Isolated Zeta Converter: Principle of Operation and Design in Continuous Conduction Mode

PIJIT KOCHCHA, SARAWUT SUJITJORN^{*} School of Electrical Engineering, Institute of Engineering Suranaree University of Technology 111 University Avenue, Muang District, Nakhon Ratchasima, 30000 THAILAND pijit_sut@hotmail.com *corresponding author: sarawut@sut.ac.th

Abstract: - The principle of operation of the zeta converter is explained in the article. Small-signal and steady-state models of the converter are presented. The design formulas for the continuous conduction mode (CCM) are given together with some design examples and simulation results. The power factor and harmonic issues are also addressed.

Key-Words: - zeta converter, continuous conduction mode (CCM), harmonic, power factor.

1 Introduction

Vast majority of power converters used nowadays employ front-end diode bridge rectifiers. Such rectifiers draw pulsating currents which leave behind a great amount of harmonics, and considerably low power factor. For a single converter of this type used with a single-phase load such as in a consumer electronic equipment, the problems may not seem serious. However, a great number of those equipments in parallel connection at a point of common coupling (PCC) to draw power simultaneously introduce some serious effects concerning reactive power and harmonic. The situations are quite common in offices and industries.

Several types of AC-DC converters have been introduced to achieve the demanded power conversion, and the less problems on harmonic and power factor [1]. To name a few, these include the Cuk converter [2], the Sepic converter [3]-[6], the combined boost with double winding flyback converter [7], and the zeta converter [8]-[13]. Among those, the zeta converter, which is originally the buck-boost type, can be regarded as a flyback type when an isolated transformer is incorporated. An isolated zeta converter has some advantages including safety at the output side, and flexibility for output

adjustment [14]-[16].

This article describes the operation principle of the isolated zeta converter in the CCM in Section 2.

Section 3 presents an overview of the state-space averaging technique, while the transfer function model can be found in Section 4. Section 5 presents the steady-state time-domain model in CCM. Section 6 provides design examples, simulation results with discussions. The harmonic, and power factor issues are also discussed. Conclusion follows in Section 7.

2 Principle of Operation

Fig.1(a) depicts the circuit diagrams of the isolated zeta converter such that its operation principle in the CCM could be readily explained.

Fig.1(b) represents the 1st region of operation in which the switch S is "on", and the diode D is "off". This region takes the time from 0 to d_1T_s seconds. The inductor L_m stores the energy received from the rectifier. The capacitor C_1 supplies energy to the load (R) via the inductor L_o , and the capacitor C_o . the currents through the inductors L_m and L_o increase linearly, while no current flows through the diode.

Fig.1(c) represents the 2^{nd} operation region in which the switch S is "off", and the diode D is "on". This region begins at the time d_1T_s seconds, and ends by d_2T_s seconds. The diode D is forward biased due to the voltage across the inductor L_m has reversed polarity, while the currents i_{Lm} and i_{Lo} decrease linearly. The stored energy in the inductor L_m is transferred to the capacitor C_1 . The load R



(a) an isolated zeta converter





(c) region 2 of operation





Fig.2 Current and voltage waveforms.

receives energy from the inductor $L_{o}.$ Hence, the current $i_{D}{=}i_{C1}{+}i_{Lo}.$

Fig.2 illustrates the steady-state current and voltage waveforms in one switching cycle. These curves will be referred to for the development of the design formulas.

3 Overview of The State-Space Averaging Technique

Considering the operation of the converter during the on- and off-time intervals denoted as t_{on} or d_1T_s , and t_{off} or $(1-d_1)T_s$, respectively, the state equations in one switching cycle can be written as

$$\dot{x} = A_s x + B_s u$$

$$y = C_s x$$
(1)

, where

$$A_{s} = [A_{1}d_{1} + A_{2}(1 - d_{1})], B_{s} = [B_{1}d_{1} + B_{2}(1 - d_{1})],$$

$$C_{s} = [C_{1}d_{1} + C_{2}(1 - d_{1})].$$

Linearization can be made to the above equations by considering small-signal perturbations such that $x = X + \tilde{x}$, $y = Y + \tilde{y}$, $d_1 = D_1 + \tilde{d}_1$, $u = U + \tilde{u}$ where $X >> \tilde{x}$, $Y >> \tilde{y}$, $U >> \tilde{u}$ and $D_1 >> \tilde{d}_1$ be substituted into Eq. (1). Under the steady-state condition of the state variables, one may write the following equations

$$X = A_{av}X + B_{av}U = 0$$

$$Y = C_{av}X$$
 (2)

