Matrix Converters : A Technology Review.

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I. INTRODUCTION

Among the most desirable features in power frequency changers are :

i. Simple and compact power circuit.
ii. Generation of load voltage with arbitrary amplitude and frequency.
iii. Sinusoidal input and output currents.
iv. Operation with unity power factor for any load.
v. Regeneration capability.

These ideal characteristics can be fulfilled by Matrix Converters and this is the reason for the tremendous interest in the topology.

The Matrix Converter is a forced commutated converter which uses an array of controlled bi-directional switches as the main power elements to create a variable output voltage system with unrestricted frequency. It does not have any dc-link circuit and does not need any large energy storage elements.

The key element in a Matrix Converter is the fully controlled four-quadrant bidirectional switch, which allows high frequency operation. The early work dedicated to unrestricted frequency changers used thyristors with external forced commutation circuits to implement the bi-directional controlled switch [1], [2], [3], [4]. With this solution the power circuit was bulky and the performance was poor.

The introduction of power transistors for implementing the bi-directional switches made the Matrix Converter topology more attractive [5], [6], [7], [8], [9]. However, the real development of Matrix Converters starts with the work of Venturini and Alesina published in 1980 [10], [11]. They presented the power circuit of the converter as a matrix of bi-directional power switches and they introduced the name “Matrix Converter.” One of their main contributions is the development of a rigorous mathematical analysis to describe the low-frequency behavior of the converter, introducing the “low frequency modulation matrix” concept. In their modulation method, also known as the direct transfer function approach, the output voltages are obtained by the multiplication of the modulation (also called transfer) matrix with the input voltages.

A conceptually different control technique based on the “fictitious dc link” idea was introduced by Rodriguez in 1983 [12]. In this method the switching is arranged so that each output line is switched between the most positive and most negative input lines using a PWM technique, as conventionally used in standard voltage source inverters. This concept is also known as the “indirect transfer function” approach [15]. In 1985/86, Ziogas et al published 2 papers [13], [40] which expanded on the fictitious dc link idea of Rodriguez and provided a rigorous mathematical explanation. In 1983 Braun [16] and in 1985 Kastner and Rodriguez [18] introduced the use of space vectors in the analysis and control of Matrix Converters. In 1989, Huber et al published the first of a series of papers [14], [41-45] in which the principles of Space Vector Modulation (SPVM) were applied to the Matrix Converter modulation problem [17].

The modulation methods based on the Venturini approach, are known as “direct methods”, while those based on the fictitious dc link are known as “indirect methods.”

It was experimentally confirmed by Kastner and Rodriguez in 1985 [18] and Neft and Schauder in 1992 [19] that a Matrix Converter with only 9 switches can be effectively used in the vector control of an induction motor with high quality input and output currents. However, the simultaneous commutation of controlled bi-directional switches used in Matrix Converters is very difficult to achieve without generating overcurrent or overvoltage spikes that can destroy the power semiconductors. This fact limited the practical implementation and negatively affected the interest in Matrix Converters. Fortunately, this major problem has been solved with the development of several multistep commutation strategies that allow safe operation of the switches. In 1989 Burany [36] introduced the later named “semi-soft current commutation” technique. Other interesting commutation strategies were introduced by Ziegler et al [22], [37] and Clare and Wheeler in 1998 [21],[38][39].

Today the research is mainly focused on operational and technological aspects: reliable implementation of commutation strategies [20]; protection issues [23], [24]; implementation of bidirectional switches and packaging [25], [26]; operation under abnormal conditions; ride-through capability [28] and input filter design [29], [30].

The purpose of this paper is to give a review of key aspects concerning Matrix Converter operation and to establish the state of the art of this technology. It begins by studying the topology of the Matrix Converter, the main control techniques, the practical implementation of bi-directional switches and commutation strategies. Finally, some practical issues and challenges for the future are discussed.

II. FUNDAMENTALS

The Matrix Converter is a single stage converter which has an array of mxn bidirectional power switches to connect, directly, an m-phase voltage source to an n-phase load. The Matrix Converter of 3x3 switches, shown in figure 1, has the highest practical interest because it connects a three-phase voltage source with a three-phase load, typically a motor.

Normally, the Matrix Converter is fed by a voltage source and for this reason, the input terminals should not be short-circuited. On the other hand, the load has typically an inductive nature and for this reason an output phase must never be opened.

Defining the switching function of a single switch as [45]

$$S_{Kj} = \begin{cases} 1, & \text{switch } \mathrm{S}_{Kj} \text{ closed} \\ 0, & \text{switch } \mathrm{S}_{Kj} \text{ open} \end{cases} \quad K = \{A, B, C\} \quad j = \{a, b, c\} \tag{1}$$

The constraints discussed above can be expressed by

$$S_{Aj} + S_{Bj} + S_{Cj} = 1 \qquad \mathrm{j} = \{a, b, c\} \tag{2}$$

With these restrictions, the 3x3 Matrix Converter has 27 possible switching states [45].

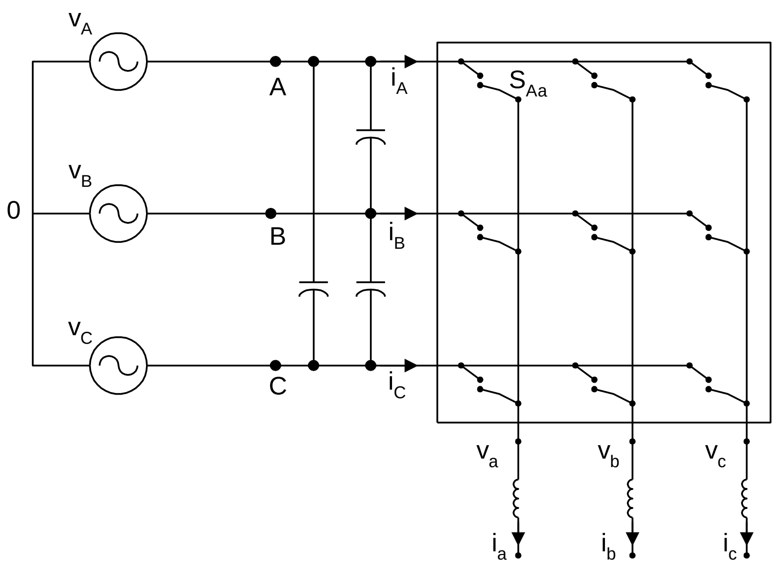


Fig. 1. Simplified circuit of a 3x3 Matrix Converter.

The load and source voltages are referenced to the supply neutral, '0' in figure 1, and can be expressed as vectors defined by:

$$\mathbf{v_o} = \begin{bmatrix} v_a(t) \\ v_b(t) \\ v_c(t) \end{bmatrix} \quad ; \quad \mathbf{v_i} = \begin{bmatrix} v_A(t) \\ v_B(t) \\ v_C(t) \end{bmatrix} \tag{3}$$

The relationship between load and input voltages can be expressed as:

$$\left.\begin{aligned} &\begin{bmatrix} v_a(t) \\ v_b(t) \\ v_c(t) \end{bmatrix} = \begin{bmatrix} S_{Aa}(t) & S_{Ba}(t) & S_{Ca}(t) \\ S_{Ab}(t) & S_{Bb}(t) & S_{Cb}(t) \\ S_{Ac}(t) & S_{Bc}(t) & S_{Cc}(t) \end{bmatrix} \begin{bmatrix} v_A(t) \\ v_B(t) \\ v_C(t) \end{bmatrix} \\ &\qquad\qquad \mathbf{v_0} = \mathbf{T} \cdot \mathbf{v_i} \end{aligned}\right\} \tag{4}$$

Where **T** is the instantaneous transfer matrix.

In the same form, the following relationships are valid for the input and output currents:

$$\mathbf{i_i} = \begin{bmatrix} i_a(t) \\ i_b(t) \\ i_c(t) \end{bmatrix} \quad ; \quad \mathbf{i_o} = \begin{bmatrix} i_A(t) \\ i_B(t) \\ i_C(t) \end{bmatrix} \tag{5}$$

$$\mathbf{i_i} = \mathbf{T}^T \cdot \mathbf{i_o} \tag{6}$$

Where $\mathbf{T}^T$ is the transpose matrix of $\mathbf{T}$.

Equations (4) and (6) give the instantaneous relationships between input and output quantities. To derive modulation rules, it is also necessary to consider the switching pattern that is employed. This typically follows a form similar to that shown in figure 2.

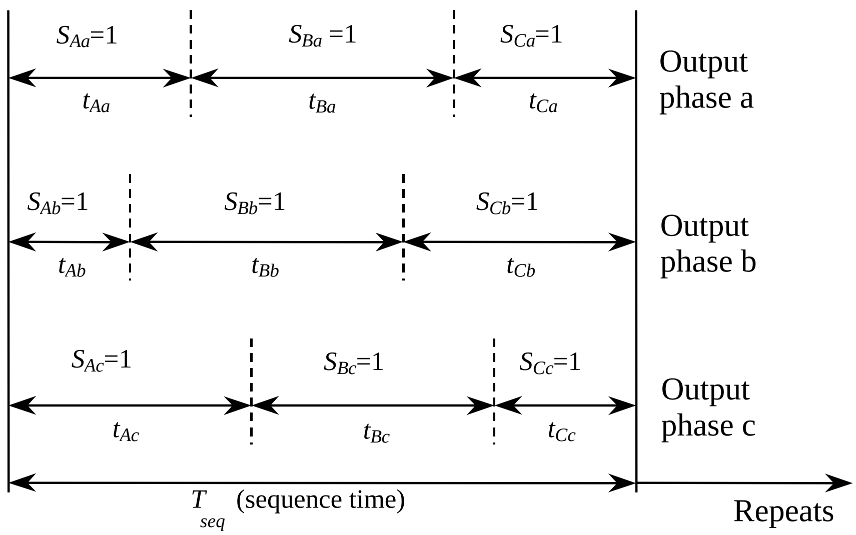


Fig. 2. General form of switching pattern.

