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# SPECIAL TRANSFORMERS FOR THE MEASUREMENT OF HARMONIC NOISE OF D.C. CURRENTS

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# Special tansformers for the measurement of harmonic noise of DC currents

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**ABSTRACT**: some design problems regarding a special current transformer are presented and discussed: the transformer must be suited for the measurement of the harmonic pollution of D.C. currents generated or absorbed by static converters; for example, situations of this kind occur in the stationary and in the on board systems of the D.C. traction plants.

## 1. INTRODUCTION.

In several D.C. plants, connected with A.C. plants by static converters, an important problem is the limitation and the measurement of the D.C. side harmonic pollution generated by the converters. On the one hand, in order to limit this pollution, especially dovoted power filters are frequently emploied; on the other hand, sometimes the verification of the levels of the harmonic content of the currents on the D.C. side requires the measurement of alternative components considerably smaller than the direct component, in a frequency range within a few Hz to some kHz.

A typical situation is that of the D.C. traction lines: for example, the limits of the harmonic noise for the on board conversion units (such as the auxiliary inverters) impose[1] that the maximum allowed amplitude of the alternative components be of the order of $10^{-3} \div 10^{-4}$ times the direct component, from a few Hz up to 10 kHz (fig.1).

Usually the precision class of the current transducers is given in terms of the maximum value of the measured current: considering that the amplitude of the harmonic components is low compared with the average value, the accuracy with which the current harmonic content can be estimated is greatly limited.

On the other hand, the presence of the direct component at the output terminals of the transducer, if not suitably filtered, implies several problems, in particular for the harmonic analysers: in fact, also their precision and resolution are related to the maximum instantaneous value of the signal to be analysed, then the instrument errors are affected by the direct component too.

The use of a current transformer (C.T.), suitably designed, as an interface device between the power circuit and the measurement system, involves certain advantages: the galvanic insulation (for safety reasons and for the de-coupling of the acquisition channels of the instruments) and the filtering of the direct component (obtained without high-pass filters added).

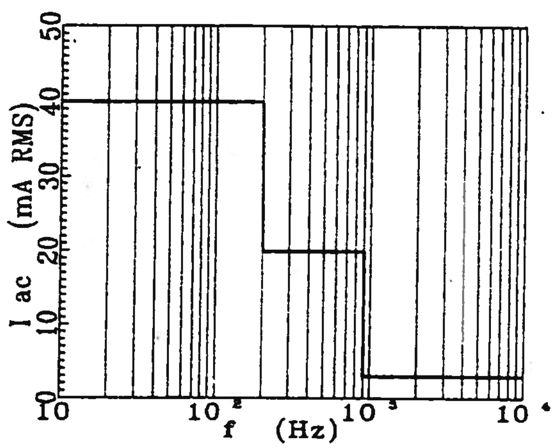


Fig.1 - Allowed limit threshold values of the harmonic pollution generated by auxiliary static converters for on board traction plants: line voltage: 3000 V d.c.; converter rating: 45 kVA.

The design problems are numerous: on one hand the high direct current requires the adoption of gapped cores, in order to prevent the magnetic saturation; on the other hand, the existence of a gap implies high values of the magnetizing components of the harmonic currents.

Leaving the winding and core data out of consideration, the highest errors occur at low frequencies, where the primary current is mostly magnetizing, and at high frequencies, where the leakage secondary reactance grows and the capacitive phenomena occur.

## 2. BEHAVIOUR OF MAGNETIC MATERIALS WITH D.C. AND A.C. FIELDS SUPERPOSED.

The presence of a great direct component in the current to be analysed causes a magnetic behaviour quite different from that in A.C. operation: as known, the A.C. operation of a magnetic material is described by the normal magnetization curve, or the curve $\mu=\mu(B)$, where $\mu$ is called normal permeability, or, simply, permeability.

On the contrary, the experience shows that, given an alternating magnetic strength $\Delta H$ and a biasing constant magnetic strength ($H_{dc}$) superposed, the alternative flux density $\Delta B$ is not $\mu$ times $\Delta H$, but sensibly lower: in this case there are asymmetrical hysteresis cycles having a slope defined by another permeability coefficient, called incremental permeability ($\mu_\Delta$): $\mu_\Delta$ decreases with the increase of $H_{dc}$ (fig.2.a), and it increases with the increase of $\Delta H$ (fig.2.b).

