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ELECTRIC MOTORS AND DRIVES IN TORSIONAL VIBRATION ANALYSIS AND DESIGN

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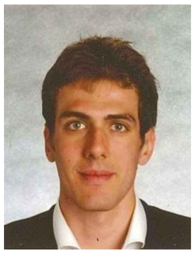

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ABSTRACT

Two types of prime movers are continuously applied for rotating machinery: mechanical and electric drives. This tutorial focuses on electric drives, which are actually composed of separate components: electric motor, frequency converter (optional), and electric network including transformers.

The main aim of this tutorial is to explain the phenomena of electric drives related to the torsional vibrations. All the components of an electric drive may affect torsional dynamics of the system. Naturally the main concern is torsional excitations. Another aspect is the electromechanical interaction in the air-gap of the motor that produces electromagnetic torsional stiffness and damping. These and other similar phenomena, related to the electric drives, have been observed on test fields and site conditions. The scope of the paper includes the most common induction, synchronous and permanent magnet motors.

First, the various rotor structures of electric motors are discussed. Then, the effects of electromagnetic stiffness and damping on the torsional dynamics are introduced. Next, the main electric excitations are presented and their effect on drive-train dimensioning is discussed. Various modelling and analysis methods are also presented for drive-trains with an electric motor.

The previous sections focus on the motor without any frequency converter. The rest of the paper deals with the variable speed drives. First the main voltage source inverter, current source inverter and load commutated inverter drive types, their operational principles and application areas are introduced. Then the harmonic and inter-harmonic torsional excitations are

explained that depend on the configuration and control of the frequency converter.

The main functionality of a variable speed drive system is the speed-control feedback loop. This loop is used to maintain the preset speed. Although this feedback loop is also a potential source of instability it can with proper tuning be used to damp low frequency torsional resonances.

The analysis guidelines and design principles are distributed throughout the paper. The paper is concluded with summary and main recommendations.

INTRODUCTION

The electrical drives are composed of an electric motor and its electric supply system. Thus analysis cannot always be limited to the electric drives but have to be extended over the whole electric drive from the shaft-end to the electric grid including the optional frequency converter.

In order to predict the vibration behaviour of any torsional system, the inertia and stiffness characteristics of all the components are needed. Typically, the electromagnetic stiffness and damping due to the electromechanical interaction in the air-gap are ignored, but they can be important when improved prediction accuracy of natural frequencies or damping is required. In addition, to dimension the drive-train components, all the excitations must be known both in the rated operation conditions and during the different failure occurrences.

In most industrial cases, the drive train components are designed and manufactured by various vendors. Usually, these vendors are well familiar with their own component, but only superficially acquainted with other components. This poses a challenge to the system integrator to successfully design his system.

The main aim of this tutorial is to explain the phenomena of electric drives related to the torsional vibrations, with a target audience of mechanical engineers with background in torsional dynamics but without previous knowledge of electrical machines. The scope of this paper will be the modelling, analysis and design of electrical drives as a part of torsional systems.

COMPONENTS AND TYPES OF ELECTRIC DRIVES

The function of electric drives is to convert electric energy into the form of mechanical energy. The electric energy is supplied by the electric grid and the mechanical energy appears as a rotating shaft end. The main parameters describing the electric drives are power and speed. The basic principles of electric drives and power conversion are presented, e.g. by Fitzgerald, et al. (2003) and Barnes (2003).

Power Conversion

The power conversion is produced by the magnetic field in the air gap between the rotor and stator. The magnetic field is characterized by the number of poles. The number of poles is defined by the motor construction. Because every north pole of a magnet has to have a corresponding south pole, the number of poles is always even. Figure 1 shows a cross-section of the electromagnetically active parts of a small electric motor. The main structural parts of an induction motor are shown in Fig. 2.

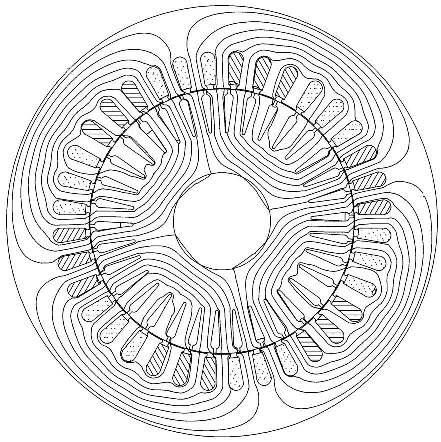

Figure 1. Magnetic Field of a Four-Pole 15 kW Induction Motor (Holopainen and Arkkio 2008).

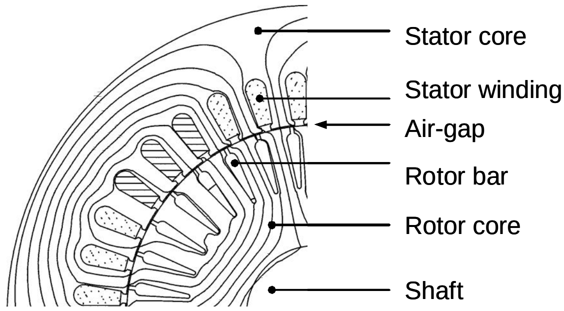

Figure 2. Main Parts in the Core Region of a Four-Pole 15 kW Induction Motor (Holopainen & Arkkio 2008).

Motor Speed

Most of the industrial motors are supplied by alternating current (AC). There are also direct current (DC) motors, but they are omitted in this paper. Two main types of AC motors exist: synchronous and asynchronous. Asynchronous motors are also called induction motors. The rotor of a synchronous motor rotates synchronously, i.e. with the same angular speed of the magnetic field. The relation between the synchronous speed n_1 and the motor supply frequency, given in revolutions per minute (rpm), is

$$n_1 = \frac{60 \cdot f_1}{p} \frac{\text{rpm}}{\text{Hz}} \tag{1}$$

where f_1 is the supply frequency in Hz, and p is the number of pole-pairs of the motor. It can be seen that the number of poles ($= 2p$) is closely related to the speed of the motor. For example, the speed of a four pole synchronous motor fed from 60 Hz grid is 60 · 60 Hz / 2 rpm/Hz = 1800 rpm.

The rotor of an asynchronous motor rotates asynchronously, i.e. with a lower angular speed than the magnetic field. The relation between the rotor speed and the motor supply frequency of an asynchronous motor in rpm is

$$n = 60 \cdot (1 - s) \cdot \frac{f_1}{p} \frac{\text{rpm}}{\text{Hz}} \tag{2}$$

where s is the slip. The slip is the relative speed difference between the magnetic field and rotor angular speed

$$s = \frac{n_1 - n}{n_1} \tag{3}$$

where n_1 is the synchronous rotational speed in rpm. The typical order of magnitude of the slip is 0.01 - 1 percent in the rated operation condition. Thus, the load only slightly affects the speed of induction motors.

Adjustment of Speed

As shown above, the rotational speed of direct-on-line (DOL) motors is quite fixed. Stepwise speed control is possible with special pole-changing induction motors that have more than one set of windings in their stator. With a 60 Hz grid, the rpm alternatives for synchronous speeds are: 3600, 1800, 1200, 900, etc. Pole-changing motors usually have two or three consecutive speeds from the previous list as their synchronous speeds.

Due to the fairly obvious advantages of stepless speed control, the application of frequency converters has increased remarkably during the last twenty years. The main task of the frequency converter is to adjust the frequency of the voltage fed to the motor in order to achieve the needed rotational speed or torque. This can be achieved with various control schemes with or without speed feedback.

Electric Grid

The last component of an electric drive is the electric grid. There are remarkable variations in electric grids used to supply electric motors. An essential feature of these distribution grids is the stability. The quality of power supply affects directly the torsional dynamics of a drive-train including a DOL motor. The unbalance between the phases and short circuits are well-known examples of electrical network disturbances appearing as extra excitations. As an additional benefit, if a frequency converter is used, most of these disturbances originated in the grid are isolated from the torsional system.

TORSIONAL STIFFNESS AND DAMPING OF ROTOR STRUCTURES

The rotor construction of an electric motor varies greatly depending mainly on the motor type, motor size and operational speed. The different constructions balance the electromagnetic and mechanical requirements of various applications. The electrical requirements are closely related to the total electric losses and the mechanical requirements to the rotor strength and thermal stability.

Induction Motors

The typical features of an induction motors are: the laminated rotor core and the rotor cage (Figure 1). The lamination is made of electrical steel sheets with thickness for example 0.5 mm. The typical materials of the cage are copper, aluminum and bronze. In addition, the rotor core of large motors usually includes axial and radial air ducts for the cooling purposes.

Synchronous Motors

The large synchronous motors have two basic designs: salient or non-salient pole rotors. The salient pole rotors are used for lower speeds and the poles are usually made of laminations of steel. However, there are some applications where solid steel poles are preferred. The armature and damper winding together with additional cooling ribs are mounted around the poles. The poles are usually removable and mounted separately on the rotor shaft.

The non-salient pole rotors are used for higher speed applications. The typical speeds are 1500 rpm and above. The core is usually made of solid steel with milled slots, but there are also laminated constructions with similar slots. The armature winding is located in the slots.

The synchronous motors with permanent magnet technology rely on magnets fitted in or on the rotor. The rotor core is usually made of steel laminations.

Rotor Inertia and Stiffness

A proper torsional analysis and design of a drive-train requires that there is adequate data describing the stiffness and inertia characteristics of the rotor. A common industrial practice is based on the rotor drawing with complementary data from motor manufacturer.

In practice, all the rotor constructions are fabricated structures consisting of possibly hundreds of separate parts. These parts are fitted together by different connection methods like dovetail joints, bolted joints, friction joints, interference fits, wedging etc. In addition, some of the fabricated rotors are resin treated, which affects the joints of the structure. Finally, due to the different thermal expansion factors of steel and winding materials like copper, the connection between the parts may change as a function of rotor temperature.

The distribution of rotor inertia is usually relatively well-known and can be easily used in the rotordynamic analyses. The determination of the torsional stiffness characteristics is more problematic. The easy part is to calculate the torsional stiffness of the solid part of the rotor without any joints. For cylindrical sections, there is a well-known simple formula for the torsional stiffness

$$k = \frac{\pi G D^4}{32L} \tag{4}$$

where G is the shear modulus, D is the diameter, and L is the length of the section. When the cross-section of a solid shaft is non-cylindrical, analytical formulas can be applied (Wilson 1956, API 684). A typical example of these is a spider shaft. In addition, the finite element method can be used for more complicated cross-section geometries and for the modeling of abrupt shaft changes in axial direction. This means that the stiffness of the solid shaft part of the rotor can be estimated with desired accuracy. The most challenging aspect is to determine the stiffening effect of the rotor core.

Stiffening Effect of Rotor Core

The stiffening effect depends on the rotor core construction. For example, the stiffening effect of a laminated core with radial air-ducts without resin treatment is naturally small. The stiffening effect can be estimated by different calculation methods. An essential observation is that the laminated rotor core is transversely isotropic and defined by the five independent elastic properties (Pilkey 1994). The material properties in transversal direction are close to those of steel. In axial direction the modulus of elasticity is much lower. Garvey et al. (2004) reported typical values for the modulus of elasticity in axial direction: E_a = 120,000 – 260,000 lbf/in.^{2} (0.8 – 1.8 GN/m^2). The potential radial air-ducts reduce this equivalent axial value even more. Further, the axial modulus of elasticity can be used to estimate the effective shear modulus that is required when the stiffening effect is evaluated.

The published information about stiffening effect is focused mainly on lateral rotordynamics (Garvey, et al. 2004). They present an approach to model the stiffening effect of laminated rotor core and a method to determine the required parameters from experimental results. This approach can be applied for the parameter estimation of stiffening effect in torsional vibrations. A common approach is to measure the natural frequencies of a free-free rotor and adjust the physical parameters to predict as many lowest modes as possible.

Mechanical Damping

Considering that the rotor of an electric motor is fabricated and there are usually plenty of friction joints, it is a bit surprising that the mechanical damping is usually low. This observation is consistent with the experience of drive-train damping. The typical damping ratio of a gearless drive train is about 1 percent (Corbo and Malanoski 1996). The typical drive-trains include electric motors and thus these values can be applied for the mechanical damping of their rotors.

Analysis Guidelines

The modeling of the solid steel part and the inertia of the rotor is relatively straightforward. It is good to remember that the torsional stiffness is proportional to the fourth power of the shaft diameter. Thus, the shaft sections close to the coupling and the drive-end journal are most flexible and dominate the motor contribution to the coupling mode (= nodal point of the mode roughly in the coupling). Anyway, the pivotal component dominating the coupling mode frequency is usually the coupling.

The stiffening effect of the rotor core varies depending on the core structure. Motor manufacturers may have experimental values for stiffening effect of different core structures. These values, if available, may improve particularly the prediction of higher frequency modes. In these modes, the torsional deformation occurs along the core region.

While the rotor is a fabricated structure with multitude of parts, a damping ratio applied for general torsional systems, e.g. 1 percent, is a good choice for rotors of electric motors.

NATURAL MODES AND FREQUENCIES

The torsional modes and frequencies of an electric motor rotor are determined by the distribution of stiffness and inertia. Typically, the main part of the inertia is roughly evenly distributed in the rotor core. In some synchronous machines, there is another remarkable inertia in the form of an exciter. A characteristic parameter related to the torsional behavior is the relative diameter of the rotor core compared to the core length. In general, the relative diameter is smaller with rotors for higher rotational speeds.

As an example, Figure 3 shows the five lowest torsional modes and frequencies of a two-pole 9400 hp (7 MW) induction motor. The relative rotor diameter is 0.46 and the rotor is rigidly fixed in the coupling end.

