SOURCE IMPEDANCE EFFECTS IN THE CONTROL OF INVERTER-INDUCTION MOTOR DRIVES

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INTRODUCTION

The development of static, adjustable frequency power converters has opened up new fields of application for alternating current (AC) motor drives. Probably the most important type of motor control is torque regulation since the torque controller forms the essential inner loop for most speed and position controllers. One of the more common types of AC drive systems used today basically consists of a voltage inverter supplying power to an AC induction motor. Such drives are being increasingly considered for applications where the extreme ruggedness of the motor together with its ability to withstand environmental contamination are desirable. A filter in the direct current (DC) supply to the inverter is necessary for satisfactory operation of the static inverter. However, the effect of this filter on the control characteristics of the drive system is usually neglected. Since applications which demand faster and more accurate regulators are increasing, it becomes necessary to evaluate the effect of the DC line filter on the drive control.

One example of such applications is the case of several drives supplied from a common DC rectifier source. The characteristics of the power system thus have a significant effect on the operation of the inverter and motor. Such a power system may consist of a rectifier supplied from 460 volts AC, thus a nominal DC potential of 600 volts is supplied.

Manuscript received 3/15/77. 1
A varying DC load is seen by the inverter during braking as a result of the continually changing receptivity of the DC bus to regenerated power due to the varying operating conditions of the other drives on the same DC source. The object of this paper is to present a method for analyzing the effect of varying DC filter impedance on the operation of an inverter-induction motor drive. The results of the analysis are presented to show how the DC filter impedance actually affects the design of the motor drive.

**SYSTEM DESCRIPTION**

The basic drive system to be studied consists of a DC to variable voltage, variable frequency AC voltage inverter using pulse width modulation for voltage control (Fig. 1). Associated with the inverter is the two-stage DC filter shown in Fig. 2. This filter serves to isolate the inverter from disturbances in the DC supply, as well as supply needed reactive current to the force-commutated inverter. Resistors $R_{B1}$ and $R_{B2}$ are used in conjunction with contactors $LB$ and $DB$ during regenerative and dynamic braking. The DC power source consists of a DC voltage $V_R$ in series with a source impedance represented by $L_R$ and $R_R$. The inverter supplies AC power to one or more AC induction motors which in turn provides the mechanical torque. Operation in both motoring and braking modes is required with good torque regulation and with no perceptible torque transients when the inverter changes operating modes. The inverter is sinusoidally pulse-width modulated at low frequencies so as to shape the AC voltage wave. At higher speeds, square wave operation is used for maximum efficiency. An intermediate transition mode of pulse width modulation that is not sine wave shaped is used for a smooth transition from the PWM mode to the square wave mode of operation. (1)

The desired motor operating conditions during motoring and braking are shown in Fig. 3. This figure shows torque, motor voltage, slip frequency and current as a function of speed. The motoring curve consists of three major portions: (1) a constant torque section to speed $V_1$; (2) a constant horsepower section from $V_1$ to $V_2$; and (3) a motoring portion at reduced flux from $V_2$ to the maximum speed. The torque limited section corresponds to the maximum current handling capacity of the inverter. The inverter operates in the PWM mode of operation during this region in order to vary the motor voltage for speeds from zero to $V_1$ in motoring and speeds from
zero to \( V_4 \) in braking. The constant horsepower section corresponds to the field weakening mode of operation of a DC motor and arises from the limitation on maximum voltage available from the inverter during motoring. The section of the motor curve from \( V_2 \) to maximum speed corresponds to operation of the motor at breakdown torque with the maximum available voltage from the inverter. Thus, the torque is inversely proportional to speed squared. In general, this section of the curve corresponds to the torque characteristic of a series DC motor although the curve for a DC motor falls somewhat below this curve because of the main field magnetic circuit saturation. The braking curve has two sections. The first section is a constant deceleration rate in the speed range from zero to \( V_3 \). The second is a constant power taper, which is the most practical characteristic for a power-limited drive system.

