

Overview of Fundamental Frequency Sensorless Algorithms for AC Motors: a Unified Perspective

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Abstract—This paper uses active flux concept to review fundamental frequency sensorless algorithms for both induction and permanent magnet motors in one framework. Fundamentally, sensorless torque estimation can be *directly* solved using voltage model (VM) estimator, or *indirectly* solved using current model (CM) estimator. The latter turns the torque estimation problem into a speed estimation problem. The stator flux in VM and the d -axis angle in CM are deemed as the two sets of original states for sensorless drive. Through change of states, the *direct* torque estimation can be realized via observer designs; whereas the speed dependency of the unknown state (e.g., active flux and emf) gives rise to a class of speed estimation methods, known as model reference adaptive system (MRAS). The idea of a general speed observer is proposed to summarize various *separate* speed estimation methods needed for *direct* torque estimation. It is suggested to adopt inherently sensorless designs such that two-way coupling between torque estimation and speed estimation is avoided. For induction motors, it turns out the unmodelled voltage in the active flux dynamics reveals current flowing in rotor bars and can be further modelled, for which the solutions to regeneration instability problem are discussed, and change of states is recommended to attain global stability. Finally discussed are the results of slow reversal test, where local weak observability of ac motors can be potentially preserved.

NOMENCLATURE

The nomenclature we use is common for permanent magnet (PM) motors, where $L_d, L_q, R, \psi_{PM}, \psi_A$ are respectively d -axis inductance, q -axis inductance, stator resistance, PM flux linkage, and active flux linkage. For induction motors, $\psi_{PM} = 0$, and L_d, L_q, ψ_A are redefined as stator inductance, stator transient leakage inductance, and inverse- Γ circuit rotor flux. All **bold** \mathbb{R}^2 vectors are in the stator stationary $\alpha\beta$ -frame, e.g., $\mathbf{i} = [i_\alpha, i_\beta]^T$ is the measured current transformed from amplitude-invariant Clarke transformation. Park transformation is $\mathbf{P}(\theta_d) = \begin{bmatrix} \cos \theta_d & \sin \theta_d \\ -\sin \theta_d & \cos \theta_d \end{bmatrix}$ with θ_d as the d -axis angle. $\mathbf{J} = \begin{bmatrix} 0 & -1 \\ 1 & 0 \end{bmatrix}$, $\mathbf{I} = \begin{bmatrix} 1 & 0 \\ 0 & 1 \end{bmatrix}$.

I. INTRODUCTION

Sensorless or self-sensing control is typically referring to the speed regulation of inverter-fed ac motors without using a mechanical sensor for speed (i.e., tachogenerator) or position (e.g., encoder, resolver, and hall sensor), which is especially desired for high speed motors (see e.g., [1]), motors with large-diameter or hollow shaft, and motors used in adverse environment [2].

Recently, there have been several excellent reviews [3]–[5] on sensorless control of PM motors with differentiated focuses on the applications [3], the saliency based invasive methods [4], and the non-ideal conditions [5]. Among these works, the fundamental frequency sensorless algorithms are covered with limited details. In our opinion, the nuanced differences and numerous variants of those algorithms can be understood fairly easily if one pays attention to the choice of state variables as well as the considerations behind the choice.

This paper reviews all fundamental frequency models that can be adopted for designing sensorless algorithms, and provides a solid foundation to understand all design variants for both PM motors and induction motors, with the aid of the active flux concept [6]. As a matter of fact, in induction motor context, the active flux

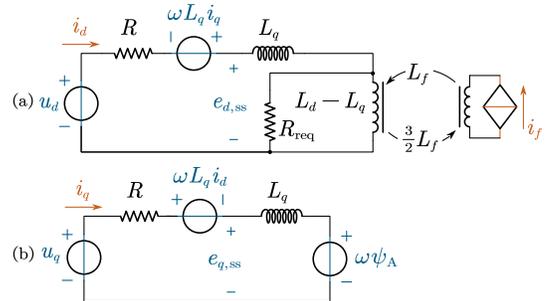


Fig. 1. The (a) d -axis and (b) q -axis equivalent circuit of the dq active flux model (6). Note that the d -axis coil is linked with the PM linkage by $L_f i_f = \psi_{PM}$.

is no new concept, and is the natural choice of state (i.e., rotor flux) with the inverse- Γ equivalent circuit [7], and it has long been found useful in analysis, e.g., of direct torque control (DTC) [8, Eq. (7)]. In salient PM motor context, active flux has other names in literature, such as fictitious PM flux [9], linear flux [10], and extended flux [11]. The key feature that makes active flux so important is the fact that it allows to model ac motor with only one inductance, L_q , that dominates how fast current can change.

A. Problem Formulation

Motor's electrical angular rotor speed ω_r is governed by Newton's second law of motion:

$$J_s n_{pp}^{-1} s \omega_r = T_{em} - T_L \quad (1)$$

where $s = \frac{d}{dt}$ is time derivative operator, J_s denotes rotor shaft inertia, n_{pp} designates pole pair number, T_L is load torque, and T_{em} is electromagnetic torque. The torque by DTC is proportional to cross product of the stator flux ψ_s and the active flux ψ_A :

$$T_{em} = \underbrace{\frac{3}{2} n_{pp} L_q^{-1} \psi_s \cdot (-\mathbf{J} \psi_A)}_{\text{DTC formulation}} = \underbrace{\frac{3}{2} n_{pp} \psi_A i_q}_{\text{FOC formulation}} \quad (2)$$

with $i_q = -i_\alpha \sin \theta_d + i_\beta \cos \theta_d$ as the q -axis current, where the field oriented control (FOC) formulation is derived by substituting the following relation between ψ_s and ψ_A [6]

$$\psi_s - L_q \mathbf{i} = \psi_A \triangleq \psi_A \begin{bmatrix} \cos \theta_d \\ \sin \theta_d \end{bmatrix} \quad (3)$$

Here, the polar coordinate definition of active flux ψ_A assumes fundamental frequency model, and the angle of active flux vector ψ_A is used as the d -axis angle for field orientation in FOC.

Sensorless control of system (1) is challenging, because neither the input T_{em} nor the output ω_r is known. The objective of a sensorless algorithm is to estimate T_{em} and ω_r . Motion equation (1) cannot be used for torque estimation because there exists an unknown term T_L , and therefore we shall seek for other models.

B. Choice of Original States

Now we are faced with the first choice of state variables. If we accept FOC formulation, making ψ_A and θ_d the states, active flux

in (3) can be described by the current-speed model (CM):

$$\psi_A = \frac{1}{\frac{(L_d - L_q)}{R_{\text{req}}} s + 1} (L_d - L_q) i_d + \psi_{\text{PM}} \quad (4a)$$

$$\theta_d = \frac{1}{s} \omega = \int (\omega_r + \omega_{\text{sl}}) dt \quad (4b)$$

with $i_d = i_\alpha \cos \theta_d + i_\beta \sin \theta_d$ as the d -axis current, R_{req} the inverse- Γ circuit rotor resistance, $\omega \triangleq s\theta_d$ the synchronous speed, and ω_{sl} the slip speed. For induction motors, there is a low pass filter (LPF) with a time constant of $\frac{L_d - L_q}{R_{\text{req}}}$ applied to the coil excitation in (4a), and only in this case the slip relation $\omega_{\text{sl}} = \frac{R_{\text{req}} i_q}{\psi_A}$ is valid. For inverter-fed PM motors, it is valid to simply put $R_{\text{req}} = \infty$, $\omega_{\text{sl}} = 0$, and $\omega = \omega_r$.

If the DTC formulation is adopted, making ψ_s the state, as per Faraday's law, the stator voltage equation in $\alpha\beta$ -frame serves as the voltage-current model (VM) of active flux:

$$\psi_A + L_q i \stackrel{(3)}{=} \psi_s = \frac{1}{s} e_s = \frac{1}{s} (\mathbf{u} - R\mathbf{i}) \quad (5)$$

where $e_s \triangleq s\psi_s$ is stator emf, and \mathbf{u} is $\alpha\beta$ -frame stator voltage.

Therefore, the torque estimation problem is solved by either:

- estimating d -axis position θ_d and ψ_A using CM (4), or
- estimating stator flux ψ_s (and ψ_A) using VM (5).

Determining d -axis angle in a feed-forward fashion as (4b) is called indirect field orientation (IFO). The IFO solves the torque estimation *indirectly* via speed estimation. Counter-intuitively, the VM in dq -frame can be used to estimate field speed ω . To see this, applying Park transformation $\mathbf{P}(\theta_d)$ to VM (5) yields:

$$s\psi_A + L_q s i_d = u_d - R i_d + \omega L_q i_q \triangleq e_{d,ss} \quad (6a)$$

$$\omega \psi_A + L_q s i_q = u_q - R i_q - \omega L_q i_d \triangleq e_{q,ss} \quad (6b)$$

where $e_{d,ss}, e_{q,ss}$ are steady-state auxiliary emfs in dq -frame, and can be used for estimating ω . Fig. 1 shows the equivalent circuit of the dq model (6), where (4a) has been elaborated as a mutual linkage and a parallel RL circuit in Fig. 1a. Fig. 1b shows the active power passing through motor air gap is $\frac{3}{2} \omega \psi_A i_q$, [cf. (2)].

