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Control of the Current-to-Voltage Converter's Transformed Power Through Magnetic Biasing

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1. Introduction

A specific problem for dissipation of the transformed secondary excess-power arises during the time of evaluation of a low-power power supply, drained from an a.c. current source.

The goal of this paper is to develop a d.c. power supply V_L, at load current I_L, drained from a main power line with frequency f, and large primary current variations in the interval $(I_{1\,min}, ..., I_{1\,max})$. The specified d.c. output power P_L should be ensured at the minimum current $I_{1\,min}$, and the dissipated electrical excess-power at $I_{1\,max} >> I_{1\,min}$ have to be acceptable.

In Fig. 1 the proposed block diagram of the above specified Current-to-Voltage Converter is shown. It is build of a: Current Transformer ***CT***, Rectifier ***Rec***, Regulated Power Supply ***LReg***, Bypassing Capacitors C_o – C_L, and Load R_L.

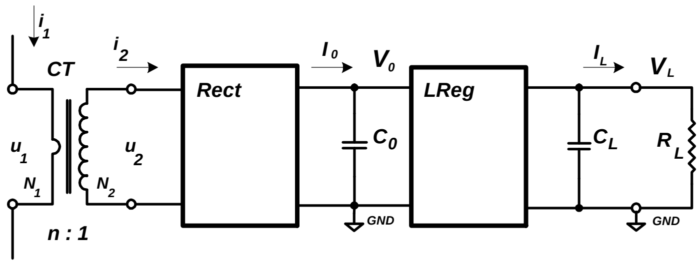

Fig. 1. Block-diagram of the proposed Current-to-voltage converter

The Converter's design and development is based on the well-known theory of the Transformers, e.g. [1, 2, 3], and the magnetic materials data, given by the manufacturers' Data Sheets [4, 5, 6].

The ***CT*** primary winding with N_1 turns is connected in series with the current loop of the Main Power Source. Through the ***CT*** the transformed secondary voltage after rectification and regulation ensures the specified Load voltage V_L at the nominal Load current I_L drained by the Load R_L.

The Converter should be optimized according to the main goals:

– ***CT*** size;

– number of turns N_1 of the secondary winding;

– the Transformation ratio n;

– the useless but unavoidable induced excess Secondary Power P_2 at strong Primary Current $I_{1\,max} >> I_{1\,min}$.

It would be enough to study the two Converter's boundary states:

– at minimum Primary Current $I_{1\,min}$, and

– at extremely strong Primary Currents $I_{1\,max} >> I_{1\,min}$.

During the **first state** the maximum-effective conversion of the energy at the primary to the specified Load voltage V_L should be ensured. As a matter of fact this case is almost independent of the approaches, perceived during the discussions on the second state.

At the **second state** the induced excess power should be limited.

The accepted approaches are different and so they are presented separately.

2. C-to-V Conversion at minimum primary current $I_1 = I_{1\,min}$

At minimum Primary Current $I_{1\,min}$ the Converter should ensure the specified Load voltage V_L at given current I_L. In Fig. 2 the Converter's Equivalent Circuit Diagram under these conditions is shown. The embedding to the Primary winding is done only for the Rectifier Circuit in order to avoid the useless analyses complication.

The used diodes are idealized: with no-resistance in forward direction, and not-conductive at reverse biasing. Their knee voltage U_k could be taken into account through the corresponding rise of the Load voltage V_L. The Rectifier output, together with the bypassing capacitor C_o, is equivalently taken into consideration by the Batteries' voltage V_o.

As usually the Transformation ratio n is defined by the relation:

(1) $$n = \frac{U_2}{U_1} = \frac{N_2}{N_1}.$$

In the case under consideration the energy losses at the Secondary Winding and the Magnetic Core are too small and so they are not taken in to account.

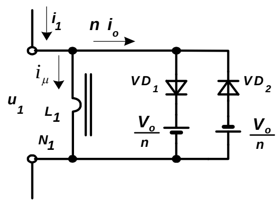

Fig. 2. Equivalent presentation of the Converter

The Converter's behavior, from the Current transformer's point of view, could be considered as ***up-critical*** or ***under-critical*** in dependence of the Transformer's secondary voltage u_2:

• At ***up-critical state*** the secondary voltage u_2 is with rectangular shape and the energy transfer to the secondary winding is continuous;

• At ***under-critical state*** the energy transfer is made during only part of the half of the main current cycle. This happens for the time-intervals, when the magnetic flux density changes are fast enough, in order to induc the necessary secondary voltage. After rectification this voltage defines the output voltage V_o, activating the Diodes $\boldsymbol{VD_1}$ and $\boldsymbol{VD_2}$ (Fig. 2).