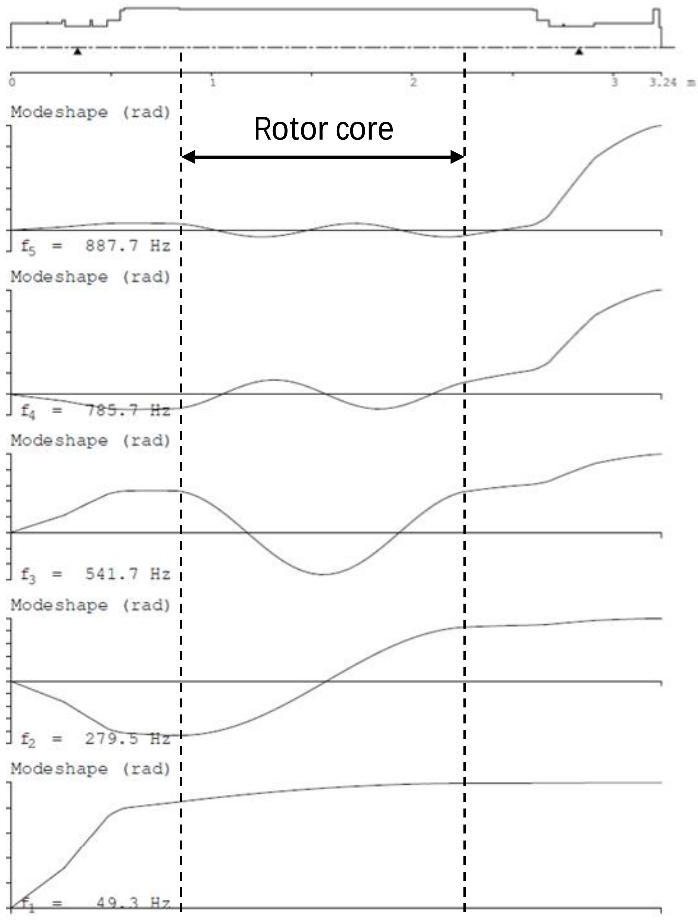

Figure. 3 The Five Lowest Torsional Modes of a Two-Pole 9400 hp Induction Motor with Rigidly Fixed Coupling End on the Left. High up, the Geometry of the Shaft without Rotor Core is Shown.

ELECTROMECHANICAL INTERACTION

Even Distribution of Torque

The electromagnetic field in the air-gap of an electric machine produces electromagnetic forces between the rotor and stator. These electromagnetic forces on the surface of the rotor can be divided between the radial and tangential components. Only the tangential component in circumferential direction, i.e. electromagnetic traction, is relevant on the torsional behavior. The distribution of this electromagnetic traction has useful regular features. Particularly, if we consider any arbitrary time instant, the amplitude of electromagnetic traction is practically constant as a function of axial coordinate and varies cyclically in the circumferential direction. This means that the distributed traction on the surface of the rotor generates an electromagnetic torque, which is evenly distributed over the length of the core. These general statements about a) the prismatic distribution of traction, and b) the even distribution of torque, are true for the distributed load, and separately for each harmonic component. Thus, the torque distribution of a) the fundamental field, b) the harmonic components, and c) the inter-harmonic components are evenly distributed over the rotor core. In addition, the distributed torque of each component has a common phase over the rotor core.

There is one exception for the prismatic distribution of traction. In some motors, the rotor or stator slots are skewed in order to reduce the harmonic components of electrical torque.

Modal Sensitivity Factor

The evenly distributed torque over the length of the core together with the variation of modal amplitudes determines the effective torque of each mode. The effective modal torque can be calculated by integrating the product of the evenly distributed torque and the mode shape over the rotor core. The comparison of modal excitation sensitivity for electromagnetic torque is more difficult. One measure to evaluate this modal sensitivity is to calculate the kinetic or strain energy in the resonance condition.

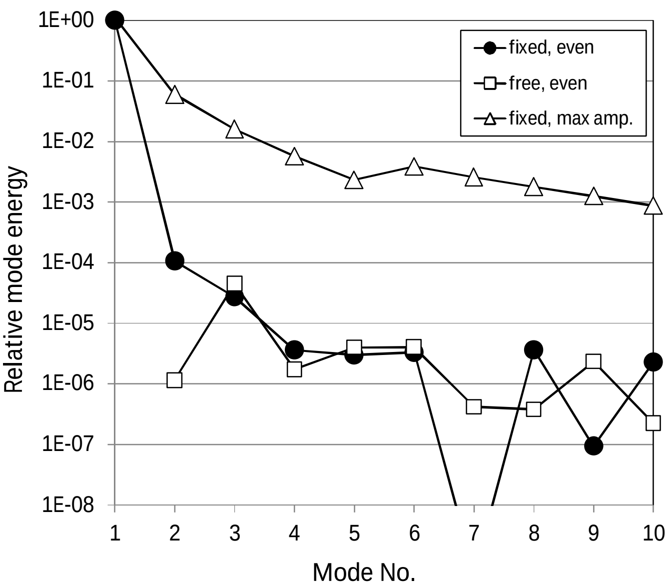

Figure 4. Relative Mode Energy in Resonance Condition Excited by an Evenly Distributed Unit Torque or by a Unit Torque in Maximum Modal Amplitude of the Core.

Figure 4 shows the relative strain/kinetic energy for the unit harmonic excitation torque with a modal damping factor (2 percent). The example rotor is the same two-pole 9400 hp rotor presented above. The effect of boundary condition is evaluated by applying fixed (=rotation restricted) or free connection (= rotation totally free) in the coupling end. It is assumed that these two extreme conditions give qualitatively the overall picture of the behavior with real torsional trains. The natural frequencies of these modes with fixed boundary condition are between 49 Hz and 2031 Hz. The five lowest modes are shown in Figure 3. The natural frequencies with free boundary condition are between 0 Hz and 1864 Hz. The zero frequency mode (No 1) is the rigid body mode and the first elastic mode has the frequency 269 Hz. Finally, Figure 4 shows also the relative strain/kinetic energy in the resonance with fixed boundary condition and excited by a unit torque in the maximum modal amplitude of the core region.

Figure 4 shows clearly that the evenly distributed torque (in the same phase) will be effectively cancelled out by the spatial variation of the second and higher modes. The effect of boundary conditions in the coupling end seems to be minor.

The cancelling effect due to torque distribution cannot be present when the torque is given in one point on the core. Thus, the decreasing curve of the unit torque response, in Figure 4, means that the relative mode energy decreases with increasing mode number. The additional decrease between the unit torque and evenly distributed torque response shows clearly the cancelling effect of the mode shapes. It can be added that the rotor stresses, determining the dimensions, are proportional to the square root of the strain energy.

The example rotor used in the example calculations was slender with relative rotor core diameter 0.46. The situation changes slightly when the rotor core is shorter or it is unsymmetrically located on the shaft. However, it can be concluded that, in general, only the lowest mode, i.e. the coupling mode, can be effectively excited by the electromagnetic torque.

In addition, if there are other drive-train modes with motor core oscillating without twisting, these modes are prone to electromagnetic excitations. One example is the vibration mode of a synchronous motor with twisting between the motor core and exciter. Usually, the excitation of higher modes, with natural frequency clearly above the coupling mode, is effectively cancelled out.

Analysis and Design Guidelines

The most prone mode for the electrical excitations is the lowest mode, i.e. the coupling mode. In addition, the electromagnetic excitation may excite drive-train modes with motor core oscillating without significant twisting. The higher modes with motor core twisting cannot be excited easily by electrical excitations. Among other things, this means that the high-order harmonics (6^{th}, 12^{th} …) of frequency converters are effectively cancelled out by electric motors.

In order to model the high-frequency behavior reliably, it is essential to use a refined model, i.e. short elements, for the rotor core and apply evenly distributed torque to describe the electromagnetic excitations. These modeling principles are necessary to model the cancelling effect of high-frequency modes.

ELECTROMAGNETIC STIFFNESS AND DAMPING

Electromechanical Interaction

The electro-mechanical conversion of power is produced by the constant component of air-gap torque. The time-harmonic components of the torque are torsional excitations. These electromagnetic excitations will be considered in the next section. In addition to the valuable torque production and the harmful excitations, there is a third phenomenon related to the time-variation of the electromagnetic torque. The source of this phenomenon is the interaction of the rotor oscillation and the electromagnetic fields. Due to this interaction an oscillating motion produces always a corresponding harmonic torque with the same frequency. This interaction turns up as an electromagnetic stiffness and damping coefficients. These resemble the negative stiffness and damping coefficients applied in lateral rotordynamics of electrical machines, but are somewhat different.

Example of Stiffness and Damping

Figure 5 shows, as an example, the calculated electromagnetic stiffness and damping of an eight-pole 1250 hp (932 kW) induction motor for blower application. The motor operates in steady-state condition with the supply frequency of 60 Hz and the stiffness and damping is determined for the oscillating motion as function of its frequency.

Figure 5 shows that the electromagnetic stiffness is relatively constant over the frequency range. The drop the stiffness is related to the supply frequency of 60 Hz. The electromagnetic damping varies strongly. The most significant is the range of negative damping between 49.7 -59.4 Hz. This range is located somewhat below the supply frequency. The damping coefficient approaches the value 10600 Nms/rad when the frequency approaches zero. This damping value at zero frequency is equivalent with the derivative of the speed-torque curve in the rated operation point. It can be concluded that the electromagnetic damping is strongly dependent on the oscillation frequency.

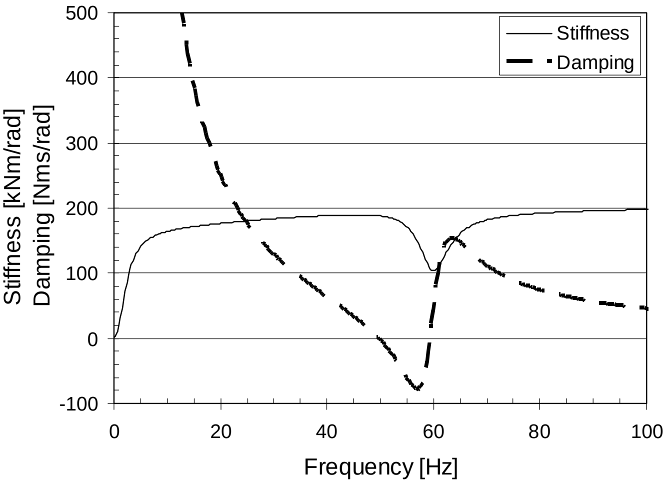

Figure 5. Torsional Stiffness and Damping of a 1250 hp Induction Motor with Supply Frequency 60 Hz (Holopainen et al. 2010).

Significance of Electromechanical Interaction

The significance of electromagnetic stiffness and damping is dependent on the application. The presented 1250 hp motor was used to drive a blower in a direct-on-line application (Holopainen et al. 2010). This drive-train is basically a two-inertia system with the lowest natural frequency at 28.1 Hz without electromechanical interaction. In the rated operation condition the supply frequency is 60 Hz and the rotational frequency 14.9 Hz, i.e. 892 rpm.

The electromagnetic interaction increased the torsional natural frequency 4.3 percent up to 29.4 Hz and the purely electromagnetic damping ratio was 0.6 percent. In addition, the originally rigid body mode at 0 Hz appears in the rated operating condition at 5.1 Hz. The damping ratio of this mode is 23 percent. This natural frequency and mode remains usually unnoticed because the electromagnetic damping is so large

It can be concluded that the significance of electromagnetic effects is dependent on the mode shape and particularly the vibration amplitude of the rotor core. The inertia ratio, i.e. the motor inertia divided by the load inertia, of the example machine is relatively small 0.57. This leads to large relative amplitude of the rotor core, and further, to more significant effects of electromechanical interaction.

Other Machines

The electromagnetic stiffness and damping presented in Figure 5 is qualitatively representative for induction machines. The general shape of the curves follows from the physical principles of induction motors. Thus, the presented behavior is inherent for all induction motors. Thus, it can be assumed that the electromagnetic stiffness is always positive and that the damping has a negative value in a frequency range just below the supply frequency.

Figure 6 shows electromagnetic stiffness and damping of a 36000 hp (27 MW) salient-pole synchronous motor for piston compressor application. The supply frequency is 50 Hz and the

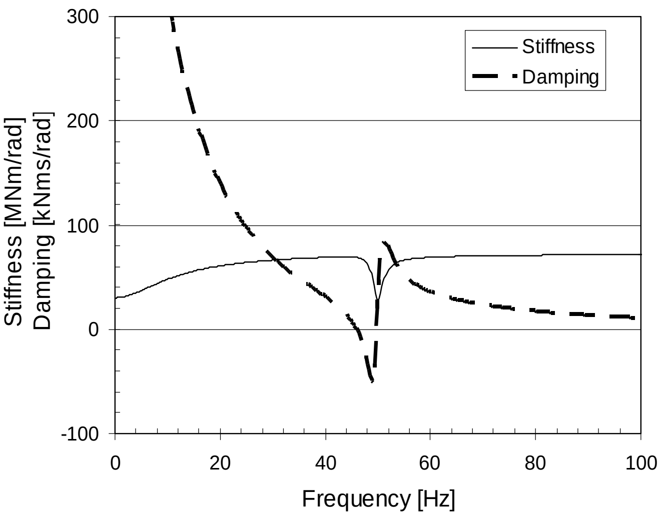

Figure 6. Torsional Stiffness and Damping of a 36000 hp Synchronous Motor with Supply Frequency 50 Hz (Holopainen et al. 2010).

rotational speed 200 rpm. As can be seen by comparing Figures 5 and 6, the behavior is qualitatively very similar. There is only one exception: the electromagnetic stiffness at zero frequency is larger than zero 27.6 MNm/rad. The frequency of originally rigid body mode increases up to 2.7 Hz and the damping ratio up to 14.0 percent due to electromagnetic interaction. It can be mentioned that this mode is sometimes critical because the piston excitations are relatively large and may coincide with this mode.