On the other hand, implementing $\theta_d = \arctan 2(\psi_{\beta A}, \psi_{\alpha A})$ is called direct field orientation (DFO). The IFO and DFO use the same information, i.e., measured voltage and current, but in different manners. The key difference is that the DFO can be made explicitly independent of motor speed and speed error, because there is no speed variable in (5), which facilitates speed-independent *direct* torque estimation.

C. Change of States

Alternatively, we may select current \mathbf{i} and active flux ψ_A as the new states for torque estimation. Taking time derivative of (3) derives the active flux dynamics in (7a), and substituting (3) into the VM (5) gives the output dynamics in (7b).

$$s\psi_A = \omega \mathbf{J} \psi_A + \underbrace{(s\psi_A)}_{\mathbf{E}_u} \begin{bmatrix} \cos \theta_d \\ \sin \theta_d \end{bmatrix} \quad (7a)$$

$$L_q s \mathbf{i} = \mathbf{e}_s - s\psi_A = \mathbf{e}_s - \omega \mathbf{J} \psi_A - \mathbf{E}_u \quad (7b)$$

where \mathbf{E}_u denotes the unmodelled dynamics. From (7), ac motor appears to be modelled with one inductance, L_q . As we will see later, ψ_A is only one example of many choices of state changes, and alternative choices include the "active emf" $e_A \triangleq s\psi_A$ or the extended emf e_E in Appendix A. Nevertheless, the key idea

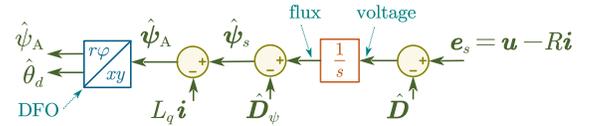


Fig. 2. Block diagram of a general stabilized VM flux estimator for DFO.

here is that by explicitly selecting the system output \mathbf{i} as state, the torque estimation problem can now be solved by designing:

- a full- or reduced-order state observer using (7); or
- a disturbance observer using only the output equation (7b).

D. Coupling with Speed Estimation

From (7), the synchronous speed $\omega \triangleq s\theta_d$ appears to be an inherent parameter of ac motors. As a result, speed estimation can be achieved by

- estimating ω_r (and T_L) as state using (1) and $s\theta_d = \omega$; or,
- estimating ω as a constant parameter (note $\omega_r = \omega - \omega_{\text{sl}}$).

Direct torque estimation using (5) offers a possibility for being independent on its ensuing *separate* speed estimation, which is preferred as stabilities of torque estimation and speed estimation can be analyzed separately. On the other hand, the speed dependencies, e.g., in model (6) and (7a), imposes to estimate torque and speed *jointly* (using either DFO or IFO), leading to complicated stability analysis. In fact, there is a convention in most sensorless PM motor literature, that is, the stability analysis of the two-way coupling between the joint estimation of torque and speed is not performed. The first part of this paper will follow this convention to help the readers better understand the literature, while rigorous analysis of the interconnected estimators is reviewed in the second part (Section V), dedicated to sensorless induction motors. Fortunately, the induction motor oriented analysis can be potentially transferred to PM motor, as is exemplified in [12].

II. DESIGN IN ORIGINAL STATES

A stator flux estimate can be obtained by integrating e_s , but the estimate $\frac{1}{s} e_s$ is unbounded in practice. There are two approaches to stabilize the integration, i.e., the time-domain approach and the frequency-domain approach, where several steady state assumptions are often resorted to, including:

- constant ψ_A assumption, thus the active flux trajectory (i.e., its Lissajous curve) is circular about origin;
- orthogonality between flux and emf: $e_A = \omega \mathbf{J} \psi_A$; and
- orthogonality between $\alpha\beta$ -axes emfs: $s e_A = \omega \mathbf{J} e_A$.

A. VM Estimator in Time-Domain

Let $\hat{\cdot}$ denote estimated value. As shown in Fig. 2, the integrator is stabilized at both its input and output:

$$\hat{\psi}_s = \hat{\psi}_A + L_q \hat{\mathbf{i}} = \int_0^t (\mathbf{u} - R\mathbf{i} - \hat{\mathbf{D}}) dt - \hat{\mathbf{D}}_\psi \quad (8)$$

where the flux compensation $\hat{\mathbf{D}}_\psi$ is used to offset the center of the estimated flux trajectory, and the voltage compensation $\hat{\mathbf{D}}$ is added to remove the unmodelled dc drift \mathbf{D} in e_s , e.g., due to thermal drift in analog devices. When there is a drift $\mathbf{D}_\psi = \int_0^t \mathbf{D} dt$ in the flux estimate, the resulting errors in position and speed estimation can be derived as in [13, Eqs. (19), (21)]. Typical generation processes of $\hat{\mathbf{D}}_\psi$ and $\hat{\mathbf{D}}$ are summarized in Fig. 3.

1) *Correction by Flux Trajectory Center*: A corollary of the constant flux amplitude assumption is that an accurate α -axis flux estimate results in a sinusoidal waveform whose maximum and

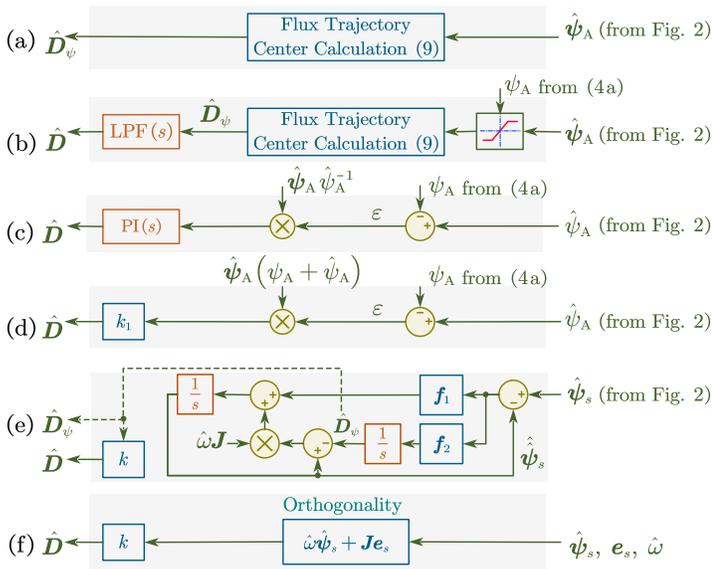


Fig. 3. Block diagrams of generation of \hat{D} and \hat{D}_ψ . Flux trajectory center calculation based (a) flux and (b) voltage compensation; (c) (d) two main variants of flux amplitude mismatch ε based voltage compensation; (e) state observer for \hat{D}_ψ with f_1, f_2 the gain matrices; (f) orthogonality based voltage compensation.

minimum add up to null. In other words, the center of the flux estimate *over one electrical cycle*, i.e.,

$$\hat{D}_\psi = \begin{bmatrix} \hat{D}_{\alpha\psi} \\ \hat{D}_{\beta\psi} \end{bmatrix} = \frac{1}{2} \begin{bmatrix} \max(\hat{\psi}_{\alpha A}) + \min(\hat{\psi}_{\alpha A}) \\ \max(\hat{\psi}_{\beta A}) + \min(\hat{\psi}_{\beta A}) \end{bmatrix} \quad (9)$$

should be $\mathbf{0}$. One can directly use \hat{D}_ψ in (9) to produce a final flux estimate as $\frac{1}{s}e_s - \hat{D}_\psi$ [13] (see Fig. 3a), or one can use it to construct a voltage compensation as $\hat{D} = \text{LPF}(s)\hat{D}_\psi$ [14] (see Fig. 3b). In fact, the voltage compensation error $D - \hat{D}$ can be exactly calculated from \hat{D}_ψ by further utilizing time information, as illustrated in [15, Fig. 4]. As is discussed in [16, Sec. III], it is better to use active flux instead of stator flux to calculate \hat{D}_ψ in (9), because stator flux components are not sinusoidal when current sudden changes.

