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A New PWM Strategy Based on a 24-Sector Vector Space Decomposition for a Six-Phase VSI-Fed Dual Stator Induction Motor

Khoudir Marouani, Lotfi Baghli, Djafar Hadiouche, Abdelaziz Kheloui, and Abderrezak Rezzoug

Abstract—This paper presents a new space vector pulsewidth modulation (SVPWM) technique for the control of six-phase voltage source inverter (VSI)-fed dual stator induction machines (DSIM). A DSIM is an induction machine which has two sets of three-phase stator windings spatially shifted by 30 electrical degrees and fed by two three-phase VSIs. Despite their advantage of power segmentation, these machines are characterized by large zero sequence harmonic currents, and in particular those of order $6k \pm 1$, which are due to the mutual cancellation between the two stator windings. The proposed SVPWM scheme, while easy to implement digitally, reduces significantly these extra stator harmonic currents. Experimental results, collected from a 15 kW prototype machine controlled by a digital signal processor are presented and discussed.

Index Terms—Dual stator induction machines (DSIM), six-phase voltage source inverter (VSI), space vector pulsewidth modulation (SVPWM).

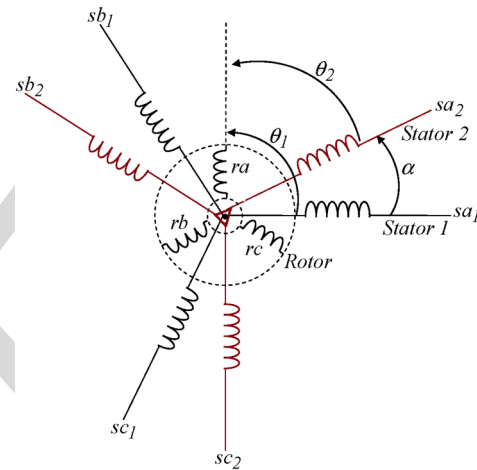


Fig. 1. DSIM windings.

I. INTRODUCTION

NOWADAYS, electrical machine drives are widely used in industrial applications and transportation systems such as electric/hybrid vehicles, traction locomotives and electric propulsion ships, where high-power levels in conjunction with high-performance requirements are more and more demanded. To achieve these high ratings, there are two possible approaches; one focuses on the converter side by increasing the number of output voltage levels and the other one on the machine side by increasing the number of phases. In the first approach, the idea is to divide the high dc bus voltage into multiple low levels and therefore to distribute the high power required among cells of reduced-voltage power switches without the problem of dynamic voltage sharing encountered in the series connection of active devices. However, increasing the number of inverter levels adds to the control complexity and may introduce some voltage imbalance problems [1]–[3]. It is a solution well suited for high-power and high-voltage utility ap-

plications. For adjustable speed drives, however, an alternative approach is to use a multiphase machine, i.e., a machine with more than three phases in the stator, since the number of phases is not imposed anyway, given that the machine is connected to the electric supply through a dc/ac converter. The number of phases could be used instead as an additional degree of freedom in the overall system design [4].

Although the answer to the question whether it is better to use a multilevel inverter-fed three-phase machine or a multiphase machine depends on the application, it is undeniable that the latter option offers several advantages which may make it appear very attractive. In fact, the most significant features are a low torque ripple, a reduction in the power per phase and fault-tolerance capability. Other interesting advantages can be pointed out, such as a better torque production per ampere for the same machine volume, higher efficiency and improved reliability [5], [6].

A common type of multiphase machine is the dual stator induction machine (DSIM), where two sets of three-phase windings, spatially phase shifted by 30 electrical degrees, share a common stator magnetic core as shown in Fig. 1.

Due to the development of fast switching power semiconductor devices, voltage source inverters (VSIs) are preferred in variable speed machine drives. As VSI-fed multiphase machines are gaining increasing interest for high-power applications, various pulsewidth modulation (PWM) techniques have been developed accordingly, as they strongly affect the overall inverter efficiency and output voltage waveform quality.

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68 In a VSI-fed DSIM, the two stator windings are mutually
69 coupled and small unbalances in the two supply voltages may
70 generate high currents [7]. Furthermore, because of the low
71 impedance seen by the voltage harmonic components generated
72 by the switched voltage waveforms, harmonic currents of high
73 level are circulating uselessly in the two stator windings, adding
74 to the overall losses and therefore to the semiconductor devices
75 ratings [8], [9].

76 To minimize these extra harmonic currents in a six-phase
77 VSI-fed DSIM, a new 24-sector PWM technique is proposed
78 in this paper and tested on a 15 kW laboratory machine. The
79 digital implementation is carried out on a DS1104 dSPACE
80 controller board. A comparative study between the proposed
81 technique and similar space vector PWM (SVPWM) techniques
82 [5], [10], based on analytical harmonic current analysis, is also
83 developed and discussed.

84 II. MACHINE MODEL

85 The machine model is based on the assumption that space
86 harmonics and magnetic saturation are negligible, and that the
87 two stator three-phase windings are identical and symmetrical
88 with the two neutrals being isolated. In order to derive a prac-
89 tical model suitable for control, a decoupling transformation
90 matrix is used, as proposed in [5]–[7]. The matrix has the
91 following form:

$$[T_s]^{-1} = \frac{1}{\sqrt{3}} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} & 0 \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} & \frac{1}{2} & \frac{1}{2} & -1 \\ \hline 1 & -\frac{1}{2} & -\frac{1}{2} & -\frac{\sqrt{3}}{2} & \frac{\sqrt{3}}{2} & 0 \\ 0 & -\frac{\sqrt{3}}{2} & \frac{\sqrt{3}}{2} & \frac{1}{2} & \frac{1}{2} & -1 \\ \hline 1 & 1 & 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 & 1 & 1 \end{bmatrix}. \quad (1)$$

92 By applying (1) to the voltage vector equations, the overall
93 machine model is transformed into three decoupled submodels,
94 written in three independent space coordinates, identified as
95 $(\alpha-\beta)$, $(x-y)$, and (o_1-o_2) , respectively.

96 The machine voltage submodel in $(\alpha-\beta)$ coordinates can be
97 written as:

$$\begin{bmatrix} v_{s\alpha} \\ v_{s\beta} \\ v_{r\alpha} \\ v_{r\beta} \end{bmatrix} = \begin{bmatrix} R_s & 0 & 0 & 0 \\ 0 & R_s & 0 & 0 \\ 0 & M\dot{\theta} & R_r & L_r\dot{\theta} \\ -M\dot{\theta} & 0 & -L_r\dot{\theta} & R_r \end{bmatrix} \begin{bmatrix} i_{s\alpha} \\ i_{s\beta} \\ i_{r\alpha} \\ i_{r\beta} \end{bmatrix} + \begin{bmatrix} L_s & 0 & M & 0 \\ 0 & L_s & 0 & M \\ M & 0 & L_r & 0 \\ 0 & M & 0 & L_r \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{s\alpha} \\ i_{s\beta} \\ i_{r\alpha} \\ i_{r\beta} \end{bmatrix} \quad (2)$$

98 where $\dot{\theta} = \Omega_m$ is the rotor mechanical speed, and $L_s = L_{ls} +$
99 $3L_{ms}$, $L_r = L_{lr} + (3/2)L_{mr}$, $M = (3/\sqrt{2})M_{sr}$. L_{ls} and L_{lr}
100 are the stator and rotor leakage inductances in $(\alpha-\beta)$ coordi-
101 nates, respectively.

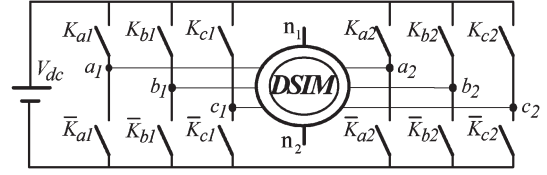


Fig. 2. Six-phase VSI fed DSIM.

The DSIM $(\alpha-\beta)$ submodel expressed in the stationary
reference frame is similar to the three-phase induction machine
model [11].

The machine voltage submodel in $(x-y)$ coordinates is
given by:

$$\begin{bmatrix} v_{sx} \\ v_{sy} \end{bmatrix} = \begin{bmatrix} R_s & 0 \\ 0 & R_s \end{bmatrix} \begin{bmatrix} i_{sx} \\ i_{sy} \end{bmatrix} + \begin{bmatrix} L_{lsxy} & 0 \\ 0 & L_{lsxy} \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{sx} \\ i_{sy} \end{bmatrix} \quad (3)$$

where L_{lsxy} is the transformed stator leakage inductance in
 $(x-y)$ coordinates.

The machine voltage submodel in (o_1-o_2) coordinates is
expressed as follows:

$$\begin{bmatrix} v_{so1} \\ v_{so2} \\ v_{ro} \end{bmatrix} = \begin{bmatrix} R_s & 0 & 0 \\ 0 & R_s & 0 \\ 0 & 0 & R_r \end{bmatrix} \begin{bmatrix} i_{so1} \\ i_{so2} \\ i_{ro} \end{bmatrix} + \begin{bmatrix} L_{lso} & 0 & 0 \\ 0 & L_{lso} & 0 \\ 0 & 0 & L_{lr} \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{so1} \\ i_{so2} \\ i_{ro} \end{bmatrix} \quad (4)$$

where L_{lso} is the transformed stator leakage inductance in
 (o_1-o_2) coordinates.

The electromagnetic torque of the DSIM is expressed only
in terms of stator and rotor $(\alpha-\beta)$ current components, since
the $(x-y)$ and (o_1-o_2) counterparts do not contribute to the
electromechanical energy conversion, as shown by (3) and (4).
The expression of the electromagnetic torque is then as follows:

$$T_e = pM(i_{s\beta}i_{r\alpha} - i_{s\alpha}i_{r\beta}) \quad (5)$$

where p is the number of pole pairs.

General Remarks

The $(x-y)$ and (o_1-o_2) current components do not con-
tribute to the air-gap flux linkages. Hence, they are limited only
by the stator resistance and leakage inductance [12], [13]. They
produce only losses and therefore must be kept equal to zero or
as small as possible.

The transformed voltage equations in the three subframes
are well decoupled and, as a result, both machine analysis and
control are greatly simplified.

III. SVPWM CONTROL OF A DOUBLE-STAR INDUCTION MOTOR

The drive system is a six-phase VSI fed DSIM, as shown
in Fig. 2. A combinatorial analysis of the inverter switch
states shows 64 switching modes. Thus, 64 different voltage
vectors can be applied to the machine. Each voltage vector is
represented by a decimal number corresponding to the binary
number $(K_{c2}K_{b2}K_{a2}K_{c1}K_{b1}K_{a1})$, which gives the state of

136 the upper switches. By using the (6×6) transformation matrix
 137 $[Ts]^{-1}$, each voltage vector can be decomposed into $(\alpha-\beta)$,
 138 $(x-y)$, and (o_1-o_2) voltages. The (o_1-o_2) ones are all equal
 139 to zero because the neutrals (n_1, n_2) of the two winding sets
 140 are isolated. So the SVPWM strategy operates in two complex
 141 planes $(\alpha-\beta)$ and $(x-y)$. Four variables need to be controlled
 142 simultaneously during each sampling period, by generating
 143 maximum $(\alpha-\beta)$ and minimum $(x-y)$ voltage amplitudes.
 144 Therefore, during each sampling period, a set of four active
 145 voltage vectors must be chosen to fulfil these two conditions,
 146 according to the reference voltage vector location. There are
 147 numerous ways for choosing such a set.

148 **A. Six-Phase SVPWM Techniques**

149 The principle of the PWM control techniques proposed in
 150 [5] and [10] is to choose switching sequences in such a way
 151 that two consecutive nonzero voltage vectors are practically
 152 opposite in phase in the $(x-y)$ plane. In this way, each change
 153 in the applied vectors leads to a sequence of increases and
 154 decreases in $(x-y)$ currents around zero. Moreover, in order
 155 to minimize $(x-y)$ harmonic currents and maintain the lowest
 156 switching frequency, there are different choices to allocate
 157 zero voltage vectors (0, 7, 56 or 63) within the switching
 158 sequences. Thus, the switching sequences presented in [10] lead
 159 to continuous and discontinuous modulation techniques and,
 160 consequently, to different harmonic distortion characteristics.
 161 A modulation technique is continuous when on/off switching
 162 occurs within every sampling period, for all inverter legs and
 163 all sectors. A modulation technique is discontinuous when one
 164 (or more) inverter leg stops switching, i.e., the corresponding
 165 phase voltage is clamped to the positive or negative dc bus for
 166 at least one sector [14].

167 **B. 12-Sector SVPWM Technique**

168 In the SVPWM technique addressed in [5], only the $(\alpha-\beta)$
 169 voltage vectors having maximum magnitude (45, 41, 9, 11,
 170 27, 26, 18, 22, 54, 52, 36, 37) are employed to synthesize
 171 the reference voltage vector $v_{s\alpha\beta}^*$. These voltage vectors divide
 172 the $(\alpha-\beta)$ plane into 12 sectors and each sector is $\pi/6$ rad,
 173 as shown in Fig. 3. For example, voltage vectors 45, 41, 9,
 174 and 11 are selected when the reference voltage vector is located
 175 in sector 1. As shown in Fig. 4, continuous and discontinuous
 176 modulation techniques can be obtained according to the switch-
 177 ing sequences given below.

178 **1) Continuous Modulation C6 ϕ SVPWM12:** For example,
 179 when the reference voltage vector is located in sector 1, a con-
 180 tinuous modulation technique (C6 ϕ SVPWM12) is obtained
 181 with the following sequence:

$$|7-45-41-56-9-11-7|7-11-9-56-41-45-7|.$$

182 **2) Discontinuous Modulation D6 ϕ SVPWM12-A:** For the
 183 same sector 1, a discontinuous modulation technique (D6 ϕ
 184 SVPWM12-A) can be obtained with the following sequence:

$$|7-45-41-9-11-7|7-11-9-41-45-7|.$$

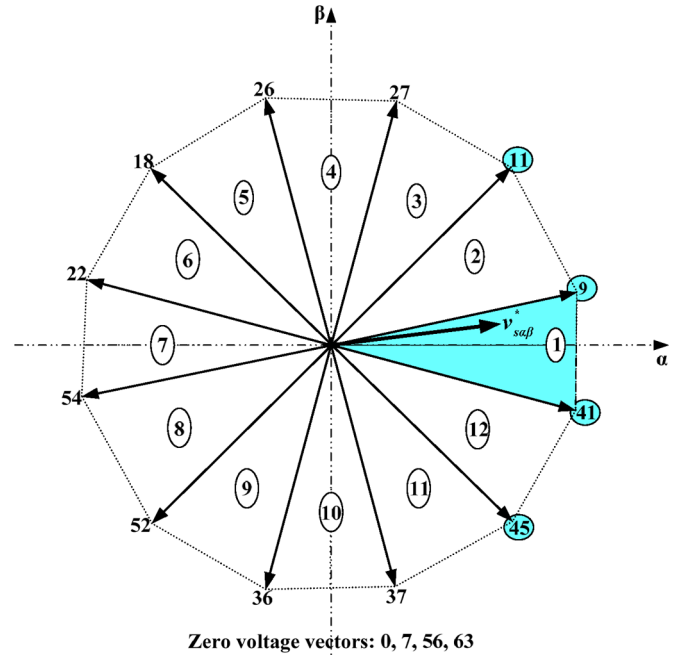


Fig. 3. Presentation of the inverter voltage vectors having maximum magnitude in $(\alpha-\beta)$ plane.

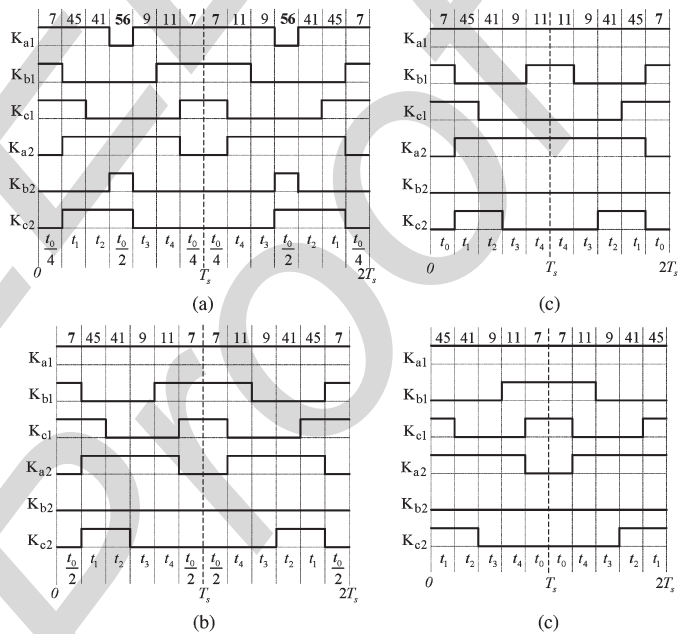


Fig. 4. Twelve-sector SVPWM switching sequences when the reference voltage vector is located in sector 1. (a) C6 ϕ SVPWM12. (b) D6 ϕ SVPWM12-A. (c) D6 ϕ SVPWM12-B1. (d) D6 ϕ SVPWM12-B2.

