Small-Signal Modeling of Nonideal Magamp PWM Switch

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Abstract—Circuit-based small-signal models of the magnetic amplifier (magamp) which include the effects of the nonideal squareness of the magamp’s core B-H curve are derived for both the voltage-reset and current-reset control techniques. Since in this modeling approach the small-signal behavior of the magamp is described by equivalent circuits, circuit simulators can be easily used to facilitate the control-loop design optimization of the magamp.

Index Terms—B-H characteristic, control-loop interactions, magnetic amplifier, small-signal modeling.

I. INTRODUCTION

THE MAGNETIC amplifier (magamp) technique is one of the most reliable and cost-effective postregulation methods for multiple-output power supplies. This is especially true for high-current postregulated outputs since at higher output currents the efficiency of linear postregulators is unacceptably low, while the complexity of more efficient switch-mode postregulators is associated with a significant cost. Although in the past the magamps were used in numerous applications, their usage has dramatically increased with the introduction of 3.3-V integrated circuits (IC’s). Namely, in today’s data-processing equipment, which is based on both 3.3- and 5-V IC’s (mixed-power designs), it is necessary to provide at least two high-current tightly regulated outputs. The most cost-effective approach to meet the tight-regulation requirements at both outputs is to directly regulate the 5-V output and postregulate the 3.3-V output using a magamp. Very often, in computer power supplies, the regulation requirements on the 12-V output, which is mainly used for driving the storage disk(s), warrant the use of a second magamp. As a result, today’s multiple-output power supplies are complex control systems which contain multiple control loops that require suitable small-signal models for the control-loop performance optimization as well as for the analysis of possible loop interactions [1].

Various issues related to the operation, design, modeling, control, and simulations of magamps were discussed in [2]–[11]. However, while the operation of magamps with nonideal magamp-core materials was analyzed in [2] and [3], the small-signal analysis, modeling, and simulations of magamps have been, so far, exclusively carried out under the assumption of the ideal squareness of the B-H curve of magamp’s core material [4]–[11].

The objective of this paper is to introduce accurate circuit-based small-signal models of magamps which include the effects of the nonideal squareness of magnetic-core materials. Since, in this approach, the small-signal behavior of magamps is described by equivalent circuits, circuit simulators such as Spice and Saber can be easily used to facilitate the control-loop design optimization of magamps as well as perform their loop-interaction analysis.

II. KEY WAVEFORMS OF NONIDEAL MAGAMP

A simplified circuit diagram of a typical, two-output forward converter with a magamp postregulator is shown in Fig. 1. In this converter, the regulation of output voltage $V_{01}$ is achieved by a pulsewidth modulation (PWM) of the duty cycle of the primary switch S, whereas output voltage $V_{02}$ is regulated by a local magamp feedback loop, which modulates the duration of the blocking time of magamp inductor $L_{MA}$. It should be noted that the output of the magamp reset circuit $x_{control}$ is control voltage $v_C$ for the voltage-type reset, and reset current $i_R$ for the current-type reset [3].

Fig. 2 shows the key waveforms of a nonideal magamp with the voltage reset, whereas Fig. 3 shows the assumed nonideal core B-H characteristics. The waveforms in Fig. 2 assume that load current $i_L = \left<i_{LF2}\right>$ is large compared to the magnetizing current of $L_{MA}$ so that $i_{sec2} = i_{MA}$ is zero during the time intervals in which $L_{MA}$ is not saturated and that $L_{F2}$ is large so that the $i_{LF2}$ ripple is negligible. It should be noted that the above assumptions greatly simplify the derivations of the models without compromising their accuracy.

