A Novel Thyristor-Based CSI Topology With Multilevel Current Waveform for Improved Drive Performance

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Abstract—Load-commutated inverters (LCIs), combined with wound-field synchronous machines (WFSMs), can be an excellent solution for high power drives, but their present technology suffers from important drawbacks related to low power factor, large torque pulsations, and poor starting performance. This paper presents a new LCI design intended to overcome the mentioned limitations. An SCR-based forced-commutation circuit is added to the common inverter topology to obtain a five-level waveform for the stator current. This leads to significantly reduced current harmonics and torque pulsations, in addition to bringing benefits in terms of lower additional losses. As a further advantage, the proposed design allows for a significant power factor enhancement. Finally, it enables the WFSM to be started with a much smoother torque compared to the traditional pulsed operating mode of conventional LCI drives. Simulation studies are conducted on a high-power drive scheme to show the aforementioned improvements. Also, a reduced-scale laboratory prototype of a WFSM drive system is tested to verify the feasibility of the proposed converter.

Index Terms—Current harmonics, five-level current source, forced commutation, load-commutated inverter (LCI)-fed drive, power factor improvement, start up, torque pulsations.

I. INTRODUCTION

CURRENT-SOURCE inverters (CSI) are still widely used in synchronous machine drives especially for medium-voltage high-power industrial and ship propulsion applications [1], [2] thanks to some key advantages they offer with respect to their voltage-source counterpart. Inherent short-circuit protection and regenerative capability, simpler structure, lower voltage stresses over the machine insulations and the semiconductor devices and the suppression of travelling wave phenomena when the motor is fed via long cables are among these advantages [3], [4]. Similar to voltage-source inverters (VSIs), pulse width modulation (PWM) techniques can be utilized in CSIs which employ self-turn-off power electronic switches such as Gate Turn-Off Thyristors (GTOs) and Insulated Gate Bipolar Transistors (IGBTs) in order to eliminate the low-ordered harmonic contents of the machine stator current [3]–[6]. In CSI-fed machine drives, however, forced commutation of the dc-link current creates voltage spikes due to motor phase inductances. In order to absorb the voltage spikes, a three-phase capacitor bank is installed at the CSI output which can also play the role of a filter for high-frequency current harmonics produced by the PWM switching [7]–[9]. A rule of thumb in designing the capacitor bank is that the lowest frequency of the stator current harmonic components must be far away from the resonance frequency of the circuits created by the capacitor bank and the machine magnetizing and leakage inductances [10]. Consequently, the capacitance needed will normally vary between 0.3 and 0.5 p.u. of the machine base impedance for switching frequencies of approximately 400 Hz [3].

On the other hand, high frequency switching of the power devices is not desirable in some high-power applications, with power ratings typically over 10 MW [11]–[13], since it decreases the converter efficiency. On the contrary, in load-commutated inverter (LCI)-fed wound-field synchronous motor (WFSM) drives, the efficient, reliable, and inexpensive thyristor (SCR) devices are naturally commutated at the machine operating frequency when the machine is working in the over-excitation mode. This load commutation results in high converter efficiency which can be higher than 99% when the drive operates under the rated conditions [3]. Furthermore, the need for bulky capacitor bank is removed in this drive.

All this said, it is a matter of fact that today’s LCI technology suffers from some well-known weaknesses, such as large line-side and motor-side current harmonics, low power factor and efficiency of the WFSM, poor start-up performance, and large torque pulsations [3]. The line-side current harmonics can be effectively reduced to acceptable levels by adopting a 12-pulse or 24-pulse configuration for the line-side converter [14], [15]. Instead, no effective solutions are presently available in LCI drive state-of-the-art to cope with the other issues that relate to motor operation. As regards the power factor, in fact, a safe motor-side converter operation with no commutation failure risk requires a firing angle usually not higher than 150°, which implies power factor values below 0.9.

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Regarding start-up, it is known that at low speeds (usually under 10% of rated) motor counter electro-motive forces are insufficient for SCR load commutation and a so-called pulsed operating mode is employed in the industry [16], [17]; this gives rise to high torque ripples that may result in severe shaft-line mechanical resonance issues [18].

Finally, in regards to torque pulsations during normal (load-commutated) operating mode, the use of dual three-phase WFSMs can bring benefits thanks to the disappearance of the main torque harmonic component (sixth order) [19]; however, the solution is not always feasible, e.g., it does not fit the cases when LCIs are used to supply grid-connected (three-phase) generators for start-up before synchronization or in pumped storage hydropower plants [20].

