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

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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 (CSCC) methods, the inductor saturation results in significant dynamic performance loss and poor steady-state current waveform quality. This paper proposes an inverse dynamic model based compensation (IDMBC) method to overcome these performance issues. Accordingly, a review of inductor saturation and core materials is performed, and the motivation on the use of saturable inductors is clarified. Then, two-phase exact modelling of the 3P3W VSC control system is obtained and the drawbacks of CSCC have been demonstrated analytically. Based on the exact modelling, the inverse system dynamic model of the nonlinear system is obtained and employed such that the nonlinear plant is converted to a fictitious linear inductor system for linear current regulators to perform satisfactorily.

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

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, statcom, storage, and uninterruptible power supply. 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]. Pulse width modulation (PWM) mode of operation is common and designers select among various PWM switching methods [2].

In standard low-voltage (400 V_{IL-RMS}) grid-tied systems, the converter design involves the well-matured silicon MOSFET and 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

Fig. 1. Three-phase three-wire two-level voltage-source converter topology.

semiconductors, it is the costliest, lossiest, and largest element in the circuit. With the present switching frequency levels, filter inductors with only several percent values of the base impedance suffice, decreasing the total size and cost of the converter.

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], [5]. However, most such inductors have deep magnetic saturation characteristics, especially when made cost-effective [5]. While the magnetic saturation of the inductor is favorable in terms of cost and size reduction, it affects the converter performance characteristics adversely by shrinking the current control bandwidth and generating low order current harmonics. Thus, the inverter current control dynamic performance and steady-state current waveform quality substantially degrade.

Recently, research has been conducted to solve the inductor saturation based performance loss issues for single-phase controlled VSCs [6]–[12]. The approach for single-phase controlled VSC systems in [11] is generalized to 3P3W VSC systems in this study. Accordingly, this paper investigates the 3P3W VSCs with highly saturable inductors in terms of their performance characteristics and proposes an inverse dynamic model based compensation (IDMBC) method to overcome the inductor saturation based performance degradation of conventional synchronous frame current control (CSCC) methods. Throughout the paper, the term saturable inductor corresponds to inductors that normally (within the rating range) enter the deep magnetic saturation region (by design, due to cost and filtering advantages).

For the inductor saturation issues in 3P3W systems, in [13], two-phase modulation is proposed. This study has the disadvantage of lacking the integrator terms in the controller which is a basic requirement for low-frequency controller performance.

In [14], a feed-forwarding approach is adopted to conventional PI type synchronous frame regulators simplifying the approach in [13], and reducing low order harmonics by 60 %. However, a symmetrical linear 3P3W system model is utilized in the work which is shown to be incorrect in Section III.

Another current controller structure for 3P3W VSC systems based on complex vectors is presented in [15], which achieves reducing the cross-coupling effects via feedback rather than decoupling directly. Such an approach is shown to exhibit improved robustness to estimated parameter errors when compared to CSCC. However, when the controller structure is linear as in the case of complex current regulator, performance degradation is inevitable in the case of inductor saturation.

In spite of the significance of the subject, relevant literature is limited; simple, complete, and effective engineering solutions to the problems due to the utilization of saturable inductors 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 paper investigates and solves the problems that arise with the use of saturable inductors in 3P3W VSCs. While the investigation is conducted for two-level 3P3W VSCs, it can be generalized to multiphase multilevel configurations with slight modifications. It is organized as follows. In Section II, the theory of inductor saturation is introduced, a brief review of inductor core materials is performed, and the motivation on the use of saturable inductors is presented. In Section III, conventional current control of 3P3W VSC systems is elaborated and analytically its inconsistency is demonstrated when saturable inductors are employed. Moreover, two-phase modelling is established which describes the exact 3P3W nonlinear system dynamics. Based on the modelling, the drawbacks of the CSCC are investigated analytically. In Section IV, to overcome the associated drawbacks, the overall inductor saturation compensation method is presented based on the inverse dynamic modelling of the nonlinear 3P3W VSC system. In Section V, practicality of the IDMBC method is discussed and in Section VI, conclusive remarks are made.

In a subsequent paper [16], the command to output, disturbance to output, and cross-coupling decoupling characteristics of the proposed IDMBC method and CSCC method are elaborated analytically. Furthermore, simulation and experimental results are provided in comparison with the CSCC method to verify the analyses and to demonstrate the superiority of the IDMBC method.

II. SATURABLE INDUCTORS

A. Theory of Inductor Saturation

The introduction of the terminal equation of an inductor is of great importance especially when it is a saturable one as there exist two definitions for inductance that can lead to confusion in the development of the system control. The first definition relates the magnetic flux λ and the current i of an inductor as

$$\lambda = \mathcal{L}_s i \quad (1)$$

where $\mathcal{L}_s$ is called secant inductance. The other definition can be derived by Faraday's law of induction and the chain rule as

$$v_{\mathcal{L}} = \frac{d\lambda}{dt} = \frac{d\lambda}{di} \frac{di}{dt} = \mathcal{L} \frac{di}{dt} \quad (2)$$

wherein $\mathcal{L}$ is the incremental inductance defined by

$$\mathcal{L} \triangleq \lim_{\delta i \rightarrow 0} \frac{\delta \lambda}{\delta i}. \quad (3)$$

When the inductor is a linear one, the definition of the secant inductance and the incremental inductance becomes the same [17]. However, in the case of saturable inductors only the incremental inductance can be used to obtain the terminal equation described by (2). Evidently, the incremental inductance of the system decreases as the magnitude of current increases and the system becomes a nonlinear one as the magnetic domains of the inductor core align [18]. With the ESR of the inductor being $\mathcal{R}$, to represent inductor core and copper losses, the terminal equation of a saturable inductor can be given by

$$v_{\mathcal{L}} = \mathcal{L} \frac{di}{dt} + \mathcal{R}i. \quad (4)$$

B. A Brief Review of Inductor Core Materials

In order to filter the PWM ripple, many filter topologies are presented in the literature [3], [19]. In all topologies, the converter side is matched by an inductor as the converter outputs rectangular voltage pulses. Among the topologies, LCL filter topology gained widespread use due to its low complexity and high PWM ripple attenuation characteristics [19]. When compared to load (grid) side inductor, the converter side inductor is lossier, more expensive, heavier, and more complex to design because the converter side experiences PWM ripple current and the fundamental current whereas the load side inductor is a low frequency one [5].

