Fast Current Trajectory Tracking Control Based on Synchronous Optimal Pulsewidth Modulation

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Abstract — State-of-the-art pulsewidth modulation techniques fail to give satisfactory results when high-bandwidth torque control is required at the low switching frequency of modern high-power ac machine drives. A novel current control method refers to precalculated optimal synchronous pulse patterns. These are used to generate specific current reference trajectories to be adapted on line to the actual dynamic torque command. A fast tracking controller, operated in parallel to a conventional PI controller, minimizes the trajectory tracking error. The optimized current trajectories ensure fast response and minimum harmonic distortion.

1. Introduction

While the technology of ac drives is getting more and more standardized with respect to power conversion techniques, drive control methods continue to develop. One of the important issues is pulsewidth modulation (PWM). The present desire to extend the application of voltage-source inverter drives to power levels in the megawatt range, using GTOs as semiconductor switches, is impeded by the switching frequency constraint of these components. Owing to the higher switching losses of GTOs, the switching frequency is effectively limited to only a few 100 Hz. These limitations make it worthwhile to employ more elaborate techniques for pulsewidth modulation in order to reduce the harmonic losses and the torque pulsations of the machine.

2. Pulsewidth Modulation and Fast Current Control

The preferred circuit topology for three-phase dc-to-ac power conversion at very high power levels is the voltage source inverter, e.g., for railway traction applications [1]. In this configuration, each of the three load terminals is switched either to the positive or the negative rail of a dc link circuit of constant voltage. The duty-cycles of switching are varied by pulsewidth modulation such that the average values of the three-phase voltages mirror their respective reference signals. The permitted frequency of switching is very low, determined by the properties of the power semiconductor switches.

For the application in high dynamic performance drives two important aspects deserve particular attention:

- The current waveforms are irregular, and their instantaneous deviations from the fundamental wave tend to be high at low switching frequency.

Filters to extract the fundamental content from the measured waveform introduce a phase lag to the current control loop. This deteriorates the dynamic performance [2]. It is therefore preferred to merge the two tasks of current control and pulsewidth modulation into a single functional structure with a view to control the fundamental current in good approximation without necessarily having to measure this quantity directly. Such current control methods have been successfully applied in low power servo drives [3].

2.1 Hysteresis control

A very simple and effective method limits the harmonic current error magnitude in each phase using hysteresis comparators, Fig. 1(a). Any change in the current reference $i_s^*$ triggers a commutation in the appropriate inverter half-bridge as soon as the permitted current error magnitude is exceeded. The error is subsequently corrected at the fastest possible rate, Fig. 1(b). While the current reference $i_s^*$ is a fundamental quantity, the scheme controls actually only the envelopes of the peak harmonic currents. As a result, the fundamental current may differ from its reference, which makes subharmonic components appear in its harmonic spectrum. Also the content of higher harmonics is fairly large owing to the absence of a mechanism which enforces predetermined time sequences of the inverter switching states [4].

The switching frequency in the steady-state depends on the fundamental voltage content $u_1$ of the switched waveform. The normalized fundamental voltage is defined as the modulation index $m = u_1/u_{1\text{ six-step}}$, where $u_{1\text{ six-step}}$ is the fundamental voltage at six-step operation [5]. As the modu-
The performance becomes worse as the switching frequency reduces. The modulation index changes; the switching frequency varies considerably. Hence the switching capability of the power devices is mostly not utilized for the reduction of switching harmonics. The same applies for modified hysteretic control schemes which have the current errors transformed to a synchronous reference frame before they act on the hysteretic comparators [6]. Also flux trajectory control schemes [7] behave the same way since the harmonics of the stator flux vector are proportional to those of the stator current vector, but for the effects of magnetic saturation.

The deficiencies of hysteretic controllers are marginal if the width of the hysteresis band is sufficiently small; they are prohibitive for slow-switching high power converters.