During the rest of the half-cycle the velocity of the magnetic flux is lower, and the induced voltage keeps the diodes in non-conductive state.

2.1. Converter at up-critical state

At high enough Core's Magnetic permittivity, i.e. when $\mu_{cr} \rightarrow \infty$, the whole Primary Current is flowing through the Load. This define the maximum-achievable Transformation Ratio n_{max}:

(2) $$n_{\max} = \frac{2\sqrt{2}}{\pi} \frac{I_{1\min}}{I_{\mathrm{L}}} .$$

Actually part of the Primary Current, the Magnetizing Current i_μ, flows through the inductance L_1 of the primary winding, shown on the equivalent presentation.

At Up-Critical state primary voltage is with rectangular shape and as a matter of fact there are no time intervals, during which the Secondary Current is zero. The Secondary Voltage u_2 polarity changes at the instant when the Primary Current i_1 becomes equal to the Magnetizing Current i_μ (Fig. 3):

(3) $$I_{\mu\, m} = \sqrt{2}\, I_{1\min} \sin \alpha,$$

where α is the instant Phase angle value, at which the currents i_1 and i_μ are equalized.

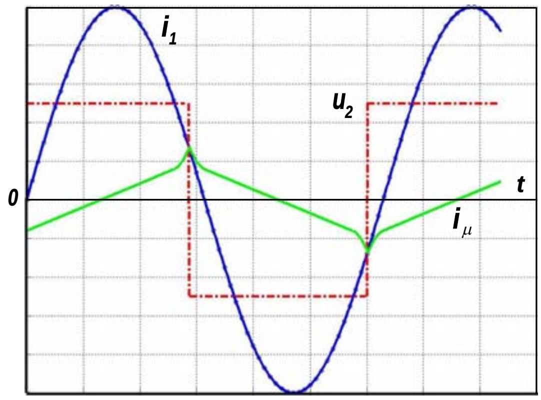

Fig. 3. Current transformer's voltages and currents $I_{1\ \min}$

Important Transformer's parameter at minimum Primary Current $I_{1\ \min}$ is the relation between γ the Magnetizing Current and the Primary Current:

(4) $$\gamma = \frac{I_{\mu m}}{\sqrt{2} I_{1\min}} = \sin\alpha .$$

Taking into account the Magnetizing Current one could write the Average Value of the current through the output after rectification:

(5) $$I_o = \frac{2\sqrt{2}}{\pi\, n} I_{1\min} \int_{\pi-\alpha}^{\alpha} \sin\theta \, d\theta = \frac{2\sqrt{2}}{\pi} I_{1\min} \sqrt{1-\gamma^2} .$$

Then the necessary Transformation Ratio will be

(6) $$n = \frac{2\sqrt{2}}{\pi} \frac{I_{1\min}}{I_o} \sqrt{1-\gamma^2} = n_{\max} \sqrt{1-\gamma^2} .$$

In Table 1 the calculations of relation $n / n_{\max} = f(\gamma)$ are displayed.

Table 1. The Transformation Relation dependence from γ

γ	0	0.2	0.3	0.4	0.5	0.75
$n/n_{\max}$	1	0.98	0.95	0.92	0.87	0.66

It should be mentioned that without significant falls of the Coefficient of Transmission, a Core could be used, characterized with Magnetizing Current up to 50% of the Primary Current.

The relation (5) is received at presumption for hysteresis-free Core relation $B(H)$, to which belong the soft-magnetic materials, which hysteresis loop is narrow and field is weakly-dependent. Such materials are characterized with high magnetic permittivity, as the well-known permaloy materials, and especially as the recent amorphous and nanocrystalline soft-magnetic materials [4].

The presence of hysteresis at $B(H)$ loop is responsible for the Coefficient of Transmission drop. The hysteresis provokes the phase-shift between the Magnetizing Current i_μ and the Magnetic Flux Density B, which becomes yet non-orthogonal with the Secondary Voltage. In the used Current Transformer with enhanced non-harmonic, practically rectangular voltage, this non-orthogonallity is displayed by the non-zero average value $I_{\mu o}$ for the voltage half-cycle. By this value, transformed to the secondary winding, the Secondary Current is decreased.

In Fig. 3 the Transformer's Current and Voltages at presence of Core hysteresis are displayed. The i_μ,s average value for the Primary Current half-cycle, displayed as shadowed area, is exactly $I_{\mu o}$. It should be mentioned that the transformers, used in the investigated Converter with high-value relation $I_{1\max}/I_{1\min}$, the conventional transformers silicon-iron alloys are useless due to their too large hysteresis loop.

