Closed-Loop Control Design for a Three-Level Three-Phase Neutral-Point-Clamped Inverter Using the Optimized Nearest-Three Virtual-Space-Vector Modulation

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Abstract—This paper presents a closed-loop control scheme for the three-level three-phase neutral-point-clamped dc-ac converter using the optimized nearest-three virtual-space-vector pulselwidth modulation, a modulation that produces a low output voltage distortion with a significant reduction of the dc-link capacitance. A specific loop modifying the modulating waveforms is introduced to rapidly control possible perturbations in the neutral-point voltage balance. The remaining part of the control is analogous to the control for a two-level converter with an appropriate interfacing to the selected modulation, including an online estimation of the load displacement angle at no extra cost. The closed-loop control is designed for the case of a renewable energy source connected to the ac mains and its performance is analyzed through simulation and experiments.

I. INTRODUCTION

Multilevel converter topologies [1]-[2] have received special attention during the last two decades due to their significant advantages in high-power medium-and-high-voltage applications. In these topologies, and compared to a two-level converter, the voltage across each semiconductor is reduced, avoiding the problems of the series interconnection of devices, reducing the harmonic distortion of the output voltage, and improving the efficiency. But a larger number of semiconductors are needed and the modulation strategy to control them becomes more complex.

Among these topologies, the three-level three-phase neutral-point-clamped (NPC) dc-ac converter [3], in Fig. 1(a), is probably the most popular. The application of conventional modulation techniques to this converter causes a low frequency (around three times the fundamental frequency of the output voltage, \( f_s \)) oscillation of the neutral-point voltage. This, in turn, increases the voltage stress on the devices and generates low order harmonics in the output voltage.

There have been many efforts to analyze this problem and define a modulation strategy to solve it [4]-[14], therefore eliminating the need to significantly increase the dc-side capacitance to minimize the voltage oscillation. Among them, the nearest-three virtual-space-vector (NTV\(^2\)) pulselwidth modulation (PWM) [13], allows controlling the neutral-point voltage over the full range of converter output voltage and for any load. The optimized nearest-three virtual-space-vector (ONTV\(^2\)) PWM [14] also allows comprehensively controlling the neutral-point voltage but with a lower output voltage harmonic distortion in the case of linear-and-balanced ac loads.

The design of a closed-loop control scheme interfacing the ONTV\(^2\) PWM and correcting possible dc-link capacitor voltage balance perturbations is not straightforward. This paper presents a proposal for such control, focusing on a particular application, and the performance of the proposed control scheme is verified through simulation and experiments.

II. OPTIMIZED NEAREST-THREE VIRTUAL-SPACE-VARIABLE PWM

Let us designate \( d_{nn}, d_{pq}, d_{np}, d_{nm}, d_{mnp}, d_{mnn} \) the six independent converter phase duty-ratios, where \( d_{nn} \) refers to the duty ratio of the phase \( x \) connection to the dc-link point \( y \). Equation (1) reproduces from [14] the six expressions that define the ONTV\(^2\) PWM in terms of these six independent converter phase duty-ratios in \( d-q-0 \) coordinates \((d_{pp}, d_{pq}, d_{pp}, d_{q0}, d_{q0}, d_{n0})\).

\[
\begin{align*}
\theta &< 2\pi/3: \quad d_{nn} = \sqrt{2} \cdot (d_{q0} \cdot \sin(\theta) - \sqrt{2} \cdot m) \\
\theta &< 2\pi/3: \quad d_{pq} = \sqrt{2} \cdot (d_{q0} \cdot \cos(\theta) + \sqrt{2} \cdot m) \\
\theta &< 2\pi/3: \quad d_{np} = \sqrt{2} \cdot (d_{q0} \cdot \cos(\theta) + \sqrt{2} \cdot m) \\
\theta &< 2\pi/3: \quad d_{nm} = \sqrt{2} \cdot (d_{q0} \cdot \cos(\theta) - \sqrt{2} \cdot m) \\
\theta &< 2\pi/3: \quad d_{mnp} = \sqrt{2} \cdot (d_{q0} \cdot \sin(\theta) - \sqrt{2} \cdot m) \\
\theta &< 2\pi/3: \quad d_{mnn} = \sqrt{2} \cdot (d_{q0} \cdot \sin(\theta) - \sqrt{2} \cdot m) \\
\end{align*}
\]