The behavior of permanent magnet motors is different. The rotor construction of these motors does not include windings. Without windings there will not be any anomaly in stiffness and damping values close to the supply frequency and thus, the frequency dependency is small. In addition, the electromagnetic damping is small compared to induction and other synchronous motors.

Design and Analysis Guidelines

The electromagnetic stiffness and damping affects the torsional dynamics of the system. Currently, the inclusion of these effects is not common. The reason for this is most probably the missing interface between the rotordynamics and electromagnetic calculations.

It is clear that the effect of electromagnetic interaction is largest in the coupling mode and particularly in applications were the load inertia is large compared to the motor inertia. In addition, it is known, and can be cautiously used, that the electromagnetic stiffness may increase the natural frequency of the coupling mode by some percent.

The electromagnetic damping can be regarded mainly as an additional bonus for the mechanical damping of the system. However, it is good to be aware of the negative damping ratio below the supply frequency. In our induction motor example the negative damping ratio frequency was 83…99 percent and in the synchronous motor example 92…100 percent of the supply frequency. The contribution of negative damping could be a problem, if there is a torsional mode close to the supply frequency. However, there are recommendations, and in some standards requirements that all the natural frequencies of a complete train must be removed at least 15 percent (API 541 Induction motors) or 10 percent (API 546 Synchronous motors) from the line frequency.

ELECTROMAGNETIC AIR-GAP LOADS

Steady State Operation

The electromagnetic excitations of an ideal electric motor in steady state operation are usually very small. The potential excitations occur at high frequencies due to slot harmonics. The meaning of these harmonics is usually minor in the design of the torsional drive-train.

If the voltage supply is unsymmetrical, there will be an excitation torque at twice-line frequency. In principle, a rotor eccentricity may also appear as a twice-line excitation source. Theoretically, there is not any excitation at line frequency for a three phase motor, and it is difficult to find any non-ideality to generate this component. However, there seems to be empirical observations that indicate the existence of line frequency excitations. It has been suggested representative torque magnitudes of 1.0 percent of rated torque for the line frequency excitations and 0.5 percent for the twice-line frequency excitations (Wachel and Szenasi 1993, Corbo and Malanoski 1996, API 684).

Starting

The direct-on-line start of an electric motor generates high transient torque fluctuations at line-frequency. The length of this air-gap transient is usually a fraction of a second. The maximum torque amplitude is dependent on the motor characteristics but the typical values are 4-7 times the rated torque.

The direct-on-line start of a synchronous motor has an additional excitation component which must be considered particularly with large motors. This torsional excitation occurs at two times the slip frequency when the motor operates in the asynchronous starting mode. The frequency of this excitation can be given by the formula

$$f_{\mathrm{exc}} = 2 f_1 \cdot \frac{n_1 - n}{n_1} \qquad (5)$$

This means that the excitation will be two times the line frequency at start and will reduce linearly with speed increasing towards the operational speed. Because the coupling mode is in practice always below the twice-line frequency, this loading will excite this resonance vibration. Though this resonance will be transient, this loading is usually critical for the dimensioning of shaft systems of large synchronous DOL motors. The starting of synchronous motors has been presented in API 684 (API 684) and discussed thoroughly by Corbo et al. (2002).

Electric Faults

A short-circuit is the most common electric fault affecting the electric motor and thus the whole drive-train. Two types of these failure cases are typically used for the dimensioning of the system: three-phase and two-phase short-circuit. Single phase-to-ground faults are much rarer and less severe for the motor and are thus usually neglected.

The transient of the three-phase short-circuit is very similar than the transient of the DOL start of the machine. The frequency of this transient is the line-frequency and the typical maximum amplitude of the fluctuation 4-7 times the rated air-gap torque.

The two-phase short-circuit includes two frequencies: the line and the twice-line components. The maximum amplitude is typically 5 – 15 times the rated air-gap torque. Figure 7 shows an example of a two-phase short-circuit torque.

It can be added that the short circuit torque amplitude is dependent on the location of a short circuit. If the fault occurs far away in the supply network, the short-circuit torque will be attenuated. In principle, the situation is worst when the short circuit occurs in the motor terminals. This worst case scenario is usually applied in the calculations.

Reclosure and Bus-transfer

The power supply of an electric motor can be interrupted by faults in the power system. The reapplication of the power is carried out by the breaker reclosure. This may involve a bus-transfer where the power transmission supply line is changed. It is important that a premature reclosure of a motor may induce a transient, which in induction motors can be as high as 15 – 20 times the rated torque (API 684). The synchronous motors have similar behavior and the reclosure torques can be even higher. The reason for these torque transients is the residual magnetic flux which remains in the motor after the voltage supply is interrupted. The transient torque is generated by the interaction of this residual flux and the new current inrush due to the reclosure. The induced torque transient is a function of the remaining residual flux and the phase difference between the residual flux and the flux generated by the inrush currents. The remedy for these transients is the adequate reclosure control (API 684, NEMA MG-1, ANSI C50.41).

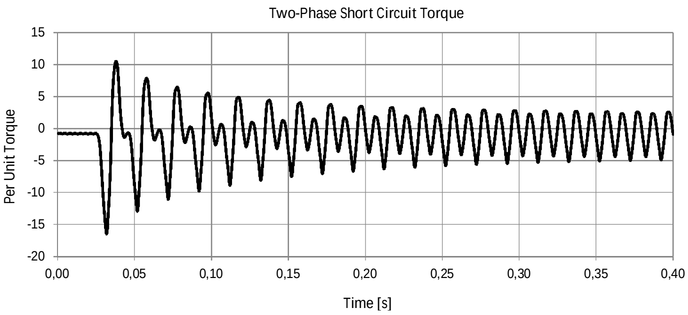

Figure 7. A Typical Two-Phase Short-Circuit Torque in the Air-Gap of a Motor.

Analysis and Design Guidelines

The steady-state excitations seem to be negligible from the torsional dynamics point of view. The most important excitation for the dimensioning of drive-train components is usually the two-phase short-circuit. The maximum air-gap amplitude of this fault condition is typically 5-15 times the rated torque of induction and form-wound synchronous motors. The maximum torque of permanent magnet motors is remarkably lower, typically 2-3 times the rated torque. The motor manufacturer provides usually this maximum amplitude for each motor.

It is significant that the excitation frequencies of main short-circuit faults occur at one and two times the grid or supply frequency, i.e. 60 Hz and 120 Hz. This means that the adverse effect of these transients can be avoided by adjusting the natural frequencies far enough from these frequencies. This will be discussed in the next section.

One remarkable challenge is the DOL start of large synchronous motors. Sometimes, the design of drive-train components is based on the limited number of starts during the lifetime of the motor.

A premature reclosure may generate a very strong transient air-gap torque. The main frequency component of this torque depends on several parameters and cannot be predicted easily. Thus, the risk for drive-train failure is large when a premature torque appears. Thus, the recommended approach is to apply an adequate reclosure control.

SHAFT-END TORQUE

Electrical calculations provide the air-gap torque in steady-state operating, starting and fault conditions. Usually, it is not reasonable to apply the air-gap torques directly to the dimensioning of the drive-train components and driven machine. A more refined dimensioning is based on the dynamics of the drive-train system. As an example, we consider in this section the dimensioning torques of the coupling.

A Simple Two-Inertia System

We assume that the drive train can be simplified to a simple two inertia system: electric motor – coupling – driven machine. This system can be presented by the formula (Ehrich 2004)

$$\frac{I_1 I_2}{I_1+I_2}\frac{d^2\psi}{dt^2}+B_{12}\frac{d\psi}{dt}+K_{12}\psi=\frac{I_2}{I_1+I_2}T_1(t) \tag{6}$$

where I_1 and I_2 are the mass moments of inertia for the motor and driven machine, K_{12} is the torsional stiffness between the inertias, B_{12} is the viscous damping coefficient, ψ is the angle of twist in the connecting shaft, and $T_1(t)$ is the air-gap torque. The amplitude of the shaft end torque, $\hat{T}_{12}$, due to the harmonic air-gap torque, $T_1(t)=\hat{T}_1\sin\omega t$, can be written in the form

$$\hat{T}_{12}=\hat{T}_1\cdot\frac{I_2}{I_1+I_2}\cdot\frac{1}{\sqrt{\left(1-\tau^2\right)^2+\left(2\zeta\tau\right)^2}} \tag{7}$$

where τ is the frequency ratio between the exciting frequency divided by the undamped natural frequency of the torsional system, and ζ is the damping ratio.

Inertia Factor

Equation (7) shows that the shaft-end torque is obtained from the harmonic air-gap torque by multiplying it with two terms: the inertia factor (= the load inertia divided by the total inertia) and the dynamic amplification factor. Figure 8 shows the effect of the ratio I_1/I_2 on the shaft-end torque. It can be seen that the quasi-static shaft end torque is always lower than the air-gap torque. For constant acceleration or deceleration of the drive the frequency ratio is zero and thus the inertia factor gives directly the amount of the shaft-end torque.

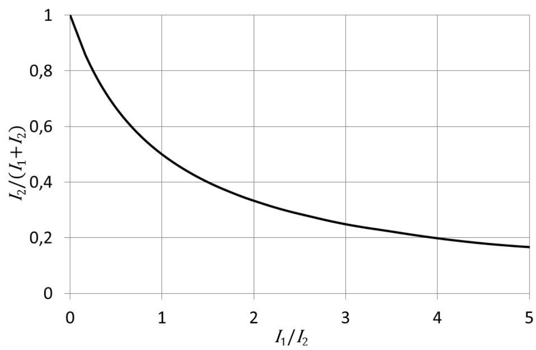

Figure 8. Effect of Ratio I_1/I_2 on the Inertia Factor.

Dynamic Amplification Factor

Figure 9 shows the effect of the dynamic amplification factor on the shaft-end torque. The excitation frequency is assumed to be 60 Hz.

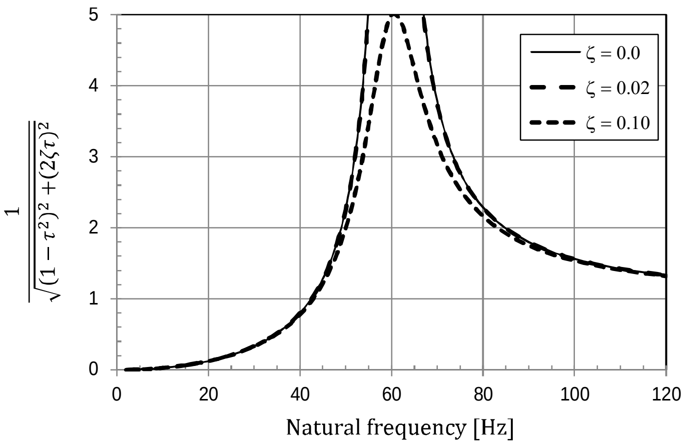

Figure 9. Dynamic Amplification Factor as Function of Natural Torsional Frequency with 60 Hz Excitation Frequency.

Figure 9 shows that the dynamic amplification factor has a remarkable effect on the shaft end torque. Usually, the torsional natural frequency is clearly lower than the supply frequency. Thus, the dynamic amplification is lower than one, e.g. with torsional natural frequency 20 Hz we get for the dynamic amplification factor 0.125. The effect of damping outside the resonance is minor.

Equation (7) gives the shaft-end amplitude for the harmonic excitation. The amplitudes of harmonic excitations in rated operating condition are very small. The recommended design values are typically 1 – 2 percent (API 684) and thus the shaft end torque oscillation will be very small assuming that the excitation does not coincide with the resonance frequency. Usually, the shaft-end torque due to electrical faults is more significant for the dimensioning of the coupling and other drive-train components. Next, we consider the two-phase short-circuit as an example of electric faults.

Two-Phase Short-Circuit

A simplified form of a two-phase short-circuit can be given by the formula (DIN 4024)

$$T_{\mathrm{sc}}(t) = 10T_0\left(\mathrm{e}^{-t/0.4}\sin\omega_1 t - \tfrac{1}{2}\mathrm{e}^{-t/0.4}\sin 2\omega_1 t\right) - T_0\left(1-\mathrm{e}^{-t/0.15}\right) \tag{8}$$

where T_0 is the rated torque of the motor and ω_1 is the angular grid or supply frequency. This transient excitation can be used to evaluate the dynamic amplification factor of two-phase short-circuit loads. An updated version of Equation (7) including the effect of the transient excitation can be written in the form

$$\hat{T}_{12,\mathrm{sc}} = \hat{T}_{1,\mathrm{sc}} \cdot \frac{I_2}{I_1+I_2} \cdot \frac{A_{\mathrm{sc}}(\tau,\zeta)}{\sqrt{\left(1-\tau^2\right)^2+\left(2\zeta\tau\right)^2}} \tag{9}$$

where $A_{\mathrm{sc}}(\tau,\zeta)$ is the factor for the short-circuit loading. Figure 10 shows this term multiplied by the dynamic amplification factor as a function of torsional natural frequency for a 60 Hz supply.