An important feature, that strongly influences the Transformer sizes, is the Magnetic Flux saturation B_s. At maximized Core usefulness towards the Magnetic Flux density one can write the equations:

(7) $$U_1 = 4 N_1 f \Phi_s \qquad V_o = 4 N_2 f \Phi_s$$

where f is the Primary Current frequency; $\Phi_s = A_c B_s$ is the Core Magnetic Flux at saturation; A_c is the Core cross area.

For the Transformers, implemented on Magnetic Core without hysteresis, Magnetic Flux density variations up to $\pm B_s$, and minimum Primary Current strength $I_1 = I_{1\min}$, the output D.C. Power P_o will be

(8) $$P_o = V_o\, I_o = 4\, n\, N_1\, f\, A_c\, B_s\, \frac{2\sqrt{2}}{\pi}\, I_{1\min}\, \sqrt{1-\gamma}\,.$$

The maximum available power, which the Transformer could deliver at $\gamma = 0$, is:

(9) $$P_{o|\gamma=0} = \frac{8\sqrt{2}}{\pi}\, f\, B_s\, n\, N_1\, A_c\, I_{1\min}\,.$$

Important features of the transformers' under discussion are the insurance of minimum volume, and minimum turns number of the Primary Winding, if possible to be used only one turn. An useful parameter at these cases is the Specific Transformer Power, defined as the power, which could be delivered by a Transformer, characterized with unity Core cross area, only one turn of the Primary Winding, at Primary current $I_{1\min} = 1$ A:

(10) $$G = \frac{P_o}{N_1\, A_c\, I_{1\min}}\,.$$

Using the weak-hysteresis' magnetic core materials, the Specific Transformer Power becomes

(11) $$G_{|H_c \to 0} = \frac{8\sqrt{2}}{\pi}\, n\, f\, B_s\,.$$

So, the G's values at too strong magnetizing current, $\gamma = 0.4$, and $f = 50$ Hz, are respectively:

$G \sim 19.8\,\dfrac{\text{mW}}{\text{A.t.cm}^2}$ – for nanocrystaline permaloy [3] at $B_s \sim 1.1$ T, and

$G \sim 24.8\,\dfrac{\text{mW}}{\text{A.t.cm}^2}$ – for wound-strip soft-magnetic material at $B_s \sim 1.5$ T.

2.2. Function at sub-critical conditions

At sub-critical conditions the energy transfer to the secondary winding is realized only during the part of the Primary Current's half-cycle. This is due to the not-sufficient primary winding's inductance L_1. Herewith are discussed two reasons, provoking the sub-critical transformer's conditions, appearing when the magnetic core transformer is saturated, due to:

- too strong Primary Current I_1;
- due to the too high Output d.c. Voltage V_o.

In Fig. 4 the shapes of the magnetizing curve $B(H)$, Primary current, and Secondary voltage of the Converter at Sub-Critical conditions are displayed.

For the time-interval at which the magnetic core is not saturated, the magnetic flux varyes linearly in time from $-\Phi_s$ to $+\Phi_s$, where $\Phi_s = A_c B_s$, and vise-versa.

The process active phase time duration τ is:

$$\tau = 2\frac{\Phi_s}{V_o} N_2 . \qquad (12)$$

It is defined by the Magnetic Flux saturation B_s and the Output Voltage V_o.

For magnetic cores, characterized with low hysteresis losses and too high magnetic core permittivity μ_{cr} , i.e. at $\mu_{cr} > 10^4$, the $B(H)$ curve could be idealized and approximated through the relation

$$B = B_s \text{ sign } H . \qquad (13)$$

This means $\mu_{cr} \to \infty$ before saturation and $\mu_{cr} \to 0$ after that.

Then the Primary Current, flowing during the time-interval τ, is transformed completely into Secondary Current, which Average means is

$$I_o = 2\frac{\sqrt{2}\, I_1}{n T}\int_0^{\tau} \sin \omega\tau \, d\tau = \frac{\sqrt{2}}{\pi\, n} I_1 \,(1 - \cos\omega\tau) . \qquad (14)$$

At these circumstances the Output Voltage V_o depends on the Load, $V_o = R_L\, I_o$ and τ is independent from the Primary Current strength.