Practical inductor core materials can basically be classified as laminated silicon steel, ferrite, and powder materials while amorphous steel and nanocrystalline material can still be an option [20]. Depending on the application requirement such as cost, efficiency, and weight, the choice is made. Among the materials, silicon steel exhibits weaknesses in terms of increased core losses due to eddy current losses despite the laminated structure [18] and due to fringing losses especially for high switching frequencies. Due to cost advantage, at very low switching frequencies (approximately less than 10 kHz) and large current ratings the magnetic design involves thin silicon steel laminations, typically multi-gapped airgap design, and multi-turn windings.

At frequencies above several tens of kHz, the choice of magnetic material becomes ferrite and low loss powder alloys. In powder cores, the power loss is directly related to the grain size. The power loss is reduced with tinier grain but the price increases [20]. When compared to powder core materials, air-gapped ferrite is the cheaper solution per weight. Moreover, the inductance characteristics can be made almost linear by adjusting the air-gap. However, even if they are not saturated fully, powder materials have certain advantages over gapped ferrite cores [4], [21], [22]:

- Powder cores do not have fringing losses whereas gapped ferrite cores suffer excessively these losses.
- Powder cores are inherently fault-tolerant due to their continued inductance characteristics whereas ferrite cores are not.
- Flux capacity of ferrite cores decreases with temperature increase whereas that of powder cores are relatively constant.
- Powder cores exhibit excellent EMI performance whereas gapped ferrite cores do not due to fringing flux, which may be prohibitive in power converters requiring low noise levels.

These advantages of powder cores make them preferable over ferrite cores. Furthermore, temperature stability, minimal aging, and minimal L-i characteristics dependency [22] make them suitable for saturable inductor applications as the inductance estimation can be performed with direct calculation or look-up tables. Therefore, the core size, weight, cost and inductor losses can be significantly reduced as discussed in the following.

C. Motivation on the Use of Saturable Inductors

The cost, weight, and loss reduction benefits of using saturable inductors may not be apparent at first sight. The purpose of this subsection is to clarify these benefits.

In terms of design parameters, the incremental inductance can be described by

$$\mathcal{L} = N^2 \mu \frac{A}{l} \quad (5)$$

where N is the number of turns, μ is relative permeability at the excitation level, A is the effective core cross-section area, and l is the effective length of the core. In (5), the ratio of the effective core cross-section area to the effective length of the core (A/l) is constant. Thus, the term $N^2 \mu$ determines the L-i characteristics of an inductor for a given core ($\mathcal{L} \sim N^2 \mu$). The increase in the inductance is drastic with the square of the turn number while the relative permeability decreases with the applied magnetomotive force Ni . The decrease in μ is slight when compared to the increase due to N^2 , especially for powder cores [23], [24]. Therefore, the effective increase in the inductance with an increased turn number can be exploited by using a smaller core to yield reduced inductor size and cost.

When smaller cores are employed to yield the same average inductance, the number of turns increases while the turn length decreases. Furthermore, the core losses per unit volume increases. However, the total core volume is reduced diminishing the total core losses. These counter-effects emphasize a reduction in the inductor losses. In [24], further discussion on the motivation on the utilization of saturable inductors is performed via design examples.

III. THREE-PHASE THREE-WIRE VSC SYSTEMS WITH SATURABLE INDUCTORS

Current control is the building-block approach in most of the 3P3W VSC applications such as grid-connected renewable energy, energy storage, and statcom systems. Even if saturable

inductors are advantageous by means of size, cost, and inductor loss reduction; their utilization brings serious issues when CSCC is employed. In this section, to build a basis for the identification of these issues and a convention for the analyses to be performed, CSCC for 3P3W VSCs is summarized. Then, the inconsistency of the symmetrical three-phase linear system approach is presented. Subsequently, two-phase modelling of 3P3W VSC systems is developed and finally, the issues are investigated analytically.

A. Conventional Synchronous Frame Current Control (CSCC) of 3P3W VSC Systems

Current control of 3P3W VSC systems is thoroughly studied in the literature and hence quite matured especially filters having linear dynamics. Despite the abundance of the filters used in the grid connection, the utilization of LCL filters has become industry standard [3], [25]. The current control of LCL filters can be performed via grid current feedback or converter current feedback. The grid current feedback requires active or passive damping techniques while the converter side current feedback is inherently stable [26], [27]. Hence, the overall grid current regulation is reduced to the inductor current regulation problem with converter side current feedback.