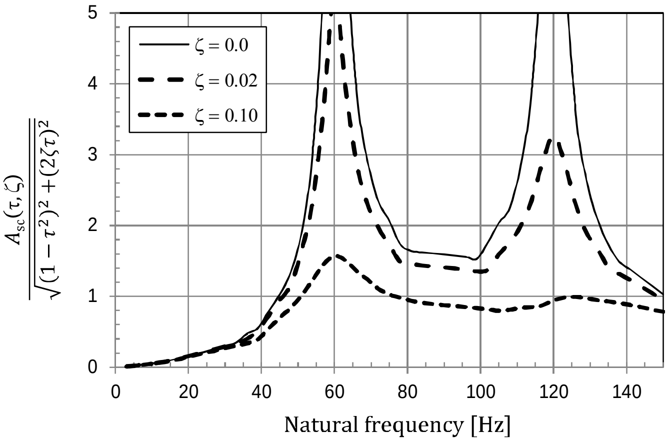

Figure 10. Amplification Factor for Transient Short-Circuit Load (DIN 4024) as Function of Torsional Frequency of the Lowest Mode with 60 Hz Grid Frequency.

By comparing Figures 9 and 10 we can observe that the amplification factor of the transient excitation is lower than that of the harmonic excitation up to about 100 Hz. When the natural frequency of the torsional system is higher than 100 Hz, the twice-line component excites this natural frequency. Figure 10 shows clearly that when the natural frequency is clearly lower than the supply or grid frequency, the amplification factor is less than one.

Other Loads and General Drive-Trains

The simple calculation example above was based on the two-inertia model of the drive train and only the harmonic and short-circuit excitations were studied. However, this simple example is representative for general drive-train systems. First, a two-inertia model with a torsional spring can be extracted from a general drive-train system. The motor inertia is, in practice, without exceptions one of these two inertias. Thus, the two-inertia model represents adequately the lowest torsional mode. Second, the fault conditions and starting are the main excitation sources. All of these loads are transient by nature. These transient loads occur at line and twice-line frequency. Thus, the lowest excitation frequency occurs at supply frequency. Third, the inertia ratio, presented in Figure 8, is determined typically by the application and drive-train arrangement. Thus, the inertia ratio may only decrease the shaft-end torque. Fourth, and finally, the transmission of all these transient excitations can be eliminated by designing the lowest natural mode clearly below the grid or supply frequency. The amplification factor is less than one when the frequency ratio is about 0.75, i.e. the natural frequency is 45 Hz with 60 Hz supply frequency.

It must be added two exceptions for the main excitation frequencies. The starting of salient-pole synchronous motors excites all the frequencies between zero and twice-line frequency and the premature breaker reclosure may generate low frequency excitations. The first one is unavoidable but the vibration amplitude can be attenuated by accelerating (or decelerating) the system as quickly as possible over the resonance frequency. The best way to handle the premature breaker reclosure problems is to resort to adequate reclosure control methods.

Analysis and Design Guidelines

The air-gap torque is determined by the electric system and the transient torque values are large compared to the rated torque. The dimensioning torque is the shaft-end or coupling torque which usually differs from the air-gap torque remarkably. There are two main parameters determining the ratio of these two torques: a) the inertia ratio, and b) the amplification factor.

The inertia ratio is usually determined by the application. The gear ratio must be included to the calculation of an equivalent inertia. The quasi-static shaft-end torque is always smaller than the air-gap torque and can be much smaller if the inertia ratio is large (Figure 8).

The amplification factor is a function of frequency ratio. The amplification factor can be decreased clearly by decreasing the frequency ratio. The frequency ratio is the ratio between the natural frequency of the coupling mode and the supply frequency (Figure 10).

LEVEL OF ANALYSIS MODEL

The example calculations presented above have been based on simple models. The design of actual drive-trains requires torsional models capable to represent the real behavior of the system. There are various alternatives to model the electromagnetic part of the system.

Electric Loads Only

The torsional analysis of a drive-train including an electric motor can be carried out using various approaches. The simplest approach is to model only the mechanics of the motor rotor. This means the modeling of stiffness and inertia and application of modal damping ratio for global modes. The electromagnetic loads can be applied either as a part of harmonic or transient analysis.

Electromagnetic Stiffness and Damping

A slightly more refined analysis can be carried out by including the electromagnetic stiffness and damping into the model. These values are dependent on the operating condition and oscillation frequency (Holopainen et al. 2010) and they must be calculated beforehand either analytically or numerically. This approach is suitable particularly for stead-state harmonic analyses.

Equivalent Circuit Model

One step forward is to model the motor with an equivalent circuit model. This is an analytic model describing the main behavior of a motor. The input variables of this model are the frequency and voltage. This equivalent circuit model is coupled to the rotordynamics by the air-gap torque and angular velocity. The angular velocity describes the rotational speed of the rotor and torsional vibrations. This equivalent circuit model can be used both for steady-state and transient analyses. The main challenge seems to be the coupling of the mechanical rotordynamic model to the electrical equivalent circuit model. The rotordynamic model is represented as a system of second order differential equations while the equivalent circuit model is described by the first order system. Usually, the rotordynamic model is transformed into the form of first order system and the models are combined directly. Another challenge is related to the input parameters of the equivalent circuit model. These parameters can be calculated by analytical or numerical methods based on the electrical design of the motors. Thus, the motor manufacturer has a key role if a reliable model is needed. It can be added that the control software of frequency converters includes the equivalent circuit model as a motor description.

Drive-Train Model Included in Electromagnetic Motor Model (FEM)

The most advanced model type is a fully integrated electromechanical rotor model. In this approach the electromagnetic behavior is modeled by Finite Element Method (FEM) and the mechanics of the drive-train is included as an integrated part of the calculations. This approach is computationally very onerous, because the electric system is non-linear and the calculations are usually carried out in time-domain. Thus, this approach is currently applicable mostly for research and development purposes.

CONVERTERS

Operational Principles

With frequency converters it is possible to control motor speed with very low losses. Thus considerable savings of energy can be gained compared to fixed speed drives with vane or other throttling control of the flow. Another advantage is smooth start of high inertia loads.

Figure 11 shows the structure of a variable speed drive system (VSDS). Note that there are many other terms and acronyms used in the literature, for example Adjustable Frequency Drives (AFD), Adjustable Speed Drives (ASD), Variable Frequency Drives (VFD), etc.

The converter adjusts the speed and torque of the AC motor by changing the frequency and magnitude of the voltages and currents fed to the motor. The attainable torque and power of a motor with rated 1 per unit current is shown in Figure 12 as a function of the speed in addition to the motor’s magnetic flux and voltage.

proportional to the speed squared, the decreasing motor torque and increasing load torque usually limit the maximum speed of the drive not to be far above the nominal speed. However, the motor is usually slightly over-dimensioned and thus, if necessary, it is possible to use field weakening range to increase the production of the equipment beyond the value that is attainable by a direct on line operation of the motor.

Main Types

The basic structure of a frequency converter is shown in Figure 13. The input section contains the devices such as circuit breakers, supply transformers and harmonic filters that may be needed for the AC grid connection.

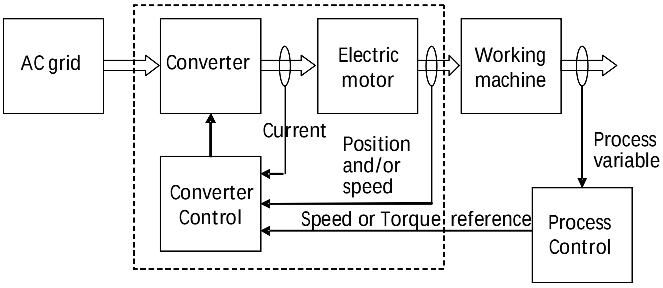

Figure 11. Structure of a Variable Speed Drive System.

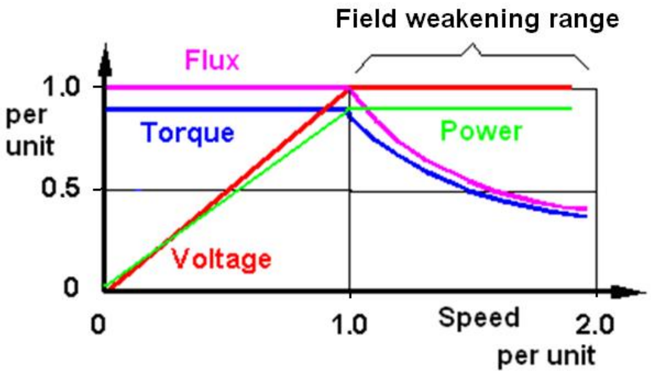

Figure 12. Attainable Torque and Power of a Typical Induction Motor Drive with Rated 1 Per Unit Current as a Function of the Speed. Magnetic Flux and Voltage of the Motor are Also Shown.

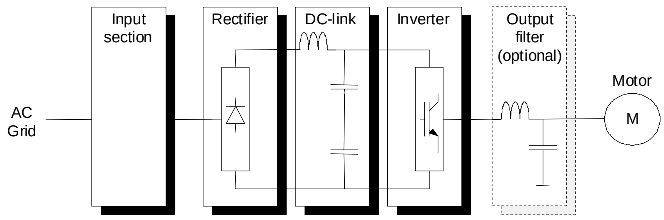

Figure 13. Structure of a Variable Speed Drive System.

The mechanical power P, voltage U, current I, speed n and torque T are related with the following approximate per unit Equation (10)

$$P = \eta U I \cos\varphi = nT \tag{10}$$

where η is the efficiency and cos φ is the power factor. Thus if the motor's efficiency is 0.95, power factor is 0.86 and the current has its rated 1 per unit value at nominal 1 per unit speed where both the voltage and the flux are 1 per unit, both the power and torque will be about 0.82 per unit*.

The linear increase of the voltage with the increasing speed means that the magnetic flux inside the motor is kept at its rated value. If the speed is increased above the nominal 1 per unit speed, the magnetic flux has to be decreased in order to prevent the motor voltage exceeding its rated value. Thus the speed range above the nominal speed is called field weakening range.

As can be seen from Figure 12 and the Equation E1 the power corresponding to the rated current will be constant in the field weakening range. Respectively the torque will decrease inversely proportional to the speed. As the load torque often is

* Note that a common convention with per unit base values is to define the rated mechanical power of the motor as 1 per unit. With this scaling the per unit current value corresponding to rated torque at rated speed will be higher than 1 per unit. In Figure 12 a different scaling is used for clarity.

The rectifier converts the AC from the grid into DC. The DC-link contains capacitors and/or inductors that are needed to smooth the ripple in the DC.

The inverter converts the DC into variable frequency and variable amplitude voltage and current that is fed to the motor. Sometimes an additional du/dt or sinus filter is needed to smooth the waveform fed to the motor.

The frequency converters can be categorized as current source and voltage source converters depending on the output, current or voltage, that the rectifier with the DC link provides for the inverter.

Current Source Converters

Two common current source converter drive schemes are shown in Figure 14. These are often called load commutated inverter (LCI) drives. The 12-pulse version is used at higher power ratings.

The rectifier controls the DC current and the inverter selects sequentially the phases of the motor where the DC current is fed. Thus phase currents of the motor consists of positive and negative blocks of DC current that have a duration of one third of the cycle. The 12-pulse version motor is wound in such a way that there is 30 degrees phase shift between the two three-phase winding sets.

Both the rectifier and the inverter of the LCI have thyristors, also known as Silicon Controlled Rectifiers (SCR), as their semiconductors. As can be seen, the basic building block is the six-pulse bridge assembly consisting of six thyristors. Six-pulse converter has two bridges and the 12-pulse converter has four.

Because thyristors cannot itself force their current down to zero after they have been turned on, the motor has to be capable to provide reactive power for this purpose. Thus LCI drives usually drive wound rotor synchronous motors.

Alternative to LCI is the Current Source Inverter (CSI) that has Insulated Gate Bipolar Transistors (IGBT) or Integrated Gate Controlled Thyristors (IGCT) instead of thyristors and thus can control all AC motor types.

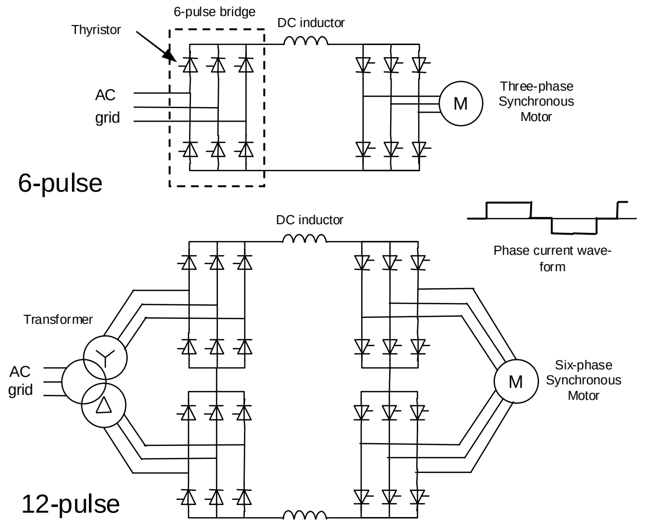

Figure 14. Six- and Twelve-Pulse Load Commutated Current Source Converter Drives and Their Ideal Phase Current Waveform.

Voltage Source Converters

The most common voltage source converter type is shown in Figure 15. The rectifier is an uncontrolled diode bridge that feeds essentially constant DC voltage to the inverter that has IGBTs with anti-parallel connected diodes.

In each phase only one transistor can conduct at any time. If the upper transistor conducts, positive voltage is fed to motor phase. Negative voltage is connected when the lower transistor is conducting. Because motor phase can have only two voltage levels this type of inverter is called two-level inverter.

