DSP Based Implementation of High Performance Flux Estimators for Speed Sensorless Induction Motor Drives Using TMS320F2812

M.Kaliamoorthy
Lecturer, Department of Electrical Engineering
IFET College of Engineering
Villupuram, India-605104
kalias_ifet@yahoo.com

S.Himavathi, A.Muthuramalingam
Assistant Professors, Department of Electrical Engineering
Pondicherry Engineering College
Pondicherry, India-605104
Himavathi@mailcity.com, amrligam@hotmail.com

***Abstract*-This paper deals with Programmable LPF (Low Pass Filter) based stator flux estimation for speed sensorless induction motor drives. This algorithm is proposed to solve the dc drift problem and errors due to stator resistance variation associated with the ideal integrator and a LPF. This algorithm has pole/gain compensator to estimate stator flux over wide speed range. Accordingly, the severity of drift problem is much reduced and the stator flux is exactly estimated in the wide speed range. The validity of the Programmable LPF algorithm is verified by rotor flux oriented speed sensorless vector control of a three phase induction motor through simulation and the simulated results are verified in an TMS320F2812 DSP processor.**

***Keywords:* Flux Estimation, induction motor drives, first order low pass filter, rotor flux oriented speed sensorless control.**

I. INTRODUCTION

The voltage model is a convenient flux estimator for sensorless induction motor drives because the motor speed is not needed for the flux estimation, and the only crucial parameter of the model is the stator résistance. The voltage model is often used in stator flux oriented control [1] and direct torque control [2], [3] schemes, but it can as well be used for rotor flux orientation control [4]. However there are two well known problems when the voltage model is used: even a small dc offset in measured currents causes drift problems if a pure integrator is used and at low speeds the model is extremely sensitive to errors in the stator resistance value and to measurement errors. . To solve the problems related with the voltage model based flux estimation, ideal integrator in the voltage model equations are replaced with low pass filters. Various low pass filter algorithms investigated are listed: Saturable feedback loop algorithm [5], Amplitude limiter algorithm [5], and Adaptive magnitude compensation algorithm. All these algorithms suffer from a basic problem of stability when the motor operates at a frequency close to zero.

The programmable LPF algorithm proposed in this paper with the phase/gain compensator estimates precisely the stator flux in a wide speed range. In addition, the pole of the programmable LPF is located far from the origin in order to decrease the time constant as the speed increases. Consequently, the stator flux is exactly estimated and the impact of drift problem is much reduced by the small time constant in a wide speed range.

II. INDUCTION MOTOR MODEL

The dynamic model corresponding to the T-equivalent circuit of the induction motor is shown in fig 1.The same developed model is used for simulations in the Matlab/Simulink environment

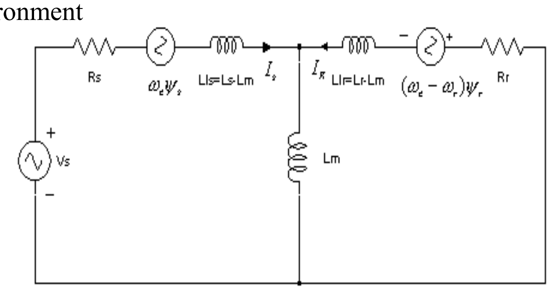

Figure.1 Induction Motor –T-Equivalent Model

The voltage equations given below are referred to general reference frame [7]. However, the stationary reference frame model is used for simulations.

$$V_S = R_s i_s + \frac{d\lambda_s}{dt} + j\omega_k \lambda_s \quad (1)$$

$$0 = R_r i_r + \frac{d\lambda_r}{dt} + j(\omega_k - \omega_m)\lambda_r \quad (2)$$

Where, ω_k is the speed of the reference frame, ω_m is the electrical angular speed of the motor shaft, V_s the stator voltage, i_s is the stator current and R_s is the stator resistance. Similarly subscript r denotes rotor parameters. The stator and rotor flux linkage equations are given as below

$$\lambda_s = L_s i_s + L_r i_r \quad (3)$$

$$\lambda_r = L_m i_s + L_r i_r \quad (4)$$

Where L_s, Lr and L_m are stator, rotor and mutual inductances respectively. The stator currents and rotor currents can be solved from the equations (3) and (4). Since the simulations are carried out in the stator reference frame the speed of the reference frame is set to zero. The electromagnetic torque developed by the motor is given by

$$T_e = \frac{3}{2} p I_m (i_s \lambda_s^*) \quad (5)$$

Where p is the number of pole pairs and the complex conjugate is marked as *.

