Research Article

Controlling Transformer Magnetizing Offset Current in Isolated Phase-Shift Full-Bridge Converters Using a Luenberger Observer

Angelika Neumann, Sönke Meynen, Ahmad Rahmoun, Daniel Ziegler, and Thomas Kirchartz

1Karlsruhe University of Applied Sciences, Karlsruhe, Germany
2Ads-tec Energy GmbH, Nürtingen, Germany
3Forschungszentrum Jülich, Jülich, Germany

Correspondence should be addressed to Angelika Neumann; angelika.neumann@h-ka.de

Received 21 June 2022; Revised 28 July 2022; Accepted 8 August 2022; Published 26 September 2022

This paper proposes a flexible digital control scheme for isolated phase-shift full-bridge (PSFB) converters. The required transformer suffers from inevitable imbalance of magnetic flux resulting in an increased magnetizing DC-offset current that threatens system reliability due to saturation effects. The paper addresses two major issues of the occurrence of a magnetizing DC-offset current. First, caused by the change of duty cycle due to output power regulation and second caused by initial manufacturer tolerances of devices. In contrast to common methods the novel control scheme uses a Luenberger observer to estimate the magnetizing current requiring only simple measurement of transformer voltages without additional and lossy auxiliary networks. The observer model, in combination with a PI-controller, directly interventions the duty cycle and removes any DC-offset current resulting from both issues. A detailed deviation of the state-space model of the transformer and a subsequently design of the observer are presented. Simulation and experimental results on a PSFB prototype verify the principal functionality of the proposed control scheme to prevent transformer saturation.

1. Introduction

Regarding the worldwide increasing climate problematic the portion of clean and renewable energy generation has increased over three-fold since the last ten years [1]. Unfortunately, renewable energies like photovoltaic (PV) or wind power reduce the reliability of power generation with its dependency on weather conditions [2, 3]. To utilize the maximum capacity of fluctuating renewable energies different storage technologies have been developed. Although lithium-ion battery systems are not the only solution to store energy, this technology has prevailed in various mobile DC-DC applications and also becomes more and more important for the high and medium power DC-DC charging sector. Typically, high power DC-DC systems use a liquid cooling system due to large heat development while charging batteries, but especially non-industrial DC-DC applications with medium power have to manage cooling without liquid cooling system and therefore a high efficiency is very important [4].

The high efficiency and power density of an isolated phase-shift full-bridge (PSFB) converter in Figure 1 make this topology very attractive for DC-DC medium and high power applications such as DC-storage battery systems for PV or battery chargers for electric vehicles [5−7]. Its phase-shift control scheme allows for zero-voltage-switching (ZVS) operation with negligible switching losses and reaches higher efficiency compared to conventional hard-switched topologies [8].

Two full-bridges on primary $H_A$ and secondary $H_B$ side make the topology suitable for bidirectional operation. The transformer TR, with ratio $r = N_p/N_s$ between primary $N_p$ and secondary $N_s$, winding, compensates a high voltage difference between input voltage $V_A$ and output voltage...
Figure 1: Phase-shift full-bridge converter topology. Two full-bridges \(H_A\) and \(H_B\) allow for bidirectional operation with power \(P\). Transformer \(TR\) partially compensates high voltage difference between \(V_A\) and \(V_B\) and provides galvanic isolation.

\[
V_B = V_A/r \cdot d_{\text{eff}}
\]

to avoid low duty cycles \(d_{\text{eff}}\) with decreased efficiency. Additionally, its naturally galvanic isolation satisfies the normative requirements for e.g. battery chargers in electric vehicles for safety reasons [9]. Nevertheless, a transformer isolated converter suffers from issues such as an imbalance of magnetic flux resulting in an increasing DC-offset current \(I_{h,\text{DC}}\) that drives the transformer into saturation [10]. To prevent damage due to high saturation currents and guarantee system reliability \(I_{h,\text{DC}}\) must be eliminated.

Common topologies connect a blocking capacitor in series to the transformer in the main power path to suppress any DC-offset \(I_{h,\text{DC}}\). However, placed in series with the transformer the blocking capacitor must be dimensioned for the maximum current of the converter at full power and additionally increases conduction losses due to its series resistance. As a consequence the blocking capacitor is not preferred for applications with high power density and efficiency [11]. Considering bidirectional operation, a blocking capacitor must be placed in series to primary and secondary side of transformer. This method needs several additional power devices compared to controller-based methods and therefore results in increased size of power stage and overall system costs [12, 13].

Without additional passive devices in the main power path a controller needs to eliminate \(I_{h,\text{DC}}\) [14]. Designing of a stable feedback control loop requires the measurement of \(I_{h,\text{DC}}\) as process variable. However, considering the transformer as a four terminal device the acquisition of \(I_{h,\text{DC}}\) is not directly possible as it overlays the transformer primary current \(I_p\) [15]. The current-mode control (CMC) is a very popular analog controller-based method to prevent the transformer from saturation. Its fast response due to analog comparators allows for stable operating conditions even at high switching frequencies \(f_{\text{sw}}\). Unfortunately, in certain operating conditions instabilities, known as sub-harmonic oscillations, occur [16, 17]. The slope compensation technique normally is used to compensate the oscillation but overcompensation may result in a response delay [18, 19]. Additionally, the CMC method is not suitable for flexible digital control schemes as it needs analog comparators [20].

Numerous research works concern the adjustment of PWM signals of power switches to balance the transformer magnetizing current in isolated DC-DC converters. However, the method of gathering the magnetizing current as process variable for a controller differs. [12] extends current-mode control with a hybrid peak and valley current control and does not need for slope compensation. The valley current detection requires AC-current sensing on the primary side of transformer with a suitable bandwidth for high switching frequencies. Considering bidirectional operation, the study in [21] uses primary and secondary transformer current to estimate magnetizing current from the previous switching period. A similar method in [22] only uses the averaged primary current to estimate the DC-offset from the previous period assuming the inductor current to be in steady state and average secondary current to be zero. [23] addresses the problem of magnetizing DC-offset causing DC-offsets in primary and secondary current. Here, two control loops need to scene also both, primary and secondary current and calculate the magnetizing current accordingly. In [24] a sensor directly measures the magnetic flux in the transformer core. The sensor requires modification of transformer core to mount the sensing coil through hole. An extended Kalman-Filter-approach to estimate the primary transformer current is presented in [25]. However, the study only concerns the discrimination of the transformer inrush current from internal faults to trigger a protective relay.