From Fig. 3 it is apparent that the motor control can be divided into three modes of operation; Mode I for speed of zero to \( V_1 \) in which slip frequency is held constant with the motor terminal voltage increasing linearly with speed; Mode II between speeds \( V_1 \) and \( V_2 \) in which voltage is held constant with slip frequency increasing with speed; and Mode III for speed \( V_2 \) to maximum speed for which both voltage and slip frequency are constant. Analogous modes also exist in braking operation. By proper sizing of the braking resistors, Mode III can be avoided in braking.

When in the PWM mode of operation (Mode I), the level of flux is regulated to a desired value. The level of flux in the motor is regulated at its maximum value in order to maintain operation of the inverter at the minimum possible switching frequency. Thus operation at any desired torque level within the limiting values shown is essentially accomplished by control of the motor slip frequency in Mode I.

Although Fig. 2 indicates only maximum motor and braking performance, steady-state operation at any torque value between these limits must be possible. However, problems arise during braking operation, since the effective source impedance can vary between wide limits due to the other loads on the DC bus also being variable. In extreme cases the power supply may be temporarily interrupted due to circuit interruptions. In addition, the DC power supply typically contains diode rectifiers which are not capable of accepting regenerated current during the braking mode of operation. The braking control system must be stable for any variation in line impedance
as well as be effective in limiting the peak DC voltage at the inverter terminals in the event of a sudden loss of the DC line.

**CONTROL STRATEGY**

Figure 4 shows a block diagram illustrating the essential features of a control scheme capable of meeting the requirements listed above. In effect, the inverter can be regarded as a power amplifier which converts the DC power to AC power for the motor. The two control inputs to this power amplifier are frequency and voltage which in turn are derived from the control system for the desired torque and flux levels. The signals are fed to the waveform generator which constructs a three-phase set of sine waves proportional to the required voltage and at the required frequency. The primary feedback loops are the motor current obtained from current transformers and the voltage obtained from sensing coils which measure the level of air gap flux in the motor.\(^{(2)}\) The outputs from these sensors are processed to obtain the motor flux and electromagnetic torque. Feedback of speed is provided to restart the system at approximately the correct frequency should power be lost but does not take an active part in the control system. A feedback of DC source voltage level is included to improve system stability when in the braking mode of operation. The two major control loops consist of flux regulation by the use of voltage control and torque regulation by means of frequency control. The flux regulator \((K_3/S)\) is an integrator in which the input is flux error and the output is equivalent to volts per Hertz. The integrator output is multiplied by stator frequency to obtain the desired motor voltage. Torque regulation is accomplished with an integral plus proportional controller \((K_1/S + K_2)\) for which the input is torque error minus the approximate derivative of the DC source voltage. The output of the control block is equivalent to motor slip frequency which is added to shaft speed to obtain stator frequency.

An additional control loop is included but not shown in Fig. 4. The added circuit replaces the torque regulator loop during overvoltage or undervoltage conditions and is used to limit the maximum inverter DC voltage during braking and the minimum inverter DC voltage during motoring. The control circuit senses DC voltage and adjusts the motor frequency to either limit the amount of regenerated current so that the maximum DC voltage of 1000 volts is not exceeded, or uses the motor as a generator to maintain a minimum DC inverter voltage so that system operation is continuous during a power interruption. This latter mode of operation is included to allow
electrical braking in the event of a DC power failure and to allow continuous system operation during dynamic braking.

**SYSTEM REPRESENTATION IN d-q COMPONENTS**

To study the dynamic behavior of the system of Fig. 1, the linearized equations of the system must be derived. For this purpose it has proved useful to develop equations for the induction machine expressed in a reference frame which rotates synchronously with the stator voltage vector.\(^{(3)}\) In this manner a balanced set of sinusoidal terminal voltages can be transformed to equivalent d-q voltages which are DC in the steady-state. As a result, the motor currents which flow in this reference frame also become DC. The equivalent circuit of an induction machine in the synchronously rotating reference frame is shown in Fig. 5. Note that since the orientation of the d- and q- axes are arbitrary, it can always be arranged so that use of the two axes (q-axis) is aligned along the voltage vector so that \(v_{qs}^e = V_s\) the amplitude of the voltage vector, and that the other axis component is zero \(v_{ds}^e = 0\).

