Current Control Techniques for Three-Phase Voltage-Source PWM Converters: A Survey

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***Abstract*— The aim of this paper is to present a review of recently used current control techniques for three-phase voltage-source pulsewidth modulated converters. Various techniques, different in concept, have been described in two main groups: *linear* and *nonlinear*. The first includes proportional integral stationary and synchronous) and state feedback controllers, and predictive techniques with constant switching frequency. The second comprises bang-bang (hysteresis, delta modulation) controllers and predictive controllers with on-line optimization. New trends in the current control—neural networks and fuzzy-logic-based controllers—are discussed, as well. Selected oscillograms accompany the presentation in order to illustrate properties of the described controller groups.**

***Index Terms*— AC motor drives, current control, inverters, power filters, pulsewidth modulation, switch-mode rectifiers.**

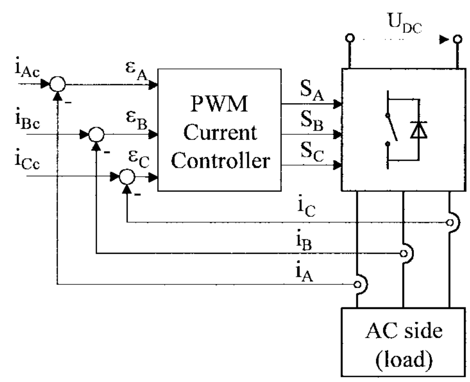

Fig. 1. Basic block diagram of CC-PWM converter.

I. Introduction

Most applications of three-phase voltage-source pulsewidth modulated (VS-PWM) converters—ac motor drives, active filters, high power factor ac/dc converters, uninterruptible power supply (UPS) systems, and ac power supplies—have a control structure comprising an internal *current feedback loop*. Consequently, the performance of the converter system largely depends on the quality of the applied current control strategy. Therefore, current control of PWM converters is one of the most important subjects of modern power electronics. In comparison to conventional open-loop voltage PWM converters, the *current-controlled PWM (CC-PWM) converters* have the following advantages:

1) control of instantaneous current waveform and high accuracy;
2) peak current protection;
3) overload rejection;
4) extremely good dynamics;
5) compensation of effects due to load parameter changes (resistance and reactance);
6) compensation of the semiconductor voltage drop and dead times of the converter;
7) compensation of the dc-link and ac-side voltage changes.

Development of PWM current control methods is still in progress. The purpose of this paper is to give a short review of the available CC techniques for the three-phase, two-level converters. The basic approaches and performance of the various methods are summarized. However, due to space limitations, a quantitative comparison of the methods under discussion is not included.

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II. Basic Concepts

A. Basic Scheme of CC-PWM

The main task of the control scheme in a CC-PWM converter (Fig. 1) is to force the currents in a three-phase ac load to follow the reference signals. By comparing the command i_{Ac} (i_{Bc}, i_{Cc}) and measured i_A (i_B, i_C) instantaneous values of the phase currents, the CC generates the switching states S_A (S_B, S_C) for the converter power devices which decrease the current errors ε_A $(\varepsilon_B, \varepsilon_C)$. Hence, in general, the CC implements two tasks: error compensation (decreasing $\varepsilon_A, \varepsilon_B, \varepsilon_C$) and modulation (determination of switching states S_A, S_B, S_C).

B. VS Converter as Power Amplifier

A three-phase VS bridge converter [Fig. 2(a)] is a discontinuously operated power amplifier, the operation of which has been extensively investigated and analyzed in literature [1]–[5], [8], [9], [16], [18], [20]. However, some basic operation constraints and limitations, which are important from the point of view of current control, are recalled below.

1) Modulation: The VS converter generates, at each output phase x $(x = A, B, C)$, a voltage u_x with a two-level rectangular waveform [Fig. 2(c)]. In conventional hard-switched VS bridge converters, there are no mutual constraints between phase switching instants, so that the pulse length can be varied

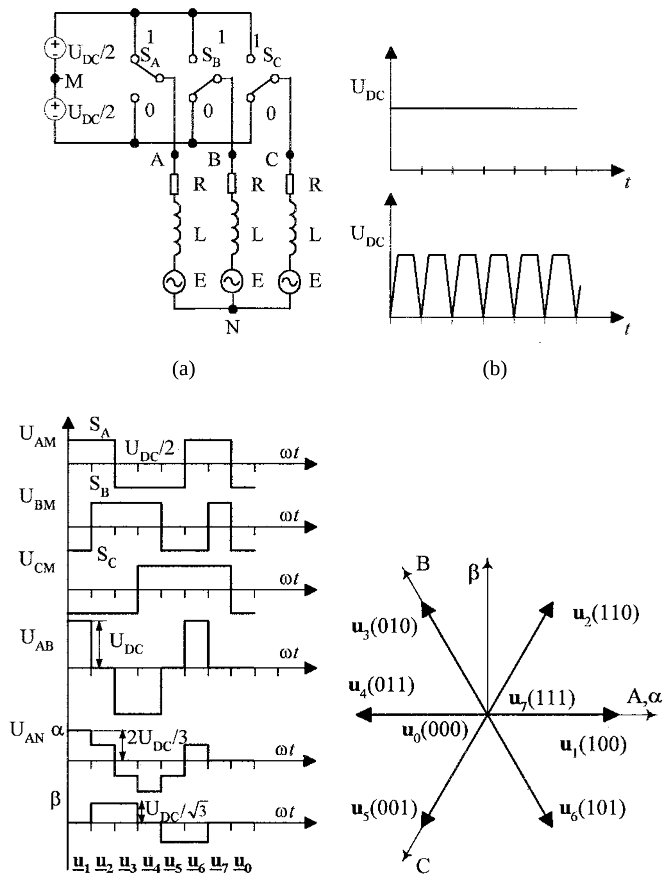


Fig. 2. Three-phase VS bridge converter. (a) Simplified main circuit topology. (b) DC-link voltage for hard and soft switching [resonant dc-link (RDCL) inverter]. (c) Time representation of the output ac voltages. (d) Vector representation of the output ac voltages.