This article presents a novel digital feedback control loop scheme for eliminating DC-offset \(I_{h,\text{DC}}\) without additional devices in the main power path. In contrast to related work a Luenberger observer is used that only requires the measurements of the transformer voltages to estimate the magnetizing current as process variable for a PI-controller. Main purpose of the study is to develop a digital controller that allows for simple application or flexible modification of existing controller structures only by adding the voltage measurement feature and to improve system reliability. At
the beginning Section 2 explains the inevitable occurrence of DC-offset $I_{h,DC}$ in isolated PSFB converters in detail. In Section 3 conventional methods, like a blocking capacitor or CMC, used to overcome this problem are introduced. Section 4 explains the structure of the proposed digital control scheme. Section 5 describes an approach of acquiring the not measurable voltage-second product in (a) causes a DC-offset $I_{h,DC}$ and its absolute value increases each period $T(n)$ till transformer saturation.

2. Transformer DC-Offset Current

The magnetizing current $I_h$ creates the magnetic field of the transformer required for power transfer from primary to secondary side and flows through the magnetizing inductor $L_h$ of the transformer. $I_h$ adds either to primary $I_p$ or secondary current $I_s$ of transformer depending on power transfer direction. The transformer is not capable to transfer any DC component of $I_h$ [26]. Therefore, the magnetizing current $I_h$ needs to be detected on primary and secondary side when operating in bidirectional direction. The following considerations refer exemplary to a power transfer direction from primary side A with $I_p$ to secondary side B with $I_s = (I_p - I_h)r$.

Figure 2 shows the typical schematic characteristics of applied voltage $V_p$ in Figure 2(a), primary current $I_p$ in Figure 2(b) and magnetization current $I_h$ in Figure 2(c) for transformer of the PSFB converter shown in Figure 1.

The power transfer cycles $PC_1$ and $PC_2$ are determined according to the effective duty cycle $d_{eff} = 2t_{on}/T$ applied to the transformer TR. For desired duty cycle $d_{eff}$ the phase-shift between lagging $S_1 + S_3$ and leading leg $S_2 + S_4$ of $H_A$ varies and applies either the positive or the negative input voltage $V_A$ to the magnetizing inductor $I_h$ of the transformer TR according to Figure 2(a). For positive voltages $V_p > 0$ the magnetizing current $I_h$ increases and for negative voltages $V_p < 0$, $I_h$ decreases respectively. During the interval $t_{off}$ no voltage is applied to $L_h$ resulting in a constant $I_h$. The alternating positive and negative voltage-second products applied to the transformer within one period $T(n)$
determines the peak value \( I_{h,peak,1} \) and \( I_{h,peak,2} \), assuming to be constant during \( T(n) \), in Figure 2(c) according to:

\[
I_{h,peak,1} (n) = \frac{1}{L_h} V_1 t_{on} (n) = \frac{1}{L_h} V_1 d_{\text{eff}} (n) \frac{T}{2} \quad (1a)
\]

\[
I_{h,peak,2} (n) = \frac{1}{L_h} V_2 t_{on} (n) = \frac{1}{L_h} V_2 d_{\text{eff}} (n) \frac{T}{2} \quad (1b)
\]

Therefore, the DC-offset \( I_{h,DC} (n) \) can approximately be defined as the average value of magnetizing current \( I_h (n) \) during one period \( T(n) \) with:

\[
I_{h,DC} (n) = \bar{I}_{\text{h}} (n) = \frac{I_{h,peak,1} (n) + I_{h,peak,2} (n)}{2} \quad (2)
\]

Figure 2(b) explains the sloped magnetizing current \( I_h \) as an additional comparable small portion to the high main primary current \( I_{p,n} \) with \( I_h = I_{p} - I_{p,n} \) [15]. As consequence \( I_h \) cannot be measured directly considering the transformer as a four terminal device. Assuming negligible variations of \( I_{p,n} \) and \( d_{\text{eff}} \) during one switching period \( T(n) \) the DC-offset \( I_{h,DC} (n) \) can be calculated from the difference between the peak values of \( I_p \) during one complete period \( T(n) \) according to:

\[
I_{h,DC} (n) = \frac{I_{p,peak,1} (n) - I_{p,peak,2} (n)}{2} \quad (3)
\]

For a present asymmetrical voltage-second product \( V_1 \neq V_2 \), the absolute value of \( I_{h,DC} (n) \) increases each period \( T(n) \) with

\[
I_{h,DC} (n+1) = I_{h,DC} (n) + \Delta I_{h,DC} \quad (4)
\]

and the resulting DC-offset \( I_{h,DC} \neq 0 \) drives the transformer into saturation. Ideally, the symmetrical structure of the PSFB topology ensures no DC-offset \( I_{h,DC} \) as positive and negative current slope of \( I_h \) balance each other. Unfortunately, various inevitable effects can cause a small asymmetrical voltage-second product on primary side of transformer [14]. One transient reason is the unbalanced duty cycle between power cycle \( PC_1 \) for the lagging leg and \( PC_2 \) for the leading leg due to ZVS-operation or the change of \( d_{\text{eff}} \) for regulating the output current \( I_B \) of the PSFB [12]. Even in steady state operation manufacturing tolerances of devices as well as circuit board layout (PCB) cause different voltage drops and apply different voltages to the magnetizing inductor \( L_h \) and increase \( I_{h,DC} \) [10]. Beside of the transformer the resulting high saturation current threatens the switching devices \( H_1 \) and \( H_2 \) and general system reliability. Additionally, the conduction loss unnecessarily increases with an unbalanced magnetizing current \( I_h \) as it does not contribute to power transfer from primary to secondary side. Therefore, system efficiency \( \eta \) is directly related to minimize \( I_{h,DC} \).

3. Conventional Methods

Several methods have been proposed either to suppress or to control DC-offset \( I_{h,DC} \) in [11, 12, 16, 17, 27, 28].

3.1. Blocking Capacitor. Connecting a blocking capacitor in series to the transformer suppresses any DC component in the transformer current \( I_h \) and allows for symmetrical operation [11]. Therefore, this method is suitable for acquiring only secondary side output current and apply proportional integral current control [27]. Additionally, in phase-shift control scheme the blocking capacitor method reduces the circulating currents during the free-wheel interval of the lagging leg and minimizes conduction loss. However, the timing of the leading leg becomes more complicated [12]. On the other hand, the blocking capacitor method simultaneously increases the conduction loss as the capacitor is located in the main power path with maximum current of the PSFB [28]. For bidirectional operation the magnetizing current \( I_h \) occurs on either side, primary and secondary, according to power direction. Therefore, this method needs two blocking capacitors and significantly increases volume and overall costs.