When the motor supply consists of an inverter supply rather than a set of sinusoidal voltages, the same transformation procedure can be carried out. However, the problem becomes more difficult since the switching of the inverter must now be transformed to a synchronously rotating reference frame. Fortunately, it has been determined that the inverter voltage harmonics have a negligible effect on the dynamic behavior of the system.\(^{(4)}\) Hence, it is only sufficient to relate the DC components of voltage and current on the DC side of the inverter to the corresponding fundamental components on the AC side. It can be shown that when the inverter is a simple square wave type then, neglecting harmonics, the required relationships are\(^{(4)}\)

\[
v_{qs}^e = V_s = \frac{2V_I}{\pi} \tag{1}
\]

\[
i_{qs}^e = I_s = \frac{\pi I_I}{3} \tag{2}
\]

and

\[
v_{ds}^e = 0 \tag{3}
\]

where \(V_I\) and \(I_I\) are the DC side voltage and current and \(V_s\) and \(I_s\) are the fundamental amplitude of stator voltage and current.

In the case of sine-wave modulated PWM inverters, the relationship between DC and AC side variables becomes even more complex, since the control inputs to the inverter now affect the amplitude of the inverter
output voltage as well as the frequency. The most popular PWM modulation scheme consists of a comparison between a sinusoidal control signal proportional to the desired output voltage and a constant amplitude triangle wave. Summation of the triangle wave and the sine wave into a comparator produces a switching of one pole of the inverter. To reduce the harmonics to a minimum, the triangle wave is kept as high as possible—say, 400 Hz—when the output frequency is low. The sine-wave control signal and triangle reference then operate asynchronously. As the motor frequency increases, the ratio of the two frequencies decreases until the ratio reaches a specified minimum, at which point the two signals are synchronized. Further increases in AC voltage merely affects the width rather than the number of inverter pulses per pole.

A plot of the fundamental component of output voltage amplitude vs sine-wave control voltage input is given by the solid line in Fig. 6. For convenience, the input control voltage $V_s^*$ is plotted as a per unit of $V_n$ the nominal (rated) AC voltage for square wave operation at rated DC voltage. The output is given as a per unit of the maximum possible AC output voltage (i.e., square wave operation). The actual AC output voltage can then be written

$$V_s = f(V_s^*/V_n) \frac{2V_I}{\pi}$$

where $f(V_s^*/V_n)$ represents the functional relationship of Fig. 6.

It can be noted that before the sine wave is synchronized to the triangle wave ($V_s^*/V_n < 0.625$) the input-output relationship is linear. However, as the sine-wave amplitude continues to increase, it exceeds the amplitude of the triangle wave and two pulses are "dropped" from the output waveform and a sudden change in slope occurs. With a chopping ratio of six, three slope changes occur before the inverter reaches square wave operation. In addition, because of practical limitations in the recovery time of the inverter thyristors, the pulses cannot be smoothly decreased until dropout occurs, but must be held at a minimum width until the intersections making up the pulse disappear entirely, a condition called "lockout."(1) This effect produces discontinuities in position as well as slope.

To eliminate the nonlinearities caused by dropping pulses, the entire modulation scheme must be changed before dropout occurs. One technique that allows smooth dropout of pulses is the level-set method in which the sine wave reference is compared with a series of preset DC levels. (1,5) Although the output now deviates markedly from a shaped sine wave, linearity
between input command and output inverter voltage can be preserved. The dashed line indicates the input-output relationship achieved when the level-set modulation is used during the pulse dropout region.