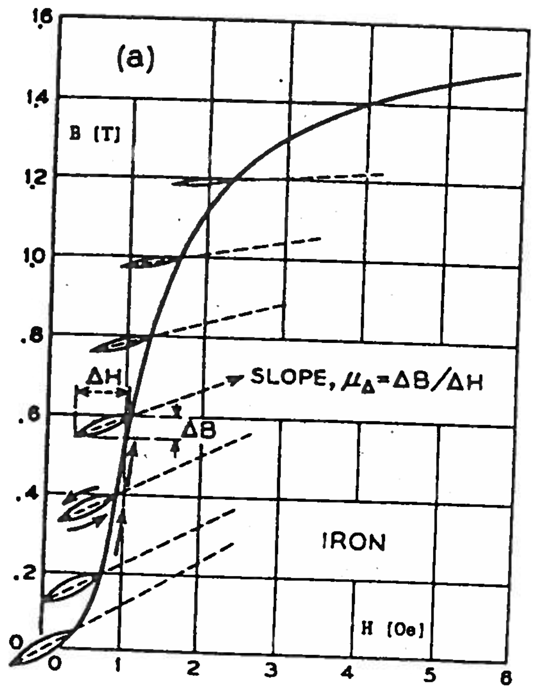


Fig.2.a.- Minor hysteresis loops in iron[2]: $\mu_\Delta$ (slope of the loops) as a function of the biasing field strenght ($H_{dc}$), being constant the alternating magnetic strenght ($\Delta H$).

1 Oe = $10^3/(4\cdot\pi)$ A/m

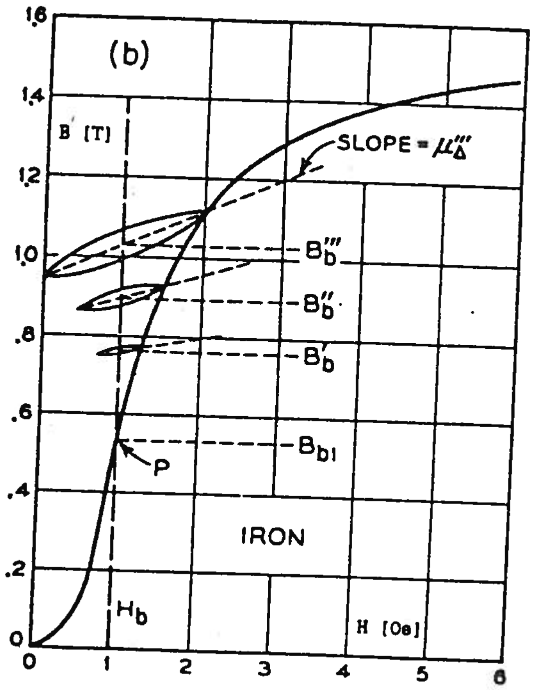


Fig.2.b.- Minor hysteresis loops in iron[2]: $\mu_\Delta$ (slope of the loops) as a function of the alternating magnetic strenght ($\Delta H$), being constant the biasing field strenght ($H_{dc}$).

It is defined reversible permeability ($\mu_{rev}$) the following quantity:

$$\mu_{rev} = \lim_{\Delta H\to 0}\left(\frac{\Delta B}{\Delta H}\right)_{H_{dc}=const} = \lim_{\Delta H\to 0}\mu_\Delta \qquad (1)$$

Considering the low amplitude of the alternative components that occur in the present case, in the following we will refer to the reversible permeability rather than to the incremental permeability.

It is defined initial permeability the quantity:

$$\mu_i = \lim_{\Delta H\to 0}\left(\frac{\Delta B}{\Delta H}\right)_{H_{dc}=0} = \lim_{\Delta H\to 0}\mu \qquad (2)$$

On the basis of some theoretical-experimental studies[2] we can assume that the reversible permeability obeys the following law:

$$\mu_{rev} \simeq \mu_i\cdot\left(1-\frac{B_{int}}{B_{sat}}\right)^k \qquad (3)$$

where:

- $B_{sat}$ is the material saturation flux density;
- $B_{int}$ is the intrinsic biasing flux density, that can be assumed equal to:

  $B_{int} = B_{dc}-\mu_o\cdot H_{dc} \simeq B_{dc}\cdot(1-\mu_o/\mu(B_{dc}))$;

  considering that usually $B_{dc}$ is quite lower than saturation value $B_{sat}$, we have $\mu(B_{dc}) \gg \mu_o$, from which: $B_{int} \simeq B_{dc}$;
- k is an exponent depending on the kind of the magnetic material: its value is included in the range 0.75÷2.

