

ISTANBUL TECHNICAL UNIVERSITY ★ GRADUATE SCHOOL OF SCIENCE
ENGINEERING AND TECHNOLOGY

**COMPARISON OF
THREE IPMSM SENSORLESS POSITION ESTIMATION METHODS
THROUGH SIMULATIONS AND EXPERIMENTS**

M.Sc. THESIS

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Department of Control and Automation Engineering

Control and Automation Engineering Programme

Thesis Advisor: Asst. Prof. Dr.Ali Fuat ERGENÇ

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İSTANBUL TEKNİK ÜNİVERSİTESİ ★ FEN BİLİMLERİ ENSTİTÜSÜ

**İÇ SABİT MIKNATISLI SENKRON MOTORLARDA ÜÇ SENSÖRSÜZ
KONUM TAHMİNİ YÖNTEMİNİN SİMÜLASYONLAR VE DENEYLER
ARACILIĞI İLE KARŞILAŞTIRILMASI**

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To my wife,

FOREWORD

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ABBREVIATIONS

AF	: Active Flux
ATO	: Angle Tracking Observer
EEMF	: Extended Electromotive Force
EKF	: Extended Kalman Filter
GDM	: Gradient Descent Based Sensorless Method
IPMSM	: Interior Permanent Magnet Synchronous Motor
PLL	: Phase-Locked Loop
PMSM	: Permanent Magnet Synchronous Motor
SPMSM	: Surface Permanent Magnet Synchronous Motor

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COMPARISON OF THREE IPMSM SENSORLESS POSITION ESTIMATION METHODS THROUGH SIMULATIONS AND EXPERIMENTS

SUMMARY

In recent years, permanent magnet synchronous motors (PMSMs) have become more widely used in numerous different areas of application. The main advantage of the PMSMs is their high torque to mass ratio, which increases the dynamic performance of the motors while enabling them to be more compact. Strong permanent magnets in the rotor produce high flux linkages and eliminate the need of rotor coils. In position sensorless operation estimators, which use applied stator winding voltages and resulting stator winding currents as their inputs, are implemented in driver software instead of a mechanical sensor to measure the position of rotor. As a result, a more cost effective and reliable construction with less parts to maintain can be achieved. This topic is nowadays a very active research area, where various solutions to this problem are proposed and many researchers in the field extensively study their effectiveness.

The focus of this study is to compare the performances of three sensorless interior permanent magnet synchronous motor position estimation methods through simulations and experiments. All of the methods selected for comparison belong to the class of sensorless methods called as the back-EMF based methods, which are also known as, the fundamental excitation methods. These methods estimate states with equal frequency to the fundamental electrical frequency of the motor, which is proportional to its rotation speed.

First step is to determine a state, which contains the rotor position information. The most frequently employed states include back-emf, stator flux, extended-emf and active flux. Once the selected state is estimated the next phase is to extract the position and speed information from that state. The lowest possible speed at which stable and accurate position estimation is maintained, while efficient and sensorless speed control operation is in progress should be determined. Moreover, the sensorless position estimation methods are expected to remain stable and maintain accurate position estimation in the face of unavoidable parameter uncertainties that are present in all motor drive systems.

In the scope of this thesis, formulation, simulation, practical implementation, and experimentation studies have been performed to select the appropriate sensorless methods and their performances have been evaluated.

The simulations based on different observers were performed at different speeds and under different loads and their behaviors in both steady and transient states were studied. The simulations performed to determine the position estimation errors of each observer while operating under parameter uncertainty conditions were also investigated.

In the simulation results, it was shown that all three methods had similar estimation capabilities within 200 to 2000 rpm speed and 0 to 2 Nm load ranges. It was observed that the position estimation error of EEMF increased significantly during rapid transitions in load. This behavior, resulting from the speed adaptive nature of the EEMF method, was not observed in the cases of the AF and GDM. In these methods, increases in error during load transitions were insignificant.

The effects of parameter uncertainty on estimation accuracy were investigated in the last part of the simulations. In the AF and GDM, strong sensitivity to variation in values of permanent magnet flux (or in short flux), stator resistance (R) and quadrature inductance L_q were observed. On the other hand, the EEMF method was found to be sensitive only to the changes in L_q , resulting in a more robust method. All three methods were insensitive to variations in L_d as expected since the simulations were performed with zero direct current reference.

The aim of first set of experiments was to determine the effects of gain selection on each of the three sensorless methods. Detailed gain selection guidelines for each method were formed according to the results of these experiments. It was found that EEMF has the most complicated gain selection procedure, because it has four gains to tune. Gain selection for AF was found to be relatively simpler since it features two gains, and GDM enjoyed the easiest gain selection procedure among the three as it contains only one gain that is needed to be tuned.

Then experiments to determine the accuracy of the estimated position of the rotor magnets were carried out under steady state conditions and during transient operations. In the first set of the experiments, a relative encoder to measure the position were used in the control functions of the driver software. Then, the experiments were repeated while operating in a completely sensorless manner at steady and transient states, where the estimated rotor position was used in control functions. Finally, additional experiments to evaluate the sensitivity of observers to uncertainties in parameters of motors were performed.

The results of the experiments showed that all three methods were capable of full sensorless operation, in which, estimated position was used in the transformations that appear in control functions of driver software, within 200 to 2000rpm speed and 0 to 2 Nm load ranges. In accordance with the results obtained in simulations, EEMF performed poorly during transient operating conditions, in comparison to the other two methods. A difference between the simulation and experimentation results of AF has been detected during experiments at steady state operation. In experiments, AF position estimation error exhibited a load dependent characteristic, especially at low speeds, which was not observed during the simulations performed with this method.

The effects of the inverter nonlinearity compensation function used in the driver software on sensorless control performance, had also been investigated, and it is determined to play a crucial role in determination of the lowest possible speed at which stable sensorless operation can be maintainable. As a result, the need to use such a compensation function in junction with each of the three sensorless methods had been confirmed.

The comparisons up to this point revealed some complementary characteristics between the EEMF and AF. The EEMF produced slightly more accurate estimations at steady state operating conditions than the AF, whereas the AF showed considerably superior performance during transient operating conditions. The EEMF also had the important robustness property that allowed its position estimation to remain accurate in the face of variations in the stator phase resistance and permanent magnet flux linkage. The AF displayed high sensitivity to variations in both of these parameters.

The complementary characteristics of the AF and EEMF revealed in this study led to the proposal of a Combined AF-EEMF method, which incorporated the strengths of both methods. Through experiments, this new method was demonstrated to produce

accurate position estimations both in steady state and in transient operating conditions. Furthermore, the Combined AF-EEMF method was found to be insensitive to variations in the stator phase resistance and permanent magnet flux linkage, as it contained the EEMF method. The Combined AF-EEMF was also shown to identify these two parameters accurately under most operating conditions.

In conclusion, this study features several benefits regarding the investigation of the sensorless control problem and the methods proposed as a solution to it. First of all, a motor drive platform, which can be used as a tool to study and compare different sensorless control techniques has been obtained. As an initial use of this tool, the three of the sensorless methods commonly used for sensorless control of the IPMSMs have been studied in detail to reveal their strengths and weaknesses. This type of study also has the use of providing a base on which new sensorless methods can be designed, via combining the parts of the compared methods in an innovative manner as proposed in the Combined AF-EEMF method. In this regard, this tool can be used as rapid implementation and testing platform to compare a larger number of sensorless methods.

During these studies, considerable differences between results of the simulations and experiments have also been observed. These differences were most likely caused by the nonlinearities in the inverter, which remained uncompensated even in the presence of a compensation function, and the unmodeled characteristics of the IPMSMs such as the spatial harmonics of rotor permanent magnet flux linkage and the stator phase inductances, which are left out of the scope of this study. Future studies should aim to take into account, the effects of these phenomena on sensorless position estimation problem with the help of software and hardware tools developed in this study.

İÇ SABİT MIKNATISLI SENKRON MOTORLARDA ÜÇ SENSÖRSÜZ KONUM TAHMİNİ YÖNTEMİNİN TASARIM, ANALİZ VE UYGULAMALARI

ÖZET

İç sabit mıknatıslı senkron motorların kullanımı son yıllarda giderek yaygınlaşmaktadır. Bu tip motorların en büyük avantajı, motorların dinamik performansını artırmada etkili olan kütle başına üretebildikleri torkun yüksek olmasıdır. Böylece motorlar daha kompakt bir yapıya sahip olurlar. Motorun rotor kısmında bulunan güçlü sabit mıknatısların oluşturduğu manyetik akı sayesinde rotor sargılarına gerek olmaksızın motorun çalışması sağlanmış olur. Senkron motorların çalışabilmesi için, rotor ve stator manyetik alanlarının birbirlerine göre eş zamanlı bir şekilde dönmeleri gerekir. Dolayısıyla, bunu sağlayabilmek için, motor hareket halindeyken her durumda mıknatıs konumlarının bilinmesi gerekmektedir. Ayrıca, diğer senkron motorlarda kullanılan komütatör ve fırçaların da sabit mıknatıslı motorlarda kullanılmalarına gerek yoktur. Bu parçaların kullanılmaması motorun ömrünü de uzatır. Sonuç olarak, sabit mıknatıslı senkron motorların kullanımı artmakta ve bu araştırma alanı giderek yaygınlaşmaktadır.

Motor kontrolünde enkoder, hall sensörü gibi konum algılayıcıların kullanılmasının birçok dezavantajı vardır. Bunların başında kullanılan konum algılayıcı ve kablolarının maliyetinin toplam sistem maliyetini artırması gelir. Ayrıca, sistemdeki toplam parça sayısı arttığı için sistemin kurulum işlemi karmaşıklaşır ve sistem arızalara karşı daha dayanıksız hale gelir. Sıklıkla tercih edilen uygun maliyetli konum algılayıcılarının sağlayabildiği konum çıkışının çözünürlüğü ise sensörsüz uygulamalar ile karşılaştırıldığında daha düşüktür. Bununla beraber, kurulum sırasında meydana gelebilecek konum hizalanması hatalarından kaynaklanan sorunların giderilmesi ancak ek kalibrasyon süreçlerinin kullanılmasıyla mümkün olur.

Bu nedenlerle iç sabit mıknatıslı motorların (IPMSM) konum sensörsüz alan kontrolü konusu yirmi yılı aşkın süredir aktif bir araştırma alanı olmayı sürdürmektedir. Kısaca sensörsüz kontrol problemi olarak bahsedilen bu problem dört alt başlıkta incelenebilir. Bunlar, konum bilgisi içeren bir sistem durumunun doğru tahmin edilebilmesi, tahmin edilen bu durumu kullanarak mıknatıs konum ve hız tahminlerinin oluşturulması, kullanılan sensörsüz konum tahmini yönteminin motor parametrelerindeki belirsizlikler karşısında doğru çalışmayı sürdürebilmesi ve motor sürücü devresinin doğrusal olmayan davranışlarının etkilerinin kompanse edilmesidir.

Literatür araştırması sonucu önerilen yöntemlerin çoğunun yüzey mıknatıslı senkron motorlar (SPMSM) için önerilmiş olduğu tespit edilmiştir. Bu tip motorlarda konum bilgisi sadece zıt emk içerisinde bulunur. Bu nedenle bu motorların sensörsüz konum kontrolü, IPMSM motorlara göre daha kolaydır. IPMSM motorlarda stator sargısı bobinlerinin endüktanslarının konuma bağlı değişimler göstermesi sensörsüz konum problemini zorlaştırmaktadır.

IPMSM için önerilen konum sensörsüz konum tahmini yöntemleri, kısaca sensörsüz yöntemler, araştırıldığında ön plana çıkan üç yöntem tespit edilmiştir. Bunlardan birincisi adaptif gözlemciler sınıfına aittir. IPMSM'deki bobin endüktanslarındaki konuma bağlı değişkenliği de hesaba katacak şekilde hayali yeni bir sistem durumu önerilmiştir. Genişletilmiş zıt-emk (extended back-EMF, EEMF) olarak adlandırılan

bir durum ortaya atılmıştır. Bu tez çalışmasında incelenen ilk yöntem EEMF durumunu hıza bağlı adaptif özellik taşıyan bir gözlemci aracılığı ile tahminini içermektedir. Tahmin edilen EEMF daha sonra açı izleyen bir gözlemciye gönderilerek (Angle Tracking Observer, ATO), konum ve hız tahminleri elde edilmektedir. İlerleyen bölümlerde bu yöntemden kısaca EEMF olarak bahsedilecektir.

Tez çalışması kapsamında karşılaştırılan ikinci yöntem aktif akı (Active Flux, AF) olarak adlandırılan başka bir hayali durumun tahmini üzerine kurulmuştur. Aktif akı, rotor mıknatısının oluşturduğu akıya ek olarak motor anisotropisinden kaynaklanan bobin endüktansları arasındaki farka bağlı olarak da değişebilen ayrı bir akı ile ilişkili bir terime de sahiptir. Bu yeni durum aracılığı ile IPMSM motorların SPMSM motorlara denk bir yapıya dönüştürülmesi sağlanmaktadır. Aktif akının kullanımı sayesinde, içerisinde hıza bağlı olan bir ifade bulunmayan bir stator akısı gözlemcisi tasarlanması mümkün olmuştur. Bu gözlemci IPMSM'lere ait stator koordinatlarında tanımlı bir gerilim modeli ve rotor koordinatlarında tanımlı bir akım modelinin birleşiminden oluşmaktadır. Stator koordinatlarında ifade edilen tahmini stator akısı kullanılarak aktif akı hesaplanır ve son olarak trigonometrik bağıntılar aracılığıyla aktif akı içerisindeki tahmini konum bilgisi çıkarılır. İlerleyen bölümlerde bu yöntemden kısaca AF olarak bahsedilecektir.

Bu karşılaştırma çalışmasında yer alan üçüncü yöntemde doğrusal olmayan bir gözlemci kullanılmaktadır. Bu gözlemci, stator akısı ve rotor akılarının genlikleri arasındaki cebri bir denklikten faydalanmaktadır. Tahmini stator akısı kullanılarak bu denkliğin sağlanıp sağlanmadığı kontrol edilir ve denkliği bozan büyüklük hata olarak stator akı gözlemcisine geri bildirilir.

Bu üç yöntem Simulink ortamında simüle edilerek ve gerçek bir motor sürücüsü üzerinde uygulanarak deneyler yapılmıştır. Simulasyon sonuçları, kullanılan her üç gözlemci yapısının ve bu yapılar için seçilen kazanç değerlerinin, 200-2000 devir/dk hız aralığında ve 0-2 Nm yük aralığında kararlı ve doğru konum tahminleri üretebildiğini göstermiştir. Yöntemlerin kararlı haldeki davranışlarının birbirine yakın olduğu gözlenirken, ani yük değişimleri esnasında EEMF yönteminin diğer yöntemlere göre daha yüksek konum tahmini hatalarına sahip olduğu görülmüştür. Simulasyonların ikinci kısmında motor parametrelerindeki belirsizliklerin sensörsüz yöntemlerin konum tahmin doğruluğu üzerine etkilerini ortaya koyacak simülasyonlar gerçekleştirilmiştir. Bunlar sonucunda L_q parametresindeki değişikliklerin her üç yönteminde konum tahminini belirgin bir biçimde etkilediği, L_d parametresindeki değişikliklerin ise bunun tam aksine üç yöntemin konum tahmin sonuçlarında da belirgin farklılıklara yol açmadığı gözlemlenmiştir. Motor sabiti (mıknatıs akısı genliği olarak da düşünülebilir) ve stator faz direncindeki belirsizliklerin ise AF ve GDM yöntemlerinin konum tahminini büyük ölçüde etkilediğini gösteren simulasyon sonuçları elde edilmiştir.

Bu çalışmanın ikinci kısmında ise AF, GDM ve EEMF isimli sensörsüz konum algılama yöntemleri, fiziksel bir deney düzeneği üzerinde gerçekleştirilmiştir. Bu deney düzeneğinin ana parçaları, 900 W gücünde iç mıknatıslı bir motor, Renesas RX62T mikrokontrolör tabanlı bir motor sürücüsü ve 0-25 Nm aralığında tork üretebilen bir hysteresis freninden oluşmaktadır. Hysteresis freninin ürettiği tork, kontrollü bir akım kaynağı tarafından kontrol edilmektedir. Ayrıca frenin miline bağlı bir algılayıcı aracılığıyla fren mili hızı ve yük torku gerçek zamanlı olarak ölçülebilmektedir. Bu ölçümler, bir NI-PXI sistemine gönderilmekte ve bu sistemdeki kontrol yazılımı yardımıyla voltaj kontrollü akım kaynağına istenen yük torku ile doğru orantılı düşük gerilim kontrol sinyalleri gönderilmektedir. NI-PXI

sistemin üzerinde LabVIEW yazılım platformunda geliştirilen bu kontrol yazılımı sayesinde, histeresis freni ile istenen zamana bağlı tork profillerinin gerçekleşmesi sağlanmaktadır.

Deney düzeneğindeki motor sürücü yazılımı, alan hizalamalı motor kontrolü (Field Oriented Control(FOC)) için gerekli tüm fonksiyonlara sahiptir. Ayrıca sensörsüz konum algılama yöntemlerinin doğruluğunu değerlendirebilmek için sisteme bir de enkoder tipi konum algılayıcı ilave edilmiştir. Sürücü yazılımındaki önemli değişkenlerin değerleri 234000 bps hızındaki bir seri haberleşme aracılığıyla gözlem ve kayıt alma amacı ile kullanılan bir PC yazılımına gönderilmektedir. LabVIEW ortamında geliştirilen bu gözlem ve kayıt yazılımı eşzamanlı olarak 10 değişkenin 200 örnek/saniye hızında veya 1 değişkenin 2000 örnek/saniye hızında örneklenmesine ve elde edilen eğrilerin kaydedilmesine olanak sağlamaktadır.

Bu çalışmada kullanılan yukarıda bahsedilen yazılımların tamamı, tüm alt fonksiyonları dahil olmak üzere, Arçelik Elektrik Motorları Arge Laboratuvarı'nda geliştirilmiştir.

Yapılan deneylerin ilk bölümünde sensörsüz yöntemlerin içinde yer alan kazanç katsayılarının tahmin doğruluğuna etkisi incelenmiştir. Bunun sonucunda her üç yöntem için de kazanç seçme prosedürleri ortaya konmuştur. Kazanç seçme prosedürlerinin karmaşıklıklarının yöntemin içerdiği kazanç sayısına bağlı olduğu tespit edilmiştir. EEMF dört kazanç ile ayarlanması en zor yöntem olarak belirlenmiş, sadece tek bir kazançla sahip GDM yönteminin ise ayarlanması en kolay yöntem olduğu ortaya çıkmıştır.

Deneylerin ikinci kısmında sabit çalışma koşullarında ve geçiş koşullarında yapılan incelemeler sonucu yöntemlerin konum tahmin yeteneklerinin karşılaştırılması hedeflenmiştir. Sonuçlara bakıldığında sabit çalışma koşullarında her üç yöntem de oldukça başarılı sonuçlar vermesine rağmen EEMF ve GDM yöntemlerinin konum tahmin hatalarındaki yüke bağlı değişimlerin AF yöntemine göre daha az olması nedeniyle, daha yüksek performansa sahip oldukları sonucuna varılmıştır. Ani hız değişimi ve ani yük değişimi içeren geçiş koşullarında ise EEMF konum tahmini hatasında belirgin artışlar meydana geldiği gözlemlenilmiştir. Bu davranış, EEMF yönteminin formülasyonu esnasında yapılan bazı varsayımlardan ve de bu yöntemin hız adaptif bir yöntem olmasından kaynaklanmaktadır. Formülasyon esnasında yapılan varsayım, hız ve akım türevlerini içeren bazı terimlerin sıfıra yakın olduğu yönündedir. Geçiş anlarında bu türevlerin ihmal edilemez büyüklüklere ulaşması EEMF'in performansını düşüren birinci etkidir. İkinci etken ise, EEMF yönteminin yapısında bulunan genişletilmiş-emk gözlemcisinin çalışabilmek için tahmini hız bilgisine ihtiyaç duymasıdır. Geçiş anlarındaki ani değişimler, tahmini hızdaki hatanın artmasına neden olmaktadır. Bu da, tahmini genişletilmiş-emk'daki ve dolayısıyla tahmini konumdaki hataların da artması anlamına gelmektedir. Bu süreç de, tahmini hızdaki hatanın daha çok büyümesine neden olarak tüm hataların artmasına neden olacak bir hata sarmalını meydana getirmektedir. Öte yandan AF ve GDM yöntemlerinin içerdiği stator akı gözlemcilerinin hız bilgisine ihtiyaç duymaması, tahmin doğruluklarını geçiş koşulları altında da koruyabilmelerini sağlamaktadır.

Üçüncü grup deneylerde, simülasyonların ikinci kısmında olduğu gibi tahmini motor parametreleri ile bu parametrelerin gerçek değerleri arasında fark oluşması halinde konum tahmin yöntemlerinin bu durumdan nasıl etkilendiği incelenmektedir. Bunların sonucunda tespit edilen parametre hassasiyetleri simülasyonlar sonucunda tespit edilenlerle örtüşmektedir.

Deneyler içerisindeki bir grup da, sürücünün voltaj üretimindeki doğrusallık bozulmalarını telafi etmek amacı ile kullanılan yazılım fonksiyonunun, konum tahmini üzerindeki etkisini ortaya çıkarmak amacı ile gerçekleştirilmiştir. Bu deneyler sonucunda bu fonksiyonun sensörsüz kontrolün kararlı olarak çalışmayı sürdürebildiği minimum hızı düşürme yönünde önemli bir faydası olduğu tespit edilmiştir. Ayrıca bu fonksiyonun devreye alınması her üç yöntem içinde, 500 devir/dk altındaki hızlarda, konum tahmini hatasında yükteki değişimlere bağlı olarak meydana gelen değişimlerin azalmasını da sağlamıştır.

Bu noktaya kadar yapılan karşılaştırmalar, AF ve EEMF yöntemlerinin birbirlerinin eksiklerinin tamamlayıcı özelliklere sahip olduklarını ortaya koymuştur. EEMF sabit çalışma koşullarında AF'den daha doğru konum tahmini sonuçları ortaya koymuştur. Ani yük ve hız geçişlerinde ise AF hatasının, EEMF'ninki ile karşılaştırıldığında çok daha düşük olduğu gözlemlenmiştir. Ayrıca EEMF'nin stator direnci ve motor sabiti parametrelerindeki değişimlere karşı duyarsız olması, bu iki yöntemin bir arada kullanılabilmesine olanak sağlayacak faydalı bir özellik olarak ortaya çıkmıştır. Bu tespitler doğrultusunda bu çalışma kapsamında "Birleştirilmiş AF-EEMF" adında yeni bir yöntem önerilmiştir. Bu yöntemin yukarı da bahsedilen her iki yönetmin de güçlü yönlerini bünyesinde barındıracağı öngörülmüştür.

Son gruptaki deneyler ile, Birleştirilmiş AF-EEMF yönteminin bu öngörülen özelliklerini doğrulanmıştır. Bu yöntemin sabit çalışma koşullarında, EEMF kadar, geçiş koşullarında ise AF kadar hassas konum tahmini yapabildiği ve de EEMF gibi stator direnci ve motor sabitindeki değişimlerden etkilenmediği gözlemlenmiştir. Son olarak AF-EEMF yönteminin aynı anda stator direnci tahmini ve motor sabiti tahmini yapmak için de kullanılabileceği doğrulanmıştır. Ancak, direnç tahmini özelliğinin değişken yük veya hız koşulları gerektirdiği ve faz akımı genliklerindeki değişimlerin çok küçük olduğu sabit çalışma koşullarında, kullanılamayacağı da belirtilmiştir.

Özetle, bu tez kapsamındaki çalışmalar sonucunda seçilen üç yöntemin de IPMSM sensörsüz kontrolü için uygun olduğu tespit edilmiş ve aralarındaki benzerlikler ve farklılıklar detaylı olarak incelenmiştir. Ayrıca çalışma kapsamında oluşturulan, yazılım ve donanım yeteneklerinin, daha çok sayıda yöntemin de benzer şekillerde karşılaştırılarak değerlendirilmesine ve Birleştirilmiş AF-EEMF yöntemi gibi inovatif çözümlerin uygulanabilirlik çalışmalarının hızlıca tamamlanmasına olanak sağlayacağı öngörülmektedir. Tüm bu çalışmaların, gelecekte de araştırmacıların ilgisini çekmeyi sürdüreceği düşünülen sensörsüz motor kontrolü probleminin daha iyi anlaşılması yönündeki ilerlemelere katkıda bulunması hedeflenmektedir.

1. INTRODUCTION

1.1 Organization of This Study

The focus of this study is to compare the performances of three sensorless interior permanent magnet synchronous motor position estimation methods through simulations and experiments. The scope of the study is limited to the operation of IPMSM motors up to rated speed and rated load. Operation in flux weakening region is excluded.

All of the methods selected for comparison belong to the class of sensorless methods called as the back-EMF based methods (also known as the fundamental excitation methods). The aim is to find a simple enough method, which is feasible to be realized in software of low cost microcontrollers commonly used in motor control applications typically seen in household appliances, while satisfying the estimation requirements. In this respect, the motor used in simulations and experiments was selected to have a power around 1kW.

The main performance criterion is determined as the lowest speed at which stable, accurate position estimation can be maintained, while efficient and sensorless speed control operation is in progress. The methods are required to fulfill this criterion at least at speeds as low as 300rpm.

Robustness is the second most important performance criterion that is investigated in this study. The sensorless position estimation methods are expected to remain stable and maintain accurate position estimation in the face of unavoidable parameter uncertainties that is present in all motor drive systems.

In order to fulfill these purposes literature review, formulation, simulation, practical implementation, and experimentation studies have been performed through which appropriate sensorless methods have been selected and their performances have been evaluated.

This study is composed of five chapters. The first chapter starts by a brief explanation of the superior performance characteristics of permanent magnet synchronous motors. Then, continues by a discussion on the motivation behind this vast area of research, known as the sensorless control of high performance PMSM drives, which has been continued in more than two decades and has still not failed to capture the interest and imagination of many researches from all around the world.

Following sections of first chapter include a detailed discussion on the parts of the sensorless position estimation problem, which aims to introduce the reader to the challenges commonly faced in sensorless motor control. The first chapter ends with a detailed review of literature, in which solutions proposed for each of the previously described challenges are addressed.

The second chapter includes a detailed formulation of each of the three sensorless position estimation methods. In order to facilitate the observer derivation, the chapter starts with a description of the motor model, its fundamental equations, representations in stationary and rotating reference frames and the coordinate transformations frequently used to alternate between these representations. Then, the states commonly utilized in sensorless methods such as stator flux, back-EMF, extended back-EMF (EEMF) and active flux are described in detail.

In the following sections of chapter two, formulations of EEMF observers in stationary and rotating reference frames, a hybrid active flux observer that uses both of the motor's voltage model in stator coordinates and current model in rotor coordinates and a gradient search based stator flux observer are discussed. At the end of the each section, gain tuning guidelines are presented for the observer studied in that section.

Chapter three details the simulations performed to compare the performances of methods described in chapter two, and presents the results of these simulations. In the introduction of this chapter, the model used for simulations in the Simulink is described in detail, where, simulations at different speeds and under different loads have been performed. Behaviors of the observers in both steady state and transient operating conditions have been studied as well. In the final section of this chapter, the simulations performed for determining the position estimation errors of each observer while operating under parameter uncertainty conditions are presented.

The subject of fourth chapter is the practical implementation of the sensorless position estimation methods discussed in the previous chapters. The chapter starts with a description of the experimental setup. It continues with presentation of the results of the experiments that were performed to determine the effects of gain selection for each of the three observers separately. Then, in the preceeding sections, experiments carried out under steady state conditions, and during transient operations are discussed. In the first set of the experiments, a relative encoder to measure the position was in the control functions of a driver software. Then, the experiments at steady and transient states were repeated while operating in a completely sensorless manner, where the estimated rotor position was used in control functions. Finally, sensitivity of observers to uncertainties in parameters of motors were evaluated through additional experiments. In each of these sections, results are represented in a comparative format to illustrate the differences and similarities between the three sensorless position estimation methods.

This chapter also includes a section on the subject of inverter nonlinearity compensation, which is a necessary feature of practical implementation of any sensorless method that is aimed to work at low speeds. In the first half of this section, the inverter nonlinearity compensation method employed in this study is explained. In the second half, the beneficial effects of using such a compensation function, on position estimation performance of the sensorless methods are presented through the experimental results.

In the last section of Chapter4 a new method is proposed, in which two of the previously studied methods, the EEMF and AF, are combined. Firstly, the observations about the behaviors of the EEMF and AF methods that led to the proposal of this new method and the advantages it is expected to have over the AF and EEMF are explained. Secondly, construction of this method, which is called as the combined AF-EEMF, is described. Then the position estimation performance of the combined AF-EEMF is demonstrated through the measurements performed during steady state and transient operating conditions. Finally, the stator phase resistance and permanent magnet flux linkage identification features of the combined AF-EEMF are presented.

In the final chapter, the experimental results and the results of simulations are summarized and their outcomes are evaluated. The chapter ends with a last section in which possible future studies, that would further the investigations carried out in this study are suggested.

1.2 Advantages of Permanent Magnet Synchronous Motors

In recent years, permanent magnet synchronous motors (PMSMs) have become more widely used in numerous different areas of application like home appliances, electric vehicles, factory lines, elevators etc.

The main advantage of the PMSMs is their high torque to mass ratio, which enables them to be more compact. High torque to mass ratio also increases the dynamic performance of the motors. Strong permanent magnets, which are capable of producing high flux linkages, in the rotor render these advantageous properties to the PMSMs[1]. Moreover, using permanent magnets in rotor eliminates the need for the rotor coils. The motor efficiency is also greatly improved without resistive loss in the rotor. Having no coils on the rotor also eliminates the need to use commutators and brushes that are essential parts of the other synchronous motors. Since wearing out of commutator brushes is no longer an issue, the lifetime of the motor is improved as well.

As a result, a more cost effective and reliable construction with less parts to maintain is achieved. The PMSM motors are gaining more popularity due to aforementioned advantages and even spreading to areas in which only induction motors have been used until now.

1.3 Motivation For Sensorless Position Control

In order to be able to take advantage of the superior properties of the PMSMs, the orientation of the permanent magnet flux of the rotor with respect to stator windings has to be known at all times. This information is needed to be able to generate a stator magnetic field that rotates with that of the permanent magnets in the rotor in synchrony. By proper control of the angle between these two magnetic fields, efficient motor operation with minimum torque ripple is achieved. Field oriented

control and direct torque control are among the most commonly used control strategies for this purpose.

In PMSMs, the rotor position information can be obtained through shaft mounted position sensors such as encoders, resolvers and hall effect sensors [1]. However, using sensors have several disadvantages, which have led to the study of position sensorless speed and position estimation of PMSMs. First of all, the presence of sensors and their related cabling increases the complexity and cost of the driver. Moreover, the commonly used cost effective sensors are usually with low angular resolution and often have a limited operating range in terms of speed. Misalignment due to variances in mounting also decreases the accuracy of the sensors. In order to reduce this effect, an additional calibration procedure is needed [2]. Sensors also need maintenance over time to confirm continuation of their initial performances. Furthermore, with the presence of sensors the drive system becomes more vulnerable to variations in environmental conditions [3].

On the other hand, in position sensorless operation there is no mechanical sensor to measure the position of rotor. Instead, estimators, which use voltages applied to stator windings and resulting stator winding currents as their inputs, are implemented in the driver software. In addition, most estimators include a model of the motor under study or some partial information regarding its parameters. By the help of this information, position and speed of the motor, which are essential for motor control can be estimated. In sensorless operation, there are only electrical connections between the motor and its driver electronics, so overall drive installation cost is reduced [4]. This topic is now a very active research area, where many solutions to this problem are proposed and many researchers in the field extensively study their effectiveness [5].

However, this way of operation is not completely sensorless. The system includes current sensors, which are needed to measure instantaneous values of currents of stator coils. These currents are used in current controllers and position estimators. From now on whenever sensorless methods or estimators are mentioned in this study what actually meant is position sensorless methods and estimators. This interchangeable naming is used frequently in the related literature.

1.4 Parts of The Sensorless Control Problem of IPMSMs

As mentioned in the previous sections, the sensorless control problem can be defined as controlling the operation of a PMSM without using a mechanical position sensor. This problem can be divided into the following parts: state estimation, speed and position extraction from estimated state, parameter identification, and compensation of inverter nonlinearities. Each of these subproblems are investigated separately in this section differing from the common practice in the literature. The solutions proposed in the literature usually aim to solve the complete problem, ending up with quite similar solutions that only differ in their approaches to one of the subproblems. Actually, it is more practical and useful to identify each part of this problem separately. This way, a set of criteria against which each proposed estimator is evaluated can be established. Moreover, parts of different sensorless methods can also be combined.

1.4.1 State estimation

First of all, a state that contains rotor position information has to be determined. The most frequently employed states include back-emf, stator flux, extended-emf and active flux. The details of the properties of these states are described in the next chapter. An observer that is able to accurately estimate the selected state is needed in order to solve this problem.

1.4.2 Position and speed extraction from estimated state

Once the selected state is estimated the next phase is to extract position and speed information from that state. For this part of the problem, trigonometric functions can be used to determine the phase of the estimated state, which can be represented as a vector in two-dimensional space. The estimated state vectors can be represented either in stationary stator or in rotating rotor reference frames. The position of the rotor can be determined by using the magnitudes of components of the estimated state vector and the arctangent function of the phase of the vector. However, this often results in noise in position estimations so other types of position extraction methods are proposed as part of sensorless control methods. Among these are phase-locked loops and angle tracking observers[6].

1.4.3 Operation under parameter uncertainty

State observers are frequently used to estimate a state, which contains position information as described in Section 1.4.1. Since such observers usually employ a motor model, the parameters of the motors have to be known for them to function properly. Unfortunately, it is not possible to know the exact values of these parameters at all times, as they are mostly dependent on operating conditions. Resistance and motor constant of PMSMs, which is proportional to the magnet flux linkage, are temperature dependent parameters. Furthermore, due to poor current control, magnets can be partially demagnetized, resulting in significant and permanent changes in the permanent magnet flux linkage parameter. Finally, quadrature inductance of PMSMs decreases when core is saturated under high loads where stator phase current is high. These types of phenomena along with the gradual wearing of the motor introduce uncertainties in parameter values. The resulting value mismatches the actual parameter and the parameter used in the state observer can cause position estimation errors or even instability.

There are two ways to resolve this issue. Firstly, the state observer can be augmented with a parameter identification scheme. In this case, even if the value of a parameter changes as described above the new value of this parameter is identified and the estimation remains unaffected. Secondly, a robust observer structure can be proposed, where the sensitivity to parameter changes is lowered.

1.4.4 Inverter nonlinearity compensation

In state estimator, the inputs used to estimate the desired state are the phase voltages and currents. The currents are measured in all drives, as they are also required in the motor control. In some motor drivers, sensors measuring phase voltages are also used, however, this significantly increases the cost of the driver. As a result, the voltage sensors are not used in most motor drivers. Instead, the reference voltages generated by the current controller are assumed to be equal to the actual phase voltages applied to the phases windings of the motor by the inverter of the motor driver. Based on this assumption, reference voltages are used in the observer calculations instead of the actual values. As a result, the need for voltage sensors is omitted.

Nevertheless, this assumption is not valid because there is always a small difference between these voltages, which is also called as the distortion voltage or the inverter nonlinearity. The position estimation error introduced by inverter nonlinearity is not significant during motor operations at medium or high speeds, but as the motor slows down, the amplitude of the phase voltages also decrease proportionally with the speed of the motor. As a result, the ratio of distortion voltage to phase voltage amplitude increases and the resulting position estimation becomes significant.

The inverter nonlinearity has several sources of origin. The most dominant source among these is the deadtime. Deadtime is defined as the time gap introduced between switching-off of a power switch and the switching on of another that is placed on the same inverter arm with the first one. During the deadtime, both switches on an inverter arm are turned off, thus voltage of the phase controlled by this arm is solely determined by the direction of the current flowing through that phase. This phenomenon causes a distortion voltage between reference and actual phase voltages.

Similarly, turning on and off times of the power switches, the conduction losses of the power switches and power diode contribute to the inverter nonlinearity.

This issue can be resolved by implementation of an inverter nonlinearity compensation function in the motor driver software. By this function, the amount of distortion voltage present due to inverter nonlinearity is calculated and added to the reference voltages in order to compensate the distortion voltage. As a result, the actual phase voltages and the reference voltages are equalized. Devising such a compensation function is another part of the sensorless control problem.

1.5 Review of Literature

Over the last two decades, a vast number of methods has been proposed for the sensorless position estimation and this area of research is now considered to be mature by many researchers [7,8,9]. In order to be able to investigate these methods the following classification is followed throughout study.

The distinction that divides the methods into two groups is based on the source of position information they use [7]. In this regard, the first group is called back-EMF based methods, where the position information stored in the orientation of rotor

permanent magnet is utilized. The position and speed of the rotor flux shapes the back-emf through law of induction, hence the name of the group.

The second group is called saliency based methods in which position information is stored in the phase inductances. In this case, the motor is required to be salient, as the name suggests. However, under saturation conditions, these methods can also be applied to non-salient motors in which a saliency is induced by operating the motors [10].

1.5.1 Back-emf based sensorless methods

Back-emf based methods are also known as fundamental excitation methods, as they aim to estimate states with equal frequency to the fundamental electrical frequency of the motor, which is proportional to its rotation speed.

In the fundamental equations of the motors, terms including the multiplication of speed and currents appear, both of which are states of the system when PMSM systems are considered. This nonlinear behaviour requires the use of a nonlinear observer to estimate the position and speed of these motors.

In most of the early proposed methods, only the case of surface mounted permanent magnet motors is considered. As SPMSMs are non-salient, the only term containing the position information is the back-emf in fundamental equations of the motor. However, in Interior Permanent Magnet Motors (IPMSMs), in addition to back-emf, stator coil inductances are also position dependent due to presence of saliency.

Saliency is defined as the position dependent change in phase inductances. When the representation of motor in rotor coordinates is considered the inductance in direct axis is called L_d and the inductance in quadrature axis which is perpendicular to the direct axis is called L_q . In salient motors L_d is not equal to L_q .

Presence of saliency further complicates the position estimation problem. The scope of this study is limited to the case of IPMSMs, therefore only sensorless methods that are proposed for salient motors and those which include nonlinear estimators are reviewed.

1.5.1.1 Speed adaptive observers

One group of nonlinear observers proposed for the position estimation in IPMSMs are called speed adaptive observers. Speed adaptive observers consist of a linear Luenberger type state observer, in which the motor speed appears as a constant parameter, and a speed adaptation scheme that gradually updates speed using the error between estimated and measured currents[11]. In [12], the estimated state is selected as stator flux, where the adaptive observer is constructed in estimated rotor reference frame.

In [13], extended back-emf (EEMF) is proposed as a new state variable that can be used in speed adaptive observers. EEMF includes position information from both back-EMF and stator inductances, thus when motor equations are expressed in terms of EEMF, it remains as the only position dependent term in the equations. The EEMF disturbance observers in stator and rotor reference frames have been shown to be able to estimate position in IPMSMs [13,14]. It has been found that both of these disturbance observers had very similar performances at steady state operating conditions regardless of the reference frame used. A least-order observer has also been proposed to estimate the EEMF [15]. In the steady state it has been shown to produce accurate speed and position estimations. However, significant position estimation errors have been observed during acceleration. These are attributed to the omission of stator current transients during the derivation of EEMF concept.