, and

$$\dot{x} = A_{av}\dot{\tilde{x}} + \left[\left(A_1 + A_2 \right) X + \left(B_1 + B_2 \right) U \right] \tilde{d} + B_{av} \tilde{u}$$

$$\tilde{y} = C_{av} \tilde{x} + \left[\left(C_1 - C_2 \right) X \right] \tilde{d}$$
(3)

, where

$$\begin{aligned} A_{av} &= A_1 D_1 + A_2 \left(1 - D_1 \right), \ B_{av} &= B_1 D_1 + B_2 \left(1 - D_1 \right), \\ C_{av} &= C_1 D_1 + C_2 \left(1 - D_1 \right). \end{aligned}$$

From Eq. (2), Eq. (4) below are obvious.

$$X = -A_{av}^{-1}B_{av}U$$

$$\frac{Y}{U} = -C_{av}A_{av}^{-1}B_{av}.$$
(4)

Under small-signal assumption, taking the Laplace transform to Eq. (3) results in

$$\dot{x}(s) = [sI - A_{av}]^{-1} [B_{av}\tilde{v}_{g}(s) + [(A_{1} - A_{2})X + (B_{1} - B_{2})V_{g}]\tilde{d}_{1}(s)]$$
(5).
$$\tilde{y}(s) = C_{av}\tilde{x}(s) + [(C_{1} - C_{2})X]\tilde{d}(s)$$

Finally, one can obtain the following transfer functions: for $n\tilde{v}_g = 0$

$$\frac{\tilde{y}(s)}{\tilde{d}_{1}(s)} = C_{av} [sI - A_{av}]^{-1} [(A_{1} - A_{2})X + (B_{1} - B_{2})V_{g}] + (C_{1} - C_{2})X$$
(6)

, and for $\tilde{d}_1 = 0$

$$\frac{\tilde{y}(s)}{\tilde{v}_{g}(s)} = C_{av} \left[sI - A_{av} \right]^{-1} B_{av}$$
(7).

4 Transfer Function Models

Referring to the circuit in Fig. 1(a), the parameters transferred from the secondary onto the primary side are

$$v_o' = v_o/n, \ L_o' = L_o/n^2, \ C_1' = n^2 C_1, \ C_o' = n^2 C_o,$$

 $R' = R/n^2.$

According to the operation in one switching cycle, the circuit equations can be written as

$$\frac{di_{Lm}(t)}{dt} = \frac{V_s}{L_m}(d_1) - \frac{v_{C1'}}{L_m}(1-d_1)$$

$$\frac{di_{Lo'}(t)}{dt} = \frac{n^2 V_s}{L_o}(d_1) + \frac{n^2 v_{C1'}}{L_o}(d_1) - \frac{n^2 v_{Co'}}{L_o}$$

$$\frac{dv_{C1'}(t)}{dt} = -\frac{i_{Lo'}}{n^2 C_1}(d_1) + \frac{i_{Lm}}{n^2 C_1}(1-d_1)$$
(8)
$$\frac{dv_{Co'}(t)}{dt} = \frac{i_{Lo'}}{n^2 C_o} - \frac{v_{Co'}}{RC_o}$$

$$v_o' = v_{Co'}.$$

Let the state, input and output vectors be represented by

$$x = \begin{bmatrix} i_{Lm'} & i_{Lo'} & v_{C1'} & v_{o'} \end{bmatrix}^T, \ u = \begin{bmatrix} v_g \end{bmatrix}, \ y = \begin{bmatrix} v_{o'} \end{bmatrix},$$

the coefficient matrices of the state-variable models can be obtained as

$$A_{av} = \begin{bmatrix} 0 & 0 & -\frac{(1-D_1)}{L_m} & 0 \\ 0 & 0 & \frac{n^2 D_1}{L_o} & -\frac{n^2}{L_o} \\ \frac{(1-D_1)}{n^2 C_1} & -\frac{D_1}{n^2 C_1} & 0 & 0 \\ 0 & \frac{1}{n^2 C_o} & 0 & -\frac{1}{RC_o} \end{bmatrix}$$
(9)

$$B_{av} = \left[\frac{D_{1}}{L_{m}} \quad \frac{n^{2}D_{1}}{L_{o}} \quad 0 \quad 0\right]^{T}$$
(10)

$$C_{av} = \begin{bmatrix} 0 & 0 & 0 & 1 \end{bmatrix}$$
(11).