### 2.2 Suboscillation control

An important improvement is made by generating the inverter switching states in a predetermined time sequence. Fig. 2 shows a signal flow structure in which the switching instants are determined by a ramp comparison, or suboscillation type of modulator. The modulator forms a subordinate system to the current control loop.

The modulator reference is not sinusoidal; it contains the harmonics of the current feedback signal. The current controller gain must be adjusted such that the slope of the reference signal never exceeds the triangular carrier slope. This sets the controller gain proportional to the switching frequency.

The modulator establishes proportionality between the fundamental components of the stator voltages and their reference signals. Accordingly, the current regulator in Fig. 2 outputs the normalized signal \( u_c \). Working further backwards to the input of the current regulator shows that a steady-state current error of fundamental frequency must exist. This is not a major problem at high switching frequency, where the gain of the current controller gain is high. The performance becomes worse as the switching frequency reduces.

The modulation index is limited to a maximum value \( m_{\text{max,1}} = 0.785 \). This is an inherent limitation of the suboscillation modulator [5]. Pulse dropping occurs when the modulator is driven beyond its voltage limit. The switching frequency is then reduced and the current distortion increases.

The dynamic performance of a suboscillation current controller is good. Steady-state current harmonics are moderate, calling for switching frequencies in the lower kilohertz range, or above. This mostly excludes the application for large power inverters. The reduced modulation range is another drawback. It entails a poor utilization of the power converter which becomes more decisive at higher power level.

### 2.3 Current controlled space vector modulation

A space vector modulator controls the time average of the three switching state vectors that form a subcycle to equal the sampled value \( u^*(t_s) \) of the reference vector. Sampling the signals \( t_s \) marks the beginning of the subcycle. Fig. 3 shows the corresponding signal flow graph. The reference voltage \( u_i^* \) is derived from the superimposed current controller.

A fundamental difference exists in comparison to the previous scheme in Fig. 2. A reference value sampled at a particular instant, including the harmonic content herein, controls the on-duration of an entire modulation subcycle. This means that an instantaneous harmonic deviation would influence the fundamental inverter output voltage. This creates an additional source of harmonic distortion which must be prevented. The problem is solved by exploiting a specific property of the space vector modulator [8]. The stator current trajectory intersects the locus of the fundamental current vector invariably when one half of the zero vector on-time has elapsed. The fundamental current can be sampled at this time instant, and the sampled value \( i_{s1}(t_s) \) is used as the current feedback signal. This is indicated in the lower portion of Fig. 3.

The current controlled space vector modulator operates in an extended modulation range up to \( m_{\text{max,2}} = 0.907 \), which yields higher output power. Also the steady-state distortions are lower than those of the previous schemes [5]. A deterioration of performance can be observed when operating at extreme low switching frequency:

- The detection of the instantaneous fundamental current by sampling is only correct when a lossless machine is assumed. This assumption fails to be true when the subcycle duration \( 1/2f_s \) comes closer to the transient machine time constant [9]. A distorted feedback signal makes the controlled current irregular.
- A large current sampling time interval \( 1/2f_s \) impairs the dynamic performance.
2.4 Predictive current control

Predictive current control is an optimal PWM technique [5]. Switching events occur whenever the current vector \( \mathbf{i}_s \) normally located within a circular boundary around the current reference vector \( \mathbf{i}_s^* \) touches the boundary line, Fig 4. The future trajectories of the current vector for each possible switching state are then computed, and predictions are made of the respective time intervals required to reach the error boundary again. These events depend also on the location of the error boundary, which is considered moving in the complex plane as commanded by the predicted current reference. The movement is indicated by the dashed circle in Fig. 4(b). The predictions of on-times are based on a simplified mathematical model of the machine. The switching state vector that produces the maximum on-time is finally selected. This effectively minimizes the switching frequency.

