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Research Paper

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Controlling the Speed of DC Motor by Applying a Fractional Order Sliding Mode Controller

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ABSTRACT

A vigorous Fractional Order Sliding Mode Controller (FOSMC) to be used to control the speed of the DC motor. It is a computer software and simulation-based research. DC motors are one of the most versatile electric machines that widely used in industries and home automations. Many control techniques have been proposed to effectively control the position and speed of DC motor for industrial as well domestic purpose. Research is directed towards the development of state-of-the-art approach for the fractional order controller as the addition of latest classic philosophy. Fractional order control is an emerging area of research, this research presents novel fractional order sliding mode controller to control the speed of the direct current motor. Parametric uncertainties are presented in this approach and un-modelled dynamics. The fractional-sliding order is defined in accordance with the fractional calculus, and stability of system is guaranteed using Lyapunov theorem. Simulations to be carried out in MATLAB Simulink and simulation results are verified through experiments. Desirable and superior performance of proposed control scheme to be observed from simulation and experimental results.

Keywords: DC motor, Sliding mode control (SMC), Fractional order controller, Robust control, Open loop response, Close response

Due to large, easy, and continuous control features, D.C. motors have been widely used in many industrial applications, such as electric vehicles, electric cranes, and home appliances. They are used to transform mechanical power into electrical power. The DC engine was discovered at the end of the 18th century. High starting torque, high power, excellent performance, controllability, continuous control characteristics and fine speed control are the advantages of direct current motors. In various industrial applications, such as electric vehicles, mixers, juicers, robots, cutting machines, and home appliances, D.C. motors are commonly used. Compared to A.C. motors,

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which are costly due to their complicated Design, D.C. motors are not costly. P.I.D. Controls are typically used to control the speed of D.C. motors[1]. To get optimal P.I.D. Parameter values, some practice is required, however, inapplications where the device is complicated and non-linearities induced by disruptions, the P.I.D. The controller does not function well[2].

The P.I.D. The controller is negatively impacted by external disturbances, load changes and parameter changes depending on its position, so its use in applications such as speed control is limited [3]. Therefore, P.I.D. Efficiency is not robust, and various strategies are used to resolve these effects, such as adaptive and self-tuning P.I.D.s[4]. A controller that provides quick response, minimum steady-state error, and minimum settling time with a minimum overshoot to maintain the D.C. motor speed at the desired value are very important in industrial applications[5]. The adaptive fractional-order S.M.C. for dc motor speed control is formulated in this article. The mathematical model is exposed to nonlinearities, parametric uncertainties, and disturbances of the oscillatory form.

For the indicated dc motor, a non-linear fractional-order adaptive S.M.C. is derived[6]. The adaptive law and stability proof was derived using the Lyapunov theorem of fractional order. Finally, compared to P.I.D. and integer S.M.C., the proposed adaptive fractional-order S.M.C.'s efficiency is thoroughly investigated. Compared with other classical techniques, the numerical results would show that the proposed algorithm is efficient and stable for dc motor speed control[7]. The engine parameters are modified due to changes in temperature, current, voltage and load, etc., or other operating conditions. As the load on the D.C. motor shifts, it causes motor speed fluctuations[8]. For speed control of the D.C. motor, a new controller is proposed. The DC motor is used in many industries for many applications. Where fine speed control is needed, it can be used[9]. Efficient techniques are necessary for implementing and designing controllers. The device is subject to disruptions and uncertainties, so a simple design technique is required to meet the performance requirements[11].

The P.W.M. generated by the controller controls the motor's input power and controls the motor's speed. In the past, several D.C. motor speed control controllers were used[12]. But their performance was not outstanding. But a high-performance controller is a fractional controller. In FOSMC, fractional integrals and derivatives are used. It has little time to settle, lessovershooting, and fast transient response. Using the Lyapunov theorem, the stability of the controller is proven. For the speed regulation of a PMSM, this article applies the fractional calculus to the sliding surface architecture and proposes a robust fractional-order sliding mode controller (FOSMC).