1.5.1.2 Active flux observers

Active flux concept has been proposed [16], to be able to transform IPMSMs to a form equivalent to SPMSMs whose position estimation problem is considerably simpler as described in the previous section. Active flux is also referred to as fictitious permanent magnet flux [17] or extended flux [18]. Active flux is an extension of permanent magnet flux by an imaginary flux, which is modeled to emulate the reluctance torque creating the motor saliency property. For this reason, it is also interpreted as the torque-producing flux in salient-pole motors. Furthermore, it is aligned to the direct rotor axis. An active flux observer and a PLL type position speed observer are constructed and shown to be able to produce accurate estimations down to 1rpm with a 6kW IPMSM that has a strong saliency [16]. It should be noted that speed estimation is not required in this active flux observer. In another active

flux observer, both the IPMSM voltage model in stationary reference frame and current model in rotating reference frame are used together [18]. These two models are linked by coordinate transformations and the estimated rotor position used in these transformations are derived from the trigonometric relation between the components of active flux represented in the stationary reference frame. Again no speed dependency is present in the observer design.

1.5.1.3 Gradient descent search based sensorless methods

Position observers based on standard gradient search have been shown to have great stability properties [19]. These observers do not require the knowledge of speed. They are constructed on an algebraic relationship between inductance related and permanent magnet related components in the stator flux. However, in salient-pole motors such as IPMSMs, inductances are also position dependent so their contribution to the stator flux cannot be computed without knowledge of rotor position. For this reason, gradient descent based methods (GDM) have been initially proposed for non-salient motors [19]. Nevertheless, they have been shown to retain most of their significant stability properties when applied to the salient motors without any modifications. Moreover, an extension to these observers has been proposed to further increase their performance when used for estimation of position in salient motors[20].

1.5.1.4 Saliency based methods

In salient PMSMs, the inductances along stationary coordinates change as the rotor position changes. This property can be used to determine position of the permanent magnet at any given moment. In a group of observers, high frequency signals are injected in to the stator signals to detect the changes in stator inductances[21]. In most cases, a high frequency voltage signal is added on top of existing voltages. This results in a high frequency current. The measured current is demodulated and phase and amplitude difference between high frequency voltage and current are used to calculate position dependent impedances from which position estimation can be calculated. There are numerous schemes of this type of high frequency injection techniques in literature, however these types of observers are out of the scope of this study.

1.5.2 Position and speed extraction

Different methods can be used to extract speed and position information from the estimated states. The simplest and the most direct method is the use of arctan function to determine the phase of an estimated stated vector[22]. Unfortunately, the trigonometric functions are known to generate noisy results. To circumvent this problem, phase-locked loop type structures are proposed to extract speed and position estimation [7] from estimated states. Extended Kalman Filters can also be used for this purpose [9]. Their robustness and ability to work with noisy signals are significant advantages but in exchange they add significantly to the computational load of the complete sensorless control application.

1.5.3 Parameter identification

In all of the speed and position estimators described above, whole or part of the motor model are utilized in calculations. Depending on the type of the observer, difference in sensitivity to mismatch different motor parameters is observed. To prevent a decrease in the estimation accuracy resulting from parameter Uncertainty, many parameter identification methods have been proposed to be used alongside with state observers in sensorless control systems. In [23], a scheme for quadrature inductance estimation is presented which is based on high frequency injection. There are numerous schemes, in which stator resistance is estimated [24]. A parameter identification scheme based on system identification theory is proposed [25]. Here, a parameter matrix is formed and a recursive least method square in which applied voltages and measured currents in rotating reference frame are used to perform the online estimation of motor parameters. In this method, direct and quadrature inductances can be estimated along with resistance while the motor is rotating.

1.5.4 Inverter nonlinearity compensation

Inverter nonlinearities are known to degrade the back-emf based sensorless methods, especially while the motor is operating at low speed as explained in the Section1.4.4. In most cases, a positive compensation voltage is determined experimentally. The amount of compensation voltage depends on the charactesistics of the inverter such as the amount of deadtime, turn-on and turn-off times of the power switches and on-state voltage drops across power switches and diodes. Once the appropriate amount of compensation voltage is determined, a compensation function is implemented, in

which, for each phase the compensation voltage is either added or subtracted from the reference voltage of that phase according to the direction of the current of the phase under consideration. Instead of directly adding or subtracting the whole compensation voltage at the points where phase currents change their signs, a ramping weighting coefficient, whose value increases upto 1 in proportion with the amount of phase current, is suggested to be used. This enables using voltages less than compensation voltage at operating conditions, in which phase currents are close to zero. This second method has been evaluated to result in more effective inverter nonlinearity compensations, where the value of compensation voltage is also determined with parameter identification methods [26].

In a different approach, an automatic search scheme is proposed, in which DC currents references at two different values are generated interchangeably in a periodic manner while the motor is at standstill [27]. Using the generated voltage references, a distortion voltage is calculated at the end of each period. Then the value of compensation voltage is automatically changed by a PI controller to diminish the distortion voltage. This closed loop scheme results in generating the necessary compensation voltage needed for different phase current levels without the need of knowledge of any inverter parameters.

2. FORMULATION OF SENSORLES METHODS

Starting point in any type of study concerning the control or state estimation of IPMSMs, is to construct a useful model of the motor. The model has to be accurate enough to represent all the fundamental behaviors of the motor, which have direct impact on the drive performance. On the other hand, it is also important to keep the details in a model at a manageable level in order to improve the feasibility of analysis, simulation and implementation studies. Therefore, it is necessary to exclude some of the secondary behaviors of the IPMSM, which have minor effects on the drive performance. These principles are taken into account in the two motor models that are presented in this section.

In this chapter, several sensorless methods are formulated. The performance of sensorless methods can be evaluated with respect to several criteria. These criteria can be various, such as the limits of stable operation conditions (speed and torque), accuracy of estimation during steady state and transient operations, ease of tuning procedure, robustness, computational resources required by the software implementation, etc.

Some of the important performance criteria are related to extending the range of operation conditions. For example, the lowest possible speed at which the observer operates in a stable manner is very critical for most applications. This is especially important for fundamental excitation-based sensorless methods since the amplitude of fundamental excitation decreases proportionally with the speed, which in turn obscures the observability of the whole system. The range of torque that results in accurate estimation is also important. Furthermore, the estimation accuracy should not deteriorate under transient operating conditions. This means that operation should be stable under rapid speed and load changes.

The topics that are discussed in this section provide a useful foundation on top of which the studies discussed in the following chapters can easily be built.

2.1 The IPMSM Model

The IPMSMs are a subcategory of the PMSMs. In PMSMs, the magnetic field of the rotor is created by the permanent magnets in the rotor. Therefore, the only currents in the motor are those of the stator phases, which create the magnetic field of the stator. The interaction between these two magnetic fields produces the electromagnetic torque that rotates the rotor. There are no rotor windings or brush contacts on the motor.

In IPMSMs the magnets are placed inside the rotor as opposed to the surface of it. The effect of this configuration is explained further along the construction of the motor model.

The differential equations that constitute the motor model can be based on electromagnetic conversion, which takes place while the motor is operating. Faraday's induction principle facilitates this conversion. The rotation of the rotor causes the magnetic flux linkages of the stator phase windings to vary in time, which in turn causes an emf to be induced in the stator windings. This phenomenon is commonly referred to as the back-emf. Fig.2.1 depicts an equivalent circuit of a single stator phase. Some of the energy absorbed by the voltage source representing the back-emf is converted to the mechanical energy via the electromagnetic torque that acts on the rotor.

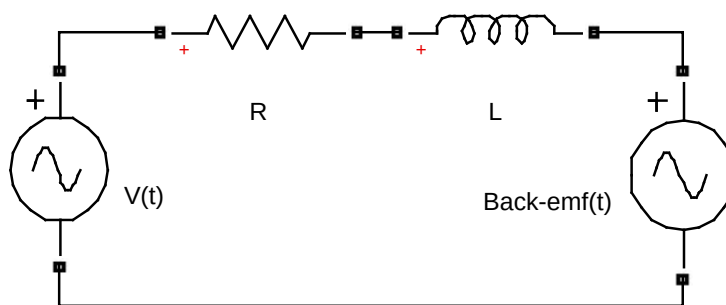


Figure 2.1 : Equivalent circuit of a stator phase winding.

Ennis obtained by applying the Kirchhoff's voltage law on the circuit in. The stator has three symmetrical stator phases, so the other phases have the same equivalent

circuits. In Eqn(2.1) the variance of signals with respect to time is explicitly notated. After this point in this study, time dependence is not explicitly shown, but always implied unless otherwise stated for brevity of the equations.

$$\begin{aligned} v_u(t) &= Ri_u(t) + L_u \frac{di_u(t)}{dt} + emf_u(t) \\ v_v(t) &= Ri_v(t) + L_v \frac{di_v(t)}{dt} + emf_v(t) \\ v_w(t) &= Ri_w(t) + L_w \frac{di_w(t)}{dt} + emf_w(t) \end{aligned} \quad (2.1)$$

As stated earlier, back-emf is the time derivative of the stator flux linkage component, which is caused by the rotor permanent magnets. In Eqn(2.1), it is shown that the stator flux linkage of each phase has another component due to its phase current. Complete phase stator flux linkages are defined in Eqn(2.2).

$$\begin{aligned} \Psi_u^s &= L_u i_u + \Psi_u^{pm} \\ \Psi_v^s &= L_v i_v + \Psi_v^{pm} \\ \Psi_w^s &= L_w i_w + \Psi_w^{pm} \end{aligned} \quad (2.2)$$

Eqn(2.1) and Eqn(2.2) can be combined to obtain Eqn(2.3), which is a more compact representation.

$$\begin{aligned} v_u &= Ri_u + \frac{d\Psi_u^s}{dt} \\ v_v &= Ri_v + \frac{d\Psi_v^s}{dt} \\ v_w &= Ri_w + \frac{d\Psi_w^s}{dt} \end{aligned} \quad (2.3)$$

In addition to these, in conventional PMSM drives the following additional equations are also satisfied.

$$\begin{aligned} v_u + v_v + v_w &= 0 \\ i_u + i_v + i_w &= 0 \end{aligned} \quad (2.4)$$

Eqn(2.4) suggests that there are actually two independent currents and voltages instead of three. In fact, the three- phase- system can be reduced to an equivalent two- phase system by Eqn(2.4). For this reduction, space orientations of the phase windings with respect to each other are considered (Fig. 2.1).

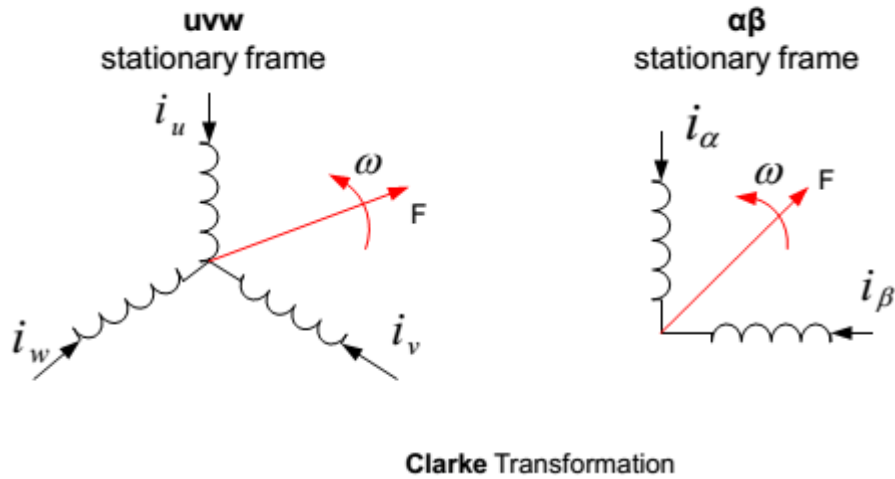


Figure 2.2 : Clarke transformation [31].

There is 120 electrical degrees of spatial phase difference between each pair of phase voltages. In order to be able to take into account these space orientations, the space vector representation is introduced. In space vector representation, the electrical quantities such as voltages, currents and flux linkages are all represented by 2 dimensional vectors. Two-dimensional vectors can also be represented by complex numbers whose real and imaginary parts correspond to the magnitude of the orthogonal components of the space vector.

The two-phase system can be represented either in the stationary reference frame of the stator or in the rotating reference frame of the rotor. Both of the representations are commonly employed in speed and position estimation schemes, so both of them are formulated in this chapter. It is customary to first transform the three-phase representation to stationary two-phase representation, which is referred to as the stator reference frame in this study. The alpha axis coincides with the u-phase of the three phase representation, and the beta axis leads alpha axis by 90° in the counter-clockwise direction (Fig.2.2).

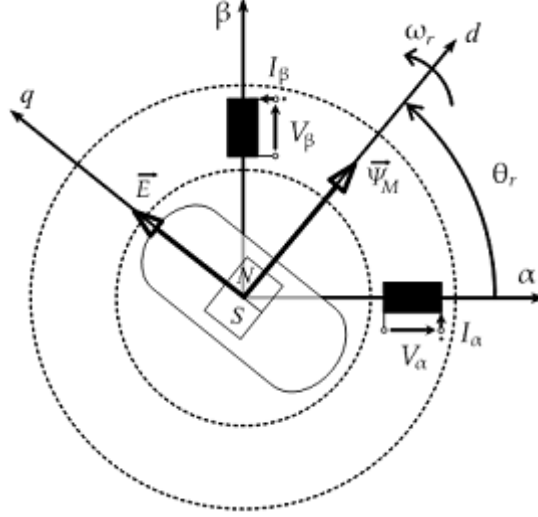


Figure 2.3 : The stationary and rotating reference frames [6].

Conversion from three-phase representation to two-phase representation is also referred to as the Clarke Transformation (Fig.2.2), which is represented in Eqn(2.5) for voltages. The same transformation equations are valid for the currents in the model.

$$\begin{aligned} v_\alpha &= v_u \\ v_\beta &= \frac{\sqrt{3}}{3}v_u + \frac{2\sqrt{3}}{3}v_v \end{aligned} \quad (2.5)$$

The inverse of this transformation, which can be called as the Inverse Clarke Transformation is presented in Eqn(2.6).

$$\begin{aligned} v_u &= v_\alpha \\ v_v &= -\frac{1}{2}v_\alpha + \frac{\sqrt{3}}{2}v_\beta \\ v_w &= -\frac{1}{2}v_\alpha - \frac{\sqrt{3}}{2}v_\beta \end{aligned} \quad (2.6)$$

The rotating reference frame, which is also called the dq, is fixed to the rotor. The d-axis coincides with the north pole of the rotor magnet, which also points to the peak of the flux created by the permanent magnet. As in the stationary reference frame, q-axis leads the d-axis by 90 degrees in the direction of angular speed. The angle between the rotating and stationary reference frames, which is commonly referred to as the rotor position and it is denoted by θ_r (or in short by θ). As a result, the rotor position has to be known in order to perform the transformations between the stationary and rotating frame representations. The transformation from stationary to rotating

reference frame is also known as the Park Transformation and it is expressed in Eqn(2.7).

$$\begin{aligned} v_d &= \cos(\theta)v_\alpha + \sin(\theta)v_\beta \\ v_q &= \cos(\theta)v_\alpha - \sin(\theta)v_\beta \end{aligned} \quad (2.7)$$

The inverse of this transformation is called as the Inverse Clarke Transformation and shown in Eqn(2.8).

$$\begin{aligned} v_\alpha &= \cos(\theta)v_d - \sin(\theta)v_q \\ v_\beta &= \cos(\theta)v_d + \sin(\theta)v_q \end{aligned} \quad (2.8)$$

With the help of Clark Transformation, phase representation of PMSM that is given in Eqn(2.1) and Eqn(2.2) can be transformed into the stationary 2-phase representation as given in Eqn(2.9).

$$\begin{aligned} v_\alpha &= Ri_\alpha + (L_0 + L_1 \cos(2\theta))\frac{di_\alpha}{dt} + L_1 \sin(2\theta)\frac{di_\beta}{dt} - \omega\Psi_{pm} \sin(\theta) \\ v_\beta &= Ri_\beta + L_1 \sin(2\theta)\frac{di_\alpha}{dt} + (L_0 - L_1 \cos(2\theta))\frac{di_\beta}{dt} + \omega\Psi_{pm} \cos(\theta) \end{aligned} \quad (2.9)$$

where,

$$\begin{aligned} L_0 &= \frac{(L_d + L_q)}{2} \\ L_1 &= \frac{(L_d - L_q)}{2} \end{aligned} \quad (2.10)$$

in which L_d , L_q and Ψ_{pm} are inductances along d and q-axis, and amplitude of permanent magnet flux respectively. If the terms containing the position dependent inductances are grouped together and matrix form notation is used, (2.9) can be expressed as,

$$v_{\alpha\beta} = Ri_{\alpha\beta} + L_0 \frac{di_{\alpha\beta}}{dt} + L_1 \frac{d(Qi_{\alpha\beta})}{dt} + \frac{d\Psi_{\alpha\beta}^{pm}}{dt} \quad (2.11)$$

where,

$$\begin{aligned}
v_{\alpha\beta} &= \begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix} \\
i_{\alpha\beta} &= \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} \\
\Psi_{\alpha\beta}^{pm} &= \begin{bmatrix} \Psi_\alpha^{pm} \\ \Psi_\beta^{pm} \end{bmatrix} \\
Q &= \begin{bmatrix} \cos 2\theta & \sin 2\theta \\ \sin 2\theta & -\cos 2\theta \end{bmatrix}
\end{aligned} \tag{2.12}$$

Eqn(2.9) can be represented in a more compact form by introducing the stator flux notation ($\Psi_{\alpha\beta}^s$):

$$\begin{aligned}
v_{\alpha\beta} &= Ri_{\alpha\beta} + \frac{d\Psi_{\alpha\beta}^s}{dt} \\
\Psi_{\alpha\beta}^s &= \Psi_{\alpha\beta}^i + \Psi_{\alpha\beta}^{pm} \\
\Psi_{\alpha\beta}^i &= (L_0 I + L_1 Q) i_{\alpha\beta}
\end{aligned} \tag{2.13}$$

where $\Psi_{\alpha\beta}^i$ is the component of stator flux due to stator currents.

When $L_d \neq L_q$, Eqn(2.9) becomes quite complex, which in turn complicates the analysis required in position estimation studies. This complexity is considerably reduced when the rotating reference frame representation, as shown in Eqn(2.14), is considered.

$$\begin{aligned}
v_d &= Ri_d + L_d \frac{di_d}{dt} - \omega L_q i_q \\
v_q &= Ri_q + L_q \frac{di_q}{dt} + \omega L_d i_d + \omega \Psi_{pm}
\end{aligned} \tag{2.14}$$

The further advantages and disadvantages of using stationary or rotating reference frames are discussed as they are employed in various position estimation methods that are examined in the later chapters.

It should be noted that the motor model described up to this point is a simplified model, in which several secondary phenomena, presented below, are omitted:

1. Harmonics of back-EMF and inductances
2. Saturation of iron core(i.e. L_d and L_q is assumed to be constant)
3. Core losses

2.2 Determination of The States To Be Estimated

In this section, several possible candidate states for the estimation are evaluated. It is possible to extract position information from these estimated states as discussed in Section 1.4.2. Only the easily implementable reference frame representation, either stationary or rotating, is presented for each state for brevity.

2.2.1 The stator flux state

The stator flux is a state commonly used in many sensorless algorithms as it contains the flux produced by the permanent magnet of the rotor. It also contains the position information. When it is considered to represent the two-phase stationary reference frame motor equations in Eqn(2.15) in the compact form as in Eqn(2.15), the following is obtained.

$$\begin{aligned} v_\alpha &= Ri_\alpha + \frac{d\Psi_\alpha^s}{dt} \\ v_\beta &= Ri_\beta + \frac{d\Psi_\beta^s}{dt} \end{aligned} \tag{2.15}$$

Eqn(2.15) can be further combined to yield,

$$v_{\alpha\beta} = Ri_{\alpha\beta} + \frac{d\Psi_{\alpha\beta}^s}{dt} \tag{2.16}$$

where, if stator flux is chosen as the system state, state space representation of Eqn(2.16) becomes,

$$\frac{d\Psi_{\alpha\beta}^s}{dt} = v_{\alpha\beta} - Ri_{\alpha\beta} \tag{2.17}$$

2.2.2 The back-EMF state

Another position dependent variable in the motor model is the back-EMF induced in the stator windings. However, in observer constructions that try to estimate the back-EMF, phase currents are selected as the state variables instead of stator windings. This is mostly used in the case of surface permanent magnet synchronous motors (SPMSM) as $L_d=L_q$, which greatly simplifies Eqn(2.18) as follows:

$$\begin{aligned} v_\alpha &= Ri_\alpha + L_s \frac{di_\alpha}{dt} - \omega \Psi_{pm} \sin(\theta) \\ v_\beta &= Ri_\beta + L_s \frac{di_\beta}{dt} + \omega \Psi_{pm} \cos(\theta) \end{aligned} \quad (2.18)$$

When currents are selected as system state, then the state space representation becomes,

$$\frac{di_{\alpha\beta}}{dt} = \frac{v_{\alpha\beta}}{L_s} - \frac{R}{L_s} i_{\alpha\beta} - \frac{emf_{\alpha\beta}}{L_s} \quad (2.19)$$

where $L_s = L_d = L_q$ is the stator synchronous inductance and e_s is the back-emf which is defined as,

$$e_s = \begin{bmatrix} e_\alpha \\ e_\beta \end{bmatrix} = \begin{bmatrix} -\omega \Phi \sin(\theta) \\ \omega \Phi \cos(\theta) \end{bmatrix} \quad (2.20)$$

Eqn(2.20) is used as the system model in disturbance observers, which can be used to estimate back-emf. However, Eqn(2.19) is only valid for SPMSMs. In IPMSMs due to saliency phase inductances are also rotor position dependent. In other words, back-emf is not the only term containing position information as presented in Eqn(2.9).

2.2.3 The extended-EMF state

As discussed in the previous section, the back-EMF cannot be directly used for IPMSM position estimation. To overcome this problem an extension to back-emf has been suggested [13], which is called as the extended-emf. Following the derivation there, Eqn(2.14) is rewritten in matrix form as,

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \begin{bmatrix} R_s + pL_d & -\omega L_q \\ \omega L_d & R_s + L_q \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} 0 \\ \omega \Psi_{pm} \end{bmatrix} \quad (2.21)$$

Eqn(2.21) can be rewritten as,

$$\begin{bmatrix} v_d \\ v_q \end{bmatrix} = \begin{bmatrix} R_s + pL_d & -\omega L_q \\ \omega L_q & R_s + L_d \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} 0 \\ (L_d - L_q)(\omega i_d + \frac{di_q}{dt}) + \omega \Psi_{pm} \end{bmatrix} \quad (2.22)$$

where p is used as the time derivative operator. Transforming Eqn(2.22) into α - β reference frame yields Eqn(2.23):

$$\begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix} = \begin{bmatrix} R_s + pL_d & \omega(L_d - L_q) \\ -\omega(L_d - L_q) & R_s + pL_d \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + ((L_d - L_q)(\omega i_d + \frac{di_q}{dt}) + \omega \Psi_{pm}) \begin{bmatrix} -\sin \theta \\ \cos \theta \end{bmatrix} \quad (2.23)$$

When Eqn(2.23) is compared with the original α - β reference frame representation in Eqn(2.9), it is seen that there is only one term containing position information in Eqn(2.23) whereas there are many position dependent terms in Eqn(2.9).

From Eqn(2.23), the extended-emf is defined as,

$$e_{\alpha\beta} = \begin{bmatrix} e_\alpha \\ e_\beta \end{bmatrix} = ((L_d - L_q)(\omega i_d + \frac{di_q}{dt}) + \omega \Psi_{pm}) \begin{bmatrix} -\sin \theta \\ \cos \theta \end{bmatrix} \quad (2.24)$$

In contrast to back-emf, extended emf can also be used in speed and position estimation of IPMSMs.

2.2.4 The active flux state

Active flux is an extension of the rotor permanent magnet flux [16]. It points towards the same direction as the rotor flux, which is the d-axis direction, and its magnitude is defined as,

$$\Psi_a = \Psi_{pm} + (L_d - L_q)i_d \quad (2.25)$$

where $L_d < L_q$ is assumed. Active flux is an imaginary concept, which does not have a direct physical correspondence. However, in order to interpret its meaning the expression of electromagnetic torque can be considered.

$$T_e = \frac{3}{2} p_{pole} (i_q \Psi_{pm} + (L_d - L_q) i_d i_q) \quad (2.26)$$

Where p_{pole} is the number of pole pairs. When Eqn(2.25) is substituted into Eqn(2.26), Eqn(2.27) is obtained.

$$T_e = \frac{3}{2} p_{pole} \psi_a i_q \quad (2.27)$$

As presented in Eqn(2.27), active flux can also be interpreted as the flux, which produces the torque.

The salient motors (IPMSM) can be treated as non-salient motors (SPMSM) with the help of the active flux concept. As a result, it allows adaptation of some of the estimation methods, which are originally proposed for non-salient motors to salient-motors with minimum modification.

2.3 Construction of The State Observers

In this section, the equations representing the state observers studied in the following Chapters 3 and 4 are derived. The topologies considered here are those of speed adaptive observers, i.e. observers based on gradient descent search and observes that utilize both dq and $\alpha\beta$ reference frame representations.

2.3.1 Speed adaptive observers

If dq frame representation of IPMSM presented in Eqn(2.14) is considered, it can be observed that there are terms, which include the multiplication of currents and speed. This in turn introduces a nonlinear behavior into the system.

Speed adaptive observers are based on the fact that speed varies considerably slower than currents [8]. Therefore, it can be interpreted as a constant parameter during the minor time intervals of the current estimation loop (which is can also be called as the current sampling time). Therefore, the speed adaptive observers can be implemented

as a combination of a fast state estimator, in which the speed is assumed to be constant, and a slower speed adaptation loop, which takes into account the gradual variations in speed. As a result, the necessary estimation accuracy can be maintained.

2.3.1.1 Extended-EMF observer in the stator reference frame

This observer is discussed in [25] and the formulation presented there is followed in this section. With the help of the extended-emf concept (Eqn(2.24)), IPMSM model in stator coordinates becomes,

$$\begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix} = \begin{bmatrix} R + pL_d & \omega(L_d - L_q) \\ -\omega(L_d - L_q) & R + pL_d \end{bmatrix} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} + \begin{bmatrix} e_\alpha \\ e_\beta \end{bmatrix} \quad (2.28)$$

if currents and extended emf are selected as the system states, then state space representation of Eqn(2.28) is,

$$\frac{d}{dt} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} = \frac{1}{L_d} \begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix} - \left(\frac{R_s}{L_d} I - \frac{\omega(L_d - L_q)}{L_d} J \right) \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} - \frac{1}{L_d} \begin{bmatrix} e_\alpha \\ e_\beta \end{bmatrix} \quad (2.29)$$

In addition, definition of extended emf can be utilized to describe its behavior. Taking the derivative of Eqn(2.29) yields,

$$\frac{d}{dt} \begin{bmatrix} e_\alpha \\ e_\beta \end{bmatrix} = \omega J \begin{bmatrix} e_\alpha \\ e_\beta \end{bmatrix} + (L_d - L_q) \left(\omega \frac{di_d}{dt} + \frac{d^2 i_q}{dt^2} \right) \begin{bmatrix} -\sin \theta \\ \cos \theta \end{bmatrix} \quad (2.30)$$

The second term on the right hand side of Eqn(2.30) adds nonlinearity to the system as it involves multiplications of current derivatives and position. It is assumed to be zero as shown in Eqn(2.31) in this observer construction.

$$(L_d - L_q) \left(\omega \frac{di_d}{dt} + \frac{d^2 i_q}{dt^2} \right) \begin{bmatrix} -\sin \theta \\ \cos \theta \end{bmatrix} \approx 0 \quad (2.31)$$

with Eqn(2.31), Eqn(2.30) is simplified as,

$$\frac{d}{dt} \begin{bmatrix} e_\alpha \\ e_\beta \end{bmatrix} = \omega J \begin{bmatrix} e_\alpha \\ e_\beta \end{bmatrix} \quad (2.32)$$

If the extended-emf in Eqn(2.29) is replaced with its estimated value, the adaptive model is obtained as,

$$\frac{d}{dt} \begin{bmatrix} \hat{i}_\alpha \\ \hat{i}_\beta \end{bmatrix} = \frac{1}{L_d} \begin{bmatrix} v_\alpha \\ v_\beta \end{bmatrix} - \left(\frac{R}{L_d} I - \frac{\omega(L_d - L_q)}{L_d} J \right) \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} - \frac{1}{L_d} \begin{bmatrix} \hat{e}_\alpha \\ \hat{e}_\beta \end{bmatrix} \quad (2.33)$$

The reference model is the motor itself. In the observer construction, the derivatives of currents are obtained from both reference and adaptive models and the error between them is multiplied by observer gain (G) and feedback to update the estimated extended-emf as,

$$\frac{d}{dt} \begin{bmatrix} \hat{e}_\alpha \\ \hat{e}_\beta \end{bmatrix} = \omega J \begin{bmatrix} \hat{e}_\alpha \\ \hat{e}_\beta \end{bmatrix} + G \left(\frac{d}{dt} \begin{bmatrix} \hat{i}_\alpha \\ \hat{i}_\beta \end{bmatrix} - \frac{d}{dt} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} \right) \quad (2.34)$$

where G is,

$$G = aL_d I + (\omega - b)L_d J \quad (2.35)$$

In Eqn(2.35), a and b are the observer's design parameters.

As Eqn(2.34) involves terms with derivatives of the currents, the following change of variables is proposed,

$$\varepsilon_{\alpha\beta} = \hat{e}_{\alpha\beta} + G i_{\alpha\beta} \quad (2.36)$$

If Eqn(2.34) and Eqn(2.36) are used in Eqn(2.29), the observer equations become

$$\begin{aligned} \frac{d}{dt} \varepsilon_{\alpha\beta} &= \left(\frac{-1}{L_d} G + \omega J \right) \varepsilon_{\alpha\beta} + \left(G - \frac{R_s}{L_d} I + \frac{\omega(L_d - L_q)}{L_d} J + \frac{1}{L_d} G - \omega J \right) i_{\alpha\beta} + \frac{1}{L_d} v_{\alpha\beta} \\ \hat{e}_{\alpha\beta} &= \varepsilon_{\alpha\beta} - G i_{\alpha\beta} \end{aligned} \quad (2.37)$$

For this observer, a design guideline for the parameters in G is proposed in [13] as,

$$\begin{aligned} \alpha &= |\omega| \nu \\ \beta &= \omega \end{aligned} \quad (2.38)$$

where,

$$\alpha = \frac{g_1}{L_d} \quad (2.39)$$

$$\beta = \omega - \frac{g_2}{L_d}$$

There is an embedded filter in this disturbance observer. The choice of the observer gains affects the poles of this filter. ν is a free design parameter and it is chosen according to a trade-off between two desired observer characteristics. When ν is increased, observer error's sensitivity to speed estimation error decreases. On the other hand when ν is decreased the assumption in Eqn(2.31) is satisfied to a greater extent due to stronger filtering of high frequency components in estimated extended emfs. The details of the derivations of Eqn(2.38) and Eqn(2.39) have been reported in [16].

2.3.1.2 Extended-EMF observer in the rotating reference frame

The extended-emf observers can also be implemented in the rotating reference frame. However, in sensorless control, the actual rotor position is not known. Therefore, in this case, the estimated rotating reference frame, which is based on the estimated rotor position has to be employed. The relation between the estimated (also called as $\gamma\delta$ coordinates) and actual rotating rotor coordinates are displayed in Figure 2.4.

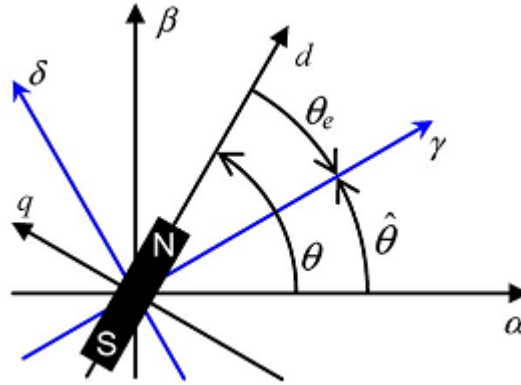


Figure 2.4 : The estimated reference frame [26].

γ -axis points in the direction of the estimated rotor position ($\hat{\theta}$) and the phase difference between the dq and $\gamma\delta$ coordinates is the position estimation error (θ_e).

Transforming IPMSM Eqn(2.28) from dq coordinates to $\gamma\delta$ coordinates results in,

$$\begin{bmatrix} v_\gamma \\ v_\delta \end{bmatrix} = \begin{bmatrix} R_s + pL_d & -\omega L_q + (\hat{\omega} - \omega)L_d \\ \omega L_q & R_s + L_d \end{bmatrix} \begin{bmatrix} i_\gamma \\ i_\delta \end{bmatrix} + \begin{bmatrix} e_\gamma \\ e_\delta \end{bmatrix} \quad (2.40)$$

Similarly, Eqn(2.24) is transformed into $\gamma\delta$ coordinates as follows,

$$\begin{bmatrix} e_\gamma \\ e_\delta \end{bmatrix} = \left((L_d - L_q) \left(\omega i_d + \frac{di_q}{dt} \right) + \omega \Psi_{pm} \right) \begin{bmatrix} -\sin \theta_e \\ \cos \theta_e \end{bmatrix} + (\hat{\omega} - \omega) L_d \begin{bmatrix} -i_\delta \\ i_\gamma \end{bmatrix} \quad (2.41)$$

The state space representation of the system represented by Eqn(2.40) and Eqn(2.41) is expressed as,

$$\frac{d}{dt} \begin{bmatrix} i_\gamma \\ i_\delta \end{bmatrix} = \left(\frac{-R}{L_d} I - \left(\frac{\omega L_q}{L_d} - \frac{d\theta_e}{dt} \right) J \right) \begin{bmatrix} i_\gamma \\ i_\delta \end{bmatrix} - \frac{1}{L_d} \begin{bmatrix} e_\gamma \\ e_\delta \end{bmatrix} + \frac{1}{L_d} \begin{bmatrix} v_\gamma \\ v_\delta \end{bmatrix} \quad (2.42)$$

$$\frac{d}{dt} \begin{bmatrix} e_\gamma \\ e_\delta \end{bmatrix} = \frac{d\theta_e}{dt} J \begin{bmatrix} e_\gamma \\ e_\delta \end{bmatrix} + (L_d - L_q) \left(\omega \frac{di_d}{dt} + \frac{d^2 i_q}{dt^2} \right) \begin{bmatrix} -\sin \theta_e \\ \cos \theta_e \end{bmatrix} \quad (2.43)$$

During the transition from Eqn(2.41)to Eqn(2.43),the derivative of extended-emf is derived. Here ω is assumed to be constant as its variation is considerably smaller than that of the extended-emf. Before constructing the observer,Eqn(2.43) is further simplified with two additional assumptions.

First,

$$\frac{d\theta_e}{dt} \approx 0 \quad (2.44)$$

is assumed because it is usually the case accept for the speed transients.

Second,

$$(L_d - L_q) \left(\omega \frac{di_d}{dt} + \frac{d^2 i_q}{dt^2} \right) \begin{bmatrix} -\sin \theta_e \\ \cos \theta_e \end{bmatrix} \approx 0 \quad (2.45)$$

is assumed. This assumption is also violated during transients, as derivatives of currents become non-zero. However, there is an embedded low-pass filter in the observer construction. This filters the high frequency disturbing effect of this term on rotor estimation in Eqn(2.43). With the assumptions in follows with the assumptions in Eqn(2.30) and Eqn(2.31), Eqn(2.43) is simplified as,

$$\frac{d}{dt} \begin{bmatrix} e_\gamma \\ e_\delta \end{bmatrix} = 0 \quad (2.46)$$

A state filter type observer is proposed in [28], for estimation of the extended emf in stator coordinates. The same approach is used here for the case of extended emf in the rotor coordinates. The following are the equations of this observer:

$$\frac{d\hat{i}_{\gamma\delta}}{dt} = \frac{1}{L_d} \left(v_{\gamma\delta} - (RI + \hat{\omega}_{filtered} L_q J) i_{\gamma\delta} - \hat{e}_{\gamma\delta} \right) \quad (2.47)$$

$$\hat{e}_{\gamma\delta} = g_1 \tilde{i}_{\gamma\delta} + g_2 \int \tilde{i}_{\gamma\delta} \quad (2.48)$$

In Eqn(2.47), $v_{\gamma\delta}$, $i_{\gamma\delta}$, $\hat{i}_{\gamma\delta}$, $\hat{e}_{\gamma\delta}$ denote the applied voltages, measured currents, estimated currents and estimated extended emf all represented in estimated rotor coordinates, respectively. In Eqn(2.48), current estimation error, denoted by $\tilde{i}_{\gamma\delta}$, is calculated as,

$$\tilde{i}_{\gamma\delta} = i_{\gamma\delta} - \hat{i}_{\gamma\delta} \quad (2.49)$$

In this observer, the current is estimated by the extended emf estimation as shown in Eqn(2.47). Then, the current estimation error is used to revise the extended emf estimation to a more correct value as presented in Eqn(2.48). In Eqn(2.48), there are two terms among which first and second terms are proportional to the current error and it's integral, respectively. This structure acts as a state filter to correctly estimate the extended emf. In Eqn(2.47) there is a term involving estimated speed ($\hat{\omega}$) therefore, to be able to estimate the current, the speed has to be estimated as well. As a result, this observer is classified as a speed adaptive observer.

An angle tracking the observer structure is proposed in [29], in which both the estimated position and speed are calculated from the components of the estimated EEMF.

First, a quantity related to the position estimation error has to be derived. For this purpose, components of the estimated extended emf are utilized. It is known that the actual extended emf is oriented towards the direction of permanent magnet flux, which is also the d-axis of rotating coordinates. If the EEMF estimation is correct,

the estimated and actual rotating coordinates should coincide, which means that the δ -axis component of estimated EEMF vanishes. Based upon this observation, the position estimation error can be computed as,

$$\tilde{\theta} = \arctan\left(\frac{\hat{\mathbf{e}}_{\delta}}{\hat{\mathbf{e}}_{\gamma}}\right) \quad (2.50)$$

The angle tracking observer can be represented with the following equations:

$$\frac{d\hat{w}}{dt} = g_3 \tilde{\theta} \quad (2.51)$$

$$\hat{\theta} = g_4 \hat{w} + \int \hat{w} \quad (2.52)$$

The estimated speed is used to calculate the estimated position in angle tracking observer, however, the actual speed changes at a lower rate than this speed estimate in most cases, as the estimated speed is affected by extended emf estimation errors, which are not related to the actual speed. To remove these high frequency components in the estimated speed, a further low pass filter is applied. The resulting filtered estimated speed is utilized in Eqn(2.47).