The combination of VSI-based active filter and the SCR-based LCI has been proposed in the literature for the high-power induction and synchronous machine drives [7]–[9], [13], [21]–[24]. These research works mainly pursue three objectives which are:

1) the removal of output capacitor bank in PWM-CSI;
2) natural commutation of the SCRs as the main switches;
3) improving the above-mentioned drawbacks of the LCI-fed machine drives.

For instance in [24], the machine could be started from standstill via the VSI keeping the LCI OFF, while at normal operating speeds the VSI acts as an active filter attenuating the stator current harmonics and reducing the torque pulsations. However, in high-voltage applications, a multilevel configuration is needed for the VSI to reduce the voltage stresses over the switches. Also, a series inductance must be connected at the VSI output acting as a controlled current source in addition to a capacitor bank to absorb the harmonics arising from the PWM switching. This LC filter could draw the resonance current from the VSI. Therefore, an active damping control scheme should be used to damp the LC resonance. Most importantly, the hybrid VSI–CSI topology tends to reduce the advantage of converters based on SCRs only. In addition, the VSI-related problems in high-voltage drives such as short-circuit protection still remain.

In this paper, a novel five-level-CSI (FLCSI) topology is presented to significantly enhance LCI drive operation in the following respects: reduction of torque harmonics in a three-phase LCI-fed WFSM; increase in WFSM power factor; strong reduction of start-up torque pulsations; reduction of motor current harmonics; and related additional losses. These advantages are obtained through the addition of suitable SCR- and capacitor-based circuits, so as to change motor currents from their usual quasi-square shape into a five-level waveform and to reduce current fundamental shift angle with respect to motor voltage (with consequent power factor increase). The effectiveness of the proposed new concept is illustrated and positively assessed through accurate numerical simulations of a high-power WFSM drive system and through experimental implementation of a low-power prototype.

II. PROPOSED FIVE-LEVEL-CURRENT-SOURCE CONFIGURATION CONCEPT

The proposed FLCSI structure is shown in Fig. 1. It consists of a usual LCI combined with a thyristor-based CSI, which is, in turn, split into a couple of SCR bridges, called CSI and CSI-A (A standing for auxiliary). Neglecting the dc-link current ripple, the output currents of both converters, i.e., $i_{LCl_a}$ and $i_{CSIA}$, taking phase $a$ as an example, feature the usual quasi-square waveform and are identical except for a phase displacement $\omega \Delta t_0$, as shown in Fig. 2 (the line-side converters are controlled to output the same dc-link current $I_{dc}/2$). In the figure, the motor phase $a$ voltage $e_{ma}(t)$ and line-to-line voltage $e_{mn}(t)$ neglecting commutation notches (so focusing the attention on the fundamentals) are considered.

The resultant waveform of $i_a(t) = i_{LCl_a}(t) + i_{CSIA}(t)$ will have five levels, i.e., $0, \pm I_{dc}/2, \text{and} \pm I_{dc}$ (see Fig. 2). In particular, if the phase displacement between $i_{LCl_a}$ and $i_{CSIA}$ is set equal to $30^\circ$, the amplitude of the fifth and seventh order current harmonics becomes 5.36% and 3.83% of the fundamental, respectively, while they would be 20% and 14.35% if the usual LCI topology with quasi-square-wave currents were employed. The $30^\circ$ displacement is selected so as to minimize the current total harmonic distortion (THD) as illustrated in Fig. 3.
Out of the three thyristor bridges included in the proposed topology, the converter indicated as LCI in Fig. 1 operates as a usual load-commutated inverter. To guarantee safe commutations, it must be operated with a suitable firing angle \( \alpha \) (adequately lower than 180°), as illustrated in Fig. 2, where thyristor \( T_1 \) is shown to turn on at point A. Conversely, CSI thyristors are force commutated as it will be explained in Section III, with no risk of commutation loss; therefore, they can be safely fired with a 30-electrical-degrees delay with respect to LCI thyristors (for instance \( S_1 \) is fired at point \( B \), which is placed \( \omega \Delta t_0 = 30 \) electrical degrees after A). Assuming \( t = 0 \) as the zero-crossing instant of the phase voltage \( e_{ac} \), from Fig. 2 it can be inferred that instant \( t_B \) is given by

\[
t_B = (-\pi + \pi / 6 + \alpha + \pi / 6) / \omega = (\alpha - 2\pi / 3) / \omega
\]

being \( (-\pi + \pi / 6) / \omega \) the negative derivative zero crossing of \( e_{ac} \), that is the reference instant for firing \( T_1 \).

