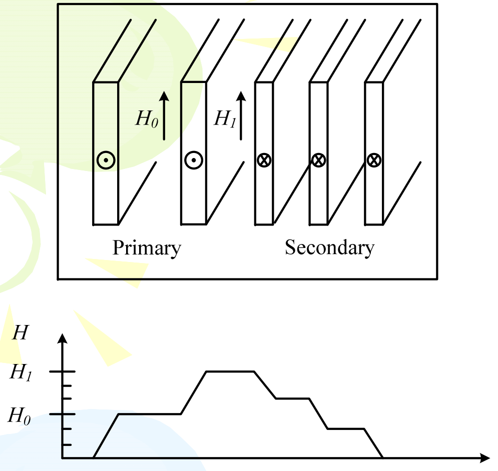


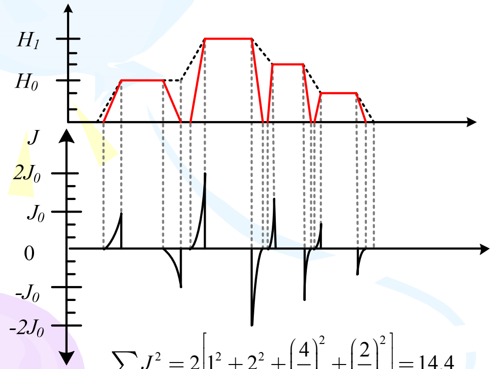


- Proximity effect occurs at strand level when wire is twisted
- Twisting cancels proximity effect at bundle level

Acero J., Lope I., Burdio J.M., Carretero C., Alonso R., “Loss Analysis of Multistranded Twisted Wires by Usinf 3D-FEA Simulation”, IEEE 15th Workshop on Control and Modelling for Power Electronics, (COMPEL), pp. 1-6, 2014

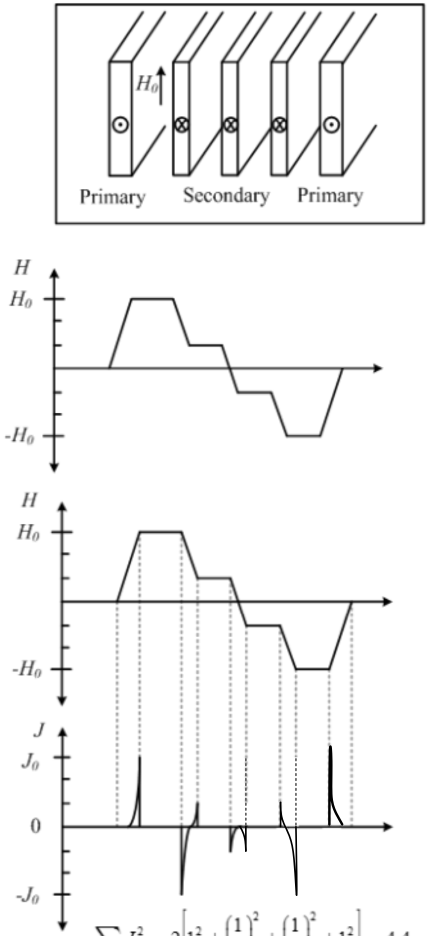
Interleaving the Windings





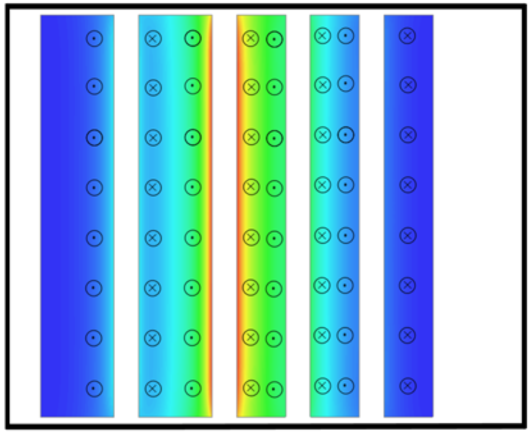
$$\sum J^2 = 2\left[1^2 + 2^2 + \left(\frac{4}{3}\right)^2 + \left(\frac{2}{3}\right)^2\right] = 14.4$$

Current density distribution before interleaving

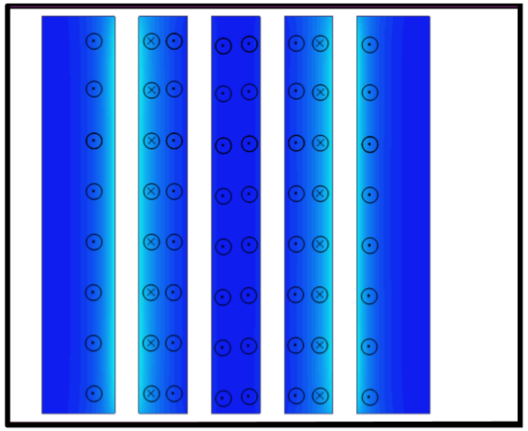


$$\sum J^2 = 2\left[1^2 + \left(\frac{1}{3}\right)^2 + \left(\frac{1}{3}\right)^2 + 1^2\right] = 4.4$$

Current density distribution after interleaving



Current density distribution before interleaving in FEA

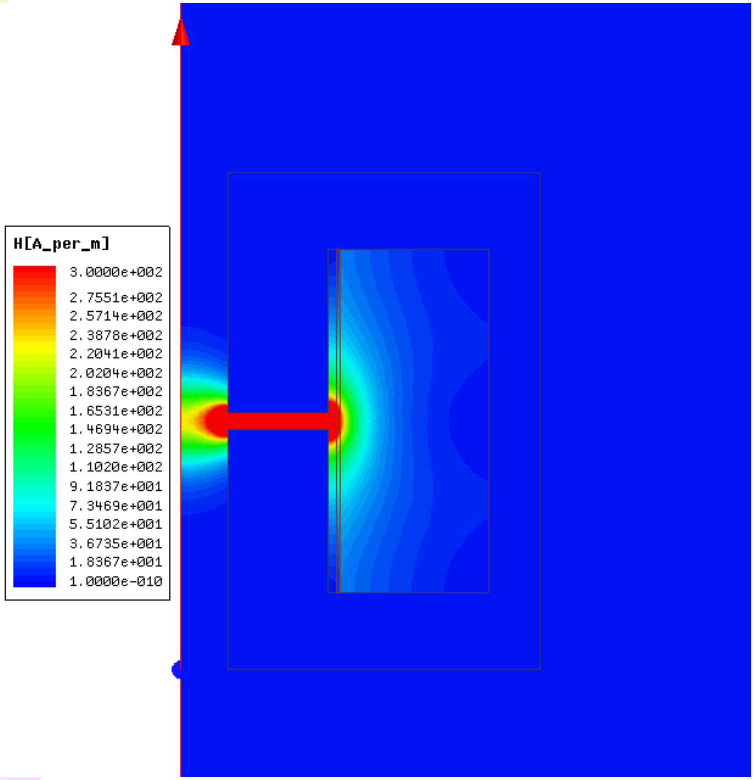


Current density distribution after interleaving in FEA

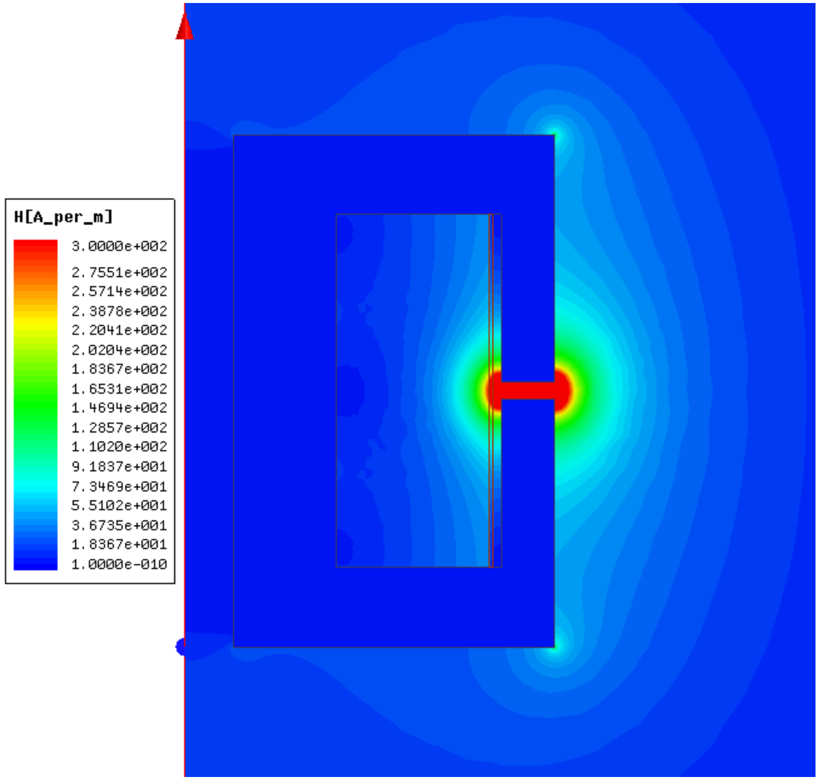
Fringing (Magnetic Field)

Frequency: 100kHz

Core: Magnetics® port core

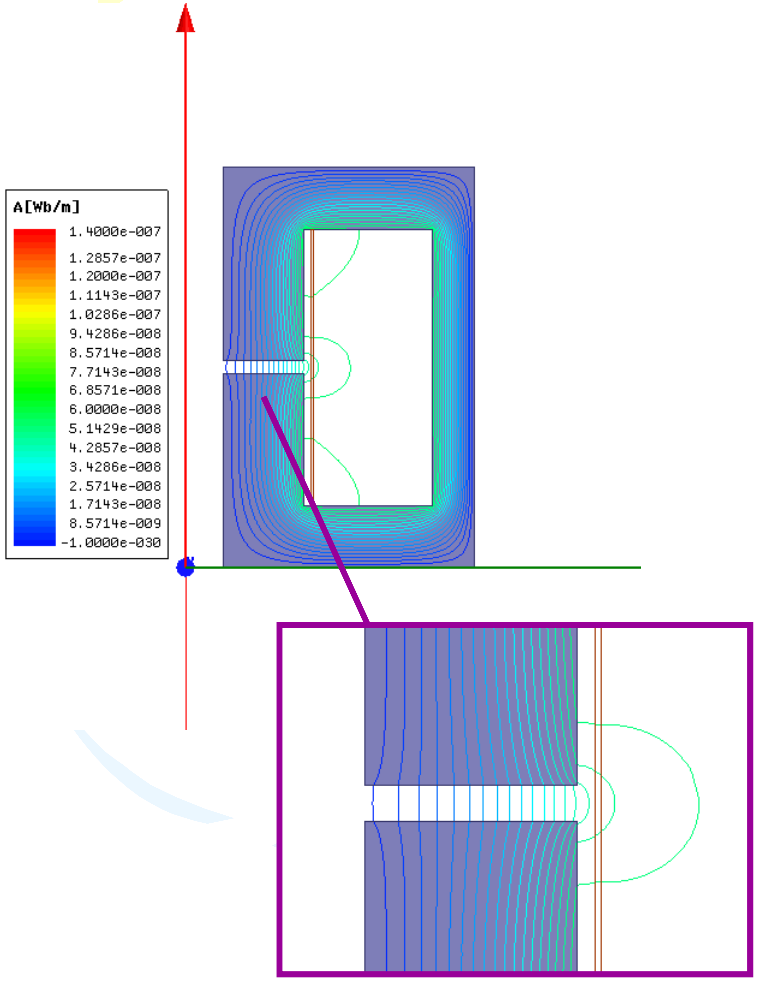


Gap in the centre leg

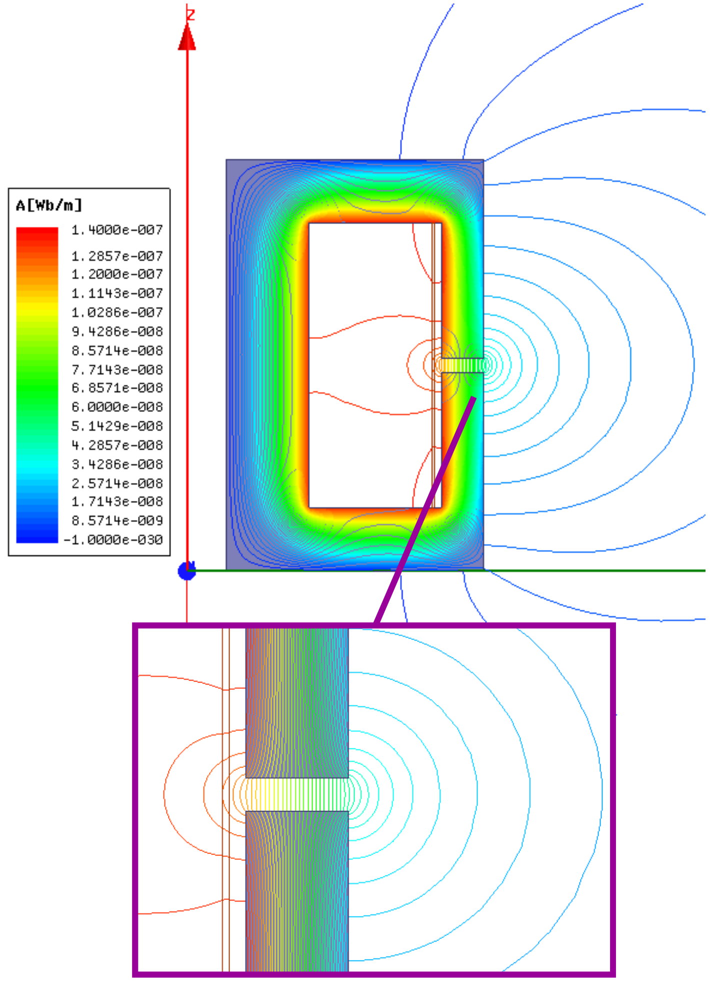


Gap in the outer leg

Fringing (Flux)



Gap in the centre leg



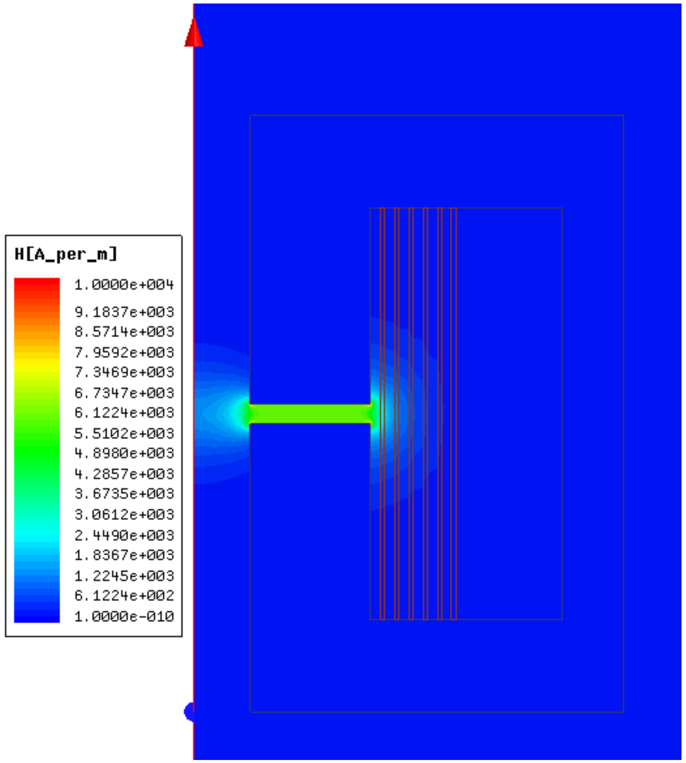
Gap in the outer leg

Fringing (Different Frequencies)

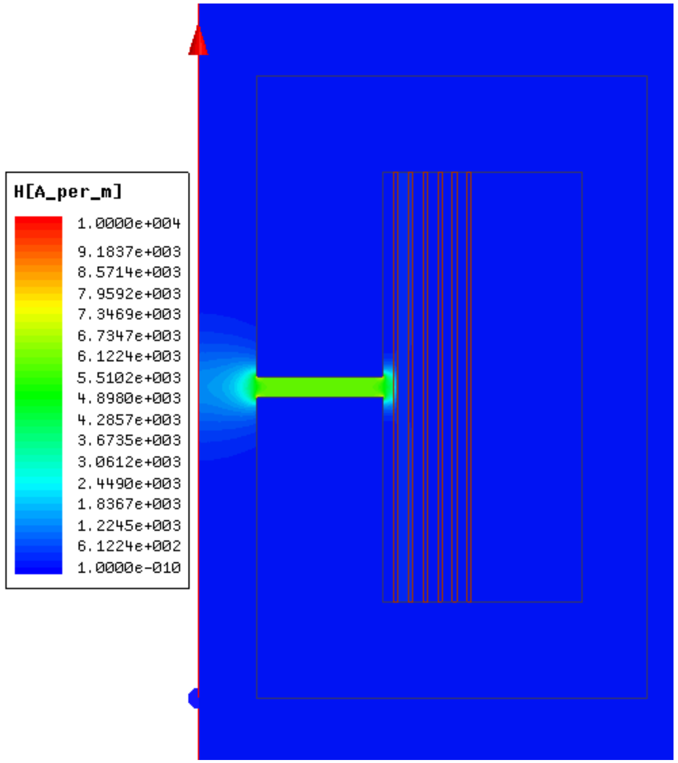
Magnetic Field Intensity

Width of conductor: 0.2mm

Core: Magnetics® port core



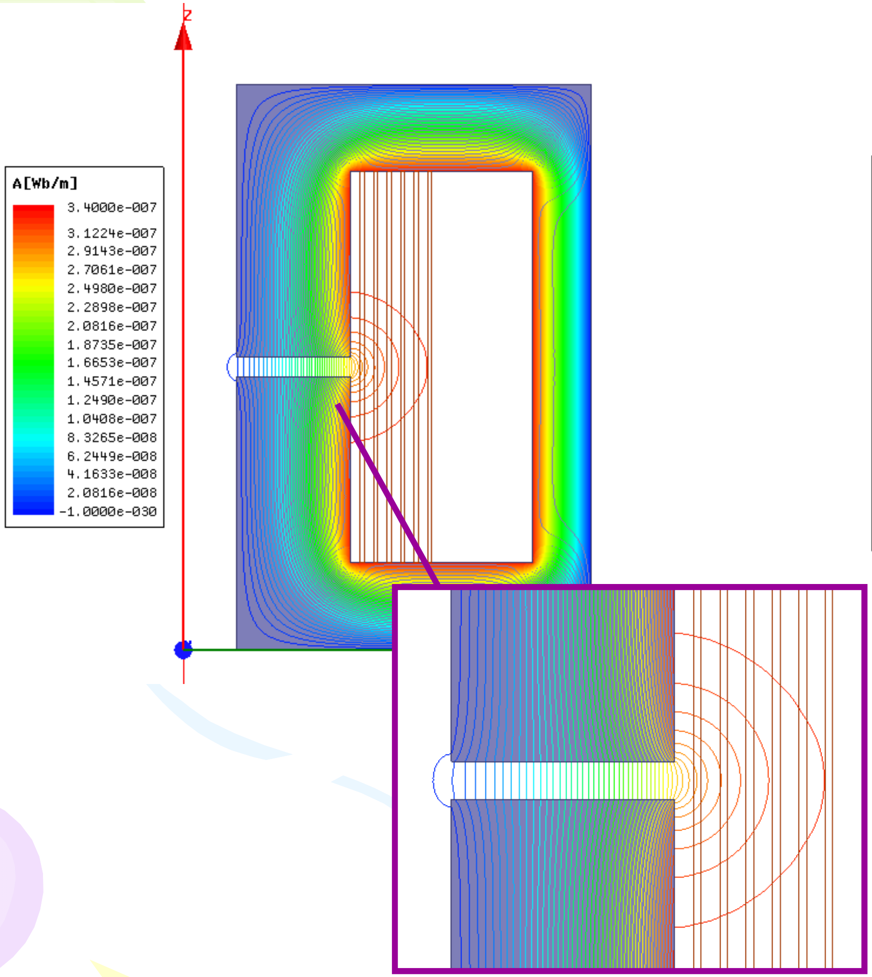
Frequency 1kHz



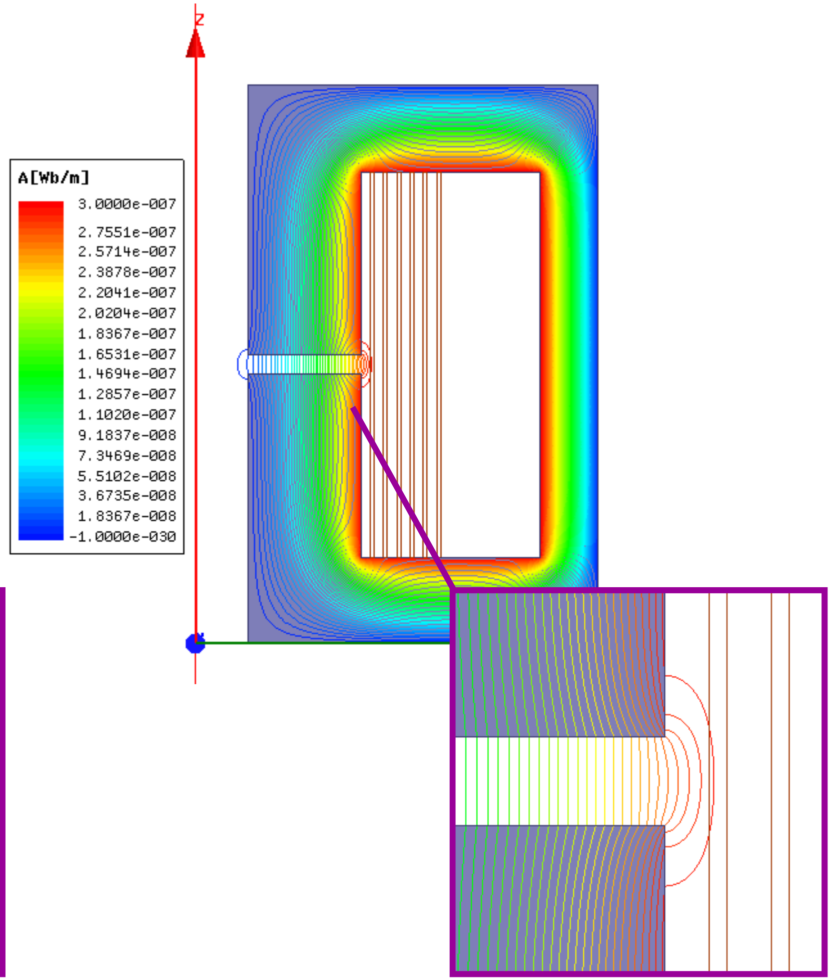
Frequency 100kHz

Fringing (Different Frequencies)

Magnetic Flux



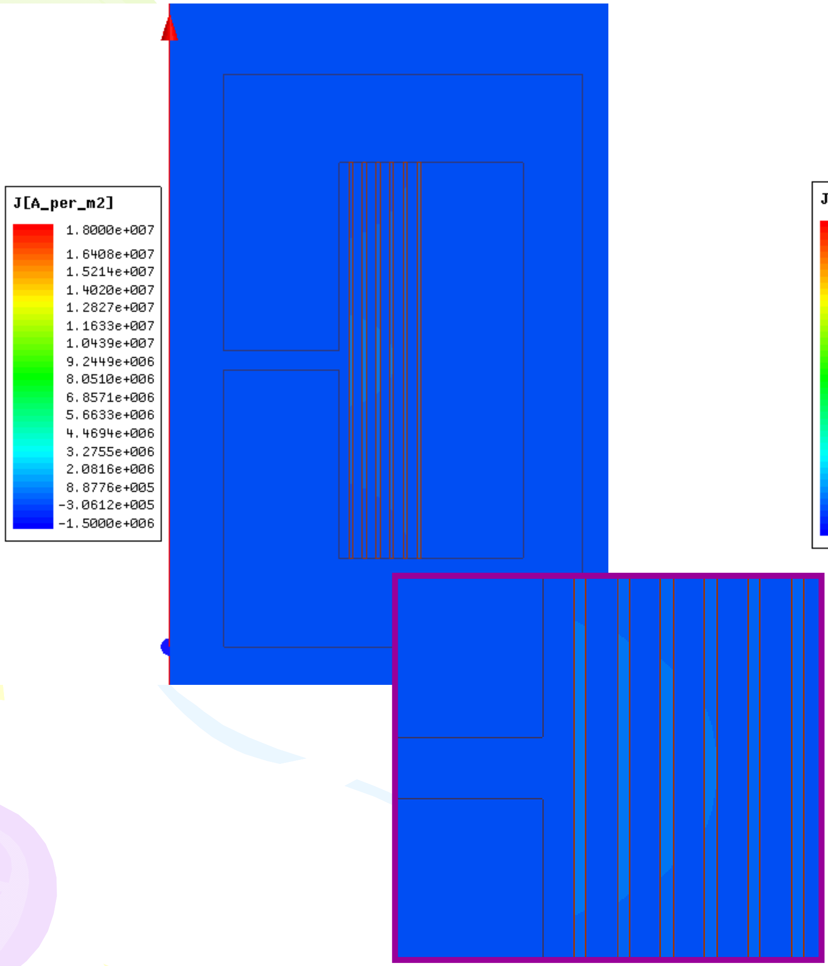
Frequency 1kHz



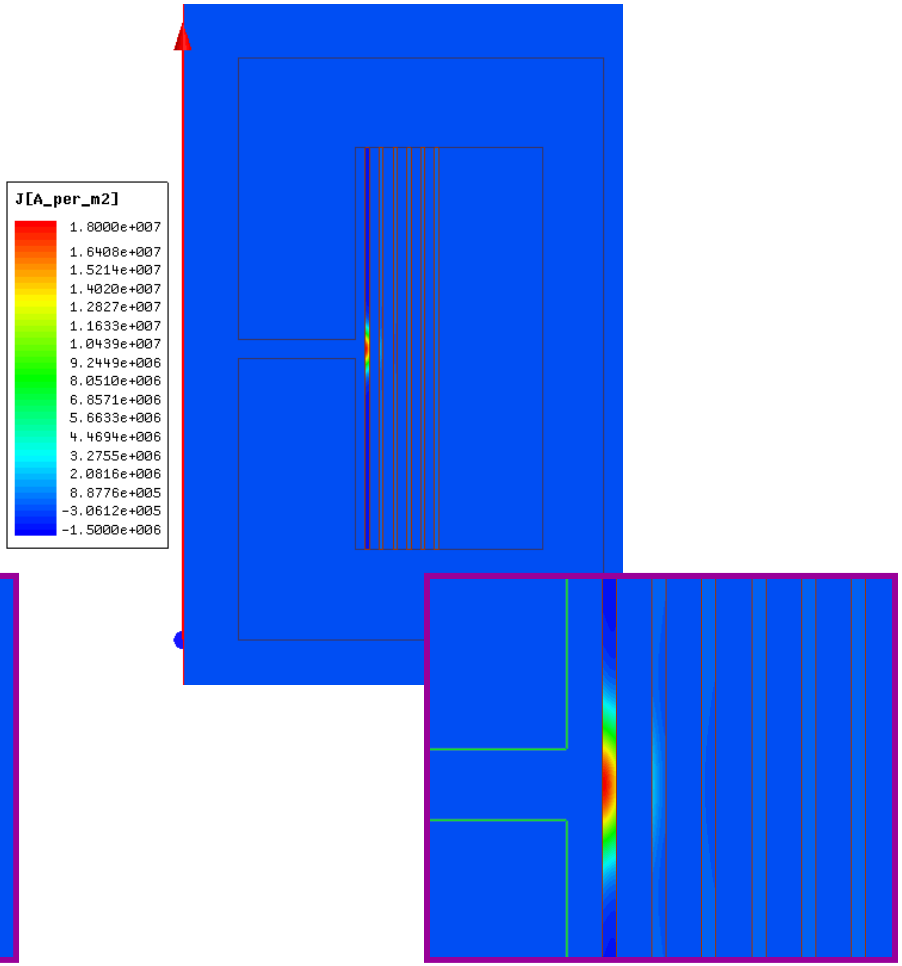
Frequency 100kHz

Fringing (Different Frequencies)

Current Density



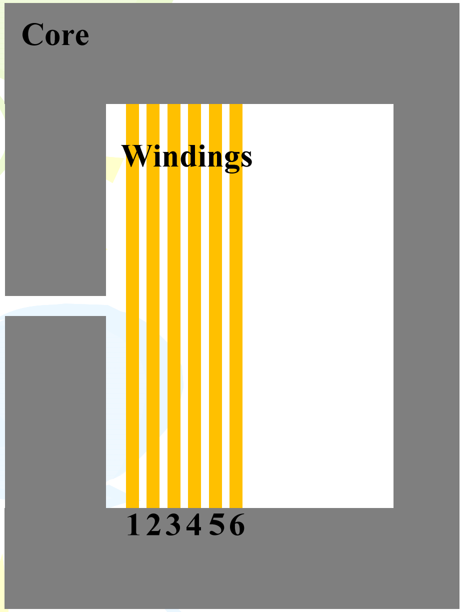
Frequency 1kHz



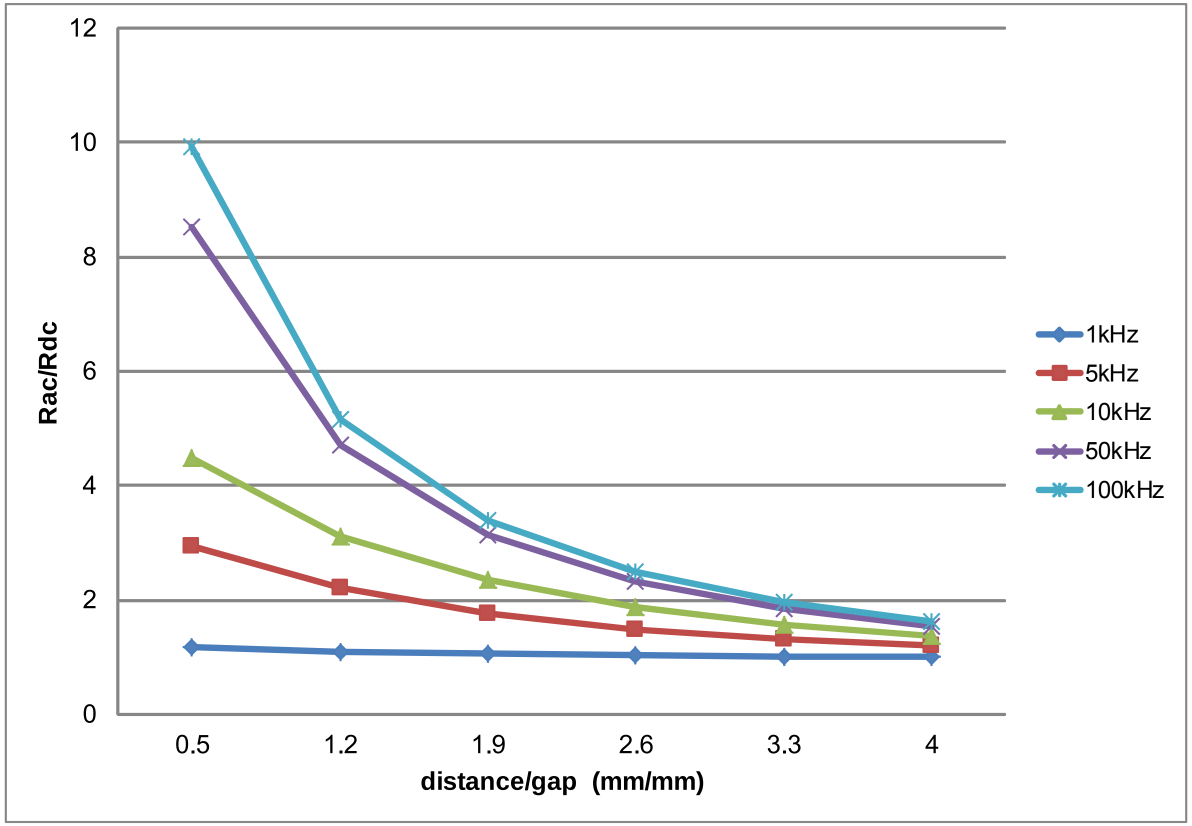
Frequency 100kHz

Winding Resistance

Winding Resistance related to the fringing effect, g=1mm



Distance 1:	0.5mm
Distance 2:	1.2mm
Distance 3:	1.9mm
Distance 4:	2.6mm
Distance 5:	3.3mm
Distance 6:	4.0mm



Analytical Models for Planar Magnetics

Advantages

- **Low profile** — planar magnetic components has a lower profile that their wire wound counterparts due to the fabrication process;

- **Automation** — it is difficult to automate the winding of conventional inductors and transformers, the processes used in planar magnetics are based on advanced computer aided manufacturing techniques. Suitable for SMT

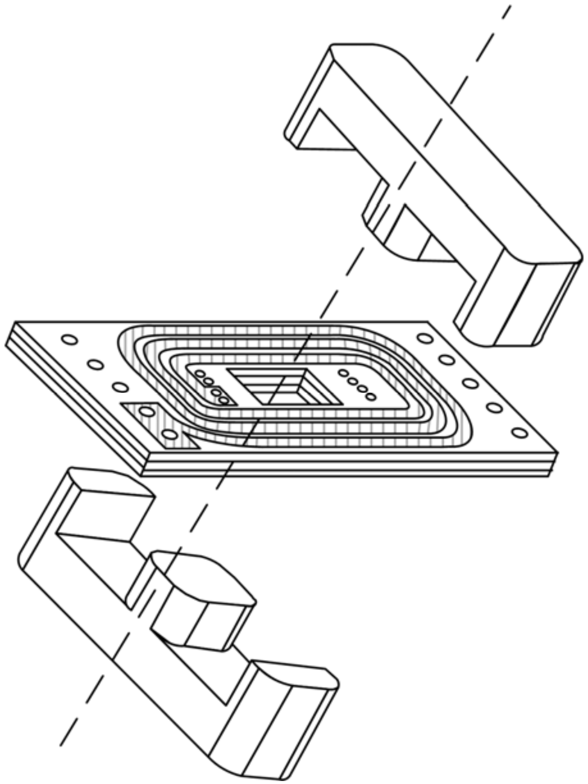
- **High power densities** — planar inductors and transformers are spread out and this gives them a bigger surface-to-volume ratio than conventional components, this enhances the thermal performance;

- **Predictable parasitics** — with planar magnetics, the windings are precise and consistent, yielding magnetic designs with highly controllable and predictable characteristic parameters.