Mechatronic systems are governed by several engineering fields, spanning mechanical, electrical, pneumatic, thermal disciplines, containing different technical components. Thus, these types of systems are typically characterized by significant nonlinearities and requirements of tight efficiency. Because of these characteristics, the key challenge in mechatronics design and analysis Accurate modelling consists of structures. Whenever it is difficult to obtain such a model accurately, the controller needs to be adequately configured to compensate for modelling uncertainties and maintain the closed loop's output at different operating points. Numerous control algorithms have been suggested to deal with the mechatronic systems' non-linear dynamics. For mechatronic linear structures,Due to its simple structure and robustness, the proportional-integral-derivative (P.I.D.) controller is often used[13]. Fractional order (F.O.) control techniques are another way of dealing with the complexities of mechatronic systems. In all mechatronic domains, control of D.C. motors is one of the most common applications. Due to the broad range of applications that involve D.C. motor control, several researchers have been interested in controlling D.C. motors.

The use of D.C. motors of various kinds. The controllers designed for these D.C. motors range from simple conventional P.I.D.s to advanced control algorithms, which have gained growing popularity for fractional-order control. In modelling and control applications, fractional calculus has been used relatively recently. Due to the extra tuning parameters, the appeal of the fractional-order P.I.D. Controllers lie in their ability to improve the closed-loop efficiency and robustness of the closed-loop framework.

As opposed to the traditional controller, the order of differentiation and integration can be used as additional tuning parameters with fractional-order controllers. Thus, more requirements can be satisfied simultaneously, including robustness to plant uncertainties, such as frequent shifts in gain and time. Using optimization routines to generate the final solutions, the fractional-order controllers' frequency domain tuning is preferred in general. Frequently, the success parameters are Stated in terms of crossover frequency of gain, crossover frequency of phase, phase margin, the margin of gain, and robustness to variations of open-loop gain.

A fractional calculus-based control strategy is proposed in this paper for speed control of a D.C. motor with changes in load. The paper's importance to the research field is the simplicity of the methodology, but it offers a robust controller that can meet the performance criteria for significant load changes. The robustness and efficiency were contrasted to an integer-order P.I. controller[14-15]. For the tests carried out, a modification in the motor loading unit is considered to test the controls' robustness. The system's gain and time constant are also changed due to the change in the brake unit. Simulations and real-time experiments examine the outputs of both the classical integer-order approach and the fractional-order approach. To explain basic time domain and frequency domain concepts, the control design method and the application are kept simple, but efficient. In contrast with the classical one, the experimental findings showed better performances of the fractional method.

Literature Review

Modern control theories have made great strides in the past decades. Control techniques have been developed, including optimal control, H H-2 control, fuzzy control, neural network control, predictive control, etc.

Considerably. However, in many industrial applications, the proportional-integral-derivative (P.I.D.) control technique has still been commonly used, such as process control, motor drives, flight control, etc. Nowadays, more and more than 90 per cent of industry control loops are P.I.D. Control. This is mainly because of the P.I.D. The controller has robust performance to meet the global shift in the industry's process, a straightforward framework that engineers can easily understand, and simplicity of design and implementation. Recently, there is a growing interest in improving the efficiency of P.I.D. Controller by using the principle of fractional calculus, where derivative and derivative orders are non-integer integrals.