As a result, not only a better current harmonic spectrum is obtained (as mentioned above), but also a power factor enhancement occurs. In fact, Fig. 2 shows that the fundamental components of \( i_{LCI}(t) \) and \( i_{CSI}(t) \) have, respectively, a leading \((\varphi_{LCI})\) and lagging \((\varphi_{CSI})\) phase shift with respect to \( e_{ac}(t) \). Consequently, the phase shift \( \varphi \) of the fundamental component of the total current \( i_0 \) will be reduced compared to \( \varphi_{LCI} \).

In addition to reducing the current distortion and increasing the power factor, the forced-commutation circuit composed of CSI and CSI-A can be used, at standstill or low speeds, as a high-performance starter, too, achieving a definitely better torque than with the usual pulsed operating mode of normal LCIs [25]. This will be discussed in more detail in Section IV.

III. FORCED-COMMUTATION CIRCUIT PRINCIPLES OF OPERATION

So far, several forced-commutation circuits have been proposed in the literature to turn-off the thyristors in high power CSI-fed induction motor drives. One of the first forced commutation schemes was proposed in [26] and utilized, for instance, in [27] and [28]. The commutation circuit is composed of one capacitor (with one terminal connected to the machine neutral point) and two auxiliary thyristors and operates only during machine starting. In this circuit, depending upon the commutation period length, pulse width of stator current is less than 120°. Moreover, at very low speeds, when the electromotive force (EMF) cannot charge the capacitor enough to ensure the commutation, the technique of delayed gating is employed, which needs a voltage sensor. One other drawback of this circuit is the reduction of electromagnetic torque to half during commutation periods since the auxiliary thyristor currents are of zero sequence and do not take part in torque production. Besides, the neutral point of the machine should be accessible. Autosequential current-fed inverters are another example of these circuits [16], [29], which are composed of six additional ac capacitors and six diodes. However, the latter are in series with the main switches, so they have to be selected with the same current rating as the SCRs and cause significant extra losses; moreover they suffer from large machine spike voltages during commutations [16]. A VSI in parallel with the main CSI can also commute the SCRs [22], but this leads to lose the advantage of converters based on SCRs only.

In this paper, a different external forced-commutation circuit is proposed. It consists of six auxiliary thyristors \( S_{1A} - S_{6A} \) and three ac capacitors \( C_1 - C_3 \), of equal capacitance \( C \) (see Fig. 1). This solution can be already found in the literature as a part of CSI-fed induction motor drives [29]. In addition to being a low cost and efficient circuit due to the inexpensive SCR switches, the current rating of the auxiliary thyristors is much less than that of the main devices since the dc-link current flows through \( S_{1A} - S_{6A} \) only during commutation transients.

In this section, details are given on how forced commutation is obtained in the CSI bridge using the auxiliary converter CSI-A. The description includes mathematical details (see Section III-A) which are however required to obtain some practical dimensioning rules regarding capacitors \( C_1, C_2 \), and \( C_3 \) as well as auxiliary thyristors \( S_{1A} - S_{6A} \) (see Sections III-B–III-D).

A. Forced Commutation Dynamic Analysis

In order to investigate the forced commutation of converters CSI and CSI-A, the WFSM is modeled as a series connection of a phase EMF—e.g., \( e_a(t) = \sqrt{2}E \sin(\omega t) \), taking phase \( a \) as an example—and a commutation inductance \( L_C \) [30], being \( \omega \) the electrical angular frequency. As a case study, let us consider the forced commutation process where \( S_1 \) turns on and \( S_5 \) turns off, which occurs at instant \( B \) (see Fig. 2). Before such instant, \( T_1, T_6, S_1 \), and \( S_5 \) are conducting and \( C_5 \) is charged with voltage \( u_{C5} \), having amplitude \( +U_{C0} \) according to the polarity shown in Fig. 1. The current and voltage waveforms over a period of the electrical quantities \( T = 2\pi / \omega \) are shown in Fig. 4 for thyristors \( S_1, S_5 \), and capacitor \( C_5 \), along with a zoomed view of the commutation process between \( S_1 \) and \( S_5 \).

In order to turn \( S_1 \) off and \( S_5 \) on, \( S_{1A} \) is fired (together with \( S_1 \)) at an instant \( t_B \) corresponding to point \( B \) in Fig. 2. As a result, the voltage \( U_{C0} \) insisting on capacitor \( C_3 \) is applied across \( S_1 \) and turns it off. This turn-off process is abrupt (practically instantaneous) because there is no inductance slowing the decay of the current in \( S_5 \). Consequently, the current \( i_{CSI} \), which is
equal to \( I_{dc}/2 \), is forced to flow through the series of \( S_{SA} \) and \( C_3 \). This causes capacitor voltage \( U_{C3} \) to linearly decay according to the following equation:

\[
U_{C3} = U_{C0} - \left[ I_{dc}/(2C) \right](t - t_B). \tag{2}
\]

From (2), by setting \( U_{C3} \) to zero, the time interval \( t_p \) during which a reverse voltage is applied to the outgoing thyristor \( S_5 \) (see Fig. 4) is obtained as

\[
t_p = 2CU_{C0}/I_{dc}. \tag{3}
\]

The value of \( t_p \) must be larger than the thyristor turn-off time \( t_q \), the condition for \( S_5 \) turn-off being \( t_p - \text{min} \geq t_q \).