From eq. (3) we can deduce that the most suited magnetic material has high values of initial permeability $\mu_i$ and saturation flux density $B_{sat}$.
During the design, the adoption of a material with high values of $B_{sat}$ and $\mu_i$ allow to choose a high biasing flux density $B_{dc}$ (so that it is not necessary to adopt an air-gap too much wide), without reducing too much the reversible permeability.

The examination of the characteristics of the materials commercially available suggests that materials having the best values of these parameters are not silicon-iron alloys, but Mumetal alloys (fig.3), that have:

$$\mu_i \simeq 25\cdot 10^3\cdot\mu_0;\ B_{sat} \simeq 0,7\ T;\ k \simeq 0,75.$$

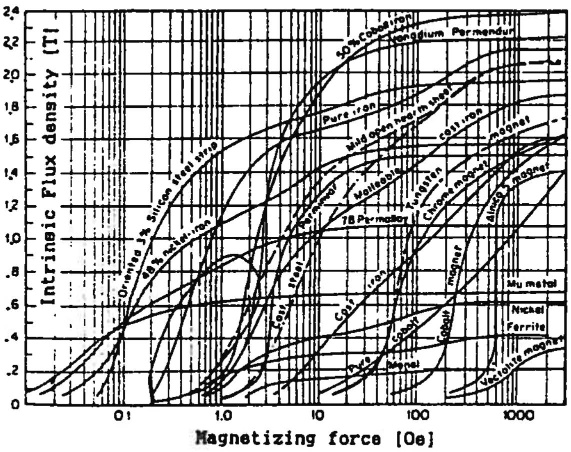


Fig.3 - Typical normal magnetization curves in a wide range of commercial magnetic materials.

## 3. THE D.C. POLARIZED C.T. IN THE OPERATION AT LOW FREQUENCY.

The high value of the direct current component in the C.T. primary winding imposes the adoption of a gapped core, in order to limit the saturation level of the material.
This structural characteristic is typical also for the current transformers designed for the measurement of transient and fault currents: about these devices, a wide literature and Standard references can be found.
In the D.C. polarized C.T. the comparative levels of the harmonic components with respect to the D.C. component implies some peculiar characteristics: because of this, the study of the design and of the model of this kind of device deserves a special analysis.

During operation at low and medium frequencies (when the effects of the parasitic capacitances are negligible) the parameters of the "T" equivalent circuit of a C.T. are, as known, the following (referred to the same number of turns):

- primary series impedance: $\bar{Z}_1 = R_1 + j\cdot\omega\cdot L_1$;
- secondary series impedance: $\bar{Z}_2 = R_2 + j\cdot\omega\cdot L_2$;
- parallel admittance: $\bar{Y}_0 = G_0 - j/(\omega\cdot L_0)$.

Called $R_m$ the value of the measurement resistance, for each harmonic component the current ratio $I_2/I_1$ (RMS values) equals:

$$\frac{I_2}{I_1} = \frac{N_1/N_2}{\sqrt{\left[1+G_0\cdot(R_2+R_m)+\frac{L_2}{L_0}\right]^2+\left[G_0\cdot\omega\cdot L_2-\frac{R_2+R_m}{\omega\cdot L_0}\right]^2}} \ . \quad (4)$$

It is evident that the error increases as the frequency decreases: then, the maximum acceptable error at a minimum reference frequency must be indicated among the design specifications; in this situation, eq. (4) can be rewritten as follows:

$$\frac{I_2}{I_1} \simeq \frac{N_1/N_2}{\sqrt{1+\left[\frac{R_2+R_m}{\omega\cdot L_0}\right]^2}} \ . \quad (5)$$

The use of eq. (5) instead of eq. (4) corresponds to neglecting the effects of the leakage flux linked with the secondary winding compared with the main flux (in terms of parameters this means to assume $L_2 \ll L_0$), and to assuming negligible the magnetic material losses compared with the magnetizing reactive power (considering the presence of the air-gap, that absorbs the prevalent part of the reactive power, the parallel branch can be practically considered purely reactive: $G_0 \ll 1/(\omega\cdot L_0)$).
Then eq. (5) is an approximate expression: nevertheless, its employment (partially not in compliance with some classical theories[3]) is acceptable for design calculations, referring to the low frequency range and considering the peculiarities of the considered device.