By considering that the bidirectional power switches work with high switching frequency, a low frequency output voltage of variable amplitude and frequency can be generated by modulating the duty cycle of the switches using their respective switching functions.

Let m_{Kj} *(t)* be the duty cycle of switch S_{Kj}, defined as $m_{Kj}(t)=t_{Kj}/T_{seq}$, which can have the following values

$$0 < m_{Kj} < 1 \qquad \mathrm{K} = \{A,B,C\} \; \mathrm{j} = \{a,b,c\} \tag{7}$$

The low-frequency transfer matrix is defined by

$$\mathbf{M}(t)=\begin{bmatrix} m_{Aa}(t) & m_{Ba}(t) & m_{Ca}(t) \\ m_{Ab}(t) & m_{Bb}(t) & m_{Cb}(t) \\ m_{Ac}(t) & m_{Bc}(t) & m_{Cc}(t) \end{bmatrix} \tag{8}$$

The low-frequency component of the output phase voltage is given by

$$\bar{\mathbf{v}}_{\mathbf{o}}(t)=\mathbf{M}(t)\cdot\mathbf{v}_{\mathbf{i}}(t) \tag{9}$$

The low-frequency component of the input current is

$$\bar{\mathbf{i}}_{\mathbf{i}}=\mathbf{M}(t)^{T}\cdot\mathbf{i}_{\mathbf{o}} \tag{10}$$

Figure 3 shows simulated waveforms generated by a Matrix Converter.

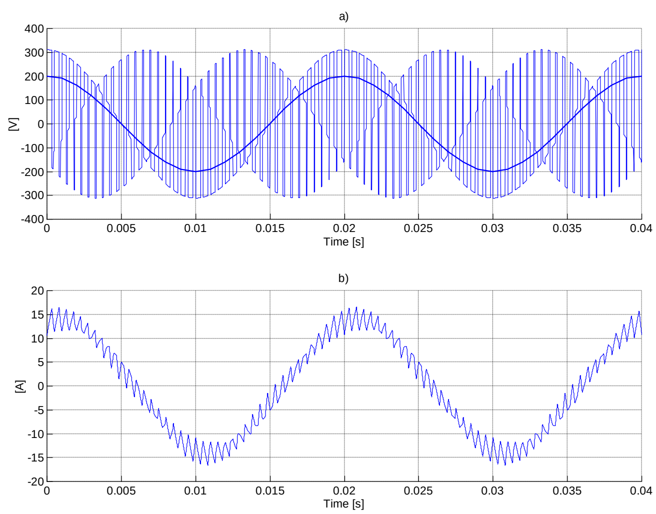


Fig. 3 : Typical waveforms : a) Phase output voltage ; b) Load current

III. THE BIDIRECTIONAL SWITCH

The Matrix Converter requires a bi-directional switch capable of blocking voltage and conducting current in both directions. Unfortunately there are no such devices currently available, so discrete devices need to be used to construct suitable switch cells.

A Realization with discrete semiconductors

The diode bridge bi-directional switch cell arrangement consists of an IGBT at the center of a single-phase diode bridge [19] arrangement as shown in figure 4. The main advantage is that both current directions are carried by the same switching device, therefore only one gate driver is required per switch cell. Device losses are relatively high since there are three devices in each conduction path. The direction of current through the switch cell cannot be controlled. This is a disadvantage, as many of the advanced commutation methods described later require this.

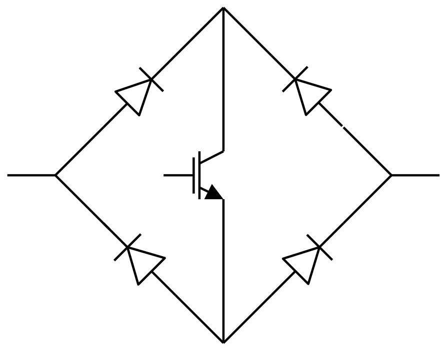


Fig. 4. Diode bridge bi-directional switch cell

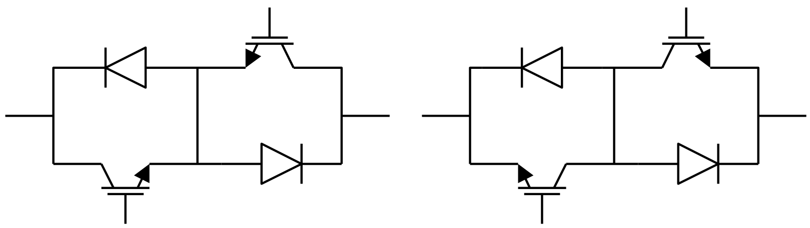


Fig. 5 Switch cell : a) Common emitter back-to-back,
b) Common collecor back-to-back.

The common emitter bi-directional switch cell arrangement consists of two diodes and two IGBTs connected in anti-parallel as shown in figure 5a. The diodes are included to provide the reverse blocking capability. There are several advantages in using this arrangement when compared to the previous example. The first is that it is possible to independently control the direction of the current. Conduction losses are also reduced since only two devices carry the current at any one time. One possible disadvantage is that each bi-directional switch cell requires an isolated power supply for the gate drives.

The common collector bi-directional switch cell arrangement is shown in figure 5b. The conduction losses are the same as for the common emitter configuration. An often quoted advantage of this method is that only 6 isolated power supplies are needed to supply the gate drive signals [46]. However, in practice, other constraints such as the need to minimise stray inductance mean that operation with only six isolated supplies is generally not viable. Therefore the common emitter configuration is generally preferred for creating the Matrix Converter bi-directional switch cells.

Both the common collector and common emitter configurations can be used without the central common connection, but this connection does provide some transient benefits during switching. In the common emitter configuration the central connection also allows both devices to be controlled from one isolated gate drive power supply.

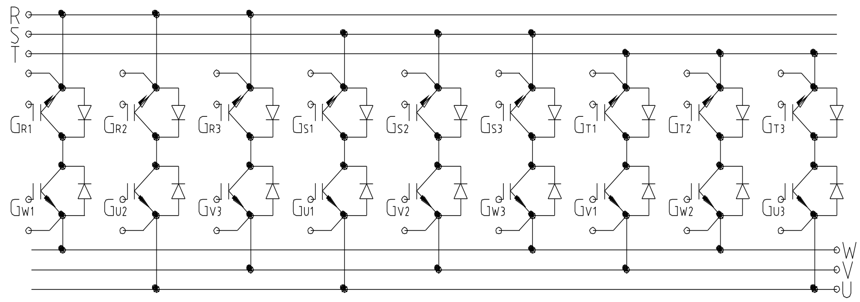


Fig. 6. Power stage of a Matrix Converter

B. Integrated power modules

It is possible to construct the common emitter bi-directional switch cell from discrete components, but it is also possible to build a complete Matrix Converter in the package style used for standard 6-pack IGBT modules. This technology can be used to develop a full Matrix Converter power circuit in a single package, as shown in figure 6. This has been done by Eupec using devices connected in the common collector configuration, see figure 7, and is now available commercially [47]. This type of packaging will have important benefits in terms of circuit layout as the stray inductance in the current commutation paths can be minimised.

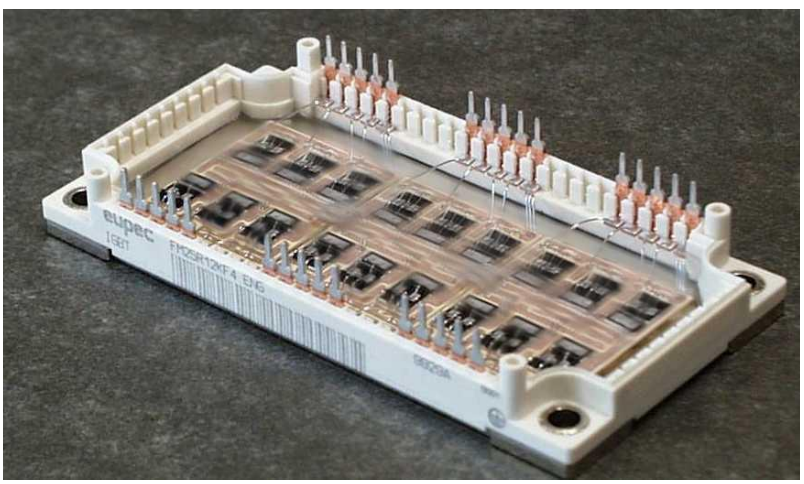
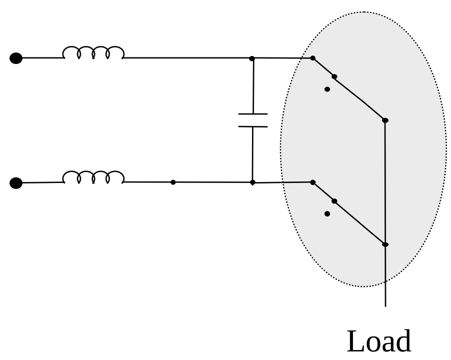


Fig. 7. The Eupec ECONOMAC Matrix Module

If the switching devices used for the bi-directional switch have a reverse voltage blocking capability, for example MTOs, then it is possible to build the bi-directional switches by simply placing two devices in anti-parallel. This arrangement leads to a very compact converter with the potential for substantial improvements in efficiency.

IV. CURRENT COMMUTATION

Reliable current commutation between switches in Matrix Converters is more difficult to achieve than in conventional voltage source inverters since there are no natural freewheeling paths. The commutation has to be actively controlled at all times with respect to two basic rules. These rules can be visualized by considering just two switch cells on one output phase of a Matrix Converter. It is important that no two bi-directional switches are switched on at any instant, as shown pictorially in figure 8a. This would result in line-to-line short circuits and the destruction of the converter due to over currents. Also, the bi-directional switches for each output phase should not all be turned off at any instant, as shown in figure 8b. This would result in the absence of a path for the inductive load current, causing large over-voltages. These two considerations cause a conflict since semiconductor devices cannot be switched instantaneously due to propagation delays and finite switching times.