High timing accuracy is obtained by the double prediction method. The optimal switching state vector is determined well before the switching instant, based on a prediction of the system state at the next error boundary. The dynamic performance of this method is excellent, even at low switching frequency. The harmonic distortion is low, although subharmonics do exist. The disadvantage is that the switching frequency varies with the modulation depth and hence the switching capability of the inverter is not properly utilized.

2.4 Synchronous optimal pulsewidth modulation

The most effective approach for minimum harmonic distortion PWM considers all switching angles in a fundamental period as the variables for a closed optimization [5]. The computations are carried out off-line, assuming steady-state operation of the machine, with the modulation index \( m \) and the pulse number \( N = f_p/f_s \) as parameters. The optimization yields improvements only at lower pulse number \( N < 15 \ldots 20 \). No subharmonics are generated as the optimal pulse patterns are in synchronism with the fundamental voltage. The optimized patterns are stored in a memory of the drive system and used for on-line inverter control.

Synchronous optimal pulsewidth modulation is inherently restricted to steady-state operation since it is hardly possible to redefine dynamic conditions. An optimal switching pattern generates a well-defined steady-state current trajectory \( \mathbf{i}_s(t) \) of minimum harmonic distortion, as the one shown in the left half of Fig. 5(a). An assumed change of the operating point at \( t = t_1 \) commands a different pulse pattern, to which a different steady-state current trajectory is associated. Since the current must be continuous, the actual trajectory resulting at \( t > t_1 \) exhibits an offset in space, Fig. 5(b). This is obvious since two optimized steady-state trajectories rarely have the same instantaneous current values at any given point of time. The offset in space is called the dynamic modulation error \( \delta(t) \) [10].

The dynamic modulation error tends to be large at low switching frequency. It is therefore almost impossible to use a synchronous optimal pulsewidth modulator as part of a fast current control system. During transient operation, the reference vector to the modulator changes its magnitude and phase angle very rapidly. Sections of different optimal pulse patterns are then pieced together to form a real-time pulse sequence in which the preoptimized balance of voltage-time area is lost. The dynamic modulation error accumulates, and overcurrents occur which cause the inverter to trip.

2.5 Trajectory Tracking Control

The combination of fast current control and synchronous optimal pulsewidth modulation requires a more refined approach in which not only the fundamental currents, but also the harmonic currents are controlled.

Harmonic current control is done by trajectory tracking. A reference current trajectory is obtained on-line from the respective pulse pattern in actual use. The reference trajectory reproduces the steady-state current which had been the basis of off-line pattern optimization. The difference between the actual current vector and the corresponding value on the reference trajectory is the dynamic modulation error. This error is reduced to zero by an adequate modification of the selected pulse pattern.

Trajectory tracking control has been demonstrated to be a viable solution for high-bandwidth torque control in slow-switching high-power ac drives [10]. Yet the method could need further improvement regarding the following:

- the dynamic modulation error results as the consequence of deviations from steady-state operation. It is measured and subsequently compensated, which takes two interrupt cycles of the digital signal processing system, totaling to a
3. Improved Scheme

3.1 Steady-state performance

Fig. 6 shows the harmonic current distortion at synchronous optimal PWM in comparison to space vector modulation and the suboscillation method. The zero db level is the harmonic distortion in the six-step mode, where \( N = 1 \). The advantage of synchronous optimal PWM in terms of harmonic distortion is particularly prominent at higher modulation index.

Note that the space vector modulation curve in Fig. 6 terminates at \( m_{\text{max},2} = 0.907 \), while the optimal modulation comes very close to the absolute limit of \( m = 1 \), even when taking the minimum on-time of a GTO inverter into account. Thus an additional 10% more output power is available when synchronous optimal PWM is used.

3.2 Parallel-channel current control

The discussion refers now to the signal flow diagram Fig. 7. The current reference \( i_s^{(F)}(i) \) is received in field coordinates (superscript \( (F) \)), representing the flux and the field producing current components. The reference is compared with the instantaneous fundamental current vector \( i_s^{(F)}(F) \). The resulting current error \( e_i \) is a fundamental quantity, free from harmonics. The fundamental current error feeds two parallel current control channels.

