Induction Machine Shaft Speed Estimation using Imaginary Power

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Abstract—This paper investigates the use of an imaginary power based speed estimation algorithm for induction machines. The algorithm investigated was first proposed by Peng and Fukao in 1994 [1], and has some very desirable characteristics such as insensitivity to knowledge of stator resistance, and does not use pure integrations. However, it has been shown that the algorithm does not operate stably under regeneration conditions, which has been the main reason that it has not seen much application. This paper investigates the reasons behind this instability, and then develops a solution approach.

I. INTRODUCTION

Estimation of the shaft speed of induction machines has been extensively researched over a number of years [1]–[9]. The motivation for this research has been the desire to eliminate the fragile speed sensor from induction machine drive systems, because it compromises the rugged characteristics of the machine. Also in some situations it is not practical to use a speed sensor because of a particularly hazardous environment, or very high shaft speeds. In addition, in some applications the cost of the speed sensor is significant.

The speed estimation techniques discussed in the literature fall into three broad categories in terms of their fundamental operational principles:

- 1) Using the machine back emf or flux to determine the rotational speed of the rotor [2], [9], [10].
- 2) Detection of rotor slot harmonics in the stator current waveforms [5], [8].
- 3) Detection of the saliency in the rotor of the machine [6], [7].

This paper considers a speed estimation algorithm published by Peng and Fukao in 1994 [1]. In terms of the categories mentioned above it is a back-emf based algorithm (i.e. Category 1). The unique feature of the algorithm is that it uses instantaneous imaginary power as the driving feedback signal for the estimator, which in principle means that it would be insensitive to the accurate knowledge of stator resistance. A further advantage of the algorithm compared to, for example some saliency techniques, is that it only requires a standard machine.

One of the significant problems with Category 1 speed estimation algorithms is that they implicitly rely on accurate knowledge of the machine stator resistance in order to calculate the back-emf accurately. An accurate estimate is essential at low shaft speeds where the stator resistance drop is very significant compared to the back-emf voltage itself. One of the complications in estimating the stator resistance is that it is not constant but varies significantly with the temperature of the machine.

The use of instantaneous imaginary power in [1] to estimate speed eliminates the stator resistance from the estimation process, since imaginary power does not involve any real power components. Therefore the algorithm, in principle, should work as well at low speeds as it does at higher speeds.

The Peng-Fukao algorithm has a Model Reference Adaptive System (MRAS) structure, similar to [9]. However, it differs in that it does not use pure integrations, as does [9] and many other speed estimation techniques. Consequently it avoids the initial condition and drift problems associated with pure integration based approaches. This property, coupled with the stator resistance independence, is the main motivation for studying the algorithm.

In [1] simulation and practical results were presented that indicated that the algorithm worked very well in all regions of operation. The authors have carried out extensive simulation studies of the algorithm and found that it produces accurate speed estimates at all speeds when the machine is motoring, but the speed estimates are incorrect under regeneration conditions.

This paper investigates and analyses the problem of incorrect speed estimates under regeneration for the Peng-Fukao algorithm, and then proposes and simulates a solution approach. The remainder of the paper is organised as follows: Section II gives a brief derivation of the Peng-Fukao estimator; Section III presents simulation results of the estimator under motoring and regeneration conditions; Section IV presents some analysis of the linearised equations of the system; Section V presents a solution approach and simulation results; and finally Section VI presents the conclusions.

II. BRIEF DEVELOPMENT OF PENG-FUKAO ESTIMATOR

Complete details of the Peng-Fukao algorithm can be found in [1]. In this section we shall develop the basic equations, using standard notation, that form the basis of the estimator (which are not derived in [1]).

Consider the following definition for the rotor magnetising current (expressed in a stationary reference frame) of an induction machine:

$$\dot{i}'_{mr} = \frac{\Psi'_r}{L_m} = \dot{i}_s + (1 + \sigma_r)\dot{i}'_r \quad (1)$$

where $\sigma_r \triangleq \frac{L_{lr}}{L_m}$ (the rotor leakage factor). Therefore $1 + \sigma_r = \frac{L_r}{L_m}$.