2.3.2 Active flux observer

Active flux is an extension of the rotor flux with the term $(L_d - L_q)i_d$. This term integrates the position dependent effect of saliency into the active flux term. As a result, in the voltage model of IPMSM, the only position dependent term is the active flux. With the active flux concept, it becomes possible to treat IPMSMs as non-salient SPMSMs. An observer, which takes advantage of this property is proposed in [18]. The formulation of this observer is examined in this section. The formulation starts with the fundamental motor model equation represented in rotor coordinates.

$$\begin{aligned} v_d &= Ri_d + L_d \frac{di_d}{dt} - \omega L_q i_q \\ v_q &= Ri_q + L_q \frac{di_q}{dt} + \omega L_d i_d + \omega \Psi_{pm} \end{aligned} \quad (2.53)$$

Eqn(2.53) can be rewritten in terms of active flux ($\Psi_a = \Psi_{pm} + (L_d - L_q)i_d$) as,

$$\begin{aligned}
v_d &= Ri_d + L_q \frac{di_d}{dt} - \omega L_q i_q + \frac{d\Psi_a}{dt} \\
v_q &= Ri_q + L_q \frac{di_q}{dt} + \omega L_q i_d + \omega \Psi_a
\end{aligned} \tag{2.54}$$

Equation (2.54) is the same as the equations defined for SPMSMs in dq-axis except for the term $\frac{d\Psi_a}{dt}$. This term only exists when i_d is not constant. Omitting this term and transforming Eqn(2.54) into the stationary reference frame results in,

$$\begin{aligned}
v_\alpha &= Ri_\alpha + L_q \frac{di_\alpha}{dt} - \omega \Psi_a \sin \theta \\
v_\beta &= Ri_\beta + L_q \frac{di_\beta}{dt} + \omega \Psi_a \cos \theta
\end{aligned} \tag{2.55}$$

Eqn(2.55) has exactly the same form as the equations of SPMSMs in stator coordinates. The differences are in the terms synchronous inductance L_s and permanent magnet flux linkage Ψ_{pm} , which are replaced with L_q and Ψ_a in Eqn(2.55), respectively. This equation can be rewritten in the matrix form and in terms of stator flux as,

$$\frac{d\Psi_{\alpha\beta}^s}{dt} = v_{\alpha\beta} - Ri_{\alpha\beta} \tag{2.56}$$

where stator flux is defined in terms of active flux as,

$$\Psi_{\alpha\beta}^s = \Psi_{\alpha\beta}^a + L_q i_{\alpha\beta} \tag{2.57}$$

Equations (2.56) and (2.57) are in the matrix form, where stator flux, active flux, stator voltages and stator currents are represented in stator coordinates are respectively denoted by $\Psi_{\alpha\beta}^s = [\Psi_\alpha^s \ \Psi_\beta^s]^T$, $\Psi_{\alpha\beta}^a = [\Psi_\alpha^a \ \Psi_\beta^a]^T$, $v_{\alpha\beta} = [v_\alpha \ v_\beta]^T$ and $i_{\alpha\beta} = [i_\alpha \ i_\beta]^T$.

In these two equations, the only term containing the position information, whose orientation is the same as that of the permanent flux is the active flux and its components can be calculated as,

$$\Psi_{\alpha\beta}^a = \begin{bmatrix} \Psi_\alpha^a \\ \Psi_\beta^a \end{bmatrix} = \begin{bmatrix} \Psi_a \cos \theta \\ \Psi_a \sin \theta \end{bmatrix} \tag{2.58}$$

It should be noted that through the utilization of the active flux concept, speed is not included in the equations(2.56) and (2.57),which constitute the voltage model of the IPMSM [18].

This observer utilizes the motor equations in both stator and rotor coordinates and both representations are linked through the stator flux term. For this purpose motor equations in rotor reference frame, given in Eqn(2.53),are rewritten in terms of the stator flux, $\Psi_{dq}^s = [\Psi_d^s \quad \Psi_q^s]^T$, as,

$$\Psi_{dq}^s = \begin{bmatrix} L_d i_d + \Psi_{pm} \\ L_q i_q \end{bmatrix} \quad (2.59)$$

Using equations (2.56), (2.57) and (2.59) a stator flux observer is constructed by the equations presented below:

$$\frac{d\hat{\Psi}_{\alpha\beta}^s}{dt} = v_{\alpha\beta} - Ri_{\alpha\beta} + G\tilde{i}_{\alpha\beta} \quad (2.60)$$

$$\hat{i}_{\alpha\beta} = T^{-1}(\hat{\theta})L^{-1}T(\hat{\theta})\hat{\Psi}_{\alpha\beta}^s - \frac{\Psi_{pm}}{L_d} \begin{bmatrix} \cos \hat{\theta} \\ \sin \hat{\theta} \end{bmatrix} \quad (2.61)$$

where $T(\hat{\theta})$ is the transformation from stator to the rotor coordinates, also known as the Park transform,

$$T(\hat{\theta}) = \begin{bmatrix} \cos \hat{\theta} & \sin \hat{\theta} \\ -\sin \hat{\theta} & \cos \hat{\theta} \end{bmatrix} \quad (2.62)$$

L is the inductance matrix,

$$L = \begin{bmatrix} L_d & 0 \\ 0 & L_q \end{bmatrix} \quad (2.63)$$

The variables with “^” designate the estimated quantities.

In Eqn(2.61), estimated stator flux, $\hat{\Psi}_{\alpha\beta}^s$, is used to current estimations $\hat{i}_{\alpha\beta}$. Then the current estimation errors calculated by $\tilde{i}_{\alpha\beta} = i_{\alpha\beta} - \hat{i}_{\alpha\beta}$, are multiplied by the observer gain G and used in Eqn(2.60) to correct the stator flux estimation.

The observer gain G is presented in Eqn (2.64).

$$G = g_1 I + g_2 J = \begin{bmatrix} g_1 & -g_2 \\ g_2 & g_1 \end{bmatrix} \quad (2.64)$$

where g_1 and g_2 are the gain parameters.

In [18], a Lyapunov stability analysis is performed for this observer and a rule is obtained for observer gain selection that would maintain stability as,

$$g_1 I + J(g_2 I - \omega L) > 0 \quad (2.65)$$

This rule provides a guide for the gain selection of this observer.

2.3.3 The gradient search based stator flux observer

In this section a non-linear observer is discussed. The observer is based on a gradient search[30]. It is implemented in the stator coordinates. The IPMSM model in stator coordinates can be written in terms of stator flux linkage is,

$$v_{\alpha\beta} = R i_{\alpha\beta} + \frac{d\Psi_{\alpha\beta}^s}{dt} \quad (2.66)$$

where,

$$\Psi_{\alpha\beta}^s = (L_0 I + L_1 Q(\theta)) i_{\alpha\beta} + \Psi_{\alpha\beta}^{pm} \quad (2.67)$$

This observer is first proposed for the SPMSMs where there is no or very little saliency[30]. Therefore, the observer is derived under $L_1 = 0$ condition. In this case, the middle term in Eqn (2.37) vanishes and an algebraic relation between the stator flux and the rotor flux is obtained as

$$\Psi_{\alpha\beta}^{pm} = \Psi_{\alpha\beta}^s - L_0 i_{\alpha\beta} \quad (2.68)$$

To construct the observer, first Euclidian norm of rotor flux is derived from Eqn(2.68) as,

$$\Psi_{\alpha\beta}^{pm} = \left\| \Psi_{\alpha\beta}^s - L_0 i_{\alpha\beta} \right\| \quad (2.69)$$

where $\Psi_{pm} \equiv \left\| \Psi_{\alpha\beta}^{pm} \right\|$, is the norm of the rotor flux. According to 2.39 an error between estimated and actual stator fluxes can be formed as,

$$\varepsilon_{gdm} = (\Psi_{pm}^2 - \|\hat{\Psi}_{\alpha\beta}^s - L_0 i_{\alpha\beta}\|^2)^2 \quad (2.70)$$

Then a gradient descent based observer, which tries to minimize ε_{gdm} , is constructed as,

$$\frac{d\hat{\Psi}_{\alpha\beta}^s}{dt} = v_{\alpha\beta} - Ri_{\alpha\beta} + \frac{g_1}{2} (\hat{\Psi}_{\alpha\beta}^s - L_0 i_{\alpha\beta}) (\Psi_{pm}^2 - \|\hat{\Psi}_{\alpha\beta}^s - L_0 i_{\alpha\beta}\|^2) \quad (2.71)$$

where g_1 is the observer gain. The division of observer gain by two, in Eqn(2.71) is redundant and it is only used to simplify the observer gain selection rule. According to the stability analysis performed in [30], it is determined that the only role of g_1 is to determine the convergence rate of the observer estimation. Therefore, it is possible to decrease g_1 as speed decreases without threatening stability. Selecting gain of this observer in speed dependent manner like this is also useful to avoid using unnecessarily large gains because increasing gain also increases the noise in the estimated position signal produced by this observer.

The observer in Eqn(2.71) is derived according to Eqn(2.38), which is only valid for motors with no saliency. This causes estimation errors when applied to an IPMSM with considerable saliency. To overcome this problem an extension to the IPMSM case is proposed in [20]. For the derivation of this extension, Eqn(2.71) should be represented in a slightly different format. First, Eqn(2.38) is rewritten in terms of the estimated stator flux,

$$\begin{bmatrix} \hat{\Psi}_\alpha \\ \hat{\Psi}_\beta \end{bmatrix} = \Psi_{pm} \begin{bmatrix} \cos \hat{\theta} \\ \sin \hat{\theta} \end{bmatrix} + L_0 i_{\alpha\beta} \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} \quad (2.72)$$

Similarly, Eqn(2.72) can be expressed as,

$$\begin{bmatrix} \hat{\Psi}_\alpha \\ \hat{\Psi}_\beta \end{bmatrix} = \Psi_{pm} \begin{bmatrix} \cos \hat{\theta} \\ \sin \hat{\theta} \end{bmatrix} + (L_0 I + L_1 \begin{bmatrix} \cos 2\hat{\theta} & \sin 2\hat{\theta} \\ \sin 2\hat{\theta} & -\cos 2\hat{\theta} \end{bmatrix}) \begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} \quad (2.73)$$

Using Eqn(2.73) in Eqn(2.71) results in,

$$\frac{d\hat{\lambda}_{\alpha\beta}}{dt} = v_{\alpha\beta} - Ri_{\alpha\beta} + \frac{\gamma}{2} (\hat{\lambda}_{\alpha\beta} - L_0 i_{\alpha\beta}) \lambda_{pm}^2 (1 - \cos \hat{\theta}^2 - \sin \hat{\theta}^2) \quad (2.74)$$

To extend this observer to the IPMSM's case, it is proposed to calculate estimated rotor position in Eqn(2.74), according to Eqn(2.73) instead of Eqn(2.72).

$$\hat{\theta} = \arctan \left(\frac{\hat{\lambda}_{\beta} - L_0 i_{\beta} - L_1 (i_{\alpha} \sin 2\hat{\theta} - i_{\beta} \cos 2\hat{\theta})}{\hat{\lambda}_{\alpha} - L_0 i_{\alpha} - L_1 (i_{\alpha} \cos 2\hat{\theta} + i_{\beta} \sin 2\hat{\theta})} \right) \quad (2.75)$$

It is claimed that this derivation, Eqn(2.45), results in a more accurate rotor position estimation as it takes into account the affect of saliency to some degree.

3. SIMULATION OF SENSORLESS SPEED AND POSITION ESTIMATION METHODS

The formulations of several speed and position observers for the IPMSMs were presented in the previous chapter. In this chapter, these observers are implemented into the Simulink simulation platform.

The simulation includes a model for the IPMSM, and models of its current and speed controllers along with models for each of the speed and position observers discussed in the previous chapter. The inputs to the observers are the voltages applied to the motor and the measured motor currents. Then, the position and speed are estimated from model of the observer as outputs. The actual motor speed and rotor position are also included in the outputs of the motor model. Through this construction, estimated and actual values of speed and position can be compared for each observer.

The performance of each observer is evaluated according to the accuracy of position and speed estimations under a range of operation scenarios. First of all, the steady state accuracy of the estimations are tested, when the motor is operated at different speeds and under different levels of load. In addition, the estimation accuracy of each observer is also evaluated during transient operating conditions. Observer behaviors during rapid changes in the speed and the load torque are studied as well.

Secondly, the effects of uncertainties in motor parameters on estimation performance are evaluated for each observer. These uncertainties arise due to temperature and current dependent nature of most motor parameters, which cause them to be time varying. For investigation of the effects of parameter uncertainties, differences between the values of motor parameters used in motor model and the corresponding parameters used in the observer model are introduced. The observer estimation accuracies are evaluated while the simulations are performed under these conditions.

Finally, the behaviors of the observers are compared for all the corresponding simulation scenarios, and the performance of each observer is evaluated according to these results.

In this chapter, the studies described above are presented starting from a detailed description of each block in the complete simulation model described in the first section. Then, the simulation results for Extended EMF, Active Flux, and Gradient Descent Method observers are presented respectively in the preceding three sections. In the last section performance of observers are evaluated by comparing these results.

3.1 Simulation Model Block Diagram

In Fig.3.1, the complete top-level block diagram of the simulation is presented. The diagram has a total of 3 closed feedback loops, one outer and two inner. In the outer loop, the speed of the motor is fed back to the speed controller block named PI_{sp} whereas, in the inner loops currents i_d and i_q are fed back to PI_{id} and PI_{iq} loops, respectively. The reference of quadrature (i_{qref}) is PI_{sp} controller's control signal and changing i_{qref} causes i_q itself to change. Since i_q is directly related to the torque of the motor, by changing i_{qref} PI_{sp} controller actually applies a torque as its control action to control the motor speed. Similarly, PI_{id} and PI_{sp} blocks apply v_d and v_q , respectively to control i_d and i_q . In addition to the controllers, the other main components of the simulation model are the IPMSM and the observer blocks. The IPMSM block is used to simulate the behavior of the motor, whereas the observer block contains the implementation of the sensorless method that is investigated.

The details of each block is explained in the following sections.

3.2 The IPMSM model

In this section, the motor model used in simulations is examined. All the simplifying assumptions made during the formulation of the sensorless methods in the previous chapter are also valid in this model (Section 2.1). This means that, the fundamental equations are used to construct the simulation model. All the secondary effects such as core losses, saturation of the core, and spatial harmonics of the permanent magnet flux linkage and spatial harmonics of stator inductances are omitted. The IPMSM block in Fig. 3.1 is a subsystem that contains the motor model. The inside diagram of this subsystem is shown in Fig. 3.2.

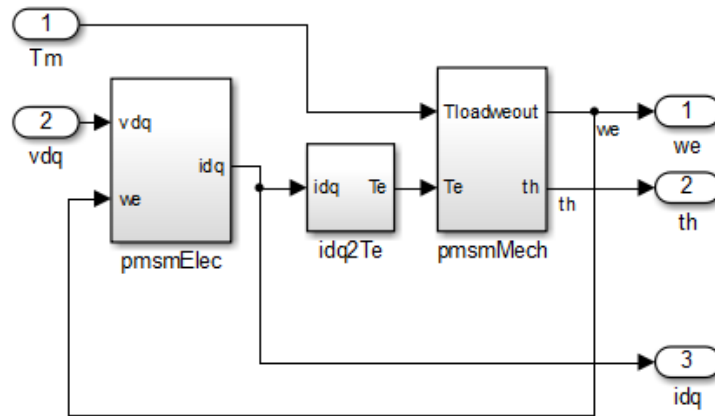


Figure 3.2 : IPMSM block diagram.

The motor model consists of two subsystems where the mechanical and the electrical parts are named as **pmsmElec** and **pmsmMech**, respectively. The details of these blocks are discussed in the next section.

There are two connections between electrical and mechanical parts. The first connection is the stator currents, which are responsible for creating the electrical torque of the motor. This relation is implemented in **idq2Te** block, and its block diagram is shown in Fig. 3.3.

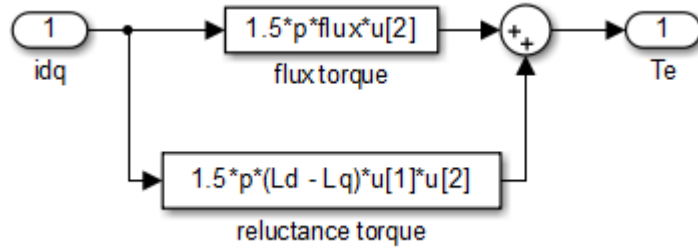


Figure 3.3 : Electromagnetic torque generation.

In IPMSMs with low saliency such as the one used in this study, i_q creates the larger part of the motor torque. On the other hand, a non-zero i_d , whose sign is opposite to that of i_q , is needed in order to be able to produce some amount of reluctance torque. In this motor model, both magnetic reaction and reluctance components are the motor torque components and are modeled in pmsmMech. This is for future use as $i_d=0$ is employed in simulation, which results in no reluctance torque.

The second connection is the electrical angular speed, which is feedback from the mechanical part to the electrical part.

The IPMSM block has three input signals and three output signals. The inputs are quadrature axis voltage, direct axis voltage and mechanical load torque, which are respectively denoted by v_d , v_q , and T_{load} in the simulations. To decrease the number of signal wires in all simulation blocks, the signals are multiplexed whenever appropriate. An example to this is the v_{dq} input in Fig3.2 that contains v_d and v_q signals in multiplexed form. The order of the signals also correspond to the order of their corresponding letters in the input name. As shown in Fig 3.4, v_d and v_q are the first and second signals in v_{dq} , respectively. This naming convention is applied similarly throughout all the block diagrams in the simulations.

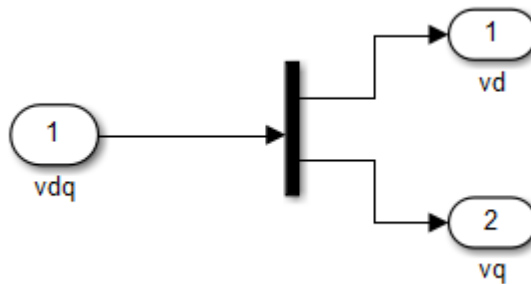


Figure 3.4 : Naming convention in multiplexing.

The IPMSM block has four outputs. These are the direct- and quadrature-axis currents, angular speed and angular position named as i_d , i_q , w_e , and θ respectively. It should be noted that θ and w_e refer to the electrical quantities.

3.2.1 The pmsmElec

In this section, the subsystem used to model the electrical part of the IPMSM is examined. This subsystem is called the pmsmElec. Its inside diagram is shown in Fig.3.5.

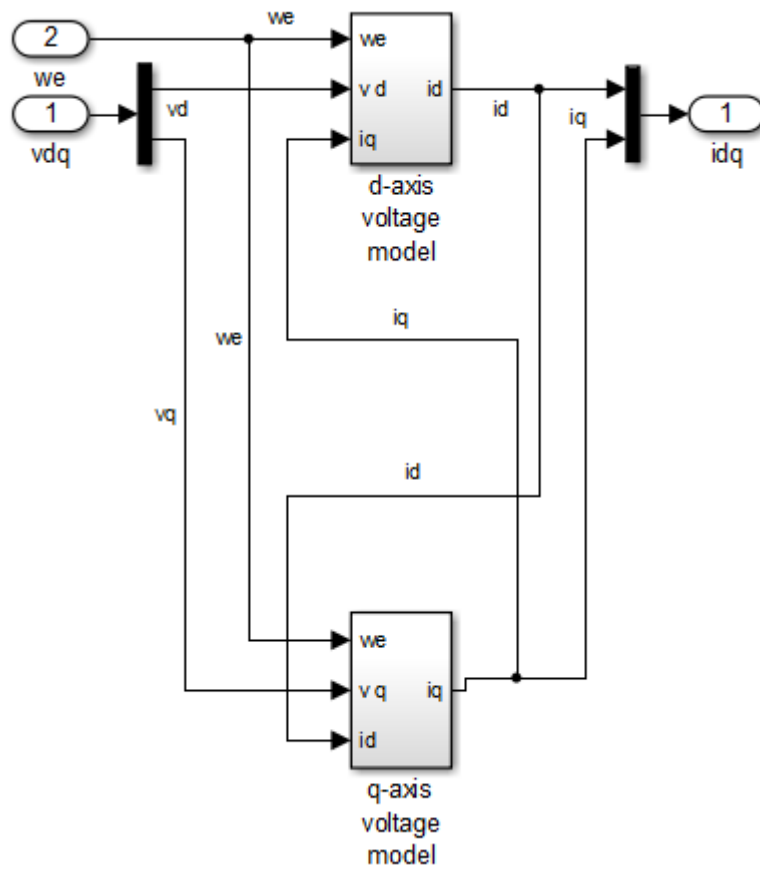


Figure 3.5 : Block diagram of the pmsmElec.

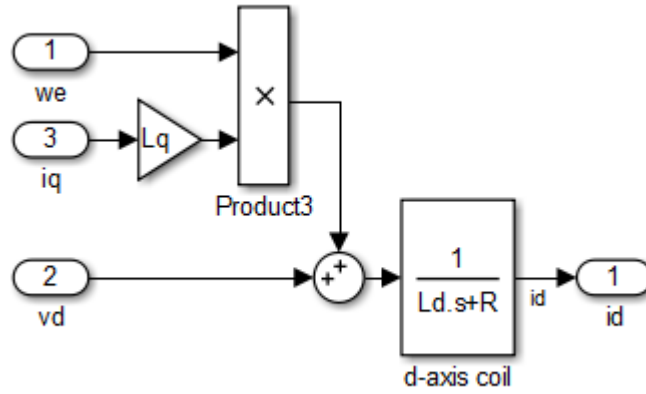


Figure 3.6 : D-axis voltage model.

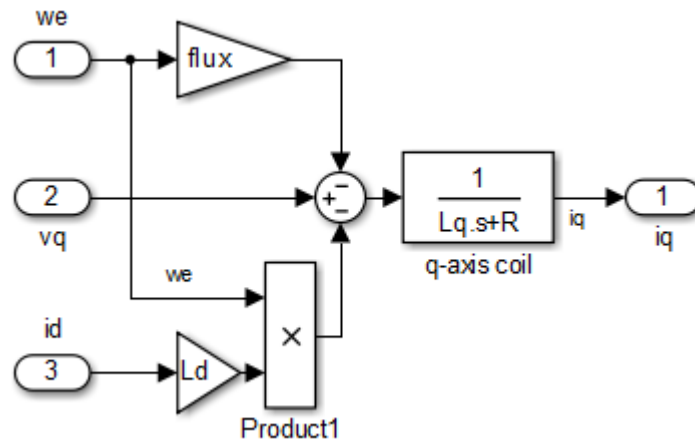


Figure 3.7 : Q-axis voltage model.

As presented in Fig.3.5, the direct and quadrature axis equations are coupled to each other. The coupling terms are $i_q L_q \omega_e$ and $i_d L_d \omega_e$. The coupling becomes stronger as speed and load increases, which correspond to an increase in ω_e and i_q , respectively. When $i_d=0$ is selected, the quadrature axis equation becomes uncoupled, as a result controlling i_q becomes easier. However, in this condition no reluctance torque is produced which has a negative effect on motor efficiency.

The d- and q- axis coils are represented as first order systems in Fig.3.6. They act as low-pass filters between voltages and currents. This in turn allows filtering of the high frequency harmonics in phase voltages that arise due to PWM switching of the inverter. It is easier to determine the time constants of the equivalent d- and q- axis coils, and use them to determine appropriate gains for current controllers by the use of this representation.

3.2.2 The pmsmMech

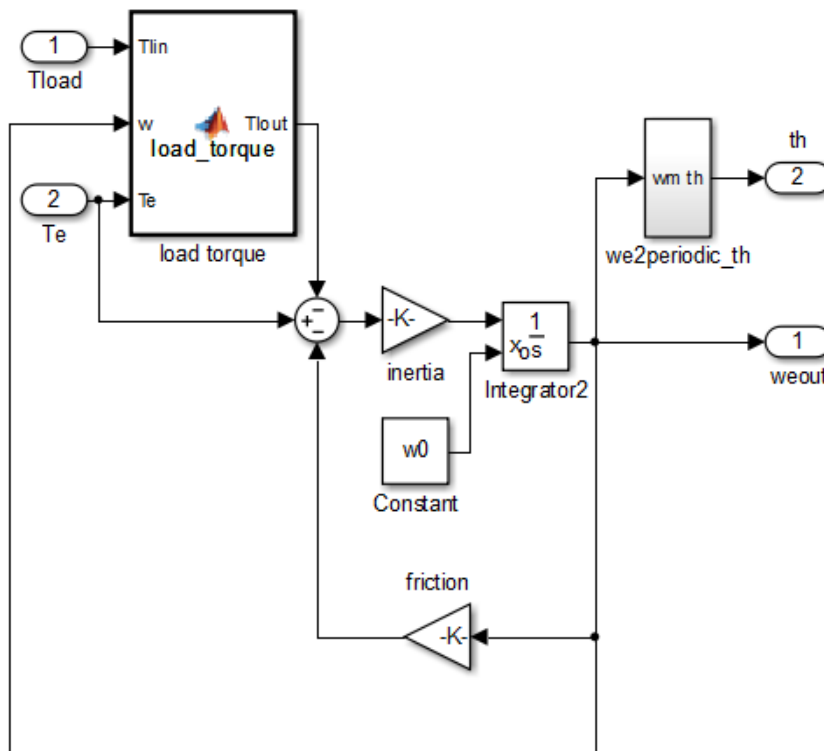


Figure 3.8 : Model of the mechanical part of the IPMSM.

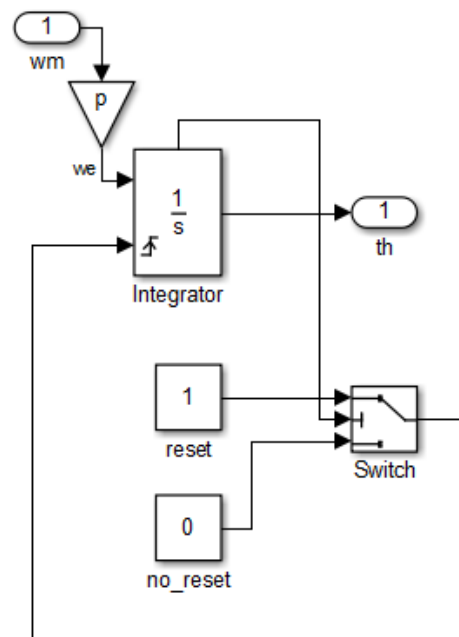


Figure 3.9 : Periodic position generator.

3.2.3 Control loops

There are three PI controllers in the simulation. These are PI_{id} , PI_{iq} and P_{isp} shown in Fig3.9 PI_{id} and PI_{iq} control i_d and i_q respectively. PI_{sp} controls the motor speed. In Figure 3.9 block diagram of PI_{id} is shown.

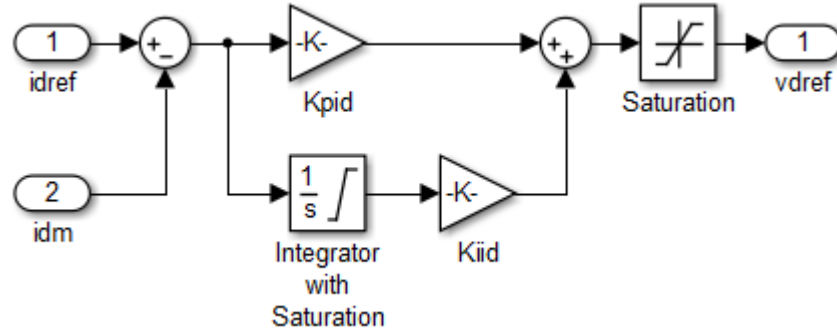


Figure 3.10 : PI_{id} .

PI_{iq} and PI_{sp} are implemented using the same block diagram as PI_{id} . The only change are the values of the parameters.

3.2.4 Observers

The observers are implemented as Embedded MATLAB function blocks in Simulink. The observer blocks use discrete time, and their time step is equal to the current sampling frequency (also the PWM frequency). This time step is selected to be used in the simulations, as it is the same observer sampling time used in the real inverter software. This way, the time between two consecutive estimation updates is same for both the real and simulated observers. In these block, the equations of observers, which are formulated in the previous chapter, were implemented. Same calculations were performed in the driver software implementations of the observers. This way the implementation of the observer equations had been first validated through simulations before they were used in experiments.

3.3 Results of The Simulations

The following parameters are used in all the simulations in this chapter:

Table 3.1 : All the parameters used in simulations.

Gain	Value
R	3.5 Ω
L_d	13.4 mH
L_d	18.6 mH
Ψ_{pm}	0.105 Vs/rad
$K_{p_{speed}}$	0.01
$K_{i_{speed}}$	1
$K_{p_{id}}$	5
$K_{i_{id}}$	200
$K_{p_{iq}}$	5
$K_{i_{iq}}$	200
EEMF/ g_1	100
EEMF/ g_2	25266
EEMF/ g_3	30000
EEMF/ g_4	0.005
GDM/ g_1	500/1000
AF/ g_1	10
AF/ g_2	30

3.3.1 The EEMF method

For simulations, a set of gains has to be selected for stable and accurate operation of the EEMF estimator, and the ATO that is needed to be used with it to provide the speed adaptation scheme needed.

The following gains were selected for the EEMF method.

Table 3.2 : The gains used in simulations with the EEMF method.

Gain	Value
g_1	100
g_2	25600
g_3	5000
g_4	0.04

The simulations were performed at 200 rpm and 2000 rpm and the results are shown in Fig.3.10 and Fig.3.11. The simulations start at no-load condition, then at $t=1\text{sec}$ the load is changed to 0.75Nm, and at $t=2\text{sec}$ load is changed to 1.5Nm.

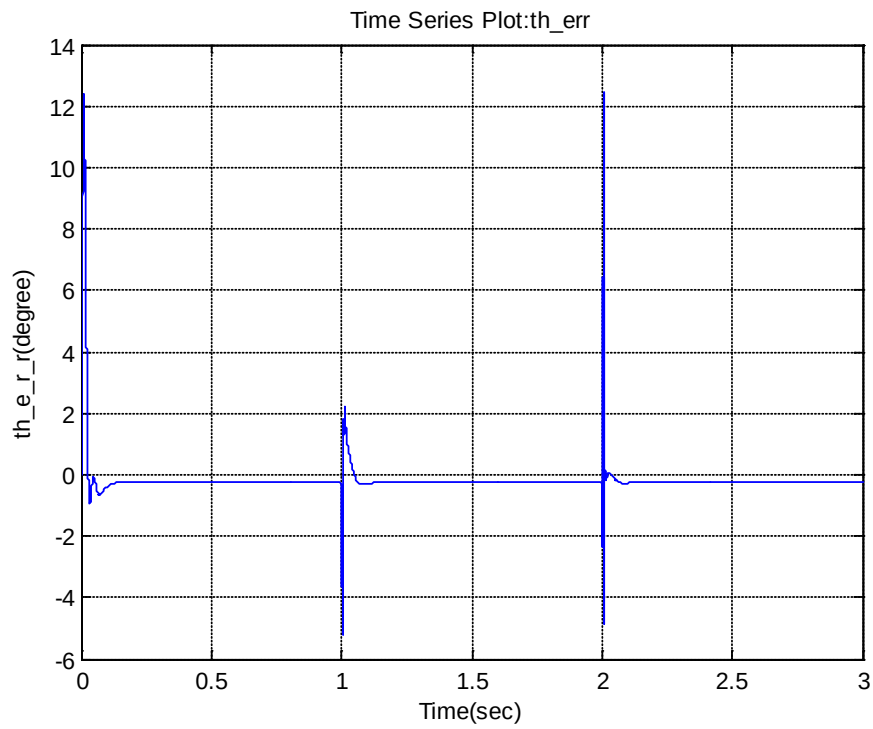


Figure 3.11 : EEMF position estimation error at 200 rpm.

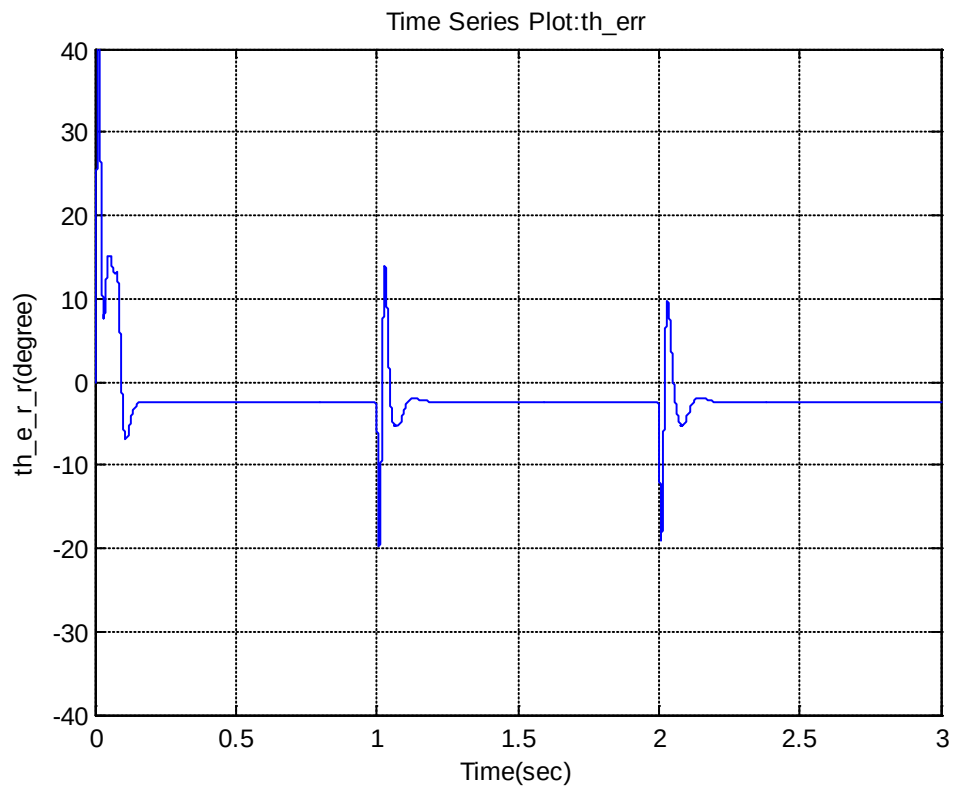


Figure 3.12 : EEMF position estimation error at 2000 rpm.

The results of these simulations showed that the structure formulated in EEMF method and the gains chosen for it were capable of providing stable and accurate position estimations under the operating conditions, for which the simulation results in Fig.3.12 and Fig.3.13 were obtained. According to these simulations the estimation error is almost zero at 200rpm at all loads. The error increases to approximately 3° at 2000rpm. The results also showed the increase in position estimation error during the load transients. At these instances, rapid changes in motor speed were observed. Thus, the behavior of the EEMF method in these times was in accordance with the speed adaptive nature of this method.

In the second part of simulations, robustness of this method under variations of motor parameters were simulated. The simulations were carried out at 200rpm and the estimated value (the value used in the EEMF calculations) of the parameter under study is changed from 70 to 130% of its actual value. The results are presented in Fig.3.13, Fig.3.14 and Fig.3.15.

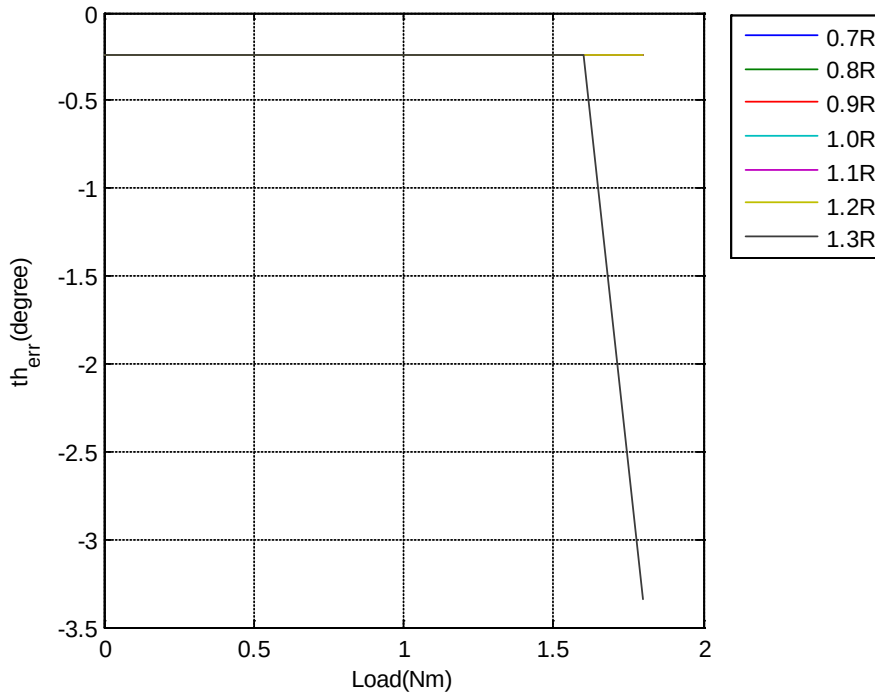


Figure 3.13 : EEMF position error under R uncertainty.

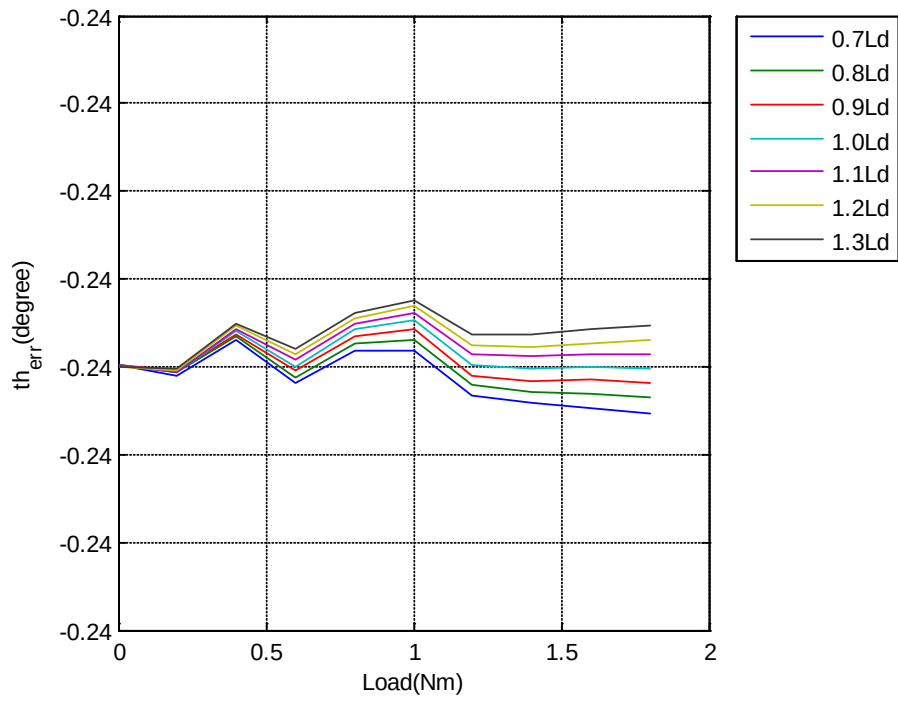


Figure 3.14 : EEMF position error under L_d uncertainty.

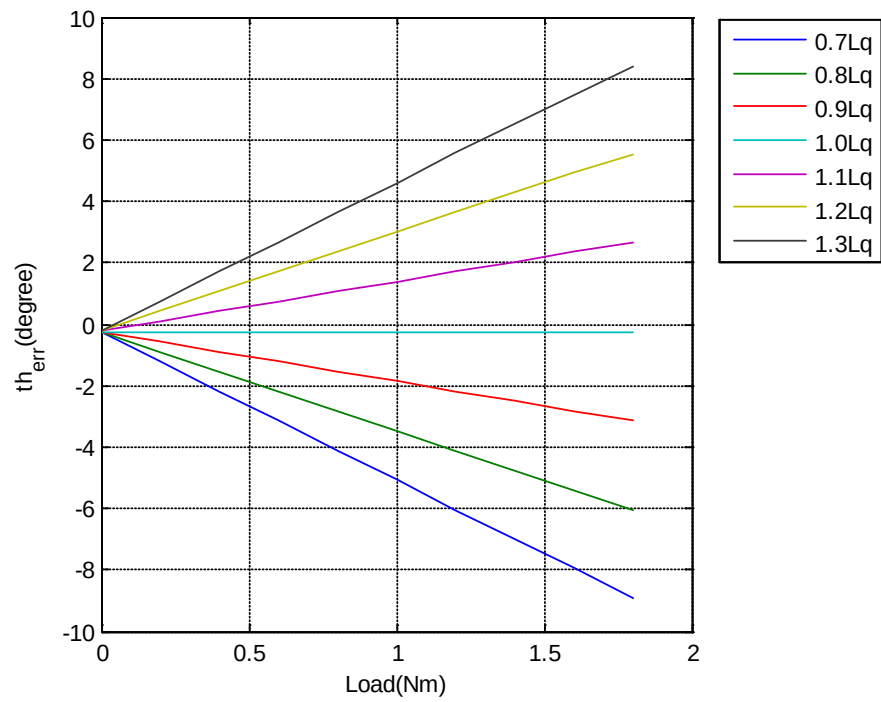


Figure 3.15 : EEMF position error under L_q uncertainty.