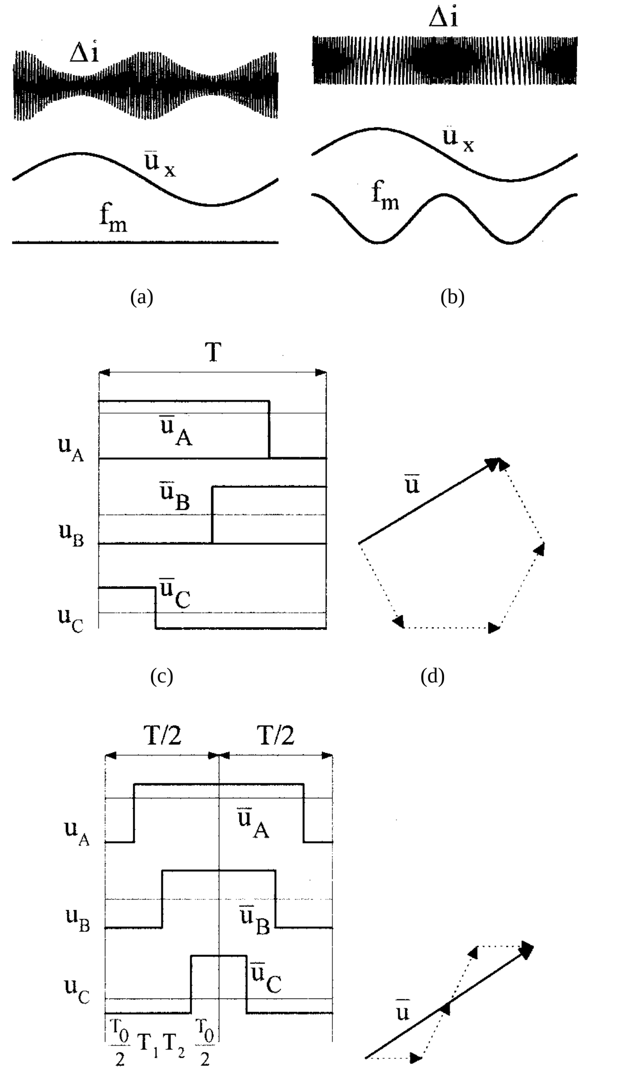


Fig. 3. (a) and (b) Ripple and modulation frequency. (c), (e) PWM pulse patterns and (d), (f) its vector representation.

continuously (PWM). In some cases, however, commutation mechanisms [*RDCL inverters*, Fig. 2(b)] or control systems (e.g., delta modulation (DM), Fig. 9) allow commutations only at fixed times. The modulation process controls the phase-switching sequence according to a given command u_{xc}, so that the phase voltage low-order harmonics result in a voltage $\overline{u}_x$ (average over the modulation period), the waveform of which should follow u_{xc} as closely as possible. Modulation generates high-order voltage harmonics, located around the switching frequency. If the latter is high enough, the two groups are quite separated from each other.

2) Current Ripple and Switching Frequency: Modulation also produces instantaneous deviations (ripple) of the current i_x from its average $\overline{i}_x$, as an effect of the voltage harmonics. Irrespective of the kind of modulation technique used, the ripple amplitude depends on the duration of the modulation period T (or on the modulation frequency $f_m = 1/T$), the supply voltage U_{DC}, the ac-side average voltage $\overline{u}_x$, and on the load parameters (R, L, E). With a purely inductive load $(R = 0)$, the peak-to-peak ripple amplitude Δi can be expressed as

$$\Delta i = \frac{T}{L}\overline{u}_x\left(1 - \frac{\overline{u}_x}{U_{\mathrm{DC}}}\right).$$

Note that, if voltage $\overline{u}_x$ varies [Fig. 3(a)], for a constant modulation period T (and frequency f_m) the ripple amplitude varies, too. However, if the ripple amplitude is kept constant, the modulation frequency must vary, as shown in Fig. 3(b).

Usually, losses put a limitation on the average switching frequency of each phase. In some cases, the control system, filtering, or other needs may also require the switching frequency to be constant.

3) Phase Interference Effect: If the neutral of the three-phase load N and the converter midpoint M (when available) are not connected [Fig. 2(a)], phase currents depend only on the voltage difference between phases. Therefore, a common term can be added to the phase voltages, thus shifting their mean value u_N, without affecting load average currents $\overline{i}_x$. The current ripple, however, is changed by the shift. This shift is often used to extend the maximum phase voltage which can be produced by the converter (third harmonic PWM) and to minimize the ripple, or to reduce the average switching frequency (flat-top PWM) [4], [8], [19], [55].

While phase voltages can be controlled independently, phase currents are determined not only by their own phase voltage, but also by those of other phases. Thus, a *phase interference* occures. This phenomenon has to be taken into account in designing CC.

4) Voltage Vector Sequence and Current Ripple: The converter output voltage can be represented as a space vector [Fig. 2(d)]. This is particularly suitable when considering the phase voltage effects on the load [4], [12], [14]. Vector sequences with the same resultant give equal mean voltages $\overline{u}_x$ and, therefore, equal average current $\overline{i}_x$ in an inductive load [Fig. 3(d) and (f)]. On the other hand, different vector paths produce different current ripples. A sensible ripple reduction, mainly at high modulation index, is obtained when phase pulses are centered and symmetrical, with a choice of u_N corresponding to Fig. 3(e). This condition results in a maximum zero-state duration and, in vector representation [Fig. 3(f)], in an equal length for states 0 and 7 [4], [8].

5) DC-Link Voltage Limit: A voltage reserve is required to force an ac-side (load) current according to its command value. For small amplitudes E of ac-side voltage, the dc-link voltage $U_{\rm DC}$ is not critical. However, as E is increased, a point is reached where the converter passes to a six-step square-wave operation and the CC is not capable of forcing the command current. Therefore, the converter requires a sufficient supply voltage reserve to force the ac line current in the entire E and load range.

C. Basic Requirements and Performance Criteria

The accuracy of the CC can be evaluated with reference to basic requirements, valid in general, and to specific requirements, typical of some applications. Basic requirements of a CC are the following:

1) no phase and amplitude errors (ideal tracking) over a wide output frequency range;
2) to provide high dynamic response of the system;
3) limited or constant switching frequency to guarantee safe operation of converter semiconductor power devices;
4) low harmonic content;
5) good dc-link voltage utilization.

Note that some of the requirements, e.g., fast response and low harmonic content, contradict each other. The specific requirements for the most important applications can be summarized as follows.

1) VS PWM inverters

a) AC motor control: This requires a wide range of output frequency, variable ac-side voltage (motor EMF), high dynamic, decoupled *d–q* control structure, operation in PWM/square-wave transient region.

b) AC power supply/UPS: This requires a narrow range of output frequency (UPS), reduced harmonic content (output filter), and fault protection.

2) VS PWM AC/DC Converters and Active Filters: These require constant ac-side (line power) frequency 50/60 Hz, nearly constant amplitude and waveform of ac-side voltage, poorly damped ac-side network, and variable dc-link voltage (ac/dc converters and power filter).