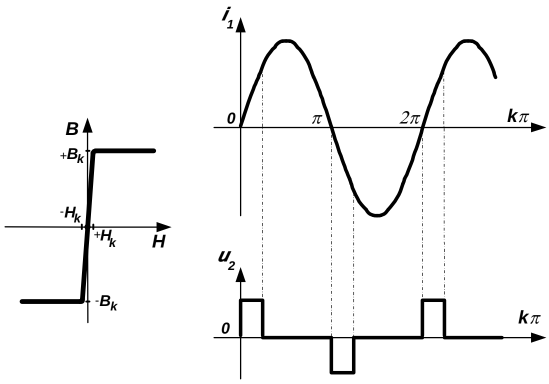

Fig. 4. Primary winding's Current i_1 , Secondary Voltage u_2, and Core's Magnetizing curve $B(H)$ of the Current Transformer ***CT*** at sub-critical conditions

For the Specific Power one could receive:

$$G = \frac{V_o\, I_o}{N_1\, A_c\, I_1} = \frac{2\sqrt{2}}{\pi}\,\frac{1 - \cos\omega\tau}{\tau}\, B_s . \qquad (15)$$

In Table 2 the relation $G = f(\omega\tau)$ is shown at B_s = 1.1 T and f = 50 Hz.

Table 2. $G = f(\omega\tau)$ at B_s = 1.1 T and f = 50 Hz.

$\omega\tau$	deg	30	45	60	70	80	90	120	150	180
G	$\frac{\text{mW}}{\text{A.t.cm}^2}$	7.96	11.6	14.7	15.2	18.4	19.8	22.3	22.2	19.8

The Specific Power is related mainly with the Magnetic flux density at saturation B_s and it is R_L-dependent through τ.

The induced power is almost linearly-dependent of the Primary Current and it is in weak relation with R_L. It should be mentioned that the maximum of G is at $\omega\tau = 135°$, but not at $\omega\tau = 180°$, corresponding to the case of sub-critical state.

In Table 3 the experimental measurements' results of the Converter are shown, using a Torroidal transformer, characterized with the following features:

$$A_c \sim 0.81 \text{ cm}^2, \quad N_1 = 2\,\text{t}, \quad N_2 = 180\,\text{t}.$$

The Primary winding's current is I_1 = 2 A, and the load R_L variations are responsible for the changes of the secondary voltage and output power. In this case the bridge-rectifier's losses are not taken into consideration.

Table 3. $P_o\ (V_o)$ at $I_1 = 2$ A, using a Torroidal Transformer

V_o	V	6	8	9	10	11	12	13	14	15.1
P_o	mW	45.3	54.4	56.7	58.5	58.3	57.6	55.9	53.9	50.6

One could mention the results follow close enough the theoretically defined relations.

In Fig. 5 the currents and voltages of a Converter are shown, implemented on a Core with typical $B(H)$ magnetizing curve.

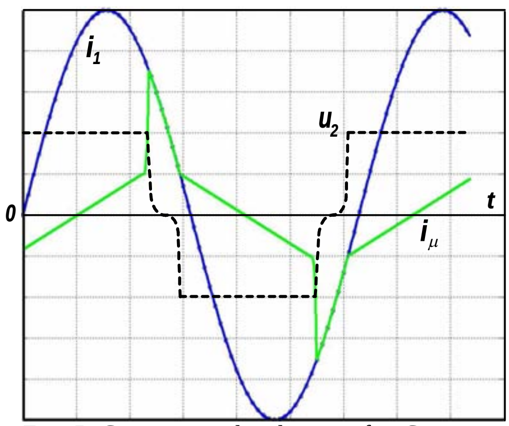

Fig. 5. Currents and voltages of a Converter, implemented on a Core with typical $B(H)$ magnetizing loop

3. Function at strong primary currents $I_1 > I_{1\min}$

At Primary currents $I_1 > I_{1\,\min}$ the transformed power P_2 exceeds the needed load power P_L and the useless power has to be dissipated in the surrounding media. At $I_1 >> I_{1\,\min}$, which cases are under discussion here, the exceed power reach unacceptable high values.

One solution of the above mentioned problem could be a Transformation Ratio n reduction at $I_1 > I_{1\ min}$, as it is done in Fig. 6. The bridge rectifier circuit ***Rect***, used at the blockdiagram from Fig. 1, is combined with controlled components, united all together as Ballast Circuit ***Blst***, which keeps the Converter under control.

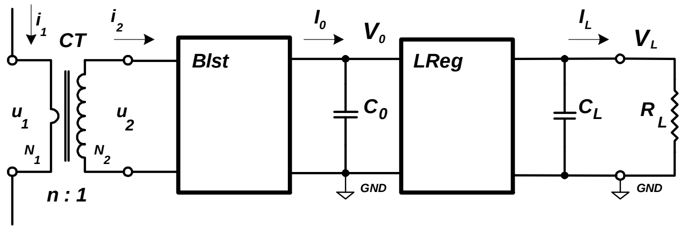

Fig. 6. Current-to-Voltag Converter, with controlled P_2 exceed power rise

The ***Blst*** could be connected continuously, or it could be switched periodically.