It was shown that the stationary frame PI current regulators eventuate significant steady-state error in the output current waveform [28], [29]. As a measure, stationary frame PI current regulator is modified as a PI+R controller providing sufficient gain, hence sufficiently small steady-state error at the fundamental frequency ω [28]. However, the synchronous frame aligned with the grid voltage space vector has become industry standard in grid-connected converters due to its small steady-state error characteristics at ω that of stationary frame PI+R stationary frame current regulator and due to its ease of d- and q-axis (real/reactive) current control. The implementation of the 3P3W VSC synchronous frame CSCC is based on the space vector representation of continuous-time real variables. For a 3P3W VSC system the space vector of a three-phase variable can be described as

$$x_{abc} = \frac{2}{3}(x_a + ax_b + a^2 x_c) \quad (6)$$

where $a = e^{j\frac{2\pi}{3}}$ and x_a , x_b , and x_c denote instantaneous values of phase variables such as voltages or currents [30], [31]. With such a transformation, three-phase variables can be represented by a single vector rotating with ω . On a synchronously rotating frame the space vector x_{abc} can be represented as dc quantities according to

$$x_{dq} = x_{abc} e^{-j\theta} \quad (7)$$

where $\theta = \omega t$. The real and imaginary components of the complex variable x_{dq} form the d- and q-axis components respectively

$$\begin{aligned} x_d &= \text{Re}\{x_{dq}\} \\ x_q &= \text{Im}\{x_{dq}\}. \end{aligned} \quad (8)$$

Fig. 2 visualizes a space vector and its synchronous frame components for better insight. With the inclusion of a zero

sequence component x_n to the synchronous frame components, the stationary-synchronous frame transformation can be described alternatively as an invertible scalar transformation by

$$\mathbf{x}_{dq} = \mathbf{F}(\theta) \mathbf{x}_{abc} \quad (9)$$

with $\mathbf{x}_{dq} = [x_d \ x_q \ x_n]^T$, $\mathbf{x}_{abc} = [x_a \ x_b \ x_c]^T$, and

$$\mathbf{F}(\theta) = \frac{2}{3} \begin{bmatrix} \cos(\theta) & \cos(\theta - 2\pi/3) & \cos(\theta + 2\pi/3) \\ -\sin(\theta) & -\sin(\theta - 2\pi/3) & -\sin(\theta + 2\pi/3) \\ \frac{1}{2} & \frac{1}{2} & \frac{1}{2} \end{bmatrix}.$$

The transformation matrix $\mathbf{F}(\theta)$ can be encountered with different coefficients in the literature while the prime function, providing a transformation between synchronous and stationary frames, is preserved [32].

Fig. 2. Illustration of a space vector rotating with ω with respect to stationary frame and its decomposition into synchronous frame components.

Fig. 3. 3P3W R-L-E circuit representing the VSC current regulation system.

The current control of the 3P3W VSC system in Fig. 1 can be modelled as shown in Fig. 3 with load voltages and other disturbance sources are not shown as they are assumed to be decoupled from the control system as usual. Assuming linear elements and balanced operation, the 3P3W L-R circuit can be described by the state equation

$$\mathbf{v}_{abc} = L \frac{d}{dt} \mathbf{i}_{abc} + R \mathbf{i}_{abc} \quad (10)$$

where $\mathbf{v}_{abc} = [v_a \ v_b \ v_c]^T$, $\mathbf{i}_{abc} = [i_a \ i_b \ i_c]^T$, and the system parameters are constants ($L = L_a = L_b = L_c$, and $R = R_a = R_b = R_c$). By making use of (9), (10) can be represented in synchronous frame by

$$\mathbf{v}_{dq} = L \frac{d}{dt} \mathbf{i}_{dq} + \mathbf{W}_{cc} \mathbf{i}_{dq} + R \mathbf{i}_{dq} \quad (11)$$

where $\mathbf{W}_{cc} = \begin{bmatrix} 0 & -\omega L \\ \omega L & 0 \end{bmatrix}$, which involves cross-coupling terms due to derivative operation in (10). Further, $\mathbf{i}_{dq} =$

$[i_d \ i_q]^T$ and $\mathbf{v}_{dq} = [v_d \ v_q]^T$. The form of (11) describes a cross-coupled control system in which $\mathbf{v}_{dq}$ is the control input. It is common practice to decouple the cross-coupling term via its estimated value $\hat{\mathbf{W}}_{cc}$. After decoupling and compensation of the error by a PI compensator, the closed-loop system dynamics in synchronous frame can be described by

$$\begin{aligned} K_p \{\mathbf{i}_{dq}^* - \mathbf{i}_{dq}\} + K_i \int_0^t \{\mathbf{i}_{dq}^* - \mathbf{i}_{dq}\} d\tau + \hat{\mathbf{W}}_{cc} \mathbf{i}_{dq} \\ = L \frac{d}{dt} \mathbf{i}_{dq} + \mathbf{W}_{cc} \mathbf{i}_{dq} + R \mathbf{i}_{dq}. \end{aligned} \quad (12)$$

wherein the current command $\mathbf{i}_{dq}^* = [i_d^* \ i_q^*]^T$ is compared to the actual current and compensated by a PI compensator having a proportional gain K_p and an integral gain K_i . In Fig. 4, the overall conventional vector controlled current regulation architecture is summarized. In the figure, 3P3W VSC is modelled as an ideal VSC which is denoted by a unity matrix $\mathbf{I}$. The nonideal constituents of the conversion such as semiconductor voltage drops, dead-time, and the load voltage are modelled by the disturbance vector $\mathbf{v}_{D3}$ having three elements each of which is associated to each phase of the VSC. It is common practice to decouple the estimated disturbance constituents ($\hat{\mathbf{v}}_{D3}$) such as load voltage and dead-time from the control loop as shown in Fig. 4 [33], [34]. Having sufficiently correct cross-coupling and disturbance estimations, the current control system of (12) becomes a decoupled one constituting a subsystem for each axis. The closed-loop s-domain representations of the decoupled d- and q-axis dynamics become the same with the same controller parameters as

$$\frac{i_d(s)}{i_d^*(s)} = \frac{i_q(s)}{i_q^*(s)} = \frac{sK_p + K_i}{s^2L + s(K_p + R) + K_i}. \quad (13)$$

As the closed-loop control structure reached up in (13) is the same as that of a single-phase controlled VSC system, the design of PI controller becomes straightforward and identical to single-phase dc current control case for an L-R load [11]. Accordingly, the PI compensator parameters can be selected as $K_p = \omega_{BW} \hat{L}$ and $K_i = \omega_{BW} \hat{R}$ wherein ω_{BW} is the design bandwidth [12], [32].

