Analysis and Design Optimization of Magnetic-Feedback Control Using Amplitude Modulation

Brian T. Irving and Milan M. Jovanović, Fellow, IEEE

Abstract—In offline ac–dc and high-voltage dc–dc power supplies, galvanic isolation between the input and output is often implemented with optocoupler feedback. However, several disadvantages exist when implementing optocoupler feedback, such as a variable loop gain due to optocoupler tolerance and sensitivity to temperature, as well as a relatively high cost. An alternative to optocoupler feedback is to use magnetic feedback, which can be designed to have insensitivity to component tolerance, and good temperature stability. Although magnetic feedback has been in use for many years, a detailed analysis and clear design procedure has not been presented in the literature. This paper presents a thorough analysis of a magnetic-feedback implementation, and provides a comprehensive design procedure that is verified on a 7-V/15-W experimental prototype.

Index Terms—Amplitude modulation, isolated feedback, magnetic feedback, primary side control, sample and hold.

I. INTRODUCTION

GALVANIC isolation between the input and output is a requirement of offline ac–dc and high-voltage dc–dc power supplies. In the power stage, galvanic isolation is typically achieved through use of a transformer. However, in order to provide regulation of the output, a galvanically isolated feedback loop is also required. Two commonly used isolation methods are primary-side control with optocoupler feedback [1]–[4] and secondary-side control, where the primary gate drive is provided through gate-drive transformers. The drawbacks of primary-side control with optocoupler feedback include variation of loop gain due to wide current transfer ratio (CTR), sensitivity to both time and temperature [5]–[8], and a high cost. A drawback of secondary-side control is the need for an additional secondary-side supply voltage, often supplied by a separate “housekeeping” converter.

An alternative to primary-side control with optocoupler feedback and secondary-side control is to use magnetic feedback [1], [2], [9], [10]. Using a very small coupling transformer, a modulator, and a sample-and-hold circuit, a signal from the secondary side can be passed to the primary using either AM or FM. A magnetic-feedback circuit can be implemented either discretely or by using an IC [10]. Although it has been used in industry for many years, very few design details appear in the literature. The goal of this paper is to analyze an AM magnetic-feedback implementation, and to provide a clear, step-by-step design procedure in order to optimize the circuit performance.

II. PRINCIPLES OF AM MAGNETIC FEEDBACK

As shown in Fig. 1, magnetic feedback can be implemented using a small dc–dc transformer, which acts as both the modulator and demodulator, to pass error voltage $V_{EA}$ from the secondary side to the primary side. The dc–dc transformer consists of a small coupling transformer and diode, which is excited by an external source to form the modulator, with its switching frequency equal to the modulator carrier frequency. The rectified output of the dc–dc transformer acts as a sample-and-hold circuit, or demodulator.

The coupling transformer ($T_C$) can be configured either as a forward-type or flyback-type converter, as shown in [9]. Both the carrier and sampling frequencies, which are synchronized, can be equal to or higher than the power-stage switching frequency, depending on the desired volume of coupling transformer $T_C$ and the desired loop performance. Generally, the loop gain of the converter is negatively impacted by the sampling delay of the demodulator.

A simplified block diagram of a general power stage with magnetic feedback implemented with a flyback-type dc–dc transformer is shown in Fig. 2. Error amplifier (EA) output $V_{EA}$ represents the amplified difference between output voltage $V_O$ and reference voltage $V_{REF}$. An isolated modulator is implemented by coupling transformer $T_C$, which is magnetized by current source $i_C$ based on an external carrier signal. While
current source $i_C$ is on, diode $D_{EA}$ is reversed biased, sampler switch $S_H$ is off, and current $i_C$ divides between magnetizing inductance $L_M$ and resistor $R_M$. Once current source $i_C$ turns off, magnetizing current $i_M$ forward biases diode $D_{EA}$, and error voltage $V_{EA}$ is reflected to the primary side with a reverse polarity. Sampler switch $S_H$, which is synchronized to the modulator, turns on during the demagnetization period of coupling transformer $T_C$, and capacitor $C_H$ holds the sampled error voltage $V^*_{EA}$, which is then compared at the pulsewidth modulator (PWM) to periodic ramp voltage $V_{RAMP}$ in order to generate the gate-drive signal to the power stage.

