Inductor Saturation Compensation in Three-Phase Three-Wire Voltage-Source Converters Via Inverse System Dynamics

Ziya Özkan, Member, IEEE, and Ahmet M. Hava, Senior Member, IEEE

Abstract—In three-phase three-wire (3P3W) voltage-source converter (VSC) systems, utilization of filter inductors with deep saturation characteristics is often advantageous due to the improved size, cost, and efficiency. However, with the use of conventional synchronous frame current control methods, the inductor saturation results in significant dynamic performance loss and poor steady-state current waveform quality. This article proposes an inverse dynamic model-based compensation (IDMBC) method to overcome these performance issues. For this purpose, two-phase exact modeling of the 3P3W VSC control system is obtained. Based on the modeling, the inverse system dynamic model of the nonlinear system is obtained and employed such that the nonlinear plant is converted to a virtual linear inductor system for linear current regulators to perform satisfactorily. Further, to control phase currents in the synchronous frame, a two-phase coordinate transformation is proposed. The IDMBC method is tested via dynamic command response and waveform quality simulations and experiments that employ saturable inductors reaching down from full inductance at zero current to 1/9th inductance at full current. The results obtained demonstrate the suitability of the method for 3P3W VSCs employing saturable inductors.

Index Terms—Bandwidth, current regulator, dynamic response, harmonic distortion, inductor saturation, three-phase, three-wire, voltage-source converter, waveform quality.

I. INTRODUCTION

VOLTAGE-SOURCE converters (VSCs) are widely utilized in ac–dc and dc–ac power conversion applications of grid-connected energy systems such as active filter, regenerative drive, renewable, static, and uninterruptible power supply (UPS). Typically, in power ratings, less than several kW single-phase converters are used. In higher power ratings, three-phase four-wire converters or three-phase three-wire (3P3W) converters are employed. In all cases, two-level converters emphasize low-cost and high-switching frequency while three-level topologies are favorable in high-efficiency applications [1]. Pulsewidth modulation (PWM) mode of operation is common, and designers select among various PWM switching methods [2].

In standard low-voltage (400 Vll−RMS) grid-tied systems, the converter design involves the well-matured silicon MOSFET and Insulated Gate Bipolar Transistor (IGBT) switches allowing switching in several tens of kHz at several kVA levels, while the switching frequency decreases to several kHz levels in hundreds of kVA rating applications. In all these applications, the converter is interfaced with the grid (or the load) through LCL ripple filter structures to meet grid code compatibility by restraining the total current harmonic distortion (THD_i) and individual harmonics [3]. The converter-side filter inductor experiences the rectangular voltage pulses and carries the load and PWM ripple current such that it is both electrically and magnetically stressed. After the power semiconductors, it is the costliest, lossiest, and largest element in the circuit.

Inductors made of distributed airgap soft magnetic materials (powder alloys such as sendust, powder iron, etc.) are often favored for their low core loss, high volumetric energy density, and acceptable cost [4]–[7]. With saturation slightly allowed, their use in 3P3W VSCs covers a vast variety of applications such as PV inverters and UPS systems where power range extends from a few kWs to several hundred kWs. However, especially when made cost- and efficiency-effective, such inductors exhibit deep magnetic saturation characteristics [5], [6] [7, Ch. 2]. While the magnetic saturation of the inductor is favorable in terms of cost, size, and loss reduction, it affects the converter performance characteristics adversely by shrinking the current

Fig. 1. Two-level 3P3W VSC topology.
control bandwidth and generating low-order current harmonics. Thus, the inverter current control dynamic performance and steady-state current waveform quality substantially degrade prohibiting extensive and effective use of saturable inductors.