The CSCC has become wide-spread for 3P3W VSCs employing linear inductors as the controller gains can be calculated such that dynamic response can be predetermined and satisfactory quality of the current waveform is ensured. Besides, the implementation of the method is quite simple and effective. However, when saturable inductors are employed within, serious drawbacks of the CSCC arise. Prior to the identification of these drawbacks, the inconsistency of the symmetrical linear three-phase system approach in 3P3W VSCs with saturable inductors is shown.

B. The Inconsistency of Symmetrical Linear System Approach

In the development of CSCC, the generic 3P3W system shown in Fig. 5(a) is employed based on the assumption that the voltage potential between the nodes $o-n$ is always zero ($v_{o-n} = 0$). Hence a connection between these two nodes

Fig. 4. Conventional synchronous frame current control architecture of 3P3W VSC systems with linear and symmetrical inductors.

(shown in Fig. 5(a) with dashed line) can be realized without any change in the physical states (voltages and currents) of the lumped elements of the circuit. Such a connection enables one to analyze the 3P3W circuit in a reduced form of a single-phase system shown in Fig. 5(b) in terms of one of the phases of the system, e.g. phase-a. Then the variables belonging to other phases can be found by adding phase shifts of $2\pi/3$ and $4\pi/3$ to the corresponding states.

While the 3P3W circuit is symmetrical and linear, this approach is adopted well to the CSCC of 3P3W VSC systems. However, when saturable inductors are employed, the nodes $o-n$ are not equipotential strictly for all time even if balanced pure sinusoidal voltages are applied. Therefore, the nodal points $o-n$ can not be connected. This is the fundamental obstacle to the assumption of the symmetrical linear 3P3W circuit approach of the CSCC method. Furthermore, (10) represents an overdetermined system as the currents of the three phases can not be controlled independently. In the next section, the exact modelling of 3P3W VSC system is performed to demonstrate the drawbacks of the CSCC and to establish a basis for the development of a high-performance current controller even deeply saturated inductors are employed.

Fig. 5. Generic circuits for analyzing 3P3W systems. (a) Generic 3P3W circuit. (b) The single-phase representation when the 3P3W system is linear and symmetric.

C. Two-Phase Modelling of 3P3W VSC Systems

Fig. 3 shows the equivalent 3P3W VSC system for current regulation wherein v_a , v_b , and v_c are the input control voltages provided by the VSC, namely pole voltages. This 3P3W system can be exactly described by the mesh equations

$$\begin{aligned} v_{ab} &= L_a \frac{di_a}{dt} - L_b \frac{di_b}{dt} + R_a i_a - R_b i_b \\ v_{bc} &= L_b \frac{di_b}{dt} - L_c \frac{di_c}{dt} + R_b i_b - R_c i_c \end{aligned} \quad (14)$$

where $v_{ab} = v_a - v_b$, $v_{bc} = v_b - v_c$. Utilizing that the sum of the phase currents is zero, $i_a + i_b + i_c = 0$, equation (14) can be rewritten as

$$\begin{aligned} v_{ab} &= L_a \frac{di_a}{dt} - L_b \frac{di_b}{dt} + R_a i_a - R_b i_b \\ v_{bc} &= (L_b + L_c) \frac{di_b}{dt} + L_c \frac{di_a}{dt} + (R_b + R_c) i_b + R_c i_a \end{aligned} \quad (15)$$

wherein the inductances and resistances are instantaneous (nonlinear) values. The equation (15) can be written in matrix form as

$$\mathbf{v}_\sigma = \mathbf{L}_\sigma \frac{d\mathbf{i}_\sigma}{dt} + \mathbf{R}_\sigma \mathbf{i}_\sigma \quad (16)$$

where $\mathbf{v}_\sigma = \begin{bmatrix} v_{ab} \\ v_{bc} \end{bmatrix}$, $\mathbf{i}_\sigma = \begin{bmatrix} i_a \\ i_b \end{bmatrix}$, $\mathbf{L}_\sigma = \begin{bmatrix} L_a & -L_b \\ L_c & L_b + L_c \end{bmatrix}$, and $\mathbf{R}_\sigma = \begin{bmatrix} R_a & -R_b \\ R_c & R_b + R_c \end{bmatrix}$.

In (16), $\mathbf{v}_\sigma$ represents the control column vector in a line-to-line voltage manner to regulate the load current vector $\mathbf{i}_\sigma$. In the current regulation process, the load voltage, nonidealities such as semiconductor voltage drops, dead-time effects, and the load (grid) voltage appear as voltage disturbance sources and denoted as two-element line-to-line voltage vector $\mathbf{v}_D$. The decoupling of $\mathbf{v}_D$ with its estimated/measured value $\hat{\mathbf{v}}_D$ corresponds to the same that of the CSCC, hence can be performed in the same manner. When the disturbance decoupling is sufficiently accurate, the VSC voltage command vector becomes approximately the applied voltage ($\mathbf{v}_\sigma^* = \mathbf{v}_\sigma$). Analyses on the case of imperfect load voltage decoupling (disturbance rejection) characteristics are elaborated in the subsequent paper.