Disadvantages

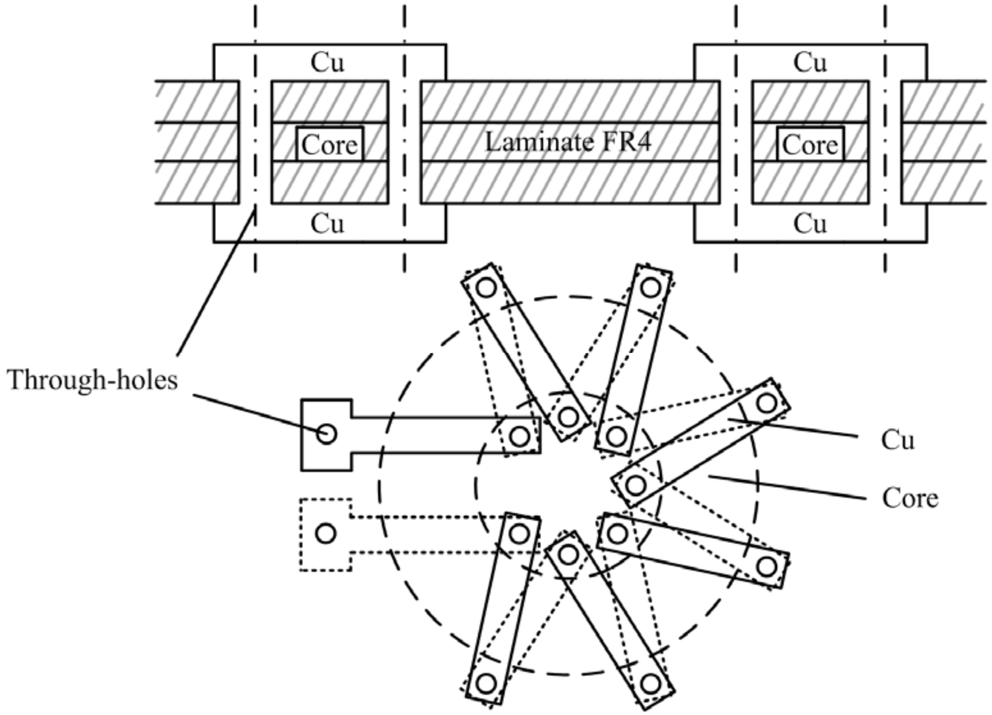
- **Turns** — the number of turns in planar device tends to be limited by the manufacturing process;
- **Footprint** — larger footprint compared with its conventional counterpart;
- **Capacitance** — interlayer capacitance introduces resonance at high frequencies;
- **Trade-off** — between magnetic core area and winding window area;
  between the path length versus the mean length of a turn.

Discrete PCB Magnetics



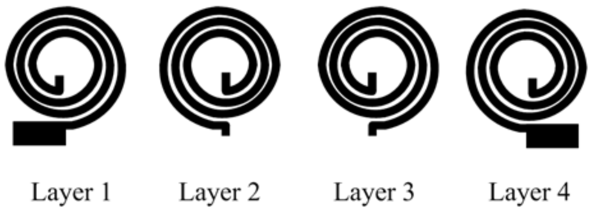
A typical planar transformer with an E-I core

Integrated PCB Magnetics

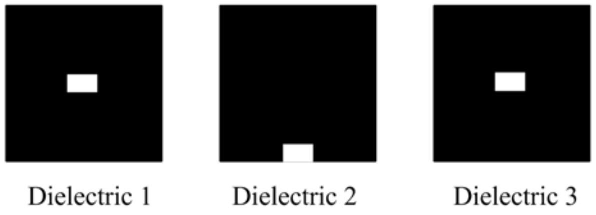


PCB integrated magnetic toroidal transformer

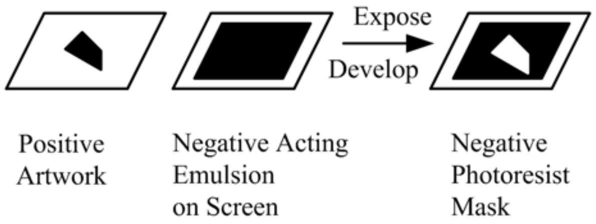
Thick Film Devices: Photoplots



Photoplots of conducting layers

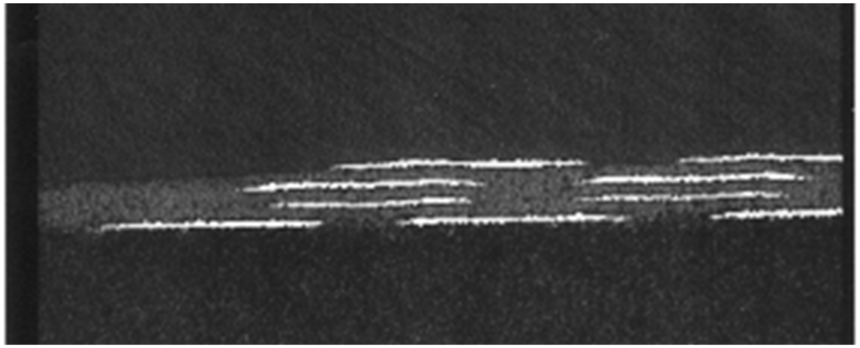


Masks for dielectric layers



Screen generation

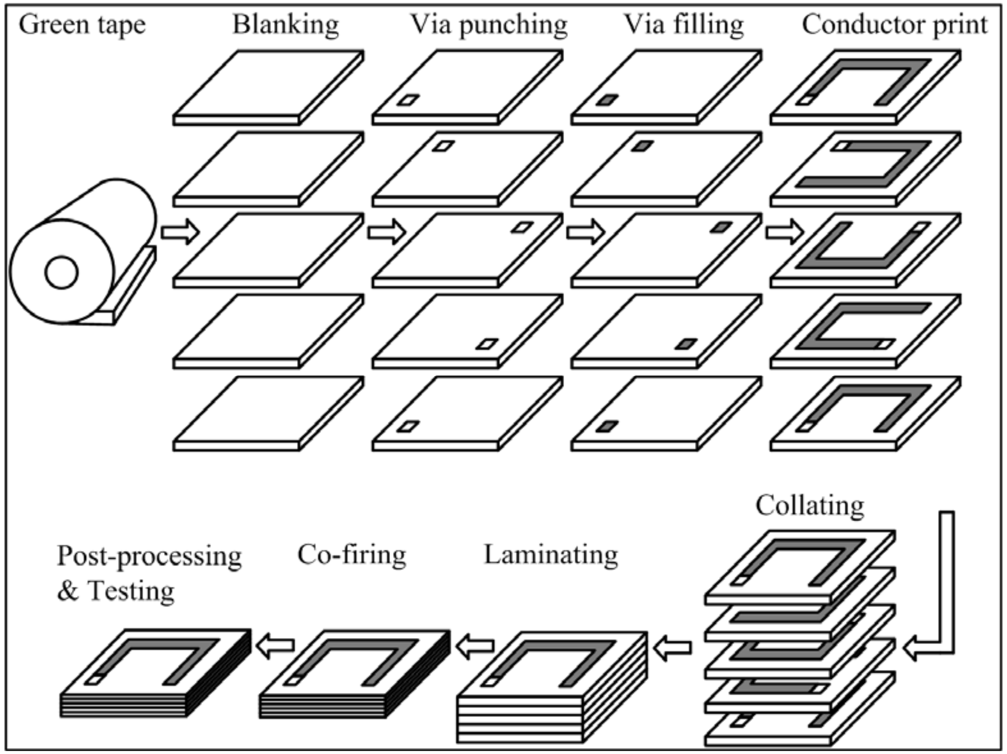
Thick Film Devices: Microsection



Optical photograph of a microsectioned device (scale 30:1) [1] Reproduced with permission from [1]. Copyright 1999 IEEE.

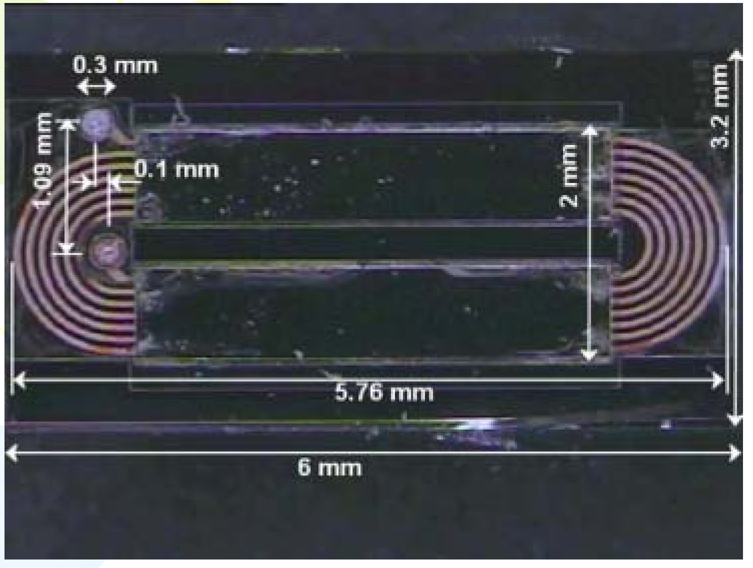
[1] W. **G. Hurley, M. C. Duffy, S. O'Reilly, and S. C. O'Mathuna, 'Impedance formulas for planar magnetic structures with spiral windings,' IEEE Transactions on Industrial Electronics, vol. 46(2), pp. 271-278, 1999.**

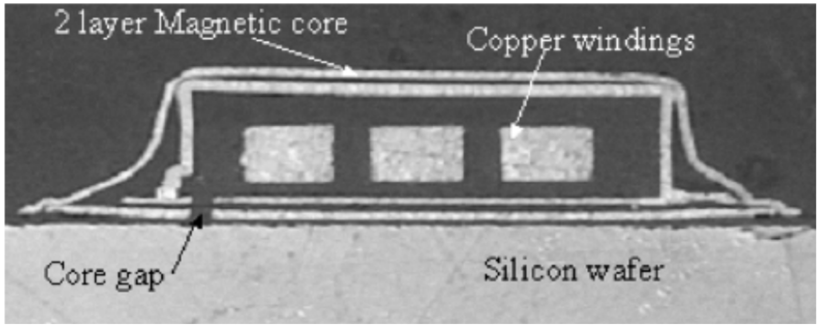
Low Temperature Co-fired Ceramics



LTCC Process flow

Silicon Integrated Microinductor



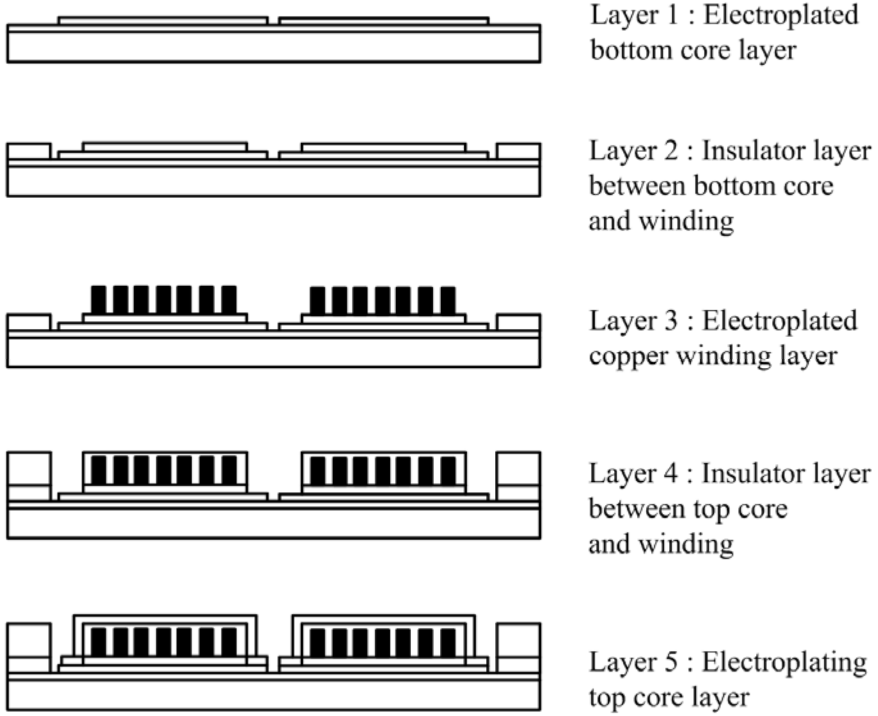


Silicon integrated microinductor: (a) Top view. Reproduced with permission from [1]. Copyright 2008 IEEE, (b) Cross-section. Reproduced with permission from [2]. Copyright 2005 IEEE

[1] **W. Ningning, T. O'Donnell, R. Meere, F. M. F. Rhen, S. Roy, and S. C. O'Mathuna, 'Thin-Film-Integrated Power Inductor on Si and Its Performance in an 8-MHz Buck Converter,' IEEE Transactions on Magnetics, vol. 44(11), pp. 4096-4099, 2008.**

[2] **S. C. O. Mathuna, T. O'Donnell, W. Ningning, and K. Rinne, 'Magnetics on silicon: an enabling technology for power supply on chip,' IEEE Transactions on Power Electronics, vol. 20(3), pp. 585-592, 2005.**

Thin Film Devices: Microfabrication



Microfabrication process flow for an inductor

Technology Comparison

Technology	Frequency (Typical)	Power (Typical)	Inductance (Typical)	Size (Typical)
PCB magnetics	20 KHz ~ 2 MHz	1 W ~ 5 kW	10 µH ~ 10 mH	100 mm^2 ~ 100's cm^2
Thick Film	< 10 MHz	< 10 W	1 µH ~ 1 mH	< 1 cm^2
LTCC	200 KHz ~ 10 MHz	< 10W	1 µH ~ 1 mH	< 1cm^2
Thin Film	> 10 MHz	< 1W	10's ~ 100's nH	< 10mm^2

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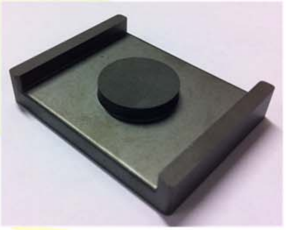
Advantages and Disadvantages

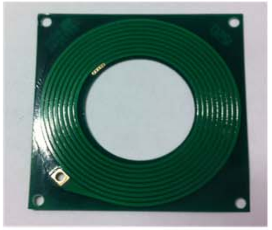
Technology	Integration method	Advantages	Disadvantage
PCB	Discrete core on laminated structure or integrated core in laminated structure Parallel or sequential process	Low cost Multilayer structure Thick copper High current High inductance	Low resolution (line width 100µm) Relatively low frequency
Thick Film	Screen printed on sintered ceramic Sequential build up of multiple layers	Low cost	Difficult to form Long process time and low yield due to sequential build-up
LTCC	Screen printed on green tapes Parallel multilayer and final cofired structure	Parallel layer process High layer counts Module reliability	Cofireability of materials
Thin Film	Sequential build up of lithographically defined layers	Precision value (line width 5µm) High tolerance High component density High frequency	Low inductance Equipment costly Limited selection on film materials/ material compatibility

Planar Transformer

Planar Core ER41

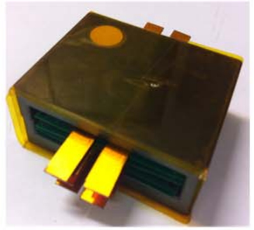


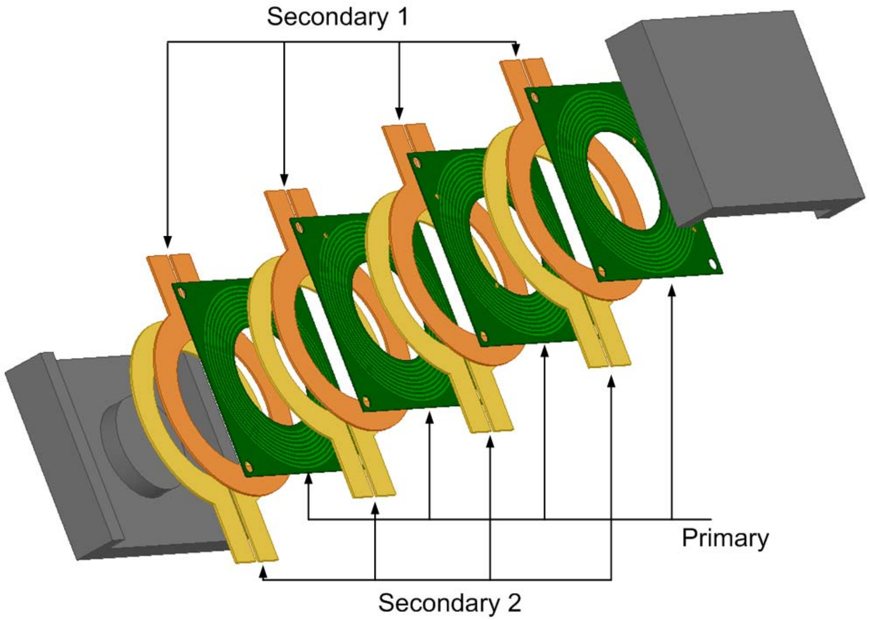
Primary PCB Windings

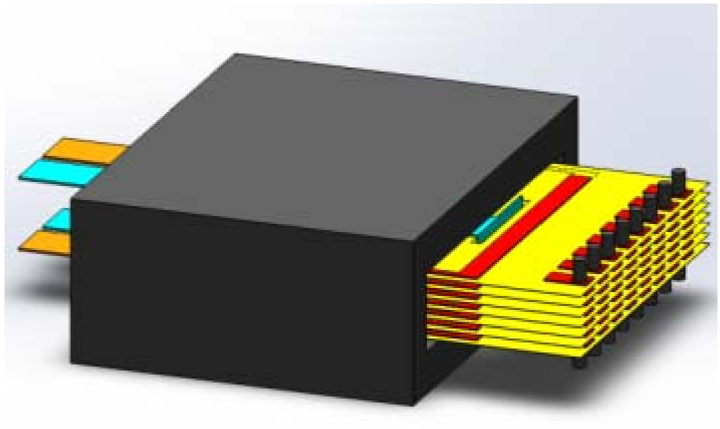


Secondary Foil Windings

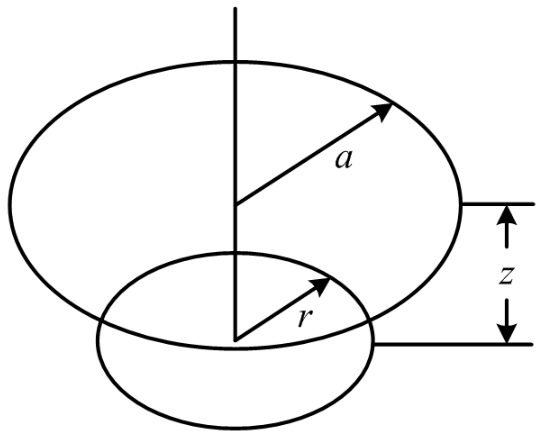
Planar Transformer







Spiral Coil in Air



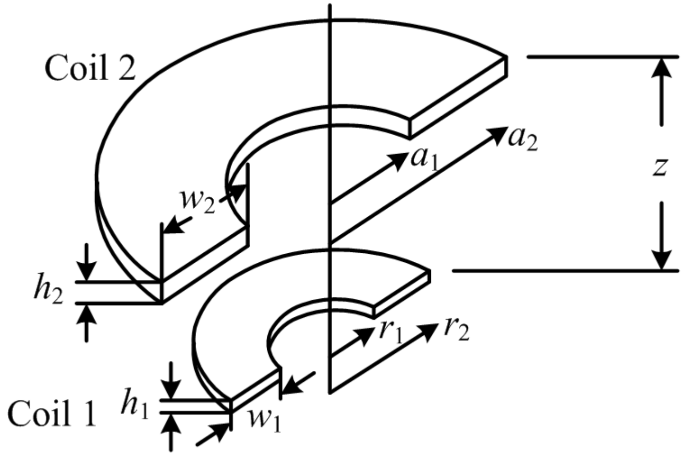
Circular concentric filaments in air

$$M = \mu_0 \sqrt{ar}\frac{2}{f}[(1-\frac{f^2}{2})K(f) - E(f)]$$

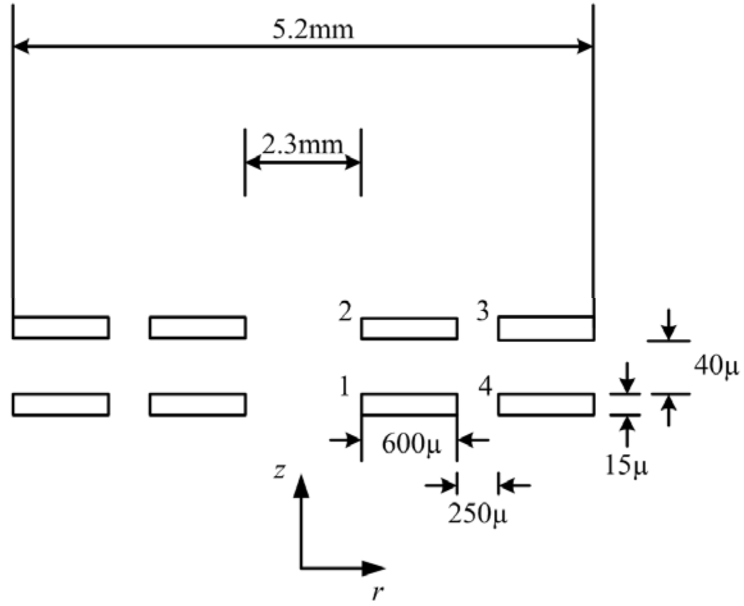
$$f = \sqrt{\frac{4ar}{z^2 + (a+r)^2}}$$

$K(f)$: complete elliptic integrals of the first kind
$E(f)$: complete elliptic integrals of the second kind

Mutual Inductance of Planar Coils



Planar coils of rectangular cross-section



$$M_{12} = \frac{\mu_0 \pi}{h_1 h_2 \ln\left(\frac{r_2}{r_1}\right) \ln\left(\frac{a_2}{a_1}\right)} \int_0^\infty S(kr_2, kr_1) S(ka_2, ka_1) Q(kh_1, kh_2) e^{-k|z|} dk$$

$$S(kx, ky) = \frac{J_0(kx) - J_0(ky)}{k}$$

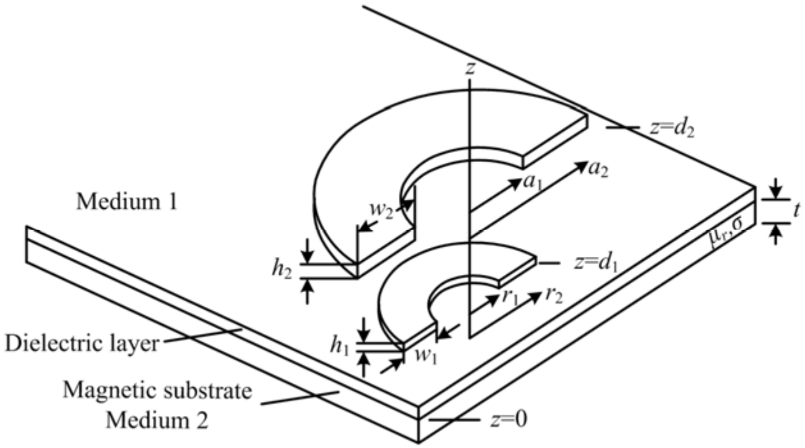
$$Q(kx, ky) = \frac{2}{k^2}\left[\cosh k\frac{x+y}{2} - \cosh k\frac{x-y}{2}\right] \qquad z > \frac{h_1 + h_2}{2}$$

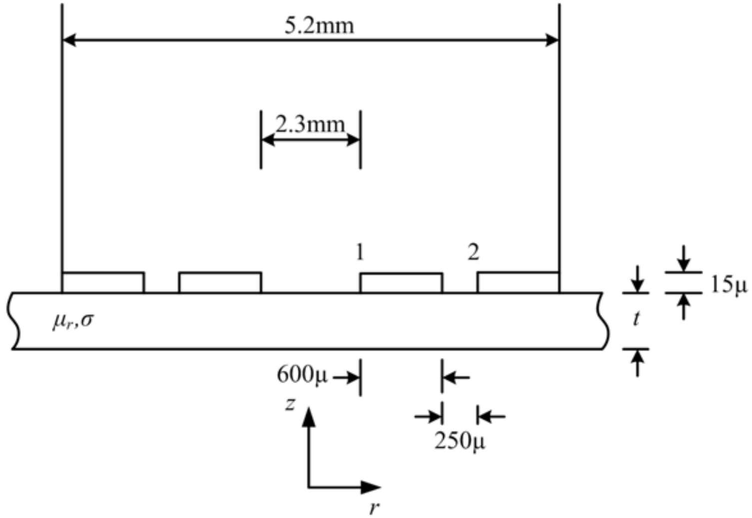
$$= \frac{2}{k}\left(h + \frac{e^{-kh} - 1}{k}\right) \qquad z = 0, x = y = h$$

Solve with MATLAB

[1] W G Hurley, M C Duffy, J Zhang, I Lope, B Kunz, W H Wölfle, , "A Unified Approach to the Calculation of Self and Mutual Inductance for Coaxial Coils in Air", IEEE Trans. on Power Electronics, vol. 30, no.11, pp. 6155–6162, November 2015.