TABLE I
PERFORMANCE CRITERIA

CRITERIA DEFINITION	COMMENTS
$\mathrm{RMS} = [1/T\int(\varepsilon_\alpha^2 + \varepsilon_\beta^2)dt]^{1/2}$	• the r.m.s. vector error
$J = 1/T\int[(\varepsilon_\alpha^2 + \varepsilon_\beta^2)]^{1/2}dt$	• the vector error integral
$N = \Sigma\, \mathrm{imp}\,\vert_{t\,\in\,\langle 0,T\rangle}$	• number of switchings (also for nonperiodical)
$I_{hrms} = [1/T\int(i_{(t)} - i_{1(t)})^2dt]^{1/2}$	the r.m.s harmonic current
$d = I_{hrms} / I_{hrms\ six\text{-}step}$	the distortion factor
$d = [\Sigma\, h_i^2(k\cdot f_1)]^{1/2} \quad k\neq 1$	synchronised PWM case
$d = [\int h_d^2(f)\, df]^{1/2} \quad f\neq f_1$	nonsynchronised PWM case

$h_i(k\cdot f_1)$ - discrete current spectra

$h_d(f)$ - density current spectra

The evaluation of CC may be done according to performance criteria which include static and dynamic performance. Table I presents the static criteria in two groups:

1) those valid also for open-loop voltage PWM (see e.g., [1], [8], [9], [16]);
2) those specific for CC-PWM converters based on current error definition (denoted by •).

The following parameters of the CC system dynamic response can be considered: dead time, settling time, rise time, time of the first maximum, and overshoot factor. The foregoing features result both from the PWM process and from the response of the control loop. For example, for deadtime, the major contributions arise from signal processing (conversion and calculation times) and may be appreciable, especially if the control is of the digital type. On the other hand, rise time is mainly affected by the ac-side inductances of the converter. The optimization of the dynamic response usually requires a compromise which depends on the specific needs. This may also influence the choice of the CC technique according to the application considered.

In general, the compromise is easier as the switching frequency increases. Thus, with the speed improvement of today's switching components [e.g., insulated gate bipolar transistors (IGBT's)], the peculiar advantages of different methods lose importance, and even the simplest one may be adequate. Nevertheless, for some applications with specific needs, like active filters, which require very fast response or high power inverters where the commutations must be minimized, the most suitable CC technique must be selected.

D. Presentation of CC Techniques

Existing CC techniques can be classified in different ways [3], [8], [9], [11]–[13], [15], [27]. In this paper, the CC techniques are presented in two main groups, linear and nonlinear controllers.

III. LINEAR CONTROLLERS

The linear controllers operate with conventional voltage-type PWM modulators [21]–[36]. In contrast to the nonlinear controllers (see Section IV), linear controller schemes have clearly separated current error compensation and voltage modulation parts. This concept allows us to exploit the advan-

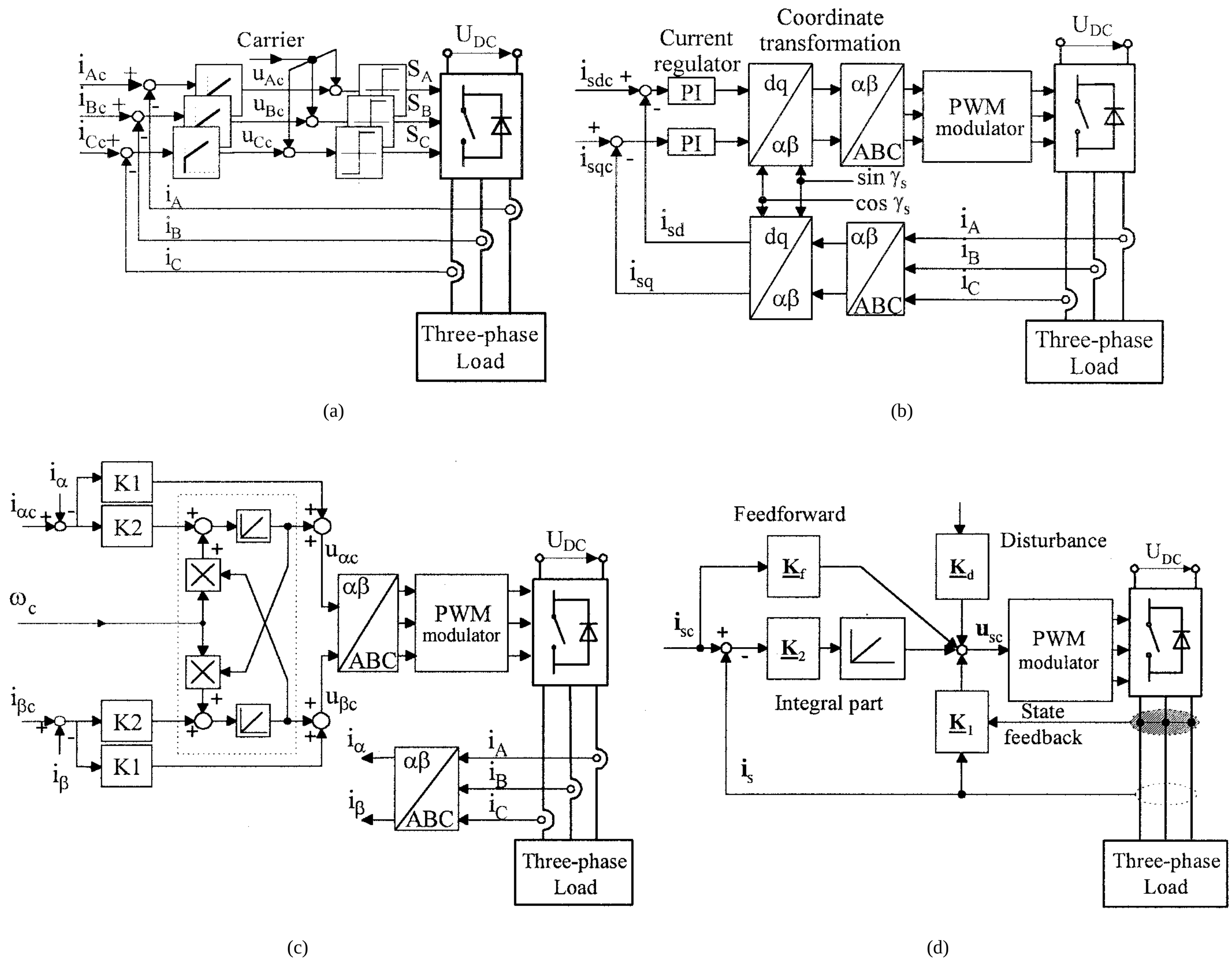


Fig. 4. Linear current controllers. (a) Stationary PI. (b) Synchronous PI working in rotating coordinates with DC components. (c) synchronous PI working in stationary coordinates with AC components. (d) State feedback controller.

tages of open-loop modulators (sinusoidal PWM, space-vector modulator, and optimal PWM) which are constant switching frequency, well-defined harmonic spectrum, optimum switch pattern, and dc-link utilization. Also, full independent design of the overall control structure, as well as open-loop testing of the inverter and load, can be easily performed. In the linear group, the following controllers are described: PI stationary and synchronous, state feedback, and predictive with constant switching frequency.

