

4. Simulation Results

An application of the computer aided control design of a starter/generator PMSM drive system discussed in Chapter 3, Figure 13, is presented in this chapter. A load torque profile is given in Figure 14. The PM synchronous motor is used to accelerate an aircraft APU's synchronous generator to its synchronous speed (15000 rpm) in a given time interval (30 sec.). The speed profile is not underscored. The control requirements are a robust system with fast and stable current and speed responses. The design is based on the simulation models from Chapter 2 and design principles and analysis from Chapter 3.

At the beginning, the examination of the load torque profile from Figure 14 brings some important design guidelines:

- 1) A higher torque is required at low speeds than at high speeds. It makes it possible to apply series-to-parallel switching of the stator windings in order to achieve a high starting torque and high speed without motor and VSI over-design.

- 2) In order to further optimize the motor drive design, flux-weakening control can be applied. Because some three-phase VSI inverter topologies [65] have the voltage-to-current phase shift limit of 30° el. for regular operation, the flux-weakening control method should be carefully chosen and designed against that limitation. It should also determine the series-to-parallel switching point.

- 3) The power rating of the PMSM starter should be as high as necessary to satisfy the system acceleration requirements, but, again, the motor drive should not be over-designed. The motor can even be shut-down at a lower speed than the targeted speed of 15000 rpm, since the load torque profile is already generative (negative) after 7000 rpm (due to the fully operating gas turbine). Again, the VSI limitations should determine the motor shut-down speed, by binding the parallel-mode flux-weakening region, as discussed above. In order to obey start-time requirements, a motor rated speed (parallel winding connection) should be about 10000 rpm (5000 rpm with series windings) with starting torque (series windings) of about 30 Nm (15 Nm maximum torque with parallel windings). It leads to the 20 kW motor power rating range, if we include power losses.

4) The biggest (negative) change of the load torque slope occurs at the speed of 5000 rpm, when the gas turbine turns on. It is close to the suggested series-mode rated speed, so that the flux-weakening can begin just below that speed. There are two reasons to support this idea. First, the significant load torque drop after 5000 rpm enables the rated speed extension of about 1000 rpm. The extended speed is close to the point (7000 rpm) where the load becomes accelerating (negative), so that the starter electromagnetic torque could be significantly reduced (series-parallel switching), but still high enough to meet the starting-time requirements. Second, the motor torque decreases with speed during the flux-weakening. That helps the speed open-loop stabilization, as discussed in Chapter 3, if the speed profile at some point becomes important. Since the motor torque must always exceed the load torque in order to produce the acceleration, with the chosen motor torque and power, the series mode base speed should be about 4500 rpm.

5) The load torque dynamics require the determination of the worst case operating points, following the criteria from Section 3.2.1. Besides the speed of 5000 rpm, two other worst-case-operating-points are at 9000 rpm and 11000 rpm. At 9000 rpm the motor should operate in parallel mode, still with the full torque, while at 11000 rpm it should already enter the flux-weakening region, but the load torque slope is steeper than at 9000 rpm. Also, the line frequency first harmonic at these two operating points is much closer to the VSI switching frequency (20-50 kHz range in order to avoid high VSI switching losses), than at 5000 rpm, so that the influence of the switching noise and sampling delay is higher.

6) Since the current (inner) loop should be much faster than the speed (outer) loop, and the speed profile is not a strict requirement, the current loop design is more critical. Consequently, the decoupling-with-back-emf-elimination concept seems to be preferable compared to the equivalent-dc-motor control method.

7) The negative load torque slope suggests a reference motor torque correction in order to provide a stable speed loop, as it is discussed in 3.2.2. However, since in the flux-weakening regions the motor torque already decelerates with speed by definition, it will not be further modified in this example, for the sake of simplicity.

8) The load torque dynamics sometimes require an adaptive change of gains of the PI regulators, with the change of the load torque slope. However, the load torque slope change is slow enough to neglect that dynamic in comparison with the other motor drive time constants.

Regarding the above observations and project requirements, a good designer's choice can be a PMSM drive with next parameters (assumed parallel windings), Table 3:

Table 3. Parameters of a PMSM drive

dc link voltage:	$V_{dc}=370V$	maximum phase current:	$I_{phmax}=50A$
stator phase resistance:	$R=0.08\Omega$	number of pairs of poles:	$p=3$
stator phase inductance:	$L=0.19mH$	back emf constant:	$k_t=0.19V/(rad/s)$
d-axis inductance:	$L_d=0.4L$	motor moment of inertia:	$J_m=0.0017kgm^2$
q-axis inductance:	$L_q=0.8L$	load moment of inertia:	$J_{load}=0.316kgm^2$
filter inductance:	$L_f=0.34mH$	friction torque approx.:	$T_{fr}=1Nm$

The chosen motor drive has the output characteristics given in Figure 23. Their examination leads to the next control design frames:

1) For $V_{dc} = 370 V$ and VSI maximum current $I_{phmax} = 50A$, which produces a maximum torque of 28.5 Nm (with $i_d = 0$ control), the motor base speed values are: $w_{bs} \approx 4700rad/s$ for series and $w_{bp} = 9250rad/s$ for parallel operating mode.

2) A motor torque must always be higher a load torque in order to produce the acceleration. A critical point is at 5000 rpm, where the load torque reaches its peak and the motor torque already decreases due to flux-weakening.

The PMSM drive system small- and large-signal models are created using the modules described in Chapter 2. Flux-weakening control simulation models are based on the three control strategies discussed in Section 3.2.2, and they follow the control-signal flow principle, defined in Chapter 2. A two-stage cascade control in d-q coordinate space, Figure 15, is chosen for this design. Two current control methods (decoupling loops with and without back emf elimination) with and without load torque slope extraction, encircled by a speed control loop, discussed in Section 3.2.1, are compared and discussed in this application. Small-signal design of the current and speed loop PI regulators is done using the procedure from Section 3.2.1. A Matlab program and Simulink GUI models were developed in order to calculate PI regulator gains at different operating points, and to perform Bode analysis of the d-q average model of the PMSM drive, for

the abovementioned control approaches (see Appendix C for a shortened program listings and the system model samples). According to the above load torque observations, the worst operating points are at $w = 5000 \text{ rpm}$ (series mode), and $w = 9000 \text{ rpm}$ (parallel mode). The initial controller design was made for these two points. Subsequently, the design was checked at several other speed points, and modified where it was necessary to significantly improve the drive performance. The results of the controller design are presented in Appendix C as a part of the earlier mentioned program listings.

After choosing the motor drive power line components and determining the worst case operating points, the controller design continues with the small-signal design of current and speed regulators in d-q coordinates. The first step is the choice of a decoupling method, which serves to simplify the system small-signal d-q model by reducing the order of its small-signal transfer functions. This is the main difference between the two control approaches discussed in Section 3.2.1. Nevertheless the chosen control method, the abovementioned benefits of having decoupling loops, are obvious from Figure 31. However, in the case of the equivalent DC motor, the instability of the i_q current control-to-output transfer function at a low frequency at a given operating point also becomes obvious. This is because the unstable electromechanical, current (torque)-to-speed, transfer function has the influence to current control-to-output transfer functions through back emf (12). The PI regulator gains are calculated following the design procedure from Section 3.2.1. The current loop-gain transfer functions are shown in Figure 32. The current controllers' effects on coupling transfer functions is evident in Figure 34. Because of a similar effect as in the loop-gain transfer function, the "loop-gain" was included in the coupling plot titles. However, they are not feedback loops. It should be noted that the current control design with full decoupling (with back emf elimination) is an easy and straight-forward single-pole plant control design of two fully decoupled subsystems, explained in Section 3.2.1.2. On the other side, the equivalent DC motor approach is full of uncertainty related to the load torque profile, even if it is fully extracted and predictable. The open loop instability can be cured only with the reference motor torque reduction with a slope at least equal to the load torque slope (assuming negative slope). It will stabilize the motor electromechanical equation (see 12), so that speed will not be a destabilizing factor in the i_q current loop any more. However, the i_q current loop will remain a second order system (equivalent DC motor).