With reference to Fig. 4, the further instants \( t_1 \) and \( t_2 \) are identified. Between \( t_B \) and \( t_1 \), \( S_1 \) is not conducting (being reversely polarized) and a constant current \( I_{dc}/2 \) flows through \( C_3 \) and \( S_{SA} \). At \( t = t_1 \), a direct voltage is applied to \( S_1 \) causing it to start conducting together with \( S_{SA} \). Finally, at \( t = t_2 \) the commutation completes with all the current \( I_{dc}/2 \) transferred from \( S_{SA} \) to \( S_1 \) causing \( S_{SA} \) to turn off. The overall transient between \( t_B \) and \( t_2 \) is governed by the equivalent circuit shown in Fig. 5 and, hence, by the following differential equation:

\[
v_{S1}(t) + u_{C3}(t) - L_i \delta_i(t)/dt - e_c(t) + e_a(t) + L_i \delta_i(t)/dt = 0 \tag{4}
\]

where \( e_a(t) - e_c(t) = e_{ac}(t) = \sqrt{6}E \sin(\omega t - \pi/6) \) is the line-to-line voltage that drives the commutation transient.

For \( t_B \leq t \leq t_1 \), \( i_{S1}(t) = i_{LC1a}(t) = I_{dc}/2 \) and \( i_c(t) = i_{CSIa}(t) = I_{dc}/2 \) [see Figs. 4(a) and 5]. Hence, \( di_c(t)/dt \) and \( di_a(t)/dt \) in (4) are equal to zero for \( t_B \leq t \leq t_1 \). Therefore, by setting \( v_{S1}(t) \) equal to zero in (4) for \( t = t_1 \) and replacing \( u_{C3}(t_1) \) from (2) into (4), \( t_1 \) is obtained from the following equation:

\[
U_{C0} - \left[ I_{dc}/(2C) \right](t_1 - t_B) = -\sqrt{6}E \sin(\omega t_1 - \pi/6). \tag{5}
\]

For \( t_1 < t < t_2 \), \( S_1 \) begins conducting together with \( S_{SA} \) [see Fig. 4(b)], which causes \( i_{CSIa} \) to increase and \( i_{CSIe} \) to decrease. At \( t = t_2 \), currents \( i_{CSIa} \) and \( i_{CSIe} \) reach \( I_{dc}/2 \) and zero, respectively, and consequently \( S_{SA} \) turns off and the commutation is completed. Hence for \( t_1 < t < t_2 \)

\[
i_{CSIa} + i_{CSIe} = I_{dc}/2, \quad i_{CSIa} = i_a - I_{dc}/2, \quad i_{CSIe} = i_c. \tag{6}
\]

Since \( i_{CSIe} = -CdU_{CSI}/dt \) (see Fig. 5), extracting \( u_{C3} \) from (4), deriving it, and using (6), the following equation is obtained for \( t_1 < t < t_2 \) [\( v_{S1} \) is zero in (4) neglecting its on-state voltage drop]:

\[
2LC^2d^2i_{CSI}/dt^2 + i_{CSIa} = I_{dc}/2 + Cd(e_c - e_a)/dt. \tag{7}
\]

\( \omega_e(t) \) in (7) can be regarded as constant and equal to its value at \( t = t_1 \), or

\[
e_{ac}(t) \equiv \sqrt{6}E \sin(\omega t_1 - \pi/6). \tag{9}
\]

Thus, (7) can be simply solved imposing \( i_{CSIa} = \int_{t_1}^{t} i_{CSIa}/dt = 0 \) at \( t = t_1 \) as initial condition. Defining \( \omega' = 1/\sqrt{2LC} \), for \( t_1 < t < t_2 \) the following solution is found:

\[
i_{CSIa} = (I_{dc}/2) \left\{ 1 - \cos[\omega'(t - t_1)] \right\}. \tag{10}
\]

\( u_{C3}(t) \) for \( t_1 < t < t_2 \) can be found from (4), (6), (9), and (10) as

\[
u_{C3}(t) \equiv -\sqrt{6}E \sin(\omega t_1 - \pi/6) - LC\omega' I_{dc} \sin\omega'(t - t_1). \tag{11}
\]