In general, for reasons due to construction needs, one can assume that the core is always provided with two air-gaps, having the same geometrical thickness $\delta$, disposed magnetically in series; moreover, the core has a geometrical lenght $\ell_n$ and a net magnetic core section $A_n$.
Defined an equivalent magnetic air-gap:

$$\delta_{eq} = \delta + (\ell_n/2)/(\mu_{rev}/\mu_0) \ , \quad (6)$$

(where $\mu_{rev}$ is the reversible permeability, evaluated by eq. (3)), the inductance $L_0$, referred to the turns number $N_2$, can be expressed as follows:

$$L_0 = N_2^2\cdot\mu_0\cdot A_n/(2\cdot\delta_{eq}) \ . \quad (7)$$

By substituting eq. (7) into eq. (5) one obtains:

$$\frac{I_2}{I_{1s}} \simeq \frac{1}{\sqrt{1+\beta^2}} \ , \quad (8)$$

where $I_{1s} = I_1\cdot N_1/N_2$ is the primary current referred to the secondary and $\beta$ equals:

$$\beta = \frac{2\cdot\delta_{eq}\cdot(R_2+R_m)}{\omega\cdot N_2^2\cdot\mu_0\cdot A_n} \ . \quad (9)$$

The quantity $\beta$ can be defined as "error function". Besides the current ratio, it is of interest to give the approximate expressions of other functional quantities:

- phase error $\gamma$:

$$\gamma = \angle \bar{I}_2 - \angle \bar{I}_1 \simeq \operatorname{arctg}(\beta) \;; \qquad (10)$$

- e.m.f. per turn $E_t$:

$$E_t = \frac{\omega \cdot \Phi}{\sqrt{2}} \simeq N_1 \cdot I_1 \cdot \frac{R_2 + R_m}{N_2^2} \cdot \frac{1}{\sqrt{1+\beta^2}} \;; \qquad (11)$$

- magnetizing current:

$$I_\mu \simeq I_1 \cdot \frac{\beta}{\sqrt{1+\beta^2}} \;; \qquad (12)$$

- voltage $V_m$ at the terminals of the measurement resistance $R_m$:

$$V_m \simeq \frac{N_1}{N_2} \cdot \frac{R_m \cdot I_1}{\sqrt{1+\beta^2}} \,. \qquad (13)$$

When $\beta$ increases, the magnetizing current and the ratio and phase errors increase too, while the e.m.f. per turn and the voltage at the measurement resistance terminals decrease: then, the more $\beta$ increases, the more the C.T. performances are getting worse.

With regard to the influence of the various parameters one can say:

- air-gap $\delta$: the presence of the air-gap is necessary in order to avoid an excessive reduction of the reversible permeability $\mu_{rev}$, caused by the direct component $I_{dc}$ of the primary current. As a matter of fact, the biasing flux density $B_{dc}$ can be practically assumed as follows:

$$B_{dc} \simeq \mu_0 \cdot N_1 \cdot I_{dc}/(2 \cdot \delta) \;; \qquad (14)$$

  on the other hand, a too high value of the air-gap increases $\beta$, worsening the performances;
- total secondary resistance: a high value of such a resistance $(R_2+R_m)$ causes greater C.T. errors; on the other hand, high values of the voltage $V_m$ can be obtained by high values of $R_m$;
- frequency: as the frequency increases, there is a decrease of the errors, at least in the low and medium frequency range (f<1 kHz, when the effects of the parasitic capacitances are negligible);
- number of turns $N_2$ of the secondary winding: the increase of $N_2$ implies a remarkable reduction of the errors, thanks to its quadratic influence; on the other hand, when $N_2$ increases, the voltage $V_m$ decreases proportionally to the inverse of $N_2$;
- section $A_n$ of the magnetic core: a high value of $A_n$ reduces the errors, but it requires a heavier, bulkier and more expensive design;
- middle geometric length $\ell_n$ of the magnetic core: it is implicit in the expression (6) of the equivalent air-gap $\delta_{eq}$; if the material is characterized by high values of $\mu_{rev}$, $\ell_n$ affects very weakly the errors: in this case the magnetic length of the core equals a few percent of the air-gap length. The definition of the geometric length $\ell_n$ depends on the area of the window necessary for the windings and on technological needs regarding the construction of the core.