Fractional calculus is non-local, allowing long-memory to be stressed mathematically. The fractional order P.I.D. controller suggested in[16] is a generalization of the fractional order P.I.D. controller. Uh. Calculus. Five parameters, i.e., the proportional gain, the integrating gain, the derivative gain, the integrating order, and the derivative order, are defined by a P.I. D controller. Multiple applications in irrigation channel control[17], temperature tracking[18], D.C. motor motion control[19], boost converter control[20], hypersonic flight vehicle control[21], servo press control system[22] have been found by P.I. D controllers in recent years. The above research results show that the P.I. D controller is better than traditional P.I.D. controllers in terms of efficiency and robustness. Although this is so, the P.I. D controller's parameter tuning is an essential and crucial problem. The P.I. D controller has two extra parameters compared to traditional P.I.D. controllers. On the one hand, it helps people to build P.I.D. controllers with more degrees of freedom, and on the other hand, it means that it is more complicated inP.I.D. controller synthesis. Many methods have been reported in the literature for developing P.I.D. controllers. It is possible to divide these approaches into two classes: empirical methods and heuristic methods. In the analytical sense, by minimizing a non-linear objective function based on the requirements imposed by the designers, the parameters of the P.I.D. the controller is tuned. A new theoretical approach for designing the P.I.D. controller by extending the input and output of the control loop signal and reference model over a piecewise orthogonal function in[23].

In this way, the fractional differential calculus is replaced by the generalized differentiation operating matrices associated with these bases. Then the tuning of the controller is built in the manner of a matrix manipulator. The tuning of Integrator and Derivation orders, however, is not considered. In[24], tuning rules were built based on a process model of first order plus time by minimizing the integrated absolute error with a maximum sensitivity constraint. Ziegler-Nichols were given in[25] as tuning rules for the P.I.D. controller. Evolutionary algorithms, including Genetic Algorithm (G.A.), Optimization of Particle Swarm (PSO), Electromagnetism, Differential Algorithm (E.M.).

The architecture of the P.I.D., the controller, is also used for evolution. Cao and Meng and Xue (2009) adopted genetic algorithms to design the P.I.D. controller by recasting the issue to an optimization problem. Particle swarm optimization was used in[26] to design the P.I.D. controller for an Automatic Voltage Regulator (A.V.R.) system by minimizing an objective function consisting of overshoot, increasing time, settling time, steady-state error, absolute error integral (I.A.E.), square input integral, gain margin, and phase margin. In[27] for the P.I.D. controller architecture and improved electromagnetism-like genetic algorithm (IEMGA) algorithm were suggested by minimizing the integrated-square-error technique (I.S.E.). A PID based on the root locus method in[28] uses enhanced differential evolution.

In[29], P.I.D. controller tuning is recast as a loop shaping problem of quantitative feedback theory (QFT). The optimization target is the high-frequency gain of the nominal loop subject to the specifications' limits. Tuning parameters of the PI Dλ μ controller by considering five step and gain margin specification criteria and sensitivity feature constraints in[30]. In terms of heuristic methods, numerous scholars have discussed rule-based methods and evolutionary algorithm-based methods. In[31], the first order and first-order plus dead time (FOPDT) Bode envelopes with the parametric uncertainty structure are successfully combined with five design parameters that have already been used in[30] to achieve a robust PI Dλ μ controller.

Variable Structure System (V.S.S.)

A closed-loop proportional (P) controller and proportional integral derivative (P.I.D.) controller[32],[35] are the most common control design techniques used by industry to control a Direct Current (D.C.) motor. Although the Design of analogue controllers is very well known, digital technology advances have led to the implementation of lightweight, inexpensive micro-controllers/microcomputers with sufficient capacity to carry out advanced control actions [36]. The complexity and versatility of microcomputer software make it simple to design and experiment with control systems engineering. However, when contrasted with analogue controllers, the microcomputer speed, interface, and memory allocation limitations should not be overlooked. P, and P.I.D. controllers, especially when the non-linearity is very high, or the uncertainty range is wide, are not perfectly capable of stabilizing the system[37]. Almost perfect disturbance rejection or control output is required in many practical problems. To obtain this output, Sliding Mode Variable Structure Controls (SMVSC) can, therefore be applied to the system.