Recently, for the inductor saturation issues in 3P3W systems, two-phase modulation is proposed [8]. This study has the disadvantage of lacking the integrator terms in the controller, which is a basic requirement for low-frequency controller performance. In [9], a feed-forwarding approach is adopted to conventional proportional-integral (PI)-type synchronous frame regulators simplifying the approach in [8] and reducing low-order harmonics by 60%. Another current controller structure for 3P3W VSC systems based on complex vectors is presented in [10], which achieves reducing the cross-coupling effects via feedback rather than decoupling directly. Such an approach is shown to exhibit improved robustness to estimated inductance errors when compared to conventional synchronous frame current control (CSCC). A simplified feedback linearization method in [11] is employed to control the grid current; however, inductor saturation is not elaborated, which can occur even in linear-inductor designs. Also, a fractional-order repetitive control method is elaborated in [5] to solve the inductor saturation-based problems in 3P3W VSCs. However, these methods employ linear controller structures and three-phase balanced system modeling. Thus, for such methods, together with a nonlinear system eventuated by inductor saturation, performance degradation is inevitable. As a nonlinear control approach for single-phase VSCs, inverse system of the physical system is used for solving inductor saturation-related issues [12]. However, as studied in this article, direct extension of such approach to 3P3W VSCs controlled in synchronous frame requires exact modeling, appropriate coordinate transformations, and modulation of 3P3W VSCs.

In spite of the significance of the inductor saturation in 3P3W VSC systems, relevant literature is limited; simple, complete, and effective engineering solutions to associated problems are absent. As a result, conservative designs (involving large inductors with limited saturation and reasonable current control performance) or low-cost designs (highly saturating inductors with poor current control performance) have been employed in the field.

This article proposes an inverse dynamic model-based compensation (IDMBC) method and an unconventional two-phase coordinate transformation (TCT) to overcome the problems that arise with the use of saturable inductors in 3P3W VSCs with CSCC and conventional three-phase/two-phase coordinate transformation (CCT) [13, App. B] [14]. Throughout the article, the term saturable inductor corresponds to inductors that are provided verifying the superiority of the proposed IDMBC method over the CSCC method while the converter-side inductors are deeply saturated. Section V concludes this article.

II. INDUCTOR SATURATION COMPENSATION IN 3P3W VSC SYSTEMS

The CSCC is the state-of-the-art technique for the grid connection of 3P3W VSCs and has been applied in the field satisfactorily with linear inductors being constituents of LCL filters [13], [15]–[17]. However, when the inductors are saturable, the control performance degrades significantly; the current waveform becomes distorted and the dynamic response characteristics become angle-dependent [7], [9], [12].

The objective of this article is to present an easy-to-apply and effective current control method to overcome the drawbacks of 3P3W VSC systems when saturable inductors are employed. The presented method employs converter-side current feedback with inherent damping characteristics [18]. Therefore, the investigated current control system is considered as a 3P3W L filtered VSC as shown in Fig. 1. If the inherent damping is not sufficient, active or passive damping methods can be incorporated [19] and then the proposed approach becomes valid.

When employs saturable inductors in the interface LCL filter, a 3P3W grid-connected VSC system exhibits nonlinear and unsymmetrical behavior. Thus, in this section, the two-phase modeling, which describes the nonlinear and unsymmetrical 3P3W VSC systems, is presented first. Then, an inverse dynamic model (IDM)-based linearization of the nonlinear system is provided. Subsequently, an integral compensator is also incorporated to yield an equivalent linear inductance system. After that, having a two-phase linear equivalent system, TCT is presented to be able to transform the ac stationary frame signals into dc synchronous frame signals. Finally, the overall current control of the linearized system is overviewed and cross-coupling decoupling and modulator details are provided.

A. Two-Phase Modeling of 3P3W VSC Systems

Fig. 2 shows the equivalent 3P3W VSC system for current regulation, wherein \(v_{ab}, v_{bc}, \) and \(v_c\) are the input control voltages provided by the VSC. This system can be exactly described by the mesh equations

\[
\begin{align*}
\frac{di_a}{dt} &= L_a \left( \frac{di_a}{dt} - R_a i_a - R_b i_b \right) \\
\frac{di_b}{dt} &= \left( L_b + L_c \right) \frac{di_b}{dt} + L_c \frac{di_a}{dt} + (R_b + R_c) i_b + R_c i_a 
\end{align*}
\]

Fig. 2. 3P3W R-L-E circuit modeling the current controlled system.
where \( v_{ab} = v_a - v_b, \ v_{bc} = v_b - v_c, \) and \( i_c = -i_a - i_b \) are employed. The inductances and resistances are instantaneous (nonlinear) values. This equation can be written in matrix form as

\[
v_{\sigma} = L_{\sigma} \frac{d}{dt} i_{\sigma} + R_{\sigma} i_{\sigma}\]

where \( v_{\sigma} = [v_{ab} \ v_{bc}]^T, \ i_{\sigma} = [i_a \ i_b]^T, \ L_{\sigma} = \begin{bmatrix} L_a & -L_b \\ L_c & L_b + L_c \end{bmatrix}, \) and \( R_{\sigma} = \begin{bmatrix} R_a & -R_b \\ R_c & R_b + R_c \end{bmatrix}. \)

