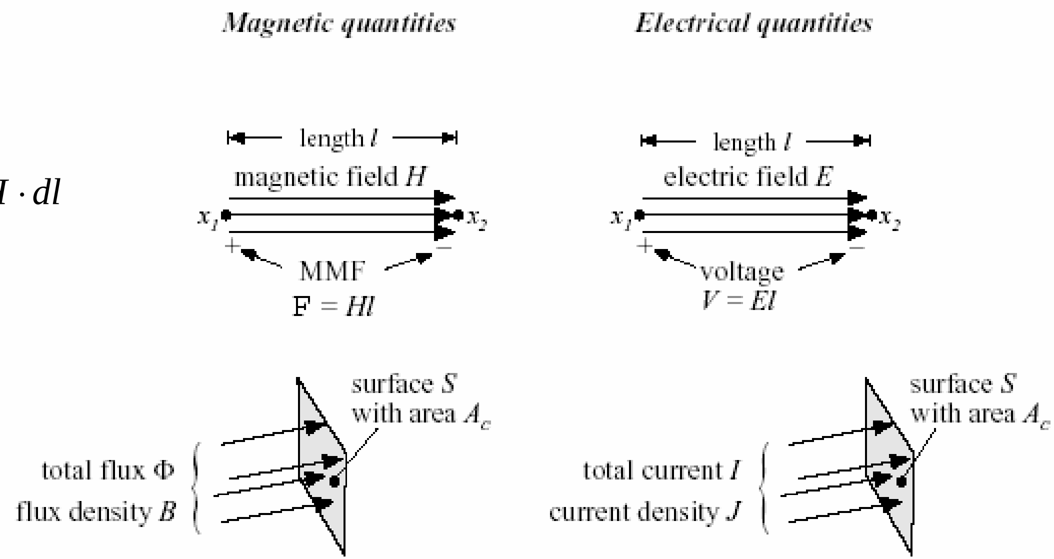
基本電磁理論

$$F = \int_{x_1}^{x_2} H \cdot dl$$

$$\Phi = \int_s B \cdot dA$$



- 磁動勢(magnetomotive force) ***F=Hl***
- 磁通量(Flux) ***Φ=BA***
- 法拉弟定理(Faraday’s law)

$$v(t) = \frac{d\Phi}{dt} = A_c \frac{dB(t)}{dt}$$

磁與電之對偶性

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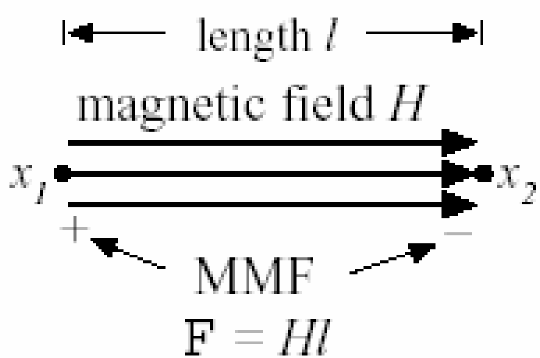
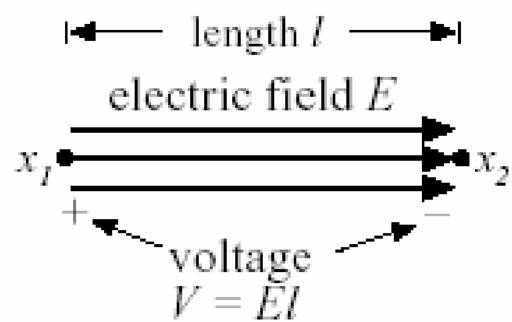
Magnetomotive force (MMF) $\mathcal{F}$ between points x_1 and x_2 is related to the magnetic field $\boldsymbol{H}$ according to

$$\mathcal{F} = \int_{x_1}^{x_2} \boldsymbol{H} \cdot d\boldsymbol{l}$$

Example: uniform magnetic field of magnitude H



Analogous to electric field of strength E, which induces voltage (EMF) V:



磁與電之對偶性

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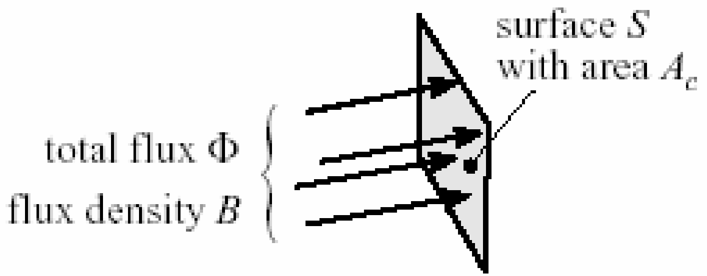
The total magnetic flux Φ passing through a surface of area A_c is related to the flux density B according to

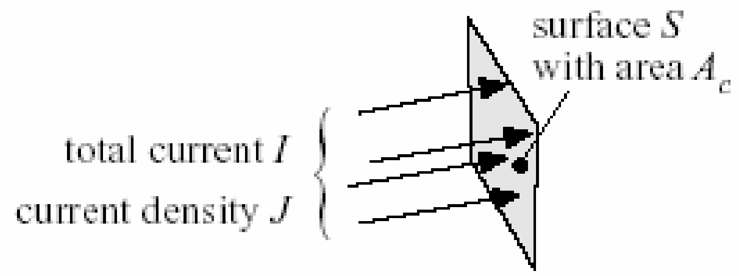
$$\Phi = \int_{surface\ S} B \cdot dA$$

Example: uniform flux density of magnitude B

$$\Phi = B\,A_c$$



Analogous to electrical conductor current density of magnitude J, which leads to total conductor current I:



磁與電之相關性

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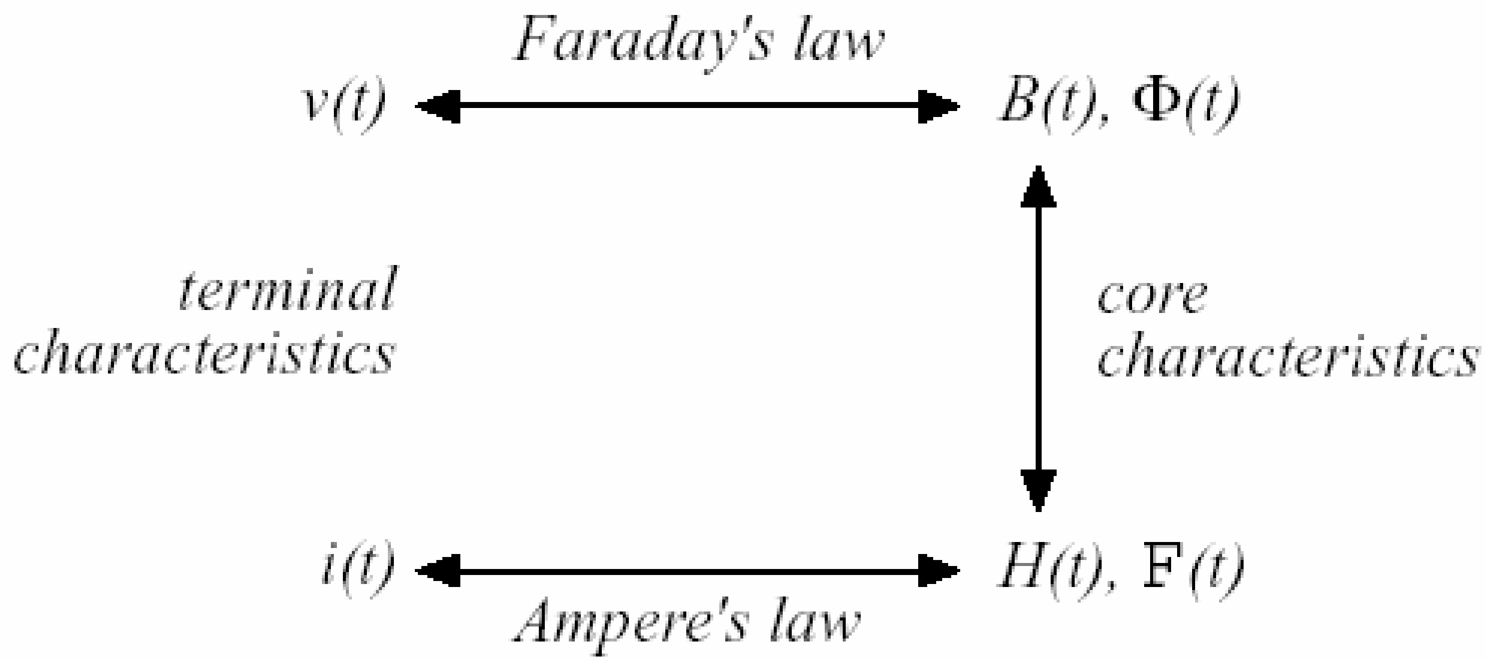

Faraday's Law

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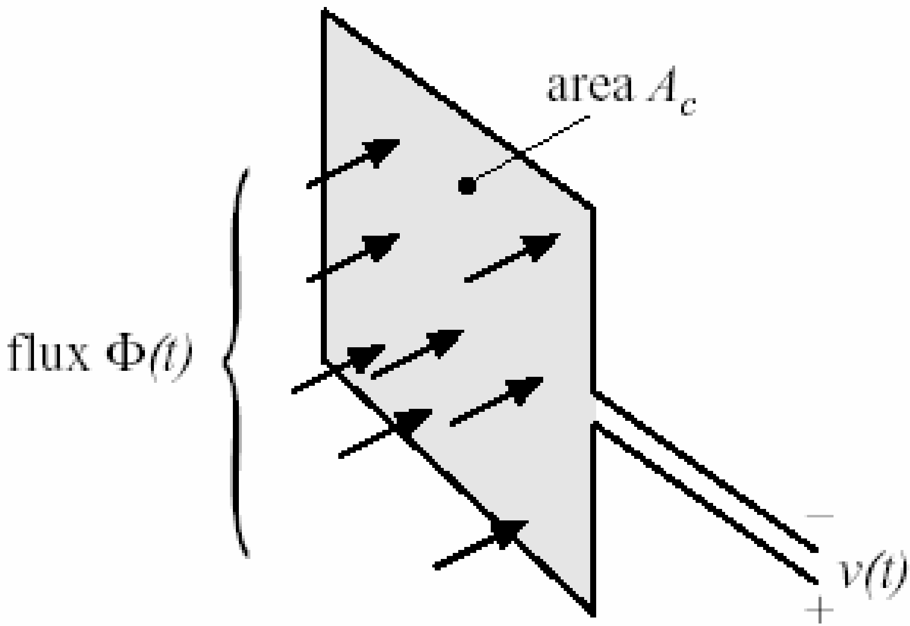
Voltage $v(t)$ is induced in a loop of wire by change in the total flux $\Phi(t)$ passing through the interior of the loop, according to

$$v(t) = \frac{d\Phi(t)}{dt}$$

For uniform flux distribution, $\Phi(t) = B(t)A_c$ and hence

$$v(t) = A_c \frac{dB(t)}{dt}$$



Lenz's Law

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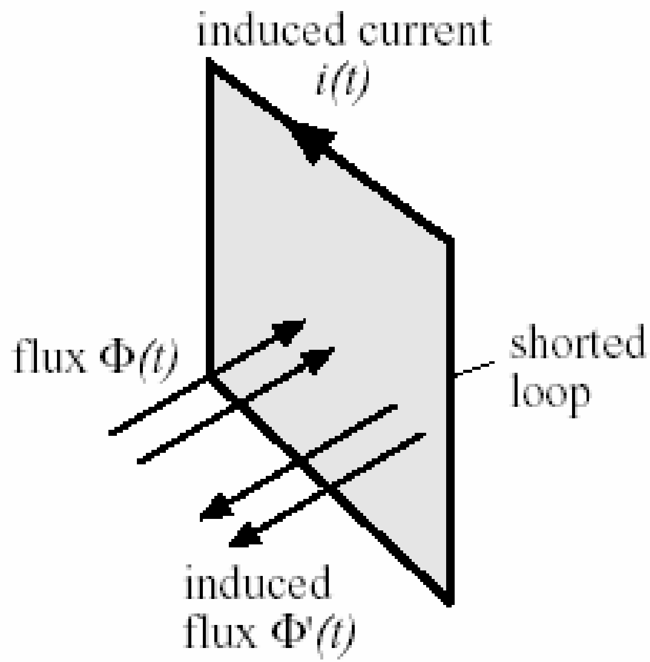
The voltage *v(t)* induced by the changing flux *Φ(t)* is of the polarity that tends to drive a current through the loop to counteract the flux change.

Example: a shorted loop of wire

- Changing flux Φ(t) induces a voltage v(t) around the loop
- This voltage, divided by the impedance of the loop conductor, leads to current i(t)
- This current induces a flux Φ'(t), which tends to oppose changes in Φ(t)



Ampere's Law

NTUST

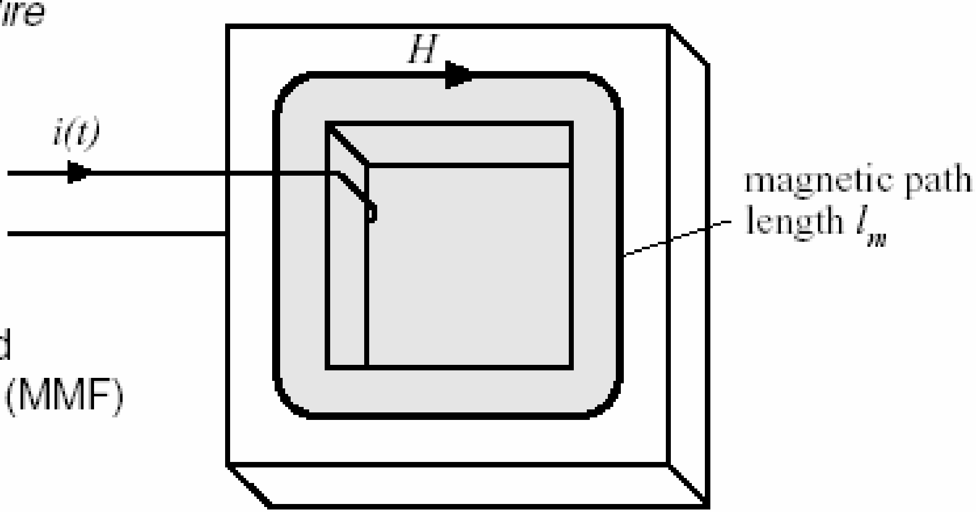
The net MMF around a closed path is equal to the total current passing through the interior of the path:

$$\oint_{closed\ path} \boldsymbol{H} \cdot \boldsymbol{dl} = total\ current\ passing\ through\ interior\ of\ path$$

Example: magnetic core. Wire carrying current $i(t)$ passes through core window.

- Illustrated path follows magnetic flux lines around interior of core
- For uniform magnetic field strength $H(t)$, the integral (MMF) is $H(t)l_m$. So

$$F(t) = H(t)\, l_m = i(t)$$



Ampere's law 解說

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- 磁場強度$H(t)$和繞組電流$i(t)$的關係
- 我們可以把繞組電流$i(t)$看做**MMF**源
- 前例中,環繞鐵心的全部**MMF**, $F(t)=H(t)l_m$就等於繞組的電流**MMF** $i(t)$
- 整個閉迴路的**MMF**應為零,指所有繞組的電流**MMF** $i(t)$之和

鐵心材料性質

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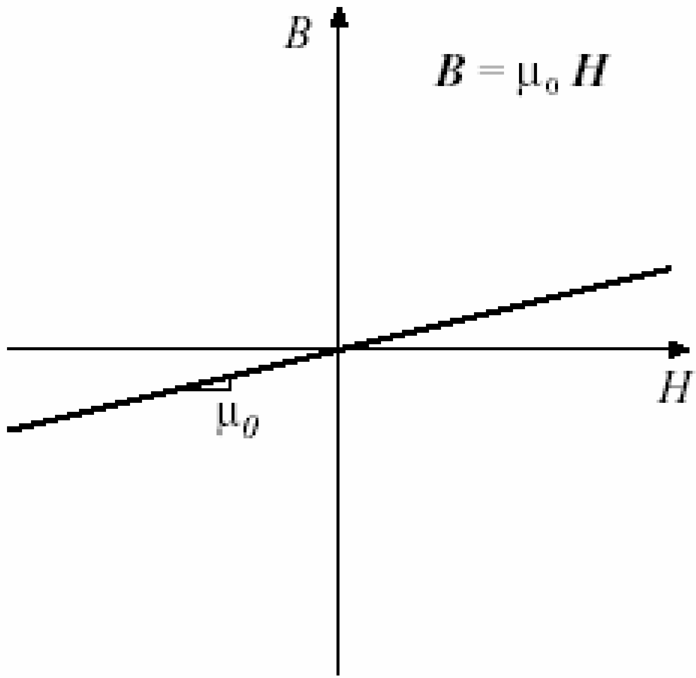
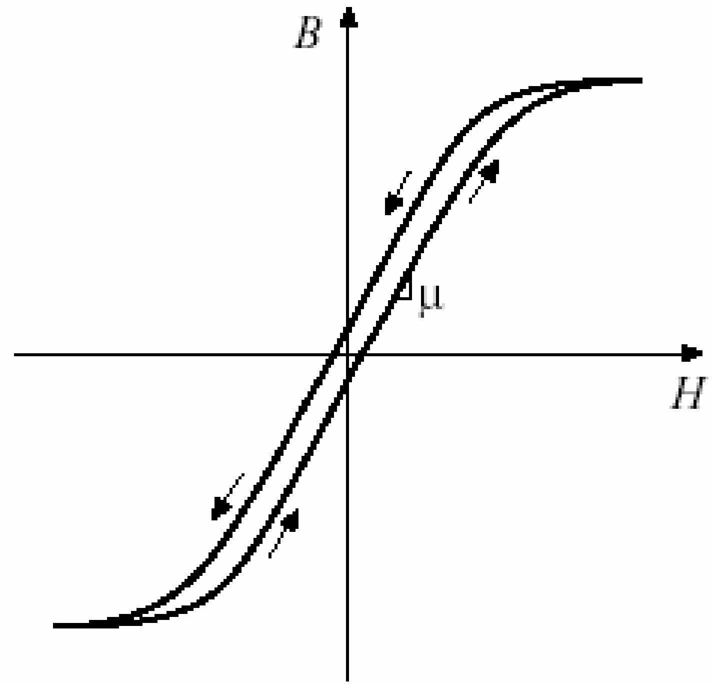
Free space