In the last seventy years, the theory of the variable structure system (V.S.S.) was established in the former Soviet Union. The principle of variable structure control/system (VSC/S) has been developed and applied based on that theory to control a wide range of plants producing missiles, chemicals, etc. The reader can refer to Utkin's survey paper[38] and the publications that followed[39-42]. Researchers such as Young[43], Zinober[44], Spurgeon[45], White, Slotine and Sastry[46] and Edwards[47] have also expanded the use of V.S.S. in the U.S.A. and Europe, covering model-following, ambiguity, and non-linear systems. VSC's basic principle is for the system to change structure at certain times to combine the beneficial properties of the system.

Each of the[48-51] structures. This versatility allows the use of structures' "good" dynamic properties that cannot be used for a long continuous period. Operating such a device in what is referred to as asliding mode makes it insensitive to external loads, external disturbances (e.g., noise) and variations of parameters applied[52]. VSC has nice transient efficiency and rapid response[53-56].

A second order system is shown in the following equations to understand the concept of V.S.S.

$$x-1 = x-2 \quad \ldots(1)$$

$$x-2 = -a-2\,x-2-a_1\,x-1+bu \quad \ldots(2)$$

$x-1$ and $x-2$ are the state variables. The control law is given in Eq (3)

$$u = \psi x-1 \quad \ldots(3)$$

Sliding Mode Control (S.M.C.)

Sliding mode control (S.M.C.) is a non-linear control technique with exceptional precision, robustness, and simple tuning and implementation characteristics. SMS systems are designed to drive system states onto a specific surface, called a sliding surface, in the state space. If the sliding surface is reached, regulation of the sliding mode holds the states on the sliding surface nearby. Therefore, the control of the sliding mode is a two-part controller style. The first component includes the construction of a sliding surface such that design requirements are met by the sliding motion. The second is concerned with the selection of a control law that will make the device state desirable to the switching surface.

The sliding mode control has two key benefits. The first is that the system's dynamic actions can be customized by the sliding function's individual preference. Secondly, some basic uncertainties become insensitive to the closed loop answer. This theory refers to small model parameter uncertainties, disturbances, and nonlinearity. S.M.C. allows for the monitoring, from a functional point of view, of non-linear processes subject to external disturbances and extreme model uncertainties[57-59].

Unknown parameters of the plant or, more generally, instability of the plant and expectations where system dynamics are intentionally described by simplifications, such as the use of a linear friction model, lead to the imprecision of the model[60]. The model inaccuracies, here used as a synonym of imprecision, are categorized into two main categories as organized and unstructured uncertainties by control engineering. The first implies inaccuracies in the model, and the second implies inaccuracies in the order of the structure (i.e., underestimated system order). Inaccuracies in modelling can adversely affect non-linear control systems. The best examples of closely coupled highly non-linear, time-varying dynamical systems[61] are robotic manipulators. In addition to structure non-linear friction, gearing compliance, sensor noise, external disturbances, and the high-frequency aspect of dynamics, these characteristics make the motion control of rigid-link manipulators a complex issue. From these organized and unstructured uncertainties, robotic manipulators suffer too much. The effect of having to deal with different uncertainties in their dynamics and the need to control the different instruments and, thus, the variance of dynamic parameters during operation makes it difficult for robots to implement a suitable mathematical model for operation.

Employing model-based techniques for regulation. Since the 1980s, the principle of conventional sliding mode control (S.M.C.) as a basic solid non-linear control device has been successfully applied to robotic manipulators. The advantages of the S.M.C. properties, such as its robustness against disturbances and parameter variation, and its fast dynamic response, have been greatly used in these studies. Two major techniques, such as robust,Regulation and adaptive control should cope with the complexity of modelling[62]. Adaptive control is effective in addressing organized and unstructured uncertainties and can sustain a consistently good performance over a limited range. Due to its simplicity and robustness against parameter variations and disruptions, S.M.C. has been favoured as a special class of variable structure systems (V.S.S.) in practical applications for over 50 years. The V.S.S. definition was first developed in Russia in the early 1960s, from the pioneering work of Emel'yanov and Barbashin[63]. V.S.S. and S.M.C. have gained a great deal of attention through control research.