In (2), \( v_{\sigma} \) represents the control column vector in a line-to-line voltage manner to regulate the load current vector \( i_{\sigma}. \)

### B. Linearization in the Large Via Inverse Dynamic Model

System inversion has been employed in the control of nonlinear systems [12], [20, Ch. 6]. However, the presented two-phase IDM-based linearization for 3P3W VSCs has not been reported in the literature. Likewise, different from CCT, a TCT is introduced in this article as the number of linearized system variables and the synchronous frame variables is two. Within the scope of this article, the term “linearization in the large” is employed to discriminate the proposed nonlinear control method from the well-known small-signal linearization at an equilibrium point.

Having the exact system model by (2), the IDM of the system is obtained via setting the input voltage \( (v_{\sigma}) \) as the replica of the nonlinear system with intermediate variable \( (u_{\sigma}) \) as

\[
L_{\sigma} \frac{d}{dt} i_{\sigma} + R_{\sigma} i_{\sigma} = \hat{L}_{\sigma} \frac{d}{dt} u_{\sigma} + \hat{R}_{\sigma} u_{\sigma}
\]

where \( \hat{L}_{\sigma} \) and \( \hat{R}_{\sigma} \) are estimated values of \( L_{\sigma} \) and \( R_{\sigma}, \) respectively. Fig. 3(a) illustrates (3) in terms of nonlinear block diagrams, where the IDM and the nonlinear physical system are in cascade. When these estimated terms are equal to the actual parameters and the initial states are identical \( (i_{\sigma}(0) = u_{\sigma}(0)) \), the equivalence \( i_{\sigma}(t) = u_{\sigma}(t) \) is obtained for \( t > 0. \) Therefore, the resulting system from the cascade of these two nonlinear systems can be considered as a linear, unity-gain, zero-phase system as shown in Fig. 3(b).

### C. Incorporation of Integral Compensator

The unity-gain zero-phase equivalent system obtained in the previous subsection by making use of the IDM has infinite bandwidth. However, the physical realization of such a system will exhibit strong deficiencies such as high-frequency oscillations and tracking error due to physical bandwidth limitations and lack of feedback, respectively. Furthermore, the derivative operator in the IDM makes the overall control system vulnerable to high-frequency noise. For these reasons, an integral compensator with a coefficient of \( 1/L_{\text{min}} \) is cascaded to the IDM as shown in Fig. 4(a). The integration term can be spread to the IDM and cancel the differentiation block as shown in Fig. 4(b), exterminating the noise amplification problem. Resultantly, the compensated system behaves as a two-input two-output decoupled linear system as shown in Fig. 4(c).

### D. Two-Phase Coordinate Transformation

The synchronous frame control of 3P3W VSCs and the utilization of linear (usually PI type) controllers in synchronous frame have become industry standard with the extensive use of microcontrollers and digital signal processors (DSPs) since 1980s. This is due to the ease of application, zero steady-state error, independent control of d- and q-axis (active/reactive) power, high performance under distortions, and low computational burden brought by such controllers [15], [23]. However,
when the physical system becomes a nonlinear one with the use of saturable inductors, two-phase exact modeling becomes essential and a distinct approach of coordinate transformation is needed to transform two-phase variables into synchronous frame variables. Therefore, a TCT achieving synchronous/stationary frame transformation will be presented now.

The linearized overall system shown in Fig. 4(c) represents two independent phases having the dynamics of

$$w_{\sigma} = L_M \frac{d}{dt} i_{\sigma}$$

(4)

where $L_M = \begin{bmatrix} L_{\min} & 0 \\ 0 & L_{\min} \end{bmatrix}$ in which $L_{\min}$ is the minimum inductance value with $i_{\sigma} = [i_a \ i_b]^T$ being the measured phase currents of the physical system and $w_{\sigma} = [w_a \ w_b]^T$ being the virtual control voltages. To describe the stationary frame relation (4) in synchronous frame, a space vector can be defined as

$$\vec{x}_{ab} = \frac{2}{3} \{(2 + a)x_a + (2a + 1)x_b\}$$

(5)