Generally, coupling transformer $T_C$ operates as a flyback converter because error voltage $V_{EA}$ is sampled during the demagnetization period of coupling transformer $T_C$. As shown in [9], it is also possible to operate coupling transformer $T_C$ as a forward converter, i.e., to sample error voltage $V_{EA}$ during the magnetization period of coupling transformer $T_C$.

III. ANALYSIS OF AM MAGNETIC FEEDBACK

An implementation of an AM magnetic-feedback circuit applied to a forward converter with synchronous rectifiers and current-mode control is shown in Fig. 3. EA is implemented with transconductance amplifier TLV431 and capacitor $C_{KA}$ that allows TLV431 to have a low output impedance at high frequencies, and resistor $R_{ST}$ and diode $D_{ST}$ are implemented to facilitate startup. A dc–dc transformer is used as part of the modulator, and has both a flyback winding for sampling error voltage $V_{EA}$ and a forward winding to supply amplifier TLV431 and provide a turn-on signal for synchronous-rectifier turn-off switch $Q_{OFF}$. Coupling transformer $T_C$ is magnetized by a simple current source consisting of p-n-p transistor $Q_1$, resistors $R_E$, $R_{B1}$, and $R_{B2}$, and primary $V_{CC}$ is supplied by auxiliary winding $N_S$ of inductor $L_F$. The current source is turned on and off by a carrier signal that is synchronized and equal to the main converter switching period $T_S$, and which has a fixed on time $T_A$. The demodulator is implemented with a simple peak detector, where diode $D_H$ acts as sampling switch $S_H$. Finally, sampled error voltage $V^*_{EA}$ is level-shifted and inverted, and compared at the PWM modulator to a ramp that is proportional to switch current $i_S$.

The AM magnetic-feedback circuit can be simplified to three topological stages, as shown in Fig. 4, for the case when the current source is on, i.e., when the carrier signal is high, as shown in Fig. 4(a), and when the current source is off, i.e., when the carrier signal is low, as shown in Fig. 4(b) and (c). Key switching waveforms of a single switching cycle are shown in Fig. 5. While the carrier signal is high, the current source is on, diodes $D_H$ and $D_S$ are reverse biased, and diode $D_2$ is forward biased. While diode $D_2$ is forward biased, voltage $V_{LM}$ across magnetizing inductance $L_M$ is equal to $V_{CV} + V_F$, where $V_{CV}$ is the voltage across capacitor $C_{CV}$, $V_F$ is the forward voltage drop of diode $D_2$, and magnetizing current $i_{LM}$ begins to increase linearly from zero. Current $i_C$ then divides between magnetizing inductance $L_M$, equivalent gate resistor $R_G$, and diode $D_2$. Meanwhile, voltage $V^*_{EA}$ across capacitor $C_H$, which in the previous switching cycle was charged to error voltage $V_{EA}$, slowly discharges through resistors $R_{IA}$ and $R_{IB}$. Once the carrier signal goes low, the current source turns off, diodes $D_1$ and $D_H$ become forward biased, diode $D_2$ becomes reverse biased, and voltage $V_{LM}$ is equal to $- (V_{EA} + V_F)$. Inductance $L_M$ begins to demagnetize, as shown in Fig. 5. Finally, hold capacitor $C_H$ peak charges to error voltage $V_{EA}$. Once magnetizing inductance $L_M$ completely demagnetizes, all diodes become reverse biased, and hold capacitor $C_H$ begins to slowly discharge through resistors $R_{IA}$ and $R_{IB}$, as shown in Fig. 4(c).

IV. DESIGN OF AM MAGNETIC FEEDBACK

A. Steady-State Design

In order to continuously sample the error voltage, coupling transformer $T_C$ must be designed with enough margin to prevent saturation. In addition, saturation of current-source transistor $Q_1$ must be avoided so that current $i_C$ is insensitive to current gain $h_{FE}$ of transistor $Q_1$, since $h_{FE}$ is sensitive to both temperature and tolerance.