III. VECTOR CONTROL SYSTEM

The system taken for the performance study of the voltage model based estimators is Rotor flux Oriented Vector Control Induction Motor Drive (RFVCIMD).The block diagram of the RFVCIMD scheme is shown in figure 2. The motor parameters used for simulations are given in table -1.

For the scheme shown in figure 2 all the different types of estimator were simulated in MATLAB/SIMULINK environment and comparison study is performed. The angular speed was estimated using the relation

$$\omega_e = \frac{1}{\hat{\psi}_r^{\,2}}\left[\left(\hat{\lambda}_{dr} * \dot{\hat{\lambda}}_{qr} - \hat{\lambda}_{qr} * \dot{\hat{\lambda}}_{dr}\right) - \frac{L_m}{T_r}\left(\hat{\lambda}_{dr} * i_{qs} - \hat{\lambda}_{qr} * i_{ds}\right)\right] \quad (6)$$

Table -1
Parameters of 2.2Kw 4-Pole 400V 50Hz Motor

Stator resistance R_s	3.67Ω
Rotor resistance R_r	2.32Ω
Magnetizing Inductance L_m	0.235H
Stator Inductance L_s	0.245H
Rotor Inductance L_r	0.248H
Inertia J_{TOTAL}	0.0126kgm^2
Rated Speed	1430 r/min
Rated Current	6.27 A
Rated Torque	14.6 Nm

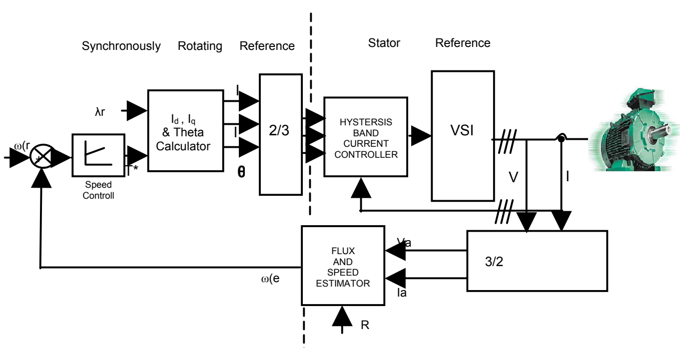

Figure.2 Rotor flux oriented vector control of Induction motor drive

IV. THEORY OF INTEGRATOR

Let's say a signal X is fed to an integrator and an integrated output signal Y will be obtained. This can be written as:

$$Y = \int X dt \quad (7)$$

for $X = A_m \sin(\omega t)$, *then*

$$Y = \frac{1}{\omega}\left(-A_m \cos(\omega t) + A_m \cos(\omega t)\Big|_{t=0}\right)$$

$$Y = \frac{1}{\omega}\left(-A_m \cos(\omega t) + A_m\right) \quad (8)$$

If the signal X is with an offset where X is represented by:

$X = A_m \sin(\omega t) + A_{dc}$

$$Y = \frac{1}{\omega}\left[\left(-A_m \cos(\omega t) + A_m \cos(\omega t)\Big|_{t=0}\right) + A_{dc}t - A_{dc}t\Big|_{t=0}\right]$$

$$Y = \frac{1}{\omega}\left(\left(-A_m \cos(\omega t) + A_m\right) + A_{dc}t\right) \quad (9)$$

In equation (8) there is an error due to initial value of the limit of integration, i.e. taking *t=0* as starting point. However, in equation (9) there is additional term, a ramp, due to a dc signal or offset in the signal that is fed at the input of the integrator. This ramp signal keeps on increasing with time. The aim is to eliminate A_m, appearing as constant value and $A_{dc}t$, appearing as a ramp signal at the output of the integrator. With time the ramp signal will easily drive the integrator into saturation whatever the magnitude of offset A_{dc} available in the input signal. The same theory of integration can be extended to voltage model based flux estimation equations as given below:

$$\hat{\lambda}_{ds} = \int\left(V_{ds} - R_{se} i_{ds}\right) dt \quad (10)$$

$$\hat{\lambda}_{qs} = \int\left(V_{qs} - R_{se} i_{qs}\right) dt \quad (11)$$

Where V_{dqs}, i_{dqs} and R_{se} are measured d and q axis voltage, current and estimated stator resistance of the machine. Figure.3 shows the actual and estimated d axis flux when a small dc error (0.1) is added with the back emf of d axis.

