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SYNTHESIS OF DESIRED AC LINE CURRENTS IN CURRENT-SOURCED DC-AC CONVERTERS

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Abstract

The paper addresses the topic of the harmonic reduction and fundamental power factor correction for three-phase current-source (CS) DC-AC converters in a unified mathematical manner. Starting only from the analytical definition of the power characterizing the line current error signal, a closed-loop general optimal control strategy is deduced valid for different CS inverters and suitable to be implemented on-line via DSP-based controllers. The proposed strategy is applied to control the two most common types of CS inverters and related simulation results are presented. The reasoning accompanying these results and the developed strategy leads also to highlight and review the intrinsic differences between these types of converters, concerning the synthesis process of the desired line currents.

1 Introduction to the problem addressed

This paper is mainly aimed at proposing an alternative view of CS inverters modulation realized by an optimal-control methodology implementing true real-time feedback. Other authors like Bowes et al. have performed extensive research on optimal PWM, according to specific chosen criterions [4-7]. Unlike the method proposed here, their solutions need the open-loop determination of optimal non-sinusoidal modulating functions, which are regularly sampled on-line by DSPs tailored for PWM modulation. The optimal control described in this paper compensates for non-linearities within the inverter such as switching delays, conduction and switching losses through the continuous measurement of proper quantities.

The most general circuit topology, which constitutes the building block of any three-phase CS DC-AC converter, is depicted in Figure 1 where ideal switches - as defined by the circuit theory - are assumed for sake of generality [12]. Table 1 lists the sole nine, out of sixty-four, possible switching configurations (states) which do not violate either the Kirchhoff’s Current Law for the current-source nor the Kirchhoff’s Voltage Law for the voltage sources. The line currents \( i_a(t), i_b(t), i_c(t) \) can be written by using the switching functions \( S_{ij}(t) \in \{0,1\}, i=\{1,2\}, j=\{a,b,c\} \) as Equation (1) shows. The usual purpose of such DC-AC converters is to synthesize three alternating line currents, \( i_a(t), i_b(t), i_c(t) \) characterized by a reduced harmonic content and whose fundamentals possess a desired amplitude and phase displacement with respect to the related sinusoidal phase voltages \( v_a(t), v_b(t), v_c(t) \) [1, 8, 11, 12].

\[
i_{\gamma}(t) = I_{dc} \cdot \left( S_{2\gamma}(t) - S_{1\gamma}(t) \right), \quad \gamma = \{a, b, c\}
\]

Table 1: The 9 possible switching states (1 = ON, 0 = OFF)

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In regenerative operation, such phase displacement is desired to be \( \pi \) radians when the current and voltages are assumed with the reference directions depicted in Figure 1.

The general goal of the topology control is to render the currents \( i_a(t), i_b(t), i_c(t) \) the “closest possible” to three sinusoidal desired reference currents: \( i_{a,ref}(t), i_{b,ref}(t), i_{c,ref}(t) \) by using a proper control methodology of the six switches \( S_{ij} \), \( i=\{1,2\}, \gamma=\{a,b,c\} \) (indicating the three phases). “Closest possible” is meant here in the sense of the Euclidean norm applied to the three infinite dimensional vectors whose elements are the Fourier components of \( i(t) - i_{ref}(t), \gamma=\{a,b,c\} \). The definitions used are shown in Equation (2) under the assumption that the currents are alternating periodic functions with pulsation \( \omega_m = 2\pi/T \). This hypothesis reduces the Fourier integral to a sum of Dirac’s Delta distributions. Equation (3) shows the Euclidean norm of the aforementioned difference vector (i.e. the “distance” between the actual current \( i(t) \) and the reference one \( i_{ref}(t) \) ) and Equation (4)
expresses its squared value by using the well-known Parseval’s identity [2, 10]. Equation (3) has been expanded considering that, by definition, only the first-order harmonic components of the desired sinusoidal reference currents are different from zero.