3) Discontinuous Modulation D6 ϕ SVPWM12-B1: In the 185
 D6 ϕ SVPWM12-B1, the zero-voltage vectors are applied at the 186
 beginning and at the end of the switching sequence as follows: 187

$$|7-45-41-9-11|11-9-41-45-7|.$$

4) Discontinuous Modulation D6 ϕ SVPWM12-B2: In the 188
 D6 ϕ SVPWM12-B2, the zero-voltage vectors are applied in 189
 the middle of the switching sequence as follows: 190

$$|45-41-9-11-7|7-11-9-41-45|.$$

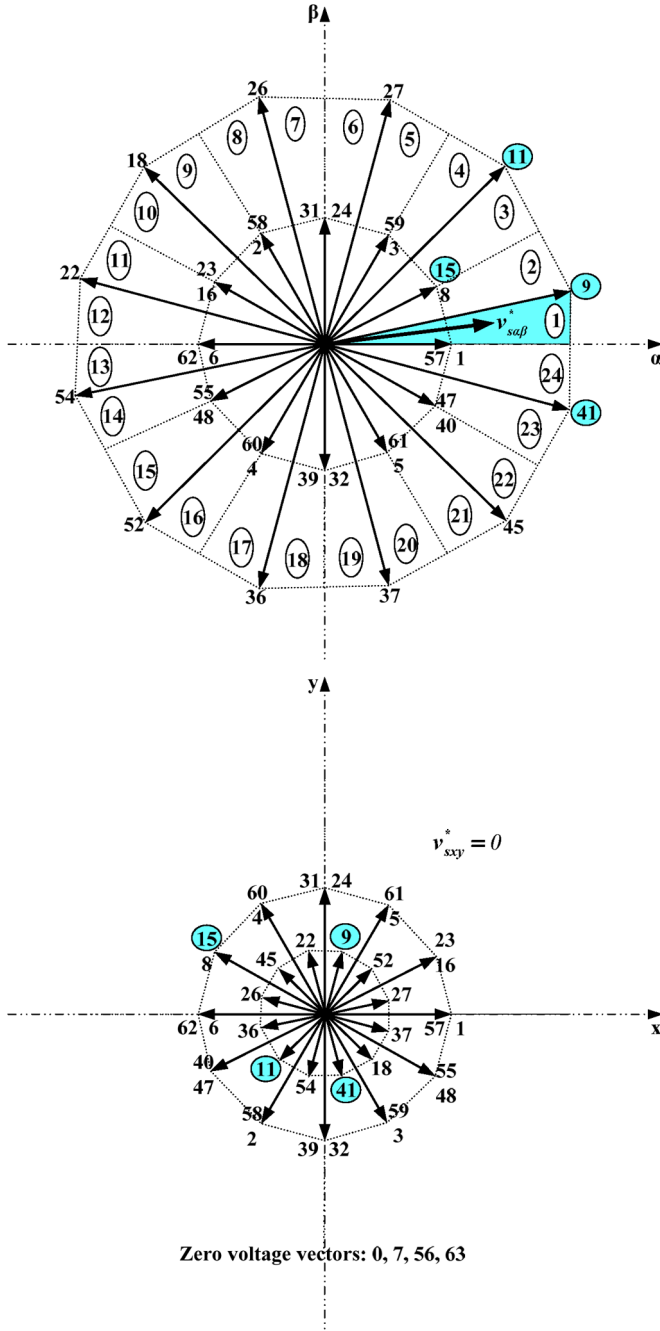


Fig. 5. Presentation of the inverter voltage vectors having maximum and half magnitude in $(\alpha-\beta)$, and $(x-y)$ planes.

191 C. Proposed 24-Sector SVPWM Technique

192 As shown in Fig. 4, the switching pattern corresponding
 193 to the switching sequences of the 12-sector PWM techniques
 194 presents asymmetrical waveforms and usually more than two
 195 transitions (from low to high or from high to low) occur on
 196 the corresponding PWM outputs within a sampling period,
 197 which increases the switching frequency of the inverter legs
 198 and causes difficulties for digital signal processor (DSP) im-
 199 plementation of these strategies. Accordingly, some additional
 200 adaptations in DSP programs are necessary to ensure successful
 201 experiments [15], [16].

202 To overcome these drawbacks, the new 24-sector SVPWM
 203 technique proposed in this paper combines the maximum

magnitude $(\alpha-\beta)$ voltage vectors and the ones with half
 204 magnitude (1, 57, 8, 15, 3, 59, 24, 31, 2, 58, 16, 23, 6, 205
 62, 48, 55, 4, 60, 32, 39, 5, 61, 40, 47) generated by one 206
 inverter. These voltage vectors divide the $(\alpha-\beta)$ plane into 207
 twenty four $\pi/12$ -rad sectors, as shown in Fig. 5. In each 208
 sampling period, the reference voltage vector is achieved by 209
 selecting a set of three voltage vectors among those having 210
 maximum magnitude and a fourth vector among the ones with 211
 half magnitude. For example, voltage vectors 41, 9, 11 and 15 212
 are selected when the reference voltage vector is located in 213
 sector 1. Then, the voltage vectors applying times: t_1, t_2, t_3 and 214
 t_4 are obtained as explained in Appendix A. For the remaining 215
 time $t_0 = T_s - (t_1 + t_2 + t_3 + t_4)$, zero state vectors (0, 7, 56 216
 or 63) are applied. Consequently, simple PWM outputs with 217
 symmetrical waveforms are obtained. As shown in Fig. 6, 218
 only two transitions or less (from low to high or from high 219
 to low) occur on the corresponding PWM outputs within a 220
 sampling period. This fact decreases the switching frequency 221
 of the inverter legs and allows easy DSP implementation. The 222
 switching sequences and the corresponding applying times for 223
 all sectors are presented in Table I. It should also be noticed 224
 that both continuous and discontinuous modulation techniques 225
 can be obtained with this new 24-sector SVPWM scheme. This 226
 is achieved by selecting the appropriate zero voltage vector 227
 locations within the switching sequence. 228

1) *Continuous Modulation C6 ϕ SVPWM24*: Continuous 229
 PWM technique (C6 ϕ SVPWM24) can be obtained for all sec- 230
 tors by selecting the switching sequences presented in Table I. 231
 As an example, when the reference voltage vector is located 232
 in sector 1, a C6 ϕ SVPWM24 is obtained by selecting the 233
 following sequence: 234

$$|56-41-9-11-15-7|7-15-11-9-41-56|.$$

2) *Discontinuous Modulation D6 ϕ SVPWM24-B1 and B2*: 235
 Two discontinuous PWM schemes can be obtained through the 236
 appropriate positioning of the zero voltage vectors. In sector 1, 237
 the examples are as follows. 238

1) *D6 ϕ SVPWM24-B1*: The first discontinuous modulation 239
 technique can be obtained by placing the zero voltage 240
 vector both at the beginning and at the end of the switch- 241
 ing sequence, as follows: 242

$$|56-41-9-11-15|15-11-9-41-56|.$$

2) *D6 ϕ SVPWM24-B2*: The second discontinuous mod- 243
 ulation technique is obtained by placing the zero volt- 244
 age vector in the middle of the switching sequence, as 245
 follows: 246

$$|41-9-11-15-7|7-15-11-9-41|.$$

3) *Switching Sequences and Applying Time Selection*: For 247
 optimal DSP implementation and low algorithm execution time, 248
 the applying times (t_1, t_2, t_3 , and t_4) computation can be 249
 simplified by an offline calculation for all sectors in the same 250
 manner as for sector 1. As a result, within each sampling period, 251
 there is a total of only 12 coefficients T_i to be calculated in (6). 252

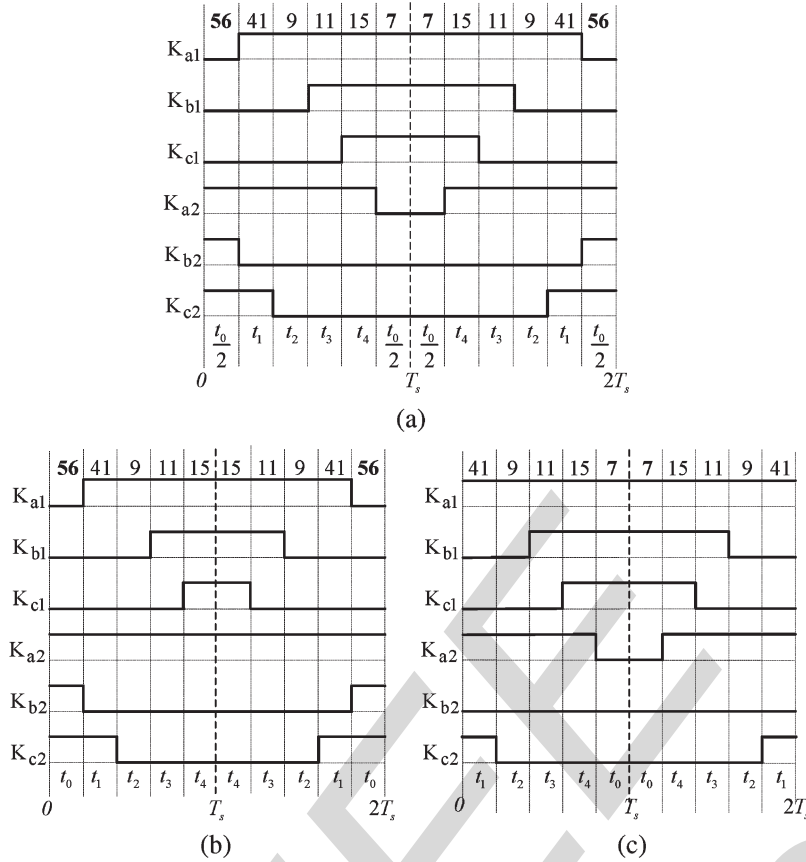


Fig. 6. Twenty-four-sector SVPWM switching sequences when the reference voltage vector is located in sector 1. (a) C6 ϕ SVPWM24. (b) D6 ϕ SVPWM24-B1. (c) D6 ϕ SVPWM24-B2.

253 Table I describes the C6 ϕ SVPWM24 switching sequences and
 254 applying time selection according to the sector number in the
 255 $(\alpha-\beta)$ plane.

$$\begin{bmatrix} T_1 \\ T_2 \\ T_3 \\ T_4 \\ T_5 \\ T_6 \\ T_7 \\ T_8 \\ T_9 \\ T_{10} \\ T_{11} \\ T_{12} \end{bmatrix} = \frac{T_s}{2V_{dc}} \begin{bmatrix} \sqrt{3}-2 & 1 \\ 1 & -\sqrt{3} \\ 1 & \sqrt{3}-2 \\ 0 & 2 \\ \sqrt{3}-1 & \sqrt{3}-1 \\ -(\sqrt{3}-1) & \sqrt{3}-1 \\ \sqrt{3} & -1 \\ 1 & -(\sqrt{3}-2) \\ -(\sqrt{3}-2) & 1 \\ 2 & 0 \\ \sqrt{3} & 1 \\ 1 & \sqrt{3} \end{bmatrix} \begin{bmatrix} v_{s\alpha}^* \\ v_{s\beta}^* \end{bmatrix}. \quad (6)$$

256 D. Maximum Modulation Index

257 The modulation index m can be defined as the ratio of
 258 the fundamental component magnitude of the line to neutral
 259 inverter output voltage V_{1m} to the fundamental component
 260 magnitude of the six-step mode voltage $V_{1m6step} = 2V_{dc}/\pi$
 261 [17]. When the inverter is operating in the linear modulation
 262 region, the sum of the applying times of the active voltage
 263 vectors is less than the switching period T_s [18], [19]. The
 264 largest value for linear output with the 12-sector and the
 265 proposed 24-sector SVPWM techniques, $m_{max} = \pi/(2\sqrt{3}) \approx$

0.907, coincides with the corresponding value for the three-
 266 phase SVPWM [20]. Note that m_{max} is obtained by solving
 267 $t_0 = T_s - (t_1 + t_2 + t_3 + t_4) = 0$ as shown in Appendix B. 268

IV. HARMONIC CURRENT ANALYSIS

269

The voltage and current waveform quality of the PWM-VSI
 270 drives is determined via the switching frequency harmonics,
 271 since they determine the switching frequency copper losses
 272 and the torque ripple of a motor load and the line current
 273 total harmonic distortion of a line-connected VSI. While the
 274 copper losses are measured over a fundamental cycle and
 275 therefore require a per fundamental cycle (macroscopic) rms
 276 ripple current value calculation, the peak and local stresses
 277 are properly investigated on a per-carrier cycle (microscopic)
 278 basis. Therefore, first a microscopic and then a macroscopic
 279 investigation is required [20]. Because, the machine model
 280 includes $(\alpha-\beta)$ and $(x-y)$ components, the harmonic current
 281 analysis must be made for the $(\alpha-\beta)$ and $(x-y)$ currents. 282

A. Normalized Harmonic Currents and Fluxes Calculation

283

The stator voltage equations in the stator coordinate system
 284 are expressed as follows: 285

$$\begin{aligned} v_{s\alpha\beta} &= R_s i_{s\alpha\beta} + \frac{d\lambda_{s\alpha\beta}}{dt} \\ v_{sxy} &= R_s i_{sxy} + L_{lsxy} \frac{di_{sxy}}{dt} \end{aligned} \quad (7)$$

TABLE I
PROPOSED C6 ϕ SVPWM24 SWITCHING SEQUENCES

Sector	Switching sequences	Voltage vectors applying times			
		t_1	t_2	t_3	t_4
1	56-41-9-11-15-7	T_2	T_5	T_4	$-T_1$
2	56-57-41-9-11-7	T_1	T_2	T_3	T_4
3	0-9-11-27-59-63	T_7	T_9	$-T_2$	$-T_6$
4	0-8-9-11-27-63	T_6	T_7	T_8	$-T_2$
5	7-11-27-26-24-56	T_{10}	T_1	$-T_7$	T_3
6	7-3-11-27-26-56	$-T_3$	T_{10}	T_5	$-T_7$
7	63-27-26-18-2-0	T_{11}	T_6	$-T_{10}$	T_8
8	63-31-27-26-18-0	$-T_8$	T_{11}	T_9	$-T_{10}$
9	56-26-18-22-23-7	T_{12}	$-T_3$	$-T_{11}$	T_5
10	56-58-26-18-22-7	$-T_5$	T_{12}	T_1	$-T_{11}$
11	0-18-22-54-62-63	T_4	$-T_8$	$-T_{12}$	T_9
12	0-16-18-22-54-63	$-T_9$	T_4	T_6	$-T_{12}$
13	7-22-54-52-48-56	$-T_2$	$-T_5$	$-T_4$	T_1
14	7-6-22-54-52-56	$-T_1$	$-T_2$	$-T_3$	$-T_4$
15	63-54-52-36-4-0	$-T_7$	$-T_9$	T_2	T_6
16	63-55-54-52-36-0	$-T_6$	$-T_7$	$-T_8$	T_2
17	56-52-36-37-39-7	$-T_{10}$	$-T_1$	T_7	$-T_3$
18	56-60-52-36-37-7	T_3	$-T_{10}$	$-T_5$	T_7
19	0-36-37-45-61-63	$-T_{11}$	$-T_6$	T_{10}	$-T_8$
20	0-32-36-37-45-63	T_8	$-T_{11}$	$-T_9$	T_{10}
21	7-37-45-41-40-56	$-T_{12}$	T_3	T_{11}	$-T_5$
22	7-5-37-45-41-56	T_5	$-T_{12}$	$-T_1$	T_{11}
23	63-45-41-9-1-0	$-T_4$	T_8	T_{12}	$-T_9$
24	63-47-45-41-9-0	T_9	$-T_4$	$-T_6$	T_{12}