3.1. P_2 exceed power limit by ***Blst*** in a switching mode

An effective control of the transformer's power transfer could be made applying a switched mode regulating ***Blst*** in the secondary transformer's side. Connected after the ***Rect*** Building block, the Ballast ***Blst*** is switched periodically on-and-off in dependence on the Primary's and Load's current variations.

The ***Blst*** connection could be **Serial** as well as **Parallel** to the Load.

At **Serial Mode** the connection between the ***CT*** 's second winding is switched, and the ***LReg*** after the rectifier circuit. This Mode provokes the appearance of high pulse voltage spicks.

At **Parallel Mode** the Transformer's secondary winding is switched either to the Load or it is short-circuited. This Mode is accompanied with flowing of too strong secondary currents.

3.2. Exceed power limit through ***Blst*** in a parallel switching mode

The Parallel-Mode ***Blst*** circuit is shown in Fig. 7.

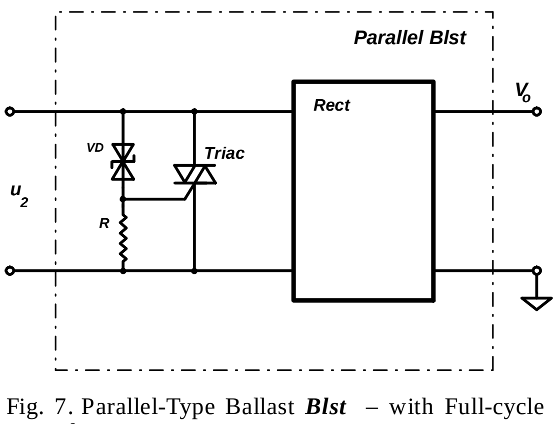

Fig. 7. Parallel-Type Ballast ***Blst*** – with Full-cycle control

The Symmetric Zener-Diode ***VD*** and a ***Triac*** limit the maximum level of Output Voltage V_o at rise of the Primary Current over its minimum value $I_{1\ lim}$.

The circuit is simple but reliable during its functioning. There are no sharp transitions, sources of disturbances. The energy, applied to the regulator ***Lreg*** input, is effectively limited, and after careful design could be made close to the minimum required.

The Transformed Power's limit is results of the Output Voltage V_o decrease, determined by the voltage drop on the ***Triac*** and the secondary winding's equivalent resistance. The Transformer ***CT*** is in linear mode, and do not went in saturation. The secondary current is defined by the Transformation Ratio n: as conventional Current Transformer the Secondary Current is in proportion to the Primary one.

In order to minimize the dissipated power the secondary winding has to be designed at minimum equivalent resistance requirements.

Appreciable induced power reduction at high overloading could be obtained also using **half-wave parallel switching** as it is shown in Fig. 8.

With the rise of the Primary Current i_1, the Secondary Voltage u_2 is rising too, and when its positive value reaches the specified threshold voltage $U_{2\,\mathrm{lim}}$, the Thyristor ***Thr*** is activated. So the ***Thr*** short-circuite the ***CT***'s Secondary Winding. The provoked Secondary Voltage u_2 cycle asymmetry led to magnetic core biasing, which rises with the rises of the Primary Current i_1 strength. At too high values of I_1 the magnetic core quiescent point is shifted deeply at saturation, and so it limits the transformed power.

Now the Converter's functioning will be discussed at too strong Primary Currents $I_1 >> I_{1\,\mathrm{min}}$. It is made under assumption of a ***CT***, implemented on a magnetic core, which magnetizing curve $B(H)$ is approximated by (13).

The network process during the activated ***Thr***'s time interval, could be approximated by the equation

$$N_2 \frac{d\Phi}{dt} = r_2 \frac{i_1}{n} + U_{\mathrm{Th}}, \tag{16}$$

where r_2 is the sum of the secondary winding's resistance and the Thyristor's equivalent dynamic resistance; U_{Th} is the Thyristor's threshold voltage.

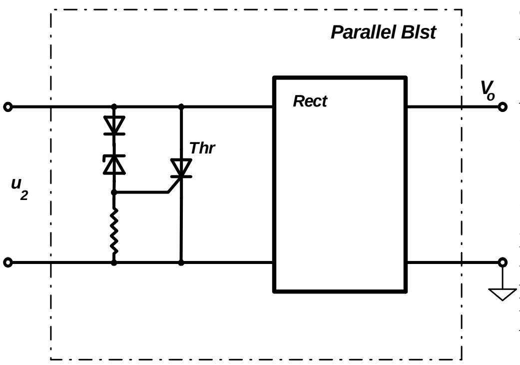

Fig. 8. Parallel-Type Ballast Blst

We assume r_2 to be too small, and the voltage drop on it comparable or smaller than U_{Th}.