2) *Correction by Amplitude Mismatch*: Assuming ψ_A and $\hat{\psi}_A$ are respectively available from (4a) and Fig. 2, the active flux amplitude mismatch $\varepsilon = \hat{\psi}_A - \psi_A$ can be used to generate a voltage compensation, as shown in Fig. 3c [6], [17]–[30]:¹

$$\hat{D} = \text{PI}(s) \frac{\hat{\psi}_A}{\hat{\psi}_A} \varepsilon, \quad \text{with } \text{PI}(s) \triangleq k_1 + \frac{k_2}{s}, \quad k_1, k_2 > 0 \quad (11)$$

Note for induction motors, the calculated ψ_A is often rendered as the flux command [31], [32]. VM estimator (8) plus correction (11) describes the so-called *hybrid flux estimator* that outputs the sum of high-pass filtered VM estimate and low-pass filtered “CM” estimate, and leads to the classical interpretation that VM is used for high speeds and “CM” is used for low speeds [18]–[21], [23]. It is worth pointing out there exists an abuse use of the term “CM”. In PM motor literatures, the “CM” estimate is often written in dq -frame and does not provide any angle information (see e.g., [23, Eq. (10)]), which is distinct from the CM (4) expressed in polar-

¹The flux amplitude mismatch $\varepsilon = \hat{\psi}_A - \psi_A$ from (11) has an equivalent current error form as shown in (10b), if a current estimate \hat{i} is introduced as (10a):

$$\hat{i} = L_q^{-1} \left(\hat{\psi}_s - \frac{\hat{\psi}_A}{\hat{\psi}_A} \psi_A \right), \quad \text{note: } \frac{\hat{\psi}_A}{\hat{\psi}_A} = \begin{bmatrix} \cos \hat{\theta}_d \\ \sin \hat{\theta}_d \end{bmatrix} \quad (10a)$$

$$i - \hat{i} = L_q^{-1} \hat{\psi}_A \left(\frac{\psi_A}{\hat{\psi}_A} - 1 \right) = -L_q^{-1} \frac{\hat{\psi}_A}{\hat{\psi}_A} \varepsilon \quad (10b)$$

which is used, e.g., in [25] to build a sliding mode variant of (11).

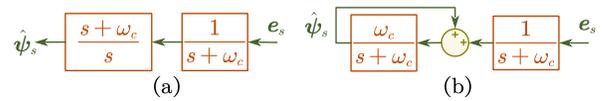


Fig. 4. (a) Multiplicative and (b) additive equivalent form of an integrator $\frac{1}{s}$. Note $\hat{\psi}_s$ is unbounded in practice for both cases.

coordinates. Discussions for selection of k_1, k_2 are documented in [17]–[19], and it is suggested in [17], [18], [31] to set k_2 to zero. Coincidentally, a contribution from the control community also suggests to not implement the k_2 integral term, but in a slightly different form from (11) as shown in Fig. 3d, to facilitate the stability analysis [33], [34]:

$$\hat{D} = k_1 \hat{\psi}_A \left(\hat{\psi}_A + \psi_A \right) \varepsilon = k_1 \hat{\psi}_A \left(\hat{\psi}_A^2 - \psi_A^2 \right) \quad (12)$$

However, note the $k_2 = 0$ design results in a non-zero bias in flux estimate in practice. The integrator $\frac{k_2}{s}$ or another dedicated “offset extractor” [32] can be used to eliminate the flux bias.

The VM estimator (8) corrected by (11) or (12) is speed-independent, but it is possible to introduce speed-dependency by transforming (8) into dq -frame. In [35], [36], the current error in (10) is used to build a speed-adaptive VM estimator in dq -frame.

3) *State Observer for Flux Offset D_ψ* : Assume there is an offset D_ψ existing in the biased VM flux estimate $\hat{\psi}_s$, which satisfies following steady state speed-dependent dynamics [37]:

$$s\hat{\psi}_s = \hat{\omega}J \left(\hat{\psi}_s - D_\psi \right), \quad sD_\psi = D \approx 0 \quad (13)$$

where it is assumed that D_ψ is slowly varying and that the orthogonality between the α - and β -axes of the flux holds. Fig. 3e shows a state observer using (13), which is also known as the *frequency-adaptive observer* in literature [23]. The observer tuning (or pole placement) in [37] is dependent on $\hat{\omega}$, thus sensitivity with respect to speed error has been analyzed. One can understand the VM along with its frequency-adaptive observer as a “single tune integrator” with a mandatory dependency on $\hat{\omega}$ [38], [39]. Readers are referred to [5, Table 4] for a review of frequency-adaptive observers for eliminating emf harmonics.

4) *Correction by Orthogonality*: The speed-dependent vector orthogonality condition $\hat{\omega}\hat{\psi}_s + J e_s = \mathbf{0}$ can be used to generate voltage compensation \hat{D} [40], [41], as shown in Fig. 3f.

B. VM Estimator in Frequency-Domain

In order to stabilize the pure integration, a high pass filter (HPF) can be added to the output of the VM estimator [42], which is equivalent to replacing the integrator with an LPF:

$$\hat{\psi}_s = \underbrace{\frac{\omega_c s}{s + \omega_c}}_{\text{HPF}} \times \frac{1}{s} e_s = \underbrace{\frac{\omega_c}{s + \omega_c}}_{\text{LPF}} e_s \quad (14)$$

which is speed-independent if a fixed cut-off frequency ω_c is used. In fact, it is reported that placing the LPF pole $-\omega_c$ to be close to zero is sufficient for zero speed operation of induction motor [43], but it is also recommended to adopt the speed-dependent tuning $\omega_c = k|\hat{\omega}|$, $k > 0$ to adjust observer damping with respect to operating speed [44], [45].

The LPF in (14) is stable but introduces undesired lag and gain. The two equivalent forms of an integrator, as shown in Fig. 4, serve as a starting point to design two types of frequency-domain compensations for (14).

1) *Multiplicative Compensation*: Let $j = \sqrt{-1}$. Allow an abuse of notation between complex number and \mathbb{R}^2 vector, i.e., $e_s =$

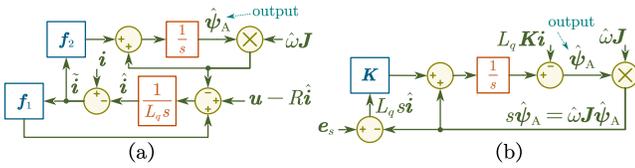


Fig. 7. (a) Full-order and (b) reduced-order state observer for active flux ψ_A .

inspired speed estimation prefers to be implemented in dq -frame. Using no correction in the CM estimator (i.e., $\hat{\theta}_d = \frac{1}{s}\hat{\omega} + 0$) is generally suggested [60], but outlier does exist, e.g., [63, Eq. (17)].

III. DESIGN VIA CHANGE OF STATES

The current dynamics (7b) are disturbed by an unknown emf disturbance $e_A \triangleq s\psi_A$. There are two ensuing ideas to deal with it, i.e., state observer and disturbance observer. The latter does not need a model for disturbance thus is potentially speed-independent; while the former becomes speed-dependent as it models the disturbance as either a flux or emf state.

A. State Observer for Flux

Using (7), a flux state observer can be implemented in full-order form [9], [10], [64], [65], or reduced-order form [66]. Fig. 7a shows that full-order observer is corrected by current error $\tilde{i} = i - \hat{i}$ with correction functions $f_1, f_2 \in \mathbb{R}^2$. Fig. 7b implies that reduced-order observer is in effect corrected by the current time-derivative error $(si - s\hat{i})$ with the gain matrix $K \in \mathbb{R}^{2 \times 2}$.

1) *Full-order Flux Observer Variants*: In [9], the observer is analyzed by Lyapunov stability theory through finding the positive-definite matrix for the Kalman-Yakubovich lemma. In [65], [67], sliding mode (SM) flux state observer that has nonlinear f_1, f_2 is proposed, and the pole placement for robustness improvement is detailed in [67]. In [64], the flux state observer is implemented in $\gamma\delta$ -frame, and in particular, the speed difference between dq -frame and $\gamma\delta$ -frame that arises during speed transients is estimated by a separate speed state observer.

2) *Reduced-order Flux Observer Design*: In [66], the reduced-order flux observer is analyzed in the misaligned $\gamma\delta$ -frame, and is described by a rotatory differential operator: $P(\theta) sP^{-1}(\theta) = sI + (s\theta)J$, where the stability analysis (that suggests $K = k_1I - k_2\text{sign}(\hat{\omega})J$) holds only if speed is already known: $\hat{\omega} = \omega$.

B. State Observer for EMF

Differentiating the active flux dynamics (7a) yields the active emf dynamics (19b) and its corresponding output equation (19a):

$$L_q s\hat{i} = u - Ri - e_A \quad (19a)$$

$$s e_A = \omega J e_A + \underbrace{(s\omega)J\psi_A + sE_u}_{\text{Unmodelled dynamics}} \quad (19b)$$

where the unmodelled dynamics appear during speed transients, but are often neglected in state observer design. The d -axis angle is recovered by integrating emf: $\int e_A dt$, or directly calculated as

$$\theta_d = -\arctan2\left(\frac{e_{\alpha A}}{\text{sign}(\omega)}, \frac{e_{\beta A}}{\text{sign}(\omega)}\right) \quad (20)$$

Note constant active flux amplitude (i.e., $s\psi_A = 0$) has been implicitly assumed in (20). Note the extended emf e_E in Appendix A can also be substituted in (20).