Add a Magnetic Substrate

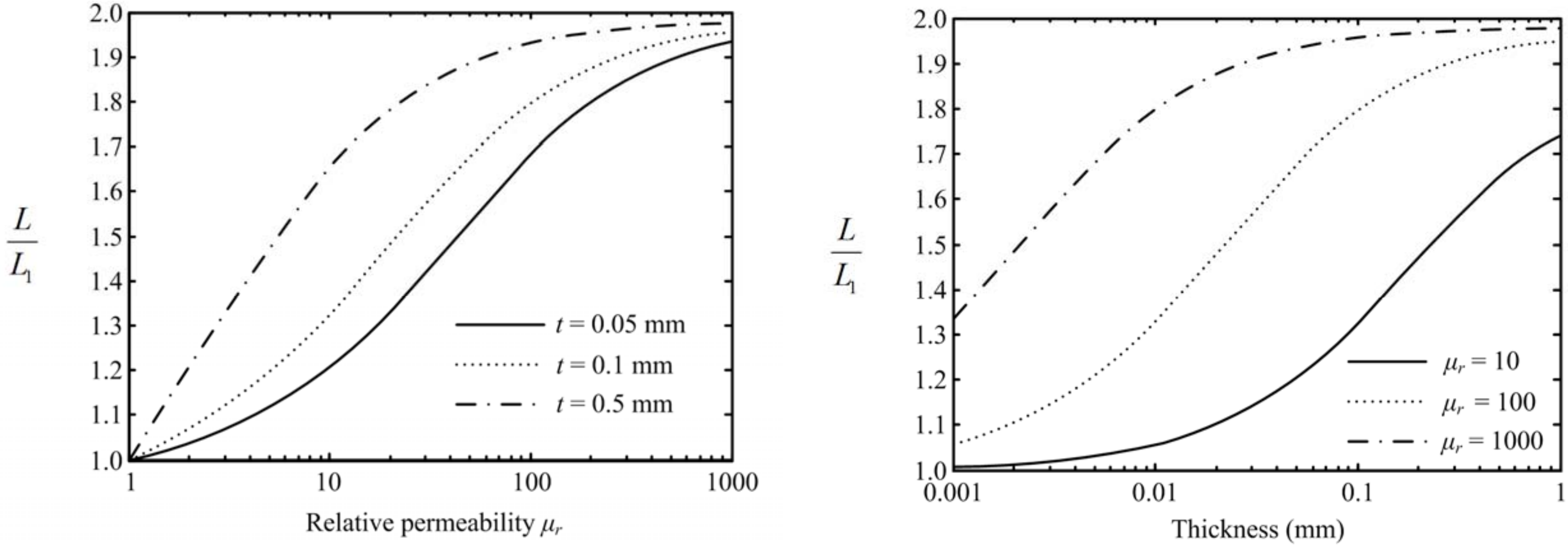




Planar coils on a magnetic substrate

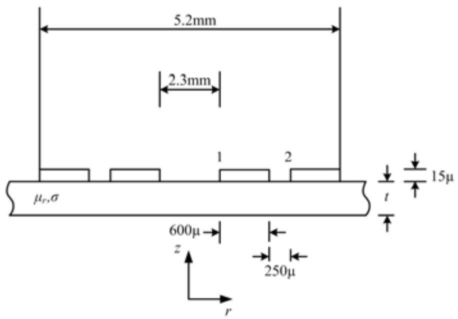
$$Z_t^p = \frac{j\omega\mu_0\pi}{h_1 h_2 \ln(\frac{r_2}{r_1})\ln(\frac{a_2}{a_1})}\int_0^\infty S(kr_2,kr_1)S(ka_2,ka_1)Q(kh_1,kh_2)\lambda(t)e^{-k(d_1+d_2)}dk$$

Effect of Permeability and Thickness

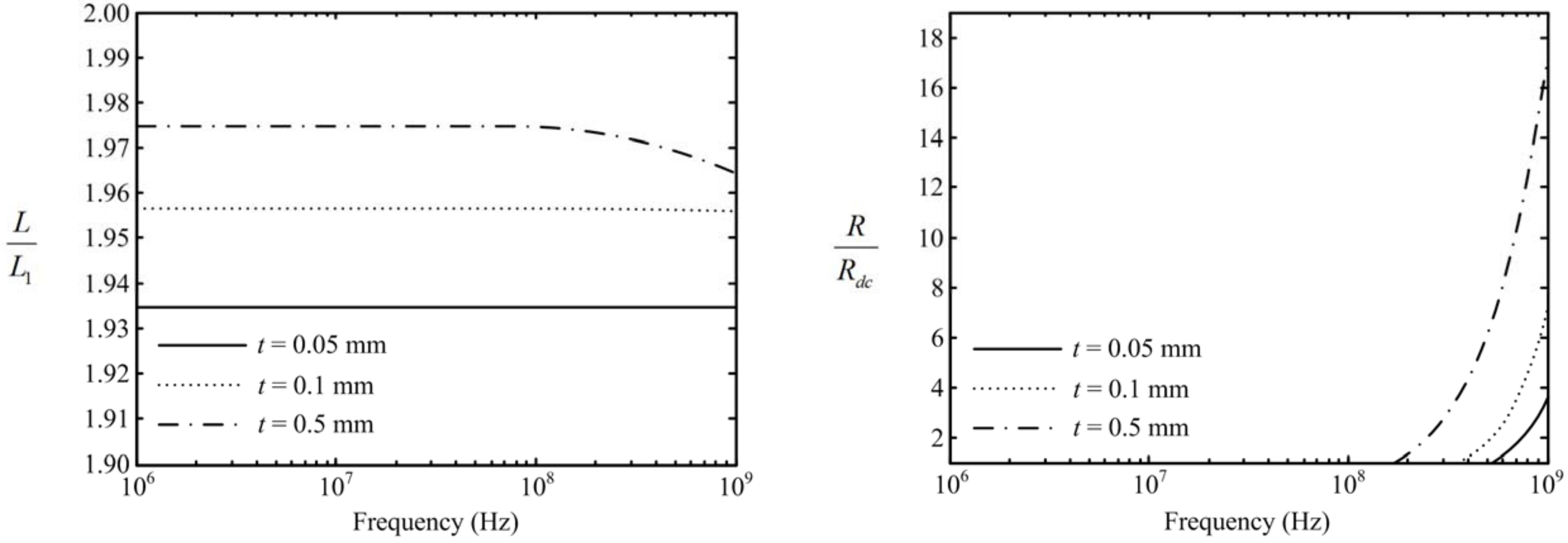


(a) (b)

Enhancement of inductance with magnetic substrate: (a) as a function of μ_r, (b) as a function of t.



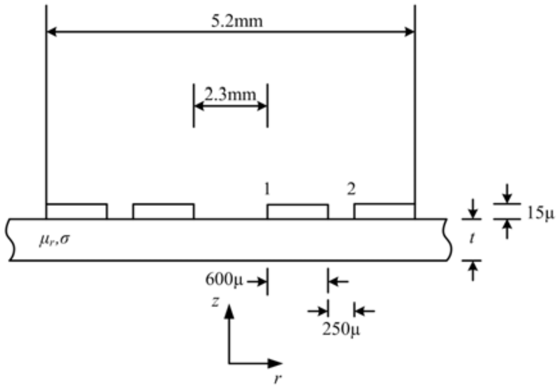
Effect of Frequency



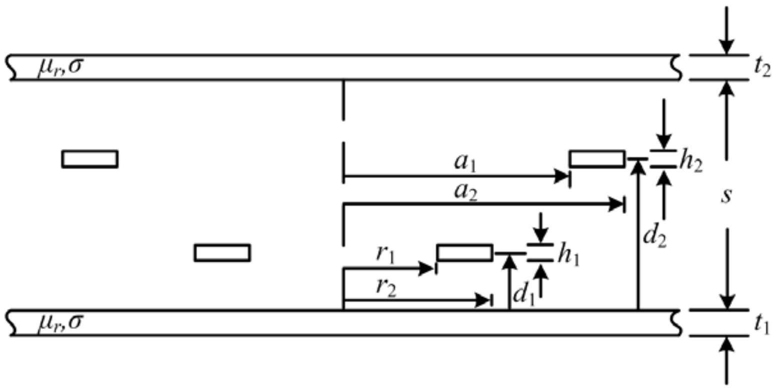
(a) (b)

Self impendence of a planar coil on a finite substrate: (a) inductance, (b) resistance.

$$g(\lambda) = \frac{2\lambda(t_1)\lambda(t_2)e^{-2ks}\cosh[k(d_2 - d_1)]}{1 - \lambda(t_1)\lambda(t_2)e^{-2ks}}$$



Sandwich Structure



Planar coils in a sandwich structure

$$Z_{s\omega}^{p} = \frac{j\omega\mu_0\pi}{h_1 h_2 \ln(\frac{r_2}{r_1})\ln(\frac{a_2}{a_1})} \int_0^\infty S(kr_2, kr_1)S(ka_2, ka_1)[f(\lambda) + g(\lambda)]Q(kh_1, kh_2)dk$$

$$f(\lambda) = \frac{\lambda(t_1)e^{-k(d_1+d_2)} + \lambda(t_2)e^{-k(d_1'+d_2')}}{1-\lambda(t_1)\lambda(t_2)e^{-2ks}}$$

$$g(\lambda) = \frac{2\lambda(t_1)\lambda(t_2)e^{-2ks}\cosh[k(d_2-d_1)]}{1-\lambda(t_1)\lambda(t_2)e^{-2ks}}$$

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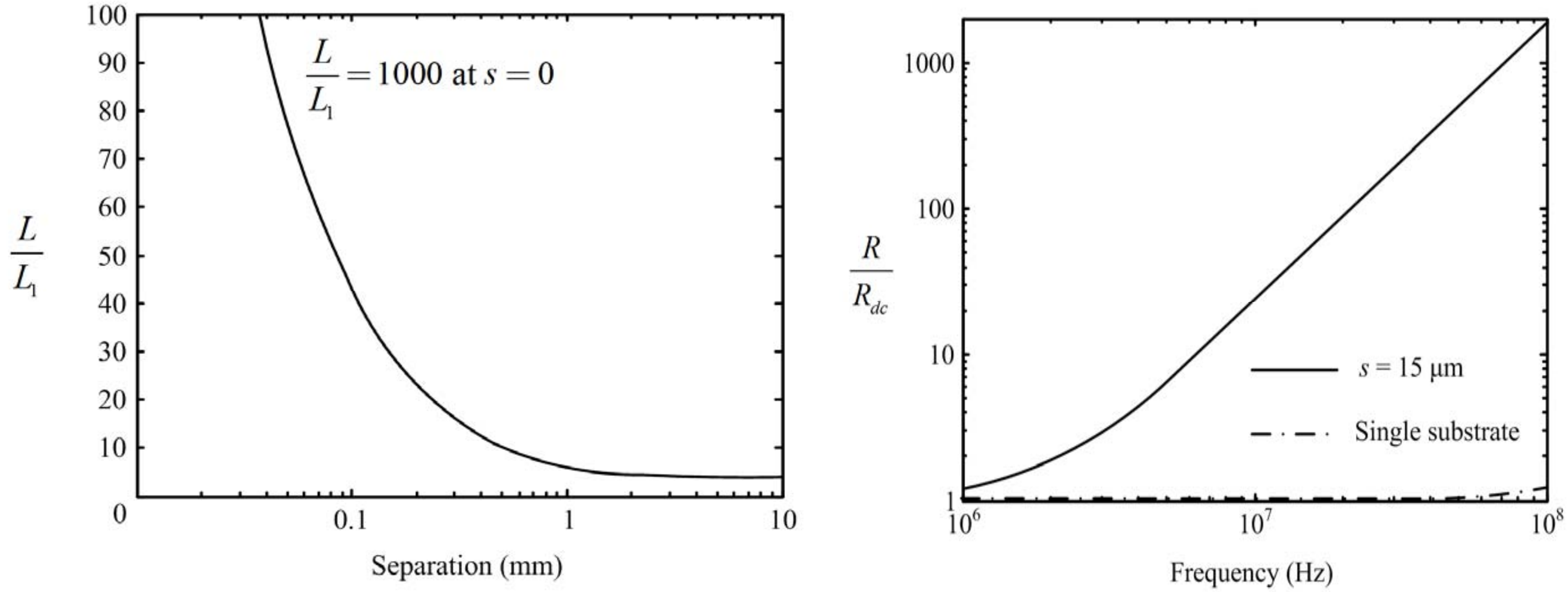
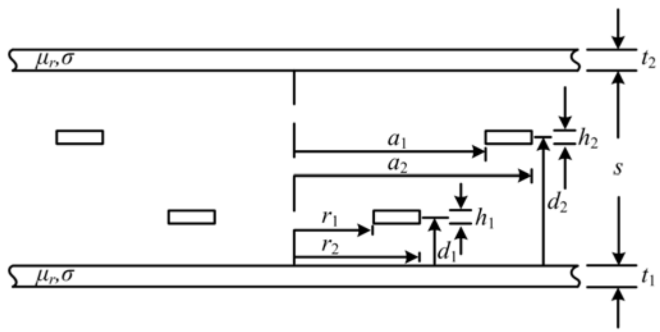
Separation and Frequency



Inductance as a function of substrate separation; L_1=17.14nH.



References

[1] C. P. Steinmetz, 'On the law of hysteresis,' Proceedings of the IEEE, vol. 72(2), pp. 197-221, 1984.
[2] W. G. Hurley and M. C. Duffy, 'Calculation of self and mutual impedances in planar magnetic structures,' IEEE Transactions on Magnetics, vol. 31(4), pp. 2416-2422, 1995.
[3] J. C. Maxwell, A Treatise on Electricity and Magnetism. Oxford: Clarendon Press, 1881.
[4] A. Gray, Absolute Measurements in Electricity and Magnetism. London: MacMillan, 1893.
[5] T. R. Lyle, Philosophical Transactions of the Royal Society of London. Series A, Containing Papers of a Mathematical or Physical Character. London: The Royal Society, 1914.
[6] H. B. Dwight, 'Some New Formulas for Reactance Coils,' Transactions of the American Institute of Electrical Engineers, vol. XXXVIII(2), pp. 1675-1696, 1919.
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[8] K. Venkatachalam, C. R. Sullivan, T. Abdallah, and H. Tacca, 'Accurate prediction of ferrite core loss with nonsinusoidal waveforms using only Steinmetz parameters,' in Proceedings of IEEE Workshop on Computers in Power Electronics, COMPEL, 2002, pp. 36-41.
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[10] F. Judd and D. Kressler, 'Design optimization of small low-frequency power transformers,' IEEE Transactions on Magnetics, vol. 13(4), pp. 1058-1069, 1977.
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[13] W. G. Hurley, D. J. Wilcox, and P. S. McNamara, 'Calculation of short circuit impedance and leakage impedance in transformer windings,' in Proceedings of the IEEE Power Electronics Specialists Conference, PESC, 1991, pp. 651-658.
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[15] W. G. Hurley, W. H. Wolfle, and J. G. Breslin, 'Optimized transformer design: inclusive of high-frequency effects,' IEEE Transactions on Power Electronics, vol. 13(4), pp. 651-659, 1998.
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[17] K. Venkatachalam, C. R. Sullivan, T. Abdallah, and H. Tacca, 'Accurate prediction of ferrite core loss with nonsinusoidal waveforms using only Steinmetz parameters,' in Proceedings of IEEE Workshop on Computers in Power Electronics, COMPEL, 2002, pp. 36-41.
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[19] E. C. Snelling, Soft Ferrites: Properties and Applications, 2nd edn. London: Butterworths, 1988.
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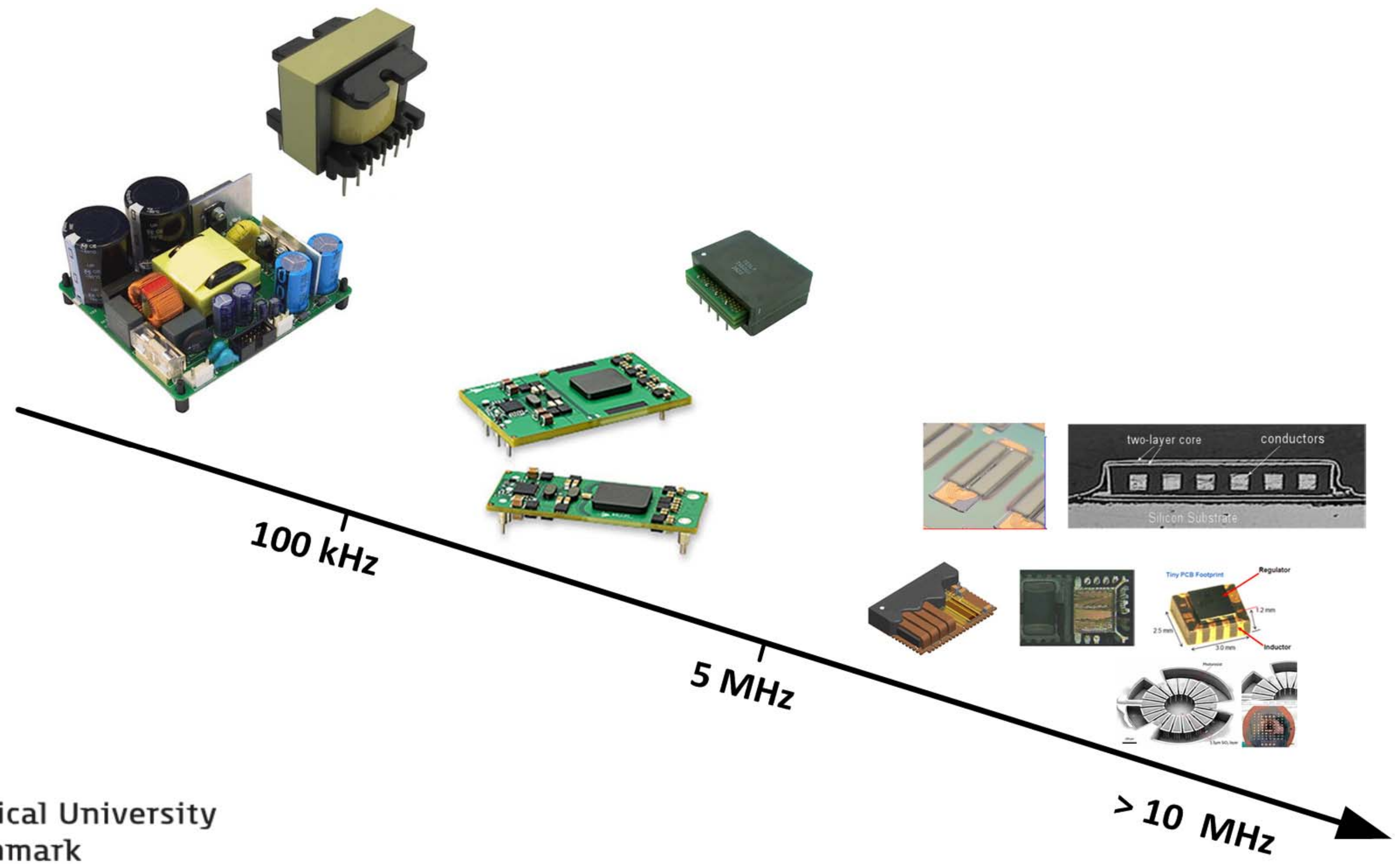
Planar Magnetics – Fundamentals

Introduction

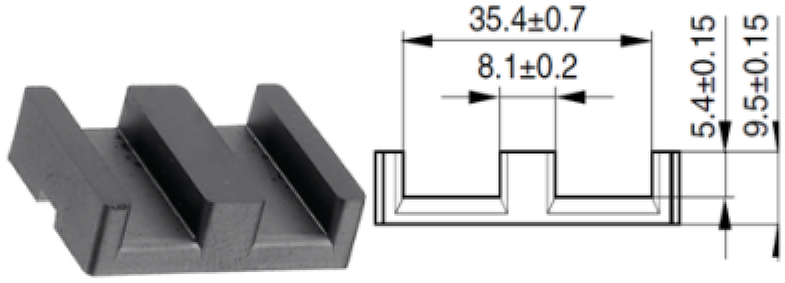
With rapidly increased frequencies, magnetics become **VERY important** factor to achieve high-efficiency and high-power-density converter.



Advantages

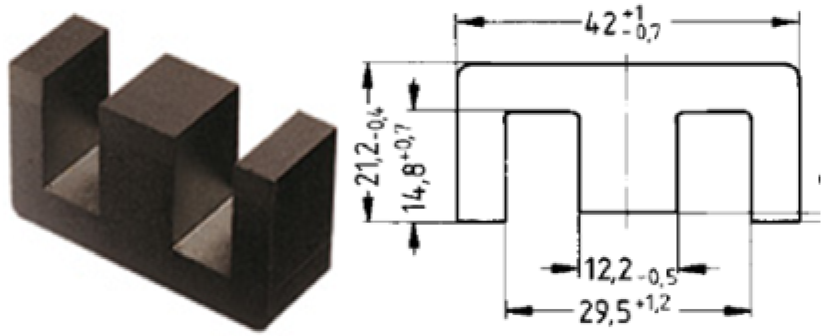
❖ **Low profile:** the height of a planar magnetic core is typically 25% to 50% the height of its wire-wound counterpart.

ELP 43/10/28



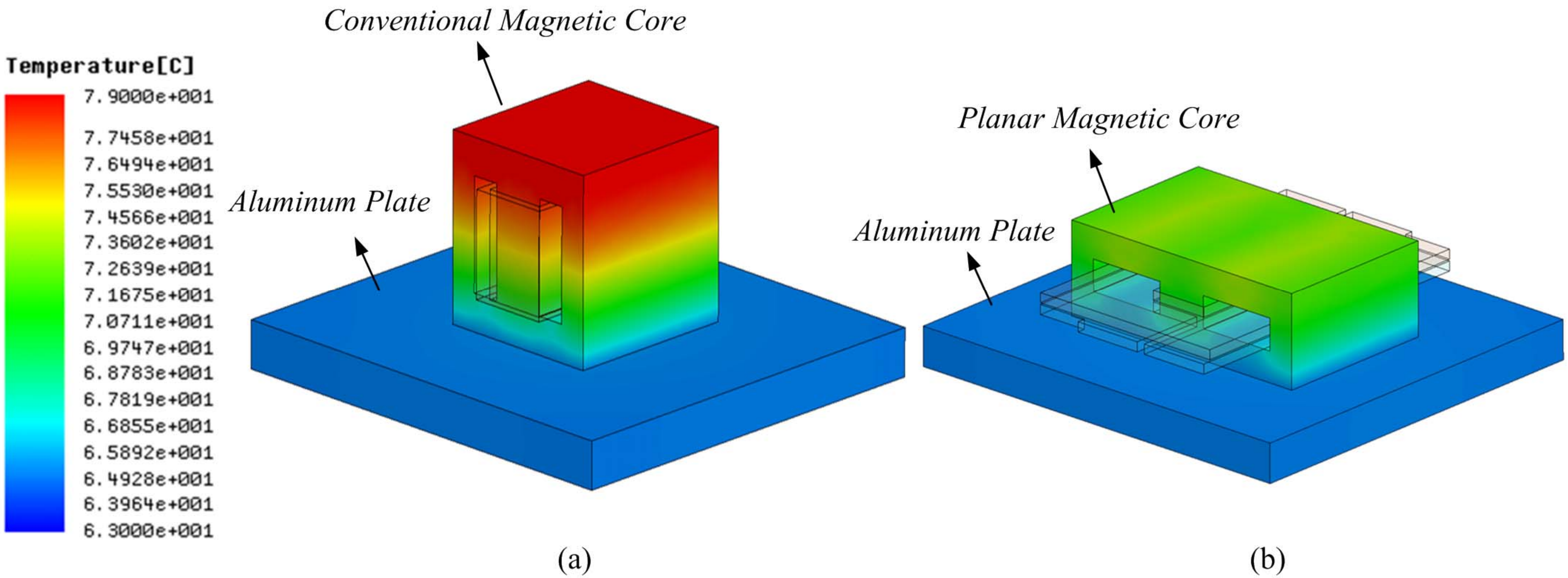
A_e = 229 mm²
A_{min} = 225 mm²
V_e = 14000 mm³

E 42/21/20



A_e = 234 mm²
A_{min} = 229 mm²
V_e = 22 700 mm³

❖ **Good thermal characteristic:** planar cores essentially have a higher surface area to volume ratio than conventional magnetic cores.



Finite Element Simulation

Advantages

- **Ease of manufacturability and cost reduction:** automation process and computer aided.

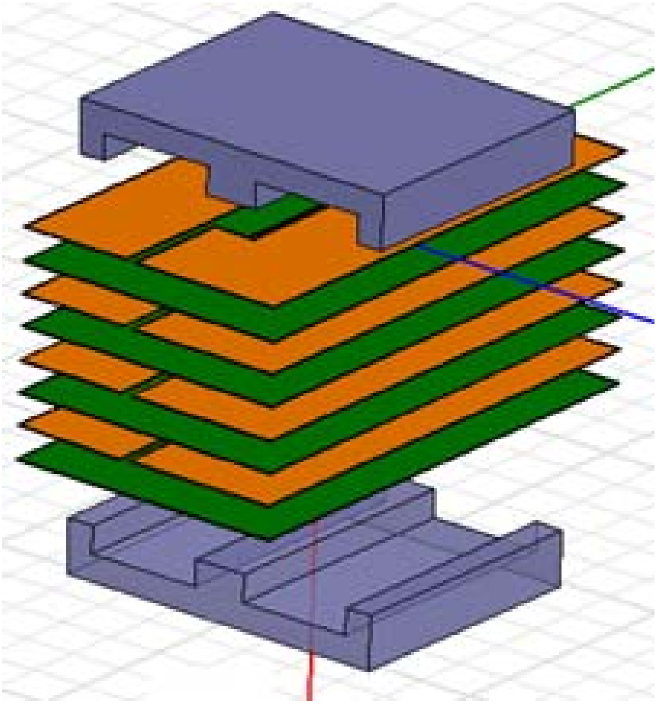
- **Modularity:** no extra connections are required

Photos are from Google

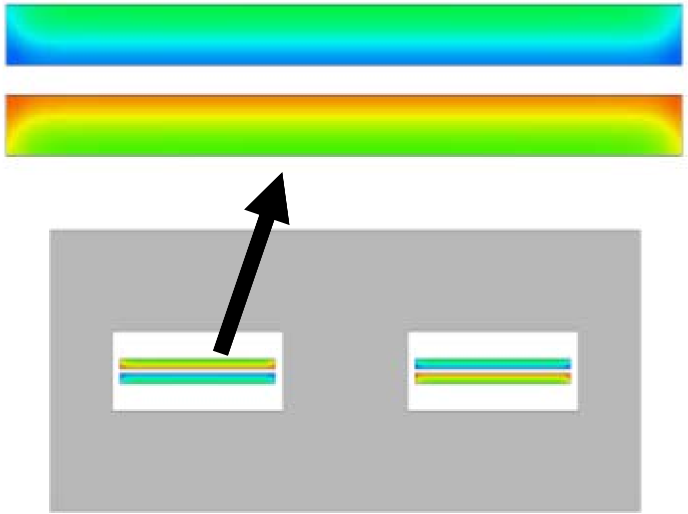
Advantages

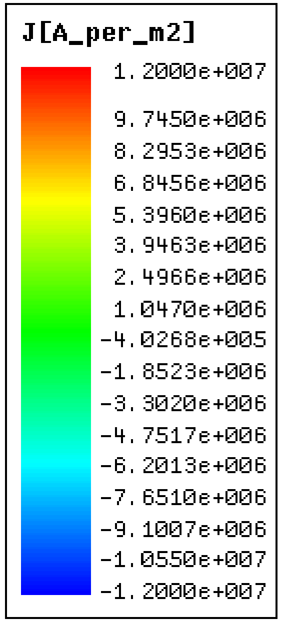
- **Predictable parasitics:** windings manufactured by PCB machines are more precise and consistent, resulting in magnetic designs with highly controllable and predictable parasitic parameters

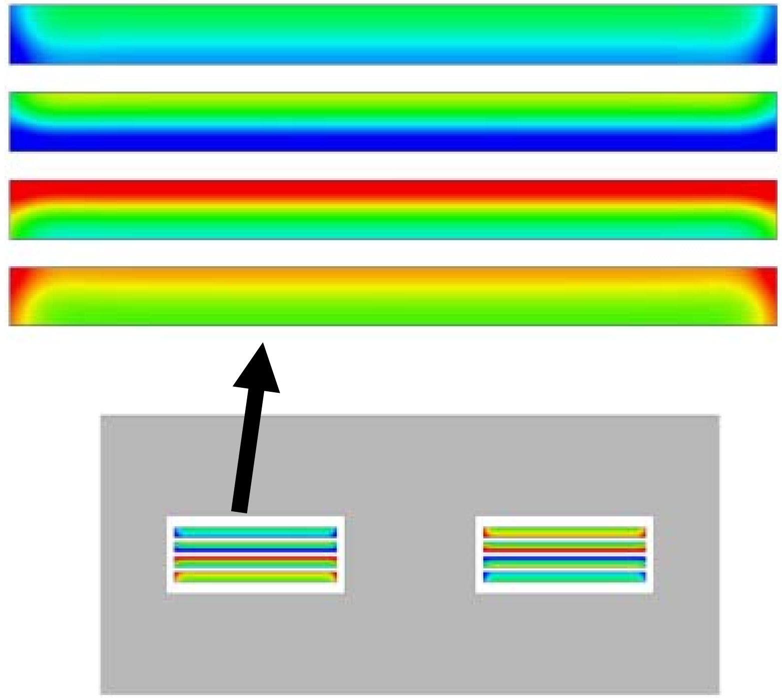
- **Ease of implementation on interleaved windings:** multi-layer PCBs allow for a interconnection between arbitrary layers.

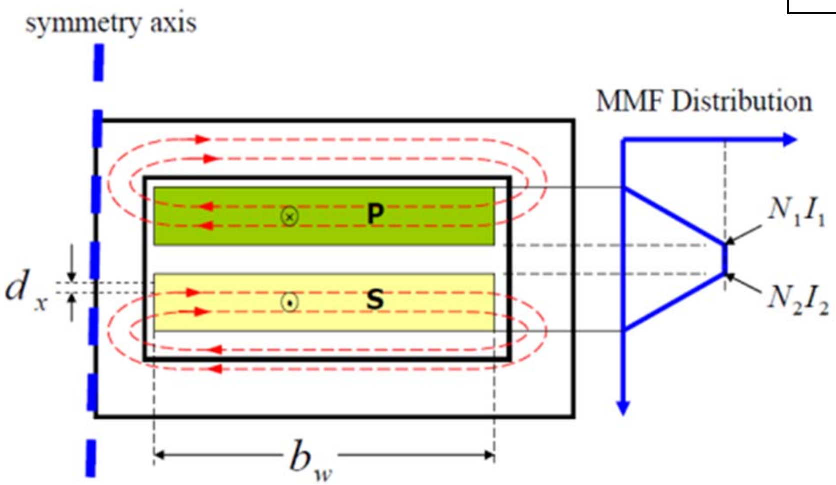


Eddy Current Effect









In typical planar transformers, most of external flux (leakage flux) is parallel to the surface of the conductors

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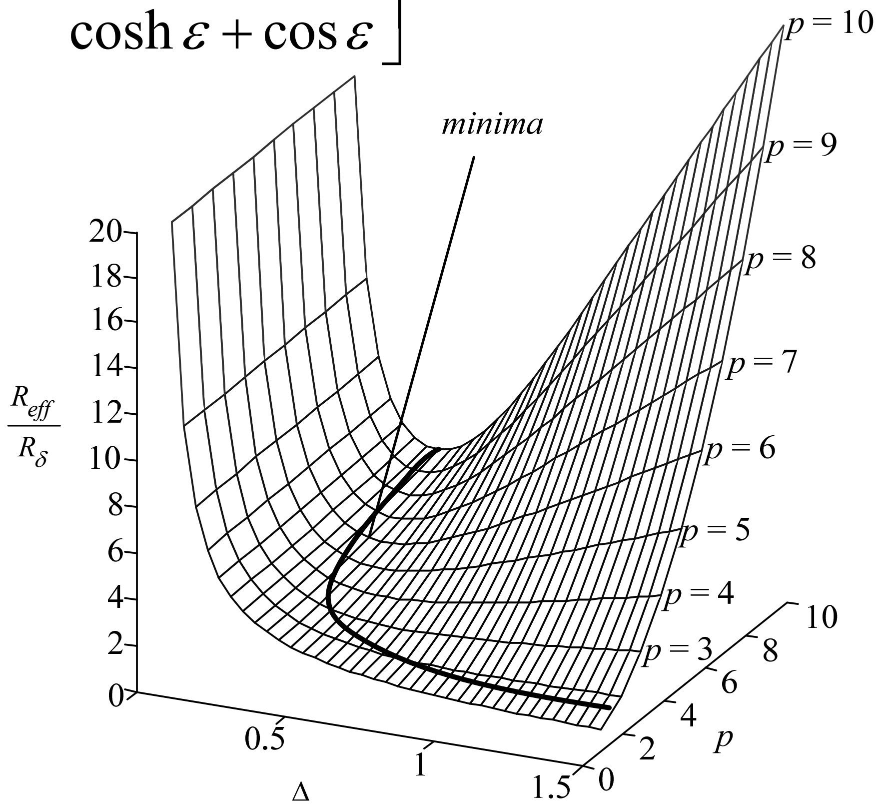
High Frequency Resistance

Modelling ac resistance in planar transformer is the same with traditional one-dimensional Dowell's analysis:

$$\frac{R_{ac}}{R_{dc}} = \varepsilon\left[\frac{\sinh 2\varepsilon + \sin 2\varepsilon}{\cosh 2\varepsilon - \cos 2\varepsilon} + \frac{2(p^2-1)}{3}\frac{\sinh\varepsilon - \sin\varepsilon}{\cosh\varepsilon + \cos\varepsilon}\right]$$

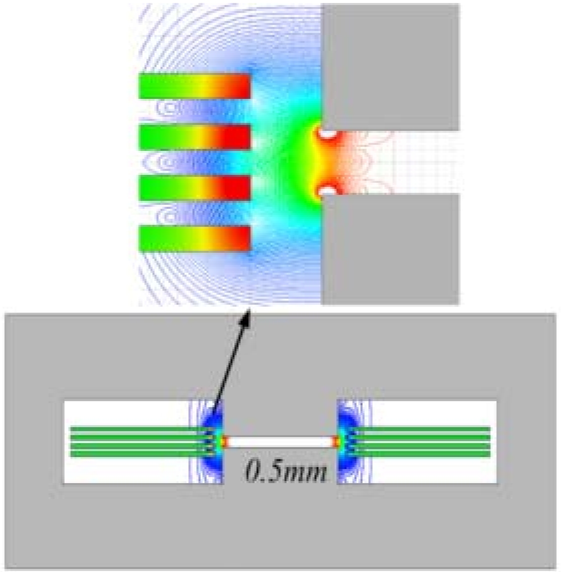
where ε is the ratio of the conductor thickness to the skin depth. p is the number of layers

Note: ***"radial current distribution"*** in planar structures may affect the dc resistance, but no effect on the ratio of ac resistance to dc resistance.