By substituting Eqs. (9)-(11) into Eq. (4), the following steady-state relations can be obtained

$$\begin{bmatrix} I_{Lm} \\ I_{Lo'} \\ V_{C1'} \\ V_{o'} \end{bmatrix} = \begin{bmatrix} \frac{n^2 D_1^2}{R(1 - D_1)^2} \\ \frac{n^2 D_1}{R(1 - D_1)} \\ \frac{D_1}{(1 - D_1)} \\ \frac{D_1}{(1 - D_1)} \end{bmatrix} \begin{bmatrix} V_g \end{bmatrix}$$
(12).

$$\frac{V_o'}{V_g} = \frac{D_1}{(1 - D_1)}$$

The small-signal models as shown in Eqs. (13)-(14) can be obtained from substituting Eqs. (9)-(11) into Eq. (5).

$$\begin{bmatrix} \tilde{i}_{Lm} \\ \tilde{i}_{Lo'} \\ \tilde{v}_{CI'} \\ \tilde{v}_{O} \end{bmatrix} = \begin{bmatrix} 0 & 0 & -\frac{(1-D_{1})}{L_{m}} & 0 \\ 0 & 0 & \frac{n^{2}D_{1}}{L_{o}} & -\frac{n^{2}}{L_{o}} \\ \frac{(1-D_{1})}{n^{2}C_{1}} & -\frac{D_{1}}{n^{2}C_{1}} & 0 & 0 \\ 0 & \frac{1}{n^{2}C_{o}} & 0 & -\frac{1}{RC_{o}} \end{bmatrix} \begin{bmatrix} \tilde{i}_{Lm} \\ \tilde{i}_{Lo'} \\ \tilde{v}_{O'} \end{bmatrix} \\ + \begin{bmatrix} \frac{D_{1}}{L_{m}} \\ \frac{n^{2}D_{1}}{L_{o}} \\ 0 \\ 0 \end{bmatrix} \begin{bmatrix} \tilde{v}_{g} \end{bmatrix} + \begin{bmatrix} 0 & 0 & -\frac{\tilde{d}_{1}}{L_{m}} & 0 \\ 0 & 0 & \frac{n^{2}\tilde{d}_{1}}{L_{o}} & 0 \\ \frac{\tilde{d}_{1}}{n^{2}C_{1}} & -\frac{\tilde{d}_{1}}{n^{2}C_{1}} & 0 & 0 \\ 0 & 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} I_{Lm} \\ I_{Lo'} \\ V_{CV'} \\ V_{o'} \end{bmatrix} \\ + \begin{bmatrix} \frac{V_{g}}{L_{m}} \\ \frac{n^{2}V_{g}}{L_{o}} \\ 0 \\ 0 \end{bmatrix} \begin{bmatrix} \tilde{d}_{1} \end{bmatrix}$$
(13)

$$\begin{bmatrix} \dot{\tilde{v}}'_{o} \end{bmatrix} = \begin{bmatrix} 0 & 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} \tilde{\tilde{i}}_{Lm} \\ \tilde{\tilde{i}}_{Lo'} \\ \tilde{\tilde{v}}_{Cl'} \\ \tilde{\tilde{v}}'_{o} \end{bmatrix}$$
(14).

The transfer function models of the isolated zeta converter can be derived from the above Eqs. (13)-(14), and obtained as

$$G_{vd}(s) = \frac{\tilde{v}'_o(s)}{\tilde{d}_1(s)} = \frac{1}{(1-D_1)^2} \frac{a_{vd}s^2 + b_{vd}s + c_{vd}}{as^4 + bs^3 + cs^2 + ds + e}$$
(15)
 $\tilde{v}'(s)$

$$G_{\nu\nu}(s) = \frac{v'_o(s)}{\tilde{v}_g(s)} = \frac{1}{(1-D_1)^2} \frac{a_{\nu\nu}s^2 + b_{\nu\nu}}{as^4 + bs^3 + cs^2 + ds + e}$$
(16)