3.2. Current-Mode Control. The current-mode control (CMC) method is well known for non-isolated converters such as buck or boost converter topologies [16–18]. For non-isolated converters the inductor current \( I_L \) serves as process variable for the control loop. Fast response analog comparators generate PWM signals and directly trigger the power switches when inductor current \( I_L \) reaches a given threshold value and provide output control. Reference [18] distinguishes CMC between peak detect, valley detect and emulated control mode. However, either method suffers from sub-harmonic oscillation when inductor current \( I_L \) does not return to its initial value by the start of next switching cycle in continuous current-mode (CCM). In peak detect CMC the sub-harmonic oscillations occur at duty cycles \( d_{\text{eff}} > 0.5 \) and for valley detect at duty cycles \( d_{\text{eff}} < 0.5 \) respectively. A slope compensation circuit needs to correct the ripple current \( \Delta I_h \) and results in a response delay for the controller [19]. In isolated converters the CMC method can additionally provide the elimination of DC-offset \( I_{h,DC} \) with small modifications. Any DC component of \( I_h \) cannot pass the transformer. Therefore, the input current \( I_h \) must serve as process variable for \( I_{h,DC} \) control with power direction from A to B side.

Figure 3 explains the compensation principle of CMC with the schematic characteristic curves of primary voltage \( V_p \) in Figure 3(a), input current \( I_A \) in Figure 3(b) and magnetizing current \( I_h \) in Figure 3(c) of transformer with a present flux imbalance. Neglecting parasitic inductors, the magnetizing current \( I_h \) slopes up with \( m_1 = V_1/I_h \) during power cycle \( PC_1 \) and with \( m_2 = V_2/I_h \) during power cycle \( PC_2 \) respectively. Parasitic effects reduce the applied voltage \( V_1 \) for power cycle \( PC_1 \) in Figure 3(a). Assuming a symmetrical duty cycle \( d_1 = d_2 = \bar{d}_{\text{eff}} \) for both power cycles the peak of input current \( I_{A,peak,1} \) does not reach the threshold value \( I_{th} \) of the analog comparator and the negative DC-offset \( I_{h,DC} < 0 \) rises [12]. The current-mode controller directly influences the duty cycle \( d_{\text{CMC}} = 2 \Delta I_{h,\text{CMC}}/T \) and extends \( t_{on,\text{CMC}} > t_{on} \) until \( I_{A,peak,1} \) reaches the threshold current \( I_{th} \). Therefore, the CMC ensures flux balance of the
transformer with $I_{h,DC} = 0$ and provides a symmetrical voltage-second product $V_1 t_{on,CMC} = V_2 t_{on}$ [12, 18].

Common control schemes prefer an analog controller with comparators compared to digital controller as the erratic increasing of $I_A$ needs high bandwidth to detect the peak current $I_{A,peak}$ exactly. In consequence, the CMC is hard to combine with a flexible digital control scheme [20]. An additional challenge for the measured process variable $I_A$ is the accuracy to properly detect the small portion of $I_h$ as Figure 2(b) explains $I_h$ as a comparable small additional sloped offset to the high input current $I_A$ or rather $I_{p,n}$.

4. Proposed Control Scheme ctrIh

The proposed control scheme ctrIh uses flexible digital design. It requires neither lossy and expensive additional passive devices in the main power path of the PSFB nor analog comparator for triggering power switches. Similar to CMC, the proposed digital control scheme ctrIh directly interventions the duty cycle $d_2$ of the leading leg to equalize the peak values of primary current $|I_{p,peak,1}| = |I_{p,peak,2}|$ and eliminates any magnetizing DC-offset $I_{h,DC}$ within a few switching cycles.

Figure 4 depicts the control loop of the PSFB converter with ctrIL for output current regulation and ctrIh for DC-offset $I_{h,DC}$ control. The output controller ctrIL is built up with a PI-controller $P_{IL}(z)$ and feedback control for the output voltage $V_B = V_{B} + AV_B$ for better adjustment of output operating point. The transfer function of the PSFB $G_{duty}$ determines the effective duty cycle $d_{eff} = d_1 = V_B/V_A r$ for desired output current $I_{B,ref}$. The proposed magnetization controller ctrIh adds or subtracts a small offset $\Delta d$ between duty cycle $d_1$ for the lagging and duty cycle $d_2 = d_1 + \Delta d$ for the leading leg until $I_{h,DC} = 0$. As $I_{p,n}$ and $d_1$ vary only a little during one switching period and also the impact of parasitic effects contribute only to a small amount, the ctrIh limits the magnitude of $\Delta d$ to a few percentage of $d_1$. Therefore, the influence of ctrIh on the output controller ctrIL is negligible [14]. According to $d_1$ and $d_2$ the phase-shift calculator provides eight separate PWM signals for power switches $S_1$–$S_8$ of PSFB power electronics. In the feedback path the main control unit (MCU) performs the acquisition of process variables $V_A$, $V_B$, $I_1$ and $I_h$ with analog digital converters (ADC). For output control the ripple inductor current $I_L$ averages to the desired output current $I_B$. Therefore, the moving average finite impulse response (FIR)
filter with $G_W(z)$ averages $I_L$ over $N$ number of periods with:

$$I_B(n) = \frac{1}{N} \sum_{i=0}^{N-1} I_L(n-i).$$  \hspace{1cm} (5)$$

Accordingly, the DC-offset $I_{h,DC}$ is defined as the average of rippled magnetization current $I_h$ over several periods.

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unbalanced flux leading to $I_{h,DC} \neq 0$. Therefore, these effects have to be considered in the transfer function $G_{h,DC}(z)$.

Figure 5 illustrates the influence of $\Delta d$ on the DC-offset $\Delta I_{h,DC}$ for analyzing the impact of unbalanced duty cycle due to output regulation $\text{ctr}_{IL}$ over two consecutive switching periods $T(n)$ and $T(n+1)$. According to phase shifted PWM signals for power switches $S_1$–$S_4$ in Figure 5(a) the voltage $V_p$ in Figure 5(b) applies to the transformer. Figure 5(c) shows the time course of magnetizing current $I_h$ resulting from unbalanced duty cycles $d_1(n) < d_1(n+1)$.