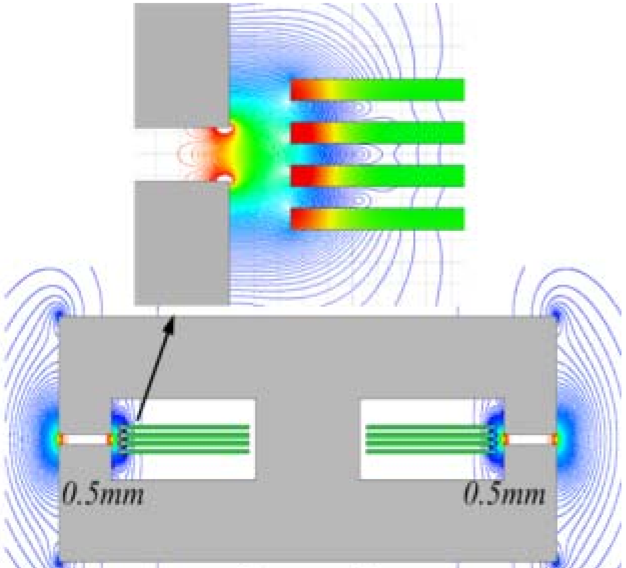


[Ref.]: W. G. Hurley, E. Gath, and J. G. Breslin, "Optimizing the AC resistance of multilayer transformer windings with arbitrary current waveforms," IEEE Trans. Power Electron., vol. 15, no. 2, pp. 369–376, Mar. 2000.

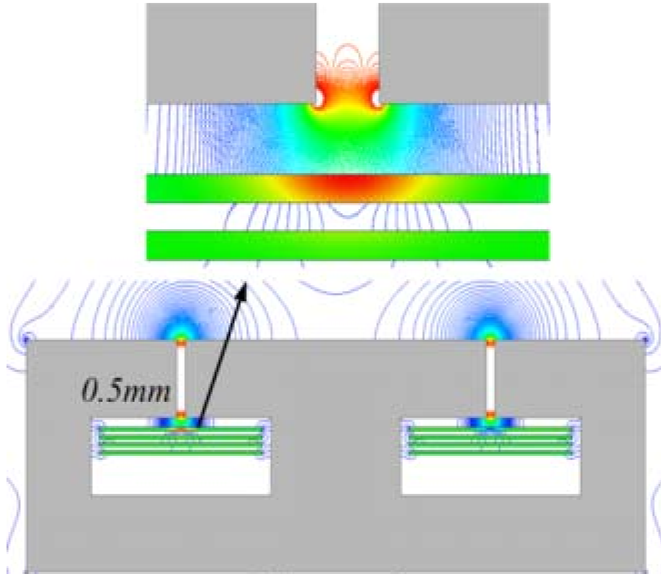
Fringing Effect



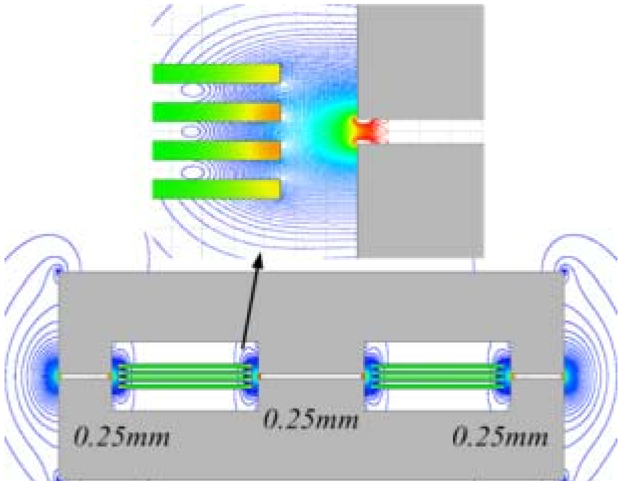
(a) case-1
Center gap



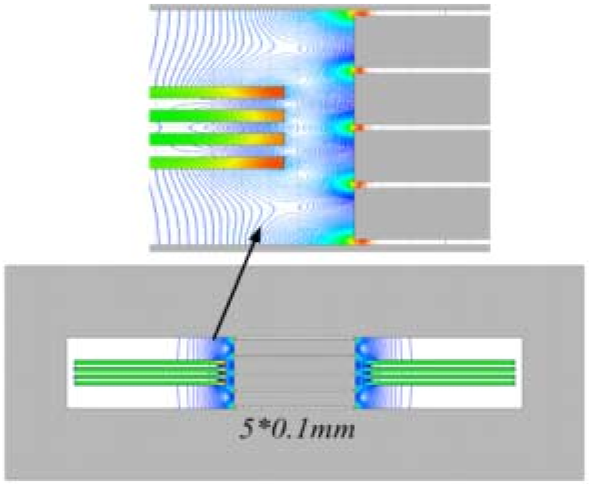
(b) case-2
Side gap



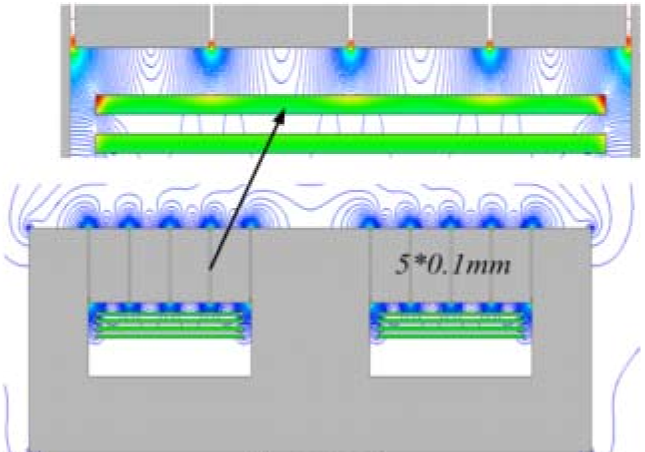
(c) case-3
Horizontal gap



(d) case-4
Equalized gap

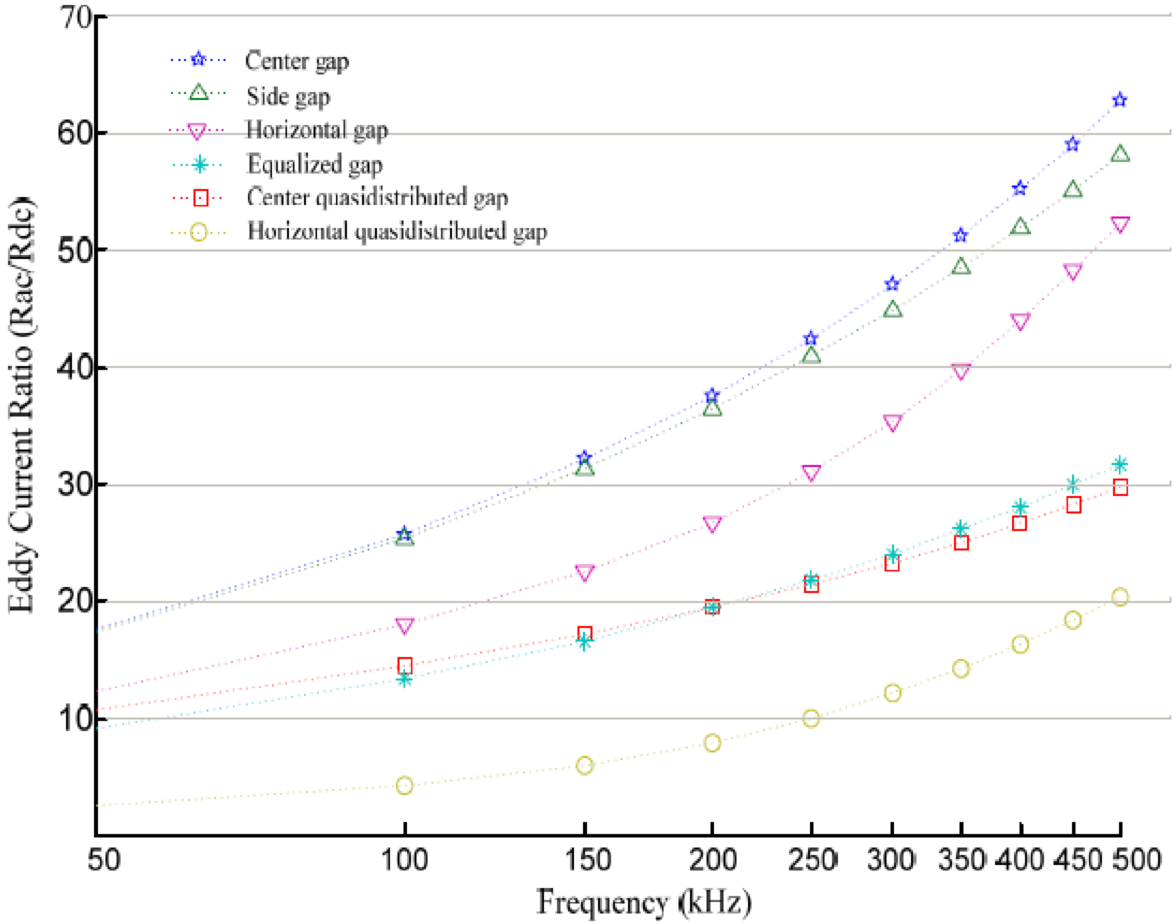


(e) case-5
Center quasidistributed gap



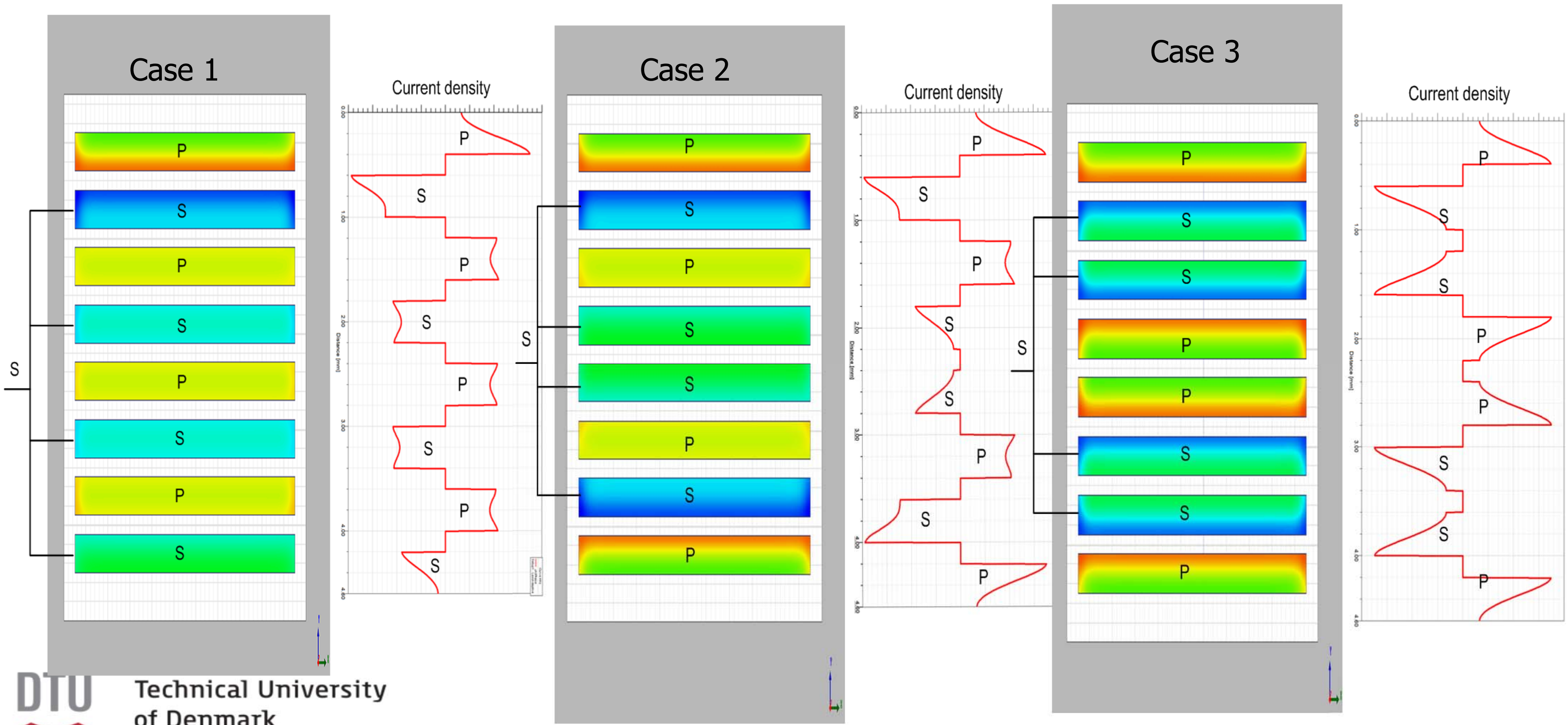
(f) case-6
Horizontal quasidistributed gap

Fringing Effect

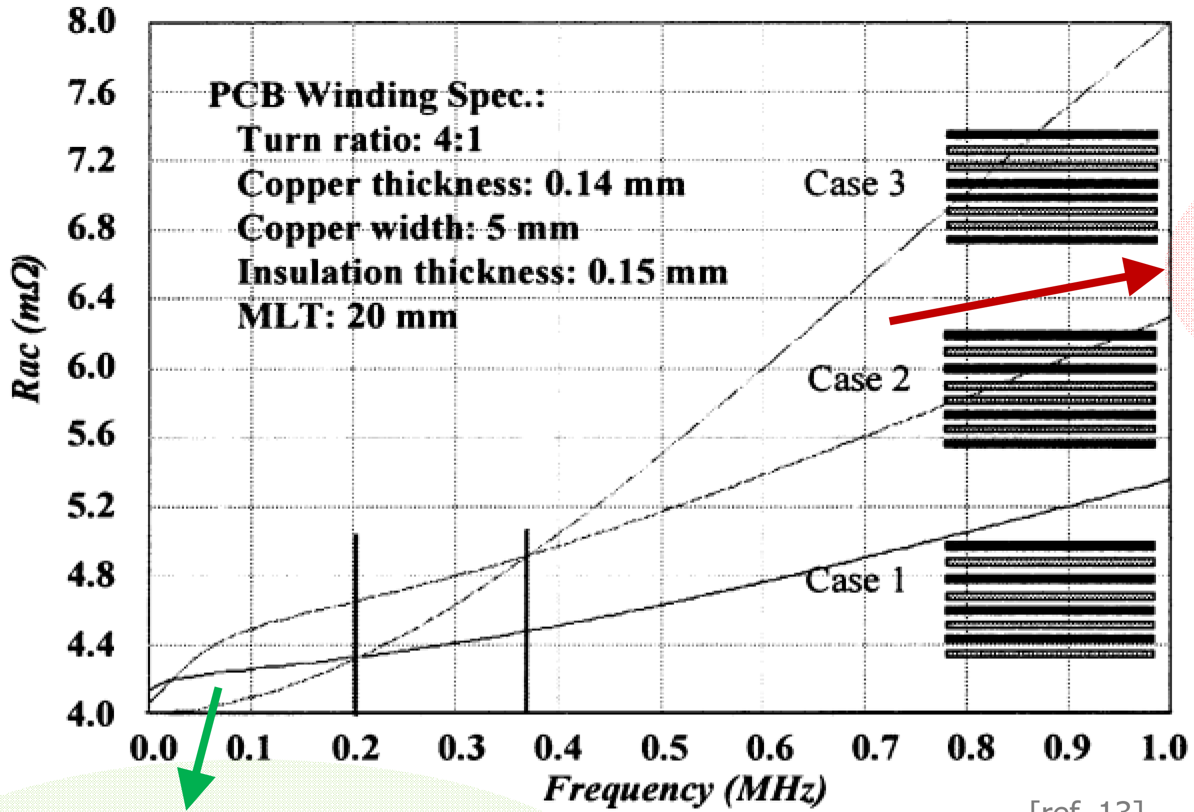


	Fringing effect on winding loss	Fringing effect on external magnetic field
Case 1: **Center gap**	Highest	Low
Case 2: **Side gap**	High	High
Case 3: **Horizontal gap**	Medium	High
Case 4: **Equalized gap**	Low	Medium
Case 5: **Center quasi-distributed gap**	Low	Low
Case 6: **Horizontal quasi-distributed gap**	Lowest	Medium

- High current applications
- Currents may not be equally distributed in the paralleled winding layers



Parallel Windings



[ref. 13]

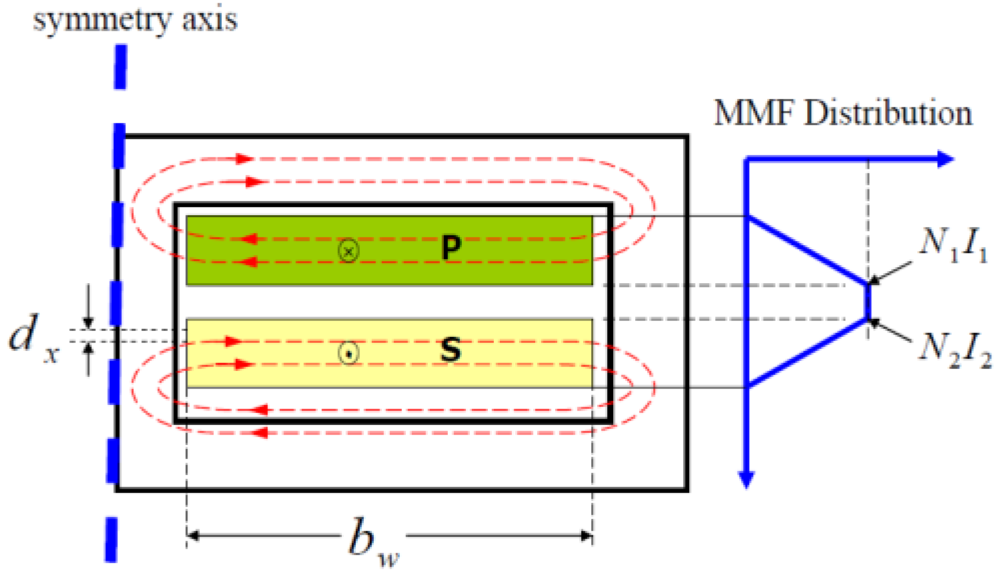
At high frequency, eddy current effect losses dominate.

At low frequency, “parallel effect losses” or ”circulating currents losses” dominates

[Ref.]: W. Chen, Y.-P. Yan, Y.-Q. Hu, and Q. Lu, “Model and design of PCB parallel winding for planar transformer,” IEEE Trans. Magn., vol. 39, no. 5, pp. 3202–3204, Sep. 2003.

Leakage Inductance

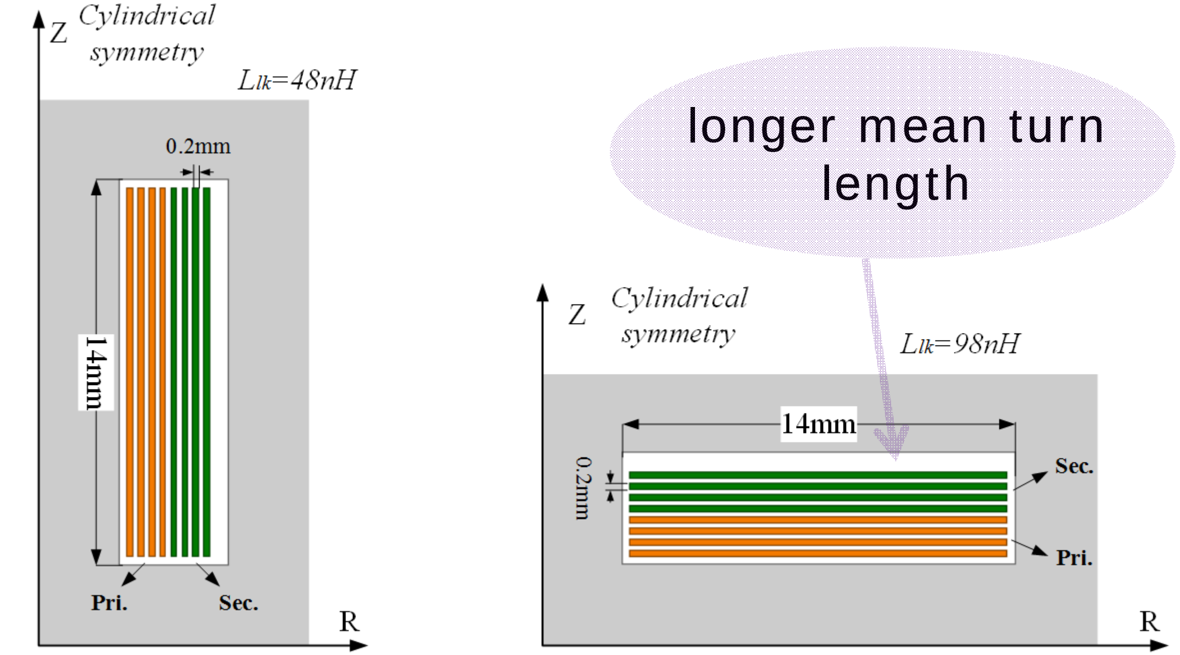
Leakage inductance is simply dependent on the energy stored in core window area:



Leakage energy stored in each elementary layer:

$$E_{energy} = \frac{\mu_0}{2} \sum \int_0^h H^2 \cdot l_w \cdot b_w \cdot dx$$

Leakage Inductance



(a) conventional core geometry with vertical-winding

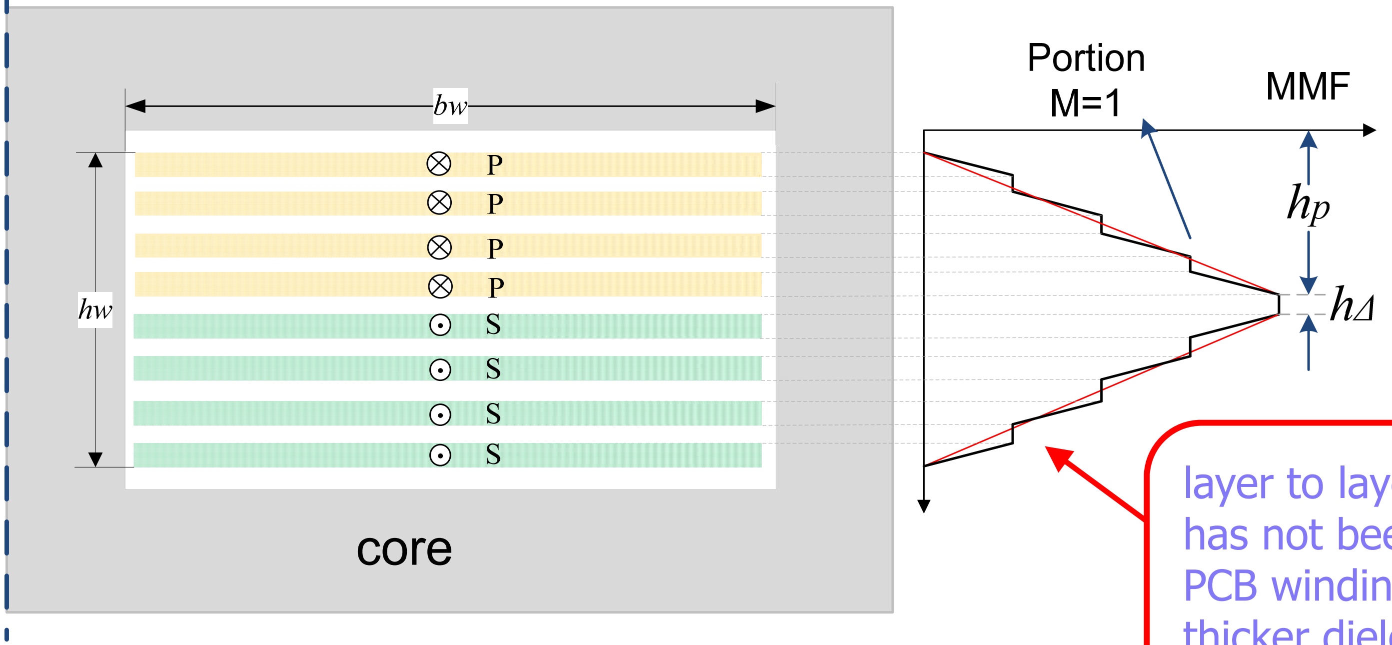
(b) planar core geomtry with horizontal-winding

- Planar structure is not intrinsically a low leakage inductance construction.
- The benefit of planar PCB transformers in this regard is the relative ease with which primary and secondary windings can be heavily interleaved

[Ref.]: Z. Ouyang and M. A. E. Andersen, "Overview of planar magnetic technology—fundamental properties," IEEE Trans. on Power Electronics, vol.29, no.9, pp.4888-4900, Sept., 2014.

Leakage Inductance

sysmetric



layer to layer insulation thickness has not been considered, which in PCB windings (usually have a thicker dielectric layer) may make a significant error

Traditional analytical expression:

$$L_{lk} = \mu_0 \frac{N^2 l_w}{M^2 b_w} \left(\frac{1}{3} \sum_{P=1}^{2M} h_P + \sum_{\Delta=1}^{M} h_\Delta \right)$$

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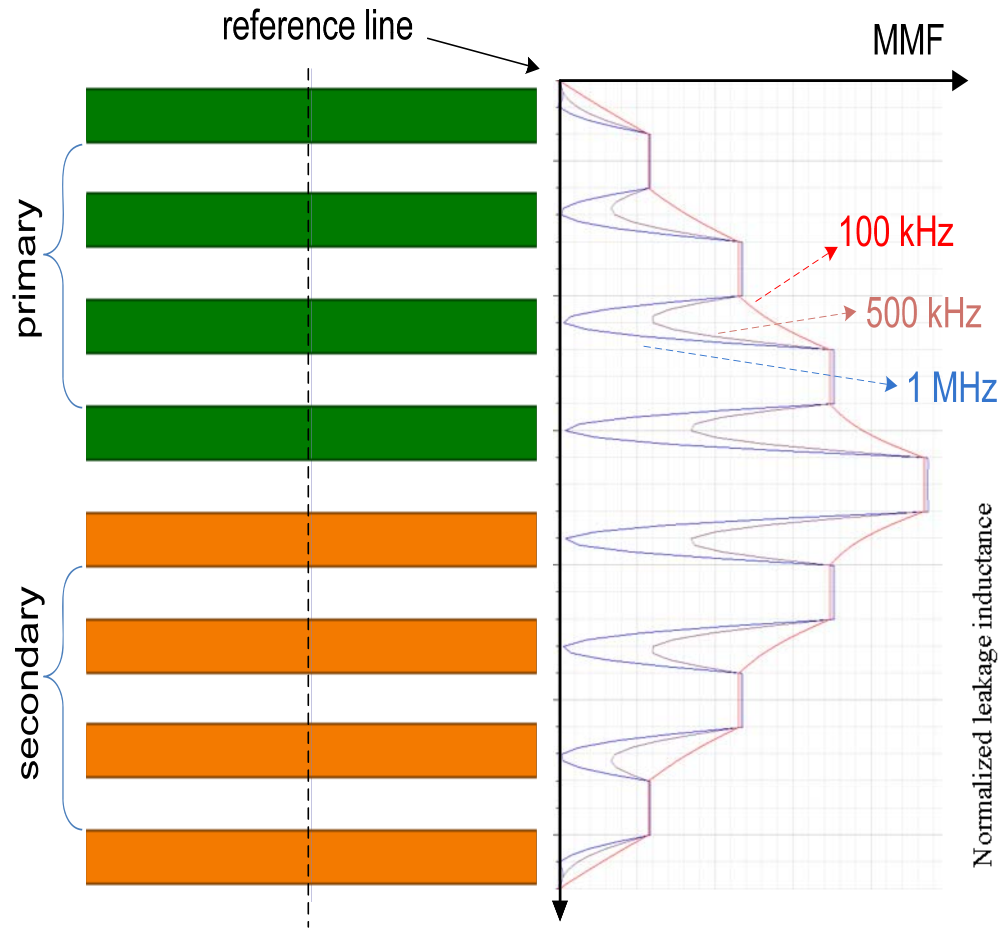
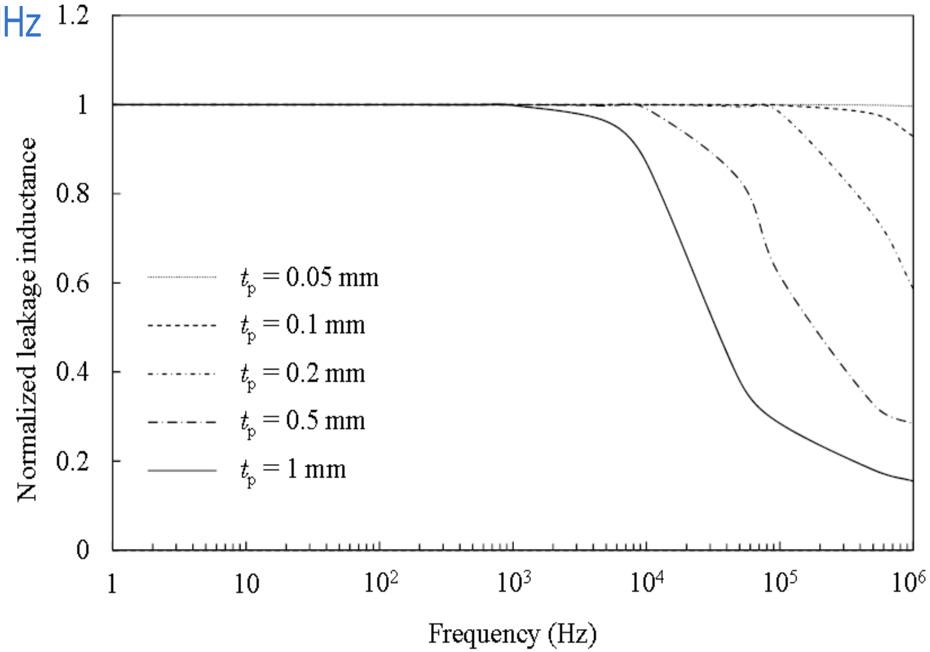
[Ref.1]: P. L. Dowell, "Effects of eddy currents in transformer windings," Proc. Inst. Elect. Eng., vol. 113, no. 8, pp. 1387–1394, Aug. 1966.