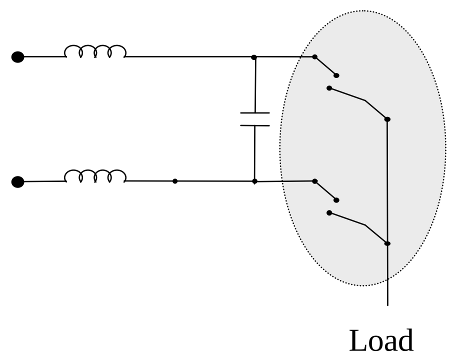


Fig. 8a. Avoid short circuits on the Matrix Converter input lines

Fig 8b. Avoid open circuits on the Matrix Converter output lines

A. Basic Current Commutation

The two simplest forms of commutation strategy intentionally break the rules given above and need extra circuitry to avoid destruction of the converter. In overlap current commutation, the incoming cell is fired before the outgoing cell is switched off. This would normally cause a line-to-line short circuit but extra line inductance slows the rise in current so that safe commutation is achieved. This is not a desirable method since the inductors used are large. The switching time for each commutation is also greatly increased which may cause control problems.

Dead time commutation uses a period where no devices are gated, causing a momentary open circuit of the load. Snubbers or clamping devices are then needed across the switch cells to provide a path for the load current. This method is undesirable since energy is lost during every commutation and the bi-directional nature of the switch cells further complicates the snubber design. The clamping devices and the power loss associated with them also results in increased converter volume.

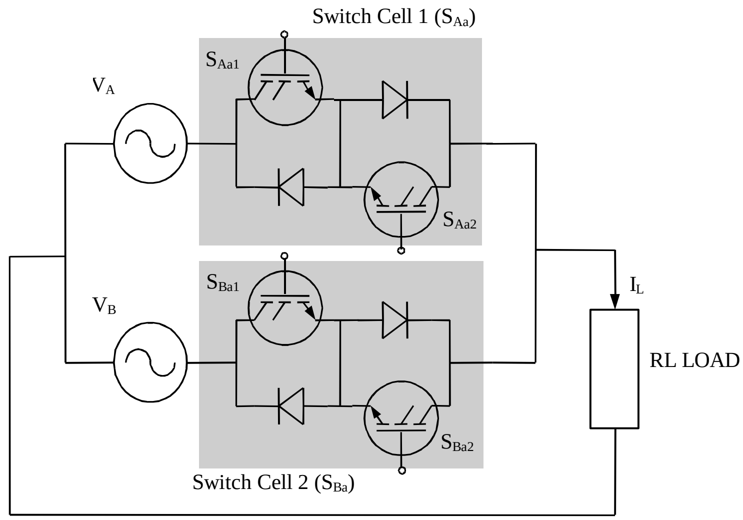


Fig 9. Two phase to single phase Matrix Converter

B. Current Direction Based Commutation

A more reliable method of current commutation, which obeys the rules, uses a four-step commutation strategy in which the direction of current flow through the commutation cells can be controlled. To implement this strategy the bi-directional switch cell must be designed in such a way as to allow the direction of the current flow in each switch cell to be controlled.

Figure 9 shows a schematic of a two phase to single phase Matrix Converter, representing the first two switches in the converter shown in figure 1. In steady state, both of the devices in the active bi-directional switch cell are gated to allow both directions of current flow. The following explanation assumes that the load current is in the direction shown and that the upper bi-directional switch (S_{Aa}) is closed. When a commutation to S_{Ba} is required, the current direction is used to determine which device in the active switch is not conducting. This device is then turned off. In this case device S_{Aa2} is turned off. The device that will conduct the current in the incoming switch is then gated, S_{Ba1} in this example. The load current transfers to the incoming device either at this point or when the outgoing device (S_{Aa1}) is turned off. The remaining device in the incoming switch (S_{Ba2}) is turned on to allow current reversals. This process is shown as a timing diagram in figure 10, the delay between each switching event is determined by the device characteristics.

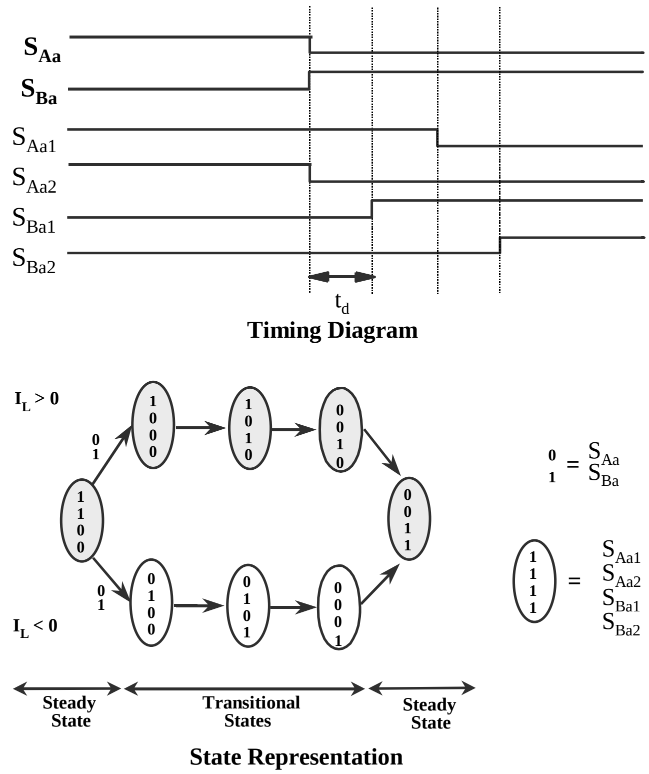


Fig 10 - Four Step Semi-Soft Current Commutation
between two bi-directional switch cells

This method allows the current to commutate from one switch cell to another without causing a line-to-line short circuit or a load open circuit. One advantage of all these techniques is that the switching losses in the silicon devices are reduced by 50% because half of the commutation process is soft switching, and hence this method is often called ‘semi-soft current commutation’ [46]. One popular variation on this current commutation concept is to only gate the conducting device in the active switch cell, which creates a two-step current commutation strategy [48].

All the current commutation techniques in this category rely on knowledge of the output line current direction. This can be difficult to reliably determine in a switching power converter, especially at low current levels in high power applications where traditional current sensors such as hall-effect probes are prone to producing uncertain results. One method that has been used to avoid these potential hazard conditions is to create a ‘near zero’ current zone where commutation is not allowed to take place, as shown for a two-step strategy in the state representation diagram in figure 11. However, this method will give rise to control problems at low current levels and at start-up.

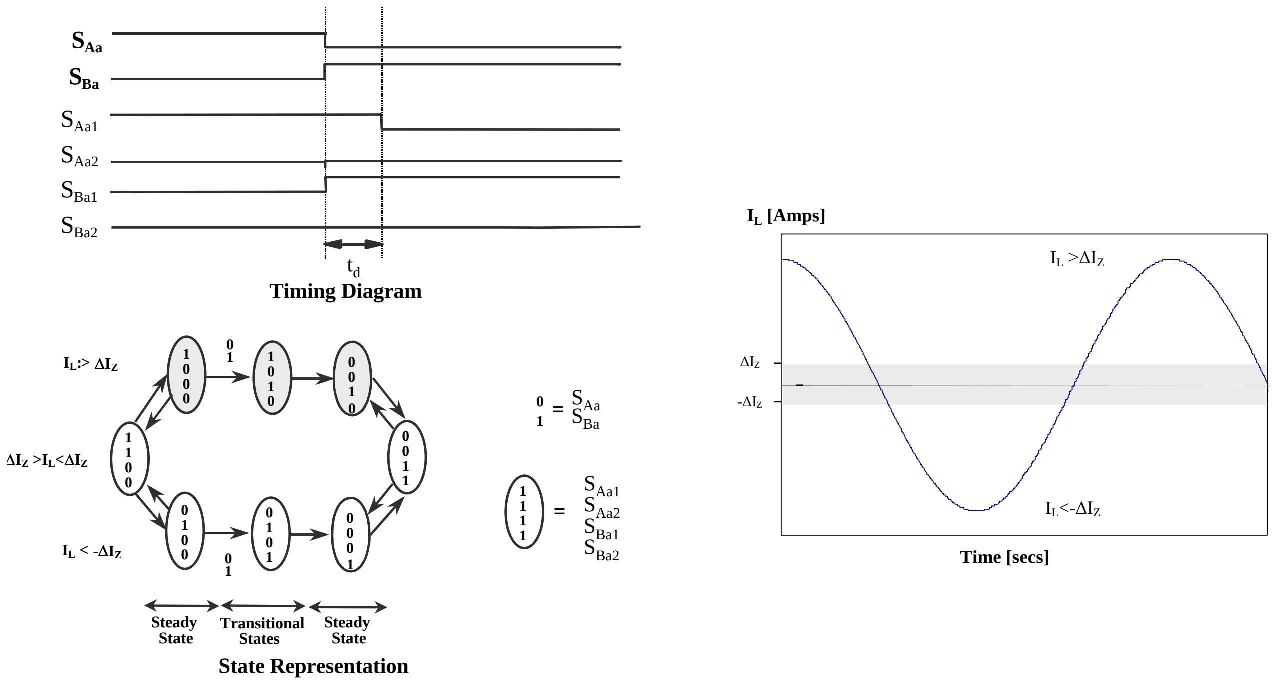


Fig. 11 - Two Step Semi-Soft Current Commutation
between two bi-directional switch cells

To avoid these current measurement problems a technique for using the voltage across the bi-directional switch to determine the current direction has been developed. This method allows very accurate current direction detection with no external sensors. Because of the accuracy available using this method, a two-step commutation strategy can be employed with deadtimes when the current changes direction, as shown in figure 12. This technique has been coupled with the addition of intelligence at the gate drive level to allow each gate drive to independently control the current commutation [21].