where $a = e^{j \frac{2\pi}{3}}$ and $x_a$ and $x_b$ denote instantaneous values of phase-a and -b variables such as voltages or currents. It is notable that relation (5) can be obtained from well-known space vector definition [13] by employing dependency of phase variables ($x_c = -x_a - x_b$). Similar to the conventional space vector maths, on a synchronously rotating frame, the rotating space vector $\vec{x}_{ab}$ can be represented as dc quantities according to

$$\vec{x}_{dq} = x_d + jx_q = \vec{x}_{ab}e^{-j\theta}$$

(6)

where $\theta = \omega t$. The relation between the stationary and synchronous variables defined by (5) and (6) can also be given in a scalar vector form as

$$x_{\sigma} = T(\theta)x_{dq}$$

(7)

where $T(\theta) = \begin{bmatrix} \cos(\theta) & -\sin(\theta) \\ \cos(\theta - 2\pi/3) & -\sin(\theta - 2\pi/3) \end{bmatrix}$, $x_{dq} = \begin{bmatrix} x_d \\ x_q \end{bmatrix}$, and $x_{\sigma} = \begin{bmatrix} x_a \\ x_b \end{bmatrix}$. Furthermore, the inverse of TCT matrix is $T^{-1}(\theta) = \frac{2}{\sqrt{3}} \begin{bmatrix} \cos(\theta - \pi/6) & \sin(\theta) \\ -\sin(\theta - \pi/6) & \cos(\theta) \end{bmatrix}$. This transformation constitutes the basis for the synchronous frame current control of the two-phase linearized system.

**E. D-Q Axis Cross-Coupling Decoupling**

In CSCC with linear system elements, cross-coupling terms appear in synchronous frame representation [23]. Owing to the cross-coupling terms, d- and q-axis currents cannot be controlled independently. Especially in transients, one axis variable affects the other. Therefore, it is common practice to decouple the cross-coupling terms [10], [23].

On the other hand, for the IDMBC method, the synchronous frame representation of the linearized system of (4) is the same as if the synchronous frame representation of the CSCC method having linear components. Applying the TCT given by (7) on the stationary frame open-loop dynamics of (4), one can obtain the synchronous frame open-loop dynamics as

$$w_{dq} = L_M \frac{d}{dt} i_{dq} + H_{cc}i_{dq}$$

(8)

where $H_{cc} = \begin{bmatrix} 0 & -\omega L_{\min} \\ \omega L_{\min} & 0 \end{bmatrix}$. Similar to the case in CSCC with linear inductors, the synchronous frame representation of the linearized system brings cross-coupling terms ($H_{cc}i_{dq}$) that weaken the dynamic behavior and independent control of d- and q-axis currents. For better performance of the control system, the cross-coupling terms should be decoupled from the control path via adding the calculated cross-coupling terms ($\hat{H}_{cc}i_{dq}$) to the control input $w_{dq}$ in (8), in the same manner as the conventional d- and q-axis cross-coupling decoupling [23].

**F. Current Control Via IDMBC**

The overall closed-loop IDM-based current control architecture is illustrated in Fig. 5. The method operates as follows. The measured currents are transformed into synchronous frame according to the TCT of (7). Then, the errors between the commanded (reference) and measured synchronous frame current values (termed as the error signals, $e_{dq} = [e_d \ e_q]^T$) are processed by PI compensators and the cross-coupling feed-forward terms ($H_{cc}i_{dq}$) are added to decouple the d- and q-axis dynamics and get the synchronous frame control voltage $w_{dq}$. The virtual control voltages in stationary frame $w_{\sigma}$ are obtained as an input to the linearized system of (4) based on the TCT given by (7). When $w_{\sigma}$ is applied, the output of the unified IDM block becomes the voltage command $\vec{v}_{\sigma} = [v_{ab} \ v_{bc}]^T$. After the decoupling of the disturbances (such as grid voltage, nonlinearities such as semiconductor voltage drops, dead-time effects;
all denoted as two-element line-to-line disturbance voltage $v_D$ by the estimated/measured disturbance voltage ($\hat{v}_D$), the voltage command for PWM ($v_{\text{ref}} = [v_{a\text{ref}}, v_{b\text{ref}}, v_{c\text{ref}}]^T$) is obtained.