Finally, at \( t = t_2 \), \( i_{CSIa} = I_{dc}/2 \) and (10) gives

\[
t_2 - t_1 = \pi/(2\omega'). \tag{12}
\]

At \( t = t_2 \), a period of commutation is finished and for reasons of symmetry \( u_{C3} \) reaches \( -U_{C0} \) (see Fig. 4) in order to ensure the turn-off of \( S_2 \) by firing \( S_{2A} \) at \( t = t_B + T/2 \). Hence, from (11) and (12) we have

\[
U_{C0} = -u_{C3}(t_2) \equiv \sqrt{6}E \sin(\omega t_1 - \pi/6) + (I_{dc}/2)\sqrt{2LC/C}. \tag{13}
\]

The same process occurs at \( t_B + T/6 < t < t_2 + T/6 \), when \( S_6 \) is turned off by firing \( S_{6A} \).

### B. Capacitor Dimensioning Considerations

The voltage \( U_{C0} \) given by (13) depends on the operating point and in particular varies with \( \alpha \). Using (1) for \( t_B \) and assuming \( t_B \geq t_1 \) we get

\[
U_{C0} \equiv \sqrt{6}E \sin(\alpha - 5\pi/6) + (I_{dc}/2)\sqrt{2LC/C} \tag{14}
\]

which shows that \( U_{C0} \) decreases as \( \alpha \) decreases (assuming that firing angles larger than 150° are not allowed). So the minimum
value of $U_{C0}$ occurs when the LCI is operating at its minimum firing angle $\alpha_{\text{min}}$. However, the minimum value of $U_{C0}$ can be also obtained from (3) in the limit case where $t_p = t_q$. Now, if we equate the minimum $U_{C0}$ obtained in this way to that given by (14) with $\alpha = \alpha_{\text{min}}$, the following algebraic relationship results:

$$t_q = t_p = (2\sqrt{6}EC / I_{dc}) \sin(\alpha_{\text{min}} - 5\pi / 6) + \sqrt{2CL_C}. \quad (15)$$

From (15), it is possible to find the minimum capacitance $C_{\text{min}}$, required to ensure safe commutation as a function of parameters $\alpha_{\text{min}}$, $t_q$, $L_C$, and $E$.

As an example, if $\alpha_{\text{min}} = 150^\circ$ from (15) we get

$$C_{\text{min}} = t_q^2 / (2L_C). \quad (16)$$

For illustration purposes, let us assume the drive data given in Appendix A, with an SCR turn-off time of 100 $\mu$s and $L_C$ approximately equal to 0.5 ($L''_d + L'_d$) = 87 $\mu$H [30] (being $L''_d$ and $L'_d$ the machine subtransient inductances along rotor $d$ and $q$ axes). In this case, $C_{\text{min}}$ can be computed to be of 57 $\mu$F according to (16).

### C. Auxiliary SCR Dimensioning Considerations

As explained above, the auxiliary thyristors ($S_{1A} - S_{6A}$) only conduct during the commutation intervals. For example, using (10) the current through $S_{5A}$ over one period $T$ is

$$i_{S_{5A}} = \begin{cases} I_{dc}/2, & t_B < t < t_1 \\ I_{dc}/2 - i_{\text{CSlA}} = (I_{dc}/2) \cos \omega(t - t_1) & t_1 < t < t_2 \\ 0 & t_2 < t < t_B + T \end{cases} \quad (17)$$

and its rms value is

$$i_{S_{5A} \text{Arms}} = (I_{dc}/2) \sqrt{((t_1 + t_2)/2 - t_B)/T} \quad (18)$$

where $t_B$, $t_1$, and $t_2$ in the above-mentioned equation are obtained according to (1), (5), and (12), respectively. The maximum value of the auxiliary-SCR rated current, normalized to that of the main thyristors ($i_{\text{Arms},n}$) of the LCI and CSI bridges (see Fig. 1), is plotted in Fig. 6(a) as a function of the capacitance for the rated operating condition, i.e., rated speed and torque of the machine specified in Appendix A. From the figure, the maximum value of the auxiliary thyristor current rating is 40% for a capacitance of 200 $\mu$F.

### D. General Dimensioning Considerations

As regards the sizing of capacitors $C_1$, $C_2$, $C_3$, their voltage $U_{C0}$ using (14) is plotted in Fig. 6(b) as a function of capacitance for rated conditions. To sum up, we can conclude that an increase in the capacitance $C$ causes an increase in the current rating of the auxiliary thyristors, while it decreases the peak capacitor voltage $U_{C0}$. Thus, a trade off should be made to have reasonable values for capacitor voltage rating together with an acceptable current capability for auxiliary thyristors (see Section III-C).