Both coupling transformer $T_C$ and current-source transistor $Q_1$ can saturate due to load variations, as shown in Fig. 6. At light load, control voltage $V_C$ is low, and therefore, voltage $V_{CV}$ is low, resulting in a low turn-on and turn-off slope of magnetizing current $i_{LM}$, and therefore, a short deadtime $T_d$ of coupling transformer $T_C$. As the load increases, control voltage $V_C$ increases, and therefore, voltage $V_{CV}$ increases, which, in turn, increases the turn-on and turn-off slope of magnetizing current $i_{LM}$ and increases deadtime $T_d$. However, as voltage $V_{CV}$ increases, current-source transistor $Q_1$ approaches saturation, as shown in Fig. 6.

Both coupling transformer $T_C$ and current-source transistor $Q_1$ can saturate due to temperature variations, as shown in Fig. 7. As the temperature of current-source transistor $Q_1$ increases, the forward voltage drop of emitter to base p-n junction decreases, leading to an increase in collector current $i_C$. This, in turn, leads to an increase in voltage $V_{CV}$, which pushes current-source transistor $Q_1$ closer to saturation. In addition, the turn-on slope of magnetizing current $i_{LM}$ increases, while the turn-off
slope remains constant since error amplifier voltage $V_{EA}$ remains constant. As a result, deadtime $T_d$ decreases, and coupling transformer $T_C$ also approaches saturation. It should be noted that although magnetic-feedback designs are sensitive to temperature, this sensitivity can be minimized through proper design, unlike optocoupler feedback designs whose widely varying temperature-dependent current transfer ratio cannot be minimized. Magnetizing inductance $L_M$ of coupling transformer $T_C$ also varies with temperature; generally, as the temperature increases, the magnetizing inductance increases, and voltage $V_{CV}$ increases since the average magnetizing current decreases. Transistor $Q_1$ approaches saturation since base voltage $V_B$ increases, while deadtime $T_d$ remains unchanged since both the turn-on and turn-off slopes of magnetizing current $i_{LM}$ change proportionally. Generally, it is recommended that the maximum value of magnetizing inductance be used throughout the calculations.

To prevent saturation of transistor $Q_1$, base voltage $V_B$ must be more than 1 p-n diode drop greater than voltage $V_{LM}$. By selecting a desired maximum voltage level of error voltage $V_{EA}$, e.g., $V_{E\text{Amax}} = 4$ V, and by selecting a reasonable value for resistor $R_{IB}$, e.g., 10–50 kΩ, resistor $R_{FB}$ can be determined

$$R_{FB} = R_{IB} \left( \frac{V_{E\text{max}}}{\Delta V_{EA}} \left( 1 + \frac{V_{E\text{Amin}}}{V_{REF2}} \right) - \frac{V_{E\text{min}}}{\Delta V_{EA}} \left( 1 + \frac{V_{E\text{Amax}}}{V_{REF2}} \right) + 1 \right)$$  

where $V_{E\text{max}} = 4.2$ V, $V_{E\text{min}} = 0$ V, $V_{E\text{Amin}} = 1.24$ V, and $\Delta V_{EA} = V_{E\text{Amax}} - V_{E\text{Amin}}$. Next, resistor $R_{IA}$ and hold capacitor $C_H$ can be calculated as

$$R_{IA} = \frac{R_{FB} R_{IB}}{R_{FB} - R_{IB}} \left( 1 + \frac{V_{E\text{Amin}}}{V_{REF2}} \right)$$  

$$C_H = \frac{10 T_S}{2 \pi (R_{IA} + R_{IB})}$$  

where the value of $C_H$ is a tradeoff between a low ripple and excessive delay introduced in the feedback loop.

Voltage $V_{CV}$ can be calculated based on the maximum steady-state error voltage $V_{EA}$ and selected coupling transformer deadtime $T_d$ and selected current source on time $T_A$ as

$$V_{CV} = - \left( 1 - \left( 1 - \frac{T_d}{T_S} \right) \frac{T_S}{T_A} \right) (V_{EA} + V_F) - V_F.$$  