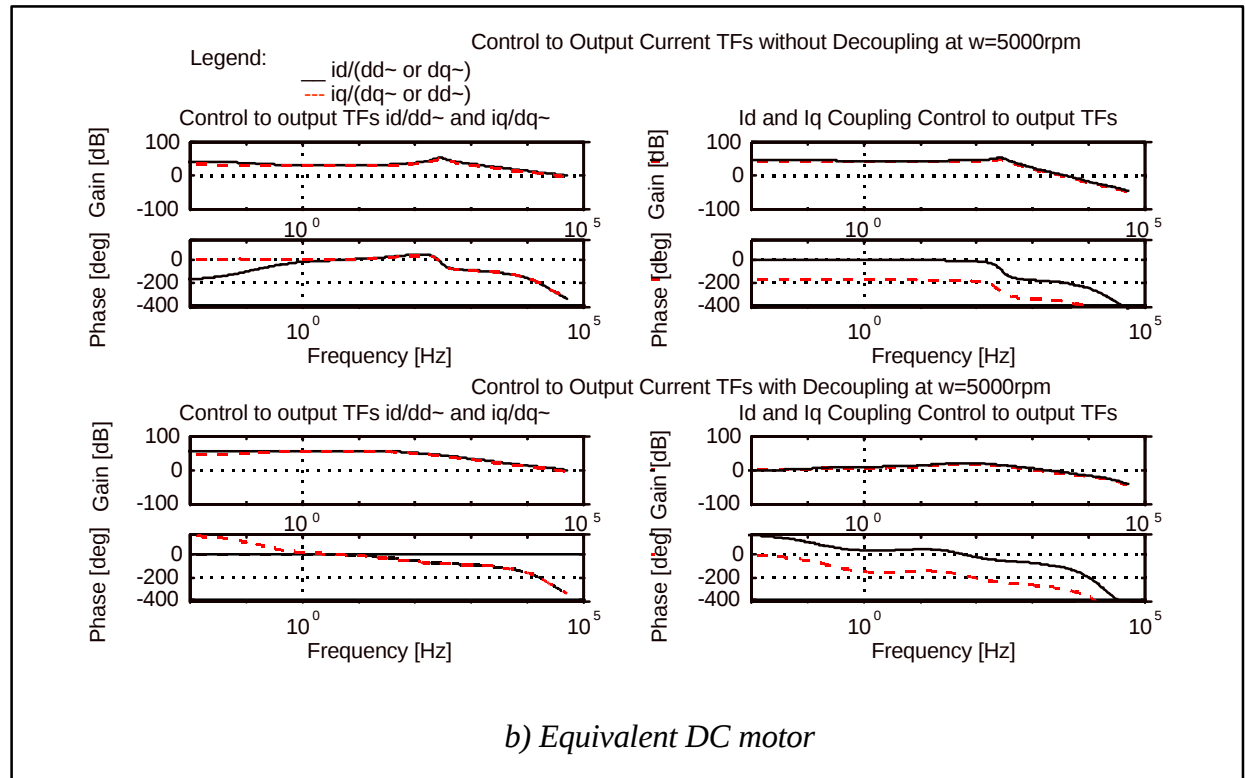
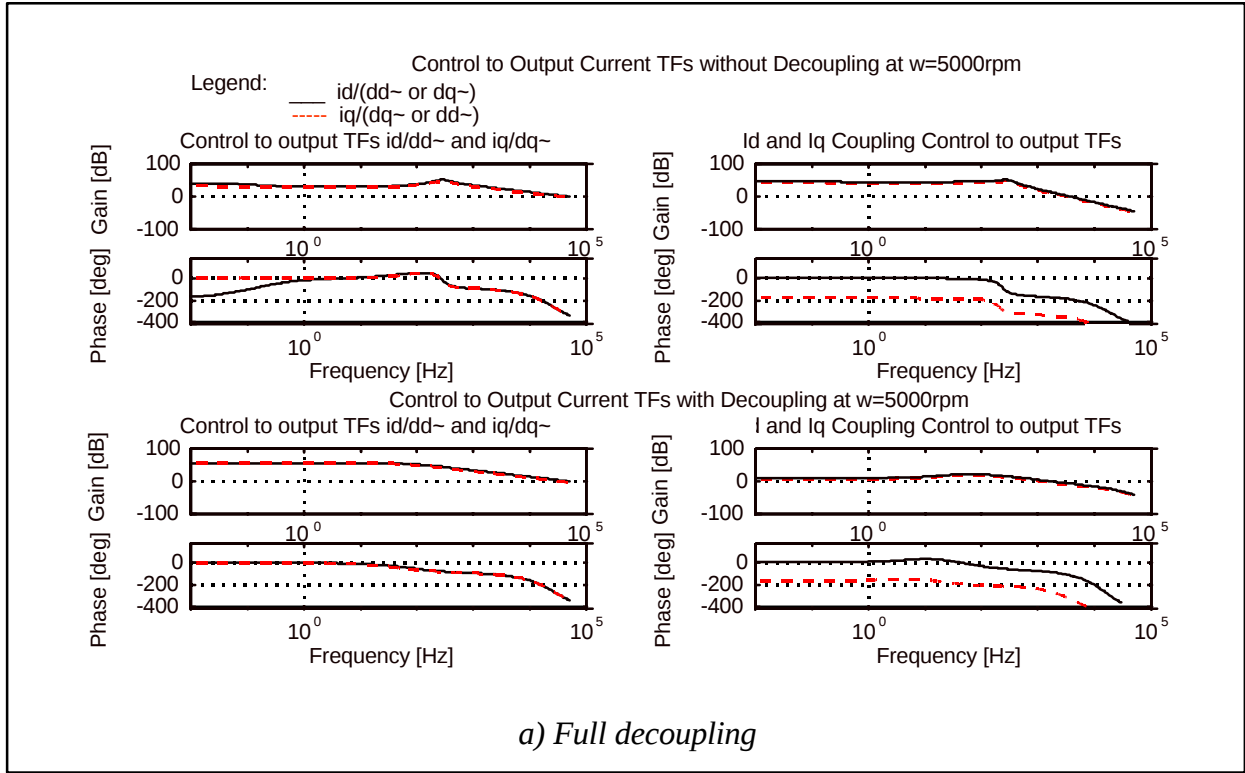


Figure 31. Decoupling Bode diagrams

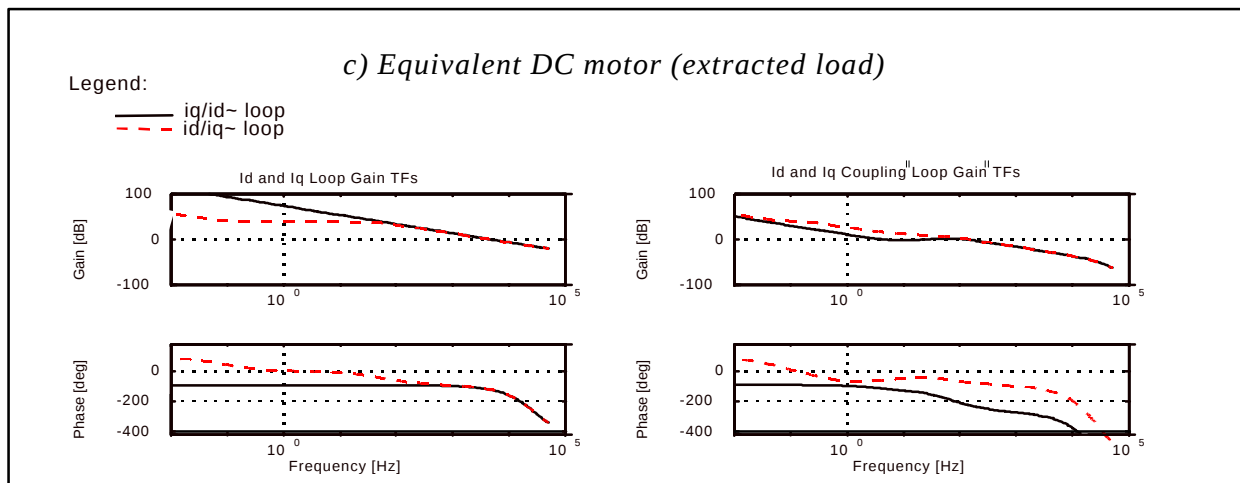
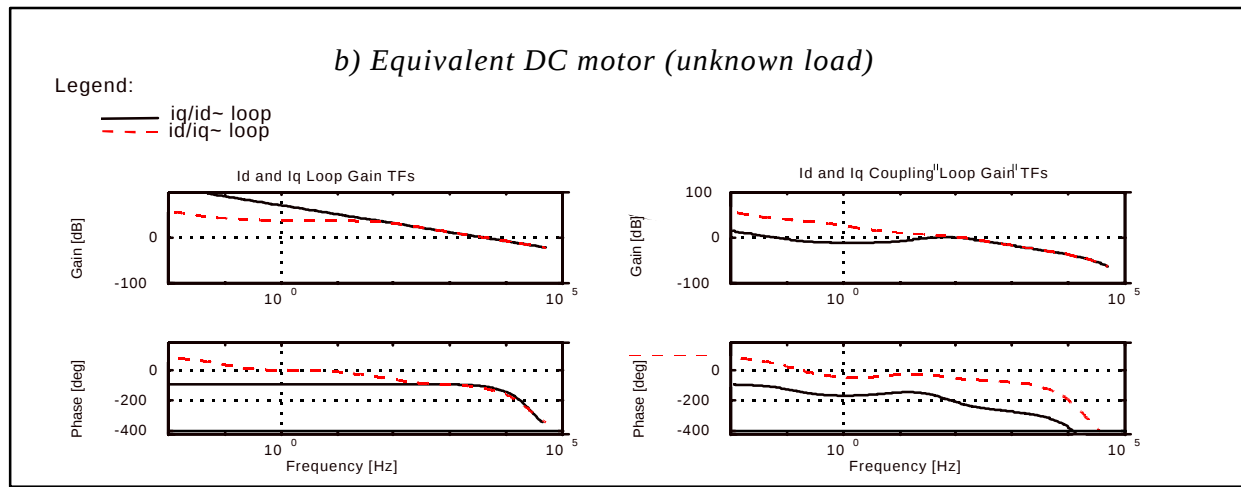
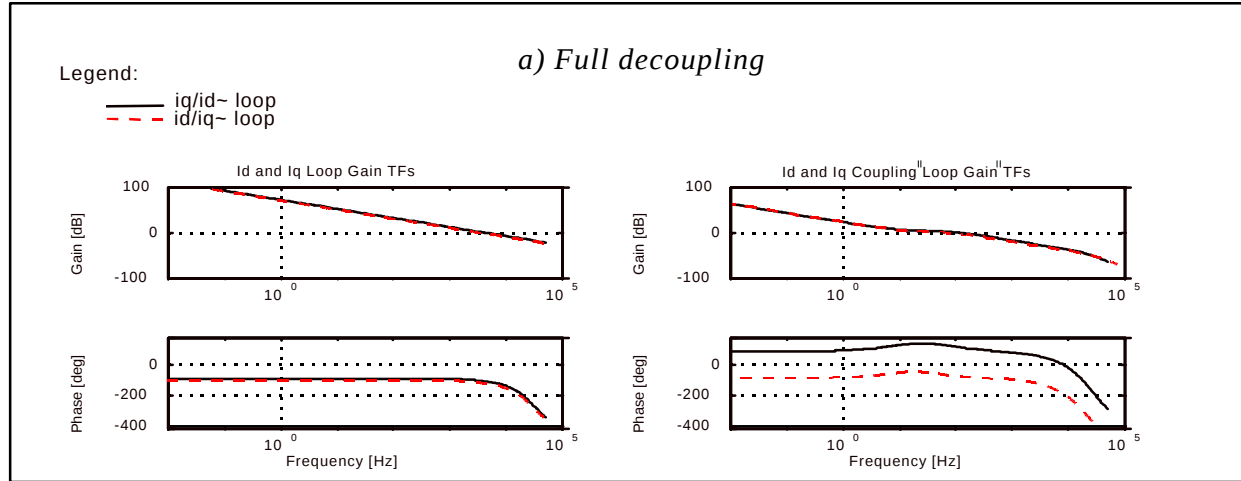


Figure 32. Current loop gain transfer functions: a) full decoupling, and equivalent DC motor with b) unknown load and c) extracted load control methods

