Virtual Prototyping of Magnetic Components

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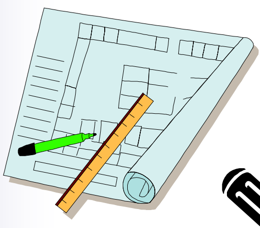
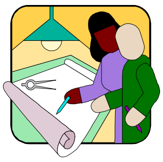
Introduction





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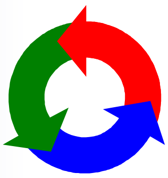
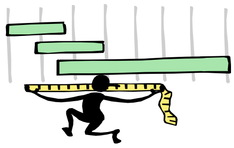
Does Virtual Prototyping make sense?

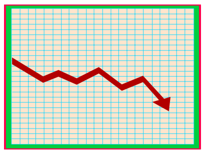
Design and prototype iterations



Project schedule delay



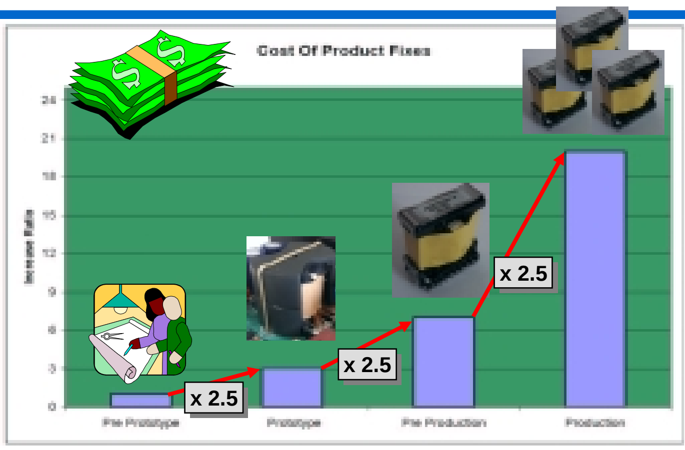
Delay of market entry



Revenue losses

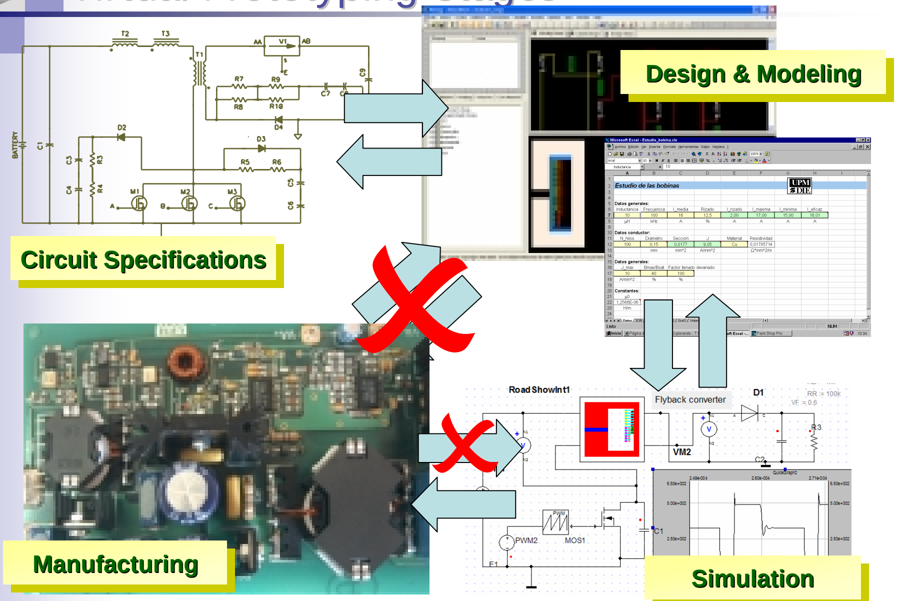
Does Virtual Prototyping make sense?

Cost of finding a problem is a function of where in the design process you find the problem



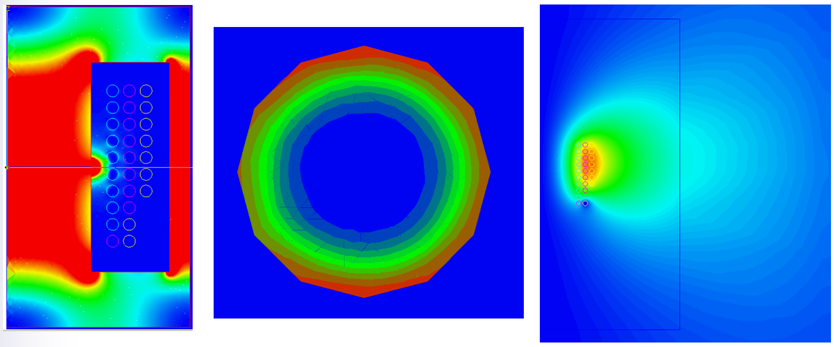
Source: Paul Duffy, Manager, E/E CAE Department, Ford Motor Company, 9/18/2002

Virtual Prototyping Stages



Seminar Contents

Basic Concepts



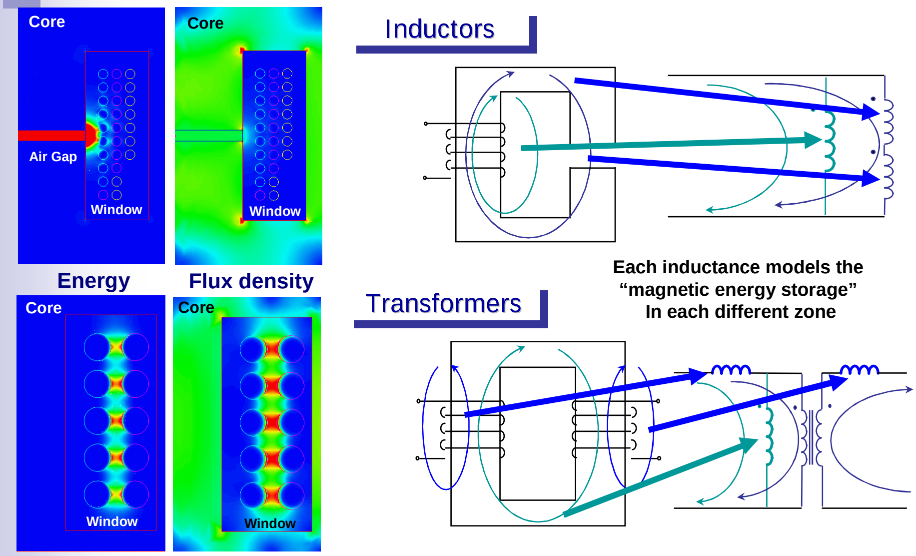
Virtual Prototyping of Magnetic Components

Basic Concepts

- There are many Applications of Magnetic components
  - ✓ Galvanic Isolation
  - ✓ Adjust Voltage Levels
  - ✓ Filters
  - ✓ Resonant Inductors
  - ✓ Measurements for Feedback and Protections
  - ✓ Pulse Transformers
  - ✓ …
- But basically, they produce
  - ✓ Energy Storage
  - ✓ Energy Transfer
  - ✓ Losses

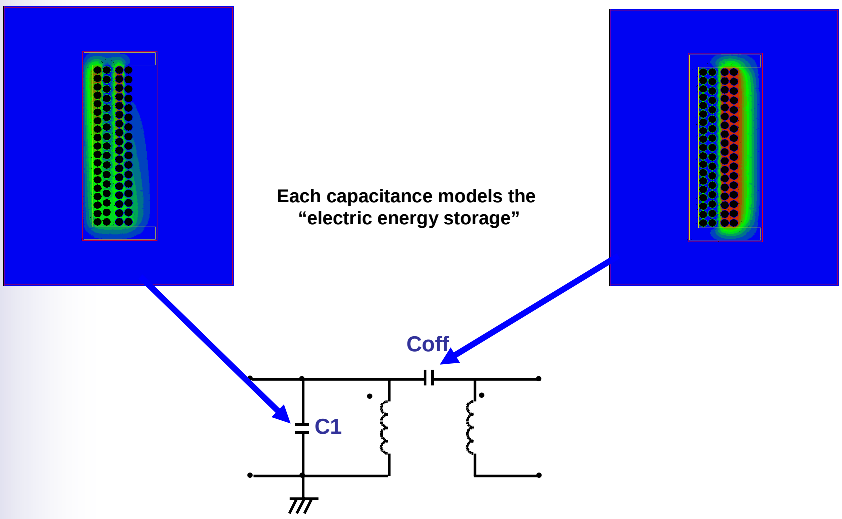
Virtual Prototyping of Magnetic Components

Magnetic Energy



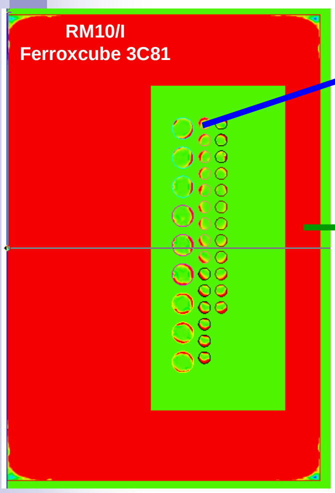
Each inductance models the "magnetic energy storage" In each different zone

Electric Energy

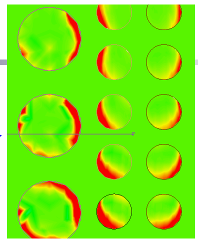


Each capacitance models the "electric energy storage"

Losses



Losses @ 1MHz



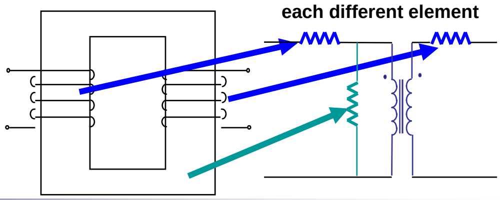
Winding Losses

- Skin effect
- Proximity effect
- Gap effect
- End effect

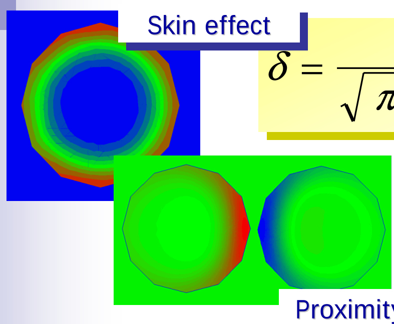
Core Losses

- Eddy Current
- Hysteretic
- Residual

Each resistance models the "losses" in each different element



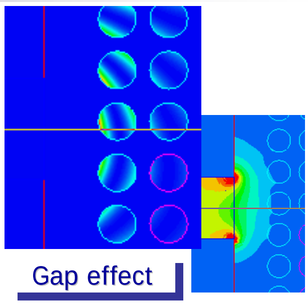
Winding Losses



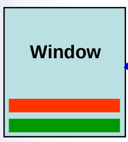
Skin effect

$$\delta = \frac{1}{\sqrt{\pi f \mu_0 \sigma}}$$

Proximity effect



Gap effect



	Different Open Circuit Resistances		Same Short Circuit Resistance
	R_{open_prim} (mΩ)	R_{open_sec} (mΩ)	R_{short_prim} (mΩ)
A	3.5	3.65	5.74
B	6.32	6.33	5.74

Core Proximity effect

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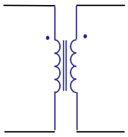
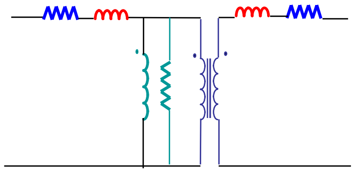
Ideal and Real Components





- Ideal transformer:

$$\frac{d\phi}{dt} = \frac{v_1}{n_1} = \frac{v_2}{n_2} = \dots = \frac{v_p}{n_p}$$

$$n_1 \cdot i_1 = n_2 \cdot i_2$$

No energy storage !!

No losses !!

- Real transformer :
  - ✓ **Common energy:**

$$L_{mag,\,i} = \frac{n_i^2}{\Re}$$

$$\sum ni = \Phi \cdot \Re$$

From any winding

  - ✓ **Self energy**
  - ✓ **Conductor losses**

Notice the difference between "transferred" and "stored" energy

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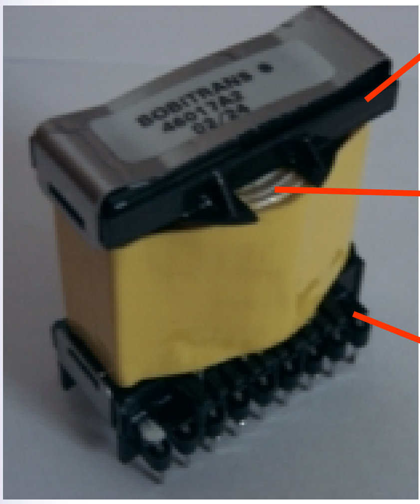
Magnetic Component Elements



Core

- ☑ **Flux Path**
- ☑ **Energy storage**

Windings

- ☑ **Electrical terminals**
- ☑ **Current path**

Bobbin

- ☑ **Windings holder**

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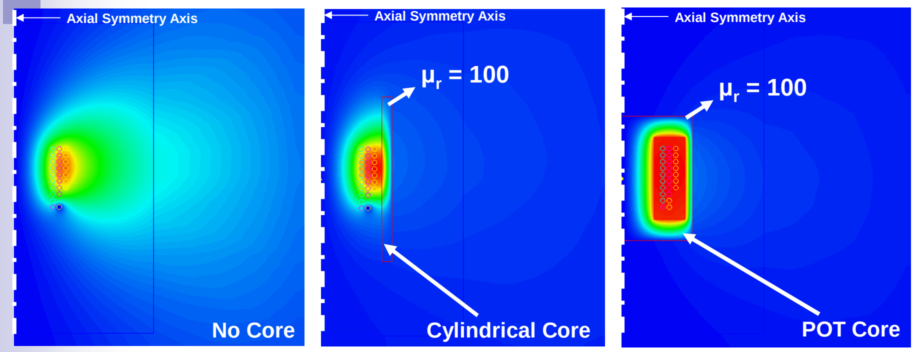
The Role of the Core I



Flux Lines

The Core helps to create a flux path

Additionally the core determines the energy and inductance

Unfortunately, the core also produces losses

The Role of the Core II

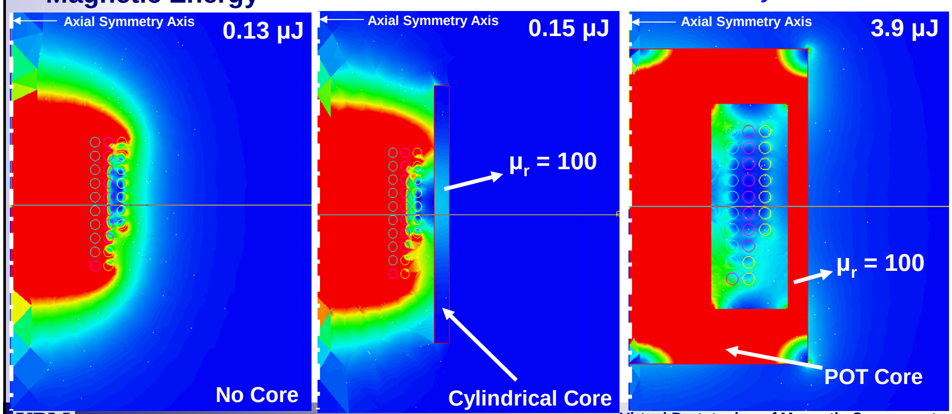
Reluctance

$$\Re = \frac{1}{\mu} \cdot \frac{\text{length}}{Area}$$

$$L = \mu_o \cdot \mu_r \cdot \frac{Area}{Length} \cdot n^2$$

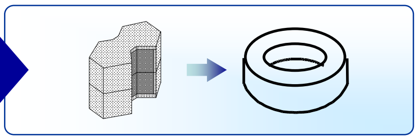
Material Geometry Turns

Magnetic Energy

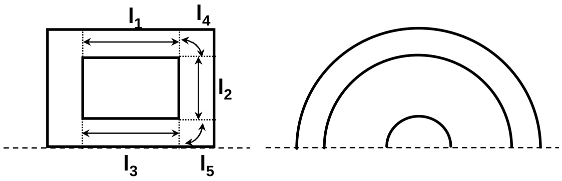


The Effective Values of the Core

Effective Values



Example: POT Core



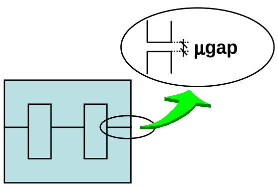
$$L = \mu_o \cdot \mu_r \cdot \frac{Area}{Length} \cdot n2$$

$$A_e = \frac{C_1}{C_2}; \quad V_e = \frac{C_1^3}{C_2^2}; \quad l_e = \frac{C_1^2}{C_2}$$

$$C_1 = \sum_1^5 \frac{l_i}{A_i} mm^{-1} \quad C_2 = \sum_1^5 \frac{l_i}{A_i^2} mm^{-3}$$

The Tolerances of the Core

GRADE	A_L (nH)	μ_e	AIR GAP (µm)
3B8 sup	160 ±3%	≈59	≈900
	250 ±3%	≈92	≈500
	315 ±3%	≈116	≈400
	400 ±3%	≈147	≈300
	630 ±3%	≈232	≈150
	4950 ±25%	≈1820	≈0

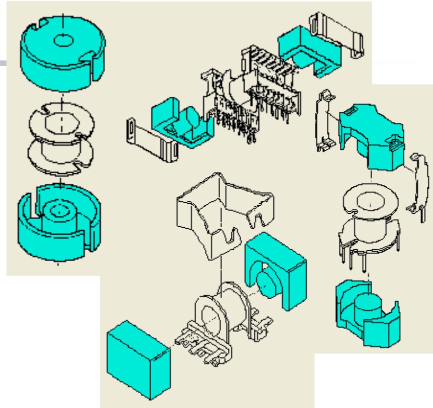


μ_e (and A_L) depends on the "micro-gap" (2.5 µm or 100 mils each union) because of mechanization process

The Core Choices

Shapes
RM, POT, EE, EI, PQ, TOROIDS…

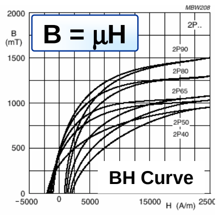
- ☑ Effective values
- ☑ Magnetic coupling
- ☑ Heat transfer
- ☑ Unit Cost
- ☑ Manufacturing Cost

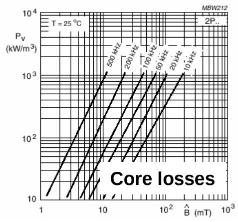


Materials
3F3, 3C85, 3C90, N67, N47…

- ☑ Core losses
- ☑ Permeability and Reluctance
- ☑ Conductivity
- ☑ Saturation flux density

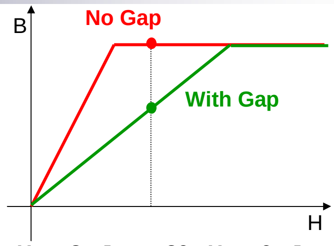
$$L = \mu_o \cdot \mu_r \cdot \frac{Area}{Length} \cdot n2$$