According to the simulation, the EEMF method was not affected by changes in resistance, except for the single case where estimated resistance is set to 130% of its actual value while the load was at 1.8Nm. On the other hand variation in L_q of the motor had a very direct impact on the estimation accuracy of EEMF. The effect of variations in rotor flux was not simulated since this parameter is not used in EEMF observer block of the simulation model. EEMF is found to be insensitive to variation in L_d , as well.

3.3.2 The AF method

For the simulations, a set of gains had to be selected for stable and accurate operation of the AF estimator. The following gains were selected for the AF method.

Table 3.3 : The gains used in simulations with the AF method.

Gain	Value
g_1	10
g_2	30

The simulations were performed at 200 rpm and 2000 rpm and the results are presented in Fig.3.15 and Fig.3.16. The simulations started at no-load condition, then at $t=1\text{sec}$ the load is changed to 0.75Nm, and at $t=2\text{sec}$ load is changed to 1.5Nm.

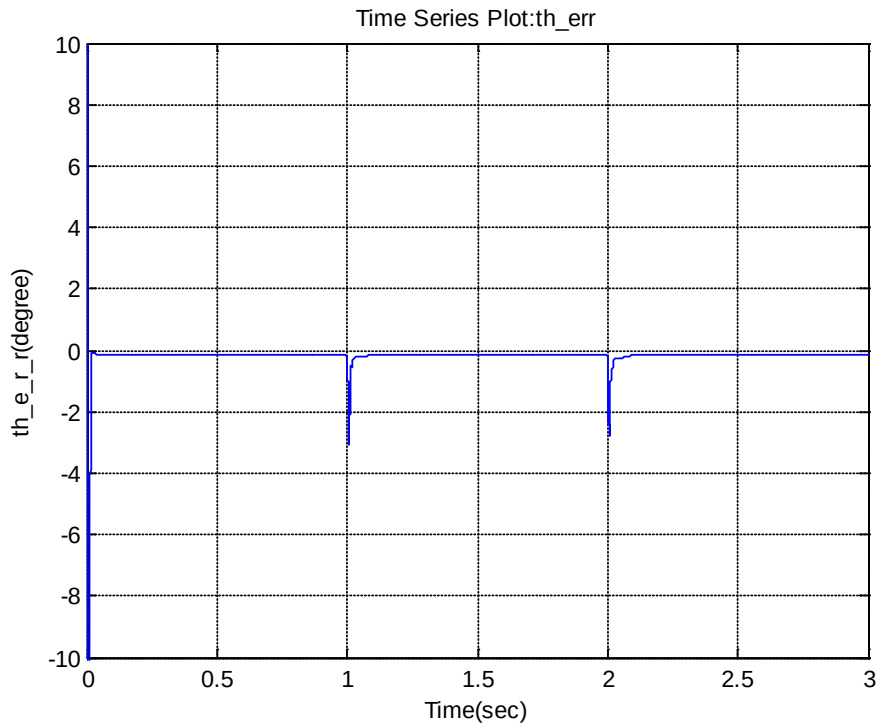


Figure 3.16 : AF position estimation error at 200 rpm.

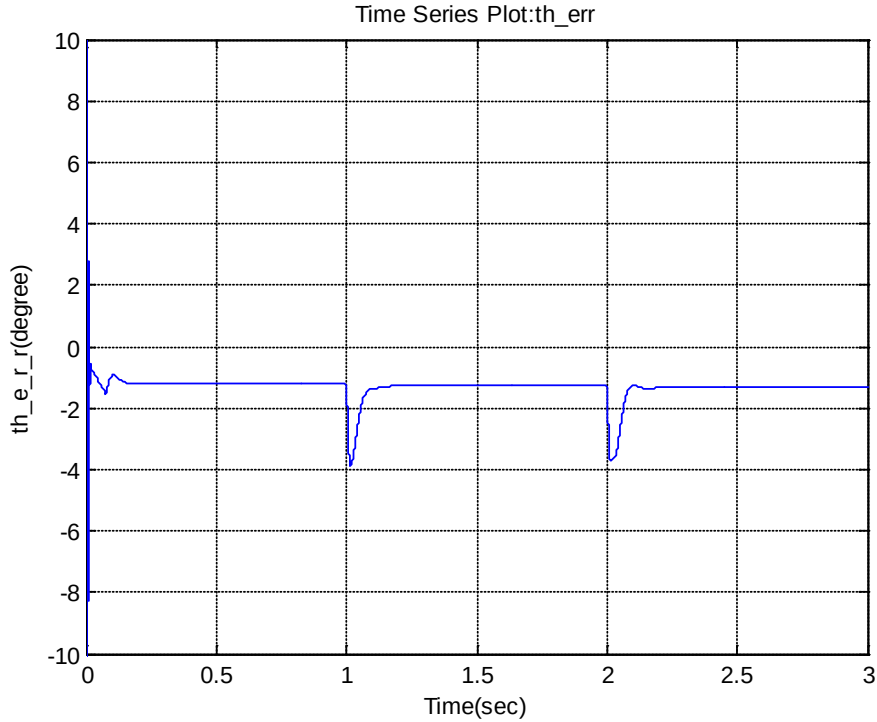


Figure 3.17 : AF position estimation error at 2000 rpm.

The results of these simulations showed that the structure formulated in the AF method and the gains chosen for it were capable of providing stable and accurate position estimations under the operating conditions, for which the simulation results in Fig.3.16 and Fig.3.17 were obtained. According to these simulations, the estimation error was very close to zero both at 200rpm and 2000rpm at all loads. The error was very slightly higher (1°) at 2000 rpm. Moreover, increasing the load had an affect towards increasing the position estimation error. The results also showed that the accuracy of position estimation was not affected significantly during the load transients, and the estimator was very fast in converging the estimation error to its steady state value.

In the second part of simulations, robustness of this method under variations of motor parameters were simulated. The simulations were carried out at 200rpm and the estimated value (the value used in the AF calculations) of the parameter under study was changed from 70% to 130% of its actual value. The results are shown in Fig.3.18, Fig.3.19, Fig.3.20 and Fig.3.21.

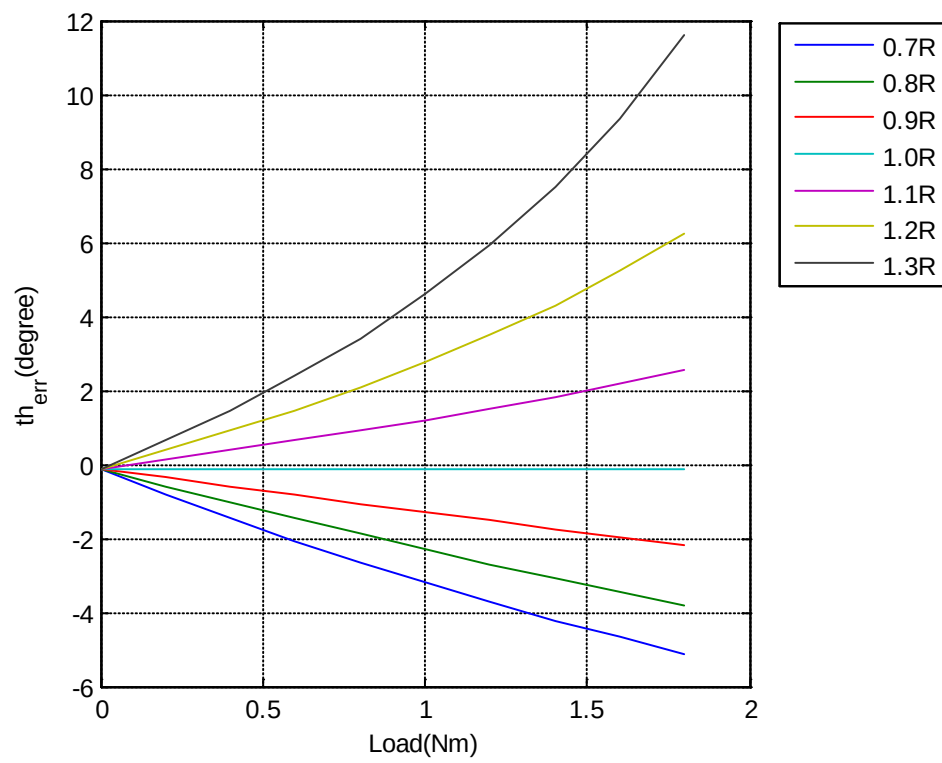


Figure 3.18 : AF position error under R uncertainty.

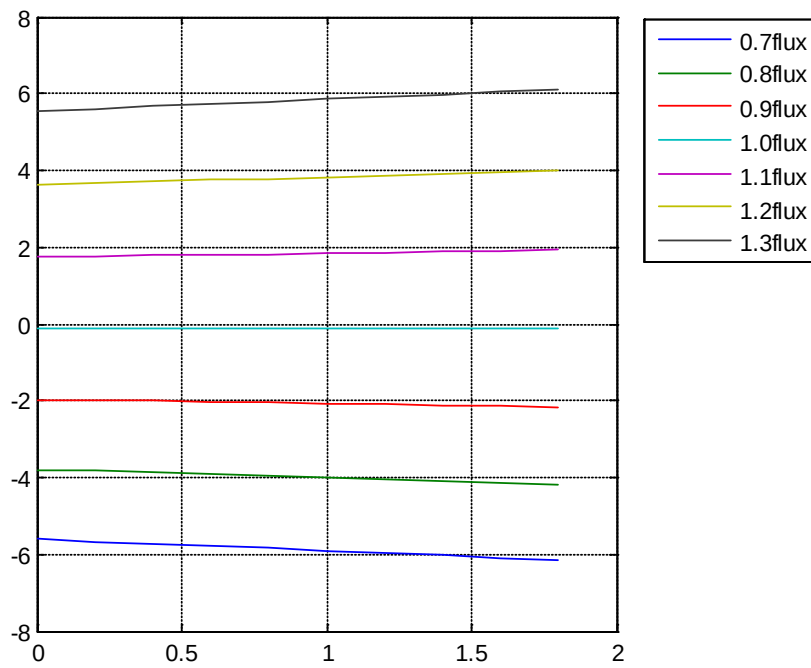


Figure 3.19 : AF position error under flux uncertainty.

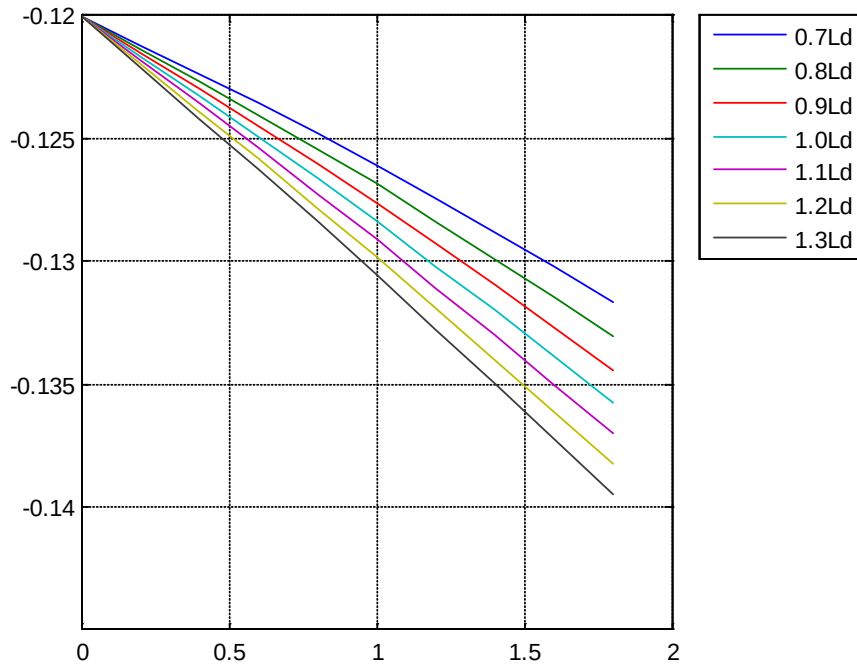


Figure 3.20 : AF position error under L_d uncertainty.

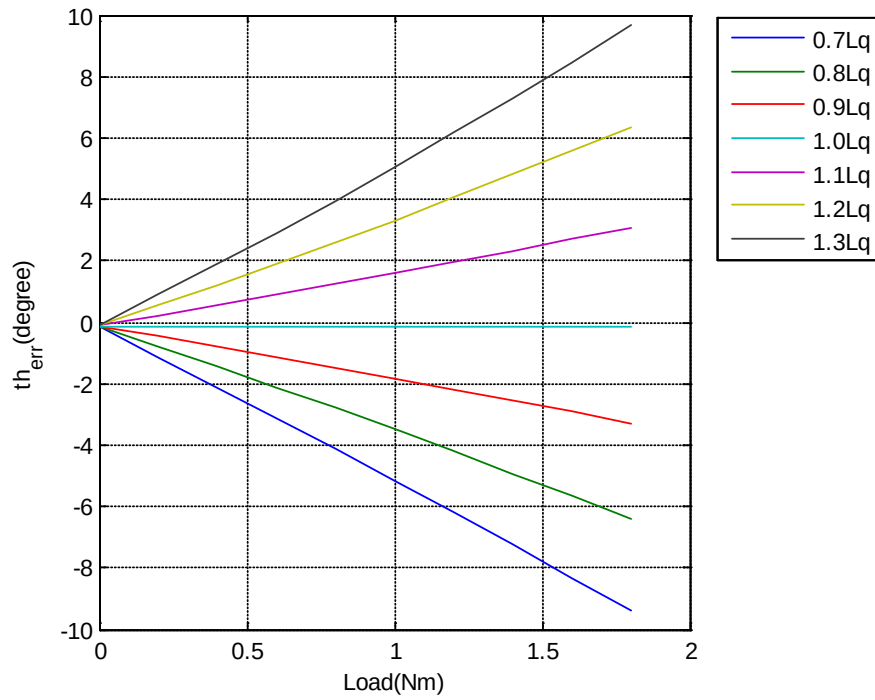


Figure 3.21 : AF position error under L_q uncertainty.

According to the results of simulations AF method was not affected by variations in L_d . All the other three parameters had significant effects on estimation errors. Variations in L_q and R had effects on the position error of this observer, which

increase linearly by the increasing load. This was expected, as there is a linear relationship between the current and load and both of the parameters L_q and R appear in equations of AF method in terms, where they are multiplied by the phase currents. Variation of flux on the other hand had a very significant effect on the estimation error of AF in spite of the amount of load present.

3.3.3 GDM method

For simulations, a set of gains had to be selected for stable and accurate operation of the GDM estimator. The following gain was selected for the GDM method.

Table 3.4 : The gain used in simulations with the GDM method.

Gain	Value
g_1	500 (at 200rpm) 1000(at 2000rpm)

The simulations are performed at 200 rpm and 2000 rpm and the results are shown in Fig.3.22 and Fig.3.23. The simulations started at no-load condition, then at $t=1\text{sec}$ the load is changed to 0.75Nm , and at $t=2\text{sec}$ load is changed to 1.5Nm .

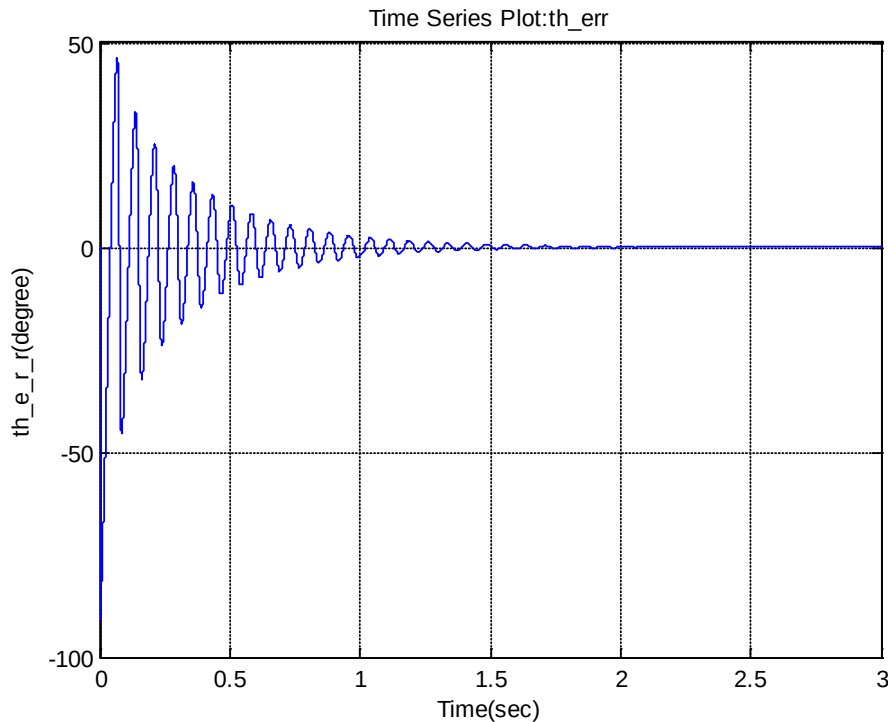


Figure 3.22 : GDM position estimation error at 200 rpm.

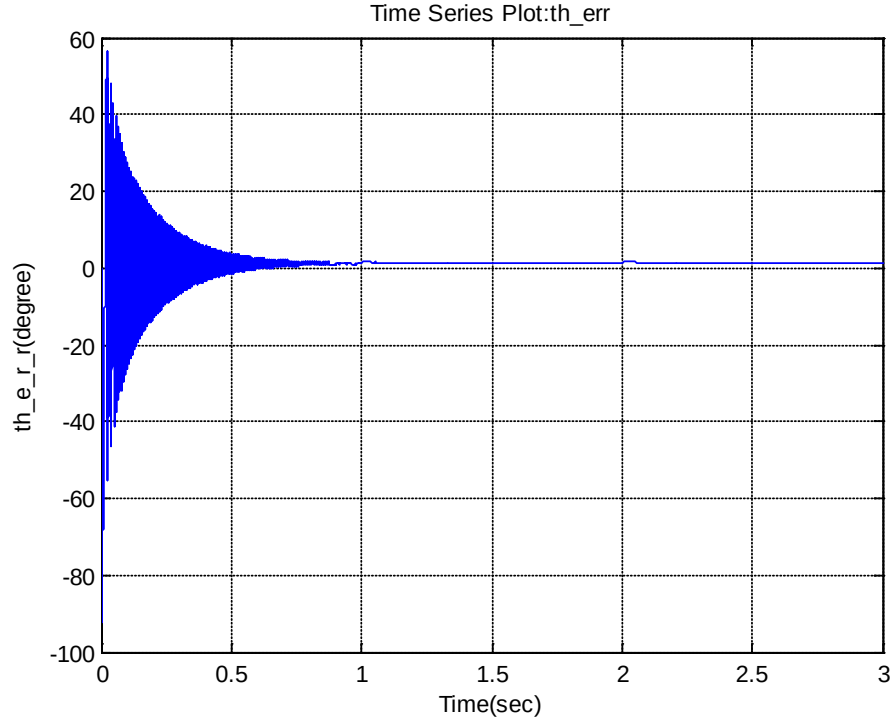


Figure 3.23 : GDM position estimation error at 2000 rpm.

The simulations of GDM results shown in Fig.3.22 and Fig.3.23, show that the convergence rate of GDM is slow than the other two methods. However it is observed that the convergence speed was nearly twice in the case of 2000 rpm than that of 200 rpm. In 200 rpm simulation g_1 is selected as 500 while in 2000 rpm simulation it is selected as 1000. Therefore the increase in the convergence speed is in accordance with the theoretical claim in which it is stated that the convergence rate of GDM is proportional to the value selected for g_1 .

In the second part of simulations, robustness of this method under variations of motor parameters was simulated. The simulations were carried out at 200rpm and the estimated value (the value used in AF calculations) of the parameter under study was changed from 70 to 130% of its actual value. The results are shown in Fig.3.24, Fig.3.25, Fig.3.26 and Fig.3.27.

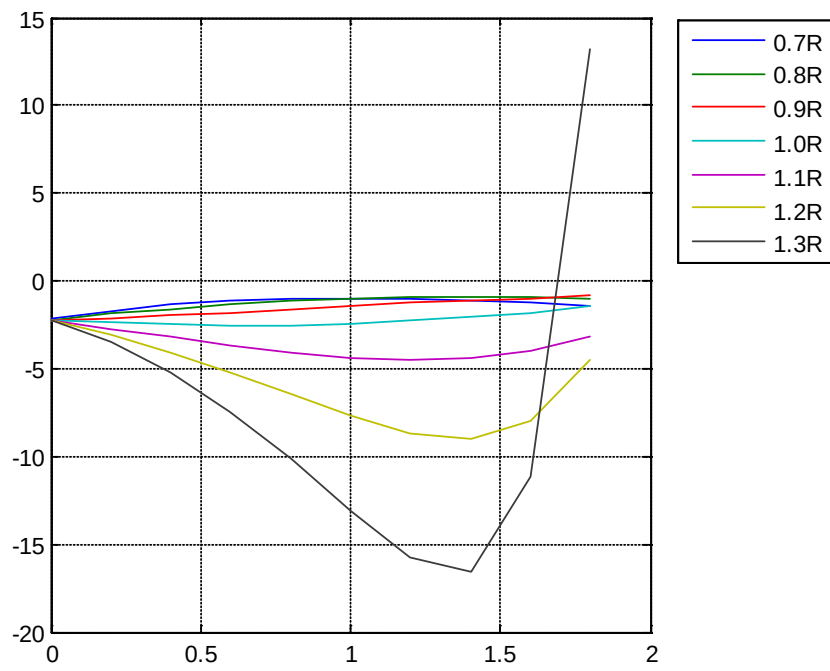


Figure 3.24 : GDM position error under R uncertainty.

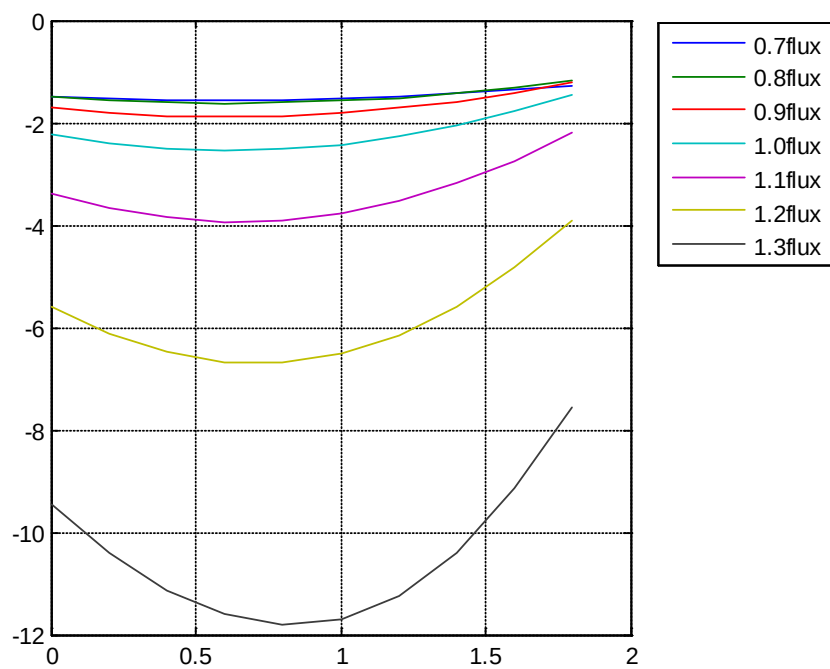


Figure 3.25 : GDM position error under flux uncertainty.

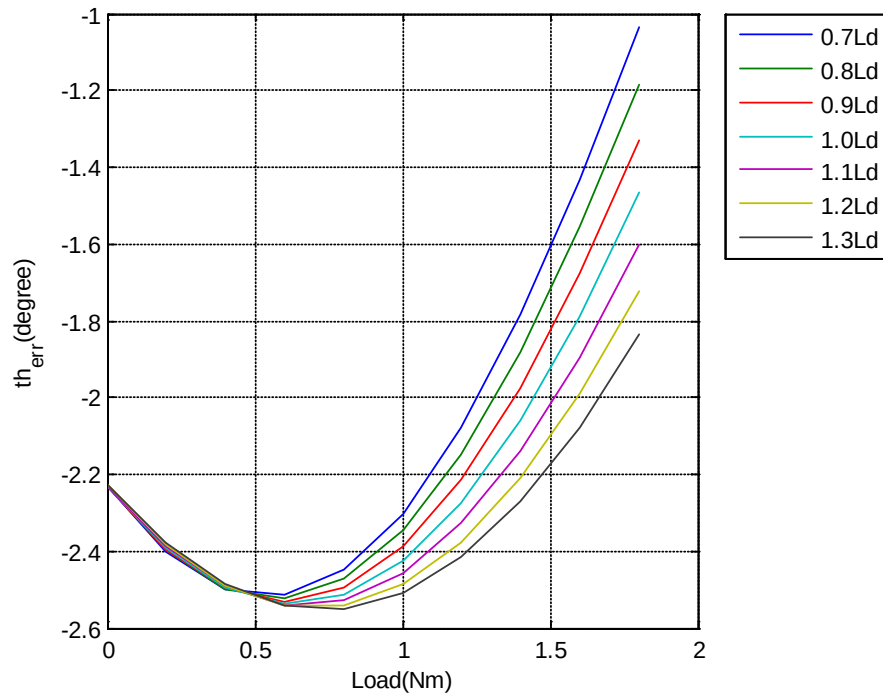


Figure 3.26 : GDM position error under L_d uncertainty.

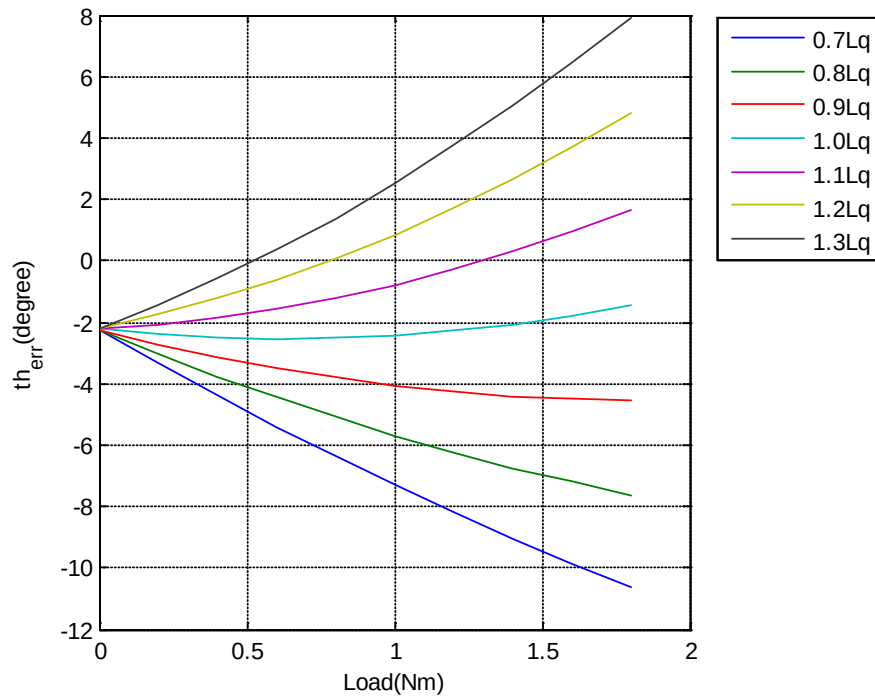


Figure 3.27 : GDM position error under L_q uncertainty.

According to the results of the simulations, the GDM method under parameter uncertainty conditions was least affected by the variations in L_d among all the motor parameters like the other method. Same with the other two methods, a strong

sensitivity to variations in L_q was determined as presented in Fig. 3.27. This method also expressed sensitivity to changes in the R and flux parameters, however, these sensitivities are observed to be smaller when the estimated value of the parameter is less than its actual value, differing from the case of the AF method.

4. EXPERIMENTS WITH SENSORLESS METHODS

In this chapter, the sensorless methods that had been analyzed and simulated in the previous chapters were tested in a real application. For this purpose, a drive test system was constructed.

The main parts of this system are an IPMSM motor, an inverter, a controlled brake, and PC monitoring softwares. First a brief description of these parts is presented. With the help of the drive test system, performances of sensorless methods under various operating conditions were tested via a series of experiments. Finally, through the results of these experiments the performances of sensorless methods were compared to each other.

4.1 The Experimental Setup

The complete test setup used for the experiments is presented in Fig4.1.

Table 4.1 : Specifications of the motor used in drive test system.

Specifications	Values
Power	900W
Nominal Load	2Nm
Nominal Speed	4000rpm
R	2.48 Ω
L _d	14mH
L _q	18mH
Ψ_{pm}	0.105Vs.rad ⁻¹



Figure 4.1 : The complete test drive system.

The specifications of the motor used in drive test system are presented in Table4.1

This motor is controlled with a Renesas microcontroller based motor control demo driver named MCRP07. MCRP07 is composed of two electronic boards (Fig.4.2).

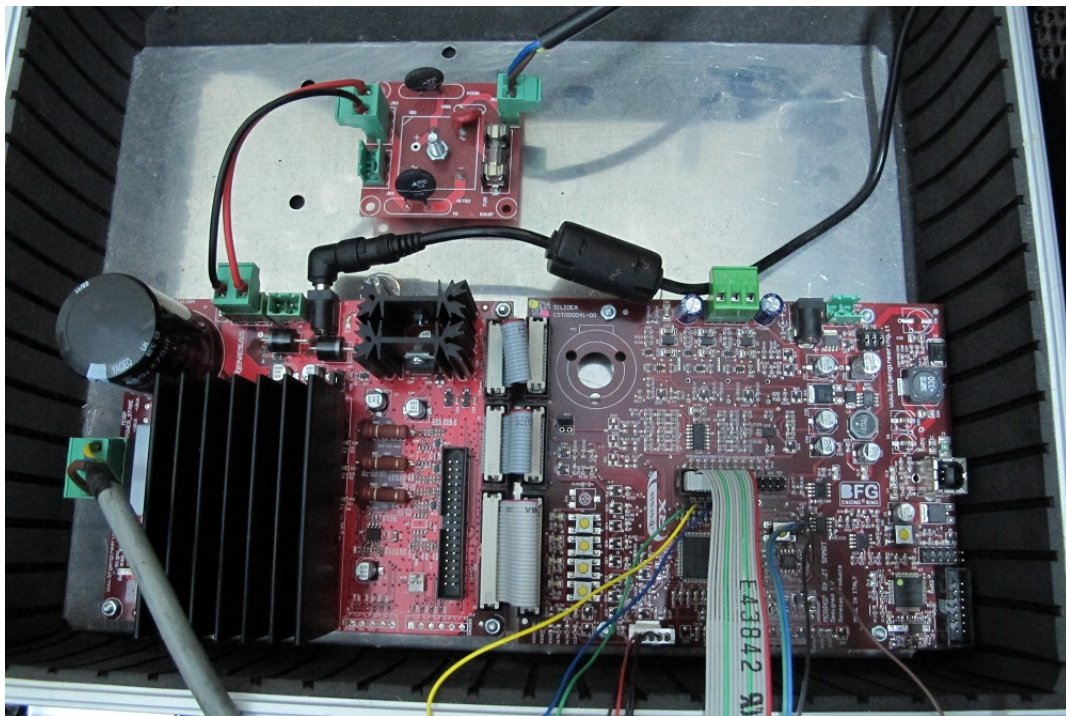


Figure 4.2 : The MCRP07 motor control kit.

The first is the controller board that is used to generate gate signals for the power switches. The second board is the power module. An intelligent power module is used in the power board, which is suitable to be used in driver applications up to 1kW. The gate signals from the control module are routed to the IPM in the power module and the output signals of IPM are connected to the IPMSM.

At the heart of the controller is a RX62T microcontroller, on which the driver software is executed. This is a microcontroller especially designed to be used in motor control applications. A summary of the peripherals available in RX62T is given in Fig.4.3.

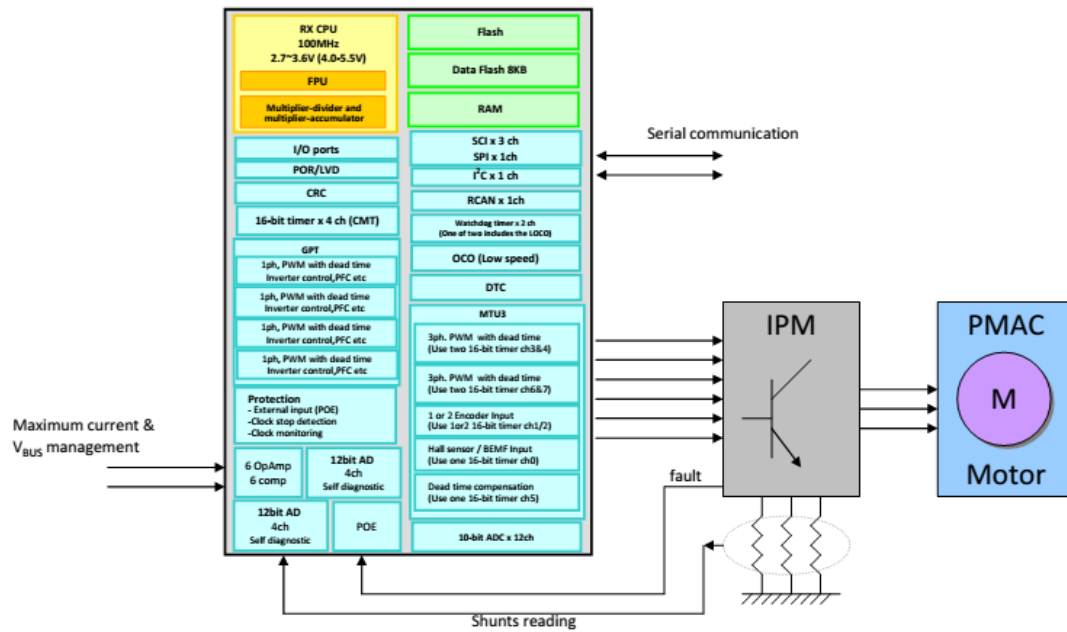


Figure 4.3 : Key features of the RX62T microcontroller [32].

The RX62T is connected to a PC that is used for monitoring and recording the values of variables in the driver software. The software that is responsible for these functions is developed in LabVIEW platform. Up to 10 selected variables can be monitored and logged at a rate of 200S/s with the help of this software.

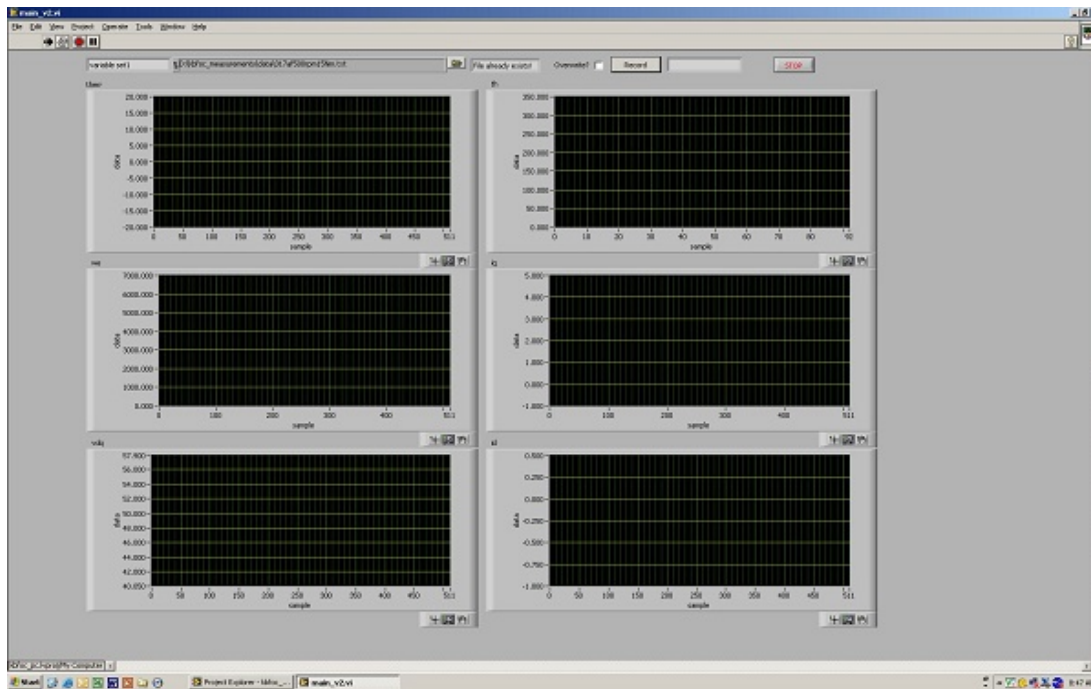


Figure 4.4 : The GUI of monitoring and data logging software.

The motor loaded with a hysteresis brake that is capable of producing torques up to 25Nm (Fig.4.5).

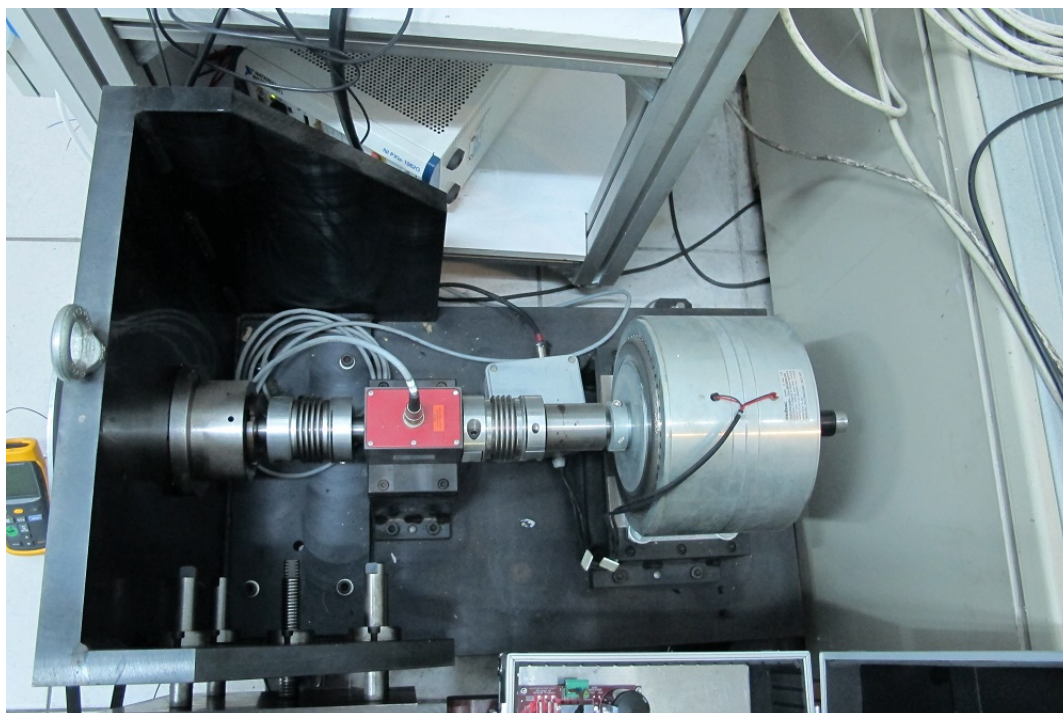


Figure 4.5 : The hysteresis brake.