The PWM signals of power switches $S_1$–$S_4$ obtain the duty cycle difference $\Delta d$ to:

$$\Delta d = d_1(n+1) - d_1(n) = \frac{t_{\text{on}}(n+1) - t_{\text{on}}(n)}{T/2}. \quad (6)$$

Assuming equal voltages $V_1 = V_2 = V_A$ for power cycles $PC_1(n)$ and $PC_2(n)$ the magnetizing current $I_h$'s slopes with the same gradient $(V_1/I_h) = (V_2/I_h) = (V_A/I_h)$. In period $T(n+1)$, the output regulator $\text{ctr}_{IL}$ increases $d_1(n+1) = d_1(n) + \Delta d$. Due to the increased duty cycle $d_1(n+1)$ the magnetizing current $I_h$ surpasses the peak value $I_{h,\text{peak}}(n)$ before the end of power cycle $PC_1(n+1)$, thus, leading to an increased DC-offset in the following period $T(n+1)$ according to:

$$\Delta I_{h,DC}\text{duty} = I_{h,DC}(n+1) - I_{h,DC}(n). \quad (7)$$

This behavior expresses also for small differences between duty cycle $d_1(n)$ and $d_2(n)$ in the same period $T(n)$.

Furthermore, $I_{h,DC}(n)$ defines as the average of magnetizing current $I_h(n)$ with a linear slope $V_A/I_h$. Therefore, $\Delta I_{h,DC}\text{duty}$ between two consecutive switching periods due to unbalanced duty cycle approximately calculates as follows:

$$
\Delta I_{h,DC}\text{duty} \approx \frac{1}{2} \frac{V_A}{I_h} \Delta d \frac{T}{2} = \frac{V_A T}{4L_h} \Delta d.
$$

As long as there is a duty cycle imbalance $\Delta d$ present, the magnetizing current $I_{h,DC}$ needs to diverge until the PSFB gets damaged if not compensated.

Another reason causing $I_{h,DC}$ are different parasitic resistances such as switch resistances or general PCB-design resulting in unequal voltages $V_1 \neq V_2$. [14] derives the influence of parasitic elements $\Delta R$ and can approximately be calculated as follows:

$$\Delta I_{h,DC}\text{res} = \frac{\Delta R T}{2L_h} I_{h,DC}(n) = K_{\text{res}} I_{h,DC}(n). \quad (9)$$

As the impact of resistance is comparable low during transition times, (9) only considers the dominant time intervals of free-wheeling and power cycle. Regarding (9) parasitic resistances $\Delta R$ contribute proportionally with $K_{\text{res}}$ to magnitude of $I_{h,DC}$. Therefore, the impact of $\Delta I_{h,DC}\text{res}$ represents an steady-state fault of $\text{ctr}_{h}$. [14]

During the transition between power $PC_a$ and free-wheeling $FC_a$ cycles mainly the switching characteristics of the power switches and dead time control influence the DC-offset $I_{h,DC}$. The dead time control determines different duty cycle $d_1$ for the lagging and $d_2$ for the leading leg according to ZVS-conditions. In [14], the detailed derivation results in:

$$\Delta I_{h,DC,\text{trans}} = K_{\text{trans}} I_{h,DC}(n). \quad (10)$$

again, with propotional gain $K_{\text{trans}}$.

For a unified transfer function $G_{h,DC}(z)$ the three main effects influencing $I_{h,DC}$ must be considered according to [14] with:

$$\Delta I_{h,DC} = \Delta I_{h,DC,\text{duty}} - \Delta I_{h,DC,\text{res}} - \Delta I_{h,DC,\text{trans}}. \quad (11)$$

The effects of resistance $\Delta I_{h,DC,\text{res}}$ and switch transition $\Delta I_{h,DC,\text{trans}}$ both prevent the development of $I_{h,DC}$ in either direction and therefore have to be subtracted while the transient $\Delta I_{h,DC,\text{duty}}$ supports the development of $I_{h,DC}$ [14].

Equation (12) calculates the transfer function of $\Delta d$ and $I_{h,DC}$ in $z$ domain as follows:

$$G_{h,DC}(z) = \frac{I_{h,DC}(z)}{\Delta d(z)} = \frac{V_A T}{4L_h} \frac{1}{z - K_{\text{trans}} - K_{\text{res}}}. \quad (12)$$

According to (12) the magnetization controller $\text{ctr}_{h}$ requires an integrator for steady state stability. For the experimental results the $\text{ctr}_{h}$ is implemented as PI-controller $PI_{h}$.  

5. Derivation of Observer Model

In the proposed digital control scheme of $\text{ctr}_{h}$, $I_{h,DC} = I_h$ represents the process variable. Considering TR as a four terminal device with $I_p$ and $V_p$ as inputs and $I_s$ and $V_s$ as outputs, $I_{h,DC}$ cannot be measured at clamps directly. Therefore, $I_{h,DC}$ needs to be calculated from a related value depending on physically measurable in- and outputs.

In technical systems, a state observer is able to estimate an internal state value, such as $I_{h,DC}$ from measurable in- and outputs of the real system [29, 30]. Figure 6 simplifies a part of the PSFB converter with four terminal devices of full-bridge $H_A$ and transformer TR. The measurable primary voltage $V_p$ and secondary voltage $V_s$ of transformer serve as in- and output for the observer model and allow for estimating the internal value of $I_{h,DC}$.
The state observer method is typically computer-implemented and therefore suitable for the proposed digital controller $\text{ctr}_{Ih}$.

The measuring method $M_{i}$ for the observer model requires only the average value of transformer voltages $V_p$ and $V_s$ and can be measured with operational amplifiers at naturally high bandwidth $BW > 1$ MHz. Additionally, the voltage measurement only needs small signal currents and therefore it is comparable lossless. The measuring method $M_{ref}$ serves as reference to compare with conventional methods with an inductor $L_{h,ref}$ connected in parallel to transformer. According to:

$$ I_h = \frac{L_{h,ref}}{L_h} I_{h,ref}, \quad (13) $$

the reference magnetization current $I_{h,ref}$ reflects the real magnetization current $I_h$ and can be measured with a current clamp.

The proposed digital control scheme $\text{ctr}_{Ih}$ uses the well-known Luenberger observer for estimation of $I_{h,DC}$ [30]. For the design of a Luenberger observer, in Section 5.1 first a stable linear state space model of the real transformer must be developed. Although the observer allows for controlling a time continuous system, a discretization is necessary for realistic applications with a digital control as described in Section 5.2. Section 5.3 examines the observability criteria according to Kalman of derived transformer model and Section 5.4 explains the design of the Luenberger gain parameters for observer model.