286 where the stator and the rotor flux equations are given by:

$$\begin{aligned}\lambda_{s\alpha\beta} &= L_s i_{s\alpha\beta} + M i_{r\alpha\beta} \\ \lambda_{r\alpha\beta} &= L_r i_{r\alpha\beta} + M i_{s\alpha\beta}.\end{aligned}\quad (8)$$

287 The stator flux equation can be rewritten as:

$$\lambda_{s\alpha\beta} = \sigma L_s i_{s\alpha\beta} + \frac{M}{L_r} \lambda_{r\alpha\beta}.\quad (9)$$

288 Substituting (9) in (7), the stator voltage equation can be
289 expressed as follows:

$$v_{s\alpha\beta} = R_s i_{s\alpha\beta} + \sigma L_s \frac{di_{s\alpha\beta}}{dt} + \frac{M}{L_r} \frac{d\lambda_{r\alpha\beta}}{dt}.\quad (10)$$

290 If only the harmonic voltages and currents are considered,
291 it will be assumed that the reference voltage vector $v_{s\alpha\beta}^*$ is
292 constant over the switching period T_s , because the switching
293 frequency f_s is much higher than the fundamental frequency
294 f_e , and that the stator and the rotor time constants are much
295 larger than the switching period, with the resistance drops being
296 neglected [21]. Under these assumptions, the voltages and

currents can be separated in the harmonic components, which
change over T_s while the fundamental components remain
constant over the same period. Thus, from (7) and (10), the
harmonic voltage equations can be expressed as follows: 300

$$\begin{aligned}\tilde{v}_{s\alpha\beta} &= \sigma L_s \frac{d\tilde{i}_{s\alpha\beta}}{dt} \\ \tilde{v}_{sxy} &= L_{lsxy} \frac{d\tilde{i}_{sxy}}{dt}\end{aligned}\quad (11)$$

where $\tilde{v}_{s\alpha\beta}$ is the harmonic voltage and is equal to the
difference between the actual voltage vector and the reference
vector $v_{s\alpha\beta}^*$. 303

Assuming that the instantaneous harmonic currents are zero
at the beginning and at the end of the carrier cycle, the
($\alpha-\beta$) and ($x-y$) harmonic stator currents per-carrier cycle can be
calculated as follows [20], [22], [23]: 307

$$\begin{aligned}\tilde{i}_{s\alpha\beta} &= \frac{1}{\sigma L_s} \int_{NT_s}^{(N+1)T_s} (V_{s\alpha\beta k} - v_{s\alpha\beta}^*) dt \\ \tilde{i}_{sxy} &= \frac{1}{L_{lsxy}} \int_{NT_s}^{(N+1)T_s} (V_{sxy k}) dt.\end{aligned}\quad (12)$$

In (12), $V_{s\alpha\beta k}$ and $V_{sxy k}$ are the inverter output voltage
vectors of the k th state. They change according to the selected
switching sequence, since for high f_s/f_e values, the $v_{s\alpha\beta}^*$ term
can be assumed as constant within a carrier cycle. Thus, the
above integral can be calculated in a closed form. 312

Because the harmonic current and harmonic flux are only
different in scale, and in order to eliminate the need for
load parameters in (12), the harmonic flux trajectories can
be investigated. Nevertheless, the ($x-y$) current components
are limited by the stator leakage inductance L_{lsxy} , which de-
pends on the coil pitch of the stator windings [10]. Conse-
quently, the harmonic characteristics of the VSI feeding DSIM
should be investigated with the introduction of the coefficient
 $k_{\sigma xy} = \sigma L_s / L_{lsxy}$, which is necessary to evaluate and com-
pare the performances of the PWM techniques. So, employ-
ing (12) and normalizing with respect to λ_b , the per-carrier
cycle rms value of the normalized harmonic current can be
calculated with: 325

$$\begin{aligned}\tilde{i}_{srms}^2(m, \theta) &= \left(\frac{\lambda_b}{\sigma L_s}\right)^2 \frac{1}{T_s} \int_{NT_s}^{(N+1)T_s} (\tilde{\lambda}_{s\alpha\beta}^2 + k_{\sigma xy}^2 \tilde{\lambda}_{sxy}^2) dt \\ &= \left(\frac{\lambda_b}{\sigma L_s}\right)^2 (\tilde{\lambda}_{s\alpha\beta rms}^2(m, \theta) + k_{\sigma xy}^2 \tilde{\lambda}_{sxy rms}^2(m, \theta)) \\ &= \left(\frac{\lambda_b}{\sigma L_s}\right)^2 (\tilde{\lambda}_{srms}^2(m, \theta))\end{aligned}\quad (13)$$

where $\lambda_b = 2\sqrt{3}V_{dc}T_s/\pi$, $\tilde{\lambda}_{s\alpha\beta} = \sigma L_s \tilde{i}_{s\alpha\beta}$, and $\tilde{\lambda}_{sxy} =$
 $L_{lsxy} \tilde{i}_{sxy}$ 327

TABLE II
SWITCHING FREQUENCY REDUCTION COEFFICIENT

N°	SVPWM Techniques	k_f	f_s
1	C6 ϕ SVPWM12	1	
2	D6 ϕ SVPWM12-A	2/3	
3	D6 ϕ SVPWM12-B1	1/2	
4	D6 ϕ SVPWM12-B2	5/12	$f_s = \frac{1}{k_f} \cdot f_{sw}$
5	C6 ϕ SVPWM24	1	
6	D6 ϕ SVPWM24-B1	5/6	
7	D6 ϕ SVPWM24-B2	2/3	

The per-fundamental cycle rms value of the harmonic current determines the waveform quality and harmonic losses. Averaging (13) over a fundamental period results in the global harmonic current calculation as follows:

$$\begin{aligned}
 \tilde{I}_{sfrms}^2(m) &= \left(\frac{\lambda_b}{\sigma L_s} \right)^2 \frac{1}{2\pi} \int_{2\pi} \tilde{\lambda}_{sfrms}^2(m, \theta) d\theta \\
 &= \left(\frac{\lambda_b}{\sigma L_s} \right)^2 \left(\tilde{\lambda}_{s\alpha\beta frms}^2(m) + k_{\sigma xy}^2 \tilde{\lambda}_{sxy frms}^2(m) \right) \\
 &= \left(\frac{\lambda_b}{\sigma L_s} \right)^2 \tilde{\lambda}_{sfrms}^2(m). \quad (14)
 \end{aligned}$$

The above integral yields a polynomial function of the modulation index m . As an example, the per-fundamental cycle rms normalized harmonic flux $\tilde{\lambda}_{sfrms}$ was calculated for all the discussed PWM techniques. This results in m dependent analytical formulas summarized in Appendix C.

B. Performance Comparison

For comparison purposes, the harmonic current analysis is performed at the same average switching frequency f_{sw} for all the PWM techniques. Therefore, a switching frequency reduction coefficient k_f is introduced for each PWM technique according to Table II. This coefficient can be determined from the ratio of the discontinuous to the continuous PWM techniques regarding the number of commutations of all legs during one sampling period. The curves of the per-fundamental cycle normalized rms harmonic flux for all the discussed PWM techniques have been plotted as a function of modulation index m , as shown in Fig. 7. It is clear that the rms value of the harmonic flux varies with the PWM technique used and according to the selected switching sequences. These curves show that the D6 ϕ SVPWM12-A has practically the best performance at a low modulation index range, while the D6 ϕ SVPWM12-B1 (B2) exhibits the best performance in the high modulation index range. However, as the modulation index increases, the C6 ϕ SVPWM12 performance rapidly degrades compared to the C6 ϕ SVPWM24 of the proposed 24-sector PWM scheme, which reveals excellent performance over the whole voltage range. While the D6 ϕ SVPWM24-B1-(B2) PWM strategies present harmonic characteristics similar to the ones obtained with the D6 ϕ SVPWM12-A-(B1) and (B2) strategies, the

proposed 24-sector PWM techniques allow a sampling frequency increase and as a result, the switching frequency can be increased by a factor of two as compared to the 12-sector PWMs. Therefore, significant harmonic current reductions can be achieved as shown in Fig. 7(h) for the worst case, when $k_{\sigma xy} = 10$.

It should be noted that discontinuous PWM techniques allow a higher sampling rate selection and can be applied in a high voltage range, while continuous ones are advantageous in the low voltage range. In addition, an optimal PWM scheme can be obtained with a transition between these SVPWM strategies to allow rms harmonic current minimization over the whole voltage range. The intersection points define the optimal transition between different PWM techniques.

Finally, a careful look at D6 ϕ SVPWM12-B1 in Fig. 4(c) shows that during one switching period T_s , two inverter legs have commutations twice. D6 ϕ SVPWM12-B2 in Fig. 4(d) shows that one inverter leg has commutations twice. This is never the case for D6 ϕ SVPWM24-B1 and B2. Therefore, 12-sector discontinuous PWMs have a maximum instantaneous switching frequency twice as big as in the case of 24-sector discontinuous PWMs. This fact has an impact on inverter switching losses [14], [20]. While it is not the aim of this paper to evaluate the performance of PWM techniques in terms of inverter switching and conduction losses, it can be expected that the 24-sector discontinuous PWMs will have a better performance than the 12-sector ones. Further works on switching and conduction losses are in progress and will be reported in a subsequent paper.

V. EXPERIMENTAL RESULTS

To confirm the feasibility of the proposed SVPWM techniques over the entire voltage range under V/f control, a set of experiments are carried out. The experimental test bench (Fig. 8) is composed of a six-phase VSI feeding a 15 kW DSIM prototype, and the whole control algorithm is tested on a dSPACE DS1104 controller board. The original 320F240 firmware does not allow the change in PWM compare registers and action registers many times during a period. So, to allow four changes within PWM period T_s , the flashed firmware is reprogrammed [15], [24]. Hence, it is possible to implement the 12-sector PWM techniques.

On the contrary, for the proposed 24-sector PWM strategies, usually at most two transitions (from low to high or from high to low) occur symmetrically on the corresponding PWM outputs within a sampling period. This allows easier DSP implementation. Then the original 320F240 firmware is used with the help of specially developed user functions that allow the synchronization of six full PWM and three simple PWM simultaneously on the DS1104 DSP board.

These PWM techniques are successfully tested and the following conclusions can be drawn from these experimental results: In the case of the 12-sector PWM strategies, usually more than two transitions (from low to high or from high to low) occur on the corresponding PWM outputs. This increases the switching frequency of the inverter legs and complicates the experimental test. However, the number of transitions

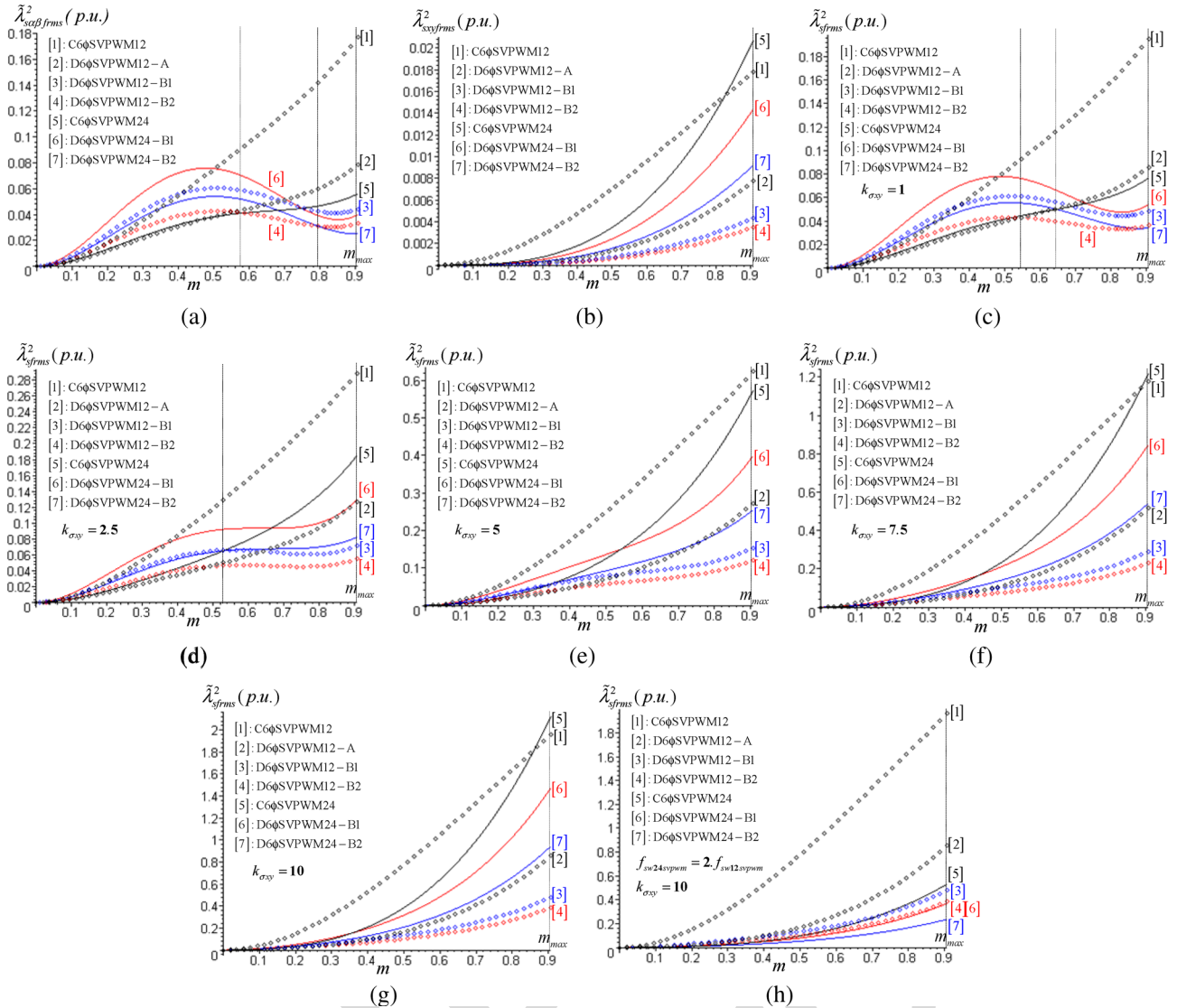


Fig. 7. Per fundamental cycle normalized rms harmonic flux as a function of modulation index m , for all the discussed PWM techniques. (a) $(\alpha-\beta)$ rms harmonic flux. (b) $(x-y)$ rms harmonic flux. (c), (d), (e), (f), (g) RMS harmonic flux at different leakage coupling ($k_{\sigma xy} = 1; 2.5; 5; 7.5; 10$), at the same average switching frequency f_{sw} . (h) RMS harmonic flux with $k_{\sigma xy} = 10$, at $f_s = 2 \cdot f_{sw}$ for the proposed 24-sector PWM techniques.

417 on the corresponding PWM outputs is reduced in the case
418 of the proposed 24-sector PWM strategies, which decrease
419 the switching frequency of the inverter legs and allow easy
420 DSP implementation. In addition, for discontinuous PWM tech-
421 niques, at least two PWM outputs remain unchanged during the
422 entire sampling period. Therefore, the carrier frequency can be
423 increased with harmonic losses reduction.

424 Experimental results under constant V/f control, for the
425 cases of C6 ϕ SVPWM12, D6 ϕ SVPWM12-B2, C6 ϕ
426 SVPWM24, and D6 ϕ SVPWM24-B2 techniques are pre-
427 sented in Fig. 9. The average switching frequency is set to
428 $f_{sw} = 5$ kHz and the motor is running at 735 r/min with a
429 connected load. As expected, these SVPWM techniques al-
430 low the control of the $(\alpha-\beta)$ and $(x-y)$ current components
431 simultaneously.