A pulley with a diameter of 300mm is attached to the shaft of the hysteresis brake. The pulley of the hysteresis brake is connected to the pulley of the motor, whose

diameter is 32mm, through a belt (See Fig.4.6). As a result, the transmission ratio between the motion of the motor shaft and the shaft of the hysteresis brake is approximately 9.38. This means that loads up to 2.5Nm can be applied to the motor with this setup.



Figure 4.6 : The pulley connected to the hysteresis brake shaft.

The braking torque of the hysteresis brake is proportional to the magnitude of DC current applied to its coils. A DC current source is connected to the coils of the hysteresis brake and the output current supplied by this source is controlled with a 0-5V analog signal that is generated through the analog output port of an NI MIO card, which is connected to a PXI system.

In addition, there is a torque and speed sensor mounted on the shaft of the hysteresis brake (Fig 4.7).

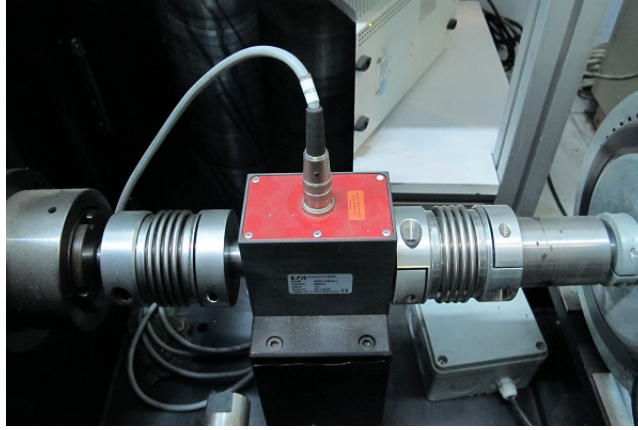


Figure 4.7 : The torque and speed sensor on the hysteresis brake shaft.

The values of the measured speed and torque of the brake shaft are sent to the PXI system as well. A control algorithm is implemented as a LabVIEW software that is running in the PXI system. The load applied to the motor can be controlled between 0-2.5Nm with the help of this software (Fig.4.8).

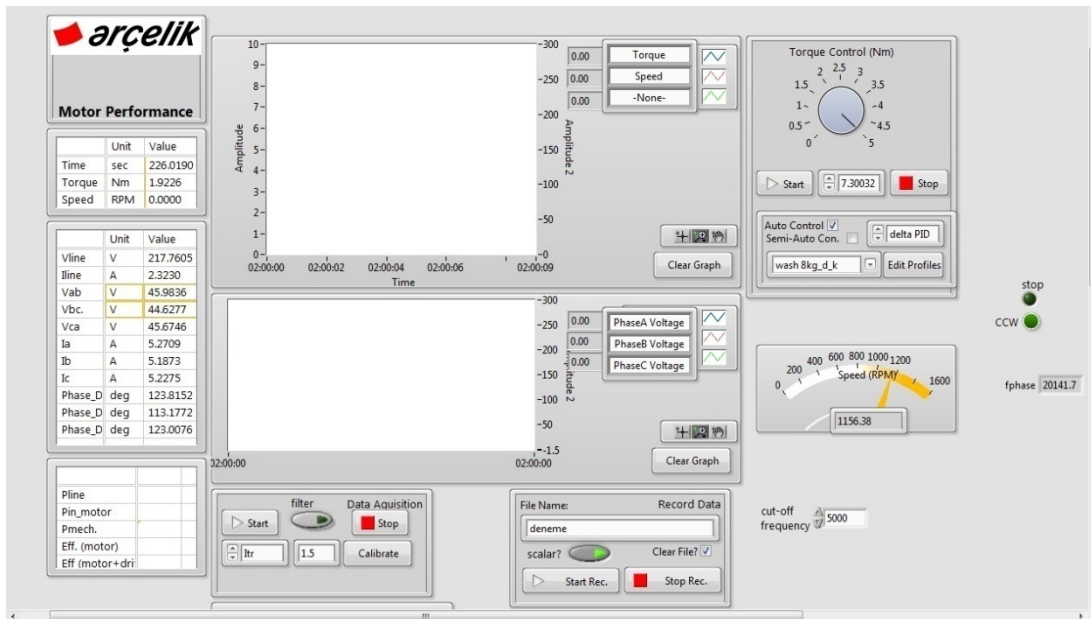


Figure 4.8 : The GUI of automatic load control software.

4.2 Methods Used in The Experiments

During the tests, driver software transmits the values of the selected variables to the PC monitoring software. The serial communication speed was 234000 bps. Up to ten variables could be sent simultaneously with a sampling rate of 200 S/s with this communication. The variables that were transmitted are the floating numbers. These

were first scaled up to an integer value. Scale gains were chosen separately for each variable according to the significant figures that are needed to be preserved.

The PC monitoring software was developed in the LABVIEW platform. This software converts the data it receives in integer format back to the original floating number format, and then monitors all the signals, which are logged for 2.5seconds.

A complete field oriented control (FOC) software library was developed as a part of this study. The three sensorless methods whose formulation were given in Chapter2 was implemented as estimation functions in the driver software. The software has the capability of switching between the estimation methods while the motor is operating. The driver software also provides options between complete sensorless control and control in which position information is measured by a relative encoder that is mounted on the motor shaft.

The software of the driver was developed in Renesas HEW software development environment. The experiments were performed in a setting where the RX62T was run in debug mode via the HEW software. This way the values of the selected variables were monitored in HEW software and manual change of operation modes was possible through this debugging environment.

Through serial communication between the driver software and the software of the monitoring, PC 10 variables were logged in real time with a sampling rate of 200S/s.

For all the experiments in this chapter parameters given in Table 3.1 are used unless otherwise is stated, with the only exception being L_q whose for which 18.6mH is used in simulations whereas 22mH is used in experiments. This change is preferred because it is detected to decrease load dependent changes in position estimation errors of EEMF and AF, during preliminary experiments.

4.3 Effects of Gain Selections

The selection of observer gains plays a crucial role on estimation stability and accuracy. The stabilizing gain selection rules have been detailed in Chapter2 for each observer. Different set of gains can be selected for each observer following these rules, each of which would result in a stable estimation error. However, even if stability is satisfied, different gain selections can result in different steady state estimation errors at different speeds and loads. Therefore, it is preferred to determine

a set of gains that result in an estimation error which remains as constant as possible in the case of changing operating conditions.

The experiments presented in this section were carried out with the aim of finding such a set of gains for each of the three observers under investigation.

4.3.1 The EEMF method

The estimation method named EEMF has four gains in its formulation, namely g_1 , g_2 , g_3 , and g_4 . First two of the gains g_1 and g_2 are in compensator block in the EEMF estimator, which has the structure of a PI controller.

During the initial measurements, it was noticed that the effects of g_1 and g_2 on position estimation error of EEMF were more observable, when the motor was operated under lower speed and lower load conditions. Thus, first experiments were performed at 200rpm and 0.5Nm with motor controlled in a sensed manner. Three sets of g_1 and g_2 values were tested. The results, presented in Figure 4.9, showed that, selecting too small values for g_1 and g_2 resulted in an oscillating position estimation error. This result was expected since the bandwidth of the PI compensator block decreases as g_1 and g_2 are decreased. Selecting 100 and 25266 for g_1 and g_2 respective was determined to yield an acceptably low oscillation level. Additionally, it was observed that the choice of g_1 and g_2 did not have a significant effect on the average position estimation error which is very close to zero for all three sets of gains (Figure 4.9).

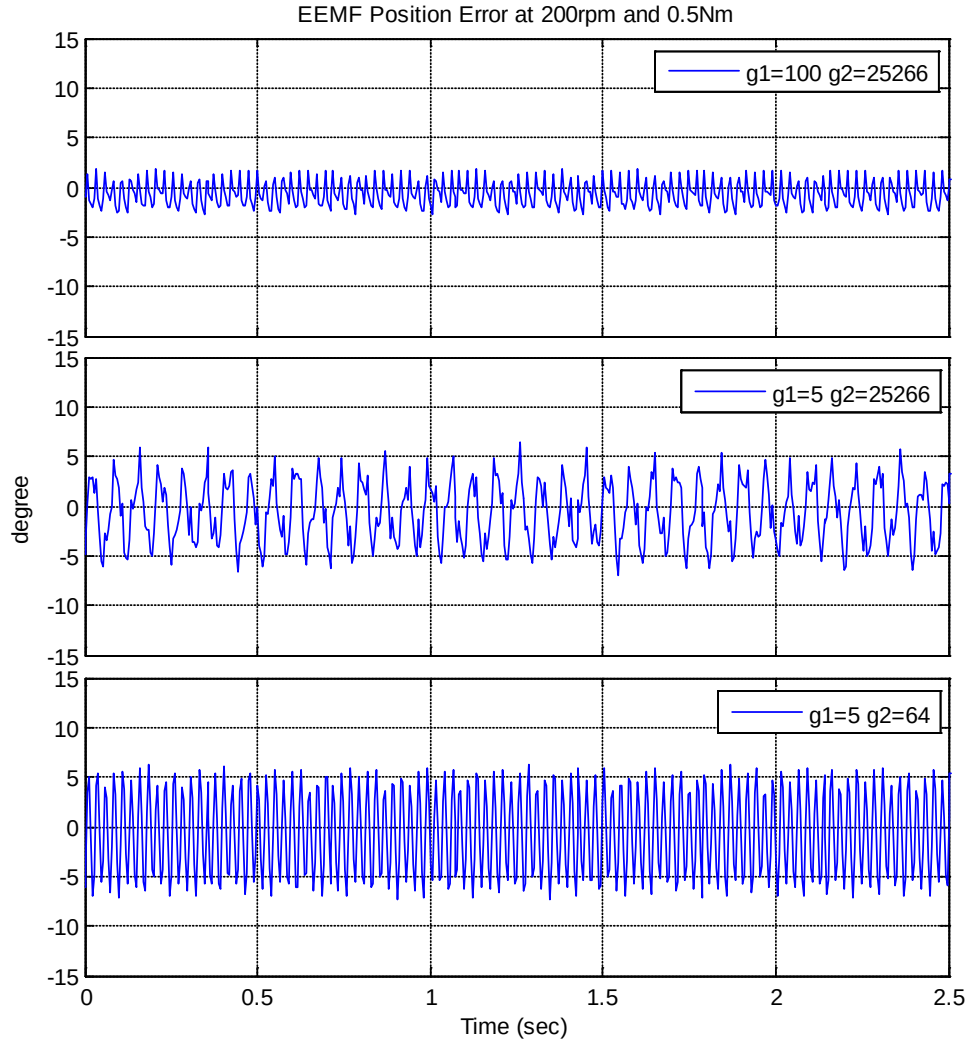


Figure 4.9 : Effects of g_1 and g_2 on EEMF position error 1.

A second group of measurements were performed at the same speed but at a near no-load condition, to test the effect of selecting very high values for g_1 and g_2 . The results, shown in Figure 4.10, indicated that this type of gain selection also increased the oscillation of position estimation. In fact, the PI compensator block of the estimator became less stable with high gains, and it should be noted that estimator became unstable for g_1 values above 500.

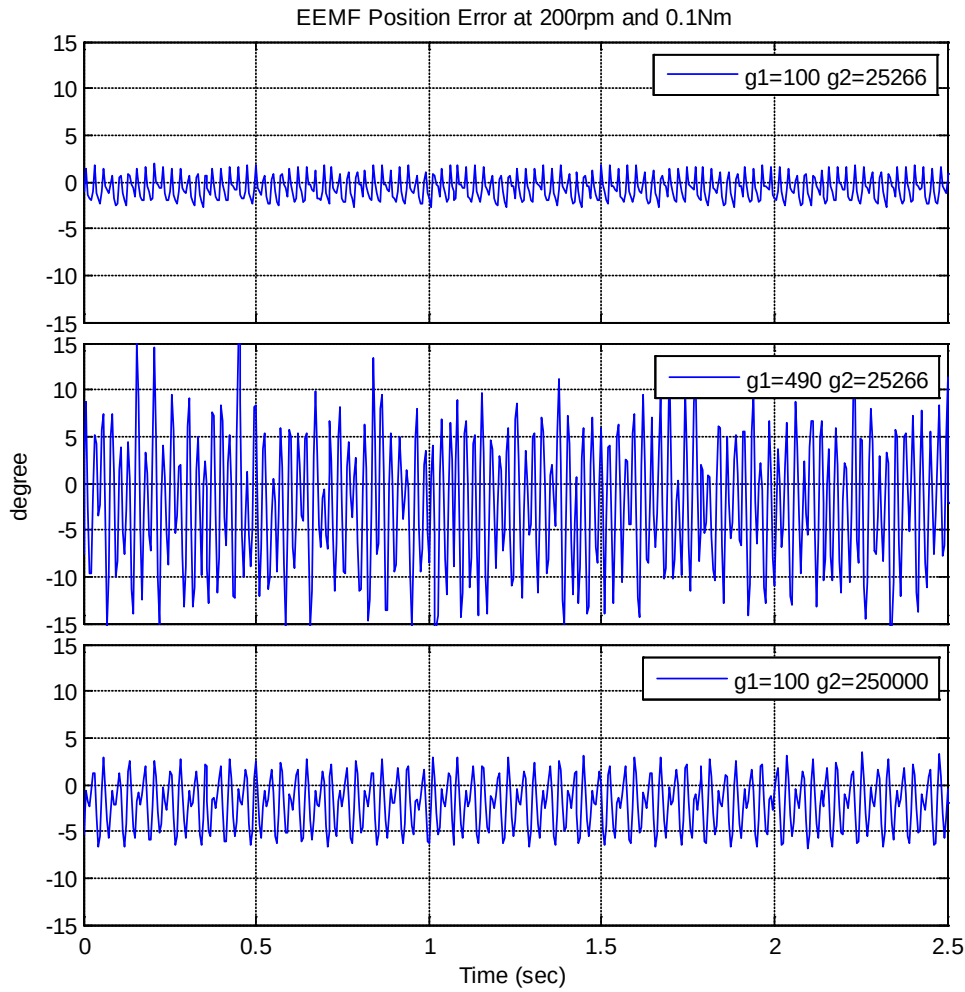


Figure 4.10 : Effects of g_1 and g_2 on EEMF position error 2.

The last two gains of EEMF method, namely g_3 and g_4 , were used in the angle tracking observer (ATO) block. This block's input was a term proportional to the position estimation error, which was calculated using the components of estimated EEMF. The outputs of ATO were the estimated position and speed. The estimated speed was in turn used in the EEMF estimator part of the method since it is a speed adaptive estimator.

ATO was very successful in producing estimation signals with very low noise content for both speed and position, due to its inherent filtering feature. However, if not tuned properly, it introduced a delay in the estimated speed (Figure 4.11). Therefore, in order to observe the effect of g_3 and g_4 , experiments were performed at 50% load (1Nm) while the motor was rapidly accelerated and decelerated in a

periodic manner. The resulting speed and position estimations for three sets of g_3 and g_4 values are given in Figure 4.11, Figure 4.12 and Figure 4.13.

It was observed that the delay between the estimated and actual motor speeds were proportional to the value of g_4 , and selecting 0.005 for this gain eliminated the delay almost completely (Figure 4.11 and 4.12). Another finding was that the position estimation error was inversely proportional to the value of g_3 and selecting 30000 for this gain limited position estimation error in -5 to 0 degrees range. Finally these three experiments showed that the stability of the position estimation error was proportional to the ratio between g_3 and g_4 . Figures 4.11 and 4.13 show that the ratio between g_3 and g_4 was the same and there was no significant oscillation of position estimation error in neither of the cases. On the other hand, in the case of Figure 4.12, the ratio of g_3 to g_4 was significantly decreased, which resulted in large oscillations in the position estimation error.

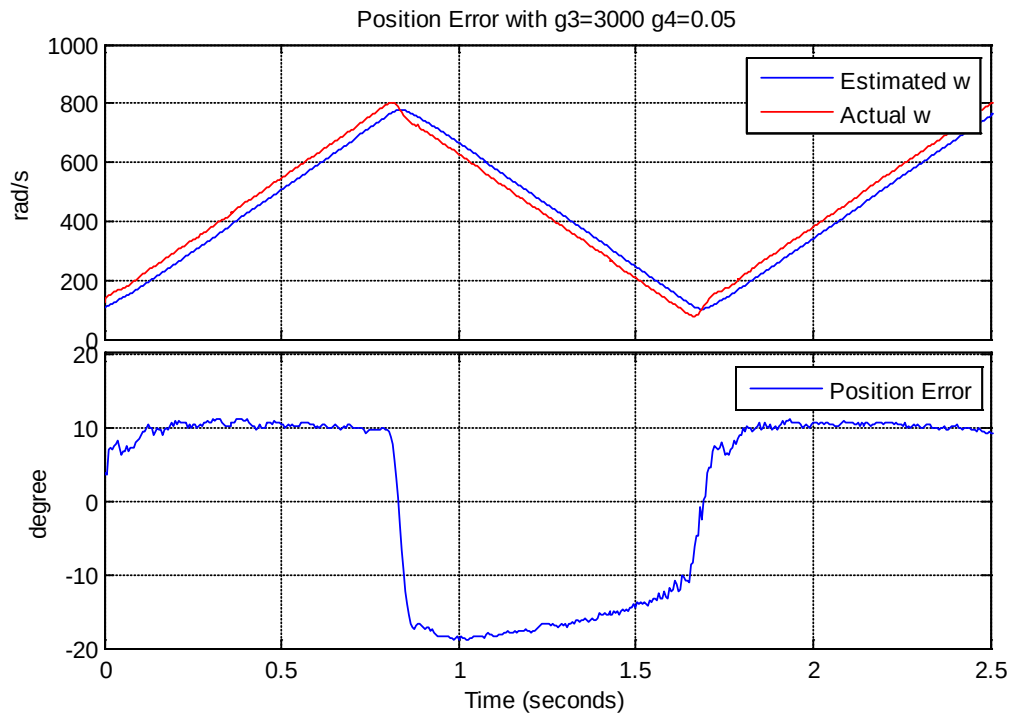


Figure 4.11 : Effects of g_3 and g_4 on EEMF position and speed errors 1.

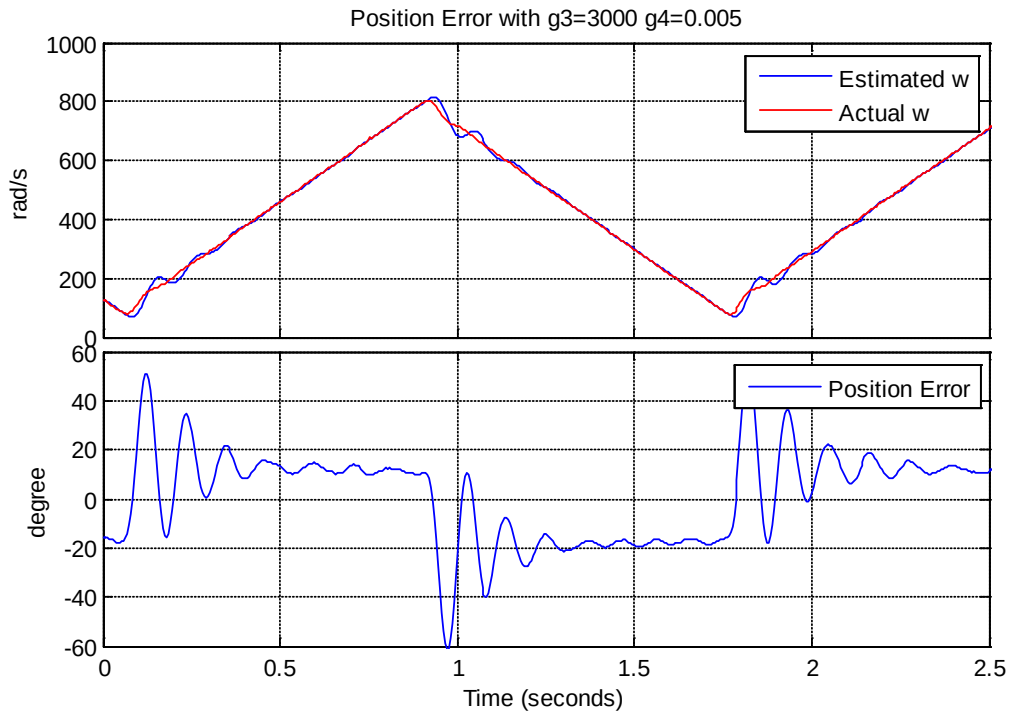


Figure 4.12 : Effects of g_3 and g_4 on EEMF position and speed errors 2.

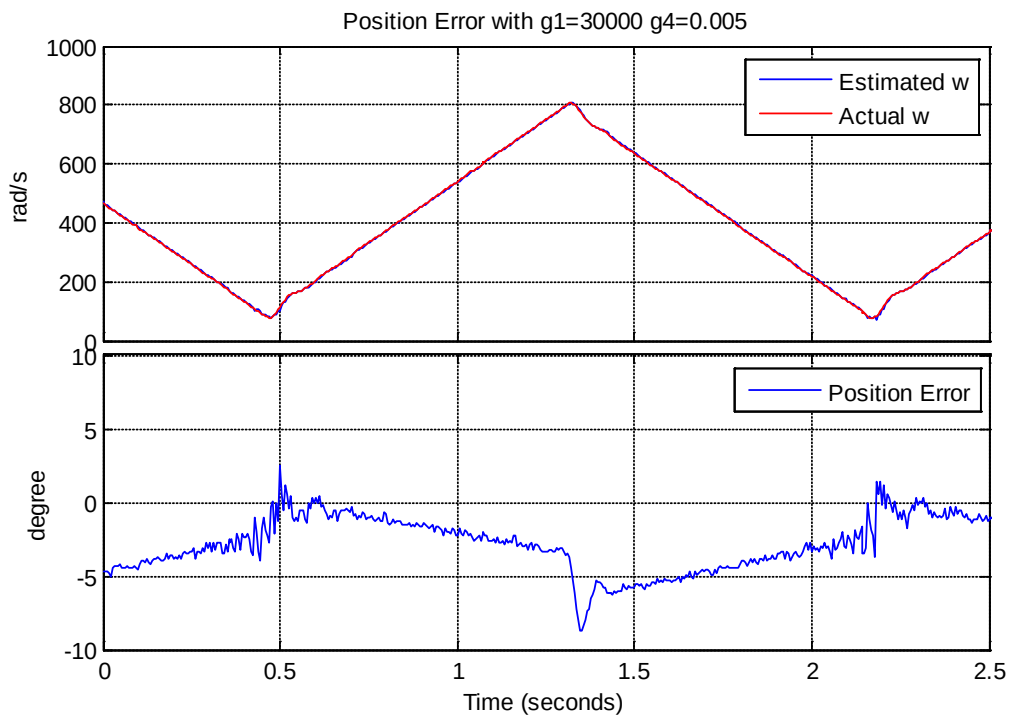


Figure 4.13 : Effects of g_3 and g_4 on EEMF position and speed errors 3.

4.3.2 The GDM method

The GDM is an estimation method in which there is only one gain, namely g_1 , which greatly simplifies the gain selection process when compared to the other two estimation methods investigated in this study. To determine the effect of this gain, experiments were performed at three different speeds while the load was at 50% and the position estimation error of GDM was measured in each experiment for different values of g_1 . The results are shown in through Figures 4.14 to 4.16.

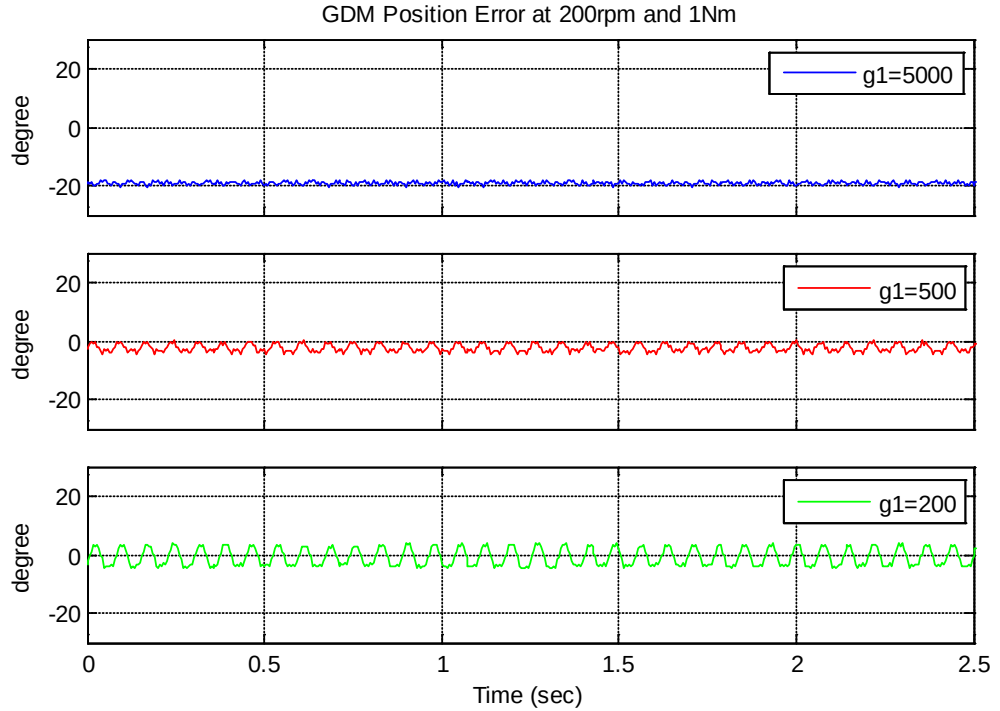


Figure 4.14 : Effects of g_1 on the GDM position error at 200 rpm.

At all three speeds, g_1 was found to be inversely proportional to the amplitude of oscillations of the position estimation error. This result is in agreement with the formulation of the GDM method, where it is stated that g_1 is proportional to the convergence rate of the stator flux estimation. As motor speed dictates the fundamental frequency of all the states in the motor, the frequency of stator flux increases proportionally with increasing speed as well. As a result at higher speeds an estimator with a faster convergence rate is required for accurate estimation of stator flux. The result of this phenomena was observed when the results for g_1 equal to 200, shown in Figure 4.12 and 4.14, were compared. At 200rpm, selecting 200 for g_1 results in an observer that was not fast enough, and the amplitude of the oscillations in estimation error was nearly 10 degrees.

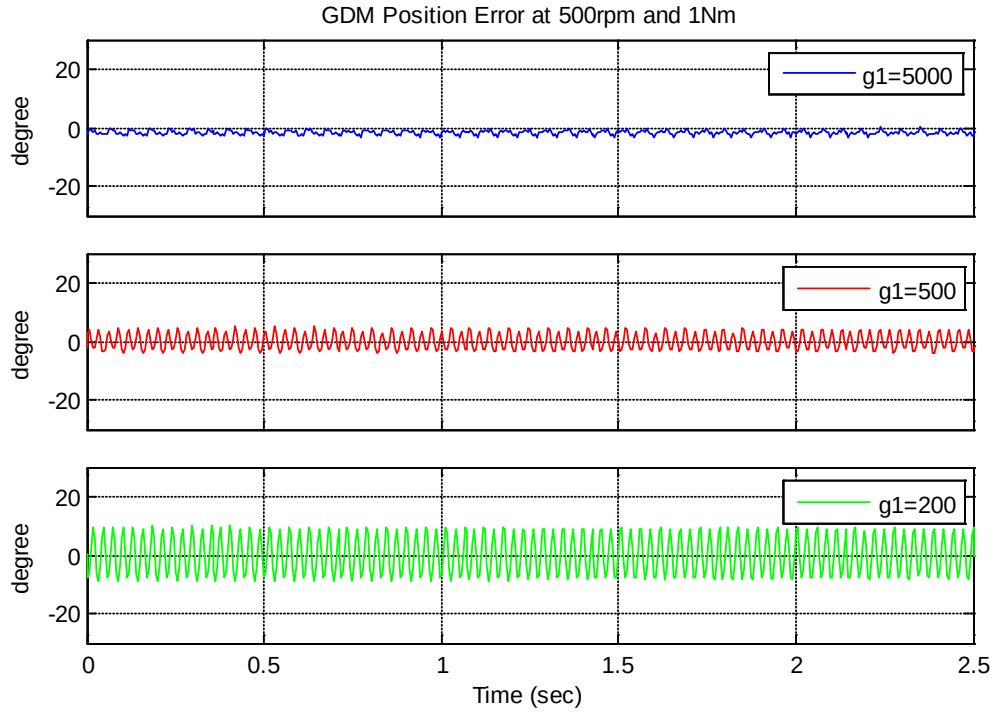


Figure 4.15 : Effects of $g1$ on GDM position error at 500 rpm.

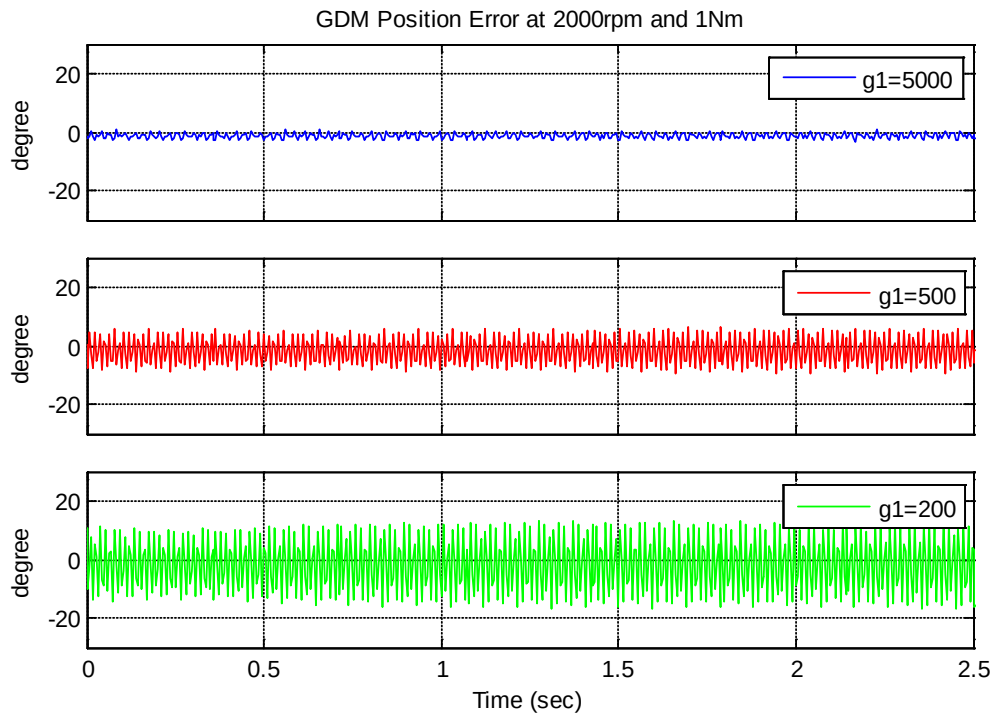


Figure 4.16 : Effects of $g1$ on GDM position error at 500 rpm.

On the other hand, selecting a very high value for $g1$ was also not preferred as it was seen to increase position estimation error at low speeds (Fig.4.14). These results

suggested that a speed dependent variable gain selection method was best suited for the GDM.

Thus, for the experiments performed in the following sections g_1 was selected as 500 when speed is below 500rpm and it was selected as 1000 when speed was above 500rpm.

4.3.3 The AF method

The active flux method includes two gains, namely g_1 and g_2 , in its stator flux estimator. In the experiments, the values selected for these gains were determined to effect position estimation accuracy considerably. To be able to compare multiple sets of values for g_1 and g_2 , experiments were performed at 200rpm and at 0.5Nm.

In Figure 4.17, seven sets of values for g_1 and g_2 were compared. Among these candidate values, $g_1=10$, $g_2=30$ and $g_1=30$, $g_2=30$ were shown to yield most accurate position estimations, with minimum noise content in the estimation waveform.

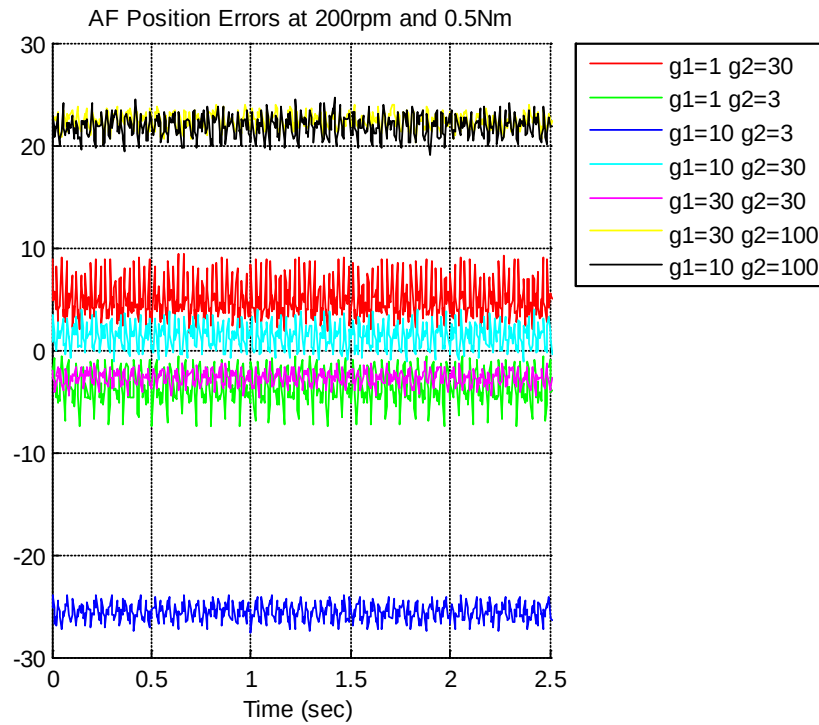


Figure 4.17 : Effects of g_1 and g_2 on AF position error 1.

In order to select one of these two sets, an additional experiment set was performed, in which the change in position estimation error with respect to speed was

investigated for both set of gains. It was determined that the position estimation error increased as speed increased in the case of $g_1=30$ and $g_2=30$, whereas when 10 and 30 were chosen for gains g_1 and g_2 , respectively. The position estimation remains almost constant as if it was independent of speed, and its value near zero at this speed range. As a result, $g_1=10$ and $g_2=30$ were used for experiments, in the following sections.

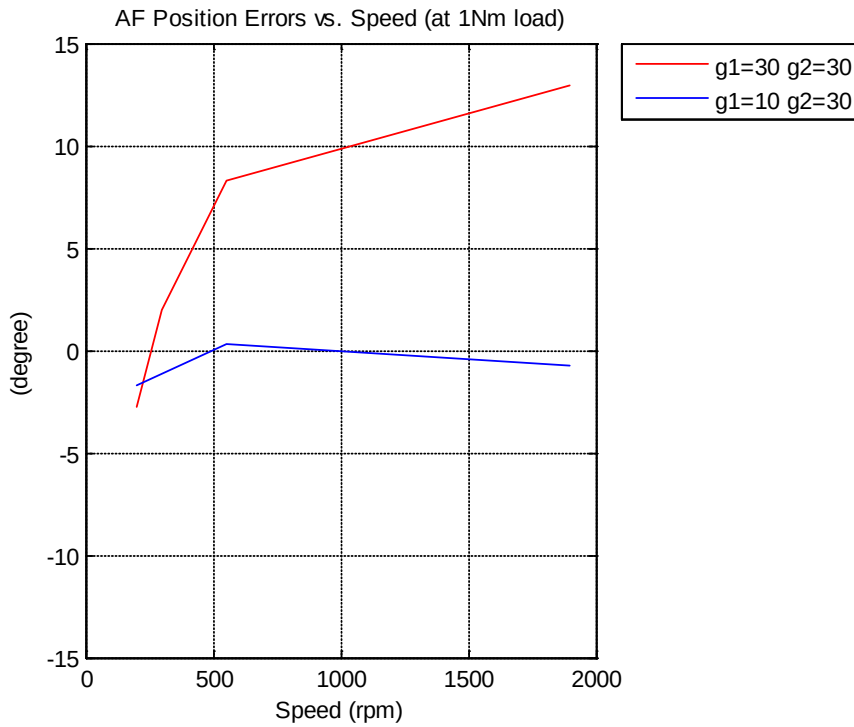


Figure 4.18 : Effects of g_1 and g_2 on the AF position error 2.

4.4 Steady State Operations

The estimation accuracies of the observers were evaluated under different operating conditions. The amount of load torque was proportional to the amplitude of phase currents, and the speed of the motor determined electrical frequencies of all the signals in the motor. Therefore, the performances of the observers had to be observed at different speed and load levels.

In this section, results of experiments performed at different motor operation points are discussed. All the measurements were taken at steady state conditions.

Experiments were grouped according to the motor speed at which they are performed. At each speed, measurements were performed at different loads.

For the values of controller gains, observer gains, and motor parameters, those given in Table 3.1 are used.

Since the driver did not have a torque sensor, the load torque could not be measured simultaneously with position estimation error by the inverter. However, the proportionality of the load torque to the quadrature current could be used as an indicator of the load torque. For this reason, results were plotted with respect to the driver software variable named i_{qref} , which held the quadrature current reference value that was the input of i_q current controller.

4.4.1 The EEMF method

The average position estimation errors obtained at steady state conditions when the EEMF was used are shown in Fig.4.19, Fig.4.20, and Fig.4.21. In each of these figures, the results are presented for both the sensed and sensorless control cases. In the sensed control encoder, position measurements were used to coordinate transformations as a part of motor control, whereas in sensorless control position estimated by EEMF is used for the same purpose.

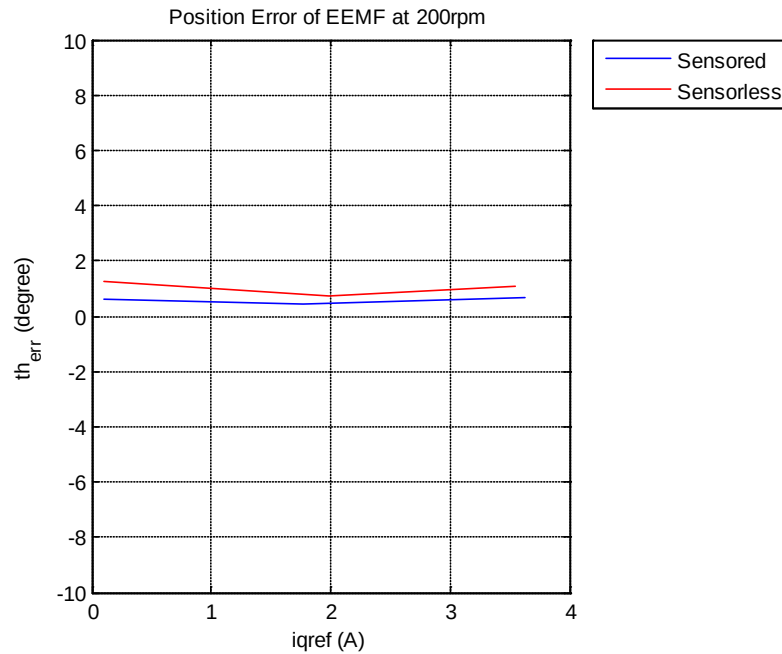


Figure 4.19 : Position estimation error of EEMF at 200 rpm.