5.1. Transformer State Space Model. The Luenberger observer requires a linear state space model of transformer in form of:

$$ \dot{x} = Ax + Bu, \quad (14a) $$

$$ y =Cx + Du. \quad (14b) $$

In (14a) and (14b) $x$ represents the state variables with input current $i_s$, output current $i_h$, and magnetization current $i_{h}$. The primary voltage $V_p$ represents the system input $u$ and the secondary voltage $V_s$ serves as measurable system output $y$ according to:

$$ x = [i_p \quad i_s \quad i_h]^T, \quad (15a) $$

$$ u = V_p, \quad (15b) $$

$$ y = V_s. \quad (15c) $$

Figure 7 shows the equivalent circuit diagram (ECD) of a real transformer. Parasitic resistances $R_p$ and $R_s,u$ and leakage inductors $L_p$ and $L_{s,u}$ depict the loss component of primary and secondary windings while $R_{Fe}$ simulates the magnetic resistance of the transformer core. The core magnetization is regarded with magnetizing inductor $L_h$.

Nonlinear behavior of magnetizing inductor $L_h$ is neglected as the controller $\text{ctr}_{Ih}$ prevents saturation of transformer. The ratio $r$ of primary winding $N_p$ to secondary winding $N_s$ represents the ideal transformer TR with:

$$ r = \frac{N_p}{N_s}. \quad (16) $$

In this model the parasitic capacitors are neglected as their time constants are very small compared to switching frequency and time constant of inductors. $Z_0$ represents the impedance of the load for gathering secondary current $i_s$.

The reference-arrow system in Figure 7 defines the sign of variables for further derivations. Secondary side values with indexes $s,u$ are referred to primary side with transformer ratio $r$ according to:

$$ v_{su} = r v_s, \quad i_{su} = \frac{1}{r} i_s, \quad i_{s,u} = \frac{1}{r} i_s, \quad r_{su} = (r_s + Z_0) r^2, \quad L_{su} = L_r r^2. \quad (17) $$

The Kirchhoff current (KCL) and voltage (KVL) laws obtain the main equations of the transformer model according to:

$$ V_p = R_p i_p + L_p \dot{i}_p + L_{h,i} \dot{i}_h, \quad (18a) $$

$$ V_s = -\frac{R_{su} i_s}{r^2} + \frac{L_{s,u} i_s}{r^2} - \frac{L_h i_h}{r} \quad (18b) $$

$$ V_s = -\frac{R_{su} i_s}{r^2} + \frac{L_{s,u} i_s}{r^2} - \frac{L_p i_p}{r^2} - \frac{R_{Fe} i_p}{r^2} + \frac{1}{r} V_s, \quad \text{with } V_s = Z_0 i_s \quad (18c) $$

$$ i_p = \frac{1}{r} i_s + i_{Fe} + i_h, \quad \text{with } i_{Fe} = \frac{L_h}{R_{Fe}} i_h. \quad (18d) $$

Reshaping (18a)–(18d) delivers the time continuous state space vector $x$ to:
Table 1: Transformer parameter for simulation and PSFB prototype.

| Parameter                     | variable | value  |
|-------------------------------|----------|--------|
| Sampling time                 | $t_s$    | 2 [μs] |
| Transformer ratio             | $r$      | 2 [-]  |
| Magnetization inductor        | $L_h$    | 5000 [μH] |
| Leakage inductor              | $L_{p}, L_s$ | 6.23 [μH] |
| Magnetic resistance           | $R_{Fe}$ | 1000 [Ω] |
| Parasitic resistance (prim)   | $R_p$    | 0.0045 [Ω] |
| Parasitic resistance (sec)    | $R_s$    | 0.0070 [Ω] |
| Reference measuring inductor  | $L_{h,ref}$ | 630 [μH] |

\[
i_p = \frac{1}{L_p} V_p - \frac{R_{Fe} + R_p}{L_p} i_p + \frac{R_{Fe} L_p}{L_p} r s + \frac{R_{Fe}}{L_p} i_h, \tag{19a}
\]
\[
i_s = \frac{R_{Fe}}{L_{s,m}} r_i p - \frac{R_{s,m} + R_{Fe}}{L_{s,m}} i_s - \frac{R_{Fe}}{L_{s,m}} r_i h, \tag{19b}
\]
\[
i_h = \frac{R_{Fe}}{L_h} r_i p - \frac{R_{Fe}}{L_h} i_s - \frac{R_{Fe}}{L_h} i_h, \tag{19c}
\]

Referred to (14a) the system $A$ and input $B$ matrices can be obtained to:

\[
A = \begin{bmatrix}
R_{Fe} + R_p & \frac{R_{Fe}}{L_p} r & \frac{R_{Fe}}{L_p} \\
\frac{R_{Fe}}{L_{s,m}} r & -\frac{R_{s,m} + R_{Fe}}{L_{s,m}} & -\frac{R_{Fe}}{L_{s,m}} r_i h \\
\frac{R_{Fe}}{L_h} r & -\frac{R_{Fe}}{L_h} & -\frac{R_{Fe}}{L_h}
\end{bmatrix}, \tag{20a}
\]

\[
B = \begin{bmatrix} 1 \\ 0 \\ 0 \end{bmatrix}^T. \tag{20b}
\]

The terms of voltage drop $(L_p/r)i_p$ and $(L_{s,m}/r^2)i_{s,m}$ in (18c) depend on the dynamic change of input and output current due to leakage inductors. However, the small time constant of transient process of leakage inductors in μH-range is not detected due to FIR filtering of signal with $G_{ih}$. Neglecting the dynamic terms in (18c) provides the equation for system output $V_s$ to:

\[
V_s = \frac{1}{r} V_p - \frac{R_{Fe} L_p}{L_p} r s - \frac{R_{Fe}}{L_p} s_i p, \tag{21}
\]

with output matrix $C$ and direct feed-through matrix $D$ from (14b) according to:

\[
C = \begin{bmatrix} \frac{R_p}{r} & \frac{R_{s,m}}{r} & 0 \end{bmatrix}, \tag{22a}
\]

\[
D = \frac{1}{r}. \tag{22b}
\]

Table 2: Poles of system matrix $A$ for the three state variables $i_p$, $i_s$ and $i_h$ of transformer model.

| state variable | poles           |
|---------------|-----------------|
| $i_p$         | $-6.4228 \times 10^5$ |
| $i_s$         | $-2.0918 \times 10^4$ |
| $i_h$         | $-3.1444$       |

Table 3: Poles of reduced system matrix $A_{red}$ for the two state variables $i_s$ and $i_h$ of transformer model.

| state variable | poles           |
|---------------|-----------------|
| $i_s$         | $-2.0916 \times 10^4$ |
| $i_h$         | $-2.4645$       |

5.2. Analysis and Discretization of the Transformer Model.

The transformer model includes three state variables $i_p$, $i_s$ and $i_h$ building a third order system with rank ($A$) = 3.