Distinctive from the conventional three-phase PWM, the modulation signals of IDMBC method are obtained in line-to-line voltage manner. These voltages can be converted into phase voltage commands prior to employ conventional three-phase PWM. Fig. 6(a) shows the modulator block diagram in a vector input vector output manner, whereas, in Fig. 6(b), the illustration of the modulator is performed via scalars. Regarding Fig. 6(b), the line-to-line voltages are converted to phase voltages ($v_{a\text{ref}}, v_{b\text{ref}},$ and $v_{c\text{ref}}$) via scalar multiplication by the matrix $K$. The matrix $K$ can be deduced from the argument that the phase voltages can be derived from the line-to-line voltages as

$$v_a = (v_{ab} + v_{a\text{ref}})/3$$
$$v_b = (v_{ba} + v_{b\text{ref}})/3$$
$$v_c = (v_{ca} + v_{c\text{ref}})/3$$

(9)

where $v_{a\text{ref}}, v_{b\text{ref}},$ and $v_{c\text{ref}}$ are the zero-sum phase voltage commands ($v_{a\text{ref}} + v_{b\text{ref}} + v_{c\text{ref}} = 0$). Hence, rearranging (9), one can obtain the final command voltages for PWM as

$$[v_{a\text{ref}}, v_{b\text{ref}}, v_{c\text{ref}}]^T = v_b + Kv_{a\text{ref}}$$

(10)

where $K = \frac{1}{3} \begin{bmatrix} 2 & -1 & -1 \\ 1 & 1 & 2 \end{bmatrix}^T$, to generate the VSC output voltages ($v_a, v_b, v_c$) to control the nonlinear physical system currents.

### G. Current Control System Stability

For the purpose of stability investigation and comparison of the IDMBC and CSCC methods, Fig. 7 shows the magnitude and phase characteristics of the d- and q-axis open-loop systems. These characteristics are obtained for the system parameters that will be given in the next section by assuming the VSC system operates at the steady state where the inductance variation is dominated by the fundamental frequency large-signal and small magnitude high-frequency command applied.

The d- and q-axis magnitude responses of IDMBC method in Fig. 7(a) and (b) show the linearized characteristics having the design control bandwidth ($\omega_{BW}$) independent of the phase angle ($\theta = \omega t$). The same figures also show that the CSCC method exhibits shrinked bandwidth with $\theta$ dependence. Similarly, as shown in Fig. 7(c) and (d), the phase margin (PM) of the IDMBC method is $90^\circ$ independent of $\theta$ while the PM of the CSCC method is less than $80^\circ$ and changes with $\theta$. These results indicate the IDMBC method is more stable and has a better utilization of the available control bandwidth when compared to the CSCC method. In [7], further details on the control system characteristics are provided.

### III. Simulation Results

The proposed IDMBC method simulation results are presented in comparison with those of the CSCC method. The current regulator attributes are investigated by means of dynamic command response, cross-coupling behavior, and steadystate waveform quality. Experimental verification of the simulation results is also provided in the subsequent section for the same configurations and tests as described here.

#### A. Simulated System Configuration

1) **Simulated System Hardware**: Two hardware configurations are employed as shown in Fig. 8 for the investigation of the performance of the current controllers. For the dynamic command response and cross-coupling behavior investigations, the VSC+inductor terminals short-circuited configuration shown in Fig. 8(a) is employed to render the sole characteristics without grid voltage interference. On the other hand, the waveform quality is investigated via grid-connected configuration shown in Fig. 8(b).
The converter-side saturable inductors are realized from a Sendust-type toroidal powder core [24] owing to the temperature stability, continuous inductance characteristics, low core losses, and minimal aging characteristics of powder core materials [25]. In Table I, the design parameters for the designed saturable inductors are listed, and in Fig. 9, associated L-i characteristics are illustrated. These characteristics are obtained from experimental measurements as performed in [26]. The selected core is deeply saturated so that the inductance is decreased with the increase in the current down to 1/9th of the initial value. Such deep saturation levels can be tolerated in 3P3W VSC systems by means of PWM ripple due to balancing effect of the inductors as the phase currents are driven by the VSC line-to-line voltages and the minimum inductance values of any two phases do not overlap in time.

The simulation/experiment parameters for hardware configurations shown in Fig. 8(a) and (b) are tabulated in Table II. The filter capacitor is selected to be 2.2 μF connected in a wye configuration. The grid-side inductor is selected to be 2 mH. As will be shown, the LCL filter provides sufficient PWM harmonic attenuation even for the minimum inductance value of the converter-side inductor which is within the design constraints provided in [16, Section III].