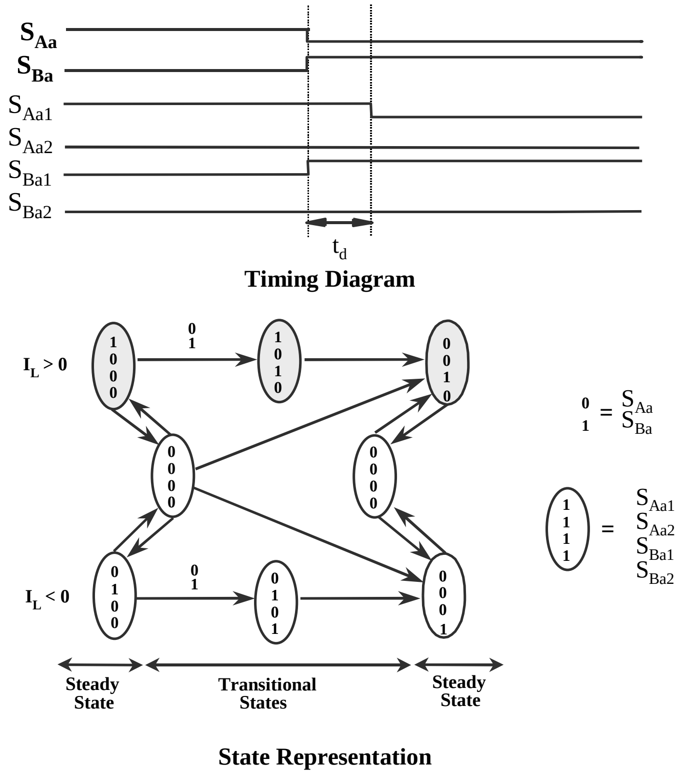


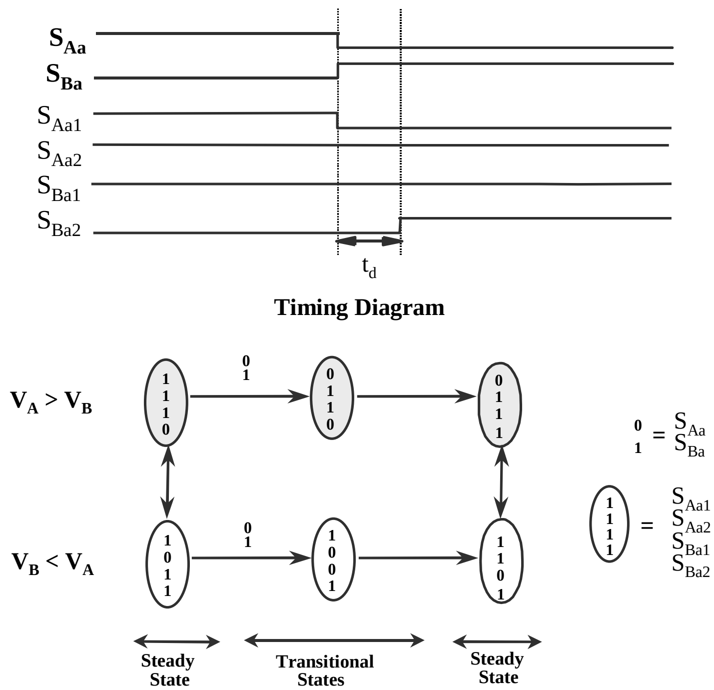
Fig 12. Two Step Semi-Soft Current Commutation
with current direction detection within the switch cell

C. Relative Voltage Magnitude Based Commutation

There have been two current commutation techniques proposed which use the relative magnitudes of input voltages to calculate the required switching patterns [37,50]. In the reduction to a two phase to single phase converter these both look identical and resulting timing and phase diagrams are shown in figure 13. The main difference between these methods and the current direction based techniques is that freewheel paths are turned on in the input voltage based methods. In ‘Metzi’ current commutation all the devices are closed except those required to block the reverse voltage [49]. This allows for relatively simple commutation of the current between phases. In [50] only one extra device is closed and the commutation process has to pass between the voltage of the opposite polarity during every commutation leading to higher switching losses. To successfully implement this type of commutation it is necessary to accurately measure the relative magnitudes of the input voltages.

C. Soft switching techniques

In many power converter circuits the use of resonant switching techniques has been proposed and investigated in order to reduce switching losses. In Matrix Converters resonant techniques have the additional benefit of solving the current commutation problem. The techniques developed fall into two categories: resonant switch circuits [51,52] and auxiliary resonant circuits [53]. All these circuits significantly increase the component count in the Matrix Converter; increase the conduction losses and most require modification to the converter control algorithm to operate under all conditions.



State Diagram

Fig. 13. Voltage based current commutation

V. MODULATION TECHNIQUES

A. Basic modulation solution

The modulation problem normally considered for the Matrix Converter can be stated as follows:

Given a set of input voltages and an assumed set of output currents

$$\mathbf{v_i} = V_{im}\begin{bmatrix} \cos(\omega_i t) \\ \cos(\omega_i t + 2\pi/3) \\ \cos(\omega_i t + 4\pi/3) \end{bmatrix}, \qquad \mathbf{i_o} = I_{om}\begin{bmatrix} \cos(\omega_o t + \phi_o) \\ \cos(\omega_o t + \phi_o + 2\pi/3) \\ \cos(\omega_o t + \phi_o + 4\pi/3) \end{bmatrix} \tag{11}$$

find a modulation matrix *M(t)* such that

$$\mathbf{v_o} = qV_{im}\begin{bmatrix} \cos(\omega_o t) \\ \cos(\omega_o t + 2\pi/3) \\ \cos(\omega_o t + 4\pi/3) \end{bmatrix} \text{ and } \quad \mathbf{i_i} = q\cos(\phi_o) I_{om}\begin{bmatrix} \cos(\omega_i t + \phi_i) \\ \cos(\omega_i t + \phi_i + 2\pi/3) \\ \cos(\omega_i t + \phi_i + 4\pi/3) \end{bmatrix} \tag{12}$$

and that the constraint equation (2) is satisfied. In (12) q is the voltage gain between the output and input voltages [10].

There are 2 basic solutions [10,11,54]:

$$\mathbf{M1} = \frac{1}{3}\begin{bmatrix} 1+2q\cos(\omega_m t) & 1+2q\cos(\omega_m t - 2\pi/3) & 1+2q\cos(\omega_m t - 4\pi/3) \\ 1+2q\cos(\omega_m t - 4\pi/3) & 1+2q\cos(\omega_m t) & 1+2q\cos(\omega_m t - 2\pi/3) \\ 1+2q\cos(\omega_m t - 2\pi/3) & 1+2q\cos(\omega_m t - 4\pi/3) & 1+2q\cos(\omega_m t) \end{bmatrix} \quad (13)$$

with $\omega_m = (\omega_o - \omega_i)$

and

$$\mathbf{M2} = \frac{1}{3}\begin{bmatrix} 1+2q\cos(\omega_m t) & 1+2q\cos(\omega_m t - 2\pi/3) & 1+2q\cos(\omega_m t - 4\pi/3) \\ 1+2q\cos(\omega_m t - 2\pi/3) & 1+2q\cos(\omega_m t - 4\pi/3) & 1+2q\cos(\omega_m t) \\ 1+2q\cos(\omega_m t - 4\pi/3) & 1+2q\cos(\omega_m t) & 1+2q\cos(\omega_m t - 2\pi/3) \end{bmatrix} \quad (14)$$

with $\omega_m = -(\omega_o + \omega_i)$

The solution in (13) yields $\phi_i = \phi_o$ giving the same phase displacement at the input and output ports whereas the solution in (14) yields $\phi_i = -\phi_o$ giving reversed phase displacement. Combining the two solutions provides the means for input displacement factor control.

This basic solution represents a direct transfer function approach and is characterised by the fact that, during each switch sequence time (T_{seq}), the average output voltage is equal to the demand (target) voltage. For this to be possible it is clear that the target voltages must fit within the input voltage envelope for any output frequency. This leads to a limitation on the maximum voltage ratio.

B. Voltage ratio limitation and optimisation

The modulation solutions in (13) and (14) have a maximum voltage ratio (q) of 50% as illustrated in figure 14.

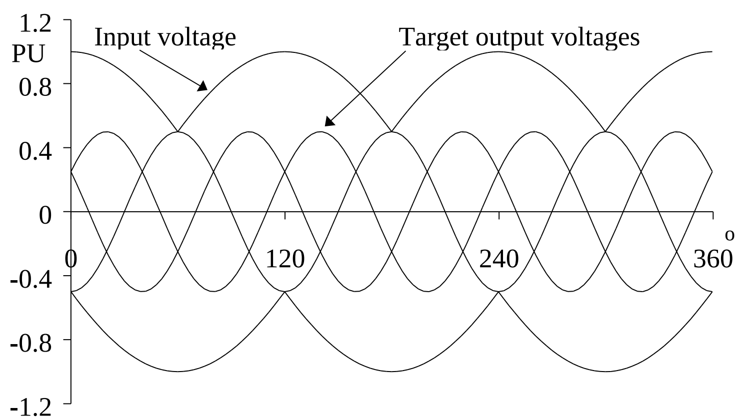


Fig. 14. Illustrating maximum voltage ratio of 50%

An improvement in the achievable voltage ratio to √3/2 (or 87%) is possible by adding common mode voltages to the target outputs as shown in equation (15).

$$\mathbf{v_o} = qV_{im}\begin{bmatrix} \cos(\omega_o t) - \frac{1}{6}\cos(3\omega_o t) + \frac{1}{2\sqrt{3}}\cos(3\omega_i t) \\ \cos(\omega_o t + 2\pi/3) - \frac{1}{6}\cos(3\omega_o t) + \frac{1}{2\sqrt{3}}\cos(3\omega_i t) \\ \cos(\omega_o t + 4\pi/3) - \frac{1}{6}\cos(3\omega_o t) + \frac{1}{2\sqrt{3}}\cos(3\omega_i t) \end{bmatrix} \quad (15)$$

The common mode voltages have no effect on the output line to line voltages, but allow the target outputs to fit within the input voltage envelope with a value of q up to 87% as illustrated in figure 15.