From the derived transformer model in the previous Section 5.1 the eigenvalues or poles of the system matrix $A$ describe the dynamic behavior of the system.

The values in Table 1 refer to the measured values from PSFB prototype or from the corresponding data from manufacturer. Analyzing the system with transformer parameter according to Table 1 delivers the poles for all three state variables $i_p$, $i_s$ and $i_h$ shown in Table 2. As all poles are in the left-hand plane (LHP), the system provides stability [31].

The pole of $i_p$ according to Table 2 is too fast for realistic sampling time $t_s = (T/5) = 2μs$ required for discretization of observer model. As the desired value of the observer only includes the state variable $i_h$, the state variable $i_p$ with the fastest pole can be ignored. The reduction of system order eliminates the first state variable $i_p$ and reconfigures the matrices $A_{red}$, $B_{red}$ and $C_{red}$ with rank ($A_{red}$) = 2 and poles according to Table 3.

Again, all poles of the system with reduced order are located in LHP with negative values. Therefore, also the reduced system is stable.

For realistic application of the observer, the reduced system matrices must be available in a discrete time form with sampling time $t_s$. The transfer of the reduced transformer model into discrete form affects only the system matrix $A_{red}$ and the input matrix $B_{red}$ [32]. One approach of discretization presented in [20] according to:

\[
A_{d,red} = e^{A_{red} t_s}, \tag{23a}
\]

\[
A_{d,red} = L^{-1} \{ (sI - A_{red})^{-1} \}, \tag{23b}
\]

delivers the discrete system matrix $A_{d,red}$ without approximation.

The MATLAB order of matrix exponential function $exp(A_{red} t_s)$ can directly perform the discretization and is suitable for the experimental approach.

With reduced discrete system matrix $A_{d,red}$ the reduced discrete input matrix $B_{d,red}$ calculates as follows:
Table 4: Poles of reduced discrete system matrix $A_{d,\text{red}}$ for the two state variables $i_s$ and $i_h$ of transformer model in $Z$-domain.

| state variable | poles       |
|----------------|-------------|
| $i_s$          | 0.9590      |
| $i_h$          | 1.0000      |

\[
B_{d,\text{red}} = \left( \int_0^t e^{A_{\text{red}}t} \, dt \right) B_{\text{red}}, \quad (24a)
\]

\[
B_{d,\text{red}} = A_{\text{red}}^{-1}(A_{\text{d,red}} - I)B_{\text{red}}. \quad (24b)
\]

The reduced discrete output matrix $C_{d,\text{red}} = C_{\text{red}}$ and reduced discrete direct feed-through matrix $D_{d,\text{red}} = D_{\text{red}}$ are not affected and remain the same.

The analysis of the reduced discrete transformer model in $Z$-domain delivers the poles of $A_{d,\text{red}}$ according to Table 4 with one stable pole at $|z| < 1$ and one critically stable pole at $|z| = 1$. Therefore, a suitable design of the observer gain factors must guarantee the stability of the system.

5.3. Observability of the Transformer Model. To use the reduced discrete transformer model and realizing a Luenberger observer the reduced system must be observable. For a completely observable system the initial state $x(t = 0) = x_0$ must be reconstructible form known input $u$ and output value $y$ within a finite time interval [31]. According to Kalman criteria of observability, the observability is proved if the observability matrix $S_{B_{d,\text{red}}}$ in (25) has the same rank $(S_{B_{d,\text{red}}})$ as the reduced system matrix $A_{d,\text{red}}$ with

\[
S_{B_{d,\text{red}}} = \begin{pmatrix} C_{d,\text{red}} & C_{d,\text{red}}A_{d,\text{red}} \end{pmatrix} = \begin{pmatrix} 94.2147 & 5.1102 \\ 90.3548 & 4.9009 \end{pmatrix}. \quad (25)
\]

According to (25) the rank of observability matrix $\text{rank}(S_{B_{d,\text{red}}}) = \text{rank}(A_{d,\text{red}}) = 2$ equals the rank of reduced discrete system matrix and so the system is completely observable. Additionally, its determinant $\det(S_{B_{d,\text{red}}}) = -7.7102 \times 10^4 \neq 0$ is not zero [30, 33].

5.4. Design of Luenberger Observer. In the structure of a Luenberger observer the derived transformer model works in parallel to real transformer. The observer compares the measured output $y$ with the calculated value of the transformer model $\hat{y}$ and feeds back the estimation error $\Delta y_{\text{error}} = L(y - \hat{y})$ with feedback matrix $L$ to the model [30].

From observer model the estimated state vector $\hat{x}$ calculates as follows:

\[
\dot{\hat{x}} = A_{d,\text{red}}\hat{x} + B_{d,\text{red}}u + \Delta y_{\text{error}}, \quad (26a)
\]

\[
\dot{\hat{x}} = A_{d,\text{red}}\hat{x} + B_{d,\text{red}}u + L(y - C_{d,\text{red}}\hat{x} - D_{d,\text{red}}u), \quad (26b)
\]

\[
\dot{\hat{x}} = (A_{d,\text{red}} - LC_{d,\text{red}})\hat{x} + (B_{d,\text{red}} - LD_{d,\text{red}})u + Ly. \quad (26c)
\]

Suitable values of feedback matrix $L$ adjust the dynamical behavior of the model to match the real transformer and react to disturbances or transient variations of duty cycle $d_1$. The design of feedback matrix $L$ with pole placement according to Ackermann [31, 33] is used and the known poles of $A_{d,\text{red}}$ from Table 4 deliver the first approach according to:

\[
L = [0.00135 - 0.02478]T. \quad (27)
\]

6. Results

The proposed digital control scheme regulates the magnetizing current $I_h$ of a PSFB to prevent transformer saturation. First approaches concern the simulation of electronics to prove principal functionality of $\text{ctrI}_h$, supported through experimental tests on a PSFB prototype with $P = 1kW$. Table 5 lists the specifications of analyzed operating point of PSFB prototype electronics, also used for simulation.

To validate the simulated results in Section 6.1 the following Sections 6.2 and 6.3 explain the experimental measuring setup up and show the achieved results in operating point according to Table 5.

6.1. Simulation of Observer. For the evaluation of proposed $\text{ctrI}_h$ in Section 4 and derived observer model in Section 5,
the PSFB topology from Figure 1 is built up in a MATLAB/ SIMULINK simulation model according to parameters in Table 1 and operating point listed in Table 5. The control scheme for $c_{trh}$ and observer model are implemented in MATLAB/SIMULINK.