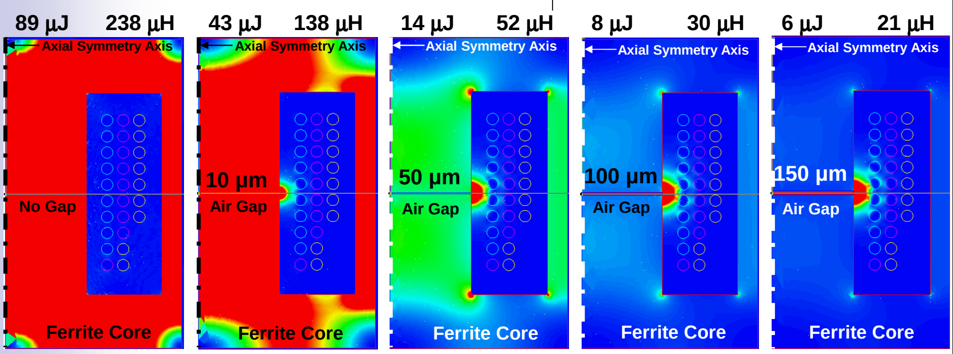




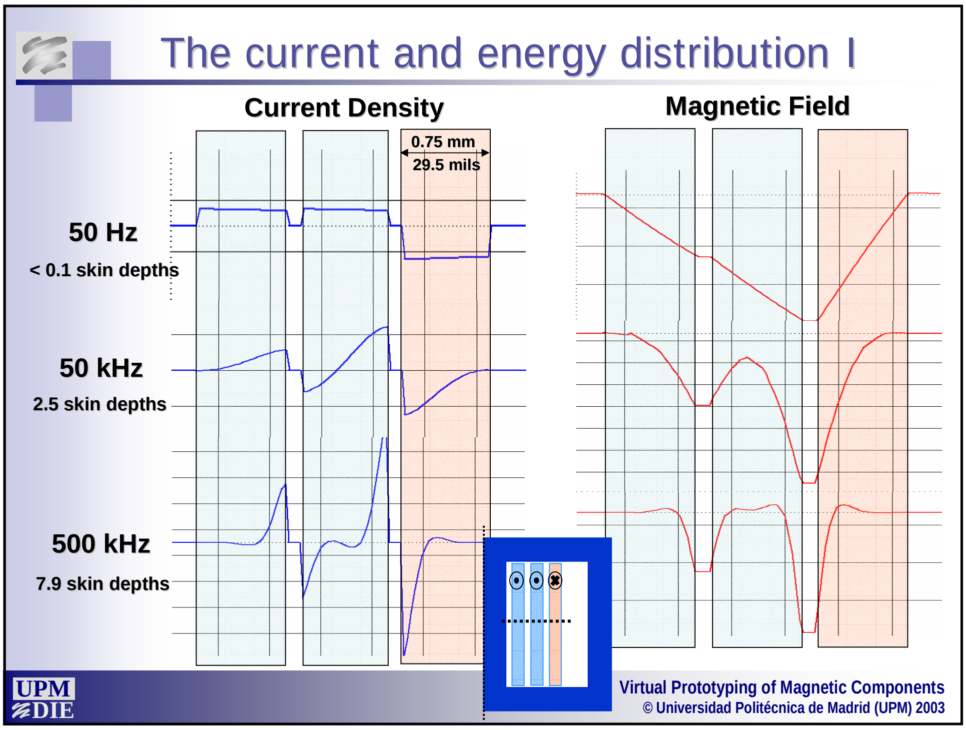
The Role of the Gap

- ☑ **Adjust the reluctance value**
- ☑ **Adjust the energy storage**
- ☑ **Avoid saturation**





Energy


The current and energy distribution I
Current Density
Magnetic Field
0.75 mm
29.5 mils
50 Hz
< 0.1 skin depths
50 kHz
2.5 skin depths
500 kHz
7.9 skin depths
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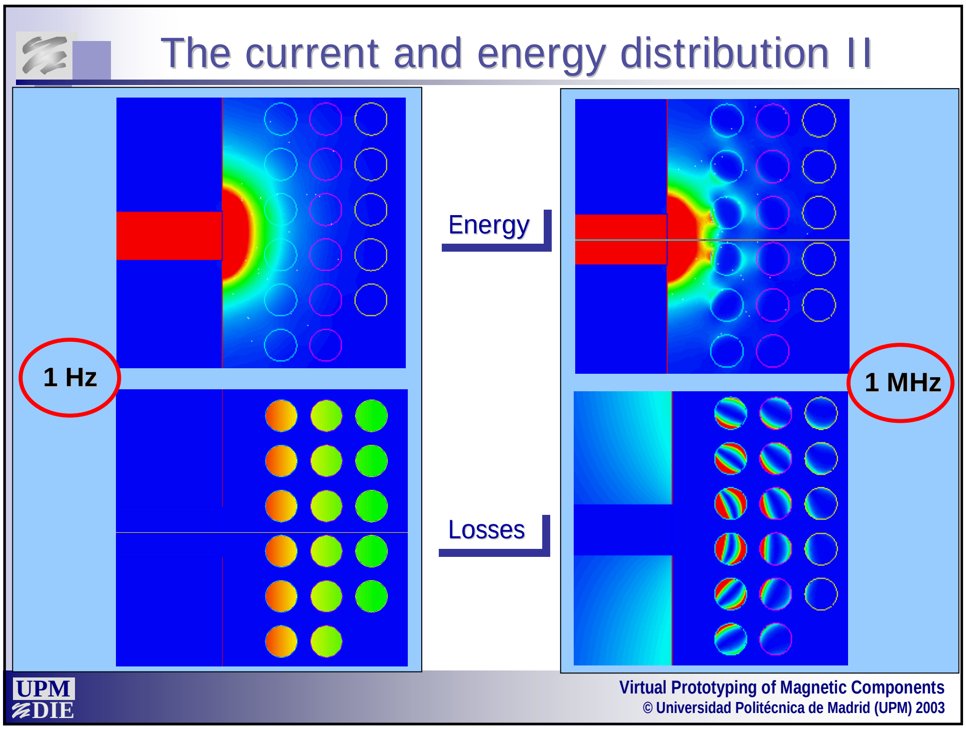
The current and energy distribution II
Energy
Losses
1 Hz
1 MHz
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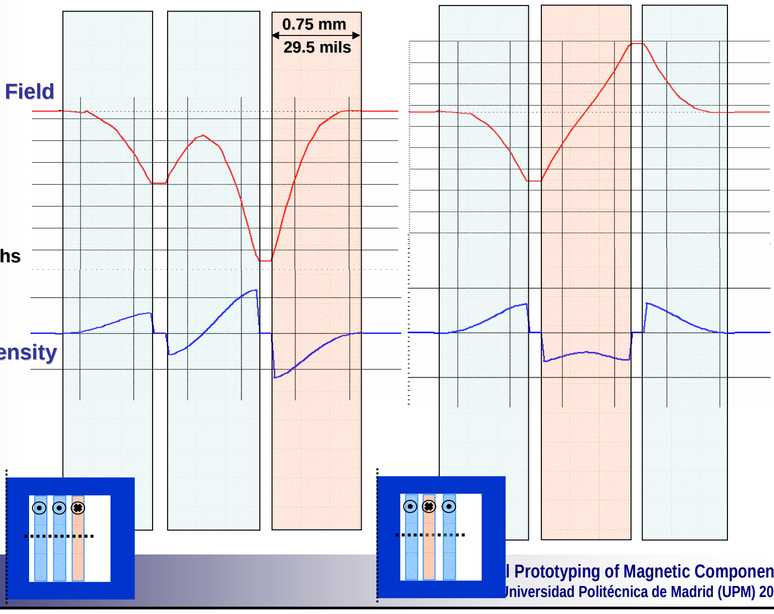
Interleaving Application

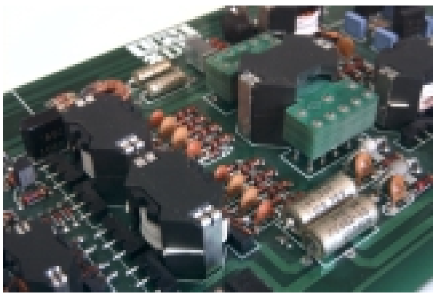
Magnetic Field

50 kHz
2.5 skin depths

Current Density



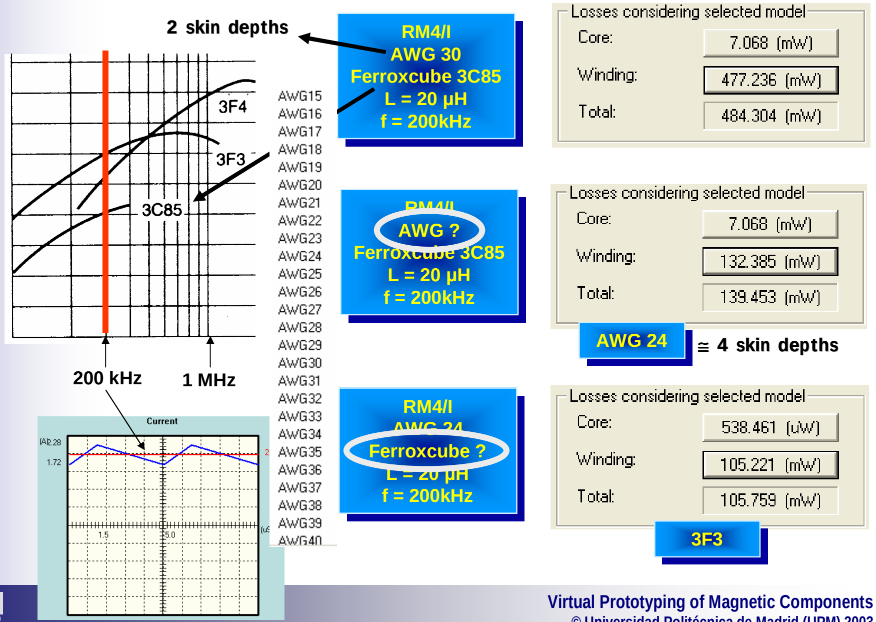
Design



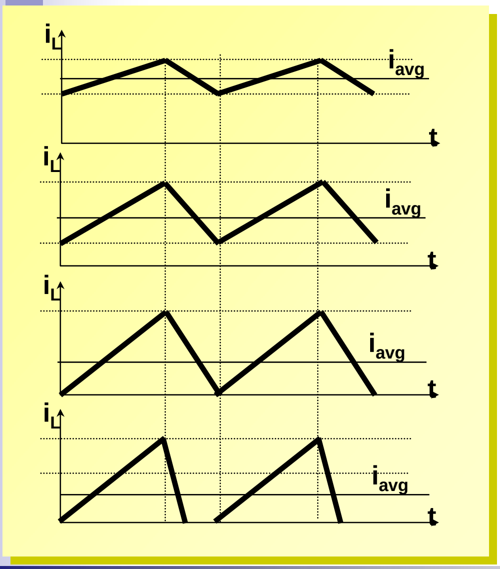
Selection of Constructive Elements

1.	CORE SHAPE	Depends on the application (see data books)
2.	CORE SIZE	Depends on the power ("area product" or data books)
3.	CORE MATERIAL	Depends on the "frequency" (see data books)
4.	CONDUCTOR TYPE	Depends on the "frequency" (skin depth) Solid wire: low frequency Litz wire and foils: high frequency
5.	CONDUCTOR AREA	Depends on the optimum losses

Example of selection of elements



The "effective frequency" concept

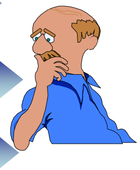


Same frequency? Yes

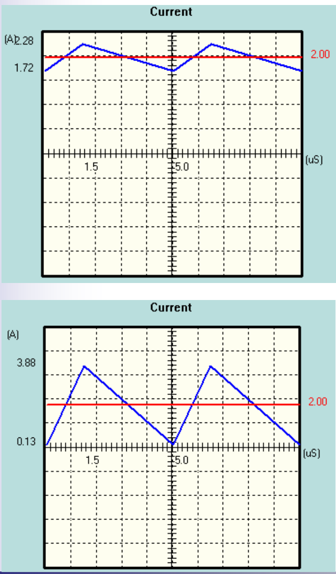
Same "effective frequency"? No

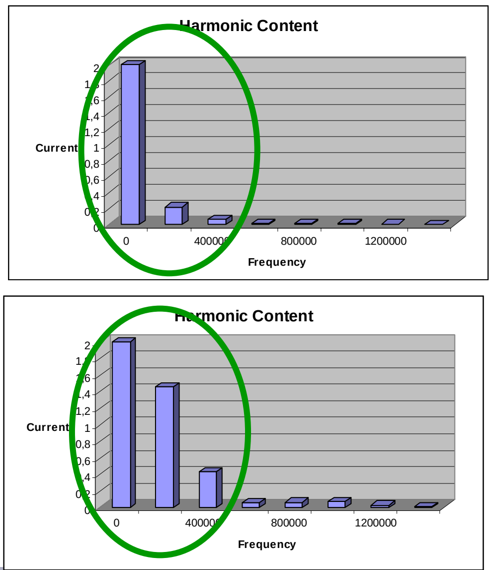
Affects the selection of:

- CORE MATERIAL
- CONDUCTOR TYPE
- CONDUCTOR AREA

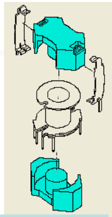


The "effective frequency" concept





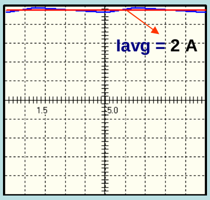
Example (Effective Frequency)



Vpos = 7.5 V; Vneg = 3.21 V; Switch. Freq = 200 kHz; Iavg = 2 A; Duty = 30%

Core shape and size: RM5/I

Wire gauge: AWG28 (370 µm or 14.5 mils diameter)

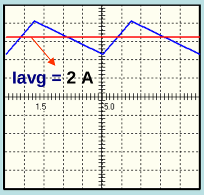


L = 100 µH;
Ripple = 112.5 mA

Ripple = 5.6% Iavg

Ferroxcube 3F35 (Ferrite)
Losses = 785 mW (32 turns)

Ferroxcube 2P90 (Iron Powder)
Losses = 577 mW (16 turns)



L = 10 µH;
Ripple = 1.125 A

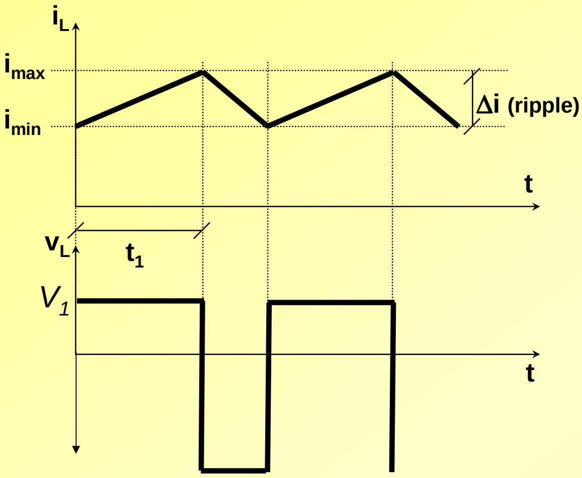
Ripple = 56% Iavg

Ferroxcube 3F35 (Ferrite)
Losses = 89 mW (4 turns)

Ferroxcube 2P90 (Iron Powder)
Losses = 371 mW (16 turns)

Inductors

Calculation of Inductance



$$V_1 = L\frac{\Delta i}{t_1}$$

$$L = \frac{V_1 \cdot t_1}{\Delta i}$$

Inductors

Calculation of initial number of turns

The minimum number of turns should be calculated in order to keep flux density under the saturation value

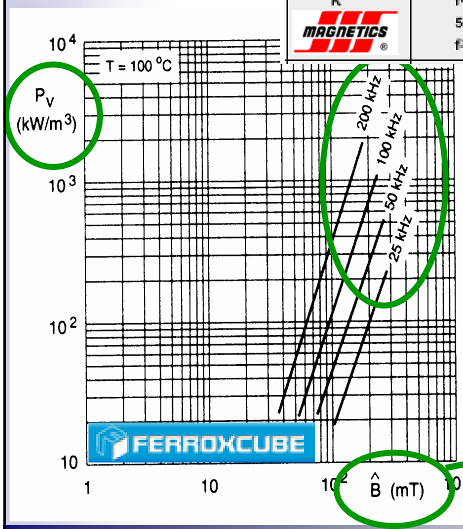
$$L\frac{\Delta i}{\Delta t} = N\frac{\Delta \Phi}{\Delta t} \rightarrow L\Delta i = N \cdot A_e \cdot \Delta B \rightarrow$$

$$\rightarrow L \cdot i_{max} = N \cdot A_e \cdot B_{max} \rightarrow N = \frac{L \cdot i_{max}}{A_e \cdot B_{max}}$$

This number of turns ensures that the inductor will not be saturated

Core Losses Calculation

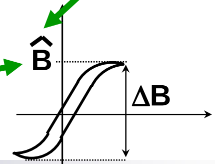
Material	Frequency	a	c	d
K MAGNETICS	f<500 kHz	0.0530	1.60	3.15
	500 kHz≤f<1 MHz	0.00113	2.19	3.10
	f≥1 MHz	$1.77 \cdot 10^{-9}$	4.13	2.98



Steinmetz Equation

FORMAT: $P_L = af^cB^d$

P_L in mW/cm^3, B in kG, f in kHz



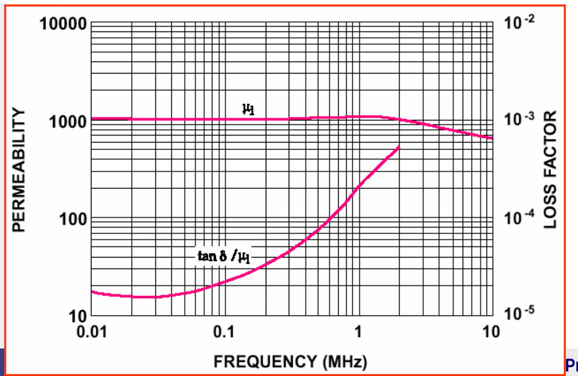
Core Losses Calculation

CORE LOSS vs PEAK AC FLUX DENSITY

FORMULA: $CL(mW/cm^3) = \dfrac{f}{\dfrac{a}{B^3} + \dfrac{b}{B^{2.3}} + \dfrac{c}{B^{1.65}}} + (d\, f^2 B^2)$

MICROMETALS IRON POWDER CORES

Material	a	b	c	d
-2	4.0×10^9	3.0×10^8	2.7×10^6	8.0×10^{-15}
-8	1.9×10^9	2.0×10^8	9.0×10^5	2.5×10^{-14}
-14	4.0×10^9	3.0×10^8	2.7×10^6	1.6×10^{-14}
-18	8.0×10^8	1.7×10^8	9.0×10^5	3.1×10^{-14}
-26	1.0×10^9	1.1×10^8	1.9×10^6	1.9×10^{-13}



Steward

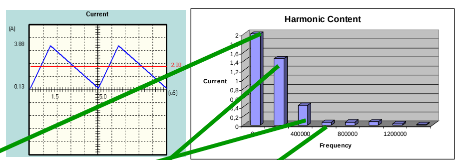
Winding Losses Calculation

1. R_{DC}

$$P = I^2_{rms} * R_{DC}$$

$$R_{DC} = \rho \cdot \frac{\text{length}}{\text{Area}}$$

2. R_{AC}



$$P = I^2_{DC} * R_{DC} + I^2_{rms\_1} * R_{AC\_1} + + I^2_{rms\_2} * R_{AC\_2} + I^2_{rms\_3} * R_{AC\_3} + \ldots$$

Winding Losses Calculation

$$P = I^2_{DC} {*} R_{DC} + I^2_{rms\_1} {*} R_{AC\_1} + $$
$$+ I^2_{rms\_2} {*} R_{AC\_2} + I^2_{rms\_3} {*} R_{AC\_3} + \ldots$$

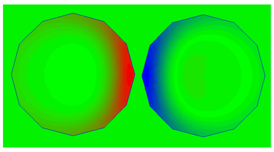
1. Only Skin

$$\delta = \frac{1}{\sqrt{\pi f \mu_0 \sigma}}$$

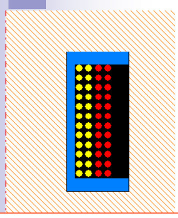
2. Analytical (i.e. Dowell)

$$K_r = \frac{R_{AC}}{R_{DC}} = 0.5y\left[M(y) + (2m-1)^2 \cdot D(y)\right]$$

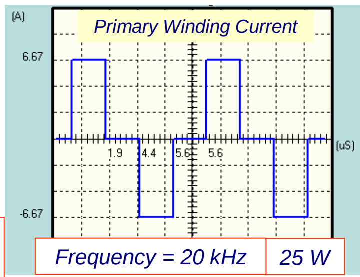
3. FEA



Winding Losses Calculation. Example 1.



RM7/I
AWG 26
(2 skin depths)
n1=n2=12 (2 parallel)



$$P = I^2_{rms} * R_{DC}$$

R_{DC} → *1.506 W*

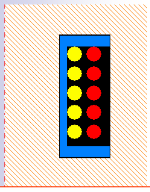
$$P = I^2_{DC} {*} R_{DC} + I^2_{rms\_1} {*} R_{AC\_1} + $$
$$+ I^2_{rms\_2} {*} R_{AC\_2} + I^2_{rms\_3} {*} R_{AC\_3} + \ldots$$

Skin → *1.529 W*

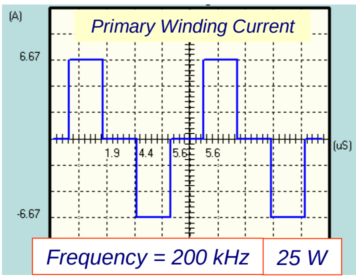
Dowell → *1.527 W*

128 harmonics

Winding Losses Calculation. Example 2.