where \( m \in [0,1] \) and \( \theta \) are the length and angle (with reference to axis \( \alpha \), aligned with vector \( V_{d1} \) corresponding to switching state \( pmn \) [13]) of the reference vector, \( V_{ref} \), a rotating vector in the converter space vector plane that represents the desired fundamental converter output three-phase voltage. The expressions in (1) assume that axis \( d \) of the \( d-q-0 \) transformation of the phase duty-ratios is aligned with \( V_{ref} \). Variable \( \varphi \) is the ac load/source displacement angle and the optimum value of parameter \( K \) is a function of \( m \) and \( \varphi \). In [14], expressions are provided to compute the value of \( K \) as a function of \( m \) and \( \varphi \). Alternatively, a look-up table as a function of \( m \) and \( \tan(\varphi) \) can be generated to select the appropriate value of \( K \) on-line.

Given the values of \( m \), \( \theta \), and \( \tan(\varphi) \), the duty ratios in \( d-q-0 \) coordinates can be obtained from (1) and the look-up table or expressions to compute \( K \). Applying the inverse
The proposed modulator and controller have been designed and implemented for the particular system in Fig. 1(a). The purpose of the system is to send the energy from a renewable energy source to the mains with unity power factor while also regulating voltage and the energy is sent to the mains, such as, for example, a wind energy conversion system where the ac generator is connected to a non-controlled boost rectifier. The diagrams of Fig. 1(b) and Fig. 1(c) summarize the proposed controller and modulator structure, discussed in detail next.

### A. Neutral-Point Voltage Control

The ONTV\(^2\) PWM guarantees no low-frequency oscillations of \(v_{\text{unb}}\) due to the loading conditions of the converter provided that the addition of line currents equals zero. Even if the load presents a severe non-linearity, this will not affect the dc-link voltage balance if we set \(K = 0\). The occurrence of neutral-point voltage perturbations should however be considered. Perturbations can occur if, for example, there is a leakage current flowing from the load neutral to ground, causing that the addition of the three-phase currents be different from zero. The non-idealities, and specially the differences, in the switching behavior of the converter devices are another possible source of disturbances.

As discussed in [15], certain modulations have the property of naturally recovering the dc-link voltage balance after a perturbation. The ONTV\(^2\) PWM with \(K = 0\) does not belong to this family of modulations. It does not affect the balance of the dc-link capacitors. If an unbalance exists at a given point in time, the ONTV\(^2\) PWM with \(K = 0\) will preserve this unbalance (see Fig. 2(a)). However, the ONTV\(^2\) with \(K > 0\) does belong to the set of modulations that naturally recover the balance (see Fig. 2(b)). The higher the value of \(K\), the faster the system recovers the balance. Still, this natural recovery process is usually slow.

The addition of discharging resistors to the dc-link capacitors also helps recovering the balance after a perturbation. Their resistance value is usually high, though, and the recovery process thanks to these resistors is also slow.

Since all preexisting possible balance recovery processes do not seem to be effective/fast enough, an appropriate perturbation of the modulating waveforms that allowed speeding-up this process would be helpful. Reference [7] shows that introducing a common-mode voltage into all three line-to-neutral output-voltage waveforms causes an unbalance in the discharging of the dc-link capacitors. This property can be used to speed-up the recovery process whenever a dc-link voltage unbalance occurs. Here, this control mechanism is adapted to the ONTV\(^2\) PWM, leading to an alternative scheme to guarantee the dc-link capacitor voltage balance in the three-level three-phase NPC dc-ac converter different from other solutions presented [16].

The introduction of a common mode voltage can be done by adding an offset (\(d_{\text{offset}}\)) to all three \(d_{\text{ap}}-d_{\text{an}}, d_{\text{bp}}-d_{\text{bn}}, d_{\text{cp}}-d_{\text{cn}}\) modulating waveforms. If we want to add an offset to the \(d_{\text{ap}}-d_{\text{an}}\) waveform, we have three options:

1. We add the offset to \(d_{\text{ap}}\).
2. We subtract the offset from \(d_{\text{an}}\).
3. We add part of the offset to \(d_{\text{ap}}\) and we subtract the remaining part from \(d_{\text{an}}\).