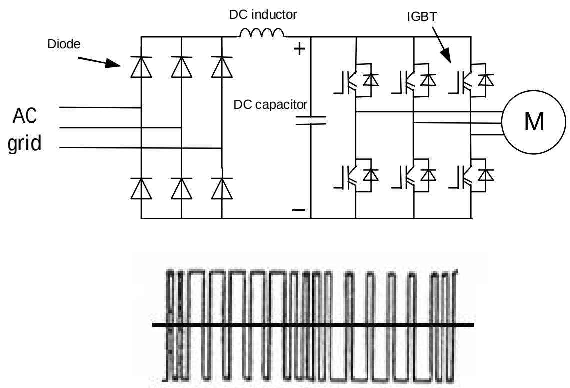

Figure 15. Two-Level Voltage Source Converter Drive that has Diodes in the Rectifier and IGBTs in the Inverter. Example of the Phase Voltage Waveform Is Also Shown.

The IGBTs can be turned on and off any time using their gate input. Due to this it is possible to feed induction motors in addition to synchronous motors. Moreover, instead of single pulse per half cycle it is possible to produce multiple voltage pulses with varying duration. By controlling the average value of the voltage pulse train by sinusoidal variation of their durations the phase currents of the motor can be made approximately sinusoidal. This control principle is called Pulse Width Modulation (PWM).

Two-level converters are almost exclusively used to drive low voltage motors up to 690 V rating. The power range extends to several megawatts.

For higher powers and medium voltage motors multilevel converters are used. The first multilevel converter was three-level neutral point clamped (NPC) converter, Figure 16, that was introduced to 3.3 kV multi-megawatt drive applications in the mid-eighties. It had Gate Turn-Off (GTO) thyristors instead of IGBTs as semiconductor switches. In addition to IGBTs IGCTs are used today in NPC drives.

As the name implies, the phase voltage of a three-level inverter can have three voltage levels: minus, zero and plus. This naturally means that it is easier to achieve good sinusoidal motor currents. Another advantage is the reduction of the steps in the phase voltage to half of the DC voltage that decreases the stress caused to motor winding insulations.

Another three-level converter scheme is the Neutral Point Piloted (NPP) that is also shown in Figure 16.

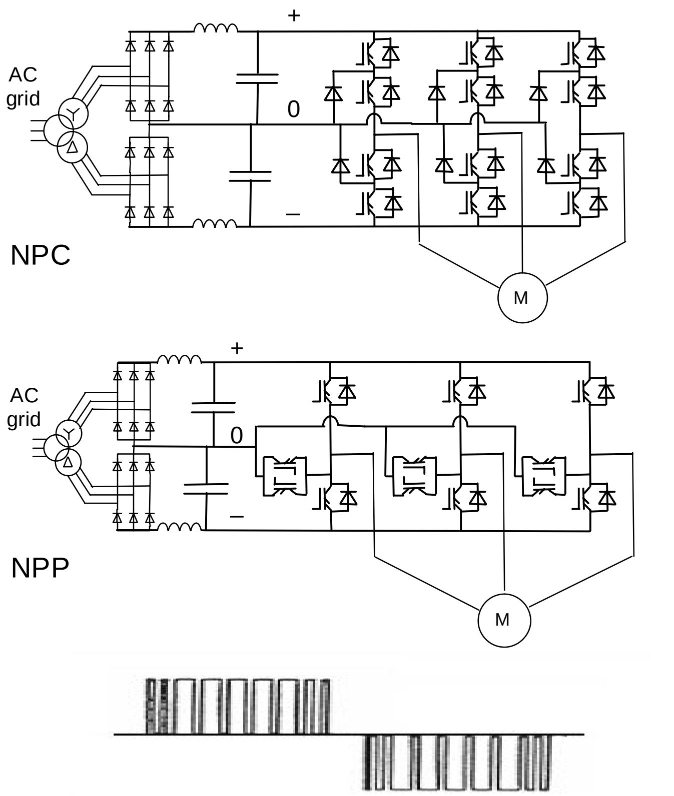

Figure 16. Examples of the Three-Level Voltage Source Converter Drives that Have Diodes in the Twelve Pulse Rectifier and IGBTs in the Inverter. Example of the Phase Voltage Waveform Is Also Shown. NPC = Neutral Point Clamped, NPP = Neutral Point Piloted.

For more than three levels the Cascaded H-Bridge (CHB) scheme shown in Figure 17 is often used. As can be seen, it consists of several converter cells that are fed separately from their transformers. The name H-bridge refers to the shape of the two phase inverter section of the converter cell. The voltage between each cell's output terminals can be minus DC, zero or plus DC voltage depending on which IGBTs are conducting. By cascading the cells in their outputs more combinations are obtained. The phase voltage of the motor fed by the five-level converter shown in Figure 17 can levels -2DC, -DC, 0, +DC and +2DC voltage.

More information about different converter topologies can be found for example from Kazmierkowski et al. (2011).

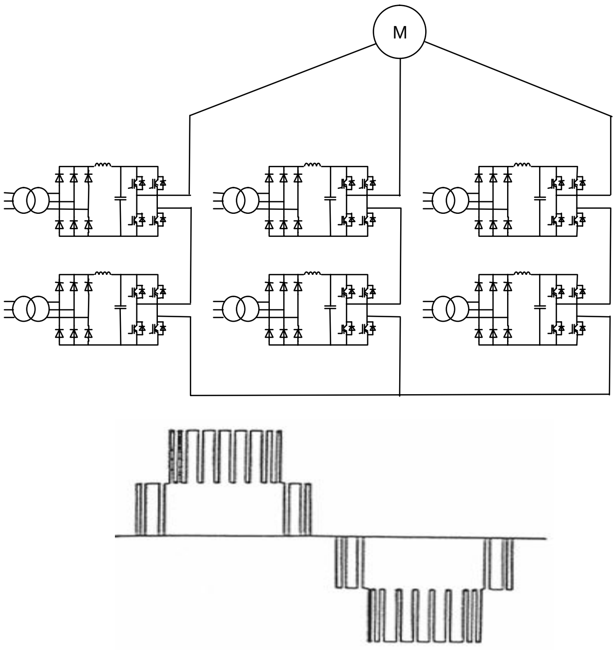

Figure 17. Five-Level Cascaded H-Bridge (CHB) Converter and Example of its Phase Voltage Waveform.

ELECROMAGNETIC TORQUE OF CONVERTER-FED MOTORS

Instead of the three phase voltages or phase-to-phase voltages a common visualization of the converter voltage output is with vectors. The IGBTs of the two-level inverter shown in Figure 15 has the functionality of changeover switches as shown in Figure 18. These switches can be set in eight ways. Two of these cause motor voltage to be zero. Thus two-level inverter can produce one zero and six non-zero vectors. Multi-level converters have more switches and thus more combinations and voltage vectors. For example, three level converter can produce 18 non-zero vectors in addition to zero vector.

When the output voltage of the inverter is fed to a three-phase motor, magnetic flux will develop inside the motor. The flux in the stationary motor frame, the stator, is called stator flux. The direction and speed of the movement of the stator flux vector tip can be influenced by the selection of voltage vector, see Figure 19. If a zero voltage vector is applied the stator flux will not change. Non-zero voltage vector will move the flux vector tip to the directions shown in the figure. Thus applying a proper sequence of voltage vectors the flux vector can be rotated as desired.

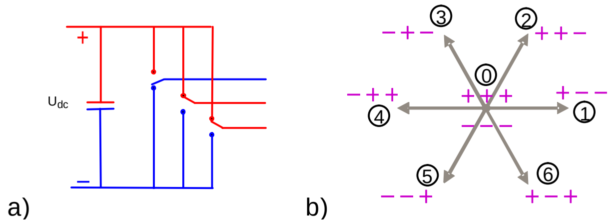

Figure 18. a) Two-level voltage source inverter represented as a set of switches, b) Voltage vectors that the inverter can output.

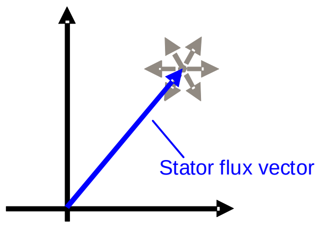

Figure 19. Influence of the Voltage Vectors to the Stator Flux.

If the motor is an induction motor, the rotating stator flux will induce currents to the short-circuited rotor winding fixed to the shaft. These rotor currents produce another flux, the rotor flux, that will rotate inside the motor. Due to the inductances and resistances of the rotor winding the rotor flux will follow sluggishly the changes in the stator flux.

The mechanical torque produced by the motor is proportional to the triangular area between the stator and rotor fluxes, see Figure 20. If the stator flux vector is rotated with the speed of the shaft both vectors will be on top of each other and the area between them, the torque, will be zero. Rotating the stator flux faster than the rotor is turning causes the rotor flux to lag behind the stator flux. Thus accelerating torque is produced. If the stator flux vector is rotated slower than the rotor is turning the stator flux will be lagging the rotor flux and thus the torque will be braking the motor.

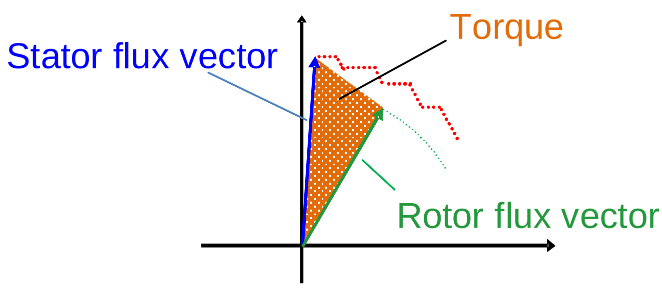

Figure 20. Torque Is Proportional to the Area of the Triangle between the Stator and Rotor Fluxes. The Path that the Tips of the Counterclockwise Rotating Vectors Have Traveled Is Shown with Dotted Lines.

Synchronous motors have either permanent magnets or externally fed rotor windings that produce the rotor flux. Thus they are able to produce torque also when the rotor rotates at the same speed than the flux. In order to change the motor torque only a temporary deviation of the stator flux vector's rotation speed from the rotor speed is necessary.

From Figure 20 one can see that torque can be increased also by increasing the flux magnitude. However, the magnetic saturation of the iron makes it impractical to use flux densities above 1 - 1.5 tesla range. Thus there is an optimum value for the flux magnitude around which the flux can be kept by optimized sequence of the voltage vectors as shown in Figure 20.

There are three basic ways to control the motor by frequency converters: Scalar control, rotor oriented vector control and direct torque control (DTC). The following subsections will give an overview on each of them.

Scalar Control

Scalar control, often referred as "constant volts per hertz control" is the simplest control method. The sequence of the voltage vectors of the inverter is produced by a control block called pulse width modulator. The transistors of the inverter are switched in such a way that both the average magnitude of the voltage vector and its rotational speed is proportional to speed command of the motor up to the rated speed. In this way the magnitude of the stator flux vector is approximately constant up to the rated speed and weakened above that as shown in Figure 12.

In its basic form the scalar control does not need any feedback from the motor. The only input is then the speed reference. Thus it is also called open loop control. In practice at least the motor phase currents are measured and used to compensate the resistive voltage drop of the motor windings that tends to decrease the flux at low speeds. This function is commonly called IR compensation (IR = current I multiplied by resistance R). The magnitude of the motor currents can further be used to roughly compensate the induction motor speed variation due to the slip of the motor.

If higher speed accuracy is needed a speed sensor (often called tachometer) and speed controller can be added. This kind of control is called closed loop scalar control. However, the control dynamics of the scalar control in general are not very good and thus its ability to actively damp torsional resonances is limited.

Rotor Oriented Vector Control

The rotor oriented vector control adds a model of the motor to the control system, Figure 21.

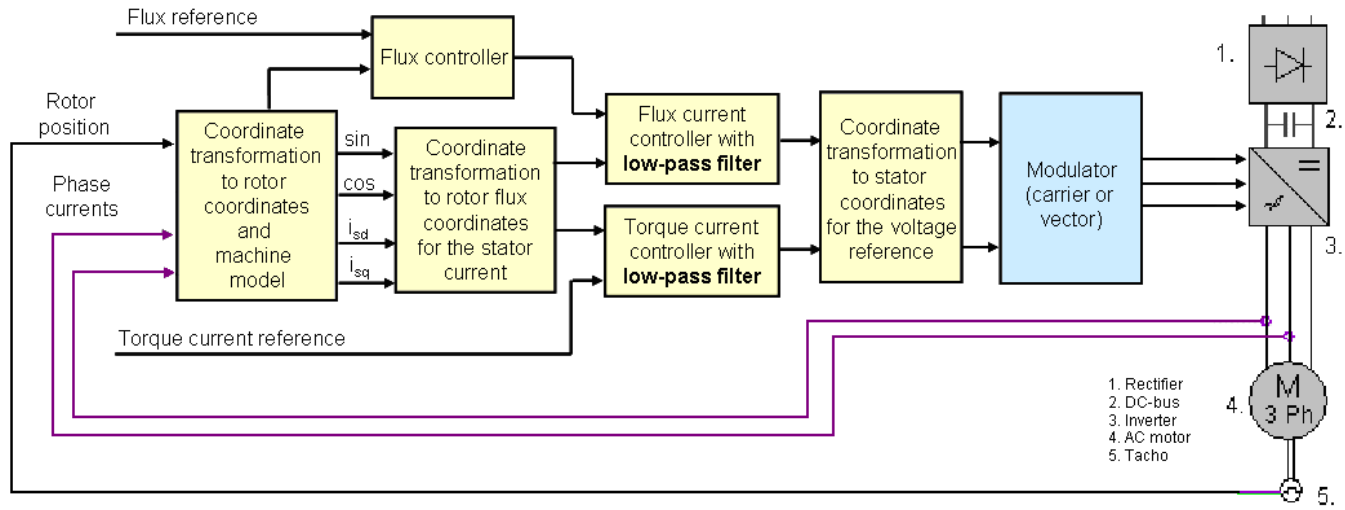

Figure 21. Rotor Flux Oriented Vector Control.