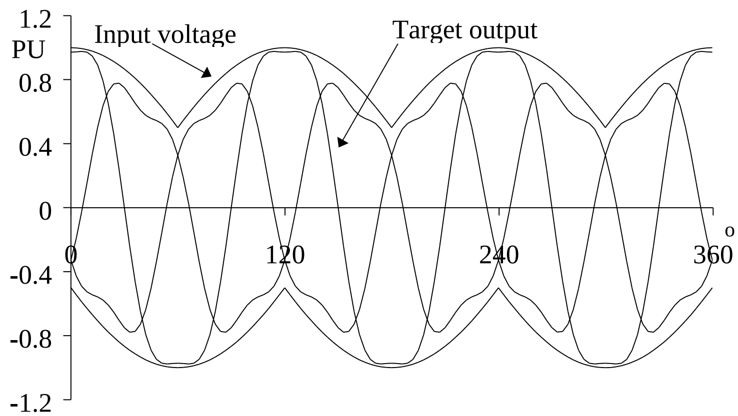


Fig. 15. Illustrating voltage ratio improvement to 87%

The improvment in voltage ratio is achieved by redistributing the null output states of the converter (all output lines connected to the same input line) and is analagous to the similar well established technique in conventional DC link PWM converters. It should be noted that a voltage ratio of 87% is the intrinsic maximum for any modulation method where the target output voltage equals the mean output voltage during each switching sequence. Venturini provides a rigorous proof of this fact in [55,56].

C. Venturini modulation methods

The first method attributable to Venturini [10,11,54] is defined by equations (13,14). However, calculating the switch timings directly from these equations is cumbersome for a practical implementation. They are more conveniently expressed directly in terms of the input voltages and the target output voltages (assuming unity displacement factor) in the form of (16).

$$m_{Kj} = \frac{t_{Kj}}{T_{seq}} = \frac{1}{3}\left[1 + \frac{2v_K v_j}{V_{im}{}^2}\right] \quad \text{for}\, K = A, B, C \,\text{and}\, j = a, b, c \quad (16)$$

This method is of little practical significance because of the 50% voltage ratio limitation.

Venturini’s optimum method [55,56] employs the common mode addition technique defined in equation (15) to achieve a maximum voltage ratio of 87%. The formal statement of the algorithm, including displacement factor control, in Venturini’s key paper [56] is rather complex and appears unsuited for real time implementation. In fact, if unity input displacement factor is required then the algorithm can be more simply stated in the form of (17).

$$m_{Kj} = \frac{1}{3}\left[1 + \frac{2v_K v_j}{V_{im}^{\ 2}} + \frac{4q}{3\sqrt{3}}\sin(\omega_i t + \beta_K)\sin(3\omega_i t)\right] \quad \text{for}\, K = A, B, C \;\text{ and }\; j = a, b, c \qquad (17)$$
$$\beta_K = 0, 2\pi/3, 4\pi/3 \text{ for } K = A,B,C \text{ respectively}$$

Note that in (17), the target output voltages, v_j, include the common mode addition defined in (15). Equation (17) provides a basis for real-time implementation of the optimum amplitude Venturini method which is readily handled by processors up to sequence (switching) frequencies of tens of kHz. Input displacement factor control can be introduced by inserting a phase shift between the measured input voltages and the voltages, v_K, inserted into (17). However, like all other methods, displacement factor control is at the expense of maximum voltage ratio.

Figure 3 shown previously illustrates typical line to supply neutral output voltage and current waveforms generated by the Venturini method.

C. Scalar modulation methods

The "scalar" modulation method of Roy [57,58] is typical of a number of modulation methods which have been developed where the switch actuation signals are calculated directly from measurements of the input voltages. The motivation behind their development is usually given as the perceived complexity of the Venturini method. The scalar method relies on measuring the instantaneous input voltages and comparing their relative magnitudes following the algorithm below.

Rule1: Assign subscript M to the input which has a different polarity to the other two.
Rule 2: Assign subscript L to the smallest (absolute) of the other two inputs. Third input is assigned subscript K

The modulation duty cycles are then given by:

$$m_{Lj} = \frac{(v_j - v_M)v_L}{1.5V_{im}^2},\ m_{Kj} = \frac{(v_j - v_M)v_K}{1.5V_{im}^2},\ m_{Mj} = 1 - (m_{Lj} + m_{Kj}) \ \text{ for } \ j = a, b, c \qquad (18)$$

Again, common mode addition is used with the target output voltages, v_j, to achieve 87% voltage ratio capability.

Despite the apparent differences, this method yields virtually identical switch timings to the optimum Venturini method. Expressed in the form of (17) the modulation duty cycles for the scalar method are given in (19).

$$m_{Kj} = \frac{1}{3}\left[1 + \frac{2v_K v_j}{V_{im}^{\ 2}} + \frac{2}{3}\sin(\omega_i t + \beta_K)\sin(3\omega_i t)\right] \qquad (19)$$

At maximum output voltage (q = √3/2), equations (17) and (19) are identical. The only difference between the methods is that the rightmost term addition is taken pro-rata with q in the Venturini method and is fixed at its maximum value in the scalar method. The effect on output voltage quality is negligible except at low switching frequencies where the Venturini method is superior.

C. Space vector modulation methods

The space vector method (SPVM) is well known and established in conventional PWM inverters. Its application to Matrix Converters is conceptually the same, but is more complex [41,42,43,45]. With a Matrix Converter, the SPVM can be applied to output voltage and input current control. A comprehensive discussion of the SPVM and its relationship to other methods is provided in another paper in this issue [49]. Here we just consider output voltage control to establish the basic principles.

The voltage space vector of the target Matrix Converter output voltages is defined in terms of the line to line voltages by (20).

$$\mathbf{V}_o(t) = \frac{2}{3}\left(v_{ab} + a v_{bc} + a^2 v_{ca}\right) \quad \text{where} \quad a = \exp(j2\pi/3) \tag{20}$$

In the complex plane, $\mathbf{V}_o(t)$ is a vector of constant length ($\sqrt{3}qV_{im}$) rotating at angular frequency ω_o. In the SPVM, $\mathbf{V}_o(t)$ is synthesised by time averaging from a selection of adjacent vectors in the set of converter output vectors in each sampling period. For a Matrix Converter, the selection of vectors is by no means unique and a number of possibilities exist [59] which are not discussed in detail here.

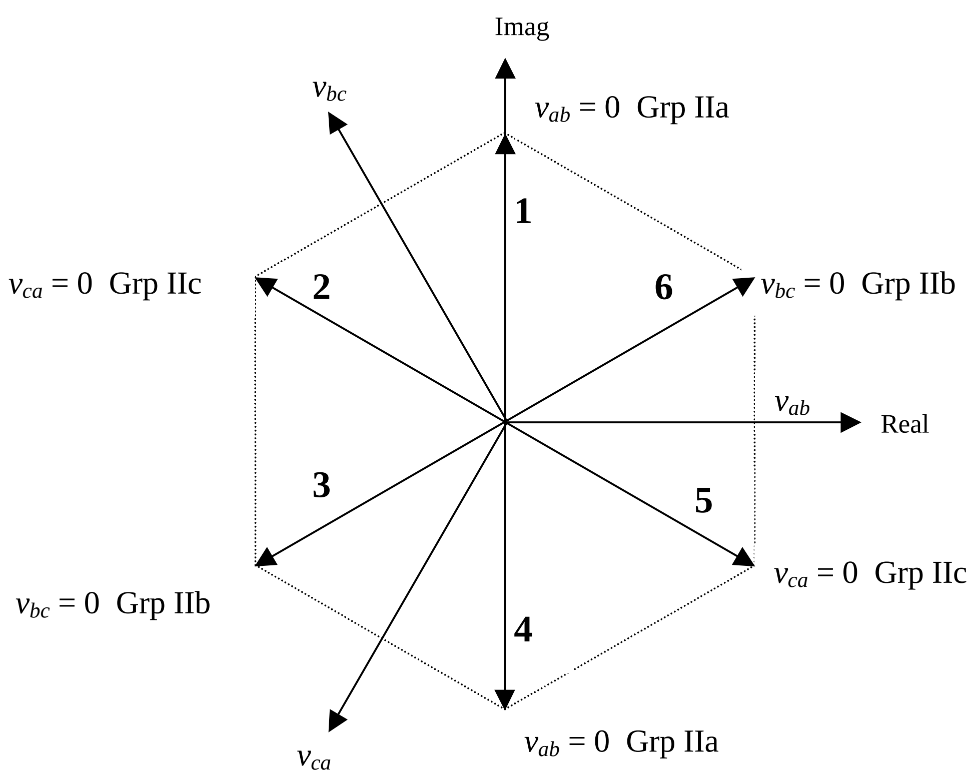


Fig. 16. Output voltage space vectors

The twentyseven possible output vectors for a three-phase Matrix Converter can be classified into three groups with the following characteristics:

- Group I: each output line is connected to a different input line. Output space vectors are constant in amplitude, rotating (in either direction) at the supply angular frequency.
- Group II: two output lines are connected to a common input line, the remaining output line is connected to one of the other input lines. Output space vectors have varying amplitude and fixed direction occupying one of six positions regularly spaced 60^0 apart. The maximum length of these vectors is $2/\sqrt{3}V_{env}$ where V_{env} is the instantaneous value of the rectified input voltage envelope.
- Group III: all output lines are connected to a common input line. Output space vectors have zero amplitude (ie located at the origin).