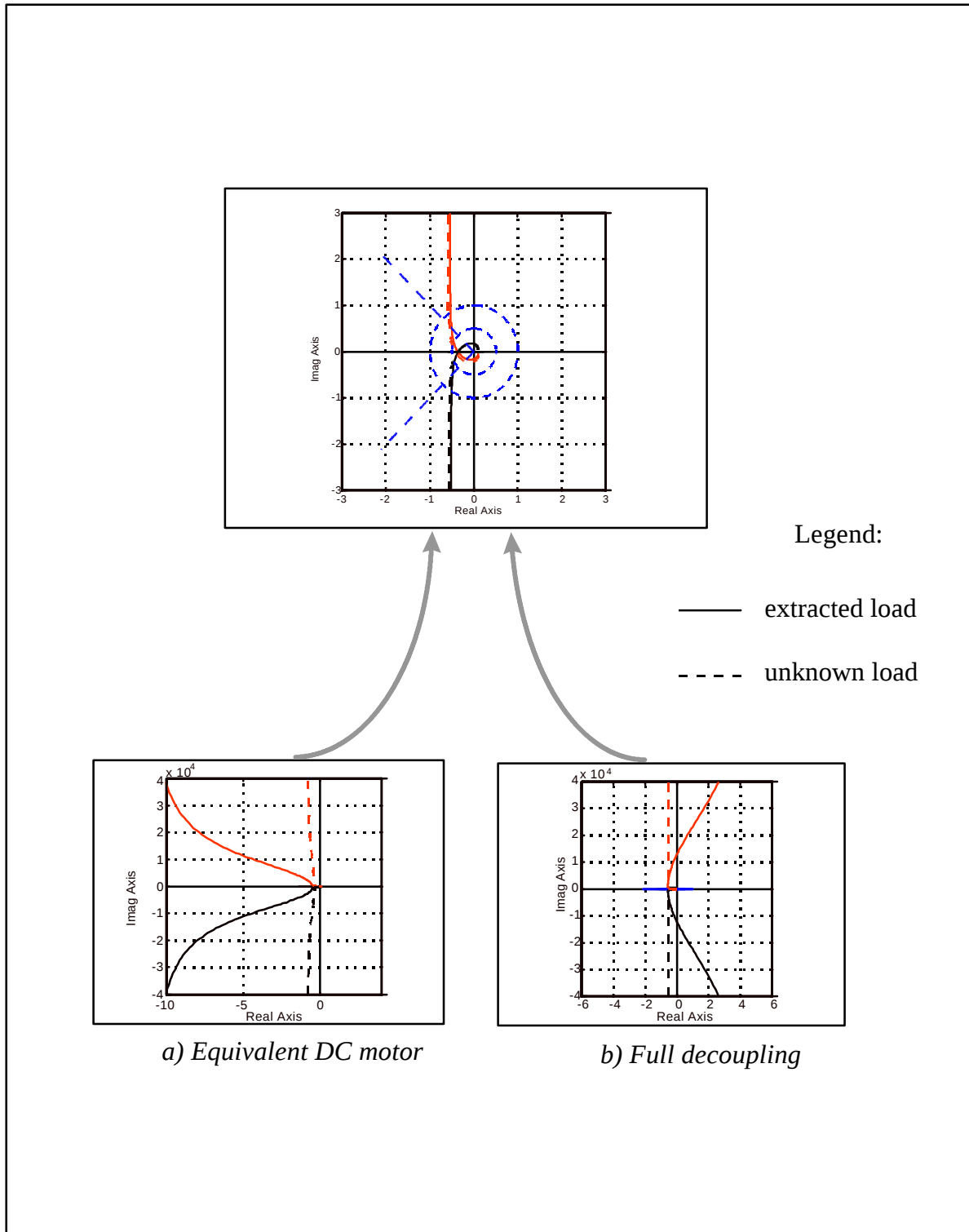


Figure 33. I_q current loop-gain transfer functions - Nyquist diagrams

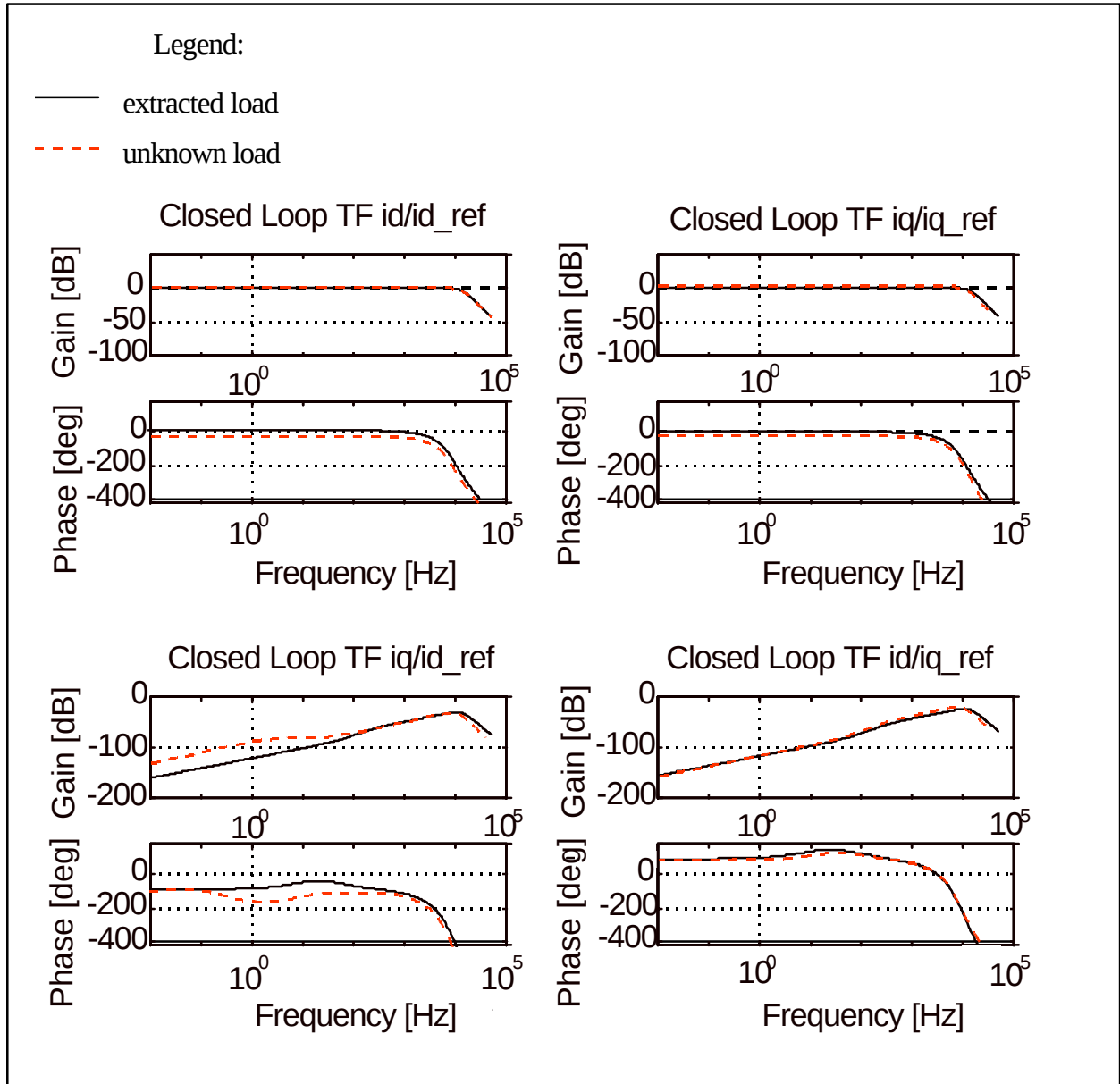
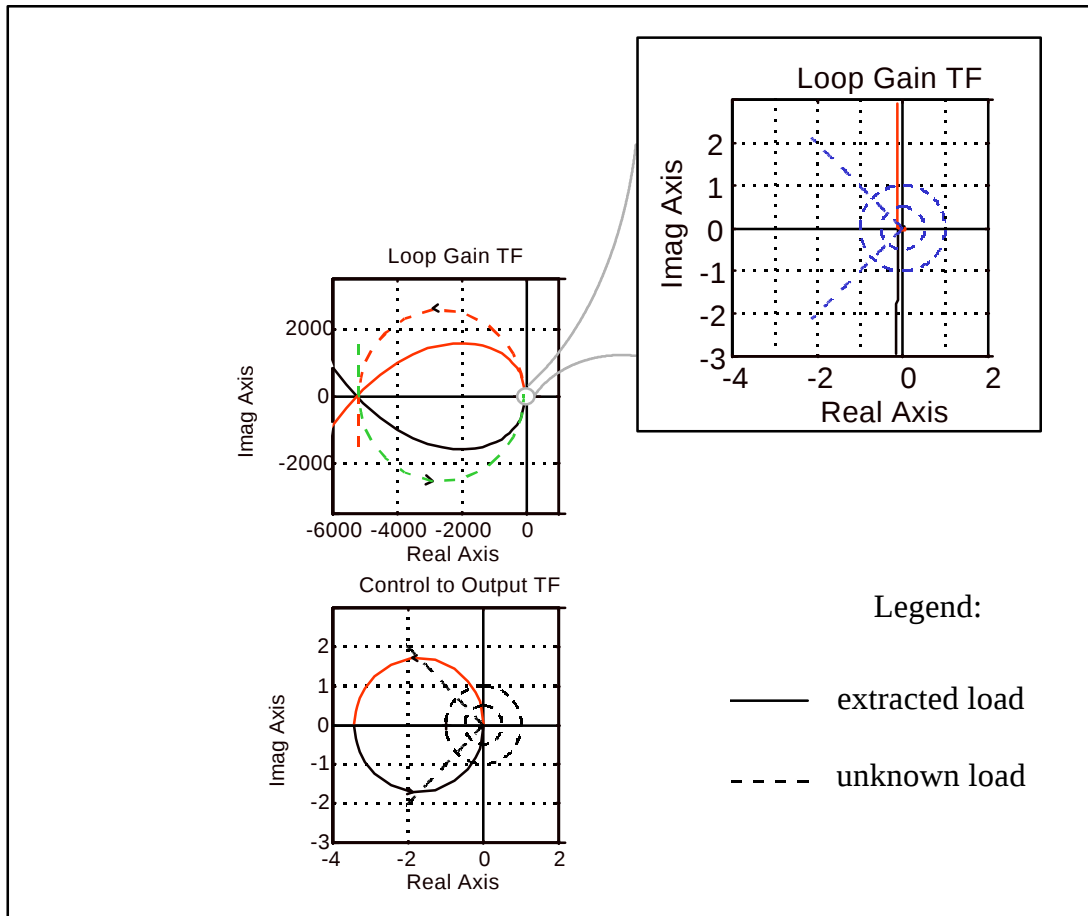
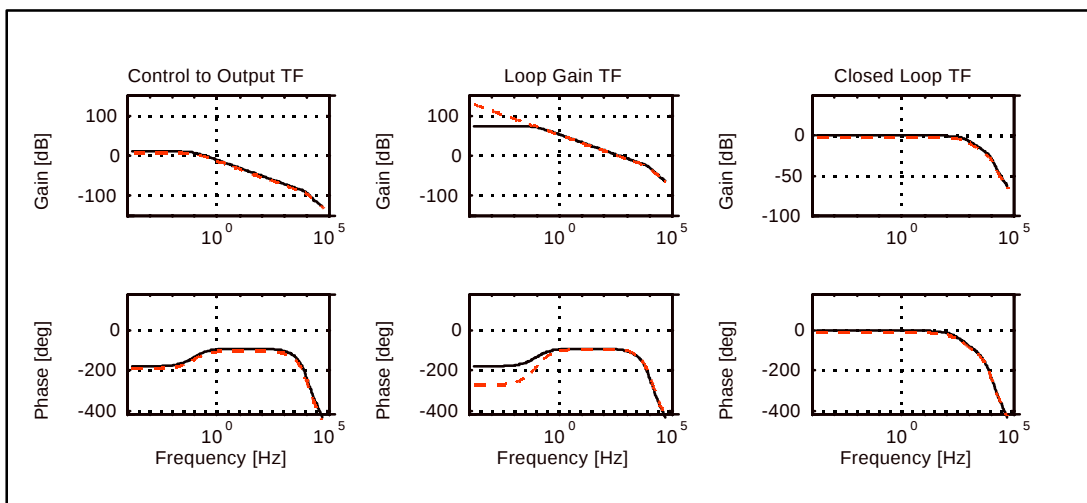