Motor model is used to estimate the torque and flux producing current vector components from the measured phase currents. These components are controlled to be equal to their references by current controllers that are usually PI (proportional and integral) type. The torque current reference is usually calculated by the speed controller that is not shown in Figure 21. The flux reference is defined by a function block from the motor speed.

The outputs of the current controllers are the stator voltage vector reference components. The pulse width modulator block is used to define a sequence of voltage vectors that has the average value equal to the reference vector.

The motor model has the rotor time constant as parameter. This time constant is the ratio of the rotor inductance to rotor resistance. These values depend on the magnetic saturation and temperature of the rotor that makes the accurate torque control challenging. Many rotor time constant estimation algorithms have been developed in order to improve the situation.

The motor model requires also the position of the rotor. However, the zero position can be arbitrary. Thus an incremental encoder (also known as pulse tachometer) is commonly used. Typically this kind of sensor gives few thousand pulses per revolution. By counting the pulses the rotor position can be calculated. Moreover these pulses can be used to calculate an estimate of the motor speed and thus a very accurate speed control can be achieved that is often capable to actively damp the torsional resonances.

However, if the accuracy and performance requirements of the speed control are not very high and there is no need to operate the drive close to zero speed for longer periods the pulse tachometer can be omitted and the rotor position and speed can be estimated instead. It is important to note that although there is then no speed sensor in the motor there is a feedback from the estimated speed calculated by the control system. Thus this is called sensorless closed loop control.

Direct Torque Control

Direct torque control (DTC) has also a motor model. The model estimates the motor torque and stator flux. However, DTC is quite immune to motor parameter variations in a wide motor speed range because the flux and torque estimations of the control system have only one and easily measurable parameter related with the motor, namely the stator resistance. Thus it is possible to drive many applications without speed or position encoders using estimated speed as actual value. An incremental encoder (pulse tachometer) is needed if continuous braking near or at zero speed is necessary.

Direct Torque Control has two hysteresis control loops, Figure 22. The flux controller keeps the stator flux magnitude in a narrow hysteresis band around the reference value. When the flux tries to escape from the hysteresis band a request for increase or decrease flux is issued.

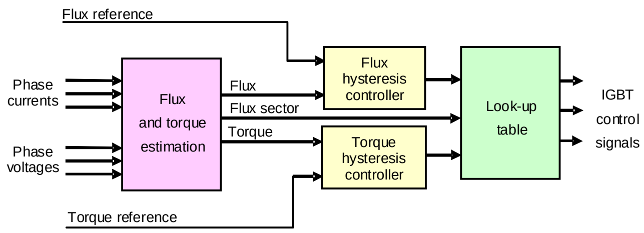

Figure 22. Principle of the DTC Control. (Courtesy of ABB)

The other hysteresis controller controls the torque in a similar manner in order to keep it within a narrow hysteresis band around the reference value.

The IGBT control signals are defined directly using a look-up table that has the requests of the flux and torque controllers as input. Thus if flux and/or torque requests are set a new voltage vector is produced that will bring the flux and torque back to their hysteresis band in shortest time.

A system based on hysteresis control does not need a separate pulse width modulator because the voltage vectors and thus the pulse widths are defined by the controller itself. Due to the compact control algorithm very fast computation, typically less than 25 microseconds, is possible in the control processor.

Because DTC control will always use the voltage vector that moves the torque fastest towards the reference, the control reaction to a speed or torque error is very fast. The response is limited only by the leakage inductance of the motor and the voltage reserve, that is, the difference between the DC link voltage and the peak voltage induced in the motor windings.

Due to the fast torque control DTC suits well for the damping of the torsional resonances. However, the difference between the DTC and PI current control performances is in practice quite small.

Control of Current Source Inverters

The control of the current source inverters (CSI) and load commutated inverters (LCI) differs from voltage source inverters (VSI) described above. In VSI drives the inverter takes care of the whole motor control and the rectifier can be an uncontrolled diode bridge. In CSI and LCI drives the rectifier controls the current that flows in the DC link. The inverter only selects the motor phases where the DC current is injected.

Due to this difference CSI and LCI drives are usually vector controlled and cannot be used to drive several parallel connected motors.

From the dynamical point of view the main difference lies in the speed with which the motor current can be changed. In VSI drives the DC link capacitor allows the current (and thus the torque) to be changed very fast. The DC link inductor of the CSI and LCI drives restricts changes in the current and thus makes it difficult to change the torque rapidly.

CSI drives usually use pulse width modulation of the current pulses in the low speed range in order to obtain smoother torque. However, LCI drives rely on the commutation of the thyristors that is driven by the motor voltage, and thus depending on the pulse number of the converter there are either six or twelve commutation instants per one cycle and no way to do PWM. This further means that the control delays in the LCI drive are quite big especially at low speeds and for a six-pulse LCI.

Moreover, at lowest speeds the motor voltage is too low for commutation. Thus the DC current has to be briefly controlled to zero in order to turn the thyristors off in the inverter. The drawback of this pulsed operation at low speeds is the increase of the torque pulsation especially when a high startup torque is required.

CONVERTER RELATED TORSIONAL EXCITATIONS IN STEADY-STATE OPERATION

One of the insidious traits of torsional vibration in a VFD system is that one may not be aware of its existence until the system fails. Direct driven equipment usually has few outward indications of high torsional vibration. Mechanical systems that have a gear may give outward indications of gear vibration if the magnitude of torsional vibration is very large relative to the mean torque. Under these conditions “clattering” of the gear teeth may be heard as the gear teeth oscillate from a loaded to an unloaded condition. This clattering of teeth may also be heard at low speed during slow speed acceleration when the mean torque is very low. Torsional oscillations can additionally transform into lateral oscillations on the pinion gear.

Sources of Torsional Excitations

All VFDs rectify alternating line current (AC), 50 or 60 Hz, to direct current, (DC), and invert the DC to variable frequency AC current. The AC-DC-AC conversion adds excitation frequencies to the system. Most of these components have varying frequency that depends on the motor speed. Thus the excitation frequencies may be even higher than the frequency of the AC grid.

Figure 23 shows how the six-pulse diode bridge rectifies the AC voltage to DC voltage. The DC voltage forms as parts of the sinusoidal phase-to-phase AC voltage waveforms of which only one is shown for clarity. Thus the result is six voltage pulses per AC cycle. Figure 13 shows ideal situation where all three phase-to-phase voltages have equal magnitude. In practice there is always some unbalance in the voltages. The resulting DC voltage waveform has ripple not only six times the grid frequency but also at twice the grid frequency.

As explained before, voltage source inverter connects the DC voltage to the motor phases as a train of varying duration pulses. Because of the ripple in the DC voltage, the height of the pulses is not ideally flat but varies according the DC voltage variation. Thus the motor phase currents are influenced both by the durations of the pulses and their height variation. The motor torque is defined by the currents and as a consequence the torque ripple has components originating both from the rectifier and the inverter operation.

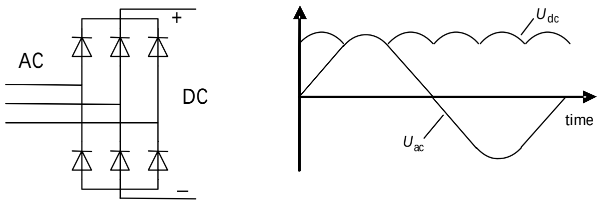

Figure 23. Six Pulse Rectifier and the Voltage Waveforms.

Naturally the DC-link filter capacitors and inductors will reduce the DC ripple but they cannot attenuate it completely. The same applies to the pulse width modulation methods of the inverter that usually strive to compensate the DC voltage variation by modifying the pulse durations accordingly.

What has been said above about voltage source inverters is also valid for the current source ones. DC current ripple is caused by the rectifier voltage and pulses of the current are fed to the motor by the inverter. The torque ripple depends on both the durations and heights of the current pulses.

Due to this duality of the sources of the torsional excitations one can divide the excitations into harmonic oscillations that are caused by the durations of the pulses and the interharmonic excitations that are due to the combined effect of the pulse durations and the variation of the pulse heights.

Harmonic Excitations

The phase current of an LCI drive shown in Figure 14 has two pulses per cycle. In a 6-pulse LCI there are three phases in the motor. Thus the motor will get six pulses per cycle causing the torque pulsation at six times the frequency the motor is fed. For example, if the motor is fed with 40 Hz, there will be torsional excitation at 6 x 40 Hz = 240 Hz. In addition to this frequency there will be multiples of six times the motor frequency in the torque as well. Thus there will be excitations at 12 x 40 Hz = 480 Hz, 18 x 40 Hz = 720 Hz, 24 x 40 Hz = 960 Hz, etc. with decreasing magnitude. Because of the excitations being at integer multiples of the motor supply frequency these excitations are called harmonic excitations.

It is important to note that contrary to the motor speed the torsional excitation frequencies do not depend on the number of the poles of the motor.

Using pulse width modulation (PWM) the number of pulses per cycle can be increased. Thus the torque ripple frequencies are shifted to higher frequency range and smoother torque is obtained. The decrease of the torque ripple by PWM is further enhanced with voltage source inverters (VSI) due to the additional smoothing of the phase currents by the inductances of the motor windings.

Though the VSI drive’s torque is very smooth when compared with LCI drives, some excitations at six times the output frequency and its multiples exist due to inaccuracy of the power semiconductor turn-on and turn-off delays. Further, the measurement errors in the motor current sensors of the drive may cause small excitations at the output frequency and twice the output frequency. Typically VSI harmonic air-gap torque components having frequencies less than 60 Hz are smaller than 1 percent of the rated torque of the motor even without speed controller damping.

The main difference between current source inverter (CSI) or LCI drives and VSI drives is the behavior of the magnitude of the torsional excitations as a function of the load. With CSI and LCI drives the magnitude depends on the load torque but with VSI drives it is quite the same at no load and at full load. Thus a VSI drive usually has much lower excitations than a LCI drive at rated torque but may have higher excitations at no load.

Interharmonic Excitations

The interharmonic or non-integer harmonic excitations are caused in a VSI drive by the non-idealities in the PWM modulation and the intermediate DC voltage's ripple. Similar phenomenon is also present in CSI and LCI drives due to ripple in the DC current.

Typically a 60 Hz fed six-pulse rectifier produces DC voltage ripple harmonics at 6 x 60 Hz = 360Hz and multiples of it. However, there are usually components at all multiples of 120 Hz (2x60 Hz) because of the asymmetry of the AC mains supply's three-phase system. If the supply frequency is 50 Hz, the components will be at multiples of 100 Hz, the most dominating being at multiples of 300 Hz. If a 12-pulse rectifier is used, the most dominating frequencies will be multiples of 600 Hz or 720 Hz with 50 or 60 Hz supply, respectively.

The intermediate circuit has a filter that smoothes the drive’s DC voltage or current. LCI and CSI drives have an inductor that filters the DC current and VSI drives have a capacitor that filters the DC voltage. Thus the higher frequency components are attenuated more than the lower ones. In practice the most important DC harmonics are 120 and 360 Hz for 60 Hz AC supply for both six-pulse and twelve pulse rectifiers. The level of the 360 Hz component may, however, be rather small with twelve pulse rectifiers, if the transformers feeding the series connected bridges have the same secondary voltages and same short circuit voltages. In practice there are, however, always some differences due to manufacturing tolerances.

As has been previously stated, the output voltage of a VSI drive consists of samples taken from the intermediate DC voltage. Thus the ripple in the DC voltage is sampled as well. Similarly the CSI or LCI drive’s inverter samples the DC current with its ripple. This sampling causes intermodulation frequencies to appear in the phase voltages and currents and thus in the torque. A general formula for integer and interharmonic frequencies is

$$f = \left| kf_1 + mf_{\text{line}} \right| \tag{11}$$

where f_1 is the inverter's output frequency, f_{line} is the incoming line frequency (50 or 60 Hz), k is a positive integer and m is a positive or negative integer. The most important excitation frequencies are plotted in Figure 24 as a function of the frequency of the inverter.

As stated before, usually the highest excitations of the most critical torsional natural frequencies are obtained for integer harmonic components with $m = 0$ and $k = 6$ or $k = 12$, that is six and twelve times the output frequency of the inverter. However, as can be seen from the figure, these are typically excited at quite low speeds and thus are not a problem if the drive is not running continuously at low speeds.

The excitations with non-zero m values are related with the DC voltage or current ripple. Most important of these interharmonics are obtained when $k = 0$, 6 or 12, and $m = -2$ or -6. Especially for $k = 6$ and $m = -6$ case when the drive output frequency approaches the line frequency, the excitation frequency reduces and has the potential to excite the first torsional mode. Thus the potential for torsional vibration associated with interharmonic distortion occurs at speeds normally above 80 percent of rated speed, where the average torque is much higher and operation sustained for long periods. Interharmonic distortion above 80 percent speeds then becomes a concern for high cycle fatigue, see for example Tripp et al. (1993) and Hudson (1992). Methodologies for analyzing systems with VFDs were presented by Hudson (1996) and Grgic et al. (1992).