RM10/I
AWG 15
(10 skin depths)
n1=n2=5



$P = I^2_{rms} * R_{DC}$

R_{DC} → *149.6 mW*

$$P = I^2_{DC} {}^* R_{DC} + I^2_{rms\_1} {}^* R_{AC\_1} + I^2_{rms\_2} {}^* R_{AC\_2} + I^2_{rms\_3} {}^* R_{AC\_3} + \dots$$

Skin → *464.9 mW*

Dowell → *750.5 mW*

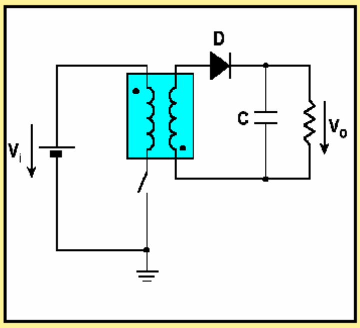
128 harmonics

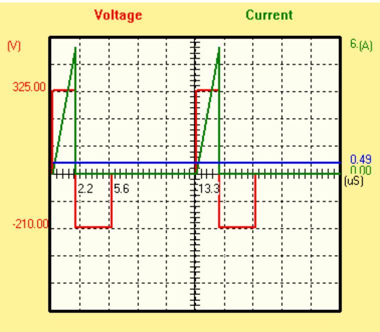
Winding Losses Calculation. Example 3

Input Voltage
325 V

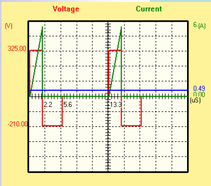
Switching Frequency
75 kHz

Output Values
Voltage: 42 V
Power: 160 W
Current: 3.81 (A)
Load: 11.03 (ohm)

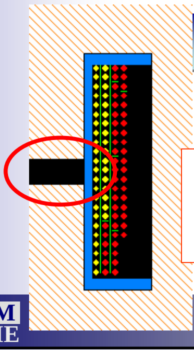




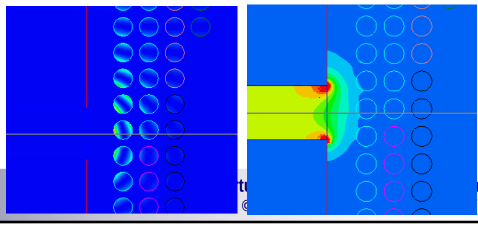
Winding Losses Calculation. Example 3



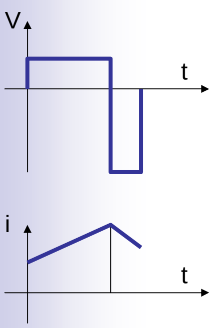
Method		
R_{DC}	Windings:	457.225 (mW)
	Temperature Rise (ºC)	6.69
Skin	Windings:	777.438 (mW)
	Temperature Rise (ºC)	10.19
FEA	Windings:	38.375 (W)
	Temperature Rise (ºC)	423.14

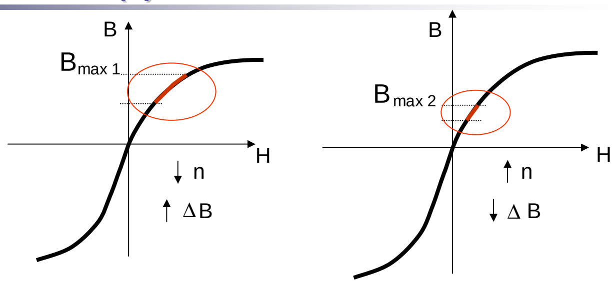


ETD39
AWG 20
n1= 40
n2=8



Optimization (I)



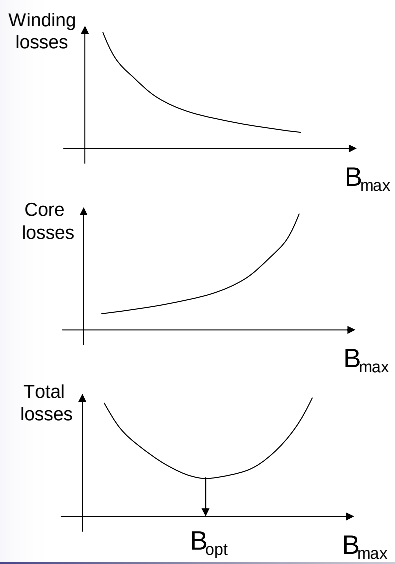


$$n{\cdot}\Delta B = \frac{V{\cdot}\Delta t}{Area} \Rightarrow n{\cdot}\Delta B = Constant$$

$$n{\cdot}B_{max} = \frac{L{\cdot}i_{max}}{Area} \Rightarrow n{\cdot}B_{max} = Constant$$

$$\uparrow B_{max} \Rightarrow \begin{Bmatrix} \downarrow n \\ \uparrow \Delta B \end{Bmatrix} \Rightarrow \begin{cases} \downarrow \text{Copper Losses} \\ \uparrow \text{Core Losses} \end{cases}$$

Optimization (II)

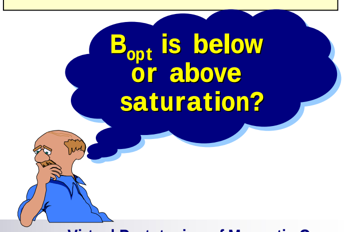


$\uparrow B_{max} \Rightarrow \downarrow n$

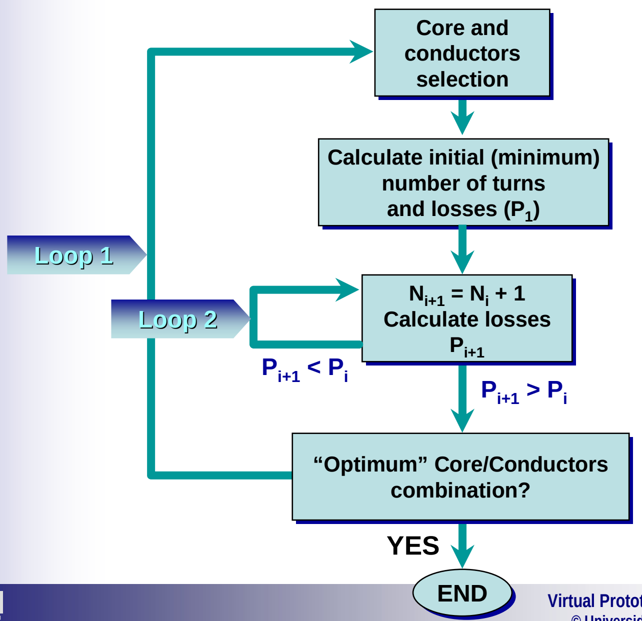
$P = I^2_{rms} \cdot n \cdot \rho \cdot \dfrac{length}{Area}$

$\uparrow B_{max} \Rightarrow \uparrow \Delta B$

$$P = V_e \cdot k \cdot f^{\alpha} \cdot \left(\frac{\Delta B}{2}\right)^{\beta}$$



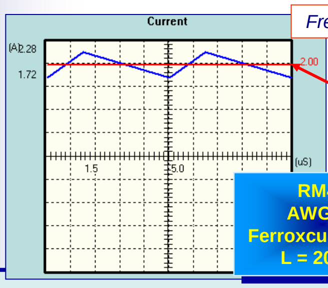
Optimization (III)



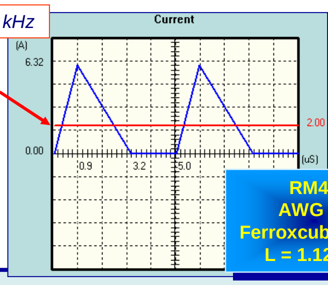
Optimization Example. Inductor

Frequency = 200 kHz

Iavg = 2 A



RM4/I
AWG 24
Ferroxcube 3F35
L = 20 μH



RM4/I
AWG 18
Ferroxcube 3F35
L = 1.12 μH

Bmax = 65% Bsat		Bmax = 65% Bsat	
Min n = 13	Core loss = 0.538 mW Wire loss = 105.2 mW Total loss = 105.76 mW	Min n = 2	Core loss = 21.3 mW Wire loss = 12.7 mW Total loss = 34 mW
n = 14	Core loss = 0.444 mW Wire loss = 114.1 mW Total loss = 114.51 mW	n = 3	Core loss = 7.4 mW Wire loss = 19.1 mW Total loss = 26.5 mW

Inductors

Calculation of the total reluctance

$$L = \frac{N^2}{\Re_T} = N^2 \cdot A_L \Rightarrow \Re_T = \frac{N^2}{L}$$

$$A_L = \frac{1}{\Re_T}$$

- $\Re_T$ is the total reluctance of the magnetic circuit.
- This reluctance may be obtained using:
  - ✓ Ferrite core + air gap
  - ✓ A material with distributed gap (low permeability) like iron powder

Inductors

Calculation of the air gap length

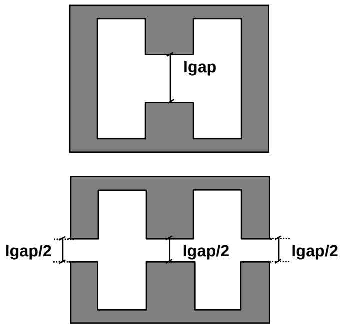
$$\Re_T = \Re_{ferrite} + \Re_{air} = \frac{1}{\mu_o \mu_e} \cdot \frac{l_e}{A_e} + \frac{1}{\mu_o} \cdot \frac{l_{gap}}{A_e}$$

$1/A_L$
(ferrite without gap)

$$\Re_T = \frac{1}{A_L} + \frac{1}{\mu_o} \cdot \frac{l_{gap}}{A_e}$$

$$\frac{1}{A_{L_T}} = \frac{1}{A_L} + \frac{l_{gap}}{\mu_o \times A_e}$$

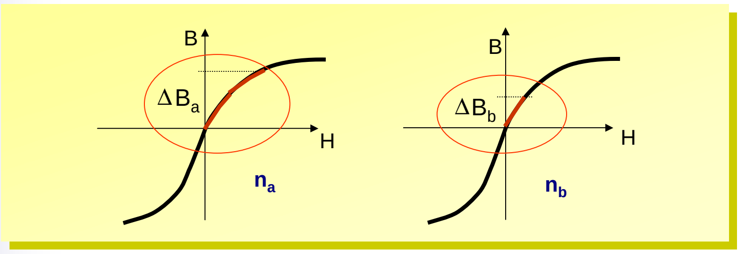
$$l_{gap} = \mu_o A_e \left[\frac{1}{A_{L_T}} - \frac{1}{A_L} \right]$$



Transformers

Number of turns

Number of turns calculated to keep maximum flux density or core losses under an appropriate value

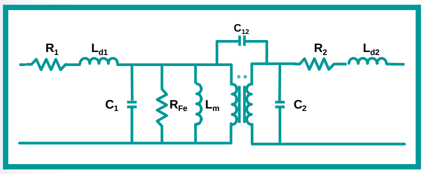


$$V_1 = N_1 \cdot \frac{\Delta\Phi}{\Delta t} = N_1 \cdot A_e \cdot \frac{\Delta B}{\Delta t}$$

$$N_1 = \frac{V_1 \Delta t}{A_e \Delta B}$$

$$a = \frac{N_1}{N_2} \Rightarrow N_2 = \frac{N_1}{a}$$

Modeling



$$\frac{\partial E(z,t)}{\partial z} = \mu \frac{\partial H(z,t)}{\partial t}$$

$$\frac{\partial H(z,t)}{\partial z} = \sigma \cdot E(z,t) + \varepsilon \frac{\partial E(z,t)}{\partial t}$$

Modeling for Virtual Prototyping?

Do I make simulations?

Do I trust them?

if not, are magnetic component models the reason?

Model

...tool to obtain MY OWN design guidelines

...gate to simulation

Modeling steps

- Physical description (equations)
- Equivalent Electrical model
- Parameters extraction
- Model solving

Physics

Physical description (equations)

Maxwell equations

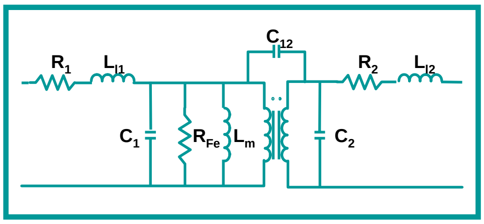
$$\nabla \cdot \boldsymbol{D} = \chi \qquad \nabla x \boldsymbol{E} = -\mu \frac{\partial \boldsymbol{H}}{\partial t}$$

$$\nabla \cdot \boldsymbol{B} = 0 \qquad \nabla x \boldsymbol{H} = \sigma \boldsymbol{E} + \varepsilon \frac{\partial \mathbf{E}}{\partial t}$$

Equivalent Circuit (I)

Equivalent Electrical model

Discrete elements



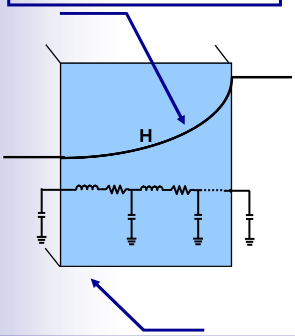
Frequency Independent (only for sinusoidal waveforms)

Equivalent Circuit (II)

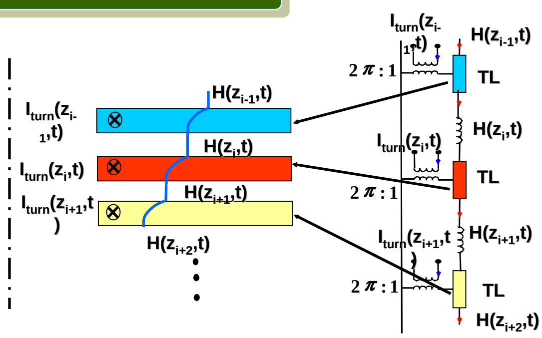
Equivalent Electrical model

Transmission lines based

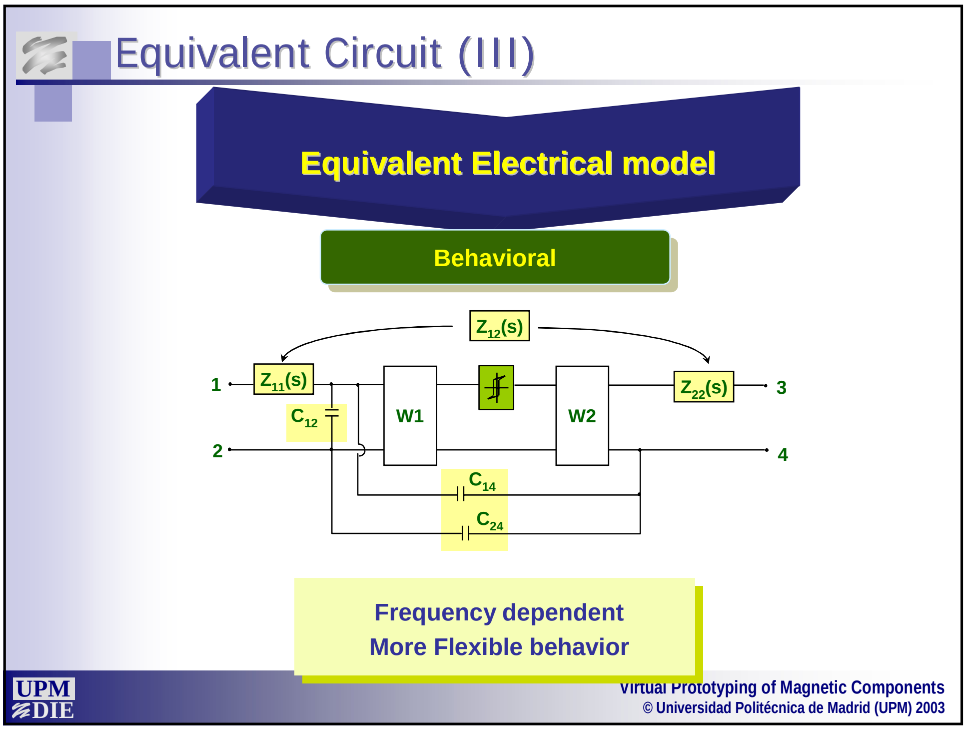
1D Maxwell equations



Transmission line equations



Frequency dependent
Magnetic/Electric Analogy based


Equivalent Circuit (III)
Equivalent Electrical model
Behavioral
Z12(s)
Z11(s)
Z22(s)
C12
W1
W2
C14
C24
1
2
3
4
Frequency dependent
More Flexible behavior
UPM
DIE
Virtual Prototyping of Magnetic Components
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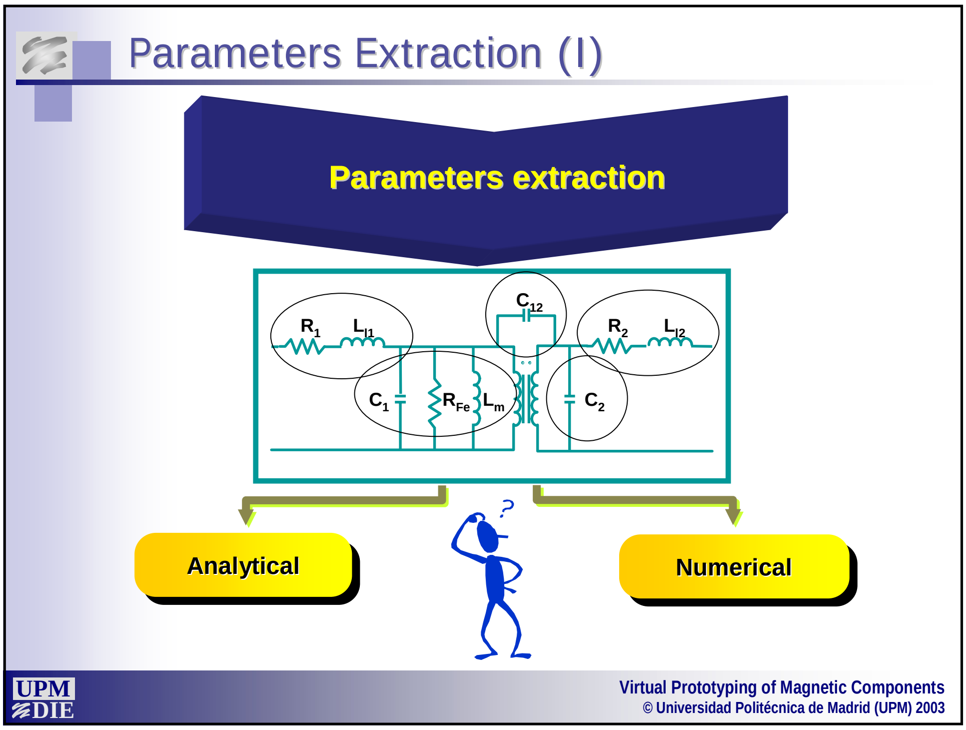
Parameters Extraction (I)
Parameters extraction
R1
Ll1
C12
R2
Ll2
C1
RFe
Lm
C2
Analytical
Numerical
?
UPM
DIE
Virtual Prototyping of Magnetic Components
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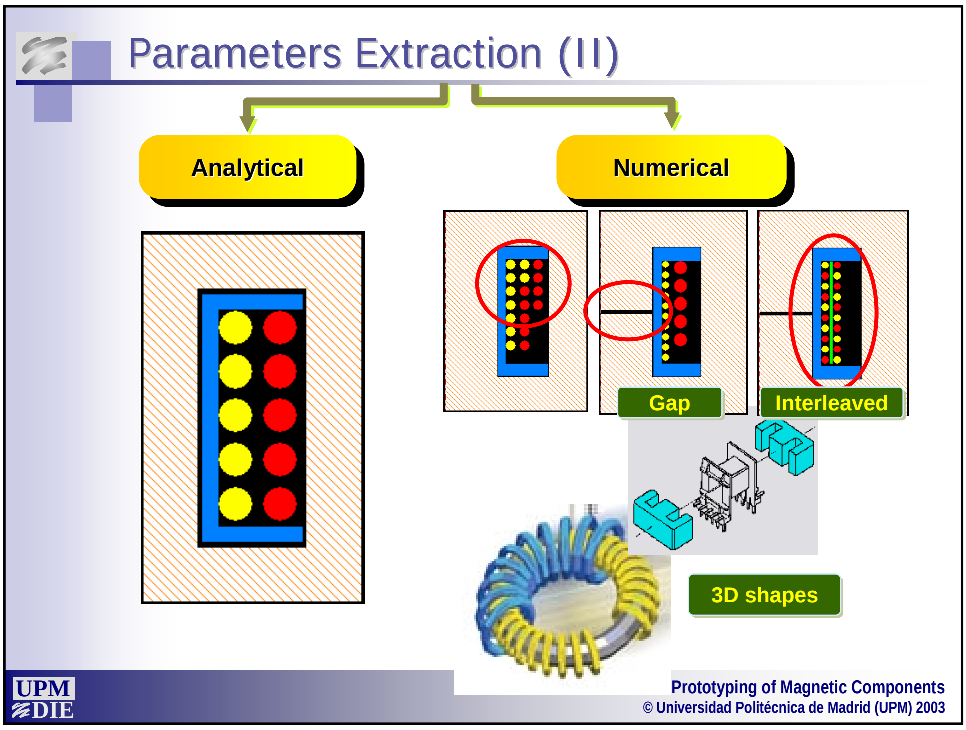
Parameters Extraction (II)
Analytical
Numerical
Gap
Interleaved
3D shapes
UPM
DIE
Prototyping of Magnetic Components
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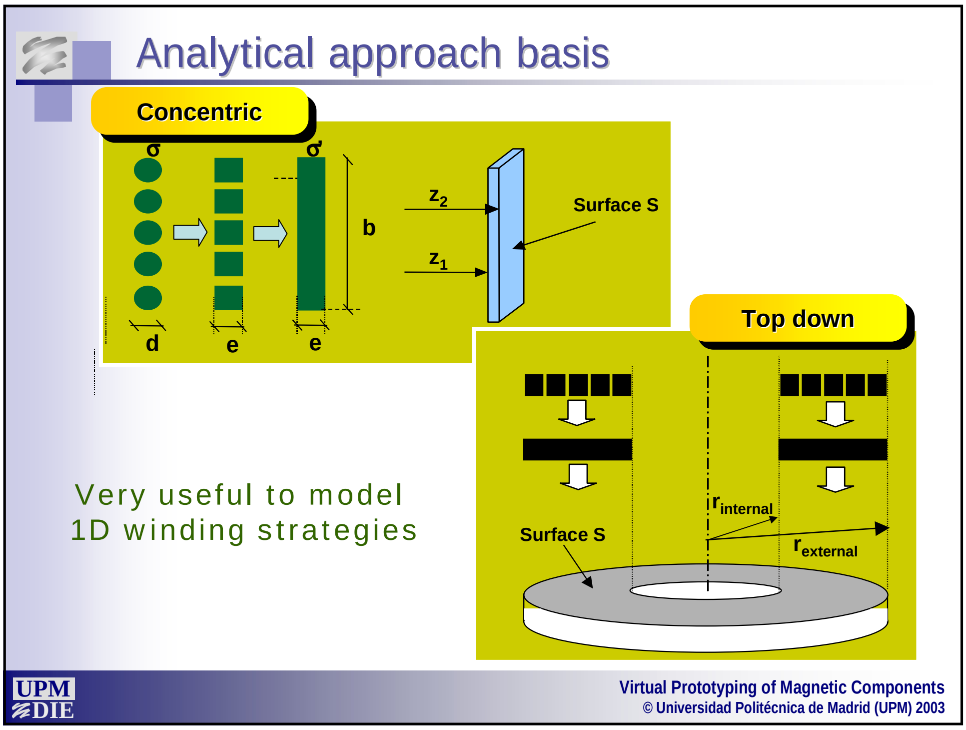
Analytical approach basis
Concentric
σ
σ'
b
d
e
e
z2
z1
Surface S
Top down
Surface S
r internal
r external
Very useful to model
1D winding strategies
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DIE
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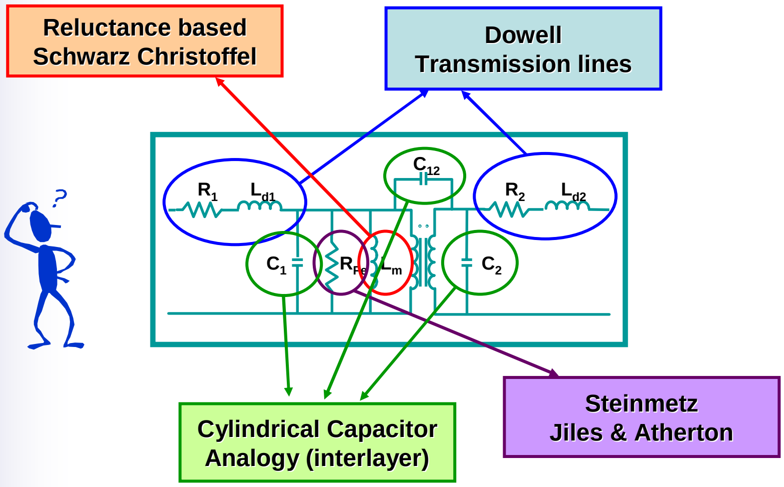
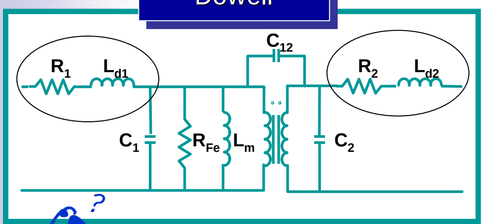
Analytical approach possibilities



