ELEMINATION OF LOW-FREQUENCY OSCILLATIONS IN THE NEUTRAL-POINT-CLAMPED CONVERTER HYBRID MODULATION STRATEGIES

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ABSTRACT--Nearest vector (NV) modulation strategies for the neutral-point-clamped converter are known to generate low frequency neutral point (NP) voltage oscillations. Non-NV strategies can eliminate these oscillations, but at the expense of higher switching losses and output voltage harmonic distortion. This letter proposes a simple way of creating hybrid strategies, as combinations of NV and non-NV strategies, which are also able to eliminate NP voltage oscillations. The approach minimizes the participation of non-NV strategies and hence their drawbacks, while it can be applied to any type of load (nonlinear and/or unbalanced). Simulations in MATLAB–Simulink are used to illustrate its operation.

I. INTRODUCTION

A well-known problem of the three-level neutral-point-clamped (NPC) converter [1] is the appearance of low frequency oscillations at the neutral point (NP) voltage $V_{np}$ (see Fig. 1). These oscillations, also known as NP voltage ripple, impose voltage stress on the converter’s dc-link capacitors and switching modules. Moreover, they require compensation by the converter modulation strategy, to prevent them from injecting low-frequency harmonics to the output [pulse width modulated (PWM)] voltage [2].

Traditionally, the modulation strategies for the NPC converter are categorized as nearest vector (NV) or non-NV strategies [3] (see Fig. 2). Examples of NV strategies are the nearest three vector (NTV) and Symmetric modulation strategies, described in [4], as well as the strategies described in [5]–[8], whereas examples of non-NV strategies are the NTV2 strategy [9] and others [10]–[12]. As shown in [13], NV strategies can only eliminate NP voltage ripple for a certain range of values of load power angle $\phi$ and converter modulation index $m$. Non-NV strategies, on the other hand, can achieve NP voltage ripple elimination throughout the converter’s operating range (i.e., for all values of $\phi$ and $m$). However, the use of non-NVs introduces additional switching steps which increase the converter’s switching frequency. In [9] and [10], two more switching steps are added per cycle; therefore, the switching frequency is increased by a third compared to NV strategies (which typically operate with six steps per cycle). Non-NVs also have a negative impact on the quality of the NPC converter’s PWM voltage waveform. Namely, during each switching period, they cause one of the converter legs to switch between $V_{dc}/2$, (possibly) $V_{np}$, and $-V_{dc}/2$. This approximates the operation of a two-level converter, therefore distorting the standard three-level PWM phase voltage waveform and increasing its harmonic distortion [total harmonic distortion (THD) and weighted total harmonic distortion (WTHD)].

The tradeoff between the NP voltage ripple produced by NV strategies on the one hand, and the increment of switching losses and output voltage harmonic distortion caused by non-NV strategies on the other, gave birth to hybrid strategies, which operate as combinations of the two. In [14], the well-known sinusoidal PWM (which is an NV) strategy is combined with a non-NV strategy, implemented as a carrier-based PWM with two carrier waveforms. A variable $D$ determines the fraction of the fundamental cycle where sinusoidal pulse width modulation modulates the converter. It is important to note that, unless $D$ is equal to zero, NP voltage ripple appears at the dc-link.

A different approach for creating a hybrid strategy can be found in [15]. There, the NTV strategy (named N3V in
with a proposed non-NV strategy, named S3V. S3V is characterized by using three vectors during each switching cycle, among which, one is non nearest. A threshold \( V_{\text{NP}, \text{max}} \) for the NP voltage is used to determine when S3V should be put into action (\( |V_{\text{NP}}| > V_{\text{NP}, \text{max}} \)), thus avoiding further NP voltage deviation (caused by the NTV). This approach also generates NP voltage ripple with amplitude \( V_{\text{NP}, \text{max}} \). Moreover, in case that \( V_{\text{NP}, \text{max}} \) is set close to zero with the aim of eliminating NP voltage ripple, successive transitions appear from the NV to the non-NV strategy and vice versa. As it will be shown later (in Section IV), this effect is undesirable since it can increase the converter’s switching losses.