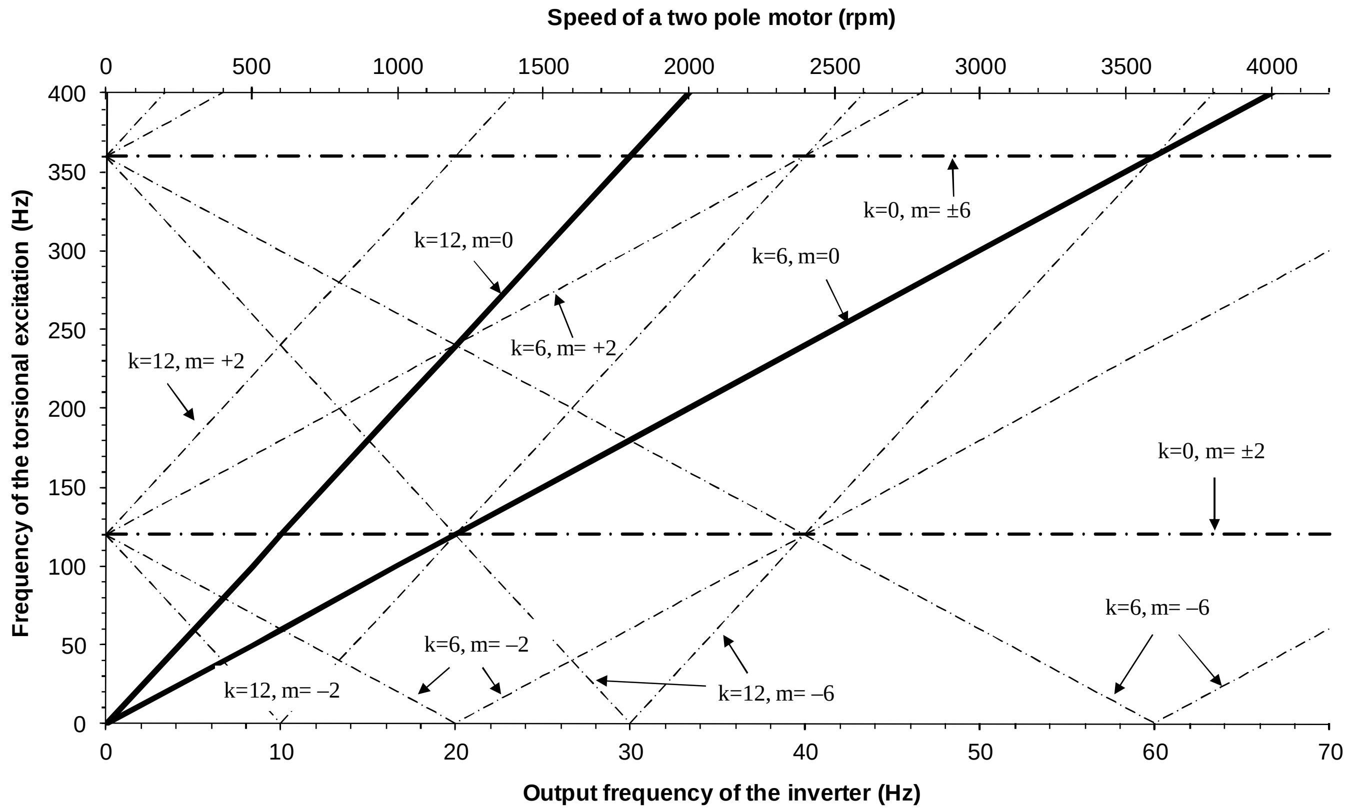

Figure 24. Most Important Harmonic and Interharmonic Excitation Frequencies of Motor Torque Represented on the So-Called Campbell Diagram when the Supply Network's Frequency Is 60 Hz.

From Equation (11) and Figure 24 one can notice that excitations for $k = 0$ have the same frequencies as the DC intermediate circuit's ripple harmonics have. It is advisable to avoid designs with ill-damped first natural mechanical resonance frequency near these frequencies.

With LCI drives it is thus quite straightforward to check the situation with harmonic and interharmonic torsional excitations using a Campbell diagram similar to Figure 24 and information of the excitation magnitudes obtained from the drive manufacturer.

The situation may be little more complicated with VSI drives although, due to the PWM modulation and smoothing of the current by the motor inductances, the excitation amplitudes generally are often so small that they can be neglected. If the pulse width modulation (PWM) used in VSI is synchronized with the output frequency, that is, the switching frequency is an integer multiple of the output frequency, all excitations are according to Equation (11). However, due to the limited maximum switching frequency of the power semiconductors, the PWM pulse pattern has to be changed at several speeds. These changes may cause small transients in the torque that temporarily cause torsional vibrations. The magnitude and decay of these transients depends on the damping of the mechanical system and the ability of the drive's speed control to damp the mechanical resonances.

Another PWM scheme is to use constant switching frequency in the whole speed range. Then the PWM pattern is not synchronized with the output frequency. This causes additional interharmonic components in the torque. This phenomenon can sometimes be seen on the measured Campbell diagram as sidebands around the main excitation frequencies. In general these components are quite small in magnitude if the switching frequency is in the kilohertz range. Thus most often these can be discarded in the torsional analysis.

Figure 25 is an example of a measured shaft torque spectrum variation of an accelerating VSI drive. These kinds of diagrams are often called as waterfall or 3D-diagrams. The spectrum measured at the start of acceleration is at the bottom of the figure and the one measured at the end of the acceleration is at the top. The frequency of the vibration can be read as the distance from the left. Thus Figures 24 and 25 are similar in their layout though the vibration frequency range of the diagram in Figure 25 is 0...250 Hz and the corresponding axis is horizontal. The magnitude of the vibration is presented in Figure 25 by the heights of the peaks in the figure.

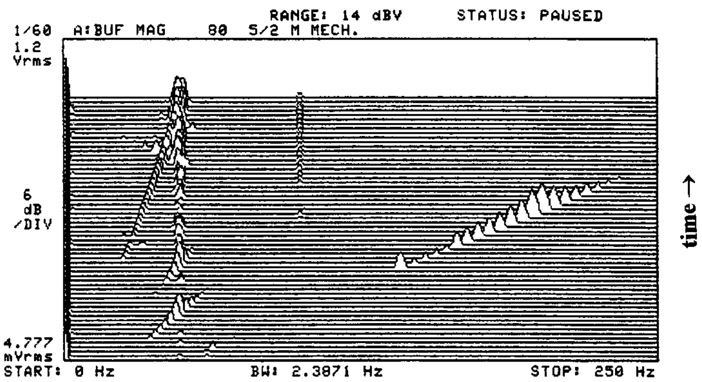

Figure 25. Example of the Shaft Torque Spectrum Variation when the Measured VSI Drive Was Accelerating. The Frequency of the Supply Network Was 50 Hz. Frequency Range of the Spectrum Analysis Was 0...250 Hz.

The accelerating torque can be seen in Figure 25 as a ridge on the left at the zero frequency. A vertical ridge at 47 Hz is due to the first torsional resonance frequency. The resonance frequency was quite high because the measurement was made with a back-to-back test setup where two motors were connected with each other. Third vertical ridge at 100 Hz is due to the intermediate DC circuit's ripple at the same frequency. This component becomes detectable only at higher speed range because magnitude of the DC ripple increases with the loading of the rectifier bridge. The main diagonal ridge is caused by the $k = 6$, $m = 0$ component, though there is a small component at the output frequency ($k = 1$, $m = 0$) and multiples of it as well. This component is typically caused by the offset error in the current measuring equipment of the drive. Some traces of interharmonics with non-zero m can be seen crossing the 47 Hz mechanical resonance near the top of the diagram.

Example

In order to clarify the relationships between output frequency of the inverter, torque harmonic excitation frequencies, mechanical resonance frequency, pole pair number and rotational frequency the following example will be considered: The torsional resonance frequency f_{res} is 20 Hz and line frequency f_{line} is 60 Hz. The potential for torsional vibration caused by harmonic torques occurs for $k = 6$ at 20 Hz / 6 ≈ 3.3 Hz, that is, at relatively low speed. If the motor has two poles (one pole pair), the corresponding speed will be only about 3.3 Hz · 60 rpm/Hz ≈ 200 rpm. If the motor has four poles (two pole pairs) the speed is about 100 rpm. Higher order harmonics would match the resonance frequency at drive output frequencies and speeds that are even lower than this. Therefore the potential for torsional vibration is transient in nature as these speeds are crossed rapidly during startup acceleration or during shutdown deceleration. Further, at such low drive output frequencies the mean torque applied to the drive train components is often quite low, and therefore, the mean stress in drive train components is low even for CSI and LCI drives that have harmonic components proportional to the mean torque.

From Figure 24 and Equation (11) one can see that the interharmonic torque component associated with $k = 6$ and $m = -6$ will be at 20 Hz when the inverter's output frequency f_1 is either $\left(-mf_{supply} - f_{res}\right)/k$ = 56.67 Hz or $\left(-mf_{supply} + f_{res}\right)/k$ = 63.33 Hz. As stated before, the pole pair number has nothing to do with these frequencies of the excitation. However, the pole pair number affects the speed of the motor on which the 20 Hz resonance excitation takes place. If the pole pair number is one, the speeds will be 3400 rpm and 3800 rpm, plus minus the slip. If the pole pair number is two, the speeds will be 1700 rpm and 1900 rpm, plus minus the slip.

Random Modulation Schemes

The modulation schemes described earlier have distinct excitation frequencies that are related with the motor speed and the frequency of the supplying AC grid. This is due to the cyclic behavior of the grid and the converter control system where the sequence of events repeats itself constantly.

However, the pulse pattern of the PWM does neither have to be synchronized with the output frequency nor the switching frequency has to be constant. This kind of PWM scheme where the pulse pattern does not repeat itself is called random modulation. This produces excitations where the spectrum does not have any distinct components. Instead, the spectrum is continuous function similar to acoustic white noise spectrum.

White noise spectrum means that there are always some excitations at every frequency but because the peaks are missing the excitations are significantly lower than with a spectrum having only few distinct frequency components.

Moreover, due to the random nature of the excitations they do not cause similar amplification due to torsional resonance as a distinct harmonic frequency. White noise may temporarily cause vibration to increase but then the exciting frequency component disappears or its phase angle changes to a damping one and the oscillation goes down again. Thus in most cases the stress in the mechanics does not develop into harmful level.

One feature of the white noise excitation is that the drive system is more robust against changes in the mechanical system properties. For example change in the torsional resonance frequency due to stiffening of the elastomer in the coupling does not cause much change in the stress level because the excitation has the same low value at any frequency. Similar stiffening with excitation spectrum having distinct harmonic and interharmonic components may cause severe consequences if the resonance frequency comes near to one of the components.

Commonly used converter control method that produces random modulation is hysteresis control. In this control scheme the controlled variable, either motor current or torque, is maintained within a narrow hysteresis band around the reference value by selecting new voltage vector, see Figure 18, every time the controlled variable tries to escape the hysteresis band. Direct Torque Control (DTC) shown in Figure 22 is one example of such systems.

A white noise spectrum can be also approximated by pseudo random modulation where the switching frequency of the frequency converter power semiconductors is changed according to a long sequence of random values.

Kocur and Muench (2011) have presented a method to model white noise excitations. However, it is important to notice that the amplitude values obtained by spectrum analysis of the measured white noise torque depend on many factors. The most influencing one is the length of the sample sequence used in the analysis. For example, if the duration of the analyzed torque measurement sample sequence is 0.5 seconds the analysis gives frequency components with 1/0.5 s = 2 Hz spacing. If the analyzed sequence is longer, the components are more densely spaced. For example an analysis based on 2 seconds sequence will give 1/2 s = 0.5 Hz resolution. With a spectrum having distinct frequencies, same magnitudes for excitation components are obtained for both 0.5 second and 2 second sequences. However, for white noise the magnitude of the components is the smaller the more densely they are calculated.

Another related factor is the use of windowing of the data, which has an impact on the amplitude of each calculated component. Thus there seems to be a need to standardize the method used to determine the torque spectrum.

Design Principles

If PWM is used in the converter the torque in the mechanical system is often smooth enough to allow unlimited speed range for most applications. However, it is a common practice to check the torsional vibration properties of the mechanics when a drive for a high power system is designed.

Usually the torsional analysis is first made to determine the mechanical resonance frequencies and resonance damping. It is important to note that the taking account only the shaft line may not be enough when the motor is not firmly on the ground. Thus care has to be taken in marine and similar arrangements where the supporting structures may not be rigid enough to be neglected.

The amplitudes of torsional vibration in the mechanical system are determined using the data of excitation frequencies and amplitudes of the converter driven motor. In some cases the load torque is not smooth and has also to be taken into account in the stress analysis. If the analysis indicates excessive stress it may still be possible to lower the amplitudes to safe level by actively damping the resonances by a suitable structure of the speed controller and dedicated tuning.

For very demanding applications a computer simulation of the complete system may be considered necessary. The computer model may take into account the nonlinearities and nonidealities of the electric, magnetic and mechanical components. Moreover, the simulation can be used to optimize the dynamics and stability of the speed control.

SPEED CONTROL AND RESONANCE DAMPING

Feedback Loop Structure

In a typical process control application, it is possible to identify the main blocks shown in Figure 26.