Analytical approach examples (I)

Dowell



$$K_r = \frac{R_{AC}}{R_{DC}} = 0.5y\left[M(y) + (2m-1)^2 \cdot D(y)\right]$$

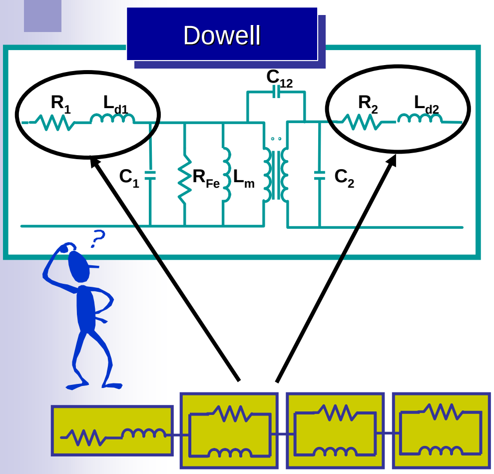
$$y = \frac{h_C}{\delta}$$

h_C = Layer thickness
δ = Skin depth
m = Number of layers of the winding sector

$$M(y) = \frac{\sinh(y) + \sin(y)}{\cosh(y) - \cos(y)} \qquad D(y) = \frac{\sinh(y) - \sin(y)}{\cosh(y) + \cos(y)}$$

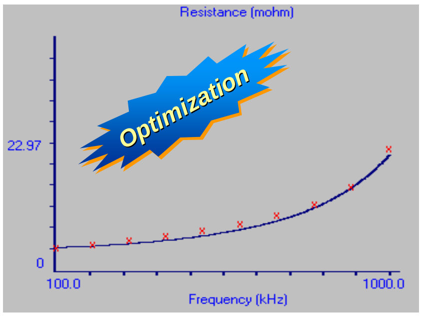
Calculation valid for a single frequency (sinusoidal)

Analytical approach examples (II)



$$Z(s) = c_0 + c_1 s + \sum_{i=1}^{NPF} \frac{c_{2i} s}{s + c_{2i-1}}$$

Foster Network



Calculation valid for the whole frequency range

Analytical approach examples (III)

Transmission line based parameters extraction

1D Maxwell equations

$$\frac{\partial E(z,t)}{\partial z} = \mu \frac{\partial H(z,t)}{\partial t}$$

$$\frac{\partial H(z,t)}{\partial z} = \sigma \cdot E(z,t) + \varepsilon \frac{\partial E(z,t)}{\partial t}$$

Transmission line equations

$$\frac{\partial V(z,t)}{\partial z} = R \cdot I(z,t) + L \frac{\partial I(z,t)}{dt}$$

$$\frac{\partial I(z,t)}{\partial z} = G \cdot V(z,t) + C \frac{\partial V(z,t)}{dt}$$

$$C = \varepsilon \cdot S$$

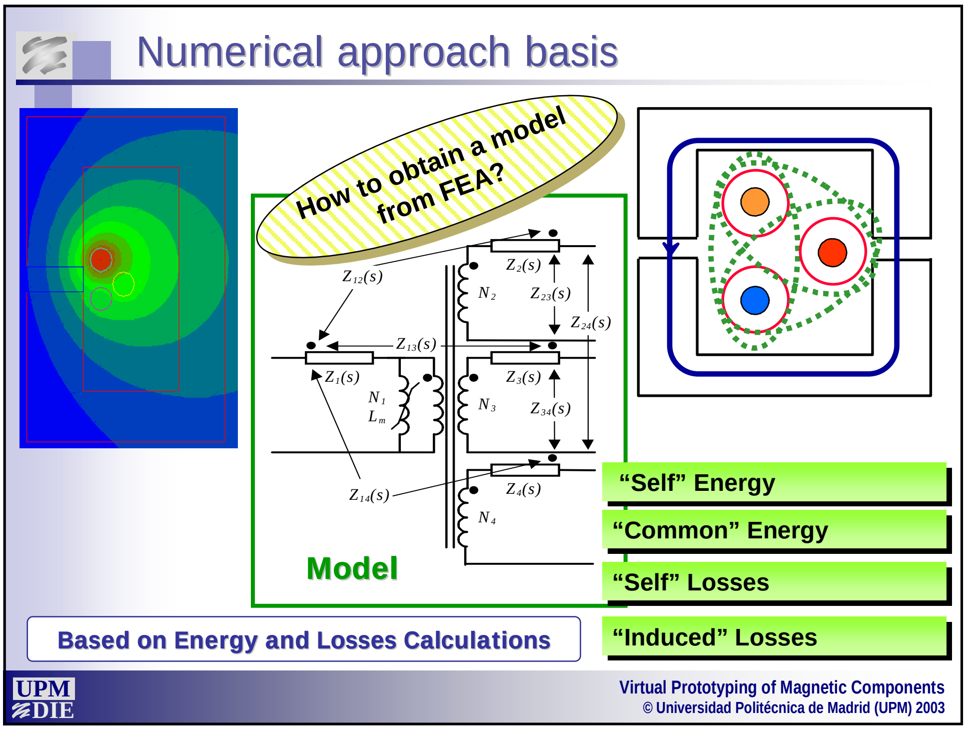
Electric energy

$$G = \sigma \cdot S$$

Losses

$$L = \frac{\mu}{S}$$

Magnetic energy



Numerical approach basis
How to obtain a model from FEA?
Model
"Self" Energy
"Common" Energy
"Self" Losses
"Induced" Losses
Based on Energy and Losses Calculations
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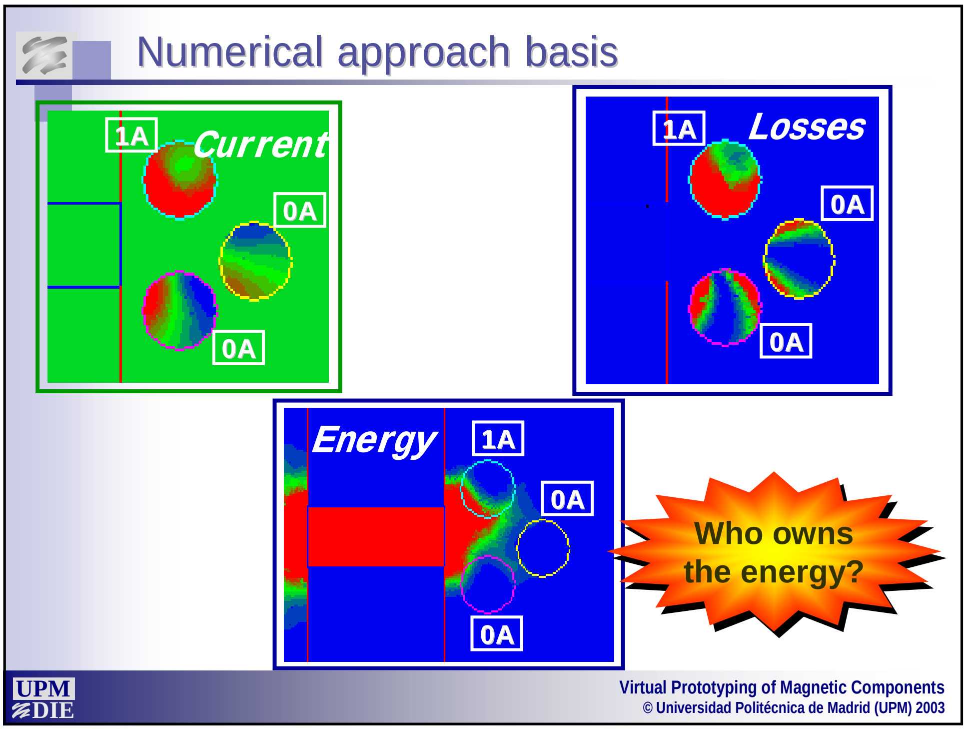

Numerical approach basis
1A
Current
0A
0A
1A
Losses
0A
0A
Energy
1A
0A
0A
Who owns the energy?
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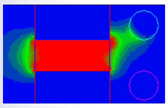
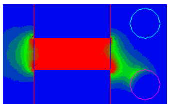
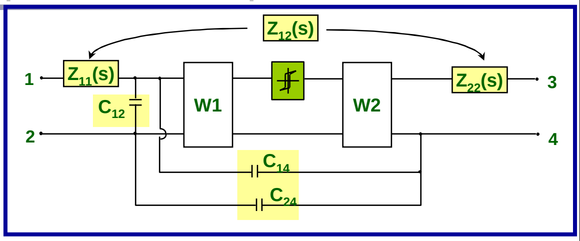
Numerical approach example



Test 1



Test 2



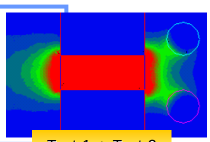
Z_{11}, Z_{22}:

✓Energy and losses due to I_1 OR I_2 INDEPENDENTLY

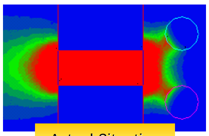
$$Z(s) = c_0 + c_1 s + \sum_{i=1}^{NPF} \frac{c_{2i} s}{s + c_{2i-1}}$$

Z_{12}: **Coupling?**

✓Energy and losses due to I_1 AND I_2 SIMULTANEOUSLY

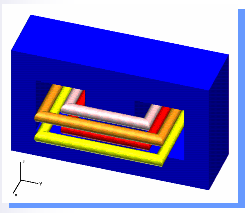


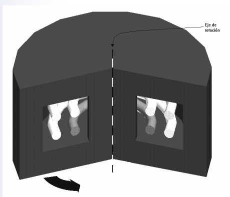
Test 1 + Test 2



Actual Situation

Numerical approach application (I)

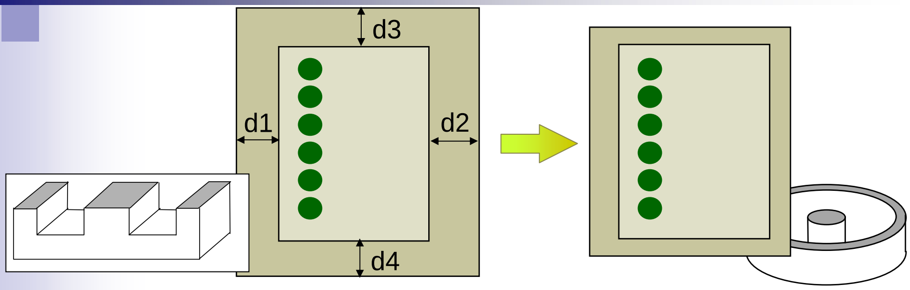




Energy and losses should be the same

- Energy in the core
- Energy in the window
- Energy in the air gap
- Losses in the core
- Losses in the conductors

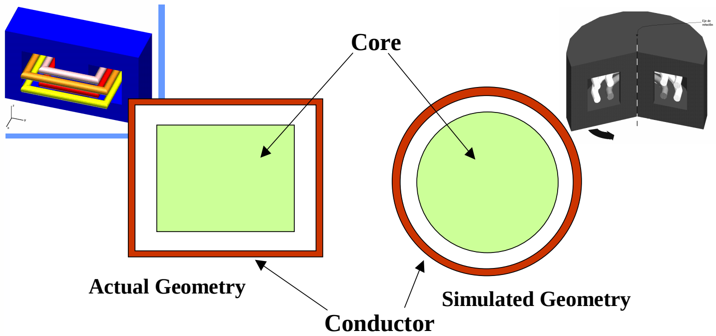
Numerical approach application (II)



- **Window Height and Window Width** are the same in both structures
- d1 is modified in order to obtain the same **central leg area** in both structures
- d2 is modified in order to obtain the same **external leg area** in both structures
- d3 is modified in order to obtain the same **core volume** in both structures

Numerical approach application (III)

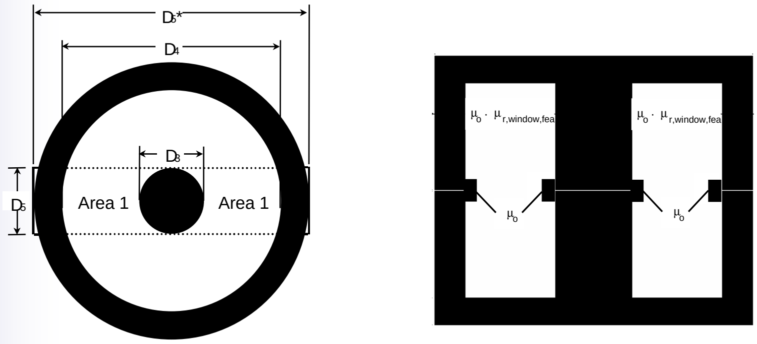
Modification of properties: Conductors Resistance



$$R_{model} = \frac{length_{actual}}{length_{model}} R_{actual}$$

Numerical approach application (IV)

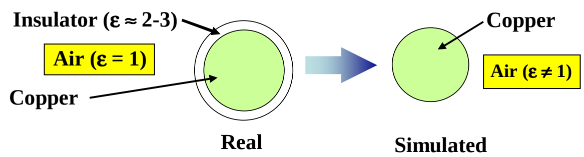
Modification of properties: Window Air Permeability



$$\mu_{r,window,fea} \cdot \mu_0 = \frac{Area\ \ 1}{Area\ \ 2} \cdot \mu_0$$

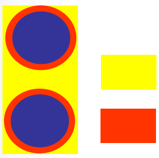
Numerical approach application (V)

Modification of properties: Window Air Permittivity



$$\varepsilon_{model} = \frac{\varepsilon_{isolator} \cdot Area_{wire\ insolator} + \varepsilon_{air} \cdot Area_{air}}{Area_{wire\ insolator} + Area_{air}}$$

$\varepsilon_{isolator} = 2.5$ **(Suggested)**

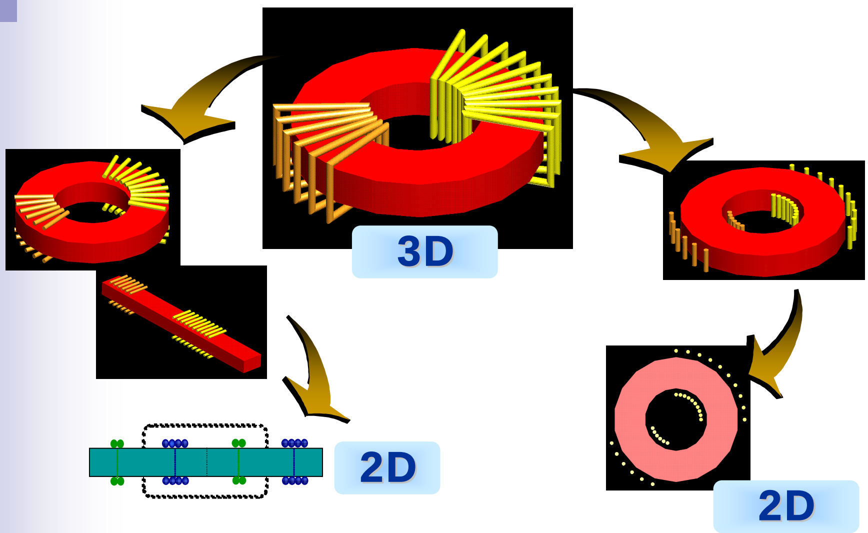


Air subwinding area

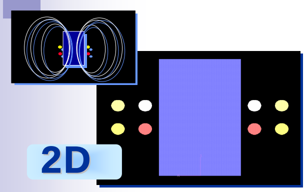
Isolator area

Considering each subwinding area

Double 2D concept: Application to Toroids

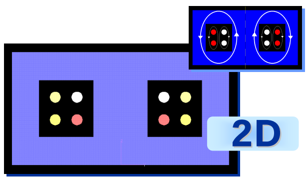


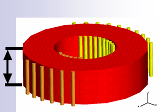
Double 2D concept



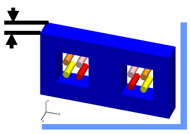
Magnetic Energy

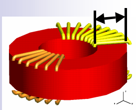
Losses



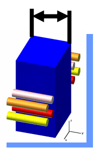


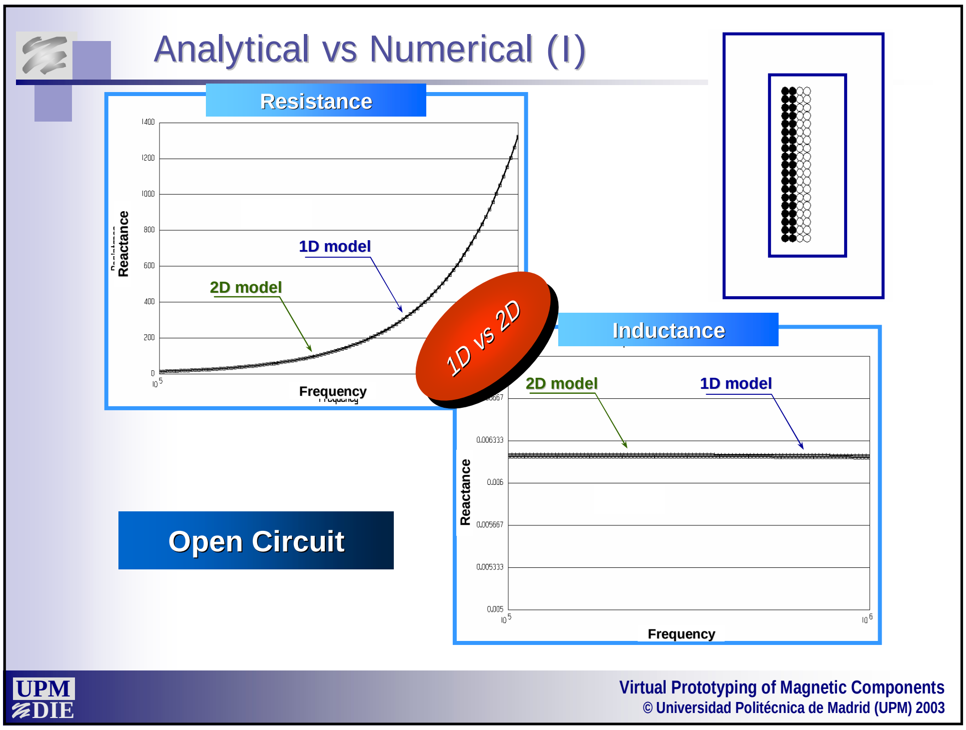
$$E_m = \frac{1}{2}\left[C_S \oint B_S \cdot H_S^* dS + C_D \oint B_D \cdot H_D^* dS\right]$$





$$P_d = C_S \oint_{A_h} \frac{J_S \cdot J_S^*}{\sigma} dS + C_D \oint_{A_h} \frac{J_D \cdot J_D^*}{\sigma} dS$$