From the perspective of state observer design, the differences between an active emf observer and an extended emf observer are

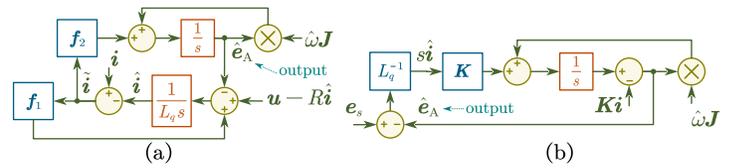


Fig. 8. (a) Full-order and (b) reduced-order state observer for active emf e_A .

minor, so there is no need to differentiate between the two, and in fact their reduced-order state observers share the same form:

$$s\hat{e}_X = \hat{\omega}J\hat{e}_X - KL_q\left(si - s\hat{i}\right) \quad (21)$$

where the subscript X is placeholder of the letter A or E. The implementation of active emf observers is shown in Fig. 8.

Full-order emf state observers are widely studied [68]–[74], while the reduced-order form stays as a classic [75], [76]. The key difference between full-order and reduced-order lies in how they treat speed error. Full-order observer reconstructs a current estimate \hat{i} , and calculate the output error \tilde{i} that is used to tune a speed estimate $\hat{\omega}$. This requires that \tilde{i} is *sensitive to speed error*. On the other hand, since current i is measured, there is no need to reconstruct an estimate \hat{i} , and the unknown state is directly estimated by (21), assuming that a speed estimate $\hat{\omega}$ is available. This needs the state estimate \hat{e}_X to exhibit *robustness against speed error* via careful observer pole placement [76].

1) *Full-order EMF Observer Variants*: Linear corrections f_1, f_2 are beneficial for analysis and tuning [64], [68]–[70]. In [68], linear system analysis is conducted to tune the observer. The $\gamma\delta$ -frame implementation of the emf observer can be found in [64], where included is an interesting study that analyzes the consequences of using constant speed assumption in emf observer. In [69], the influence of dc offset on emf estimation is analyzed. In [70], observer steady-state errors are analyzed considering current/voltage error, parameter uncertainty and filtering. Nonetheless, nonlinear SM corrections are also proposed [71]–[74], among which time-varying gain is used in [73] to reduce chattering. It is worth noting that excessive correction in current estimate dynamics reduces \tilde{i} 's sensitivity to speed error. In extreme case, \tilde{i} is forced to be zero (e.g., trapped in SM surface), so that there is no way to extract speed information from \tilde{i} (but is still possible to extract from the correction terms).

2) *Reduced-order EMF Observer Designs*: By assuming that $s e_A = 0$, Tomita *et al.* [75] proposed a reduced-order observer, for which the speed-dependent pole placement was mandatory, and was derived by analyzing the H_∞ norm of transfer function from the unmodelled dynamics to the estimated emf error. Similar H_∞ norm based pole placement is used in [76] to design a reduced-order observer for the extended emf model (37) to preserve its robustness against speed error.

C. Disturbance Observer

The idea of classical disturbance observer (DO or DOB) is to simply solve for the unknown term using the output equation (7b) or (19a), and to avoid pure differentiation, an LPF is inserted:

$$\hat{e}_A = \text{LPF}(s)(e_s - L_q s\hat{i}) \quad (22)$$

Reduced-order state observer can be deemed as a special case of (22), if LPF(s) is generalized to an $\mathbb{R}^{2 \times 2}$ transfer function matrix:

$$\hat{e}_A = \underbrace{[sI - (\hat{\omega}J - K)]^{-1}K}_{\text{In replace of the LPF}(s) \text{ in (22)}}(e_s - L_q s\hat{i}) \quad (23)$$

where the speed dependency has been introduced.

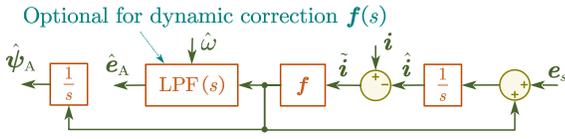


Fig. 9. Disturbance observer in output correction form for estimating active emf.

In addition to the classical form (22), DO has an alternative form being a current observer with output correction $\mathbf{f}(\tilde{\mathbf{i}})$:

$$L_q s \hat{\mathbf{i}} = \mathbf{u} - R \hat{\mathbf{i}} + \mathbf{f}(\tilde{\mathbf{i}}), \quad \text{with } \tilde{\mathbf{i}} \triangleq \mathbf{i} - \hat{\mathbf{i}} \quad (24)$$

where the correction term $\mathbf{f}(\cdot)$ or its filtered version serves as the emf estimate, as shown in Fig. 9. In this DO, the active emf e_A is treated as a disturbance with assumed dynamics, and the correction \mathbf{f} shall be modified accordingly when the assumed dynamics change. As shown in Fig. 10, the PI correction assumes e_A is constant, the proportional-resonant (PR) correction assumes e_A is sinusoidal with an angular speed of $\hat{\omega}$, and the SM correction in Fig. 10c/d assumes e_A is bounded (and the bound is known).

To our interest, we shall introduce a general form of the dynamic correction function $f_\alpha(s) \in \mathbb{R}$ for α -axis as follows

$$f_\alpha(s) = k_1 |\tilde{i}_\alpha|^{\kappa_1} \text{sign}(\tilde{i}_\alpha) + k_2 \int |\tilde{i}_\alpha|^{\kappa_2} \text{sign}(\tilde{i}_\alpha) dt \quad (25)$$

The correction f_α is said to be *non-dynamic* if $k_2 = 0$ and said to be *dynamic* if $k_2 > 0$. “Dynamic” means f_α has internal state. In fact, since the integral term in (25) can be considered as an “extended state”, the DO with dynamic correction is also called extended state observer (ESO). If $\kappa_1 < 1$, the correction is nonlinear function of \tilde{i}_α , causing undesired dither in emf estimate, known as chattering.

1) *Linear DO*: From (25), assuming constant emf disturbance $se_A = \mathbf{0}$ and putting $\kappa_1 = \kappa_2 = 1$, result in linear DO. In [77], non-dynamic correction is implemented. In [78], dynamic PI correction is used. The linear DO is deemed to be poorly damped if constant correction gains k_1, k_2 are used [5]. Frequency domain formulation of the DO with PI correction is proposed in [79], which is further generalized for salient PM motors in [69].

The constant emf assumption $se_A = \mathbf{0}$ is not true [see (19b)], but it becomes reasonable if one transforms (24) into dq -frame, such that dq -frame emf is modelled as dc disturbance and the correction \mathbf{f} becomes $\mathbf{P}\mathbf{f}$. In [80]–[82], $\mathbf{P}\mathbf{f}$ is implemented as a PI regulator. In [83], classical DO form (22) with LPF is used to estimate extended emf e_E in dq -frame². Note speed dependency is introduced in dq -frame linear DO.

2) *SMDO*: The working principle of SMDO is to use the correction term \mathbf{f} to force $\hat{\mathbf{i}}$ to track \mathbf{i} . To make this happen, the norm of \mathbf{f} should be larger than that of e_A no matter how small the output error $\tilde{\mathbf{i}}$ is. For example, the classical SMDO uses a non-dynamic *SM control law*: $\mathbf{f} = k_1 \text{sign}(\mathcal{S})$ with simple *SM surface* $\mathcal{S} = \tilde{\mathbf{i}}$ and constant *SM gain* k_1 , such that the norm $\|\mathbf{f}\|$ equals to k_1 as long as $\tilde{\mathbf{i}} \neq \mathbf{0}$. Since $\|e_A\|$ is proportional to ω , and to assure a wide operating speed range, large SM gain k_1 is mandatory, implying DO performance varies as operating speed changes.

By using large but constant SM gain k_1 , SMDO is speed-independent, but there are excessive noises in the “raw” estimated emf disturbance \mathbf{f} , known as chattering issues. To mitigate the chattering, it is proposed to use “milder” SM control law than the discontinuous correction $\text{sign}(\cdot)$, such as saturation function [84], and sigmoid function [85]–[87]. Another idea is to design \mathbf{f} as a linear combination of SM correction and proportional correction

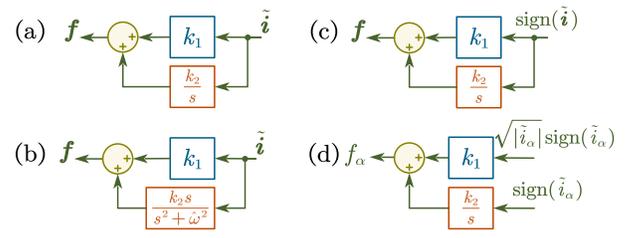


Fig. 10. Example correction function \mathbf{f} used in disturbance observer. (a) PI, (b) PR, (c) discontinuous SM, and (d) continuous SM correction.