Using (1) the standard stationary frame equations for an induction machine can be manipulated into the following form. Let us firstly consider the new form of the rotor equation for a squirrel cage machine (i.e. $\underline{u}'_r = 0$):

$$\frac{d\underline{i}'_{mr}}{dt} = j\omega_r \underline{i}'_{mr} - \frac{1}{T_r} \underline{i}'_{mr} + \frac{1}{T_r} \underline{i}_s \quad (2)$$

which can also be written as:

$$\frac{d\underline{i}'_{mr}}{dt} = \underline{\omega}_r \otimes \underline{i}'_{mr} - \frac{1}{T_r} \underline{i}'_{mr} + \frac{1}{T_r} \underline{i}_s \quad (3)$$

using the fact the $j\omega_r \underline{i}'_{mr} = \underline{\omega}_r \otimes \underline{i}'_{mr}$.

The stator equation can be written as:

$$\underline{u}_s = R_s \underline{i}_s + L_s \frac{d\underline{i}_s}{dt} + L_m \frac{d\underline{i}'_{mr}}{dt} \quad (4)$$

Rearranging (1) in terms of $\underline{i}'_{mr}$ one can write (4) as:

$$\underline{u}_s = R_s \underline{i}_s + \sigma L_s \frac{d\underline{i}_s}{dt} + (1 - \sigma) L_s \frac{d\underline{i}'_{mr}}{dt} \quad (5)$$

where:

$$\sigma = 1 - \frac{L_m^2}{L_s L'_r} \quad (6)$$

$$(7)$$

The back-emf of an induction machine is the voltage behind the transient reactance of the machine, and this can be written as:

$$\underline{e}_m = \underline{u}_s - R_s \underline{i}_s - \sigma L_s \frac{d\underline{i}_s}{dt} \quad (8)$$

By comparison with (5) one can see that:

$$\underline{e}_m = (1 - \sigma) L_s \frac{d\underline{i}'_{mr}}{dt} \quad (9)$$

$$\therefore \underline{e}_m = \frac{L_m^2}{L'_r} \frac{d\underline{i}'_{mr}}{dt} \quad (10)$$

Clearly one can substitute (3) into (10) to give:

$$\underline{e}_m = \frac{L_m^2}{L'_r} \left(\underline{\omega}_r \otimes \underline{i}'_{mr} - \frac{1}{T_r} \underline{i}'_{mr} + \frac{1}{T_r} \underline{i}_s \right) \quad (11)$$

The instantaneous imaginary power is:

$$\underline{q}_m = \underline{i}_s \otimes \underline{e}_m \quad (12)$$

Substituting (11) into (12) we can write after some manipulation:

$$\underline{q}_m = \frac{L_m^2}{L'_r} \left(\underline{\omega}_r (\underline{i}'_{mr} \odot \underline{i}_s) + \frac{1}{T_r} (\underline{i}'_{mr} \otimes \underline{i}_s) \right) \quad (13)$$

Figure 1 shows the block diagram of the Peng-Fukao MRAS estimator with the relevant equations indicated.

Remark 1: Note that in Figure 1 and the following equations a $\hat{x}$ symbol indicates that x is an estimated variable. ♦

Fig. 1. MRAS structure of the Peng-Fukao shaft speed estimator.

III. SIMULATION RESULTS OF PENG-FUKAO ESTIMATOR

The estimator shown in Figure 1 was programmed into the Saber[®] mixed mode simulation package.

The simulation was configured so that the induction machine was driven from a fixed frequency supply and the shaft of the machine was controlled by a controllable speed source. This allowed the induction machine to be manipulated into various modes of operation simply, with all the conditions of operation known. Prior to establishing this simulation configuration the estimator was used in conventional field oriented control (FOC) simulations. However the conditions of operation were not well enough controlled, and it was difficult to establish what was happening to the estimator.

Figure 2 shows the results of a simulation. The top two plots are the dq stationary frame rotor magnetising currents for the machine – i.e components of $\underline{i}'_{mr}$. Variables, marked in the figure as “imd” and “imq” are the true values of the rotor magnetising current, and “imd_prime” and “imq_prime” are the estimated values of the same variables. The voltage and speed of the machine are ramped up for the first second of operation (not shown here). After $t = 1$ second the speed is held at a constant value that is just below the synchronous speed of the machine (synchronous speed is 377 rad/sec as the machine is being driven from a constant 60Hz supply) – i.e. the machine is motoring up to $t = 1.4$ seconds.