As can be seen from Fig. 3, due to a nonideal squareness (SQ $\equiv B_{f}/B_{s}<1$) of the B-H curve, the magamp exhibits a small residual inductance when it is saturated. As a result, the commutation of output-filter inductor current $i_{LF2}$ between forward diode $D_{F2}$ and freewheeling diode $D_{FW2}$ is not instantaneous as in the case of an ideal magamp (SQ $= 1$). Namely, as seen from Fig. 2, magamp current $i_{MA}$ takes time to ramp up after $L_{MA}$ becomes saturated at $t = t_1$, and also to ramp down after secondary voltage $v_{sec2}$ becomes negative at $t = t_3$. In fact, during the commutation time $[t_3, t_4]$, during which $i_{LF2}$ commutes from $D_{F2}$ to $D_{FW2}$, the magamp core receives additional reset (additional volt seconds compared to ideal magamp) $\Delta_{mag}$, as indicated in Fig. 2. This additional reset is caused by the commencement of the conduction of $D_{FW2}$ at $t = t_3$ which clamps voltage $v_A$ to zero during the entire commutation interval $[t_3, t_4]$.
With $v_A = 0$, the core receives more reset volt seconds during $[t_3, t_5]$ interval compared to the ideal case where $v_A = x_{\text{Control}} = v_C < 0$ from $t = t_3$ to $t = t_5$. This additional reset not only limits the minimum blocking time, i.e., the maximum duty cycle of the magamp, but also has a significant effect on the small-signal behavior of the magamp regulator.

It should be noted that the reverse recovery of forward diode $D_{F2}$, or its capacitance in case of a Schottky diode, has the identical effect on the magamp behavior as the nonideal squareness of the magamp core. Namely, once $i_{MA}$ reaches zero at $t = t_4$, it continues to flow in the reverse (negative) direction while $D_{F2}$ recovers. Since during the recovery period $v_A = 0$, the core receives yet another additional reset. Because this additional reset can be minimized by employing a simple circuit described in [2] and [5], the effect of the reverse-recovery characteristic of diode $D_{F2}$ on small-signal behavior of the magamp is neglected.

The operation of the magamp with the current reset is similar to that of the magamp with the voltage reset, except for differences in the $v_A$ and $v_{MA}$ voltage waveforms. Since $i_{MA}$ and $v_B$ waveforms, which are used in the small-signal derivations of magamp power stage, are identical for both reset techniques, the current reset will be addressed in the next section, where the small-signal models for the voltage and current resets are derived.

III. SMALL-SIGNAL MODELING

The small-signal models of the magamp were derived using the PWM-switch approach introduced and described in [12]. In this approach, the small-signal characteristics of a converter power stage are derived by perturbing the average currents and voltages of a three-terminal PWM switch. Therefore, to apply this approach, it is necessary to identify the three-terminal PWM-switch structure of the magamp postregulator.

As can be seen from Fig. 2, the duty cycle of the voltage waveform $v_B$ at the input of the output filter, assuming $t_{\text{rise}}^{MA} = 0$ (i.e., ideal squareness), is a function of the primary-switch duty cycle $d = T_{ON}/T_S$ and the blocking duty cycle $d_B = T_{B}/T_S$. This dependence on $d$ and $d_B$ can be represented by an equivalent magamp PWM switch shown in Fig. 4(d), obtained by equivalent substitutions given in Fig. 4(a)–(c). As indicated in Fig. 4(b), generally, the blocking duty cycle $d_B$ of the magamp is a function of control signal $x_{\text{Control}}$, reflected transformer reset voltage $v_R = v_{\text{reset}}/n_2$, and duty cycle $d$. The instantaneous (denoted with ˜) and average (denoted with lower case letters) terminal voltages and currents of the three-terminal magamp PWM switch in Fig. 4(d) are shown in Fig. 5.

A. Power Stage Models

From Fig. 5, the relationships between the average terminal voltages and currents are

$$v_B = v_C(d - d_B) - v_C \frac{t_{\text{rise}}^{MA}}{T_S}$$  \hspace{1cm} (1)

$$i_{MA} = i_{LF}(d - d_B) - \frac{i_{LM}}{2} \left[ \frac{t_{\text{rise}}^{MA}}{T_S} - \frac{t_{\text{fall}}^{MA}}{T_S} \right]$$  \hspace{1cm} (2)