A. Stationary Controller PI

The *stationary controller*, also called *the ramp comparison current controller*, uses three PI error compensators to produce the voltage commands u_{Ac}, u_{Bc}, u_{Cc} for a three-phase sinusoidal PWM [Fig. 4(a)] [5]. In keeping with the principle of sinusoidal PWM, comparison with the triangular carrier signal generates control signals S_A, S_B, S_C for the inverter switches. Although this controller is directly derived from the original triangular suboscillation PWM [19], the behavior is quite different, because the output current ripple is fed back and influences the switching times. The integral part of the PI compensator minimizes errors at low frequency, while proportional gain and zero placement are related to the amount of ripple. The maximum slope of the command voltage u_{Ac} (u_{Bc}, u_{Cc}) should never exceed the triangle slope. Additional problems may arise from multiple crossing of triangular boundaries. As a consequence, the controller performance is satisfactory only if the significant harmonics of current commands and the load EMF are limited at a frequency well below the carrier (less than 1/9 [4]). The main disadvantage of this technique is an inherent tracking (amplitude and phase) error. To achieve compensation, use of additional phase-locked loop (PLL) circuits [24] or feedforward correction [29], [38] is also made.

B. Synchronous Vector Controller (PI)

In many industrial applications, an ideally impressed current is required, because even small phase or amplitude errors cause incorrect system operation (e.g., vector-controlled ac motors). In such cases, the control schemes based on the space-vector approach are applied. Fig. 4(b) illustrates the *synchronous controller*, which uses two PI compensators of current vector components defined in *rotating synchronous coordinates* d–q [5], [12], [14], [31], [32], [35]. Thanks to the coordinate transformations, i_{sd} and i_{sq} are dc components, and PI compensators reduce the errors of the fundamental component to zero.

Based on work in [34] (where it has been demonstrated that is possible to perform current vector control in an arbitrary coordinates), a *synchronous controller* working in the *stationary coordinates* α-β with ac components has been presented [33]. As shown in Fig. 4(c) by the dashed line, the inner loop of the control system (consisting of two integrators and multipliers) is a variable-frequency generator, which always produces reference voltages $u_{\alpha c}, u_{\beta c}$ for the PWM modulator, even when, in the steady state, the current error signals are zero.

In general, thanks to the use of PWM modulators, the linear controllers make a well-defined harmonic spectrum available, but their dynamic properties are inferior to those of bang-bang controllers.

C. State Feedback Controller

The conventional PI compensators in the current error compensation part can be replaced by a state feedback controller working in stationary [29] or synchronous rotating coordinates [13], [25], [27], [28], [30]. The controller of Fig. 4(d) works in *synchronous rotating coordinates* d–q and is synthesized on the basis of linear multivariable state feedback theory. A feedback gain matrix $\underline{K} = [\underline{K}_1, \underline{K}_2]$ is derived by utilizing the pole assignment technique to guarantee sufficient damping. While with integral part $(\underline{K}_2)$ the static error can be reduced to zero, the transient error may be unacceptably large. Therefore, feedforward signals for the reference $(\underline{K}_f)$ and disturbance $(\underline{K}_d)$ inputs are added to the feedback control law.

Because the control algorithm guarantees the dynamically correct compensation for the EMF voltage, the performance of the state feedback controller is superior to conventional PI controllers [27], [28].

D. Predictive and Deadbeat Controllers

This technique predicts at the beginning of each sampling (modulation) period the current error vector on the basis of the actual error and of the ac-side (load) parameters R, L, E. The voltage vector to be generated by PWM during the next modulation period is thus determinated, so as to minimize the forecast error [60], [102], [105], [107]–[109].

Hybrid CC combining predictive and hysteresis techniques have also been proposed [99].

1) Constant Switching Frequency Predictive Algorithm: In this case, the predictive algorithm calculates the voltage vector commands $\boldsymbol{u}_{sc}(T)$ once every sample period T. This will force the current vector according to its command $\boldsymbol{i}_{sc}$ [Fig. 5(a)]. The inverter voltage $\boldsymbol{u}_s(T)$ and EMF voltage $\boldsymbol{e}(T)$ of the load is assumed to be constant over the sample period T. The calculated voltage vector $\boldsymbol{u}_{sc}(T)$ is then implemented in the PWM modulator algorithm, e.g., space vector [60], [86], [100], [102] or sinusoidal modulator [107], [108]. Note that, while the current ripple is variable, the inverter switching frequency is fixed $(1/T)$. The disadvantage of this algorithm is that it does not guarantee the inverter peak current limit.

2) Deadbeat Controllers: When the choice of the voltage vector is made in order to null the error at the end of the sample period, the predictive controller is often called a *deadbeat controller* [85], [94], [95], [97]. Among the additional information given to the controller, nonavailable state variables (e.g., flux and speed) can be included. Their determination can require the use of observers or other control blocks, which often may be shared with the control of the entire scheme, as in the case of ac drives [83], [97].

IV. NONLINEAR CONTROLLERS

The nonlinear CC group includes hysteresis, DM, and on-line optimized controllers. To avoid confusion, current controllers for the RDCL topology are presented separately. Also, neural networks (NN's) and fuzzy logic controllers (FLC's) belong to the class of nonlinear CC.

A. Hard-Switched Converters

1) Hysteresis Current Controllers: Hysteresis control schemes are based on a nonlinear feedback loop with two-level hysteresis comparators [Fig. 6(a)] [61]. The switching signals S_A, S_B, S_C are produced directly when the error exceeds an assigned tolerance band h [Fig. 6(b)].

a) Variable switching frequency controllers: Among the main advantages of hysteresis CC are simplicity, outstanding robustness, lack of tracking errors, independence of load parameter changes, and extremely good dynamics limited only by switching speed and load time constant. However, this class of schemes, also known as freerunning hysteresis controllers [16], has the following disadvantages.

1) The converter switching frequency depends largely on the load parameters and varies with the ac voltage.
2) The operation is somewhat rough, due to the inherent randomness caused by the limit cycle; therefore, protection of the converter is difficult [56], [57].

It is characteristic of the hysteresis CC that the instantaneous current is kept exact in a tolerance band, except for systems without neutral leaders where the instantaneous error can reach double the value of the hysteresis band [3], [54] (Fig. 7). This is due to the interaction in the system with three independent controllers. The comparator state change in one phase influences the voltage applied to the load in two other phases (coupling). However, if all three current errors are considered as space vectors [60], the interaction effect can be compensated, and many variants of controllers known as space-vector based can be created [41], [48], [50], [58], [63], [68]. Moreover, if three-level comparators with a lookup table are used, a considerable decrease in the inverter switching frequency can be achieved [37], [48], [50], [58], [63]. This is possible thanks to appropriate selection of zero-voltage vectors [48] [Fig. 6(c)].