In the SPVM, the group I vectors are not used. The desired output is synthesised from the group II active vectors and the group III zero vectors. The hexagon of possible output vectors is shown in figure 16, where the group II vectors are further sub-divided dependent on which output line to line voltage is zero.

Figure 17 shows an example of how $\mathbf{V}_o(t)$ could be synthesised when it lies in the sextant between vector 1 and vector 6. $\mathbf{V}_o(t)$ is generated through time averaging by choosing the time spent in vector 1 (t_1) and vector 6 (t_6) during the switching sequence. Here it is assumed that the maximum length vectors are used, although that does not have to be the case. From figure 17 the relationship in (21) is found.

$$t_1 = \frac{|\mathbf{V}_o|}{V_{env}} T_{seq} \sin(\theta), \qquad t_6 = \frac{|\mathbf{V}_o|}{V_{env}} T_{seq} \sin(60-\theta), \qquad t_0 = T_{seq} - (t_1 + t_6) \tag{21}$$

where t_0 is the time spent in the zero vector (at the origin).

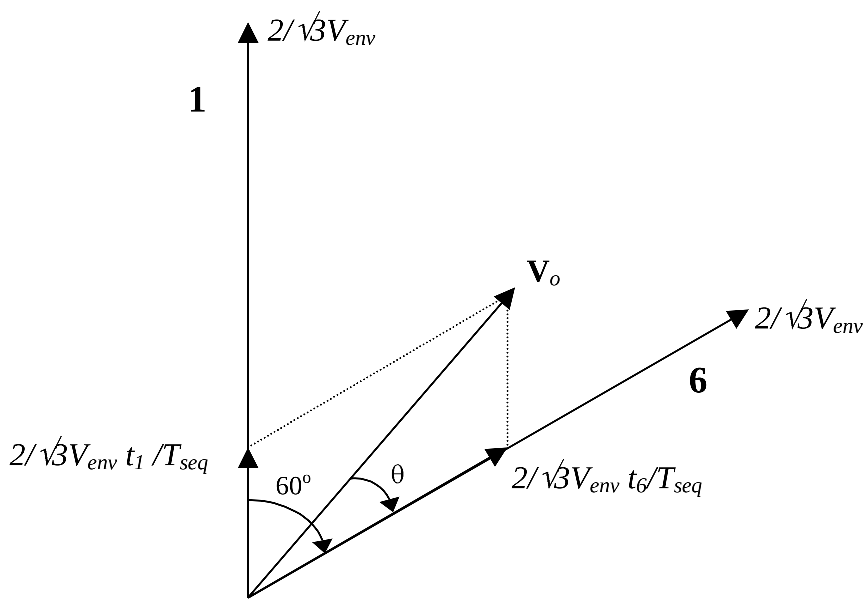


Fig. 17. Example of output voltage space vector synthesis

There is no unique way for distributing the times (t_1, t_6, t_0) within the switching sequence. One possible method is shown in figure 18.

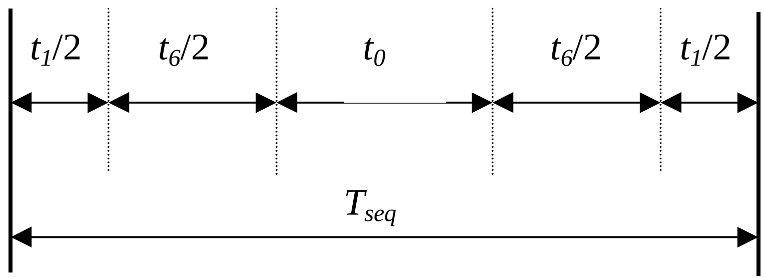


Fig. 18. Possible way of allocating states within switching sequence

For good harmonic performance at the input and output ports it is necessary to apply the SPVM to input current control and output voltage control. This generally requires four active vectors in each switching sequence, but the concept is the same. Under balanced input and output conditions, the SPVM technique yields similar results to the other methods mentioned earlier. However, the increased flexibility in choice of switching vectors for both input current and output voltage control can yield useful advantages under unbalanced conditions [59].

D. Indirect modulation methods

These methods aim to increase the maximum voltage ratio above the 86.6% limit of other methods [13, 40]. To do this the modulation process defined in (9) is split into two steps as indicated in (22).

$$\mathbf{v_o} = (\mathbf{A}\,\mathbf{v_i})\,\mathbf{B} \qquad (22)$$

In (22), pre-multiplication of the input voltages by **A** generates a "fictitious DC link" and post multiplication by **B** generates the desired output by modulating the fictitious DC link. **A** is generally referred to as the "rectifier transformation" and **B** as the "inverter transformation" due to the similarity in concept with a traditional rectifier/DC link/inverter system. **A** is given by (23).

$$\mathbf{A} = K_A \begin{bmatrix} \cos(\omega_i t) \\ \cos(\omega_i t + 2\pi/3) \\ \cos(\omega_i t + 4\pi/3) \end{bmatrix}^{\mathrm{T}} \qquad (23)$$

Hence:

$$\mathbf{Av_i} = K_A V_{im} \begin{bmatrix} \cos(\omega_i t) \\ \cos(\omega_i t + 2\pi/3) \\ \cos(\omega_i t + 4\pi/3) \end{bmatrix}^{\mathrm{T}} \begin{bmatrix} \cos(\omega_i t) \\ \cos(\omega_i t + 2\pi/3) \\ \cos(\omega_i t + 4\pi/3) \end{bmatrix} = \frac{3K_A V_{im}}{2} \qquad (24)$$

B is given by (25).

$$\mathbf{B} = K_B \begin{bmatrix} \cos(\omega_o t) \\ \cos(\omega_o t + 2\pi/3) \\ \cos(\omega_o t + 4\pi/3) \end{bmatrix} \tag{25}$$

Hence:

$$\mathbf{v_o} = (\mathbf{A}\mathbf{v_i})\mathbf{B} = \frac{3K_A K_B V_{im}}{2} \begin{bmatrix} \cos(\omega_o t) \\ \cos(\omega_o t + 2\pi/3) \\ \cos(\omega_o t + 4\pi/3) \end{bmatrix} \tag{26}$$

The voltage ratio $q = 3K_AK_B/2$. Clearly the **A** and **B** modulation steps are not continuous in time as shown above but must be implemented by a suitable choice of the switching states. There are many ways of doing this which are discussed in detail in [13, 40].

To maximise the voltage ratio the step in **A** is implemented so that the most positive and most negative input voltages are selected continuously. This yields $K_A = 2\sqrt{3}/\pi$ with a fictitious DC link of $3\sqrt{3}V_{im}/\pi$ (the same as a 6-pulse diode bridge with resistive load). K_B represents the modulation index of a PWM process and has the maximum value (squarewave modulation) of $2/\pi$ [13]. The overall voltage ratio q therefore has the maximum value of $6\sqrt{3}/\pi^2 = 105.3\%$.

The voltage ratio obtainable is obviously greater than that of other methods but the improvement is only obtained at the expense of the quality of either the input currents, the output voltages or both. For values of $q > 0.866$, the mean output voltage no longer equals the target output voltage in each switching interval. This inevitably leads to low frequency distortion in the output voltage and/or the input current compared to other methods with $q < 0.866$. For $q < 0.866$, the indirect method yields very similar results to the direct methods.

Figure 19 shows typical line to line output voltage and current waveforms obtained with the indirect method generating an output voltage with a frequency of 50Hz.

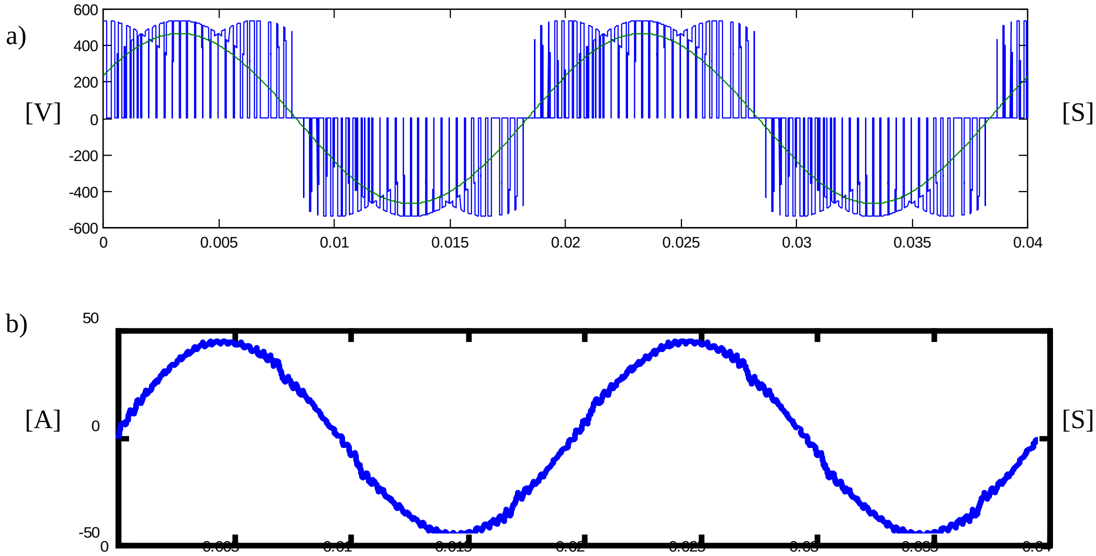


Fig. 19. Line to line voltage and current in the load with the indirect method.
Output frequency of 50 [Hz]

VI. PRACTICAL ISSUES

A. Input filters

Filters must be used at the input of the Matrix Converters to reduce the switching frequency harmonics present in the input current. The requirements for the filter are [30] :

i. To have a cut-off frequency lower than the switching frequency of the converter.
ii. To minimise its reactive power at the grid frequency.
iii. To minimise the volume and weight for capacitors and chokes.
iv. To minimise the filter inductance voltage drop at rated current in order to avoid a reduction in the voltage transfer ratio.