[84], [87]. However, those “milder” SM control laws break the Lyapunov stability of SMDO.

Alternatively, without modifying the SM control law, an additional LPF can be applied to \mathbf{f} , before $\hat{\theta}_d$ is extracted. The phase delay caused by the LPF should be compensated for different speeds [88]–[90]. Given a constant SM gain, it is suggested that the pole of the LPF should rely on $\hat{\omega}$ [84], as indicated in Fig. 9.

To avoid speed-dependency, it is recommended to implement \mathbf{f} as a dynamic correction that puts signum inside an integral like the k_2 term in (25). This integral introduces an additional state, resulting in a second-order SMDO, and the newly introduced state can be interpreted as an estimate of the emf appearing in the current dynamics. For example, the super-twisting algorithm can be used to estimate the emf [91]–[93] [94, Sec. V], and the resulting *continuous* α -axis correction is $f_\alpha = k_1 |\tilde{i}_\alpha|^{\frac{1}{2}} \text{sign}(\tilde{i}_\alpha) + k_2 \int \text{sign}(\tilde{i}_\alpha) dt$, where the integral term serves as a continuous estimate of the emf, as shown in Fig. 10d. As a comparison, the second-order SMDO proposed in [95] uses *discontinuous* α -axis correction as $f_\alpha = k_1 \text{sign}(\tilde{i}_\alpha) + k_2 \int \text{sign}(\tilde{i}_\alpha) dt$, as shown in Fig. 10c. Note the continuity of f_α is decided by the k_1 term, and also note applying variable gain $k_1 |\tilde{i}_\alpha|^{\frac{1}{2}}$ to $\text{sign}(\tilde{i}_\alpha)$ does not eliminate chattering because the derivative of $|\tilde{i}_\alpha|^{\frac{1}{2}}$ is infinite when $\tilde{i}_\alpha = 0$. The chattering is reduced because of using smaller k_1 value and a nonzero k_2 term (that takes advantage of the history of output error $\tilde{\mathbf{i}}$).

So far, all SMDOs reviewed above use current error $\tilde{\mathbf{i}}$ as the SM surface \mathcal{S} . Improvement is expected by further designing SM surface. For example, SM surface involving the integral of current error $\int \tilde{\mathbf{i}} dt$ can eliminate the reaching phase [96]–[98] [99, p.287]. SM surfaces involving the derivative of current error $s\tilde{\mathbf{i}}$ promise “convergence in finite time”, such as the fast terminal SM manifold [100] and the non-singular terminal SM manifold [89]. Note the derivative can be reconstructed by super-twisting algorithm [89].

Besides, the SM gain can be designed to be a function of \mathcal{S} [98], which is also known as the reaching law design. A less common idea to reduce SM gain is to execute the numerical integral of SMDO at a frequency that is three times as high as the pulse-width modulation (PWM) frequency [88], [101].

While most SMDO designs are speed-independent, it is possible to introduce speed dependency. First, similar to dq -frame linear DO, one can implement SMDO in dq -frame, and the dq -frame correction $\mathbf{P}\mathbf{f}$ in [102] is implemented as a combination of proportional correction and SM correction. Second, in [103], in addition to the dynamic correction $\mathbf{P}\mathbf{f}$, a speed-dependent SM surface that consists of current error and its integral is further selected. Third, the SMDO in [104] is designed based on the speed-dependent output equation of the extended emf model (37a). Finally, there are also examples of using speed-dependent tuning for second-order SMDO [93, (10)].

²The analysis in [83] is in $\gamma\delta$ -frame but the implementation is in dq -frame. For an example of $\gamma\delta$ -frame observer implementation, see [64, Eqs. (17)].

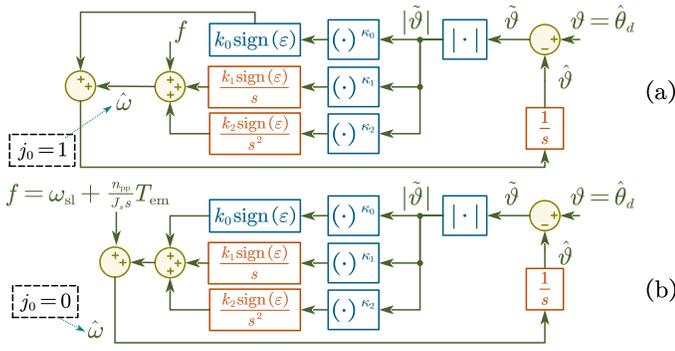


Fig. 11. The same 3rd-order position observer that generates two different rotor speed estimates. (a) Luenberger observer and (b) PID-like acceleration ($s\hat{\omega}$).

IV. SEPARATE SPEED ESTIMATION

This section focuses on DTC or DFO controlled drive, such that torque estimation is already done, and an estimate of some unknown state ($\hat{x}_x, \hat{\psi}_A, \hat{\theta}_d$) is available. This estimate serves as the reference input (denoted by ϑ or ϑ) to the *separate* speed estimation. ‘‘Separate’’ means the speed estimation design assumes the reference input signal is independent of the speed estimate, i.e., it is *inherently sensorless* [45], neglecting speed error’s influence on reference input.

A. Direct Calculation

Steady state dynamics can be used to directly calculate speed, which provides a static but instantaneous speed estimate.

1) *Direct Calculation from Emf*: Speed is simply the emf amplitude divided by the flux amplitude [89], [91]. Note this direct calculation motivates the CM estimator (16).

2) *Direct Calculation from Position*: Speed estimate can be calculated by the forward difference of the flux angle $\hat{\theta}_d$ [10], [13]. An additional LPF is embedded to reduce the amplified noise [13].

3) *Direct Calculation from Flux Orthogonality*: Multiplying $-(\mathbf{J}\hat{\psi}_s)^\top$ to both sides of the steady-state dynamics $s\hat{\psi}_s = \hat{\omega}\mathbf{J}\hat{\psi}_s$ gives a speed estimate: $\hat{\omega}(\hat{\psi}_s^\top \hat{\psi}_s) = -(\mathbf{J}\hat{\psi}_s)^\top s\hat{\psi}_s$, where the derivative $s\hat{\psi}_s$ can be substituted by the forward difference of flux [105], or the calculated stator emf as $e_s = \mathbf{u} - R\mathbf{i}$ [44], [106], or the voltage/emf estimate by SMDO [100], [107]–[111].

B. General Observer for Speed

Assuming a scalar reference signal denoted as $\vartheta \triangleq \hat{\theta}_d$ is available from prior flux/emf estimation, then the following $(n + 1)$ -th order general state observer for ϑ summarizes a wide class of speed estimation methods:

$$s\hat{\vartheta} = f(\vartheta) + \sum_{j=0}^n k_j \frac{|\tilde{\vartheta}|^{\kappa_j} \text{sign}(\tilde{\vartheta})}{s^j}, \quad \text{with } k_j, \kappa_j \geq 0 \quad (26)$$

$$f(\vartheta) = \omega_{sl} + \frac{n_{pp}}{sJ_s} T_{em}$$

where the output error is $\tilde{\vartheta} = \vartheta - \hat{\vartheta}$. Similar generalization like (26) has been studied in literature, known as generalized PI observer (GPIO) [112] or generalized ESO (GESO) [113].

The position observer (26) does not define a speed estimate, and there are potentially $n + 1$ variants of speed estimates [114]:

$$\hat{\omega}_{j_0} = f(\vartheta) + \sum_{j=j_0}^n k_j \frac{|\tilde{\vartheta}|^{\kappa_j} \text{sign}(\tilde{\vartheta})}{s^j} \quad \text{with } j_0 = 0, 1, 2, \dots, n \quad (27)$$

Table I gives a list of example implementations of (26). Fig. 11 shows block diagrams of (26) and (27) when $n = 2$, $j_0 = 0, 1$.

1) *State Observer Variants*: Position state observer (26) often uses motion equation (1) to render an inertia dependent $f(\vartheta)$ term. Linear corrections ($\kappa_j = 1$) are used in Luenberger observers [64], [80], [115], [116] [117, Fig. 9] and ESOs [90], [118], but ESO also allows nonlinear corrections [119], [120].