![Fig. 7: Parallel-channel current control scheme](image)

Fig. 6: Synchronous optimal pulswidth modulation; harmonic distortion versus modulation index, maximum switching frequency \( f_{s_{\text{max}}} = \text{param.} \); upper curves: space vector modulation and suboscillation method at \( f_s = 300 \text{ Hz} \)

3.2.1 Control of the fundamental current

A PI controller for the fundamental current is provided in the upper channel in Fig. 7. It contributes only minor corrections to the voltage reference vector \( u^* \), which, in essence, is provided by the feedforward signal \( u^* \), \( u^* \) is obtained from a saturation dependent machine model, Fig. 8. The signal is transformed to stator coordinates using the field transformation angle \( \Phi(t) \). A lead rotation term \( \exp(j \omega_s \tau) \) is included to account for the delay caused by time discrete signal processing. \( \hat{\Phi}_{s_{ss}} \) is the steady-state stator frequency and \( \tau \) is the cycle time required for the current control loops.

The resulting complex signal serves as the input to the optimal pulswidth modulator. The magnitude \( u^* \) represents the modulation index \( m \); it is used to select the actual switching pattern \( P(m, N) \). The phase angle \( \phi_s \) and the steady-state stator frequency \( \omega_{s_{ss}} \) are used in the modulator to form the time basis for the reconstruction of the switching instants from the selected pattern of switching angles.

The steady-state stator frequency \( \omega_{s_{ss}} \) is obtained from the machine model, Fig. 8, as a signal that approximates the stator frequency as if the machine was not in a transient state. \( \omega_{s_{ss}} \) is simply obtained adding a delayed rotor frequency signal to the mechanical speed. This impedes that a torque transient, which instantaneously changes \( \omega_r \) in the machine model, would have an immediate effect on the timing of the optimal switching sequence. Of course, modelling the mechanical system by a first-order delay \( \tau_1 \) is not very accurate. A phase displacement is therefore likely to occur in the fundamental machine voltage. As \( \omega_{s_{ss}} \) changes only slowly, the trajectory controller can accurately counteract, using its regular algorithm. It reestablishes the original pulse pattern by compensating the phase displacement of the reference voltage \( u^* \).

The fundamental current error \( e_i \) is also fed to the lower channel in Fig. 7. The error results from changes of the commanded current or the load current; it exerts an immediate influence on the switching pattern. The error is routed through a trajectory controller to modi-
3.2.2 Control of the harmonic current

The instantaneous harmonic current is controlled separately from the fundamental current. This task is performed by the error estimator and the trajectory controller shown in the lower portion of Fig. 7. The deviation of the instantaneous harmonic current from its reference value is the dynamic modulation error \( \delta(t) \) (Section 2.4). In a strict sense, this error is caused by improper operation of the pulsewidth modulator. The elimination of the dynamic modulation error is therefore possible within the modulator itself. The machine parameters are not required for this purpose, although it is primarily a deviation of the machine currents that indicates the presence of a modulation error, Fig. 5.

In a more general sense, the nonideal operation of the inverter due to the dead-time phenomenon [3], and the effect of other disturbances like the fluctuations and the ripple content of the dc link voltage [11] can be also treated as modulation errors. Such disturbances are taken into account by using the actual stator voltages \( u_s \) at the machine terminals as feedback signals for error detection.

Fig. 9 shows the signal flow schematic for the estimation of the dynamic modulation error. The reference voltage vector \( u^* \) of the optimal pulsewidth modulator represents the fundamental machine voltage. It is subtracted from the measured stator voltage vector \( u_s \) so as to form the vector \( u_h \) of the instantaneous harmonic stator voltage. This voltage is passed through a harmonic machine model to yield an estimated harmonic machine current \( i_h \). The model parameter \( \hat{\tau}_e \) approximates the transient machine time constant; it will be shown that the exact value is not required.

