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TECHNICAL BULLETIN

Axiom Drive Technology Tutorial

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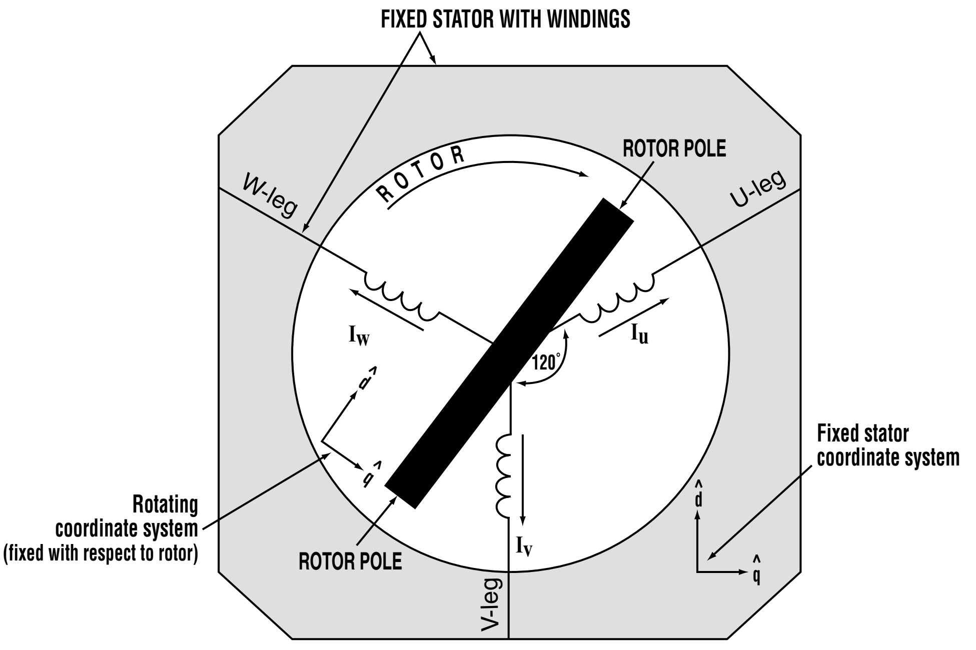
Currently, digital drive control for brushless 3-phase servo motors is based on the concept of supplying a 3-phase voltage, commutated sinusoidally with respect to the angular position of the rotor. Flux vector current control is the most efficient way to accomplish that. It involves taking encoder feedback to track rotor position, and independently processing control algorithms for both the torque-producing and waste current vectors to the motor. This could not even be accomplished before DSP technology. This paper will outline how flux vector control works.

The brushless 3-phase servo motor has become the predominant choice of automation system designers. Its superior torque and speed capability versus size, compared to the brush-commutated servo motor and stepper motor simply make it the best solution for most applications. Other advantages include mechanical reliability (no brushes to wear out), low rotor inertias, and better thermal properties. The brushless servo motor, when driven properly, can provide high torque over a considerable speed range, allowing machine designers high bandwidth control of acceleration, speed, and position.

Figure 1 demonstrates the principle of torque production in a brushless servo motor and its dependence on the relative orientation of the permanent magnetic field developed by the rotor magnets to the field produced by current in the stator windings. The motor configuration represented has two rotor magnetic poles and wye stator windings. 360 degrees of physical rotor travel represents one complete motor magnetic cycle for this configuration. A four-pole motor traverses a complete magnetic cycle every 180 degrees of rotor travel – a 6-pole motor every 120 degrees – and so on. Regardless of the number of rotor poles, the principles of torque production represented in Figure 1 apply.

Typical configuration of 2-pole, permanent magnet brushless servo-motor



NOTE: Current polarities I_U, I_V, and I_W are defined such that the resulting magnetic fields have the directions shown. In other words, positive u-leg current will produce a magnetic field vector oriented as shown by the arrow labeled I_U.