Alternatively, one can design a state observer for q -axis current by redefining ϑ and f in (26) accordingly, for either PM motors [18] or induction motors [106]. Such scalar output observer can be considered as a reduced-order implementation of the original natural observer [121], or it can be understood as an ELO implementation of the q -axis current observer [61]. The difference is that the natural observer for $\vartheta = i_q$ utilizes motor’s active power error $|u_q|\tilde{i}_q$ as output error, which, however, makes observer tuning to be dependent on q -axis voltage u_q thus time-varying [18].

2) *Phase-Locked Loop (PLL)*: PLL is widely used for extracting frequency information from d -axis angle error. During formulation of PLL, various types of position error signals can be exploited, such as the q -axis voltage [48], the angle of $\gamma\delta$ -frame extended emf [81], d -axis current error [61], and the forward difference of high frequency components of $\alpha\beta$ -frame current [116]. PLL for induction motors would need to add an additional slip relation [45] such that $f(\vartheta) = \omega_{sl}$. Speed error during speed transients is inevitable because typical PLL is 2nd-order system that assumes constant speed [122], which is known as a type-2 system [37]. The transfer function from actual position to estimated position can be found in [83, Eq. (18)] and [81, Eq. (39)]. The PLL speed error during speed transients is analyzed in [78]. In [26], an additional PI term driven by torque error is further added to the PLL based speed estimate. PLL can be generalized for higher order to track time-varying speed [66, Eq. (24)], and see also [30] and [83, Eq. (22)] for an application of type-3 PLL for tracking ramp speed signal.

3) *Extended Kalman Filter (EKF)*: In [71], [72], EKF is applied to a 3rd-order system with position, speed and acceleration as states, without needing inertia parameter J_s . This EKF is similar to type-3 PLL but has time-varying gains.

C. Model Reference Adaptive System (MRAS)

Two different concepts of MRASs are shown in Fig. 12. Fig. 12a shows an MRAS whose reference and adjustable model

TABLE I
EXAMPLES OF THE GENERAL POSITION AND SPEED STATE OBSERVER (26)

Speed estimation	Order n	Gain design	Alternative names	Assumed motion dynamics
Linear correction	2	Const. ($k_2 = 0$)	PLL, Adaptive system	$\omega_r = \text{Const.}, f = \omega_{sl}$
	3	Const.	Generalized PLL	$\omega_r = kt, f = \omega_{sl}$
	3	Time-varying	EKF, ELO	$\omega_r = kt, f = \omega_{sl}$
	3	Const.	Speed observer, LESO, PI observer	$T_L = \text{Const.}, f = \omega_{sl} + \frac{n_{pp}}{sJ_s} T_{em}$
	4	Const.	Generalized LESO, Generalized PI observer	$T_L = kt, f = \omega_{sl} + \frac{n_{pp}}{sJ_s} T_{em}$
Nonlinear correction	1	Const. ($k_1 = k_2 = 0$)	1st-order SMDO ($\kappa_0 = 0$)	ω is bounded
	2	Const. ($k_2 = 0$)	2nd-order SMDO ($\kappa_0 = \frac{1}{2}, \kappa_1 = 0$)	T_L is bounded
	3	Const.	Nonlinear ESO ($\kappa_0 = 1, \kappa_1 = \frac{1}{2}, \kappa_2 = \frac{1}{4}$)	$T_L = \text{Const.}, f = \omega_{sl} + \frac{n_{pp}}{sJ_s} T_{em}$

PLL: phase locked loop. EKF: extended Kalman filter. LESO: linear extended state observer. ELO: extended Luenberger observer.

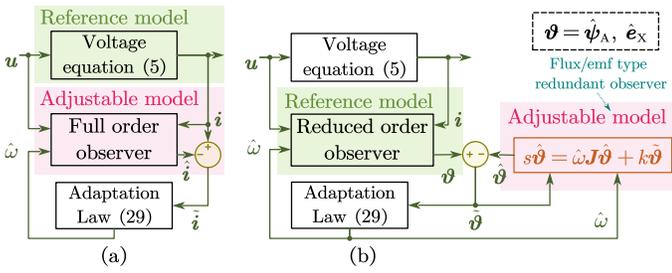


Fig. 12. MRAS with (a) full-order observer as adjustable model; (b) reduced-order observer as reference model and redundant observer as adjustable model. The key difference between the two examples of MRAS is whether its reference model exhibits explicit dependency on speed estimate $\hat{\omega}$.

are respectively the actual motor and the speed-adaptive full-order state observer. Fig. 12b shows an MRAS whose reference and adjustable model are respectively the reduced-order state observer and a speed-adaptive redundant observer.

1) *Redundant Observer*: Assuming an accurate estimate of the unknown states, denoted by ϑ (recall $\vartheta = \hat{e}_x, \hat{\psi}_A$), is available, a redundant observer reconstructs the estimate only for extracting speed information from ϑ via an adaptation law. Redundant observer is corrected by the mismatch $\tilde{\vartheta} \triangleq \vartheta - \hat{\vartheta}$:

$$s\hat{\vartheta} = \hat{\omega}J\hat{\vartheta} + k(\vartheta - \hat{\vartheta}) \quad \text{with } \vartheta = \hat{\psi}_A, \hat{e}_A, \hat{e}_E \quad (28)$$

Fig. 12b shows how the redundant observer is used as an adjustable model in MRAS, including the emf-type [69], [76], [77], [86], [87], [97], [98], and the flux-type [65].

2) *Speed Adaptation Law*: Adaptation law for field speed is

$$\hat{\omega} = \text{PI}(s) \times \text{Regressor} \times \text{Output error} \quad (29)$$

where the output error can be the current error \hat{i} from full-order observer (Fig. 12a), the flux/emf mismatch $\tilde{\vartheta}$ from redundant observer (Fig. 12b), or any other mismatch, e.g., the active flux amplitude mismatch ε in (10b) [62]. Speed adaptation assumes speed is a constant parameter, but an inertia-dependent speed-adaptation law that includes an additional torque term is also proposed [61], [63], [105]. The selection of the regressor is critical and it often depends on how speed appears in the adopted model: example choices include active flux [9] and extended emf [68]. It is shown in [62] that decoupling of analysis between flux estimation and speed estimation is achieved via linearized model based observer tuning and wise selection of speed regressor. However, the fact is that the generation of speed regressor offers a degree-of-freedom to possibly find Lyapunov function, hence eliminating the need of linearization, which will be reviewed in the next section.

V. LOSE THE CONSTANT ψ_A ASSUMPTION

As per (4a), ψ_A is time-varying as long as $(L_d - L_q)i_d$ varies, and a change in ψ_A leads to unmodelled dynamics E_u in (7a) and (19b), which will affect the performance of state observer as well as any other method relying on constant ψ_A assumption. In practice, time-varying ψ_A could be beneficial for improving operating efficiency or dynamic performance. For example, at high speeds, faster speed dynamic process can be achieved if the flux amplitude is first weakened such that more dc bus voltage becomes available for producing torque current [123].

For PM motors, the strategy is mainly to compensate the unmodelled dynamics. In [11], an angle compensation that takes into account the unmodelled dynamics in (19b) when ψ_A varies is proposed. In [9], the si_d in the unmodelled dynamics E_u in (7a) is compensated by its estimate.

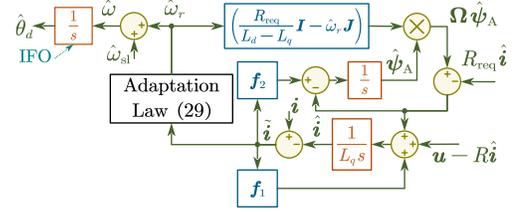


Fig. 13. Adaptive full-order observer for induction motor model (31).

For induction motors, the voltage E_u is closely related to the voltage drop on the rotor resistance, and hence it can, in fact, be modelled. This section is dedicated for induction motors.

A. Induction Motor Model with Time-Varying ψ_A

For induction motors, the unmodelled dynamics E_u satisfy

$$E_u = P^{-1}(\theta_d) \begin{bmatrix} s\psi_A \\ 0 \end{bmatrix} \stackrel{(4a)}{=} -R_{\text{req}} \underbrace{\left(\frac{\psi_A}{L_d - L_q} - i \right)}_{\text{the rotor current}} - \omega_{sl} J \psi_A \quad (30)$$

in which (4a) and (3) have been substituted, and from which the slip relation $\omega_{sl} = \frac{R_{\text{req}} i_q}{\psi_A}$ is derived. The model with stator current and active flux as states is derived from (5), (7a) and (30):

$$L_q s i = u - R i - s \psi_A \quad (31a)$$

$$s \psi_A = R_{\text{req}} i + \frac{-R_{\text{req}}}{L_d - L_q} \psi_A + (\omega - \omega_{sl}) J \psi_A \quad (31b)$$

$$\triangleq R_{\text{req}} i - \Omega \psi_A \quad (31c)$$

which is exactly the inverse- Γ circuit induction motor model, with active flux ψ_A being equivalent to the rotor flux, L_d denoting stator inductance, L_q designating transient leakage inductance, and $\Omega \triangleq \frac{R_{\text{req}}}{L_d - L_q} I - \omega_r J$. There are two facts that make induction motors unique:

- 1) Active flux amplitude ψ_A is not constant and is maintained by stator excitation for non-PM motors.
- 2) There is a slip speed ω_{sl} difference between field speed ω and rotor speed ω_r , i.e., $\omega_r = \omega - \omega_{sl}$.