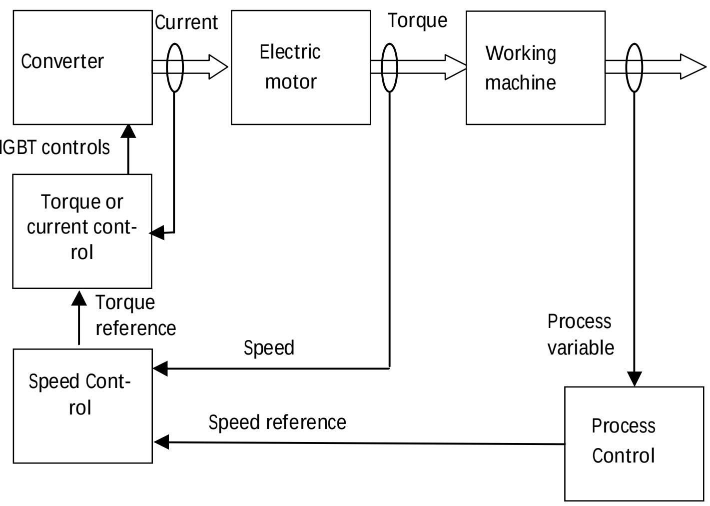

Figure 26. Block Diagram of Drive System Control.

The process controller maintains the controlled process variable, for example pressure, at its reference value by changing the speed reference accordingly. The speed controller of the converter adjusts the torque reference in order to keep the speed equal to the reference speed. Finally the torque is controlled by the torque or current controllers to be equal to the reference given by the speed controller.

Attenuation of Low Frequency Torsional Vibrations

The mechanical system of the driven equipment is usually quite complex consisting of several rotating bodies connected with shafts. Example of a three inertia system is shown in Figure 27.

The moments of inertia and the torsional stiffnesses of the system define the torsional resonances of the drive train. For the speed controller the most important thing is how the air-gap torque produced by the magnetic forces between the stator and rotor of the motor will influence the speed of the motor. This can be visualized by a transfer function of which an example is shown in Figure 28. The parameters of the system are:

- $J_1 = 5.13 \cdot 10^6$ lbm-in.^{2} (= 1500 kgm^2)
- $J_2 = 3.25 \cdot 10^6$ lbm-in.^{2} (= 950 kgm^2)
- $J_3 = 14.0 \cdot 10^6$ lbm-in.^{2} (= 4100 kgm^2)
- $K_{12} = 566 \cdot 10^6$ in.-lbf/rad (= 64 MNm/rad)
- $K_{23} = 257 \cdot 10^6$ in.-lbf/rad (= 29 MNm/rad)
- $c_{12} = 180 \cdot 10^3$ in.- s-lbf/rad (= 20 kNm·s/rad)
- $c_{23} = 80 \cdot 10^3$ in.- s-lbf/rad (= 9 kNm·s/rad)

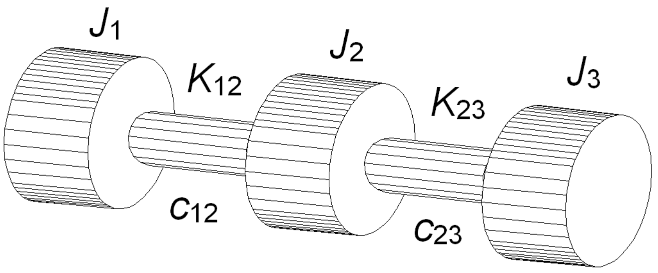

Figure 27. Example of a Drive Train Consisting of Three Moments of Inertia and Two Shafts with their Torsional Stiffnesses and Dampings.

The torsional resonance frequencies are seen as peaks at about 20 Hz and 60 Hz. In addition to these there are frequencies around 12 Hz and 50 Hz where the transfer function has very low values. These are called antiresonance frequencies.

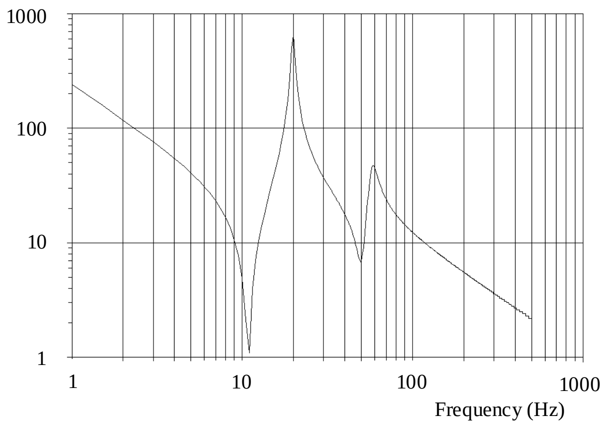

Figure 28. Example of a Transfer Function from Air-Gap Torque to Motor Speed for a Three Inertia Rolling Mill System.

The low value of the transfer function at antiresonance frequencies means that it is very difficult for the motor to oscillate the motor speed with these frequencies. As a consequence the shortest possible response time that the speed control can reach is defined by the lowest antiresonance frequency. As a rule of thumb the response time is about the inverse of that frequency. Thus in the case shown in Figure 28 the achievable response time of the speed control is about 1/12 Hz ≈ 80 ms. It is important to realize that this value is defined only by the mechanical system and thus if there are special requirements for the control dynamics it has to be taken into account in the design of the mechanical system.

Regarding the excitations, it is important to notice that the high frequency excitations well above resonance frequencies are attenuated (torque amplification less than one) by the mechanical system as can be seen from Figure 29. Thus especially in high power systems excitations above 100 Hz can often be neglected.

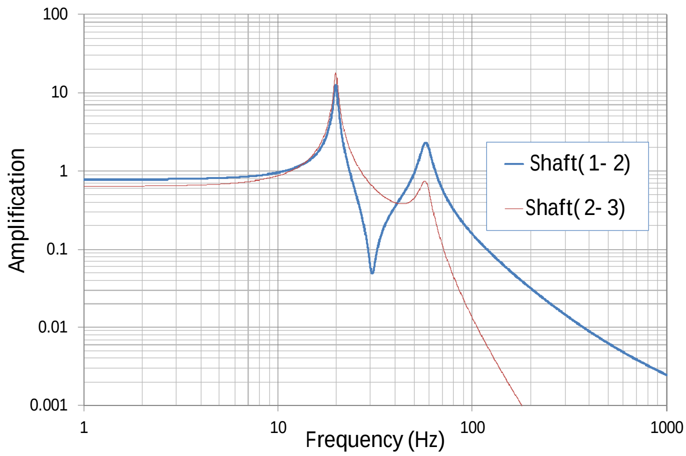

Figure 29. Example of the Amplification of the Air-Gap Torque Excitations Calculated for the Shafts of a Three Inertia Rolling Mill System with Same Parameters than in Figure 28.

The VFD control can be used to actively damp the mechanical oscillations occurring on the shaft, provided that they are within the VFD control bandwidth. Typically the lowest natural torsional resonance frequency has to be lower than half of the torque control bandwidth in order to successfully damp the resonance, see Annex B of IEC 61800-4 (2002). With DTC typically more than 70...100 Hz control bandwidth is achieved. Vector control can reach almost the same.

Usually the damping can be done by tuning the speed controller properly. Myszynski and Deskur (2010) presented rules for tuning the commonly used PID speed controller to damp resonances in a three moments-of-inertia mechanical system. Often the limiting factor in the ability to damp torsional resonances is the delay caused by the speed measurement. It should also be noted, that in-appropriate speed controller tuning can also further contribute negatively to torsional damping.

Some VFD vendors propose also the usage of real-time observers to estimate the torque or speed at different sections of the mechanical system and thus be able to react on those. This practice is quite common in high dynamic applications such as rolling mills and metal manufacturing in general as exemplified by Vauhkonen (1989) and Valenzuela et al. (2005), but is not yet used much in the Oil & Gas business.

With the torsional damping control, the reasons of having a very slow speed control are losing part of their motivations. By allowing a faster speed controller, the system is able to react faster to load changes. In this way, the oscillations after load or speed changes are quickly attenuated and continuous excitations coming either from the motor or the driven load are prevented to be amplified by the mechanical resonances.

Design Principles

Above considerations result in the following best practices:

- Tune the process control in such a way that it is slower than the speed control
- Have the speed control tuned to be slower than the VFD torque control. This assumes big importance especially in big LNG trains.
- Make sure that the speed reference or torque reference given by the process control is free of noise and other interfering signals because these will be reproduced in the motor torque and may excite mechanical resonances
- Make sure that the speed sensor is properly installed. For example, eccentric mounting causes erroneous shaft rotation frequency component in the measured speed.
- Further, make sure that lateral vibrations of the motor are not affecting the speed sensor operation. Any erroneous fluctuation in the measured signal may become amplified in the speed control and excite torsional resonances. It is naturally possible to attenuate this kind of interfering signals by filtering but then it may be impossible to use the speed control to damp resonances.
- Sometimes problems with speed sensors can be avoided by using the VFD in torque controlled mode and integrating the drive directly in the process control. The VFD can still be used for speed monitoring and fast protection.

SHORT CIRCUIT LOADING WITH VFDs

If a direct-on-line connected AC machine is short-circuited, it will produce high torques at the line frequency and twice-line frequency.

Short circuits frequently occur in the supply network. Thus it has been the practice to avoid mechanical resonance frequencies at these frequencies when the motor is directly fed from the supply grid. However, when the motor is supplied with a frequency converter, its input frequency will vary and it may not be possible to keep mechanical resonances outside the frequency range of continuous operation.

Luckily the short circuiting of converter fed motors is very rare because the converter isolates the motor when there is a fault in the supply network. Design principles have been dealt with in detail in the paper by Grgic et al. (1992). Usually, short circuits are considered not to pose any significant threat to the mechanical system when frequency converters are used.

CONCLUSIONS

This paper presented the modeling, analysis and design of electric drives as a part of torsional systems. The subject was reviewed from a broad perspective aiming to serve rotordynamic professionals with background outside of electric drives. The analysis and design guidelines were proposed throughout the paper. As a conclusion, the main points are repeated here.

One of the important observations is that only the coupling mode, and to some extent, modes with their oscillation node within the driven equipment are prone to electromagnetic excitations. The high-frequency excitations cancel effectively out due to the shape of the higher modes within the electric motor.

The second observation is that the electromechanical interaction, i.e. electromagnetic stiffness and damping, may affect significantly the coupling mode and its damping ratio, if the load inertia is large compared to the motor inertia.

The third observation is that the dimensioning torque is usually the short-circuit torque. The transient air-gap torque of this fault can be easily calculated. The corresponding shaft-end torque is determined by the inertia ratio and the frequency ratio. The inertia ratio is determined by the application and cannot be easily changed. By decreasing the frequency ratio, i.e. the frequency of the coupling mode divided by the supply frequency, the shaft end loading can be decreased significantly.

Variable frequency drives bring considerable advantages from a process control point of view. In addition, they improve the conditions for the motor and drive-train during starting and are able to protect the torsional system from disturbances in the grid, such as unbalances and short circuits. However, due to their inherent operating principle, the torque provided by variable frequency drives contains harmonics which need to be taken into account during the system design of drive-trains, especially for high power systems

Some variable frequency drives modulation schemes produce torques with distinct excitation frequencies, while others produce a spectrum which is a continuous function similar to acoustic white noise. In both cases, the designer of the shaft line shall take into account the data of excitation frequencies and amplitudes of the converter driven motor and use them to feed the torsional mode shapes of the mechanical system. Doing so, one has to keep in mind, that closed-loop control can change the torsional damping of the rotating equipment. This can be used to actively damp torsional resonances, however similarly, unintentional negative damping needs to be avoided by choosing control parameters accordingly.

For very demanding applications, especially in the high-speed-high-power range, closed loop electromechanical simulations may bring additional confidence in the system design as they consider nonlinearities of electric, magnetic and mechanical components.

The different possibilities to study the design leave the system integrator with the challenge to identify the appropriate level of detail, which he needs to successfully design his system. He will have to revert to the component suppliers, not least the drives and motor supplier, in order to get the necessary support in this decision making process, and then later the required input data to his design.

NOMENCLATURE

A_{sc}	= Factor for short-circuit loading
B_{12}	= Viscous damping coefficient
D	= Diameter
G	= Shear modulus
I	= Current
I_1	= Motor mass-moment of inertia
I_2	= Load-machine mass-moment of inertia
K_{12}	= Torsional stiffness coefficient
L	= Shaft section length

P = Power
T = Torque
T_{sc} = Short-circuit torque
T_0 = Rated torque
T_1 = Air-gap torque
T_{12} = Shaft-end torque
$\hat{T}$ = Maximum amplitude of torque T
U = Voltage
f_{exc} = Excitation frequency
f_{line} = Grid frequency, 50 Hz or 60 Hz
f_{res} = Torsional resonance frequency
f_1 = Motor supply frequency; output frequency of the inverter
k = Torsional stiffness; integer multiplier for inverter frequency
n = Speed
n_1 = Synchronous rotational speed
m = integer multiplier for grid frequency
p = Number of pole-pairs of the motor
s = Slip of the rotor
ζ = Damping ratio
η = Efficiency
τ = Frequency ratio
$\cos\varphi$ = Power factor
Ψ = Angle of twist
ω_1 = Motor supply frequency in radians

AC = Alternating current
AFD = Adjustable frequency drive
ASD = Adjustable speed drive
CHB = Cascaded H-bridge
CSI = Current source inverter
DC = Direct current
DOL = Direct on line
DTC = Direct torque control
FEM = Finite element method
GTO = Gate turn-off
IGBT = Insulated gate bipolar transistor
IGCT = Integrated gate controlled thyristor
LCI = Load commutated inverter
NPC = Neutral point clamped
NPP = Neutral point piloted
PI = Proportional-integral
PWM = Pulse width modulation
SCR = Silicon controlled rectifier
VFD = Variable frequency drive
VSDS = Variable speed drive system
VSI = Voltage source inverter

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