432 However, these experimental results demonstrate that the
433 continuous PWM techniques produce a larger amplitude of
434 the harmonic currents in the $(x-y)$ plane as compared to
435 the discontinuous ones where the amplitude of these currents

is minimized. Moreover, the phase current presents a pure
436 sinusoidal shape and the trajectory of the $(\alpha-\beta)$ stator cur-
437 rent components is a circle for all these PWM techniques,
438 confirming that these currents are controlled as well as $(x-y)$
439 harmonic currents. Indeed, it should be remembered that the
440 carrier frequency can be increased by a factor of two for a 50%
441 reduction of harmonic losses in the case of C6 ϕ SVPWM24,
442 or by a factor three for a 66% reduction in harmonic losses
443 with the D6 ϕ SVPWM24-B2 technique. In addition, due
444 to the simplicity and regularity of the proposed 24-sector
445 PWM techniques, low-cost DSPs for motor control may easily
446 be used. 447

VI. CONCLUSION

448
449 In this paper, a new SVPWM technique based on the vector
450 space decomposition suitable for six-phase VSI-fed DSIM has
451 been presented. The switching sequences presented lead to
452 continuous and discontinuous modulation strategies, according

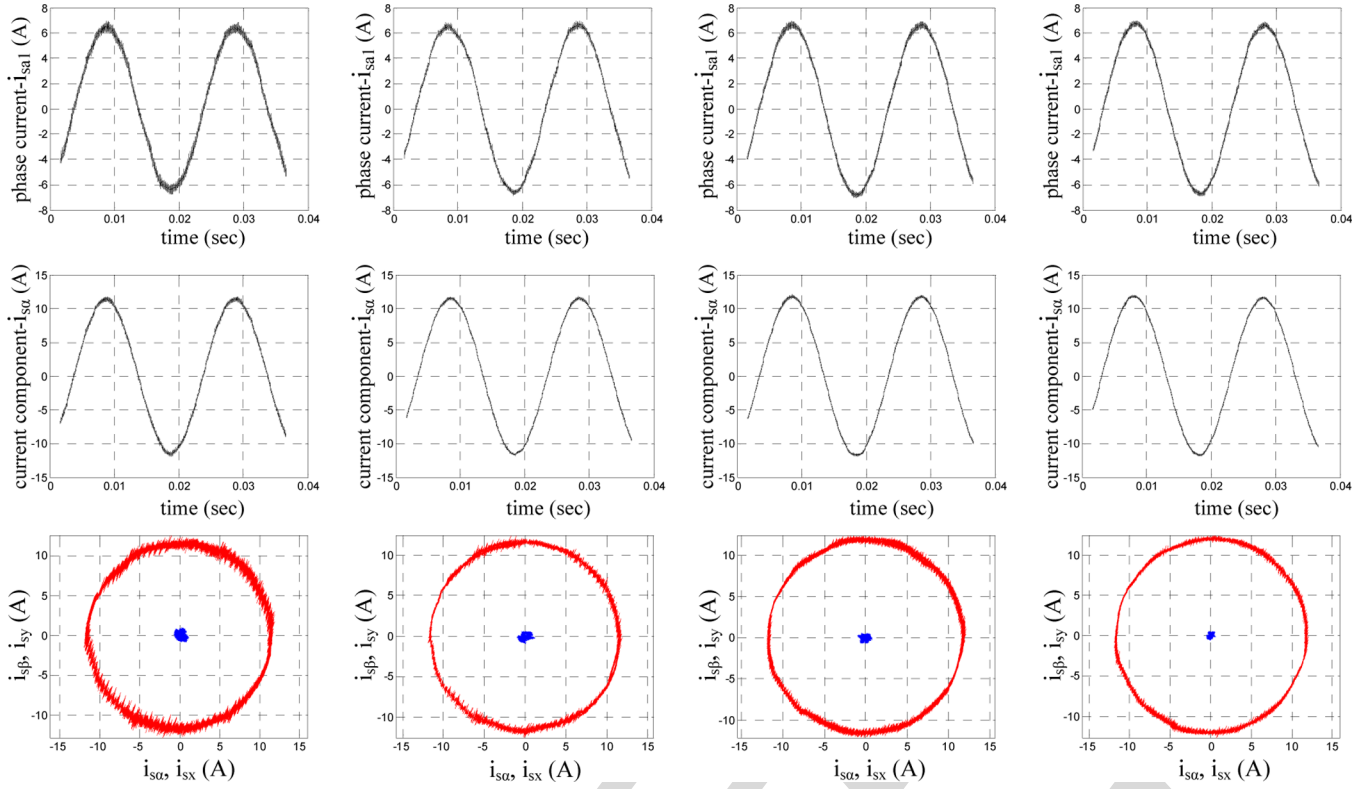


Fig. 8. Experimental test bench.

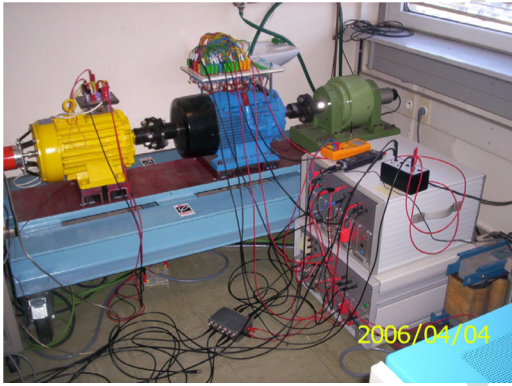


Fig. 9. Experimental results of the SVPWM techniques with the motor operating under a constant V/f control with connected load for $f_e = 50$ Hz, at 735 r/min and for the same average switching frequency f_{sw} . From top to bottom: i_{sa1} phase current, $i_{s\alpha}$ current component, $(\alpha-\beta)$ and $(x-y)$ plane current trajectories. From left to right: C6 ϕ SVPWM12, D6 ϕ SVPWM12-B2, C6 ϕ SVPWM24, and D6 ϕ SVPWM24-B2.

453 to the position of zero voltage vectors during each sampling
454 period. It is shown that the harmonic current rms values vary
455 according to the selected switching sequence and the voltage
456 range. Likewise, from this point of view, the continuous PWM
457 technique has an advantage in the low and medium voltage
458 range, while the discontinuous PWM strategy is advantageous
459 in the high voltage range. Thus, the combination of these
460 strategies provides the best harmonic current performance over
461 the whole voltage range. It has been demonstrated that the pro-
462 posed 24-sector SVPWM techniques, while easy to implement
463 digitally, allow a switching frequency increase with significant
464 extra stator harmonic currents reduction.

APPENDIX A

465

VOLTAGE VECTORS APPLYING TIMES CALCULATION

466

The inverter output voltage vectors are represented in Figs. 3, 467
and Fig. 5 by decimal number k equivalent to the binary number 468
formed by the instantaneous values of the switching functions 469
defined as: 470

$$k = K_{a1} \times 2^0 + K_{b1} \times 2^1 + K_{c1} \times 2^2 + K_{a2} \times 2^3 + K_{b2} \times 2^4 + K_{c2} \times 2^5. \quad (A1)$$

The instantaneous values of the six-phase VSI output voltage 471
vectors $(v_{a1}, v_{b1}, v_{c1}, v_{a2}, v_{b2}, v_{c2})$ can be determined by using 472
the inverter connection matrix $[Mc]$ as follows: 473

$$[v_{a1} \ v_{b1} \ v_{c1} \ v_{a2} \ v_{b2} \ v_{c2}]^T = [Mc][K_{a1} \ K_{b1} \ K_{c1} \ K_{a2} \ K_{b2} \ K_{c2}]^T \quad (A2)$$

where

474

$$[Mc] = \frac{V_{dc}}{3} \begin{bmatrix} 2 & -1 & -1 & 0 & 0 & 0 \\ -1 & 2 & -1 & 0 & 0 & 0 \\ -1 & -1 & 2 & 0 & 0 & 0 \\ \hline 0 & 0 & 0 & 2 & -1 & -1 \\ 0 & 0 & 0 & -1 & 2 & -1 \\ 0 & 0 & 0 & -1 & -1 & 2 \end{bmatrix}. \quad (A3)$$

The inverter output voltage vectors are transformed into 475
 $(\alpha-\beta)$, $(x-y)$, and (o_1-o_2) planes by means of the transfor- 476
mation matrix $[Ts]^{-1}$ given in (1), as: 477

$$[v_{s\alpha k} \ v_{s\beta k} \ v_{sxk} \ v_{syk} \ v_{o1k} \ v_{o2k}]^T = [Ts]^{-1}[v_{a1} \ v_{b1} \ v_{c1} \ v_{a2} \ v_{b2} \ v_{c2}]^T \quad (A4)$$

478 For example, when the reference voltage vector is located in
479 sector 1, voltage vectors 41, 9, 11, and 15 are selected and their
480 equivalent binary numbers are defined as:

$$\begin{aligned} 41 &= [1 \ 0 \ 1|0 \ 0 \ 1] \\ 9 &= [0 \ 0 \ 1|0 \ 0 \ 1] \\ 11 &= [0 \ 0 \ 1|0 \ 1 \ 1] \\ 15 &= [0 \ 0 \ 1|1 \ 1 \ 1]. \end{aligned} \quad (\text{A5})$$

481 Then the $(\alpha-\beta)$, and $(x-y)$ voltages can be calculated by
482 using (A2), and (A4), as follows:

$$\begin{aligned} &\begin{bmatrix} V_{s\alpha 41} & V_{s\alpha 9} & V_{s\alpha 11} & V_{s\alpha 15} \\ V_{s\beta 41} & V_{s\beta 9} & V_{s\beta 11} & V_{s\beta 15} \\ V_{sx 41} & V_{sx 9} & V_{sx 11} & V_{sx 15} \\ V_{sy 41} & V_{sy 9} & V_{sy 11} & V_{sy 15} \end{bmatrix} \\ &= \frac{V_{dc}}{2\sqrt{3}} \begin{bmatrix} 2 + \sqrt{3} & 2 + \sqrt{3} & 1 + \sqrt{3} & \sqrt{3} \\ -1 & 1 & 1 + \sqrt{3} & 1 \\ 2 - \sqrt{3} & 2 - \sqrt{3} & 1 - \sqrt{3} & -\sqrt{3} \\ -1 & 1 & 1 - \sqrt{3} & 1 \end{bmatrix}. \end{aligned} \quad (\text{A6})$$

483 The voltage vectors applying times: t_1 , t_2 , t_3 and t_4 are
484 obtained as:

$$\begin{aligned} \begin{bmatrix} t_1 \\ t_2 \\ t_3 \\ t_4 \end{bmatrix} &= \begin{bmatrix} V_{s\alpha 41} & V_{s\alpha 9} & V_{s\alpha 11} & V_{s\alpha 15} \\ V_{s\beta 41} & V_{s\beta 9} & V_{s\beta 11} & V_{s\beta 15} \\ V_{sx 41} & V_{sx 9} & V_{sx 11} & V_{sx 15} \\ V_{sy 41} & V_{sy 9} & V_{sy 11} & V_{sy 15} \end{bmatrix}^{-1} \begin{bmatrix} v_{s\alpha}^* T_s \\ v_{s\beta}^* T_s \\ v_{sx}^* T_s \\ v_{sy}^* T_s \end{bmatrix} \\ t_0 &= T_s - (t_1 + t_2 + t_3 + t_4). \end{aligned} \quad (\text{A7})$$

485 Substituting (A6) in (A7), the applying times can be calcu-
486 lated as follows:

$$\begin{aligned} \begin{bmatrix} t_1 \\ t_2 \\ t_3 \\ t_4 \end{bmatrix} &= \frac{T_s}{2V_{dc}} \begin{bmatrix} 1 & -\sqrt{3} & -1 & -\sqrt{3} \\ \sqrt{3}-1 & \sqrt{3}-1 & \sqrt{3}+1 & \sqrt{3}+1 \\ 0 & 2 & 0 & v-2 \\ -(\sqrt{3}-2) & -1 & -(\sqrt{3}+2) & 1 \end{bmatrix} \begin{bmatrix} v_{s\alpha}^* \\ v_{s\beta}^* \\ v_{sx}^* \\ v_{sy}^* \end{bmatrix}. \end{aligned} \quad (\text{A8})$$

487 With $v_{sx}^* = v_{sy}^* = 0$, (A8) can be written as:

$$\begin{aligned} \begin{bmatrix} t_1 \\ t_2 \\ t_3 \\ t_4 \end{bmatrix} &= \frac{T_s}{2V_{dc}} \begin{bmatrix} 1 & -\sqrt{3} \\ \sqrt{3}-1 & \sqrt{3}-1 \\ 0 & 2 \\ -(\sqrt{3}-2) & -1 \end{bmatrix} \begin{bmatrix} v_{s\alpha}^* \\ v_{s\beta}^* \end{bmatrix} = \begin{bmatrix} T_2 \\ T_5 \\ T_4 \\ -T_1 \end{bmatrix}. \end{aligned} \quad (\text{A9})$$

488 APPENDIX B 489 MAXIMUM MODULATION INDEX CALCULATION

490 The maximum modulation index m_{\max} can be obtained
491 by solving $t_0 = T_s - (t_1 + t_2 + t_3 + t_4) = 0$. For example in
492 sector 1, the sum of the applying times of the active voltage
493 vectors is calculated from (A9) as:

$$\theta \in \left[0, \frac{\pi}{12}\right] \quad t_1 + t_2 + t_3 + t_4 = \frac{T_s}{V_{dc}} v_{s\alpha}^* \quad (\text{B1})$$

494 where $v_{s\alpha}^* = \sqrt{3}V_{1m} \cos(\theta)$, $v_{s\beta}^* = \sqrt{3}V_{1m} \sin(\theta)$

$$V_{1m} = mV_{1m6\text{step}} = m2V_{dc}/\pi.$$

When

$$t_0 = 0 : T_s = t_1 + t_2 + t_3 + t_4 = 2\sqrt{3}\frac{T_s}{\pi}m \cos(\theta) \quad (\text{B2})$$

From (B2), the modulation index equation can be given as: 496

$$m = \frac{\pi}{2\sqrt{3} \cos(\theta)}. \quad (\text{B3})$$

To determine the angle θ corresponding to m_{\max} , (B3) is 497
derived: 498

$$\frac{dm}{d\theta} = \frac{\pi \sin(\theta)}{2\sqrt{3} \cos^2(\theta)}. \quad (\text{B4})$$

Equation (B4) is solved for $\theta = 0$. Thus, replacing θ in (B3): 499

$$m_{\max} = \frac{\pi}{2\sqrt{3} \cos(0)} = \frac{\pi}{2\sqrt{3}} \approx 0.907. \quad (\text{B5})$$

APPENDIX C ANALYTICAL FORMULAS OF THE RMS HARMONIC FLUX 501

The per-fundamental cycle rms normalized harmonic flux 502
 $\tilde{\lambda}_{s\text{frms}}$ was calculated for all the discussed PWM techniques. 503
Only the analytical formulas of the proposed 24-sector PWMs 504
are presented here and the ones of the 12-sector PWMs can be 505
found in [10]. These formulas are given below. 506

A. Continuous Modulation C6 ϕ SVPWM24 507

$$\begin{aligned} \tilde{\lambda}_{s\alpha\beta\text{frms}}^2(m) &= \frac{1}{48}m^2 + \frac{1}{144\pi^2} \\ &\quad \times (56\sqrt{3} + 63\sqrt{6} - 57\sqrt{2} - 228)m^3 \\ &\quad + \frac{1}{32\pi^3}(24\pi + 27 - 21\sqrt{3} - 8\sqrt{3}\pi)m^4 \\ \tilde{\lambda}_{sxy\text{frms}}^2(m) &= \frac{1}{144\pi^2}(63\sqrt{6} + 18 - 52\sqrt{3} - 57\sqrt{2})m^3. \end{aligned} \quad (\text{C1})$$

B. Discontinuous Modulation D6 ϕ SVPWM24-B1 508

$$\begin{aligned} \tilde{\lambda}_{s\alpha\beta\text{frms}}^2(m) &= \frac{25}{432}m^2 - \frac{25}{5184\pi^2} \\ &\quad \times (633\sqrt{2} + 408 - 56\sqrt{3} - 387\sqrt{6})m^3 \\ &\quad - \frac{25}{576\pi^3}(15\sqrt{3} + 8\sqrt{3}\pi - 24\pi - 45)m^4 \\ \tilde{\lambda}_{sxy\text{frms}}^2(m) &= \frac{25}{5184\pi^2}(63\sqrt{6} + 18 - 52\sqrt{3} - 57\sqrt{2})m^3. \end{aligned} \quad (\text{C2})$$

C. Discontinuous Modulation D6 ϕ SVPWM24-B2 509

$$\begin{aligned} \tilde{\lambda}_{s\alpha\beta\text{frms}}^2(m) &= \frac{1}{27}m^2 - \frac{1}{324\pi^2} \\ &\quad \times (129\sqrt{2} + 45\sqrt{6} + 48 - 56\sqrt{3})m^3 \\ &\quad + \frac{1}{6\pi^3}(2\pi + 3 - \sqrt{3})m^4 \\ \tilde{\lambda}_{sxy\text{frms}}^2(m) &= \frac{1}{324\pi^2}(63\sqrt{6} + 18 - 52\sqrt{3} - 57\sqrt{2})m^3. \end{aligned} \quad (\text{C3})$$

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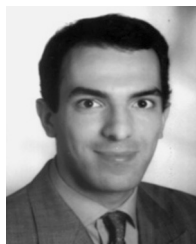
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AUTHOR PLEASE ANSWER ALL QUERIES

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A New PWM Strategy Based on a 24-Sector Vector Space Decomposition for a Six-Phase VSI-Fed Dual Stator Induction Motor

Khoudir Marouani, Lotfi Baghli, Djafar Hadiouche, Abdelaziz Kheloui, and Abderrezak Rezzoug

Abstract—This paper presents a new space vector pulsewidth modulation (SVPWM) technique for the control of six-phase voltage source inverter (VSI)-fed dual stator induction machines (DSIM). A DSIM is an induction machine which has two sets of three-phase stator windings spatially shifted by 30 electrical degrees and fed by two three-phase VSIs. Despite their advantage of power segmentation, these machines are characterized by large zero sequence harmonic currents, and in particular those of order $6k \pm 1$, which are due to the mutual cancellation between the two stator windings. The proposed SVPWM scheme, while easy to implement digitally, reduces significantly these extra stator harmonic currents. Experimental results, collected from a 15 kW prototype machine controlled by a digital signal processor are presented and discussed.