Worldwide culture after an article published in 1977[64]. The S.M.C. technique is used to design a control law requiring all system trajectories, the so-called sliding surface S, to converge on a surface in state space. The designer selects this surface's dynamics such that all trajectories converge to the set point asymptotically. The system works in the so-called sliding mode when the trajectory lies inside the sliding surface and is vulnerable to parametric variations and external disturbances. V.S.S. and S.M.C. have gained a great deal of attention through control research.

Sliding mode control operation is essentially discontinuous and can stimulate high-frequency dynamics[65]. The discontinuous nature of the control action is used by moving between two incomparably different system structures to preserve the resulting superior system output of V.S.S. and S.M.C. Mode with this function is often referred to as a new form of device motion in a manifold or, in other significant words, in the vicinity of a prescribed switching manifold, the velocity vector of the regulated state trajectories is often guided by such motion induced by the application of discontinuous control behaviour towards the switching manifold. This output of the system is expected to exhibit insensitivity to parameter variations as well as show complete rejection of disturbance. S.M.C. suffers from a very common and well-known chattering problem, which is typically interpreted as motion oscillating around the predefined switching manifold(s), despite its advantages such as simplicity and robustness.

There are two explanations behind the chattering phenomenon: first, in the absence of switching non-idealities such as delays, that is, in a situation in which the switching system switches preferably at an infinite frequency, the presence of parasitic dynamics in series with the plant induces a high-frequency oscillation of small amplitude to occur around the sliding manifold. In the open loop model used for control design in control applications, these parasitic dynamics that reflect the fast actuator and sensor dynamics are often ignored if the closed loop pole positions are well established, or the closed loop poles are well allocated with the help of the pole placement design technique. The movement of the actual system is usually closer to the ideal. A mechanism where the parasitic dynamics are ignored and a rapid decline is shown by the discrepancy between the ideal and the actual motion, which is at negligible time constants.

Generation of the control signal for S.M.C.

Sliding mode control (S.M.C.) is a non-linear control method in control systems that alters the dynamics of a non-linear system by applying a discontinuous control signal (or more rigorously, a set-value control signal) that causes the system to slide along the normal behaviour of the system over a cross-section of the system. The law on state-feedback control is not a continuous time function. Instead, depending on the current location in state space, it may move from one continuous structure to another. Sliding mode control is, therefore, a form of variable structure control. The various control structures are structured such that trajectories often travel into an adjacent region with a different control structure, such that within one control structure, the ultimate trajectory will not exist entirely. Instead, it will slip around the control structures' boundaries. A sliding mode[1] is called the motion of the device as it slides along these boundaries, and the geometrical locus consisting of the boundaries iscalled the sliding (hyper) surface. Any variable structure system, such as a system under S.M.C., can be seen as a special case of a hybrid dynamical system in the sense of modern control theory, as the system also flows through a continuous state space but also passes through various discrete modes of control.

For its robust monitoring capabilities, the S.M.C. is well established. In two phases, the S.M.C. control law is constructed; first, the sliding surface is designed for the desired dynamics of the system and second, a switching control law is created that forces the system states to satisfy the reaching condition. It is possible to describe the traditional S.M.C. control law as:

$$U = \sin\left(s\left(x\right)\right)\begin{Bmatrix} +1 \ if \ s\left(x\right) \geq 0 \\ -1 \ if \ s\left(x\right) \leq 0 \end{Bmatrix} \quad \ldots(4)$$

Where, $s(x)$ is termed as the sliding surface. The control law, which is also a switching function shows that the output of the controller depends upon the sliding surface function.

1. SIMULINK design of S.M.C.

The S.M.C. scheme for the motor speed control designed in Simulink is shown in Fig. 1.