Finally, in [16], a strategy that mitigates the drawbacks of the NTV2 [9] is described. The strategy is classified here as hybrid, because the line voltage (PWM) waveforms in [16] indicate that it can partly operate as an NV strategy. Its formulation is based on a WTHD minimization process, which should be performed offline for nonlinear or unbalanced loads. For the case of linear and balanced loads, the results of the previous process can be approximated by analytical equations. However, as explained in [17], an online estimator of the load power angle, as well as a detector of the linear and balanced nature of the load is still required.

This letter proposes a straightforward way of creating hybrid strategies for the NPC converter, which have the following characteristics:

1) Eliminate NP voltage ripple;
2) can operate with nonlinear or unbalanced loads;
3) can be built as combinations of any NV and non-NV strategy;
4) use the non-NV strategy to the minimum possible extent, thus minimizing the converter’s switching losses and output voltage harmonic distortion for the selected strategy combination.

**II. LOCALLY AVERAGED NP CURRENT FOR NV STRATEGIES**

The space vector diagram for a three-level NPC converter with balanced dc-link capacitors is illustrated in Fig. 2. The triplets used as vector names denote the states (0 for \( -V_{dc}/2 \), 1 for \( V_{\text{NP}} \), or 2 for \( +V_{dc}/2 \)) of phase voltages \( a, b, \) and \( c \), respectively. The magnitude of the voltage reference vector \( V_{\text{REF}} \) is equal to \( m \), while its angle is the reference angle \( \theta \). The diagram can be divided into six 60° sextants, for voltage reference angle, \( \theta = [0°, 60°), [60°, 120°), etc. In each sextant, there are two long vectors (\( L_0 \) and \( L_1 \)), one medium vector (\( M \)), four small vectors (\( S_0, S_1, S_2, \) and \( S_3 \)) and three zero vectors (\( Z_1, Z_2, \) and \( Z_3 \)), arranged as shown in Fig. 3.

The NP voltage ripple that appears at the dc-link is generated by the NP current \( i_N \) shown in Fig. 1. Each of the space vectors in Fig. 2 results in a certain NP current, shown in parentheses [9], [15]. It can be observed that only medium and small vectors produce a non-zero NP current. The function \( I_{\text{NP}}(V) \) will be used to denote the NP current that corresponds to vector \( V \). If \( I_{\text{LM}} \) is the locally averaged

![Fig. 1. Three-level NPC converter.](image1)

![Fig. 2. Space vector diagram for a three-level NPC converter.](image2)

![Fig. 3. Space vectors in one sextant of a three-level converter.](image3)
taken from the NP due to a medium vector $M$ (see Fig. 3), then

$$i_M = d_M I_{NP}(M) \quad (1)$$

Where $d_M$ is the duty cycle of the medium vector (expressions for the duty cycles can be found in [13]). Furthermore, if $i_S$ is the locally averaged NP current that can be taken from the NP due to small vectors, then $i_S$ depends on the distribution of the (total) small vector duty cycles, $d_{S0}$ and $d_{S1}$, among the vectors of the respective pair (see $S_0$/$S_0$ and $S_1$/$S_1$ in Fig. 3). This is because the two small vectors of each pair produce opposite values of NP current, as shown in Fig. 2. If $x_{S0}$, $x_{S1}$ $\in \{-1, 1\}$ are used as shown in [13] to adjust the distribution of $d_{S0}$ and $d_{S1}$, respectively, then the duty cycles of $S_0$, $S_0$, $S_1$, and $S_1$ are given by

$$d_{S0,1} = d_{S0} \frac{1+x_{S0}}{2}, d_{S0,2} = \frac{1-x_{S0}}{2}$$
$$d_{S1,1} = d_{S1} \frac{1+x_{S1}}{2}, d_{S1,2} = \frac{1-x_{S1}}{2} \quad (2)$$

The above distinction can provide the basis for creating hybrid strategies which can eliminate NP voltage ripple. The converter with a medium vector $\Phi = 30^\circ$ and $m = 0.9$ (for $I_o = 1$ A).