, where

$$\begin{aligned} a_{vd} &= n^2 R L_m C_1 V_g \left(1 - D_1 \right), \ b_{vd} &= -n^2 D_1^2 L_m V_g \,, \\ c_{vd} &= R V_g \left(1 - D_1 \right)^2 \\ a_{vv} &= n^2 D_1 R L_m C_1, \ b_{vv} &= D_1 \left(1 - D_1 \right) R \\ a &= n^2 R L_m L_o C_1 C_o, \ b &= n^2 L_m L_o C_1 \\ c &= R L_o C_o \left(1 - D_1 \right)^2 + n^2 D_1^2 R L_m C_o + n^2 R L_m C_1 \\ d &= L_o \left(1 - D_1 \right)^2 + n^2 D_1 L_m, \ e &= R \left(1 - D_1 \right)^2 . \end{aligned}$$

5 Steady-State Models

Referring to the waveforms in Fig. 2 and the analysis principle presented in [17], the steady-state models of the isolated zeta converter in CCM can be developed as follows.

The turn-on and turn-off times (t_{on} and t_{off}) can be expressed as

$$t_{on} = d_1 T_s = \frac{L_m \Delta i_{Lm}}{V_g} = \frac{L_o \Delta i_{Lo}}{n V_g}$$
(17)

$$t_{off} = (1 - d_1)T_s = \frac{nL_m \Delta i_{Lm}}{V_{C1}} = \frac{L_o \Delta i_{Lo}}{V_o} \quad (18).$$

Since the voltage $V_{C1} = nV_g d_1/(1-d_1) = V_o$, one can obtain the voltage gain

$$M = \frac{V_o}{V_g} = \frac{nd_1}{(1 - d_1)}$$
(19)

, and the duty cycle

$$d_1 = \frac{M}{n+M} \tag{20}.$$

Due to the fact that $T_s = \frac{1}{f_s} = t_{on} + t_{off}$, and from Eqs. (17)-(19), one may derive the current ripple terms obtained as

$$\Delta i_{Lm} = \frac{V_g V_{C1}}{f_s L_m \left(V_{C1} + n V_g \right)} = \frac{V_g d_1}{f_s L_m}$$
(21)

$$\Delta i_{Lo} = \frac{nV_gV_o}{f_sL_o\left(V_o + nV_g\right)} = \frac{nV_gd_1}{f_sL_o}$$
(22).



Fig. 3 The current i_{Co} waveform.

The voltage ripple ΔV_{C1} can be expressed as

$$\Delta V_{C1} = \frac{1}{C_1} \int_0^{d_1 T_s} i_{C1} dt = \frac{1}{C_1} \int_0^{d_1 T_s} i_o dt = \frac{i_o d_1 T_s}{C_1} \quad (23).$$

From Eq. (23) and $i = \frac{V_o}{R}$, one can obtain

$$C_1 = \frac{V_o d_1}{f_s R \Delta V_{c_1}} \tag{24}$$

Referring to the waveforms of i_{Co} and i_{Lo} in Figs. (2)-(3), and the fact that $\Delta V_{Co} = \Delta Q/C_o$, one may derive

$$C_{o} = \frac{V_{o} \left(1 - d_{1}\right)}{8f_{s}^{2}L_{o}\Delta V_{Co}}$$
(25).

6 Design Examples

Designing the isolated zeta converter in the CCM requires the knowledge of critical inductances denoted as L_{mc} and L_{oc} .

Based on the assumptions of lossless and the average load current being equal to the average current in L_o , the terms L_{oc} and L_{mc} can be found as follows:

$$L_{oc} = \frac{(1-d_1)R}{2f_s}$$
(26)

$$L_{mc} = \frac{(1-d_1)^2 R}{2n^2 f_s d_1}$$
(27)

The design criteria are that $L_o \ge L_{oc}$, $L_m \ge L_{mc}$, $C_1 \ge \frac{V_o d_1}{f_s R \Delta V_{C1}}$, and $C_o \ge \frac{V_o (1 - d_1)}{8 f_s^2 L_o \Delta V_{Co}}$ to achieve $\% \gamma \le 1\%$ and $\Delta V_{C1} = \Delta V_{Co} = 2\sqrt{3}\gamma V_o$. The formulas dignify the smallest components possible. In practice, the output qualities must be considered for a proper component sizing.