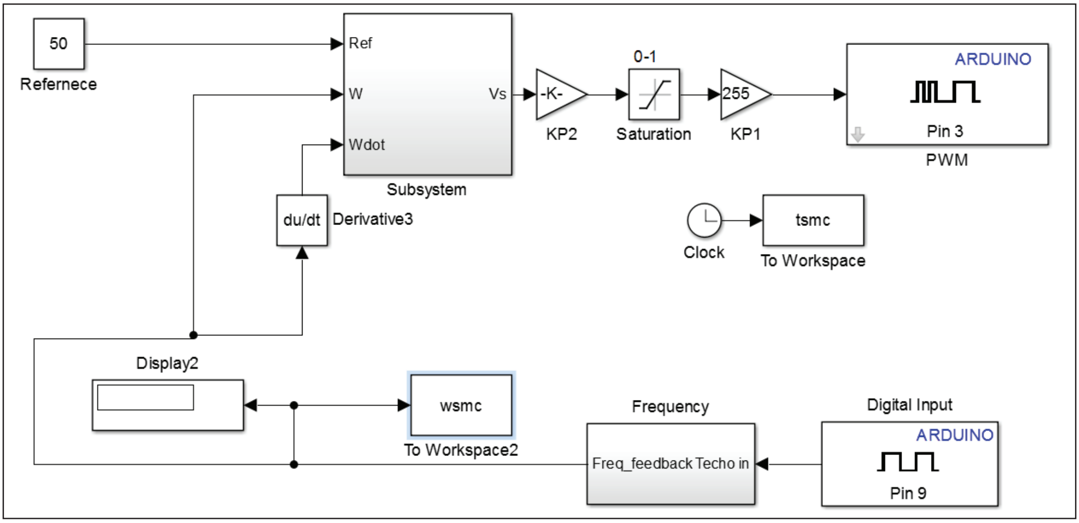

Fig. 1: S.M.C. scheme for speed control

Mathematical Model of D.C. Motor

To derive the mathematical model of the D.C. motor, consider the D.C. motor model shown in Fig. 2.

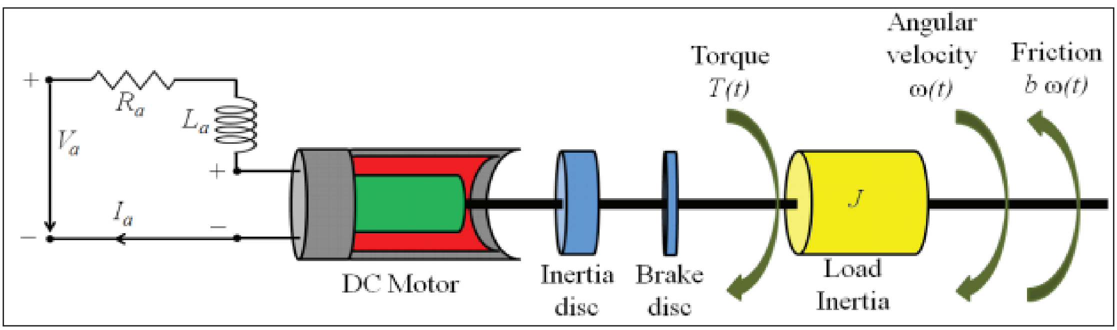

Fig. 2: D.C. motor Model

The armature voltage $V_a(s)$ controls the angular velocity $w(s)$ of the motor shaft. From a mathematical model, the transfer function is given by in Eq. 5.

$$\frac{w(s)}{va(s)} = \frac{\frac{kt}{JL}}{S^2 + \frac{R}{L}S + kt\ kb\frac{1}{JL}} \qquad \ldots(5)$$

By neglecting armature losses and rotational losses, the mechanical power output of the motor is equal to armature input. Hence, $K\_m = K\_b$ and $K\_r = 0$ The specifications of motor used in this paper are $R = 2.5\Omega$; $Km = Kb = 0.0955$ V s/rad; $L = 0.74$ H; $J = 0.0092$ Kg m^2/s^2. The transfer function of D.C. motor using above specified motor constants is given as

$$\frac{w(s)}{va(s)} = \frac{(7.68152)}{(s^2 + 3.378383s + 7.33585)} \qquad \ldots(6)$$