Remark 2: The estimated speed is very accurate from $t = 0$ to $t = 1.4$ seconds. ♦

At $t = 1.4$ seconds the machine shaft speed is increased through the synchronous speed and drives the machine into regeneration. One can see that the estimated and actual speed match for a short period after this transition and then the estimated speed diverges from the true speed and settles back to a speed that indicates that the machine is again motoring. Meanwhile the true speed remains constant until $t = 1.6$ seconds, when it is again decreased below synchronous speed. One can see as the true speed drops below synchronous speed the estimated speed relocks to the true speed.

Remark 3: When the estimated speed diverges from the true speed there is a phase shift in the estimated $\hat{\underline{i}}'_{mr}$

Fig. 2. Close-up of currents and speed estimate for the Peng-Fukao estimator.

from the true $\hat{i}'_{mr}$ current. If one views the $\underline{i}_s$ vector at this time it is constant. The phase change in the $\hat{\underline{i}}'_{mr}$ can be seen in Figure 2. ♦

Remark 4: The Peng-Fukao algorithm was extensively tested on a FOC system and the same properties were observed – when the machine regenerates the estimated speed diverges from the true speed. ♦

Remark 5: Also in Figure 2 one can see that the $q_m - \hat{q}_m$ error during the estimated speed divergence does not change. Consequently there is no feedback to correct the estimation error that occurs. ♦

IV. ANALYSIS OF PENG-FUKAO ESTIMATOR

From equations (3) and (13) it can be seen that estimator is non-linear – it has bilinear terms if one considers the input into (3) to be $\hat{\omega}_r$. Note that the algebraic output equation (13) has even more complex non-linearities. Therefore in order to examine the observability, controllability and stability of the estimator equations one needs to linearise them around the operating points of interest.

In order to simplify the analysis of the equations we shall convert (1) and (13) into a synchronously rotating reference frame aligned with the stator current vector. The resultant equations are:

$$\frac{d\hat{\underline{i}}'_{mrs}}{dt} = j(\hat{\omega}_r - \omega_s)\hat{\underline{i}}'_{mrs} - \frac{1}{T_r}\hat{\underline{i}}'_{mrs} + \frac{1}{T_r}|\underline{i}_s| \quad (14)$$

$$\hat{q}_m = \frac{L_m^2}{L_r'} \left(\hat{\omega}_r \hat{i}'_{mrds} - \frac{1}{T_r} \hat{i}'_{mrqs} \right) |\underline{i}_s| \quad (15)$$

where ω_s is the rotational angular velocity of the stator current vector, and $\hat{i}'_{mrds}$ and $\hat{i}'_{mrqs}$ are the dq components of the $\hat{\underline{i}}'_{mrs}$ vector. The “s” subscript indicates that the quantities are with reference to the stator current vector.

Equation (14) can be further manipulated to represent it in dq components:

$$\frac{d\hat{i}'_{mrds}}{dt} = (\omega_s - \hat{\omega}_r)\hat{i}'_{mrqs} - \frac{1}{T_r}\hat{i}'_{mrds} + \frac{1}{T_r}|\underline{i}_s| \quad (16)$$

$$\frac{d\hat{i}'_{mrqs}}{dt} = -(\omega_s - \hat{\omega}_r)\hat{i}'_{mrds} - \frac{1}{T_r}\hat{i}'_{mrqs} \quad (17)$$

Remark 6: The following linearisation will assume that the $\underline{i}_s$ vector is not changing – i.e. $\frac{d|\underline{i}_s|}{dt} = 0$ and $\frac{d\omega_s}{dt} = 0$. This simplifies the equations, and is representative of the condition where the Peng estimator misbehaves – i.e. under a condition where the stator current is fixed and the rotor speed is fixed, but under regeneration. ♦