The harmonic reference vector \( i_{h\,ss} \) moves on a particular steady-state trajectory which is associated to the selected optimal pulse pattern. Each pulse pattern has a different steady-state trajectory, and the transitions between them, when the pattern changes, are discontinuous. To arrive at the harmonic reference vector, a steady-state switching sequence \( u_{ss} \) is obtained from the selected switching pattern \( P(m, N) \) in a first step. The reference voltage \( u^* \) is then subtracted, yielding the steady-state voltage harmonics \( u_{h\,ss} \). A reference machine model serves to generate the steady-state harmonic current \( i_{h\,ss} \) using a well determined initial current value. The model receives a new initial setting \( i_{h\,ss}(t_c) \) at every time instant \( t_c \) at which another optimal switching pattern becomes active.

The steady-state harmonic current \( i_{h\,ss} \) is a synthesized waveform containing only ac components. Contrasting to that, the actual harmonic current \( i_h \) includes the dc offsets that occur at transient operation of the machine. The transient machine currents decay as commanded by the transient machine time constant \( \tau_e \). The same behavior is given to the estimated harmonic current \( \hat{i}_{h} \) in Fig. 9 by a feedback term \( \hat{i}_{h} \) in the harmonic model.

The dynamic modulation error \( \hat{\delta}(t) \) is now obtained as the difference between the harmonic reference trajectory \( i_{h\,ss} \) and the harmonic current \( i_h \). The error \( \hat{\delta}(t) \) is fed to a trajectory controller which, at nonzero error, modifies the selected pulse pattern using a dead-beat algorithm for fast dynamic response. The signal flow scheme is shown in Fig. 7.

The modification of the actual pulse pattern in use must depend on the transient machine time constant. Here, \( \hat{\tau}_e \) controls the conversion of the dynamic modulation error into a complex volt-second difference; in a reversed way, the modulation error itself is computed from a volt-second unbalance using the same time constant. An inaccuracy in the selected value of \( \hat{\tau}_e \) is therefore ineffective, and the control of the harmonic current is parameter independent. However, the intermediate signal \( \hat{\delta}(t) \) is marked as estimated as it does depend on the preset value of \( \hat{\tau}_e \).

3.2.3 Fundamental current extraction

Although the harmonic content of the current waveforms is minimized by synchronous optimal pulsewidth modulation, the distortion is still too high such that high-bandwidth fundamental current control is impeded. This is a consequence of the extreme low switching frequency. The resulting problem is the following: The current controller must react fast on the fundamental current error while the harmonics must be ignored. On the other hand, the controlling signal for the optimal pulsewidth modulator must be smooth and ripple-free in the absence of externally commanded transients. Only then can the precalculated optimal pulse sequences be reproduced and the harmonics minimized.
The problem is solved by extracting the instantaneous fundamental $i_{s1}$ component from the measured current waveforms. The estimated harmonic content $i_h$ of the stator current is already available as a signal which serves for computing the dynamic modulation error. It is generated in the signal flow structure Fig. 9 at the output of the harmonic machine model. The parameter of this model is arbitrarily set as it was found to have no influence on the modification of pulse patterns. However, the actual machine parameter depends on saturation; it was found to vary in a range of $1:2.5$ in the laboratory drive used for testing.

The estimated harmonic current $i_h$ is therefore scaled so as to equal the actual harmonic current $i_{h}$. The scaling factor $k_s$ is the ratio between the preset value of $\tau_\sigma^s$ in the harmonic machine model and the actual transient machine time constant $\tau_i$; hence $k_s = \tau_\sigma^s / \tau_i$. The value $k_s$ is determined by correlating the measured machine current $i_s$ and the modeled harmonic current $i_h$. The correlation is executed as a background program. It repetitively updates a table from which the conversion factor $k_s(i_{s1})$ can be instantaneously retrieved as a function of the fundamental machine current.