1. Fractional Calculus

Since the invention of the integer order calculus, fractional calculus has been recognized, but it has been treated as a single mathematical issue for a long time. The history of fractional calculus research is almost as long as the theory of integral calculus has been established. L'Hospital wrote to Leibniz at the beginning of 1695 to discuss a practical fractional derivative, but it was not until 1819 that Lacroix first presented the findings of a basic fractional derivative function. At the beginning of the theory, the progress of fractional calculus theory was very slow, due to the absence of backgrounds in practical application. Professor Mandelbrot first suggested "fractal" theory in the 1970s and pointed out that there are a number of fractal dimensions in nature and many science and technology issues, that there is a self-similar phenomenon, and that there are fractional Brownian motion and fractional calculus between the whole and the component. Fractional calculus theory and FDE theory have evolved rapidly as the dynamic basis of fractal geometry and fractal dimension and have become a hot research subject in the world.

A good method to define physical memory and heredity is given by fractional calculus. Many areas such as flabby, oscillation, stochastic diffusion theory and wave propagation, biological materials, control and robotics, viscoelastic dynamics, and quantum mechanics have been applied to fractional order calculus. The development of the theory of fractional calculus has also been accelerated by those applications. The theory of fractional order calculus and chaos and dissipative structure theory is known as the current theory of non-linear science. It is possible to extend traditional integer order differentiation and integration to impose orders that are not necessarily integer-order. Fractional differential equations are led by non-integer differentiation and integration of real functions. Such principles were transferred to control engineering as a new control approach called fractional order control. These controllers are the extended version of traditional integer order controllers that have certain additional parameters that need to be more accurately tuned and the process for the control design is more complex than integer order controllers. In short, higher performance is assured with more parameters. Fractional order models have more degrees of freedom and can approximate processes with fewer parameters.

Compared with standard P.I.D. controllers, Poldubny has shown that fractional order P.I.D. controllers denoted by PIλDμ have a better response.

Fractional calculus is the generalization of non-integer orders for differentiation and integration. It has developed by extending ordinary differential equations (ODE) to fractional-order differential equations from ordinary calculus (FODE). Similarly, an extension of a regular (integer) P.I.D. the controller is a fractional-order proportional-integral-derivative (FOPID) controller; its output is a linear input combination, a fractional input derivative and a fractional input integral. These fractional derivatives provide an excellent method to explain the memory and inherited characteristics of different materials and processes. In fractional calculus, which is part of mathematics, fractional derivatives and integrals are discussed. There are several uses for the fractional calculus that can be applied. The integral, differential fractional order is denoted by the operator $aD_t^{\propto}$. It is defined as below:

$$aD_t^{\propto} = \begin{cases} \frac{d_x}{dt^x} & \text{Re} \propto > 0 \\ 1 & \text{Re} \propto = 0 \\ \int_{\alpha}^{t} (d\tau)^{-\propto} & \text{Re } \alpha < 0 \end{cases} \qquad \ldots(7)$$

Where a and t are the limits of the operation, ∝ is the order of the operation, and generally ∝ ∈ R and ∝ can be a complex number. The two most used definitions for the general fractional differentiation and integration are the Riemann-Liouville (R.L.) definition and the Grunwald-Letnikov (G.L.) definition. According to Riemann Liouville fractional differential integral is defined as:

$$aD_t^{\propto} = \frac{1}{\Gamma(m-a)} \left(\frac{d}{dt}\right)^m \left(\int_a^t \frac{f(t)}{\left((t-\tau)^{1-(m-a)}\right)} d(\tau) \right) \qquad \ldots(8)$$