The simulation results in Figure 8 demonstrate the behavior of the PSFB magnetizing current $I_h$ at operating point in Table 5 for the different disturbances mentioned in Section 2, dynamic change of output current $I_d$ through duty cycle $d_t$ and initial static differences due to manufacturing process of devices. Fluctuations of $d_t$ due to $c_{trh}$ thereby influence the magnetization current $I_h$ in a similar way as $d_t$. The simulation starts at $t_0$ with a desired output current $I_R = 9.3A$ and a corresponding duty cycle $d_1 (t_0) = 0.76$. The series resistance $R_{S2} = R_{S1} + \Delta R = 0.2\Omega$ of switch $S_2$ is doubled compared to other switches with $R_{S1,3,4} = 0.1\Omega$, simulating an initial voltage-second product imbalance. Figure 8(a) shows the simulated average values for magnetizing current $I_h$ and estimated observer current $I_{h,obs}$. Without $c_{trh}$, the magnetizing current $I_{h,0}$ steadily raises due to the unbalanced voltage-second product till saturation of transformer. In Figure 8(a) the observer current $I_{h,obs}$ reflects the dynamic behavior of simulated current $I_h$ with the model error $\Delta I_{h,\text{error}} = I_h - I_{h,obs}$ in Figure 8(b) fluctuating around zero with a DC-offset $\Delta I_{h,\text{error,DC}} = 11mA$. The influence of $c_{trh}$ on controlled variable $\Delta d = |d_1 - d_2|$ ≈ 0.07 is shown in Figure 8(c).

The increased resistance $R_{S2}$ forces $c_{trh}$ to increase $d_2 > d_1$ and compensates the voltage-second product imbalance. After initial transition process the $c_{trh}$ is able to correct the error resulting from static difference due to $R_{S2}$ within $t_1 \approx 7.7ms$. Either change due to $d_1$ step at $t_3$ or $d_2$ control at $t_4$ and $t_2$ causes fluctuations in $I_h$. The estimated observer current $I_{h,obs}$ follows the dynamics of $I_h$ with a small delay $\Delta t_{\text{delay}} = 0.5ms$ mainly resulting from computation and FIR filter. According to $I_{h,obs}$ the $c_{trh}$ adjusts $d_2$ for regulating $I_{h,\text{DC}} = 0A$. Figure 9 explains the dynamic changes exemplary on $d_1$ in detail. At simulation time $t_3 = 50ms$ the desired output current raises to $I_R = 14.1A$ resulting in a dynamic change of duty cycle $d_1 (t_3) = 0.8$. For $t < t_3$ the magnetizing current $I_{h,0}$ approximately slopes up with $\frac{d}{dt}I_h = 3.2As^{-1}$. For time $t_3 \leq t < t_3 + 50\mu s$ the magnetizing current $I_{h,0}$ steps up according to equation (8) and (9), raising $I_{h,\text{DC}}$. The step expresses as slope with $\frac{d}{dt}I_h = 4.7mA (50\mu s^{-1})$ due to FIR filter $G_{trh}$ in MCU. With increased magnetizing current $I_{h,0}$ the rising slope of magnetizing current $\frac{d}{dt}I_h = 6.6As^{-1}$ due to $R_{S2}$ enlarges after $t_3 + 50\mu s$ encouraging saturation of transformer. The controlled estimated observer current $I_{h,obs}$ matches the dynamics of magnetizing current $I_h$ and the change can be controlled out by $c_{trh}$ with $I_{h,\text{DC}} = 0A$. According to applied to voltage $V_p$ to $L_h$ and moving average filter, the magnitude value results to $I_{h,pp} = I_{h,obs,pp} = 0.2A$.

For analyzing the resistance of $c_{trh}$, $I_{h,obs,\text{fail}}$ exemplary represents the observer current with a model failure in magnetizing inductance $L_{h,\text{fail}} = 1.1L_p$. The model failure produces a time delay $\Delta t_{\text{fail}} = 1ms$ for $I_{h,\text{fail}}$ as the bigger magnetizing inductance reduces the slope of magnetizing current and enlarges the initial time $t_1$ for regulating the initial differences. Nevertheless, the $c_{trh}$ can handle the change in $d_1$ as well as in $d_2$. A similar result is achieved with an examined model failure in primary resistance $R_p$ and load impedance $Z_0$.

6.2. Measurement Application. In the proposed control scheme the $c_{trh}$ only needs the average values of voltage $V_p$ and $V_s$ as measuring inputs to control the magnetization DC-offset $I_{h,\text{DC}}$. Typically, voltage measurements only need a network of low-cost resistors and operational amplifiers with naturally high bandwidth and low self-consumption. Therefore, voltage measurements reach higher efficiency than current measurements with high loss due to shunt resistors or hall sensors with limited bandwidth.

Figure 10 shows the measuring application for acquiring the input $V_p$ and output $V_s$ for the observer model. A simple voltage divider sets the transformer voltages to required 0 . . . 5V level of the main control unit MCU with $V_p'|$ and $V_s'$. Depending on the switching state of full-bridge $H_k$ the potential of the transformer voltage $V_p$ is floating. As in single ended mode the MCU refers its input signals to $GND_A$, an operational amplifier $OP_p$ performs the differential measurement $V_p'| = V_p'|_1 - V_p'|_2$ and provides stable and resilient signals for further processing in MCU.
Additionally, the differential measurement does not require a floating supply voltage $V_{CC}$ for amplifiers. To maintain isolation of the PSFB the voltage measurement on secondary side $V_s'$ needs to be decoupled with an isolated amplifier $OPs'$ after differential measurement.

Fhe observer model only needs the average of voltage signals $V_p$ and $V_s$. Therefore, the low pass filters $LP_p'$ and $LP_s'$ smooth the signals $V_p'$ and $V_s'$ and minimize the measuring noise due to rising edges with high $dV/dt$. In the MCU, the digital FIR filters $GV_p'(z)$ and $GV_s'(z)$ calculate the moving average over 1000 samples for averaging the signals. According to the observer model the MCU estimates $I_{h,DC}$ according to observer model for ctrIh.

### 6.3. Experimental Results.

Figure 11 shows the PSFB prototype used for the validation of the proposed control scheme. The modular structure allows for flexible testing of different combination of power device.

In the experimental test setup a National Instrument (NI) Data Acquisition (DAQ) system acquires the values from the measuring circuit board with 500MS/s and performs the calculations of the MCU. Therefore, the NI system includes the observer model, ctrIh control scheme with PI-controller and phase-shift calculator. The resulting eight PWM signals are generated on the counter outputs of NI system and directly connected to the drivers for full-bridge switches $S_1$–$S_8$.