The signal flow scheme is shown in Fig. 10. The input signals $i_s$ from the current transducers and $i_h$ from the harmonic machine model are transformed to field coordinates. Sets of 64 consecutive samples are accumulated and stored in the data memories M1 and M2. The correlation is repetitively done on each completed set of data, separately in the d- and the q-axis. The locations of these data are shown in Fig. 11. They were recorded at different steady-state operating points at positive and negative torque, and at no-load. The correlation yields

- the magnitude ratio $k_s$ between the harmonic components $i_s$ and $i_h$, expressed by the slope of the correlation line,
- the fundamental current $i_{s1}$, given by the point of intersection of the correlation line with the abscissa.

To demonstrate the strong influence of saturation, steady-state current trajectories $i_s(t)$ are oscillographed in Fig. 12 using the same pulse pattern at different load conditions. The different sizes and the distorted shapes of the trajectory patterns indicate the varying saturation of the leakage inductance. This translates into the different slopes of the correlation lines in Fig. 11.

During a change of operating point, the resulting trajectory becomes less uniform, and the correlation coefficient $k_c$ falls below a threshold limit. Only valid correlation data are used to update the table $k_s = f(i_{s1})$ in Fig. 10. Random errors are suppressed by giving the update $k_s$ a low weight $\lambda$ as compared to an already existing value $k_s$.

The updated table is used to eliminate the harmonic components from the measured current vector. Fig. 13 shows that the table is addressed by the current reference vector $i_s^*$. This yields the actual state of saturation, permitting the computation of the instantaneous harmonic current $i_h$ from the estimated vector $i_h$. The instantaneous fundamental current $i_{s1}$ is then extracted from the measured waveform of $i_s$ by subtracting its predetermined harmonic content $i_h$.

### 3.2.4 Interaction between the parallel control channels

The principal task of the trajectory controller is to eliminate the dynamic modulation error $\delta(t)$. Having the property of a dead-beat controller, it is best suited to eliminate the fundamental current error as well. Hence the trajectory controller is given the additional task of providing the required system dynamics. For this purpose, it receives the error signal $\varepsilon_i$ which is scaled to the appropriate level using the saturation coefficient $k_s(i_{s1})$. Step changes of the current reference or of the load current are then accurately executed within a single computation cycle of the controller.

At step excitation, a current error $\varepsilon_i$ exists only temporarily for the duration of one computation cycle $T_c$. The error must be converted into a persistent displacement $\Delta i_q$ in space of the current vector. This is done by feeding the voltage signal $-\varepsilon_i i_q / k_s T_c$ to the error estimator in Fig. 7. The estimator output then comprises the dynamic modula-
tion error $\delta$ and the transient current component $\Delta i_d$.

With the trajectory controller performing in a dead-beat fashion to eliminate both types of dynamic current errors, and with the machine model providing a smooth reference signal for the optimal pulsewidth modulator, the activity of the PI current controller is restricted to the correction of minor errors resulting from parameter deviations.

### 3.2.5 Machine model

The field transformation angle $\varphi(t)$ is obtained from a rotor model of the machine, Fig. 8. The model also provides the principal control signal $u^*$ for the optimal pulsewidth modulator. The influence of varying saturation of the leakage inductance is compensated by parameter adaptation using the saturation dependent factor $k_\sigma$. Saturation of the main inductance $l_h$ need not be considered in the base speed range.

### 4. Experimental Verification

The implementation of the fast current trajectory control scheme using synchronous optimal pulsewidth modulation was laboratory tested on a 30-kW induction machine drive fed from a three-level inverter. The maximum switching frequency is 200 Hz and the leakage inductance $l_\sigma$ of the machine is 0.19 (unsaturated) and 0.075 (saturated).