It must be noticed that this filter does not need to store energy coming from the load. Several filter configurations like simple LC and multi-stage LC have been investigated [29]. It has been shown that simple LC filtering, as shown in figure 20, is the best alternative considering cost and size [29], [30].

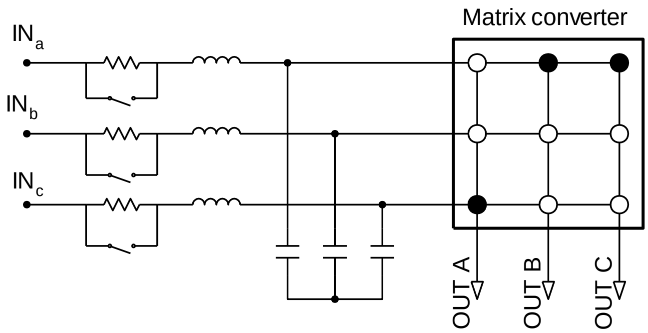


Fig. 20. Matrix converter with LC filter.

The Matrix Converter is expected to be the “pure silicon converter,” because it does not need large reactive elements to store energy. However, a recent study revealed that a Matrix Converter of 4kW needed a larger volume for reactive components than a comparable DC-link inverter [30], although this solution had not been optimised for volume. Some preliminary research works have been reported concerning the size reduction of the input filter [25].

Due to the LC configuration of the input filter, some problems appear during the power-up procedure of the Matrix Converter. It is well known that an LC circuit can create overvoltage during transient operation. The connection of damping resitors, as shown in figure 20, to reduce overvoltages is proposed in [31]. The damping resistors are shortcircuited when the converter is running. The use of damping resistors connected in parallel to the input reactors is proposed in [30].

B. *Overvoltage protection*

In a Matrix Converter overvoltages can appear from the input side, originated by line perturbations. Also dangerous overvoltages can appear from the output side, caused by an overcurrent fault. When the switches are turned off, the current in the load is suddenly interrupted. The energy stored in the motor inductance has to be discharged without creating dangerous overvoltages. A clamp circuit, as shown in figure 21, is the most common solution to avoid overvoltages coming from the grid and from the motor [32]. This clamp configuration uses twelve fast recovery diodes to connect the capacitor to the input and output terminals.

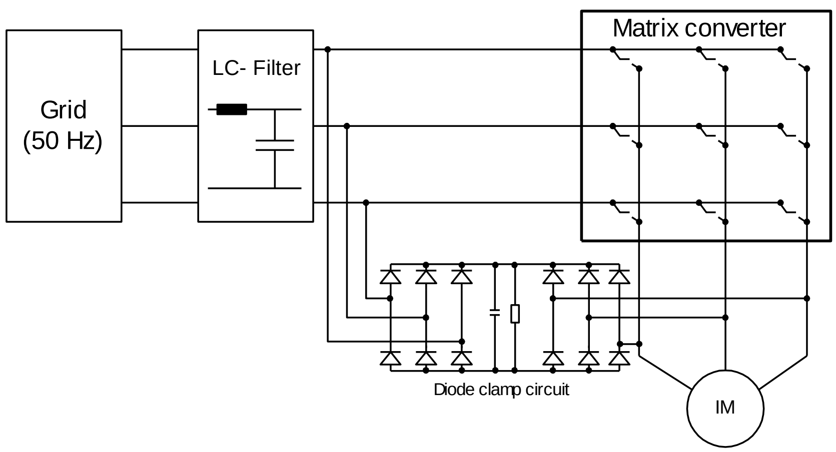


Fig. 21. Matrix converter with clamp.

A new clamp configuration uses six diodes from the bidirectional switches to reduce the extra diodes to six [23]. A different overvoltage protection strategy replaces the clamp by varistors connected at the input and at the output terminals, plus a simple extra circuit to protect each IGBT [24].

In [33], controlled shutdown of the converter without using a clamp is proposed. This strategy uses controlled freewheeling states to reduce the motor current to zero, avoiding the generation of overvoltages.

C. *Ride-through capability*

Ride-through capabability is a desired characteristic in modern drives [34],[35]. A common solution is to decelerate the drive during power loss, receiving energy from the load inertia to feed the control electronics and to magnetize the motor. This is achieved by maintaining a constant voltage in the DC-link capacitor. Matrix Converters do not have a DC-link capacitor and for this reason the previously mentioned strategy cannot be used.

Figure 22 shows a configuration proposed to provide short term ride-through capability to a Matrix Converter using the clamp capacitor as the source for a switch mode power supply which feeds the converter control circuit [28]. After detection of a perturbation in the power supply, the motor is disconnected from the grid, but the switches of the Matrix Converter do not interrupt the motor currents. By applying the zero voltage vector (short circuit of the motor leads), the stator currents and the energy stored in the leakage inductance increases. The disconnection of the active switches originates the conduction of the clamp capacitor. This energy is then used to feed the control circuits. A flux and speed observer is used to restart the drive from non zero flux and speed conditions in the shortest time.

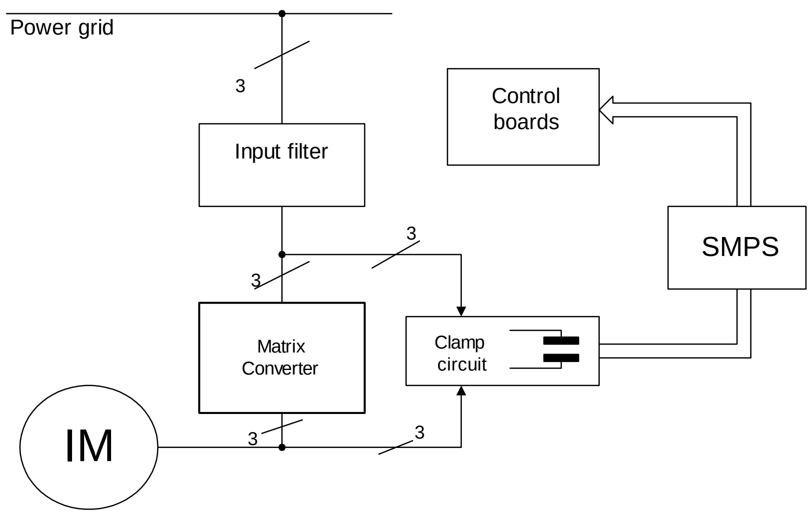


Fig. 22. Configuration to achieve ride-throug capability.

VII. COMMENTS AND CONCLUSIONS.

After two decades of research effort several modulation and control methods have been developed for the Matrix Converter, allowing the generation of sinusoidal input and output currents, operating with unity power factor using standard processors. The most important practical implementation problem in the Matrix Converter circuit, the commutation problem between two controlled bi-directional switches, has been solved with the development of highly intelligent multistep commutation strategies. The solution to this problem has been made possible by using powerful digital devices that are now readily available in the market.

Another important drawback that has been present in all evaluations of Matrix Converters was the lack of a suitably packaged bi-directional switch and the large number of power semiconductors. This limitation has recently been overcome with the introduction of power modules which include the complete power circuit of the Matrix Converter. However, research work has shown that the Matrix Converter is not a “pure silicon converter” and that passive elements in form of input filters are needed. More work must be done in order to optimise the size of these filters.

Twenty years ago the Matrix Converter had the potential to be a superior converter in terms of its performance. Now, the Matrix Converter faces a very strong competition from the voltage source inverter (VSI) with a three phase Active Front End (AFE). This fully regenerative VSI-AFE topology has similar operating characteristics of sinusoidal input and output currents and adjustable power factor. In addition, the technology is mature and well established in the market. The real challenge for the Matrix Converter is to be accepted in the market. In order to achieve this goal the Matrix Converter must overcome the VSI-AFE solution in terms of costs, size and reliability.

VIII. REFERENCES.

[1] L. Gyugi, and B. Pelly, *Static Power Frequency Changers : Theory, Performance and Applications*, John Wiley and Sons, 1976.

[2] A. Brandt, "Der Netztaktumrichter," Bull, ASE 62(1971) 15, 24 Juillet, pp. 714-727.

[3] W. Popov, "Der Direktumrichter mit zyklischer Steuerung," Elektrie 29 (1975) H.7, pp 372-376.

[4] E. Stacey, "An unrestricted frequency changer employing force commutated thyristors," in *Conf. Rec. PESC'76*, pp. 165-173.

[5] V. Jones, and B. Bose, "A frequency step-up cycloconverter using power transistors in inverse-series mode," *Int. Journal Electronics*, Vol. 41, no. 6, pp. 573-587, 1976.

[6] M. Steinfels, and P. Ecklebe, "Mit Direktumrichter Gespeiste Drehstromantriebe für den Industriellen Einsatz in einem Weiten Leistungsbereich," Elektrie 34 (1980) H.5, pp. 238-240.

[7] P. Ecklebe, "Transistorisierter Direktumrichter für Drehstromantriebe," Elektrie 34 (1980) H.8, pp. 413-433.

[8] A. Daniels, and D. Slattery, "New power converter technique employing power transistors," IEE Proc. Vol. 125 no. 2, pp. 146-150, February 1978.

[9] A. Daniels, and D. Slattery, "Application of power transistors to polyphase regenerative power converters,", IEE Proc. Vol. 125, no. 7, pp. 643-647, July 1978.

[10] M. Venturini, "A new sine wave in sine wave out, conversion technique which eliminates reactive elements," *Proc. POWERCON 7*, 1980 pp. E3_1-E3_15.

[11] M. Venturini and A. Alesina, "The generalised transformer: A new bidirectional sinusoidal waveform frequency converter with continuously adjustable input power factor", in *Conf. Rec. IEEE PESC'80*, pp. 242-252.