The synchronous frame to/from stationary frame transformation is performed via complex vector theory [32] or equivalently (inverse) Park and Clarke transformations [30], [31] for CSCC systems with linear symmetrical circuit parameters. In the case of two-phase modelling presented here, the synchronous frame to/from stationary frame transformations are distinctly performed for $\mathbf{i}_\sigma$ and $\mathbf{v}_\sigma$. Considering that phase-c current can be written in terms of phase-a and phase-b currents in a 3P3W system and that the line-to-line voltages can be expressed by line to neutral voltages, the synchronous and stationary currents and voltages can be related by

$$\begin{aligned} \mathbf{i}_\sigma &= \mathbf{T}(\theta) \mathbf{i}_{dq} \\ \mathbf{v}_\sigma &= \mathbf{H}(\theta) \mathbf{v}_{dq} \end{aligned} \quad (17)$$

wherein $\mathbf{T}(\theta) = \begin{bmatrix} \cos(\theta) & -\sin(\theta) \\ \cos(\theta - 2\pi/3) & -\sin(\theta - 2\pi/3) \end{bmatrix}$ and $\mathbf{H}(\theta) = \mathbf{T}(\theta) - \mathbf{T}(\theta - 2\pi/3)$.

Based on (16) and (17), two-phase column vector block diagram representation of 3P3W VSC conventional closed-loop current control system is illustrated in Fig. 6. According to (6) and associated transformations described by (17), the closed-loop current control dynamics can be described in synchronous frame by

$$\begin{aligned} \mathbf{H}(\theta) \{ K_p (\mathbf{i}_{dq}^* - \mathbf{i}_{dq}) + K_i \int_0^t (\mathbf{i}_{dq}^* - \mathbf{i}_{dq}) d\tau + \hat{\mathbf{W}}_{cc} \mathbf{i}_{dq} \} \\ = \mathbf{L}_\sigma \frac{d}{dt} \{ \mathbf{T}(\theta) \mathbf{i}_{dq} \} + \mathbf{R}_\sigma \mathbf{T}(\theta) \mathbf{i}_{dq}. \end{aligned} \quad (18)$$

Any zero sequence injection voltage signal appears neither in equation (18) nor the nonlinear control system in Fig. 6 as the 3P3W system dynamics are not influenced by the zero-sequence voltage whether the system is linear or nonlinear.

D. Problems of CSCC of VSC Systems with Saturable Inductors

The employment of conventional vector control in 3P3W VSC systems with saturable inductors brings serious issues regarding the current control performance. This is because of that the CSCC is inherently developed for linear symmetrical 3P3W VSC systems, however, the 3P3W VSC systems with saturable inductors are unsymmetrical and nonlinear. These issues are addressed in the following.

1) *Cross-Coupling Decoupling*: It is common practice in CSCC systems with linear and symmetric plant parameters that the cross-coupling terms of d- and q-axis elements are decoupled via simple feedforwarding of the calculated cross-coupling terms via $\hat{\mathbf{W}}_{cc}$ as shown in Fig. 6. The elements of this matrix consist of constant a prior estimated inductances multiplied by the fundamental frequency ($\omega \hat{L}$ and $-\omega \hat{L}$). However, as the control system becomes nonlinear with the utilization of saturable inductors, the cross-coupling decoupling becomes erroneous due to the correspondence

$$\mathbf{H}(\theta) \hat{\mathbf{W}}_{cc} = \omega \mathbf{L}_\sigma \mathbf{P}(\theta) \quad (19)$$

is not met where $\mathbf{P}(\theta) = \mathbf{T}(\theta + \pi/2)$ due to derivative operator of the RHS of (18). The error in (19) depends on the depth of inductor saturation because of the time-invariant (constant) constituent of the cross-coupling decoupling matrix of the CSCC system ($\hat{\mathbf{W}}_{cc}$) and the time-varying constituent of the physical plant inductances ($\mathbf{L}_\sigma$). Therefore, the difference between the left and right sides of (19) appears as a voltage error on the closed-loop current-controlled plant and is zero when the system is linear, balanced, and the estimated inductances in $\hat{\mathbf{W}}_{cc}$ are the same as actual inductances. The decoupling with large errors on the d- and q-axis can adversely affect the waveform quality and independent current control of each axis.

2) *Bandwidth Shrinkage*: The bandwidth of a system is defined as the frequency range that the magnitude of the closed-loop does not drop below -3 dB [35]. Accordingly, a higher bandwidth implies a better ability to reproduce the input signal in a faster manner. However, the bandwidth of a system is confined by physical constraints. Hence, it cannot be increased infinitely. For power electronics systems, carrier frequency, controller speed, and noise rejection requirements set upper bound for achievable current control bandwidth [11], [12], [24], [32]. Via parameters of the compensators, current control bandwidth can be adjusted. For 3P3W VSC systems with acceptably linear inductors, it is common to set $K_p = \omega_{BW} \hat{L}$ [32]. In the case of the employment of saturable inductors, the predefined current control bandwidth can not be preserved with linear PI controllers solely as the plant parameters change.

Assuming a perfect cross-coupling decoupling is achieved and the fundamental frequency is much smaller than the investigated system dynamics (bandwidth), the closed-loop system representation of (18) can be arranged in s-domain as

$$\mathbf{i}_{dq}(s) = \mathbf{G}(s) \mathbf{i}_{dq}^*(s) \quad (20)$$

wherein the command and output vectors are related by $\mathbf{G}(s) = \{ (s\mathbf{L}_\sigma + \mathbf{R}_\sigma) \mathbf{T}(\theta) + \mathbf{H}(\theta) \mathbf{K}_c(s) \}^{-1} \mathbf{H}(\theta) \mathbf{K}_c(s)$ where $\mathbf{K}_c(s) = \begin{bmatrix} K_p + K_i/s & 0 \\ 0 & K_p + K_i/s \end{bmatrix}$. The structure of $\mathbf{G}(s)$ describes a two-input two-output system which have a bandwidth matrix denoted by

$$\omega_{\mathbf{BW}} = \begin{bmatrix} \omega_{BW-dd} & \omega_{BW-dq} \\ \omega_{BW-qd} & \omega_{BW-qq} \end{bmatrix} \quad (21)$$

where the elements are the command to output single-input single-output subsystem bandwidths (or equivalently cutoff frequencies).