Figure 34. Closed current loop transfer functions



a) Nyquist plots

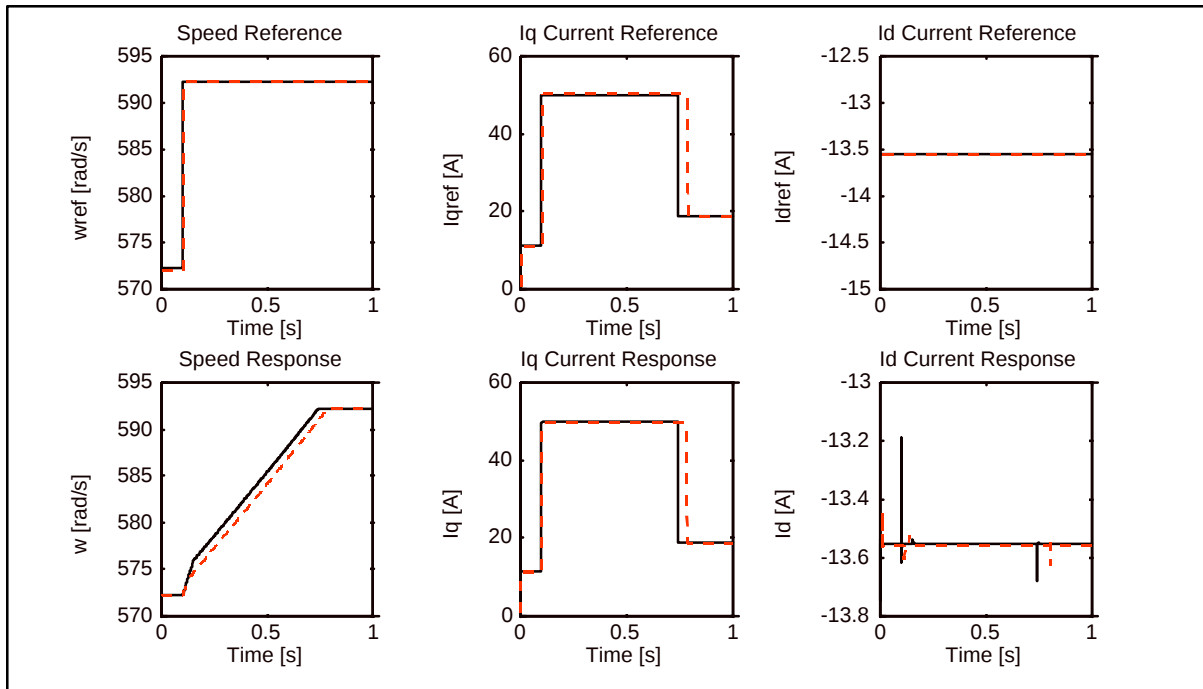


b) Bode plots

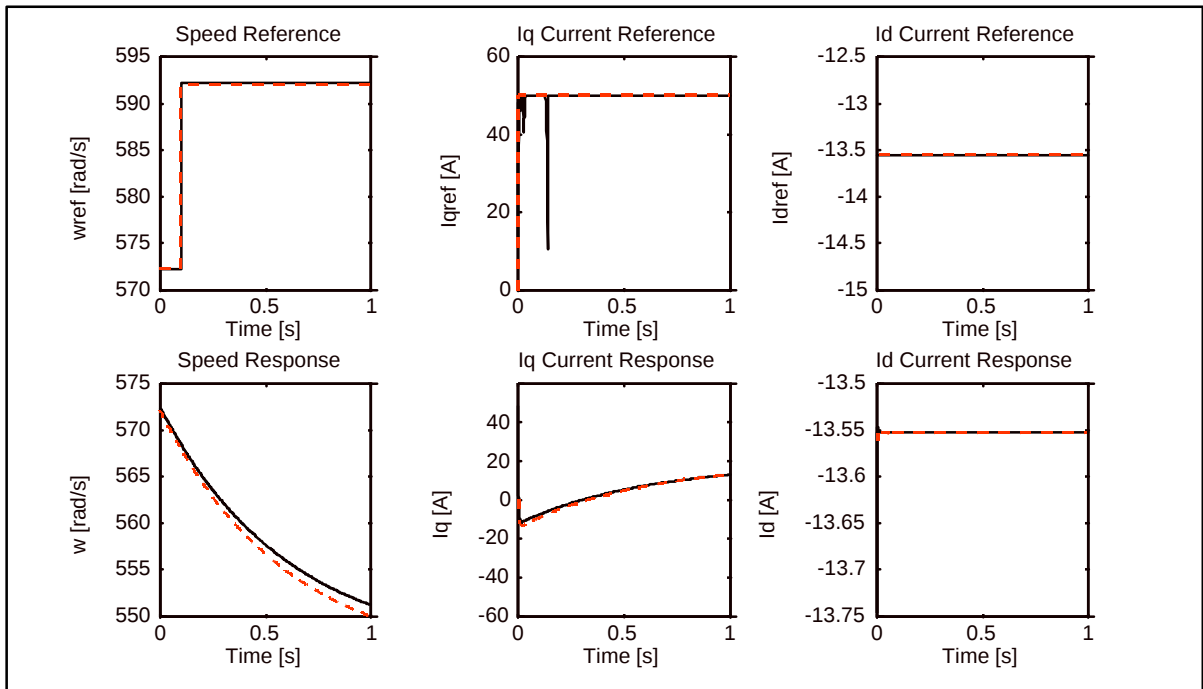
Figure 35. Speed loop transfer functions

Usually, the small-signal analysis considers a frequency lower than 1 Hz negligible and treats it as a DC operating point. However, it can be a dangerous overlook for systems with large mechanical time constants, as shown in Figure 33. The correlation with the respective Bode plots from Figure 32 should be noted. The instability occurs at the frequency below 0.1 Hz, which is directly readable from the Bode plots. A hidden danger of the apparent security of stable closed loops, Figures 34 and 35, is emphasized in Figure 36. It shows the current and speed step responses when the outputs of the regulators hit their limits, thus breaking the control loops. The local instability of the i_q current open-loop transfer function with the equivalent DC control, Figure 36.b, is again uncovered. It should be noted that a high system inertia didn't attenuate this instability. Both current and speed did not respond accordingly to their control signals. The stress is on the current instability, as can be seen from the step response of the system with a full decoupling control, Figure 36.a. There, a stable i_q current loop forced a stable speed response with the help of current and voltage limiters. The faster response with a load torque adaptive speed regulator should also be noted. It emphasizes the importance of the load torque profile extraction, whether by the method described in 3.2.1, or somehow else. Finally, the small-signal control design in this example did not take care of the speed profile, since it was not a strict requirement. However, in some more sensitive applications, that requirement could be emphasized. In that case, the load torque slope extraction becomes mandatory if the cascade feedback control is to be applied.

After the small-signal PI regulator design, the controller design proceeds with a large-signal design of non-linear controller components - limiters, VLPI compensator, reference motor torque and flux-weakening control. The limiter design will not be discussed here because its complexity requires much more space and attention, as can be assumed from the short discussion in Section 3.2.2, Table 2, than can be allowed in this work. The VLPI compensator design is well known in practice [79, 80], so it will not be discussed here. Because the system works most of the time in the open speed loop mode (saturated speed regulator output), that loop was not considered in the large-signal design. The reference motor torque is given in Figures 41 and 45, and will be mentioned later as a part of the flux-weakening discussion. Finally, the flux-weakening control design occupies the central part of the large-signal design and simulations in this example, as it did in the previous theoretical discussion (see Section 3.2.3).



a) Full decoupling



b) Equivalent DC motor

Legend:

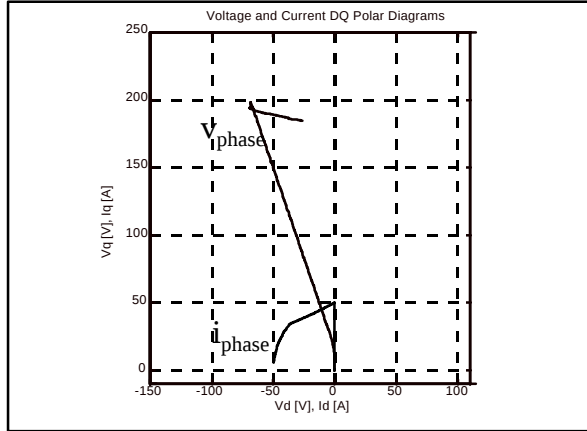
— extracted load

- - -

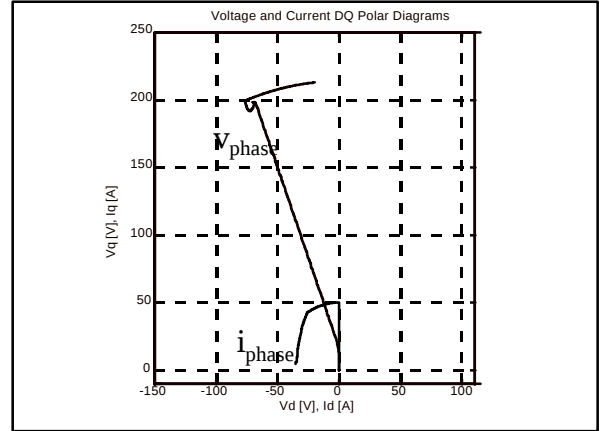
Figure 36. Speed and current step responses

Two sets of simulation results with three flux-weakening control methods, discussed in Section 3.2.3, are presented in this chapter: the first one is related to the “full-length” flux-weakening, with the maximum speed extension in series mode, while the second one represents the system simulation during the whole start-up period with a 30° voltage-to-current phase shift limit implemented in the flux-weakening control algorithms.