In the synchronous rotating d–q coordinates, the error field is rectangular, and the controller offers the opportunity of independent harmonic selection by choosing different hysteresis values for the d and q components [49], [62]. This can be used

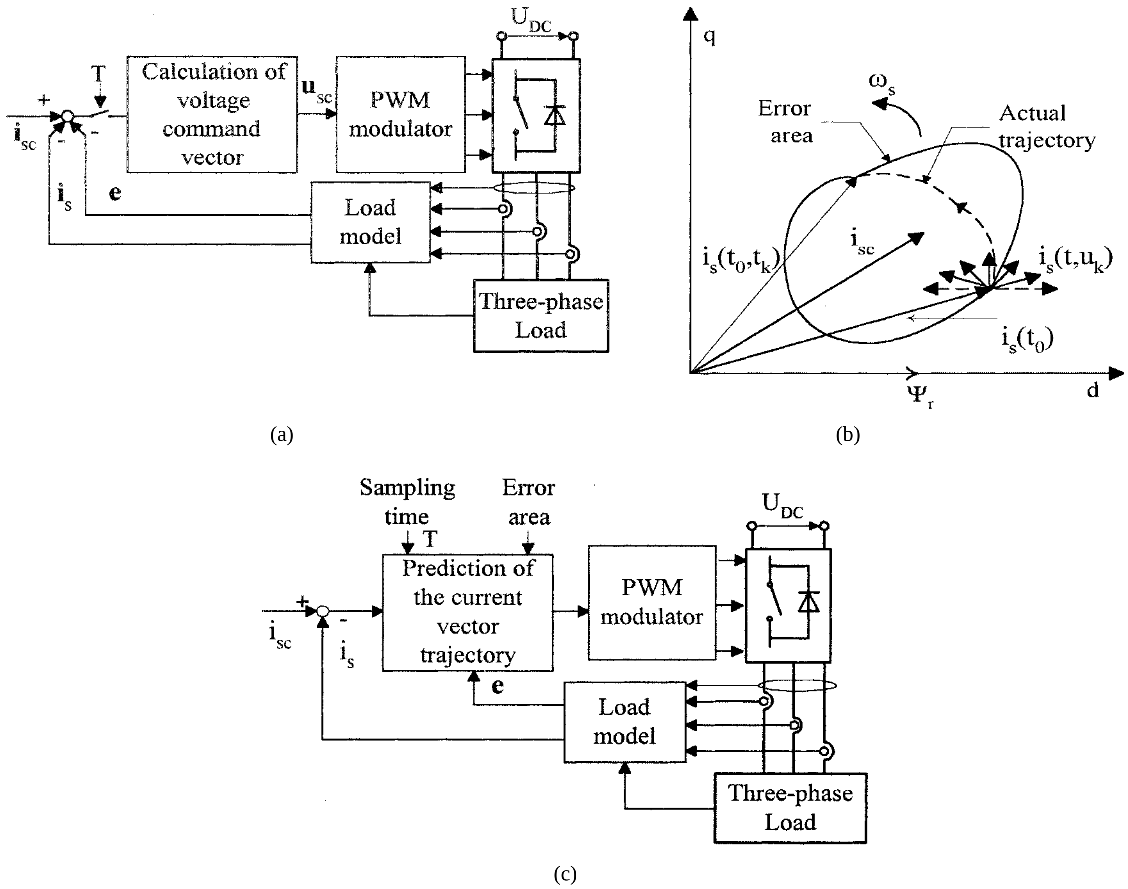


Fig. 5. Predictive current controllers. (a) Linear constant switching frequency controller. (b) Example of error area. (c) Minimum switching frequency controller.

for torque-ripple minimization in vector-controlled ac motor drives (the hysteresis band for the torque current component is set narrower than that for the flux current component) [49], [96].

Recent methods enable limit cycle suppression by introducing a suitable offset signal to either current references or the hysteresis band [45], [65], [67].

b) Constant switching frequency controllers: A number of proposals have been put forward to overcome variable switching frequency. The tolerance band amplitude can be varied, according to the ac-side voltage [39], [43], [47], [53]–[55], [57], [59], [69], [103], or by means of a PLL control (Fig. 8).

An approach which eliminates the interference, and its consequences, is that of decoupling error signals by subtracting an interference signal derived from the mean inverter voltage u_N (Fig. 8) [54]. Similar results are obtained in the case of "discontinuous switching" operation, where decoupling is more easily obtained without estimating load impedance [55]. Once decoupled, regular operation is obtained, and phase commutations may (but need not) be easily synchronized to a clock.

Although the constant switching frequency scheme is more complex and the main advantage of the basic hysteresis control—namely, the simplicity—is lost, these solutions guarantee very fast response together with limited tracking error. Thus, constant frequency hysteresis controls are well suited for high-performance high-speed applications.

2) Controllers with On-Line Optimization: This class of controllers performs a real-time optimization algorithm and requires complex on-line calculations, which usually can be implemented only on microprocessors.

a) Minimum switching frequency predictive algorithm: The concept of this algorithm [92] is based on space-vector analysis of hysteresis controllers. The boundary delimiting the current error area in the case of independent controllers with equal tolerance band $+h$ in each of three phases makes a regular symmetrical hexagon [Fig. 6(b)]. Suppose only one hysteresis controller is used—the one acting on the current error vector. In such a case, the boundary of the error area (also called the switching or error curve) might have any form [Fig. 5(b)]. The location of the error curve is determined by the current command vector $\boldsymbol{i}_{sc}$. When the current vector $\boldsymbol{i}_s$ reaches a point on the error curve, seven different trajectories of the current are predicted, one for each of seven possible (six active and zero) inverter output voltage vectors. Finally, based on the optimization procedure, the voltage vector which minimizes the mean inverter switching frequency is selected [Fig. 5(c)]. For fast transient states, the strategy which minimizes the response time is applied.

b) Control with field orientation: The minimum frequency predictive CC can be implemented in any rotating or stationary coordinates. Like the three-level hysteresis controller working in d–q field-oriented coordinates [49], a further switching frequency reduction can be achieved by