Since the modulation now becomes linear below \( V_s^*/V_n \leq 1 \), Eq. (4) can be reduced to the form

\[
V_s = V_s^* \frac{2V_I/\pi}{V_n} \quad V_s^*/V_n < 1 \quad (5)
\]

\[
V_s = 2V_I/\pi \quad V_s^*/V_n > 1 \quad (6)
\]

Because \( V_n \) corresponds to rated AC voltage, the quantity \((2/\pi)(V_I/V_n)\) can be regarded as the DC inverter voltage expressed in per unit.

Since the inverter is considered as a lossless device, then by conservation of power

\[
V_I I_I = \frac{3}{2} V_s I_s. \quad (7)
\]

By substituting Eqs. (3) and (4) into Eq. (5), it can be shown that the corresponding currents on the AC and DC sides of the inverter are related by

\[
I_I = \frac{3}{\pi} \frac{V_s^*}{V_n} I_s \quad V_s^*/V_n < 1 \quad (8)
\]

\[
I_I = \frac{3}{\pi} I_s \quad V_s^*/V_n > 1 \quad (9)
\]

Equations (5), (6), (8), and (9) can be used to construct the block diagram model of the inverter, Fig. 7. Note the product and saturation nonlinearities in the block diagram. It can be observed that delay in the frequency channel has also been included since, in general, sampling effects are present as a result of the discrete switching of the inverter. This delay is approximately represented by a dead time \( e^{-Ts} \) where \( T \) is the period of the triangle wave reference. Fortunately, the delay is typically small and does not have an appreciable effect on the dynamic behavior of the inverter. This sampling delay will be neglected henceforth in this paper.

By aligning the q-axis of the synchronously rotating reference frame with the voltage vector defined by the inverter, the inverter can be effectively attached to the equivalent circuit of the induction machine. Because of the nonlinearities in the inverter block diagram the entire equivalent circuit is difficult to construct. However, the overall set of system equations defining behavior of the filter, inverter, and induction motor are given in Appendix 1.
EFFECTS OF THE DC FILTER ON THE MOTOR TRANSFER FUNCTIONS

To show the effect of the DC filter on the motor, a series of system transfer functions have been calculated using the system equations of Appendix 1. The parameters used for these computations are summarized in Appendix 2 and correspond to an actual drive prototype. The method used to compute the transfer functions has already been presented in Refs. 7 and 8. In general, the filter is usually designed by two steady-state criteria. First, the output capacitor must be sized large enough to handle the ripple current without excessive voltage variation because of motor operation at a lagging power factor. Secondly, the resonant frequencies of the inductance and capacitance of the filter should not be excited by the six times motor frequency current ripple which exists when the inverter is in the square wave mode of operation.

Although often neglected, the presence of the DC line filter also has a considerable effect on the system control design. Since the control system is basically a torque regulator, the basic transfer function of interest is the change in electromagnetic torque output divided by the change in per unit inverter frequency as an input. Figure 8 shows the variation in the motor transfer function $\Delta T_e/\Delta \omega_e(pu)$ for changes in motor operating speed at a constant speed at a constant load torque of 1017 n·m. The effect of the DC filter has been neglected ($V_I = 600 V$). The figure shows only the upper half of the complex frequency plane since the lower half is the complex conjugate. Also, it should be recalled that the poles of the transfer function (roots of the system characteristic equation) are the same for the transfer function between any two-system variables. However, the zeros (numerator of the transfer function) depend on the particular input and output variables chosen.