Let us consider the requirement that the converter produces the average output voltage of 105 V_{dc}, and the current of 2.1 A_{dc} from the ac input of 311 V_{pk}, 50 Hz. The first step is to calculate the gain $M = V_o/V_g = 105/311 = 0.3376$, the duty cycle d₁=0.6280, and the turns ratio n. Then select the switching frequency f_s. The expressions (26) and (27) are used next to obtain the inductances L_o and L_m as follows:

$$L_{o} \ge \frac{(1-d_{1})R}{2f_{s}} = \frac{(1-0.6280)(50)}{2(50 \times 10^{3})}$$
$$= 186 \mu H$$

$$L_m \ge \frac{(1-d_1)^2 R}{2n^2 f_s d_1} = \frac{(1-0.6280)^2 (50)}{2(0.2)^2 (50 \times 10^3) (0.6280)}$$
$$= 2.75 mH.$$

Thus, the chosen inductances are $L_o = 200 \,\mu H$, and $L_m = 3mH$.

The capacitances C_o and C_1 are calculated according to Eqs. (24) and (25), and obtained as

$$C_{o} \geq \frac{V_{o}(1-d_{1})}{8f_{s}^{2}L_{o}\Delta V_{Co}} = \frac{(105)(1-0.6280)}{8(50\times10^{3})^{2}(200\times10^{-6})(1.82)}$$
$$= 5.36\mu F$$

$$C_{1} \geq \frac{V_{o}d_{1}}{f_{s}R\Delta V_{C1}} = \frac{(105)(0.6280)}{(50 \times 10^{3})(50)(1.82)}$$
$$= 14.49 \mu F$$

The design results are summarized by Table 1.

Table 1 Design results	(converter w	ith an R	load).
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Parameters	Values	
peak input voltage (Vg)	311 V 50 Hz	
avg. output voltage (V _o)	105 V	
avg. output current (I _o)	2.1 A	
switching frequency (f _s)	50 kHz	
L _m , L _o	3 mH, 200 uH	
C ₁ , C _o	20 uF, 1.41 mF	
turns ratio	0.2	
of hf transformer (n)	0.2	
duty cycle (d ₁)	0.6280	
М	0.3376	







(b) output voltage and current waveforms

Fig. 4 Simulation results (R load, without LC filter).

The simulation results using PSIM are illustrated in Fig. 4. Very large transient voltage can be observed, and later suppressed by an LC filter. The average output voltage and current are 105.48 V_{dc} , and 2.11 A_{dc}, respectively. A high current harmonic of 123.53 % THD_i, and a low power factor of 0.6270 (lagging) are observed at the PCC.



(a) zeta converter with the LC-filter



(b) source voltage and current waveforms



(c) output voltage and current waveforms

Fig. 5 Simulation results (R load, with LC filter).

An LC filter designed by a conventional method ($L_f = 100 \text{ mH}$, $C_f = 137 \text{ nF}$) is incorporated into the converter as shown by the diagram in Fig. 5(a). Figs

5(b)-(c) show the simulation results indicating that the output transient peaks are well suppressed, and the outputs become steady in a short time. The average outputs of 105.35 V_{dc} , and 2.11 A_{dc} are obtained. At PCC, the current harmonic distortion is reduced to 30.81 % THD_i, and the power factor is as high as 0.9096 (lagging).

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Let's consider another design example where the load of the converter is a series RL. This can be a good representation of a motor load in steady-state operation. The zeta converter has 220V (311 peak voltage), 50Hz supply. Its load is assumed to be 23.7 mH series with an R, 800W, 11A rated. Firstly, the equivalent load resistance, and the average output voltage can be calculated as $P_o/I_o^2 = 800/11^2$ = 6.6116 Ω , and $P_o/I_o = 800/11 = 72.73$ V_{dc}, respectively. The voltage gain and the duty cycle are $M = V_o/V_g = 72.73/311 = 0.2339$, and $d_1 = 0.5391$, respectively. The expressions (26) and (27) are used next to obtain the inductances L_o and L_m as follows:

$$L_{o} \geq \frac{(1-d_{1})R}{2f_{s}} = \frac{(1-0.5391)(6.6116)}{2(50\times10^{3})}$$
$$= 30.47\,\mu H$$
$$L_{m} \geq \frac{(1-d_{1})^{2}R}{2n^{2}f_{s}d_{1}} = \frac{(1-0.5391)^{2}(6.6116)}{2(0.2)^{2}(50\times10^{3})(0.5391)}$$
$$= 651.31\mu H.$$

Thus, the chosen inductances are $L_o = 40 \mu H$, and $L_m = 700 \mu H$.