Analytical vs Numerical (I)
Resistance
Reactance
1D model
2D model
Frequency
1D vs 2D
Inductance
2D model
1D model
Reactance
Frequency
Open Circuit
Virtual Prototyping of Magnetic Components
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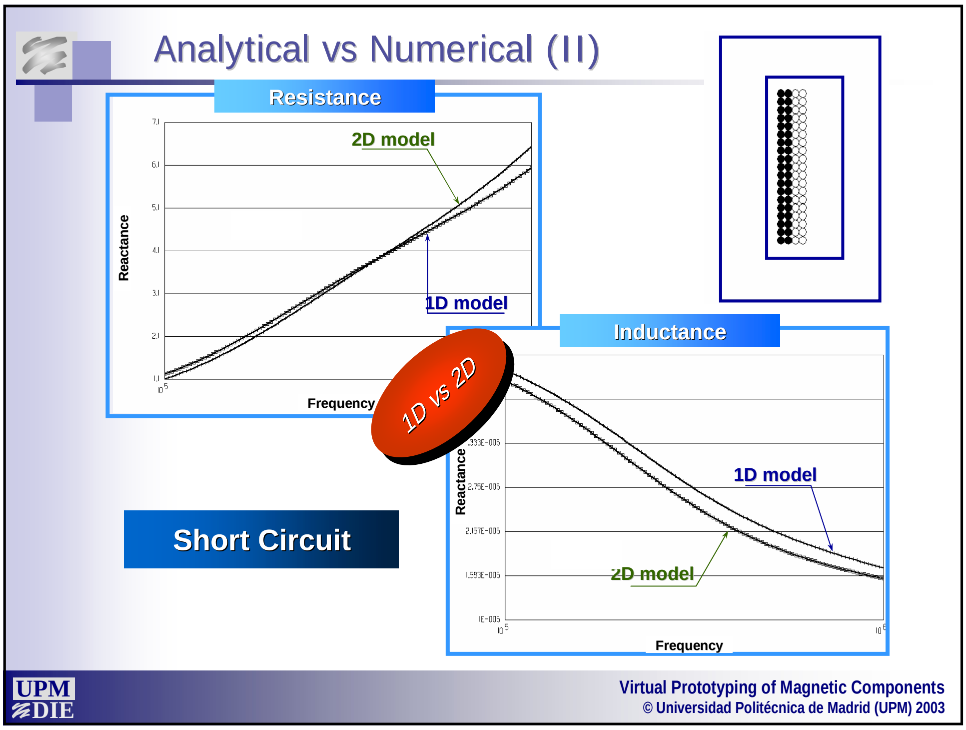
Analytical vs Numerical (II)
Resistance
Reactance
2D model
1D model
Frequency
1D vs 2D
Inductance
Reactance
1D model
2D model
Frequency
Short Circuit
Virtual Prototyping of Magnetic Components
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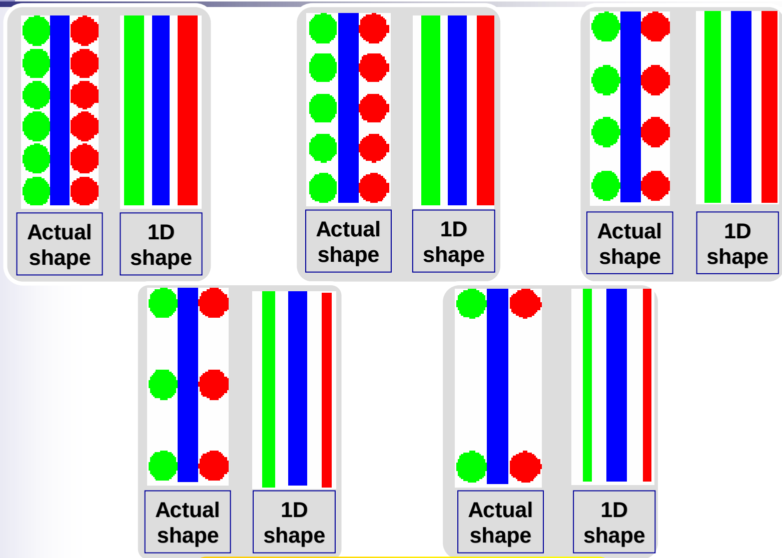
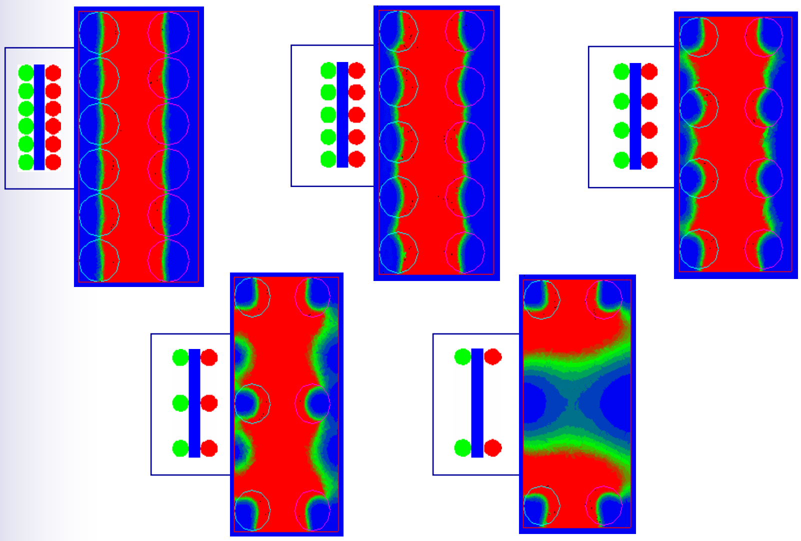
Analytical vs Numerical (III)

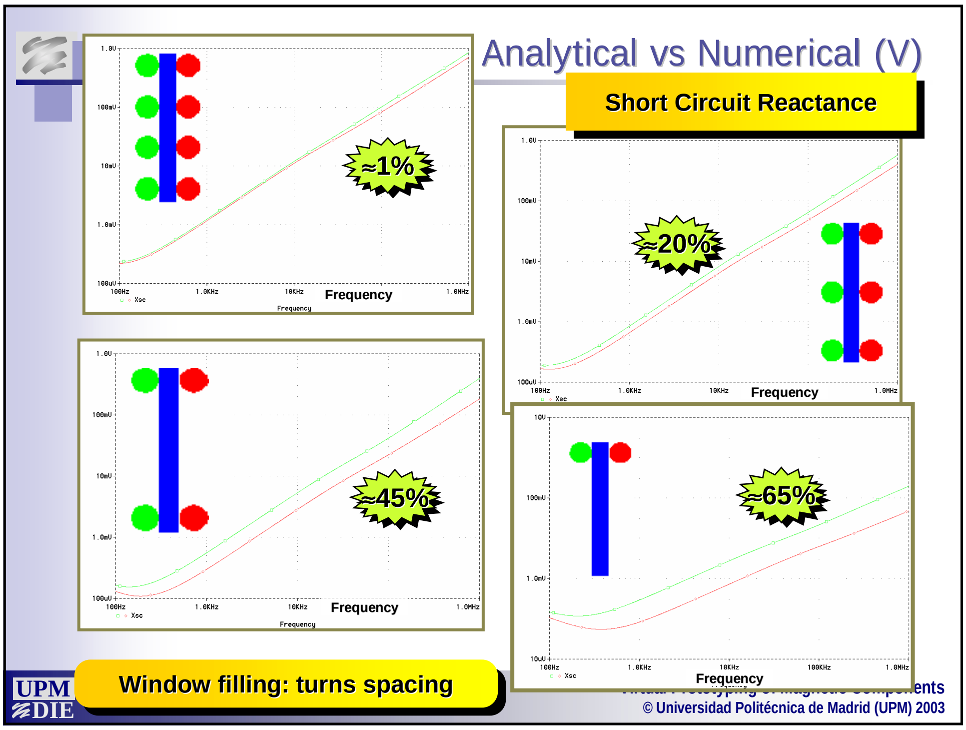


Window filling: turns spacing

Analytical vs Numerical (IV)



Window filling: turns spacing



Analytical vs Numerical (V)
Short Circuit Reactance
≈1%
≈20%
≈45%
≈65%
Frequency
Window filling: turns spacing
UPM
DIE
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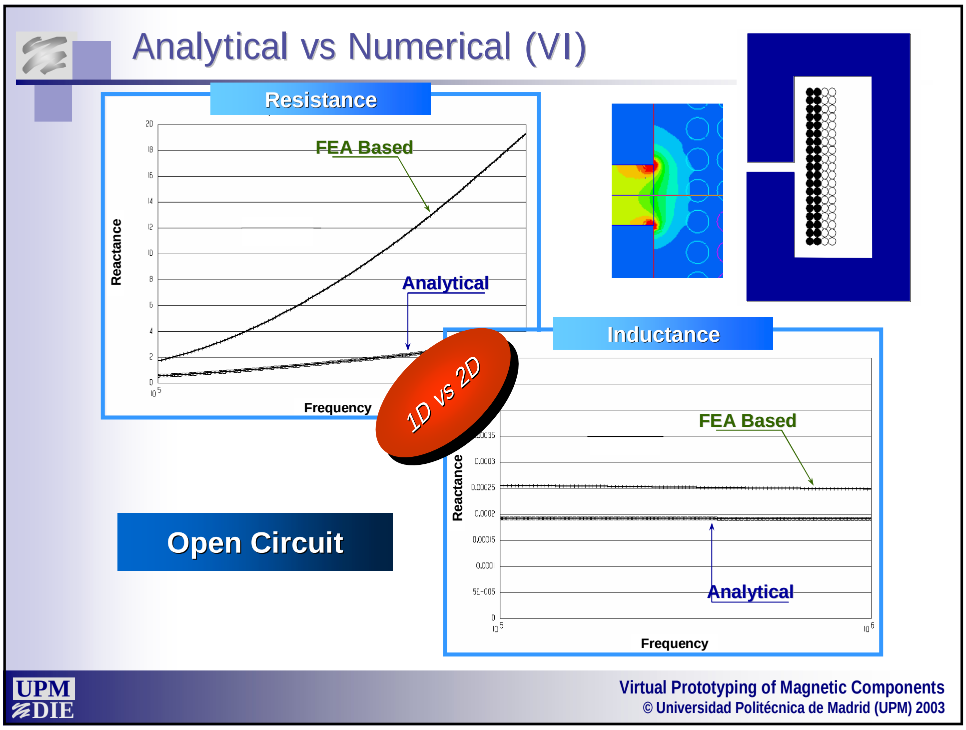

Analytical vs Numerical (VI)
Resistance
FEA Based
Analytical
Reactance
Frequency
1D vs 2D
Inductance
FEA Based
Analytical
Reactance
Frequency
Open Circuit
UPM
DIE
Virtual Prototyping of Magnetic Components
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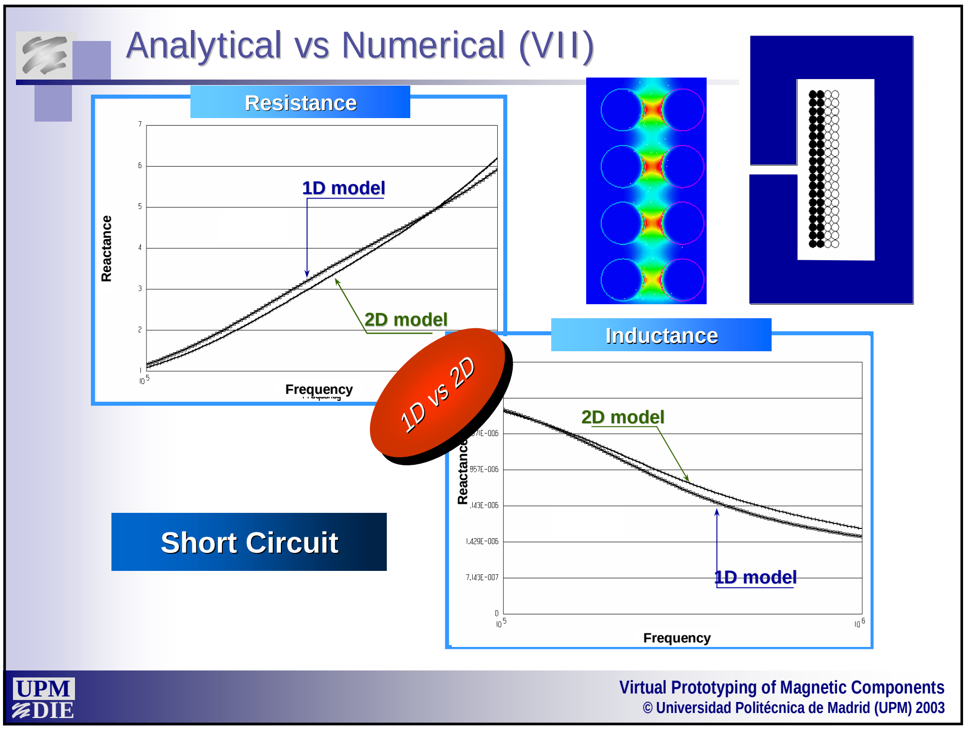
Analytical vs Numerical (VII)
Resistance
Reactance
1D model
2D model
Frequency
1D vs 2D
Inductance
2D model
Reactance
1D model
Frequency
Short Circuit
UPM
DIE
Virtual Prototyping of Magnetic Components
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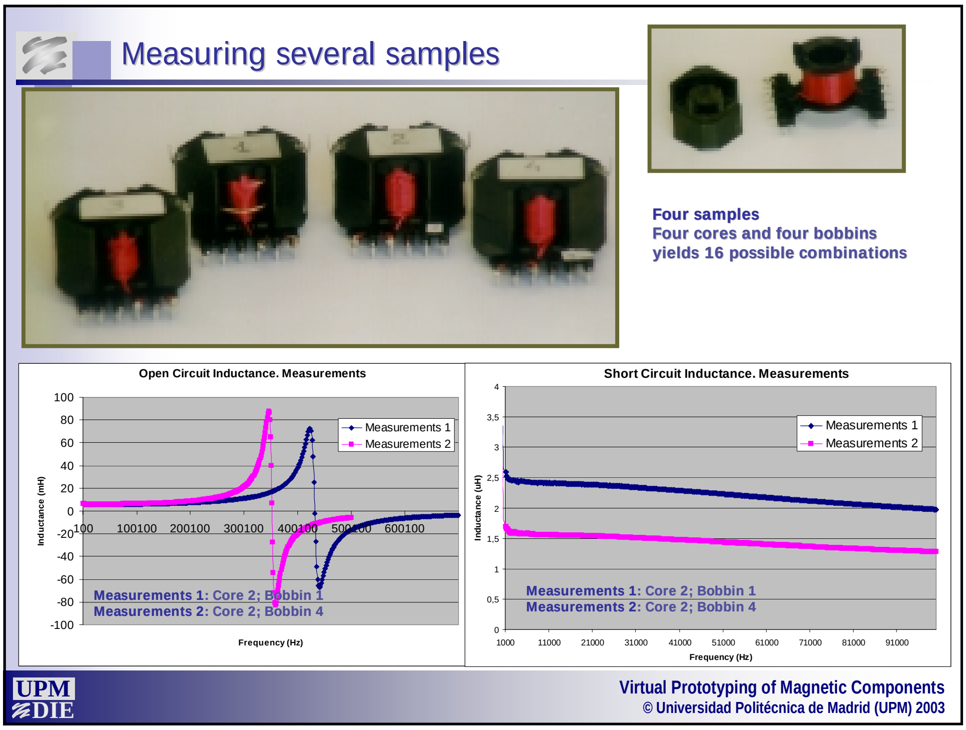
Measuring several samples
Four samples
Four cores and four bobbins
yields 16 possible combinations
Open Circuit Inductance. Measurements
Measurements 1
Measurements 2
Inductance (mH)
Frequency (Hz)
Measurements 1: Core 2; Bobbin 1
Measurements 2: Core 2; Bobbin 4
Short Circuit Inductance. Measurements
Measurements 1
Measurements 2
Inductance (uH)
Frequency (Hz)
Measurements 1: Core 2; Bobbin 1
Measurements 2: Core 2; Bobbin 4
UPM
DIE
Virtual Prototyping of Magnetic Components
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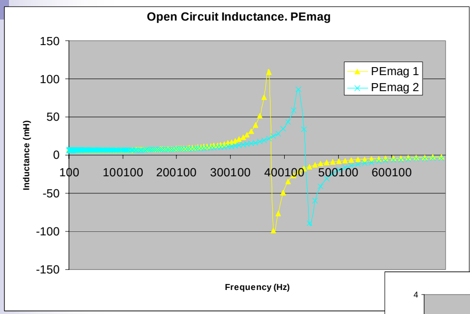
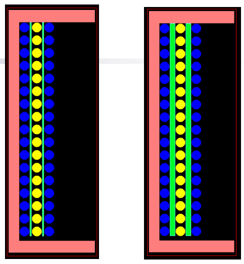
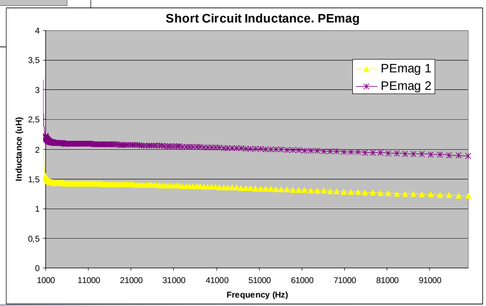
Generating the model

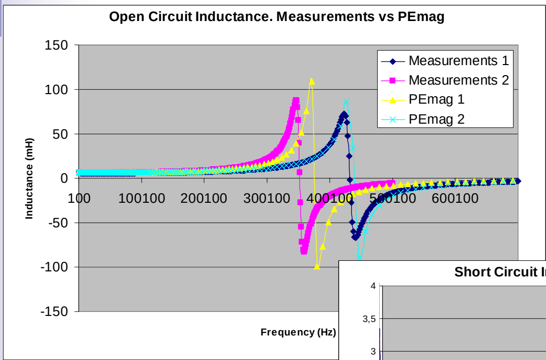


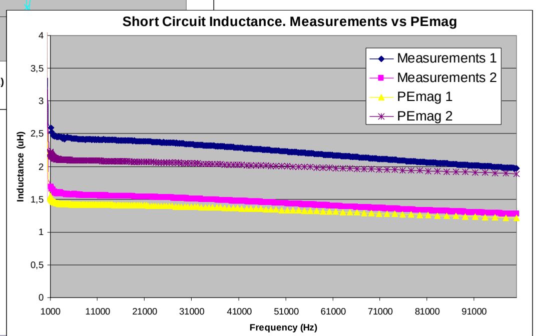


Case 1: separation 125 µm
Case 2: separation 375 µm



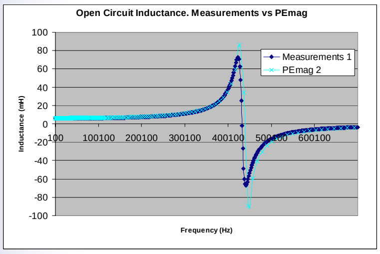
Comparing results

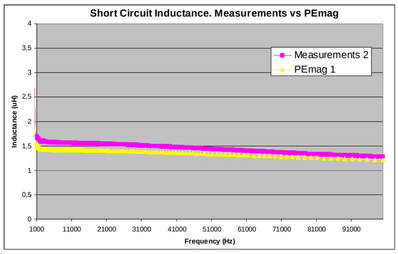




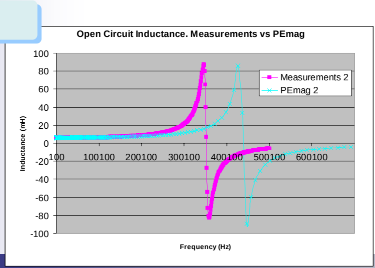
Comparing results

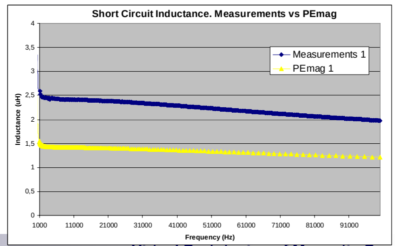
If only one real sample and only one model case are compared, results could be....





or....