Index Terms—Dual stator induction machines (DSIM), six-phase voltage source inverter (VSI), space vector pulsewidth modulation (SVPWM).

I. INTRODUCTION

NOWADAYS, electrical machine drives are widely used in industrial applications and transportation systems such as electric/hybrid vehicles, traction locomotives and electric propulsion ships, where high-power levels in conjunction with high-performance requirements are more and more demanded. To achieve these high ratings, there are two possible approaches; one focuses on the converter side by increasing the number of output voltage levels and the other one on the machine side by increasing the number of phases. In the first approach, the idea is to divide the high dc bus voltage into multiple low levels and therefore to distribute the high power required among cells of reduced-voltage power switches without the problem of dynamic voltage sharing encountered in the series connection of active devices. However, increasing the number of inverter levels adds to the control complexity and may introduce some voltage imbalance problems [1]–[3]. It is a solution well suited for high-power and high-voltage utility ap-

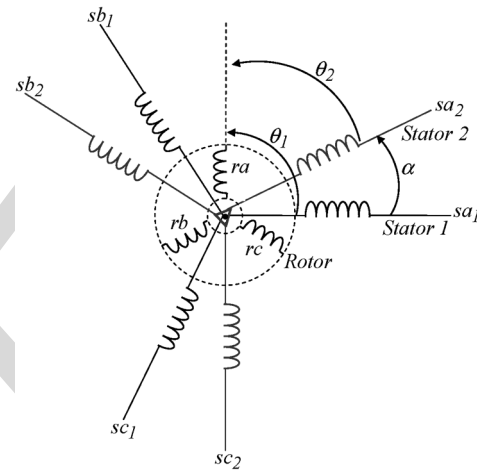


Fig. 1. DSIM windings.

lications. For adjustable speed drives, however, an alternative approach is to use a multiphase machine, i.e., a machine with more than three phases in the stator, since the number of phases is not imposed anyway, given that the machine is connected to the electric supply through a dc/ac converter. The number of phases could be used instead as an additional degree of freedom in the overall system design [4].

Although the answer to the question whether it is better to use a multilevel inverter-fed three-phase machine or a multiphase machine depends on the application, it is undeniable that the latter option offers several advantages which may make it appear very attractive. In fact, the most significant features are a low torque ripple, a reduction in the power per phase and fault-tolerance capability. Other interesting advantages can be pointed out, such as a better torque production per ampere for the same machine volume, higher efficiency and improved reliability [5], [6].

A common type of multiphase machine is the dual stator induction machine (DSIM), where two sets of three-phase windings, spatially phase shifted by 30 electrical degrees, share a common stator magnetic core as shown in Fig. 1.

Due to the development of fast switching power semiconductor devices, voltage source inverters (VSIs) are preferred in variable speed machine drives. As VSI-fed multiphase machines are gaining increasing interest for high-power applications, various pulsewidth modulation (PWM) techniques have been developed accordingly, as they strongly affect the overall inverter efficiency and output voltage waveform quality.

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68 In a VSI-fed DSIM, the two stator windings are mutually
69 coupled and small unbalances in the two supply voltages may
70 generate high currents [7]. Furthermore, because of the low
71 impedance seen by the voltage harmonic components generated
72 by the switched voltage waveforms, harmonic currents of high
73 level are circulating uselessly in the two stator windings, adding
74 to the overall losses and therefore to the semiconductor devices
75 ratings [8], [9].

76 To minimize these extra harmonic currents in a six-phase
77 VSI-fed DSIM, a new 24-sector PWM technique is proposed
78 in this paper and tested on a 15 kW laboratory machine. The
79 digital implementation is carried out on a DS1104 dSPACE
80 controller board. A comparative study between the proposed
81 technique and similar space vector PWM (SVPWM) techniques
82 [5], [10], based on analytical harmonic current analysis, is also
83 developed and discussed.

84 II. MACHINE MODEL

85 The machine model is based on the assumption that space
86 harmonics and magnetic saturation are negligible, and that the
87 two stator three-phase windings are identical and symmetrical
88 with the two neutrals being isolated. In order to derive a prac-
89 tical model suitable for control, a decoupling transformation
90 matrix is used, as proposed in [5]–[7]. The matrix has the
91 following form:

$$[T_S]^{-1} = \frac{1}{\sqrt{3}} \begin{bmatrix} 1 & -\frac{1}{2} & -\frac{1}{2} & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} & 0 \\ 0 & \frac{\sqrt{3}}{2} & -\frac{\sqrt{3}}{2} & \frac{1}{2} & \frac{1}{2} & -1 \\ \hline 1 & -\frac{1}{2} & -\frac{1}{2} & -\frac{\sqrt{3}}{2} & \frac{\sqrt{3}}{2} & 0 \\ 0 & -\frac{\sqrt{3}}{2} & \frac{\sqrt{3}}{2} & \frac{1}{2} & \frac{1}{2} & -1 \\ \hline 1 & 1 & 1 & 0 & 0 & 0 \\ 0 & 0 & 0 & 1 & 1 & 1 \end{bmatrix}. \quad (1)$$

92 By applying (1) to the voltage vector equations, the overall
93 machine model is transformed into three decoupled submodels,
94 written in three independent space coordinates, identified as
95 $(\alpha-\beta)$, $(x-y)$, and (o_1-o_2) , respectively.

96 The machine voltage submodel in $(\alpha-\beta)$ coordinates can be
97 written as:

$$\begin{bmatrix} v_{s\alpha} \\ v_{s\beta} \\ v_{r\alpha} \\ v_{r\beta} \end{bmatrix} = \begin{bmatrix} R_s & 0 & 0 & 0 \\ 0 & R_s & 0 & 0 \\ 0 & M\dot{\theta} & R_r & L_r\dot{\theta} \\ -M\dot{\theta} & 0 & -L_r\dot{\theta} & R_r \end{bmatrix} \begin{bmatrix} i_{s\alpha} \\ i_{s\beta} \\ i_{r\alpha} \\ i_{r\beta} \end{bmatrix} + \begin{bmatrix} L_s & 0 & M & 0 \\ 0 & L_s & 0 & M \\ M & 0 & L_r & 0 \\ 0 & M & 0 & L_r \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{s\alpha} \\ i_{s\beta} \\ i_{r\alpha} \\ i_{r\beta} \end{bmatrix} \quad (2)$$

98 where $\dot{\theta} = \Omega_m$ is the rotor mechanical speed, and $L_s = L_{ls} +$
99 $3L_{ms}$, $L_r = L_{lr} + (3/2)L_{mr}$, $M = (3/\sqrt{2})M_{sr}$. L_{ls} and L_{lr}
100 are the stator and rotor leakage inductances in $(\alpha-\beta)$ coordi-
101 nates, respectively.

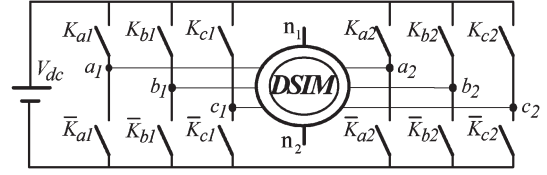


Fig. 2. Six-phase VSI fed DSIM.

The DSIM $(\alpha-\beta)$ submodel expressed in the stationary
reference frame is similar to the three-phase induction machine
model [11].

The machine voltage submodel in $(x-y)$ coordinates is
given by:

$$\begin{bmatrix} v_{sx} \\ v_{sy} \end{bmatrix} = \begin{bmatrix} R_s & 0 \\ 0 & R_s \end{bmatrix} \begin{bmatrix} i_{sx} \\ i_{sy} \end{bmatrix} + \begin{bmatrix} L_{lsxy} & 0 \\ 0 & L_{lsxy} \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{sx} \\ i_{sy} \end{bmatrix} \quad (3)$$

where L_{lsxy} is the transformed stator leakage inductance in
 $(x-y)$ coordinates.

The machine voltage submodel in (o_1-o_2) coordinates is
expressed as follows:

$$\begin{bmatrix} v_{so1} \\ v_{so2} \\ v_{ro} \end{bmatrix} = \begin{bmatrix} R_s & 0 & 0 \\ 0 & R_s & 0 \\ 0 & 0 & R_r \end{bmatrix} \begin{bmatrix} i_{so1} \\ i_{so2} \\ i_{ro} \end{bmatrix} + \begin{bmatrix} L_{lso} & 0 & 0 \\ 0 & L_{lso} & 0 \\ 0 & 0 & L_{lr} \end{bmatrix} \frac{d}{dt} \begin{bmatrix} i_{so1} \\ i_{so2} \\ i_{ro} \end{bmatrix} \quad (4)$$

where L_{lso} is the transformed stator leakage inductance in
 (o_1-o_2) coordinates.

The electromagnetic torque of the DSIM is expressed only
in terms of stator and rotor $(\alpha-\beta)$ current components, since
the $(x-y)$ and (o_1-o_2) counterparts do not contribute to the
electromechanical energy conversion, as shown by (3) and (4).
The expression of the electromagnetic torque is then as follows:

$$T_e = pM(i_{s\beta}i_{r\alpha} - i_{s\alpha}i_{r\beta}) \quad (5)$$

where p is the number of pole pairs.

General Remarks

The $(x-y)$ and (o_1-o_2) current components do not con-
tribute to the air-gap flux linkages. Hence, they are limited only
by the stator resistance and leakage inductance [12], [13]. They
produce only losses and therefore must be kept equal to zero or
as small as possible.

The transformed voltage equations in the three subframes
are well decoupled and, as a result, both machine analysis and
control are greatly simplified.

III. SVPWM CONTROL OF A DOUBLE-STAR INDUCTION MOTOR

The drive system is a six-phase VSI fed DSIM, as shown
in Fig. 2. A combinatorial analysis of the inverter switch
states shows 64 switching modes. Thus, 64 different voltage
vectors can be applied to the machine. Each voltage vector is
represented by a decimal number corresponding to the binary
number $(K_{c2}K_{b2}K_{a2}K_{c1}K_{b1}K_{a1})$, which gives the state of

136 the upper switches. By using the (6×6) transformation matrix
 137 $[Ts]^{-1}$, each voltage vector can be decomposed into $(\alpha-\beta)$,
 138 $(x-y)$, and (o_1-o_2) voltages. The (o_1-o_2) ones are all equal
 139 to zero because the neutrals (n_1, n_2) of the two winding sets
 140 are isolated. So the SVPWM strategy operates in two complex
 141 planes $(\alpha-\beta)$ and $(x-y)$. Four variables need to be controlled
 142 simultaneously during each sampling period, by generating
 143 maximum $(\alpha-\beta)$ and minimum $(x-y)$ voltage amplitudes.
 144 Therefore, during each sampling period, a set of four active
 145 voltage vectors must be chosen to fulfil these two conditions,
 146 according to the reference voltage vector location. There are
 147 numerous ways for choosing such a set.

148 A. Six-Phase SVPWM Techniques

149 The principle of the PWM control techniques proposed in
 150 [5] and [10] is to choose switching sequences in such a way
 151 that two consecutive nonzero voltage vectors are practically
 152 opposite in phase in the $(x-y)$ plane. In this way, each change
 153 in the applied vectors leads to a sequence of increases and
 154 decreases in $(x-y)$ currents around zero. Moreover, in order
 155 to minimize $(x-y)$ harmonic currents and maintain the lowest
 156 switching frequency, there are different choices to allocate
 157 zero voltage vectors (0, 7, 56 or 63) within the switching
 158 sequences. Thus, the switching sequences presented in [10] lead
 159 to continuous and discontinuous modulation techniques and,
 160 consequently, to different harmonic distortion characteristics.
 161 A modulation technique is continuous when on/off switching
 162 occurs within every sampling period, for all inverter legs and
 163 all sectors. A modulation technique is discontinuous when one
 164 (or more) inverter leg stops switching, i.e., the corresponding
 165 phase voltage is clamped to the positive or negative dc bus for
 166 at least one sector [14].

167 B. 12-Sector SVPWM Technique

168 In the SVPWM technique addressed in [5], only the $(\alpha-\beta)$
 169 voltage vectors having maximum magnitude (45, 41, 9, 11,
 170 27, 26, 18, 22, 54, 52, 36, 37) are employed to synthesize
 171 the reference voltage vector $v_{s\alpha\beta}^*$. These voltage vectors divide
 172 the $(\alpha-\beta)$ plane into 12 sectors and each sector is $\pi/6$ rad,
 173 as shown in Fig. 3. For example, voltage vectors 45, 41, 9,
 174 and 11 are selected when the reference voltage vector is located
 175 in sector 1. As shown in Fig. 4, continuous and discontinuous
 176 modulation techniques can be obtained according to the switch-
 177 ing sequences given below.

178 1) Continuous Modulation C6 ϕ SVPWM12: For example,
 179 when the reference voltage vector is located in sector 1, a con-
 180 tinuous modulation technique (C6 ϕ SVPWM12) is obtained
 181 with the following sequence:

$$|7-45-41-56-9-11-7|7-11-9-56-41-45-7|.$$

182 2) Discontinuous Modulation D6 ϕ SVPWM12-A: For the
 183 same sector 1, a discontinuous modulation technique (D6 ϕ
 184 SVPWM12-A) can be obtained with the following sequence:

$$|7-45-41-9-11-7|7-11-9-41-45-7|.$$

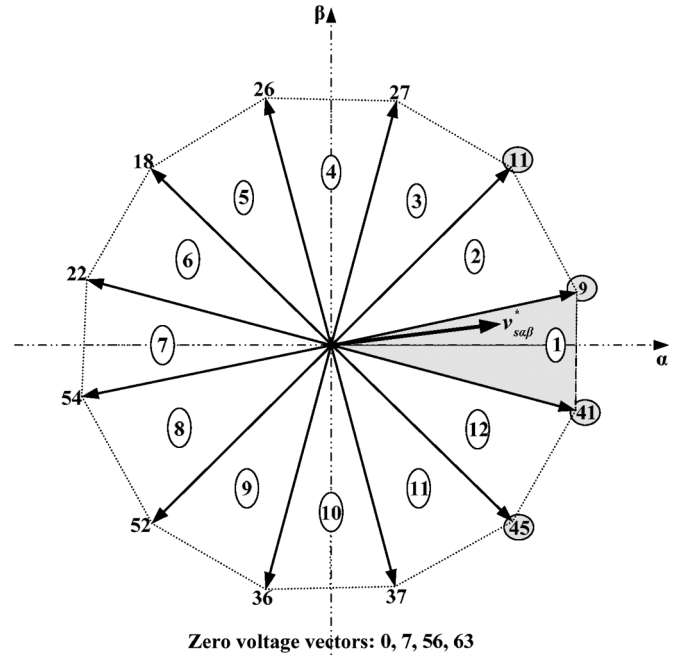


Fig. 3. Presentation of the inverter voltage vectors having maximum magnitude in $(\alpha-\beta)$ plane.