\[
I_R(\omega) = 2\pi \sum_{n=-\infty}^{\infty} a_{I_R}(n) \delta(\omega - n \cdot \omega_m) \quad \text{where:} \\
a_{I_R}(n) = \frac{1}{T} \int_0^T I_R(t) e^{-j n \omega_m t} dt \quad \alpha_{I_R}(n) = C \\
d_{I_R, i_{R, ref}} = \sqrt{2} \left[ a_{I_R}(1) \cdot a_{I_R, ref}(1) \right]^2 + \sum_{n=2}^{\infty} \left| a_I(n) \right|^2 \\
\frac{1}{T} \int_0^T \left[ I_R(t) - i_{R, ref}(t) \right]^2 dt = \left( d_{I_R, i_{R, ref}} \right)^2 \quad \gamma = \{a, b, c\} 
\]

(2) \hspace{1cm} (3) \hspace{1cm} (4)

2 The variational nature of the problem and its transposition into an optimal control strategy

The right-hand side (RHS) of the Parseval’s identity in Equation (4) represents also the average power of the periodic “error” signal \( i_R(t) - i_{R, ref}(t) \) as defined by the signal theory [2, 10]. This observation leads to the proposed control technique where the optimal controller synthesizes the “closest possible” \( i_R(t) \) by choosing the suitable sequence of switching states able to minimize the average power of this error signal.

The computation of \( i_R(t) \) considered “optimal” in accordance with this criterion, turns into the minimization of the left-hand side (LHS) in Equation (4). This constitutes a typical variational problem under the constraints that the solution function, i.e. the current \( i_R(t) \), be piecewise constant over a finite number of time intervals and assume only a finite set of values \{1, 0, -1\}. As Equation (1) shows, the current \( i_R(t) \) depends on the specific switching functions \( S_R(t) = \{1, 2\}, \gamma = \{a, b, c\} \) only, thus the ultimate solution of the posed variational problem is exactly the requested sequence of switching states, optimal in the sense defined above. In principle, since the durations of the time intervals where \( i_R(t) \) \( \gamma = \{a, b, c\} \) is kept constant are chosen, such a variational problem can also be solved off-line for each specified \( i_{R, ref}(t) \).

The main drawback of such off-line computations resides in requiring the storage of a potentially large number of solutions inside a look-up table. Additionally the controller shows limited flexibility in the cases where the actual operating conditions differ considerably from the ones whose solutions are computed off-line in advance. The authors propose here a true feedback optimal controller still centered on the variational formulation of the problem, but capable of acting on-line by using the actual current and voltage measurements without the need of relying on any pre-computed solution.

One can observe that, as \( t \to +\infty \), each one of the three non-negative terms \( \psi(t) \) in the summation constituting the time-varying functional \( \psi(t) \) defined by Equation (5), converges asymptotically to the integral at the LHS of Equation (4) written for the corresponding line current.

\[
\psi(t) = \sum_{\gamma=\{a, b, c\}} \frac{1}{t} \int_0^t \left[ I_R(t) - i_{R, ref}(t) \right]^2 dt \quad \psi_{\gamma}(t) \text{ and } t \in (0, +\infty) 
\]

(5)

This is true under the hypothesis that the line currents be alternating periodic signals. Equation (6) clarifies this statement and Equation (7) shows the first-order total time-derivative of \( \psi(t) \) highlighting also the instantaneous squared current error \( \psi(t) \) contained in its expression.

\[
\lim_{t \to +\infty} \psi(t) = \frac{1}{T} \int_0^T \left[ I_R(t) - i_{R, ref}(t) \right]^2 dt 
\]

(6)

\[
\frac{d}{dt} \psi(t) = \sum_{\gamma=\{a, b, c\}} \left[ \psi_{\gamma}(t) + \frac{\left[ i_R(t) - i_{R, ref}(t) \right]^2}{t_{term A}} + \frac{\left[ i_R(t) - i_{R, ref}(t) \right]^2}{t_{term B}} \right] 
\]

(7)

The proposed optimal controller acts on a discrete-time basis ruled by a properly chosen clock signal of period \( T_c = 1/\omega_c \) during which the switching state and, consequently, the currents \( i_a(t), i_b(t), i_c(t) \) are held constant. In principle - at each instant \( t_c \) - defined by the acknowledged triggering event in each clock period - the controller selects the optimal switching state that will remain valid for the whole specific interval \( [t_c, t_c + T_c] \). The optimal state is by definition the one that via Equation (1) - generates the values for \( i_a(t), i_b(t), i_c(t) \) able to minimize the derivative \( d\psi(t)/dt \) at a specific instant \( t_{end} \in [t_c, t_c + T_c] \). In case this a minimum value achieved for \( d\psi(t)/dt \) is negative then, at \( t_{end} \), \( \psi(t) \) is driven towards its minimum.