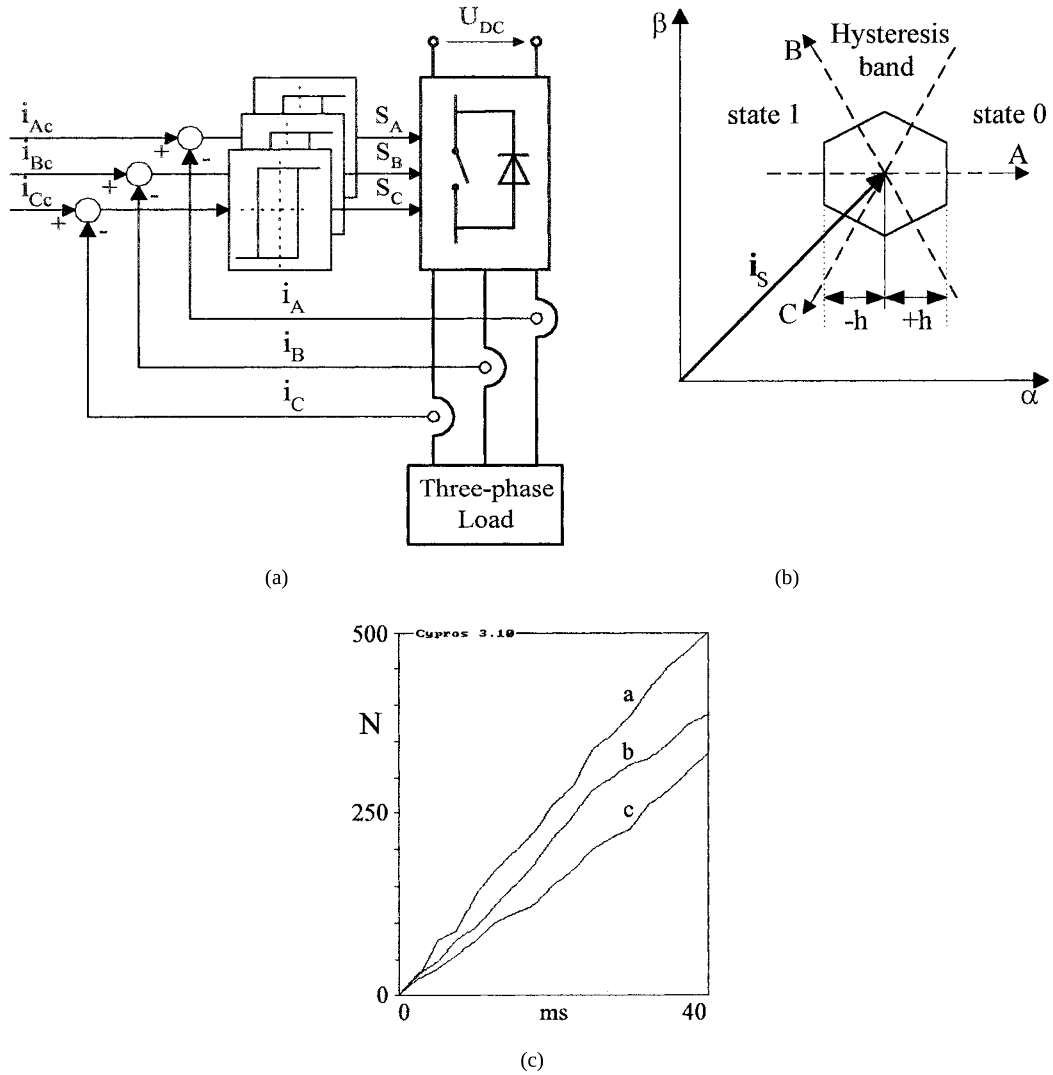


Fig. 6. Two-level hysteresis controller. (a) Block scheme. (b) Switching trajectory. (c) Number of inverter switchings N for *a*: three two-level hysteresis comparators, *b*: three-level comparators and lookup table working in the stationary, and *c*: rotating coordinates.

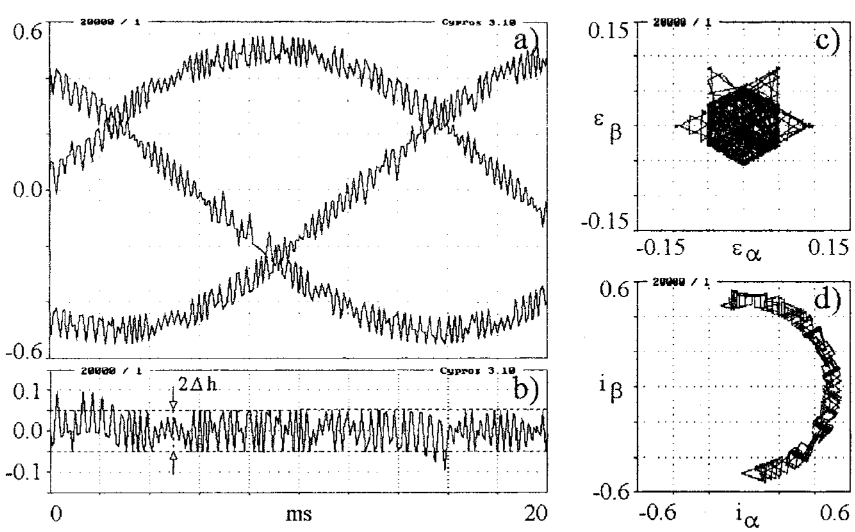


Fig. 7. Hysteresis controller ($h = 0.05$). (a) Output currents. (b) Phase current error. (c) Vector current area. (d) Output vector current loci.

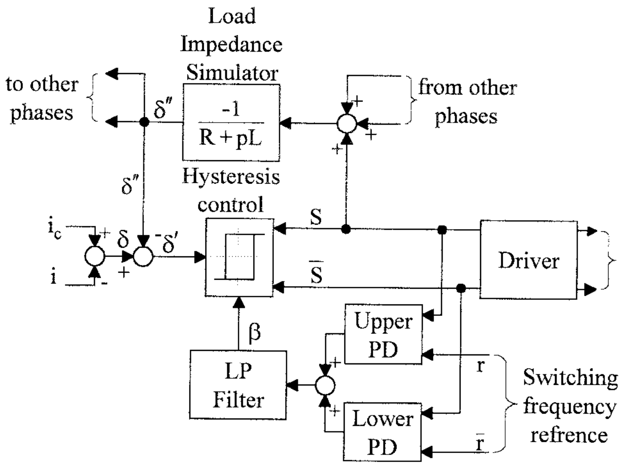


Fig. 8. Decoupled, constant average switching frequency hysteresis controller [54].

the selection of a rectangular error curve with higher length along rotor flux direction [96].

In practice, the time needed for the prediction and optimization procedures limits the achieved switching frequency. Therefore, in more recently developed algorithms, a reduced set of voltage vectors consisting of the two active vectors adjacent to the EMF vector and the zero voltage vector are considered for optimization without loss of quality [8].

c) Trajectory tracking control: This approach, proposed in [89] and [90], combines an *off-line* optimized PWM pattern for steady-state operation with an *on-line* optimization to compensate for the dynamic tracking errors of converter currents. Such a strategy achieves very good stationary and dynamic behavior even for low switching frequencies.

B. Soft-Switched RDCL Converters

In soft-switched RDCL three-phase converters with zero-voltage switching (ZVS), the commutation process is restricted to the discrete time instants T when the dc-link voltage pulses are zero [Fig. 2(b)]. Therefore, special techniques called *DM* or *pulse density modulation (PDM)* are used [70]–[82].