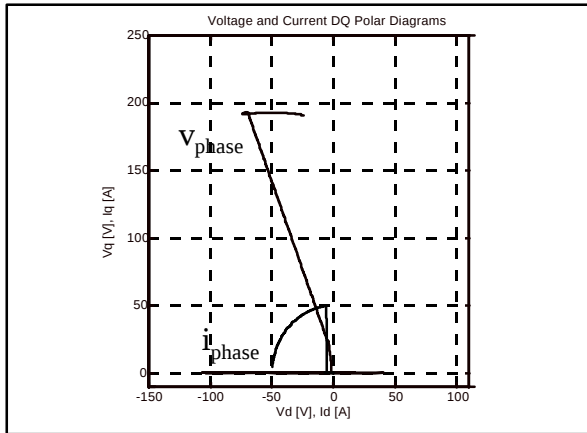
The principles of constant power (constant voltage and constant current) and optimum current vector control (with maximum torque and zero i_d current out of the flux-weakening region) methods are observable from the voltage and current (overlapped) polar d-q diagrams, shown in Figure 37.a-d, respectively. The correlation with Figures 28-30 should be noted. The critical speed points, after which the constant power strategies are not applicable anymore, are noticeable in Figure 37.a and 37.b, as current break points. The current phase shift in Figure 37.c is due to the maximum torque control, which requires some optimized value of the i_d current out of the flux-weakening region [70]. The voltage vector difference in the optimum current vector (OCV) control is due to the same base speed used for both maximum torque and $i_d=0$ controls. It is a usual design miscalculation, which can lead to losing of the benefit of having the maximum motor torque before the flux-weakening was applied. A key point is the earlier mentioned phase shift at the base speed with the maximum torque control. It shortens the available flux-weakening region and, if the base speed is the same as in the $i_d=0$ control case, the maximum speed extension is lower. However, if the base speed is left to be the rated speed with the given voltage and current vectors, which is higher than in the $i_d=0$ control (see Eq.s 12 in Section 2.2.2), the maximum speed extension is the same in both cases. The difference is only in the acceleration time. Figure 38 shows the voltage and current phase shifts in time domain, while Figure 39 shows the voltage and current time domain diagrams. The smoothness of curves and constant phase current and voltage in OCV control in contrast with constant power control strategies should be noted. This is because of the absence of the above mentioned critical speed points, characteristic for the constant power strategies, when the OCV flux-weakening control is applied. The differences in the active power used in the above discussed flux-weakening strategies are shown in Figure 40. Point A indicates the beginning of the flux-weakening region, and points B₁ and B₂ show when the critical speed is reached with constant power (CVCP and CCCP) control methods, respectively.



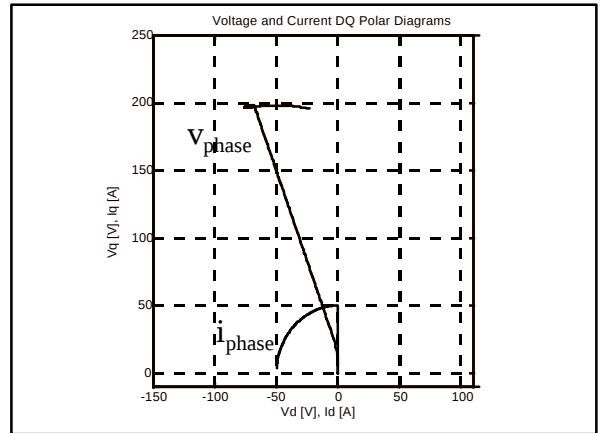
a) Constant Voltage
Constant Power Control



b) Constant Current
Constant Power Control

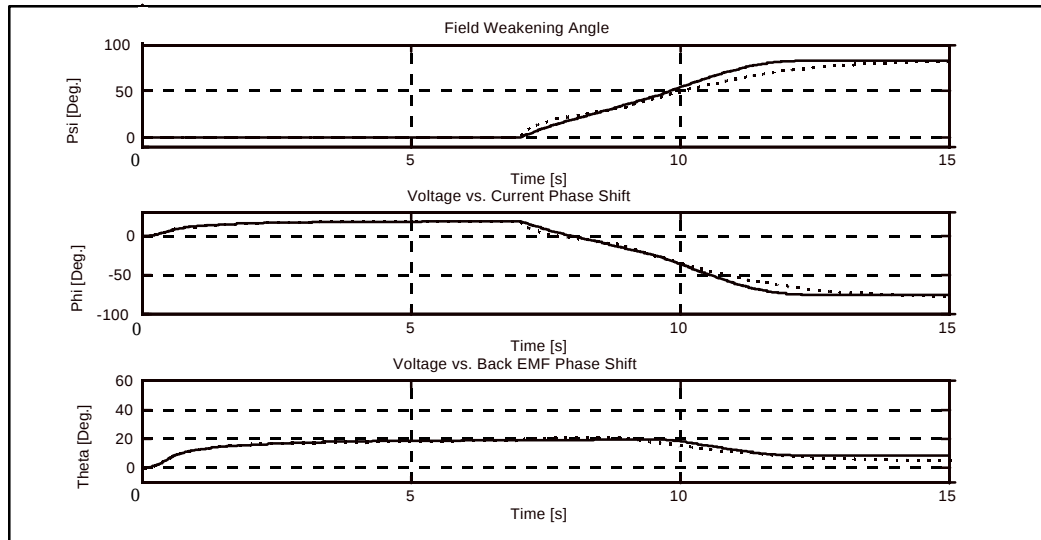


c) Optimum Current Vector Control,
 $T_m = T_{\text{max}}$



d) Optimum Current Vector Control,
 $i_d = 0$

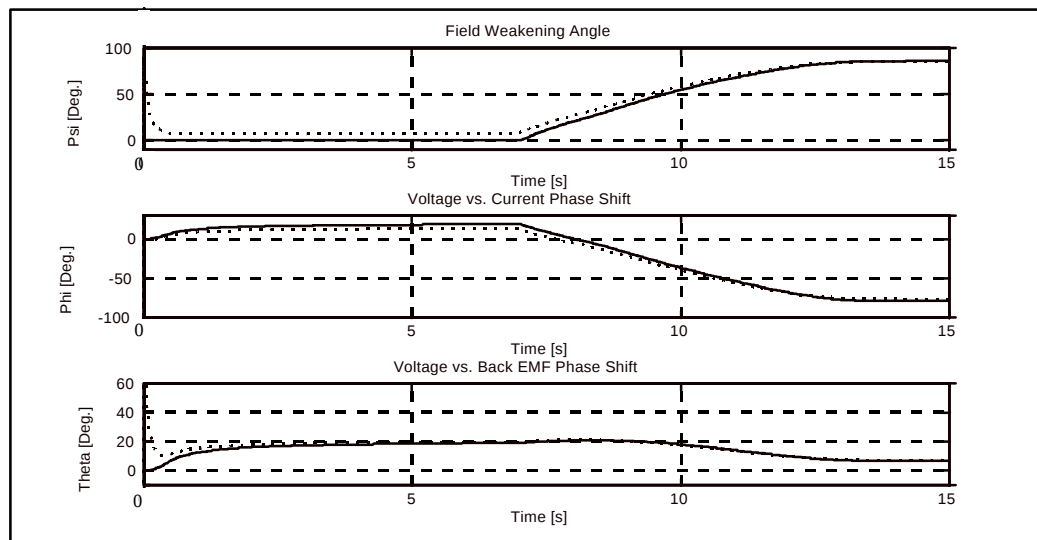
Figure 37. "Full-length" flux-weakening: voltage and current d-q polar diagrams



Legend:

- Constant Voltage Control
- - - Constant Current Control

a) Constant Power Control

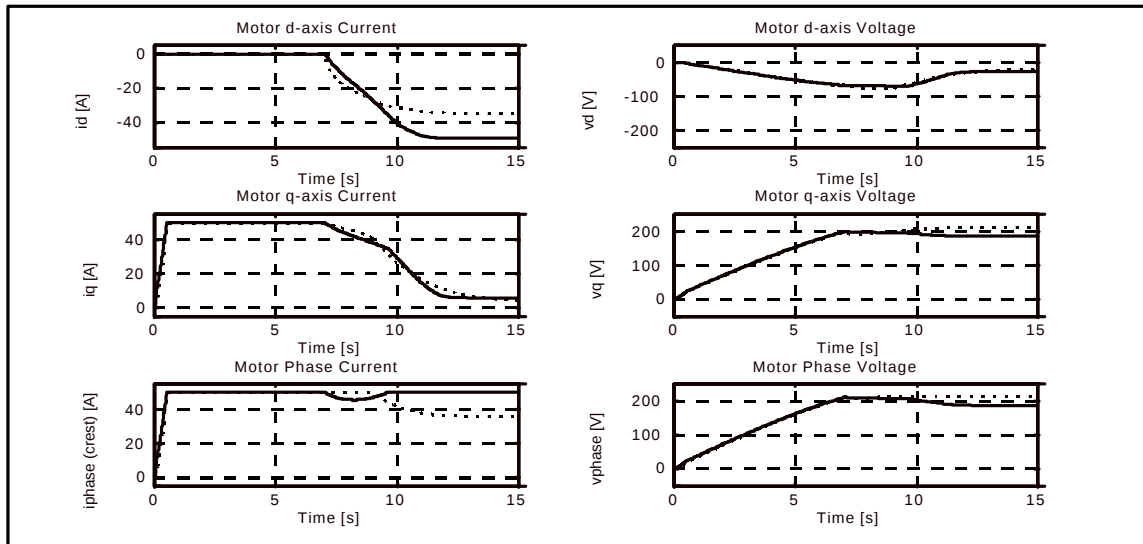


Legend:

- $I_d=0$ Control
- - - Maximum Torque Control

b) Optimum Current Vector Control

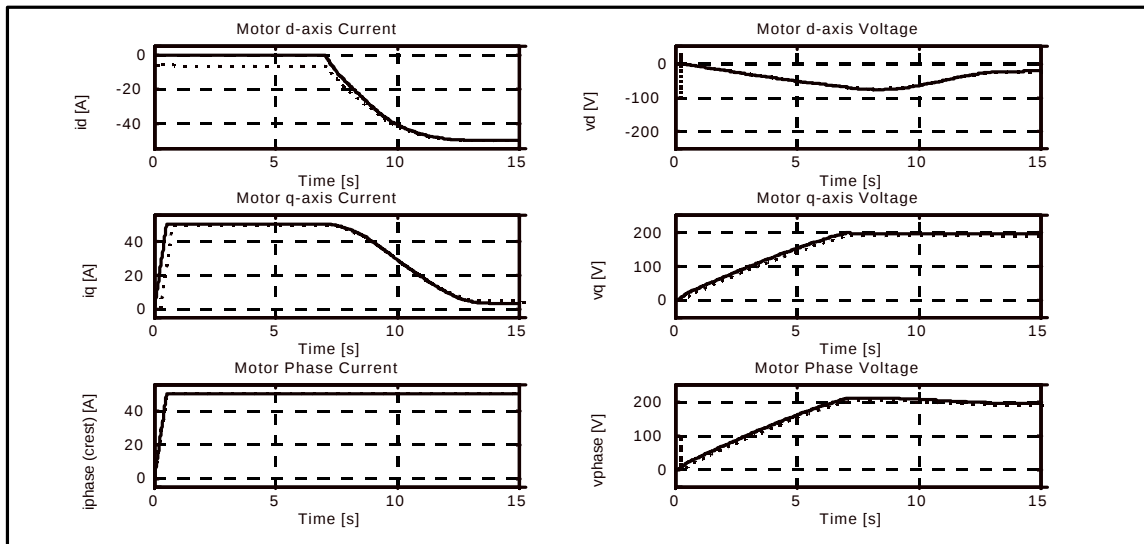
Figure 38. “Full-length” flux-weakening: voltage and current phase shifts



Legend:

- Constant Voltage Control
- - - Constant Current Control

a) Constant Power Control

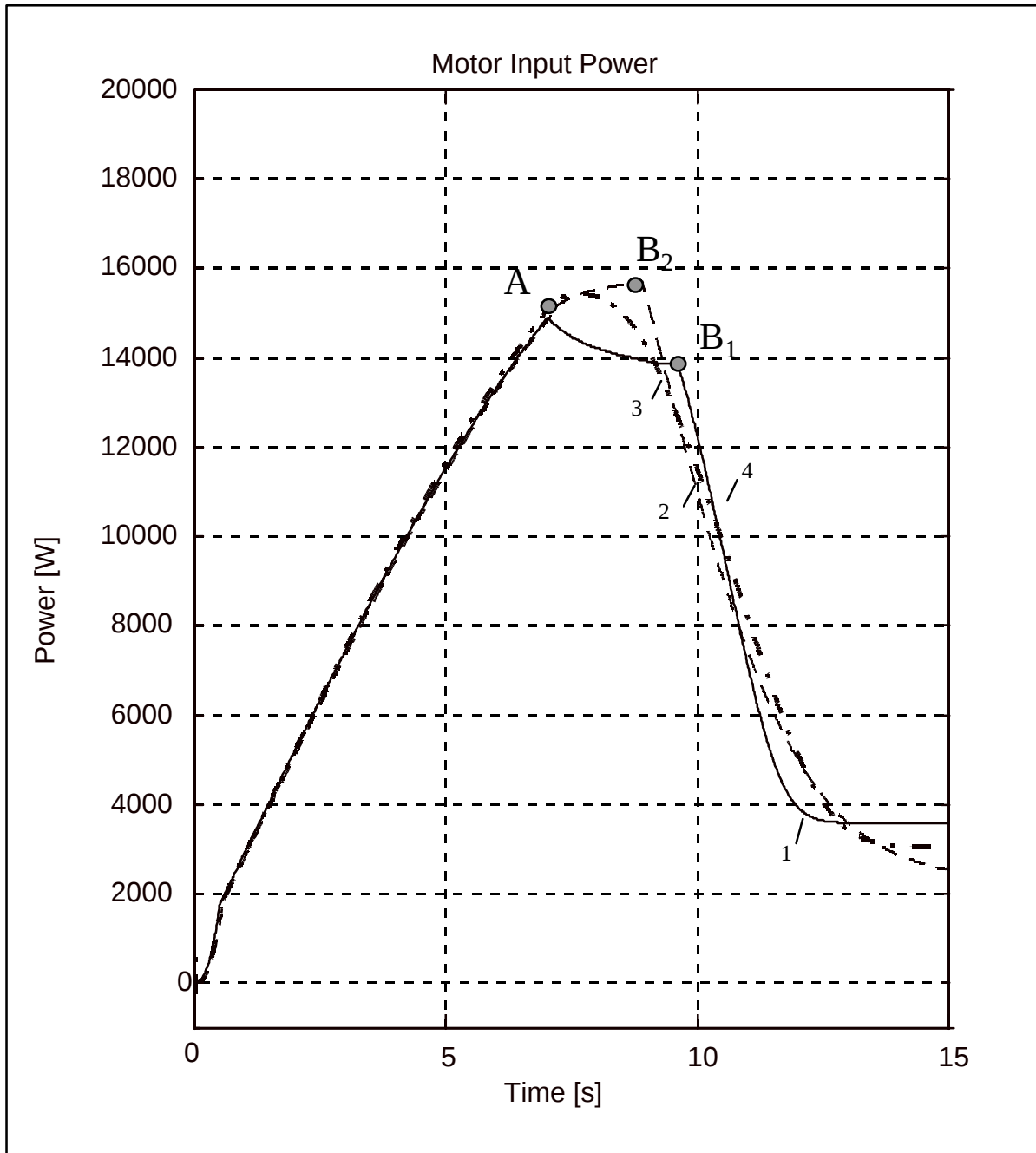


Legend:

- $I_d=0$ Control
- - - Maximum Torque Control

b) Optimum Current Vector Control

Figure 39. “Full-length” flux-weakening: voltage and current time diagrams



Legend:

- 1-Constant Voltage Constant Power Control
- 2-Constant Current Constant Power Control
- 3-Optimum Current Vector Control, $T_m = T_{mmax}$
- 4-Optimum Current Vector Control, $I_d = 0$

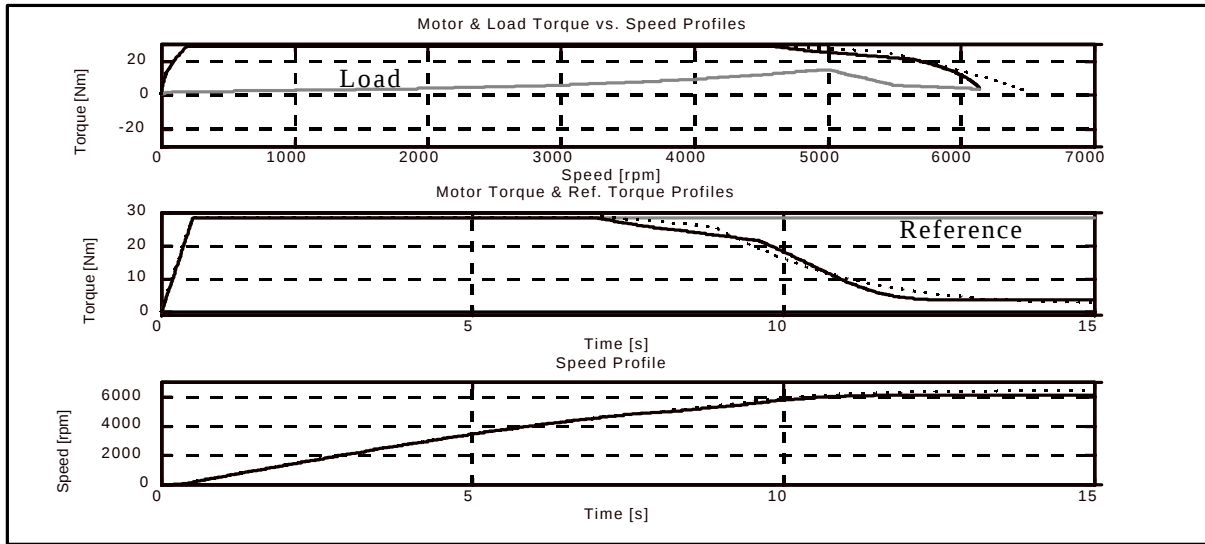
Figure 40. “Full-length” flux-weakening: input power diagrams

The superiority of the OCV control against the constant power control methods is the difference in power levels coming from the constraint of the constant power demand, which doesn't exist in the OCV control strategy. A significant power drop in the CVCP control is due to the voltage drops across the stator windings, since this control method is based on the steady-state motor equations, where the dynamics are neglected (see Eq. (77) in Section 3.2.3). Again, the lower applied power with the OCV control with maximum torque control (curve 3) than with the OCV with $i_d=0$ control (curve 4) is a consequence of the same base speed, as was discussed earlier. Table 4 shows the maximum reachable speed and corresponding time points, as well as the time comparison at the lowest maximum speed (with CVCP control). The degradation of the performance of the OCV control with maximum torque control, due to the (bad) low base speed choice, is noticeable.

Table 4. Flux-weakening control method comparison by maximum speed extensions

Control Method	Max. Speed, $\omega_{\max}$	Time @ $\omega_{\max}$	Time @ $\omega=6138$ rpm
CVCP	6138 rpm	13.3 s	13.3 s
CCCP	6467 rpm	15 s	10.8 s
OCV with $T = T_{\max}$	6377 rpm	14.5 s	10.8 s
OCV with $i_d = 0$	6559 rpm	14.6 s	10.5 s

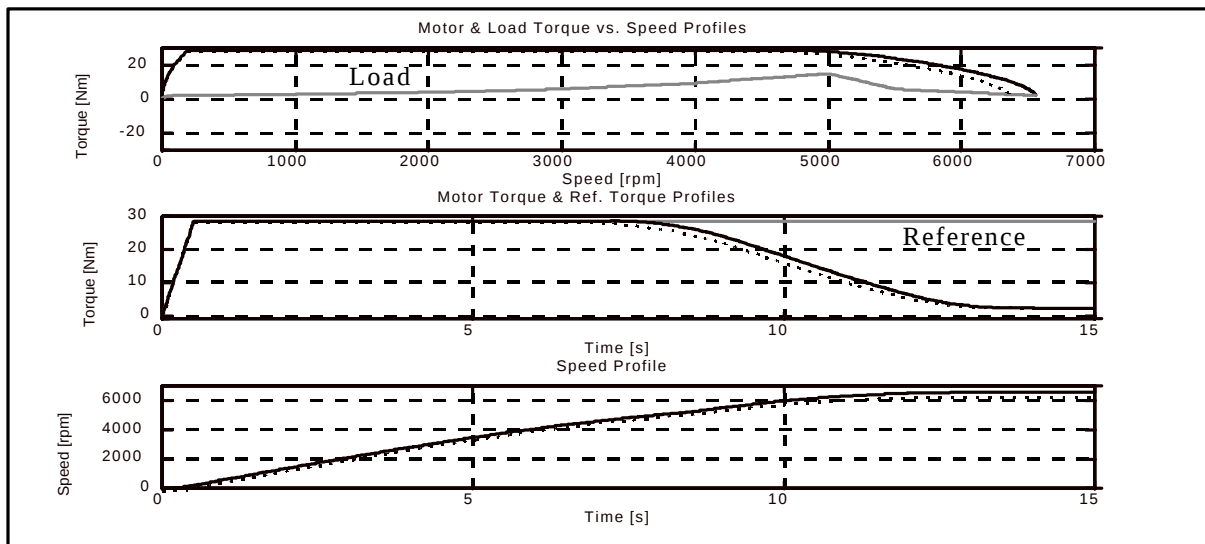
Motor electromagnetic torque and speed profiles for the four applied control methods, with the load and reference torque as referrals, are shown in Figure 41. The less constrained torque profile with the CCCP control than with the CVCP control, Figure 41.a should be noted. It is because that, besides the abovementioned power drop in CVCP strategy, the constant power constraint affects current, which is directly proportional to torque (which produces the acceleration) in the case of the CVCP control, while it affects voltage, which is proportional to speed (not to acceleration) in CCCP control. A faster maximum torque control method (in comparison with the $i_d=0$ control) loses its benefits at the flux-weakening region when the choice of the flux-weakening base speed is less than optimal, Figure 41.b.