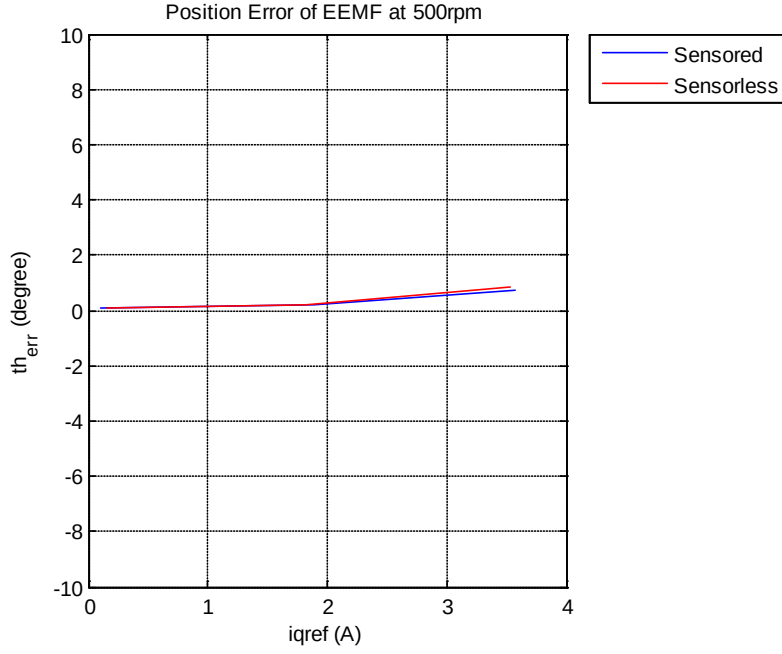


Figure 4.20 : Position estimation error of EEMF at 500 rpm.

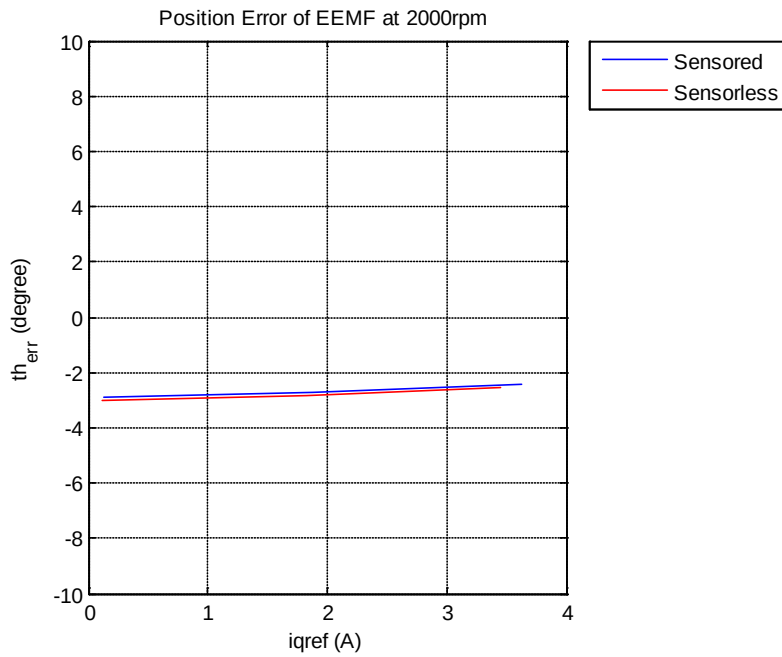


Figure 4.21 : Position estimation error of EEMF at 2000 rpm.

According to the measurements, the position estimation error of EEMF remained nearly unchanged as load (i_{qref}) was increased. This behavior was observed at all speeds. However, when speed was increased it was observed that position estimation error shifted towards negative direction. Nevertheless, the EEMF position estimation error was found to be less than 4 degrees in all operating points tested in this section.

It was also noticed that the position estimation accuracy is nearly the same in both sensed and sensorless operations with the difference being always less than 2 degrees.

4.4.2 The GDM method

The average position estimation errors obtained at steady state conditions when the GDM was used are presented in Fig. 4.22, Fig. 4.23, and Fig. 4.24. Similar to the case of EEMF, in each of these figures the results are given for both the sensed and sensorless control cases.

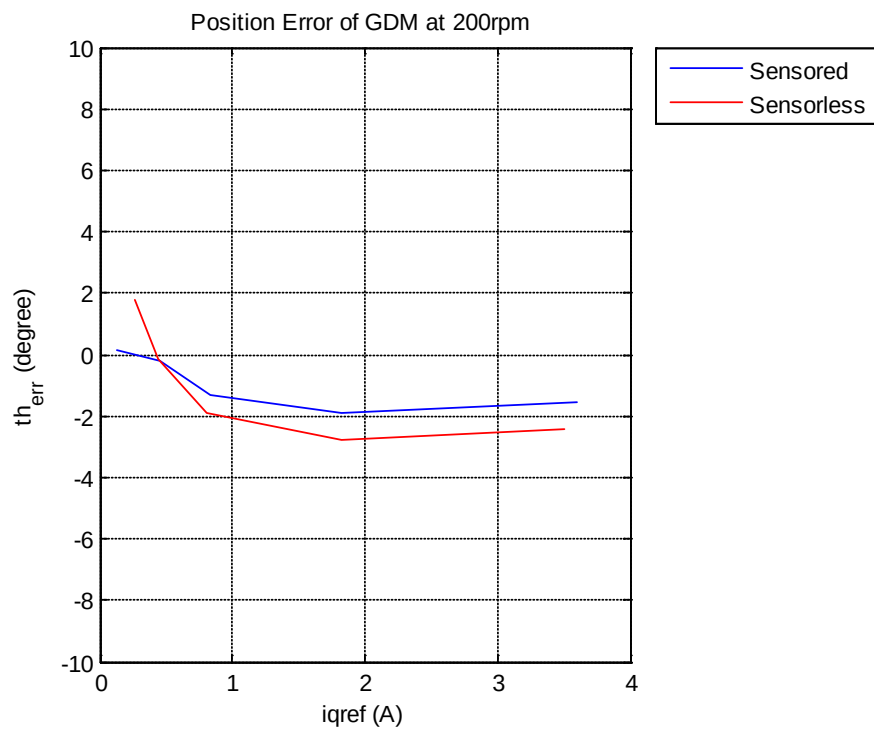


Figure 4.22 : Position estimation error of GDM at 200 rpm.

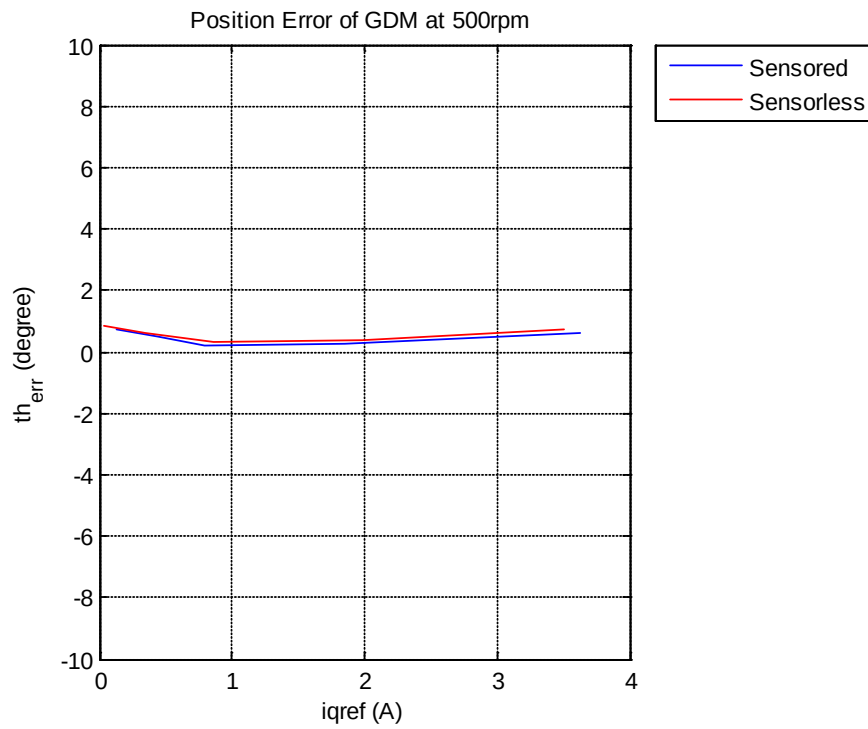


Figure 4.23 : Position estimation error of GDM at 500 rpm.

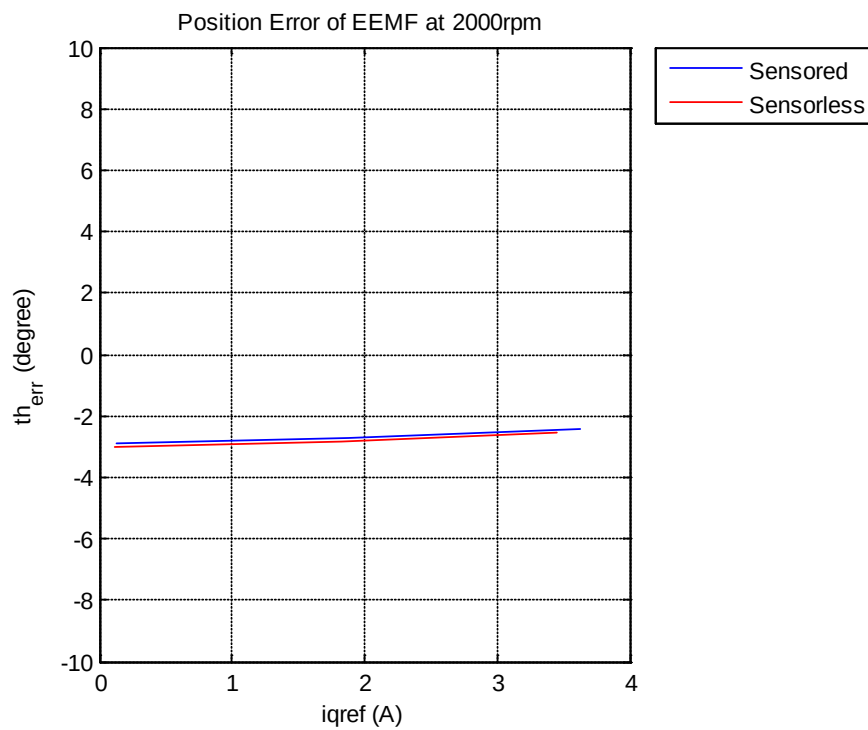


Figure 4.24 : Position estimation error of GDM at 2000 rpm.

The GDM observer had a very good estimation accuracy but it should be noted that in these measurements motor inductances used were larger than the actual inductances. These values were determined considering a strong dependency to motor inductances observed at preliminary experiments and the inductances that give the most accurate estimations. Especially, increasing L_q significantly decreased the estimation error.

Table 4.2 : Inductance values used in experiments with the GDM observer.

Inductance	Actual Value (mH)	Value in GDM (mH)
L_d	13.4	15
L_q	18.6	30

The value of the estimation error was almost independent from both the speed and load operating conditions. This was achieved by the speed dependent gain used in the observer, which were initially 500. However, as motor speed increased above 500 rpm it is also increased to 1000. A small dependency of error to the load was determined at 200rpm for loads smaller than 0.5Nm (when $i_{q_{ref}}$ was approximately 1A) as there was a difference of 4° between no-load and 0.5Nm load conditions. As the estimation accuracy is high, switching to sensorless operation did not affect the position estimation error significantly. The difference between sensed and sensorless operation was less than 1° at all points of operation except the no load condition at 200rpm, which was the worst condition for the GDM. Even at this point, the difference between estimation errors at sensed and sensorless operation remained below 2° .

4.4.3 The AF method

The average position estimation errors obtained at steady state conditions when AF was used are presented in Fig.4.25, Fig.4.26, and Fig.4.27. Similar to the cases of the previous observers, the results are shown for both the sensed and sensorless control cases in each of these figures.

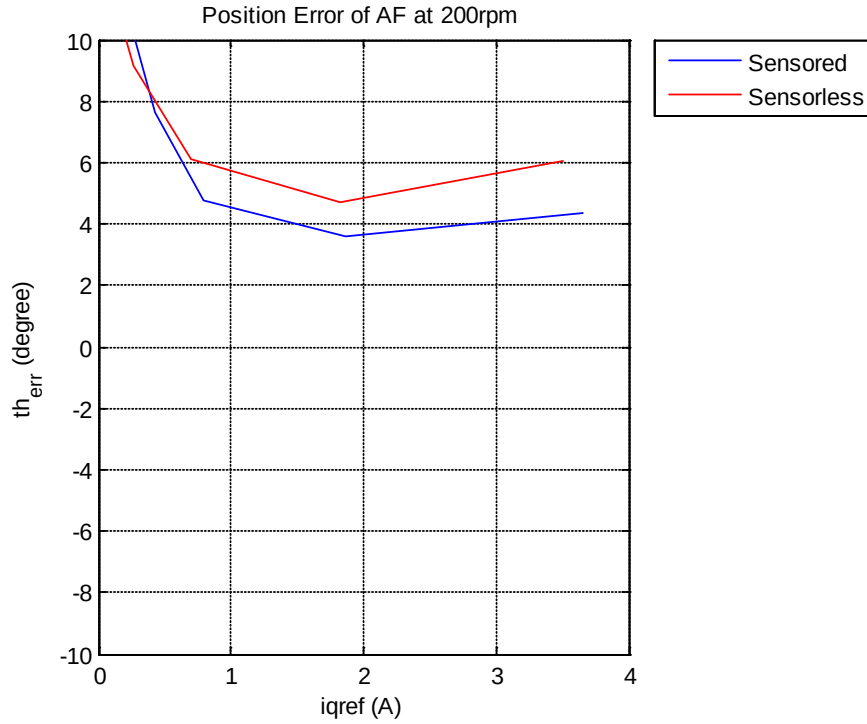


Figure 4.25 : Position estimation error of AF at 200 rpm.

As a result, it was observed that AF had a 5° position estimation error regardless of the operating point at all the three speeds. This error increased to almost 12° for the no load case at 200rpm (Figure 4.25). Another observation common to all three speeds was that for low load region ($i_{qref} < 1.5A$), the position estimation error decreased with increasing load but for high load region ($i_{qref} > 1.5A$), it increased with increasing load.

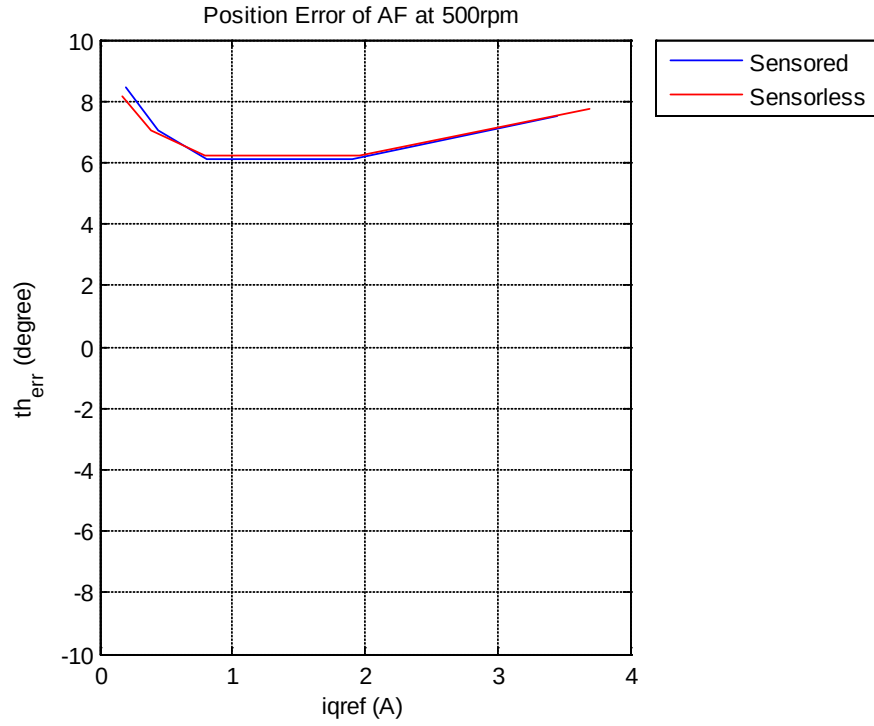


Figure 4.26 : The position estimation error of AF at 500 rpm.

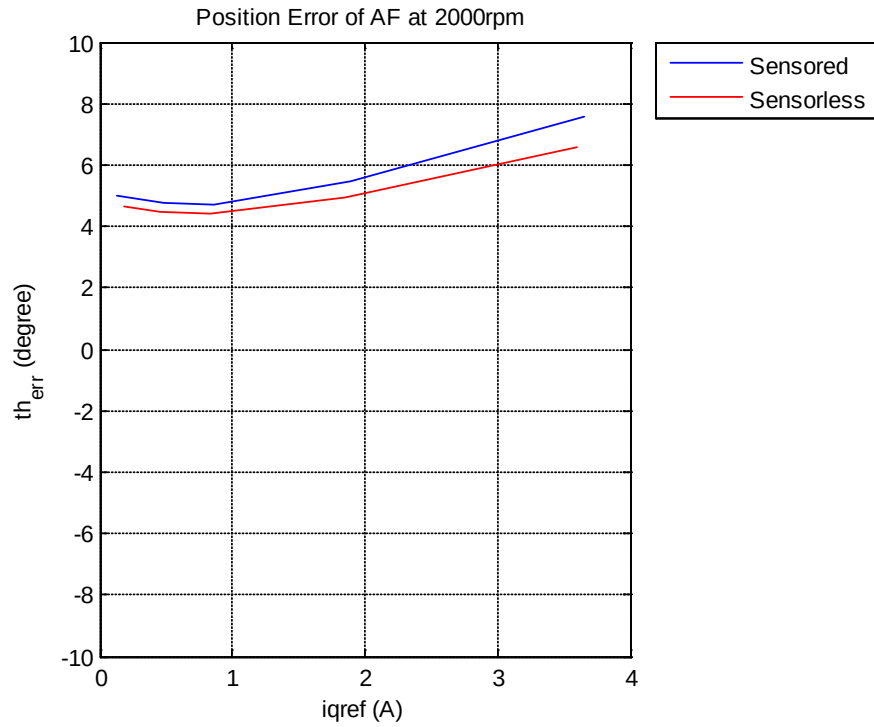


Figure 4.27 : The position estimation error of AF at 2000 rpm.

Furthermore, for lower speeds, the strength of the dependency of position estimation error to load was low in the high load regions (Fig.4.25). On the other hand, for higher speeds this dependency was low in the low load regions.

4.4.4 Comparison of three sensorless methods

In this section, the results obtained in the previous three sections are compared. All the results presented in this section were measured during sensorless motor operation.

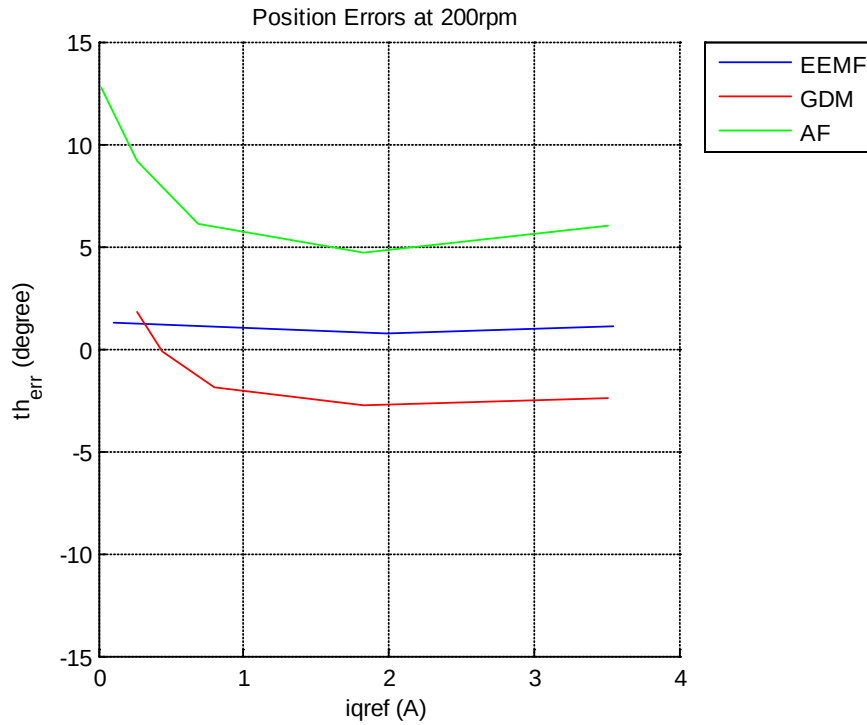


Figure 4.28 : Position errors of three estimators at 200 rpm.

When the results presented in Fig.4.28, Fig. 4.29 and in Fig.4.30 are compared, it can be observed that there is a shift towards negative direction for all three methods when the speed is increased. This is an expected result, since the methods were implemented in a discrete manner and there was a finite time interval between two estimation updates, which equals to PWM switching period of 50 microseconds in this test setup. As motor speed increased, the delay in estimation due to finite update intervals became a considerable factor affecting the estimation accuracy.

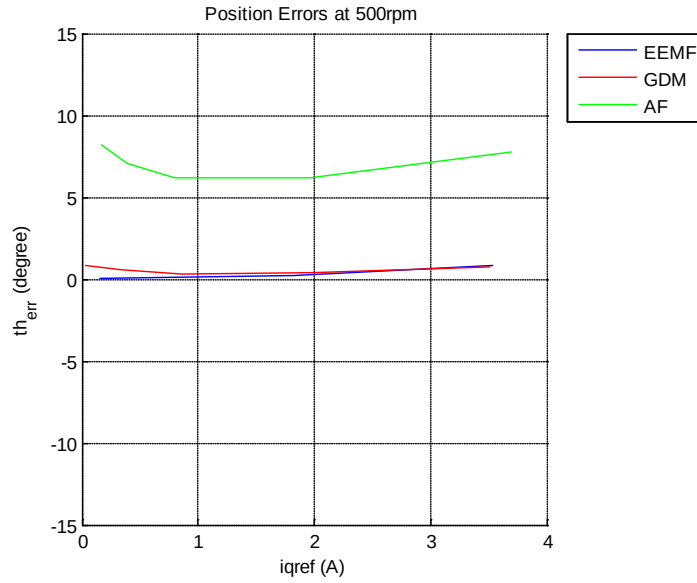


Figure 4.29 : Position errors of three estimators at 500 rpm.

The second observation valid for all three speeds is that the estimation accuracy of EEMF and GDM were closer to each other and also more accurate than AF by approximately 5° - 7° .

In summary, the best steady state results were obtained by the EEMF observer, as its estimation error was the least load dependent. On the other hand, the worst results were obtained by the AF, as it was the most load dependent observer.

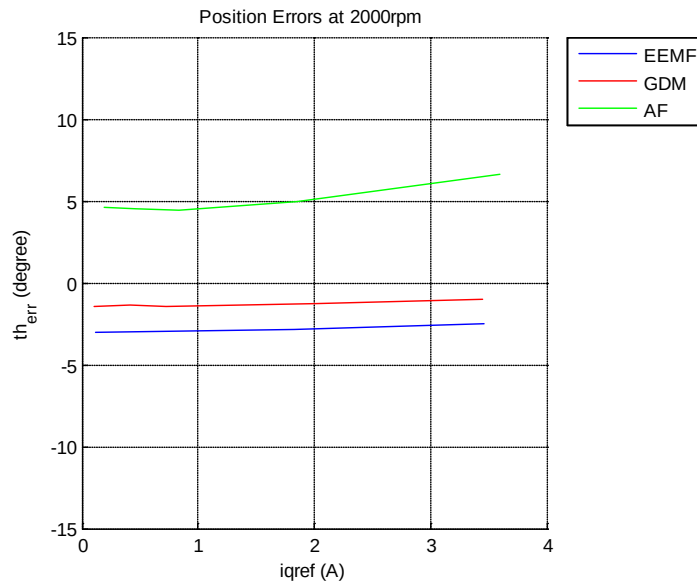


Figure 4.30 : Position errors of three estimators at 2000 rpm.

4.5 Transient Operation During Rapid Changes in Torque

The experiments in this section were aimed to assess the position estimation performance of observers during a rapid load increase or decrease. The torque increase case was examined in more detail, as it was the more complicated of the two scenarios.

For the torque increase experiments, motor was operated at a near no-load torque at the beginning of the measurements. Then, a rapid increase in load was applied and the measurement was continued until a second steady state was reached.

As all the three methods are fundamental excitation methods, a low speed of 300rpm was selected for the experiments. At this speed, the performances of these methods are more prone to degradation during torque increase, which has an initial decelerating affect on the motor.

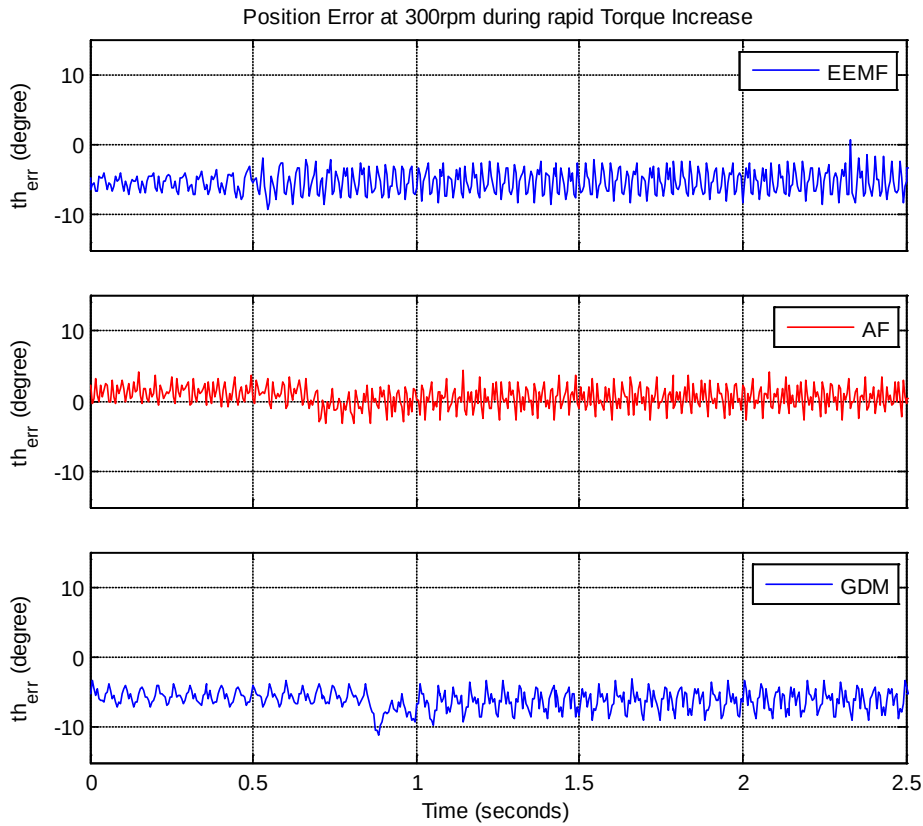


Figure 4.31 : Position error at load increase (300 rpm, $K_{psp}=0.01$ $K_{isp}=1$).

According to the results of this experiment, shown in Fig.4.31, the position estimation accuracy of all the three methods remained nearly unchanged during torque increase. This is due to the fact that the high speed controller gains were used. The speed and the current controller gains used in all experiments up to this point are given in Table3.1. High speed gains allow speed controller to quickly react to torque changes produced by the hysteresis brake in the test setup. This way, the amount of initial deceleration that was observed upon sudden torque increase was minimized, which in turn helps the position estimation methods to perform better during load transitions.

The results presented in Fig.4.31 are not suitable to compare methods. To be able to amplify the differences between the methods during torque increase, a slower speed controller was used and the experiments were repeated. With these new controller gains, it was necessary to make modifications in the gain selection of current controllers and position estimation methods as well. Table 4.3 shows the values of all the parameters used for obtaining the results shown in Figures 4.32 to 4.35.

Table 4.3 : Parameter set of the slower speed controller.

Gain	Value
R	3.5 Ω
L_d	13.4 mH
L_d	18.6 mH
Ψ_{pm}	0.105 Vs/rad
$K_{p_{speed}}$	0.005
$K_{i_{speed}}$	0.1
$K_{p_{id}}$	50
$K_{i_{id}}$	40000
$K_{p_{iq}}$	50
$K_{i_{iq}}$	40000
EEMF/ g_1	50
EEMF/ g_2	25266
EEMF/ g_3	5000
EEMF/ g_4	0.04
GDM/ g_1	Equal to speed (rpm)
AF/ g_1	15
AF/ g_2	30

Fig.4.32 and Fig.4.34, show the position estimation errors during torque increase at 300 rpm and 2000rpm respectively, whereas, Fig.4.33 and Fig. 4.32 show the position estimation errors during torque decrease, again at 300 rpm and 2000rpm respectively.

It was observed that the observer least affected by the rapid changes in load at both speeds was the active flux observer. In general, the AF and the GDM were less affected by these sorts of transients since they are not speed dependent observers. The EEMF position error increased considerably during the transient when compared to other two at 2000rpm.

At the lower speed of 300rpm, a decrease in the performance of the GDM was observed during load increase. This decrease can be attributed to the speed dependent observer gain used in this method, in which the gain is decreased to a lower value during operation at 300rpm. However, even at this speed, the largest transient position estimation error was again that of the EEMF when the load was decreased. This is a result of the speed adaptive nature of the EEMF. In estimating the extended EMF, the estimated motor angular speed is used as a parameter in the observer equations. The observer also features an angle tracking observer that adaptively updates the estimated speed. During fast torque increase, the motor speed decreases considerably before the speed controller is able to respond to this change.

At the same time, an increase in the speed estimation error of the EEMF observer was observed. This, in turn, resulted in an increase in position estimation. On other hand, the AF and GDM were not affected by a rapid increase in torque at 2000 rpm since the estimated speed is not used in observer equations.

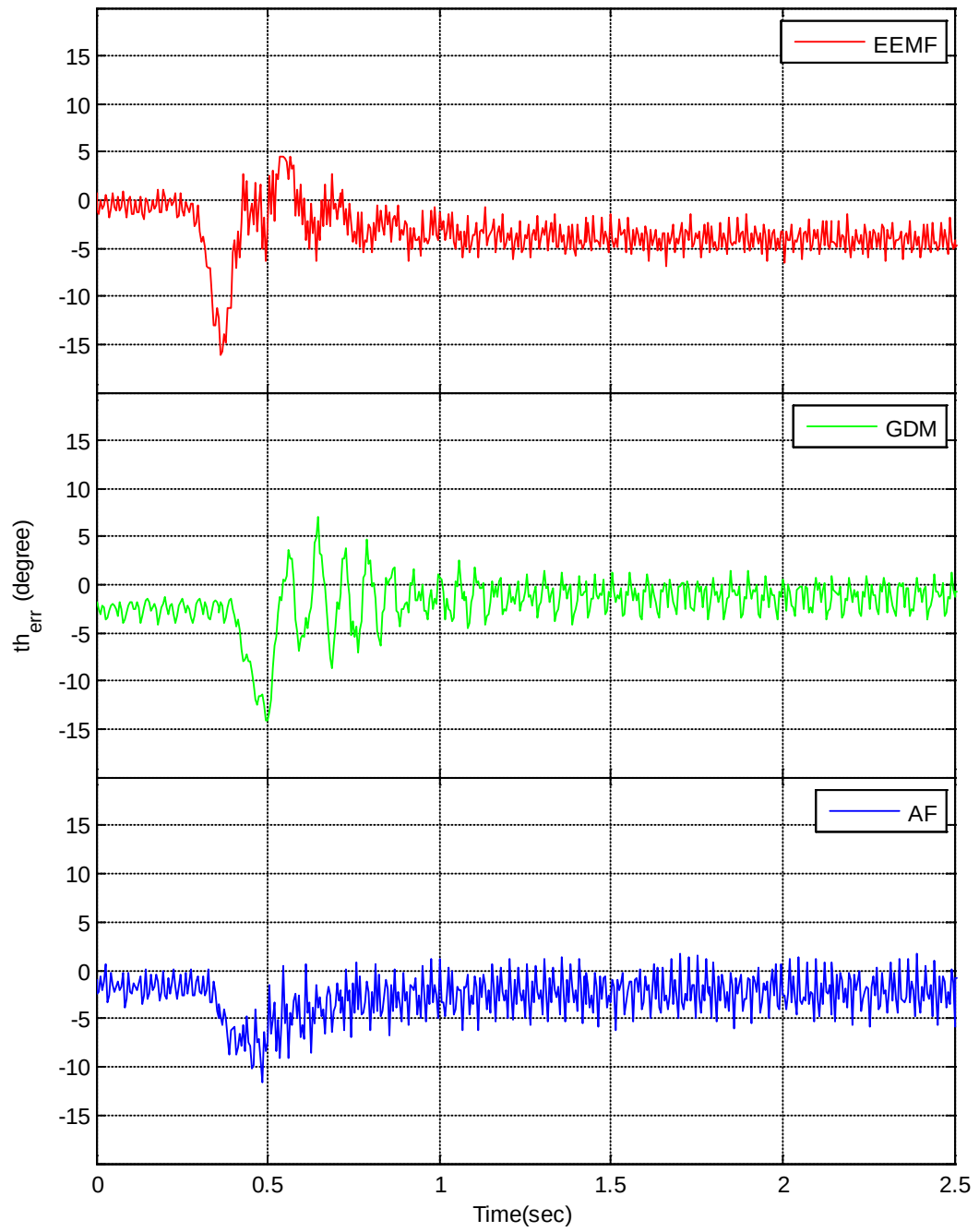


Figure 4.32 : Position error at load increase (300 rpm, $K_{psp}=0.005$, $K_{isp}=0.01$).

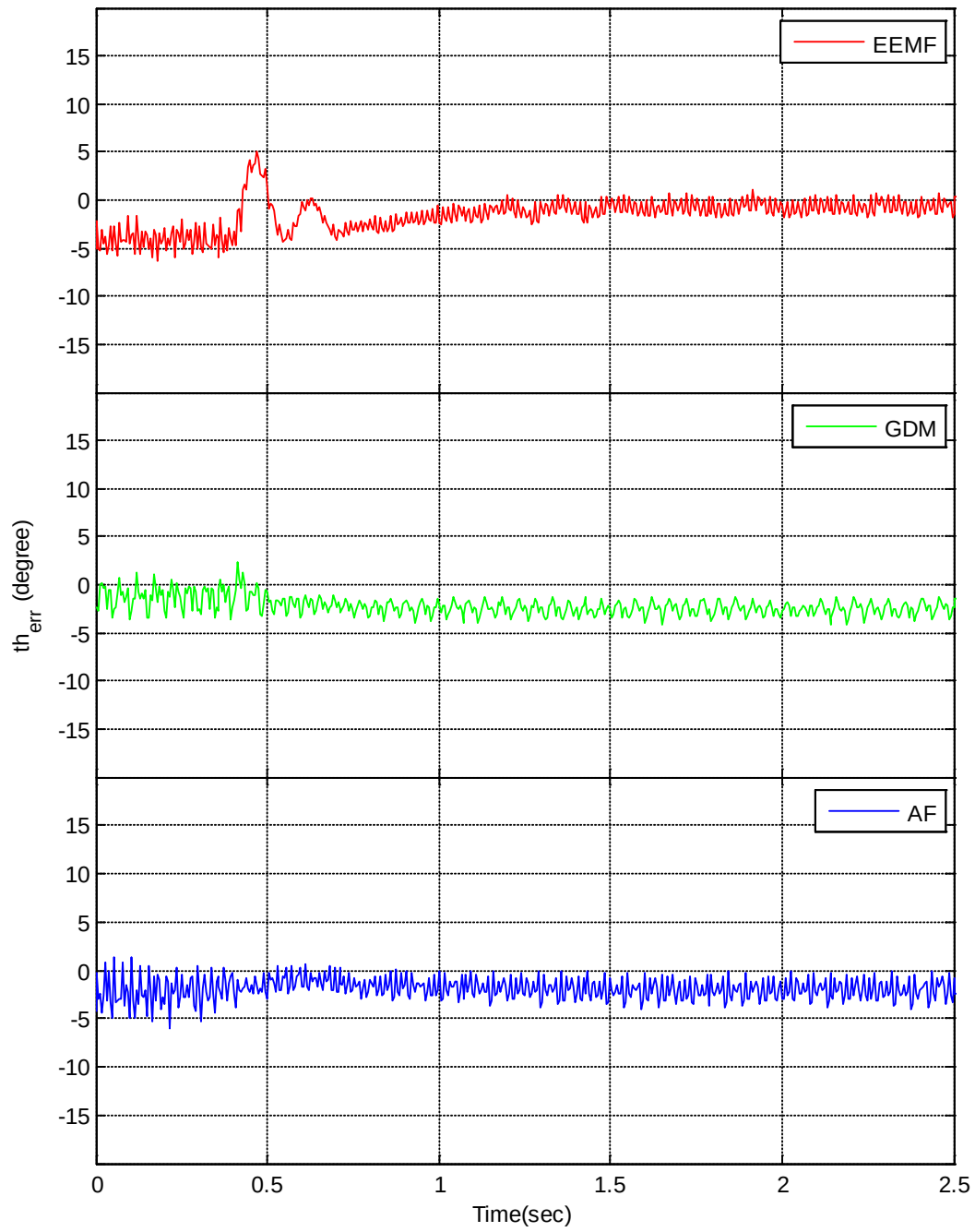


Figure 4.33 : Position error at load decrease(300rpm, $K_{psp}=0.005$, $K_{isp}=0.01$).