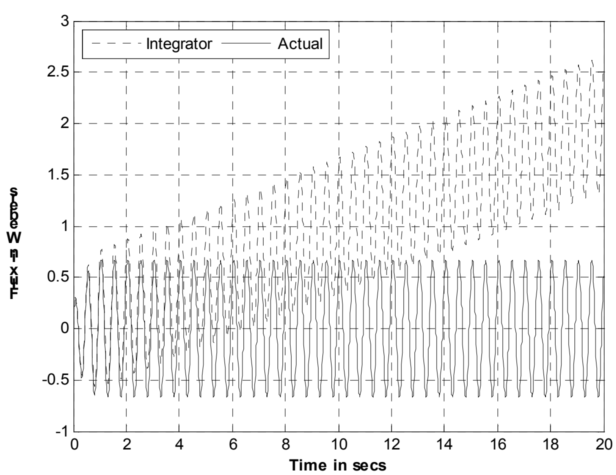

Figure.3 Actual and Estimated (Voltage Model when dc error is added) Flux

V. Problems Due To Stator Resistance

If the value of R_{se} is the actual value that of the machine, then equations 10 and 11 gives ideal estimate of the stator flux. But when there is a small error in stator resistance estimate then flux estimation using voltage model equations (10 and 11) are not ideal. Figure 4, 5 and 6 shows the actual and estimated flux at various frequencies (0.05 Hz, 0.5Hz and 5Hz) when stator resistance has 5% error.

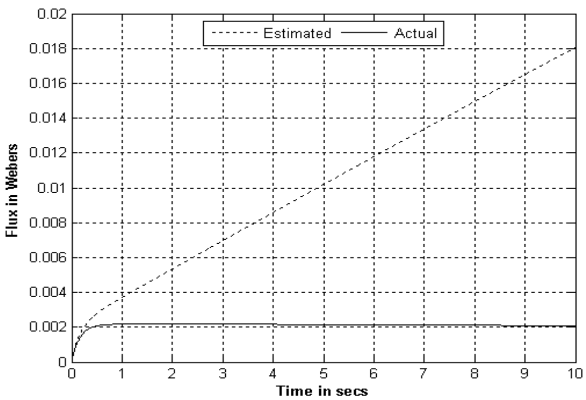

Figure.4 Actual and Estimated Flux at 0.05Hz and Rse=0.95*Rs

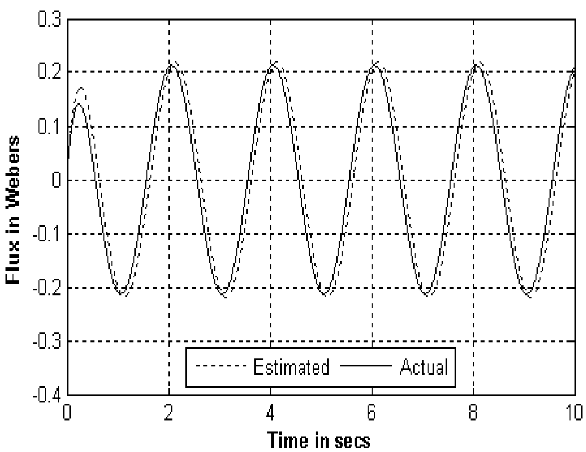

Figure.5 Actual and Estimated Flux at 0.5Hz and Rse=0.95*Rs

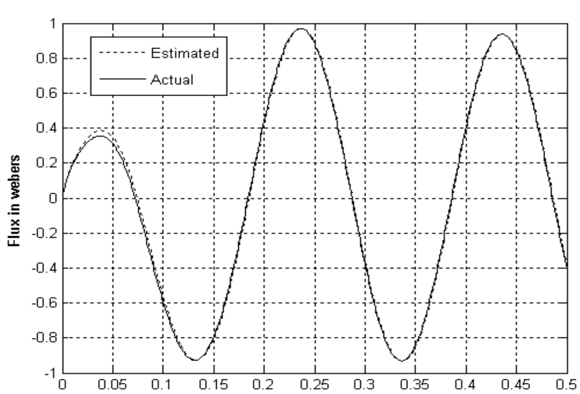

Figure.6 Actual and Estimated Flux at 5Hz and Rse=0.95*Rs

Thus from the above figures 4,5 and 6 it is found that stator resistance error dominates at low frequencies and vanishes completely when the frequency increases.