### IV. Simulation Results

The proposed FLCSI is simulated applying it to the supply of the WFSM described in Appendix A. The field-oriented control strategy proposed in [31] is implemented to control the speed. In order to limit the capacitor voltage on one side and the current rating of the auxiliary thyristors on the other, a capacitance of 150 $\mu$F is selected. Fig. 7 illustrates the phase voltage and current of the machine when it is driven by a conventional LCI and the proposed FLCSI, at the speed of 1500 r/min and with the same load torque. It can be seen that the displacement angle of the fundamental current component with respect to the voltage is nearly zero for the proposed topology (meaning almost unity power factor operation) and 23$^\circ$ for the conventional LCI design (see Section II).

The amplitude of the phase current harmonic components relative to the fundamental $i_{1\text{max}}$ is shown in Table I for the FLCSI and the conventional LCI. It can be seen from Table I...
that the power factor improvement yields lower dc-link current for a given equal torque ($I_{dc}/2$ is around 300 A for the FLCSI while $I_{dc}$ is around 670 A for the LCI). As a further beneficial effect, also the field current (and hence field losses) decreases (from 138 to 123 A) in the new topology thanks to the better power factor.

Torque pulsations are also decreased considerably as shown in Fig. 8, where the amplitude of 6th and 12th harmonics of the motor electro-magnetic torque $T_{em}$ are, respectively, 8% and 4% of the average torque for the FLCSI and 15% and 5% for the conventional LCI. The average torque is equal to 1602 Nm for both cases.

Finally, as previously mentioned, the machine can be started via the forced-commutation circuit (CSI and CSI-A), omitting to release the firing pulses for LCI thyristors. In Fig. 9, the machine electromagnetic torque waveforms at starting are shown for both the proposed scheme and the known pulsed method at the speed of 0.04 p.u. As it can be seen, the torque notches of the pulsed method are removed in the proposed scheme.

V. EXPERIMENTAL RESULTS

The proposed multilevel current source inverter is experimentally implemented to assess its capability in controlling three-phase WFSM. A picture of the laboratory prototype is shown in Fig. 10. Thyristors of the type SKKT 57/22 E are employed as the main switches (in rectifiers, LCI and CSI) and BT152-800R thyristors for the auxiliary circuit (CSI-A). The control algorithm described in [31] is implemented on dsPIC30F4011 microprocessors.

In Fig. 11, the switching pattern for the main thyristors in CSI and the auxiliary thyristors in CSI-A is shown. Considering the fact that the switching of the thyristors is carried out according to the rotor position, the pulses distinguished in blue are released to the CSI-A thyristors during normal and starting operation. The red pulses in Fig. 11 are provided to appropriately charge the capacitors at starting from standstill to guarantee the forced-commutation approach. For instance, assume that at standstill capacitors $C_1$, $C_2$, $C_3$ are discharged, i.e., $U_{C0} = 0$. At $\theta = \theta_0$ thyristors $S_1$ and $S_0$ are fired and start conducting. According to Fig. 11, at this instant $S_{3A}$ is also fired to charge $C_2$ to $-U_{C0}$ (see Fig. 1). Therefore, at the next commutation instant, i.e., at $\theta = \theta_0 + \pi/3$, $S_0$ can be turned off by firing $S_{6A}$. $S_{4A}$ is also fired at $\theta = \theta_0 + \pi/3$ in addition to $S_2$ to charge $C_1$ enough to make it ready to apply the negative voltage across $S_1$ by firing $S_{1A}$ at $\theta = \theta_0 + 2\pi/3$. After starting the machine, the red pulses can be removed.

As shown in Fig. 10(b), a WFSM coupled to a separately-excited dc generator is used in the experiments. Ratings and parameters of the synchronous motor are reported in Appendix B. According to these parameters, the subtransient inductance is equal to 19.5 mH. Furthermore, the turn-off time $t_{\phi}$ of the main thyristors in CSI is equal to 80 $\mu$s. Consequently, (16) shows that a minimum capacitance of 0.16 $\mu$F is needed to guarantee the safe commutation. However, a capacitance [see Fig. 10(a)] equals to 2 $\mu$F is employed in order to reduce $U_{C0}$ at full-load operating condition, because, as it will be explained later, it impacts the voltage ratings of the auxiliary thyristors.

In Fig. 12, the stator voltage and current waveforms are shown for the starting operation (frequency of 2 Hz or rotor speed of 40 r/min). As it can be seen from this figure, the stator voltage is not adequate to guarantee a good commutation (with no commutation failures) in the LCI, especially when the starting torque is high or, in other words, the starting current is high. Therefore, the stator current is supplied by the CSI circuit and there is no need to implement the conventional dc-link pulsing method.