μ_0 = permeability of free space

= $4\pi \cdot 10^{-7}$ Henries per meter

A magnetic core material



Highly nonlinear, with hysteresis and saturation

片段線性

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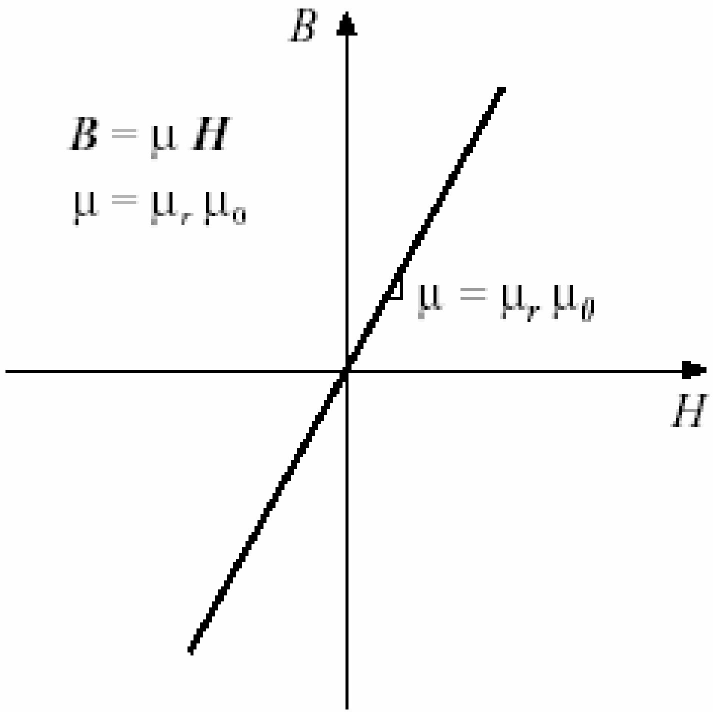
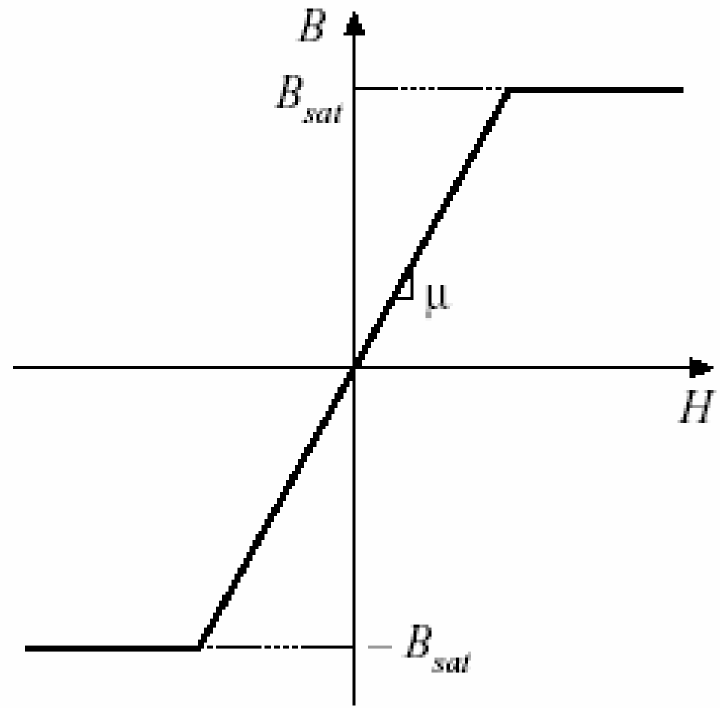
No hysteresis or saturation



Typical $\mu_r = 10^3 - 10^5$

Saturation, no hysteresis



Typical B_{sat} = 0.3-0.5T, ferrite

0.5-1T, powdered iron

1-2T, iron laminations

單位及換算

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Table 12.1. Units for magnetic quantities

quantity	*MKS*	*unrationalized cgs*	*conversions*
core material equation	$B = \mu_0 \mu_r H$	$B = \mu_r H$	
B	Tesla	Gauss	$1\text{T} = 10^4\text{G}$
H	Ampere / meter	Oersted	$1\text{A/m} = 4\pi \cdot 10^{-3}\ \text{Oe}$
Φ	Weber	Maxwell	$1\text{Wb} = 10^8\ \text{Mx}$ $1\text{T} = 1\text{Wb} / \text{m}^2$

簡單電感例

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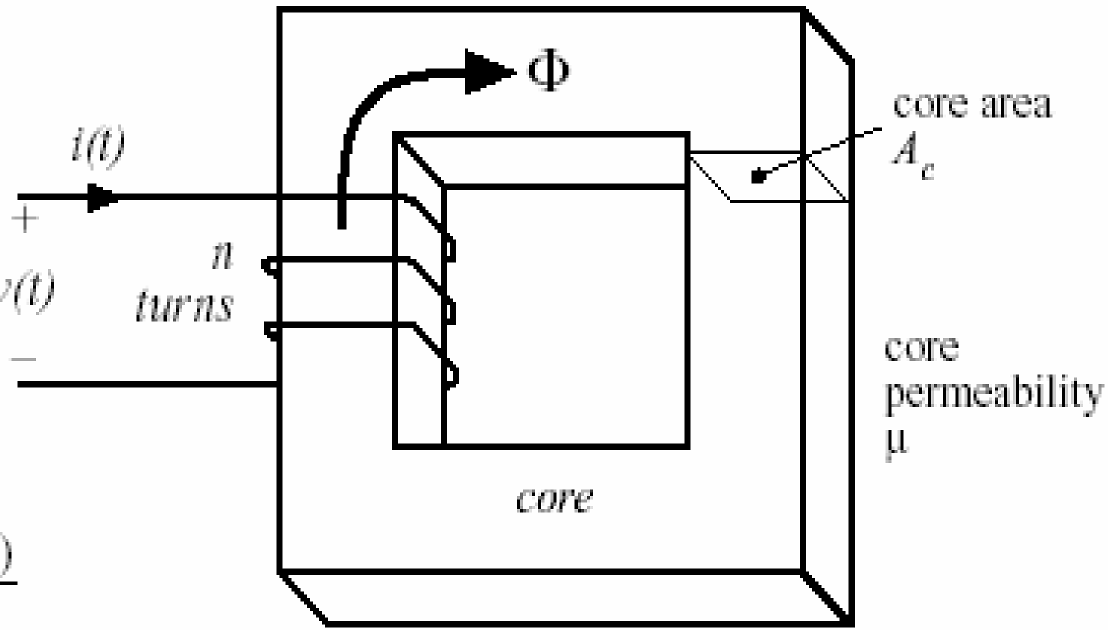
Faraday's law:

For each turn of wire, we can write

$$v_{turn}(t) = \frac{d\Phi(t)}{dt}$$



Total winding voltage is

$$v(t) = n\, v_{turn}(t) = n \frac{d\Phi(t)}{dt}$$

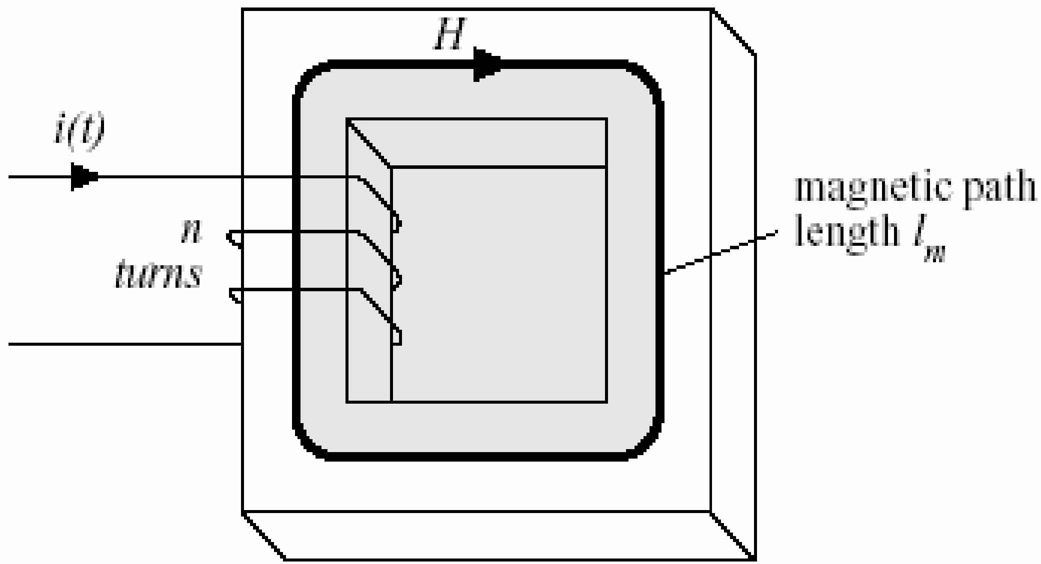
Express in terms of the average flux density $B(t) = \Phi(t)/A_c$

$$v(t) = n\, A_c \frac{dB(t)}{dt}$$

電感器例:Ampere's law

Choose a closed path which follows the average magnetic field line around the interior of the core. Length of this path is called the *mean magnetic path length* l_m.

For uniform field strength $H(t)$, the core MMF around the path is $H\ l_m$.



Winding contains n turns of wire, each carrying current $i(t)$. The net current passing through the path interior (i.e., through the core window) is $ni(t)$.

From Ampere's law, we have

$$H(t)\ l_m = n\ i(t)$$

電感器例:鐵心模式

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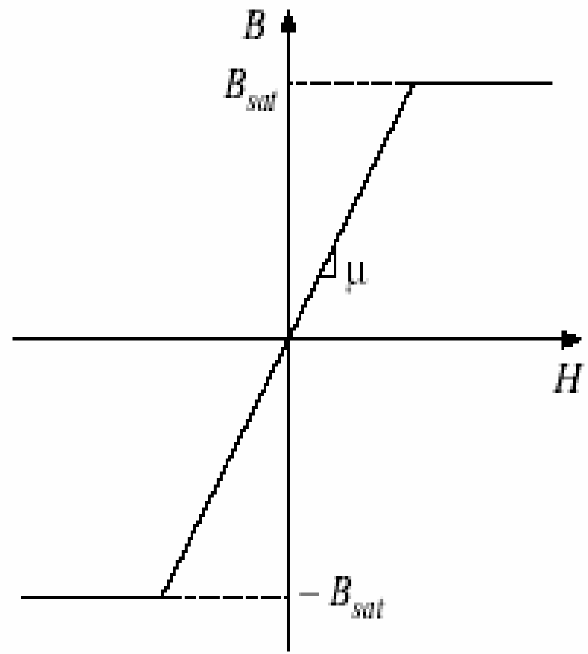
$$B = \begin{cases} B_{sat} & \text{for } H \ge B_{sat}/\mu \\ \mu H & \text{for } |H| < B_{sat}/\mu \\ -B_{sat} & \text{for } H \le B_{sat}/\mu \end{cases}$$



Find winding current at onset of saturation: substitute $i = I_{sat}$ and $H = B_{sat}/\mu$ into equation previously derived via Ampere's law. Result is

$$I_{sat} = \frac{B_{sat}\, l_m}{\mu\, n}$$

電特性

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We have:

$$v(t) = n\,A_c\,\frac{dB(t)}{dt} \qquad H(t)\,l_m = n\,i(t) \qquad B = \begin{cases} B_{sat} & \text{for } H \ge B_{sat}/\mu \\ \mu H & \text{for } |H| < B_{sat}/\mu \\ -B_{sat} & \text{for } H \le B_{sat}/\mu \end{cases}$$

Eliminate B and H, and solve for relation between v and i. For $|\,i\,| < I_{sat}$,

$$v(t) = \mu\,n\,A_c\,\frac{dH(t)}{dt} \quad \longrightarrow \quad v(t) = \frac{\mu\,n^2 A_c}{l_m}\,\frac{di(t)}{dt}$$

which is of the form

$$v(t) = L\,\frac{di(t)}{dt} \qquad \text{with} \qquad L = \frac{\mu\,n^2 A_c}{l_m}$$

—an inductor

For $|\,i\,| > I_{sat}$ the flux density is constant and equal to B_{sat}. Faraday's law then predicts

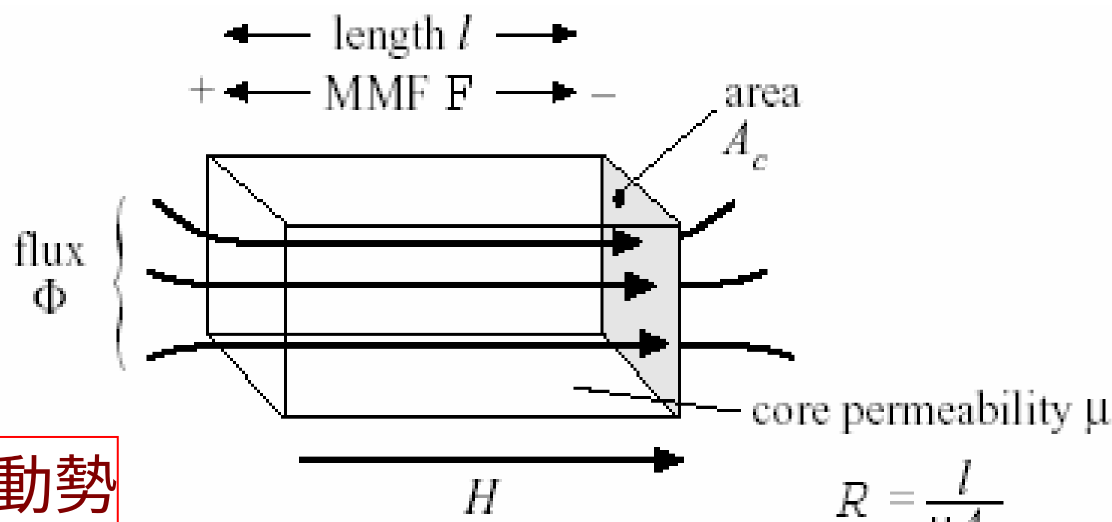
$$v(t) = n\,A_c\,\frac{dB_{sat}}{dt} = 0$$

—saturation leads to short circuit

磁路

NTUST

Uniform flux and magnetic field inside a rectangular element:



MMF between ends of element is $F = H\,l$ 磁動勢

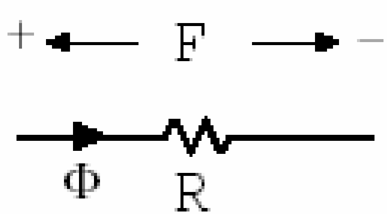
$$\mathcal{R} = \frac{l}{\mu A_c}$$

磁阻

Since $H = B / \mu$ and $B = \Phi / A_c$, we can express $\mathcal{F}$ as

$$F = \frac{l}{\mu\, A_c}\Phi \qquad \text{with} \qquad \mathcal{R} = \frac{l}{\mu\, A_c}$$

A corresponding model:



$\mathcal{R}$ = reluctance of element

磁路結構

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- 用磁阻表示電感元件
- 繞阻為**MMF**源
- **MMF**→電壓; 磁通Φ →電流
- 使用**Kirchoff's law** 解磁路

Kirchoff's current law在磁路上的應用

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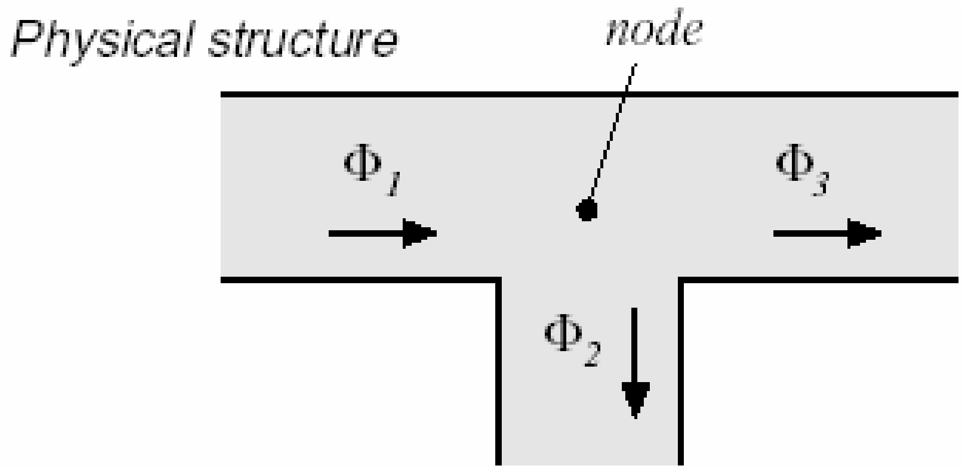
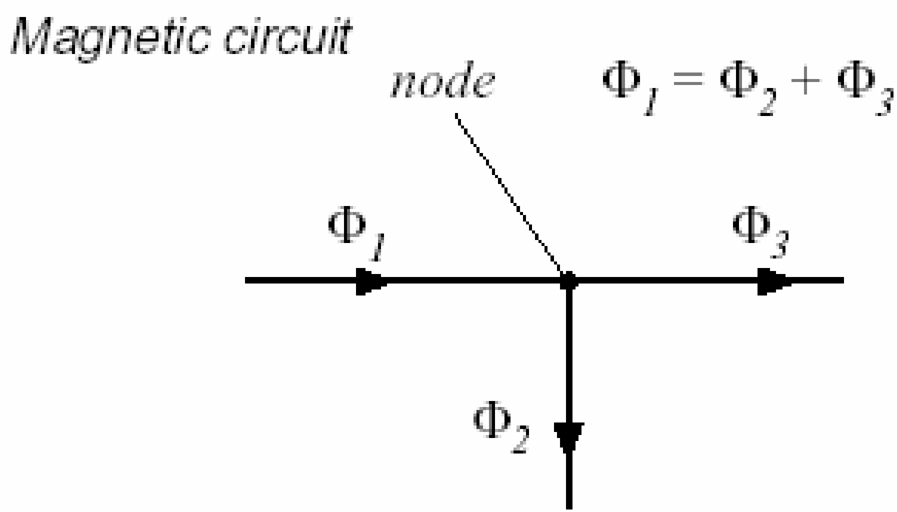
Divergence of $\boldsymbol{B} = 0$

Flux lines are continuous and cannot end

Total flux entering a node must be zero





Kirchoff's voltage law在磁路上的應用

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Follows from Ampere's law:

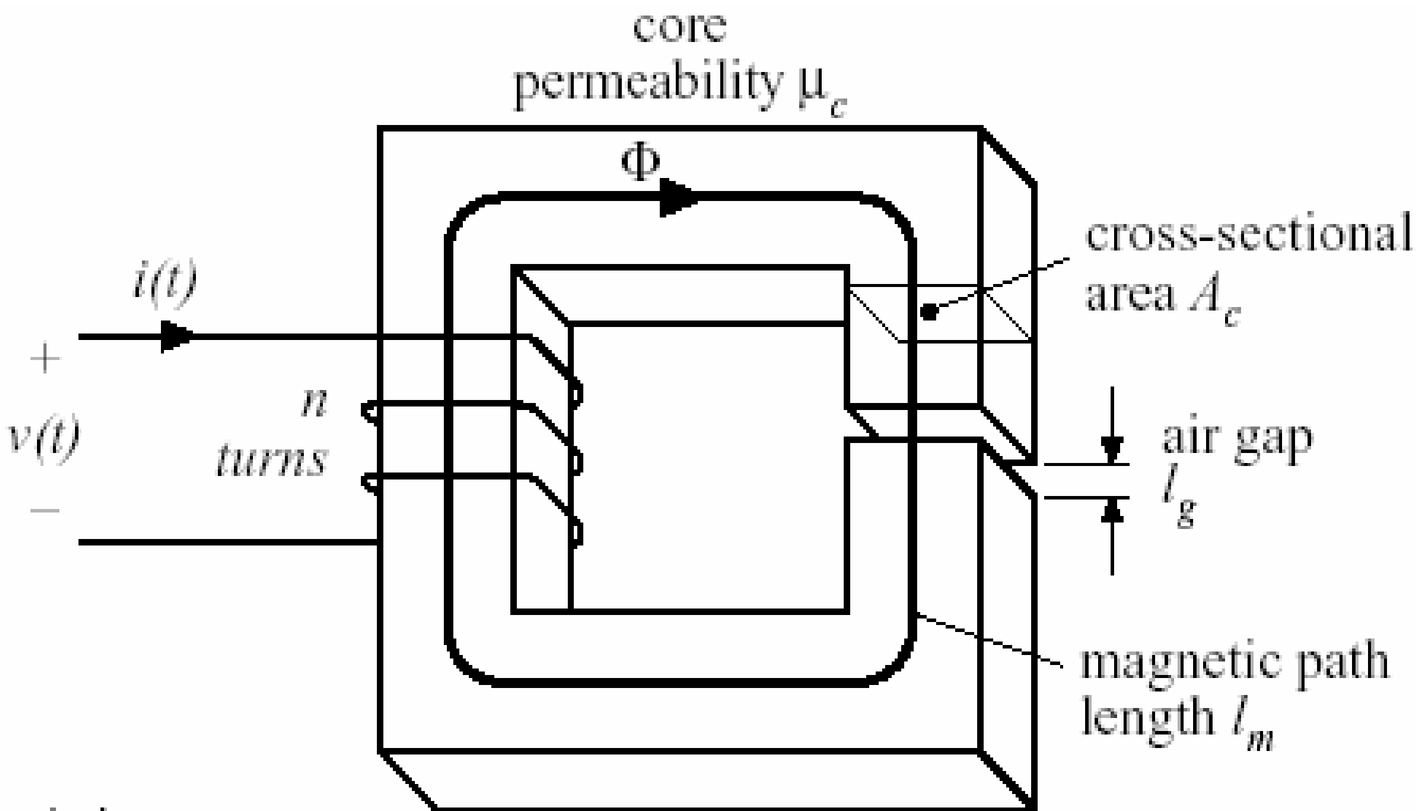
$$\oint_{closed\ path} \boldsymbol{H} \cdot d\boldsymbol{l} = total\ current\ passing\ through\ interior\ of\ path$$

Left-hand side: sum of MMF's across the reluctances around the closed path

Right-hand side: currents in windings are sources of MMF's. An n-turn winding carrying current $i(t)$ is modeled as an MMF (voltage) source, of value $ni(t)$.