Legend:

- Constant Voltage Control
- - - Constant Current Control
- Load & Reference Torque Profiles

a) Constant Power Control



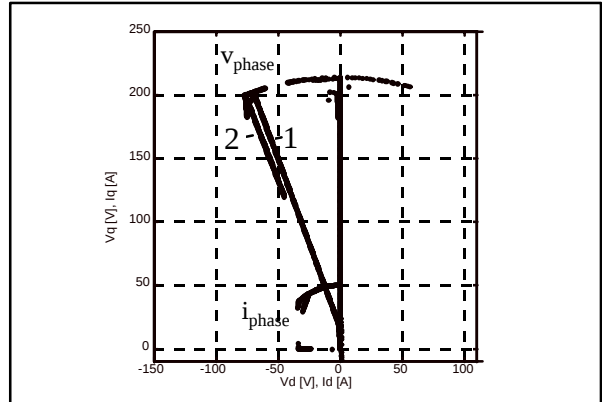
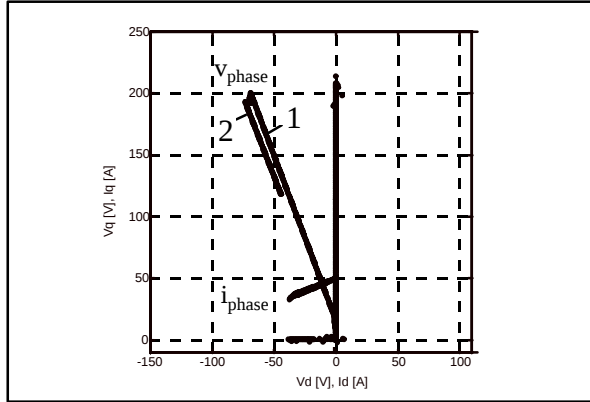
Legend:

- $I_d=0$ Control
- - - Maximum Torque Control
- Load & Reference Torque Profiles

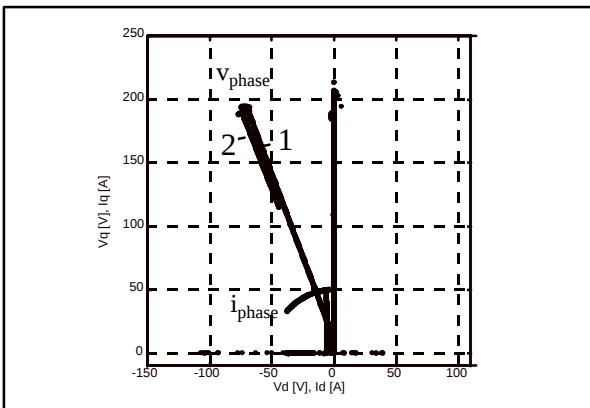
b) Optimum Current Vector Control

Figure 41. “Full-length” flux-weakening: motor torque and speed diagrams

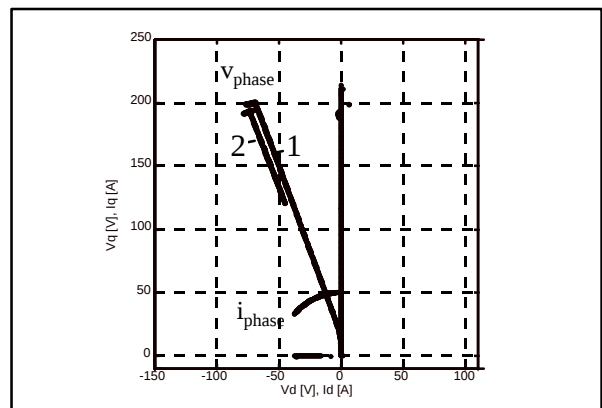
The system full start-up simulations release some other aspects of the applied flux-weakening methods. Voltage polar d-q diagrams show that the q-axis voltage component is smaller in a parallel (curve 2) than in a series operating mode (curve 1), since the back emf is lower, Eq. (72). However, that difference is attenuated when the maximum torque control is applied, because of the i_d current coupling element in Eq.s (12). In other words, the motor behavior is about the same in both operating modes. Voltage and current time-domain diagrams show the optimal phase current profile with OCV control, Figure 43.c and 43.d, and reached critical speed points with CVCP and CCCP control methods, Figure 43.a and 43.b, respectively, as well as the points of series-to-parallel switching, and motor shut-down, due to reaching the phase shift limit of -30° el. It should also be noticed that the i_d and i_q current change is much slower in CCCP control than in the other exposed control methods. The active power comparison, Figure 44, reveals that the significant differences in constant power methods, as well as OCV methods, at lower speed (series operating mode) diminish at a higher speed (parallel operating mode). The main reason is the change in the stator inductance, which is the main source of these discrepancies. The motor torque and speed profiles are shown in Figure 45. There are three things that should be noted: the highest speed for the motor shut-down with the CCCP control, the biggest divergence from the reference torque with CVCP control, and completion of the starting time requirements (30 seconds) with all four methods, with a small difference in the total acceleration time (within one second). However, Table 5 shows a significant difference in total consumed energy, where the OCV control method with maximum torque control appears to be the most economical one.



*b) Constant Current
Constant Power Control*

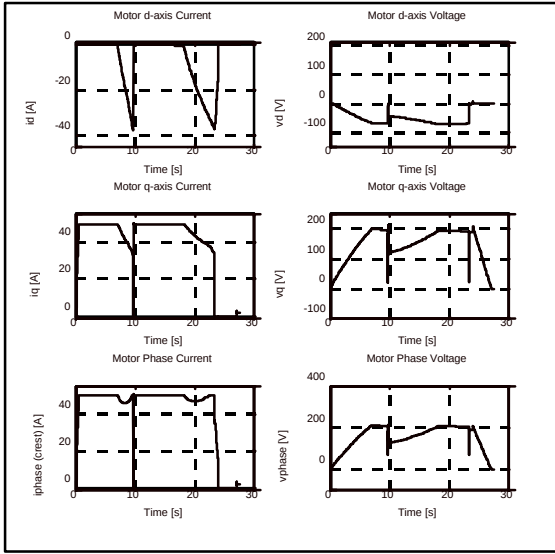


*c) Optimum Current Vector Control,
 $T_m = T_{mmax}$*

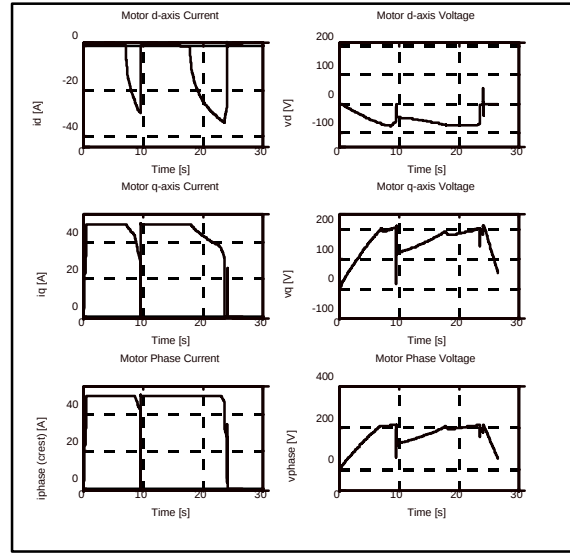


*d) Optimum Current Vector Control,
 $I_d = 0$*

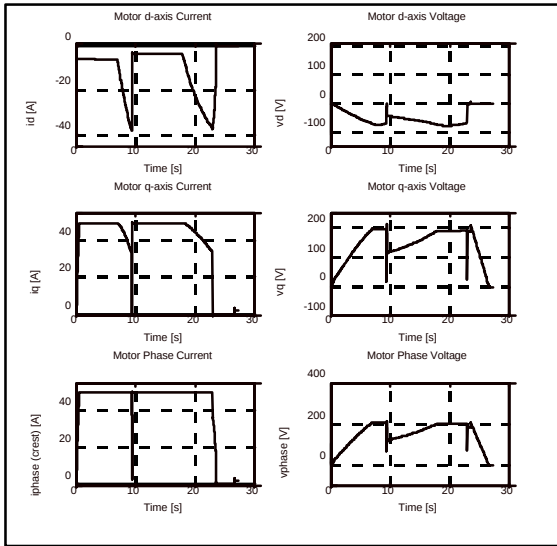
Figure 42. Full start-up: voltage and current dq polar diagrams