Applying standard techniques for linearising non-linear differential equations, under the constant $\underline{i}_s$ condition, allows us to write:

$$\begin{aligned} \frac{d(\delta\hat{i}'_{mrds})}{dt} &= (\omega_s - \hat{\omega}_r)(\delta\hat{i}'_{mrqs}) \\ &\quad - \frac{1}{T_r}(\delta\hat{i}'_{mrds}) - \hat{i}'_{mrqs}(\delta\hat{\omega}_r) \end{aligned} \quad (18)$$

$$\begin{aligned} \frac{d(\delta\hat{i}'_{mrqs})}{dt} &= \hat{i}'_{mrds}(\delta\hat{\omega}_r) \\ &\quad - (\omega_s - \hat{\omega}_r)(\delta\hat{i}'_{mrds}) - \frac{1}{T_r}(\delta\hat{i}'_{mrqs}) \end{aligned} \quad (19)$$

$$\begin{aligned} \delta\hat{q}_m &= \frac{L_m^2}{L_r'} |\underline{i}_s| \left[\hat{i}'_{mrds}(\delta\hat{\omega}_r) + \hat{\omega}_r(\delta\hat{i}'_{mrds}) \right. \\ &\quad \left. - \frac{1}{T_r}(\delta\hat{i}'_{mrqs}) \right] \end{aligned} \quad (20)$$

where δx denotes a small signal state variable or control input or system output. The above equations can be written in conventional state space form as follows:

$$\frac{d(\delta\hat{\mathbf{i}}'_{mrs})}{dt} = \mathbf{A}(\delta\hat{\mathbf{i}}'_{mrs}) + \mathbf{B}(\delta\hat{\omega}_r) \quad (21)$$

$$\delta\hat{q}_m = \mathbf{C}(\delta\hat{\mathbf{i}}'_{mrs}) + \mathbf{D}(\delta\hat{\omega}_r) \quad (22)$$

where:

$$(\delta\hat{\mathbf{i}}'_{mrs}) = \begin{bmatrix} \delta\hat{i}'_{mrds} \\ \delta\hat{i}'_{mrqs} \end{bmatrix} \quad (23)$$

$$\mathbf{A} = \begin{bmatrix} -\frac{1}{T_r} & (\omega_s - \hat{\omega}_r) \\ -(\omega_s - \hat{\omega}_r) & -\frac{1}{T_r} \end{bmatrix} \quad (24)$$

$$\mathbf{B} = \begin{bmatrix} -\hat{i}'_{mrqs} \\ \hat{i}'_{mrds} \end{bmatrix} \quad (25)$$

$$\mathbf{C} = \frac{L_m^2}{L_r'} |\underline{i}_s| \left[\hat{\omega}_r \quad -\frac{1}{T_r} \right] \quad (26)$$

$$\mathbf{D} = \frac{L_m^2}{L_r'} |\underline{i}_s| \left[\hat{i}'_{mrds} \right] \quad (27)$$

To check for controllability and observability the following matrices must have full rank:

$$\text{CM: } [\mathbf{B} \quad \mathbf{AB} \quad \mathbf{A}^2\mathbf{B} \quad \dots \quad \mathbf{A}^{n-1}\mathbf{B}] \quad (28)$$

$$\text{OM: } \begin{bmatrix} \mathbf{C} \\ \mathbf{CA} \\ \mathbf{CA}^2 \\ \vdots \\ \mathbf{CA}^{n-1} \end{bmatrix} \quad (29)$$

where $n \triangleq$ the number of states.

Let us consider the observability matrix. Substituting in the values for $\mathbf{C}$ and $\mathbf{A}$ and realising that $n = 2$, we get the following matrix:

$$\begin{bmatrix} \mathbf{C} \\ \mathbf{CA} \end{bmatrix} = \frac{L_m^2}{L_r} |\dot{i}_s| \begin{bmatrix} \hat{\omega}_r & -\frac{1}{T_r} \\ -\frac{\hat{\omega}_r}{T_r} + \frac{(\omega_s - \hat{\omega}_r)}{T_r} & \hat{\omega}_r(\omega_s - \hat{\omega}_r) + \frac{1}{T_r^2} \end{bmatrix} \quad (30)$$

If this matrix does not have full rank then determinant of the matrix will be zero. The determinant is:

$$\det \begin{bmatrix} \mathbf{C} \\ \mathbf{CA} \end{bmatrix} = \frac{L_m^2}{L_r} |\dot{i}_s| (\omega_s - \hat{\omega}_r) [\hat{\omega}_r^2 + \frac{1}{T_r^2}] \quad (31)$$

Clearly this can only be zero if $(\omega_s - \hat{\omega}_r) = 0$ – i.e. the “slip” is zero¹. This is the only condition that the system is unobservable.