When the VSC system is a linear and symmetrical one, the d- and q-axis bandwidths (self bandwidths, ω_{BW-dd} and ω_{BW-qq} respectively) of $\omega_{\mathbf{BW}}$ remains constant at their preset design values with conventional synchronous frame current control. Fig. 7(a) illustrates the spatial variation of the d- and q-axis bandwidths based on (20) for a linear and symmetrical inductance case of 450 μH with a design bandwidth of 500 Hz ($K_p = 1.41$ according to [32]). The d- and q-axis bandwidths of this design remain at their preset value as intended.

On the other hand, the spatial variation of the d- and q-axis current control bandwidths are displacement angle (θ_0) and time-dependent. In Fig. 7(b), these bandwidths for a design employing saturable inductors with an inductance characteristic decreasing from 4.5 mH to 450 μH linearly with zero to full load current are illustrated ($K_p = 0.141$ as if the VSC system has linear inductors with an inductance of 450 μH per phase and the design bandwidth is 500 Hz). In the figure, two displacement angles corresponding to unity power factor operation and fully reactive power injection cases are examined ($\theta_0 = 0$ and $\theta_0 = \pi/2$) at rated current. When the inductors have such nonlinear characteristics the higher current of an axis implies higher bandwidth and the other axis has a

Fig. 6. 3P3W VSC conventional closed-loop current control system in two-phase column vector block diagram representation.

lower bandwidth. It should be noted that the maximum design bandwidth is not reached as the instantaneous inductances of each of the three phases are distinct with distinct instantaneous currents.

Fig. 7. Spatial variation of closed-loop current control bandwidths of d- and q-axis with CSCC. (a) Linear inductor case. (b) Saturable inductor case.

3) *Waveform Quality*: In VSC applications, high quality of the load current waveform is essential due to detrimental effects of the harmonic currents such as excessive energy losses, power rating derating, grid voltage distortion, and electromagnetic interference [24], [36]. Accordingly, well-known grid codes confine the amount of individual harmonic current injection to the utility grid such as IEEE 1547, IEEE 519, IEC 61727, and EN 50160 [3], [10], [36], [37].

In the utilization of saturable inductors with the CSCC in 3P3W VSCs, the inductance of each phase varies with the associated phase current, yielding a nonlinear and time-varying system. In contrast to linear systems, the steady-state response of such nonlinear systems may involve low-order harmonics in addition to the fundamental frequency component even if the system is excited by single-frequency voltage sources [10], [11], [14]. In Fig. 8, the open-loop response of a 3P3W VSC system employing saturable inductors with the linearly decreasing L-i characteristics described above is investigated. Even the voltage sources in Fig. 8(a) supply pure sinusoidal 50 Hz voltage to the system, the phase currents are not pure sines in the steady-state as shown in Fig. 8(b). The harmonic content of the line currents involve 5th and 7th harmonics dominantly. Besides, zero-order harmonics such as 3rd and 9th are obsolete in the spectrum due to the lack of a path for zero-sequence current to flow. Since the harmonics caused by the nonlinearity of the system are much smaller than the PWM carrier frequency, control-based solutions are reasonable

to apply as presented in the following section.

Fig. 8. Spatial variation of closed-loop current control bandwidths of d- and q-axis with CSCC. (a) Simulated 3P3W circuit. (b) Time variation of applied voltages, resultant phase currents, and the phase inductances. (c) Harmonic spectrum of one of the phase currents.

IV. INDUCTOR SATURATION COMPENSATION IN 3P3W VSC SYSTEMS

An IDM-based compensation method is presented in this section to solve the aforementioned issues of the utilization of saturable inductors in 3P3W VSC systems with CSCC. Accordingly, the system is linearized in the large via IDM at first. An integral compensator is then incorporated to yield an equivalent linear inductance system. Finally, the overall current control of the linearized system with adaptation to the synchronous frame is described.

A. Linearization in the Large via Inverse Dynamic Model

Having the exact system model for 3P3W VSC systems in Section III-C, the inverse dynamic model of the system is obtained. Fig. 9(a) depicts the nonlinear physical system and its IDM in cascade. In the figure, the load voltage and known disturbances, such as the dead-time effect, are assumed to be decoupled. Since the system is nonlinear, in this representation, the order of operators is important (unlike linear systems). With the nonlinear time-domain structure, one can write the equivalence of nonlinear state equations of the cascaded system as

$$\mathbf{L}_\sigma \frac{d}{dt} \mathbf{i}_\sigma + \mathbf{R}_\sigma \mathbf{i}_\sigma = \hat{\mathbf{L}}_\sigma \frac{d}{dt} \mathbf{u}_\sigma + \hat{\mathbf{R}}_\sigma \mathbf{u}_\sigma \quad (22)$$

Fig. 9. Linearization of nonlinear 3P3W L-R system. (a) Time-domain vector representation of nonlinear L-R physical system and its IDM, (b) the simplified equivalent linear system.

where $\hat{L}_{\sigma}$ and $\hat{R}_{\sigma}$ are estimated values of L_{σ} and R_{σ} respectively. 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. 9(b).

B. Incorporation of Integral Compensator

The unity-gain zero-phase equivalent system obtained in the previous subsection and shown in Fig. 9 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 $1/L_{min}$ is cascaded to the IDM as shown in Fig. 10(a). The integration term can be spread to the IDM and cancel the differentiation block as shown in Fig. 10(b), exterminating the noise amplification problem. Resultantly, the compensated system behaves as a two-input two-output decoupled linear system as shown in Fig. 10(c).