One can observe that, from the qualitative point of view, some of these parameters have an influence on the design and on the operation of the C.T. that is similar to all the other kinds of C.T.; nevertheless, in this case this influence shows particular quantitative characteristics, due to the special-operation conditions.

## 4. GENERAL DIMENSIONING CRITERIA FOR THE DESIGN OF THE D.C. POLARIZED C.T..

The design begins from the specification of the ratio error, that can be defined as follows:

$$\varepsilon = 1 - \frac{I_2}{I_{1e}} \,. \qquad (15)$$

In particular, the error $\varepsilon$ to be considered is that $(\tilde{\varepsilon})$ corresponding to a certain minimum reference angular frequency $(\tilde{\omega})$; by manipulating eq. (8) and (15), one obtains the corresponding value of the error function $\tilde{\beta}$:

$$\tilde{\beta} = \frac{\sqrt{2 \cdot \tilde{\varepsilon} - \tilde{\varepsilon}^2}}{1 - \tilde{\varepsilon}} \,. \qquad (16)$$

With regard to the lamination thickness, the need of a linear response of the transducer in a frequency range up to some kHz requires the adoption of a maximum value of thickness equal 0.3 mm, in order to limit the dissipative and demagnetizing effects of the eddy currents.

With regard to the core design, the possible solutions could be numerous: the toroidal shape would be the best one (negligible leakage fluxes and immunity to disturbances); on the other hand, for constructive reasons it is better to adopt a "double C" shaped core. Besides, the two configurations are not so different each other (the "double C" shaped core is a toroidal core slightly squashed). For this reason, in the following a toroidal core, equivalent to the "double C" shaped one, will be considered: the equivalence means equal value of section and equal middle length. In this way, it is easy to obtain simple expressions of the quantities of interest.

A toroidal magnetic core having a rectangular section is characterized by the internal core diameter $(D_n)$ and by core widths in the radial direction $(a_n)$ and in the axial direction $(b_n)$. Let be $K_L = b_n/a_n$ the width ratio, $K_n = a_n/D_n$ the core coefficient and $K_{st}$ the stacking factor; the section of the core $A_n$ equals:

$$A_n = K_{ot} \cdot K_L \cdot K_n^2 \cdot D_n^2 \quad , \tag{17}$$

and the middle geometric length $\ell_n$ of the magnetic core equals:

$$\ell_n = \pi \cdot (1 + K_n) \cdot D_n \quad . \tag{18}$$

Eq. (17) and (18) show that the core shape is defined by the internal diameter $D_n$ and by the adimensional factors $K_L$ and $K_n$.

By referring the magnetic semilength of the core to the value of the air-gap $\delta$ by means of the coefficient $\lambda_n$, called core magnetic factor:

$$\lambda_n = [(\ell_n/2)/(\mu_{rev}/\mu_o)]/\delta, \tag{19}$$

from the comparison with eq.(6) one can write:

$$\delta_{eq} = (1 + \lambda_n) \cdot \delta \quad . \tag{20}$$

From eq. (19) and (20) one can observe that the parameter $\lambda_n$ has the same meaning of the saturation ratio, usually emploied for the study of the most kind of gapped magnetic circuits.

On the other hand, by substituting eq. (18) in eq. (19) one obtain an expression of $\lambda_n$, useful for the design:

$$\lambda_n = \frac{\pi \cdot (1+K_n)}{2 \cdot (\mu_{rev}/\mu_o)} \cdot \frac{D_n}{\delta} \quad . \tag{21}$$

In order to limit the influence that the amplitudes of the direct and alternative components have on the current ratio, the core magnetic factor $\lambda_n$ must be suitably lower than unity (for example: $\lambda_n \simeq 0.1$).