Now let us check on controllability.

$$\begin{bmatrix} \mathbf{B} & \mathbf{AB} \end{bmatrix} = \begin{bmatrix} -\hat{i}'_{mrsq} & \frac{\hat{i}'_{mrsq}}{T_r} + \hat{i}'_{mrds}(\omega_s - \hat{\omega}_r) \\ \hat{i}'_{mrds} & \hat{i}'_{mrsq}(\omega_s - \hat{\omega}_r) - \frac{\hat{i}'_{mrds}}{T_r} \end{bmatrix} \quad (32)$$

This matrix should be full rank for the system to be controllable. This means that the determinant should be non-zero. Taking the determinant we get:

$$\det \begin{bmatrix} \mathbf{B} & \mathbf{AB} \end{bmatrix} = -(\omega_s - \hat{\omega}_r) (\hat{i}_{mrds}^2 + \hat{i}_{mrsq}^2) \quad (33)$$

Clearly the determinant is equal to zero when:

$$(\omega_s - \hat{\omega}_r) = 0 \quad (34)$$

and therefore the system is uncontrollable under this condition.

Let us now consider the small signal stability of the estimator. Using the linearised equations (21) and (22) with steady state operating points for both motoring and regeneration we can plot the poles and zeros of the open loop and closed loop transfer functions for the estimator.

Let us firstly consider the open stability of the estimator from q to $\hat{q}_m$. The pole-zero plot for PI gains of 100 is shown in Figure 3. One can conclude that the system is stable, with the zero and pole near to origin due to the PI controller.

If the estimation loop is now closed, with the machine motoring and gains of $K_p = K_i = 100$ the pole-zero plot of Figure 4 results. Again the estimator is stable and is in fact stable for all feedback gains in the motoring mode of operation.

If the process is repeated for regeneration very different results are obtained. For PI gains of 100 (as used in the motoring mode analysis) the open loop pole zero plot of Figure 5 results. The estimator is open loop stable again with a pole and zero near the origin due to the PI controller.

Remark 7: Note that the linearised estimator is non-minimum phase with a zero in the right half plane. ♦

Closing the loop gives a pole-zero plot as shown in Figure 6. The estimator in regeneration is unstable with these gains (see Remark 8). In fact the estimator

¹Strictly speaking $(\omega_s - \hat{\omega}_r)$ is not the slip frequency, since transiently ω_s is not the same frequency as ω_{mr} . However, under the test conditions they are the same.

Fig. 3. Estimator open loop pole-zero plot for $K_p = K_i = 100$, motoring mode.

Fig. 4. Estimator closed loop pole-zero plot for $K_p = K_i = 100$, motoring mode.

is unstable in closed loop for all gains with negative feedback. Root locus plots have been generated to confirm this but are not shown here for conciseness.

Remark 8: One can see that there is an apparent pole zero cancellation in the right half plane of Figure 6. However if one expands the plot here one can see that the pole and zero are not co-incident and therefore there would be a small unstable residual here. ♦

The closed loop system in the generating mode can be small signal stabilised if the sign of the feedback is changed *and* the gain is reduced enormously. Figure 7 shows a closed loop pole-zero plot of the estimator with a feedback sign change and proportional and integral gains of 0.1. With this very low gain however the estimator is almost completely useless due to very poor dynamic performance.

Remark 9: Overall, one can conclude, at the operating point of the linearisation, that motoring is stable for all feedback gains if negative feedback is applied. In regeneration the estimator is unstable for all gains if

Fig. 5. Estimator open loop pole zero plot for $K_p = K_i = 100$, generating mode.

Fig. 6. Estimator closed loop pole zero plot for $K_p = K_i = 100$, generating mode.

negative feedback is utilised. The manifestation of instability in generating mode is “slow” if high gains are used because there is almost a pole-zero cancellation at the right half plane pole. However, for modest gains (≈ 0.1) regeneration can be stabilised by reversing the sign of the feedback. The estimator certainly has stability problems, and will require additional feedback in order to stabilise it. ♦

V. AUGMENTED ESTIMATOR AND SIMULATION RESULTS

A number of techniques were investigated to generate the necessary feedback to stabilise the estimator. Observations made from the simulation studies indicated that the instability in the estimator resulted in the $\hat{i}_{mr}$ space vector angle moving, under regeneration conditions, relative to the stator current space vector i_s . Therefore all the techniques developed attempted to tackle this phase shift issue by providing corrective action. We shall only consider one technique in this paper.