The shaded areas correspond to uncontrollable intervals. While $S$ is given by

$$i_S = x_{S0} d_{S0} I_{NP}(S_0) + x_{S1} d_{S1} I_{NP}(S_1) \quad (3)$$

For each value of $\theta$, $i_S$ can reach a certain highest (maximum) value $i_{S,hi}$ as a result of setting

$$x_{S0} = \text{sign}(I_{NP}(S_0)) \text{and} x_{S1} = \text{sign}(I_{NP}(S_1)) \quad (4)$$

Use of

$$x_{S0} = -\text{sign}(I_{NP}(S_0)) \text{and} x_{S1} = \text{sign}(I_{NP}(S_1)) \quad (5)$$

Results in $i_S$ taking a respective lowest (minimum) value $i_{S,lo}$. Since $I_{NP}(V)$ is zero for long and zero vectors, the locally averaged NP current $i_{NP}$ is the sum of $I_M$ and $I_S$. NP strategies cannot control $I_M$, hence, the highest $i_{NP,hi}$ and lowest $i_{NP,lo}$ values of $i_{NP}$ that they can achieve are

$$i_{NP,hi} = i_M + i_{S,hi} \text{and} i_{NP,lo} = i_M + i_{S,lo}. \quad (6)$$

It is noted that certain NV strategies follow switching sequences that do not allow $x_{S0}$ and $x_{S1}$ to take the values of $\pm 1$ for all values of $\theta$ (for example, see Symmetric strategy in [4]). In these cases, $i_{NP,hi}$ and $i_{NP,lo}$ are derived using the permissible values of $x_{S0}$ and $x_{S1}$ that are closest to the ones dictated by (4) and (5), respectively.

III. PROPOSED HYBRID STRATEGIES

The NP voltage (ripple) in the NPC inverter is determined by the integral of the NP current. In a given switching cycle, deviation of $v_{NP}$ can be avoided by an NV strategy if, by adjusting $x_{S0}$ and $x_{S1}$, $i_{NP}$ can be made equal to zero. Since $i_{NP}$ can only take values between $i_{NP,lo}$ and $i_{NP,hi}$, this is possible when

$$i_{NP,lo} \leq 0 \text{ and } i_{NP,hi} \geq 0 \quad (7)$$

Fig. 4 plots $i_{NP,lo}$ and $i_{NP,hi}$ according to (1) and (3)–(6), for $m = 0.9$, during a fundamental cycle. A balanced set of sinusoidal phase currents with rms value $I_o = 1$ A is assumed, which lag the phase voltages by $30^\circ$ ($\Phi = 30^\circ$). It can be observed that $i_{NP,lo}$ and $i_{NP,hi}$ cross the zero axis, dividing the fundamental cycle ($\theta = [0^\circ, 360^\circ]$) into two types of interval: 1) intervals in which (7) holds, which will be referred to as controllable intervals, since during them $i_{NP}$ can become equal to zero or be adjusted to control the NP voltage to some extent; and 2) intervals in which (7) does not hold (shaded in Fig. 4), which will be referred to as uncontrollable intervals. During them, an NV strategy cannot make $i_{NP}$ equal to zero, thus generating NP voltage ripple.

![Flowchart for the proposed hybrid strategies](image)

The above distinction can provide the basis for creating hybrid strategies which can eliminate NP voltage ripple. Given an NV strategy $X$ and a non-NV strategy $Y$, a hybrid strategy $H_{X,Y}$, that combines the two can be built according to the flowchart in Fig. 5. Namely, the converter
controllable intervals, since during them, \( X \) is capable of holding \( v_{NP} \) to zero. During the uncontrollable intervals, on the other hand, \( Y \) should be put into action to avoid \( v_{NP} \) voltage deviation.