where $t_{\text{rise}}^{MA}$ and $t_{\text{fall}}^{MA}$ are the rise and fall times of $i_{MA}$, respectively. From the $i_{MA}$ waveform in Fig. 2, $t_{\text{rise}}^{MA}$ and $t_{\text{fall}}^{MA}$ can be written (using the notation in Fig. 2) as

$$t_{\text{rise}}^{MA} = \frac{n_2 i_{LF} L_{MA} f_S}{u_{\text{in}}} = \frac{I_0 Z_S}{u_C}$$  \hspace{1cm} (3)

$$t_{\text{fall}}^{MA} = \frac{n_2 i_{LF} L_{MA} f_S}{u_{\text{reset}}} = \frac{I_0 Z_S}{u_R}$$  \hspace{1cm} (4)

where $Z_S = L_{MA} f_S$ is the saturation impedance of the magamp, $f_S = 1/T_S$ is the switching frequency of the primary switch, $I_0 = i_{LF}$ is the output current, $u_C = u_{\text{in}}/n_2$ is the positive secondary-winding voltage at the magamp input (Fig. 2), and $u_R = u_{\text{reset}}/n_2$ is the negative (reset) secondary-winding voltage at the magamp input (Fig. 2). It
Fig. 2. Key waveforms of nonideal magamp with voltage reset ($v_{\text{reset}} = v_C$).

Fig. 3. Nonideal B-H characteristic. Squareness of characteristic defined as $\text{SQ} = B_r/B_s$, where $B_r$ is measured at $H_c = 80$ A/m. Slopes of the B-H curve are proportional to unsaturated $L_{\text{MA,unsat}}$ and saturated $L_{\text{MA,sat}}$ inductances of the magamp.

should be noted that for the RCD-reset method of the forward-converter transformer core, reset voltage $v_R$ is constant, i.e., it is independent of input voltage $v_G$. On the other hand, for the active-clamp reset technique, $v_R$ is dependent on $v_G$. To simplify the analysis, in the following derivations, it is assumed that $v_R$ is constant, i.e., $v_R = V_R$. As a result, the derived models represent approximate models for magamps operating in the forward converters with active-clamp reset.

Substituting (3) and (4) in (1) and (2), the average $v_B$ and $i_{MA}$ can be written as

$$v_B = v_G(d - d_B) - i_LZ_S$$

$$i_{MA} = i_L(d - d_B) - \frac{1}{2}i_L^2Z_S \left[ \frac{1}{v_G} - \frac{1}{V_R} \right].$$

By perturbing the average quantities in (5) and (6) around their dc values (denoted with uppercase letters), i.e., by setting $v_G = V_G + d_G, v_B = V_B + d_B, d = D + d, d_B = D_B + d_B, i_{MA} = I_{MA} + i_{MA},$ and $i_L = I_L + i_L,$ where
(a) Fig. 4. Reduction of magamp to equivalent three-terminal PWM switch.

(b) Fig. 5. Relationships between instantaneous and average terminal voltages and currents.

\[ \tilde{v}_G(t) = v_G(t) \]
\[ \tilde{v}_B(t) = v_B(t) \]
\[ \tilde{i}_{LF}(t) = i_{LF}(t) \]
\[ \tilde{i}_{MA}(t) = i_{MA}(t) \]

\( ^{\wedge} \) denotes a small-signal perturbation, the dc and small-signal models can be obtained.

From the perturbed (5) and (6), after a number of simple algebraic operations, the dc model of the magamp can be extracted as

\[ I_{MA} = I_{LF}(D - D_B) - \frac{1}{2} \frac{L_{LF}}{V_G} Z_S \left( \frac{1}{V_G} - \frac{1}{V_R} \right) \]  
\[ V_B = V_G(D - D_B) - I_{LF} Z_S \]  

whereas the small-signal model, after the second-order terms are neglected, is

\[ \hat{i}_{MA} = I_{LF}(\hat{d} - \hat{d}_B) + \hat{i}_{LF}(D - D_B) \]
\[ -I_{LF} Z_S \left( \frac{1}{V_G} - \frac{1}{V_R} \right) \hat{i}_{LF} + \frac{1}{2} \left( \frac{I_{LF}}{V_G} \right)^2 Z_S \hat{g}_G \]
\[ \hat{v}_B = V_G(\hat{d} - \hat{d}_B) + \hat{v}_G(D - D_B) - Z_S \hat{i}_{LF}. \]

The set of equations describing the dc model can be represented by the equivalent circuit shown in Fig. 6, whereas the set of equations describing the small-signal model can be represented by the equivalent circuit shown in Fig. 7.