Next, the required average magnetizing current $I_{LM}$ can be calculated as

$$I_{LM} = \frac{V_{CV} + V_F}{2 L_M} T_A \left( 1 - \frac{T_d}{T_S} \right).$$  

From the minimum TLV431 current, average diode current $I_{S2}$ can be selected, resistor $R_K$ can be calculated as

$$R_K = \frac{(V_{CV} - V_{EA})}{I_{S2}}$$
and average diode current $I_{S1} + I_1$ can be calculated as

$$I_{S1} + I_1 = -\frac{V_{CV} + V_F}{V_{EA} + V_F} \frac{T_A}{T_S} \left( -\frac{(V_{CV} + V_d)T_A}{2L_M} + \frac{V_{KA} + V_d}{R_{GD}} \right).$$  \hspace{1cm} (7)

The required average collector current $I_C$ can now be determined since

$$I_C = I_{S2} - I_{S1} - I_1 + I_{LM}$$  \hspace{1cm} (8)

and finally, the peak collector current $i_{pk}$ can be determined as

$$i_{pk} = I_C T_S.$$  \hspace{1cm} (9)

Since saturation of transistor $Q_1$ must be avoided, base voltage $V_B$ should be set approximately 2 V higher than voltage $V_{CV} + V_F$. Resistor $R_E$ can then be calculated as

$$R_E = \frac{V_{CC} - V_B - V_{ebf}}{i_{pk}(1 + 1/h_{FE})}.$$  \hspace{1cm} (10)

where voltage $V_{ebf}$ is the base–emitter voltage of $Q_1$. It should be noted that voltage $V_{CC}$, which should be selected greater than voltage $V_B$, may be excessively high based on the selection of maximum error voltage $V_{EAmx}$. Additional iterations may be needed to find both the optimal value of voltages $V_{EAmx}$ and $V_{CC}$. By selecting a reasonable value of base resistor $R_{B2}$, e.g., 1–10 kΩ, base resistor $R_{B1}$ can be calculated assuming that
current gain $h_{FE} \gg 1$

$$R_{B1} = R_{B2} \left( 1 - \frac{V_B}{V_{CC}} \right). \quad (11)$$

B. Small-Signal Design

As long as coupling transformer $T_C$ does not saturate, the effect of the modulator/demodulator on the loop gain is simply to add a sample-and-hold delay. Generally, parasitic such as leakage inductance of coupling transformer $T_C$ have no affect on the loop gain.

Small-signal modeling of converters implemented with current-mode control typically consists of an inner current loop and an outer voltage loop [11]–[21]. A small-signal block diagram of a converter operating with current-mode control, originally proposed in [11], is shown in Fig. 8, whereas the transfer functions are given in Table I of the Appendix. Power-stage transfer functions were derived using the model of the PWM switch [12]. Current loop $T_i$ is defined as the product of power-stage transfer function $G_{id}$, equivalent sensing resistor $R_S$, sampling gain $H_e$, and modulator gain $F_M$, i.e.

$$T_i = G_{id} R_S H_e F_M \quad (12)$$

whereas voltage loop $T_v$ is defined as the product of control-to-output transfer function $G_{vc}$, sensing gain $K_d$, error amplifier transfer function $G_{EA}$, sample-and-hold transfer function $G_{SH}$, and transfer function $G_E$, i.e.

$$T_v = G_{vc} K_d G_{EA} G_{SH} G_E. \quad (13)$$

Fig. 7. Effect of temperature variation on AM magnetic-feedback implementation.

Fig. 8. Small-signal block diagram of forward converter with current-mode control and magnetic feedback.

Fig. 9. Compensation of voltage loop using pole-zero cancellation and straight-line approximations.

Transfer function $G_{vc}$ is the control-to-output transfer function of the power stage with current loop $T_i$ closed, i.e.

$$\frac{\hat{v}_o}{\hat{v}_c} = \frac{F_M G_{vd}}{1 + T_i} \approx \frac{G_{vd}}{R_S G_{id}} \left| \frac{R_L + 1 + s/\omega_{zc}}{R_S + 1 + s/\omega_p} \right|_{T_i \gg 1}. \quad (14)$$

Generally, subharmonic oscillation can occur in designs that have a duty cycle greater than or equal to 50%. However, this can be overcome by a proper selection of compensation ramp $S_c$, as discussed in [11].