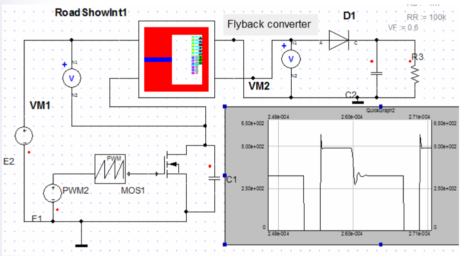


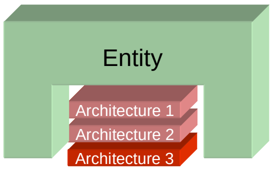


Seminar Contents

- Introduction
- Basic Concepts
- Design
- Modeling
- **Simulation**
- **Virtual Prototyping Examples**

Simulation

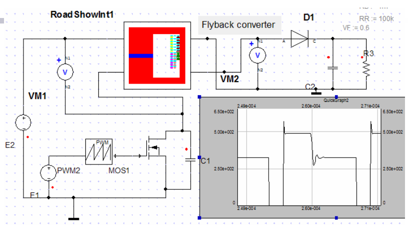





```
package electrical_system IS
        SUBTYPE voltage IS real;
        SUBTYPE current IS real;

        NATURE electrical IS
                voltage ACROSS
                current THROUGH
                ground  REFERENCE;
end package electrical_system;
```


Simulation features



Simulation Goals

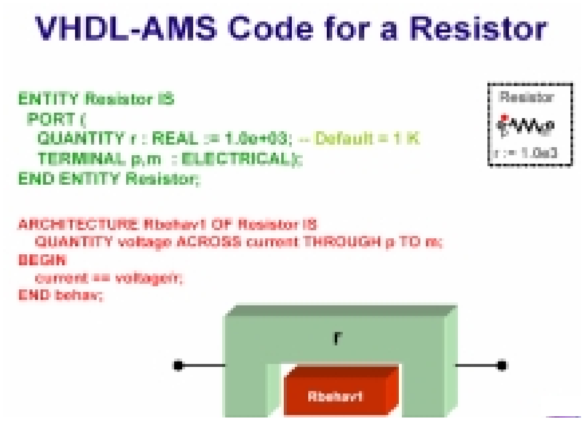
- Validation of the circuit operation
- Exploration of the waveforms
- Quantification of magnitudes (losses, stresses,…)
- Better understanding of the circuit operation

Simulation Challenges

- Lack of “accurate enough” models **(Magnetic, Semiconductors, Layout…)**
- Convergence problems (many “mathematical” parameters)
- Learning curve
- “Hard” transition from the design stage

Simulator types

- Circuit simulation (i.e. SPICE based)
- System simulation (VHDL-AMS, MAST, SMD,…)



Seminar Contents

- Introduction
- Basic Concepts
- Design
- Modeling
- Simulation
- **Virtual Prototyping Examples**
  - ✓ **Example 1: Coupling in Flyback transformer**
  - ✓ **Example 2: Coupling in Multi-winding transformers**
  - ✓ **Example 3: Multiphase Buck Inductors**
  - ✓ **Example 4: Flyback transformer**

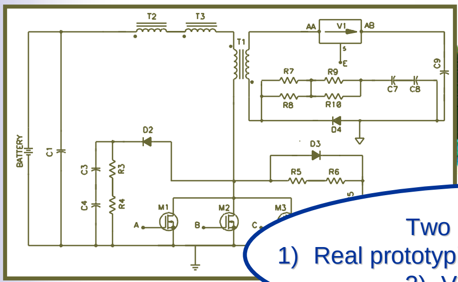
Model + Simulation

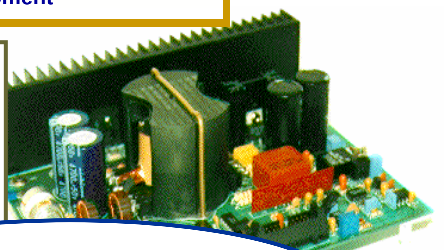
Analytical vs Numerical

Example 1: Coupling in Flyback Transformer

FLYBACK With Hysteretic Control

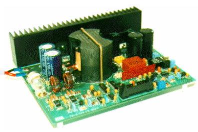
- Capacitive Load: 0.5 F
- Output Voltage: 0 to 340 V
- Output Power: up to 1kW
- Input Voltage: 16 to 32 V
- Switching Frequency: 1kHz to 40 kHz
- Application: X ray equipment





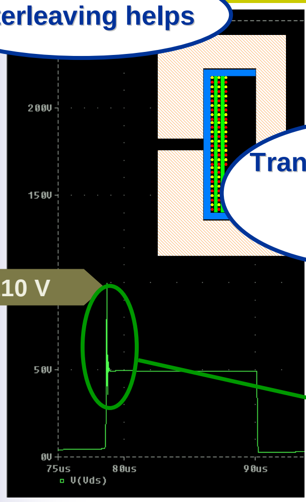
Two design methods:
1) Real prototyping: Design + Measurements
2) Virtual prototyping

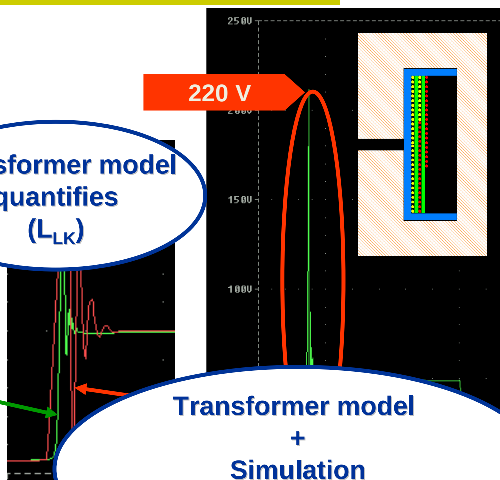
Example 1: Flyback Converter



Vds in MOSFET

Interleaving helps





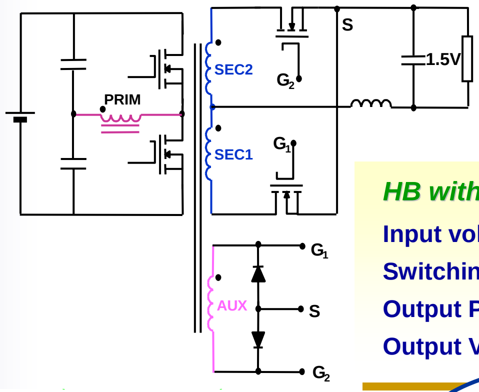
110 V

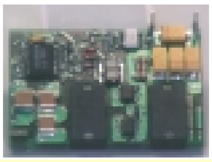
220 V

Transformer model quantifies (L_{LK})

Transformer model + Simulation provides the impact on the converter

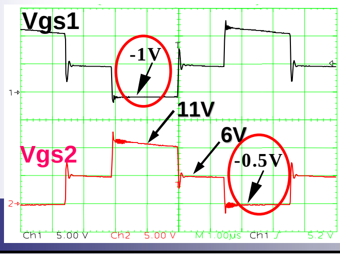
Example 2: Coupling in Multi-winding transformers





HB with SDSR

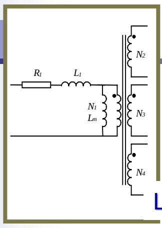
Input voltage:	**36 to 72 V**
Switching Frequency:	**150 kHz**
Output Power:	**15 W**
Output Voltage:	**1.5 V, 10 A**

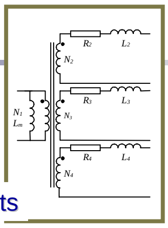


Two design methods:
1) Real prototyping
2) Virtual prototyping

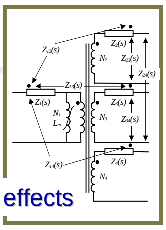
How to model?