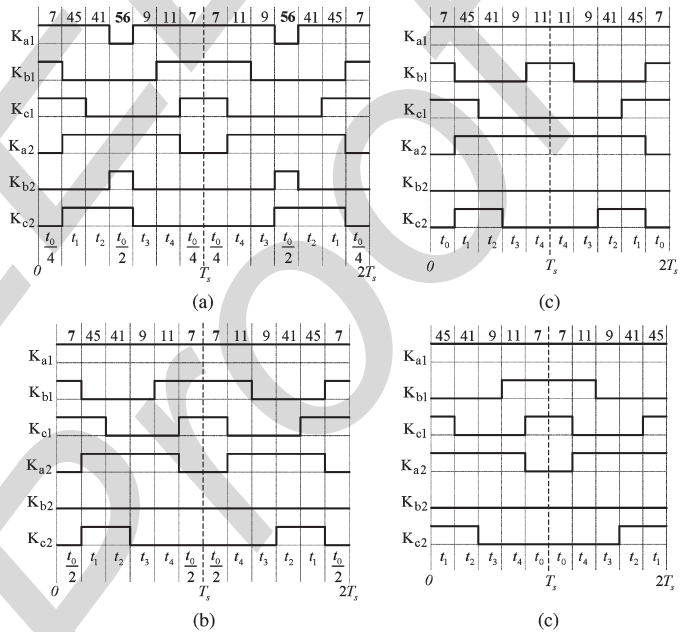


Fig. 4. Twelve-sector SVPWM switching sequences when the reference voltage vector is located in sector 1. (a) C6 ϕ SVPWM12. (b) D6 ϕ SVPWM12-A. (c) D6 ϕ SVPWM12-B1. (d) D6 ϕ SVPWM12-B2.

3) Discontinuous Modulation D6 ϕ SVPWM12-B1: In the 185
 D6 ϕ SVPWM12-B1, the zero-voltage vectors are applied at the 186
 beginning and at the end of the switching sequence as follows: 187

$$|7-45-41-9-11|11-9-41-45-7|.$$

4) Discontinuous Modulation D6 ϕ SVPWM12-B2: In the 188
 D6 ϕ SVPWM12-B2, the zero-voltage vectors are applied in 189
 the middle of the switching sequence as follows: 190

$$|45-41-9-11-7|7-11-9-41-45|.$$

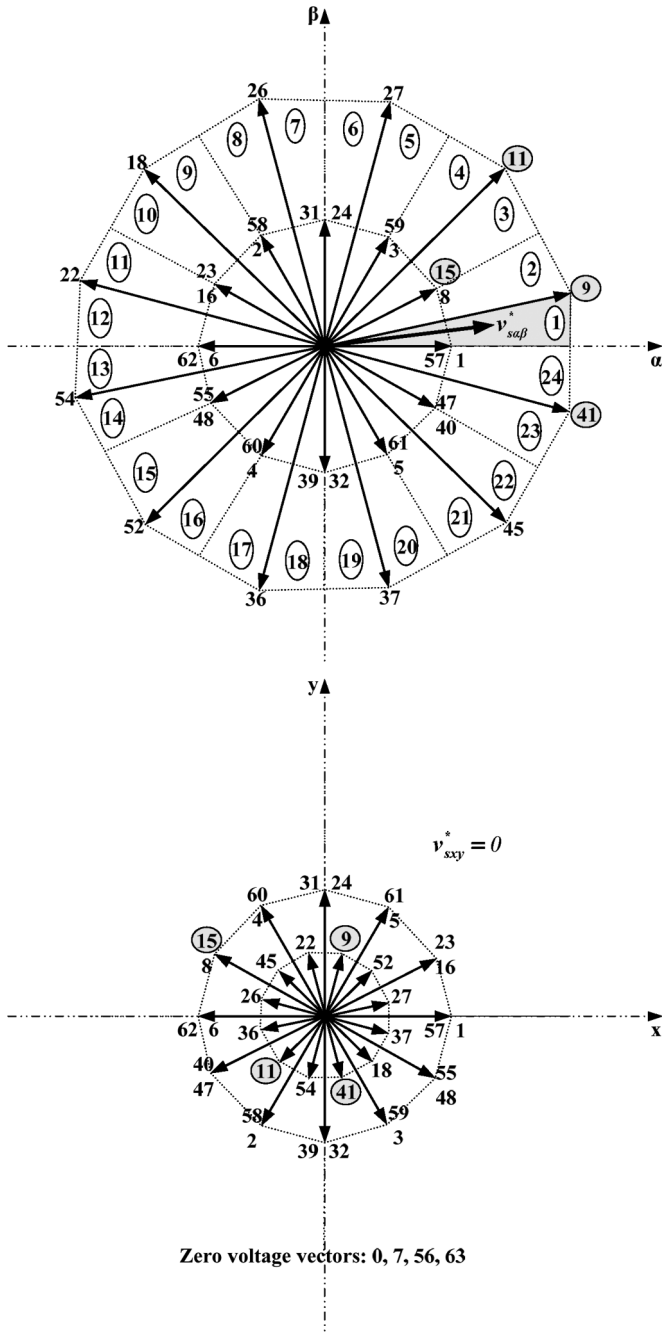


Fig. 5. Presentation of the inverter voltage vectors having maximum and half magnitude in $(\alpha-\beta)$, and $(x-y)$ planes.

191 C. Proposed 24-Sector SVPWM Technique

192 As shown in Fig. 4, the switching pattern corresponding
193 to the switching sequences of the 12-sector PWM techniques
194 presents asymmetrical waveforms and usually more than two
195 transitions (from low to high or from high to low) occur on
196 the corresponding PWM outputs within a sampling period,
197 which increases the switching frequency of the inverter legs
198 and causes difficulties for digital signal processor (DSP) im-
199 plementation of these strategies. Accordingly, some additional
200 adaptations in DSP programs are necessary to ensure successful
201 experiments [15], [16].

202 To overcome these drawbacks, the new 24-sector SVPWM
203 technique proposed in this paper combines the maximum

magnitude $(\alpha-\beta)$ voltage vectors and the ones with half
204 magnitude (1, 57, 8, 15, 3, 59, 24, 31, 2, 58, 16, 23, 6, 205
62, 48, 55, 4, 60, 32, 39, 5, 61, 40, 47) generated by one 206
inverter. These voltage vectors divide the $(\alpha-\beta)$ plane into 207
twenty four $\pi/12$ -rad sectors, as shown in Fig. 5. In each 208
sampling period, the reference voltage vector is achieved by 209
selecting a set of three voltage vectors among those having 210
maximum magnitude and a fourth vector among the ones with 211
half magnitude. For example, voltage vectors 41, 9, 11 and 15 212
are selected when the reference voltage vector is located in 213
sector 1. Then, the voltage vectors applying times: t_1, t_2, t_3 and 214
 t_4 are obtained as explained in Appendix A. For the remaining 215
time $t_0 = T_s - (t_1 + t_2 + t_3 + t_4)$, zero state vectors (0, 7, 56 216
or 63) are applied. Consequently, simple PWM outputs with 217
symmetrical waveforms are obtained. As shown in Fig. 6, 218
only two transitions or less (from low to high or from high 219
to low) occur on the corresponding PWM outputs within a 220
sampling period. This fact decreases the switching frequency 221
of the inverter legs and allows easy DSP implementation. The 222
switching sequences and the corresponding applying times for 223
all sectors are presented in Table I. It should also be noticed 224
that both continuous and discontinuous modulation techniques 225
can be obtained with this new 24-sector SVPWM scheme. This 226
is achieved by selecting the appropriate zero voltage vector 227
locations within the switching sequence. 228

1) *Continuous Modulation C6 ϕ SVPWM24*: Continuous 229
PWM technique (C6 ϕ SVPWM24) can be obtained for all sec- 230
tors by selecting the switching sequences presented in Table I. 231
As an example, when the reference voltage vector is located 232
in sector 1, a C6 ϕ SVPWM24 is obtained by selecting the 233
following sequence: 234

$$|56-41-9-11-15-7|7-15-11-9-41-56|.$$

2) *Discontinuous Modulation D6 ϕ SVPWM24-B1 and B2*: 235
Two discontinuous PWM schemes can be obtained through the 236
appropriate positioning of the zero voltage vectors. In sector 1, 237
the examples are as follows. 238

1) D6 ϕ SVPWM24-B1: The first discontinuous modulation 239
technique can be obtained by placing the zero voltage 240
vector both at the beginning and at the end of the switch- 241
ing sequence, as follows: 242

$$|56-41-9-11-15|15-11-9-41-56|.$$

2) D6 ϕ SVPWM24-B2: The second discontinuous mod- 243
ulation technique is obtained by placing the zero volt- 244
age vector in the middle of the switching sequence, as 245
follows: 246

$$|41-9-11-15-7|7-15-11-9-41|.$$

3) *Switching Sequences and Applying Time Selection*: For 247
optimal DSP implementation and low algorithm execution time, 248
the applying times (t_1, t_2, t_3 , and t_4) computation can be 249
simplified by an offline calculation for all sectors in the same 250
manner as for sector 1. As a result, within each sampling period, 251
there is a total of only 12 coefficients T_i to be calculated in (6). 252

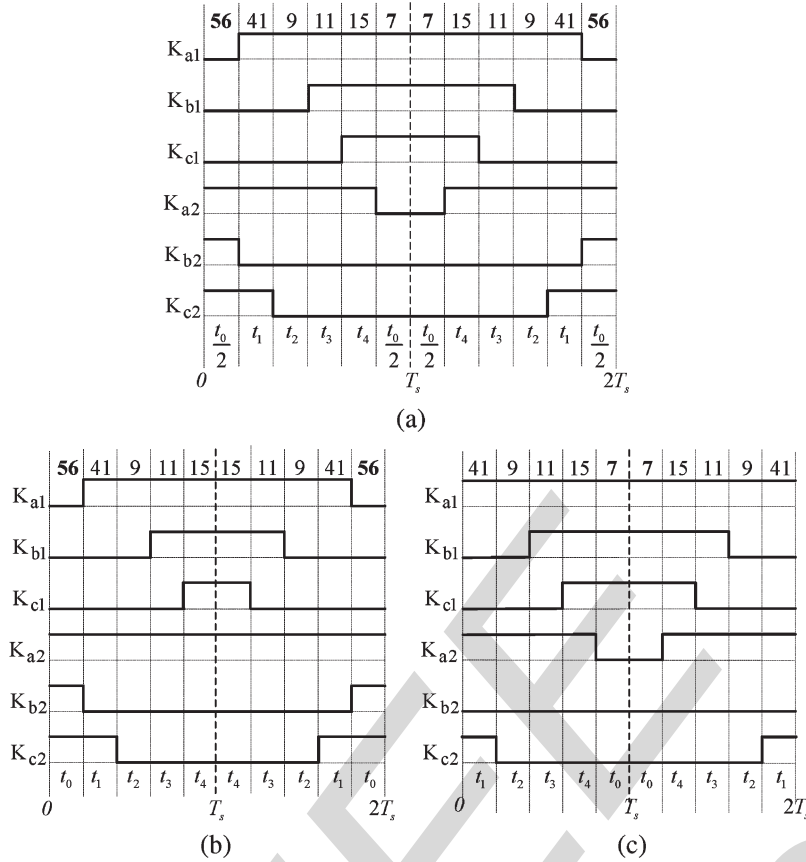


Fig. 6. Twenty-four-sector SVPWM switching sequences when the reference voltage vector is located in sector 1. (a) C6 ϕ SVPWM24. (b) D6 ϕ SVPWM24-B1. (c) D6 ϕ SVPWM24-B2.

253 Table I describes the C6 ϕ SVPWM24 switching sequences and
 254 applying time selection according to the sector number in the
 255 $(\alpha-\beta)$ plane.

$$\begin{bmatrix} T_1 \\ T_2 \\ T_3 \\ T_4 \\ T_5 \\ T_6 \\ T_7 \\ T_8 \\ T_9 \\ T_{10} \\ T_{11} \\ T_{12} \end{bmatrix} = \frac{T_s}{2V_{dc}} \begin{bmatrix} \sqrt{3}-2 & 1 \\ 1 & -\sqrt{3} \\ 1 & \sqrt{3}-2 \\ 0 & 2 \\ \sqrt{3}-1 & \sqrt{3}-1 \\ -(\sqrt{3}-1) & \sqrt{3}-1 \\ \sqrt{3} & -1 \\ 1 & -(\sqrt{3}-2) \\ -(\sqrt{3}-2) & 1 \\ 2 & 0 \\ \sqrt{3} & 1 \\ 1 & \sqrt{3} \end{bmatrix} \begin{bmatrix} v_{s\alpha}^* \\ v_{s\beta}^* \end{bmatrix}. \quad (6)$$

256 D. Maximum Modulation Index

257 The modulation index m can be defined as the ratio of
 258 the fundamental component magnitude of the line to neutral
 259 inverter output voltage V_{1m} to the fundamental component
 260 magnitude of the six-step mode voltage $V_{1m6step} = 2V_{dc}/\pi$
 261 [17]. When the inverter is operating in the linear modulation
 262 region, the sum of the applying times of the active voltage
 263 vectors is less than the switching period T_s [18], [19]. The
 264 largest value for linear output with the 12-sector and the
 265 proposed 24-sector SVPWM techniques, $m_{max} = \pi/(2\sqrt{3}) \approx$

0.907, coincides with the corresponding value for the three-
 266 phase SVPWM [20]. Note that m_{max} is obtained by solving
 267 $t_0 = T_s - (t_1 + t_2 + t_3 + t_4) = 0$ as shown in Appendix B. 268

IV. HARMONIC CURRENT ANALYSIS

269

The voltage and current waveform quality of the PWM-VSI
 270 drives is determined via the switching frequency harmonics,
 271 since they determine the switching frequency copper losses
 272 and the torque ripple of a motor load and the line current
 273 total harmonic distortion of a line-connected VSI. While the
 274 copper losses are measured over a fundamental cycle and
 275 therefore require a per fundamental cycle (macroscopic) rms
 276 ripple current value calculation, the peak and local stresses
 277 are properly investigated on a per-carrier cycle (microscopic)
 278 basis. Therefore, first a microscopic and then a macroscopic
 279 investigation is required [20]. Because, the machine model
 280 includes $(\alpha-\beta)$ and $(x-y)$ components, the harmonic current
 281 analysis must be made for the $(\alpha-\beta)$ and $(x-y)$ currents. 282

A. Normalized Harmonic Currents and Fluxes Calculation

283

The stator voltage equations in the stator coordinate system
 284 are expressed as follows: 285

$$\begin{aligned} v_{s\alpha\beta} &= R_s i_{s\alpha\beta} + \frac{d\lambda_{s\alpha\beta}}{dt} \\ v_{sxy} &= R_s i_{sxy} + L_{lsxy} \frac{di_{sxy}}{dt} \end{aligned} \quad (7)$$

TABLE I
PROPOSED C6 ϕ SVPWM24 SWITCHING SEQUENCES

Sector	Switching sequences	Voltage vectors applying times			
		t_1	t_2	t_3	t_4
1	56-41-9-11-15-7	T_2	T_5	T_4	$-T_1$
2	56-57-41-9-11-7	T_1	T_2	T_3	T_4
3	0-9-11-27-59-63	T_7	T_9	$-T_2$	$-T_6$
4	0-8-9-11-27-63	T_6	T_7	T_8	$-T_2$
5	7-11-27-26-24-56	T_{10}	T_1	$-T_7$	T_3
6	7-3-11-27-26-56	$-T_3$	T_{10}	T_5	$-T_7$
7	63-27-26-18-2-0	T_{11}	T_6	$-T_{10}$	T_8
8	63-31-27-26-18-0	$-T_8$	T_{11}	T_9	$-T_{10}$
9	56-26-18-22-23-7	T_{12}	$-T_3$	$-T_{11}$	T_5
10	56-58-26-18-22-7	$-T_5$	T_{12}	T_1	$-T_{11}$
11	0-18-22-54-62-63	T_4	$-T_8$	$-T_{12}$	T_9
12	0-16-18-22-54-63	$-T_9$	T_4	T_6	$-T_{12}$
13	7-22-54-52-48-56	$-T_2$	$-T_5$	$-T_4$	T_1
14	7-6-22-54-52-56	$-T_1$	$-T_2$	$-T_3$	$-T_4$
15	63-54-52-36-4-0	$-T_7$	$-T_9$	T_2	T_6
16	63-55-54-52-36-0	$-T_6$	$-T_7$	$-T_8$	T_2
17	56-52-36-37-39-7	$-T_{10}$	$-T_1$	T_7	$-T_3$
18	56-60-52-36-37-7	T_3	$-T_{10}$	$-T_5$	T_7
19	0-36-37-45-61-63	$-T_{11}$	$-T_6$	T_{10}	$-T_8$
20	0-32-36-37-45-63	T_8	$-T_{11}$	$-T_9$	T_{10}
21	7-37-45-41-40-56	$-T_{12}$	T_3	T_{11}	$-T_5$
22	7-5-37-45-41-56	T_5	$-T_{12}$	$-T_1$	T_{11}
23	63-45-41-9-1-0	$-T_4$	T_8	T_{12}	$-T_9$
24	63-47-45-41-9-0	T_9	$-T_4$	$-T_6$	T_{12}