The design of error amplifier $G_{EA}$ is based on transfer functions $G_{vc}, K_d, G_{SH}$, and $G_E$, and straight-line approximations are shown in Fig. 9. It is beneficial to include capacitor $C_{FB}$ across feedback resistor $R_{FB}$ for noise immunity, which introduces an additional pole ($f_{p1}$) that should be placed well below switching frequency $f_S$. In fact, pole $f_{p1}$ can be used to cancel equivalent series resistance (esr) zero $f_{zc}$ of control-to-output transfer function $G_{vc}$. In addition, it should be noted that
sample-and-hold delay function $G_{SH}$ introduces a phase delay that should be considered before finalizing the loop design.

Design of loop gain $T_V$ should be done at full load since pole $f_P$ of control-output transfer function $G_{vc}$ increases as load resistor $R_L$ decreases, thereby resulting in the maximum crossover frequency. Selection of crossover frequency $f_{CV}$ should be well below switching frequency $f_S$, i.e., $f_{CV} \ll f_S$. In fact, crossover frequency $f_{CV}$ may be further limited due to the low open-loop gain $A_{VO}$ of TLV431. Finally, crossover frequency $f_{CV}$ is further limited by the sample-and-hold delay.

From Fig. 9, an integrator is needed to provide a high gain at low frequencies for good load regulation, while zero $f_{z1}$ is needed to cancel out control-to-output transfer function pole $f_p$. Switching ripple is attenuated by both pole $f_p$ of transfer function $G_E$ and open-loop gain $A_{VO}$ of TLV431

$$|G_{EA}(f = f_{CV})| = \frac{1}{|K_dG_{vc}(f = f_{CV})||G_{E}(f = f_{CV})|}$$

where the gain of sample-and-hold transfer function $G_{SH}$ is equal to unity. Since $f_{CV} > f_{z1}$

$$|G_{EA}(f = f_{CV})| = \frac{R_F}{R_I}.$$  \hspace{1cm} (16)

By selecting feedback resistor $R_F$ within a reasonable range (e.g., 50–200 k), resistor $R_I$ can be calculated as

$$R_I = R_F AB.$$  \hspace{1cm} (17)

By setting zero $f_{z1}$ equal to pole $f_p$, capacitor $C_{FS}$ can be calculated as

$$C_{FS} = \frac{1}{2\pi R_F f_{z1}}.$$  \hspace{1cm} (18)

Once compensation components are calculated, error amplifier transfer function $G_{EA}$ should be checked with open-loop gain $A_{VO}$ included to see if the design is optimal. Generally, open-loop gain $A_{VO}$ changes with dc operating current $I_K$ as well as the amplitude of input signal $V_{OSC}$, as shown in Fig. 10. It should be noted that the test circuit was obtained from the TLV431B on Semiconductor datasheet. Fig. 11 shows the measured open-loop gain including resistor $R_{KA}$, capacitor $C_{KA}$, and $R_K = 3\, \text{k}\Omega$.

V. EXPERIMENTAL RESULTS

To validate the design procedure, a 7-V/15-W laboratory prototype of a forward converter with magnetic feedback and amplitude modulation was designed for an input voltage range $35 < V_{IN} < 72$, and an ambient operating range of $-40^\circ\text{C}$ to $100^\circ\text{C}$. The key component values are shown in Fig. 12.

In the experimental circuit, maximum error voltage $V_{EAmax}$ was selected as 4 V, and the minimum cathode current $I_K$ of TLV431 was set to 4 mA. On-time $T_A$ of current source $i_C$ was set to 500 ns, and feedback transformer deadtime $T_d$ was designed to be half of switching period $T_S$. As a result, voltage $V_{CV}$ was approximately 11.5 V, and base voltage $V_B$ of transistor $Q_1$ was set approximately 2 V higher, ensuring that transistor
Q₁ does not saturate. This required voltage V(CC) to be greater than 13.5 V, and therefore, V(CC) was set to 16 V. Finally, the carrier frequency was set equal to switching frequency fₛ, where fₛ = 285 kHz.