When \( T_c \) is chosen sufficiently small and the overall operating conditions of the converter allow the existence of a succession of states - selected in accordance with the previous criterion - capable of rendering negative the average of the minimum values assumed by \( d\psi(t)/dt \), then \( \psi(t) \) reaches its minimum asymptotically. As consequence of Equations (6), (4) and (3), it follows that the synthesized line currents \( i_a(t), i_b(t), i_c(t) \) converge to the desired ones \( i_{a, ref}(t), i_{b, ref}(t), i_{c, ref}(t) \) in the sense stated previously.

It is possible to deduce that the simultaneous convergence of each one of the three terms \( \psi_{\gamma}(t) \) towards the related LHS of Equation (4) – as shown in Equation (6) - implies also the reduction of the Total Harmonic Distortion (THD) factor for each of the line currents \( i_a(t), i_b(t), i_c(t) \). Indeed recalling the definition of the THD factor for the currents \( i_{\gamma}(t), \gamma = \{a, b, c\} \) as
given in Equation (8), it is possible to rewrite Equation (6) in the form shown by Equation (9).

\[
\text{THD}_\gamma = \sqrt{\sum_{n=2}^{\infty} |a_{\gamma n}(n)|^2 \over |a_{\gamma 1}(1)|^2} \quad \gamma = \{a, b, c\} 
\]  

\[
\lim_{t \to \infty} \Psi_f(t) = 2 \left| a_{\gamma 1}(1) \right|^2 \left( {\left| a_{\gamma 1}(1) \right|^2 \over \left| a_{\gamma 1}(1) \right|^2 + \text{THD}_\gamma^2} \right) 
\]  

From Equation (9) it can be easily inferred that the convergence of \( \Psi(t) \) towards its absolute minimum implies also a reduction of the THD factors since \( \Psi(t) \) is sum of the three non-negative \( \Psi_f(t) \). Additionally, when the overall operating conditions permit the matching between the fundamental components of \( i_l(t) \) and \( i_{\text{ref}}(t) \), \( \gamma = \{a, b, c\} \), each one of the terms \( \Psi_f(t) \) is exactly proportional to the squared THD factor of the corresponding line current. Consequently the optimal control policy reduces the THD factors for the line currents as much as possible compatible with the operating conditions.

3 Comparison with other optimal methods

It was already mentioned that authors such as Bowes et al. [4-7], proposed different PWM optimal modulation methodologies. One interesting possibility exploits fast regular-sampling PWM controllers capable of generating PWM patterns whose switching instants are close to the exact off-line solutions of the well known off-line transcendental equations to minimize harmonic content. This is made possible by using proper non-sinusoidal waveforms as modulating functions. The “optimal” structure of such waveforms is obtained by proper optimizations that must be performed off-line. Consequently such a method is not fully closed-loop.

The regular-sampled PWM modulation of the optimal waveform used in Bowes’s method implies that the switching instants belong to an almost continuous set. Conversely, the use of a fixed frequency clock in the method proposed here leads to the discretization of that set. Such discretization allows the classification of the technique proposed by the authors into the group of Pulse Density Modulators (PDM). The PWM modulator can be considered as a limit case of the general PDM and their comparison can be found in many publications such as [9].

Conceptually the proposed method relies neither on a-priori assumed waveform symmetries nor on a specific class of parametric solutions, as it is the case for other optimized PWM modulations. Additionally, the measurement of proper quantities in the system renders it a true closed-loop algorithm capable of compensating for converter non-linearities and some inaccuracies in the parameters values. The optimal solutions (i.e. switching instants) derive intrinsically from the variational formulation of the “closest possible waveform” problem according to the criterion stated in paragraph 2. The “real-time” minimization of the functional \( \Psi(t) \) follows a method conceptually similar to the Lyapounov techniques used in system theory.