A simple and interesting strategy is to apply all the offset to the duty ratio to be reduced. Since duty ratios must be greater than zero, in case we reach the value of zero, the other duty ratio will be increased an amount corresponding to the part of the offset still not applied. This strategy allows minimizing the number of commutations, since it will maximize the occasions where a non-zero duty-ratio becomes zero. With reference to Fig. 1(c), this strategy can be formulated for phase \(x\) as

\[
\begin{align*}
\text{if} \quad (d_{\text{offset}} \geq 0) \quad & \{ \quad \\
\text{if} \quad (d_{\text{ap}} > d_{\text{an}}) \quad & \{ \quad \\
\quad d_{\text{an}} = d_{\text{an}} - d_{\text{offset}} \\
\quad d_{\text{ap}} = d_{\text{ap}} \\
\} \quad \text{else} \quad \\
\quad d_{\text{an}} = 0 \\
\quad d_{\text{ap}} = d_{\text{ap}} + (d_{\text{offset}} - d_{\text{an}}) \quad (2)
\}
\end{align*}
\]

The value of \(d_{\text{offset}}\) to be applied is determined from the dc-link capacitor voltage unbalance \(v_{\text{unb}} = f_1(v_{C1}, v_{C2}) = (v_{C2} - v_{C1}) / 2\) by a compensator. This compensator must have a low-pass characteristic, in order to only react to perturbations in the dc-link voltage balance with frequencies lower than the switching frequency.
Fig. 1. System block diagram. (a) Power stage plus control. (b) Controller structure. (c) Modulator structure.
average model of the three-level converter becomes
and the dedicated control presented above. Then, the
neutral-point voltage. We can assume that the neutral-point
converter is that the latter introduces the dynamics of the
difference between a two-level and a three-level dc-ac
converter to obtain, from the sensed variables, the reference
vector length $\omega$

For the particular application considered here, the
selected control scheme is shown in Fig. 1(b). First, the dc-
link voltage $V_{\text{pm}} = \sqrt{3}/2 (V_{C1}, V_{C2}) = V_{C1} + V_{C2}$ is compared to the
desired command. The error is then processed by a
compensator to produce the $i_d$ command. Line currents $i_a$
and $i_b$ are sensed and d-q transformed. We do not need to
sense $i_c$ since we know $i_c = -i_a - i_b$. The transformation is
performed to axes $d_2-q_2-\theta_2$, where axis $d_2$ is in phase with the vector of line-to-neutral mains voltages ($V_{L-N}$). This
vector has an angle $\psi$ with reference to axis $\alpha$ (Fig. 3):

$$\psi = f_7(t) = \omega \cdot t - 2\pi/3 = 2\pi \cdot f_o - 2\pi/3.$$  

(3)

The $d_2$ and $q_2$ components of the current are compared to
their corresponding command. The $i_{\alpha2}$ command is zero to
achieve a unity displacement factor for the power
transferred to the mains. Both $d$ and $q$ channel errors are
processed by their specific compensators. Finally, both
channels are decoupled through $f_2$ and $f_3$:

$$f_2 (i_{q2}, V_{pm}) = \left(\sqrt{2} \cdot i_{q2} \cdot \omega_n \cdot L_2 \right)/V_{pm} \quad (4)$$

$$f_3 (i_{d2}, V_{pm}) = \left(\sqrt{2} \cdot i_{d2} \cdot \omega_n \cdot L_1 \right)/V_{pm} .$$

The outcome of both channels is the $d_2$ and $q_2$
components of $V_{\text{ref}}$. Through $f_2$ we obtain the $V_{\text{ref}}$ length $m$
(modulation index) and angle $\phi$ (with reference to axis $d_2$):

$$m = f_{41} \left( V_{\text{refd}} \cdot V_{\text{refq}} \right) = \sqrt{V_{\text{refd}}^2 + V_{\text{refq}}^2}$$

$$\phi = f_{42} \left( V_{\text{refd}} \cdot V_{\text{refq}} \right) = \tan^{-1} \left( V_{\text{refq}} / V_{\text{refd}} \right).$$  

(5)

Angle $\theta$ can then be obtained by simply adding $\psi$ and $\phi$
(Fig. 3). This is the angle that will be used to perform the
transformation of the duty ratios from $d_1-q_1-\theta_1$ to $a-b-c$
coordinates within the modulator.