In other words, the induction motor is not a PM motor nor synchronous motor. These two facts correspond to the two unique features of sensorless induction motors:

- 1) There is a chance for estimated flux amplitude $\hat{\psi}_A$ to collapse [124]. There is also a chance for a change in flux amplitude being mis-interpreted as a change in flux angle, which is revealed in the ‘‘basic relations for sensorless flux estimation’’ [45, Eq. (13)]. As a result, unstable sensorless operation that is not observed in PM motors might happen (e.g., at low speed regenerating operation).
- 2) Zero speed operation under load does not mean zero frequency operation for induction motors.

B. The Common Act of Choosing Active Flux as State

Many researches have been devoted for achieving speed estimation using the model (31). Note the estimated flux is seldom used for Park transformation, and in most cases, the IFO d -axis angle $\frac{1}{s}\hat{\omega}$ is used to define the dq -frame of an induction motor.

1) *State Observer Design Attempts*: Using (31), speed-adaptive observer can be designed in $\alpha\beta$ -frame [67], [105], [125]–[131] or dq -frame [63], [132]. Linear correction is widely used, but SM correction can also be used [67], [105], [130]. Fig. 13 shows a typical implementation. Time-varying correction functions are used in EKF and ELO, where speed is treated as an extended state so that linearization is required [133, Sec. 4.5.3.5].

For better dynamic performance, ω_r is further treated as observer state in dq -frame SM state observer [134] with d -axis

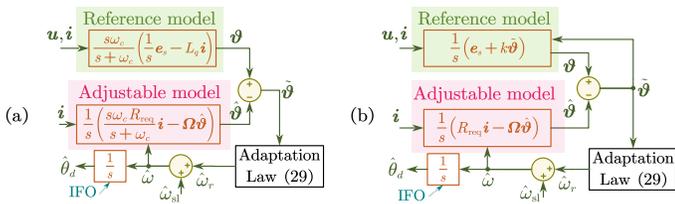


Fig. 14. Flux type MRAS for IFO of sensorless induction motor, where the VM is stabilized by (a) the HPF $\frac{s\omega_c}{s+\omega_c}$, or (b) the mismatch $\tilde{\vartheta}$. Note reference model has no explicit speed dependency (cf. Fig. 12b).

current error chosen as output error. A different approach, however, is to put all the correction terms into the dynamics of the load torque estimate \hat{T}_L and uses an exact copy of motor dynamics (31) and (1) to establish a so-called natural observer [121].

2) *Stability Challenges at Low Speed Regeneration*: It is quite a challenge to stabilize the full-order state observer for (31) in low speed regeneration even when speed is observable. Classical designs are flawed. For example, the stable speed adaptation law derived from Lyapunov function [127] depends on the unknown flux error; and the hyper-stability analysis in [125] depends on the assumption that the ratio between flux error norm and current error norm has finite upper bound (so that the unknown flux error can be replaced by current error). Careful observer gain designs based on the linearized model (see, e.g., [128], [131], [135], [136]) and the positive real property [137] are proposed for extending stable operating region. Alternatively, it is also effective to re-design the speed adaptation law [129], [138], [139] [140, Eq. (26)]. Readers are referred to [141] for a survey of the stabilization methods.

3) *MRAS*: In the field of sensorless induction motor drive, MRAS is a jargon for a system with the VM (5) being reference model and the new CM (31c) being adjustable model. The pioneer work on MRAS in [42] (see also the follow-up works [142], [143]) chooses VM estimator in (14) as reference model, and implement CM with high-pass filtered i as adjustable model, as shown in Fig. 14a. As a result, the obtained flux estimate has significant delays with respect to the actual flux thus it cannot be used for DFO. This is a good example showing the spirit of the IFO nature of sensorless induction motors, that an accurate estimate of flux/emf is not needed, because the IFO d -axis angle estimate only depends on the speed estimate. However, there are also attempts to compensate the gain and lag introduced in (14) [144].

Alternatively, one can stabilize VM estimator using the mismatch $\tilde{\vartheta}$ between outputs of VM and CM estimators as correction. Fig. 14b shows an example of proportional correction $k\tilde{\vartheta}$. [145]. The MRAS implementation in [19] has made it clear that the amplitude mismatch is used to stabilize VM estimator, while the angle mismatch is used to tune $\hat{\omega}$ used in CM estimator. In [146], the VM is transformed to an estimated dq -frame to derive a generalized slip relation in terms of the VM correction gains. The dual reference observer proposed in [147] is also an MRAS, and is an example of implementing the amplitude correction (11) in its current error form (10b) in a time-varying ψ_A model. In [143], the speed adaptation law is implemented as an SM control law. In

[148], the correction in VM is replaced with a super-twisting based dynamic correction. Readers are referred to [149] for a dedicated review of MRAS variants having different choices of output errors.

C. State Observer Design via Change of States

To attack the regeneration instability challenge, an ideal solution is to find a globally stable speed-adaptive observer design. In the classics of adaptive observer design [150]–[152], stable observer design is often proposed for a class of state-affine systems, which do not describe the induction motor dynamics (31). In fact, it has been shown in [153] that in order to find a Lyapunov function for the full-order state observer with (i, ψ_A) as states, the observer coefficients must be dependent on the actual speed, implying globally stable design does not exist. Thus to obtain global stability, one needs to either find a proper state transformation that can describe the induction motor as a system in the *adaptive observer form* [151]

$$s \begin{bmatrix} i \\ x \end{bmatrix} = \begin{bmatrix} \mathbf{0} & \mathbf{I} \\ \mathbf{0} & \mathbf{0} \end{bmatrix} \begin{bmatrix} i \\ x \end{bmatrix} + \Phi(i, u)\omega_r + \Phi_0(i, u) \quad (32)$$

where $\Phi \in \mathbb{R}^4$ is the regressor of speed that consists of only known signals (i, u) , and x is the new unknown state; or develop advanced observer design applicable to a wider class of systems that allow unknown speed regressor $\Phi(x)$ [154], [155].

1) *Change of States*: A list of state transformations for induction motors is given in Table II. Model 1 is (31). Integrator backstepping based observer design is proposed for model 2 ($x = e_A$) in [94, Sec. III], where speed is assumed as a known signal in the stability proof. In [140], an attempt to apply adaptive observer design proposed in [150] to model 3 ($x = -e_A$) has ended with requiring infinite gain to attain asymptotical stability, where the signal si is obtained using a state variable filter—which can be avoided if model 4 ($x = \Omega\psi_A$) is used instead [140]. The research using model 5 ($x = \psi_s$) is summarized in the monograph [156], and the key property of choosing stator flux as the unknown internal state is that there is no unknown variable ω_r appearing in the dynamics of unknown state x . Finally, model 6 ($x = \Omega\psi_s$) is in adaptive observer form (32) and therefore, the existence of globally stable speed-adaptive observer for induction motor is an established fact [157]–[159], where filtering the original speed regressor is found essential [159].

2) *Advances in High Gain Observer Design*: The advances in high-gain observer (HGO) design [154], [155] accept a wider class of systems than (32), and application of this HGO to model 6 is studied in [160]. Moreover, the speed ω_r and load torque T_L can be treated as states, so that model 4 has been extended with an third \mathbb{R}^2 state as $z = -s(\omega_r \mathbf{J} \psi_A)$ [161]. The requirement on the partitioned matrix needed in [154], [161] is later removed in [155], which allows one to design HGO in terms of the two scalar states ω_r and T_L instead of the \mathbb{R}^2 vector z [155]. Note the speed regressor redesign from [152] has facilitated the above Lyapunov stability based HGO design [154], [155].