Total MMF's around the closed path add up to zero.

電感含有Gap

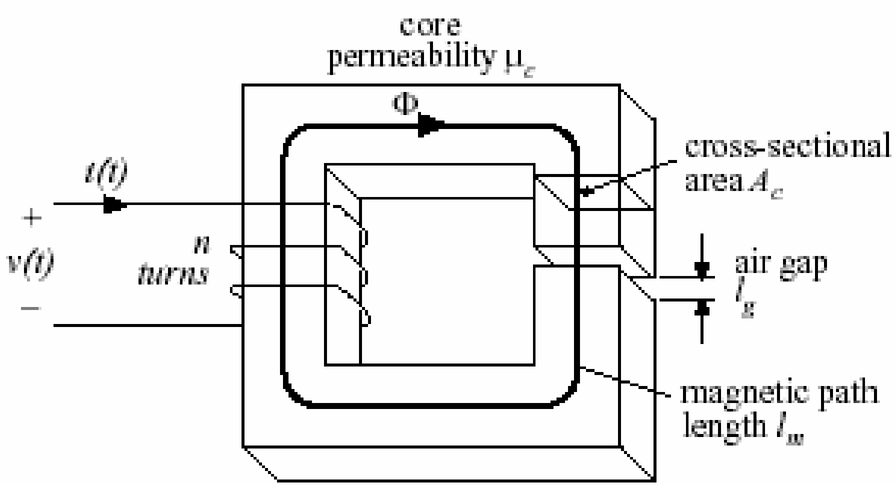


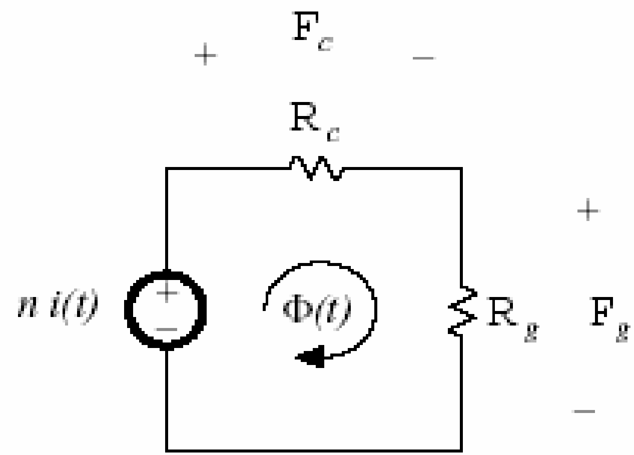
Ampere's law:

$F_c + F_g = n\,i$

$$\Phi_i = \Phi_g = \Phi$$

磁路模式





$$F_c + F_g = n\,i$$

$$R_c = \frac{l_c}{\mu A_c}$$

$$n\,i = \Phi\left(R_c + R_g\right)$$

$$R_g = \frac{l_g}{\mu_0 A_c}$$

模式之解

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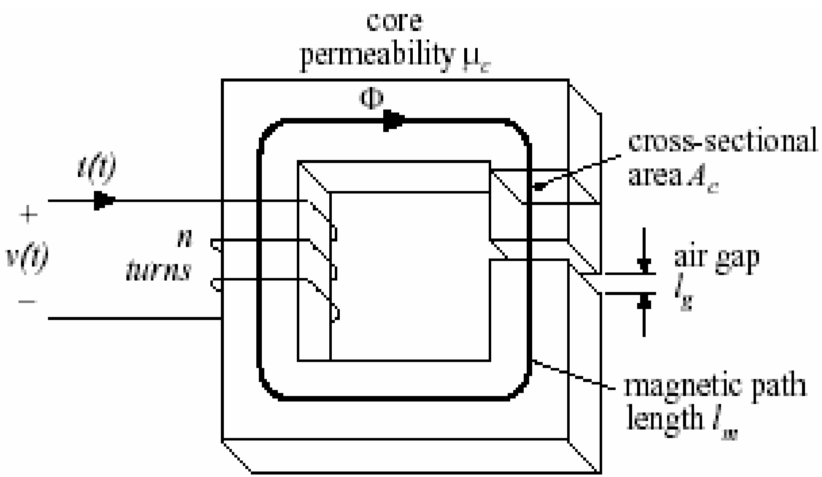
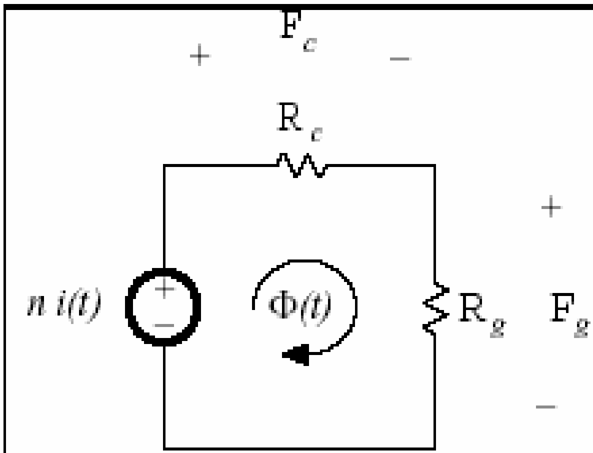




$$n\,i = \Phi\left(R_c + R_g\right)$$

$$R_c = \frac{l_c}{\mu A_c}$$

$$R_g = \frac{l_g}{\mu_0 A_c}$$

Faraday's law: $v(t) = n\dfrac{d\Phi(t)}{dt}$

Substitute for Φ: $v(t) = \dfrac{n^2}{R_c + R_g}\dfrac{di(t)}{dt}$

Hence inductance is

$$L = \frac{n^2}{R_c + R_g}$$

Air Gap的影響

$$ni = \Phi\left(R_c + R_g\right)$$

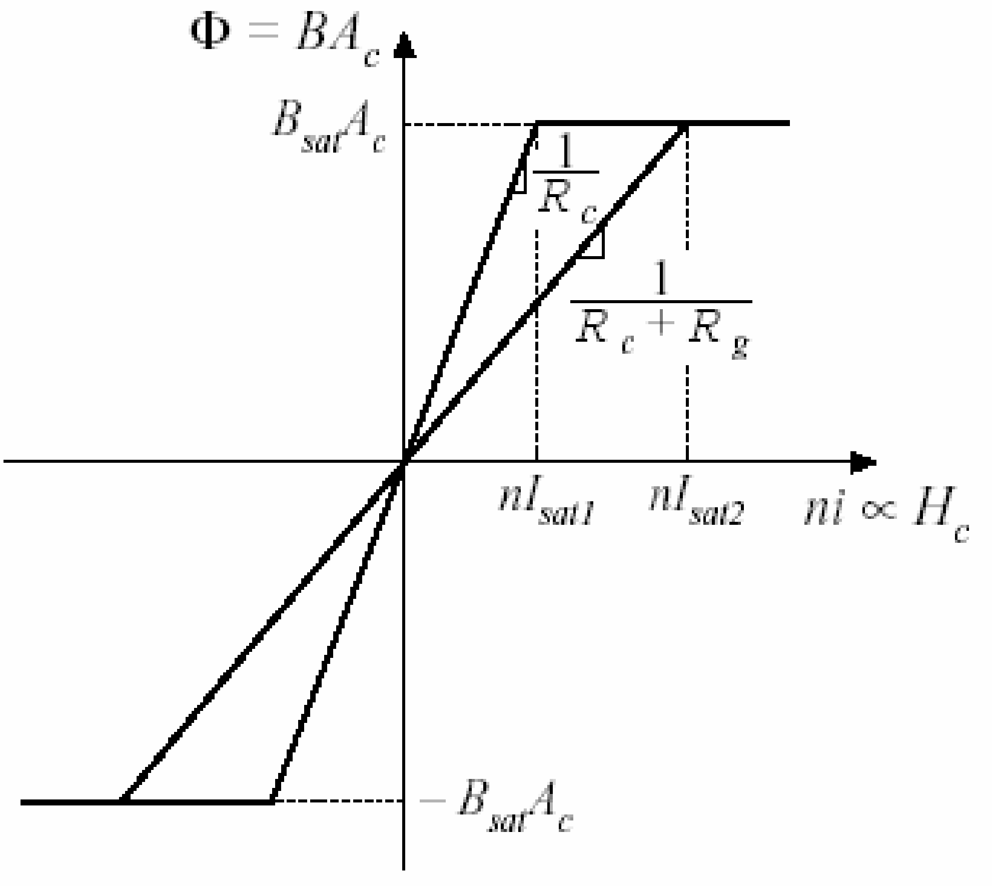
$$L = \frac{n^2}{R_c + R_g}$$

$$\Phi_{sat} = B_{sat} A_c$$

$$I_{sat} = \frac{B_{sat} A_c}{n}\left(R_c + R_g\right)$$

Effect of air gap:

- decrease inductance
- increase saturation current
- inductance is less dependent on core permeability



變壓器的模型

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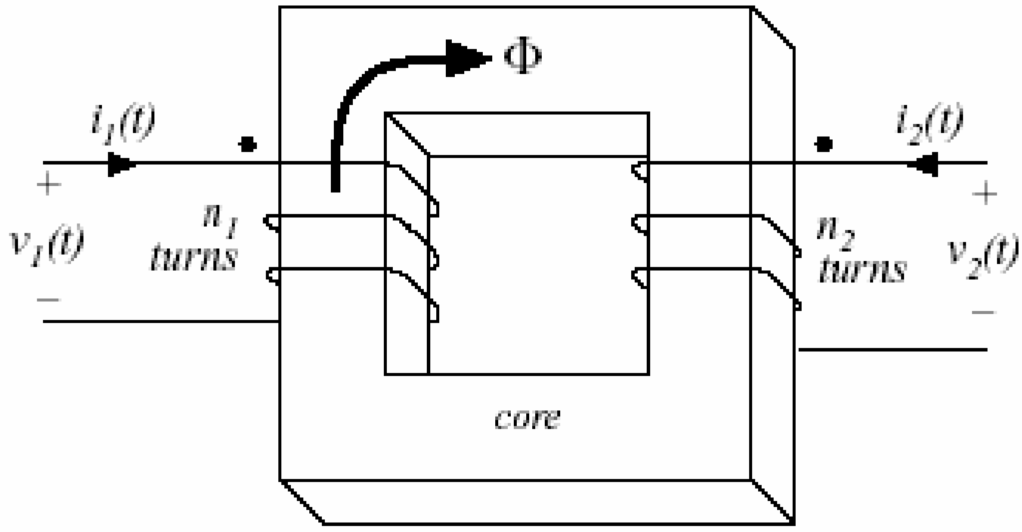
Two windings, no air gap:

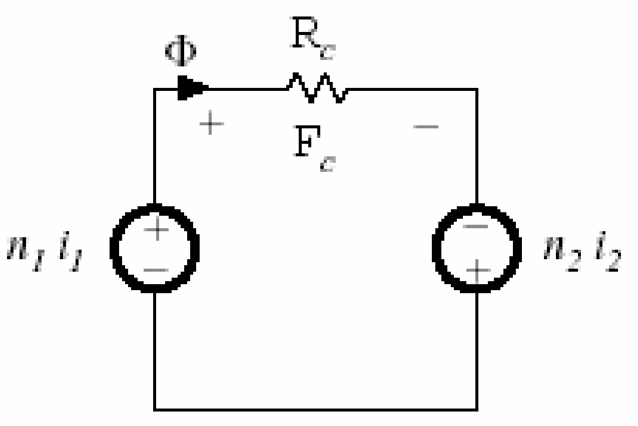
$$R = \frac{l_m}{\mu A_c}$$

$$F_c = n_1 i_1 + n_2 i_2$$

$$\Phi R = n_1 i_1 + n_2 i_2$$



Magnetic circuit model:



理想變壓器模型

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In the ideal transformer, the core reluctance R approaches zero.

MMF $F_c = \Phi$ R also approaches zero. We then obtain

$$0 = n_1 i_1 + n_2 i_2$$

Also, by Faraday's law,

$$v_1 = n_1 \frac{d\Phi}{dt}$$

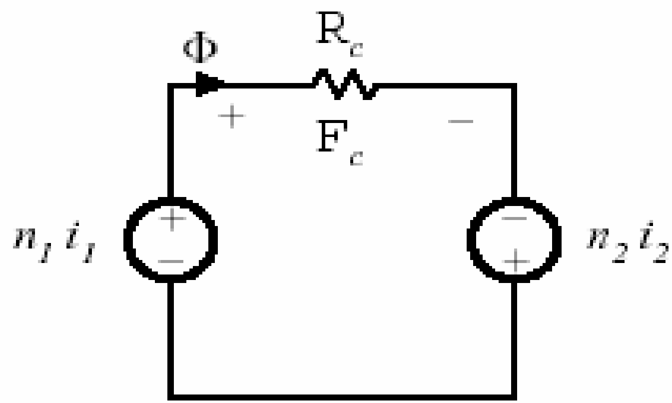
$$v_2 = n_2 \frac{d\Phi}{dt}$$

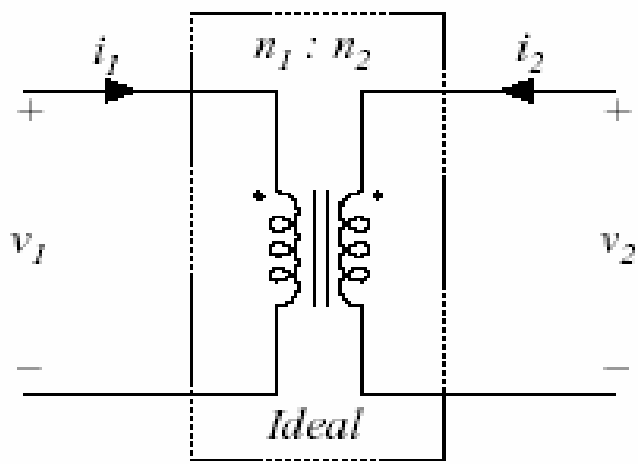
Eliminate Φ :

$$\frac{d\Phi}{dt} = \frac{v_1}{n_1} = \frac{v_2}{n_2}$$

Ideal transformer equations:

$$\frac{v_1}{n_1} = \frac{v_2}{n_2} \quad \text{and} \quad n_1 i_1 + n_2 i_2 = 0$$

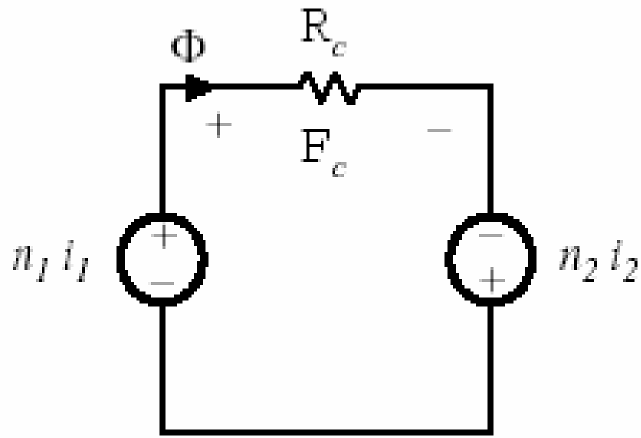




磁化電感

For nonzero core reluctance, we obtain

$$\Phi R = n_1 i_1 + n_2 i_2 \quad \text{with} \quad v_1 = n_1 \frac{d\Phi}{dt}$$



Eliminate Φ:

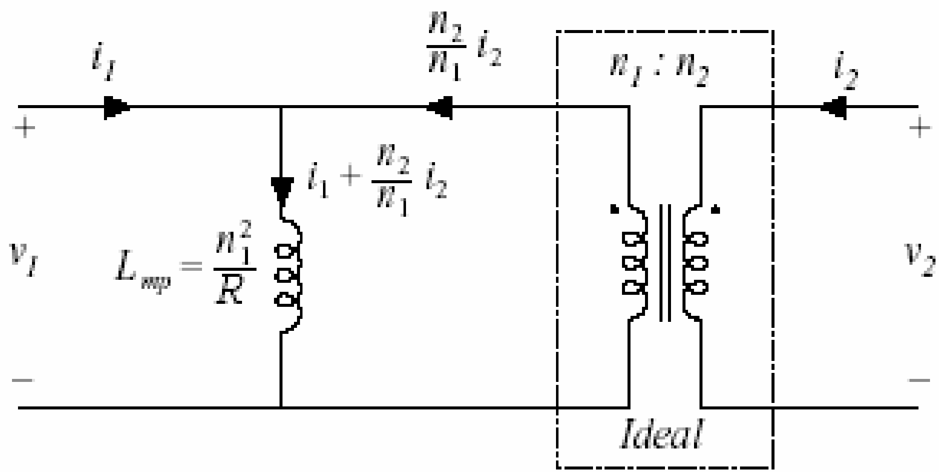
$$v_1 = \frac{n_1^2}{R} \frac{d}{dt}\left[i_1 + \frac{n_2}{n_1} i_2\right]$$

This equation is of the form

$$v_1 = L_{mp} \frac{d\, i_{mp}}{dt}$$

with $\quad L_{mp} = \dfrac{n_1^2}{R}$

$$i_{mp} = i_1 + \frac{n_2}{n_1} i_2$$



磁化電感解說

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- 實際電感器具有飽和和磁滯現象
- 若次級繞阻置空,則初級繞阻特性形同一電感器,此電感器即為初級繞阻的磁化電感
- 磁化電流會使繞阻電流比和圈數比不同

變壓器飽和

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- 當鐵心之磁通密度$B(t)$超過飽和磁通密度B_{sat}就發生飽和現象
- 當鐵心飽和,磁化電流變大,磁化電感的阻抗變小,繞阻形同短路
- 大的繞阻電流$i_1(t)$和$i_2(t)$不必導入飽和.如果$0=n_1i_1+n_2i_2$則磁化電流為零且鐵心內沒有淨磁存在
- 飽和是起因於過多的外加volt-second

飽和和外加volt-second的關係

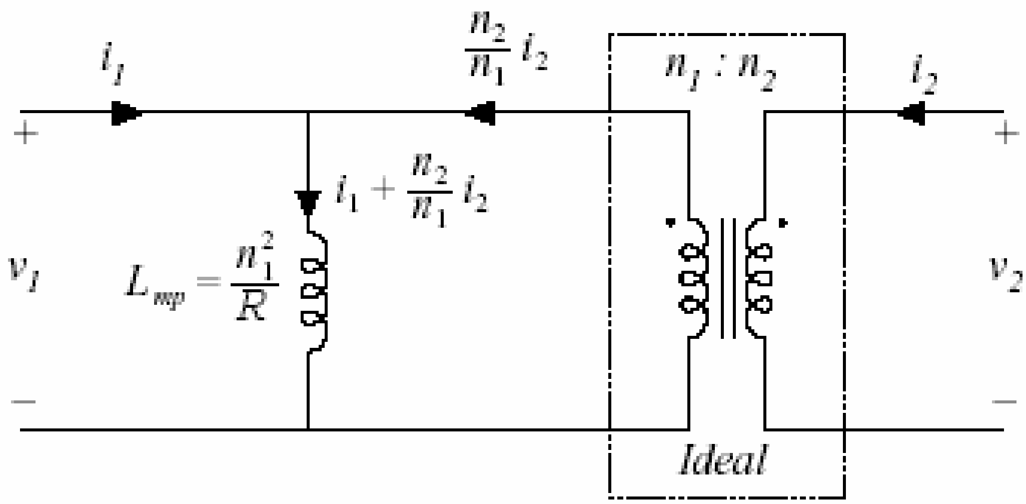
NTUST

Magnetizing current depends on the integral of the applied winding voltage:

$$i_{mp}(t) = \frac{1}{L_{mp}} \int v_1(t)\, dt$$

Flux density is proportional:

$$B(t) = \frac{1}{n_1 A_c} \int v_1(t)\, dt$$



Flux density befcomes large, and core saturates, when the applied volt-seconds λ_1 are too large, where

$$\lambda_1 = \int_{t_1}^{t_2} v_1(t)\, dt$$

limits of integration chosen to coincide with positive portion of applied voltage waveform