As can be seen from Fig. 6, the effect of the nonideal squareness of the core on the dc behavior of the magamp is modeled by the saturated inductance \( Z_S / (D - D_B)^2 \) and the current source \( 1/2 Z_S (1/V_G - 1/V_R) I_{LF}^2 \). Since impedance \( Z_S \) is connected in series with the input and output terminals of the model, it makes the dc output voltage of the nonideal magamp dependent on the output current. Similarly, as can be seen from Fig. 7, the effect of the nonideal squareness of the core on the small-signal behavior of the magamp is modeled by impedance \( Z_S / (D - D_B)^2 \), current source \( I_{LF} Z_S (1/V_G - 1/V_R) \), and conductance \( g_i = 1/2 (I_{LF}/V_G)^2 Z_S \). Due to \( Z_S \) and \( g_i \), the small-signal transfer functions of the nonideal magamp are damped. This damping is not accounted for when ideal magamp core characteristics are assumed. It should be noted that the ideal dc and small-signal magamp models can be obtained from the models in Figs. 6 and 7 by setting \( Z_S = 0 \).

The small-signal equivalent circuit shown in Fig. 7 can be modified to include control variable \( \delta_{Control} \) instead of \( \hat{d}_B \). To make this modification, it is necessary to find the functional...
relationship between control signal $x_{\text{Control}}$ and blocking duty cycle $d_B$. Generally, this functional relationship depends on the type of the magamp core reset circuit. The magamp reset circuit shown as a block in Fig. 1 can be implemented either by using the voltage reset of the magamp core, shown in Fig. 8(a), or the current reset, shown in Fig. 8(b). In the voltage-reset circuit, control voltage $v_C$, which is proportional to the error voltage $v_{E,2}$, is applied to the magamp core to obtain the desired core reset, i.e., to obtain the desired blocking time $t_B$. In the current-reset circuit, reset current $i_R$, which is proportional to the error voltage, is used to control the magamp reset.

B. Voltage Reset

From the $v_{MA}$ waveform in Fig. 2, which represents the voltage across the magamp inductor with the voltage reset, the volt-second (flux) balance during the blocking and resetting time intervals yields

$$\frac{v_{MA}}{n_2}T_B + \frac{v_{MA}}{n_2}T_{\text{reset}} = \left(\frac{v_{\text{reset}}}{n_2} + v_C\right)T_{\text{reset}} - v_C\frac{i_{\text{in}}}{L_{\text{in}}}$$

where $T_{\text{reset}}$ is the reset time of the transformer core. This reset time can be calculated from the volt-second balance requirement of the transformer core, i.e., from

$$v_{\text{in}}T_{\text{can}} = v_{\text{reset}}T_{\text{reset}}$$

After substituting $i_{\text{can}}$ and $i_{\text{fall}}$ from (3) and (4) and $T_{\text{reset}}$ from (12) in (11), the relationship between $d_B$ and $i_C$ can be derived as

$$d_B = \frac{1 + V_C}{V_R}i + \frac{V_C}{V_R}\frac{Z_S}{V_C}i_{\text{can}} - \left(1 + \frac{V_C}{V_R}\right)\frac{Z_S}{V_C}i_{\text{fall}} + \frac{D_B}{V_C + V_R}i_C.$$  \hspace{1cm} (13)

Using the relationship given in (13) to eliminate $i_B$ in the equivalent circuit in Fig. 7, the equivalent circuit, small-signal model of the magamp with voltage reset shown in Fig. 9 is obtained. Since all the dependent current and voltage sources in this model are controlled by currents or voltages, the implementation of this model in circuit simulators is easier compared to the model in Fig. 7, whose implementation requires additional modeling steps.