Fig. 10. Incorporation of integral compensator to the linearized 3P3W system. (a) Cascading of the integration term (with a coefficient of $1/L_{min}$) to the linearized unity-gain zero-phase 3P3W system, (b) the equivalent model when integration is reflected to IDM (c) the simplified equivalent linear system in s-domain.

C. Current Control via IDMB

The linearized overall system shown in Fig. 10(c) is nothing but two independent phases having the dynamics of

$$\begin{bmatrix} w_a \\ w_b \end{bmatrix} = \begin{bmatrix} L_{min} & 0 \\ 0 & L_{min} \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_a \\ i_b \end{bmatrix} \quad (23)$$

where i_a and i_b are the actual phase currents of the physical system and w_a and w_b are the fictitious input voltages. Apparently, this system represents two phases of a hypothetical balanced 3P3W VSC circuit such as described by (10) which is used in the development of the CSCC. Adopting this approach to the linearized system of (23), the CSCC or other types of linear controllers such as stationary frame controllers can be employed in the current regulation.

Having obtained the two-input two-output linearized equivalent system described by equation (23), the overall closed-loop current control scheme is illustrated in Fig. 11. The controller in the figure is a conventional synchronous frame PI type. However, it could have been in the stationary frame resonant type [38], each providing sufficiently small steady-state error at the fundamental frequency. The IDMB-based current control shown in the figure operates as follows.

The measured current vector $i_{\sigma} = [i_a \ i_b]^T$ is fed back to the synchronous frame conversion to its synchronous frame column vector representation, wherein only the phase-a and phase-b quantities are sufficient for the conversion as the phase-c current can be derived from the phase-a and phase-b currents ($i_c = -i_a - i_b$). The error between the synchronous frame current command and the measured current ($e_{dq} = i_{dq}^* - i_{dq}$) is processed by the PI type compensator. The feed-forward term $\hat{W}_{cc} i_{dq}$ is added to decouple the cross-coupling of the d- and q-axis yielding the synchronous frame voltage output v_{dq} for the linearized system wherein $\hat{W}_{cc}$ is formed based on L_{min} . This output (v_{dq}) is converted to stationary frame three-phase quantities via Park and Clarke transformations [30], [31] with the outputs are denoted by w_a , w_b , and w_c . However, only two of these outputs $w_{\sigma} = [w_a \ w_b]^T$ are utilized in the control architecture as the linearized plant is a two-input two-output decoupled one as described by (23). Such an approach functions quite well considering the argument that the system (23) can be expanded to that of a linear, balanced, and decoupled system of equation (10) by inserting a third hypothetical equation relating to phase-c. However, phase-c current is strictly dependent on phase-a and phase-b currents. Therefore, phase-c current is controlled by controlling phase-a and phase-b currents automatically. Accordingly, omitting the calculations for the phase-c, the reduced form of the Park and Clarke transformations, denoted as $T(\theta)$, is employed.

When w_{σ} column vector is applied, the output of the unified IDM block becomes the column vector command $v_{\sigma}^* = [v_{ab}^* \ v_{bc}^*]^T$ that is the input for the nonlinear physical system line-to-line voltages after the modulation process "M". These line-to-line output voltages of the VSC is provided via phase voltages as the state-of-the-art 3P3W VSC technology is built on the independent modulation of each leg with high-performance PWM techniques with zero-sequence injection [2], even if two-phase modulation is also possible

Fig. 11. Three-phase three-wire nonlinear VSC system linearization and control in synchronous frame.

[39]. Accordingly, the modulator block in Fig. 11 converts the line-to-line output voltages to three-phase voltages while being equivalent to 2×2 unity matrix “ $\mathbf{T}$ ” for dynamic system representation.

Fig. 12. Three-phase three-wire system modulator block. (a) Column vector representation. (b) Scalar representation.

Fig. 12 details the function of the modulator. Fig. 12(a) shows the modulator block diagram in a vector input vector output manner, whereas, in Fig. 12(b), the illustration of the modulator is performed via scalars. Regarding Fig. 12(b), the line-to-line voltages provided by the unified integral IDM is converted to phase voltages (v_a^* , v_b^* , and v_c^*) via scalar multiplication by the matrix $\mathbf{K}$. The matrix $\mathbf{K}$ can be deduced from the argument that the phase voltages can be derived from the line-to-line voltages as

$$\begin{aligned} v_{ab}^* + v_{ac}^* &= v_a^* - v_b^* + v_a^* - v_c^* = 3v_a^* \\ v_{ba}^* + v_{bc}^* &= v_b^* - v_a^* + v_b^* - v_c^* = 3v_b^* \\ v_{ca}^* + v_{cb}^* &= v_c^* - v_a^* + v_c^* - v_b^* = 3v_c^* \end{aligned} \quad (24)$$

where v_a^* , v_b^* , and v_c^* are the zero-sum phase voltage commands ($v_a^* + v_b^* + v_c^* = 0$) with respect to the midpoint of the dc-link with no zero-sequence component. Therefore the following relation can be obtained from (24) as

$$\begin{aligned} v_a^* &= \frac{v_{ab}^* + v_{ac}^*}{3} \\ v_b^* &= \frac{v_{ba}^* + v_{bc}^*}{3} \\ v_c^* &= \frac{v_{ca}^* + v_{cb}^*}{3} \end{aligned} \quad (25)$$

Rearranging (25) and noting that $v_{ab}^* = -v_{ba}^*$, $v_{bc}^* = -v_{cb}^*$, $v_{ca}^* = -v_{ac}^*$ and that $v_{ab}^* + v_{bc}^* + v_{ca}^* = 0$, one can obtain the relation

$$\begin{bmatrix} v_a^* \\ v_b^* \\ v_c^* \end{bmatrix} = \mathbf{K} \begin{bmatrix} v_{ab}^* \\ v_{bc}^* \end{bmatrix} \quad (26)$$

$$\text{where } \mathbf{K} = \frac{1}{3} \begin{bmatrix} 2 & 1 \\ -1 & 1 \\ -1 & -2 \end{bmatrix}.$$

After obtaining phase command voltages with no zero-sequence component, the zero-sequence voltage can be injected in order to improve the overall performance of the VSC. It can be noted that even if the 3P3W VSC system is a nonlinear one, the injection of zero-sequence voltage (v_0) will not create any zero-sequence current as there is no path for the zero-sequence current to flow.