### 2) Simulated System Control

A small step size (0.02 μs) is used for the simulations of nonlinear VSC systems with saturable inductors. The synchronous frame PI controllers (for the d- and q-axis) are tuned to provide a bandwidth of $2\pi$500 rad/s as if the system is a linear one (having constant converter-side inductance of $L_{\text{min}}$). Accordingly the PI gains are tuned as $K_p = \omega_{\text{BW}} L_{\text{min}}$ and $K_i = \omega_{\text{BW}} R$ for both methods [13]. The cross-coupling decoupling is performed as the same for both controllers by using $L_{\text{min}}$ to form the cross-coupling decoupling matrix $\hat{H}_{cc}$. Simulation results are obtained as the compensator gains are kept the same for each configuration, current control method, and test.

#### B. Dynamic Command Response

The dynamic command response characteristics of a current regulator are one of the figures of merit to evaluate and compare the controller performances being a direct indicator of the system bandwidth at the operating condition of the converter [26]. The faster the response implies the higher the control bandwidth.

To examine the dynamic command response characteristics of the current controllers, simulations are performed by providing 1 A step changes on the d-axis current command whereas the q-axis current command is set to zero with VSC+inductor terminals short-circuited [Fig. 8(a)]. In Fig. 10, these responses are illustrated for CSCC and IDMBC methods. As the q-axis currents are set to zero, they are not illustrated in the figure. For the CSCC method, the response (or rise) time to a 1 A step increase is highly dependent on the current bias. When the current dc-bias is zero, the rise time is approximately 3–4 ms whereas when the d-axis current dc-bias level is 10 A, and the

### Table I

| Parameter                        | Value     |
|----------------------------------|-----------|
| Core material/shape              | Sendust/toroid |
| Initial permeability             | 90 μ       |
| Core outer diameter              | 47.6 mm    |
| Core weight                      | 130 g      |
| Turn number                      | 144        |
| Wiring                            | 2 × AWG22  |

### Table II

| Parameter                        | Value     |
|----------------------------------|-----------|
| dc-link voltage ($V_{dc}$)       | 350 V     |
| Rated current                    | 10 Apeak/phase |
| Grid voltage                     | 115 $V_{RM}$/phase |
| Fundamental frequency ($\omega$) | 2π50 rad/s |
| Carrier frequency ($f_c$)        | 10 kHz    |
| Modulation method                | SVPWM     |
| Sampling period ($T_s$)          | 50 μs     |
| Inverter dead-time               | 3 μs      |
| $L_{\text{max}}/L_{\text{min}}$ (converter side) | 4.45/0.5 mH |
| $L_{\text{g}}$ (grid side)       | 2 mH      |
| $R$ (converter-side equivalent resistance) | 0.5 Ω     |
| $C_F$ (LCL filter capacitor)     | 2.2 μF    |
rise time of the CSCC method is around 1 ms. Such an operating point-dependent dynamic command response characteristics is due to the distinct bandwidth of the d-axis system that changes with the inductance.

On the other hand, the response time of the IDMBC method is the same for all current bias levels (1 ms) exhibiting that the bandwidth is conserved throughout various operating conditions of the VSC.

C. Cross-Coupling Behavior

In addition to the dynamic command response performance, it is desired for a 3P3W VSC current control system to have independent control of the d- and q-axis currents without any coupling effects on each other. To examine the cross-coupling behavior of the current controllers, the current command value for one axis is changed instantly while the actual current of the other axis is observed.

In Fig. 11, the waveforms regarding the observations are illustrated. For the CSCC method, when the d-axis or q-axis is excited, the current of the other axis responds to the change which is undesirable for a current controller. On the other hand, in the case of the IDMBC method, the two axes seem decoupled from each other almost perfectly as when one axis current is changed, the other axis current does not exhibit any considerable change as a desirable feature of a high-performance current controller.

D. Waveform Quality

The performance investigation with grid-connected simulations shows that the current waveform quality can be well improved on the grid side. At rated load, the grid-side current distortion is reduced from 5.45 to 1.82% with the use of the proposed method.