漏感

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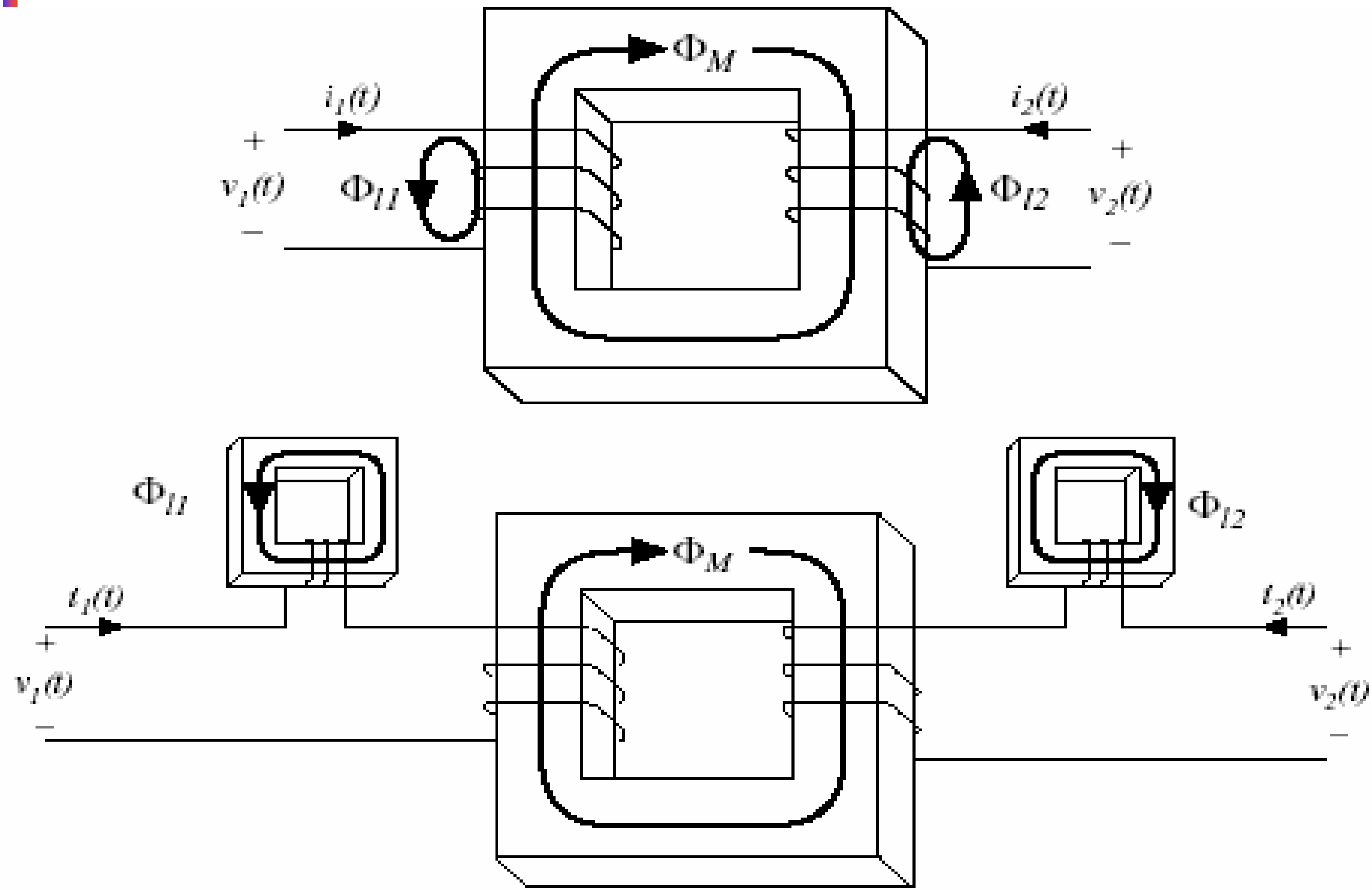

含漏感的變壓器模型

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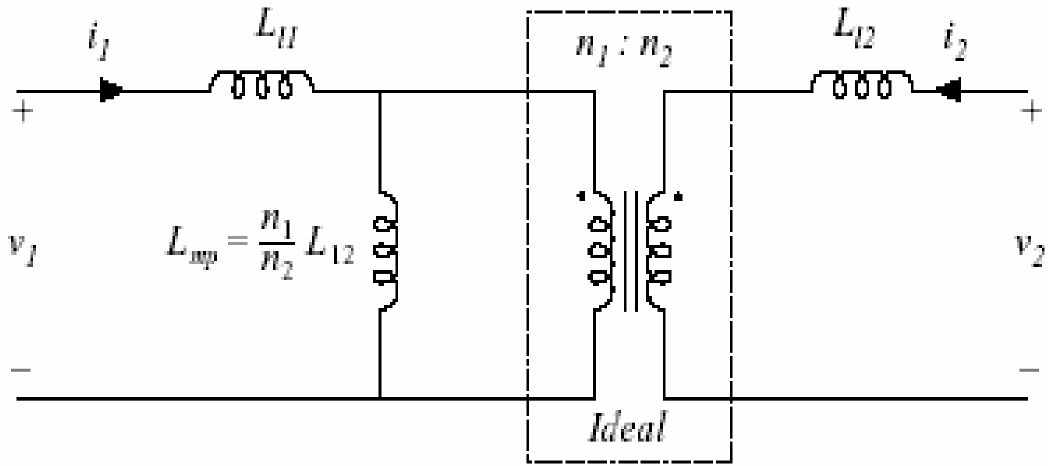
$$\begin{bmatrix} v_1(t) \\ v_2(t) \end{bmatrix} = \begin{bmatrix} L_{11} & L_{12} \\ L_{12} & L_{22} \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_1(t) \\ i_2(t) \end{bmatrix}$$



mutual inductance

$$L_{12} = \frac{n_1 n_2}{R} = \frac{n_2}{n_1} L_{mp}$$

primary and secondary self-inductances

$$L_{11} = L_{l1} + \frac{n_1}{n_2} L_{12}$$

$$L_{22} = L_{l2} + \frac{n_2}{n_1} L_{12}$$

effective turns ratio $\quad n_e = \sqrt{\frac{L_{22}}{L_{11}}}$

coupling coefficient $\quad k = \frac{L_{12}}{\sqrt{L_{11} L_{22}}}$

磁性材料的損失類型

NTUST

Low-frequency losses:

- Dc copper loss
- Core loss: hysteresis loss

High-frequency losses: the skin effect

- Core loss: classical eddy current losses
- Eddy current losses in ferrite cores

High frequency copper loss: the proximity effect

- Proximity effect: high frequency limit
- MMF diagrams, losses in a layer, and losses in basic multilayer windings
- Effect of PWM waveform harmonics

鐵心損

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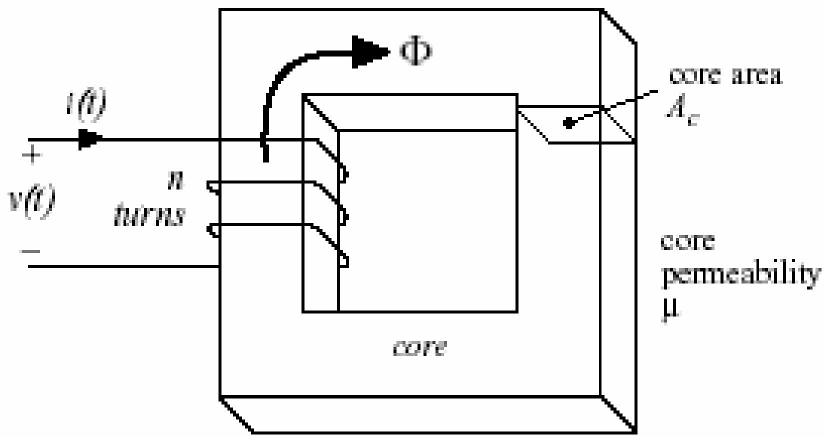
Energy per cycle W flowing into n-turn winding of an inductor, excited by periodic waveforms of frequency f:



$$W = \int_{\text{one cycle}} v(t) i(t) dt$$

Relate winding voltage and current to core B and H via Faraday's law and Ampere's law:

$$v(t) = n A_c \frac{dB(t)}{dt} \qquad H(t)\, l_m = n\, i(t)$$

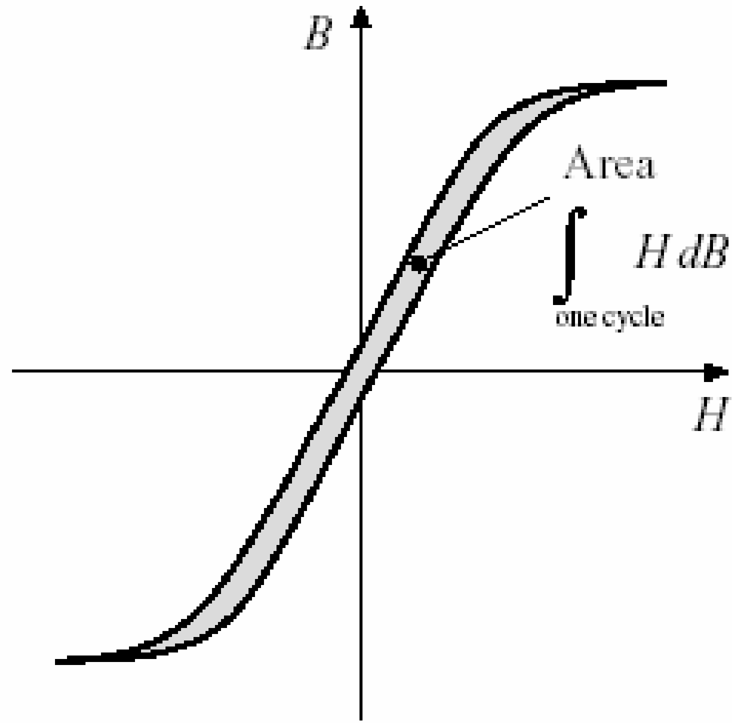
Substitute into integral:

$$W = \int_{\text{one cycle}} \left(n A_c \frac{dB(t)}{dt} \right) \left(\frac{H(t) l_m}{n} \right) dt$$

$$= \left(A_c l_m \right) \int_{\text{one cycle}} H\, dB$$

磁滯損

$$W = (A_c l_m) \int_{\text{one cycle}} H\,dB$$



The term $A_c l_m$ is the volume of the core, while the integral is the area of the B-H loop.

(energy lost per cycle) = (core volume) (area of B-H loop)

$$P_H = (f)(A_c l_m) \int_{\text{one cycle}} H\,dB$$

Hysteresis loss is directly proportional to applied frequency

磁滯損的模式化

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- 磁滯損直接受外加頻率影響
- 磁滯損和最大磁通密度有關:***B-H***迴路如何和最大磁通密度有關?經驗公式**(Steinmetz equation)**

$$P_H = K_H f B_{\max}^{\alpha} (Core\, volumn)$$

- 其中參數k_H和α依實驗而定
- P_H和B_{max}關係由電磁理論定之

渦流損

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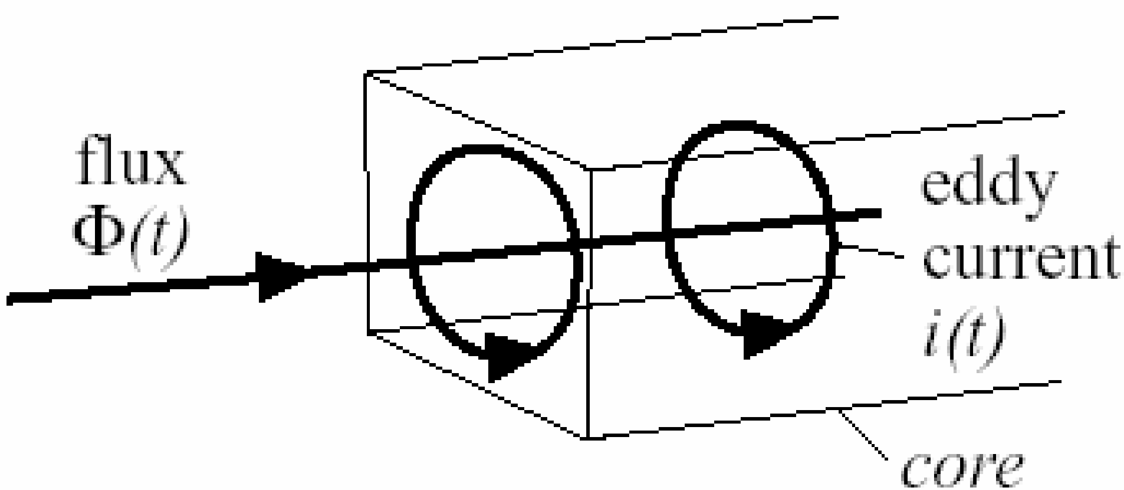
Magnetic core materials are reasonably good conductors of electric current. Hence, according to Lenz's law, magnetic fields within the core induce currents ("eddy currents") to flow within the core. The eddy currents flow such that they tend to generate a flux which opposes changes in the core flux $\Phi(t)$. The eddy currents tend to prevent flux from penetrating the core.



Eddy current loss $i^2(t)R$

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渦流損解説

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全鐵心損

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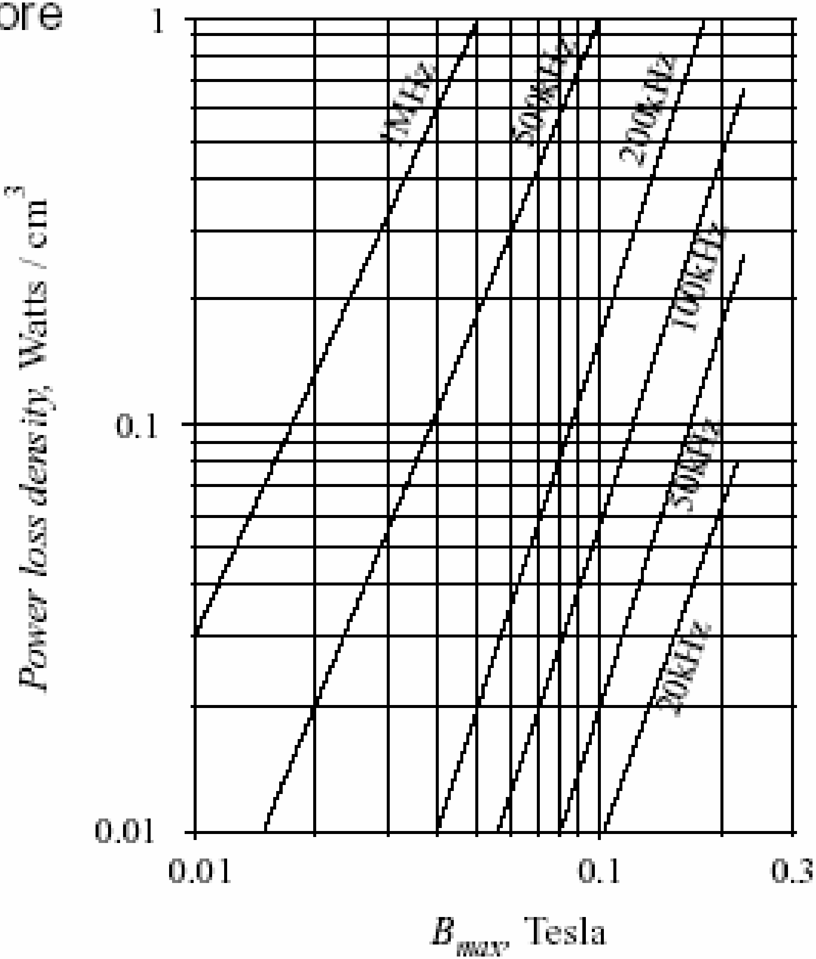
Ferrite core material



Empirical equation, at a fixed frequency:

$$P_{fe} = K_{fe} B_{max}^{\beta} A_c l_m$$

鐵心材料型式

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Core type	B_{sat}	Relative core loss	Applications
Laminations iron, silicon steel	1.5 - 2.0 T	high	50-60 Hz transformers, inductors
Powdered cores powdered iron, molypermalloy	0.6 - 0.8 T	medium	1 kHz transformers, 100 kHz filter inductors
Ferrite Manganese-zinc, Nickel-zinc	0.25 - 0.5 T	low	20 kHz - 1 MHz transformers, ac inductors

磁性材料原理及應用

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低頻銅損

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線的直流電阻

$$R = \rho \frac{l_b}{A_\omega}$$

A_w為裸線截面積l_b為限長

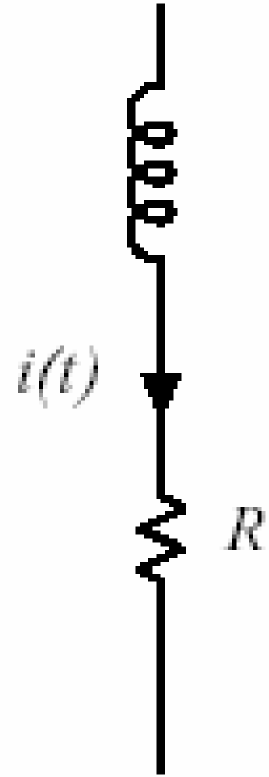
電阻係數

$\rho = 1.724 \cdot 10^{-6}\Omega cm$ (25°C)

$\rho = 2.3 \cdot 10^{-6}\Omega cm$ (100°C)

線電阻的功率損失

$P_{cu} = I^2_{rms} R$



繞組導體的渦流(Eddy Current)

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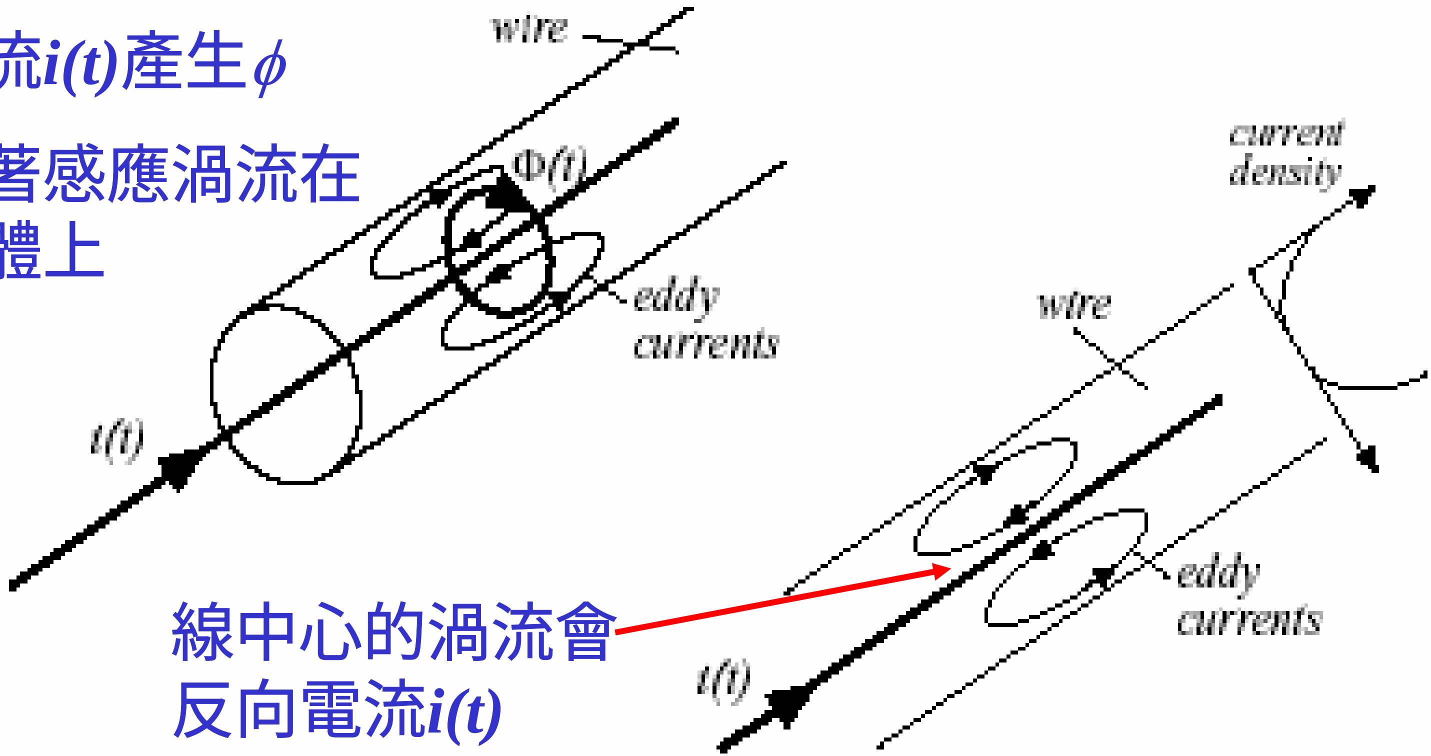
Skin effect:

電流*i(t)*產生ϕ

接著感應渦流在導體上



線中心的渦流會反向電流*i(t)*

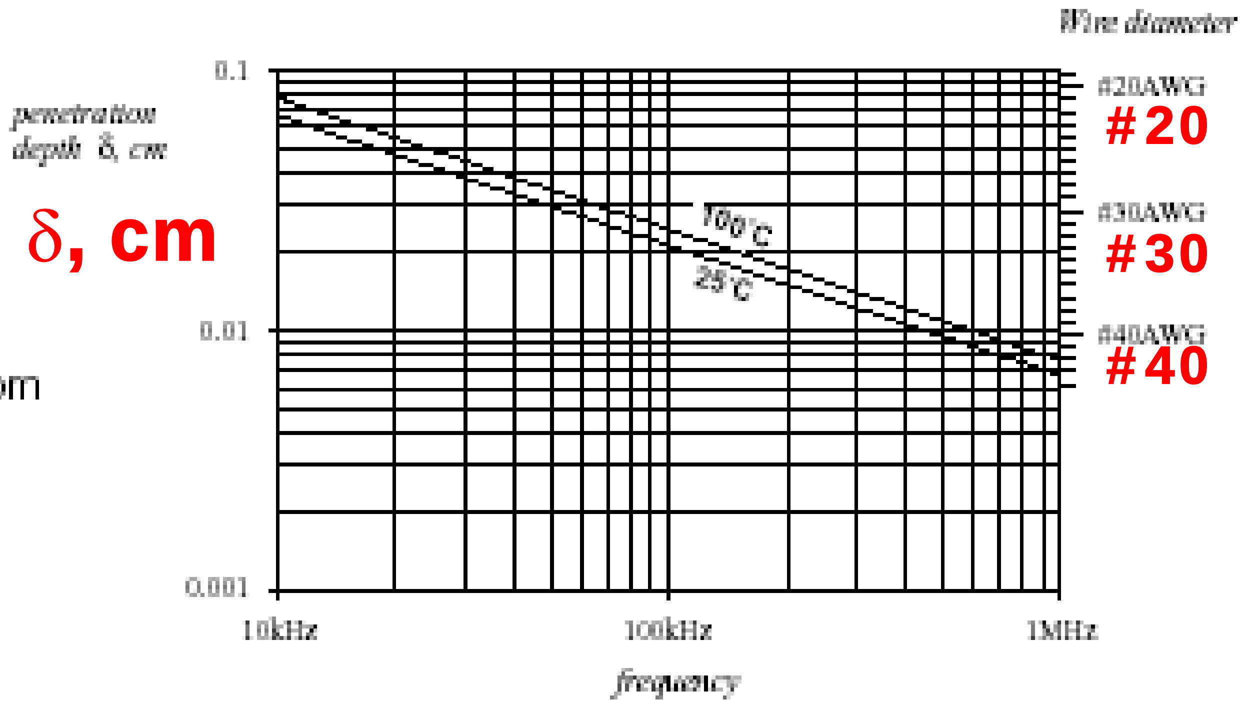
集膚深度δ

For sinusoidal currents: current density is an exponentially decaying function of distance into the conductor, with characteristic length δ known as the *penetration depth* or *skin depth*.

$$\delta = \sqrt{\frac{\rho}{\pi \mu f}}$$

For copper at room temperature:

$$\delta = \frac{7.5}{\sqrt{f}} \text{ cm}$$



集膚深度定義

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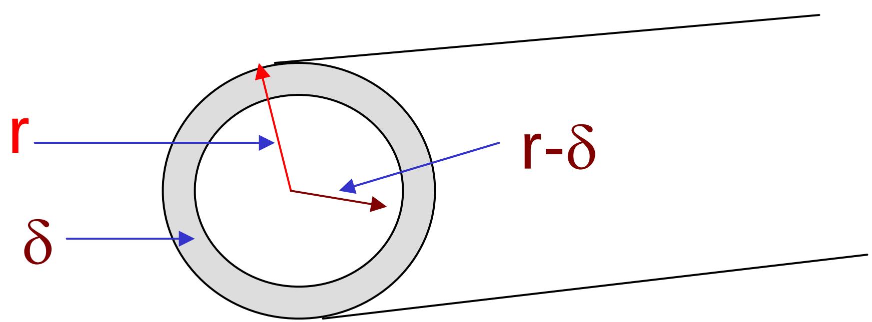
- 集膚深度定義為線表面電流密度降至其37% (or 1/e)時對應至表面的距離
- 銅在70ºC時,穿入深度$\delta = 2837/f^{-1/2}$ (mil)
- $R_{ac} = R_{dc} + \Delta R$ (Due to skin effect)

$$\frac{R_{ac}}{R_{dc}} = \frac{\pi r^2}{\pi r^2 - \pi (r-\delta)^2} = \frac{(d/2\delta)^2}{(d/2\delta)^2 - (d/2\delta - 1)^2}$$



集膚效應

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- 集膚效應和頻率成正比
- 穿入深度$\delta=(\rho/\pi\mu f)^{-1/2}$, ρ為電阻係數
- 對銅線言, $\mu=\mu_o$, 在100°C時,銅線的穿入深度可表示為$\delta=7.5/f^{-1/2}$
- AWG #40線在500kHz時,其$d/\delta=1$,直徑=穿入深度, (d_{max}=0.07987mm)
- AWG #22線在10kHz時,其$d/\delta=1$,直徑=穿入深度, (d_{max}=0.6438mm)
- 高頻電流不流線中心,而流經表面,有效截面積因而減小

接近效應(Proximity effect)

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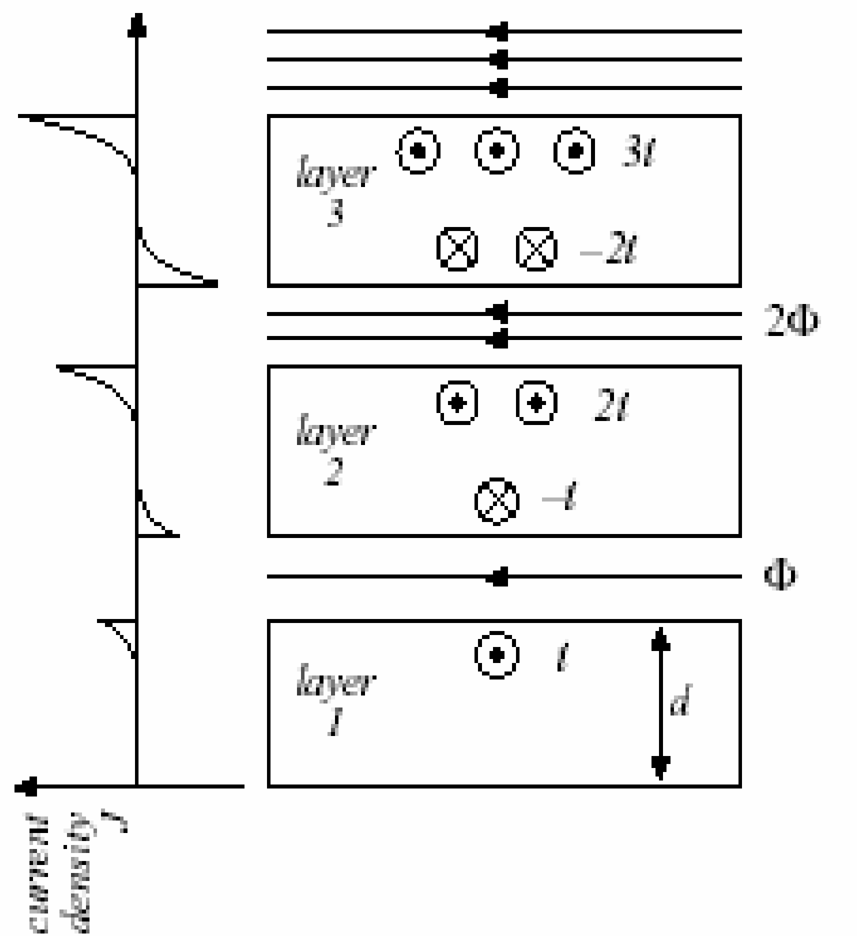
當交流在導體中產生eddy current,在相鄰的導體中會產生高頻銅損,此過程稱為接近效應(Proximity effect)



接近效應損失:高頻限制

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Let P_1 be power loss in layer 1:

銅損 $$P_1 = I_{rms}^2 \left(R_{dc} \frac{d}{\delta}\right)$$

Power loss P_2 in layer 2 is:

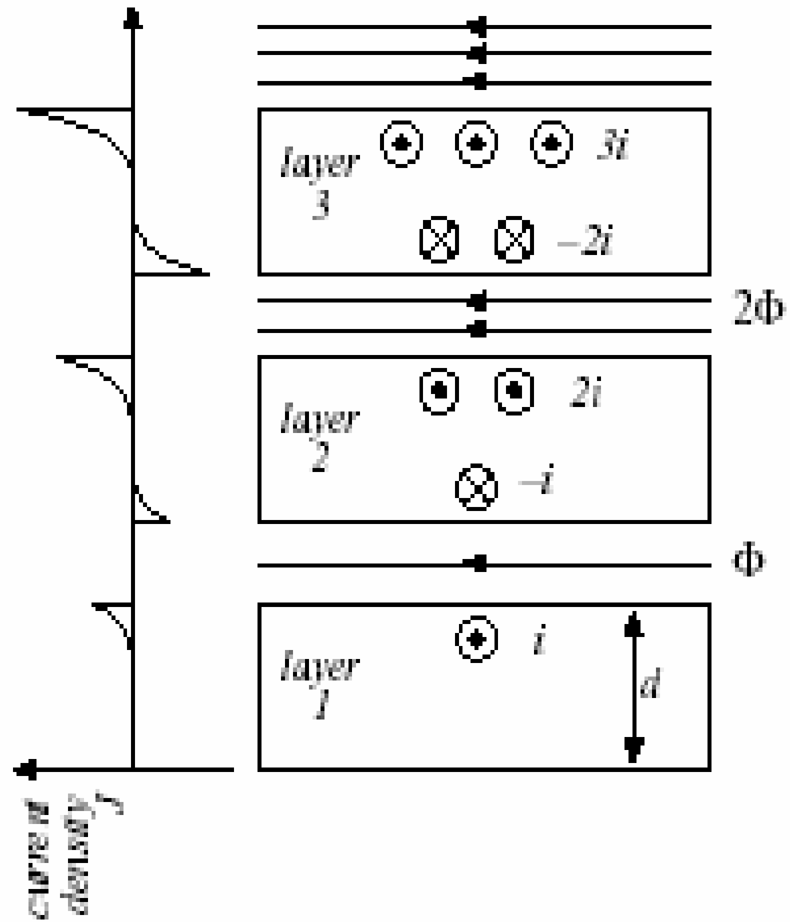
銅損 $$P_2 = I_{rms}^2 \left(R_{dc} \frac{d}{\delta}\right) + \left(2I_{rms}\right)^2 \left(R_{dc} \frac{d}{\delta}\right) = 5P_1$$

Power loss P_3 in layer 3 is:

銅損 $$P_3 = \left(2I_{rms}\right)^2 \left(R_{dc} \frac{d}{\delta}\right) + \left(3I_{rms}\right)^2 \left(R_{dc} \frac{d}{\delta}\right) = 13P_1$$

Power loss P_m in layer m is:

銅損 $$P_m = ((m-1)^2 + m^2)\, P_1$$



M-層繞組的全損失:高頻限制

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Add up losses in each layer:

$$P_w\big|_{d>>\delta} = \sum_{j=1}^{M} P_j = \frac{M}{3}\left(2M^2+1\right)P_1$$ M層繞組的全銅損

Compare with dc copper loss:

If foil thickness were $d = \delta$, then at dc each layer would produce copper loss P_1. The copper loss of the M-layer winding would be

$$P_{w,dc}\big|_{d=\delta} = M\,P_1$$

若d=δ時,在dc或低頻,接近效應可略,每層的銅損只有P_1,故M層繞組的全銅損為MP_1

For foil thicknesses other than $d = \delta$, the dc resistance and power loss are changed by a factor of d/δ. The total winding dc copper loss is

若d≠δ $$P_{w,dc} = M\,P_1\frac{\delta}{d}$$ M層繞組在低頻的全銅損

So the proximity effect increases the copper loss by a factor of

$$F_R\big|_{d>>\delta} = \frac{P_w\big|_{d>>\delta}}{P_{w,dc}} = \frac{1}{3}\frac{d}{\delta}\left(2M^2+1\right)$$ 增加銅損的因素

圓形導體的近似型

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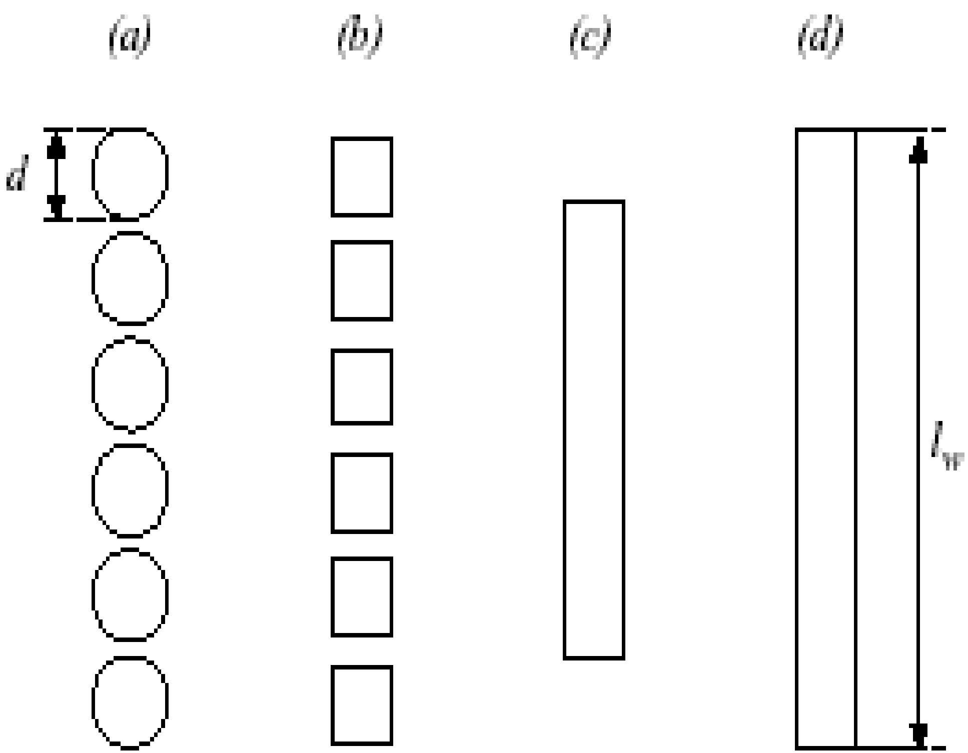


Conductor spacing factor:

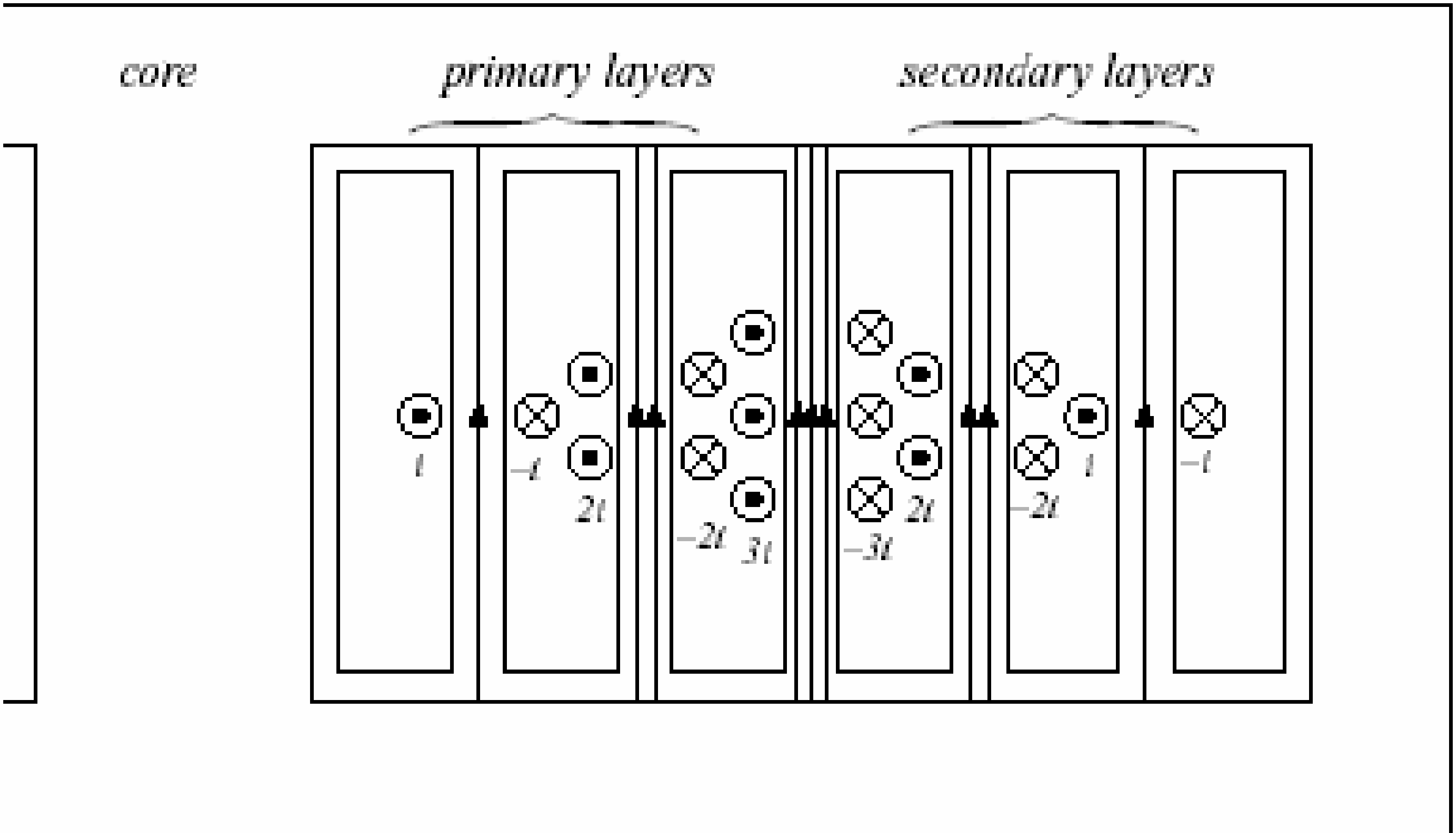
$$\eta = \sqrt{\frac{\pi}{4}}\, d\, \frac{n_l}{l_w}$$