Due to the Primary Current's i_1 high speed rise we assume that the Thyristor is activated during the time of the whole positive half-wave, neglecting the u_2 rise-time to $U_{2\,\mathrm{lim}}$. In result because of the low value u_2, the magnetic core is not saturated. The current flowing through the Thyristor is defined by the Transformation Ratio:

$$i_{\mathrm{Th}} = \sqrt{2}\,\frac{I_1}{n}\cos\theta \;, \quad \text{where} \quad -\frac{\pi}{2} \le \theta \le \frac{\pi}{2}\,.$$

The power, dissipated on the Thyristor will be

$$P_{\mathrm{Th}} = \frac{I_1}{2n}\left(\frac{2\sqrt{2}}{\pi}U_{\mathrm{Th}} + r_2\frac{I_1}{n}\right), \tag{17}$$

and it has to be dissipated in the surrounding media.

One important requirement toward the ***CT*** is its Secondary Winding's equivalent resistance to be low.

At the end of the half-cycle the Magnetic Flux rises up to

$$\Phi\left(\frac{T}{2}\right)=\frac{U_{\mathrm{Th}}}{N_2}\cdot\frac{T}{2}+\frac{\sqrt{2}\,r_2\,I_1}{n\,N_2}. \tag{18}$$

After this value during the second Primary-Current's half-cycle the Magnetic Flux start drops-down in reverse direction and will go deeply in magnetic saturation. The duration's time-interval of the current, flowing towards the Regulator ***LReg***, is

$$\tau=\frac{\Phi_{\mathrm{s}}\,N_2}{V_{\mathrm{o}}}. \tag{19}$$

It is short, disposed around the Primary Current's i_1 zero-crossing, and could be assumed it is linearly time-dependent

$$i_1=\frac{\sqrt{2}\,I_1}{n}\,\omega\,t\,,$$

where $0\le t\le\tau$, and its Average value is

$$I_{1\,\mathrm{av}}\sim I_{\mathrm{o}}=\frac{\sqrt{2}}{2}\,\frac{I_1}{n}\,2\pi\left(\frac{\tau}{T}\right)^2 \tag{20}$$

where $T=1/f$ is the Primary Current's i_1 period.

Using (17) and (18) one could write for V_{o}

$$V_{\mathrm{o}}^2=\sqrt{2}\,\pi\,\frac{I_1}{n\,I_{\mathrm{o}}}\,(N_2\,\Phi_{\mathrm{s}}\,f)^2 \tag{21}$$

and for the Output Power P_{o}:

$$P_{\mathrm{o}}=I_{\mathrm{o}}\,V_{\mathrm{o}}=\sqrt{\sqrt{2}\,\pi\,\frac{I_{\mathrm{o}}\,I_1}{n}}\,N_2\,\Phi_{\mathrm{s}}\,f\,. \tag{22}$$

This power is higher than the power P_{L}, delivered to the Load and the difference has to be absorbed by the Regulator ***LReg***. This excess power over the needed Load power is caused by the rise of the Output voltage V_{o}. During the flowing of too strong Primary Currents I_1 it is possible V_{o} to exceed the Regulator's Breakdown Voltage and Rated Power. In such cases complementary restrictive measures have to be taken.

At the end it should be mentioned that the real magnetizing loop $B(H)$ is quite different from the idealized one, used in this analysis. But in this case of a large value of the Coefficient of current overloading $\beta=I_1/I_{1\min}$, which is >>50, it could be found, that the steepness of $B(H)$ at saturation is high enough for Converter functioning at half-wave variations of the Magnetic flux density in the Core. So this could be favorable for some complementary fall of the transformed power under the conditions of extremely strong Primary currents I_1.

4. Experimental results

In order to verify the theoretical conclusions a Converter was implemented, shown in Fig. 9.

The Current Transformer ***CT*** is made on Torroidal core with A_c ~0.96 cm , using Vitroperm 800 F of Vacuumschmelze [4], with four windings: N_1 = 1 t, N_2=N_3=120 t, and N_4=420 t.

The winding N_1 is used for N_2 losses measurement through compensation of induced in N_2 voltage.

N_4 winding, galvanically isolated from another windings, is used for Magnetic flux density measurement. The used R-C integrated network with Time-constant $\tau = 0.39$ s, ensures sensitivity $k = \dfrac{0.39}{2 \times 420 \times 0.96} \sim 4.84\ \dfrac{\mathrm{T}}{\mathrm{V}}$.