The motor poles consist of two pairs of complex roots. One pair is located on the $j\omega$ axis at a frequency corresponding to the stator excitation frequency. The distance to the left along the negative real axis is a function of the stator resistance. A second pair of poles results from the rotor time constant which do not change appreciably with the motor speed. One zero is located at about -50 radians/sec on the negative real axis, thus tending to cancel the effect of one of the rotor poles. The multiplier in the voltage channel (Fig. 1) has the effect of introducing a pair of complex zeros that track the operating frequency and minimize
the effect of the stator poles. Unfortunately, these zeros disappear when the inverter operating mode shifts from pulse width modulation to square wave operation for which no voltage control is possible. The pole-zero location of this transfer function permits closing the torque regulation control loop with gain only. A stable system will result since the net transfer function can be considered as a single pole at about -50 on the real axis. No major problems will occur with this transfer function except at very low speeds as shown in Ref. 8, Fig. 16. It should be noted that the transfer function using torque as an output is essentially the same as using the real component of stator current as in Ref. 8.

Figure 9 shows the effect of adding the DC filter on the motor transfer function. The variation in the motor and DC filter poles and zeros is shown as the operating speed of the motor is varied at a constant torque of 1017 nt-m. It can be noted that there are two complex poles and two complex zeros added as a result of the filter. At low speeds where the PWM ratio is low, the effect of the filter is small and the transfer function is basically that of the motor alone. The location of one of the filter poles without the loading of the inverter is shown for reference. At low speeds the higher frequency pole and zero will converge to this point and their effects cancel. However, at higher speeds the filter severely influences the motor transfer function and makes the problem of control of torque much more difficult. For example, at the base speed of 1.0 pu, the torque regulator control loop has to be limited to a lower gain than before, due to the presence of the extra complex pole near the real axis and complex zero in the right half plane. Without the zeros introduced due to the multiplier, the situation would be even more severe.

Because of the detrimental effect of the DC filter on the motor transfer function, it is useful to observe the effect of varying the DC filter size. Figure 10 shows the effect of varying the filter impedance on the motor transfer function at a speed of 1.282 per unit in square wave operation. For the purpose of determining the effect of the filter on the motor, the filter resonant frequencies were maintained constant and only the filter impedance has been varied. For a low per unit filter impedance (0.025), the transfer function is basically a pair of complex poles at about -25+j300 with the effect of the extra filter pole and zero tending to cancel. This result is expected since the effect of the filter should be negligible for an infinitely large filter of zero impedance. As the filter impedance increases, the effect of the filter becomes more extreme until a pair of
complex poles migrate into the right half plane and the system becomes unstable. The rotor poles and zeros are little affected by the DC filter. There is a practical limit on the reduction of the impedance of the filter since the DC power source itself contains inductance and resistance. Careful coordinated power circuit and control circuit design can reduce the problems associated with the effect of the filter. However, the DC link filter always tends to reduce the system stability in motoring.

Figure 11 shows the variation in the \( \frac{\Delta T_e}{\Delta \omega_{e(pu)}} \) transfer function as torque is varied from braking (generating) to motoring. Note that the zeros due to the filter are very sensitive to load, and move relatively far into the left half plane for braking (negative torque), whereas the system poles are not significantly affected. A closure of the torque regulation loop with a simple gain is satisfactory for braking at the relatively large braking torque shown.

In summary, an examination of the effect of the DC link filter on the operation of an induction motor demonstrates that the filter tends to reduce stability in motoring but improve stability during braking.

OVERVOLTAGE CONTROL DESIGN

The drive used as the example in the previous section operates with a power source having a nominal 600 volts DC. It is desired to supply braking energy back to the DC source. However, because of the nature of the power system, the impedance of the DC source will vary and it may even be suddenly interrupted. Using transfer function techniques, a control system has been developed which senses the DC filter voltage across capacitor \( C_{F1} \) in the middle of the filter, compares this voltage to a reference voltage, and varies the inverter frequency in such a manner as to limit the rise of voltage during braking. The transfer function required to accomplish this overvoltage protection is \( \frac{\Delta V_1}{\Delta \omega_{e(pu)}} \) where \( V_1 \) is the voltage across capacitor \( C_{F1} \) in the DC filter.