Subsequently, experimental results are compared at normal operating conditions for two cases: 1) the WFSM is supplied by the FLCSI and 2) the WFSM is supplied by the conventional LCI. The rotor speed is 450 r/min and the load is the same in the two cases. In the second case, the firing angle is set to 150$^\circ$ (higher firing angles result in commutation failure); therefore, the fundamental power factor angle is 30$^\circ$ (leading). On the other hand, setting $\alpha = 150^\circ$ for the LCI in the first case (even if a slightly higher values of $\alpha$ would be allowed because the commutating current is halved) results in a fundamental power factor angle of 15$^\circ$ (leading). This yields to lower dc-link current ($I_{dc}/2 = 2.8$ A) of the proposed configuration compared to the traditional LCI case ($I_{dc} = 6.1$ A).

The average value of the field current is 1.8 A with the FLCSI topology, while in the case in which the WFSM is driven by the LCI it is 2.1 A. This proves a certain advantage of the proposed FLCSI configuration also in terms of lower field current and consequently lower field losses compared to the case of conventional LCI-fed drives.

The stator phase voltage and current waveforms for the first case (the WFSM is supplied by the FLCSI) are shown in Fig. 13. From Fig. 13(b), a firing angle of about 150$^\circ$ is observable for the LCI. Thus, the fundamental component of $i_{f_LCI}$ leads the stator phase voltage by $\varphi_{LCI} = 30^\circ$. On the other hand, the firing angle for CSI switches is 180$^\circ$, which leads to zero displacement angle between the fundamental component of $i_{fCSI}$ and the phase voltage ($\varphi_{CSI}$ in Fig. 13(c)). As a result, the fundamental power factor angle, i.e., $\varphi$ in Fig. 13(d) is 15$^\circ$ (leading).
In Fig. 14, the harmonic spectrum of the quasi-square wave LCI current shown in Fig. 13(b) is compared to that of the five-level waveform shown in Fig. 13(d). Results show that the THD is remarkably reduced in a five-level current (11.5%) with respect to the conventional three-level current waveform (26%). Also, the amplitude of the fifth and seventh components of the stator current is reduced significantly (respectively for the fifth and seventh harmonic orders: 3.7% and 4.1% of the fundamental component in a five-level waveform compared to 21.9% and 10.3% of the fundamental component in a three-level waveform).

The voltage across the capacitors is shown in Fig. 15 for the same operating condition of Fig. 13, where the phase rms voltage is 100 V [see Fig. 13(a)], \( I_{Lc}/2 \) is about 2.8 A [see Fig. 13(b) and (c)], \( \alpha \) is equal to 150° [see Fig. 13(b)]. According to (14), \( U_{C0} \) is calculated approximately equal to 420 V.
Fig. 14. Percentage of $i_a$ and $i_c$ (shown in Fig. 13) harmonic components normalized to the fundamental one.

Fig. 15. Experimental waveform of the voltage across the forced-commutation capacitors captured under the same operating conditions of Fig. 13.

Fig. 16. Experimental and analytical waveforms for the voltage across an auxiliary thyristor, captured under the same operating conditions of Fig. 13.

For this condition, which is in accordance to the experimental results shown in Fig. 15. In addition, the waveform of the capacitor voltage shown in Fig. 15 is the same as the one shown in Fig. 4(a) for $u_{C3}$.

At the same operating condition, the experimental waveform of the voltage across the auxiliary thyristors is compared in Fig. 16 to the analytical one that is obtained from the explanations presented in [25] (see [25, eqs. (20)–(23)]). A good matching between the waveforms can be found in this figure. The ripples present in the experimental waveform are due to the dc-link current ripples reflected in the resistive and inductive voltage drops, while the dc-link current is assumed to be smooth in [25]. According to the analysis presented in [25] and considering 2 μF for the capacitors in the forced-commutation circuit, the rms voltage rating of the auxiliary thyristors is calculated equal to nearly 700 V for the full-load operating condition, i.e., speed of 1000 r/min and rms value of the stator current fundamental component equal to 6.1 A. A voltage of 700 V is reasonable according to the specifications of BT152-800R thyristors.