[Ref.2]: E. C. Snelling, Soft Ferrites—Properties and Applications, 2nd ed. London, U.K.: Butterworth, 1988.

Leakage Inductance



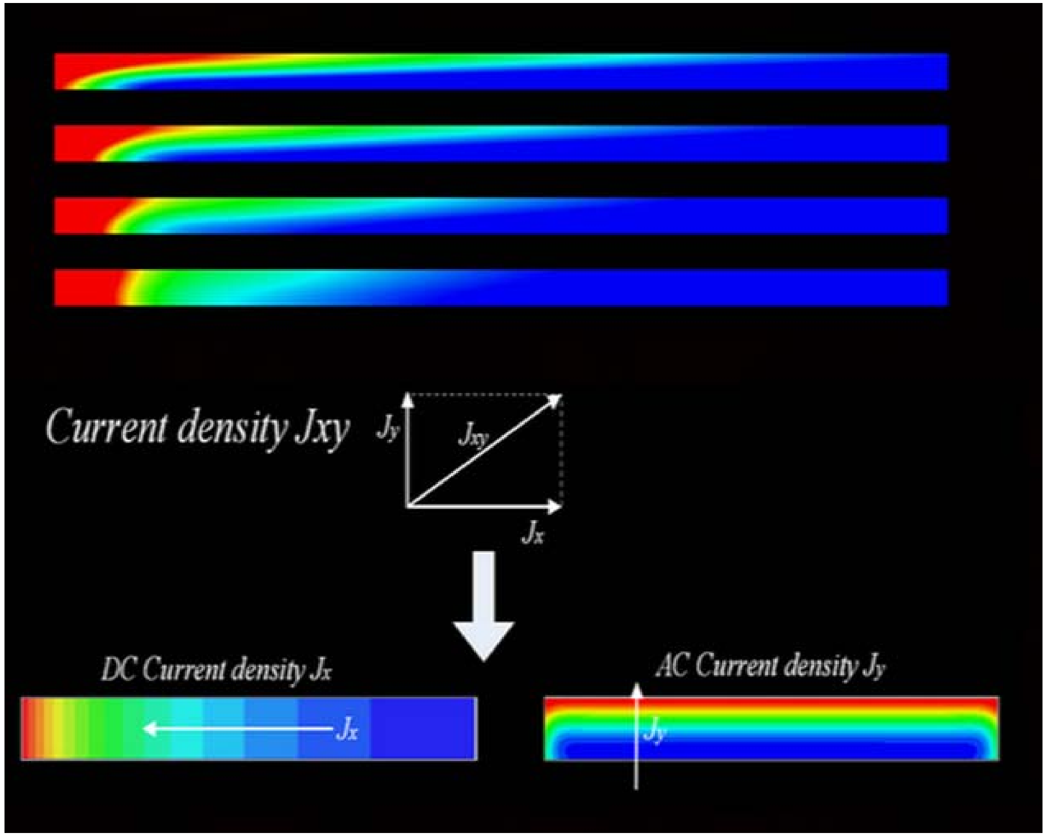
Leakage Inductance is frequency dependent

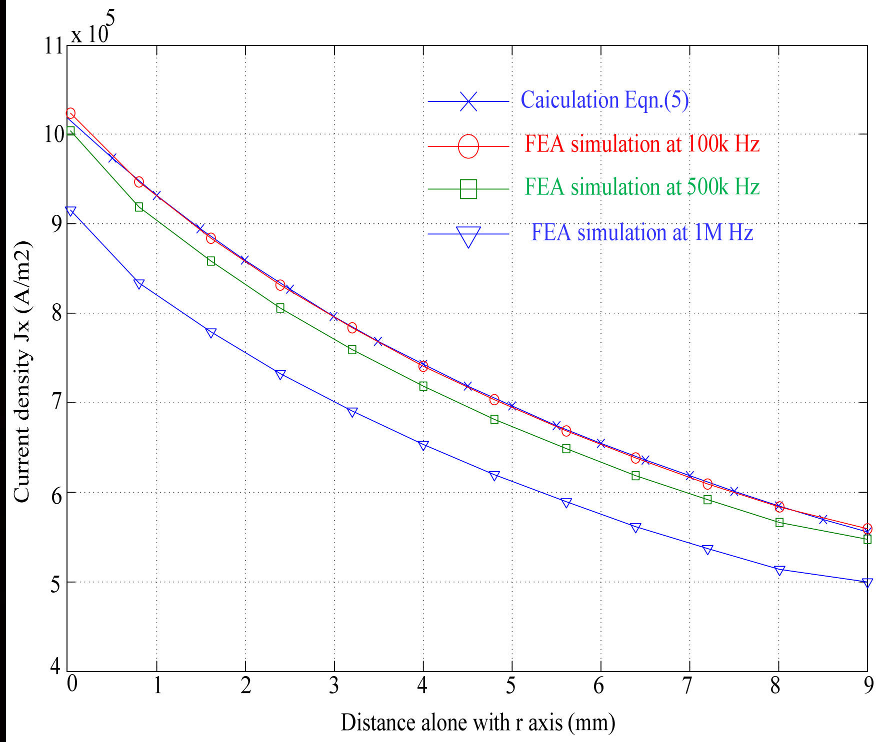


[Ref.1]: Z. Ouyang, J. Zhang, and W. G. Hurley, "Calculation of leakage inductance for high frequency transformers," IEEE Trans. on Power Electronics, vol.30, no.10, pp.5769-5775, Oct. 2015
[Ref.2]: J. Zhang, "Analysis and design of high frequency gapped transformers and planar transformers in LLC resonant converters", PhD thesis, National University of Ireland, 2015

Leakage Inductance

- "**radial current distribution**" due to high aspect ratio of width to height of a section, b_w/h_w.





New model for planar transformer leakage inductance:

- The energy stored in the primary/secondary winding is:

$$E_p = \sum_{i=1}^{n_p} E_i = \frac{\mu_0 \cdot \pi \cdot I_p{}^2 \cdot n_p \left[k_1\left(2n_p{}^2 + 1\right) + 4k_2(n_p{}^2 - 1)\right]}{In(\frac{r_2}{r_1}) \cdot 12 \cdot \gamma \sinh^2(\gamma h_p)}$$

where,

$$k_1 = \sinh(2\gamma h_p) - 2\gamma h_p$$
$$k_2 = \gamma h_p \cosh\left(\gamma \cdot h_p\right) - \sinh\left(\gamma \cdot h_p\right)$$

- The energy stored in the dielectric layer is:

$$E_d = \frac{1}{2} \cdot \mu_0 \cdot h_i \cdot \int_{r_1}^{r_2} H(r,h)^2 \cdot 2\pi r \cdot dr = \frac{\mu_0 \cdot \pi \cdot h_i}{In(\frac{r_2}{r_1})}\left[I_p{}^2 \sum_{i=1}^{n_p} i^2 + I_s{}^2 \sum_{i=1}^{n_s - 1} i^2\right]$$
$$= \frac{\mu_0 \cdot \pi \cdot h_i}{6In(\frac{r_2}{r_1})}\left[I_p{}^2 \cdot n_p\left(n_p + 1\right)\left(2n_p + 1\right) + I_s{}^2 \cdot n_s(n_s - 1)(2n_s - 1)\right]$$

Leakage Inductance

Giving an example that has only one turn in each layer, the turns ratio $n = \frac{n_s}{n_p}$ is defined, and all windings' thickness are the same ($h_p = h_s$), then the total leakage inductance is:

$$L_{lk} = \frac{\mu_0 \cdot \pi \cdot n_p}{3In(\frac{r_2}{r_1})} \left\{ \frac{{n_p}^2(k_1 + 2k_2)(n+1)}{\gamma\sinh^2(\gamma h_p)} + \frac{(k_1 - 4k_2)(n+1)}{2n\gamma\sinh^2(\gamma h_p)} + \left[2(1+n) \cdot {n_p}^2 + \frac{1}{n} + 1 \right] h_i \right\}$$

not applicable to complex interleaved cases such as primary and secondary windings on the same layer where 2-D consideration may be needed.

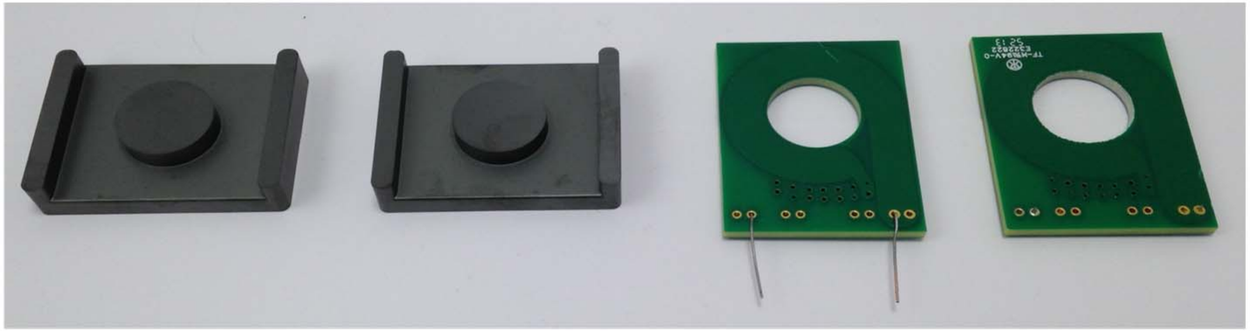
Leakage Inductance

identical

3C96 ER51 cores

Primary winding (8 layers)

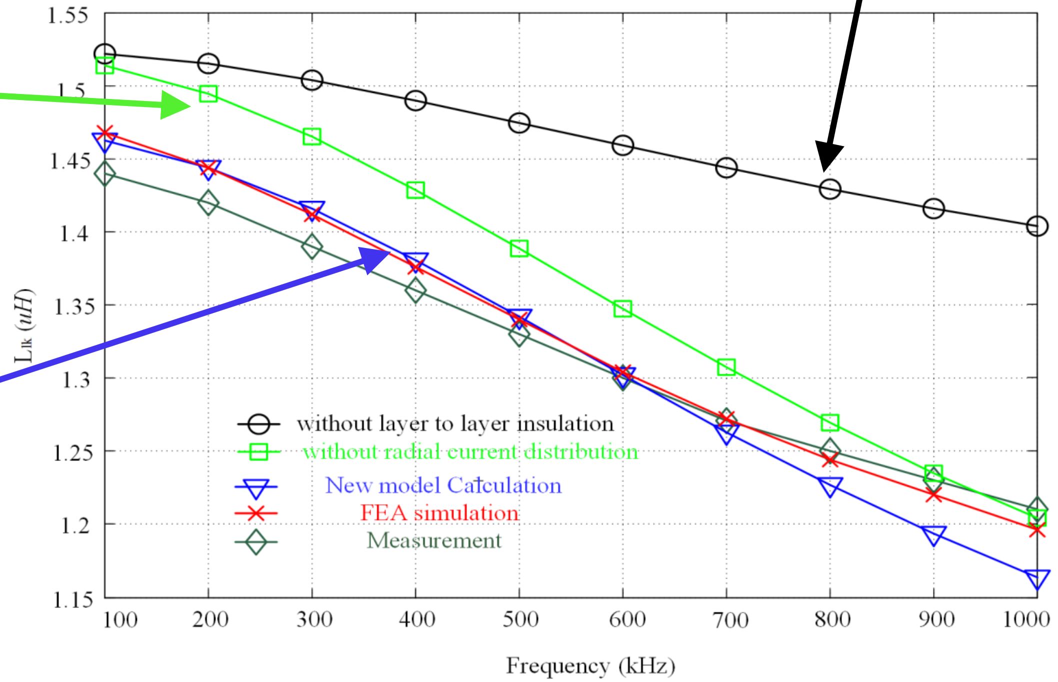
Secondary winding (8 layers)



without the consideration of layer to layer insulation

without the consideration of radial current distribution

New model calculation



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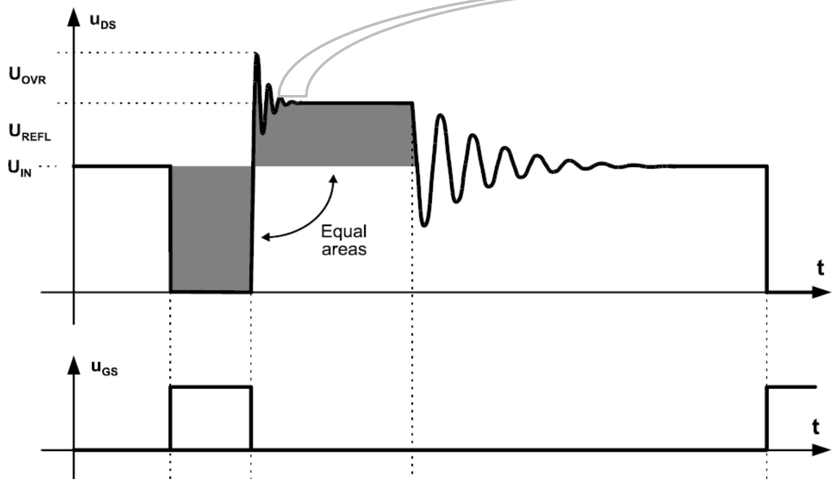
Leakage Inductance



Small Leakage Inductance is expected in most of power converters

- ❖ **Reduce the number of turns.** (core saturation and higher core loss)
- ❖ **Reduce the thickness of conductors and insulators.** (high winding resistance and high interwinding capacitance)
- ❖ **Reduce the mean turn length.**
- ❖ **Increase the window width.** (high interwinding capacitance)
- ❖ **Interleaving winding arrangement.**

Note : "trade-off"

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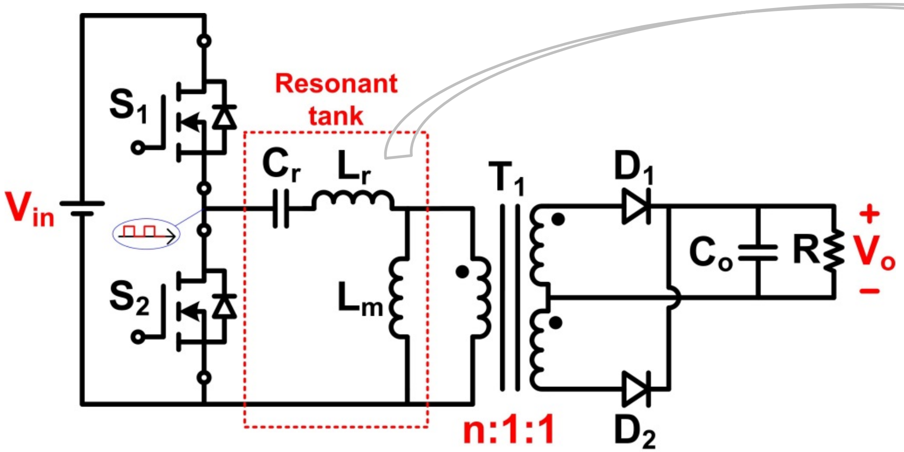
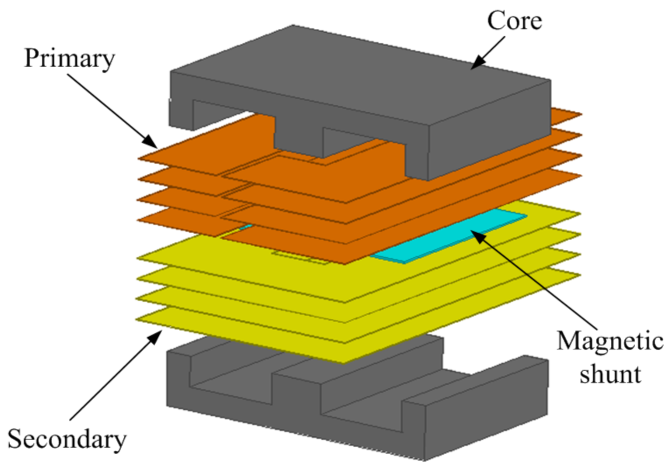
[Ref.]: Z. Ouyang, O. C. Thomsen and M. A. E. Andersen, "Optimal design and tradeoff analysis of planar transformer in high-power dc–dc converters," IEEE Trans. on Ind. Electronics, vol.59, no.7, pp.2800-2810, July, 2012.

Leakage Inductance



Higher Leakage Inductance is expected in resonant converters such as LLC, DAB etc.

- Insertion of **magnetic shunt** (f.x. ferrite polymer composites FPC)



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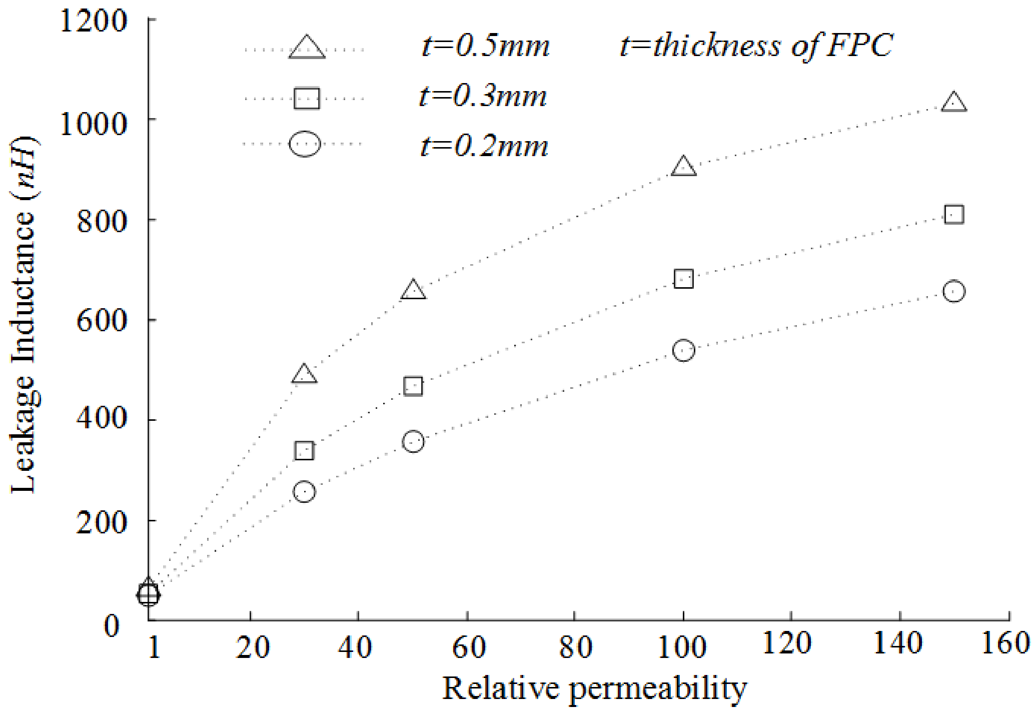
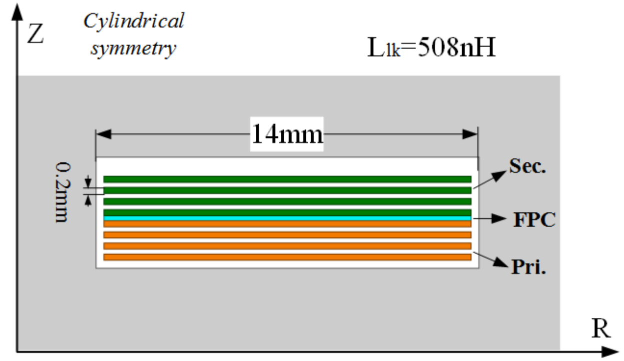
[Ref.]: J. Zhang, Z. Ouyang, M. C. Duffy, M. A. E. Andersen and W. G. Hurley, "Leakage inductance calculation for planar transformers with a magnetic shunt," IEEE Transactions on Industry Applications, vol.50, no.6, pp.4107-4112, Nov.-Dec. 2014

Leakage Inductance

- Thickness of magnetic sheet
- Permeability of magnetic sheet





planar transformer with FPC material in between primary and secondary

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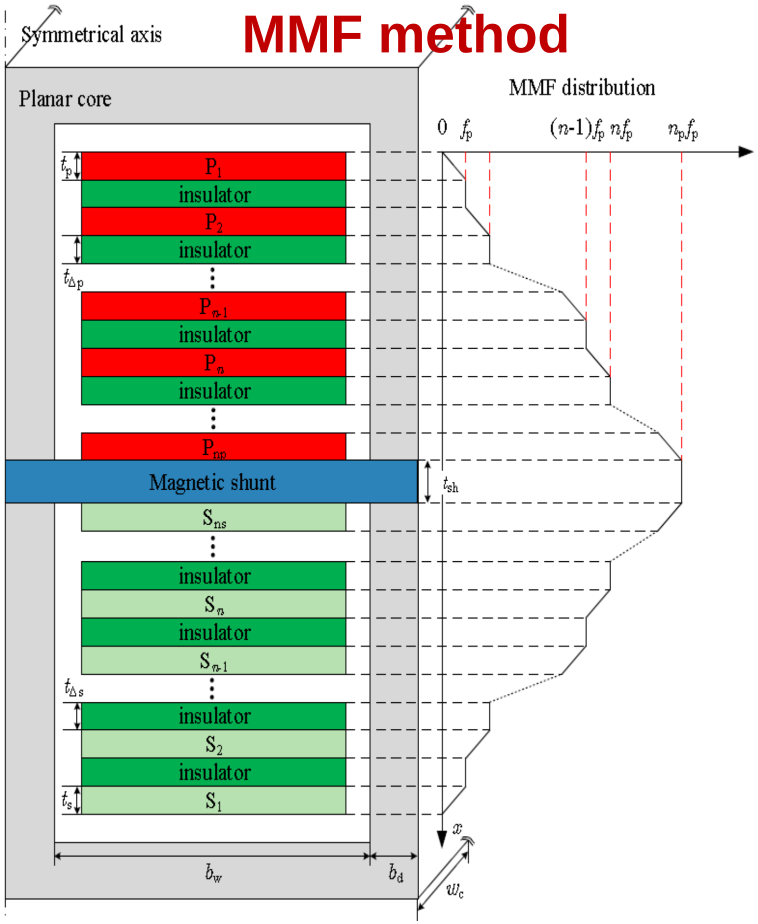
[Ref.]: Z. Ouyang, "Advances in planar and integrated magnetics," Ph.D. dissertation, Dept. Electr. Eng., Tech. Univ. Denmark, Kgs Lyngby, Denmark, 2011.

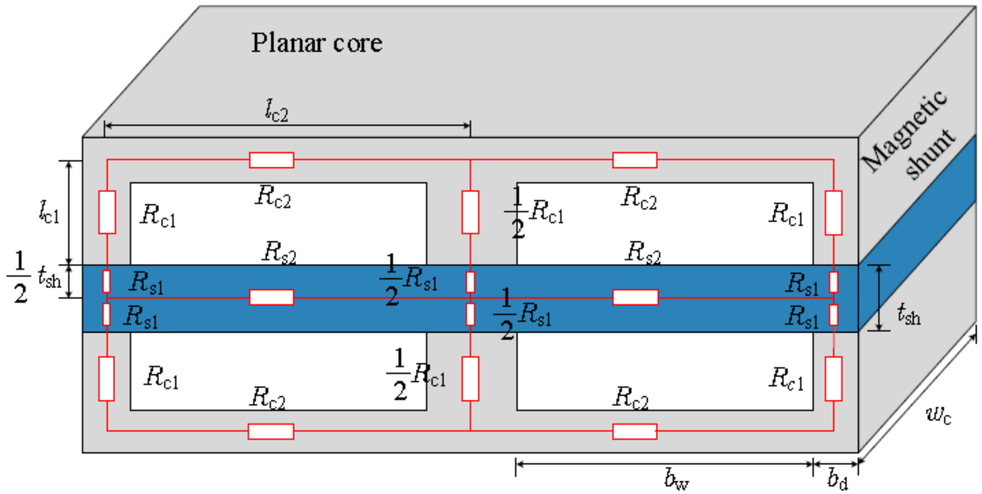
APEC

Leakage Inductance

MMF method



Reluctance method



$$L_{lk}=\frac{1}{3}\mu_0 n_{\mathrm{Lp}}^2 n_{\mathrm{p}}^2\frac{w_{\mathrm{c}}}{b_{\mathrm{w}}}\left\{\begin{aligned}&\sum_{i=\mathrm{p,s}}\left[2n_i\left(t_i+t_{\Delta i}\right)-3t_{\Delta i}+\frac{t_{\Delta i}}{n_i}\right]+\\&6\mu_s a_{\mathrm{R}}^2\left(t_{\mathrm{sh}}+\frac{2b_{\mathrm{d}}b_{\mathrm{w}}R_{\mathrm{s1}}^2}{t_{\mathrm{sh}}R_{\mathrm{s2}}^2}\right)+\\&\frac{3}{4}\mu_{\mathrm{r}}a_{\mathrm{R}}^2\frac{b_{\mathrm{w}}A_{\mathrm{c}}}{w_{\mathrm{c}}l_{\mathrm{c}}R_{\mathrm{s2}}^2}\left(R_{\mathrm{c1}}+\frac{1}{2}R_{\mathrm{c2}}\right)^2\end{aligned}\right\}.$$

$$L_{lk}=\frac{\mu_0 w_{\mathrm{c}} n_{\mathrm{Lp}}^2 n_{\mathrm{p}}^2}{3b_{\mathrm{w}}}\left\{\sum_{i=\mathrm{p,s}}\left[2n_i\left(t_i+t_{\Delta i}\right)-3t_{\Delta i}+\frac{t_{\Delta i}}{n_i}\right]+6\mu_s t_{\mathrm{sh}}\right\}.$$

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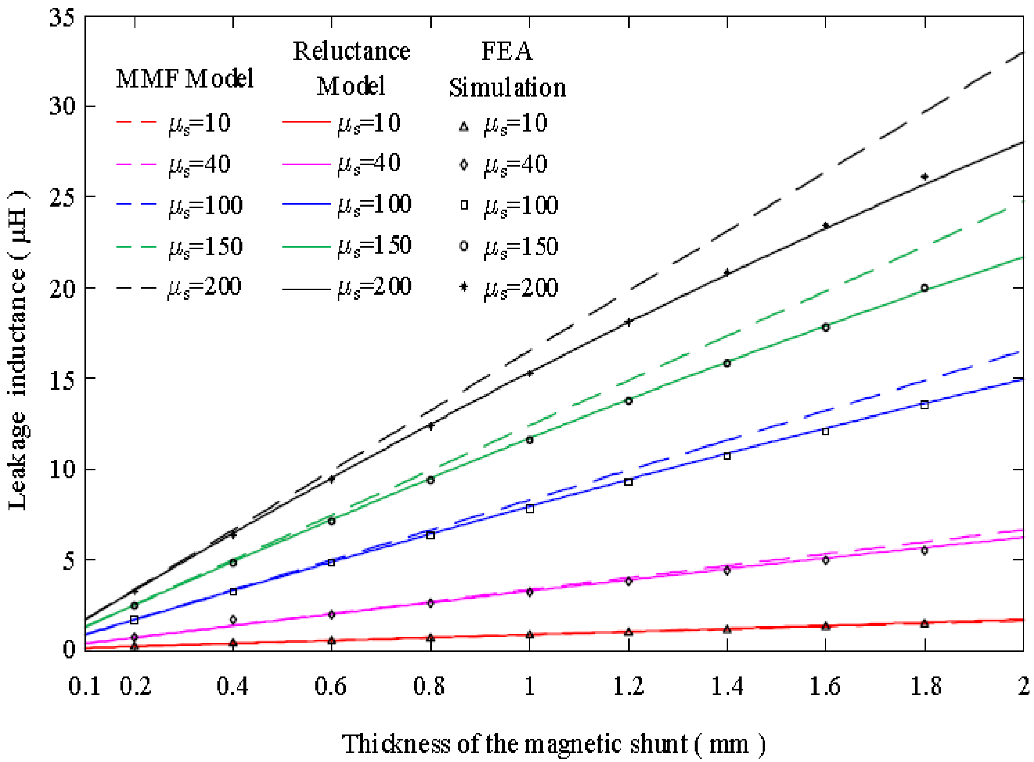
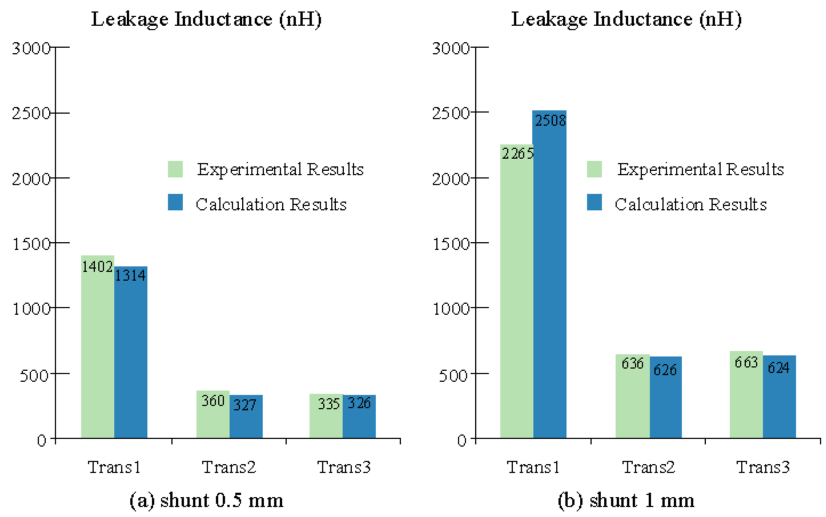
[Ref.]: J. Zhang, Z. Ouyang, M. C. Duffy, M. A. E. Andersen and W. G. Hurley, "Leakage inductance calculation for planar transformers with a magnetic shunt," IEEE Transactions on Industry Applications, vol.50, no.6, pp.4107-4112, Nov.-Dec. 2014