While computationally intensive such an approach is becoming feasible, thanks to the availability of increasingly fast microprocessors. Because of the radically different type of computations required by Equation (7), the optimal PWM in [4-7] may be more suitable for currently existing DSPs specifically dedicated to the regular-sampled PWM modulation. Nevertheless it must be observed that faster DSPs continuously emerge allowing so increasingly smaller clock periods \( T_c \). As a consequence the gap between a virtually continuous and a discrete set of switching instants will be narrowed.

4 Application of the proposed control strategy to two common CS DC-AC topologies

An optimal controller based on the principle explained previously has been implemented in Matlab / Simulink via a proper code and applied to the synthesis of the Mains line currents \( i_{a_m}, i_{b_m}, i_{c_m} \) of two frequently used CS DC-AC converter types shown in Figures 2 and 3. Although both converters are conceptually reducible to the circuit topology in Figure 1, they intrinsically differ in the manner of synthesizing \( i_{a_m}(t), i_{b_m}(t), i_{c_m}(t) \). The converter shown in Figure 2 bases such synthesis on algebraic operations only, involving \( i_l(t), i_{\text{ref}}(t), i_{\text{ref}}(t) \). This remains true even in presence of a possible intermediate transformer which can be seen just as an algebraic operator on the currents. Conversely, the converter in Figure 3 relies on a first intermediate integration, involving also \( i_l(t), i_{\text{ref}}(t), i_{\text{ref}}(t) \), performed by the Y-connected capacitors. Subsequently, the joint action of the capacitor voltages and of the Mains phase voltages produces \( i_{a_m}(t), i_{b_m}(t), i_{c_m}(t) \) through a second integration step performed by the provided series inductances.

![Figure 2: "Direct synthesis" CS converter](image2)

![Figure 3: "Indirect synthesis" CS converter](image3)
Such difference in the analytical dependences of $i_{m,n}(t)$, $i_{m,n}(t)$, $i_{m,n}(t)$ from $i_d(t)$, $i_q(t)$, $i_d(t)$ allows one to classify the converters in Figure 2 and 3 as "direct synthesis" and "indirect synthesis" topologies respectively. Additionally, conceptual care in computing $d\Psi(t)/dt$ is required to model the dynamics of the two systems properly. Indeed the switching states can influence $i_d(t)$, $i_q(t)$, $i_d(t)$ only while $i_d(t)$, $i_q(t)$, $i_d(t)$ constitute the argument of the functional $\Psi(t)$ now. This reality requires explicit consideration of the analytical dependence of $i_{m,n}(t)$, $i_{m,n}(t)$, $i_{m,n}(t)$ from $i_d(t)$, $i_q(t)$, $i_d(t)$ in the computation of $d\Psi(t)/dt$.

From Equation (7) it can be inferred that at the specific instant of time $t_{c}$, when the clock signal triggers a new quest for the optimal switching state in $(t_{c}, t_{c}+T_{c})$, the three terms A in $d\Psi(t)/dt$ each involves an integral evaluated in $(0, t_{c})$ - are already determined and the controller has no longer influence on them. In the case of the "direct synthesis" topology the controller performs the minimization of $d\Psi(t)/dt$ at the instant $t_{md} = t_{c}$ through the terms B only. Specifically, the controller selects at $t_{c}$ the switching state to which are associated the constant current values $i_{m,n}(t)$, $i_{m,n}(t)$, $i_{m,n}(t)$ able to minimize $d\Psi(t)/dt$ computed in $t_{md} = t_{c}$. These current values are indirectly obtained from the knowledge of: a) $i_d(t)$, $i_q(t)$, $i_d(t)$, deriving from the state chosen among the nine allowed. b) the structure of the input transformer and its related turns ratio.