The capacitances C_o and C_1 are calculated according to Eqs. (24) and (25), and obtained as

$$C_{o} \geq \frac{V_{o}(1-d_{1})}{8f_{s}^{2}L_{o}\Delta V_{Co}} = \frac{(72.73)(1-0.5391)}{8(50\times10^{3})^{2}(40\times10^{-6})(1.26)}$$
$$= 43.66\mu F$$

$$C_{1} \ge \frac{V_{o}d_{1}}{f_{s}R\Delta V_{C1}} = \frac{(72.73)(0.5391)}{(50 \times 10^{3})(6.6116)(1.26)}$$
$$= 94.14\mu F$$

Table 2 summarizes the case.

Parameters	Values	
peak input voltage (Vg)	311 V 50 Hz	
avg. output voltage (V _o)	72.73 V	
avg. output current (I _o)	11 A	
switching frequency (f _s)	50 kHz	
L _m , L _o	700 uH, 40 uH	
C_1, C_o	100 uF, 4.4 mF	
turns ratio	0.2	
of hf transformer (n)		
duty cycle (d ₁)	0.5391	
М	0.2339	

Table 2 Design results (converter with an RL load).



(a) source voltage and current waveforms



(b) output voltage and current waveforms



The voltage and current waveforms obtained from PSIM are illustrated in Fig. 7. With the assumed RL load, the average output voltage and current of the converter are 72.83 V_{dc} , and 10.93 A_{dc}, respectively, according to the expectation. Without a filter, the current harmonic is as high as 143%, and the power factor is 0.5844 (lagging). An LC filter having $L_f = 50$ mH and $C_f = 137$ nF is added to the input frontend. The duty cycle is also increased to 0.5861 to compensate for the associated voltage drop. The simulation results indicate that the average output voltage is $72.34 V_{dc}$, the average output current is 10.86 A_{dc}, with 28.83 % THDi and power factor of 0.8842 (lagging). It is observed that the LC filter significantly decreases the pulsation of the source current, in turn drastically reduces the harmonic distortion, and increases the power factor of the circuit.







(b) output voltage and current waveforms

Fig. 8 Simulation results (RL load, with LC filter).

7 Conclusion

This paper has explained the principle of operation of the zeta converter with the design formulas for the continuous current mode (CCM) of operation. It also presents the development of the state-variable, and the transfer function models of the isolated zeta converter. Two design examples are given for the cases of R, and series RL loads. Detailed simulation results are presented, and the following conclusions are drawn: (i) the output voltage and current contain a considerable amount of ripples which have to be limited in practice, (ii) the converter without an input filter produces a great deal of current harmonic, and possesses a low power factor, and (iii) a simple LC input-filter can be used at first sight to reduce the harmonic, and to increase the factor. The issues concerning power discontinuous current mode (DCM), the control of the output power, the power factor correction, and the harmonic reduction are under investigations.

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LIST OF SYMBOLS

- d_1 = "on" duty cycle of switch S
 - = "off" duty cycle of switch S
- $f_s = switching frequency (Hz)$
- i_{C1} = current of capacitor C_1 (A)
- i_{Co} = current of capacitor $C_o(A)$
- i_D = current of diode D (A)

 d_2

ig	=	output current of rectifier (A)
i_{Lm}	=	current of inductance $L_m(A)$
i_{Lo}	=	current of inductance $L_{o}(A)$
i _o	=	output current (A)
i _p	=	primary current of transformer (A)
i _s	=	input current (A)
n	=	turns ratio of high frequency
		transformer
V_{C1}	=	voltage of capacitor C_1 (V)
V _o , V _o	=	output voltage of rectifier (time
0. 0		domain), peak value (V)
V _{Lm}	=	voltage of inductance $L_m(V)$
VLO	=	voltage of inductance L_0 (V)
v_0, V_0	=	output voltage (time domain),
		average value (V)
Vs	=	sinusoidal source voltage (V)
$G_{vd}(s)$	=	transfer function of duty cycle to
, a.v. ,		output voltage
$G_{vv}(s)$	=	transfer function of input to output
		voltages
Μ	=	voltage gain of peak input to output
		voltages
THDi	=	total harmonic current distortion
Ts	=	switching period (s)
Δi_{Lm}	=	current ripple of inductor $L_m(A)$
Δi_{Lo}	=	current ripple of inductor L_0 (A)
ΔV_{C1}	=	voltage ripple of capacitor $C_1(V)$
ΔV_{Co}	=	voltage ripple of capacitor C _o (V)
γ	=	output ripple factor
~		
,	=	small-signal perturbation
-	=	transferred circuit parameters from
	-	secondary to primary
		secondary to primary