One of the objectives is to reproduce the preoptimized pulse patterns in the steady-state. The proof is obtained by observing the current trajectory in field coordinates, where it must reproduce six times per fundamental period. A reference pattern Fig. 14(a) was recorded first with the modulator operated in open-loop, using a pure sinusoidal reference $u^*$. The pattern should exhibit symmetry with respect to its own $d$-axis; it is considerably distorted, however, which indicates substantial changes of leakage saturation although the current varies only in the limited range of its harmonic amplitudes. The observed jitter in the trajectory is attributed to spatial unbalances of the machine. Although these act as disturbances at closed-loop control, the trajectory controller shows little irritation, Fig. 14(b).

The process of acceleration from stand-still at maximum torque is shown in Fig. 15. It is initiated by a peak of the reference voltage $u^*$ to build up the active current. Following that, the signal $u^*$ has a curved tendency which is owed to operation at constant rotor flux. Trace (d) at the bottom is high at low speed indicating the use of space vector modulation. The algorithms for instantaneous fundamental current extraction and dynamic modulation error estimation are kept active with this type of modulation so as to ensure full dynamic capability. The control then switches to synchronous optimal modulation; a series of markers in the bottom trace indicate the changes between different types of pulse patterns.

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![Fig. 15: Acceleration from stand-still at maximum torque; (a) instantaneous fundamental current, (b) speed, (c) reference voltage $u^*$, (d) markers indicating pattern changes](image)

The oscillogram Fig. 16(a) shows a reversal of torque between negative and positive nominal values. The transition is executed within the shortest possible time: The new reference value is reached after 1 ms. The drive then accelerates at rated torque; the harmonic content in the current waveform is seen to change as speed picks up and the optimal PWM patterns get changed. The extracted funda-
mental current waveform is shown in Fig. 16(b). The residual steady-state ripple has three causes. First, the saturated leakage inductance is very low and the current harmonics are high. Second, the computation cycle of the DSP monics are high. Second, the very low and the currenthar-
torous steady-state ripple has shown in Fig. 16(b). The re-
mental current waveform is by a sequence of diode commutations [12].

While the leakage saturation changes with the harmonic am-
plitude, only the saturation value for the respective current sample is considered. Third, the magnetic unbalance of the machine generates disturbanc-
es. In contrast, the residual ripple is very low at $t < 0$ in Fig. 15, with only 0.3 p.u. magnetizing current flowing.

The corresponding current trajectories in field coordinates are shown in Fig. 17. A single pulse pattern exists in the steady-state before the step, while changing pulse patterns prevail thereafter at acceleration. The curved transient portion in Fig. 17(a) is reduced to a straight line in Fig. 17(b) after extraction of the fundamental: a near-ideal response.

Using optimal synchronous PWM without trajectory tracking control, the peak-to-peak modulation error results so high that it exceeds the nominal machine current at constant torque operation, as shown in the lower portion of Fig. 18. After a step change, the high dynamic modulation error causes an overcurrent trip of the inverter. Thereafter, the current reduces to zero on a picewise linear trajectory caused by a sequence of diode commutations [12].

5. Summary

The existing current control methods for high-bandwidth servo drives loose performance quality when applied to slow switching ac drives in the megawatt power range. Also the methods of pulselength modulation need be optimized in order to prevent excessive harmonic distortion and torque pulsations. The dynamic modulation error is another source of substantial nonperiodic distortion. From a comparative study, synchronous optimal pulselength modulation with current trajectory tracking control evolves as the undisputed preference for high-power drives with dynamic performance requirements. This modulation method increases the available output power by 10%.

The paper demonstrates the advantage of using a dead-beat tracking controller not only for the elimination of the dynamic modulation error, but also for fast torque control. The large-signal rise-time equals that of a hysteresis-band controller. An additional signal delay depends on the computation cycle time. This interval is 500 $\mu$s in the test set-up but can be considerably smaller when using modern signal processing hardware. Optimal steady-state operation is instantaneously regained after execution of a commanded transient.

6. References