Leakage Inductance

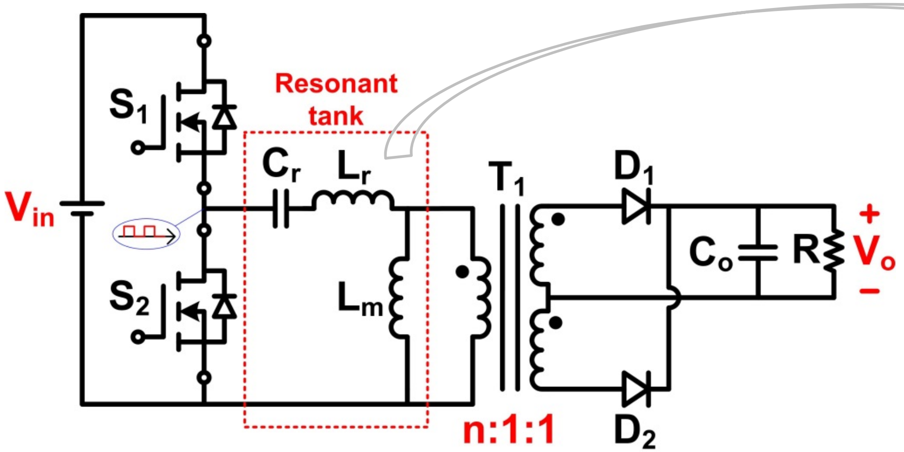




Reluctance method provides a better prediction

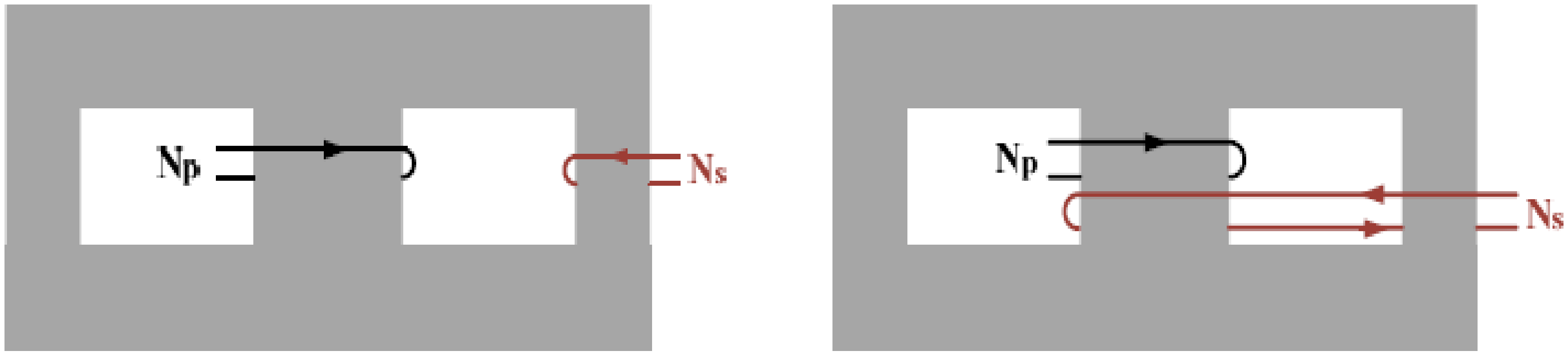
[Ref.]: J. Zhang, Z. Ouyang, M. C. Duffy, M. A. E. Andersen and W. G. Hurley, "Leakage inductance calculation for planar transformers with a magnetic shunt," IEEE Transactions on Industry Applications, vol.50, no.6, pp.4107-4112, Nov.-Dec. 2014

Leakage Inductance



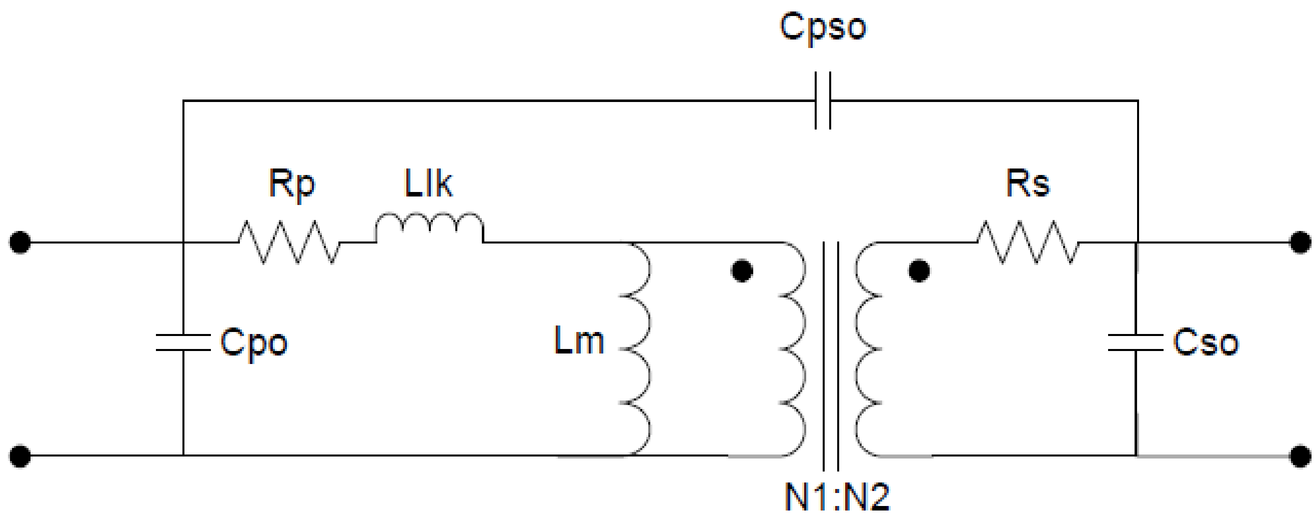
Higher Leakage Inductance is expected in resonant converters such as LLC, DAB etc.

❖ Fractional turn



(a) (b)

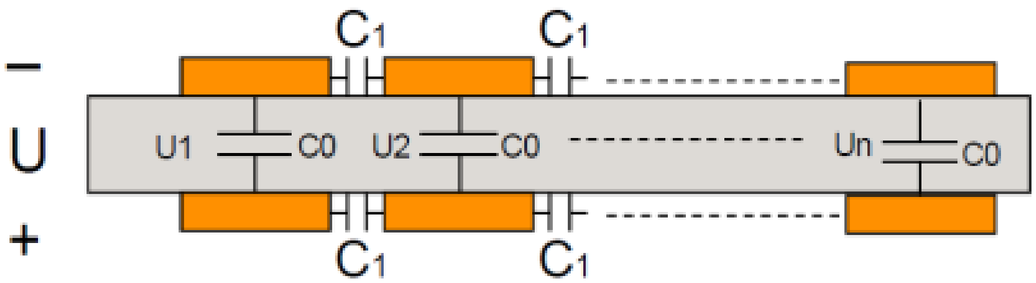
Winding Capacitance



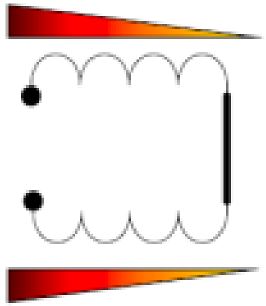
C_{po}, C_{so} are self-capacitances of the primary and the secondary windings, respectively.

C_{pso} *is* the mutual capacitance between the two windings.

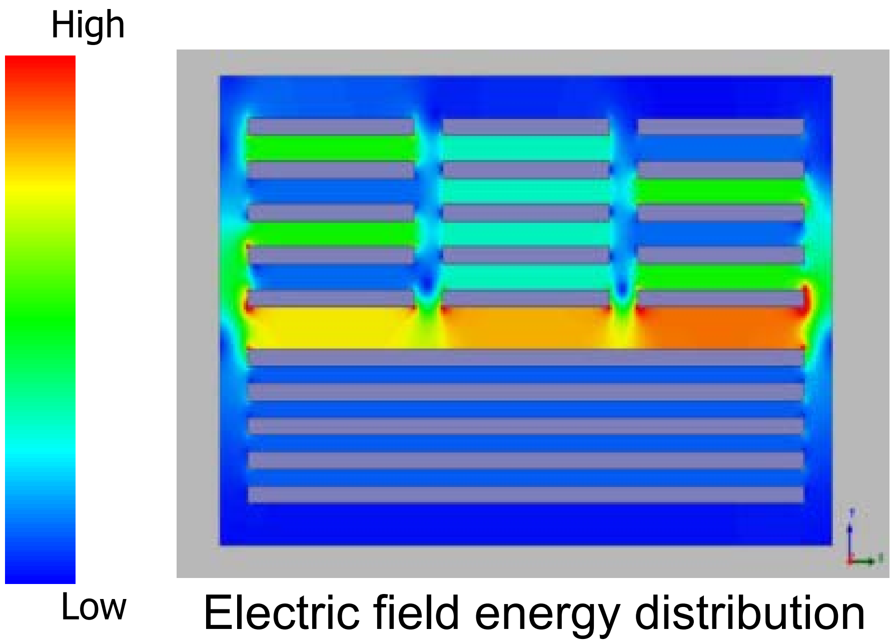
Winding Capacitance



(a)



U-type winding scheme



Electric field energy distribution

$$E = \sum_{i=1}^{n} E_i = \frac{1}{2} C_0 U^2 \sum_{i=1}^{n} \left(\frac{n+1-i}{n} \right)^2$$

$$= \frac{(n+1)(2n+1)}{12n} C_0 U^2.$$

$$C_d = \frac{2E}{U^2} = \frac{(n+1)(2n+1)}{6n} C_0.$$

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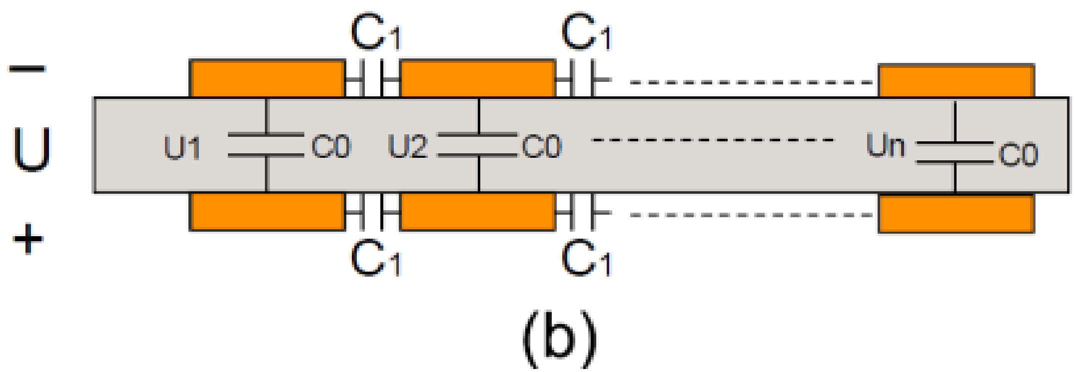
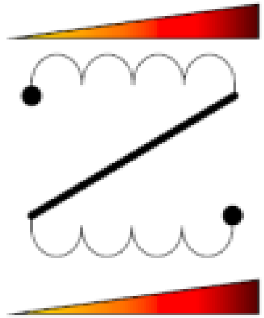
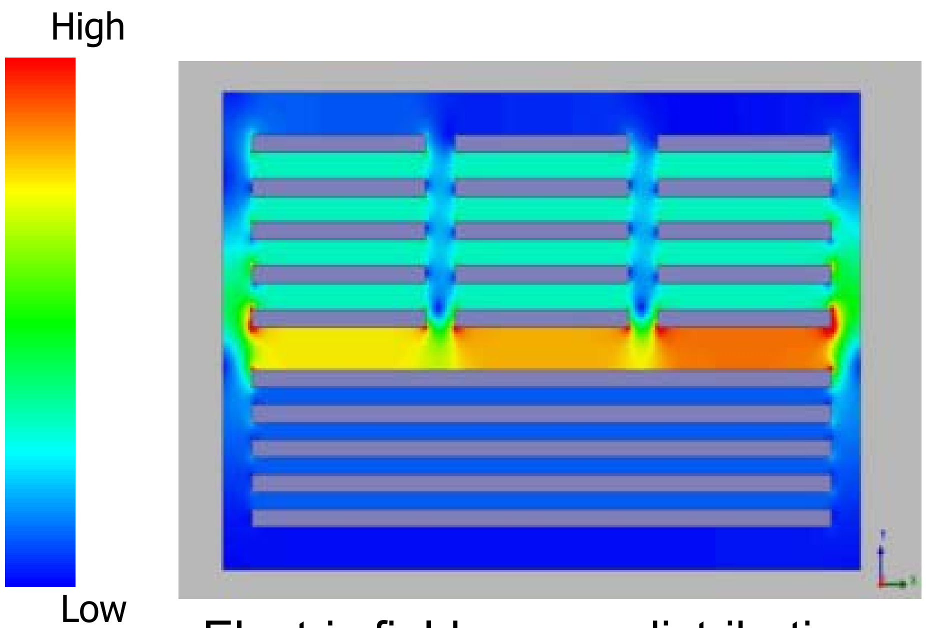
Winding Capacitance





Z-type winding scheme



Electric field energy distribution

$$C_d = (n/4) \cdot C_o$$

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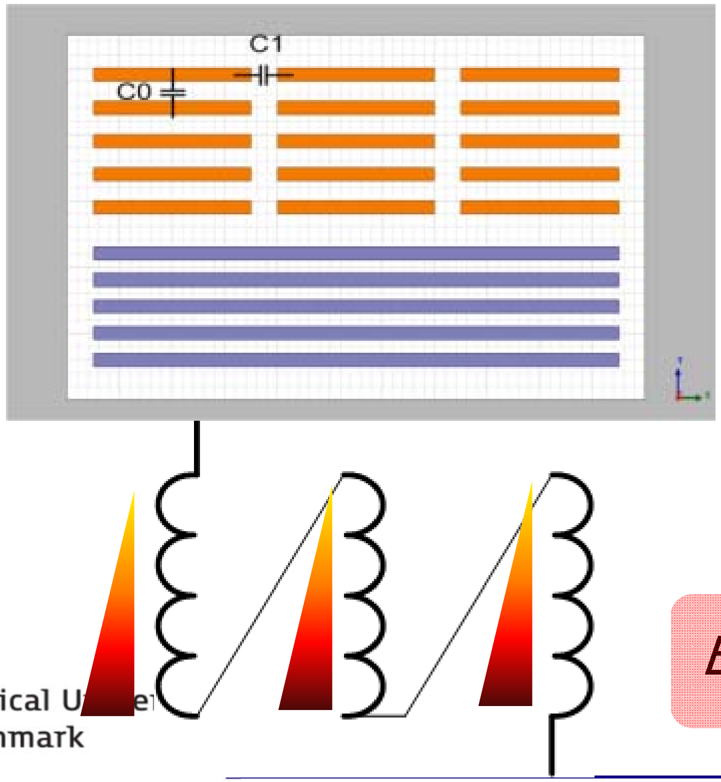
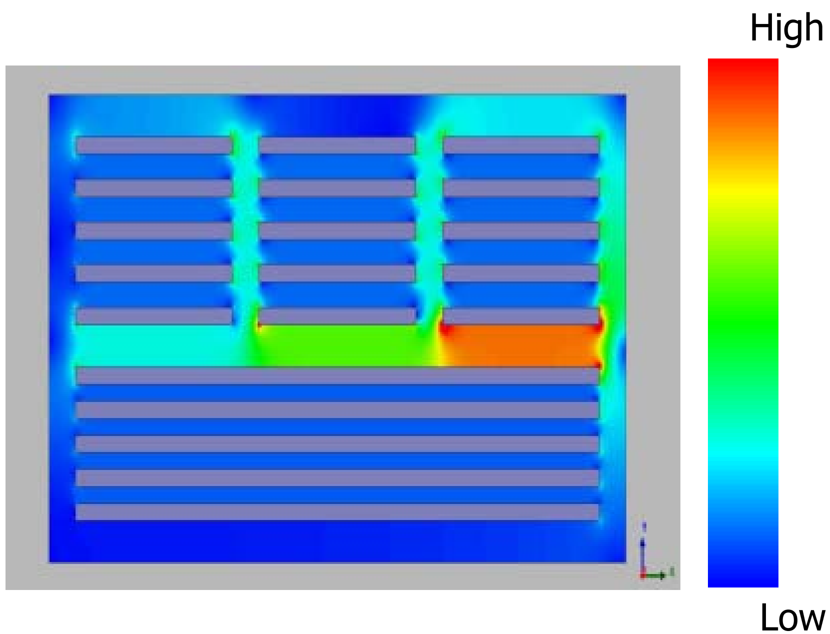
Winding Capacitance

Due to a higher ratio of the width to the thickness of the conductors (intrinsic property of PCB magnetics), C_1 is much lower than C_0. So, vertical winding scheme leads to a lower electric potential energy.





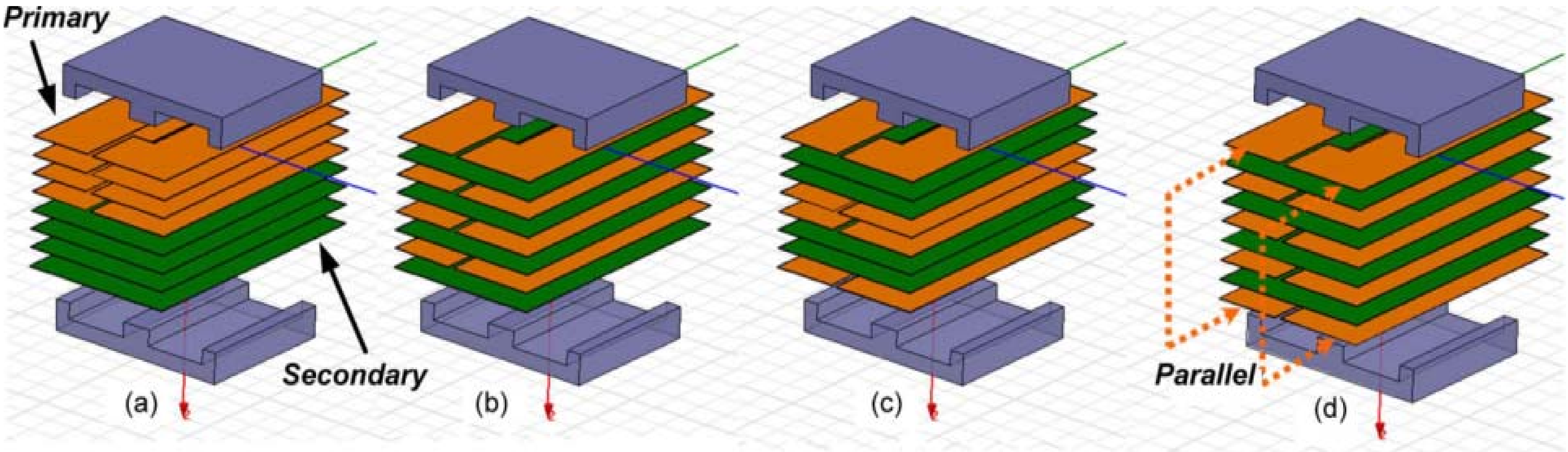
Easier implementation in PCB

Interleaved Winding

- Reduce the ac winding resistance;
  (not for the flyback converter)

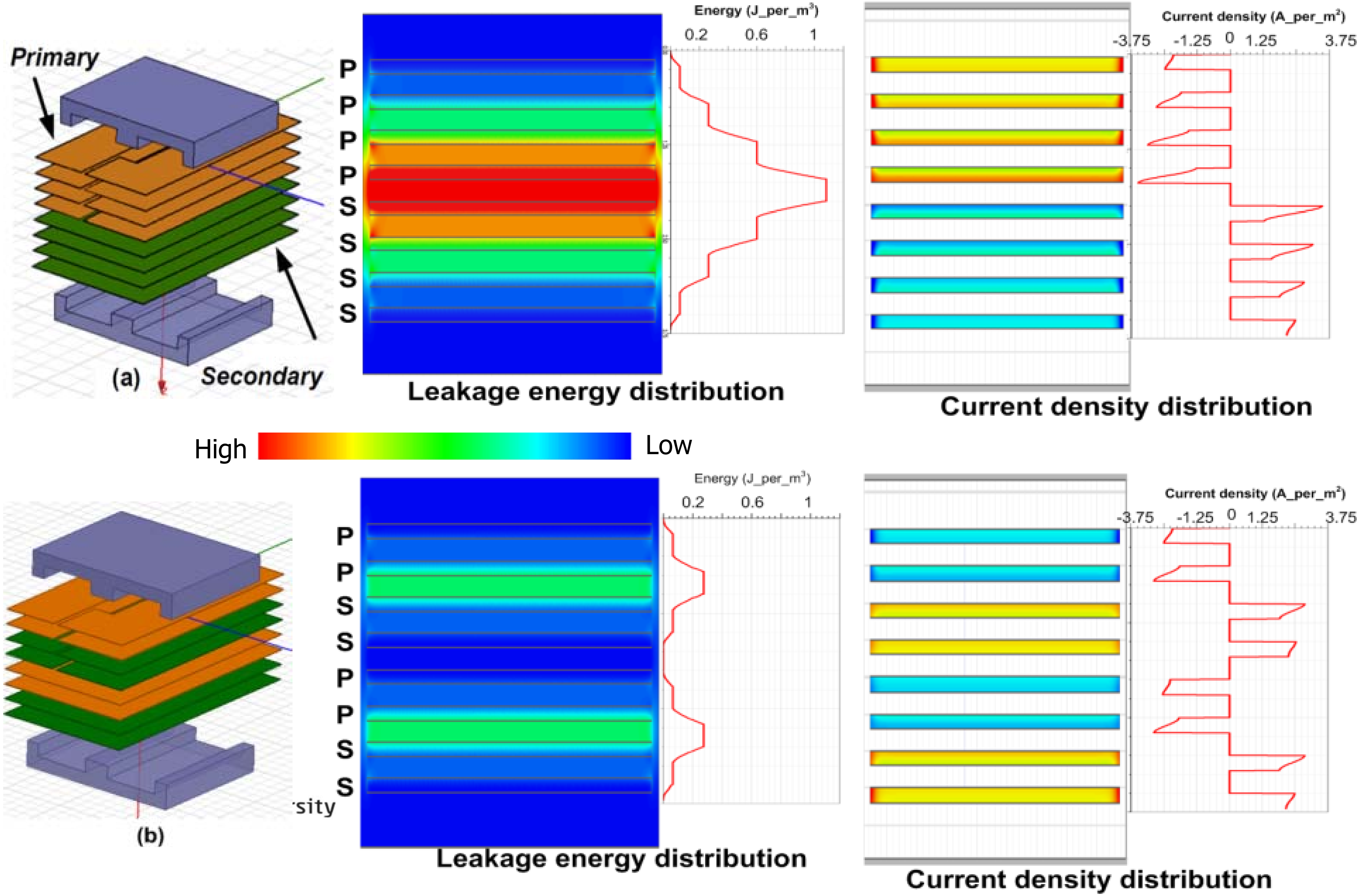
- Reduce the leakage inductance;

- Increase the interwinding capacitance.

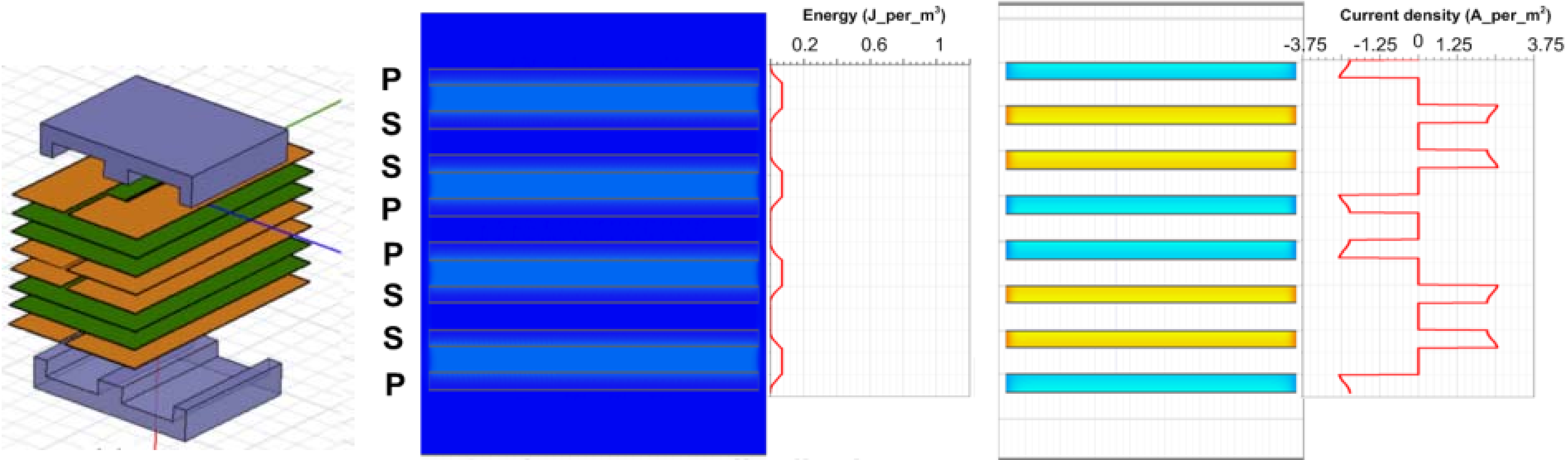


[Ref.]: Z. Ouyang, O. C. Thomsen and M. A. E. Andersen, "Optimal design and tradeoff analysis of planar transformer in high-power dc–dc converters," IEEE Trans. on Ind. Electronics, vol.59, no.7, pp.2800-2810, July, 2012.