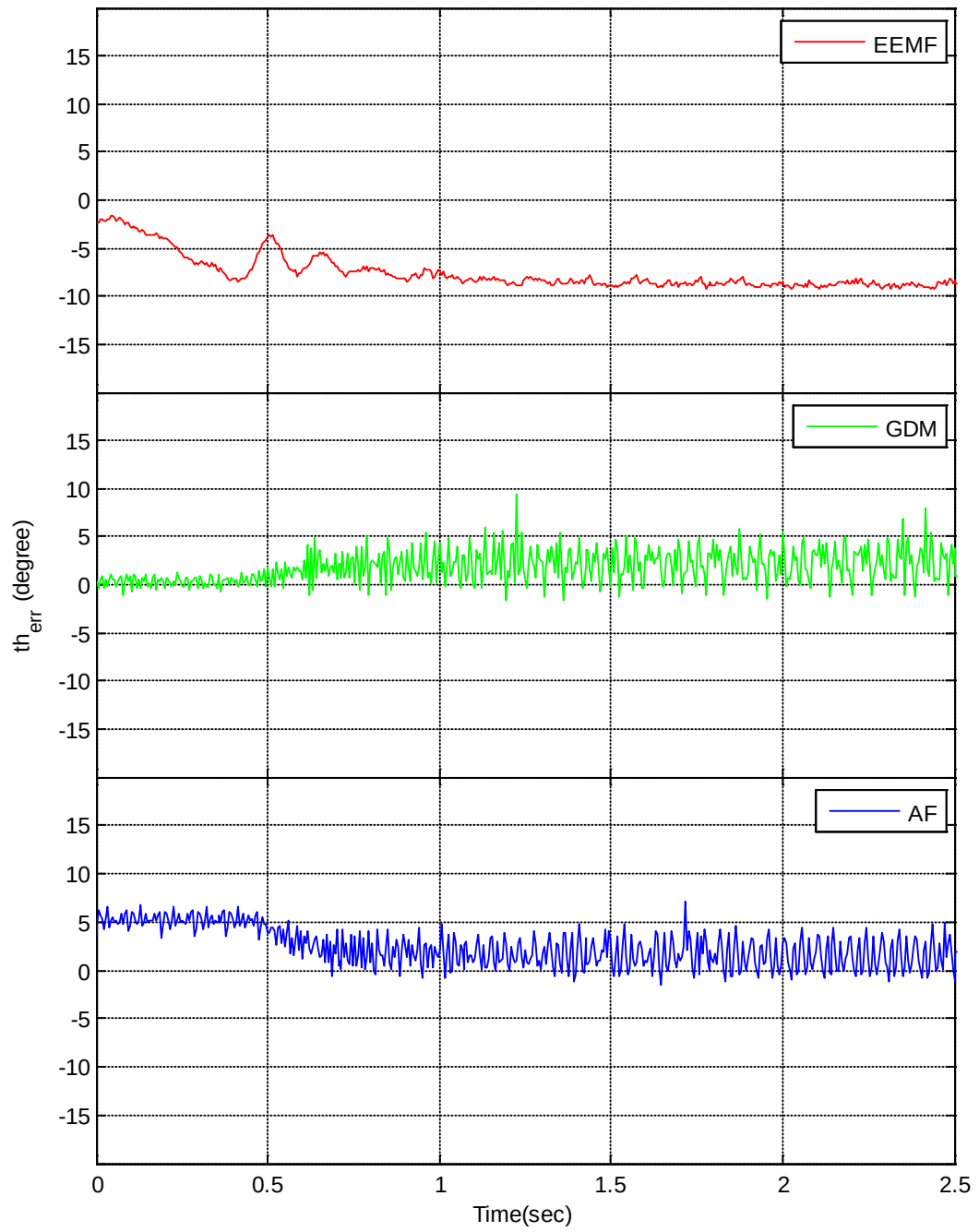


Figure 4.34 : Position error at load increase(2000rpm, $psp=0.005$, $isp=0.01$).

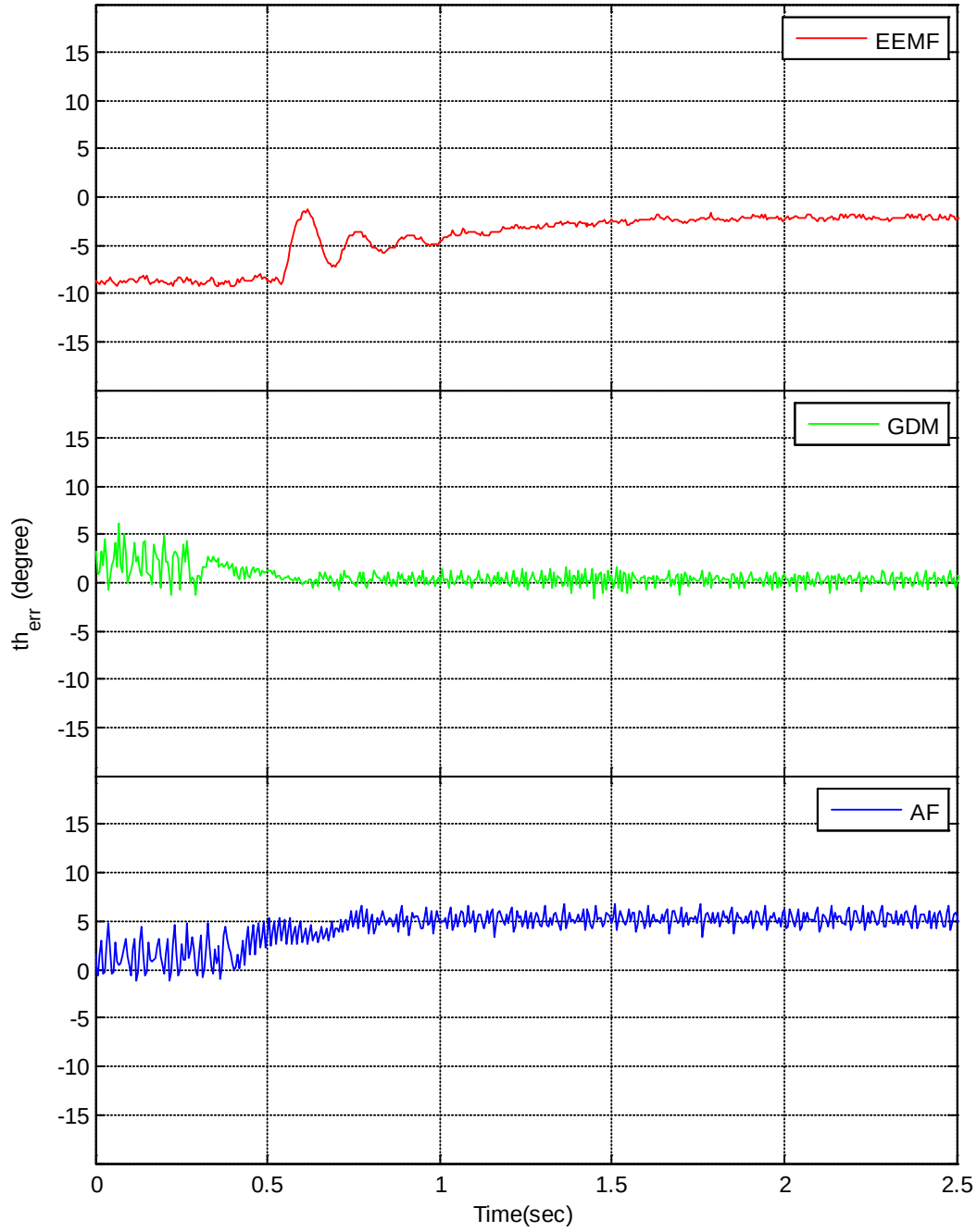


Figure 4.35 : Position error at load decrease(2000rpm, $K_{psp}=0.005$, $K_{isp}=0.01$).

4.6 Transient Operation During Rapid Changes in Speed

In this section, experiments were performed while the motor speed was changing. The motor was accelerated and decelerated periodically under 50% load condition and the behaviours of the three observers were measured. During the experiments, gains given in Table 3.1 were used for the controllers and observers.

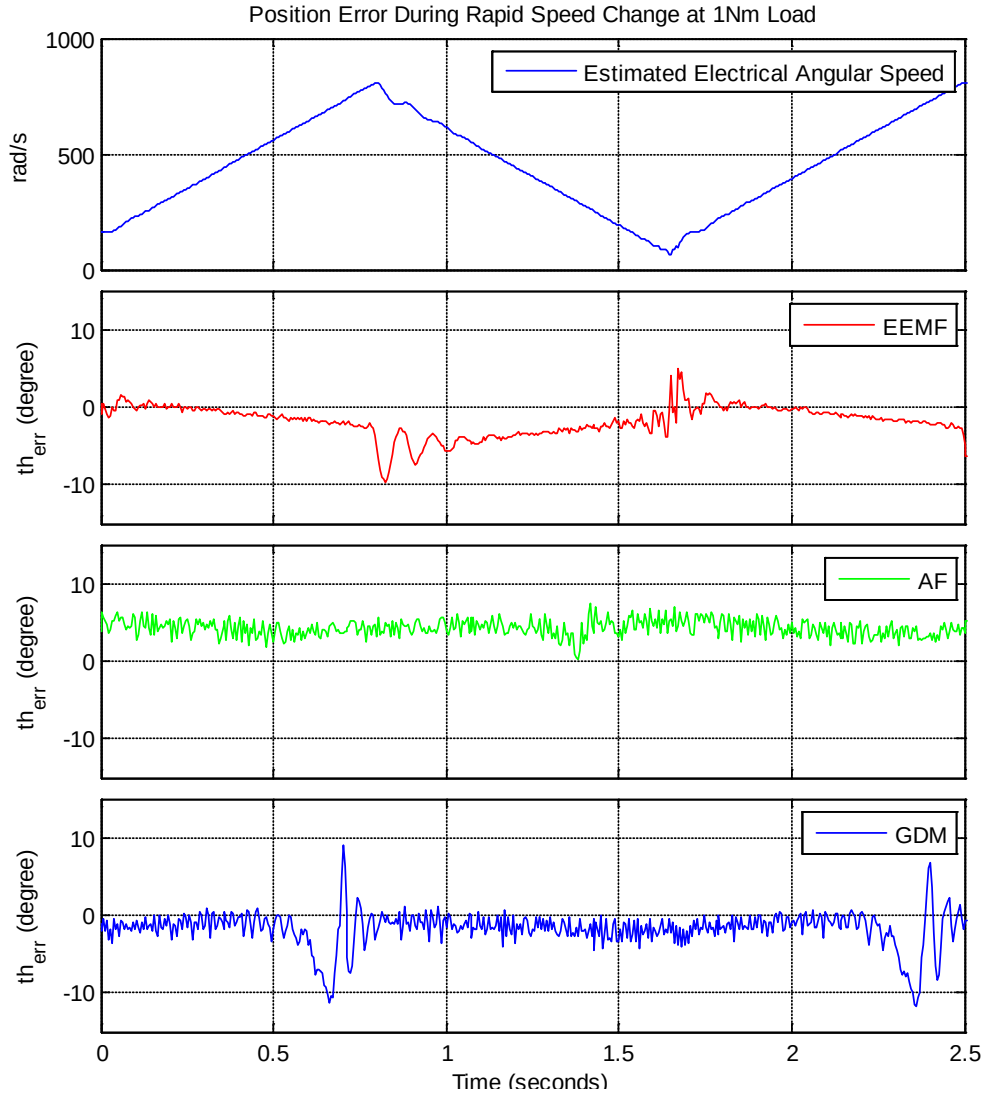


Figure 4.36 : Position errors during speed change (1Nm).

The measurements were taken during sensorless operation. The obtained results showed that the AF performed the best during rapid changes in speed. Due to the speed adaptive nature of the EEMF and the approximations performed during its derivation, its position estimation error exhibited a speed dependent nature during both acceleration and deceleration phases both of which are shown in Fig. 4.36. The GDM also appeared to have performed poorly according to the results in Fig. 4.36. The speed estimation calculation performed using the components of the estimated flux in the AF was also used in the GDM. This calculation which was originally proposed for the AF resulted in a very noisy speed estimation when applied to the GDM. Having an oscillatory and noisy speed estimation signal decreased the stability of the system and caused the degraded GDM performance observed in Figure 4.36. When actual speed was used instead of the estimated speed,

the performance of the GDM under these circumstances was greatly improved. Therefore, a more appropriate speed estimation scheme suitable for GDM should be employed in future uses of this method.

4.7 Operation Under Parameter Uncertainty

In this section, each parameter was set to 75% or 125% percent of its actual value and the steady state position estimation errors at different loads were studied for each observer. All the experiments in this section were performed at 300rpm. For the experiments in this section, parameters in Table 4.3 were used. It was expected to obtain similar result with the parameters in Table 3.1, since the sensitivity or insensitivity of an estimation method to an uncertainty in one of the motor parameters is mainly determined by the structure of the observer equations. This expectation was further confirmed because the results obtained in this section using the parameters of Table 4.3 were very similar to the corresponding simulation results, which were obtained using the parameters given in Table 3.1.

4.7.1 Operation under resistance uncertainty

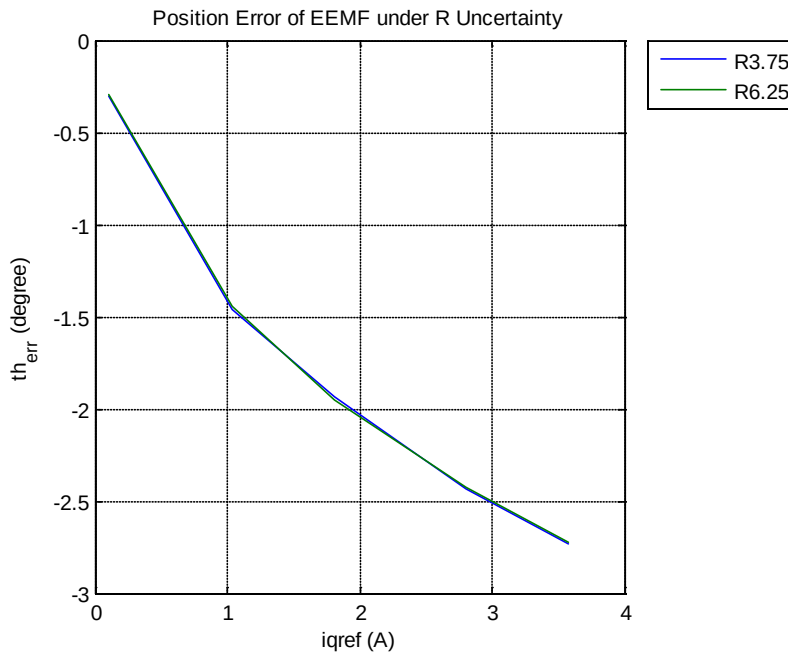


Figure 4.37 : Position error of EEMF under R uncertainty.

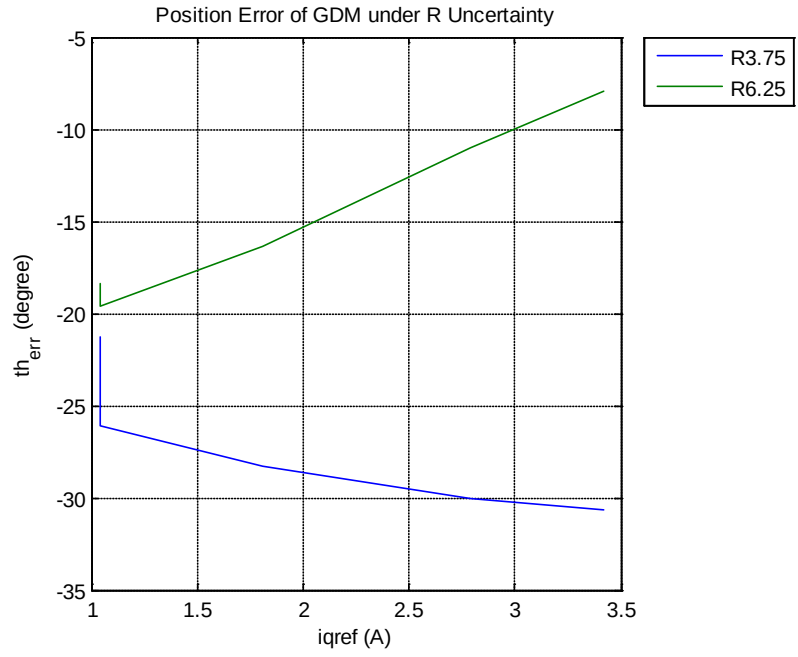


Figure 4.38 : Position error of GDM under R uncertainty.

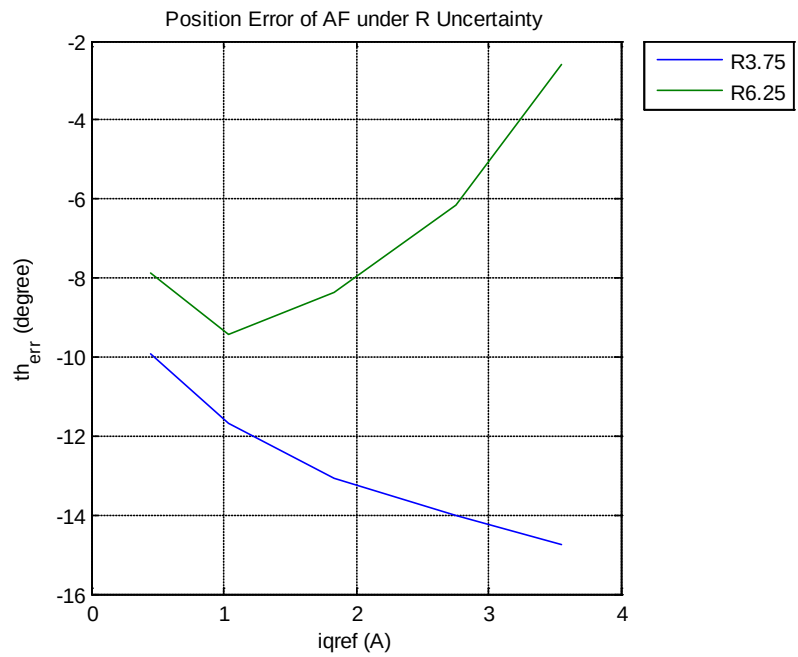


Figure 4.39 : Position error of AF under R uncertainty.

It can be seen that the AF and GDM observers were very sensitive to the changes in the resistance value, whereas the EEMF was not affected by uncertainties in this parameter. The effect of the resistance mismatch further increased with increasing

load as expected, since the proportion of the resistive voltage drop to the phase voltage amplitude also increases in this case.

4.7.2 Operation under flux uncertainty

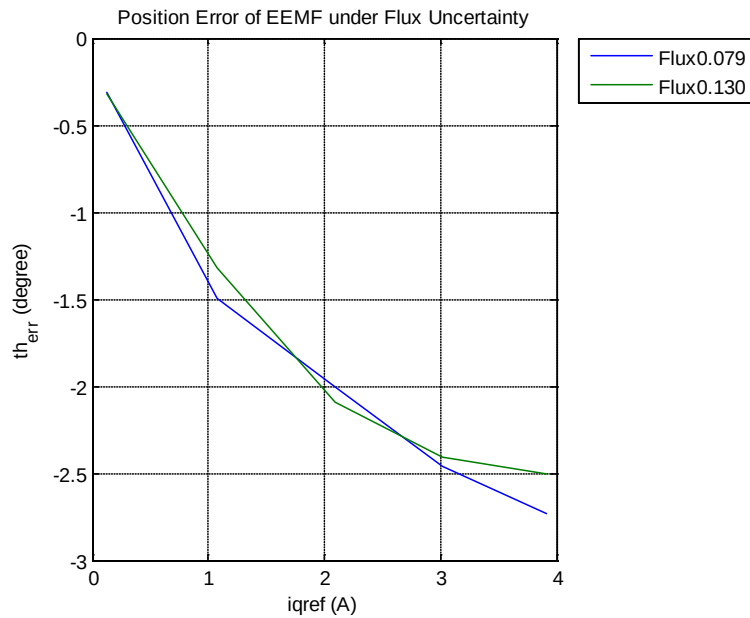


Figure 4.40 : Position error of EEMF under flux uncertainty.

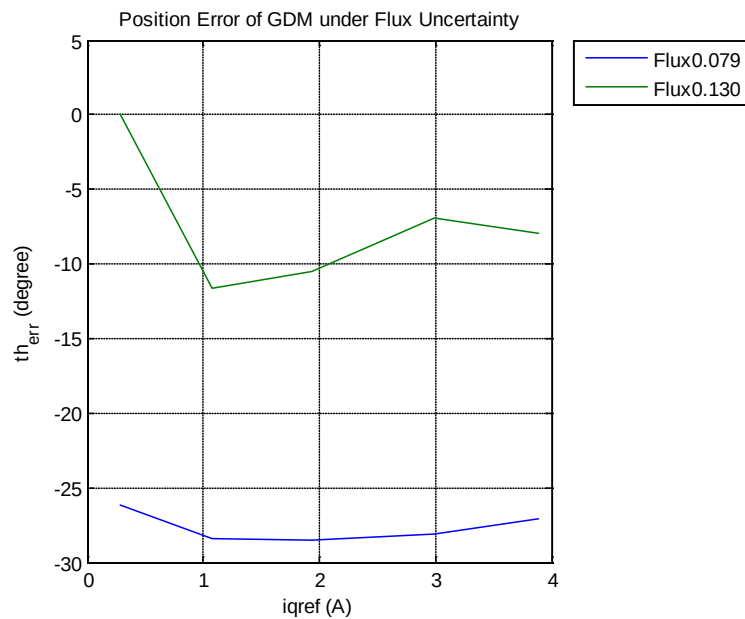


Figure 4.41 : Position error of GDM under flux uncertainty.

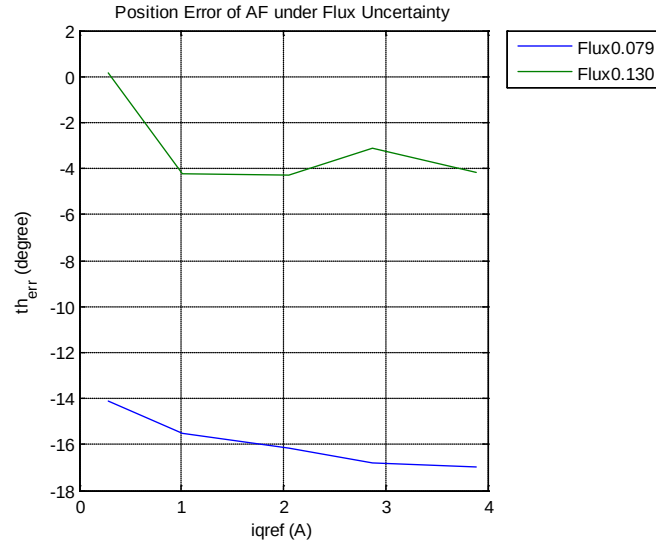


Figure 4.42 : Position error of AF under flux uncertainty.

The EEMF was found to be insensitive to flux changes in these experiments. This is due to the fact that flux did not appear in the equations of the EEMF observer. On the other hand the effect of using a wrong value for flux greatly affected the position estimation errors of the GDM and AF in which flux plays a central role in estimation equations. Unlike resistance parameters, the effect of flux variation did not change considerably with changing load because it is not multiplied by the current in the observer equations.

4.7.3 Operation under L_q uncertainty

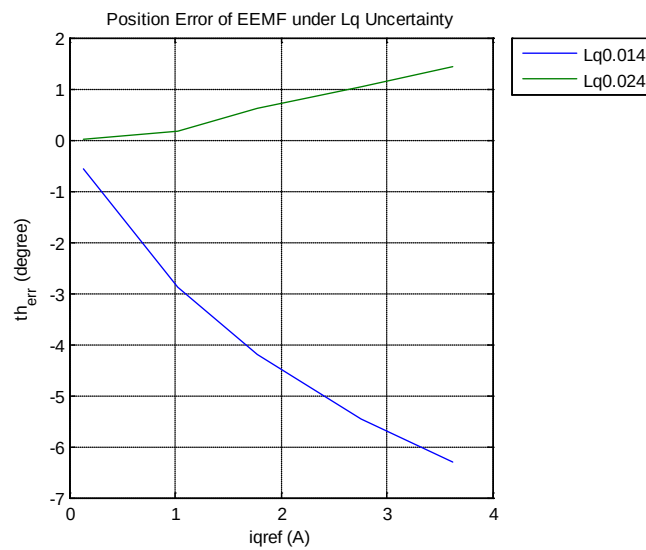


Figure 4.43 : Position error of EEMF under L_q uncertainty.

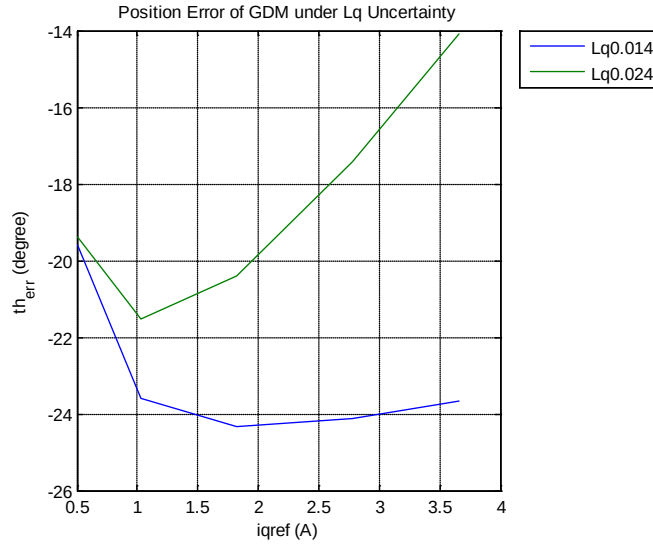


Figure 4.44 : Position error of GDM under L_q uncertainty.

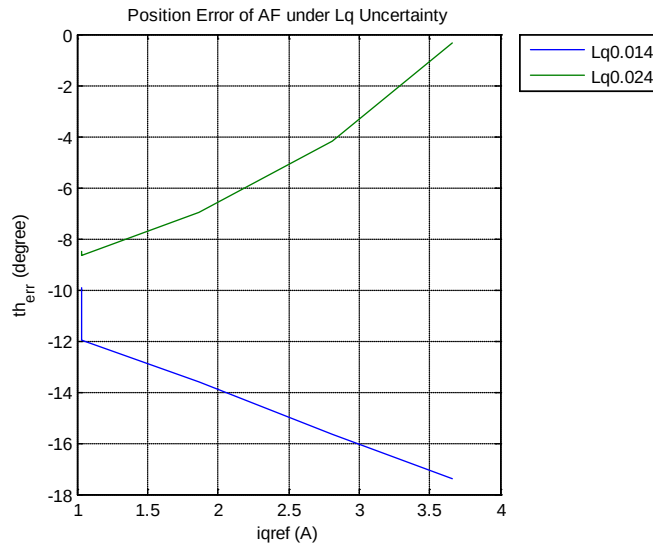


Figure 4.45 : Position error of AF under L_q uncertainty.

Under L_q variation the position estimation accuracy of all the three observers were affected greatly. However, it was observed that using a value of L_q had a positive effect on the estimation accuracy of all the observers, especially for the accuracy of the EEMF. The accuracy of the EEMF became independent of load changes when a higher than actual value was used for L_q . On the other hand, the AF and GDM were affected from L_q changes similarly.

4.7.4 Operation under L_d uncertainty

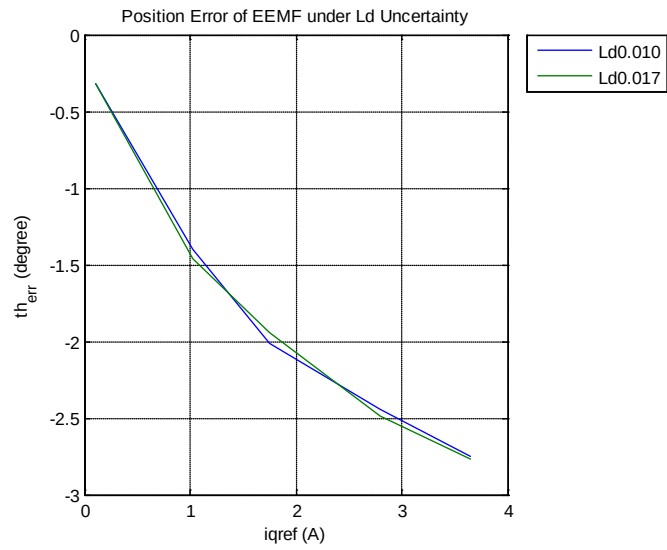


Figure 4.46 : Position error of EEMF under L_d uncertainty.

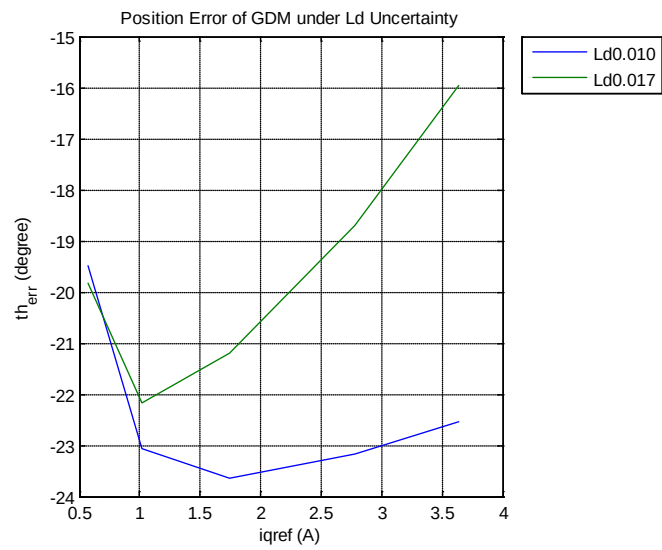


Figure 4.47 : Position error of GDM under L_d uncertainty.

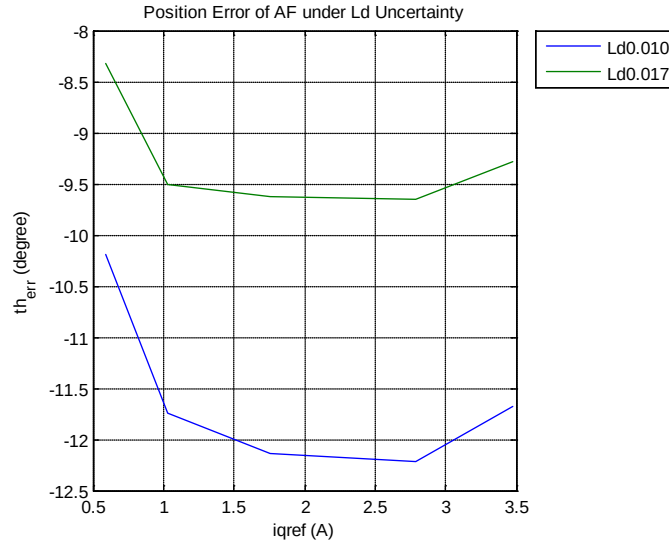


Figure 4.48 : Position error of AF under L_d uncertainty.

The EEMF was not affected considerably by variation of the L_d parameter. The AF was observed to have increased accuracy when a value higher than actual was used for the L_d . Similarly GDM was affected greatly by increases in the L_d under varying load conditions, but the affect of the decreases in the L_d did not produce considerable load dependent changes in the GDM position estimation accuracy.

4.8 Effects of Inverter Nonlinearity Compensation

As discussed in Section1.4.4, nonlinearity involved in the inverter voltage synthesis operation is an important practical aspect of the sensorless position estimation problem. If there is difference between the reference voltages used in position estimation calculations and the actual voltages applied to the stator phases, the performance of the position estimator is decreased, especially when the motor is operating in the low speed region. This difference is also known as the distortion voltage of the inverter.

In Section1.5.5, some of the inverter nonlinearity compensation methods in the literature were discussed. In this section a compensation strategy similar to the one used in [26] is presented. In [26], for each phase a compensation voltage is added to the reference voltages then duties of the PWM signals driving the gates of the power switches in the inverter were calculated using these compensated reference voltages. The polarity of the compensation voltage for a phase was same as the current polarity in that phase. Additionally, when phase current was zero, compensation voltage

was also zero and it was increased upto a maximum value in proportion with the increasing phase current (Fig.4.49).

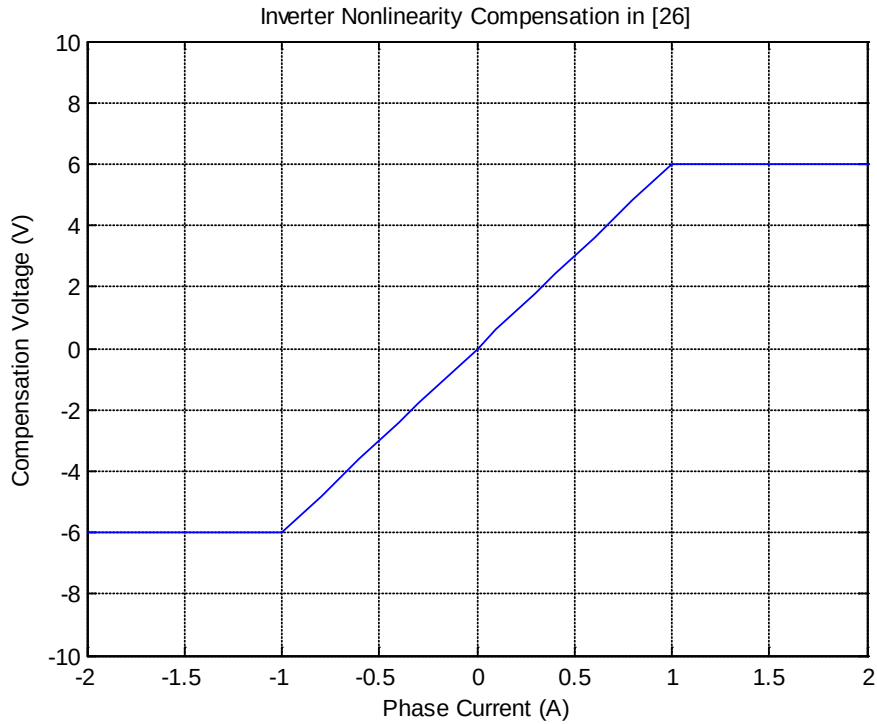


Figure 4.49 : Inverter nonlinearity compensation function in [26].

In this study, a function with the same shape was utilized, but instead of computing a compensation voltage for each phase, the function computed a compensation duty in terms of microseconds (Fig. 4.50). First, the reference voltages were sent to a Space Vector PWM (SVPWM) function. Then compensation duties for each phase were added to duty computed by SVPWM for that phase. Both of the discussed methods which are depicted in Fig.4.49 and Fig.4.50 were implemented in the drive software and the one shown in Fig.4.50 was found to yield better position estimation results in all the three methods that were being studied.

The effectiveness of this inverter nonlinearity compensation function was first tested with the EEMF estimation method. At 500rpm and 0.5Nm, the position estimation errors with and without the compensation function were measured and it was observed that the lack of compensation function decreased the stability of the estimation and caused position estimation to oscillate with a large amplitude (Fig. 4.51).

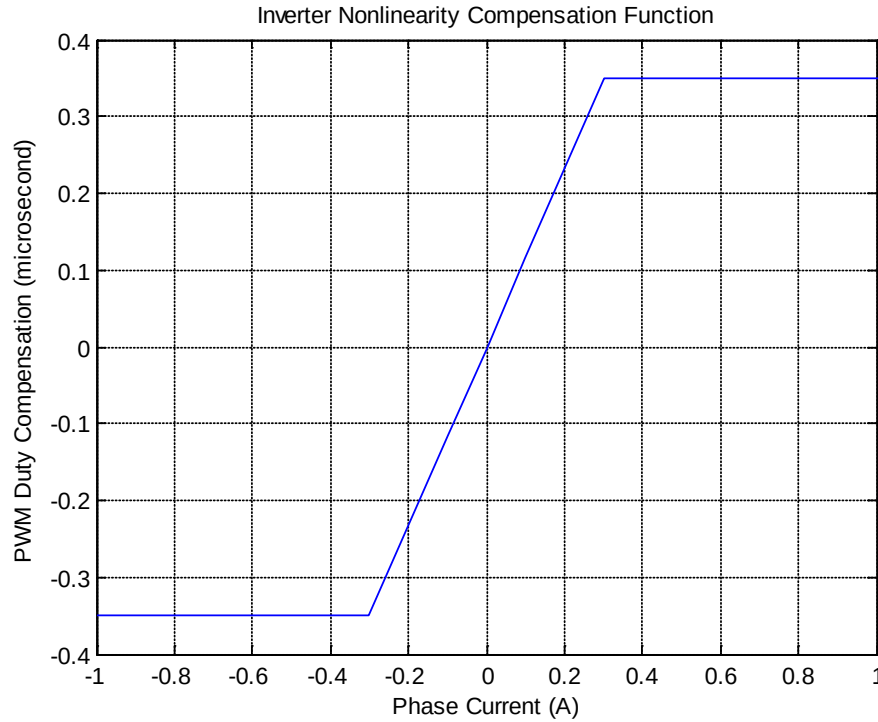


Figure 4.50 : Nonlinearity compensation function used in this study.

In fact, it was also observed that without inverter nonlinearity, compensation function EEMF became instable below 300rpm which resulted in motor tripping and sensorless operation failing, whereas, sensorless operation at 200 rpm was tested with compensation function operation and found to be successful.

Another effect of the inverter nonlinearity compensation function was demonstrated using AF and GDM methods. The position estimation errors at different speed were measured with and without compensation function, while the load was at 0.5Nm. The results showed that, the effect of compensation function is hard to notice above 500 rpm. On the other hand, when compensation function is not used below 500 rpm, the position estimation error became dependent on the load and decreasing the load caused the absolute position estimation error to increase. This performance decrease was more severe in the case of the GDM, in which the absolute position estimation error reached to 15°. When inverter nonlinearity compensation function was used, the curves of position estimation error with respect to speed remained almost flat in the whole 200-2000 rpm region.

In conclusion, the use of the inverter nonlinearity compensation function increased the performance of the position estimation error. For this reason, it was used in all measurements that are presented in the other parts of this study.

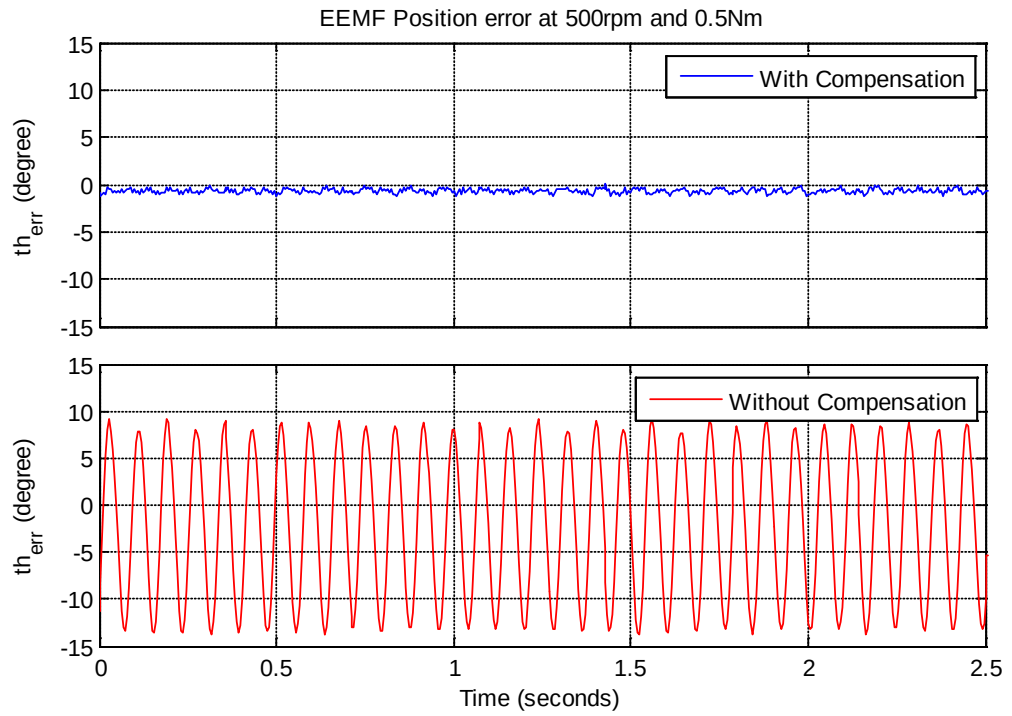


Figure 4.51 : Effect of compensation on EEMF (500 rpm, 0.5Nm).