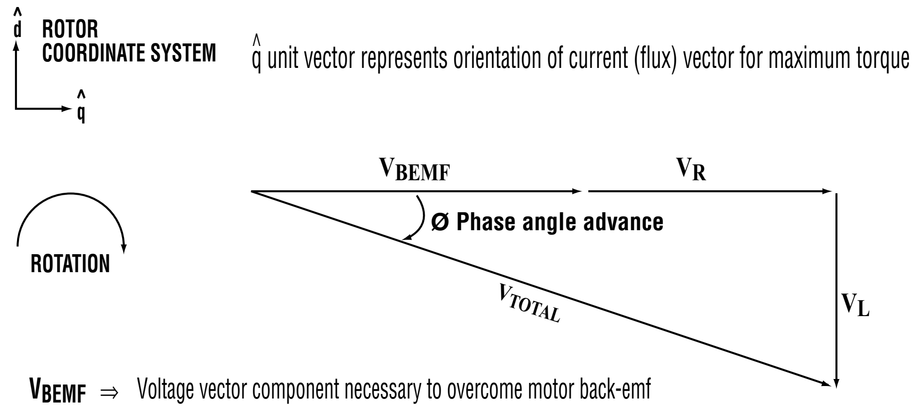
Figure 1

Looking first at the stator winding representation of Figure 1, the vectors **Iu**, **Iv**, and **Iw** represent the magnitude and direction of current flow in the respective **U**, **V**, and **W** windings. For our purposes, these vector orientations also represent the direction of the magnetic field vector produced by each current vector. Thus, a current of positive polarity in winding **U** will produce a magnetic field vector with the same orientation as **Iu**, and with a magnitude proportional to the magnitude of **Iu**. The vector sum of the three magnetic fields associated with the three individual winding currents forms a magnetic flux vector which can interact with the permanent magnetic field of the rotor to produce torque. In this article, the controlled magnetic field vector produced by the stator windings is understood to be equivalent to the current vector sum multiplied by a scalar value. The strength of this controlled field versus motor current and the strength of the rotor's permanent magnetic field in large part determine the torque constant of the motor, a useful parameter relating motor torque to winding current.

In order to produce the greatest torque with given stator and rotor magnetic fields, the stator field orientation must be perpendicular to the rotor field direction. A stator field vector that "leads" the rotor field vector by 90 degrees of motor magnetic cycle will produce the maximum positive torque for a given motor current vector magnitude. Likewise, a stator field orientated –90 degrees of magnetic cycle from the rotor field will produce the maximum negative torque. Any component of the stator magnetic field vector in parallel with the rotor field vector will produce zero torque, and thus the corresponding component of the overall stator current vector represents wasted current. Controlling the angle of the current vector (and thus the resultant stator field vector) to maintain the perpendicular condition for optimum torque production is generally termed "commutation".

Since brushless servo motors must be electrically commutated by the motor drive, it is obviously necessary for the drive to independently vary the voltage to each of the three legs of the motor windings to produce individual winding currents with a vector sum that will maintain its perpendicular relationship with the rotor field regardless of the rotor's position or angular speed. From AC induction motor theory, it is known that applying three separate sinusoidal voltage values phased 120 degrees apart (in time) will produce an overall resultant voltage vector of constant magnitude which will rotate at the sinusoidal frequency. This voltage vector will produce a rotating current vector, also of constant magnitude, for a three-phase, balanced load. It follows that sinusoidally varying the three leg voltages of a brushless motor will produce the same effect, i.e. a rotating stator current (field) vector can be produced. The principal of sinusoidal commutation of brushless motors involves creating this rotating vector through commutation with respect to the rotor field angular position within one motor magnetic cycle. If the angle of this commutation is varied directly with the angle of rotor field, and a phase offset added to achieve the perpendicular condition, then a stator field of a desired magnitude can be produced and forced to rotate so as to always interact with the rotor field to produce motor torque. In practice, there are several problems with this approach concerning the ability to dynamically control the stator field vector magnitude and direction at a high bandwidth and in conjunction with rapidly changing rotor

Vector components of commutated voltage at a given motor rotational velocity



V_{BEMF} ⇒ Voltage vector component necessary to overcome motor back-emf
— proportional to motor speed

V_R ⇒ Voltage vector component necessary to overcome resistive voltage drop in motor windings
— proportional to current

V_L ⇒ Voltage vector component necessary to overcome reactive impedance due to motor winding inductance
— proportional to current and motor speed

Ø ⇒ Phase angle advance — represents the angle by which the commutated voltage vector must lead the rotor $\hat{q}$ vector for maximum efficiency

V_{TOTAL} ⇒ Vector component of voltage necessary to produce properly phased current vector (and thus magnetic field vector) at a given motor speed. It is assumed that winding parameters are balanced

Figure 2

speeds. Figure 2 shows a simple representation of the primary voltage vector components necessary to produce the rotating current vector.