Controllable and uncontrollable intervals can be identified in practice using the instantaneous (sampled) values of the phase currents as values of \( I_{NP}(V) \) in (1) and (3)–(5). In this way, operation of hybrid strategies according to Fig. 5 is equally achievable for converter loads that draw non-sinusoidal or unbalanced currents.

IV. SIMULATION RESULTS

An NPC inverter with \( V_{dc} = 1.8 \text{ kV}, C_1 = C_2 = 0.5 \text{ mF}, \) and \( f = 50 \text{ Hz} \) is simulated using MATLAB–Simulink (SimPower Systems Toolbox). The simulation figures illustrate the locally averaged currents \( i_{NP,lo} \) and \( i_{NP,hi} \) the applied modulation strategy, the line–line voltage \( V_{ab} \) and phase current(s), and the capacitor voltage \( v_{C1} \).

In Fig. 6, the inverter is switched at 8 kHz, and supplies a linear and balanced load with \( \phi = -30^\circ \) (power factor of 0.866). Also, \( m \) is set to 0.9 to show the resemblance of the waveforms of \( i_{NP,lo} \) and \( i_{NP,hi} \) with Fig. 4 (here, \( I_o = 200 \text{ A} \)). These waveforms are shown for the entire simulation, but are only used by the simulated hybrid strategy, according to Fig. 5. This strategy, HNTV-S3V, combines the (NV) NTV strategy [4], with the (non-NV) S3V strategy, proposed in [15]. In order to demonstrate its operation as compared to the combined strategies, the inverter is modulated for one fundamental period (0.02 s) by each strategy, as follows: from 0 to 0.02 s by the S3V, from 0.02 to 0.04 s by the NTV, and from 0.04 to 0.06 s by the HNTV-S3V strategy.

It can be observed that the low-frequency NP voltage ripple that appears when the NTV strategy modulates the inverter is eliminated by the HNTV-S3V. This happens even though the NTV is still applied in place of the S3V for a significant part (48.5%) of the fundamental cycle. Moreover, the line voltage waveform generated by the HNTV-S3V can be seen to be enhanced (i.e., closer to the five level waveform generated by the NTV strategy) as compared to the S3V strategy. The line voltage THD is decreased accordingly, from 0.55 for the S3V to 0.50 for the HNTV-S3V, while it is equal to 0.32 for the NTV strategy.

The level of THD decrement, however, will depend on the converter operating point, as explained in discussion.

Combining the NTV with the S3V according to [15] and using a voltage threshold \( v_{NP,max} \) of 5 V has the effect shown in Fig. 7. Due to the NTV strategy, the NP voltage quickly reaches the threshold and varies around it, thus causing multiple transitions between the two strategies. Such transitions, however, can introduce additional switching steps to the converter. For example, for \( V_{REF} \) as in Fig. 2, the NTV strategy may need to use the switching sequence “210-110-100-110-210” (from the switching-frequency-optimized set of sequences for the NTV strategy, according to [4]). The S3V, on the other hand, would avoid the (ripple-generating) medium vector 210 and use “220-200-100-200-220” for a similar \( V_{REF} \). As a consequence, for each transition between the two strategies in that area of \( V_{REF} \), an additional switching step will be introduced to the converter (to switch from 220 to 210 or vice versa).

Fig. 6. Simulation of the NPC inverter modulated successively by the S3V, NTV, and HNTV-S3V strategies. (a) Locally averaged currents \( i_{NP,lo} \) and \( i_{NP,hi} \). (b) Applied modulation strategy. (c) Line voltage \( V_{ab} \) and current \( 5 \times I_o \). (d) Capacitor voltage \( v_{C1} \).
When using the previous approach, the switching sequences of the combined strategies should therefore be redesigned with the aim of minimizing the added steps and their impact on the converter’s switching losses.