286 where the stator and the rotor flux equations are given by:

$$\begin{aligned}\lambda_{s\alpha\beta} &= L_s i_{s\alpha\beta} + M i_{r\alpha\beta} \\ \lambda_{r\alpha\beta} &= L_r i_{r\alpha\beta} + M i_{s\alpha\beta}.\end{aligned}\quad (8)$$

287 The stator flux equation can be rewritten as:

$$\lambda_{s\alpha\beta} = \sigma L_s i_{s\alpha\beta} + \frac{M}{L_r} \lambda_{r\alpha\beta}.\quad (9)$$

288 Substituting (9) in (7), the stator voltage equation can be
289 expressed as follows:

$$v_{s\alpha\beta} = R_s i_{s\alpha\beta} + \sigma L_s \frac{di_{s\alpha\beta}}{dt} + \frac{M}{L_r} \frac{d\lambda_{r\alpha\beta}}{dt}.\quad (10)$$

290 If only the harmonic voltages and currents are considered,
291 it will be assumed that the reference voltage vector $v_{s\alpha\beta}^*$ is
292 constant over the switching period T_s , because the switching
293 frequency f_s is much higher than the fundamental frequency
294 f_e , and that the stator and the rotor time constants are much
295 larger than the switching period, with the resistance drops being
296 neglected [21]. Under these assumptions, the voltages and

currents can be separated in the harmonic components, which
change over T_s while the fundamental components remain
constant over the same period. Thus, from (7) and (10), the
harmonic voltage equations can be expressed as follows: 300

$$\begin{aligned}\tilde{v}_{s\alpha\beta} &= \sigma L_s \frac{d\tilde{i}_{s\alpha\beta}}{dt} \\ \tilde{v}_{sxy} &= L_{lsxy} \frac{d\tilde{i}_{sxy}}{dt}\end{aligned}\quad (11)$$

where $\tilde{v}_{s\alpha\beta}$ is the harmonic voltage and is equal to the
difference between the actual voltage vector and the reference
vector $v_{s\alpha\beta}^*$. 303

Assuming that the instantaneous harmonic currents are zero
at the beginning and at the end of the carrier cycle, the
($\alpha-\beta$) and ($x-y$) harmonic stator currents per-carrier cycle can be
calculated as follows [20], [22], [23]: 307

$$\begin{aligned}\tilde{i}_{s\alpha\beta} &= \frac{1}{\sigma L_s} \int_{NT_s}^{(N+1)T_s} (V_{s\alpha\beta k} - v_{s\alpha\beta}^*) dt \\ \tilde{i}_{sxy} &= \frac{1}{L_{lsxy}} \int_{NT_s}^{(N+1)T_s} (V_{sxy k}) dt.\end{aligned}\quad (12)$$

In (12), $V_{s\alpha\beta k}$ and $V_{sxy k}$ are the inverter output voltage
vectors of the k th state. They change according to the selected
switching sequence, since for high f_s/f_e values, the $v_{s\alpha\beta}^*$ term
can be assumed as constant within a carrier cycle. Thus, the
above integral can be calculated in a closed form. 312

Because the harmonic current and harmonic flux are only
different in scale, and in order to eliminate the need for
load parameters in (12), the harmonic flux trajectories can
be investigated. Nevertheless, the ($x-y$) current components
are limited by the stator leakage inductance L_{lsxy} , which de-
pends on the coil pitch of the stator windings [10]. Conse-
quently, the harmonic characteristics of the VSI feeding DSIM
should be investigated with the introduction of the coefficient
 $k_{\sigma xy} = \sigma L_s / L_{lsxy}$, which is necessary to evaluate and com-
pare the performances of the PWM techniques. So, employ-
ing (12) and normalizing with respect to λ_b , the per-carrier
cycle rms value of the normalized harmonic current can be
calculated with: 325

$$\begin{aligned}\tilde{i}_{srms}^2(m, \theta) &= \left(\frac{\lambda_b}{\sigma L_s}\right)^2 \frac{1}{T_s} \int_{NT_s}^{(N+1)T_s} \left(\tilde{\lambda}_{s\alpha\beta}^2 + k_{\sigma xy}^2 \tilde{\lambda}_{sxy}^2\right) dt \\ &= \left(\frac{\lambda_b}{\sigma L_s}\right)^2 \left(\tilde{\lambda}_{s\alpha\beta rms}^2(m, \theta) + k_{\sigma xy}^2 \tilde{\lambda}_{sxy rms}^2(m, \theta)\right) \\ &= \left(\frac{\lambda_b}{\sigma L_s}\right)^2 \left(\tilde{\lambda}_{srms}^2(m, \theta)\right)\end{aligned}\quad (13)$$

where $\lambda_b = 2\sqrt{3}V_{dc}T_s/\pi$, $\tilde{\lambda}_{s\alpha\beta} = \sigma L_s \tilde{i}_{s\alpha\beta}$, and $\tilde{\lambda}_{sxy} =$
 $L_{lsxy} \tilde{i}_{sxy}$ 327

TABLE II
SWITCHING FREQUENCY REDUCTION COEFFICIENT

N°	SVPWM Techniques	k_f	f_s
1	C6 ϕ SVPWM12	1	
2	D6 ϕ SVPWM12-A	2/3	
3	D6 ϕ SVPWM12-B1	1/2	
4	D6 ϕ SVPWM12-B2	5/12	$f_s = \frac{1}{k_f} \cdot f_{sw}$
5	C6 ϕ SVPWM24	1	
6	D6 ϕ SVPWM24-B1	5/6	
7	D6 ϕ SVPWM24-B2	2/3	

The per-fundamental cycle rms value of the harmonic current determines the waveform quality and harmonic losses. Averaging (13) over a fundamental period results in the global harmonic current calculation as follows:

$$\begin{aligned}
 \tilde{I}_{sfrms}^2(m) &= \left(\frac{\lambda_b}{\sigma L_s} \right)^2 \frac{1}{2\pi} \int_{2\pi} \tilde{\lambda}_{sfrms}^2(m, \theta) d\theta \\
 &= \left(\frac{\lambda_b}{\sigma L_s} \right)^2 \left(\tilde{\lambda}_{s\alpha\beta frms}^2(m) + k_{\sigma xy}^2 \tilde{\lambda}_{sxy frms}^2(m) \right) \\
 &= \left(\frac{\lambda_b}{\sigma L_s} \right)^2 \tilde{\lambda}_{sfrms}^2(m). \quad (14)
 \end{aligned}$$

The above integral yields a polynomial function of the modulation index m . As an example, the per-fundamental cycle rms normalized harmonic flux $\tilde{\lambda}_{sfrms}$ was calculated for all the discussed PWM techniques. This results in m dependent analytical formulas summarized in Appendix C.

B. Performance Comparison

For comparison purposes, the harmonic current analysis is performed at the same average switching frequency f_{sw} for all the PWM techniques. Therefore, a switching frequency reduction coefficient k_f is introduced for each PWM technique according to Table II. This coefficient can be determined from the ratio of the discontinuous to the continuous PWM techniques regarding the number of commutations of all legs during one sampling period. The curves of the per-fundamental cycle normalized rms harmonic flux for all the discussed PWM techniques have been plotted as a function of modulation index m , as shown in Fig. 7. It is clear that the rms value of the harmonic flux varies with the PWM technique used and according to the selected switching sequences. These curves show that the D6 ϕ SVPWM12-A has practically the best performance at a low modulation index range, while the D6 ϕ SVPWM12-B1 (B2) exhibits the best performance in the high modulation index range. However, as the modulation index increases, the C6 ϕ SVPWM12 performance rapidly degrades compared to the C6 ϕ SVPWM24 of the proposed 24-sector PWM scheme, which reveals excellent performance over the whole voltage range. While the D6 ϕ SVPWM24-B1-(B2) PWM strategies present harmonic characteristics similar to the ones obtained with the D6 ϕ SVPWM12-A-(B1) and (B2) strategies, the

proposed 24-sector PWM techniques allow a sampling frequency increase and as a result, the switching frequency can be increased by a factor of two as compared to the 12-sector PWMs. Therefore, significant harmonic current reductions can be achieved as shown in Fig. 7(h) for the worst case, when $k_{\sigma xy} = 10$.

It should be noted that discontinuous PWM techniques allow a higher sampling rate selection and can be applied in a high voltage range, while continuous ones are advantageous in the low voltage range. In addition, an optimal PWM scheme can be obtained with a transition between these SVPWM strategies to allow rms harmonic current minimization over the whole voltage range. The intersection points define the optimal transition between different PWM techniques.

Finally, a careful look at D6 ϕ SVPWM12-B1 in Fig. 4(c) shows that during one switching period T_s , two inverter legs have commutations twice. D6 ϕ SVPWM12-B2 in Fig. 4(d) shows that one inverter leg has commutations twice. This is never the case for D6 ϕ SVPWM24-B1 and B2. Therefore, 12-sector discontinuous PWMs have a maximum instantaneous switching frequency twice as big as in the case of 24-sector discontinuous PWMs. This fact has an impact on inverter switching losses [14], [20]. While it is not the aim of this paper to evaluate the performance of PWM techniques in terms of inverter switching and conduction losses, it can be expected that the 24-sector discontinuous PWMs will have a better performance than the 12-sector ones. Further works on switching and conduction losses are in progress and will be reported in a subsequent paper.

V. EXPERIMENTAL RESULTS

To confirm the feasibility of the proposed SVPWM techniques over the entire voltage range under V/f control, a set of experiments are carried out. The experimental test bench (Fig. 8) is composed of a six-phase VSI feeding a 15 kW DSIM prototype, and the whole control algorithm is tested on a dSPACE DS1104 controller board. The original 320F240 firmware does not allow the change in PWM compare registers and action registers many times during a period. So, to allow four changes within PWM period T_s , the flashed firmware is reprogrammed [15], [24]. Hence, it is possible to implement the 12-sector PWM techniques.

On the contrary, for the proposed 24-sector PWM strategies, usually at most two transitions (from low to high or from high to low) occur symmetrically on the corresponding PWM outputs within a sampling period. This allows easier DSP implementation. Then the original 320F240 firmware is used with the help of specially developed user functions that allow the synchronization of six full PWM and three simple PWM simultaneously on the DS1104 DSP board.

These PWM techniques are successfully tested and the following conclusions can be drawn from these experimental results: In the case of the 12-sector PWM strategies, usually more than two transitions (from low to high or from high to low) occur on the corresponding PWM outputs. This increases the switching frequency of the inverter legs and complicates the experimental test. However, the number of transitions

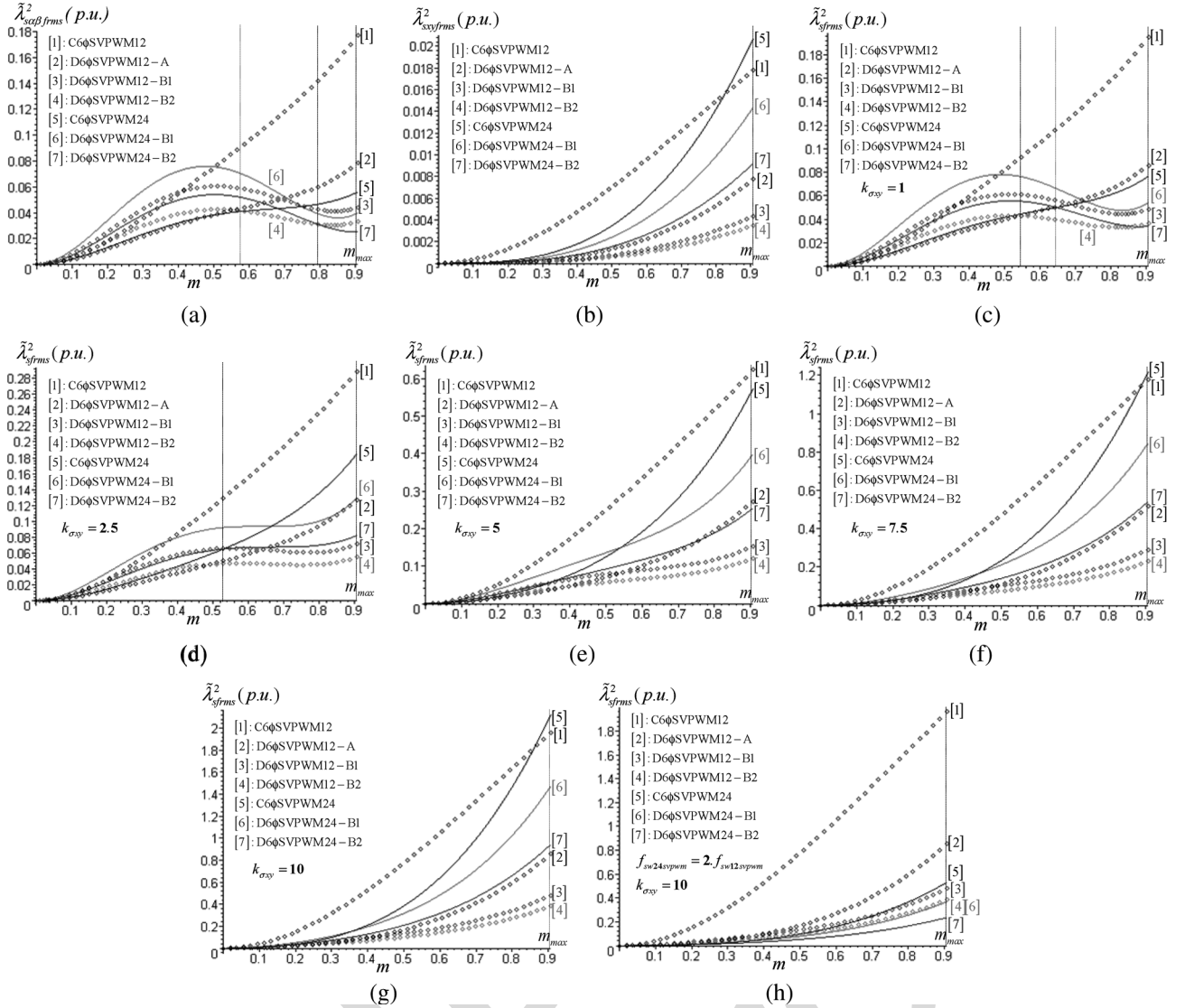


Fig. 7. Per fundamental cycle normalized rms harmonic flux as a function of modulation index m , for all the discussed PWM techniques. (a) $(\alpha-\beta)$ rms harmonic flux. (b) $(x-y)$ rms harmonic flux. (c), (d), (e), (f), (g) RMS harmonic flux at different leakage coupling ($k_{\sigma xy} = 1; 2.5; 5; 7.5; 10$), at the same average switching frequency f_{sw} . (h) RMS harmonic flux with $k_{\sigma xy} = 10$, at $f_s = 2 \cdot f_{sw}$ for the proposed 24-sector PWM techniques.

417 on the corresponding PWM outputs is reduced in the case
418 of the proposed 24-sector PWM strategies, which decrease
419 the switching frequency of the inverter legs and allow easy
420 DSP implementation. In addition, for discontinuous PWM tech-
421 niques, at least two PWM outputs remain unchanged during the
422 entire sampling period. Therefore, the carrier frequency can be
423 increased with harmonic losses reduction.

424 Experimental results under constant V/f control, for the
425 cases of C6 ϕ SVPWM12, D6 ϕ SVPWM12-B2, C6 ϕ
426 SVPWM24, and D6 ϕ SVPWM24-B2 techniques are pre-
427 sented in Fig. 9. The average switching frequency is set to
428 $f_{sw} = 5$ kHz and the motor is running at 735 r/min with a
429 connected load. As expected, these SVPWM techniques al-
430 low the control of the $(\alpha-\beta)$ and $(x-y)$ current components
431 simultaneously.