Gate voltages with a proper coupling

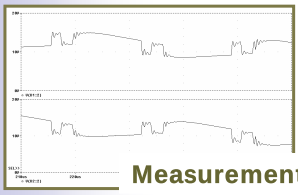


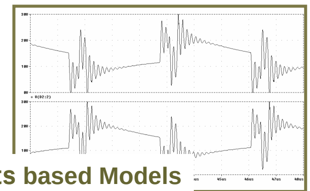


Lumped effects

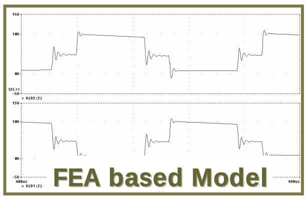


Distributed effects



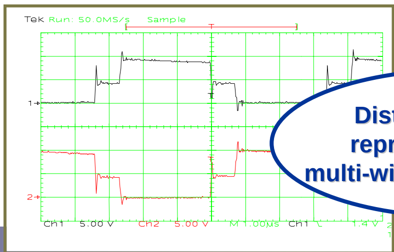


Measurements based Models

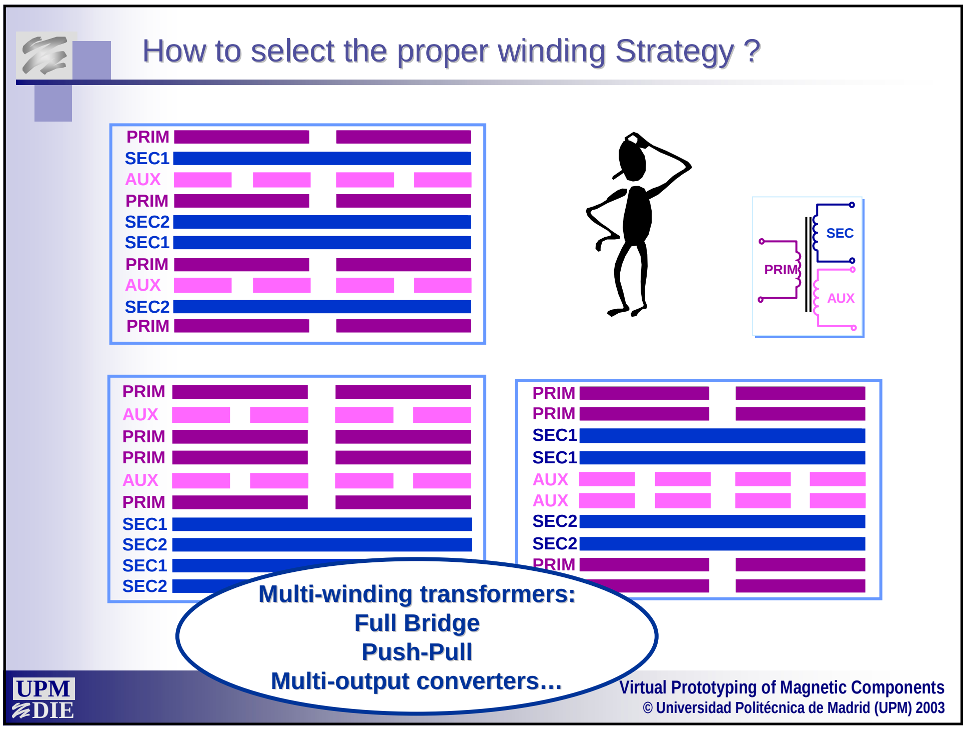


FEA based Model

Measurements



Distributed models represent better the multi-winding transformers


How to select the proper winding Strategy ?
PRIM
SEC1
AUX
PRIM
SEC2
SEC1
PRIM
AUX
SEC2
PRIM
SEC
PRIM
AUX
PRIM
AUX
PRIM
PRIM
AUX
PRIM
SEC1
SEC2
SEC1
SEC2
PRIM
PRIM
SEC1
SEC1
AUX
AUX
SEC2
SEC2
PRIM
Multi-winding transformers:
Full Bridge
Push-Pull
Multi-output converters…
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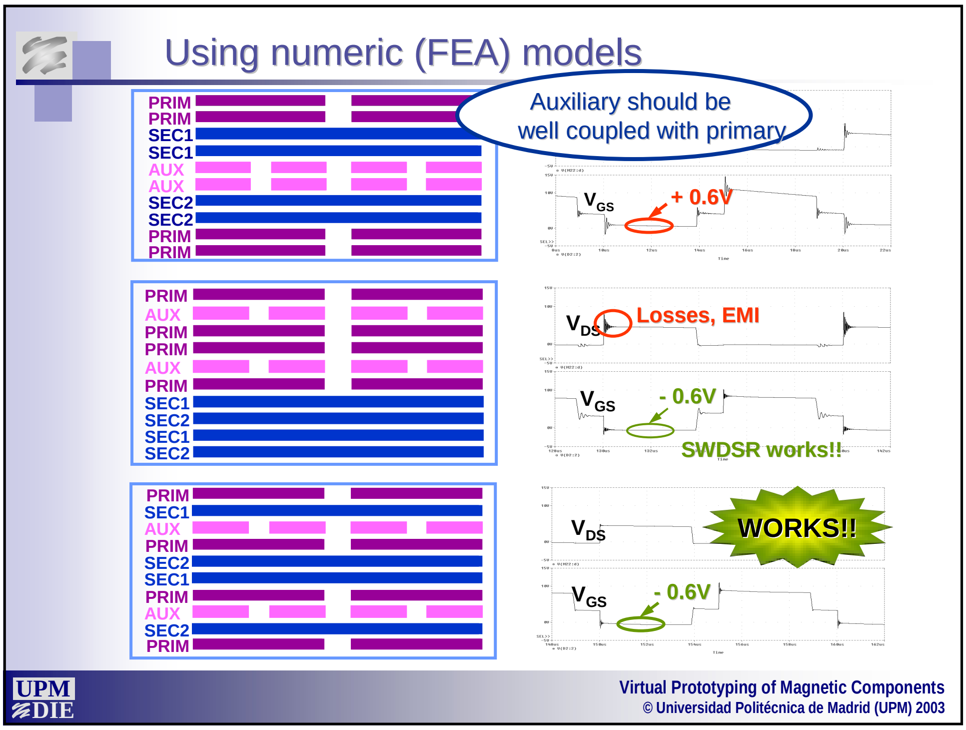
Using numeric (FEA) models
Auxiliary should be well coupled with primary
PRIM
PRIM
SEC1
SEC1
AUX
AUX
SEC2
SEC2
PRIM
PRIM
VGS
+ 0.6V
PRIM
AUX
PRIM
PRIM
AUX
PRIM
SEC1
SEC2
SEC1
SEC2
VDS
Losses, EMI
VGS
- 0.6V
SWDSR works!!
PRIM
SEC1
AUX
PRIM
SEC2
SEC1
PRIM
AUX
SEC2
PRIM
VDS
WORKS!!
VGS
- 0.6V
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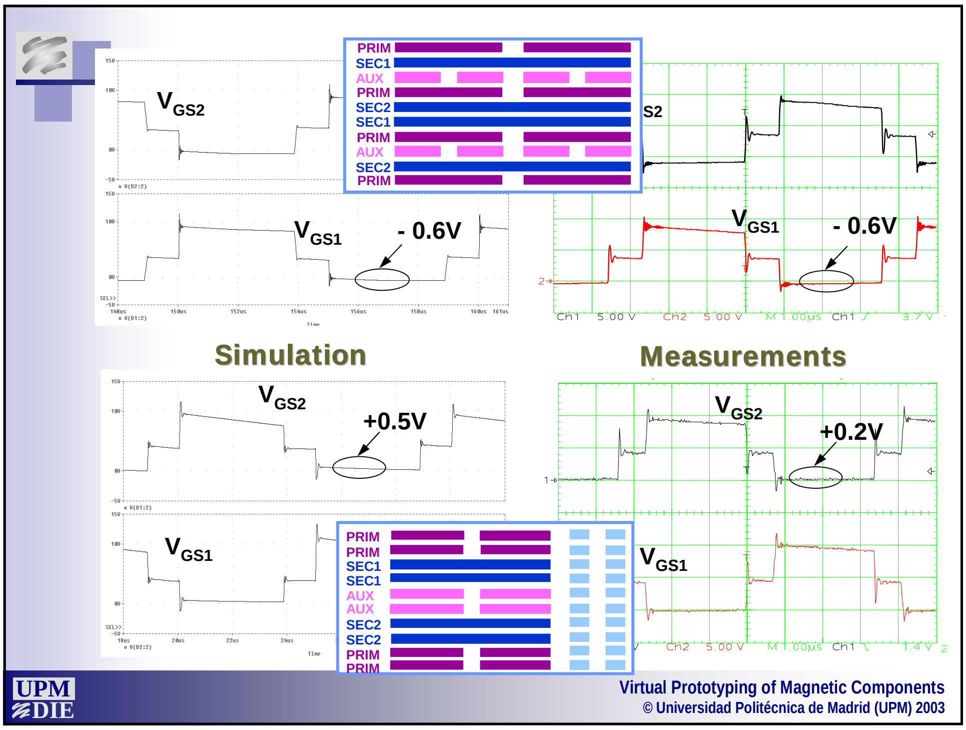
PRIM
SEC1
AUX
PRIM
SEC2
SEC1
PRIM
AUX
SEC2
PRIM
V_{GS2}
V_{GS1}
- 0.6V
- 0.6V
Simulation
Measurements
V_{GS2}
+0.5V
V_{GS2}
+0.2V
V_{GS1}
V_{GS1}
PRIM
PRIM
SEC1
SEC1
AUX
AUX
SEC2
SEC2
PRIM
PRIM
UPM DIE
Virtual Prototyping of Magnetic Components
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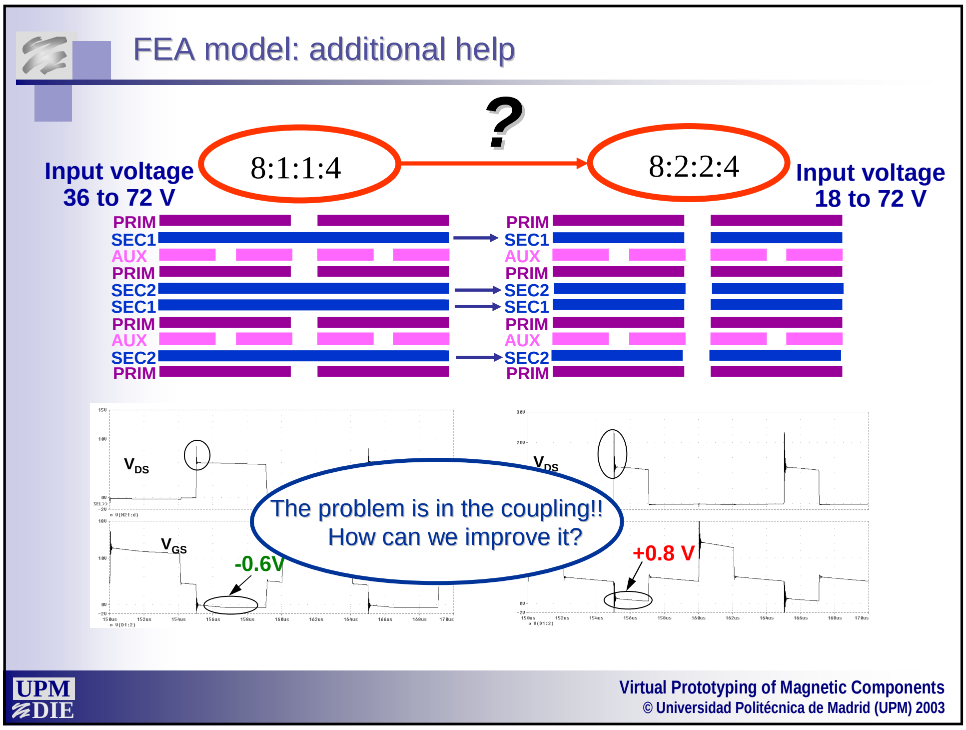
FEA model: additional help
?
Input voltage 36 to 72 V
8:1:1:4
8:2:2:4
Input voltage 18 to 72 V
PRIM
SEC1
AUX
PRIM
SEC2
SEC1
PRIM
AUX
SEC2
PRIM
V_{DS}
V_{GS}
-0.6V
V_{DS}
+0.8 V
The problem is in the coupling!!
How can we improve it?
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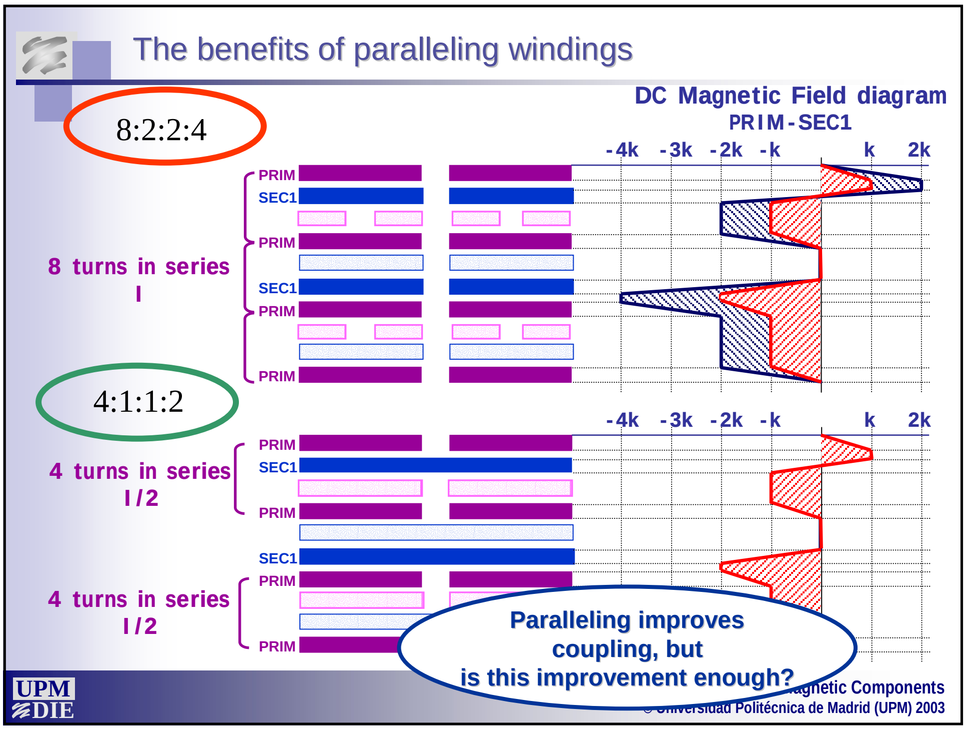
The benefits of paralleling windings
8:2:2:4
DC Magnetic Field diagram
PRIM-SEC1
-4k -3k -2k -k k 2k
8 turns in series
I
PRIM
SEC1
PRIM
SEC1
PRIM
PRIM
4:1:1:2
-4k -3k -2k -k k 2k
4 turns in series
I/2
PRIM
SEC1
PRIM
SEC1
PRIM
4 turns in series
I/2
PRIM
Paralleling improves
coupling, but
is this improvement enough?
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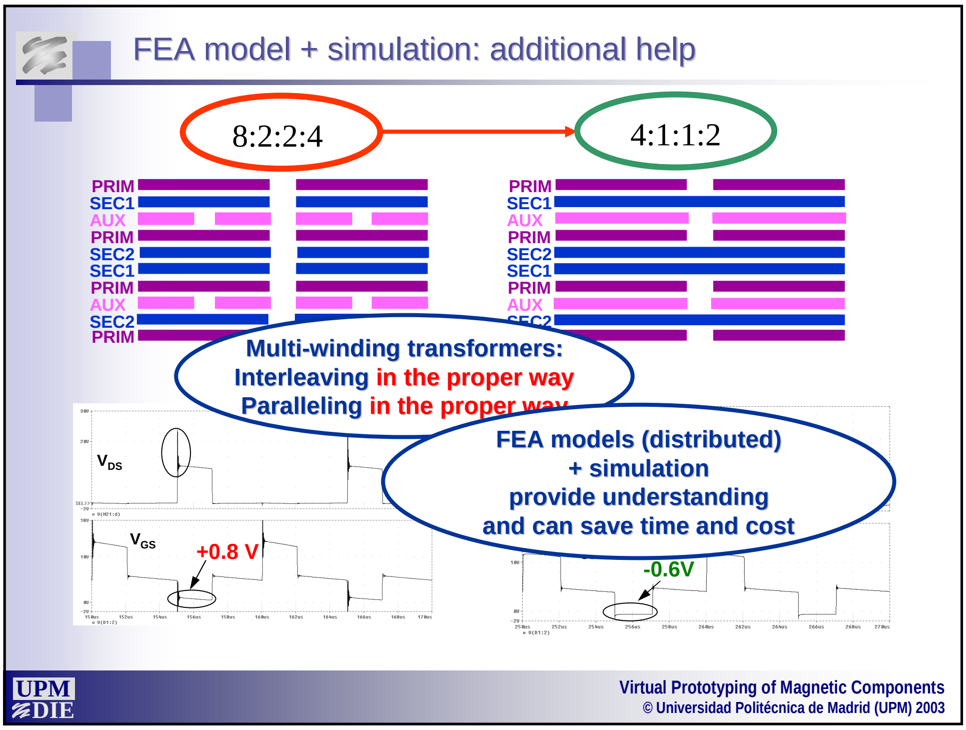
FEA model + simulation: additional help
8:2:2:4
4:1:1:2
PRIM
SEC1
AUX
PRIM
SEC2
SEC1
PRIM
AUX
SEC2
PRIM
PRIM
SEC1
AUX
PRIM
SEC2
SEC1
PRIM
AUX
SEC2
Multi-winding transformers:
Interleaving in the proper way
Paralleling in the proper way
V_{DS}
V_{GS}
+0.8 V
FEA models (distributed)
+ simulation
provide understanding
and can save time and cost
-0.6V
UPM
DIE
Virtual Prototyping of Magnetic Components
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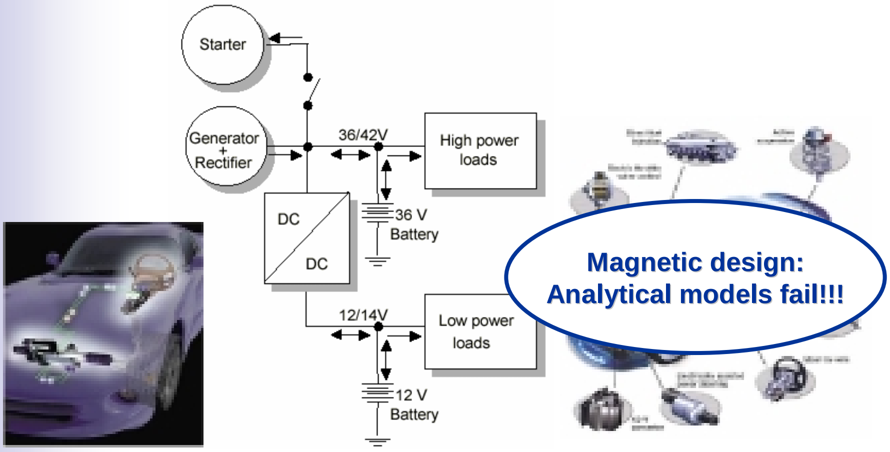
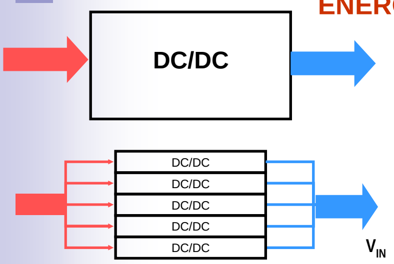
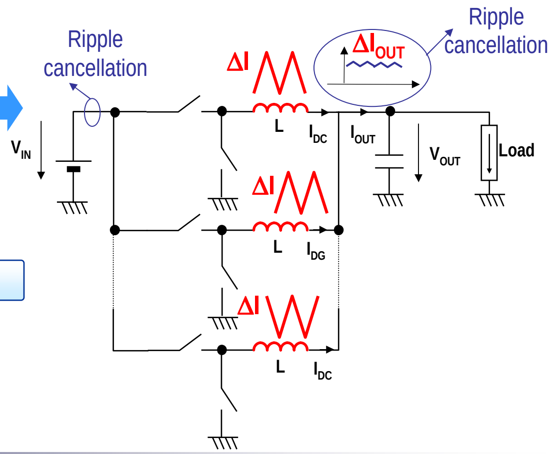
Example 3: Multiphase Buck Inductors Automotive Application



Multiphase converters



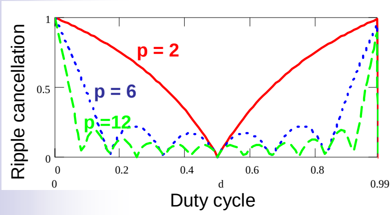
ENERGY is distributed ⇨ Distributed losses
Distributed current
Smaller ripple & filter



Applications

- ✓ VRMs (μprocessors)
- ✓ Automotive
- ✓ Portable systems

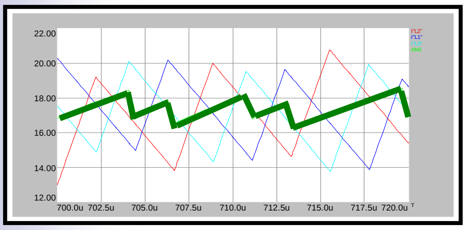
Multiphase converters



Ripple cancellation is determined by:

1. Number of phases: p
2. V_{OUT}/V_{IN} ratio (duty cycle): d

Low current ripple: small filter



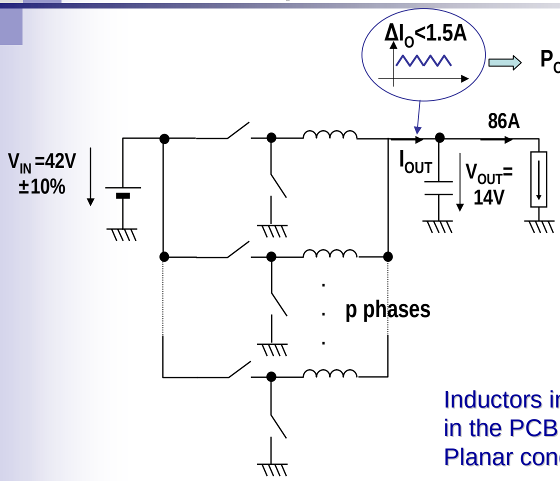
Design

- ✓ Number of phases: p
- ✓ Switching frequency: f_{SW}
- ✓ Inductance per phase: L
- ✓ Current ripple

Complex design with
a big number of combinations
(p inductors)

Converter specifications

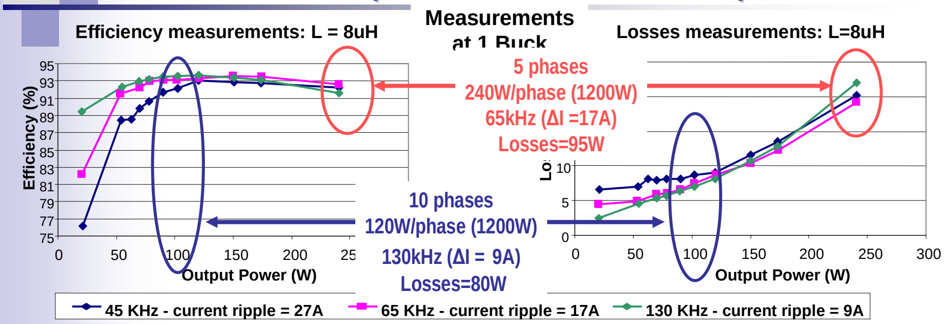
Automotive application



Inductors integrated
in the PCB:
Planar conductors

Magnetics are 40%-60%
of converter size

Which is the optimum number of phases?



- Depending on
  I_{DC}/phase, or Power/phase, or Number of phases
  the minimum losses are obtained at very different f_{sw}
- A high number of phases allows for a high f_{sw} with lower losses
  Smaller inductor size per phase but high number of phases ⇒ Size?

Trade-off: size - efficiency

Design of inductors: analytical

- **Losses calculation accounts for:**
  - ✓ DC resistance
  - ✓ Skin effect } → Winding losses
  - ✓ Core losses ($P=kf^{\alpha}B^{\beta}$) → Steinmetz
- Gap effect (2D) is not considered

Considered design options

Nº of phases	f_{sw} (kHz)	Cores	Material
6 10 16	100 200	RM6LP RM7LP RM8LP RM10LP RM12LP	3C90 3F3 3F4

Designs of inductors: 16 phases, L=5µH

Core	DC losses / inductor (mW)	AC losses / inductor (mW)	Overall Losses (W)	Overall Area (mm²)	Overall Height (mm)	Overall Volume (mm³)

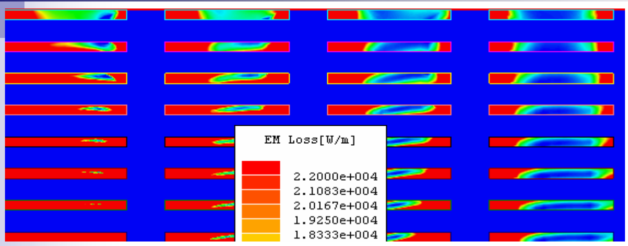
Total losses per inductor without considering AC effects

Total losses per inductor considering AC effects (AC + DC effects)

All inductors' total losses

All inductors' total dimensions

AC current distribution in planar conductors



4 turns
4 turns/layer
20 layers in parallel

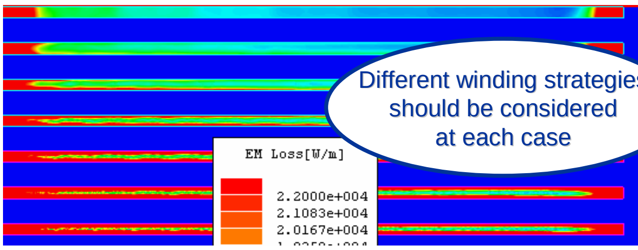
P_{DC} = 160mW

P_{AC} = 210mW

4 turns
1 turn/layer
5 layers in parallel

P_{DC} = 150mW

P_{AC} = 425mW



Different winding strategies should be considered at each case

Skin and proximity effects are stronger in wide planar conductors than in narrow ones

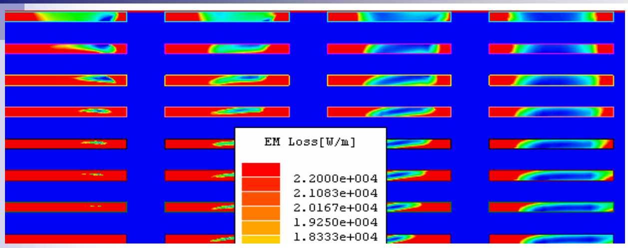
Designs of inductors: 16 phases, L=5µH

Core	DC losses / inductor (mW)	AC losses / inductor (mW)	Overall Losses (W)	Overall Area (mm^2)	Overall Height (mm)	Overall Volume (mm^3)
100kHz, I_{DC} = 5.4A, ΔI = 19A						
16*RM7LP	420	425	6.8	5,010	9.8	49,000
16*RM8LP	275	285	4.5	6,300	11.6	73,125
16*RM10LP	90	155	2.4	9,660	13	125,600
200kHz, I_{DC} = 5.4A, ΔI = 9.5A						
16*RM6LP	250	300	4.9	3,550	9	31,970
16*RM7LP	215	235	3.7	5,010	9.8	49,000
16*RM8LP	130	150	2.4	6,300	11.6	73,125

Material
Turns
Winding strategy
Gap

Hundreds of combinations!!

AC current distribution in planar conductors



4 turns
4 turns/layer
20 layers in parallel

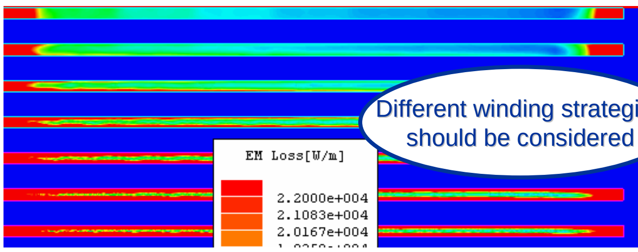
P_{DC} = 160mW

P_{AC} = 210mW

4 turns
1 turn/layer
5 layers in parallel

P_{DC} = 150mW

P_{AC} = 425mW



Different winding strategies should be considered

Skin and proximity effects are stronger in wide planar conductors than in narrow ones

Designs of inductors: 10 phases, L=8μH

Core	DC losses / inductor (mW)	AC losses / inductor (mW)	Overall Losses (W)	Overall Area (mm²)	Overall Height (mm)	Overall Volume (mm³)
100kHz, I_{DC} = 8.6A, ΔI = 11.5A						
10*RM7LP	Too many losses					
10*RM8LP	740	790	7.9	3,940	11.6	45,700
10*RM10LP	250	325	3.2	6,040	13	78,520
200kHz, I_{DC} = 8.6A, ΔI = 5.8A						
10*RM7LP	950mW (100ºC) Too many losses					
10*RM8LP	300	350	3.5	3,940	11.6	45,700
10*RM10LP	190	200	2	6,040	13	78,520

Designs of inductors: 6 phases, L=9.5μH

Core	DC losses / inductor (mW)	AC losses / inductor (mW)	Overall Losses (W)	Overall Area (mm²)	Overall Height (mm)	Overall Volume (mm³)
100kHz, I_{DC} = 14.4A, ΔI = 9.8A						
6*RM10LP	Too many losses					
6*RM12LP	680	710	4.3	5,610	16.8	94,298
200kHz, I_{DC} = 14.4A, ΔI = 4.9A						
6*RM10LP	735	760	4.5	3,625	13	47,110

Selected solutions based on analytical model

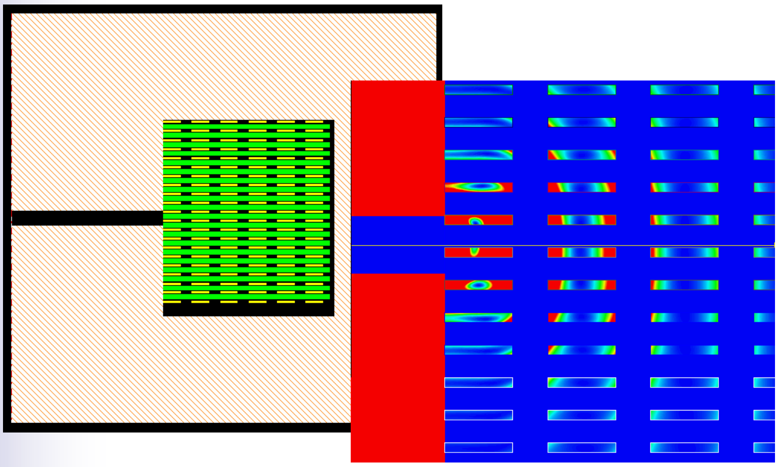
Solution	Overall Losses (w)	A (mm²)	h (mm)	V (mm³)
16 x RM7LP, 100kHz	6.8	5,010	9.8	49,000
16 x RM6LP, 200kHz	4.9	3,550	9	31,970
10 x RM8LP, 200kHz	3.5	3,940	11.6	45,700
6 x RM12LP, 100kHz	4.3	5,610	16.8	94,298
6 x RM10LP, 200kHz	4.5	3,625	13	47,110

- **Final selection depends on:**
  - ✓ Losses of the whole converter
  - ✓ Size of the whole converter.
    Inductor size has a strong influence on overall size.
  - ✓ Height requirements

Is the analytical model good enough for this case?

Automotive application: gap effect

How does the gap affect the current distribution and the losses of the selected solutions?



Automotive application: gap effect

Solution	Gap (µm)	ΔI /inductor (A)	Analytical model: Losses /inductor (W)	Analytical model: ΔT in inductor (ºC)	FEA model: Losses /inductor (W)	FEA model: ΔT in inductor (ºC)
16*RM7LP, 100kHz	380	19	0.32	45	1.7	180
16*RM6LP, 200kHz	200	9.5	0.3	40	0.87	135
10*RM8LP, 200kHz	310	6	0.35	39	1.5	175
6*RM12LP, 100kHz	410	10	0.71	40	1.9	115
6*RM10LP, 200kHz	440	5	0.35	39	1.5	175

All the selected solutions based on analytical models are NOT FEASIBLE due to the gap effect!!!