Effective ratio of conductor thickness to skin depth:

$$\varphi = \sqrt{\eta}\ \frac{d}{\delta}$$

繞組導體相鄰的磁場強度:mmf

Two-winding transfomer example



變壓器例:磁力線分布

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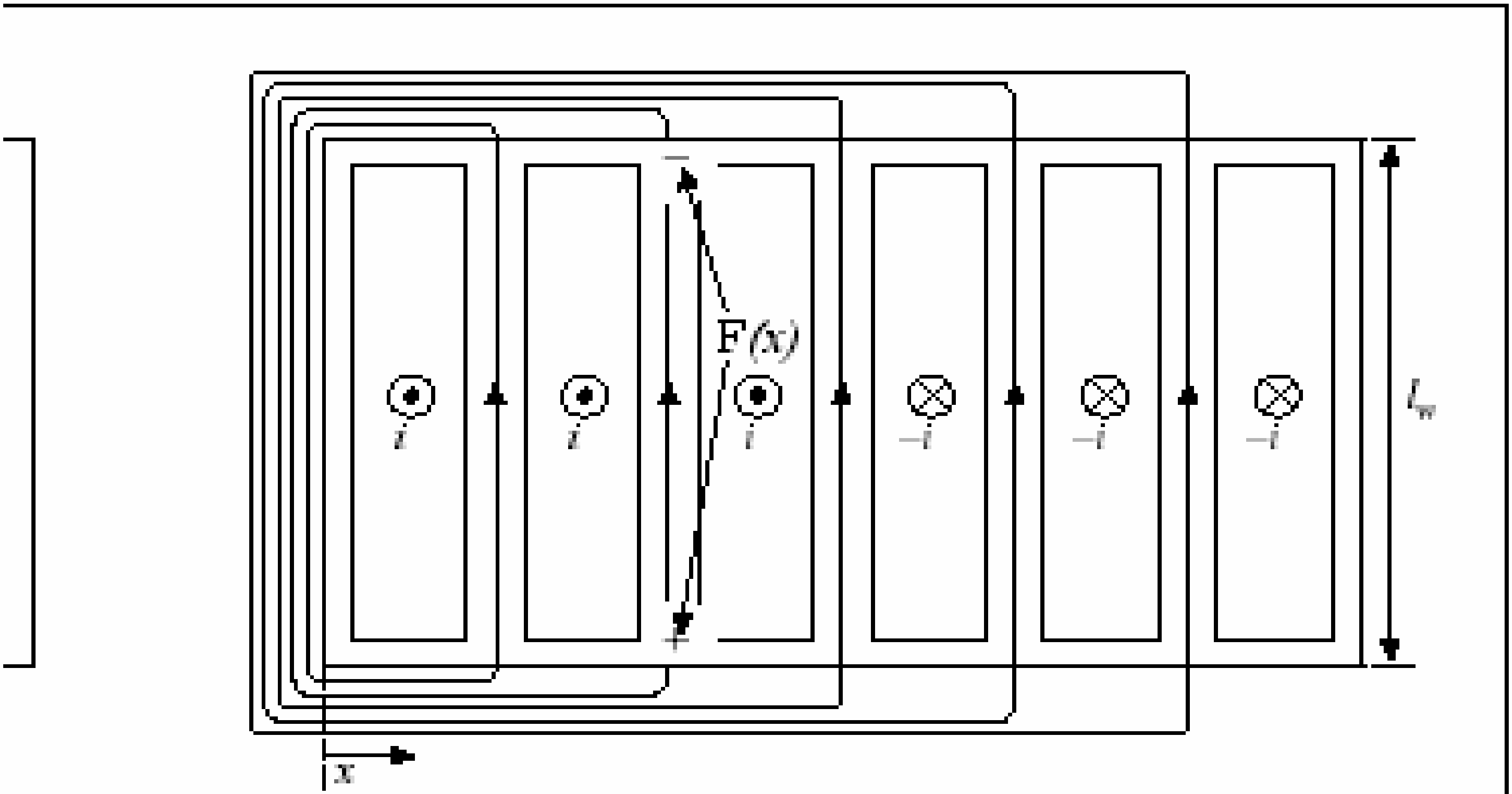
$$(m_p - m_s)\, i = F(x) \qquad H(x) = \frac{F(x)}{l_w}$$



安培定律和磁動勢圖

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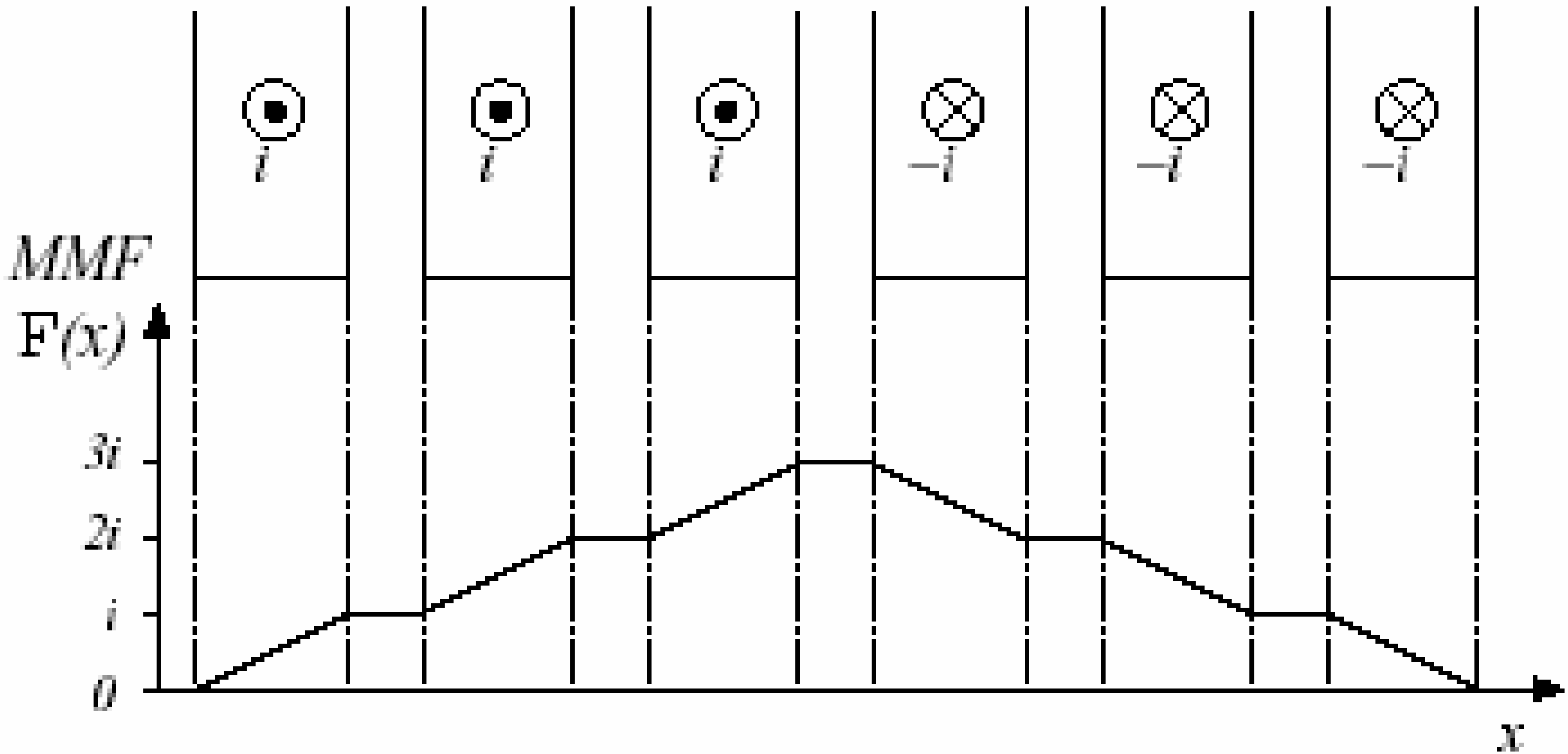

$$\left[m_p - m_s\right] i = F(x)$$

$$H(x) = \frac{F(x)}{l_w}$$

線直徑大於集膚深度的MMF圖

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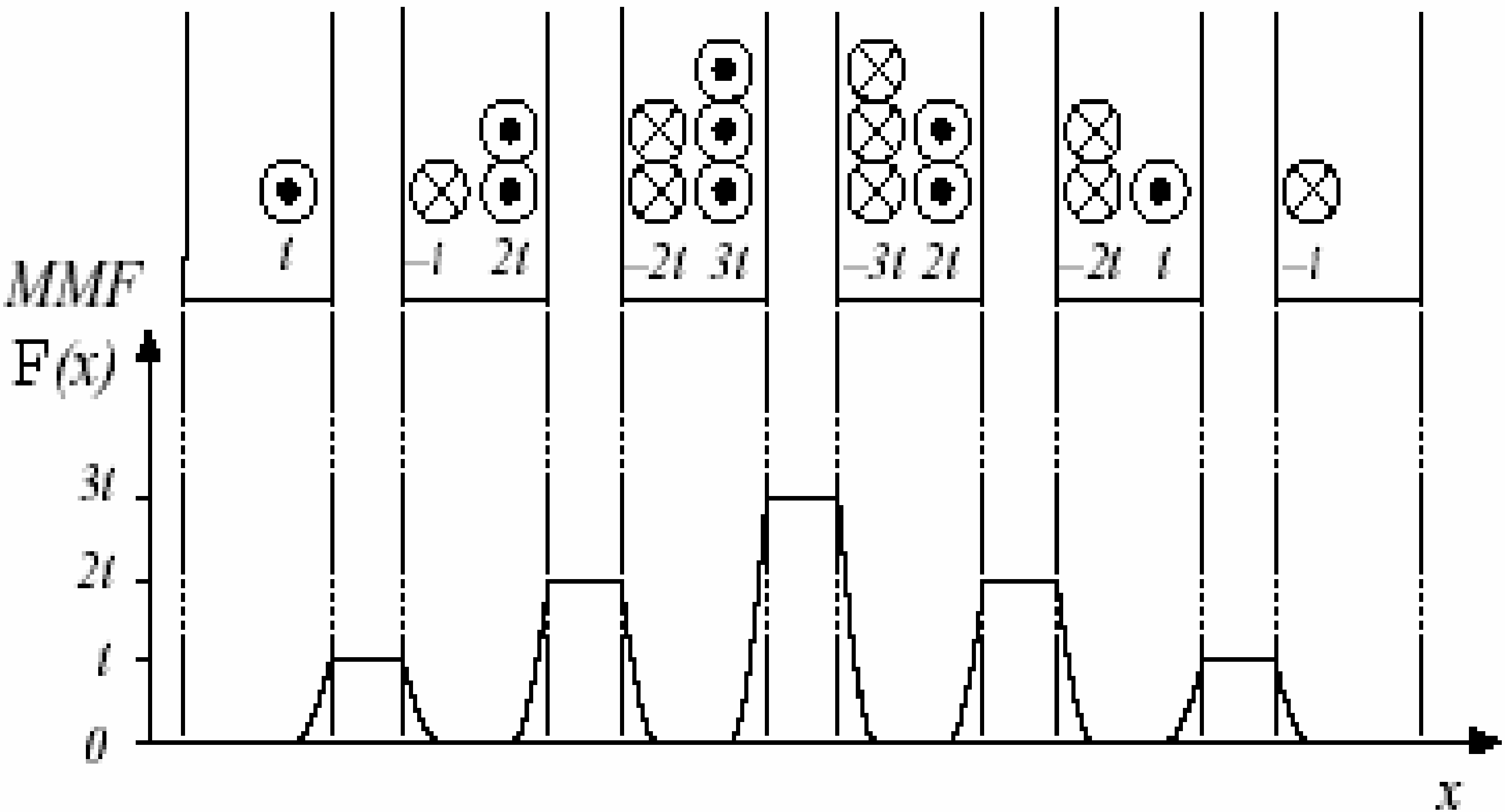

交錯繞組(Interleaved windings)

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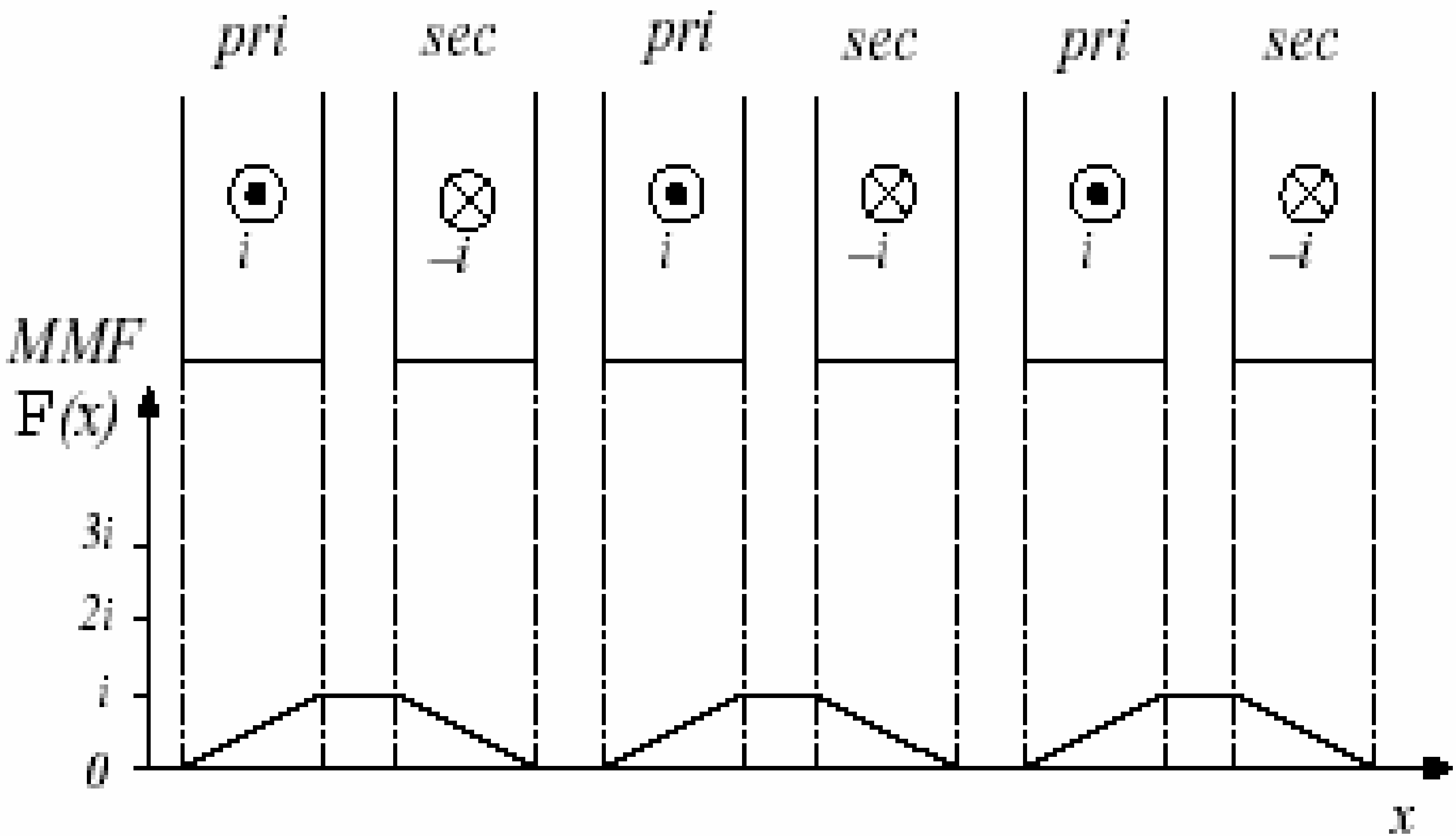

部分交錯繞組
(Partially-interleaved windings)

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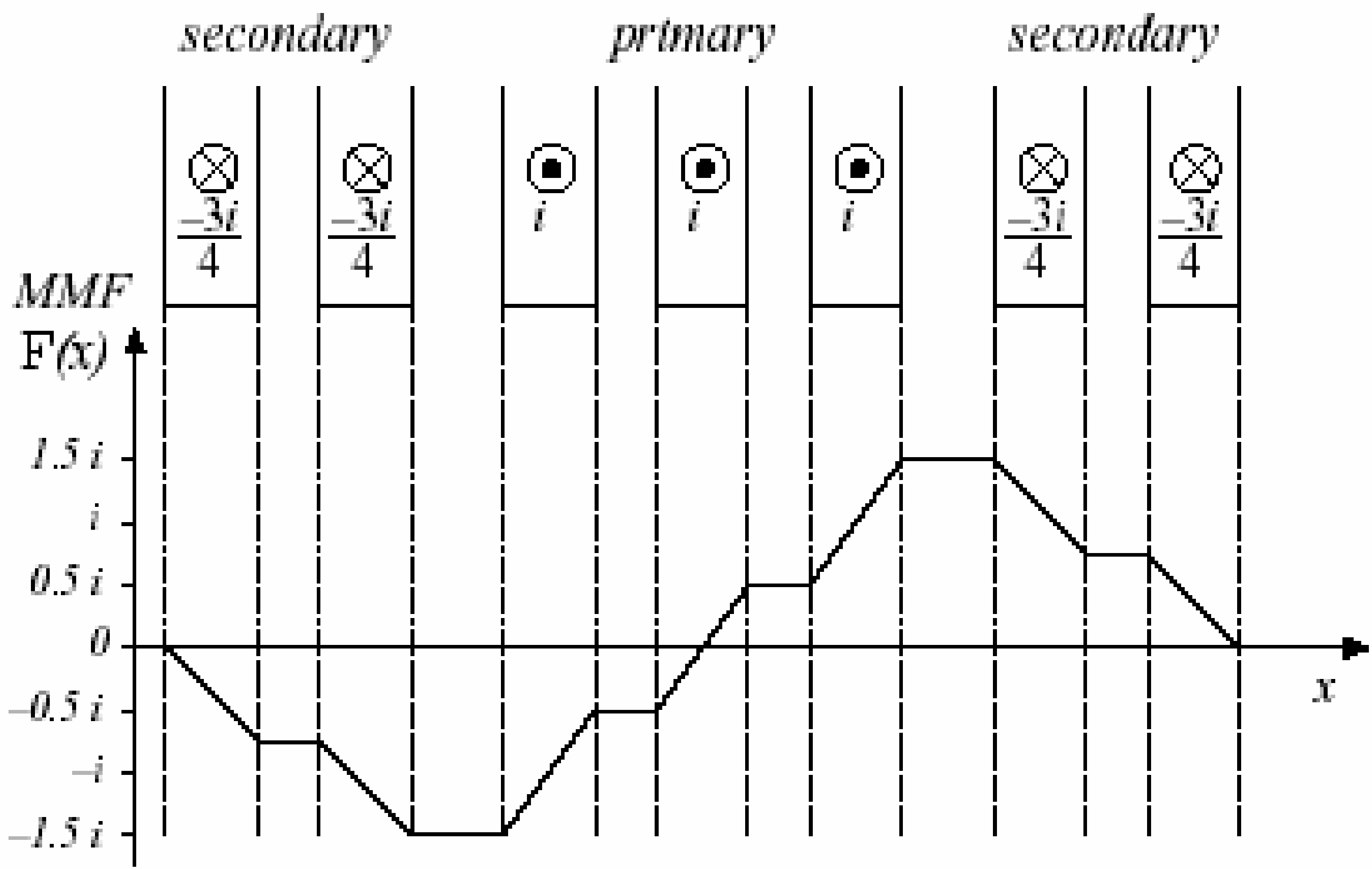