Fig. 12(a) and (b) shows the rated load converter-side current, voltage command signal, zero-sequence signal, and zero-sequence added voltage command signal (the signal provided to triangle comparator) of the CSCC and IDMBC methods (here and hereafter, when the term “scale” is used, the graph quantity is obtained by multiplying the physical quantity with the scale number). On the other hand, Fig. 12(c) and (d) shows the grid-current and the inductance of the converter-side inductor, decreasing almost to 1/9th of its zero current value at full load, as expected. As filtered by the LCL filter, almost all PWM caused high-frequency current harmonics are seen to be cleared at the grid side for both methods. Fig. 12(e) shows the harmonic spectra of the methods around the low-frequency region. Based on the spectra, the most dominant harmonics are observed to be fifth and seventh for both methods. The low-order harmonics in the grid-side current in the CSCC method are significantly suppressed with the use of the IDMBC method.

IV. EXPERIMENTAL RESULTS

The hardware implementation for the dynamic command response and cross-coupling behavior investigation of the current control methods is performed as VSC+inductor terminals short-circuited configuration shown in Fig. 8(a) to mitigate the grid-side effects on the measurements. On the other hand, for waveform quality investigation, grid-connected configuration in Fig. 8(b) is employed. Accordingly, the experimental system configuration by means of hardware and control implementation is provided.
A. Experimental System Configuration

1) Experimental System Hardware: The experimental system parameters are kept the same as tabulated in Table II. The power module PM75RL1A120 is utilized in the realization of the VSC power semiconductors. For safety reasons, in the grid-connected experiments, the ac-grid voltage is stepped down to 115 V RMS/phase via three single-phase transformers connected in a wye–wye configuration, each having a leakage inductance of 0.25 mH. Supplemental linear filter inductors as grid side have been inserted also to form grid-side induc-

tors \( L_g-a, L_g-b, L_g-c \) together with the transformer leakage inductance. The converter-side saturable inductors are wound to sendust powder material toroid cores [24] yielding the inductance characteristics given in Fig. 9. Fig. 13 shows this three-phase grid-connected VSC system hardware.

2) Control Implementation: A 150-MHz floating-point DSP is used to realize the control functions in discrete time. The interrupt cycle is selected as 50 \( \mu s \) to yield a double update for a PWM cycle of 100 \( \mu s \) (10 kHz). At every interrupt cycle, the controller executes the grid synchronization via a three-phase vector PLL method [13], current control algorithms including anti-windup, software VSC protection, and PWM signal generation. A dead-time of 3 \( \mu s \) is utilized and its compensation is performed. The backward Euler method is utilized for the discretization of the continuous-time controllers.

All system control parameters are selected the same as in the case of simulations. The L-i characteristics of the converter-side inductors are modeled as second-order polynomials [26] to establish an inverse dynamic model of the nonlinear load. As the converter-side currents are double sampled [13], the d- and q-axis currents obtained via the DSP readings are almost PWM ripple-free. Both the readings from the DSP and the oscilloscope readings are plotted via a computer for consistency.

B. Dynamic Command Response

As in the simulations, 1-A step changes are applied to the d-axis current command for both methods at distinct bias levels while the q-axis current command is set to zero. Fig. 14 shows the associated dynamic command response waveforms. When the d-axis current is zero, the rise time of the d-axis current is quite long in the CSCC method when compared that of the IDMBC method even if the associated PI controller parameters are the same aiming the same bandwidth for the methods. When the d-axis current offset is increased, the rise time of the CSCC method starts to decrease on account of the bandwidth shrinkage phenomenon in the CSCC method associated to the changing converter-side inductance. However, the dynamic command response characteristics of the IDMBC method are bias-independent as initially designed to be.

C. Cross-Coupling Behavior

The experimental cross-coupling behavior of the CSCC and IDMBC methods are investigated by applying a step change
to the current command of an axis and then observing the actual current of the other axis. In Fig. 15(a), the d-axis current command is increased from 1 to 2 A while the q-axis command is kept at 0 A. For the CSCC method, the q-axis current is not stationary but it deviates from its reference value with the change in the d-axis current indicating the imperfect cross-coupling decoupling. When the d- and q-axis currents are exchanged [Fig. 15(b)], the same phenomenon again appears in a similar way but distinct in sign with a reasoning that the cross-coupling terms have opposite signs.