It has been already observed that the error function $\tilde{\beta}$ decreases when the number of turns $N_2$ increases, following a quadratic law: then, it is convenient to increase $N_2$ as much as possible. On the other hand, eq. (9) contains the resistance $R_2/N_2^2$: this resistance, indicated with $R_2^*$, equals:

$$R_2^* = \frac{R_2}{N_2^2} = \rho \cdot \frac{\ell_{m2} \cdot N_2}{A_{c2}} \cdot \frac{1}{N_2^2} = \rho \cdot \frac{\ell_{m2}}{A_{r2}} \quad , \tag{22}$$

where $\ell_{m2}$ is the length of the middle turn of the secondary winding, $A_{c2}$ is the conductor section, $A_{r2}$ is the total copper section of this winding: by keeping constant $A_{r2}$ during the variation of $N_2$, nor the current density of the secondary winding neither its gross utilization area varies.

Considering the variation of the design parameters, in particular the resistance $R_m$, a need to be observed is to obtain a suited value of the voltage $V_m$ at the terminals of the measurement resistance: eq. (13) shows that, neglecting the effect of $\beta$, this voltage does not vary considerably if the ratio $R_m/N_2$ (indicated with $r_m$) is mantained constant:

$$V_m \simeq r_m \cdot \frac{N_1 \cdot I_1}{\sqrt{1 + \tilde{\beta}^2}} \quad . \tag{23}$$

By manipulating eq. (9), the error function $\tilde{\beta}$ changes as follows:

$$\tilde{\beta} = \frac{2 \cdot \delta_{eq}}{\tilde{\omega} \cdot \mu_o \cdot A_n} \cdot \left[R_2^* + \frac{r_m}{N_2}\right] \quad . \tag{24}$$

Eq. (24) points out the important fact that $\tilde{\beta}$ decreases when the number of turns $N_2$ increases (with $r_m$ constant): then the C.T. errors decrease too, being unvaried the amplitude of the measurement signal $V_m$.

From eq. (24), considering that $R_2^* \ll r_m/N_2$, and by the eq. (14), (16), (17), (21), (23), one obtains an expression of the product $N_2 \cdot D_n^2$ particularly interesting:

$$N_2 \cdot D_n^2 = \frac{1 + \lambda_n}{\sqrt{2 \cdot \tilde{\varepsilon} - \tilde{\varepsilon}^2} \cdot \tilde{\omega}} \cdot \frac{1}{K_{ot} \cdot K_L \cdot K_n^2} \cdot \frac{V_m}{B_{dc}} \cdot \frac{I_{dc}}{I_1} \quad . \tag{25}$$

As a matter of fact, this expression includes the main functional and constructive quantities, that affect the product $N_2 \cdot D_n^2$:

- the core magnetic factor $\lambda_n$ has a weak influence;
- the choice of the reference angular frequency $\tilde{\omega}$ and of the corresponding current ratio error $\tilde{\varepsilon}$ is very important: this choice can make the design very heavy, in case of too severe specifications;
- the characteristics of the magnetic material are included in the value of the core magnetic factor $\lambda_n$ (inversely proportional to $\mu_{rev}$) and in the flux density $B_{dc}$ (the higher is $B_{sat}$, the higher can be $B_{dc}$);
- the product $N_2 \cdot D_n^2$ increases as the signal $V_m$ increases;
- the product $N_2 \cdot D_n^2$ increases as the direct component $I_{dc}$ increases, while it decreases as the alternative component $I_1$ increases.

The quantities not explicitly expressed in eq. (25) are the air-gap and the primary number of turns. With regard to the air-gap, from eq. (18) and (19) one can obtain:

$$\delta = \frac{\pi}{2} \cdot \frac{1 + K_n}{\mu_{rev}/\mu_o} \cdot \frac{D_n}{\lambda_n} \quad ; \tag{26}$$

then the air-gap increases with the internal core diameter $D_n$ and with the core coefficient $K_n$; the higher is the relative reversible permeability, the smaller is the air-gap: in practise, above all for construction reasons, it is better to fix the ratio $\delta/D_n$ at the beginning of the design.

With regard to the primary number of turns, by solving eq. (14) one obtains:

$$N_1 \simeq \frac{2}{\mu_o} \cdot B_{dc} \cdot \frac{\delta}{I_{dc}} \quad . \tag{27}$$

Finally, the observation of eq. (25), (26), (27) suggests the possibility to employ the same core for measurements of harmonic noise of currents having quite different average components $I_{dc}$: this corresponds to the use of the same transformer for different nominal values.