1) DM: The basic scheme, the *DM current controller (DM-CC)* [74], [82], is shown in Fig. 9(a). It looks quite similar to that of a hysteresis CC [Fig. 6(a)], but the operating principle is quite different. In fact, only the error sign is detected by the comparators, the outputs of which are sampled at a fixed rate, so that the inverter status is kept constant during each sampling interval. Thus, no PWM is performed; only basic voltage vectors can be generated by the converter for a fixed time. This mode of operation gives a discretization of the inverter output voltage, unlike the continuous variation of output voltages which is a particular feature of PWM.

One effect of the discretization is that, when synthesizing periodic waveforms, a nonnegligible amount of subharmonics is generated [74], [76], [77]. Thus, to obtain comparable results, a DM should switch at a frequency about seven times higher than a PWM modulator [76]. However, DM is very simple and insensitive to the load parameters. When applied to three-phase inverters with an insulated-neutral load, the mutual phase interference and the increased degree of freedom in the choice of voltage vector must be taken into account. Therefore, instead of performing independent DM in each phase control, output vectors are chosen depending not only on the error vector, but also on the previous status, so that the zero vector states become possible [73].

Due to the sample-and-hold (S&H) block applied after the ideal comparator, the switching frequency is limited to the sampling frequency f_s. The amplitude of the current harmonics is not constant, but is determined by the load parameters, dc-link voltage, ac-side voltage, and sampling frequency. If the sampling signal in the three-phase system is shifted 120° in each S&H block [Fig. 9(b)], only one of the inverter legs will change its state during the sample period $1/f_s$. This guarantees only adjacent and zero voltage vector selection and, consequently, a better quality of current waveform [lower rms, J (for definitions, see Table I)] at this same sampling frequency f_s [Fig. 9(c)] [71].

It is noted that the DM-CC can also be applied in the space-vector-based controllers working in either stationary or rotating coordinates [75], [79], [81].

The main advantages of DM-CC are extremely simple and tuning-free hardware implementation and good dynamics.

2) Optimal Discrete Modulation Algorithm: For the RDCL converters, an optimal algorithm selects the voltage vector which minimizes the rms current error for each resonant pulse [80], [93], [106]. As shown in [106], this is equivalent to selecting the nearest available voltage vector commands $u_{sc}(T)$. So, instead of the PWM algorithm [Fig. 5(a)] only the voltage vector selector is required [Fig. 10(a)]. However, errors and subharmonics typical of the discrete modulation are obtained. The typical waveforms for discrete DM and optimal (minimum rms error) CC are shown in Fig. 11(a) and (b), respectively.

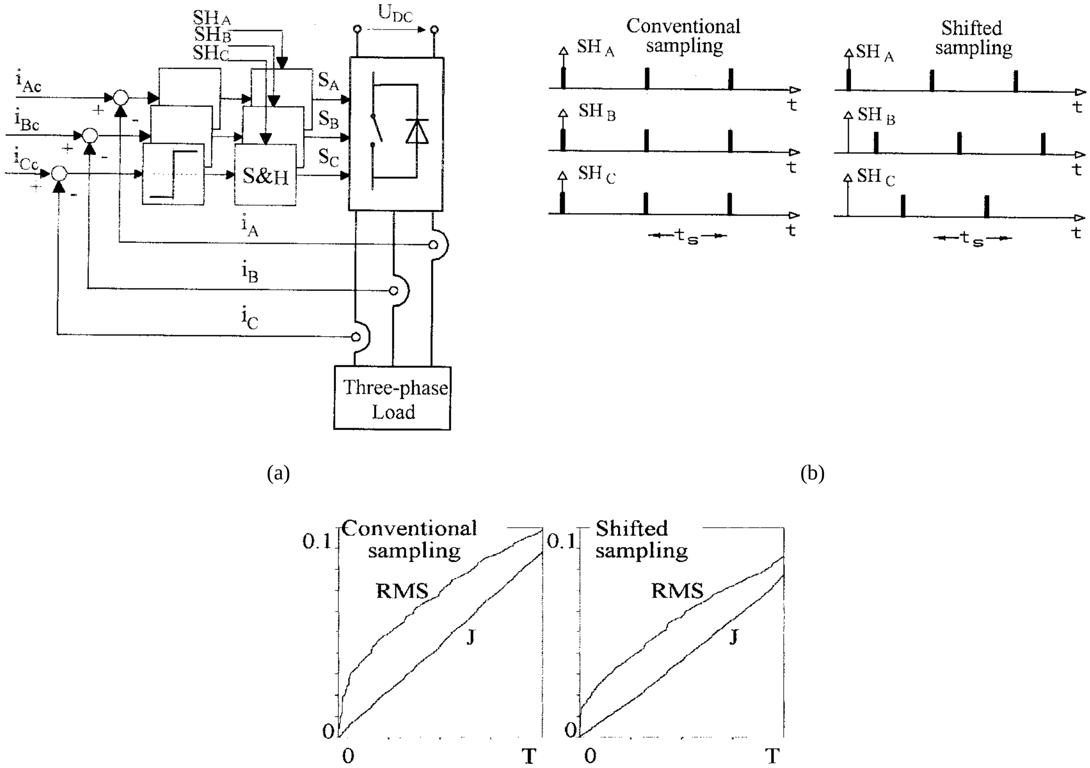
C. NN's and FL-Based Controllers

Recently, new emerging technologies such as NN's and FL methods have been applied to PWM current control.

1) NN's Controllers: The main advantages of NN are parallel processing, learning ability, robustness, and generalization. They can be effectively used for CC [110]–[112], [115]–[117], [120].

A simple example, which allows for the elimination of the on-line calculations needed to implement the optimal discrete CC of Fig. 10(a), is shown in Fig. 10(b) [117]. The three layers of the feedforward NN with sigmoidal nonlinearity—before using as a controller—were trained using a back propagation algorithm with randomly selected data from the output pattern of the optimal controller of Fig. 10(a). After training, the performance of the three-layer (architecture: 5-10-10-3) NN-based controllers differs only slightly from that of the optimal regulator [Fig. 11(c)]. Thus, the NN-based controller can be used to regulate PWM converter output current without a need for the on-line calculation required for an optimal controller.

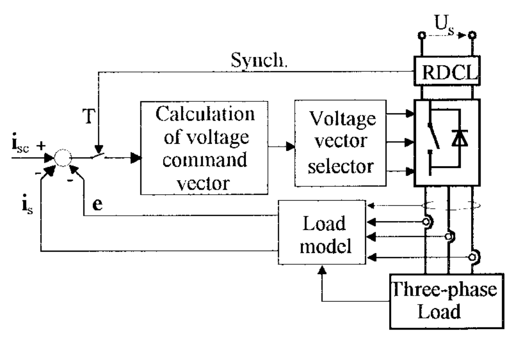
With this approach, however, no further training of the NN is possible during controller operation. Therefore, the



(a) (b)

(c)

Fig. 9. DM current controller. (a) Basic scheme. (b) Sampling techniques. (c) Quality factors.