In Fig. 15(c) and (d), the dc-bias levels on both axes are set to 4 A. Then the d- and q-axis current commands are increased to and decreased from 5 A. In these cases, even though the sixth-order harmonic blurs the decoupling effect, it can still be identified that the decoupling is imperfect when the rising and falling regions of the current waveforms are considered. On the other hand, for all tests, the IDMBC method exhibits superior cross-coupling decoupling performance as expected.

D. Waveform Quality

The experimental waveform quality investigation is conducted via grid-connection of the VSC through LCL filter [see Fig. 8(b)]. In Fig. 16(a) and (b), the VSC output voltages, and the phase-a converter-side inductor currents of the CSCC and the IDMBC methods are illustrated, respectively. Further, Fig. 16(c) and (d) shows the capacitor voltages and the grid currents. In these figures, the converter-side PWM harmonics are seen to be well suppressed in the grid-side current, demonstrating the effectiveness of the LCL filter above the carrier frequency.

On the other hand, the low-frequency harmonic characteristics of the CSCC and IDMBC methods are quite different from each other. In Fig. 17(a) and (b), the three-phase grid currents of the methods are illustrated. The THD value for the CSCC method is measured to be 9.6% which is higher than the 5% THD limit set by grid connection standards such as IEEE 519. On the contrary, the THD value of the IDMBC method is measured to be 2.58% which is well below the limit. The measured THD values in the grid-connected experiments are slightly higher than their simulated counterparts because of the secondary effects such as imperfect cross-coupling decoupling; measurement errors and
delays; and imperfect decoupling of the grid voltage, which may also have harmonics and imbalances.

In Fig. 17(c), the harmonic spectra belonging to the grid currents shown in Fig. 17(a) and (b) are illustrated. In the spectra, the zero-sequence current harmonics are minimal as there is no return path for these harmonics to flow. For the CSCC method, the fifth and the seventh harmonics flow abundantly, making this spectrum can be attributed to poor disturbance rejection characteristics, imperfect cross-coupling decoupling, and the inability of the CSCC method to respond load nonlinearity. In the case of IDMBC method, the spectrum has acceptably low fifth and seventh harmonic current magnitudes, demonstrating the suitability of the method for grid-connected applications with saturable inductors.

V. CONCLUSION

This article presented an IDMBC method to enable 3P3W VSC systems to employ saturable inductors as converter-side inductors in the grid interface LCL filters. The main aim of such utilization was higher efficiency, lower cost, and lower volume. For this purpose, a two-phase model was obtained first. Then, the linearization of the two-phase model was performed in the large. To control the linearized two-phase system in synchronous frame, an unconventional TCT was derived as well. The developed IDMBC method was tested in simulations and experiments in comparison with the CSCC. It was seen that CSCC method suffered both the low-frequency current waveform quality and current control bandwidth issues. On the other hand, the IDMBC method exhibited superior performance with uniform current control bandwidth and high steady-state current waveform quality that met grid codes. With the proposed IDMBC method, deeper saturation levels of inductors with soft magnetic materials can be reached, and hence these inductors in 3P3W VSCs can be extensively used in cost-, efficiency-, and weight-critical applications such as grid-connected renewables, storage, electric vehicle chargers, statcoms, and UPS systems.

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Ziya Özkan (Member, IEEE) was born in Canakkale, Turkey, in 1986. He received the B.S., M.Sc., and Ph.D. degrees from Middle East Technical University, Ankara, Turkey, in 2009, 2012, and 2019, respectively, all in electrical and electronics engineering. His research interests include design and control of power electronics circuits.

Ahmet M. Hava (Senior Member, IEEE) received the B.S. degree from Istanbul Technical University, Istanbul, Turkey, in 1987, and the M.Sc. and Ph.D. degrees from the University of Wisconsin, Madison, WI, USA, in 1991 and 1998, respectively, all in electrical engineering. In 1995, he was with Rockwell Automation-Allen Bradley Company, Mequon, WI, USA. From 1997 to 2002, he was with Yaskawa Electric America, Inc., Waukegan, IL, USA. From 2002 to 2021, he was with the Middle East Technical University, Electrical and Electronics Engineering Department, Ankara, Turkey, where he was a Professor. Since September 2021 he has been with the Işık University, Electrical and Electronics Engineering Department, Istanbul, Turkey, where he is a Professor. His research interests include power electronics, motor drives, power converter design and control, power quality, and renewable energy systems.