For validating the observer model and simulation the effects of initial resistance differences and variable operating points are observed. Due to layout and manufacturing tolerances of devices, initial differences are not avoidable and...
present in prototype electronics. As the effect of resistance \( \Delta R \) in real prototype is much smaller than in simulation, the rising slope of magnetizing current is smaller in experiment and allows for reacting before the prototype gets damaged.

First, Figure 12 compares the measured primary currents of the PSFB prototype in a steady state operating point according to Table 5. Figure 12(a) shows the primary current \( I_{p,0} \) without any control of magnetization current while Figure 12(b) illustrates \( I_{p,obs} \) controlled by \( \text{ctr}_{\text{ih}} \) with observer model. Figure 12(c) explains the according courses of \( I_{p,0} \) and \( I_{p,obs} \) at higher temporal resolution. The initial differences cause an unbalanced voltage-second product and lead to an asymmetrical course of \( I_{p,0} \) with a negative average offset value of \( T_{p,0} \approx -848 \text{mA} \). Without compensation this imbalance drives the transformer into saturation for continuous operating time. The \( \text{ctr}_{\text{ih}} \) detects the initial imbalance and is able to regulate the average offset to a constant value of \( T_{p,obs} \approx -42 \text{mA} \). Thus, the \( \text{ctr}_{\text{ih}} \) removes the offset difference \( \Delta T = T_{h,0} - T_{p,obs} \approx -806 \text{mA} \) and improves the system reliability and efficiency.

Figure 13 illustrates the results of the PSFB prototype controlled with \( \text{ctr}_{\text{ih}} \) switching between different operating points. Therefore, during the experiment duty cycle \( d_i \) varies each time \( t_0 - t_5 \) resulting in changed output current \( I_h \) shown in Figure 13(d).

In Figure 13(a) the magnetizing current \( I_h = 0.126 I_{h,\text{ref}} \) is calculated from measurable reference current \( I_{h,\text{ref}} \) with parallel inductor \( L_{\text{ref}} \) according to (13). Although the magnitude of observer current \( I_{h,obs} \) is scaled due to feedback matrix \( L \), the dynamical behavior reflects the magnetizing current \( I_h \) and keeps the model error \( \Delta I_{h,\text{error}} = I_h - I_{h,obs} \) in Figure 13(b) around zero. Equal to theory and simulation in Figure 13(c) the \( \text{ctr}_{\text{ih}} \) adjusts \( d_1 \) with a constant offset \( \Delta d = 0.1 \) due to initial resistance differences and controls variations of \( d_1 \). The exact measurement of the initial differences in the full-bridges of the prototype is very difficult and also vary due to temperature effects. Therefore, there is a difference between simulated duty offset \( \Delta d = 0.07 \) and prototype offset. However, the difference does not affect the functionality of the \( \text{ctr}_{\text{ih}} \). The magnetizing current in Figure 13(a) oscillates around its DC-offset \( I_{h,DC} = -30 \text{mA} \) while the \( \text{ctr}_{\text{ih}} \) shows a better performance on \( I_{h,obs,DC} = 1 \text{mA} \). This difference results from measuring voltage offset due to operational amplifier and is the main drawback of the control scheme. Therefore, the measuring offset must be minimized by using offset compensated operational amplifiers. Compared to simulation the magnitude of magnetizing current is reduced. In real prototype there are also parasitic resistances in the PCB-Layout and in the supply lines from the source. As consequence the given input voltage \( V_p \) drops till the transformer with \( V_p < V_A \) and decreases the slope of magnetizing current \( I_h \). The magnitude of observer current \( I_{h,obs} \) and magnetizing current \( I_h \) vary each time \( t_0 - t_5 \) the operating point changes, as the controlling parameter for PI-controller \( P_{\text{il}}(z) \) depend on \( I_h \) and were not modified during the experiment. Nevertheless, the effect on average values of \( T_h \) or observer current \( T_{h,obs} \) is negligible and saturation can be prevented. Therefore, the experimental results confirm the results of simulation and prove the functionality of \( \text{ctr}_{\text{ih}} \) for observed disturbances.

7. Conclusion

This paper presents a digital control scheme \( \text{ctr}_{\text{ih}} \) for preventing saturation of the transformer due to flux...
imbances in isolated PSFB converters. As the magnetizing current $I_{h}$ of a transformer, which is responsible for the saturation problem, cannot be measured directly, the proposed $\text{ctr}_{Ih}$ uses a Luenberger observer for estimating the magnetizing current. Conventional methods overcome this problem either with increased costs and system size due to additional passive devices or need for lossy and bandwidth limited current measuring. As measuring inputs, the proposed $\text{ctr}_{Ih}$ only needs the average values of transformer voltages $V_p$ and $V_s$ that can be measured with simple operational amplifiers with low self-consumption and naturally high bandwidth. The digital design of $\text{ctr}_{Ih}$ allows for flexible implementation in the main control unit MCU of PSFB converters. Therefore, the MCU includes the $\text{ctr}_{Ih}$ with Luenberger observer model and PI-controller, the output regulator $\text{ctr}_{IL}$ as well as a phase-shift calculator and directly interventions the PWM signals for all power switches $S_1\text{–}S_8$ via driver.

The paper identifies two inevitable effects as main reasons for increased magnetizing DC-offset current $I_{h,DC}$. First, initial resistance differences due to manufacturing tolerances of devices and second dynamic changes of duty cycle for regulating the output power. It is shown that without any control both effects would lead to transformer saturation and threaten system reliability. The proposed $\text{ctr}_{Ih}$ with Luenberger observer model is evaluated by simulation and experimental results on a PSFB prototype with power $P = 1\text{ kW}$ for the identified effects. The $\text{ctr}_{Ih}$ is able to compensate initial resistance differences within 7.7ms by independently controlling the duty cycles of leading and lagging leg of the full-bridge. Also dynamic changes in duty cycle can be controlled out and guarantee a stable continuous operation with a magnetizing DC-offset $|I_{h,DC}| = 30\text{ mA}$ at $V_A = 200\text{V}$. The differences between simulation and experiment can be explained through the additional voltage drop in the supply lines and PCB-layout that result in a smaller magnitude of $I_{h}$ for the real prototype. The main drawback of the proposed scheme can be eliminated by using offset compensated operational amplifiers.

### Data Availability

The datasets supporting the research are available from the corresponding author upon reasonable request.

### Conflicts of Interest

The authors declare no conflicts of interest.

### Acknowledgments

The authors would like to thank Hochschule Karlsruhe-University of Applied science and ads-tec Energy GmbH for facilitating this research.

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