Furthermore, the current rating of the auxiliary thyristors for the rated working condition can be determined according to (18). Considering that each thyristor in the CSI, $S_1 - S_6$ in Fig. 1, conducts the dc-link current in one-third of a cycle, in order to compare the rated rms current of the auxiliary thyristors with that of the main ones, the value obtained by (18) is normalized by dividing it by $I_{dc}/(2\sqrt{3})$. For this purpose, the quantities $t_B$, $t_1$, and $t_2$ need to be computed. Assuming $E = 220$ V, $I_{dc}/2 = 3.4$ A, $\omega = 100$ rad/s, and $\alpha = 150^\circ$, $t_B$ is equal to 1.70 $\times$ $10^{-3}$ s according to (1). Then, $U_{C30}$ is equal to 475 V using (14) and, consequently, $t_1$ will be obtained as 2.01 $\times$ $10^{-3}$ s by solving (13). Also, $t_2$ is equal to 2.45 $\times$ $10^{-3}$ s according to (12). It should be mentioned that $\omega'$ is equal to 3580 rad/s for the forced commutation circuit which, compared to 314 rad/s for the pulsations of the phase voltages, confirms the assumption made in (8). Finally, (18) shows that the normalized rated current of the auxiliary thyristors is equal to 0.28. Despite the short duration of the commutation transient, the auxiliary thyristor must tolerate the dc-link current which could be relatively large. However, this cannot be counted as an issue thanks to the property of the SCRs of being able to withstand large surge currents. For instance, BT152 devices are capable of conducting surge current ten times their rated rms value.

### A. Comparison of Experimental Results

In order to highlight the higher performance of the proposed FLCSI versus the conventional LCI, experimental results obtained for the two configurations are summarized in Table II, where $\varphi$ is the fundamental power factor angle, $I_{dc}$ is dc-link current, and $I_f$ is the WFSM field current.

Comparison of the experimental data reported in Table II shows the benefits brought by the proposed configuration in terms of displacement power factor, dc-link current and therefore motor stator current, motor field current, and motor stator current distortion.

### VI. Conclusion

A novel SCR-based converter concept is presented in this paper for the supply of high-power WFSM as a high-performance alternative of the traditional LCI. The new concept is based on introducing an additional forced-commutation bridge in parallel to the load-commutated inverter typical of LCI drives. This enables to improve motor current waveform that changes from the usual quasi-square-wave to a multilevel (five-level) shape with reduced harmonic distortion and lower shift angle with respect to motor phase voltages. Resulting benefits can be identified in...
a higher power factor, reduced torque ripple, lower field current, and smaller additional and excitation losses. Furthermore, the proposed converter topology is capable of starting the machine with a much better torque than achievable with the usual LCI pulsed operation. An extensive analytical treatment is proposed in the paper to investigate the forced commutation mechanism and to dimension the capacitors and the auxiliary thyristors needed for it. The overall merits of the proposed solution are illustrated by means of numerical simulations in comparative terms with respect to traditional LCI-fed WFSM drives. An experimental set-up based on a small-scale LCI drive prototype has been also prepared and experimental results prove the feasibility and the claimed benefits of the proposed converter system.

APPENDIX A

Next the characteristic data are provided for the equipment assumed in simulations.

Synchronous machine ratings: 1.02 MVA, 570 V, 1034 A, 100 Hz, 4 poles.

Stator circuit parameters: \( r_s = 1.8 \text{ m}\Omega \) (resistance), \( L_{ld} = 0.07 \text{ mH} \) (leakage inductance), \( L_{md} = 0.6316 \text{ mH} \), \( L_{mq} = 0.5356 \text{ mH} \) (magnetizing inductances). Field circuit parameters: \( r_f = 0.6 \text{ m}\Omega \) (resistance), \( L_f = 0.0504 \text{ mH} \) (leakage inductance). Damper parameters: \( r_{kd} = 5.1 \text{ m}\Omega \), \( r_{kq} = 8.6 \text{ m}\Omega \) (resistances), \( L_{kd} = 0.0147 \text{ mH} \), \( L_{kq} = 0.0243 \text{ mH} \) (leakage inductances).

DC-link inductor: 10 mH.

APPENDIX B

Next the characteristic data are provided for the equipment used in the experiments.

Synchronous machine ratings: 4 kVA, 380 V, 6.1 A, 50 Hz, 6 poles.

Stator circuit parameters: \( r_s = 1.45 \Omega \) (resistance), \( L_{ld} = 13 \text{ mH} \) (leakage inductance), \( L_{md} = 79.2 \text{ mH} \), \( L_{mq} = 49 \text{ mH} \) (magnetizing inductances). Field circuit parameters: \( r_f = 189 \text{ m}\Omega \) (resistance), \( L_f = 3.33 \text{ mH} \) (leakage inductance). Damper parameters: \( L_{kd} = 9.6 \text{ mH} \), \( L_{kq} = 13.5 \text{ mH} \) (leakage inductances).

DC-link inductors: 78 mH.

REFERENCES


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