VI. Programmable Low Pass Filter

In all the above three algorithms, the pole of the low pass filter is fixed. Due to the fixed pole, the LPF filter has large time constant and this introduces a large error in the computation of synchronous angle, when the motor frequency is lower than the cut-off frequency of the filter. To solve this problem a programmable low pass filter is used in which the pole of the LPF is varied in accordance with the motor frequency. The block diagram of the programmable low pass filter based flux estimation is shown in the figure 7.

The ideal integration of (10 and 11) involves drift and saturation problems. To solve this problem the pure integrator is replaced by a LPF. The estimated stator flux by the LPF is given as

$$\frac{\hat{\lambda}_{sl}}{V_e} = \frac{1}{s + a} \tag{12}$$

Where ^ =estimated value, λ_{sl} – estimated stator flux by LPF, a= pole and V_e is the backemf (V_s-$i_s R_s$).

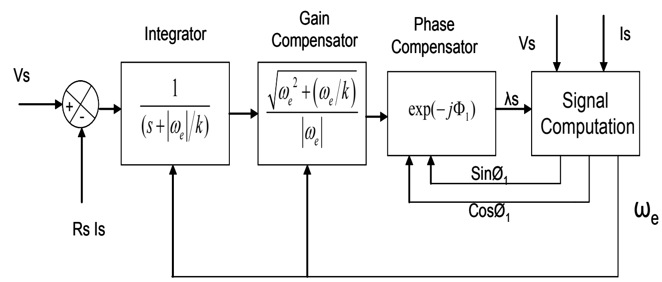

Figure.7 Block Diagram of Programmable Low Pass Filter

The phase lag and gain of (12) can be given as

$$\varphi = \tan^{-1}\left(\frac{\omega_e}{a}\right) \tag{13}$$

$$M = \left|\frac{\hat{\lambda}_{sl}}{V_e}\right| = \frac{1}{\sqrt{\hat{\omega}_e^{2} + a^2}} \tag{14}$$

Fig. 8 shows the phase lag λ_{sl} of estimated by the LPF and the phase lag of λ_s estimated by the pure integrator. The phase lag of λ_s is 90 and the gain is $1/|\omega_e|$. However, the phase lag of the LPF is not 90 and the gain is not $1/|\omega_e|$. Consequently, an error will be produced by this effect of the LPF. When the motor frequency is lower than the cutoff frequency of the LPF, the error is more severe. In order to remove this error, the LPF in (12) should have a very low cutting frequency.

There still remains the drift problem due to the very large time constant of the LPF. For the exact estimation of the stator flux, the phase lag and the gain of λ_{sl} in (12) have to be 90 and $1/|\omega_e|$, respectively. In addition, to solve the drift problem, the pole should be located far from the origin.

In this algorithm, the decrement in the gain of the LPF is compensated by multiplying a gain compensator (G), in (15) and the phase lag is compensated by multiplying a phase compensator (P), in (16). The new integrator with the gain and the phase compensator can be given as (17).

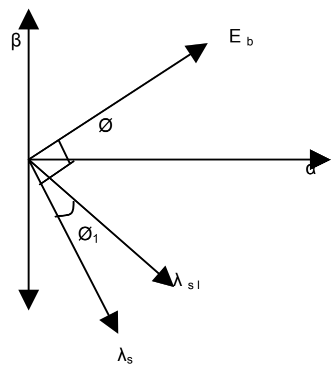

Figure.8 Vector Diagram of Programmable Low Pass Filter and Pure Integrator

$$Gain\ compensator(c) = \frac{\sqrt{\left(\omega_e^2 + a^2\right)}}{|\omega_e|} \quad (15)$$

$$Phase\ Compensator(p) = \exp(-j\phi_1) \quad (16)$$

$$\frac{\lambda_{sl}}{v_e} = \frac{1}{s+a}\left(\frac{\sqrt{\left(\omega_e^2 + a^2\right)}}{|\omega_e|}\right)\exp(-j\phi_1) \quad (17)$$

When the programmable LPF is transformed into the sampled-data model using the difference approximation, the sampled-data model has a modeling error, which, in turn, produces an error in the stator flux estimation. This error is more severe when the motor frequency is lower than the cut-off frequency of the LPF. Accordingly, the cut-off frequency cannot be located at fixed point far from the origin. If the pole is varied proportionally to the motor speed, the proportion of the motor frequency to the cut-off frequency of the LPF is constant. If the proportion is large, the estimation error will be very small. Consequently, the pole a is determined to be varied proportionally to the motor speed as (18). Therefore, the pole is located close to the origin in very low speed range and far from the origin in high speed.