Conversely, in the case of the "indirect synthesis" topology, the controller has no longer influence even on the term B computed at $t_{c}$ because $i_{m,n}(t)$, $i_{m,n}(t)$, $i_{m,n}(t)$, result from the solutions of three (one for each phase) second-order differential equations whose driving functions are $i_d(t)$, $i_q(t)$, $i_d(t)$ and the measured $V_{m,n}(t)$, $V_{m,n}(t)$, $V_{m,n}(t)$, for $t \in [0, t_{c}]$. In this case $d\Psi(t)/dt$ is minimized at the NEXT trigger instant (i.e., $t_{md} = t_{c}+T_{c}$) as Equation (10) shows by highlighting the integral contribution $C_t(t_{c}, t_{c})$ in the interval $(t_{c}, t_{c}+T_{c})$.

\[
\begin{align*}
\frac{d\Psi(t)}{dt} & \bigg|_{t = t_{c} + T_{c}} = \\
& = \sum_{\gamma = (a,b,c)} \left[ \begin{array}{c}
\text{term A} \\
\text{term B}
\end{array} \right]
\end{align*}
\]

This choice of $t_{md}$ necessarily requires the estimations of $i_{m,n}(t)$, $i_{m,n}(t)$, $i_{m,n}(t)$ in $(t_{c}, t_{c}+T_{c})$ which are computed by the controller using the analytical expressions of the aforementioned differential equations solutions programmed as part of the computer code. The controller uses the measures of two among $i_{m,n}(t)$, $i_{m,n}(t)$, $i_{m,n}(t)$ and two among $V_{m,n}(t)$, $V_{m,n}(t)$, $V_{m,n}(t)$ to derive the initial conditions for the solutions. The driving functions $V_{m,n}(t)$, $V_{m,n}(t)$, $V_{m,n}(t)$ are assumed constant in $(t_{c}, t_{c}+T_{c})$ and equal to their measured values $V_{m,n}(t)$, $V_{m,n}(t)$, $V_{m,n}(t)$ respectively. Provided that $T_{c}$ is chosen properly small in comparison with the Mains fundamental period, this assumption is well justified in $(t_{c}, t_{c}+T_{c})$ by the fact that $V_{m,n}(t)$, $V_{m,n}(t)$, $V_{m,n}(t)$ are the Mains voltages and therefore it is unlikely that they vary significantly in such interval.

The same reasoning applies to the sinusoidal reference currents $i_{m,n}(t)$, $i_{m,n}(t)$, $i_{m,n}(t)$ assumed constant at their values $i_{m,n}(t)$, $i_{m,n}(t)$, $i_{m,n}(t)$ respectively. The other forcing functions $i_d(t)$, $i_q(t)$, $i_d(t)$ are constant as well in $(t_{c}, t_{c}+T_{c})$ because they are linked to the switching state which is kept unchanged during such a time interval. These observations simplify considerably the analytical expressions of the already straightforward second order differential equations solutions and allow the analytical computation of the integral contribution $C_t(t_{c}, T_{c})$ in Equation (10). This achievement reduces the computational load on the DSP processor.

5 Simulation results and reflections on the PWM modulation

This paragraph presents the most significant results of the simulations performed on the developed Matlab / Simulink model. They show the effectiveness of the control strategy applied to both types of converters previously illustrated when -1 fundamental power factor is desired at the Mains terminals. Figures 4 and 5 concern the "direct synthesis" converter while Figures 6 and 7 the "indirect synthesis" one. For both converters the Mains frequency is 60 Hz and the amplitude of the Mains phase voltages $V_{m,n}(t)$, $V_{m,n}(t)$, $V_{m,n}(t)$ (purely sinusoidal waveforms in Figures 4 and 6) is 375 V corresponding to 460 V RMS for the line-to-line voltages. The other most significant parameters defining the specific operating conditions are listed below.

a) For the "direct synthesis" converter of Figure 2 without an intermediate transformer: $I_d = 40$ A, $f = 2400$ Hz and the results for two desired fundamental amplitudes (34 A and 24 A) of $i_{m,n}(t)$, $i_{m,n}(t)$, $i_{m,n}(t)$ are shown in Figure 4 where the "solid" piecewise constant waveform is the result related to 34 A while the "dashed" one to 24 A.

b) For the "indirect synthesis" converter of Figure 3: $I_d = 40$ A, desired amplitude of $i_{m,n}(t)$, $i_{m,n}(t)$, $i_{m,n}(t)$ is 34 A, $f = 4800$ Hz, Capacitance = 47 µF , Series inductance = 10 mH. The resulting smoothed waveform shown in Figure 6 represents the Mains line current $i_{m,n}(t)$ while the piecewise constant waveform is the line current $i_d(t)$.