C. Current Reset

The major difference between the current reset and voltage reset of the magamp is seen in the magamp voltage waveform $v_{MA}$ during the magamp reset interval. Fig. 10 shows the $v_{\text{sec2}}$, $v_{MA}$, and $i_{\text{MA}}$ waveforms of a magamp with the current reset, along with the equivalent circuits of the magamp during different phases of the reset. The other waveforms of the magamp with the current reset are identical with those for the voltage reset shown in Fig. 2. It should be noted that in Fig. 10(b) the current reset circuit shown in Fig. 8(b) is modeled by an ideal current source $i_R$ with output resistance $R_I$ in parallel.

As can be seen from Fig. 10(a), the reset period can be divided into three intervals. During the $[t_3 - t_4]$ interval, $i_{\text{can}}$ linearly decreases from $i_0$ toward zero and $v_{\text{MA}} = -\frac{v_{\text{reset}}}{n_2}$. In this interval, the behavior of the magamp with current reset is identical to that of the voltage reset. After $i_{\text{can}}$ reaches zero at $t = t_4$, diode $D_f$ in the reset circuit in Fig. 8(b), continues to conduct $i_R$ because the current through unsaturated $L_{\text{MA}}$ can not increase (in the negative direction) instantaneously. Because of the diode conduction, $v_{\text{MA}} = -\frac{v_{\text{reset}}}{n_2}$, as shown in Fig. 10. When $i_{\text{can}}$ reaches $i_R$ at $t = t_5$, diode $D_f$ ceases conducting, and voltage $v_{\text{MA}}$ starts exponentially decreasing with time constant $\tau = L_{\text{MA reset}}/R_I$, as shown in Fig. 10(a). At $t = t_5$, the transformer reset is finished so that $v_{\text{MA}} = 0$. It should be noted that the current reset requires that unsaturated inductance $L_{\text{MA reset}}$ of magamp be finite, i.e., that the slope of
B-H curve in the unsaturated region be less than $\infty$ (nonideal core).

Assuming that $\tau \ll T_{\text{sec}}$, i.e., assuming that the reset circuit in Fig. 8(b) approaches an ideal current source, the approximation of the volt-second balance on $L_{MA}$ can be obtained from the $u_{MA}$ waveform in Figs. 2 and 10(a) as

$$\frac{v_{\text{sec}}}{n_2} t_B + \frac{v_{\text{sec}}}{n_2} t_{MA} = \frac{V_{\text{sec}}}{n_2} L_{MA} + \frac{V_{\text{sec}}}{n_2} t_{CR}$$

(14)

where $t_{CR}$ is calculated from Fig. 10(b) when diode $D_I$ is conducting as

$$t_{CR} = \frac{L_{\text{MA}}}{V_{\text{sec}}/n_2} t_R = \frac{L_{\text{MA}}}{V_{\text{sec}}/n_2} t_{R \text{f}}.$$

(15)

Substituting $t_{CR}$ and $t_{MA}$ from (3) and (4), and $t_{CR}$ from (15) into (16), the small-signal relationship between $\hat{d}_B$ and $\hat{q}_R$ can be derived as

$$\hat{d}_B = \frac{Z_M}{V_G} \hat{i}_R - \frac{D_B}{V_G} \hat{q}_R.$$

(16)

Using the relationship in (16) to eliminate $\hat{d}_B$ in equivalent circuit in Fig. 7, the equivalent circuit, small-signal model of the magamp with current reset shown in Fig. 11 is obtained.

It should be noted that the presented models do not take into account the phase shift of the magamp modulator [4]. As a result, they are only accurate at frequencies at which the modulator phase shift can be neglected. Since for practical magamp implementations the modulator phase shift becomes important at frequencies above the crossover frequencies of the magamp loop, the derived models are quite accurate in predicting the close loop behavior of practical magamps. If necessary, the modulator phase shift as well as the B-H curve dynamic resistance [4], [5] can be incorporated into the models.