Figure 12 shows the variation in the transfer function \( \frac{\Delta V_1}{\Delta \omega_{e(pu)}} \) as the DC line impedance increases, thereby allowing less braking current to flow. The speed is fixed at a value of 3 pu. The DC line impedance in ohms is shown as a parameter. The zero on the real axis starts at -72 and moves right. The system poles move to the right as the DC line impedance increases until they reach the points marked for an open circuit. Thus the system stability decreases as line impedance increases. As in the
previous section, the poles of the system move as a function of speed. Figure 13 shows the changes in the same transfer function when speed is a parameter. This transfer function is plotted for an open circuit on the DC source side of the filter, which is the worst case for stability. Note that the poles on the real axis are essentially independent of speed. The complex poles are dominated by the filter resonance and do not significantly move for speeds greater than three per unit.

It is desired to design an overvoltage controller to regulate the inverter voltage at a maximum value of 1000 V which will be stable even under the worst case situation of an open-circuited DC line. It is necessary to estimate the required control system response time so as to determine the requirement on the position of the closed-loop poles. A reasonable requirement is to assume that the system is regenerating 1000 A of current at the maximum voltage of 1000 V. The maximum voltage permitted on the inverter is 1200 V; thus the regenerated current must be reduced to zero before the line filter charges more than 200 V. Assuming linear charging, this time will be given by

\[ T = \frac{V_C}{I} = 4.8 \times 10^{-3} \text{ sec.} \]

Since the current will decrease as the voltage increases due to the action of the control, a time of about \(7 \times 10^{-3}\) sec can be allowed. This yields an approximate bandwidth of 200 radians/sec. Reference to Fig. 13 shows that there is no hope of achieving this result without loop compensation since the response is dominated by the low-frequency poles on the real axis. Although a lead compensator can be used it is not sufficient. Fortunately, a compensator consisting of complex zeros plus a lead will yield results that are satisfactory. Figure 14 shows one arrangement of compensator zeros to achieve the desired result. The complex zeros must be located so that the dominant poles are at high frequency and sufficient gain can be achieved. Perfect compensation is desired, but is not really necessary, since the inverter switching losses which are not included. Also, a back-up shut down of the drive is supplied.

A complete PWM inverter induction motor drive of a 600 kW continuous, 1200 kW intermittent rating has been built and tested. Figure 15 shows the experimental results of operation of the overvoltage control system. The system is operated in regenerative braking at a speed which corresponds to a speed of 2.5 per unit. The shaft torque is 732 n-m

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per motor with two motors in use with a filter impedance corresponding to 0.5 pu in Fig. 10. The line breaker (LB in Fig. 2) is opened to simulate a power interruption. The rise time of the DC voltage is about 8 \cdot 10^{-3} \text{ sec} and the overshoot is 10 percent over the desired maximum value of 900 V. Note that the motor slip frequency signal undergoes a considerable transient in the motoring direction. The noise present on the slip signal when the overvoltage control circuit is operative disappears during normal torque regulation.

CONCLUSION

The DC input filter can be a severe limitation to the speed of response of a PWM inverter-induction motor drive. The analysis in this paper shows that this effect typically is maximum around the motor rated speed and may actually cause the system to be open-loop unstable during motoring. The location of the poles and zeros introduced by the filter tends to influence the location of the motor stator poles, thus complicating the design problem. The filter severely limits the feedback gain which can be achieved with conventional feedback signals. Nonetheless, by properly identifying the system poles and zeros by means of transfer function techniques, suitable control signals can be synthesized. In particular, the design of a high response overvoltage control was discussed. Experimental results verify that such a practical high-speed control can indeed be realized.

