Abstract-A predictive control scheme suitable for the torque and flux control of multilevel inverter-fed induction machines is presented in this paper. The control strategy combines the use of a PI controller to obtain a good steady-state behavior and a predictive controller to achieve a fast dynamic torque response. In this way, torque and stator flux references can be reached in the next sample period. With the use of Multilevel Space Vector Modulation (MSVM), low torque and flux ripple are possible with fixed sample rate. Experimental and simulation results are presented in order to demonstrate the effectiveness of the proposed strategy software.

I. THE INVERTER TOPOLOGY

A cascaded multi-cell inverter topology is considered in this work as it is shown in Fig. 1. Each cell is composed of a diode rectifier, a DC-link capacitor and a single phase H-inverter. It can be easily verified that an inverter with 1 cell synthesizes stepped waveforms with 3 voltage levels $U_{dc}$, $0$, $-U_{dc}$ and each extra cell adds two extra levels per phase [8].
II. MATHEMATICAL MODEL

For the modeling of the induction machine, the following assumptions were made:

- The neutral point is not connected.
- There are no eddy currents or core losses in stator and rotor.
- Only the fundamental wave of the air gap field is considered for the calculation of the inductances.

Using these assumptions, common in the modeling of electrical machines, the induction machine will be described by the following well-known set of complex equations in the stator reference frame:

\[ u_1 = i_1 \cdot R_1 + \frac{d}{dt} \psi_1 \]  
\[ 0 = i_2' \cdot R_2' - j \cdot \dot{\psi}_2 + \frac{d}{dt} \psi_2' \]  

The fluxes and the currents are related by:

\[ \psi_1 = L_1 \cdot i_1 + L_h \cdot i_1' \]  
\[ \psi_2' = L_h \cdot i_1 + L_2 \cdot i_2' \]  

and the electromagnetic torque is given by

\[ M = \frac{3}{2} \cdot p \cdot \frac{L_h}{\sigma \cdot L_1 \cdot L_2} \left| \psi_1 \right| \left| \psi_2' \right| \sin(\theta_{12}) \]  

With the mechanical equation

\[ M - M_L = \frac{d}{dt} J \cdot \dot{\Omega} = \frac{J}{p} \cdot \frac{d}{dt} \Phi \]  

the mathematical model of the induction machine is fully described.

III. THE CONTROL STRATEGY

A. Basic Operation Principle

The presented control strategy, like all "Direct Torque Control" methods, is based on the control of the magnitude and phase of the stator flux space phasor through the proper choice of the stator voltage space phasor. From eq. (5) and under the assumption of constant magnitude of stator and rotor flux, it is clear that a fast torque change can be achieved with a change of the angle \( \theta_{12} \) between both fluxes. Since the sample time of the controller is smaller than the rotor time constant, the variation in the trajectory of the rotor flux during the sample time interval can be neglected. Hence, it is the stator flux space phasor the main quantity influencing the torque development.

An example of the control principle can be seen in fig. 2. If an increase of the torque is needed, the angle variation of the stator flux \( \Delta \theta_1 \) must be greater than the angle variation of the rotor flux \( \Delta \theta_2 \) in a sample time and vice versa.
Finally the calculated voltage space phasor is synthesized using a Space Phasor Modulator. Since a modulator does not work with the instantaneous values, but with the mean values over a sample time, this method should not be considered as “direct control”.

Fig. 3 shows the basic structure of the controller. The torque reference is subtracted from an estimated value calculated with the phase currents and electrical rotor angle. This error is supplied to a PI controller which delivers the variation of the angle $\Delta \vartheta_{12}$. The task of the controller is relieved by the addition of the term $\Delta \vartheta_2(t_k)$ as a feed-forward [9],

$$\Delta \vartheta_2(t_k) = (\dot{\vartheta} + \frac{2R_2\cdot M_1^*}{3p\cdot L_1\cdot L_2^*})T_s.$$ (10)

Although a PI controller has advantages like zero steady state error and robustness, this controller has only a moderate dynamic performance in comparison with a DTC strategy. In order to improve the dynamic behavior, the difference $\Delta \vartheta_{12}(t_{k+1})$ can be predicted to reach the torque reference in the quickest way possible.

### B. Dead-beat Operation

A dead-beat control aims to the quickest possible dynamic response of a system. That means, after a time equal to the sample period, the controlled variables should reach the set-point and remain there [10]. In order to obtain a dead-beat like system response, eq. (5) can be used to calculate the slip angle $\vartheta_{12}(t_{k+1})$ which fulfills the requirement of torque in the next sample time as follows:

$$\vartheta_{12}(t_{k+1}) = \arcsin\left(\frac{2}{3} p \cdot L_1^* \cdot \psi_1^*(t_{k+1}) \cdot \psi_2^*(t_k)\right).$$ (11)

Thus, the output for the dead beat control is determined by:

$$\Delta \vartheta_{12}(t_k) = \vartheta_{12}(t_{k+1}) - \vartheta_{12}(t_k)$$ (12)

By assuming a good model of the inverter-motor system, the torque reference should be reached in the next sample time. However, if torque variation is big, especially for high rotor speeds, the stator voltage calculated by equation (9) could have a magnitude beyond the maximum output voltage of the inverter. In this case the controller operate with the maximum voltage possible, to obtain a torque in the next sample time as close as possible to the reference. As soon as the torque reach a window with size $\varepsilon$ around the torque reference:

$$-\varepsilon \leq M - M^* \leq \varepsilon,$$ (13)

the operation of the PI controller can be resumed. A soft transition between booth operating modes can be achieved if the integral part of the PI controller is loaded with a predicted angle variation for the next sample time.