Interleaved Winding

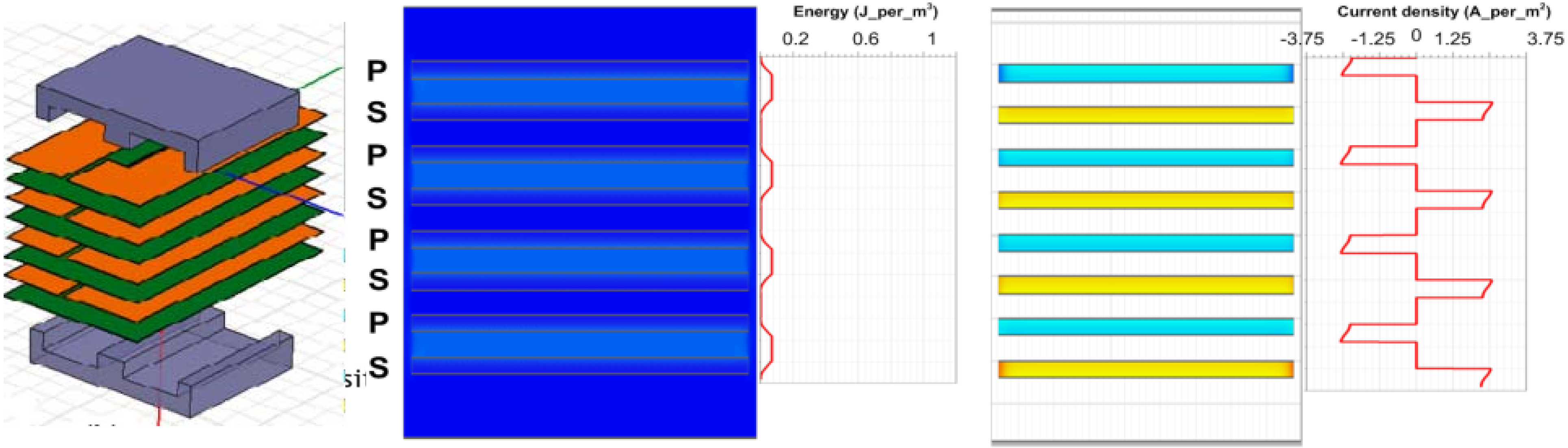


Interleaved Winding



Leakage energy distribution

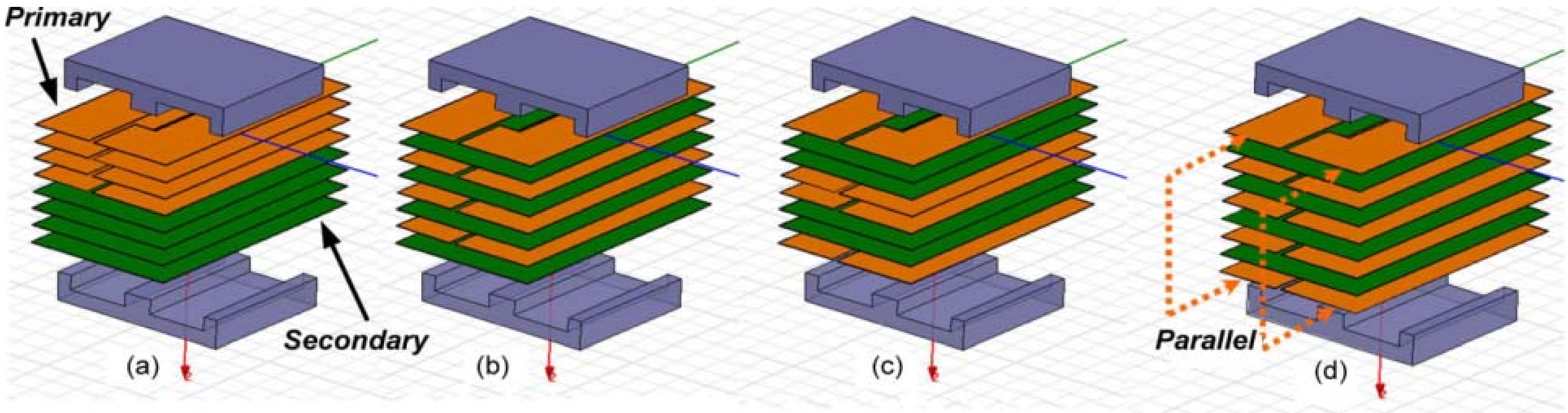
Current density distribution

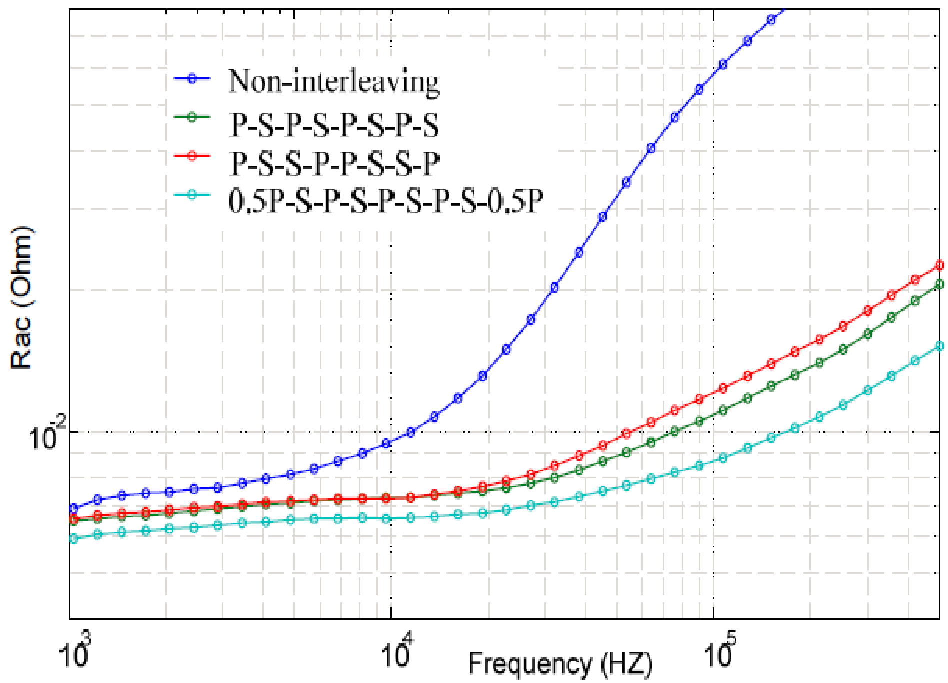


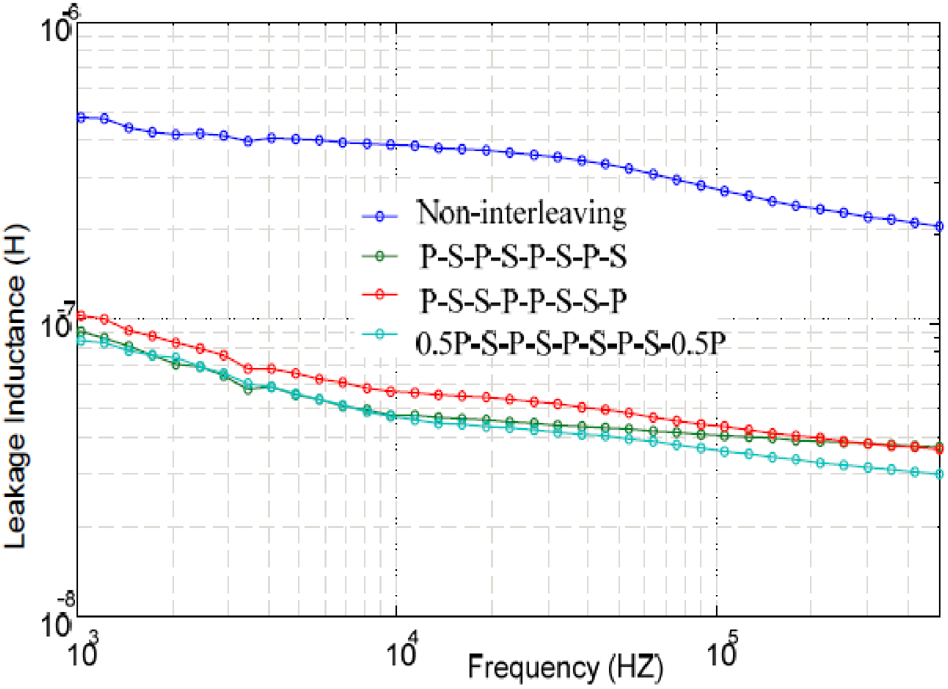
Leakage energy distribution

Current density distribution

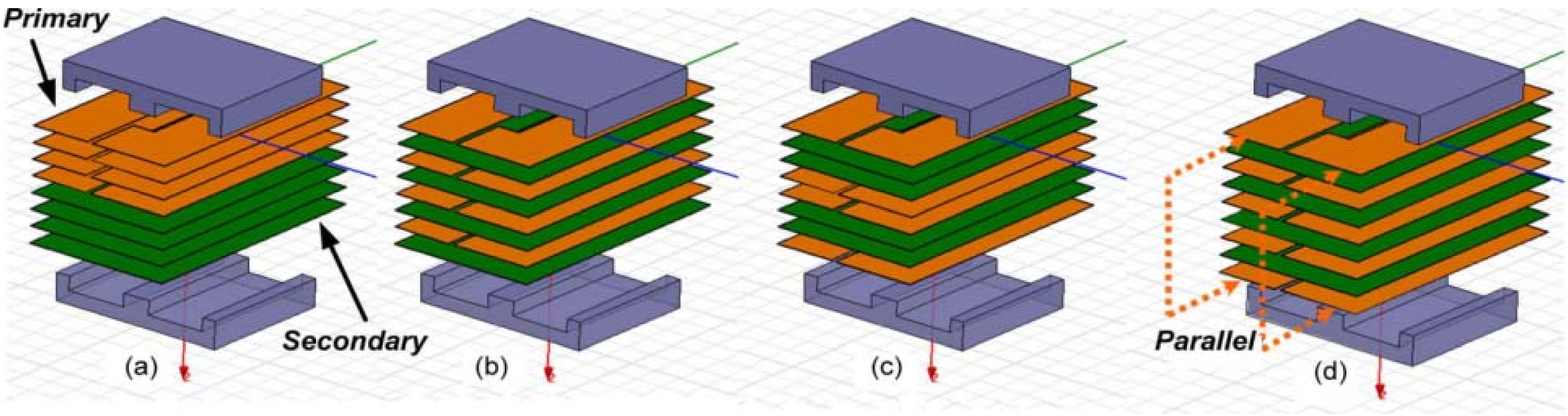
Interleaved Winding

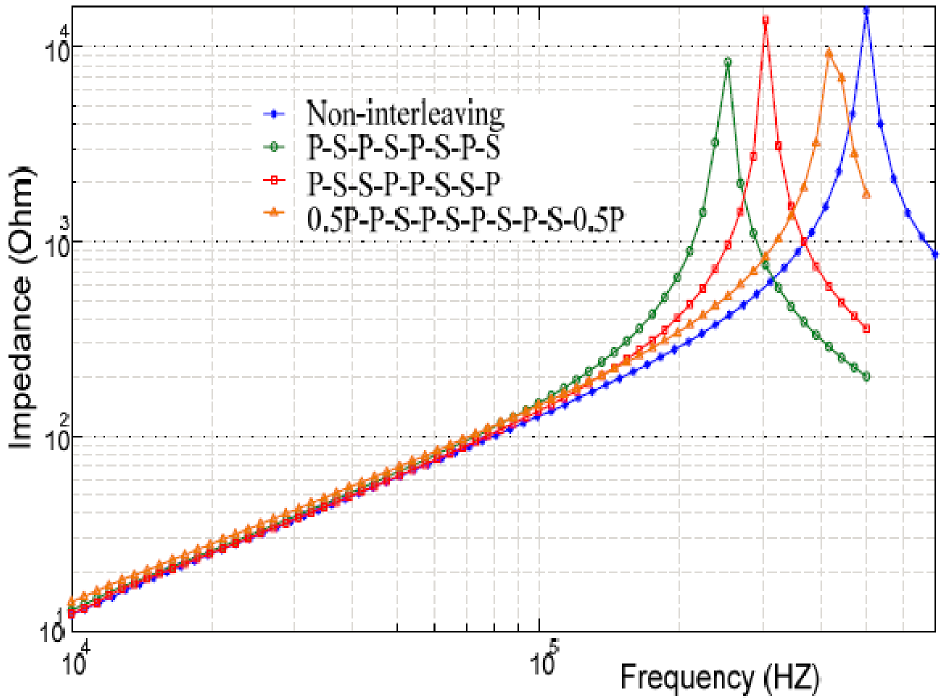






Interleaved Winding





Frequency is 50 kHz	R_{ac}/R_{dc} (mOhm)	L_{leak} (nH)	C_{str} (nF)
Non-interleaving	32.2/6.55	324	1.15
P-S-P-S-P-S-P-S	9.14/6.14	43.8	9.41
P-S-S-P-P-S-S-P	10.1/6.21	47.4	4.24
P/2-S-P-S-P-S-P-S-P/2	7.68/5.69	40.6	3.89

Planar Magnetic Components Integration

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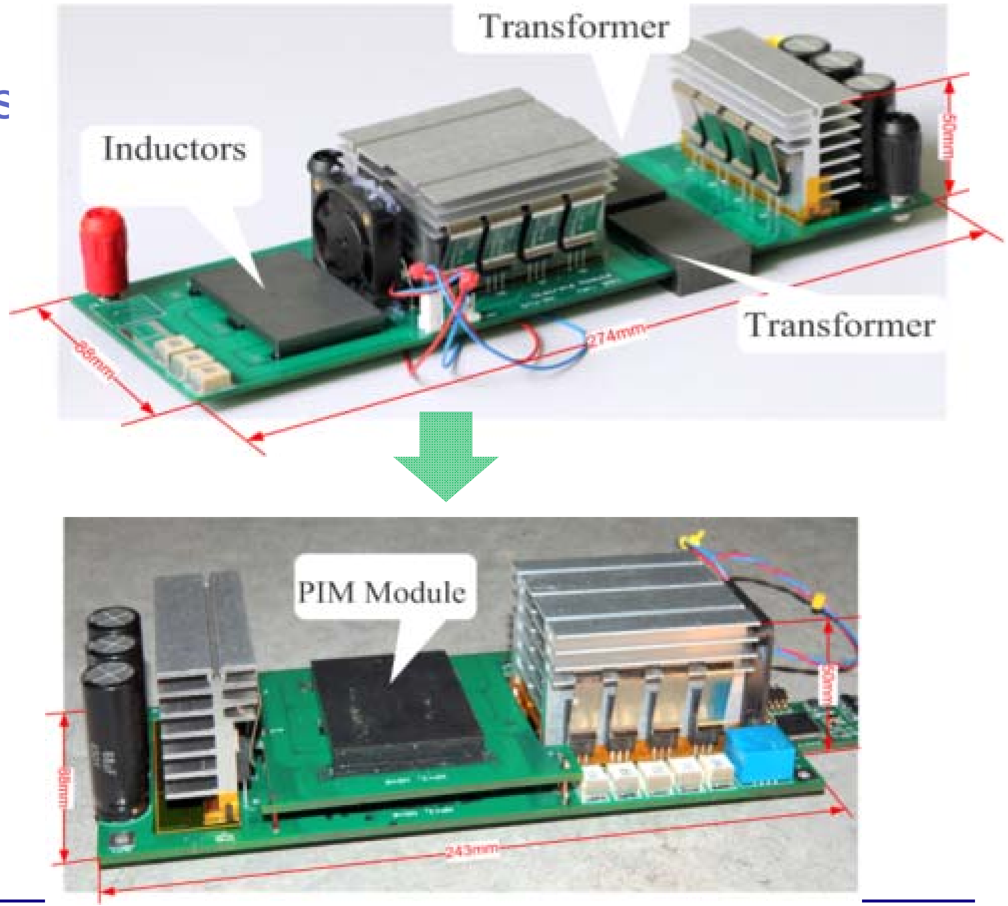
Introduction

functional devices integration, in which discrete magnetic devices with different functions are assembled as one integrated magnetic device.

Mixing the functions of transformers and inductors in:

- Current doubler rectifiers
- LLC resonant converters
- Integrated EMI filters etc.



Introduction

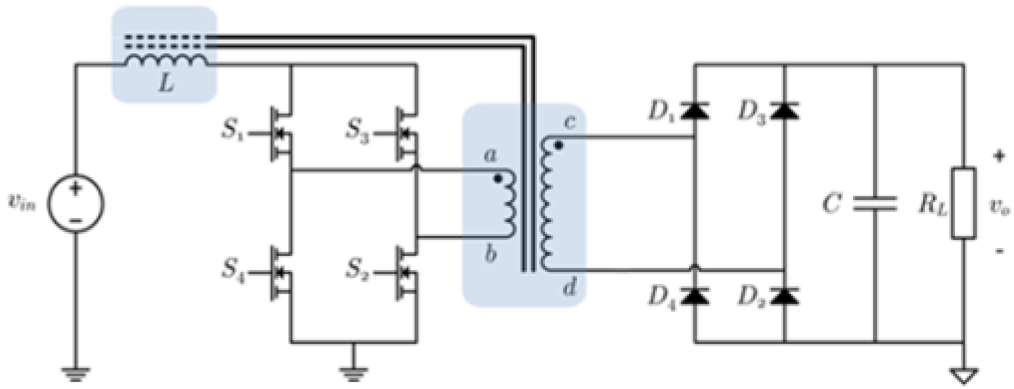
Major advantage:

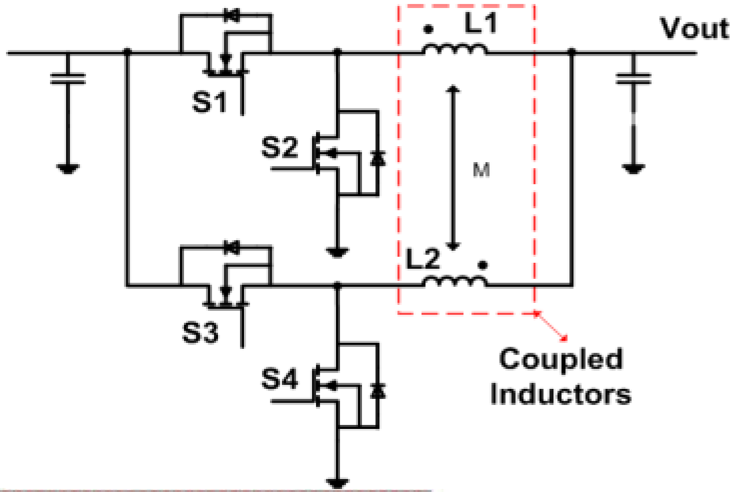
- Smaller size
- higher power efficiency
- lower core loss (spark interest into the light load conditions with the integrated magnetics)

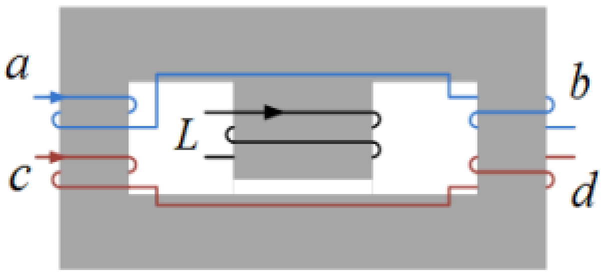
It is often used in applications where space is highly restricted such as computer systems, data center, automotive electrical systems and space applications.

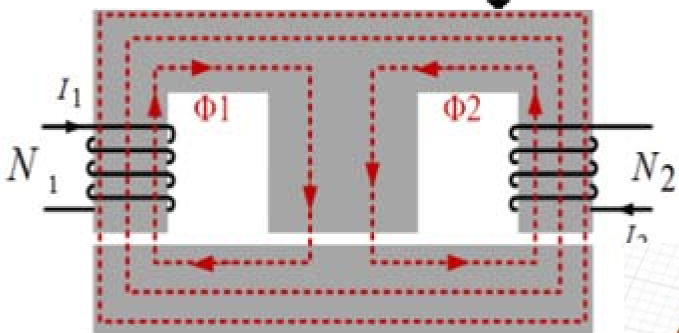
❖ Magnetic core sharing

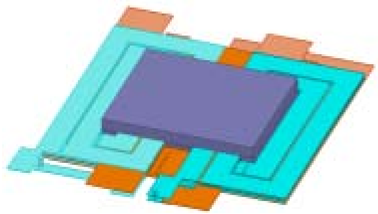
(Only the core is shared, and the windings are not shared)



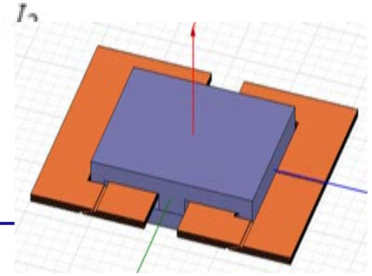






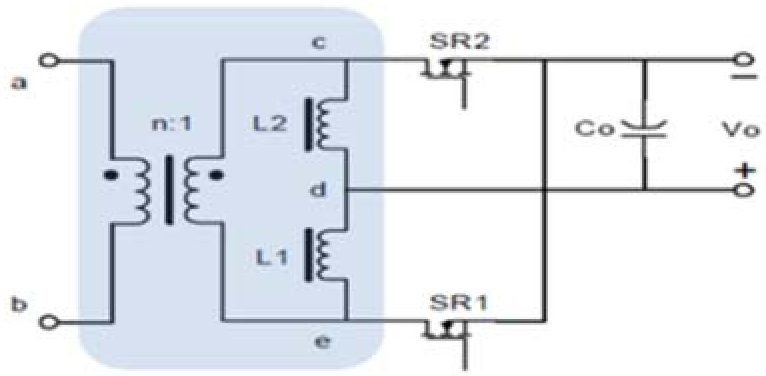


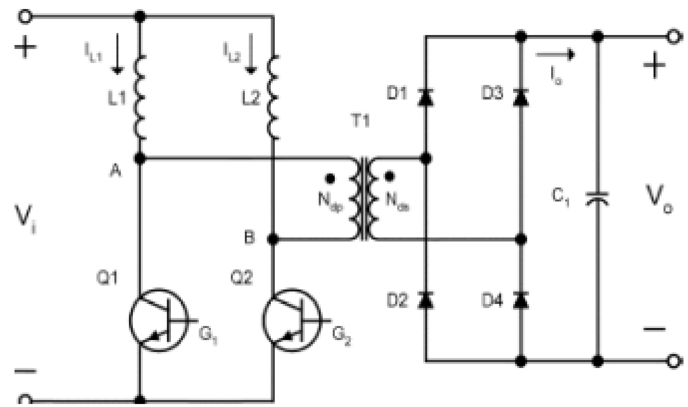
ersity

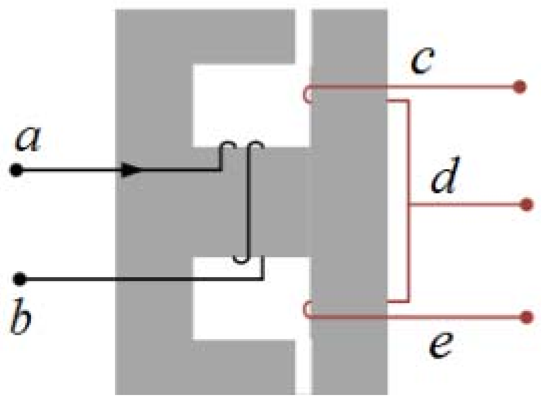


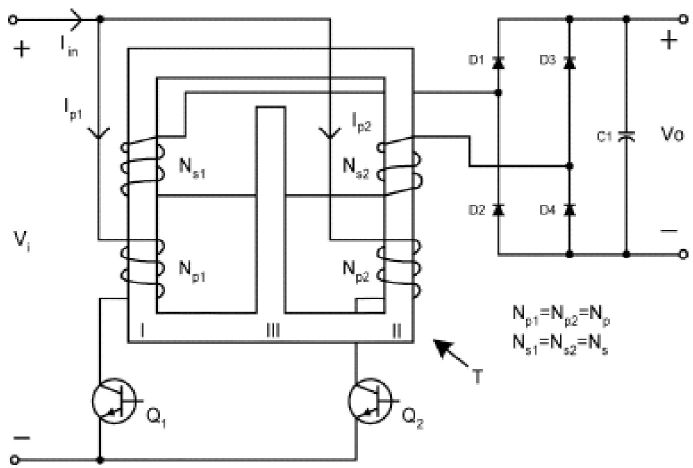
❖ Both winding and magnetic core sharing

(Fully integration)





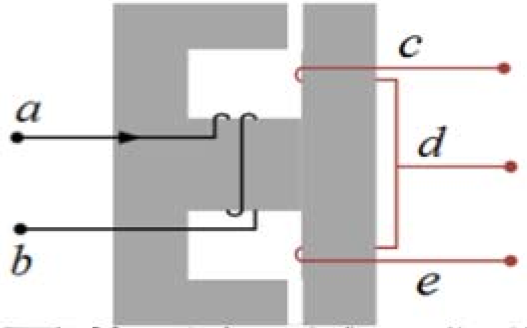


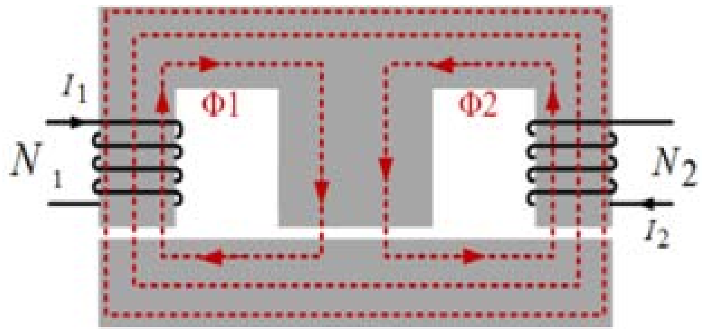


Coupled Integration

The magnetic field produced by one coil passes through the other coil, changing their effective inductances due to the mutual relationship:

- Coupled Inductors
- Multi-winding transformers
- …

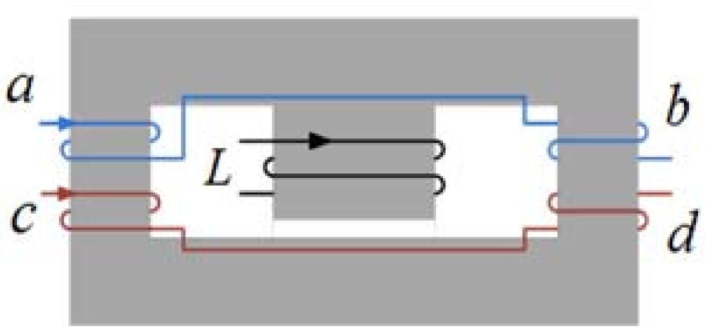


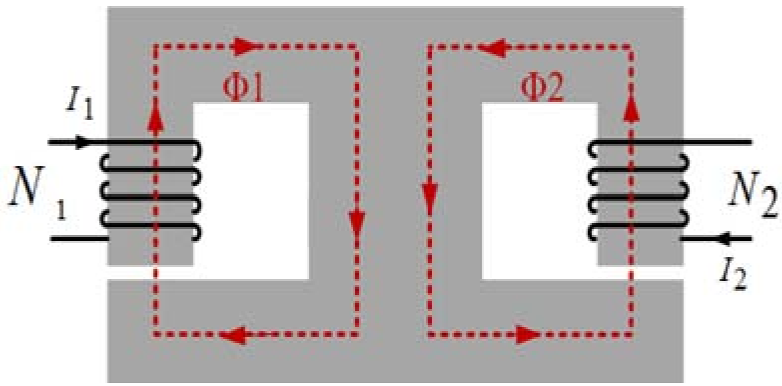


many advantages,…
but can **NOT** be applied in all circuit topologies

Decoupled Integration

The magnetic field produced by one coil does not pass through the other coil or cancel in the other coil.

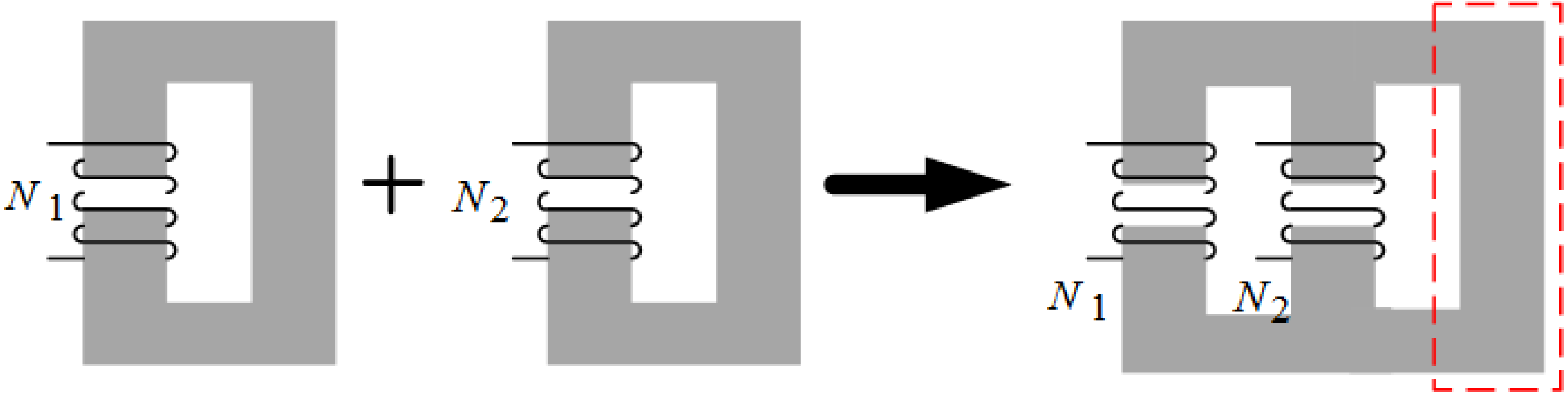




can be used in nearly all circuit topologies due to their independent operation behaviors

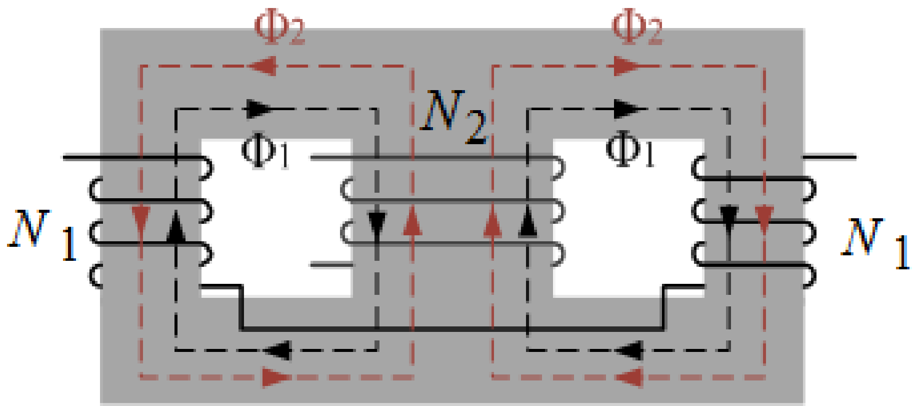
Decoupling Methods

❖ Shared low core reluctance path



(a) shared low reluctance path

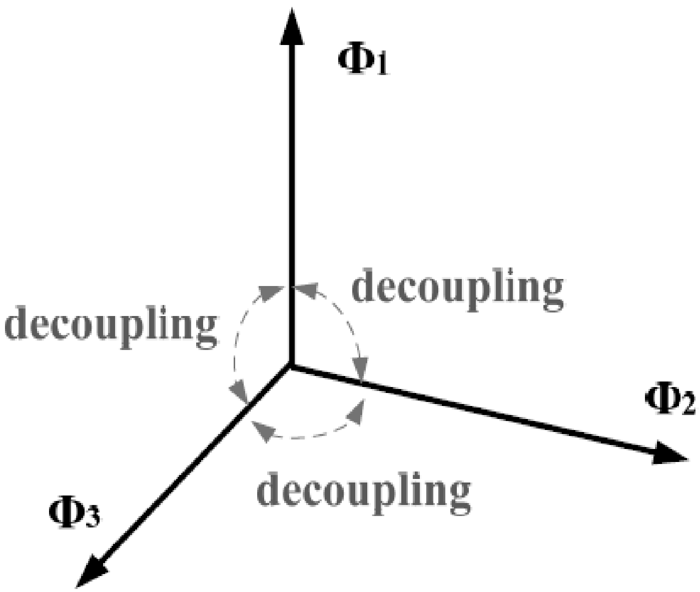
❖ Flux cancellation

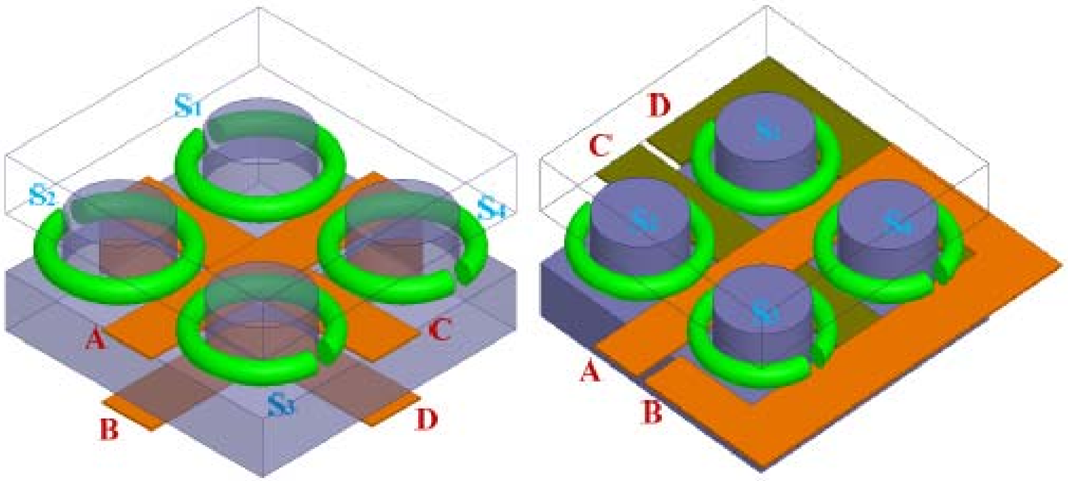


(b) flux cancellation

Decoupling Methods

❖ Orthogonal flux path





(a) (b)

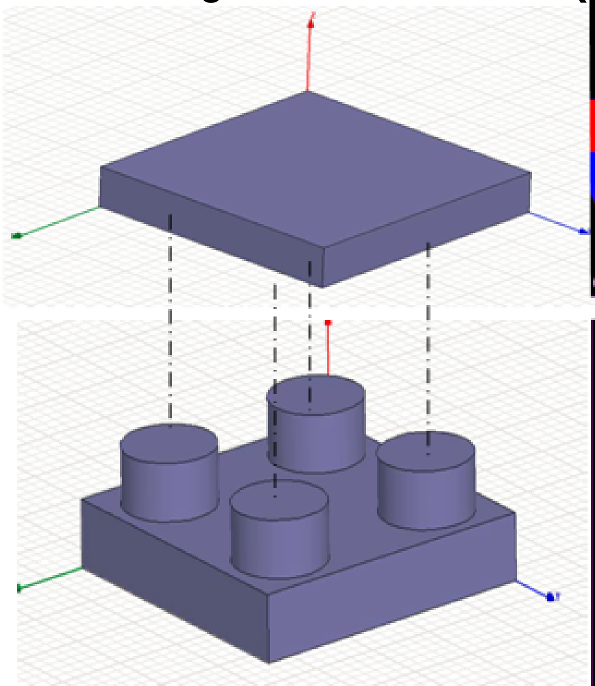
[Ref.]: Z. Ouyang, M. A. E. Andersen and Z. Zhang, "An integrated magnetics component," PCT/EP2012/067422, published no.: WO 2013037696 A1, Sept., 2011.