APPENDIX 1 -- SYSTEM EQUATIONS

Upon writing the differential equations of the machine in the synchronously rotating reference frame and utilizing the block diagram of Fig. 5 for the inverter, the system equations can be arranged in matrix form. During motoring the equations are:

\[
\begin{bmatrix}
V'_{R} \\
0 \\
V^*_{R1} \\
V^*_{R2} \\
V^*_{Q1} \\
V^*_{Q2}
\end{bmatrix} = \begin{bmatrix}
R'_{CL} + R'_{L1} + R'_{R} & 1 & -R'_{C1} & 0 \\
-1 & 0 & 1 & 0 \\
-R'_{C1} & -1 & R'_{CL} + R'_{L2} + R'_{C2} & 1 \\
0 & 0 & -1 & 0
\end{bmatrix} \times \begin{bmatrix}
L'_{L1} \\
V'_{C1} \\
V'_{L2} \\
V'_{C2}
\end{bmatrix} \times \begin{bmatrix}
L'_{F1} + L'_{R} & 0 & 0 \\
0 & C'_{F1} & 0 \\
0 & 0 & L'_{T2} & 0 \\
0 & 0 & 0 & C'_{T2}
\end{bmatrix} \times \begin{bmatrix}
I'_{L1} \\
I'_{C1} \\
I'_{L2} \\
I'_{C2}
\end{bmatrix}
\]

(A-1)
In Eqs. (A-1) and (A-2), the superscript "e" denotes machine variables in the synchronously rotating reference frame. The primed filter variables have have been employed in order to normalize DC side variables to the d-q axis of the rotating reference frame. In particular,

\[ V'_R = \frac{2V_R}{\pi} \quad \text{(A-3)} \]

\[ I'_{L1} = \frac{\pi I_{L1}}{3} \quad \text{(A-4)} \]

\[ R'_{L1} = \frac{6}{\pi^2} R_{L1} \quad \text{(A-5)} \]

All other primed voltages, currents, and impedances on the DC side of the inverter are modified by the same factors.

**APPENDIX 2 -- SYSTEM PARAMETERS**

**DC Filter**
- \( R_{B1} \) Regenerative braking resistor \( 0.2\Omega \)
- \( R_{B2} \) Dynamic braking resistor \( 0.8\Omega \)
- \( L_{F1}, L_{F2} \) \( 400\mu\text{h} \)
- \( C_{F1} \) \( 8,000\mu\text{fd} \)
- \( C_{F2} \) \( 16,000\mu\text{fd} \)
- \( R_{C1}, R_{C2} \) \( 0.03\Omega \)
- \( R_{L1}, R_{L2} \) \( 0.007\Omega \)
Motor

$\omega_{base} = 314 \text{ rad/sec}$

$r_s = 0.0217\Omega$

$r_r' = 0.0329\Omega$

$L_{ks} = 0.0003235 \text{ hy}$

$L_{kr} = 0.0003685 \text{ hy}$

$L_m = 0.013507 \text{ hy}$

REFERENCES


Fig. 1 System block diagram.

Fig. 2 DC line filter and source.

Fig. 3 Motor operating conditions in per unit.
Fig. 4 Control block diagram.

Fig. 5 Induction motor equivalent circuit.

Fig. 6 AC fundamental component of inverter voltage vs control signal $V^*_{S}$. 
Fig. 7 Inverter equivalent circuit.

Fig. 8 Locus of the roots of the transfer function $\Delta T_e/\Delta \omega_e(\text{pu})$ as per unit speed is increased (motor only).

Fig. 9 Locus of the transfer function $\Delta T_e/\Delta \omega_e(\text{pu})$ as per unit speed is varied (motor plus filter).

Fig. 10 Transfer function $\Delta T_e/\Delta \omega_e(\text{pu})$ with changes in per unit filter impedance as the parameter (motor plus filter).
Fig. 11 Transfer function $\Delta T_e/\Delta \omega_e(\text{pu})$ for changes in torque load (motor plus filter).

Fig. 12 Variation of the transfer function $\Delta V_1/\Delta \omega_e(\text{pu})$ as the DC line impedance increases.

Fig. 13 The transfer function $\Delta V_1/\Delta \omega_e(\text{pu})$ as per unit speed increases.

Fig. 14 Root locus plot of over-voltage control system as gain is changed.
Fig. 15 Test of overvoltage control system.