In Fig. 8, the simulated hybrid strategy HSym-S3V S3V. The HSym-S3V is given as a second example, demonstrating the applicability of the proposed concept on different combinations of NV and non-NV strategies. Furthermore, in this simulation, the switching frequency is reduced to 2 kHz, and the load is nonlinear and unbalanced (a bidirectional switch, switching at 400 Hz with a duty cycle of 50%, was connected in series with phase a of the load of the previous simulations). Due to the lower switching frequency, the amplitude of the high-frequency capacitor voltage ripple is increased, and the effect of phase current sampling on the wave forms of $i_{NP,lo}$ and $i_{NP,hi}$ becomes more evident. At the same time, the shape of these wave forms significantly differs from that in Fig. 6(a), as a consequence of the distorted phase currents. Nevertheless, no deterioration of the performance of the hybrid strategy can be observed; the HSym-S3V again eliminates the (low-frequency) NP voltage ripple generated by the Symmetric strategy. The latter is still applied for approximately 50% of the fundamental cycle, decreasing the line voltage THD from 0.55 for the S3V to 0.44 for the HSym-S3V, which is closer to the symmetric’s own value of 0.32.

V. DISCUSSION

For hybrid strategies created according to Fig. 5, depicts the percentage duration of applying the selected non-NV strategy. It can be observed that, for $\phi = -90^\circ$ and $m > 0.7$, this percentage reaches close to 100%. Thus, for purely reactive loads, the proposed hybrid strategies offer no benefit compared to non-NV strategies, since they operate as such (the same has been observed in [16]). On the other hand, for less reactive loads, the participation of non-NV strategies is lower than 100% and decreases with $m$. For low values of $m$, it drops to zero; therefore, the hybrid strategies operate as NV strategies.

An estimate for these switching losses can be obtained by an analysis similar to that included in [14]. Such an analysis can be performed owing to the fact that the intervals of the fundamental cycle where each strategy is applied can be specified analytically (as in Fig. 4); this is not the case in the approaches of [15] and [16]. In comparison to [15], the proposed approach also has the advantage of avoiding multiple transitions between the combined strategies. As a consequence, a converter designer can readily apply it to any implemented (NV–non-NV) pair of strategies, without the need for modification of their switching sequences. Computationally, the selection between the combined strategies according to Fig. 5 incurs an additional cost. However, this cost is comparable to that of NV strategies, since these strategies also perform determination of the NVs and calculation of duty cycles (as in Fig. 5), as well as duty cycle distribution according to (2). Implementations of the proposed hybrid strategies can therefore be expected not to exceed the limits of modern microcontrollers. The approaches in [14] and [15], on the other hand, have the advantage of even lower computational requirements. Moreover, they are able to
adjust the amplitude of NP voltage ripple. This feature is helpful in cases that a sufficient amount of ripple can be accepted by the converter, but cannot be used effectively for ripple elimination (as explained in Section I).

A last comment refers to the NP balancing capabilities of the proposed strategies after possible NP voltage deviations. As shown in [4] and [8], NV strategies can achieve “natural” NP balancing, that is, decrease NP voltage deviation (by determining the values of $x_{S0}$ and $x_{S1}$ to adjust iNP) directly using measurements of the phase currents and dc-link capacitor voltages. This is not the case for certain non-NV/hybrid strategies (see [9] [16] and [10]), which need to implement additional control loops (described in [17] and [18], respectively) to achieve the balancing task. In the proposed approach, however, such loops are not required, since the selected NV strategy has the chance to adjust $i_{SP}$ in favor of capacitor balancing during the controllable intervals of the fundamental cycle (a similar comment can be found in [15]). A simulation detail illustrating the balancing operation of HSym-S3V can be observed in Fig. 8 (see $v_{C1}$ being driven to 900 V at 0.04 s, when the HSym-S3V is applied).

VI. CONCLUSION

The proposed approach inherently guarantees minimum participation of non-NV strategies, since non-NV strategies are only applied when an NP voltage deviation cannot be prevented by NV strategies. Hence, for a given combination of strategies, the approach yields an NP-voltage-ripple-eliminating strategy with minimum switching losses and output voltage harmonic distortion.

REFERENCES


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