[12] J. Rodriguez, "A new control technique for AC-AC converters", IFAC Control in Power Electronics and Electrical Drives Lausanne Switzerland, 1983, pp. 203-208.

[13] P. D. Ziogas, S.I. Khan and M. H. Rashid, "Analysis and design of forced commutated cycloconverter structures with improved transfer characteristics," *IEEE Trans. Ind. Electron.*, vol. IE-33, no. 3, pp. 271-280. Aug. 1986.

[14] L. Huber, D. Borojevic and N. Burany, "Voltage space vector based PWM control of forced commutated cycloconvertors", *IEEE IECON*, 1989, pp106-111.

[15] J. Oyama, T. Higuchi, E. Yamada, T. Koga and T. Lipo, "New control strategy for Matrix Converter," in *Conf. Rec. IEEE PESC'89*, pp. 360-367.

[16] M. Braun, and K. Hasse, "A direct frequency changer with control of input reactive power", in *IFAC Control in Power Electronics and Electrical Drives*, Lausanne Switzerland, 1983, pp187-194.

[17] E. Wiechmann, J. Espinoza, L. Salazar, and J. Rodriguez, "A direct frequency converter controlled by space vectors," in *Conf. Rec. IEEE PESC'93*, pp314-320.

[18] G. Kastner, and J. Rodriguez, "A forced commutated cycloconverter with control of the source and load currents," in *Proc. EPE'85,* pp. 1141-1146.

[19] C. L. Neft, and C. D. Schauder, "Theory and design of a 30-HP Matrix Converter," *IEEE Trans. Ind. Applicat.*, vol. 28, no. 3, pp. 546-551, May/June 1992.

[20] J.H. Youm, and B.H Kwon, "Switching technique for current-controlled AC-to-AC converters," *IEEE Trans. Ind. Applicat.*, vol. 46, no. 2, April 1999.

[21] L. Empringham, P. Wheeler, and J. Clare, "Intelligent commutation of Matrix Converter bi-directional switch cells using novel gate drive techniques," in *Conf. Rec. IEEE PESC'98*, pp. 707-713.

[22] M. Ziegler, and W. Hofmann, "Performance of a two steps commutated Matrix Converter for ac-variable-speed drives," in *Proc. EPE'99,* CD-ROM.

[23] P. Nielsen, F. Blaabjerg, and J. Pedersen, "Novel solutions for protection of Matrix Converter to three phase induction machine," in *Conf. Rec. IEEE Ind. Applicat. Soc. Annu. Meeting, 1997,* pp.1447-1454.

[24] J. Mahlein, and M. Braun, "A Matrix Converter without diode clamped over-voltage protection," in *Conf. Proc. IPEMatrix Converter'2000,* pp. 817-822.

[25] C. Klumpner, P. Nielsen, I. Boldea, and F. Blaabjerg, "New steps towards a low-cost power electronic building block for Matrix Converters," *Conf. Rec. IEEE Ind. Applicat. Soc. Annu. Meeting, 2000,* CD-ROM.

[26] J. Chang, T. Sun, A. Wang, and D. Braun, "Medium power AC-AC converter based on integrated bidirectional power modules, adaptive commutation and DSP control," *Conf. Rec. IEEE Ind. Applicat. Soc. Annu. Meeting, 1999,* CD-ROM.

[27] D. Casadei, G. Serra, A. Tani, and P. Nielsen, "Theoretical and experimental analysis os SVM-controlled Matrix Converters under unbalanced supply conditions," Electromotion 4 (1997), pp. 28-37.

[28] C. Klumpner, I. Boldea, and F. Blaabjerg, "Short term ride through capabilities for direct frequency converters," in *Conf. Rec. IEEE PESC'00*, CD-ROM.

[29] P. Wheeler, H. Zhang, and D. Grant, "A theoretical and practical consideration of optimised input filter design for a low loss Matrix Converter", IEE PEVD, September 1994, pp. 363-367.

[30] C. Klumpner, P. Nielsen, I. Boldea, and F. Blaabjerg, "A new Matrix Converter-motor (Matrix ConverterM) for industry applications," *Conf. Rec. IEEE Ind. Applicat. Soc. Annu. Meeting, 2000,* CD-ROM.

[31] C. Klumpner, F. Blaabjerg, "The Matrix Converter: overvoltages caused by the input filter, bidirectional power flow, and control for artificial loading of induction motors," *Electric Machines and Power Systems*, vol. 28, pp. 129-242, 2000.

[32] U.S. Patent Application 4. 697.230, 1986, Neft/Westeinghouse Electric Corp.

[33] A. Shuster, "A Matrix Converter without reactive clamp elements for an induction motor drive system," in *Conf. Rec. IEEE PESC'98,* pp. 714-720.

[34] J. Holtz, W. Lotzkat, "Controlled AC drives with ride-through capability at power interruption," *IEEE Trans. Ind. Applicat.*, vol. 30, no. 5, pp1275-1283, xxx/yyy 1994.

[35] A. Von Jouanne, P. Enjeti, B. Banerjee, “Assessment of ride-through alternatives for adjustable speed drives,” *Conf. Rec. IEEE Ind. Applicat. Soc. Annu. Meeting 1998*, pp. 1538-1545.

[36] N. Burany, “Safe control of four-quadrant switches”, *Conf. Rec. IEEE Ind. Applicat. Soc. Annu. Meeting 1989*, pp. 1190-1194.

[37] M. Ziegler and W. Hofmann, “Semi natural two steps commutation strategy for Matrix Converters”, in *Conf. Rec. IEEE PESC’98*, pp727-731.

[38] L. Empringham P. Wheeler and J. Clare, “Bi-directional switch current commutation for Matrix Converter applications”, *Conf. Rec. PEMatrix Converter Prague, September 1998.*, pp42-47.

[39] L. Empringham, P. Wheeler and J. Clare, “Matrix converter bi-directional switch commutation using intelligent gate drives” *Conf. Rec. IEE PEVD, London, 1998*, pp626-631.

[40] P. Ziogas, S. Khan and M. Rashid, “Some improved forced commutated cycloconverter structures”, *IEEE Trans. Ind. Applicat.*, vol. 1A-21, no. 5, pp. 1242-1253, Sept/Oct 1985.

[41] L. Huber and D. Borojevic, “Space vector modulator for forced commutated cycloconverters”, *Conf. Rec. IEEE IAS, 1989*, pp871-876.

[42] L. Huber, D. Borojevic and N. Burany, “Analysis design and implementation of the space-vector modulator for forced-commutated cycloconvertors”, *IEE Proceedings-B Vol. 139 No.2, March 1992*, pp103-113.

[43] L. Huber, D. Borojevic, X. Zhuang and F. Lee, “Design and implementation of a three-phase to three-phase Matrix Converter with input power factor correction”, *IEEE APEC Conf, 1993*, pp860-865.

[44] L. Huber, D. Borojevic and N. Burany N, “Digital implementation of the space vector modulator for forced commutated cycloconverters”, *IEE PEVD Conf , 1990* , pp63-65.

[45] L. Huber and D. Borojevic, “Space vector modulated three phase to three phase Matrix Converter with input power factor correction,” *IEEE Trans. Ind. Applicat.*, vol. 31, no. 6, pp. 1234-1246, November/December 1995.

[46] P. Wheeler and D.Grant, “Optimised Input Filter Design and Low Loss Switching Techniques for a Practical Matrix Converter”, *IEE Proceedings Part B, Vol. 144, No. 1, January 1997*, pp 53-60.

[47] M. Munzer, ”EconoMac – the First All In One IGBT Module for Matrix Converters“, *Drives and Control Conference, Section 3, London, 2001*.

[48] T. Svensson and M. Alakula, “The Modulation and Control of a Matrix Converter Synchronous Machine Drive”, *Conf. Rec. EPE’91, Firenze, 1991*, pp 469-476.

[49] Casadei D, Serra G, Tani A and Zarri L, "Matrix Converter Modulation Strategies: A New General Approach Based on Space Vector Representation of the Switch States", This issue.

[50] Kwon B.H., Min B.H. and Kim J.H., "Novel Commutation Technique of AC-AC converters", *IEE Proceedings Part B, July 1998*, pp295-300.

[51] Pan C.T. Chen T.C. and Shieh J.J., "A Zero Switching Loss Matrix Converter", *IEEE PESC Conf, 1993*, pp545-550.

[52] Villaça M.V.M. and Perin A J., "A Soft Switched Direct Frequency Changer", *IEEE IAS, 1995*, pp2321-2326.

[53] Cho J.G. and Cho G.H., "Soft Switched Matrix Converter for High Frequency Direct AC-to-AC Power Conversion", *Conf. Rec. EPE’91, Firenze, 1991*, pp4-196-4-201.

[54] Alesina A. and Venturini M.G.B., "Solid-State Power Conversion: A Fourier Analysis Approach to Generalized Transformer Synthesis", *IEEE Transactions on Circuits and Systems Vol. Cas-28 No.4, April 1981*, pp319-330.

[55] Alesina A. and Venturini M., "Intrinsic Amplitude Limits and Optimum Design of 9-Switches Direct PWM AC-AC Converters", *Conf. Rec. IEEE PESC, April 1988,* pp1284-1291.

[56] Alesina A. and Venturini M.G.B., "Analysis and Design of Optimum-Amplitude Nine-Switch Direct AC-AC Converters", *IEEE Transactions on Power Electronics Vol. 4 .No.1., January 1989*, pp101-112.

[57] Roy G. Duguay L. Manias S. and April G.E., "Asynchronous Operation of Cycloconverter with Improved Voltage Gain by Employing a Scalar Control Algorithm", *Conf. Rec. IEEE IAS, 1987*, pp889-898.

[58] Roy G. and April G. E., "Cycloconverter Operation Under a New Scalar Control Algorithm", *Conf. Rec. IEEE PESC*, *1989*, pp368-375.