Design of inductors: numerical (FEA) model

- **It accounts for:**
  - ✓ DC resistance
  - ✓ Skin effect
  - ✓ Proximity effect
  - ✓ Gap effect
  - ✓ Core losses

Considered design options

Nº of phases	f_{sw} (kHz)	Cores	Material
6 10 16	100 125 150 200	RM6LP RM7LP RM8LP RM10LP RM12LP	3C90 3F3 3F4

Design of inductor: 16 phases, 200kHz, RM7LP

	L (µH)	ΔI / inductor (A)	ΔI_{OUT} (A)	Gap (µm)	Analytical Losses / inductor (W)	Numerical Losses / inductor (W)
Material	2.5	19	1.2	170	0.28	2.2
Turns	3.6	13	0.8	215	0.22	2.0
Winding strategy	6.2	7.4	0.5	190	0.21	0.85
	9.4	4.9	0.3	260	0.33	0.63
	12.6	3.7	0.2	250	0.33	0.76

In this case, the optimum gap-ΔI combination is obtained for a small ΔI_{OUT} (0.3A)

The analytical model gives a very high ΔI = 13A

This analysis is repeated for each selected case…

Feasible solutions: Numerical Model

5000 combinations analyzed!!!

From the point of view of the inductors:

Solution	ΔI_{OUT} (A)	Overall Losses (W)	Overall Area (mm²)	Overall Height (mm)	Overall Volume (mm³)
16 phases, RM7LP, 200kHz	0.3	10.1	5,010	9.8	49,000
16 phases, RM8LP, 125kHz	0.3	7.5	6,340	11.6	73,125
10 phases, RM10LP, 150kHz	1.2	6.9	6,040	13	78,520
6 phases, RM12LP, 125kHz	0.8	9	5,610	16.8	94,298

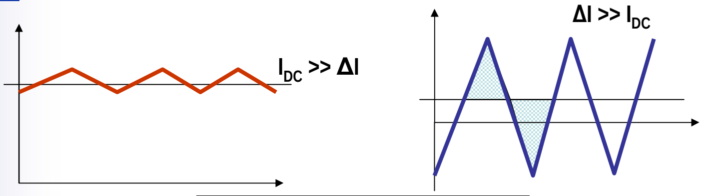
ΔI_{OUT} < 1.5A

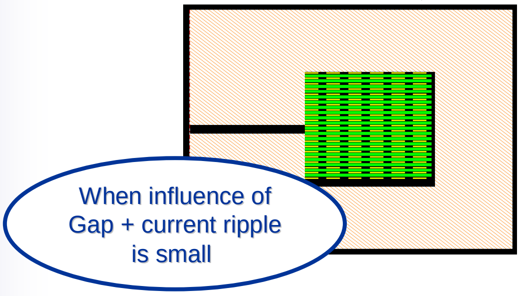
- Virtual prototyping has allowed:
  - ✓ To study 5000 cases, learning what is happening
  - ✓ To obtain design rules
  - ✓ To quantify and optimize

When are Analytical models useful?

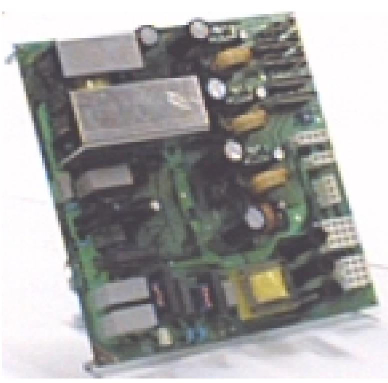
A I_{DC} is predominant

B ΔI is predominant

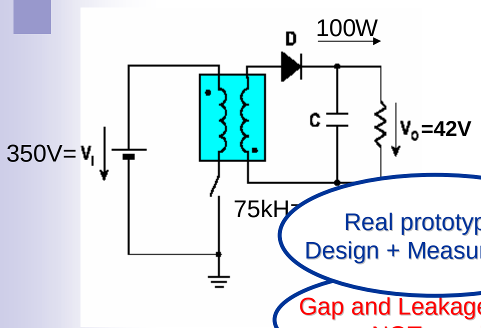




Example 4:
Flyback Transformer



Design example: Flyback



Real prototyping: Design + Measurements

Gap and Leakage inductance NOT considered

Transformer A
Analytical design

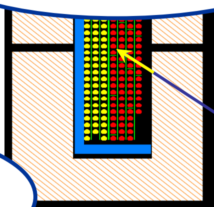
ETD 39, 3F3

Air Gap 2.2mm (86.6mils)

Primary
- # turns: 30
- Wire Gauge: AWG22 (0.7mm, 27mils, 3 x skin depth)
- Parallel: 3

Secondary
- # turns: 6
- Wire Gauge: AWG22 (0.7mm, 27mils, 3 x skin depth)
- Parallel: 18

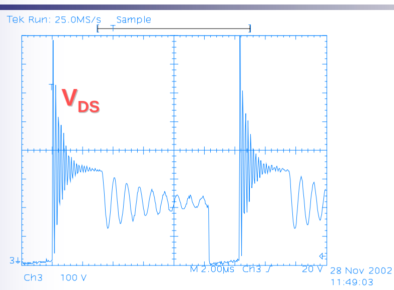
Analytical results
- ✓ Core losses: 450mW
- ✓ Winding losses: 200mW
- ✓ ΔT =7ºC

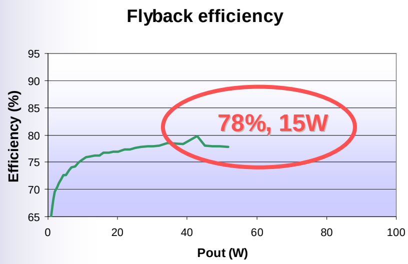


High window filling

DC resistance
Skin effect
Core losses

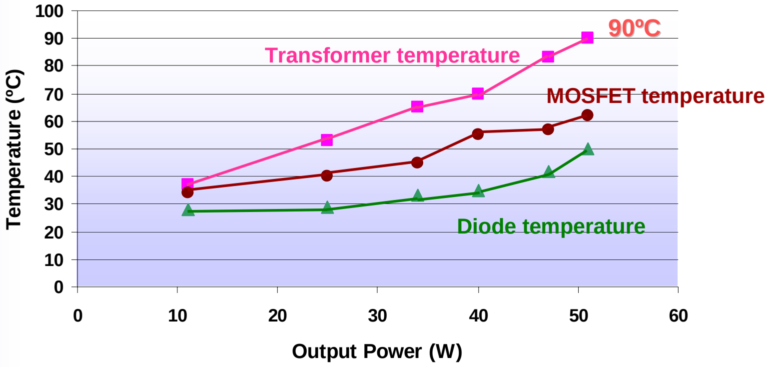
Transformer A: Measurements on Flyback prototype





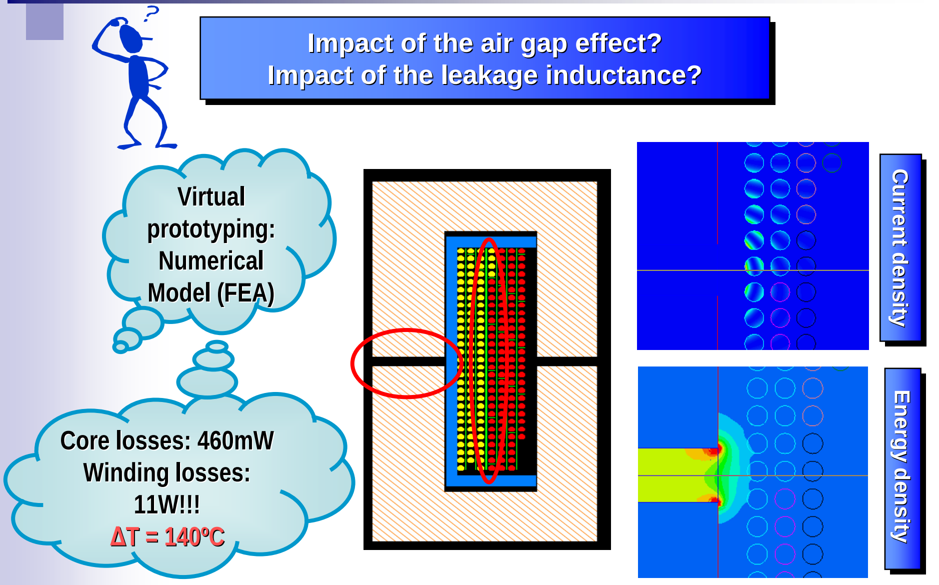
Too many losses
What is happening???

Transformer A: Measurements on Flyback prototype



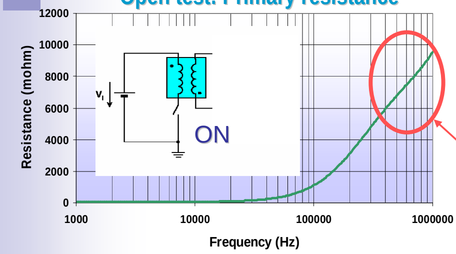
The hottest component is the transformer, full power (100W) can not be achieved due to the transformer losses

What can we do?

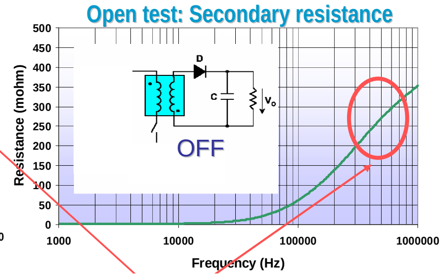


Transformer A: Numerical (FEA) model

Open test: Primary resistance

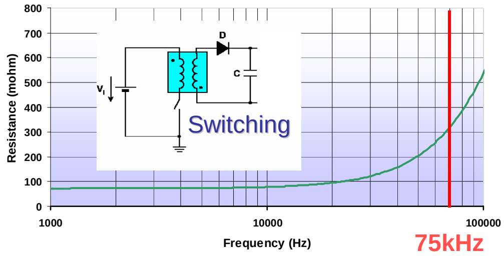


Open test: Secondary resistance



Gap effect!!

Short-circuit test
Primary resistance
(Numerical model)



Design based on numerical model (FEA)

Transformer B

ETD 44, 3F3

Air Gap 2.2mm

Primary

- # turns: 30
- Wire Gauge: AWG22
- (0.7 mm, 27mils, 3 x skin depth)
- Parallel: 1

Secondary

- # turns: 6
- Wire Gauge: AWG22
- (0.7 mm, 27mils, 3 x skin depth)
- Parallel: 6

Compared with Tr. A:
Reduces gap effect
Improves coupling
Increases DC resistance

RESULTS

Core losses:	720mW
Winding losses:	6.2W
ΔT =	60ºC

Transformer C

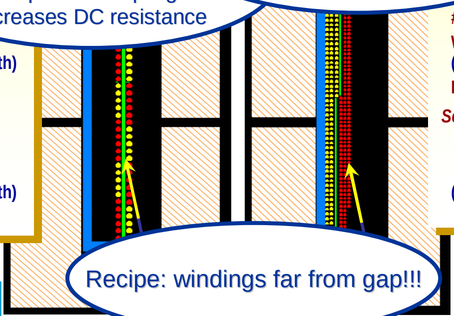
mm (86.6mils)

- # turns: 30
- Wire Gauge: AWG25
- (0.5mm, 19.7mils, 2 x skin depth)
- Parallel: 4

Secondary

- # turns: 6
- Wire Gauge: AWG25
- (0.5mm, 19.7mils, 2 x skin depth)
- Parallel: 24

Compared with Tr. A:
Reduces gap effect
Similar coupling
Increases DC resistance



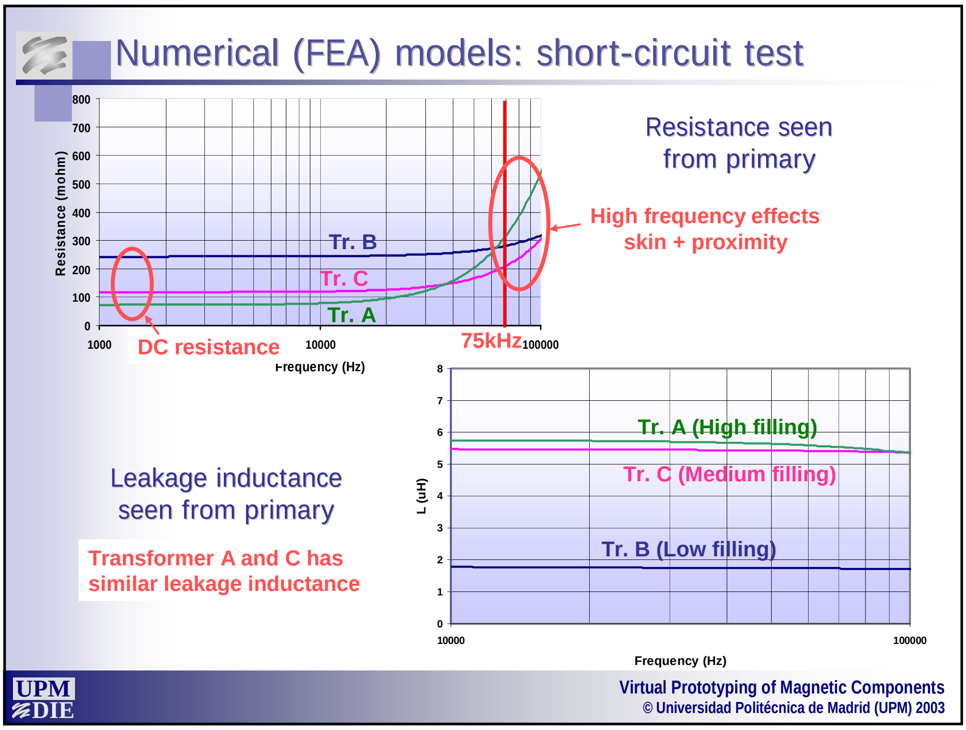
Recipe: windings far from gap!!!

Low window filling

Medium window filling

RESULTS

Core losses:	710mW
Winding losses:	9.8W
ΔT =	85ºC


Numerical (FEA) models: short-circuit test
Resistance seen from primary
High frequency effects skin + proximity
Tr. B
Tr. C
Tr. A
DC resistance
75kHz
Resistance (mohm)
Frequency (Hz)
Leakage inductance seen from primary
Transformer A and C has similar leakage inductance
Tr. A (High filling)
Tr. C (Medium filling)
Tr. B (Low filling)
L (uH)
Frequency (Hz)
UPM DIE
Virtual Prototyping of Magnetic Components
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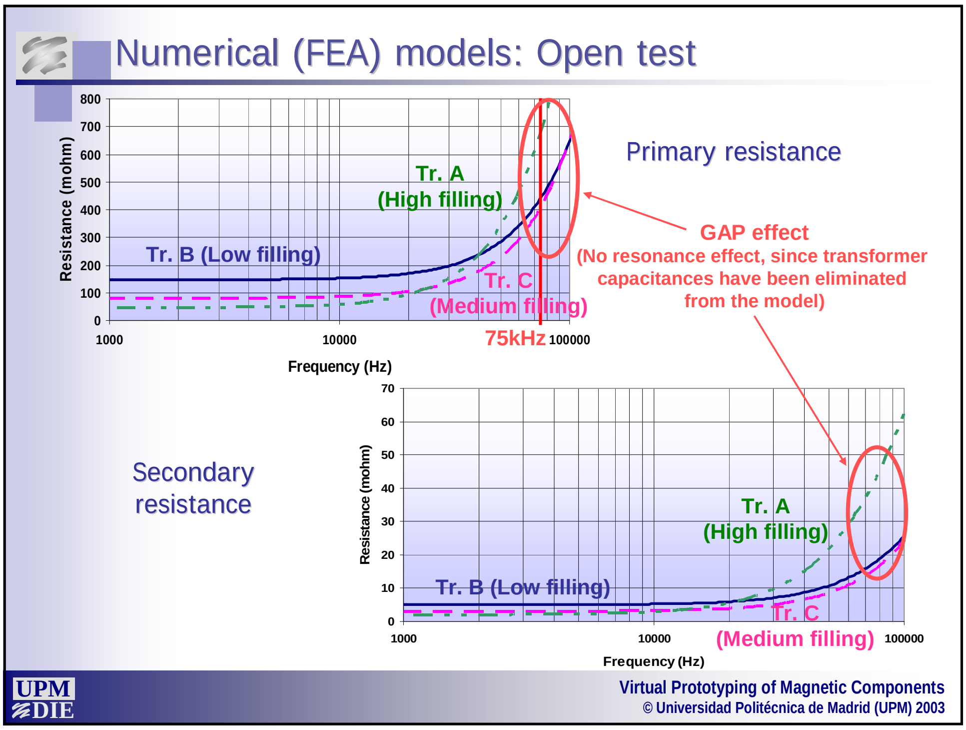
Numerical (FEA) models: Open test
Primary resistance
Tr. A (High filling)
Tr. B (Low filling)
Tr. C (Medium filling)
75kHz
GAP effect
(No resonance effect, since transformer capacitances have been eliminated from the model)
Resistance (mohm)
Frequency (Hz)
Secondary resistance
Tr. A (High filling)
Tr. B (Low filling)
Tr. C
(Medium filling)
Resistance (mohm)
Frequency (Hz)
UPM DIE
Virtual Prototyping of Magnetic Components
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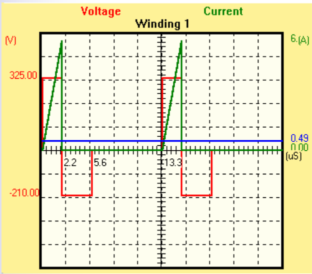
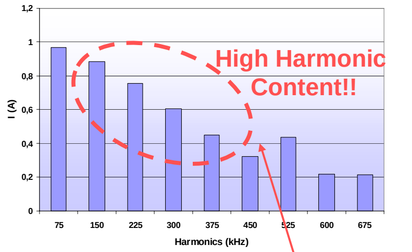
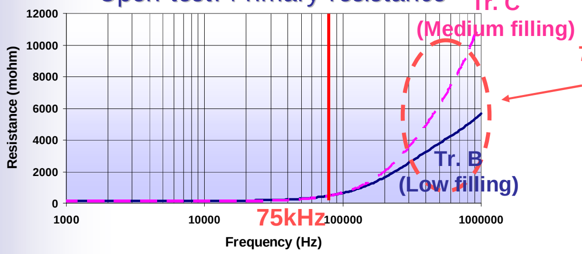
Studying the problem



Primary current: Harmonic content



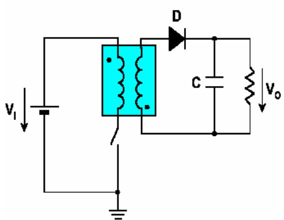
Open test: Primary resistance



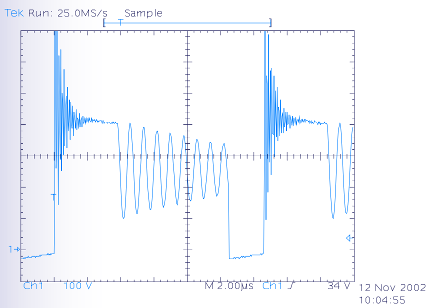
Similar resistance at 75kHz, but Tr. C is more sensitive at gap effect

Tr. B should be more efficient than Tr. C in the converter

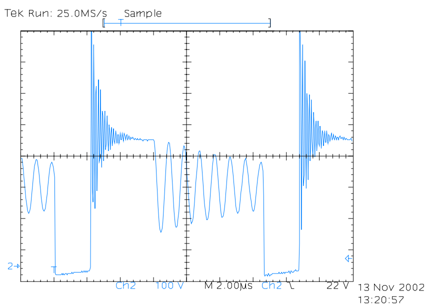
Flyback prototype: Measurements



Transformer B

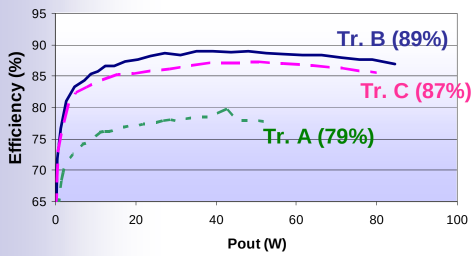


Transformer C



Flyback prototype: Measurements

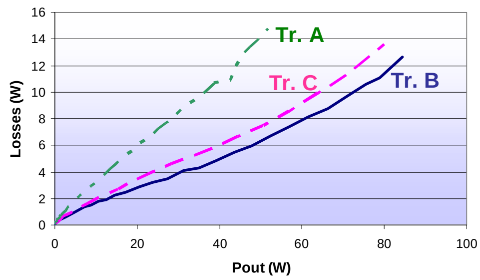
Flyback efficiency



Transformer A and C has similar inductance, however converter losses are very different ⇨GAP effect!!

Losses with Transformer C are 27% higher (3W higher) than with Transformer B

Flyback losses



Virtual prototyping accounts for 2D effects:

- ✓ problems can be properly understood
- ✓ optimum designs can be proposed:
  - ✓Not too many wires but properly placed and winded