$$a = \frac{|\omega_e|}{k} \quad (18)$$

Where k = constant. The time constant of the programmable LPF, $k/|\omega_e|$, is decreased with the increase of the motor speed. The value of K is found to be 5.5. Figure 9 and 10 shows the actual and estimated Flux and speed of the machine using programmable low pass filter algorithm. Figure 11 and 12 shows the flux vector trajectory due to fixed pole algorithm suggested in [5] and flux vector trajectory due to programmable low pass filter. It can be observed that stator fluxes are more stable in Programmable LPF when compared with fixed pole algorithms especially in 4-5 secs in the figure.

Thus from the above discussions, in Programmable low pass filter algorithm only one tuning parameter completely solves the problem due to DC drift and errors due to stator resistance estimate. More over PLPF algorithm solve stability problem due to fixed poles. The MATLAB/SIMULINK environment was used for simulations. An example of simulation results is given in figure 4.3. The speed reference was initially set to 2Hz. and a speed reversal to 10Hz. was applied (t = 0.5 s). The speed reference was changed to 10Hz. (t = 1.5 s) and a rated-load torque step was applied (t = 2.5s). Finally, the speed reference was lowered to 2Hz. (t = 3.5 s) while the rated load torque was still applied.

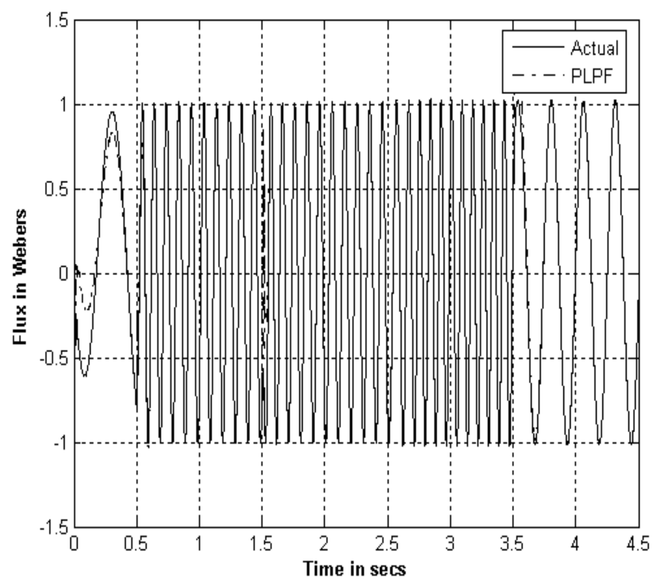

Figure.9 Actual and Estimated Flux PLPF

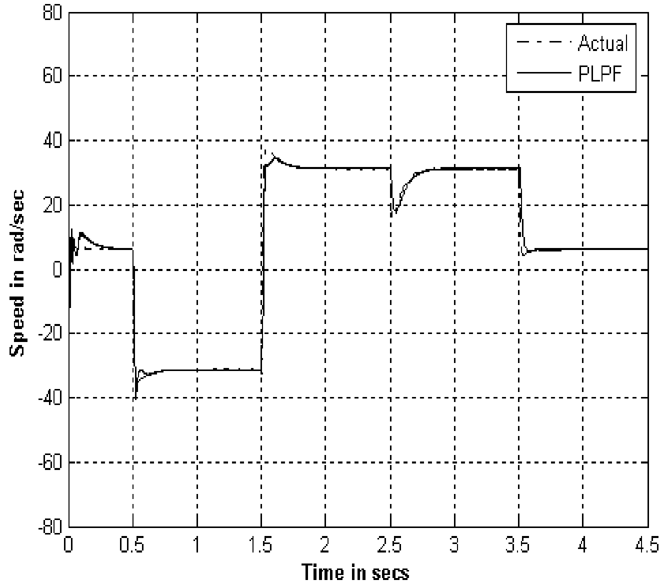

Figure.10 Actual and Estimated speed PLPF

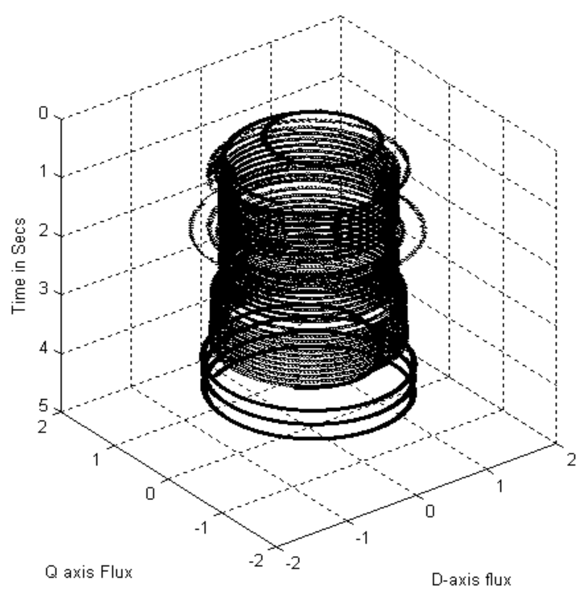

Figure.11 Stator Flux Vector Trajectory Due to Fixed Pole

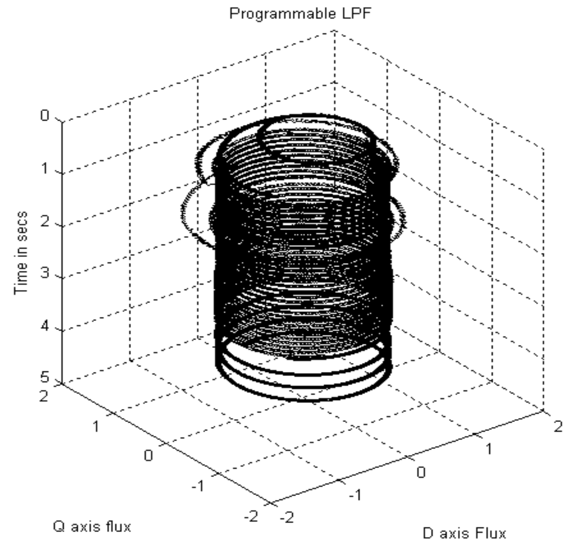

Figure.12 Stator Flux Vector Trajectory Due to Programmable LPF

VII. DIGITAL IMPLEMENTATION

The PLPF method of flux estimation is implemented in Texas instrument digital signal processor TMS320F2182 using Code Composer studio (a real time debugger for Texas instrument processors) and interfaced with MATLAB/SIMULINK environment. The DSP results are compared with simulation results and presented below. Figure 13 shows Flux estimated using DSP. Figure 14 shows Flux estimated using DSP and Matlab simulation. Figure 15 shows comparison of speed from Matlab simulation and DSP. In the implemented DSP algorithm, the time taken to process each sample is about 6.46 μ sec. The sampling frequency used in the implementation is 2 KHz. The program memory used in this implementation is just 18 bytes of the total available 32KB of memory.

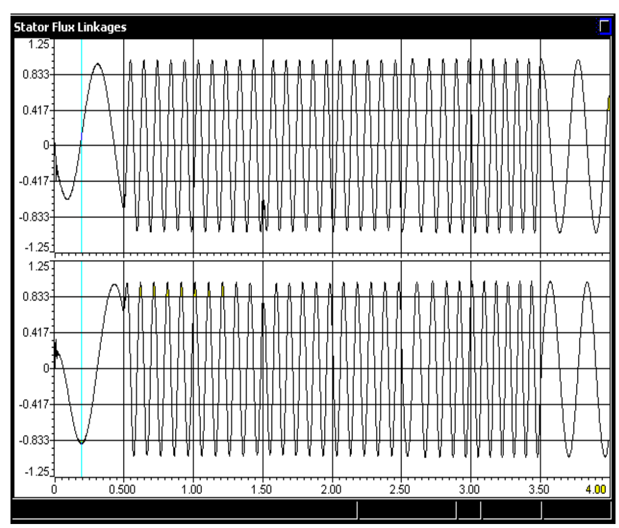

Figure.13 Flux Estimated using Code Composer studio (TMS320F2812)