(a)

(b)

Fig. 10. (a) Optimal (mode) discrete modulation controller for RDCL converter. (b) NN discrete modulation controller for RDCL converter.

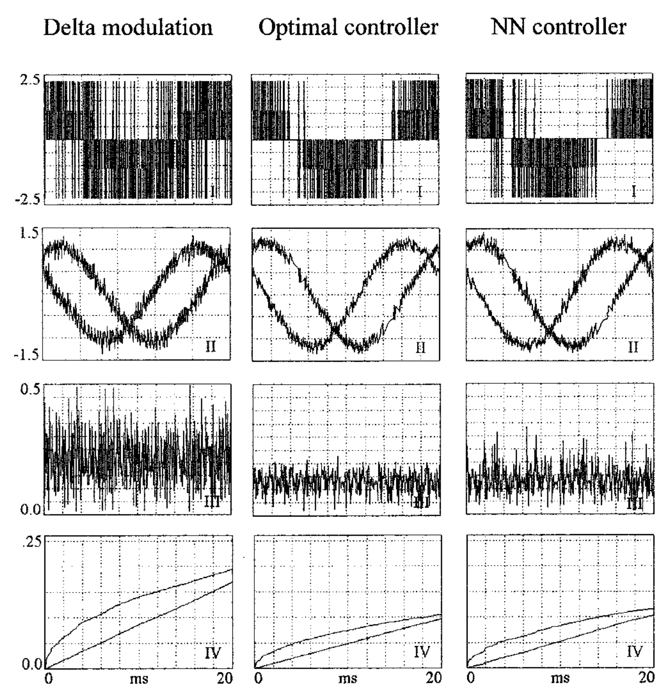


Fig. 11. Current control in RDCL based on discrete modulation. From the top: I—line-to-line voltage u_β; II—current vector components i_α, i_β; III—current error $(\epsilon_\alpha^2 + \epsilon_\beta^2)^{1/2}$; IV—rms and J of current error $\epsilon(t)$.

performance of such an *off-line* trained NN controller depends upon the amount and quality of training data used and is also sensitive to parameter variations. For systems where parameters variations have to be compensated, an *on-line* trained NN controller can be applied [111], [112], [116]. In [112], an NN induction motor CC with parameter identification

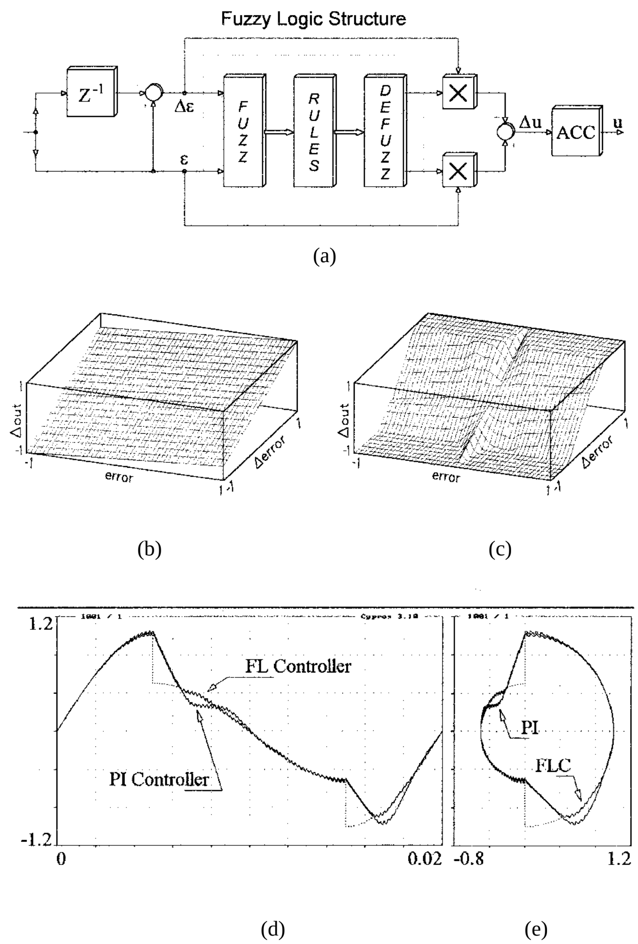


Fig. 12. (a) Block scheme of FL controller, (b) control surface of conventional PI controller, (c) control surface of FL controller; comparison of current-tracking performance with PI and FL controller: (d) current waveform, and (e) current vector loci.

was proposed. To achieve very fast on-line training (8 μs for one training cycle) a new algorithm called random weight change (RWC) is applied. This algorithm allows us to identify and control the motor currents within a few milliseconds.

2) FL-CC's: In basic applications, the FLC is used as a substitute for the conventional PI compensator [114], [118]. The block scheme of the FL-CC is similar to the system of Fig. 4(a), where, instead of PI, FL self-tuned PI controllers are used. The basic block scheme of an FL-tuned discrete PI controller, including the fuzzy inference mechanism, is shown in Fig. 12(a). The current error ε and its derivative $\Delta\varepsilon$ are the FL controller input crisp values. The reference voltage for the PWM modulator are the FL-CC crisp output commands u. When an FL controller is used as a current controller, the tracking error and transient overshoots of PWM current control can be considerably reduced [Fig. 12(d) and (e)]. This is because—in contrast to the conventional PI compensator—the control surface of the FL controller can be shaped to define appropriate sensitivity for each operating point [Fig. 12(b) and (c)]. The FL-tuned PI controller can easily be implemented as an off-line precalculated three-dimensional lookup table consisting of the control surface [114]. However, the properties of the FL controller are very sensitive to any change of fuzzy sets shapes and overlapping. Therefore, the design procedure and resulting performance depend strongly on the knowledge and expertise of the designer.

V. Conclusions

CC techniques for VS converters can be divided into two groups: 1) *linear*, i.e, stationary, synchronous, and predictive deadbeat controllers and 2) *nonlinear*, i.e., hysteresis, DM, and on-line optimized controllers. The basic principles and the latest developments of these techniques have been systematically described in this paper. The advantages and limitations have been briefly examined, and the application field where each technique is particularly suited has been indicated.

Recently, the research trend favors fully digital control. Thus, the methods which allow digital implementation are preferred, even with some sacrifice in accuracy and dynamic performance. In particular, for low-performance applications with large diffusion (e.g., pumps, blowers and fans, and retrofit applications), digitally implemented PI regulators are adequate. Use of linear predictive and on-line optimized CC is growing fast in medium- and high-performance systems, especially for traction and high power units. Hysteresis CC, in their improved versions, are well suited to fast, accurate conversion systems (e.g., power filters and UPS's).

It is possible that NN's and FL-based CC techniques can offer a new interesting perspective for future research. At present, however, they represent only an alternative solution to existing CC methods, and their specific applications areas cannot be clearly defined.

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