The winding N_2 is implemented with a large enough cross-section's lacquer-wire in order to decrease the power losses in it. Its d.c. resistance is ~0.4 Ω.

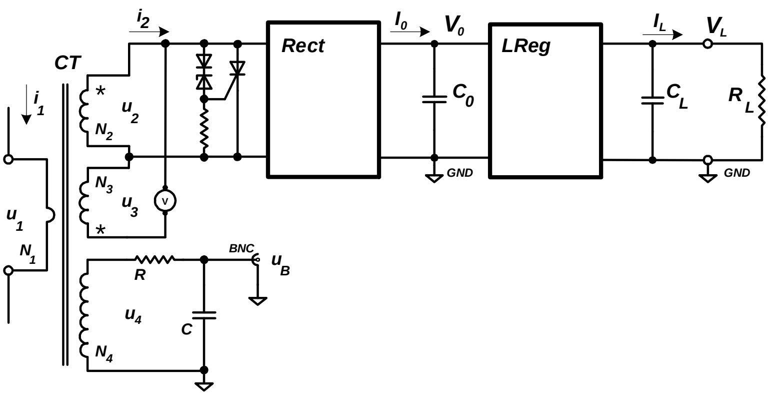

Fig. 9. Experimental Current-to-Voltage Converter

The presented below measurements results are received at $I_{1\,\min}$ = 2 A in order to check the maximum power, which could be transformed into the secondary loop, as well as at $I_{1\,\max}$ = 141 A, in order to estimate the induced excess power.

In Table 4 the received results at $I_{1\,\min}$ = 2 A are displayed.

Table 4. Experimental results

I_L	mA	1.5	2	3	4	5	6	7	8	9	10	11	12	13
V_o	V	7.12	6.24	5.20	4.44	3.91	3.47	3.1	2.79	2.52	2.29	2.05	1.81	1.56
P_o	mW	10.5	12.5	15.6	17.8	19.6	20.8	21.7	22.8	22.7	22.9	22.6	21.8	20.3
P_T	mW	12.6	15.1	19.5	23.0	26.0	28.6	30.8	32.7	34.4	35.9	36.8	37.3	37.2
θ_a	deg	47	56	65	78	87	95	105	113	121	130	139	149	160

P_T is the Total power, delivered by the transformer ***CT***, taking into account the losses in the bridge rectifier ***Rect***. It is assumed that the voltage drop over the serially connected diodes in the Bridge is ~ 1.3 V.

With θ_a the active part's duration of every half cycle is denoted.

It could be seen from Table 4 that the maximum power is at $\theta_a \approx 140°$, which is quite close to the theoretically defined value.

The investigated function's part is selected in a way that the magnetic flux penetrates into the saturation domain. The voltage u_B magnitude over the integrated group is within the limits ±(250, ..., 253)mV, which corresponds to Magnetic flux density variations B_m ~ 4.85 (0.250, ..., 0.253) ~ (1.21, ..., 1.22)T. It should be mentioned that these B_m values are in compliance with the data, given by the producer.

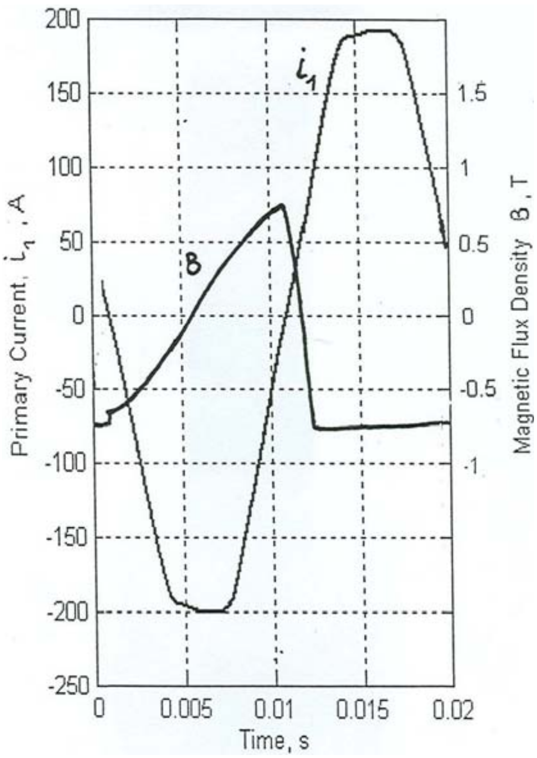

Fig. 10. Time domain variations of i_1

If one takes into account the measured Total power P_T = 37.3 mW then the Specific power could be defined:

$$G = \frac{37.3}{I_1 N_1 A_c} = \frac{37.3}{2 \times 1 \times 0.96} = 19.4 \frac{\text{mW}}{\text{At.cm}^2},$$

which is close to the theoretically defined result.