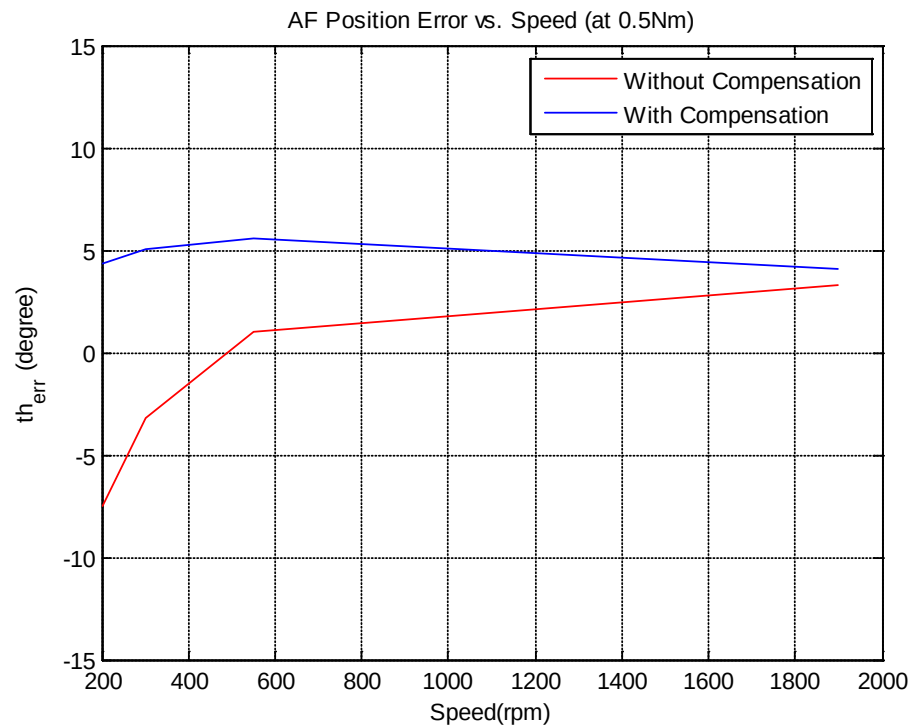


Figure 4.52 : Effect of nonlinearity compensation on AF (0.5 Nm).

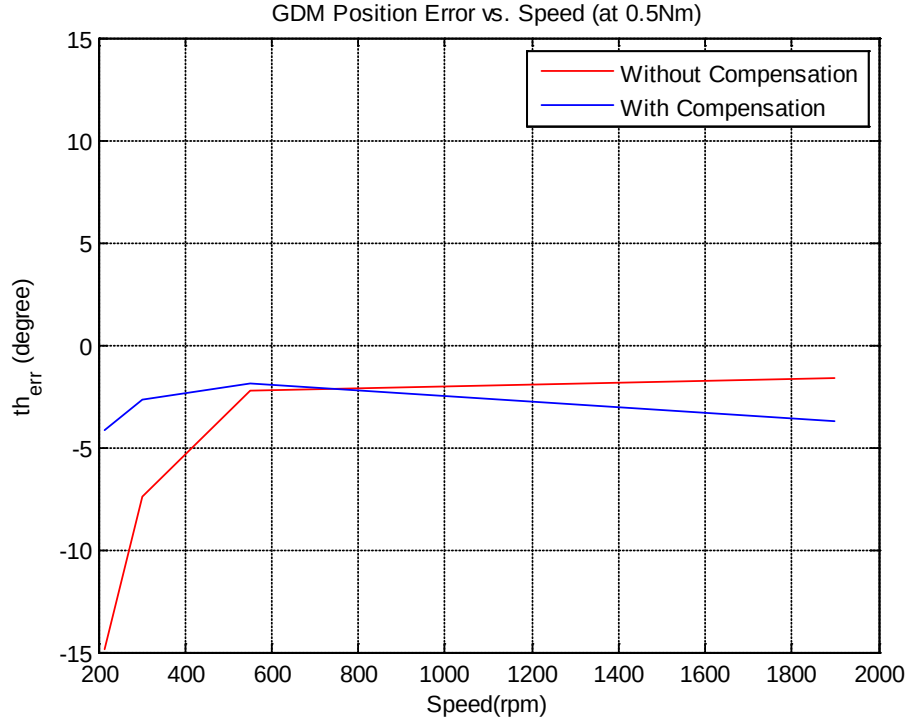


Figure 4.53 : Effect of nonlinearity compensation on GDM (0.5 Nm).

4.9 The Combined AF-EEMF Method

In the light of all the experiments and simulations performed up to this point, it was concluded that AF and EEMF had complementary characteristics. Two sets of findings had led to this conclusion.

The first set of findings were concerned with the behaviours of the observers under motor parameter sensitivities. The EEMF method's position estimation error has no sensitivity to the variations of permanent magnet flux linkage, because flux is not a part of the motor model equations used in the formulation of the EEMF method. Furthermore, according to the results obtained in the simulations and experiments, the EEMF method's sensitivity to uncertainties in resistance was observed to be very low as well. On the other hand, in the same studies, the AF method was found to be highly sensitive to both flux and resistance parameters especially at low speeds. Additionally, it was observed that the position estimation error of the AF method caused by resistance mismatch increased as the load increased, whereas the position estimation error of the AF method in the case of flux mismatch was independent of the load.

The second set of findings were related to the performances of the two estimation methods during transient operating conditions. During fast transients in the motor load, which were close to a step change in torque, the EEMF position estimation was found to oscillate considerably, especially at low speeds. Similarly, the position estimation error of the EEMF increased during rapid accelerations and decelerations. Performance decreases as such during transient operating conditions were not observed in the case of the AF method. The source of this difference between the two methods, was accounted for the differences between their formulations. During the derivation of the EEMF state and derivation of the state observer used to estimate the EEMF, certain terms containing derivatives of the speed and currents were assumed to be negligible and they were omitted, whereas no assumptions of this kind were made in the AF method formulation. When motor was operating under transient conditions the assumptions made in the case of the EEMF became less valid because of the increases of derivatives of the speed and currents. As a result, the EEMF performed poorly in comparison to the AF at transient operating conditions.

According to the two sets of findings described above, during transient operations it is advantageous to utilize the AF method and under flux and resistance uncertainty, it is preferable to use EEMF because of its robustness against variations in these parameters. Therefore, due to the complementary nature of these two methods, it is possible to construct a combined AF_EEMF method which is superior than both AF and EEMF methods.

In this section, a Combined AF-EEMF is proposed following the above conclusion. First the construction of this method is explained. Then the experimental results of its implementation are presented to evaluate the performance of the Combined AF-EEMF method. In the last part of this section, the resistance and flux estimation feature of the Combined AF-EEMF is demonstrated.

4.9.1 Construction of the combined AF-EEMF method

In the Combined AF-EEMF, both the AF and EEMF methods are executed simultaneously. The position and speed estimation of AF is used in sensorless control functions of the driver software. Thus, this method can also be interpreted as an augmented version of AF method. The position estimation of the EEMF is compared with that of AF in a real time manner, and the result of this comparison is

used to update the values of resistance and flux parameters used in the AF calculations. Through updating values of these parameters, the EEMF is used to increase the performance of the AF and the resulting Combined AF-EEMF observer's performance is expected to contain the strengths of both methods and perform better than the AF and the EEMF methods.

In this method, the position estimations of the AF and EEMF are compared at every PWM period, which is 50 microseconds in this application. The average of the difference of the position estimations between the AF and EEMF is calculated after accumulating the sum this difference for 100 milliseconds. The resulting averaged position estimation difference, $\Delta\hat{\theta}_{avg}$, forms the first part of a measurement, which is named as a comparison measurement in this algorithm. Simultaneously, the average of iq reference for the same 100 milliseconds time interval is calculated, $iqref_{avg}$, which forms the second part of the comparison measurement. This means that a comparison measurement result is obtained at every 0.1seconds. After two comparison measurements, the values of flux and resistance in the AF equations are updated according to the equations given below:

$$\begin{aligned} r(k) &= r(k-1) + \alpha_r \left(\frac{\Delta\hat{\theta}_{avg}^2 - \Delta\hat{\theta}_{avg}^1}{iqref_{avg}^2 - iqref_{avg}^1} \right) \\ \Psi^{pm}(k) &= \Psi^{pm}(k-1) + \alpha_{\Psi^{pm}} \left(\Delta\hat{\theta}_{avg}^2 + \Delta\hat{\theta}_{avg}^1 \right) \end{aligned} \quad (4.1)$$

where $(iqref_{avg}^1, \Delta\hat{\theta}_{avg}^1)$ and $(iqref_{avg}^2, \Delta\hat{\theta}_{avg}^2)$ are the two consecutive comparison measurements, and α_r and $\alpha_{\Psi^{pm}}$ are the smoothing coefficients of r and Ψ^{pm} respectively.

4.9.2 Steady state performance

In this section, position estimation errors obtained by the Combined AF-EEMF method was measured while the motor was operating under steady state conditions. The results for three different motor speeds are shown in Fig.4.54, Fig.4.55 and Fig.4.56. the results are almost identical to those obtained for the EEMF method. As the steady state position estimation performance of the EEMF was superior than that of the AF, the Combined AF-EEMF method was demonstrated to show the superior performance in comparison to its constituent methods as expected. This was

achieved through the change of Ψ^{pm} parameter in a manner to counter the difference between the AF and EEMF estimations as shown in Eqn(4.1).

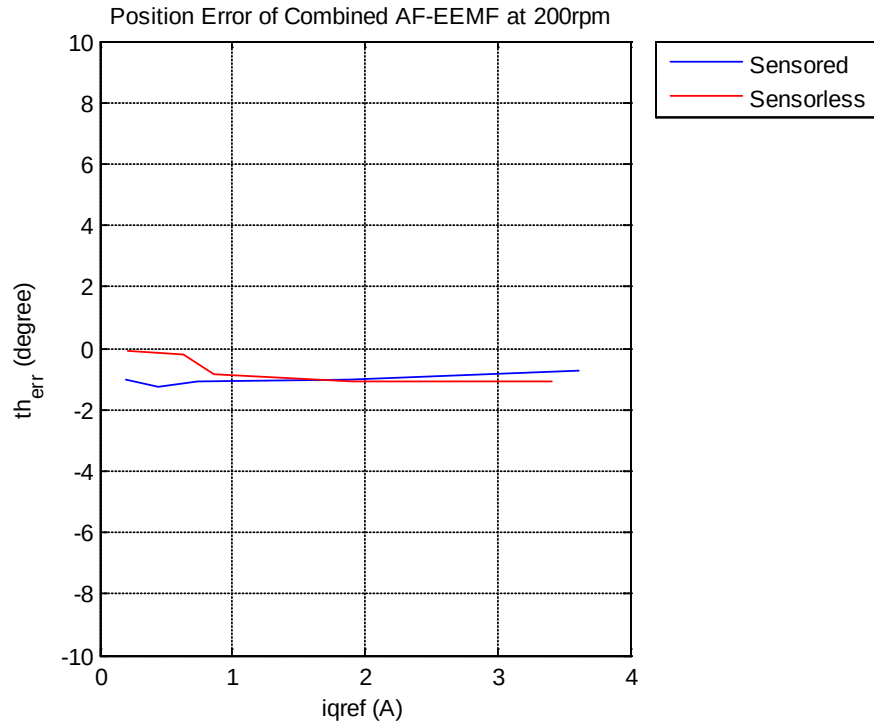


Figure 4.54 : Position error of combined AF-EEMF at 200 rpm.

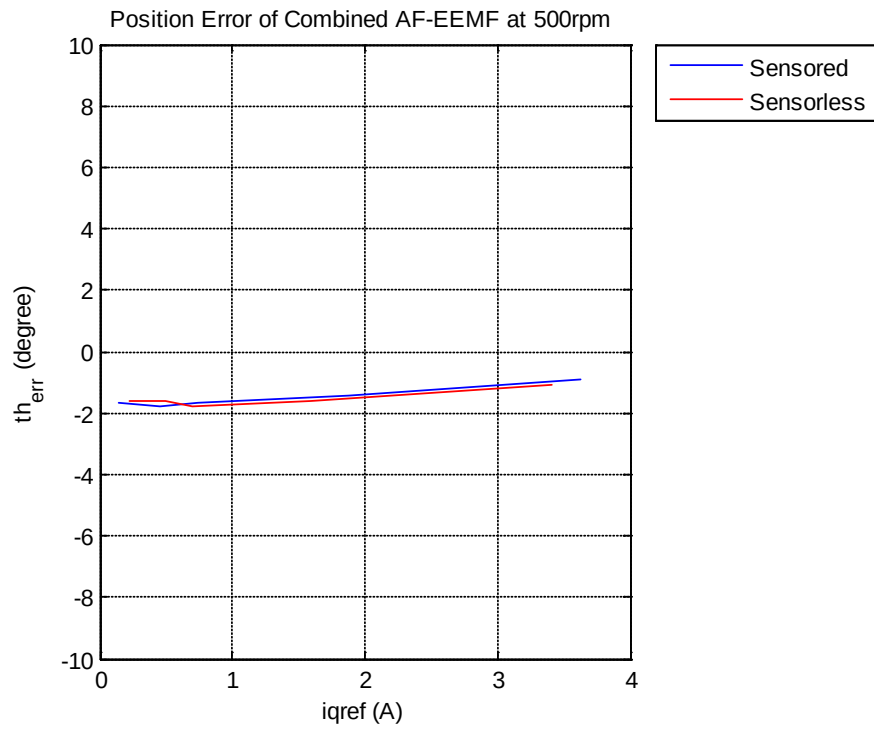


Figure 4.55 : Position error of combined AF-EEMF at 500 rpm.

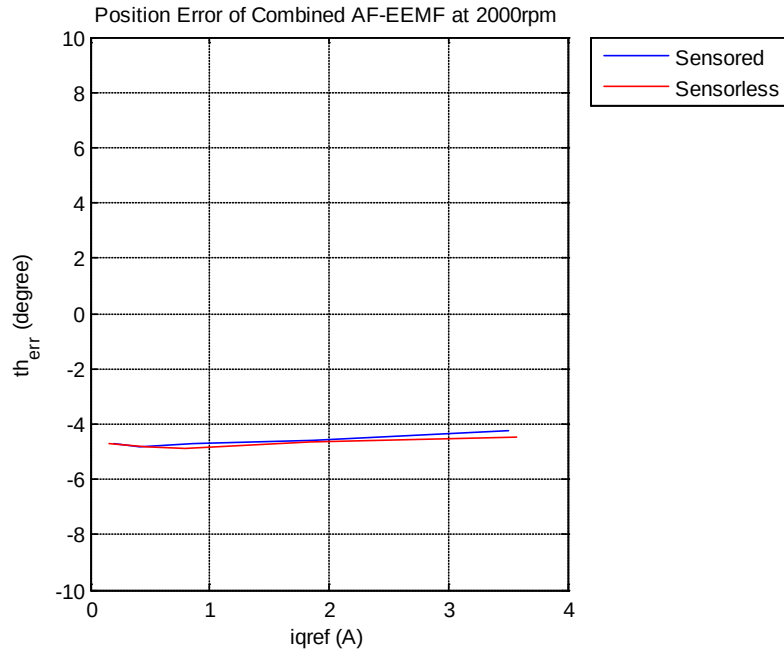


Figure 4.56 : Position error of combined AF-EEMF at 2000 rpm.

4.9.3 Transient torque performance

In this experiment, the behaviour of the Combined AF-EEMF during rapid torque increase, at 300 rpm was investigated.

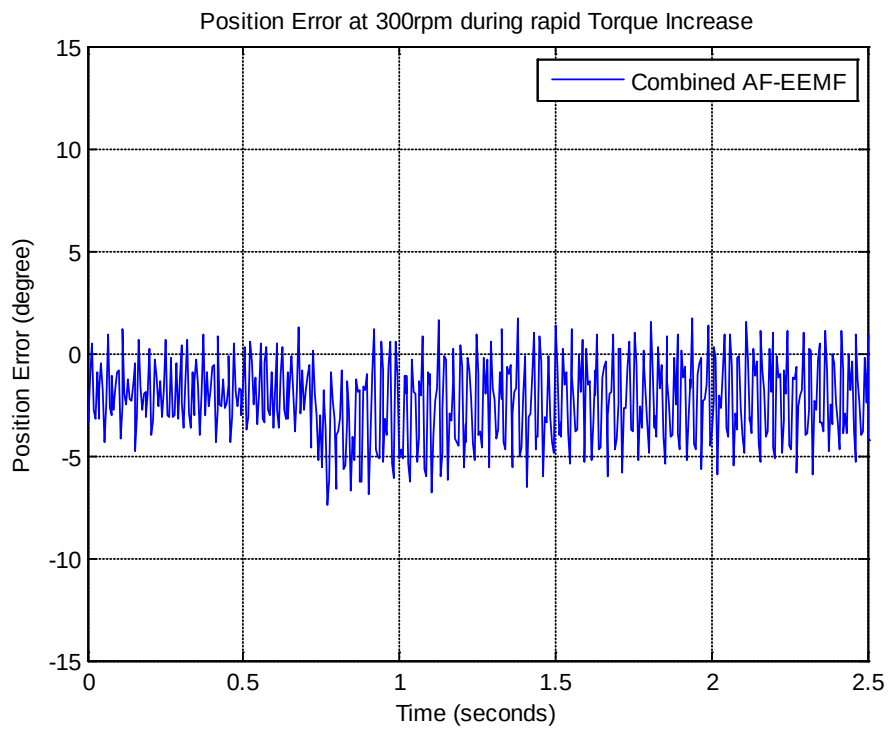


Figure 4.57 : Position error of combined AF-EEMF (0.3 - 1.8 Nm, 300 rpm).

When the results are compared to those of the AF and EEMF, it was observed that the better performance seen in the case of the AF is also retained in Combined AF-EEMF. This demonstrates another case in which the better one among the performances of AF and EEMF has been transferred to the Combined AF-EEMF method.

4.9.4 Transient speed performance

When behaviour of the AF-EEMF during rapid acceleration and deceleration conditions were investigated, the method was found to produce accurate position estimations, where error remains below 5° at all times. However, it should be mentioned that during fast speed transients, the link between the AF and EEMF had to be severed in order to prevent the EEMF's poor performance under these conditions to affect the AF estimation. This was accomplished by inferring the speed change rate from speed reference signal at the start of the acceleration or deceleration and decreasing the values of both α_r and $\alpha_{\psi^{pm}}$, accordingly. In the measurements below the value of these coefficients are set to zero at the beginning of the measurements.

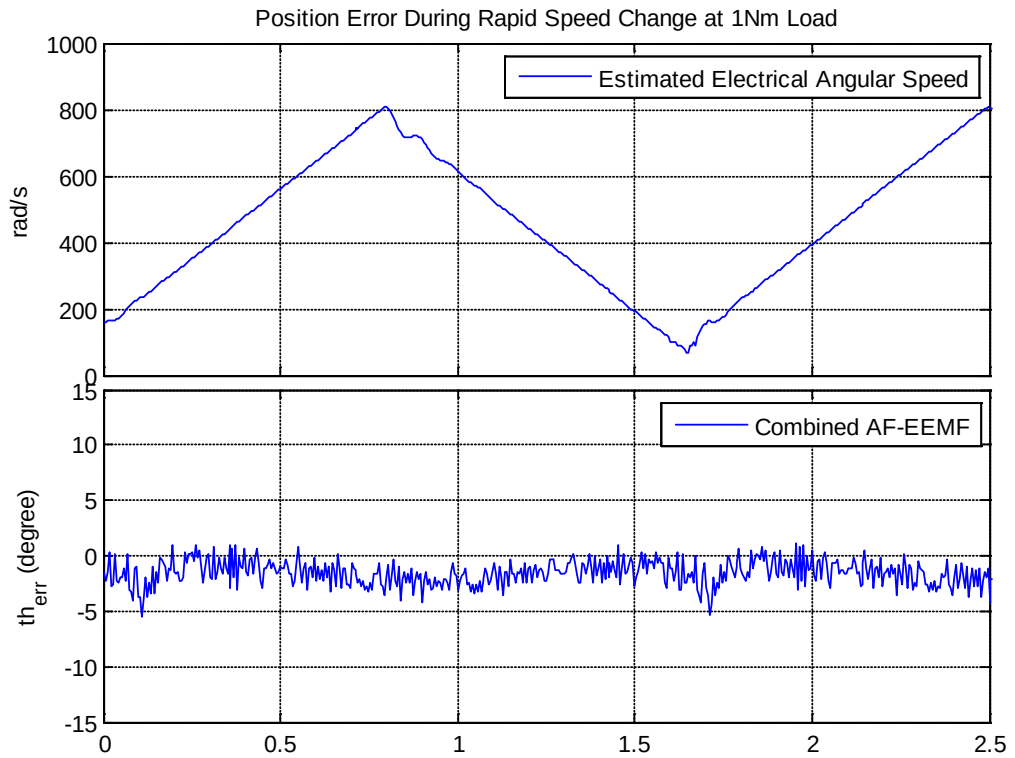


Figure 4.58 : Position error of combined AF-EEMF at speed change (1 Nm).

4.9.5 Resistance and flux estimation of the combined AF-EEMF method

As explained earlier, the Combined AF-EEMF takes advantage of EEMF's insensitivity to variations in r and Ψ^{pm} . Therefore, the Combined AF-EEMF also has the useful feature of correcting r and Ψ^{pm} , even if they are initially set to wrong values. To demonstrate this feature, both of the r and Ψ^{pm} were initially set to either 50% or 150% of their actual values and updated values of r and Ψ^{pm} were recorded until r was stabilized. The results are shown through figures 4.59 to 4.64. It was observed that the large part of the error in Ψ^{pm} was corrected within the first few seconds. This was due to the high sensitivity of the AF to Ψ^{pm} , which caused slight variations in this parameter to generate observable $\Delta \hat{\theta}_{avg}$. However, after these initial seconds, the correct value Ψ^{pm} was not completely reached, since there was still a significant error in r and the estimation of these two parameters were interdependent from each other.

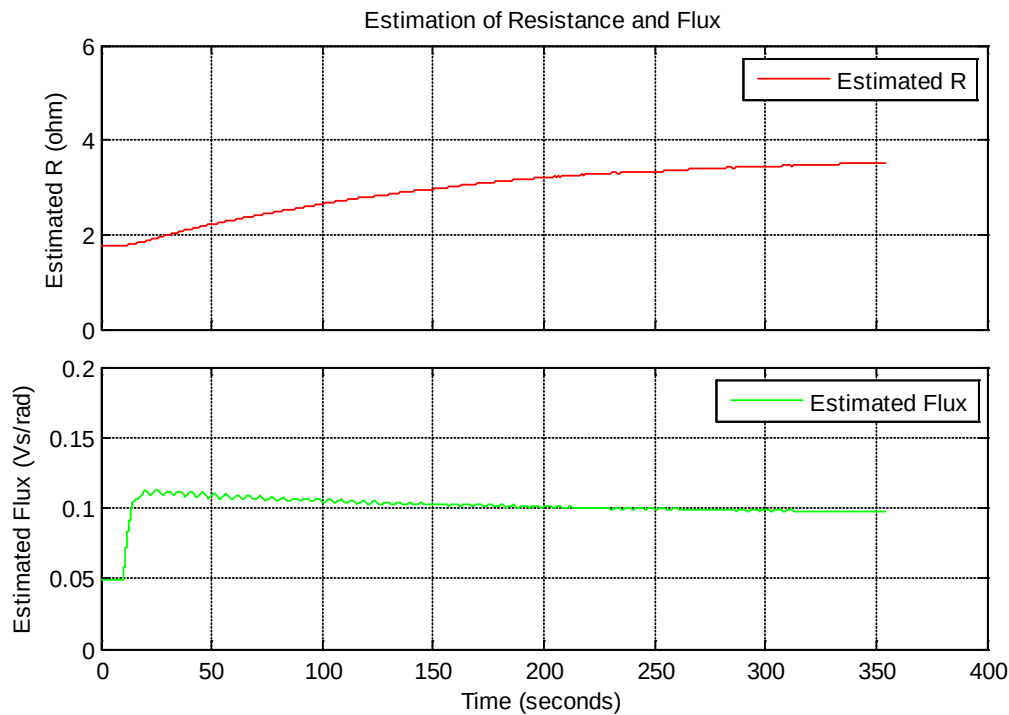


Figure 4.59 : Estimation of R and flux (at $t=0$, $R=1.75 \Omega$, flux= 0.05 Vs/rad).

Besides, if the load and speed remained unchanged the term $(iqref_{avg}^2 - iqref_{avg}^1)$ became very small.

As this term appears in a denominator of a division in Eqn4.1, it can cause large fluctuations in r which would result in instability of the whole system.

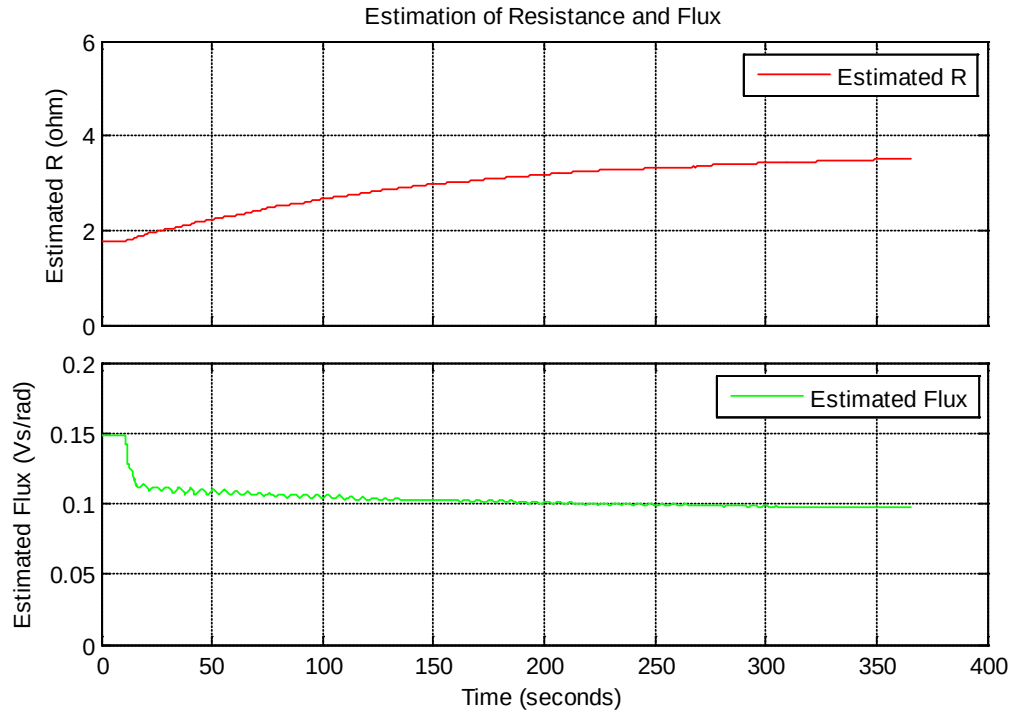


Figure 4.60 : Estimation of R and flux (at $t=0$, $R=1.75 \Omega$, flux= 0.15 Vs/rad).

To prevent this unstable condition, a minimum limit was introduced for the term $(iqref_{avg}^2 - iqref_{avg}^1)$, under which the last two comparison measurements were discarded and r was left unchanged. This means that, remaining at the same speed and load conditions prevented this method from correcting the error in the value of r . For this reason, the load was increased and decreased slowly in a periodic manner during the measurements performed in this section. As a result, both r and Ψ^{pm} were reset to their actual values in all the four cases presented through the figures 4.59 to 4.62.

In order to eliminate constant estimation errors, due to large errors in the r value, during operations in which both the load and speed remain unchanged for long periods of time, these rare operation scenerios are included in the algorithm of the Combined AF-EEMF method. When the algorithm detects such an operation scenerio, the position estimation of the EEMF is used in control functions instead that of the AF, in order to eliminate the constant position estimation error introduced by utilization of a wrong value for r .

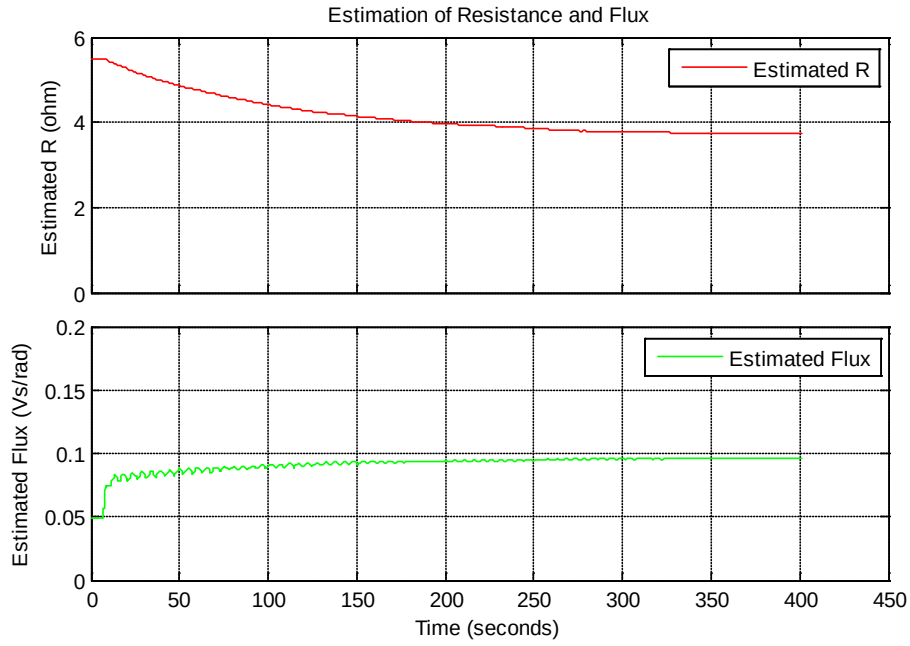


Figure 4.61 : Estimation of R and flux (at $t=0$, $R=5.5\Omega$, $\text{flux}=0.05$ Vs/rad).

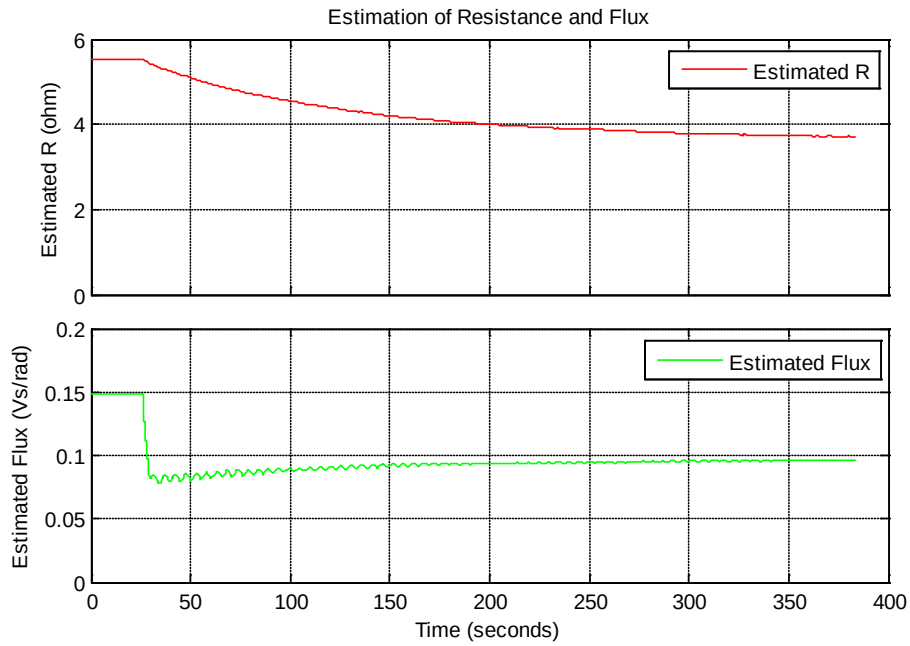


Figure 4.62 : Estimation of R and flux (at $t=0$, $R=5.5 \Omega$, $\text{flux}=0.15$ Vs/rad).

5. CONCLUSION

In this study, three methods for motion-sensorless position estimation of the IPMSM have been studied. The properties of the sensorless control problem and the challenges involved have been discussed before studying the details of the each estimation method.

The first methods under study is a speed adaptive EEMF observer implemented in the estimated rotor coordinates in which an angle tracking observer is used to produce the estimated position and speed signals from the estimated EEMF signals. The second method takes the advantage of the active flux (AF) concept that allows the construction of a hybrid stator flux observer, which links the voltage model of the IPMSM in the stator reference frame and its current model in the rotor reference frame. The result is an estimation method capable of estimating rotor position without using the estimated rotor speed in any of the calculations. The third method is based on an algebraic constraint between the magnitudes of the rotor and stator fluxes. Any inequality violating this algebraic constraint is interpreted as an error in estimated stator flux, and corrected by the help of an observer, which has the form of a standard descent method (GDM).

The formulations of the equations used in these three methods were presented along with the guidelines for selecting the gains appearing in these equations. Then, the behavior of each method was studied through simulations in the Simulink environment. The simulation results showed that, all three methods were capable of accurate estimations within 200 to 2000 rpm speed and 0 to 2Nm load ranges. It was observed that the position estimation error of the EEMF increased significantly during the rapid transients in the load. This behavior was caused by the speed adaptive nature of the EEMF method, and it was not observed in the cases of the AF and GDM, where increases in error during load transients were insignificant. In the second part of the simulations, the effects of parameter uncertainty on estimation accuracy were investigated. The results of the AF and GDM were almost identical to

each other. In these methods strong sensitivity to variation in values of the permanent magnet flux (or in short flux), stator resistance (R) and quadrature inductance L_q . On the other hand, the EEMF method was found to be sensitive only to the changes in L_q , which improved the robustness of this method. All three methods were found to be insensitive to variations in L_d . This was expected as simulations were performed with zero direct current reference.

In the second part of this study, the methods that had been previously examined, were implemented in an actual motor drive application. For this purpose, a comprehensive experiment setup, which provided precise control of the load torque, and detailed monitoring and logging of variables in the driver software was developed. The experiment platform also had the advantage of a 100MHz microcontroller with a dedicated floating point unit, which was capable of performing arithmetic operations with floating numbers in a single clock cycle. Using this microcontroller as the heart of the drive system, a complete FOC software library was developed. This provided a suitable platform on top of which any number of sensorless position estimation algorithms could be implemented.

The first set of experiments were aimed at determining the effects of the gain selection for each observer. For the EEMF, lower and upper bounds of g_1 and g_2 were determined and it was shown that accurate position estimations could be obtained as long as g_1 and g_2 were selected within these boundaries. It was also observed that changing g_1 and g_2 , within these boundaries did not introduce constant position estimation errors but instead affected the noise content of the position estimation error signal. Furthermore, experiments revealed that the selection of angle tracking observer gains, g_3 and g_4 , which is a part of the EEMF method to be more critical for position and speed estimation accuracies. A thorough gain selection procedure had been defined for g_3 and g_4 .

In terms of gain selection, the GDM was found to be the most advantageous method among the three because it had only one gain, g_1 , to tune. During the formulation of the GDM it was stated that the main effect of g_1 was to determine the convergence rate of the estimator in GDM. The simulations and experiments confirmed this fact and it was observed that selecting a too low value for g_1 caused unwanted oscillations in the position estimation error signal. On the other hand, it was found that selecting a too high g_1 causes constant position estimation errors. In light of these findings, it

was determined that a speed dependent g_1 value, which increased with increasing the speed was most suited for the GDM.

In the case of the AF, there were two gains, g_1 and g_2 , to be selected. The experiments showed that the selection of g_1 and g_2 affected the steady state position estimation error at lower speeds. It was also found that different g_1 and g_2 values effect the change in the position estimation error as load increased. These two effects were considered when appropriate g_1 and g_2 values were determined.

The next two sets of experiments aimed to comparing the estimation accuracy at different steady state and transient operation conditions, and robustness under parameter variation. The results of the experiments showed that all three methods were capable of full sensorless operation, in which the estimated position was used in the transformations appearing in the control functions of the driver software, in the 200 to 2000rpm speed range with loads varying from 0 to 2Nm. It was found that the EEMF and GDM had exceptionally lower position estimation errors throughout the covered operating region, the EEMF being the superior among the two. The position estimation errors of the EEMF and GDM were found to show a little change at different load levels. On the other hand, the position error of the AF was determined to have a stronger dependency on the load. At lower speeds, the load dependency of the AF estimation accuracy increased. The largest steady state position estimation error in this study was observed to be around 15° , which occurred when the AF was used at 200 rpm under the no load condition.

When the case of operation during rapidly changing speed were considered, it was found that the EEMF resulted in significant position estimation errors up to 10° . The error of EEMF during load transients were found to be significantly higher than its GDM and AF counter parts, as well. This inaccuracy in the EEMF position estimations were determined to be avoidable by the use of a faster speed controller, whose reaction to changes in load was considerably quicker. However, even if such a controller was used, large errors in the EEMF position estimation, during rapid accelerations and decelerations were still observed. The GDM and AF had shown considerably better results during the experiments under transient conditions than the EEMF since the estimated speed was not used in stator flux estimators of these methods. Similar results were observed in the simulations as well.

In the next set of experiments, parameter sensitivity of the position estimation error of the three methods were tested. The observed results were similar to the simulation results, in which all the three methods showed high sensitivity to the changes in L_q , whereas only the GDM and AF showed sensitivity to variations in permanent magnet flux linkage and stator phase resistance. The EEMF was found to be insensitive to both of these parameters, which is an important and useful feature of EEMF. Furthermore, in both the AF and GDM, the position estimation error introduced due to a mismatch between actual and estimated resistance increased as the load increased. On the other hand, the position estimation error caused by the flux variation remained constant even if the load was changed.

In experimental studies, the effect of the inverter nonlinearity compensation function on position estimation accuracy was also investigated. The results showed that when compensation function was turned on, the EEMF could operate accurately at 200 rpm. In contrast, when it was turned off, the EEMF estimation became unstable under 300 rpm and the sensorless operation failed. In the AF and GDM, using the inverter nonlinearity compensation function had resulted in reduced load dependency of position estimation error at speeds below 500 rpm. Thus, the effectiveness of the compensation function was confirmed for all of the three sensorless methods.

The comparisons up to this point revealed some complementary characteristics between the EEMF and AF. The EEMF produced slightly more accurate estimations at steady state operating conditions than the AF, whereas the AF showed considerably superior performance during transient operating conditions. The EEMF also had the important robustness property that allowed its position estimation to remain accurate in the face of variations in the stator phase resistance and permanent magnet flux linkage. The AF displayed high sensitivity to variations in both of these parameters, however, as explained above the nature of these sensitivities showed difference concerning their relation to changing load conditions.

The complementary characteristics of the AF and EEMF revealed in this study led to the proposal of a Combined AF-EEMF method, which incorporated the strengths of both methods. Through experiments, this new method was demonstrated to produce accurate position estimations both in steady state and transient operating conditions. Furthermore, the Combined AF-EEMF method was found to be insensitive to variations in the stator phase resistance and permanent magnet flux linkage, as it

contained the EEMF method. The Combined AF-EEMF was also shown to identify these two parameters accurately under most operating conditions.

In conclusion, this study features several benefits in regarding the investigation of the sensorless control problem and the methods proposed as a solution to it. First of all, a motor drive platform, which can be used as a tool to study and compare different sensorless control techniques has been obtained. As an initial use of this tool, the three of the sensorless methods commonly used for sensorless control of the IPMSMs have been studied in detail to reveal their strengths and weaknesses. This type of study also has the use of providing a base on which new sensorless methods can be designed, via combining the parts of the compared methods in an innovative manner as proposed in the Combined AF-EEMF method. In this regard, this tool can be used as rapid implementation and testing platform to compare a larger number of sensorless methods.

During these studies, considerable differences between results of the simulations and experiments have also been observed. These differences were most likely caused by the nonlinearities in the inverter, which remained uncompensated even in the presence of a compensation function, and the unmodeled characteristics of the IPMSMs such as the spatial harmonics of rotor permanent magnet flux linkage and the stator phase inductances, and saturation of the stator core, which are left out of the scope of this study. Future studies should aim to take the effects of these phenomena on sensorless position estimation problem into account with the help of software and hardware tools developed in this study.

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