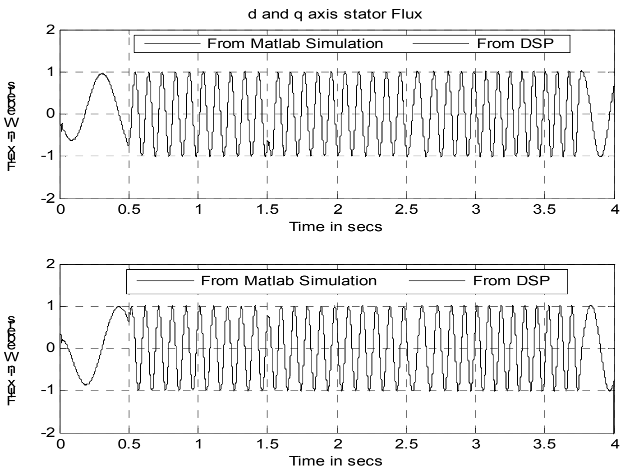

Figure-14 Comparison of Flux Estimated using Code Composer studio (TMS320F2812) and Matlab

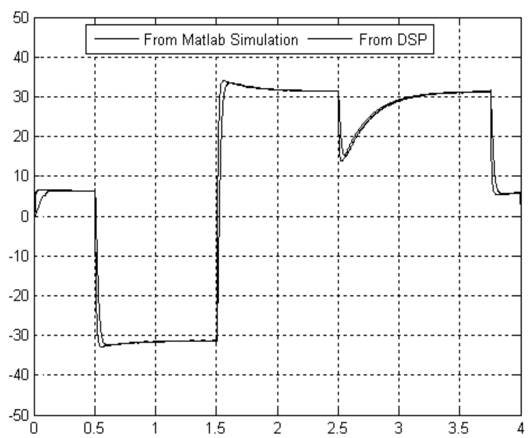

Figure-15 Comparison of Speed Estimated using Code Composer studio (TMS320F2812) and Matlab

VIII. CONCLUSIONS

This paper investigated the DC drift and errors due to stator resistance estimate in voltage model based flux estimation of speed sensorless induction motor drives. The stator flux was exactly estimated by the phase/gain compensator of the proposed programmable LPF for a wide speed range. In addition drift and stator resistance error estimate was much improved by the small time constant of the programmable LPF. Consequently the speed sensorless drive system is found to more stable than the system with the LPF with fixed pole. The validity of programmable LPF was verified through simulations.

REFERENCES

[1].I.Takahashi and T.Noguchi,"A new quick-response and high efficiency control strategy of Induction motor", *IEEE Transactions on Industrial application*, vol. 22, no 5, pp 820-827, 1986.

[2].M.Depenbrock, "Direct self-control (DSC) of inverter fed induction machine".*IEEE Transaction on power electronics*, vol. 3, no 4, pp 420-429, 1988.

[3].K.D.Hurst, T.G.Habetler, G.Griva and F.Profumo,"Zero speed tacholess IM torque control: simple a matter of stator voltage integration,". *IEEE Transactions on Industrial application*, vol. 34, no 4, pp 790-795, 1998.

[4].B.K.Bose and N.R.Patel,"A Programmable cascaded low pass filter based flux synthesis for a stator flux oriented vector control of induction motor drive," *IEEE Transactions on Industrial electronics*, vol. 44, no 1, pp 140-143, 1997.

[5].L.E.B.Da Silva, B.K.Bose and J.O.P.Pinto,"Recurrent neural networks based implementation of a programmable cascaded low pass filter used in stator flux synthesis of vector controlled of induction motor drive". *IEEE Transactions on Industrial application*, vol. 46, no 3, pp 662-665, 1999.

[6].M.H.Shin, D-S-Hyun, S-B Cho and S-Y Choe,"An Improved stator flux estimation for speed sensorless stator flux orientation control of induction motors". *IEEE Transaction on power electronics*, vol. 15, no 2, pp 312-318, 2000.

[7].Hinkkanen, M. and Luomi.J."Modified integrator for voltage flux estimation of Induction motors". *IEEE Transactions on Industrial electronics*, 50(4), pp 818-820.2003.

[8].J.Hu and B.Hu,"New integration algorithms for estimating motor flux over a wide speed range", *IEEE Transaction on power electronics*, vol. 44, no 5, pp 969-977, 1998

[9]."Induction machine Speed Estimation-Observations on Observers "Bo Peterson.Ph.D Thesis, Lund University.

[10].J.Jiang and J.Holtz,"High dynamic speed sensorless AC drive with online parameter tuning and steady state accuracy", *IEEE Transactions on Industrial electronics*, vol-44, no-2, pp 240-246.1997.