The generation of the necessary feedback signal to implement the phase locking function is simple, and can

Fig. 7. Estimator closed loop pole zero plot for $K_p = K_i = 10^{-1}$ and positive feedback

be expressed as:

$$\begin{aligned} \hat{e}_m \otimes e_m &= |\hat{e}_m| |e_m| \sin(\tilde{\theta}_{mr} - \hat{\theta}_{mr}) \approx \\ &|\hat{e}_m| |e_m| (\tilde{\theta}_{mr} - \hat{\theta}_{mr}) = |\hat{e}_m| |e_m| \delta \tilde{\theta}_{mr} \end{aligned} \quad (35)$$

for small $\tilde{\theta}_{mr} - \hat{\theta}_{mr}$, where $\tilde{\theta}_{mr}$ denotes that this is the angle of the back-emf, as opposed to the i'_{mr} current.

Clearly one can generate a normalised version of (35) as follows:

$$\delta \tilde{\theta}_{mr} \approx \frac{\hat{e}_m \otimes e_m}{|\hat{e}_m| |e_m|} \quad (36)$$

The proposed solution is to use the phase difference between the back-emf derived from the reference model via (8) and that derived from estimator equation (11). The idea in this case is to “phase lock” the angle of $\hat{e}_m$ to e_m , and thereby lock $\hat{i}'_{mr}$ to i_{mr} .

Remark 10: This approach still has the problem that there is an implicit dependence on knowledge of R_s , since the phase angle of e_m will vary considerable from the true version of this variable at low speeds. Therefore the $\hat{e}_m$, and hence $\hat{i}'_{mr}$ will be forced towards an incorrect phase. On the plus side, this approach uses expressions that are already part of the estimator, and the approach does not rely on pure integrations. ♦

There are a variety of approaches for feeding back the phase information. In this paper we will feed the output of (35) as an augmentation of the $\hat{\omega}_r$ signal back into the estimator. In addition there is a feed-forward term of $\hat{\omega}_{r1}$ which is summed into the feedback path. This allows the gain of the PI controller to be lower. The block diagram of this structure is shown in Figure 8. As can be seen the raw value of the approximate phase error between the $d\hat{i}_{mr}/dt$ vector and the e_m vector is fed back into the speed feedback loop of the estimator.

The simulation results of this strategy appear in Figure 9. As can be seen from this figure the instability problems evident in Figure 2 do not occur. One can see that the phase of the estimated magnetising currents remain locked to the true values. The gain of the feedback is low – in the case of the simulation shown it is 0.1. In

Fig. 8. Modified Peng speed estimator with phase locking.

Fig. 9. Simulation result of phase lock feedback with actual R_s .

Fig. 10. Simulation result for phase lock feedback with 30% error in R_s .

this simulation the stator resistance used to derive e_m is the stator resistance of the machine.

A second simulation was carried out with the stator resistance used for e_m equal to $1.3R_s$. The results of this simulation are shown in Figure 10. As can be seen from this figure the error in the stator resistance leads to a significant error in the estimated speed. However, the estimator still remains stable in the sense that the estimated speed indicates that the machine is regenerating

when indeed it is regenerating.

Remark 11: From Figure 9 and 10 one can see that the $\hat{i}'_{mr}$ currents are in phase with the true i'_{mr} current. ♦

Remark 12: Simulations were also carried out using the normalisation indicated in (36) with similar results. ♦

VI. CONCLUSIONS

The following conclusions can be drawn from this paper:

- A linearised version of the original Peng-Fukao estimator has been shown to be stable for all feedback gains in motoring mode and unstable in regeneration unless the sign of the feedback is reversed and the feedback gain is very low.
- Simulations of the full non-linear estimator confirm the linearised instability.
- The Peng-Fukao estimator is both observable and controllable, which indicates that, in principle, the instability problem can be remedied with appropriate feedback.
- The estimator can be stabilised by adding auxiliary feedback which “phase locks” the $d\hat{i}'_{mr}/dt$ vector to e_m . This inevitably leads to dependence on knowledge of the stator resistance. An augmented algorithm using this approach has been devised and simulated, and the estimator is stable.

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