A measurement is made also at $I_{1\,max\,eff}$ ~ 141 A, i.e. at $I_{1\,max\,m}$ ~ 200 A, flowing through the ***CT***'s Primary winding in order to define the Converter's dissipated power. As useful power the power is assumed, delivered to the Load, at V_L = 5 V, and I_L = 10 mA.

In Fig. 10 the shapes of the measured Primary current i_1 and the Core's Magnetic flux density ***B*** at for one main's current cycle are displayed.

The next results are received through these measurements:

Output d.c. voltage = 15.2 V at I_o ~ 10.3 mA.

The i_2 shape during the:

– Passive phase	Half-sinusoid with magnitude 1.46 A. The same current also flows through the Thyristor;
– Active phase	Saw-tooth shape, with duration of 1 ms and magnitude 0.42 A.
Voltage over the Thyristor at activated state:	d.c. component of 0.75 V, with superposed half-sinusoid at magnitude 0.23 V.

Pick-to-pick voltage u_B 150 mV

Based on these data the next losses are calculated as follows:

Losses in

– Voltage regulator ***LReg*** $V_o = 15.2$ V at $I_o \sim 10.3$ mA ~ 0.10 W;

– Secondary winding N_2 $0.72\times\frac{1}{2}\times\frac{1}{2}\times 1.46^2 \sim 0.38$ W;

– Thyristor $0.75\times\frac{1.46}{\pi}+1.46\times 0.23\times\frac{1}{4} \sim 0.43$ W;

– Rectifier circuit $1.3\times 10\times 10^{-3} \sim 0.13$ W;

Total Losses: ~1.04 W.

The Core's Magnetic flux density variations: $\Delta B = \frac{0\times 39\times 0.15}{420\times 0.96\times 10^{-4}} \sim 1.45$ T.

The Core's Magnetic flux density limits: (– 0.24, …, +1.21) T.

This asymmetry is the main reason for the so small "useless" power, transformed to the Secondary winding at the huge, 70-time primary current overcharging in the primary winding.

During the implemented experiments the Main voltage shape was too distorted, with almost trapezoidal shape, which influenced the Primary current too. But we consider this does not change significantly the assessments and conclusions we have made.

5. Conclusions

A Current-to-Voltage Converter was investigated, characterized with huge $I_{1\,max}/I_{1\,min}$ relation, keeping small dissipated excess power, small size, and minimum number of turns at the primary winding. The too specific requirements make probably the Converter a little bit "out-of-stream" of the currently discussed Converters. Too exiguous are the reports in the available issues. This gave us the reason for a more detailed presentation of the implemented investigation.

One immediate reason for the carried out research is the development of the power supplies-repeaters of the communication systems, based on the Power line communication, as well as special requirements of Electronic Electrical Energy Meters, power supplying from the voltage as well as from the current loop.

The next main conclusions could be summarized:

- the received experimental results are too close to the theoretically predicted;
- the proposed enhanced asymmetrical magnetic biasing is able to limit effectively the transformed useless excess power;
- the proposed Converter could be used as a power supply operational source in the Main power networks and control cables of different power consumers, in measurements and information exchange units.

The Converter's development and investigation was made at IIT, BAS, as well as at Multiprocessor Systems Ltd., Sofia, Bulgaria.

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Управление трансформированной мощности в конверторе ток-напряжение через поляризирование магнитопровода

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(Р е з ю м е)

В настоящей работе предлагаются метод и схемы для реализации компактных и высокоэффективных генераторов стабилизированного оперативного напряжения, питаемых гальванично отделенным токовым источником с исключительно большой динамикой, 40 dB и больше.

Исследована работа конверторов, реализованных согласно предложенного метода и показаны основные процессы и полученные экспериментальные результаты.

Основным модулем в конверторе является токовым трансформатором.

Приведены основные зависимости между геометрическими параметрами, магнитными свойствами магнитопровода и выходной мощностью. Показано, что существует максимально достижимое переводное отношение, при котором получается заданное выходное напряжение, и что выходная мощность определяется магнитопроводом и она слабо зависима от переводного отношения.

Чтобы ограничить лишно индуктиранную мощность при больших первичных токов, которая должна рассеяться, предлагается переход в режим поляризирования магнитопровода и его возрастающая степень насыщения или периодическое включение и выключение вторичной намотки.