The Figures 5 and 7 show the minimum $d\Psi(t)/dt$ (lower graph) computed by the controller at the clock triggering
Figure 4: Phase a Mains voltage and line current for the transformerless “direct synthesis” converter

Figure 5: The functional $\Psi(t)$ (above) and its $d\Psi(t)/dt$ (below) in the “direct synthesis” converter (for the case 34 A)

Figure 6: Phase a Mains voltage, Mains and bridge line currents for the “indirect synthesis” converter

Figure 7: The functional $\Psi(t)$ (above) and its $d\Psi(t)/dt$ (below) in the “indirect synthesis” converter

A potential solution for $i_{L_{m}}(t)$, $i_{L_{m}}(t)$, $i_{L_{m}}(t)$ characterized by several “notches” (i.e. more similar to a PWM modulated waveform in linear modulation region) would act against the reduction of the THD factor in this topology because a PWM modulated waveform has a very high THD factor in general.

Indeed a primary goal of inserting “notches” in the waveform describing the time evolution of a physical quantity is not to minimize its THD but to reduce some of its frequency components and/or separate them as much as possible [3]. This eases the reduction of the unwanted frequency components - not in the modulated quantity itself - but in other quantities which descend from its integration in time as performed by physical processes intrinsically present in the system.

The integrated quantities are characterized by a good THD because the integral operator enhances - in the frequency domain - the low frequency portion of the power spectrum and reduces the high frequency one. As an example one should observe that in voltage sourced DC-AC converters the voltage is PWM modulated but the quantities for which a good THD is desired are the currents flowing in inductances and magnetic fluxes. Such quantities descend indeed from a time-integration of the PWM modulated voltage.

As a general principle, when PWM techniques are chosen to obtain certain desired quantities with good THD, it is necessary to use physical systems where different quantities - proportional to the time-derivatives of the desired ones - could be PWM modulated. Then, the time-integration performed by the phenomena intrinsic in these systems will bring a good harmonic content in the desired quantities. This integration process is present in the “indirect synthesis” converter indeed, and it is performed mainly by the Y-connected capacitors.

As further confirmation of the previous reasoning, the line current $i_L(t)$ (piecewise constant waveform in Figure 6) at the output of the switch matrix results much more PWM modulated when compared to the “direct synthesis” case in Figure 4 where $i_L(t) = i_L(t)$. The exploitation of the time-integration intrinsically present in the “indirect synthesis” topology leads also to a better minimization of $\Psi(t)$ as one can observe by comparing Figure 5 and Figure 7.

The previous reasoning highlights the reason why, usually, the “indirect synthesis” converters can provide better Mains line currents using a reduced number of fast fully controllable switches (like IGBTs) and simpler converter structures when compared to the “direct synthesis” implementations. These simpler inverters usually use slow devices (like Thyristors) and need complex phase-shifting transformers to interconnect several three-phase bridge topological units. The algebraic harmonic cancellations occurring among these interconnections are the key to obtain spectrum improvements in the Mains line currents. Nevertheless, in principle, the “direct synthesis” topologies can achieve good spectrums as well, but provided that a sufficiently large number of properly phase-displaced three-phase bridges are used.

6 Conclusions

A methodology aimed at obtaining the desired AC Mains line currents using a current-source DC-AC topology has been proposed and applied to the control of the two most common classes of converters belonging to this category. The principle on which the methodology is based derives from the observation of properties arising from signal theory and leads to a true closed-loop control strategy suitable to be implemented on-line on DSP based controllers without the need of storing any a-priori computed look up tables.

It should be noted however that due to the potentially high computational requirements posed on the controller, proper DSP processors must be carefully selected. The faster the DSP processor, the higher the clock frequency $f_c$ can be chosen with the positive consequence that smaller values for the Y-connected capacitances and series inductances are possible in the “indirect synthesis” topology. The reasoning presented highlights - from an alternative point of view - the theoretical duality linking the CS converters belonging to the “indirect synthesis” type, and the voltage-source PWM-operated converters when these last ones are aimed at synthesizing currents and magnetic fluxes with good harmonic content.

References