The voltage vectors in Figure 2 are referenced to a coordinate system that is fixed with respect to the rotor magnetic field direction, i.e. the coordinate system rotates with the permanent rotor field. The direction of the **q** unit vector corresponds to the direction of stator field vector necessary for maximum torque. As shown, to produce a stator current (field) vector of a given magnitude at the proper orientation, it is necessary to supply a voltage vector of proper magnitude and phase to overcome the back emf of the motor and the resistive and reactive impedance of the motor windings. As rotor speeds increase, the impedance of the winding inductance at the frequency of commutation increases, which means that the magnitude and phase of the applied voltage vector must be adjusted accordingly or the current vector will "lag" its intended position. Any deviation of the stator current vector from its desired angular position with respect to the rotor field reduces the torque produced. The total current supplied to the motor will remain the same, but as mentioned earlier, a component of the resultant current vector is "wasted". If the stator current vector (which we are equating with the stator magnetic field vector) is decomposed into two components, one in the direction of the q unit vector and the other in the direction of the d unit vector, then the **q** component can be thought of as torque-producing, and the **d** component as waste. With inefficient current commutation, a significant portion of the total current supplied to the motor by the drive is utilized to create this **d** component. This current is then providing no useful torque but still causing heat in the motor windings and drawing on the drives current-sourcing capability. In addition, at a given drive current, the actual torque produced varies depending on the current phase error.

It is obvious that to produce accurate and high-bandwidth control of motor-torque, which is the fundamentally most important task of the drive, the non-torque producing portion of the current vector must be kept small and the torque-producing (**q**-vector) component varied according to the magnitude of the desired, or commanded, torque. A sinusoidally-commutated drive must continually calculate and supply a "phase-advance" to the commutation voltages as motor speed changes in an effort to maintain maximum efficiency. Looking again at Figure 2, this calculated phase-advance depends on the back emf of the motor, as well as the resistance and inductance of the windings. The back emf is proportional to rotor rotational speed and can be estimated, but there is considerable tolerance from motor to motor in the constant of this proportionality. Also, motor inductance and resistance values specified for a given motor typically have a tolerance of 10%, and winding resistance can easily change up to 40% as stator temperature varies over the full operating range of the motor. All these factors combine to reduce the accuracy with which the phase-angle advance can be calculated. Some newer-generation sinusoidal drives attempt to use higher feedback gains to reduce the effects of inductive "lag". But because the back-emf voltage vector is of considerable magnitude and changes with speed, the corresponding high voltages applied create considerable d-component currents (waste) with small commutation inaccuracies. All the afore-mentioned problems associated with sinusoidal commutation become more complex and detrimental under dynamic conditions of quickly changing motor load and velocity. In effect, because the sinusoidal drive does not have direct, high-bandwidth control of the torque-producing current vector, efficiency, torque bandwidth, and accuracy suffer, particularly in the higher-horsepower region of the