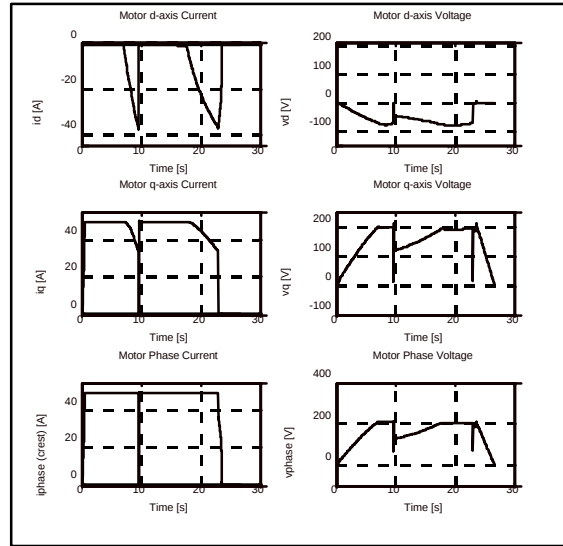
a) Constant Voltage
Constant Power Control



b) Constant Current
Constant Power Control

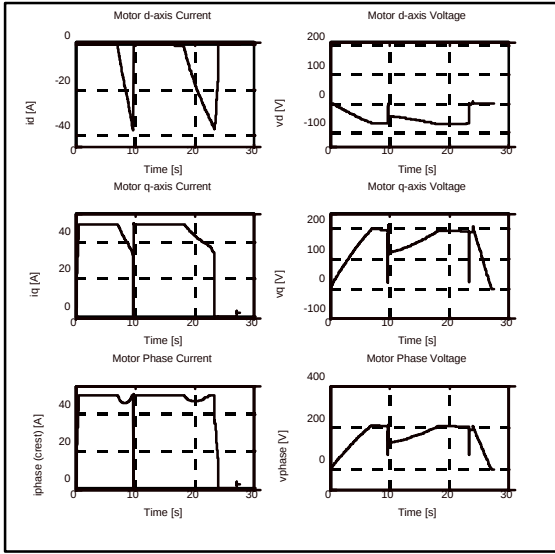


c) Optimum Current Vector Control,
 $T_m = T_{mmax}$

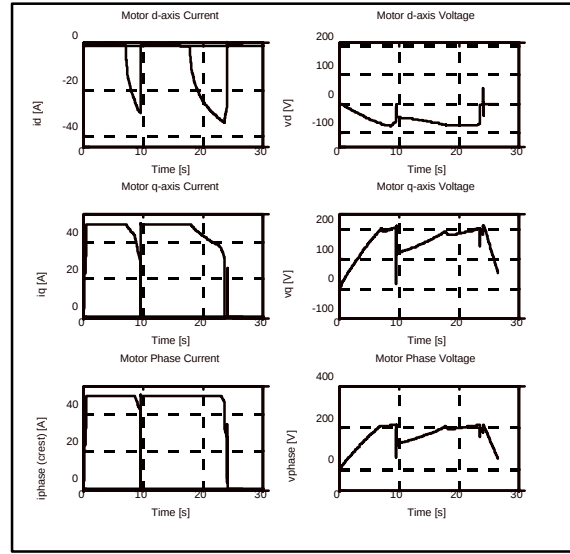


d) Optimum Current Vector Control,
 $I_d = 0$

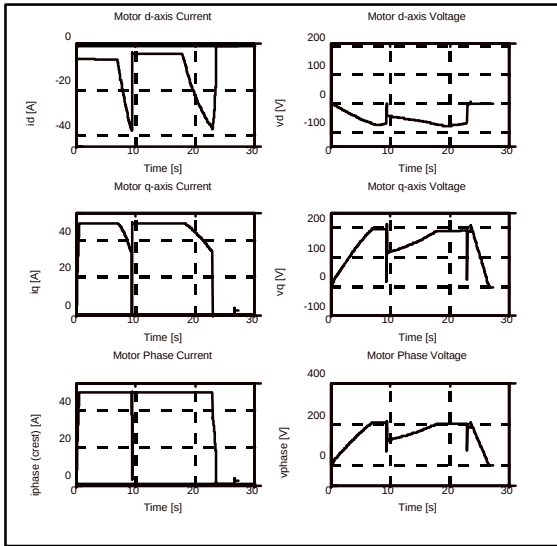
Figure 43. Full start-up: voltage and current time diagrams



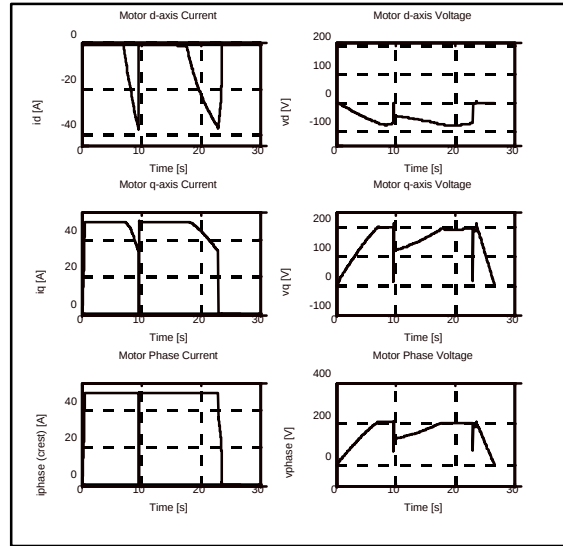
a) Constant Voltage
Constant Power Control



b) Constant Current
Constant Power Control

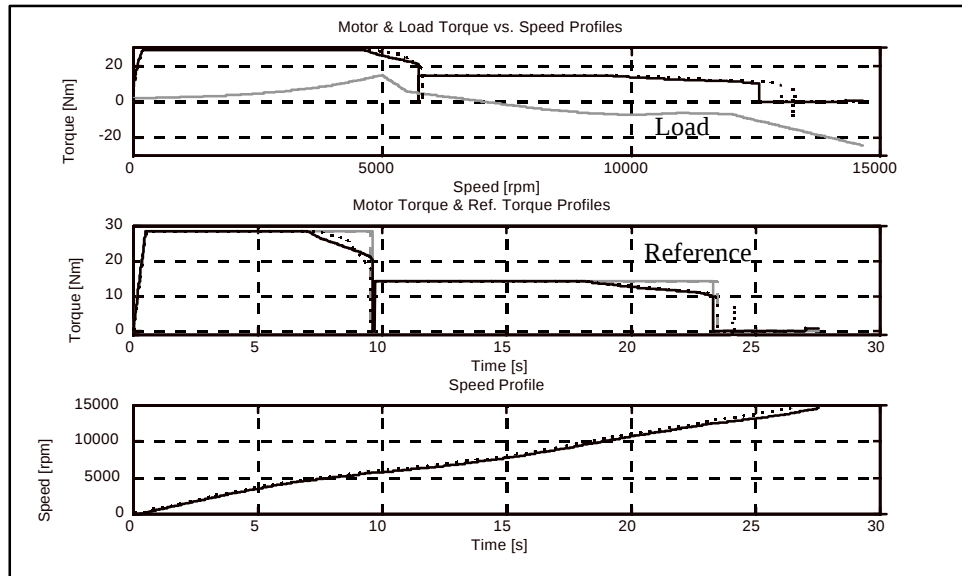


c) Optimum Current Vector Control,
 $T_m = T_{max}$



d) Optimum Current Vector Control,
 $I_d = 0$

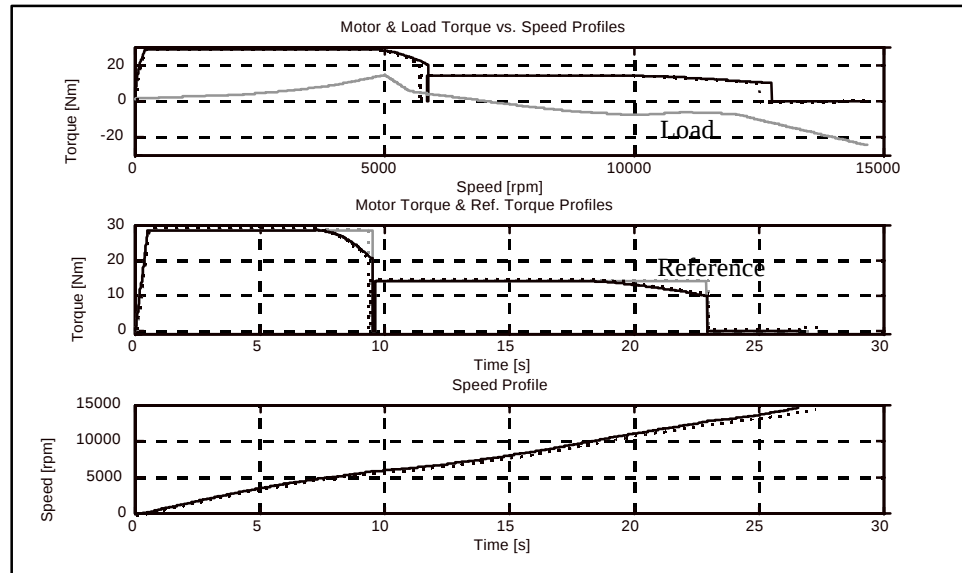
Figure 44. Full start-up: input power diagrams



Legend:

- Constant Voltage Control
- - - Constant Current Control
- Load & Reference Torque Profiles

a) Constant Power Control



Legend:

- $i_d = 0$ Control
- - - $T_m = T_{mmax}$ Control
- Load & Reference Torque Profiles

b) Optimum Current Vector Control

Figure 45. Full start-up: motor torque and speed profiles

Table 5. Flux-weakening control methods: starting time and energy consumption

Control Method	Time @ $\omega_{\max} = 14650$ rpm	Consumed Active Power
CVCP	27.51s	238.4 kJ
CCCP	26.52s	257.8 kJ
OCV with $T = T_{\max}$	27.27s	228.9 kJ
OCV with $i_d = 0$	26.58s	241.1 kJ

A few notes as a conclusion of this chapter:

- A motor drive system simulation with a verified system model has a role in a pre-prototype control design. It can predict the system behavior with a high level of accuracy, help the understanding system at a high-level, and save significant time and effort in a design process.
- Small-signal control design of the motor drive system with active load should be closely related not only to the application requirements, but also, if not even more, the load torque (dynamic) profile, which is not easily predictable in most cases.
- A starter feedback control design should take into account that the system will not work in closed loop all the time during a start-up, due to the regulator saturation, in which case the locally (open-loop) unstable system components can destabilize the whole system. In order to stabilize them, the load torque (slope) should be extracted, and/or a more suitable control method should be applied.
- Two control methods, one with back emf elimination in the decoupling loops and the other without it (equivalent DC motor), of a two-stage digital feedback controller in d-q coordinates served for comparison and were simulated in the application of a PMSM starter/generator for an aircraft APU. The simulation results proved the theoretical discussion from the previous chapter.
- A central part of the large-signal design simulations took the comparison of the four highly effective (and popular) control methods in the above mentioned application. Again, the simulation results follow the conclusions of the theoretical discussion from the previous chapter. A choice of a flux-weakening technique should depend on the system parameters and limitations,

and design requirements. The most ineffective method in one application could show up as the most effective one in another.

- Two working conditions of the abovementioned motor drive system were simulated: one with a full flux-weakening until the maximum speed extension, and the other with the full system start-up with 30° phase shift limitation. A lousy method (although not the worst one) in the first application, appeared to be the most effective in the other. A design oversight from the first application shows up as a proper design judgment in the other. These simulations proved that there is no universal optimal flux-weakening method, as could be concluded from some previously published scientific papers.