The generation of the voltage commands upto the triangle comparators (v_a^{**} , v_b^{**} , and v_c^{**}) can be summarized as

$$\begin{bmatrix} v_a^{**} \\ v_b^{**} \\ v_c^{**} \end{bmatrix} = v_0 + \mathbf{K} \begin{bmatrix} v_{ab}^* \\ v_{bc}^* \end{bmatrix}. \quad (27)$$

The injection of the zero-sequence voltage can be made among various PWM methods [2], [40] depending on the performance criteria. Having obtained the zero-sequence signal injected voltage commands for the inverter legs, one can generate the pulse patterns and hence the VSC output voltage for each leg with respect to the midpoint of the dc-link $\mathbf{v}_{abc} = [v_a \ v_b \ v_c]^T$. Although these pole voltages are with respect to the midpoint of the dc-link, the 3P3W load sees line-to-line voltages ($\mathbf{v}_\sigma^*$), and the current is controlled accordingly.

V. DISCUSSION

The nonlinear dynamics of 3P3W VSC current regulation systems with saturable inductors in Fig. 3 can be described by two-phase representation as in (16) which was formerly elaborated in [13]. Based on this representation, the CSCC becomes as shown in Fig. 6 wherein the current/voltage stationary/synchronous frame conversion matrices differ from each other due to phase and line-to-line representation of these variables respectively. Via the two-phase modelling, the cross-coupling, bandwidth shrinkage, and waveform quality issues are indicated and an IDMB method is proposed to cope with these issues. In the method, the nonlinear plant dynamics are converted to a linear one via the inverse dynamic model. The error compensation in the method is performed in the

synchronous frame via PI controllers as in the case of the CSCC method.

The proposed IDMBBC method has a complexity and computational burden similar to the CSCC method. Hence a wide utilization can be expected. However, the IDMBBC method requires system L-R parameters to establish the inverse dynamic model of the system. When powder core materials are employed, L-i characteristics become almost aging and temperature-independent and can be embedded into a digital signal processing platform in the form of a polynomial of current (as performed in the subsequent part of this study) or a look-up table. On the other hand, when other types of core materials are employed with temperature- and aging-dependent inductance characteristics, online and robust inductance estimation methods seem to be needed.

VI. CONCLUSION

An inverse dynamic model based compensation (IDMBBC) method for 3P3W VSC systems employing saturable inductors is presented. In the first place, saturable inductors are reviewed by means of theory, core types, and motivation on their utilization. The conventional synchronous frame current control (CSCC) method is then provided for linear inductors. After that, the exact nonlinear two-phase system modelling is performed and the inconsistency of the conventional approach is discussed. Furthermore, the dynamic performance and waveform quality degradation issues of the CSCC method are indicated analytically. Based on the exact modelling, the IDMBBC method is developed to yield a fictitious linear inductor load for current control. In a subsequent paper, it is shown that by means of analyses, simulations, and experiments that this fictitious linear inductor system current can be controlled with uniform bandwidth and high steady-state waveform quality.

Considering the wide-spread and large-volume applications of the current-controlled 3P3W VSCs, the cost, size, and efficiency improvements that can be obtained by the utilization of saturable inductors are fairly significant. Therefore, this paper is expected to improve better utilization of world resources.

VII. APPENDIX

A. Discussion on the Two-Phase Current and Voltage Transformations

The relation (17) describes synchronous frame to stationary two-phase frame current and voltage transformations via $\mathbf{T}(\theta)$ and $\mathbf{H}(\theta)$ respectively. The current transformation $\mathbf{T}(\theta)$ can be derived from the space vector to phase variable relations

$$\begin{aligned} i_a &= \text{Re}\{i_{dq}e^{j\theta}\} \\ i_b &= \text{Re}\{i_{dq}a^2e^{j\theta}\} \\ i_c &= \text{Re}\{i_{dq}ae^{j\theta}\} \end{aligned} \quad (28)$$

wherein only the phase-a and phase-b relations are involved in the transformation to $\mathbf{i}_s$. Similarly, the two-phase voltage transformation $\mathbf{H}(\theta)$ can be obtained from the relation between line-to-line voltages and the space vector as

$$\begin{aligned} v_{ab} &= \text{Re}\{v_{dq}e^{j\theta}\} - \text{Re}\{v_{dq}a^2e^{j\theta}\} \\ v_{bc} &= \text{Re}\{v_{dq}a^2e^{j\theta}\} - \text{Re}\{v_{dq}ae^{j\theta}\} \\ v_{ca} &= \text{Re}\{v_{dq}ae^{j\theta}\} - \text{Re}\{v_{dq}e^{j\theta}\} \end{aligned} \quad (29)$$

wherein only the first two equations are essential to obtain line-to-line voltage vector $\mathbf{v}_s$. Accordingly, the equivalence $\mathbf{H}(\theta) = \mathbf{T}(\theta) - \mathbf{T}(\theta - 2\pi/3)$ holds.

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