D. Online Estimation of $\tan(\phi)$

To implement the ONTV$^2$ PWM we need an estimate of $\tan(\phi)$. Angle $\phi$ corresponds to the angle between the vector
of fundamental line currents and $V_{\alpha}$ref. Hence, the value of
$\tan(\phi)$ can be computed online by simply sensing the line
currents, applying the $d_1-q_1$ transformation, and using (6) (in
the case of non linear-and-balanced loads, the dc values of
the $d_1$ and $q_1$ components can be used). Since, in general,
the controller already requires sensing the line currents, the
implementation of the ONTV$^2$ PWM does not require
additional sensors.

$$\tan(\phi) = f_5 (i_d, i_q) = -i_q / i_d.$$  

(6)

For the particular application considered here, the
selected control scheme is shown in Fig. 1(b). First, the dc-
link voltage $V_{\text{pm}} = \sqrt{3}/2 (V_{C1}, V_{C2}) = V_{C1} + V_{C2}$ is compared to the
desired command. The error is then processed by a
compensator to produce the $i_d$ command. Line currents $i_a$
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sense $i_c$ since we know $i_c = -i_a - i_b$. The transformation is
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vector has an angle $\psi$ with reference to axis $\alpha$ (Fig. 3):

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their corresponding command. The $i_{\alpha2}$ command is zero to
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transferred to the mains. Both $d$ and $q$ channel errors are
processed by their specific compensators. Finally, both
channels are decoupled through $f_2$ and $f_3$:

$$f_2 (i_{q2}, V_{pm}) = \left(\sqrt{2} \cdot i_{q2} \cdot \omega_n \cdot L_2 \right)/V_{pm} \quad (4)$$

$$f_3 (i_{d2}, V_{pm}) = \left(\sqrt{2} \cdot i_{d2} \cdot \omega_n \cdot L_1 \right)/V_{pm} .$$

The outcome of both channels is the $d_2$ and $q_2$
components of $V_{\text{ref}}$. Through $f_2$ we obtain the $V_{\text{ref}}$ length $m$
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Angle $\theta$ can then be obtained by simply adding $\psi$ and $\phi$
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the $d_1$ and $q_1$ components can be used). Since, in general,
the controller already requires sensing the line currents, the
implementation of the ONTV$^2$ PWM does not require
additional sensors.

$$\tan(\phi) = f_5 (i_d, i_q) = -i_q / i_d.$$  

(6)
IV. SIMULATION AND EXPERIMENTAL RESULTS

Simulations have been carried out in Simulink with closed-loop switching and average models of the system. For the experimental validation, the controller and modulator blocks have been implemented using a PowerPC (dSPACE 1103) and the distributor block using an FPGA (Altera EPF10K70).

Fig. 4(a) demonstrates the advantage of introducing the neutral-point voltage control. The simulation has been performed in the same conditions as in Fig. 2(a). Fig. 4(b) shows the perturbed duty-ratio pattern during the transient. Fig. 5 depicts the good performance achieved with this control in the experiments. In Fig. 5(a), the control effort is negligible: $d_{\text{offset}} \approx -0.001$. In Fig. 5(b), $d_{\text{offset}} \approx 0.01$.

The complete control has been tested through simulation in the conditions of Fig. 6. These results have been obtained with the complete closed-loop switching model for a step in voltage command $v^*$, from 800 V to 750 V. We can observe that there is no unbalance of the dc-link capacitor voltages before, during, and after the step transient, as expected (Fig. 6(a)). Since the ONTV$^2$ PWM already guarantees this balancing, the effort of the dedicated control is minimal ($-0.001 < d_{\text{offset}} < 0.001$). In Fig. 6(b) we can observe that the power conversion is achieved with unity displacement factor, as intended.

Fig. 7 shows the experimental results obtained with an emulator of a wind energy conversion system connected to the ac mains (Fig. 1(a)). The voltage of a wind mill generator is connected to the inverter dc-link through a set of three line inductances ($L_s$) and a three-phase diode rectifier. A constant torque of 10 Nm is applied to the generator shaft to emulate the wind torque. A step in $v^*$ from 340 V to 300 V is forced, producing a variation of the neutral-point voltage control. The simulation has been performed in the same conditions as in Fig. 2(a). Fig 4(b) depicts the good performance achieved with this control effort is negligible: $d_{\text{offset}} \approx -0.001$. In Fig. 5(b), $d_{\text{offset}} \approx 0.01$.