Finally, it should be noted that the modulator gain of the primary switch and its circuit implementations are discussed in [13] and [14].

IV. MODEL VERIFICATION

The experimental verifications of the derived models were performed on an off-line 100-W two-output power supply implemented with a 100-kHz forward-converter power stage. The main output, implemented with the current-mode control, is rated at $V_{G1} = 5$ V and $I_{G1} = 16$ A, whereas the rating of the magamp-regulated output is $V_{G2} = 3.3$ V and $I_{G2} = 6$ A.

The magamp inductor was implemented with the MS 12 $\times$ 8 $\times$ 4.5-W Toshiba core with six turns of AWG#18 magnet wire. The current-reset method is used to reset the magamp core.

Fig. 12 shows the measured and simulated (ideal and non-ideal) control-to-output characteristics of the experimental magamp with current reset. The values of the relevant circuit parameters and the parameters of the nonideal and ideal models are given in Tables I and II. The saturated and unsaturated inductances of the magamp, $L_{\text{MA}}$ and $L_{\text{MA}}$, were estimated from the measured B-H characteristic at 100 kHz. As it can be seen from Fig. 12, the measured amplitude and phase characteristics are in good agreement with the corresponding simulated characteristics obtained using the proposed nonideal magamp small-signal model. The proposed nonideal model accurately models the damping in the control-to-output transfer function. The discrepancy between the measured phase and the predicted phase which is observed at frequencies above 5 kHz is caused by the phase shift of the modulator, which is not included in the model given in Fig. 11. Because the crossover frequency of the main loop of 8.5 kHz is higher than that of the magamp loop which is approximately 2.5 kHz, no measurable interaction between the loops is noticeable in Fig. 12.

If the crossover frequency of the main loop is reduced below the crossover frequency of the magamp loop, the two loops exhibit strong interactions. Fig. 13 shows the simulated and measured control-to-output transfer functions of the experimental converter when the crossover frequency of the main loop is reduced to 0.1 kHz, i.e., below the crossover
frequency of the magamp loop. As can be seen from Fig. 13, the main loop affects the control-to-output transfer function of the magamp by reducing its gain in the 80–500-Hz range. The agreement between the simulated and measured gain is very good up to 10 kHz, whereas the simulated and measured phase show good agreement up to approximately 6 kHz. Above 6 kHz, the modulator phase lag becomes significant.

It should be noticed that besides the saturated impedance \( Z_S \) of the magamp, the load resistance, the resistances of the output filter, and the parasitic interconnect resistances on the secondary side, the winding loss and the core loss of the magamp also contribute to the damping in the control-to-output transfer function. The winding loss of the magamp can be easily included in the model by adding the winding resistance in series with the saturation impedance in series with the saturation impedance of the magamp. In the experimental circuit, the ac resistance of the magamp winding was approximately 3 mΩ and it was neglected compared to \( Z_S = 54 \text{ mΩ} \). The core loss, which is usually dominated by the winding loss at larger load current, has not been taken into account in the developed model. However, if necessary, the core loss can be incorporated into the model as proposed in [5]. It should be also noticed that the parasitic inductances in series with the magamp such as the leakage inductance of the transformer and layout inductances have the same effect on the small-signal model of the magamp as saturated inductance of the magamp. Since in the experimental circuit these parasitic inductances were much smaller than the saturated inductance of the magamp, they were neglected. If necessary, the effect of the leakage inductance of the transformer and interconnect inductance can be easily taken into account by increasing the value of saturated magamp inductance \( L_{\text{sat}}^{\text{mag}} \) for the amount of the parasitic inductances.

V. SUMMARY

Circuit-based small-signal models of magamps which are suitable for the analysis of control-loop interactions in multiple-output power supplies are derived. The models include the effects of the nonideal squareness of the magnetic-switch-core B-H curve and also account for the differences in small-signal characteristics of the voltage-reset and current-
reset techniques. Since the small-signal behavior of magamps is described by equivalent circuits, circuit simulators can be easily used to facilitate the control-loop design optimization of the magamp.

**REFERENCES**


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