TABLE II
LIST OF STATE TRANSFORMATIONS FOR INDUCTION MOTORS ASSUMING CONSTANT SPEED

Model No.	Output state dynamics	Unknown state	Unknown state dynamics	Speed regressor Φ
1	$L_q s i = -s x + u - R i$	$x = \psi_A$	$s x = -\Omega x + R_{\text{req}} i$	$[(-\mathbf{J}x)^\top, (\mathbf{J}x)^\top]^\top$
2	$L_q s i = -x + u - R i$	$x = e_A$	$s x = -\Omega x + R_{\text{req}} s i$	$[\mathbf{0}, (\mathbf{J}x)^\top]^\top$
3	$L_q s i = x + u - R i$	$x = -e_A$	$s x = -\Omega x - R_{\text{req}} s i$	$[\mathbf{0}, (\mathbf{J}x)^\top]^\top$
4	$L_q s i = x + u - (R + R_{\text{req}}) i$	$x = \Omega \psi_A$	$s x = -\Omega (x - R_{\text{req}} i)$	$[\mathbf{0}, (\mathbf{J}(x - R_{\text{req}} i))^\top]^\top$
5	$L_q s i = \Omega (x - L_q i) + u - (R + R_{\text{req}}) i$	$x = \psi_s$	$s x = u - R i$	$[(-\mathbf{J}(x - L_q i))^\top, \mathbf{0}]^\top$
6	$L_q s i = x + u - \left(\frac{L_d R_{\text{req}}}{L_d - L_q} + R\right) i + \omega_r \mathbf{J} L_q i$	$x = \Omega \psi_s$	$s x = \Omega (u - R i)$	$[\mathbf{J}y, -\mathbf{J}(u - R i)]^\top$

Note: $\Omega = \frac{R_{\text{req}}}{L_d - L_q} \mathbf{I} - \omega_r \mathbf{J}$.

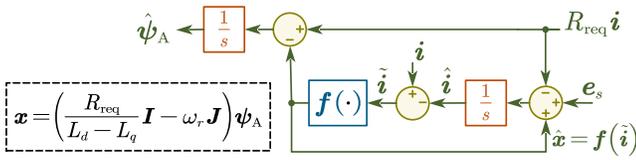


Fig. 15. SMDO design for induction motor that estimates the disturbance \mathbf{x} .

D. Disturbance Observer Design via Change of States

Substituting (31c) into (31a) will introduce a disturbance variable denoted as $\mathbf{x} = \Omega\psi_A$. Fig. 15 shows the general DO for estimating this disturbance, where SM correction is widely used [107]–[111], and linear dynamic correction is also studied [162]. Alternatively, one can design a full-order SMDO for model 1 or model 5 with the speed-dependent term $\omega_r \mathbf{J}\psi_A$ or $\omega_r \mathbf{J}\psi_s$ being treated as disturbance [147].

VI. CONSIDERATIONS FOR LOW SPEED OPERATIONS

The sufficient conditions for preserving ac motor observability provide guidance for stable operation near $s\theta_d = 0$. The state vector $[i_\alpha, i_\beta, \omega, \theta_d]^T$ of a PM motor is locally weakly observable, if the following inequality holds [163]:

$$s\theta_d = \omega \neq s \tan^{-1} \left(\frac{L_d - L_q}{\psi_A} i_q \right) \quad (33)$$

The state vector $[i_\alpha, i_\beta, \psi_{\alpha A}, \psi_{\beta A}, \omega, T_L]^T$ of an induction motor is locally weakly observable, if the inequality holds [161], [163]:

$$\omega \neq s \tan^{-1} \left(\frac{L_d - L_q}{R_{req}} \omega_r \right) = s \tan^{-1} \left(\frac{L_d - L_q}{R_{req}} \omega - \frac{L_d - L_q}{\psi_A} i_q \right) \quad (34)$$

where the slip speed has been substituted. To preserve the observability at $\omega = 0$, both (33) and (34) suggest that an auxiliary observability vector in dq -frame³

$$\psi_O^{dq} \triangleq \begin{bmatrix} \psi_A \\ -(L_d - L_q) i_q \end{bmatrix} \quad (35)$$

should rotate or change its direction with respect to the d -axis. In particular, the local weak observability of a surface mounted PM motor holds with non-zero acceleration at $\omega = 0$, i.e., zero frequency crossing [164].

Slow speed reversal test has been recommended to test stability margin of a sensorless drive near zero frequency operation [141]. The CM estimator can survive the test with the help of the load-dependent i_d excitation [50], where the d -axis angle error still shows trends of divergence during zero frequency crossing. It is suggested to use high frequency signal injection to assist zero frequency crossing for the dq -frame VM estimator in [35]. However, the $\alpha\beta$ -frame VM estimator shows smooth d -axis angle waveform even during zero frequency crossing [18], [32], [37], [106]. Note only voltage compensation is needed in [18], [32], [106], while the speed-dependent flux compensation is used in [37]. The SMDO proposed in [84] is also found to be able to perform the test, but shows increased noises in speed estimate as compared to its VM counterpart [18]. The globally stable speed-adaptive observer also shows speed divergence near zero frequency crossing because the persistency of excitation is lost at dc excitation [158]. Typical acceleration rate of the slow reversal is between 50–100 r/min/s, whereas a successful 5 r/min/s test is reported in [106], [158], and it can go down to 2 r/min/s if deliberate flux weakening is allowed [158]. As a final note, compared with operating at standstill (see, e.g., [18], [37]), operating extremely near zero frequency (e.g., 1 r/min [6]) is considered as a more challenging working condition.

³It is interesting to point out that this auxiliary flux vector is used to design a regressor for the speed adaptation law of the dq -frame VM estimator in [62].

VII. CONCLUDING REMARKS

The act to use the same torque expression for all ac motors, provides us a unified perspective to review fundamental frequency sensorless algorithms based on the model of torque-producing flux, i.e., active flux. The resulting equivalent circuit models the effect of the PM, the saliency and the rotor conductors.

The torque estimation problem can be solved directly or indirectly, which corresponds to a DFO or IFO sensorless drive. The *direct* torque estimation includes the VM estimator and the observers, such that d -axis angle is computed from $\alpha\beta$ components of flux or emf vector. The *indirect* torque estimation problem is turned into speed estimation problem via the CM estimator. The CM estimator enables speed estimation in IFO dq -frame, but any other speed estimation method such as MRAS and speed-adaptive state observer can be used—which is often seen in IFO controlled induction motors.

No method is perfect. VM is critically stable, CM estimator has algebraic loop issues, state observer is disturbed by estimated speed error, and disturbance observer suffers from undesired dither in estimation. Nevertheless, advances to each method have been made in literature:

- 1) The VM estimator is analyzed in the dq frame to derive operating speed dependent stability results [34], implying stability is guaranteed with speed dependent tuning.
- 2) The CM estimator is linearized for analysis, from which “complete stability” is attained with proper tuning [50], [165].
- 3) Challenges have been reported in designing stable state observer for model using active flux as states, while the researches that involve *change of states* are fruitful, e.g., globally stable speed-adaptive observer (except zero stator frequency condition) [12], [155], [158].
- 4) Dynamic SM correction of higher order has long been found able to eliminate chattering in SMDO (see e.g., [166, Ch. 6]).

Unlike IFO drive, a DFO drive needs to estimate speed *separately*. A general observer is proposed to summarize various speed estimation algorithms including PLL, ELO, EKF, GPIO, GESO, and SMDO, which differ in terms of linear or nonlinear corrections, time-invariant or time-varying gains, and their assumed motion dynamics. It is worth mentioning that for the same general observer of order n , there are potentially $n + 1$ variants of speed estimates having different characteristics in terms of dynamic performance and noise attenuation.

The applicability of sensorless algorithms depends. Speed estimation methods assuming constant speed are more suited for motors with large inertia, while high dynamic performance can be obtained with higher order speed estimation. For ultra-high speed applications with high switching frequencies, CM estimator is recommended for its minimum computational cost. For high-saliency motors and non-PM motors, as the active flux amplitude is regulated in real-time, designs that take into account of the dynamic active flux amplitude should be considered. For low speed applications, the local weak observability has suggested to adopt high saliency PM motor or induction motor from a theoretical point of view.

APPENDIX A

THE HISTORY OF EXTENDED EMF MODEL

By manipulating inductance, the dq -model (6) is rewritten

$$\underbrace{\omega\psi_A - (L_d - L_q) \frac{d}{dt} i_q}_{\text{Amplitude of the extended emf } e_E} + L_d s i_d = u_d - R i_d + \omega L_q i_q \quad (36)$$

Transforming (36) to $\alpha\beta$ -frame yields extended emf model: [76]

$$L_d s \mathbf{i} + \mathbf{e}_E + \omega \mathbf{J} (L_d - L_q) \mathbf{i} = \mathbf{u} - R \mathbf{i} \quad (37a)$$

$$\text{with } s \mathbf{e}_E = \omega \mathbf{J} \mathbf{e}_E + \text{Unmodelled Dynamics} \quad (37b)$$

Historically, it is not easy to design sensorless algorithm for salient-pole PM motors, owing to the position-dependent inductance that leads to a highly nonlinear model. The salient-pole PM motor can be treated as a non-salient one if one assumes $\|(L_d - L_q) \mathbf{i}\| \ll \psi_{PM}$ [76]. The idea of extended emf was timely proposed to relax the need of this low saliency assumption. Note the current dynamics are now explicitly dependent on ω in (37a).

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