As a matter of fact, eq. (25) contains the ratio $I_{dc}/I_1$: then, if the harmonic content is constant in per unit when referred to the $I_{dc}$ component, the product $N_2 \cdot D_n^2$ does not vary as $I_{dc}$ increases; from eq. (26) one can deduce that the air-gap does not change too. Then, according to eq. (27), the number of turns $N_1$ must decrease as $I_{dc}$ increases, with a law of inverse proportionality. In conclusion, within certain limits and provided that $N_1 \geq 1$, the same core with the same number of turns $N_2$ can be emploied, also for performing measurements regarding fluctuating direct currents having amplitude different from the value considered for the design of the current transformer.

## 5. CONCLUSIONS.

In this paper it has been presented and discussed a new type of current transformer, suited to perform the measurement of small amplitude harmonic noise associated to high values of direct current components, absorbed or generated by static converters, for example, the converters used in D.C. traction lines.

The design criteria of the transformer are quite different from those adopted for A.C. current measurements, either for the choice of the magnetic material or for the dimensioning of the core and windings.

The studies continue, in order to improve the design and the high frequency model of the C.T., also by means of experimental tests using prototypes.

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## LIST OF SYMBOLS

- $A_{c2}$ conductor section of the secondary winding
- $a_n$ radial width of the core section
- $A_n$ net magnetic core section
- $A_{r2}$ total section of the secondary winding
- $B_{dc}$ D.C. biasing flux density
- $B_{int}$ intrinsic D.C. biasing flux density
- $B_{sat}$ magnetic material saturation flux density
- $b_n$ axial length of the core section
- $\beta$ error function (generic value, see eq. (9))
- $\tilde{\beta}$ error function (design value, see eq. (16))
- $D_n$ internal diameter of the toroidal core
- $\delta$ geometrical air-gap
- $\delta_{eq}$ equivalent magnetic air-gap (see eq. (6))
- $E_t$ e.m.f. per turn
- $\varepsilon$ current ratio error (generic value)
- $\tilde{\varepsilon}$ current ratio error (design value)
- $\Phi$ magnetic flux in the core
- $G_o$ parallel conduttance of the equiv. circuit
- $\gamma$ phase error
- $H_{dc}$ biasing D.C. magnetic strength
- $I_{dc}$ D.C. component of the primary current
- $I_\mu$ magnetizing current
- $I_1$ A.C. component (single harmonic) of the primary current (RMS)
- $\dot{I}_{1s}$ A.C. primary current, referred to secondary
- $I_2$ secondary current (single harmonic)
- $K_L$ ratio of the core widths ($K_L = b_n/a_n$)
- $K_n$ core coefficient ($K_n = a_n/D_n$)
- $K_{st}$ lamination stacking factor
- $\ell_{m2}$ length of the middle secondary turn
- $\ell_n$ geometric length of the magnetic core
- $L_o$ parallel inductance of the equiv. circuit
- $L_1$ primary series induct. of the equiv. circuit
- $L_2$ secondary series induct. of the eq. circuit
- $\lambda_n$ core magnetic factor (see eq.s. (19) e (20))
- $\mu$ normal permeability
- $\mu_i$ initial permeability
- $\mu_\Delta$ incremental permeability
- $\mu_o$ vacuum permeability
- $\mu_{rev}$ reversible permeability
- $N_1$ number of turns of the primary winding
- $N_2$ number of turns of the secondary winding
- $r_m$ specific measurement resist. ($r_m = R_m/N_2$)
- $R_m$ measurement resistance
- $R_1$ primary series resist. of the equiv. circuit
- $R_2$ secondary series resistance of the C.T. equivalent circuit
- $R_2^*$ specific secondary resistance, referred to 1 secondary global equiv. turn (see eq. (22))
- $\rho$ copper resistivity
- $V_m$ measurement voltage at the terminals of $R_m$
- $\omega$ angular frequency of the A.C. quantities (single harmonic)
- $\tilde{\omega}$ angular frequency corresponding to the error $\tilde{\varepsilon}$ (design value)
- $\bar{Y}_o$ parallel admittance of the equiv. circuit
- $\bar{Z}_1$ primary series impedance of the "T" shaped C.T. equivalent circuit
- $\bar{Z}_2$ secondary series impedance of the "T" shaped C.T. equivalent circuit.