It should be noted that the equations were expressed only one step in the future, assuming that the calculation time can be neglected. Considering that the calculation of the control algorithm needs some time, the value predicted in the real algorithm correspond actually to $t_{k+2}$. Thus, the output voltage is calculated for the next sample time $t_{k+1}$, and the references values of torque and flux are reached in $t_{k+2}$. A diagram of the
control strategy is presented in fig 4. The control principle can be applied in a converter of any number of levels.

C. Multilevel Modulator

The last block in the control algorithm is a modulator which synthesizes the reference space voltage phasor $u_1^*$ using three nearest inverter space voltage phasors. As illustrated in fig. 5 for a 3 level inverter, these space phasors corresponds to the vertices of the triangle which contains the reference $u_1^*$. Due to the high number of available space voltage phasors in multilevel topologies, this election can be a time consuming process.

Based on the algorithm presented in [11], the transformation in eq. 14 was used to locate $u_1^*$ in a new coordinate system, where the position of each inverter voltage space phasor is described only by integer numbers.

$$u_j(t_k)^* = \begin{bmatrix} 3/2 & -\sqrt{3}/2 \\ 0 & \sqrt{3} \end{bmatrix} u_i(t_k)^*$$  (14)

Thus, the coordinates for the three nearest inverter space voltage phasors $u_a$, $u_b$, $u_c$ can be easily found using the elementary approximation functions ceil$(x)$ and floor$(x)$ with the real and imaginary part of $u_1^*$. Once these space phasors are determined, the duty cycle is calculated using the standard method for two level inverters. Finally, the three-phase output voltage combination $(u_{UN}, u_{VN}, u_{WN})$ corresponding to each voltage space phasor were calculated considering the criterion of common mode voltage minimization. That means, the modulator selects only a combination which produces the lower common mode voltage defined by

$$u_{cm} = \frac{u_{UN} + u_{VN} + u_{WN}}{3}.$$  (15)

The reduction of the common mode voltage allows the mitigation of problems like bearing failures, and EMI [7]. Considering Fig. 5 as example, from the two alternatives to generate the voltage phasor $u_a$, the second one $(0, 0, -U_{dc})$ generates a common mode voltage with lower amplitude. Despite the advantages that the common mode voltage reduction offers, the number of commutations for some voltage phasor transitions with this algorithm is not minimum. Furthermore it must be used in combination with an over-modulation strategy, if the performance of the DTC in the high speed range wants to be reached.

IV. SIMULATION AND EXPERIMENTAL RESULTS

A 15kW prototype of a multilevel inverter was developed in order to verify the proposed control method. Each cell is composed of a three-phase rectifier, a DC-link of 250V and 4 IGBTs IRG4PH40UD of 21A nominal current with their respective Drivers (fig. 6). A floating point DSP platform ADSP-21062 of Analog Devices was used for the implementation of the digital control. The modulator, the timers and the incremental encoder were programmed in a FPGA included in the board. The sample period was set up in 200µs, far more than a conventional DTC. The multilevel inverter was connected to a 5.5kW induction machine for the first tests.

![Fig. 5: Space phasor diagram of a three level inverter](image)

![Fig. 6: A cell of the multilevel Inverter, a) Top Side b) Bottom side.](image)
Fig. 7a presents a simulation of the dynamic performance of the control algorithm. A step in the torque reference of 2Nm was applied with a DC-link voltage of 120V and with the machine running at 240min⁻¹. It can be observed that the proposed control strategy has a short torque rise time and the magnitude of the stator flux presents no change. Fig. 7b shows the same test in the experimental set-up with a three level inverter. The result is similar to the simulation but a little overshoot due to an imperfect prediction of the fluxes can be seen in the torque.

The dynamic behavior of the dead beat controller depends strongly on the capacity of the converter to deliver enough voltage. For some operating conditions, like in the high speed range, the controller could deliver a reference voltage $u_1^*$ with a modulation index greater than the maximum. In this case a stator voltage phasor with maximum amplitude is utilized. Fig. 8 illustrates this situation. For a rotor speed of 1280min⁻¹, the control strategy uses the maximum output voltage but the torque reference value is not reached in one step. Therefore an extra sample period is necessary to reach the reference value. In this saturation condition, the dead beat strategy offers no special advantages respect to the conventional PI-Controller solution.

V. CONCLUSION

An application of a predictive control strategy for induction machines in multilevel inverters has been presented. The proposed control technique combines the use of a PI controller to obtain a good steady-state behavior and a predictive controller to achieve a fast dynamic torque response, like the DTC strategies. Experimental results for a 3 level converter were obtained, showing low torque ripple, fast
dynamics and almost fixed switching frequency. The strategy can also be applied to a n-level inverter.

Since the torque controller uses predicted values of the torque and fluxes to feed the control loop, discrepancies will be observed if this prediction is not performed with enough precision.

It should be noted that the use of a modulator reduces the performance of the proposed method in the high speed range, compared with a DTC strategy. For this reason, the strategy must be used in combination with an over-modulation, if the performance of the DTC in the high speed range has to be reached. In further works this aspect will be considered.

REFERENCES