432 However, these experimental results demonstrate that the
433 continuous PWM techniques produce a larger amplitude of
434 the harmonic currents in the $(x-y)$ plane as compared to
435 the discontinuous ones where the amplitude of these currents

is minimized. Moreover, the phase current presents a pure
436 sinusoidal shape and the trajectory of the $(\alpha-\beta)$ stator cur-
437 rent components is a circle for all these PWM techniques,
438 confirming that these currents are controlled as well as $(x-y)$
439 harmonic currents. Indeed, it should be remembered that the
440 carrier frequency can be increased by a factor of two for a 50%
441 reduction of harmonic losses in the case of C6 ϕ SVPWM24,
442 or by a factor three for a 66% reduction in harmonic losses
443 with the D6 ϕ SVPWM24-B2 technique. In addition, due
444 to the simplicity and regularity of the proposed 24-sector
445 PWM techniques, low-cost DSPs for motor control may easily
446 be used. 447

VI. CONCLUSION

448
449 In this paper, a new SVPWM technique based on the vector
450 space decomposition suitable for six-phase VSI-fed DSIM has
451 been presented. The switching sequences presented lead to
452 continuous and discontinuous modulation strategies, according

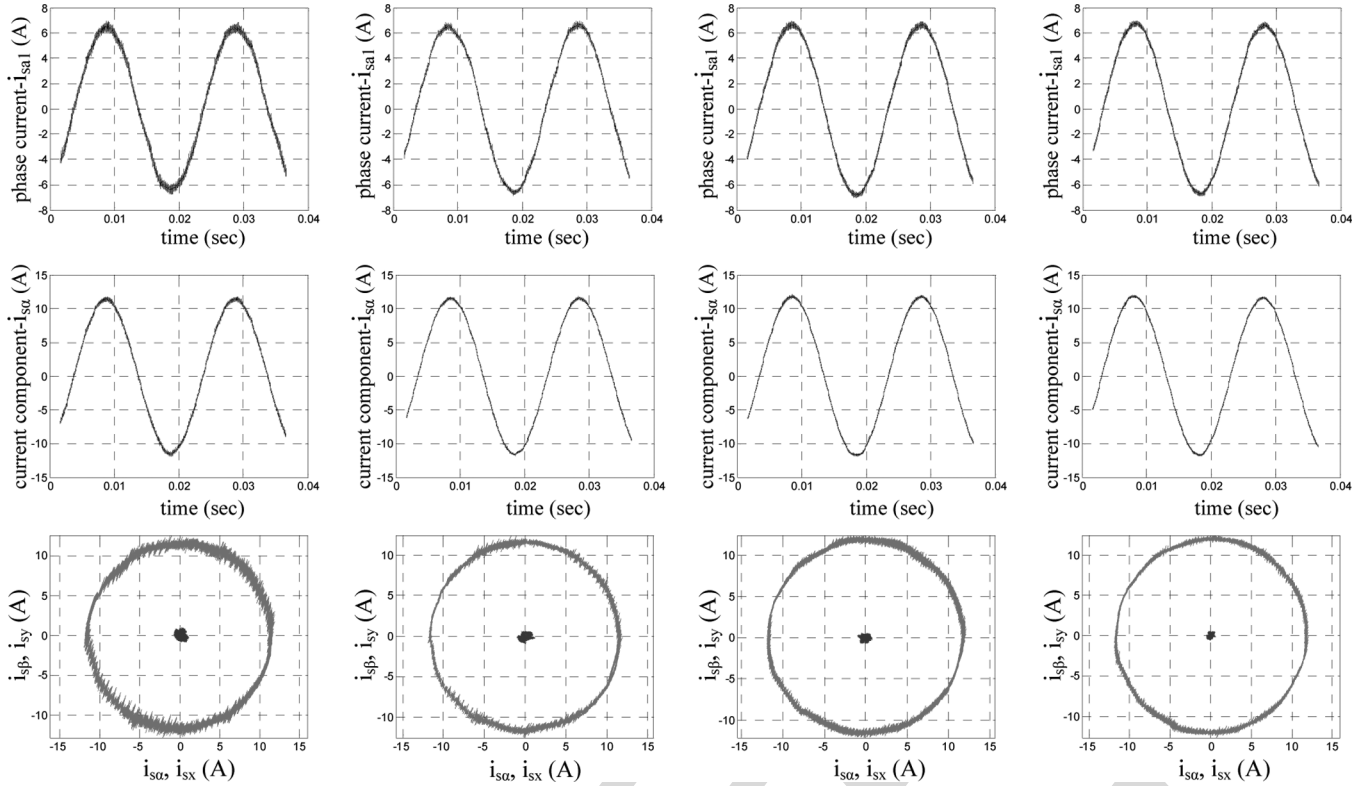


Fig. 8. Experimental test bench.

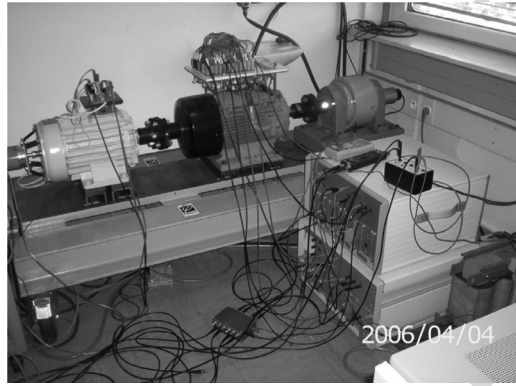


Fig. 9. Experimental results of the SVPWM techniques with the motor operating under a constant V/f control with connected load for $f_e = 50$ Hz, at 735 r/min and for the same average switching frequency f_{sw} . From top to bottom: i_{sa1} phase current, $i_{s\alpha}$ current component, $(\alpha-\beta)$ and $(x-y)$ plane current trajectories. From left to right: C6 ϕ SVPWM12, D6 ϕ SVPWM12-B2, C6 ϕ SVPWM24, and D6 ϕ SVPWM24-B2.

453 to the position of zero voltage vectors during each sampling
454 period. It is shown that the harmonic current rms values vary
455 according to the selected switching sequence and the voltage
456 range. Likewise, from this point of view, the continuous PWM
457 technique has an advantage in the low and medium voltage
458 range, while the discontinuous PWM strategy is advantageous
459 in the high voltage range. Thus, the combination of these
460 strategies provides the best harmonic current performance over
461 the whole voltage range. It has been demonstrated that the pro-
462 posed 24-sector SVPWM techniques, while easy to implement
463 digitally, allow a switching frequency increase with significant
464 extra stator harmonic currents reduction.

APPENDIX A

465

VOLTAGE VECTORS APPLYING TIMES CALCULATION

466

The inverter output voltage vectors are represented in Figs. 3, 467
and Fig. 5 by decimal number k equivalent to the binary number 468
formed by the instantaneous values of the switching functions 469
defined as: 470

$$k = K_{a1} \times 2^0 + K_{b1} \times 2^1 + K_{c1} \times 2^2 + K_{a2} \times 2^3 + K_{b2} \times 2^4 + K_{c2} \times 2^5. \quad (A1)$$

The instantaneous values of the six-phase VSI output voltage 471
vectors $(v_{a1}, v_{b1}, v_{c1}, v_{a2}, v_{b2}, v_{c2})$ can be determined by using 472
the inverter connection matrix $[Mc]$ as follows: 473

$$[v_{a1} \ v_{b1} \ v_{c1} \ v_{a2} \ v_{b2} \ v_{c2}]^T = [Mc][K_{a1} \ K_{b1} \ K_{c1} \ K_{a2} \ K_{b2} \ K_{c2}]^T \quad (A2)$$

where

474

$$[Mc] = \frac{V_{dc}}{3} \begin{bmatrix} 2 & -1 & -1 & 0 & 0 & 0 \\ -1 & 2 & -1 & 0 & 0 & 0 \\ -1 & -1 & 2 & 0 & 0 & 0 \\ \hline 0 & 0 & 0 & 2 & -1 & -1 \\ 0 & 0 & 0 & -1 & 2 & -1 \\ 0 & 0 & 0 & -1 & -1 & 2 \end{bmatrix}. \quad (A3)$$

The inverter output voltage vectors are transformed into 475
 $(\alpha-\beta)$, $(x-y)$, and (o_1-o_2) planes by means of the transfor- 476
mation matrix $[Ts]^{-1}$ given in (1), as: 477

$$[v_{s\alpha k} \ v_{s\beta k} \ v_{sxk} \ v_{syk} \ v_{o1k} \ v_{o2k}]^T = [Ts]^{-1}[v_{a1} \ v_{b1} \ v_{c1} \ v_{a2} \ v_{b2} \ v_{c2}]^T \quad (A4)$$

478 For example, when the reference voltage vector is located in
479 sector 1, voltage vectors 41, 9, 11, and 15 are selected and their
480 equivalent binary numbers are defined as:

$$\begin{aligned} 41 &= [1 \ 0 \ 1|0 \ 0 \ 1] \\ 9 &= [0 \ 0 \ 1|0 \ 0 \ 1] \\ 11 &= [0 \ 0 \ 1|0 \ 1 \ 1] \\ 15 &= [0 \ 0 \ 1|1 \ 1 \ 1]. \end{aligned} \quad (A5)$$

481 Then the $(\alpha-\beta)$, and $(x-y)$ voltages can be calculated by
482 using (A2), and (A4), as follows:

$$\begin{aligned} &\begin{bmatrix} V_{s\alpha 41} & V_{s\alpha 9} & V_{s\alpha 11} & V_{s\alpha 15} \\ V_{s\beta 41} & V_{s\beta 9} & V_{s\beta 11} & V_{s\beta 15} \\ V_{sx 41} & V_{sx 9} & V_{sx 11} & V_{sx 15} \\ V_{sy 41} & V_{sy 9} & V_{sy 11} & V_{sy 15} \end{bmatrix} \\ &= \frac{V_{dc}}{2\sqrt{3}} \begin{bmatrix} 2 + \sqrt{3} & 2 + \sqrt{3} & 1 + \sqrt{3} & \sqrt{3} \\ -1 & 1 & 1 + \sqrt{3} & 1 \\ 2 - \sqrt{3} & 2 - \sqrt{3} & 1 - \sqrt{3} & -\sqrt{3} \\ -1 & 1 & 1 - \sqrt{3} & 1 \end{bmatrix}. \end{aligned} \quad (A6)$$

483 The voltage vectors applying times: t_1 , t_2 , t_3 and t_4 are
484 obtained as:

$$\begin{aligned} \begin{bmatrix} t_1 \\ t_2 \\ t_3 \\ t_4 \end{bmatrix} &= \begin{bmatrix} V_{s\alpha 41} & V_{s\alpha 9} & V_{s\alpha 11} & V_{s\alpha 15} \\ V_{s\beta 41} & V_{s\beta 9} & V_{s\beta 11} & V_{s\beta 15} \\ V_{sx 41} & V_{sx 9} & V_{sx 11} & V_{sx 15} \\ V_{sy 41} & V_{sy 9} & V_{sy 11} & V_{sy 15} \end{bmatrix}^{-1} \begin{bmatrix} v_{s\alpha}^* T_s \\ v_{s\beta}^* T_s \\ v_{sx}^* T_s \\ v_{sy}^* T_s \end{bmatrix} \\ t_0 &= T_s - (t_1 + t_2 + t_3 + t_4). \end{aligned} \quad (A7)$$

485 Substituting (A6) in (A7), the applying times can be calcu-
486 lated as follows:

$$\begin{aligned} \begin{bmatrix} t_1 \\ t_2 \\ t_3 \\ t_4 \end{bmatrix} &= \frac{T_s}{2V_{dc}} \begin{bmatrix} 1 & -\sqrt{3} & -1 & -\sqrt{3} \\ \sqrt{3}-1 & \sqrt{3}-1 & \sqrt{3}+1 & \sqrt{3}+1 \\ 0 & 2 & 0 & v-2 \\ -(\sqrt{3}-2) & -1 & -(\sqrt{3}+2) & 1 \end{bmatrix} \begin{bmatrix} v_{s\alpha}^* \\ v_{s\beta}^* \\ v_{sx}^* \\ v_{sy}^* \end{bmatrix}. \end{aligned} \quad (A8)$$

487 With $v_{sx}^* = v_{sy}^* = 0$, (A8) can be written as:

$$\begin{aligned} \begin{bmatrix} t_1 \\ t_2 \\ t_3 \\ t_4 \end{bmatrix} &= \frac{T_s}{2V_{dc}} \begin{bmatrix} 1 & -\sqrt{3} \\ \sqrt{3}-1 & \sqrt{3}-1 \\ 0 & 2 \\ -(\sqrt{3}-2) & -1 \end{bmatrix} \begin{bmatrix} v_{s\alpha}^* \\ v_{s\beta}^* \end{bmatrix} = \begin{bmatrix} T_2 \\ T_5 \\ T_4 \\ -T_1 \end{bmatrix}. \end{aligned} \quad (A9)$$

488 APPENDIX B 489 MAXIMUM MODULATION INDEX CALCULATION

490 The maximum modulation index m_{max} can be obtained
491 by solving $t_0 = T_s - (t_1 + t_2 + t_3 + t_4) = 0$. For example in
492 sector 1, the sum of the applying times of the active voltage
493 vectors is calculated from (A9) as:

$$\theta \in \left[0, \frac{\pi}{12}\right] \quad t_1 + t_2 + t_3 + t_4 = \frac{T_s}{V_{dc}} v_{s\alpha}^* \quad (B1)$$

494 where $v_{s\alpha}^* = \sqrt{3}V_{1m} \cos(\theta)$, $v_{s\beta}^* = \sqrt{3}V_{1m} \sin(\theta)$

$$V_{1m} = mV_{1m6step} = m2V_{dc}/\pi.$$

When

$$t_0 = 0 : T_s = t_1 + t_2 + t_3 + t_4 = 2\sqrt{3}\frac{T_s}{\pi}m \cos(\theta) \quad (B2)$$

From (B2), the modulation index equation can be given as: 496

$$m = \frac{\pi}{2\sqrt{3} \cos(\theta)}. \quad (B3)$$

To determine the angle θ corresponding to m_{max} , (B3) is 497
derived: 498

$$\frac{dm}{d\theta} = \frac{\pi \sin(\theta)}{2\sqrt{3} \cos^2(\theta)}. \quad (B4)$$

Equation (B4) is solved for $\theta = 0$. Thus, replacing θ in (B3): 499

$$m_{max} = \frac{\pi}{2\sqrt{3} \cos(0)} = \frac{\pi}{2\sqrt{3}} \approx 0.907. \quad (B5)$$

APPENDIX C ANALYTICAL FORMULAS OF THE RMS HARMONIC FLUX 501

The per-fundamental cycle rms normalized harmonic flux 502
 $\tilde{\lambda}_{sfrms}$ was calculated for all the discussed PWM techniques. 503
Only the analytical formulas of the proposed 24-sector PWMs 504
are presented here and the ones of the 12-sector PWMs can be 505
found in [10]. These formulas are given below. 506

A. Continuous Modulation C6 ϕ SVPWM24 507

$$\begin{aligned} \tilde{\lambda}_{s\alpha\beta frms}^2(m) &= \frac{1}{48}m^2 + \frac{1}{144\pi^2} \\ &\quad \times (56\sqrt{3} + 63\sqrt{6} - 57\sqrt{2} - 228)m^3 \\ &\quad + \frac{1}{32\pi^3}(24\pi + 27 - 21\sqrt{3} - 8\sqrt{3}\pi)m^4 \\ \tilde{\lambda}_{sxy frms}^2(m) &= \frac{1}{144\pi^2}(63\sqrt{6} + 18 - 52\sqrt{3} - 57\sqrt{2})m^3. \end{aligned} \quad (C1)$$

B. Discontinuous Modulation D6 ϕ SVPWM24-B1 508

$$\begin{aligned} \tilde{\lambda}_{s\alpha\beta frms}^2(m) &= \frac{25}{432}m^2 - \frac{25}{5184\pi^2} \\ &\quad \times (633\sqrt{2} + 408 - 56\sqrt{3} - 387\sqrt{6})m^3 \\ &\quad - \frac{25}{576\pi^3}(15\sqrt{3} + 8\sqrt{3}\pi - 24\pi - 45)m^4 \\ \tilde{\lambda}_{sxy frms}^2(m) &= \frac{25}{5184\pi^2}(63\sqrt{6} + 18 - 52\sqrt{3} - 57\sqrt{2})m^3. \end{aligned} \quad (C2)$$

C. Discontinuous Modulation D6 ϕ SVPWM24-B2 509

$$\begin{aligned} \tilde{\lambda}_{s\alpha\beta frms}^2(m) &= \frac{1}{27}m^2 - \frac{1}{324\pi^2} \\ &\quad \times (129\sqrt{2} + 45\sqrt{6} + 48 - 56\sqrt{3})m^3 \\ &\quad + \frac{1}{6\pi^3}(2\pi + 3 - \sqrt{3})m^4 \\ \tilde{\lambda}_{sxy frms}^2(m) &= \frac{1}{324\pi^2}(63\sqrt{6} + 18 - 52\sqrt{3} - 57\sqrt{2})m^3. \end{aligned} \quad (C3)$$

510

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AQ1 = “1972” was assumed as the author’s birth year. Please check if appropriate.

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