The complete control has been tested through simulation and experiments. In Fig. 5(a), the control effort is negligible: $d_{\text{offset}} \approx -0.001$. In Fig. 5(b), $d_{\text{offset}} \approx 0.01$.

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The results show an overall good performance of the complete control, which guarantees the dc-link capacitor voltage balance and unity displacement factor during the entire transient.

$$H_1(s) = -\frac{2\left(s + 2\pi \cdot 0.01\right)}{s \cdot (s + 2\pi \cdot 25)}$$
$$H_2(s) = -\frac{1000\left(s + 2\pi \cdot 5\right)}{s \cdot (s + 2\pi \cdot 2500)}$$
$$H_3(s) = \frac{500 \cdot (s + 2\pi \cdot 25)}{s \cdot (s + 2\pi \cdot 2500)}$$
$$H_4(s) = \frac{1000 \cdot (s + 2\pi \cdot 5)}{s \cdot (s + 2\pi \cdot 2500)}$$

$$d_{\text{offset min}} = -0.1, \quad d_{\text{offset max}} = 0.1$$
$$[i_{d2 \text{ min}}, i_{d2 \text{ max}}] = [-10, 10]$$
$$[m_{\text{min}}, m_{\text{max}}] = [0, 1]$$

V. CONCLUSIONS

A closed-loop control scheme for the three-level three-phase NPC dc-ac converter using the ONTV$^2$ PWM has been presented. The selected modulation allows using small dc-link capacitors leading to an improved performance of the closed-loop system. A specific control loop has been designed to speed-up the recovery from neutral-point voltage perturbations. The remaining part of the control is analogous to the control for a two-level converter, with an appropriate interfacing to the selected modulation, including an online estimation of the load angle at no extra cost. The good performance of the proposed control has been verified through simulation and experiments.

REFERENCES

Fig. 4. Dc-link capacitor voltage balance recovery transient in the same conditions as in Fig. 2(a) but with the dedicated neutral-point voltage control activated. (a) Dc-link voltages $v_{C1}$ and $v_{C2}$. (b) Phase $a$ independent duty-ratios.

Fig. 5. Neutral-point voltage control performance in the following conditions: ONTV2 PWM, $K = 0$, $V_{pn} = 140$ V, $m = 0.75$, $f_o = 50$ Hz, $f_s = 5$ kHz, $C_1 = C_2 = 1.1$ mF, $R_L = 16.5$ $\Omega$, and $L_s = 5$ mH. (a) Control tuned to achieve $v_{unb} = 0$ V. (b) Control tuned to achieve $v_{unb} = -5$ V.

Fig. 6. Simulation results for a step in command $v_{pn}^*$ at time = 0.1 s in the following conditions: $I = 12.5$ A, $C_1 = C_2 = 400$ $\mu$F, $L_s = 5$ mH, $V_{aN} = V_{bN} = V_{cN} = 230$ Vrms, $f_o = 50$ Hz, and $f_s = 5$ kHz. (a) Dc-link voltages $v_{C1}$ and $v_{C2}$. (b) Mains voltage $v_{aN}$ and line current $i_a$. 

Fig. 7. Experimental results for a step in command \( v^*_{pn} \) from 340 V to 300 V (variation of rotor speed from 1577 rpm to 1396 rpm). Conditions: Rectified wind mill generator voltage, constant torque applied to the generator shaft = 10 Nm, \( L_s = 12.5 \) mH, \( C_1 = C_2 = 1.1 \) mF, \( L_L = 10 \) mH, \( V_{aN} = V_{bN} = V_{cN} = 75 \) Vrms, \( f_o = 50 \) Hz, and \( f_s = 5 \) kHz. (a) Dc-link voltages \( v_{C1} \) and \( v_{C2} \) [25 V/div]. (b) Generator line-to-line voltages \( v_{uv} \), \( v_{vw} \), and \( v_{wu} \) [150 V/div]. (c) Generator currents \( i_u \), \( i_v \), and \( i_w \) [2 A/div]. (d) \( i_a \), \( i_b \), and \( i_c \) [3 A/div]. (e) Mains voltage \( v_{aN} \) [30 V/div] and line current \( i_a \) [5 A/div].