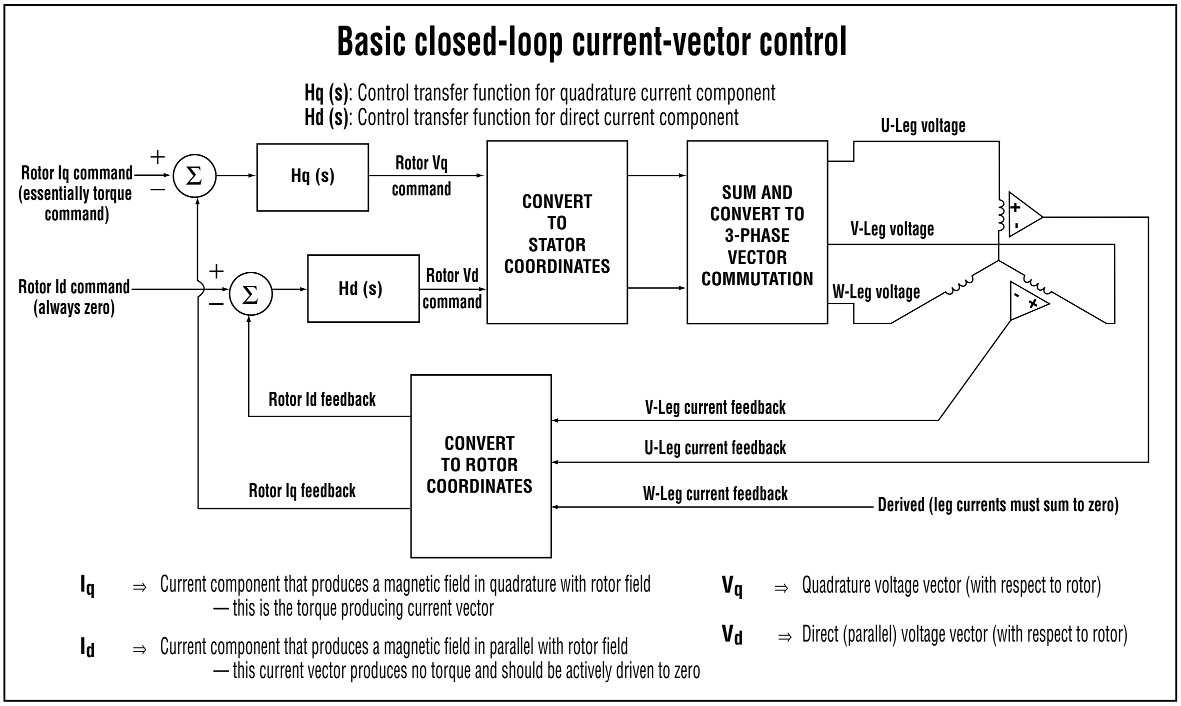


Figure 3

theoretical speed-torque envelope. This will produce higher-than-expected motor operating temperature for a given load and duty cycle, and limit accelerations at higher speeds.

At present, the best means to overcome the limitations of traditional, sinusoidally commutated drives in producing efficient and high-bandwidth control of torque involves performing the current control effort directly against both the **q** and **d** components of the total stator current vector. Referred to as "flux-vector control", this approach has been developed to allow smooth torque control of AC induction motors down to very low speeds. Figure 3 shows a simplified block diagram of this type of current control effort as applied to permanent magnet, brushless motors. The vector transformation functions necessary to implement this algorithm require encoder feedback so that rotor position can be tracked. The encoder feedback is not shown in Figure 3, which focuses only on the fundamental actions of the current vector control feedback loops. As shown, two separate current commands are provided. The **Iq** command corresponds to the desired motor torque-producing current vector magnitude. The **Id** command corresponds to the desired magnitude of the non-torque-producing current vector. The **Id** command will always be set to zero. The significance of this is that the "waste" current component in the motor windings is always being driven to zero under high-bandwidth, closed loop control. Thus, the **Id** control loop supplies voltages necessary to eliminate error in the angle of the total current vector under dynamic conditions.

The current vector commands are summed with feedback values defining the actual **q** and **d** current vector components. In order to provide these feedback values, two legs of motor current are sensed directly. The current in the third leg can then be derived, since the values of the three currents must sum to zero. A vector conversion is then performed to decompose the total stator current vector into two components—the **q** component corresponding to a current vector component which produces a magnetic field in perpendicular with the rotor field—and a **d** component corresponding to the current vector component which produces no motor torque. Since these vectors are defined according to a coordinate system that is fixed with respect to the rotor, encoder feedback is necessary to track rotor position. With these feedback values calculated and the corresponding two error signals derived, control effort can be applied to these two current vectors independently.

The characteristics of the load seen by each current loop are relatively easily controlled if sampling bandwidth is high enough. Execution of the entire current control algorithm at 10kHz allows both **q** and **d** current loop bandwidths of 1.5kHz + with good stability margins and minimal overshoot responses to a step input. The back emf voltage can be easily cancelled by providing integral gain and/or feedforward as part of the **q** current control. Once the perpendicular and direct voltage vectors are calculated, another vector transformation must be performed to calculate the equivalent output bridge switching patterns necessary to apply these voltage vectors to the stator windings. Again, this conversion requires knowledge of rotor position obtained with encoder feedback.

By independently processing control algorithms for both the perpendicular (torque-producing) and direct (waste) current vectors, the flux-vector method allows consistent and accurate control of motor torque with maximum efficiency under dynamic conditions of varying load and speed. The algorithm does not depend on velocity feedback, only on encoder positional feedback to track rotor position, which can be practically and accurately sampled at high execution rates. The net result is that both the torque bandwidth and horsepower of a motor-drive combination are significantly improved. Most noticeable are the gains in torque at higher speeds and under conditions of rapidly changing speeds. The efficiency gains allow motors to run at higher duty cycles while still operating within winding temperature limits. The calculations which accomplish this algorithm are much more mathematically intense and potentially time-consuming than those necessary for sinusoidal commutation. However, DSP microprocessors are now being produced which can process this type of algorithm at the rates necessary for top performance. Some of these microprocessors also incorporate peripheral hardware functions catered specifically to servo motor drives, such as quadrature-decoding, vectored PWM output logic, etc.

The previously described technology: flux vector current control and DSP technology are utilized in the Axiom series of brushless servo drives manufactured by Tol-O-Matic, Inc. Learn more about this exciting technology advancement at: www.tolomatic.com

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