Fig. 13 shows oscillograms of two different control designs of the experimental prototype. In Fig. 13(a), current-source transistor Q₁ was designed to operate in the saturation region (i.e., voltage V.CV > base voltage V.B), and in Fig. 13(b), current-source transistor Q₁ was designed to operate in the active region. For both designs, deadtime T_d was compared at room and high ambient temperatures. Fig. 13(a) shows that on-time T_A of the current source is significantly longer at high ambient temperature than at room ambient temperature. This is due to the fact that on-time T_A is dependent on current gain h₉E of transistor Q₁ when Q₁ operates in the saturation region. As a result, the magnetizing energy of coupling transformer T_C increases and deadtime T_d decreases. As the ambient temperature increases, deadtime T_d decreases until it reaches zero, which results in the loss of output-voltage feedback.

Fig. 13(b) shows that on-time T_A is nearly constant because transistor Q₁ operates in the active region. Generally, on-time T_A is independent of current gain h₉E when transistor Q₁ operates in the active region.

A comparison between the measured and calculated loop gain is shown in Fig. 14, which demonstrates a 6-kHz bandwidth, 60° phase margin, and 20-dB gain margin. The open-loop gain A_VO was measured using the test circuit shown in Fig. 11 for an oscillation input of 50 mV, and used in the voltage-loop gain calculations.

Generally, it is desirable to have a crossover frequency greater than one-tenth of switching frequency fₛ. At one-tenth of switching frequency fₛ, the phase lag due to the sample and hold is 36°, permitting at best a phase margin less than 54°, assuming that the voltage-loop gain crosses 0 dB with a −1 slope. It was found that although switching frequency fₛ was set very high (i.e., fₛ = 285 kHz) to achieve a phase margin greater than 45°, the voltage-loop crossover frequency was limited to less than 10 kHz by the low open-loop gain of TLV431.
TABLE I
KEY SMALL-SIGNAL TRANSFER FUNCTIONS

<table>
<thead>
<tr>
<th>Function</th>
<th>Expression</th>
</tr>
</thead>
<tbody>
<tr>
<td>$G_{d}$</td>
<td>$\frac{\dot{V}<em>{d}}{d} = \frac{V</em>{in}}{N} L \frac{1}{\omega_{a}}$</td>
</tr>
<tr>
<td>$F_{M}$</td>
<td>$\frac{\dot{V}<em>{M}}{V</em>{M}} = \frac{1}{(S_{M} + S_{H})}$</td>
</tr>
<tr>
<td>$G_{k}$</td>
<td>$\frac{\dot{V}<em>{k}}{V</em>{k}} = \frac{R_{FB} \cdot R_{L}}{R_{M} + R_{L} + R_{J}}$</td>
</tr>
<tr>
<td>$H_{e}$</td>
<td>$I + \frac{s}{Q_{e} \cdot \omega_{a} - \frac{s^{2}}{\omega_{a}^{2}}}$</td>
</tr>
</tbody>
</table>

$\omega_{a} = \frac{1}{r_{C} C_{F}}$  $\omega_{L_{F}} = \frac{1}{R_{L} C_{L_{F}}}$  $\omega_{o} = \frac{1}{R_{o} C_{o}}$

$T_{S} = \frac{1}{Q_{1}} = 2$  $\omega_{p} = \frac{1}{L_{F} C_{FB}}$

VI. SUMMARY

A forward converter with magnetic feedback and amplitude modulation was thoroughly analyzed, and a comprehensive steady-state and small-signal design procedure was presented. The design procedure was verified with a 7-V/15-W experimental prototype, and steady-state and small-signal measurements were provided.

APPENDIX

(See Table I at the top of the page)

REFERENCES


Brian T. Irving was born in Ossining, NY, in 1973. He received the B.Sc. degree in electrical engineering from the University of Binghamton, Binghamton, NY, in 1998.

From 1996 to 1998, he was an Engineer with Celestica, Inc., Endicott, NY. In 1998, he joined the Power Electronics Laboratory, Delta Products Corporation, Research Triangle Park, NC, where he is currently a Senior Member of R&D Staff. His current research interests include low-harmonic rectification, control techniques, current sharing, modeling, and simulation.

Milan M. Jovanović (F’01) received the Dipl. Ing. degree in electrical engineering from the University of Belgrade, Belgrade, Serbia.

He is currently the Chief Technology Officer of the Power Systems Business Group, Delta Electronics Inc., Taipei, Taiwan.