APEC

Four Quadrants Integrated Transformer

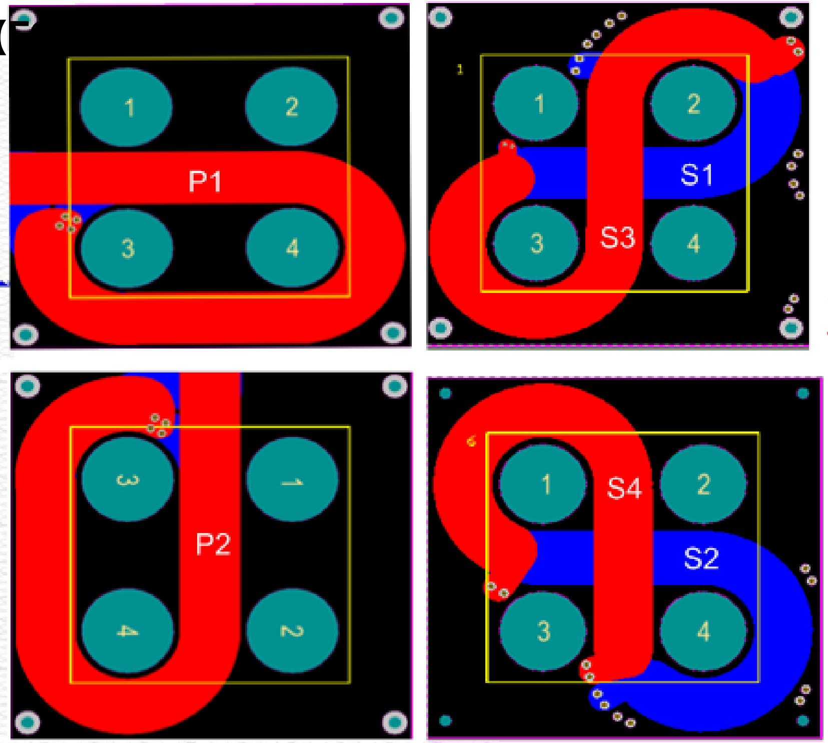
Core geometry

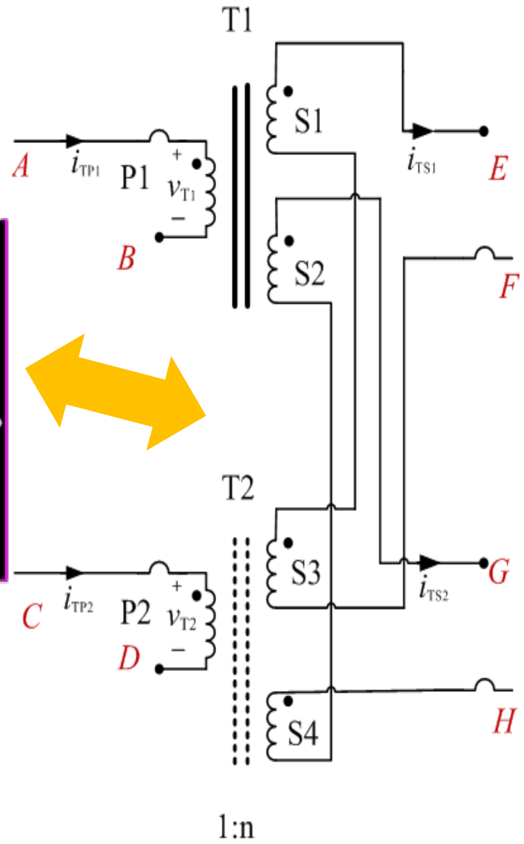
Integrated transformer (

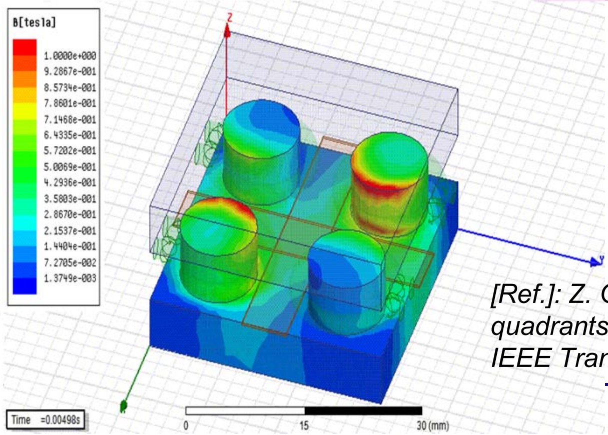


Primary turns

Secondary turns

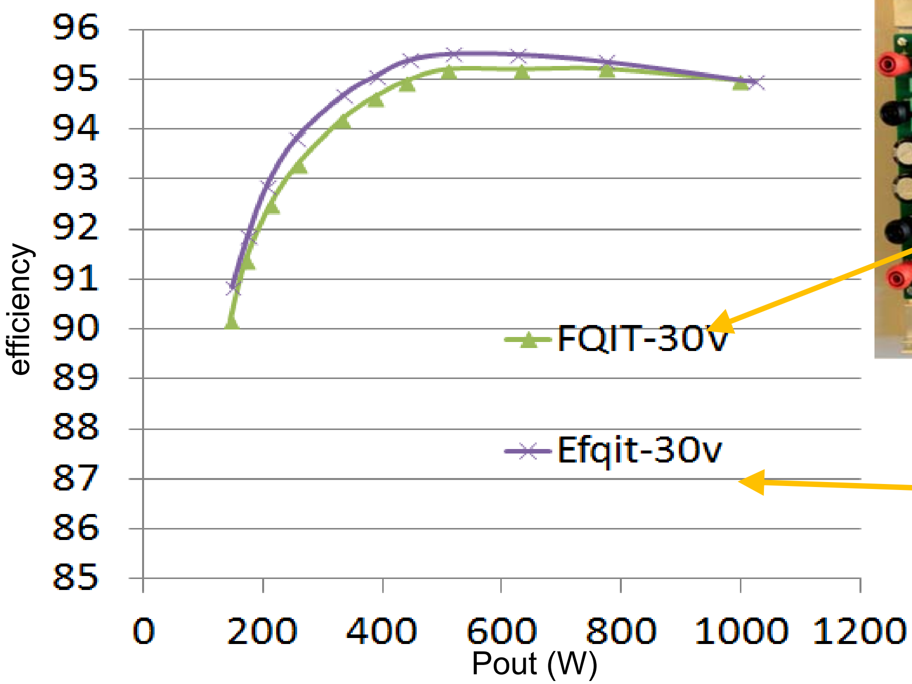


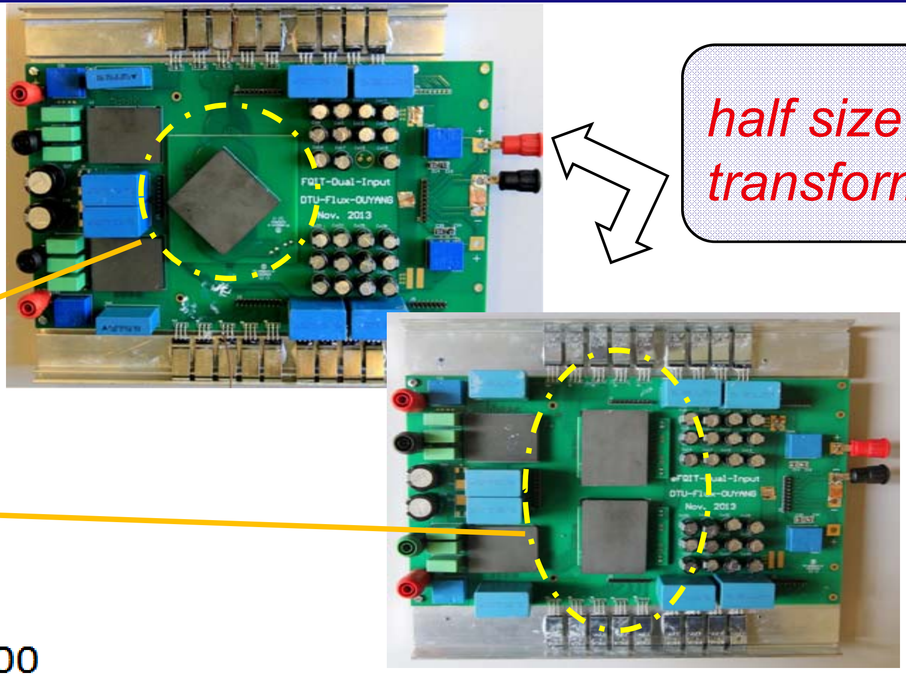




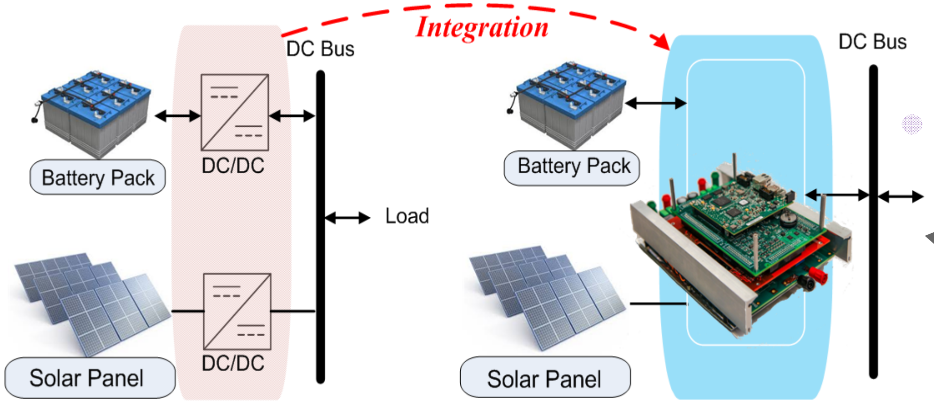
[Ref.]: Z. Ouyang, Z. Zhang, M. A. E. Andersen and O. C. Thomsen, “Four quadrants integrated transformers for dual-input isolated dc-dc converters,” IEEE Trans. on Power Electron. vol.27, no.6, pp.2697-2702, June 2012

Four Quadrants Integrated Transformer





half size of transformer



Advantages

- ✓ wide input range
- ✓ simple control and communication
- ✓ high reliability
- ✓ low overall system cost
- ✓ efficient thermal management
- ✓ compact packaging

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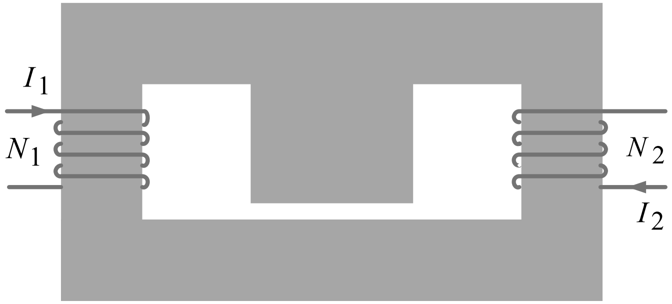
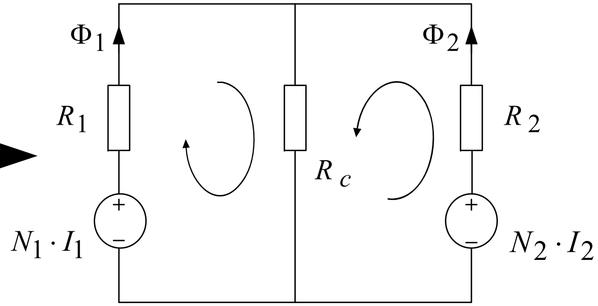
[Ref.]: Z. Ouyang and M. A. E. Andersen, "Three-port DC-DC converter with new integrated transformer for DC Distribution Systems," in the 2014 international power electronics conference-PCIM Asia, Shanghai, June, 2014.

Magnetic Reluctance Modeling

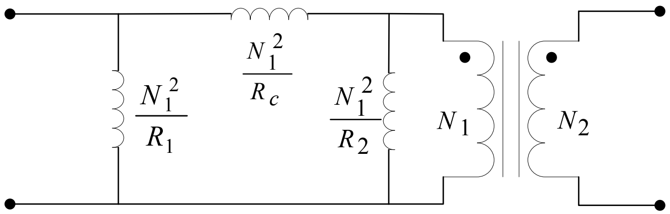
Applying reluctance-resistance analogy

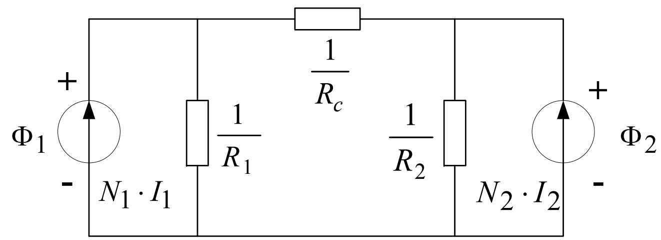




Reluctance Model

duality transformation rules



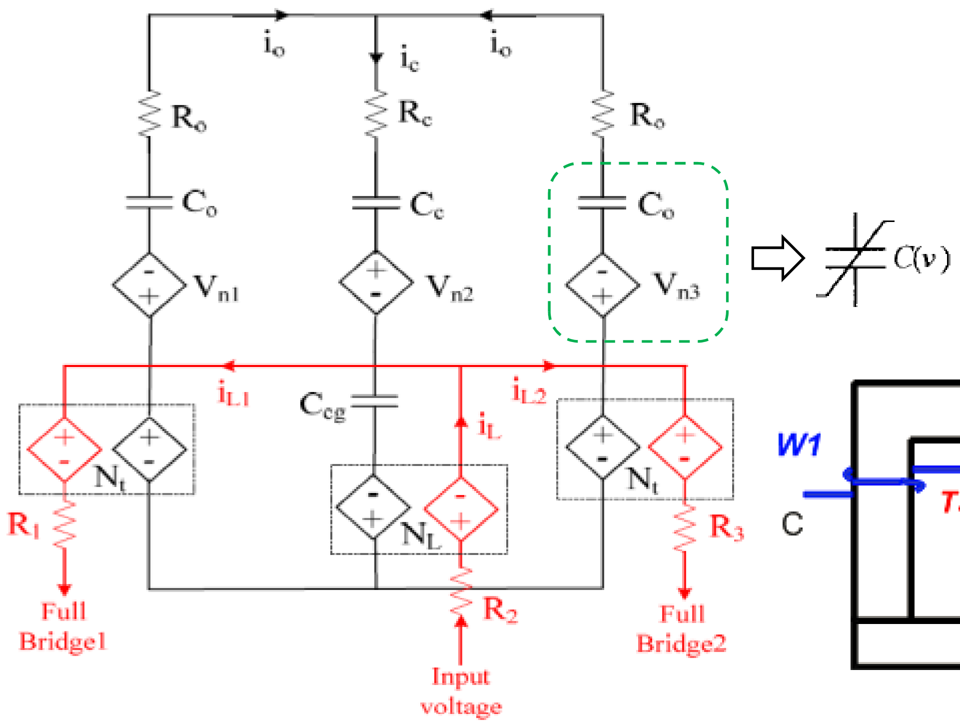


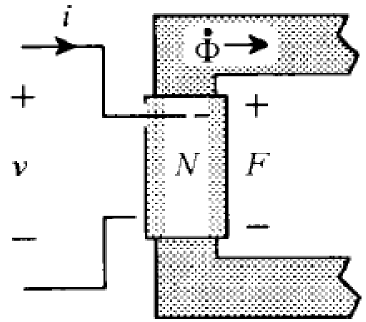
Inductance Model

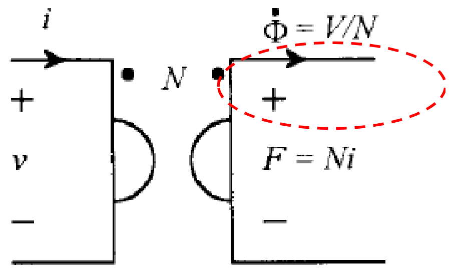
For a practical interest that the magnetic and electrical circuits interact

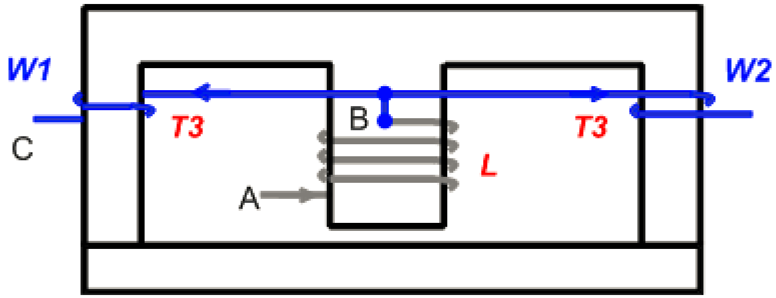
Gyrator-Capacitor Modeling

- ❖ Energy interchange between windings and cores
- ❖ Understand energy relations and dynamics in the context of power electronics



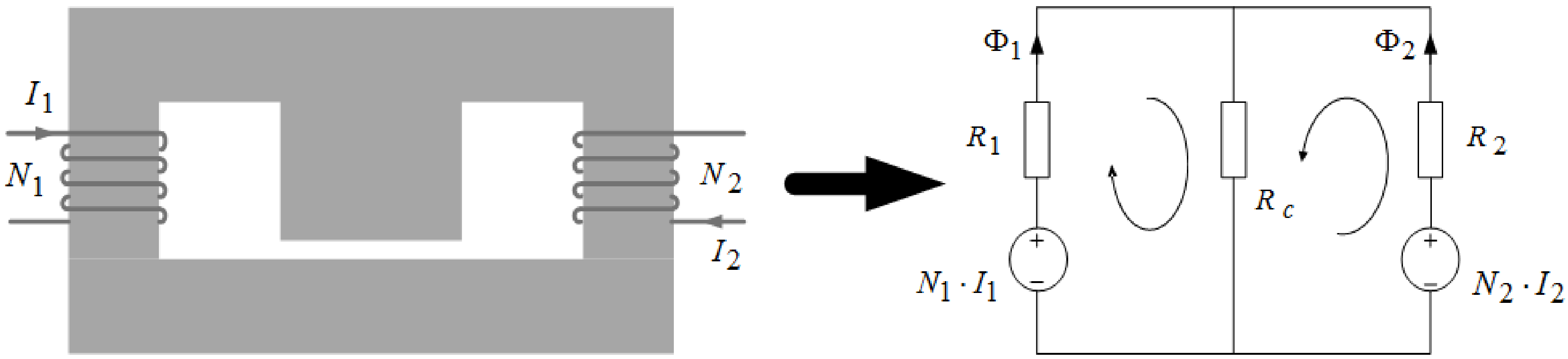






[Ref.]: G. Sen, Z. Ouyang, O. C. Thomsen, M. A. E. Andersen and L. Møller, "Integrated current balancing transformer for primary parallel isolated boost converter," in Proc. EPE, Birmingham, U.K, Sept. 2011.

Applying reluctance-model:



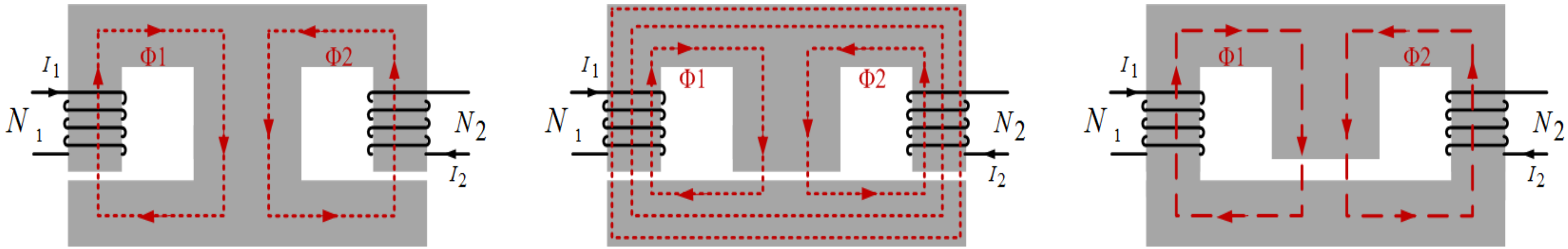
Since Kirchhoff's laws and Faraday's law,

$$\begin{bmatrix} v_1 \\ v_2 \end{bmatrix} = \begin{bmatrix} \frac{N_1^2 \cdot (R_2 + R_c)}{\Delta} & -\frac{N_1 \cdot N_2 \cdot R_c}{\Delta} \\ -\frac{N_1 \cdot N_2 \cdot R_c}{\Delta} & \frac{N_2^2 \cdot (R_1 + R_c)}{\Delta} \end{bmatrix} \cdot \begin{bmatrix} \frac{di_1}{dt} \\ \frac{di_2}{dt} \end{bmatrix} = \begin{bmatrix} L_{11} & L_M \\ L_M & L_{22} \end{bmatrix} \cdot \begin{bmatrix} \frac{di_1}{dt} \\ \frac{di_2}{dt} \end{bmatrix}$$

where, $\Delta = R_1 \cdot R_2 + R_1 \cdot R_c + R_2 \cdot R_c$

Coupling coefficient: *(assuming N1 =N2)*

$$k = \frac{L_M}{\sqrt{L_{11} \cdot L_{22}}} \xRightarrow{L_{11} = L_{22}} \boldsymbol{k} = \frac{L_M}{L_{11}} = -\frac{N_2 \cdot R_C}{N_1 \cdot (R_2 + R_C)}$$



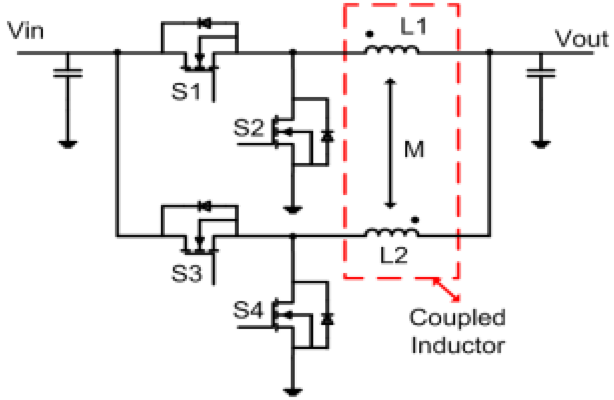
(a) (b) (c)

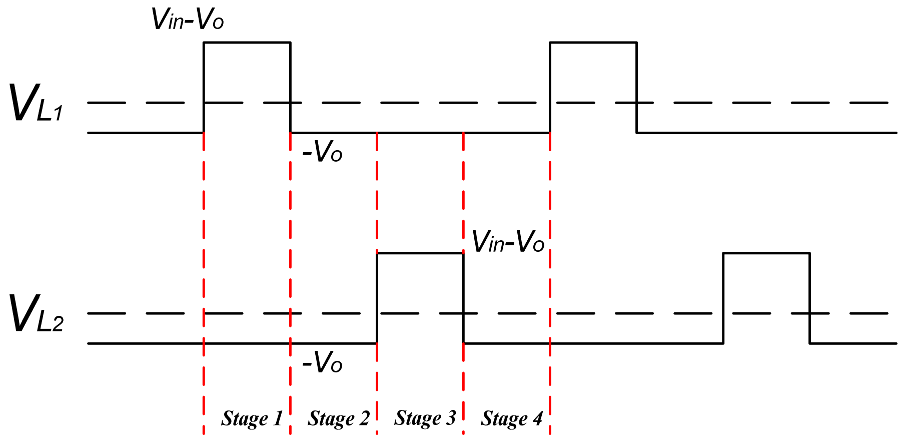
$R_2 >> Rc$ $\Rightarrow$ $k \approx 0$

$R_2 = 2*Rc$ $\Rightarrow$ $k = \frac{1}{3}$

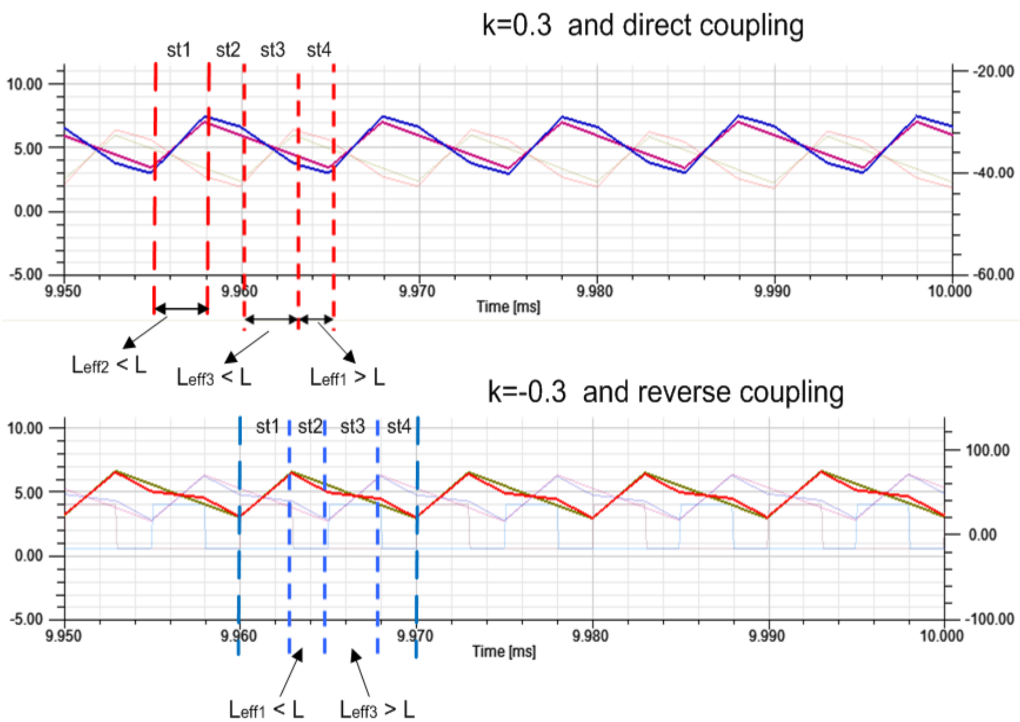
$R_2 << Rc$ $\Rightarrow$ $k \approx 1$

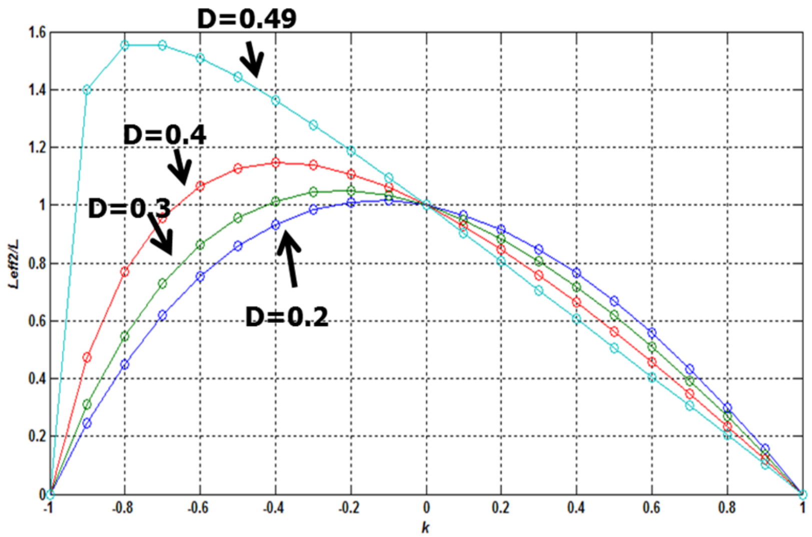
Coupled inductors





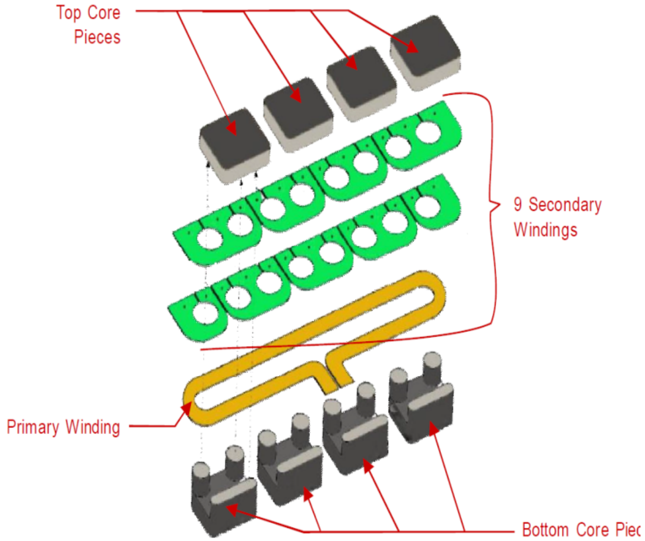
Effective inductances are dependent of circuit operation

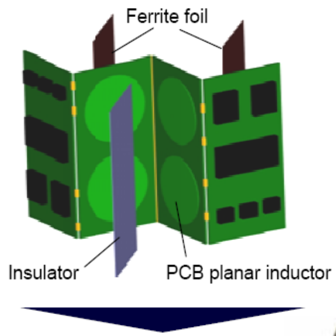


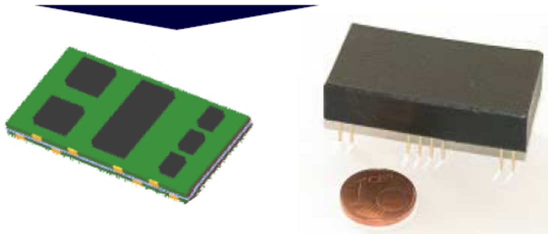


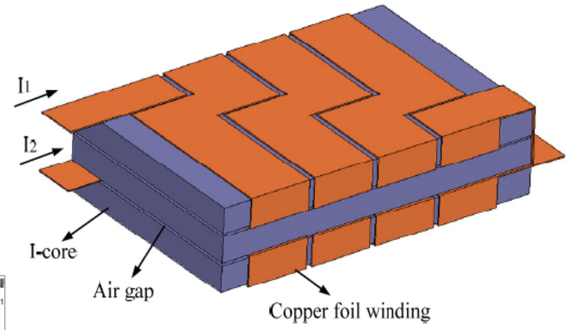
[ref.]: P.-L. Wong, Q.-Q. Wu, P. Xu, B. Yang, and F. C. Lee, "Investigating coupling inductors in the interleaving QSW VRM," in Proc. IEEE Appl. Power Electron. Conf. Expo., 2000, pp. 973–978.

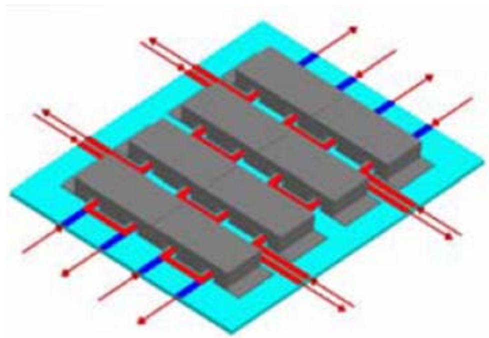
❖ Integrated magnetics is actually an "open" technology, and may create "innovation" ideas.











Thin isolating low-power converters with folded planar transformer, e.g. for gate drive applications (patent pending DE 10 2004 026 052)

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Summary

- Integrated magnetics has advantages **less number of components**, **smaller size**, and potentially **higher power efficiency**.

- Integrated magnetics are not all advantageous. The main issue is to produce **unwanted parasitic capacitances** among the inductive elements, and a **limited power capability**.

- Must understand the principle of circuit operation.

- **New geometry core** may create something interesting.

Conclusion and Trends

- Planar magnetics still gain its popularity due to low profile and easy manufacture

- Planar magnetics towards very high frequency (so-called micro-inductor/transformer) would be interesting. Accurate 2D/3D models are emergency

- High frequency magnetic materials are emergency

- New winding and core geometries may create something innovative

Thank you very much!

Questions?

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