繞組層的功率損失

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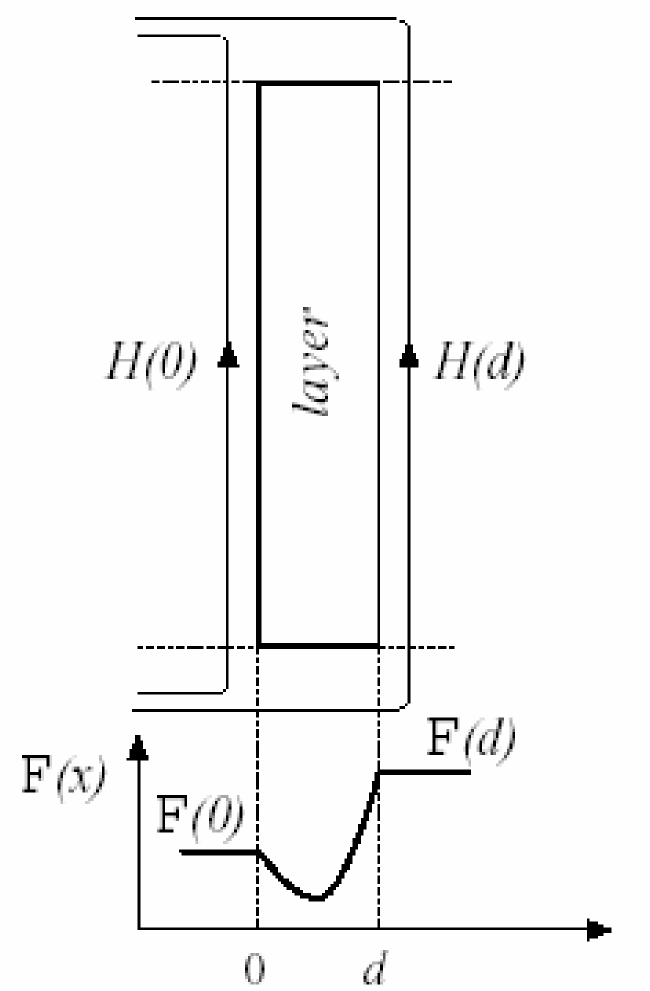
- 假設層的表面磁場為均勻,*H(0)*及*H(d)*
- *H(o)*及*H(d)*是由*F(0)F(d)*所驅動
- *H(o)*及*H(d)*是同相



繞組層的銅損解

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由Maxwell方程式解得

$$P = R_{dc}\,\frac{\varphi}{n_l^2}\left[\left(F^2(d) + F^2(0)\right) G_1(\varphi) - 4\,F(d)F(0)\,G_2(\varphi)\right]$$

where

$$R_{dc} = \rho\,\frac{(MLT)\,n_l^2}{l_w\,\eta\,d}$$

$$G_1(\varphi) = \frac{\sinh(2\varphi) + \sin(2\varphi)}{\cosh(2\varphi) - \cos(2\varphi)}$$

$$G_2(\varphi) = \frac{\sinh(\varphi)\cos(\varphi) + \cosh(\varphi)\sin(\varphi)}{\cosh(2\varphi) - \cos(2\varphi)}$$

$$\varphi = \sqrt{\eta}\,\frac{d}{\delta} \qquad \eta = \sqrt{\frac{\pi}{4}}\,d\,\frac{n_l}{l_w}$$

n_l = number of turns in layer,
R_{dc} = dc resistance of layer,

(MLT) = mean-length-per-turn, or circumference, of layer.

繞組電流 *I*, 每層有 n_1圈

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If winding carries current of rms magnitude I, then

$$F(d) - F(0) = n_1 I$$

Express F(d) in terms of the winding current I, as

$$F(d) = m\, n_1 I$$ m為mmf *F(d)*對 $n_1 I$之比值

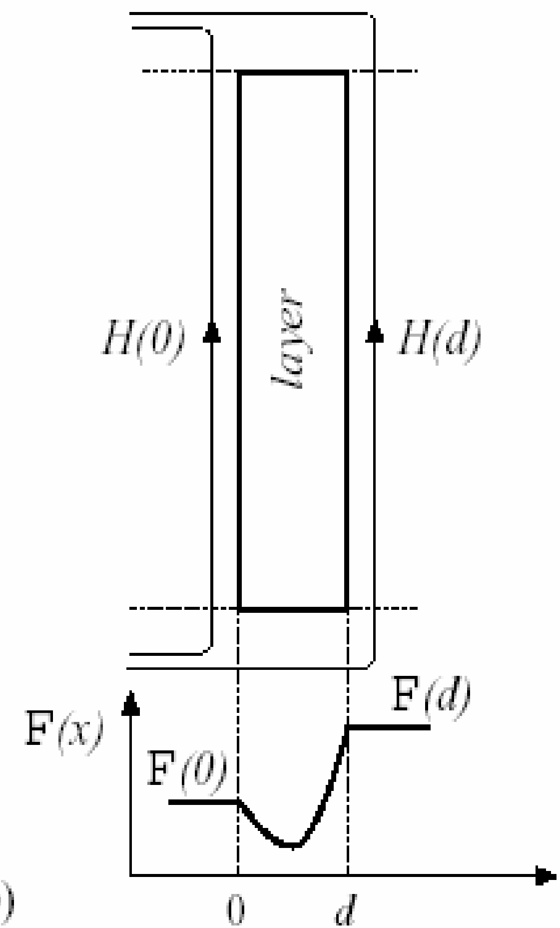
The quantity m is the ratio of the MMF F(d) to the layer ampere-turns $n_1 I$. Then,

$$\frac{F(0)}{F(d)} = \frac{m-1}{m}$$

Power dissipated in the layer can now be written

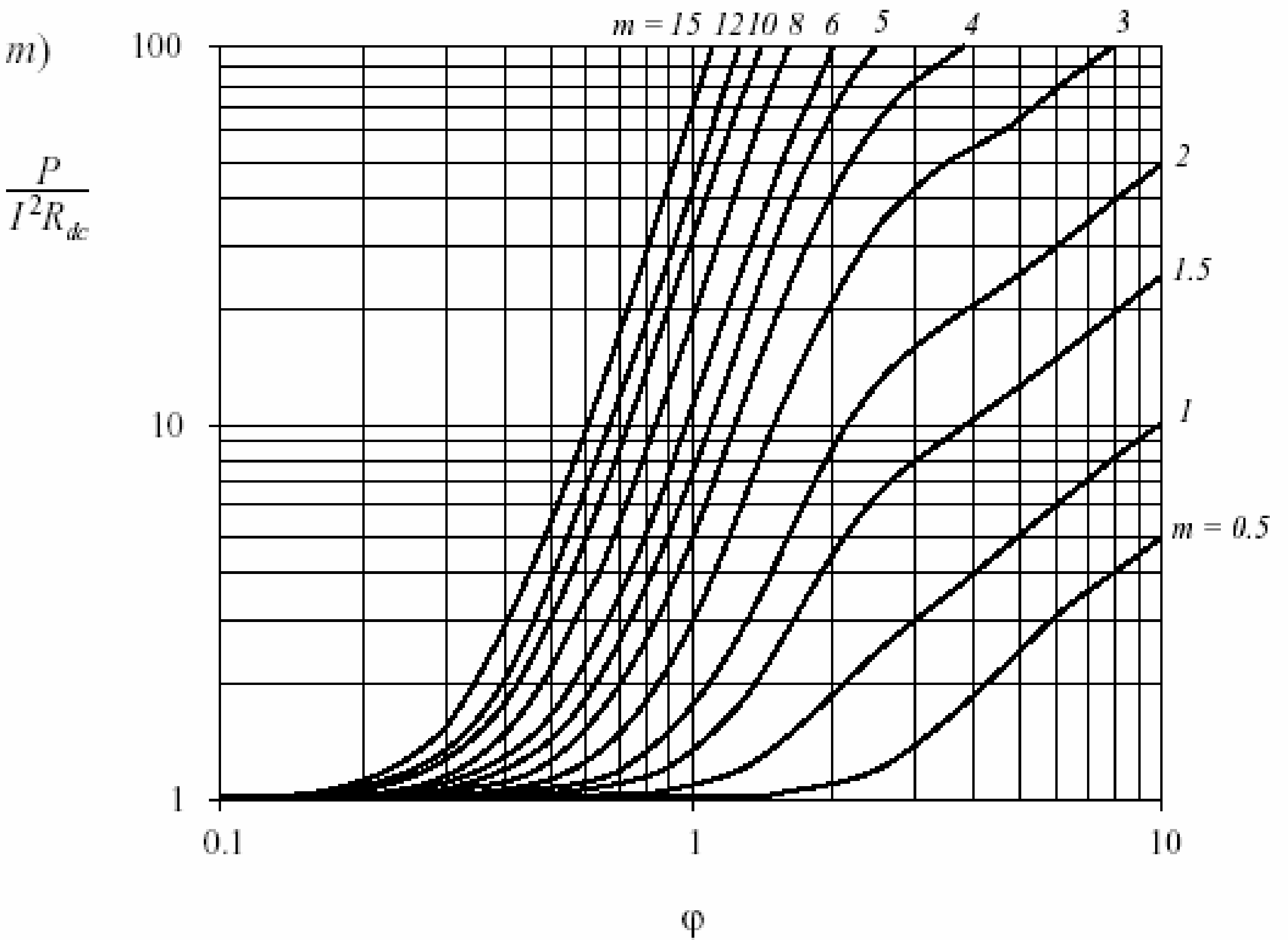
$$P = I^2 R_{dc}\, \varphi\, Q'(\varphi, m)$$

$$Q'(\varphi, m) = \left(2m^2 - 2m + 1\right) G_1(\varphi) - 4m\left(m-1\right) G_2(\varphi)$$



繞組層的銅損增量

$$\frac{P}{I^2 R_{dc}} = \varphi\, Q'(\varphi, m)$$



層銅損對層厚度

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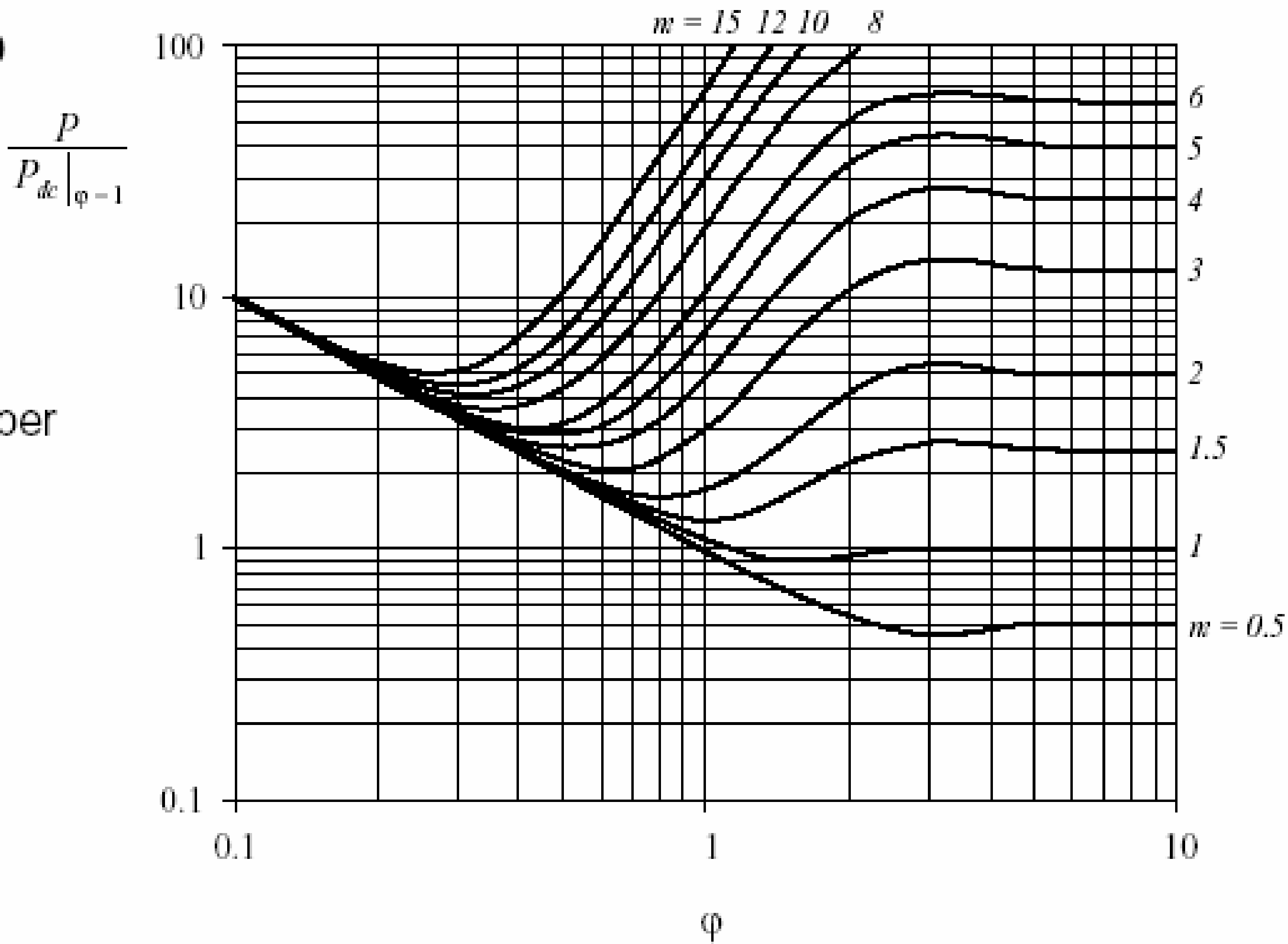
$$\frac{P}{P_{dc}\big|_{d=\delta}} = Q'(\varphi, m)$$



Relative to copper loss when $d = \delta$

例:在變壓器繞組的功率損失

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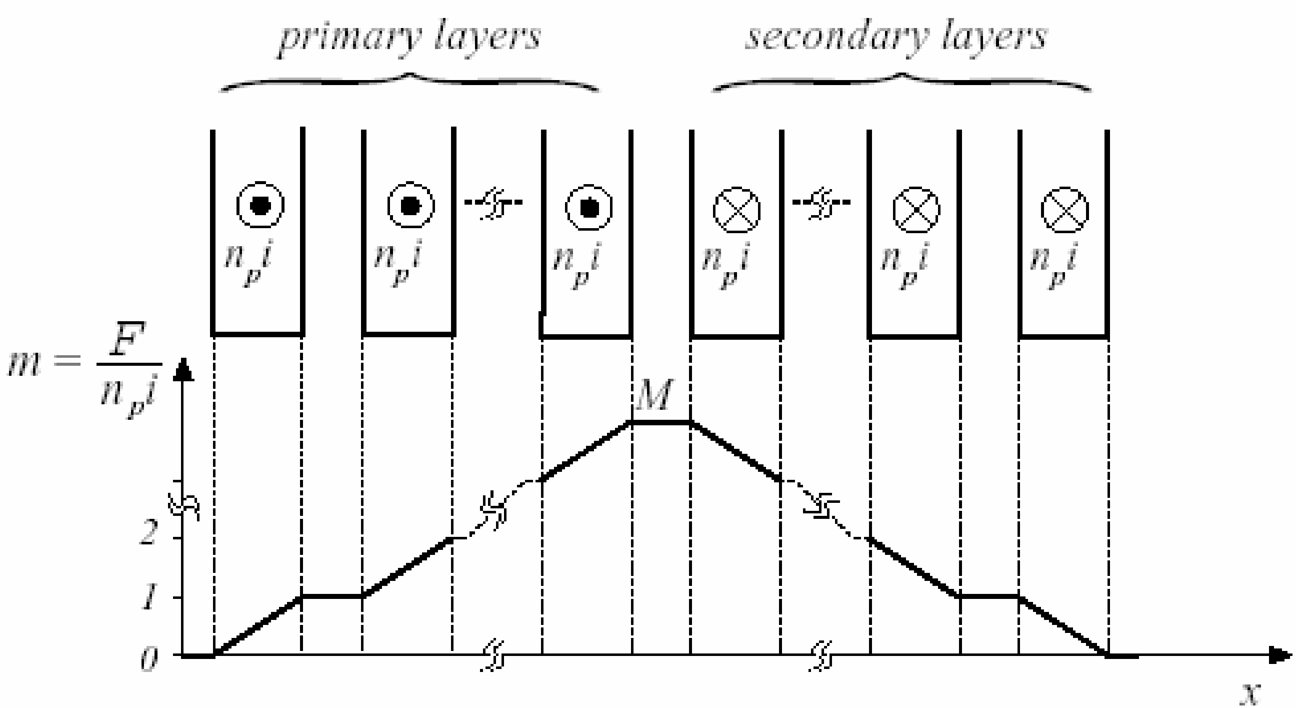
因Proximity effect, 第m層的銅損因數 $\frac{P}{I^2 R_{dc}} = \varphi\, Q'(\varphi, m)$

整個初級層的損失 $F_R = \frac{P_{pri}}{P_{pri,dc}} = \frac{1}{M} \sum_{m=1}^{M} \varphi\, Q'(\varphi, m)$



增加的全部繞組損失 $F_R = \varphi\left[G_1(\varphi) + \frac{2}{3}\left(M^2 - 1\right)\left(G_1(\varphi) - 2G_2(\varphi)\right)\right]$

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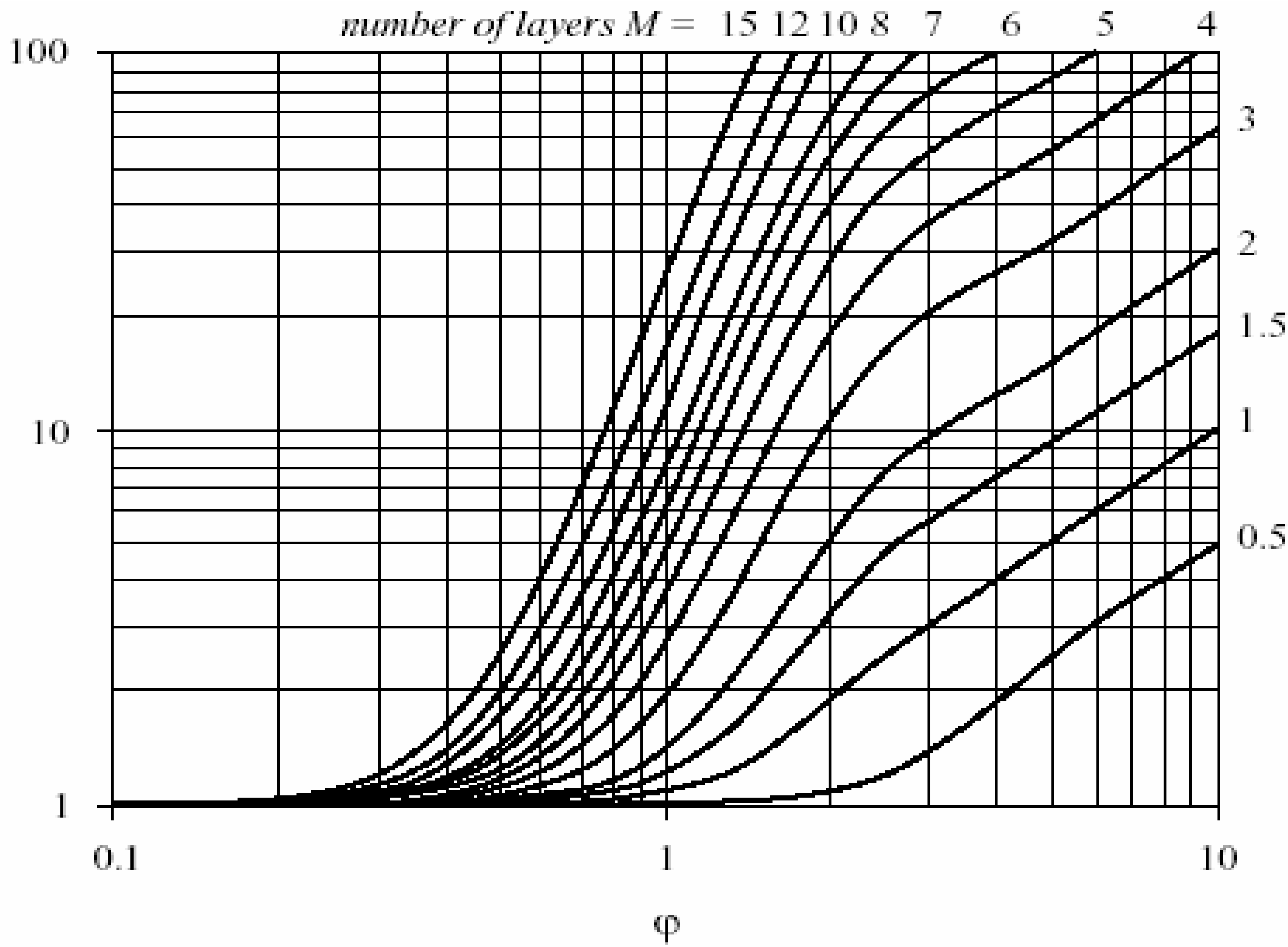
$$F_R = \frac{P_{pri}}{P_{pri,dc}}$$



全部繞組損失

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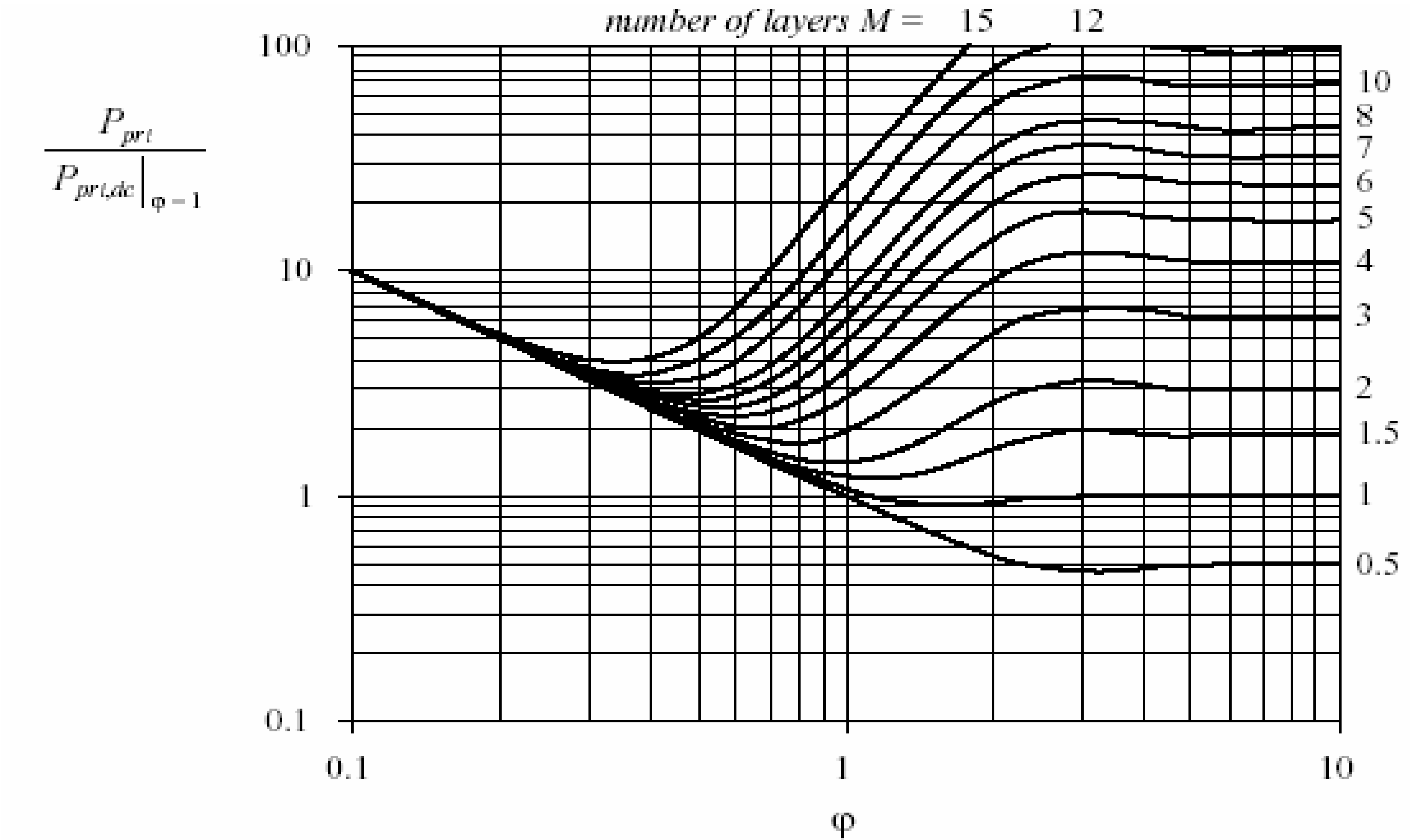


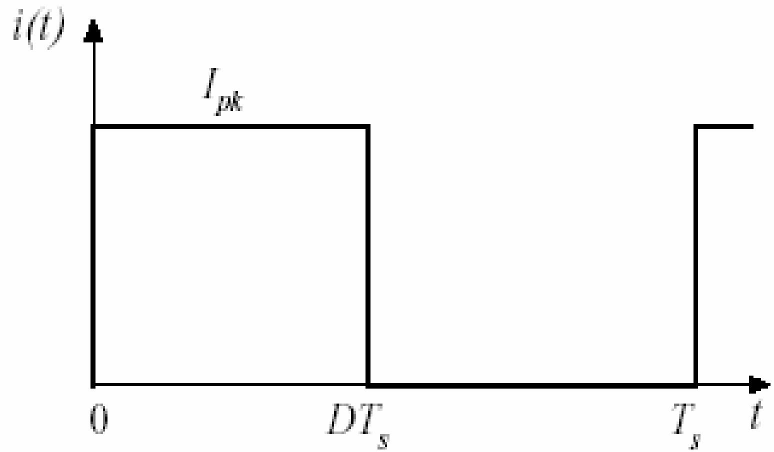
PWM的諧波

Fourier series:

$$i(t) = I_0 + \sum_{j=1}^{\infty} \sqrt{2}\, I_j \cos(j\omega t)$$

with

$$I_j = \frac{\sqrt{2}\, I_{pk}}{j\,\pi} \sin(j\pi D) \qquad I_0 = D I_{pk}$$



Copper loss:

Dc $\quad P_{dc} = I_0^2 R_{dc}$

Ac $\quad P_j = I_j^2 R_{dc} \sqrt{j}\, \varphi_1 \left[G_1(\sqrt{j}\, \varphi_1) + \frac{2}{3}\left(M^2 - 1\right)\left(G_1(\sqrt{j}\, \varphi_1) - 2G_2(\sqrt{j}\, \varphi_1)\right) \right]$

Total, relative to value predicted by low-frequency analysis:

$$\frac{P_{cu}}{D\, I_{pk}^2 R_{dc}} = D + \frac{2\varphi_1}{D\pi^2} \sum_{j=1}^{\infty} \frac{\sin^2(j\pi D)}{j\sqrt{j}} \left[G_1(\sqrt{j}\, \varphi_1) + \frac{2}{3}\left(M^2 - 1\right)\left(G_1(\sqrt{j}\, \varphi_1) - 2G_2(\sqrt{j}\, \varphi_1)\right) \right]$$

諧波損失因數

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諧波效應: F_H為全部的AC銅損對基本銅損

$$F_H = \frac{\sum_{j=1}^{\infty} P_j}{P_1}$$

全部繞組銅損:

$$P_{cu} = I_0^2 R_{dc} + F_H F_R I_1^2 R_{dc}$$

由D=0.5 PWM諧波引起的接近損失

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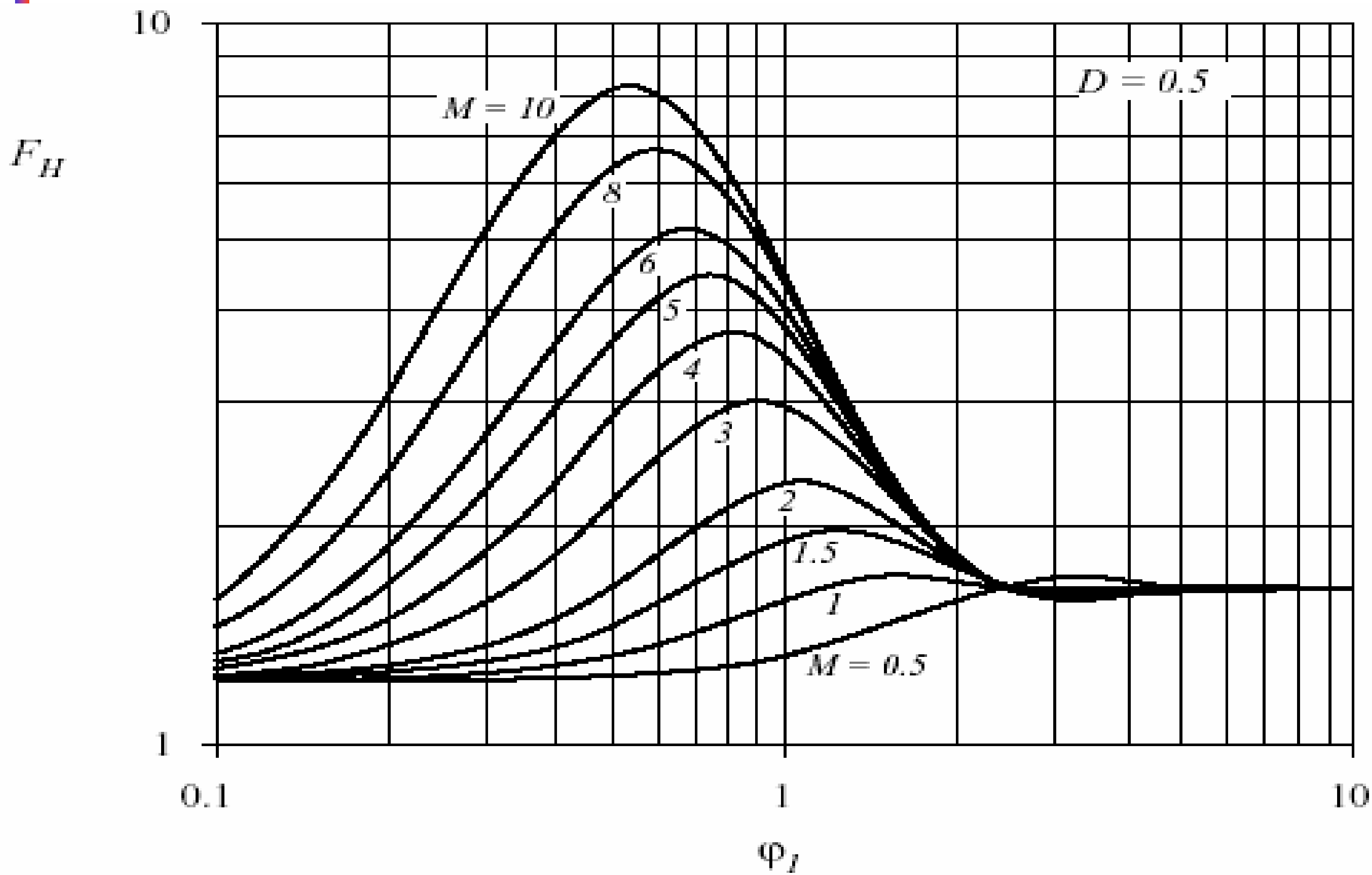

由D=0.3 PWM諧波引起的接近損失

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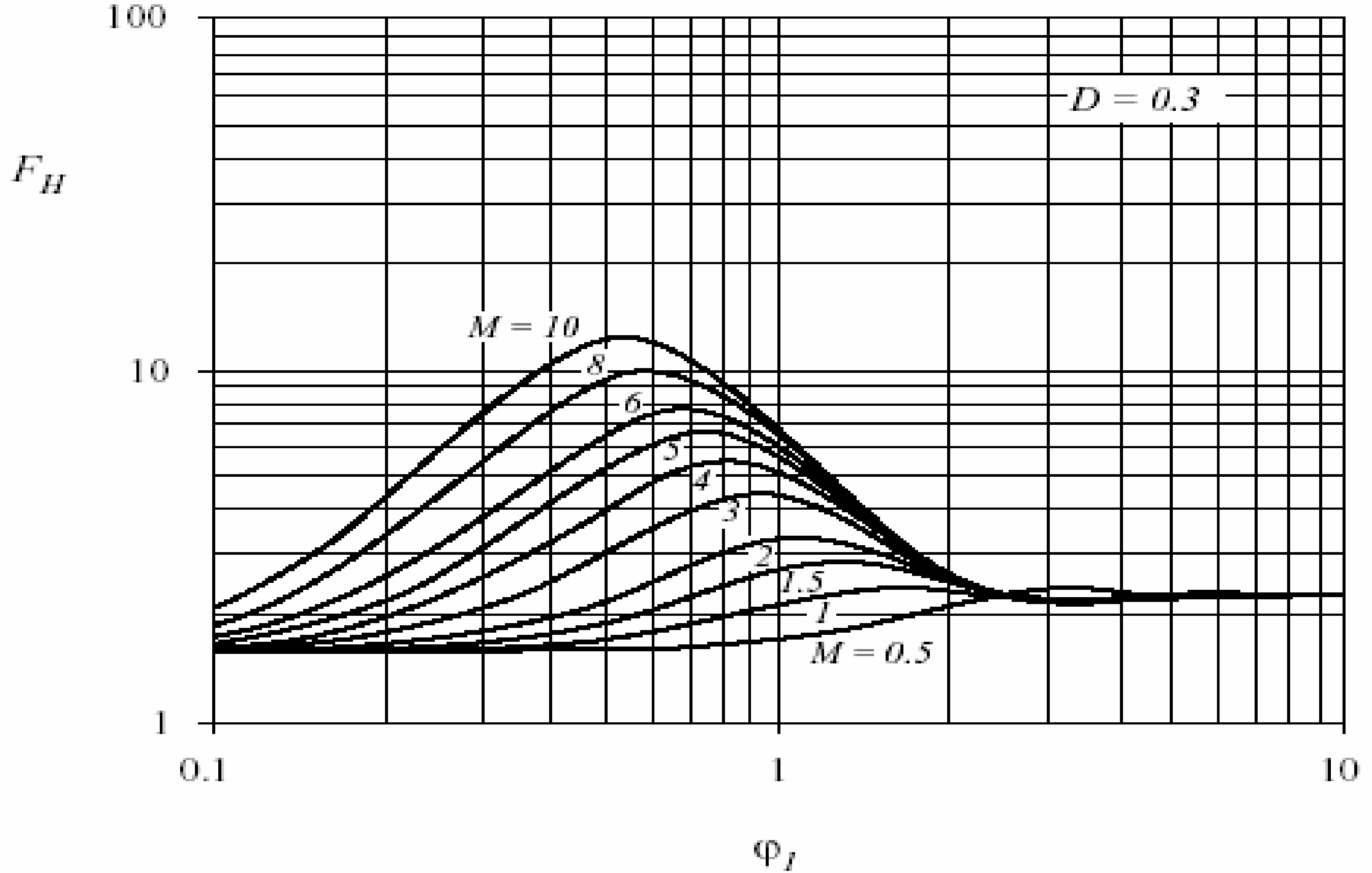

由D=0.1 PWM諧波引起的接近損失

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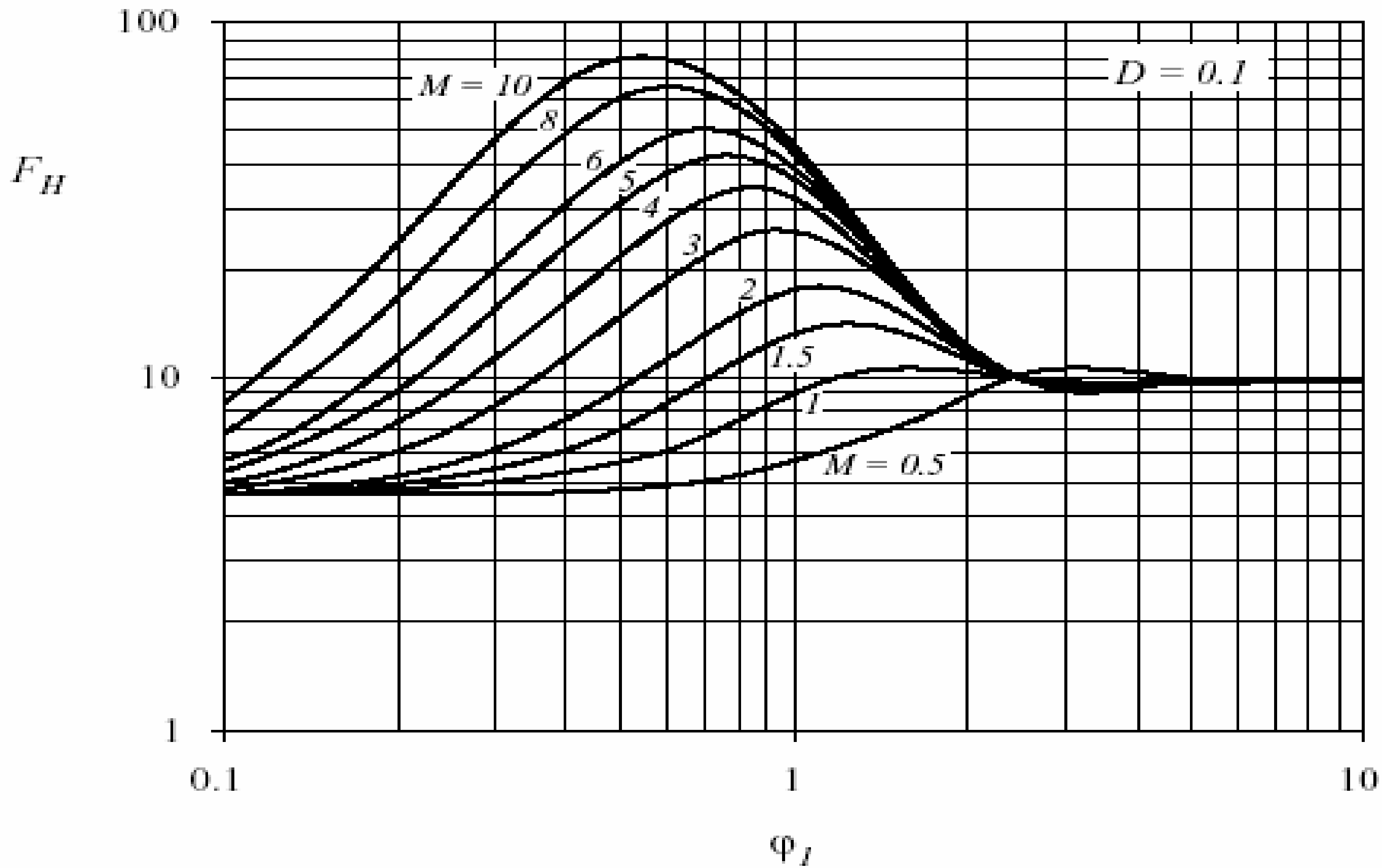

摘要

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- 在已知最大電流流經電感時,電感中加氣隙可防止飽和
- 理想變壓器其磁阻為零;磁阻不為零時,有激磁電感存在
- 當外加電壓秒(**Volt-second**)過大時,傳統變壓器會飽和,增加氣隙無助於改善飽和;只有增加鐵心的截面積或增加初級圈數才能防止飽和.
- 磁性材料的鐵心損是由**B-H**磁滯損和渦流損引起的.
- 實用的鐵心必須在高B_{sat}和高鐵心損P_{fe}間做一項取捨

摘要con’t

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- **Laminated iron alloy core: highest B_{sat}和 highest P_{fe}**
- **Ferrite cores:lowest B_{sat}和lowest P_{fe}**
- **Powdered iron alloy core 和amorphous alloy core**責介於前兩者之間
- **Skin和Proximity effect**會導致繞組導體有渦流,此現象會在高電流高頻磁性裝置中增加銅損
- 當一導體其厚度接近或大於穿入深度時,該導體附近的磁場會感應渦流在導體上