Where n −1 < λ < n and Γ(.) is Euler's gamma function. Having zero initial conditions, and is defined as;

$$\Gamma(x) = \int_0^{\infty} e^{-t} t^{(x-1)} dt, x > 0 \qquad \ldots(9)$$

Definition by Grunwald Letnikov for fractional derivative integral as;

$$aD_t^{\propto} f(t) = \lim_{h \to o} \frac{1}{h^{\alpha}} \sum_{k \to 0}^{\frac{t-a}{h}} (-1)^k \binom{a}{k} f(t-kh) \qquad \ldots(10)$$

2. Design of a Fractional Order Controller

Fractional calculus represents the generalization of the integration and differentiation to an arbitrary order. The Laplace transform is given in Eq. 11.

$$\mathrm{LhI}^{\alpha}f(t)i = s^{-\alpha}F(s) \quad \ldots(11)$$

for the fractional-order integral, while for the fractional-order derivative, the equation is:

$$\mathrm{LhD}^{\alpha}f(t)i = s^{\alpha}F(s) \quad \ldots(12)$$

with $\alpha \in (0,1)$.

The transfer function of the FO-PI controller is given as in Eq. 13.

$$H_{F0} - PI(s) = k_p\left(1 + \frac{k_i}{S^{\mu}}\right) \quad \ldots(13)$$

where k_p and k_i are parameters of the fractional-order P.I. controller, while μ represents the fractional order. The Design of the FO-PI controller is usually based on a phase margin and a gain crossover condition, to which a third criteria may be added in order to uniquely determine the three controller parameters. In order to tune the fractional order P.I. controller, the open-loop transfer function needs to be computed as in Eq. 14:

$$Hopen{-}loop(s) = HFO{-}PI(s)HP(s) \quad \ldots(14)$$

where H.P. (s) is the process to be controlled. The tuning of the FO-PI controller implies the computation of the three parameters k_p, k_i and μ according to three performance specifications imposed. The gain crossover frequency - ωcg - implies that the modulus of the open-loop transfer function obeys the following:

$$|\mathrm{Hopen{-}loop}(j\omega gc)| = 1 \quad \ldots(15)$$

while the phase margin - φ_m - specification sets a condition upon the phase of the open-loop system at the gain crossover frequency, mathematically written as:

$$Hopen{-}loop(j\omega gc) = -\pi + \phi m \quad \ldots(16)$$

The performance specifications given above may be rewritten as:

$$|H_{F0} - PI\left(jw_{gc}\right)| = \frac{1}{|H_p\left(jw_{gc}\right)|} \quad \ldots(17)$$

$$\angle HFO{-}PI(j\omega gc) = -\pi + \phi m - \angle HP(j\omega gc) \quad \ldots(18)$$

which can be further detailed as:

$$\left| k_p \left[1 + k_i w_{gc} - \mu \left(\cos\frac{\pi\mu}{2} - j\sin\frac{\pi\mu}{2} \right) \right] \right| = \frac{1}{\left| H_p\left(jw_{gc} \right) \right|} <$$

$$\left[\left[1 + k_i w_{gc} - \mu \left(\cos\frac{\pi\mu}{2} - j\sin\frac{\pi\mu}{2} \right) \right] \right] = -\pi + \varphi_m - \angle HP\left(j\omega gc \right)$$

Since the FO-PI controller has three independent parameters, these can be adequately tuned to meet three performance specifications. Thus, apart from imposing a certain gain crossover frequency and a certain phase margin, which naturally imply a certain settling time and overshoot, a third condition may be added to the design problem. This third condition can refer to high frequency noise rejection, good output disturbance rejection, robustness to variations in the gain of the plant, etc.

3. Review of P.I.D. Controller

The P.I.D. controller is a linear controller and is very common. It is used in many applications due to its simple nature. It can be used for various applications, such as temperature, speed control, pressure, and flow. As a result of these three coefficients, there are 3 simple parameters or gains that are modified to get an optimal answer. The coefficient for K.P. is compounded by error. The Integral variable adds the time interval error. This term will gradually increase by a tiny mistake duration. To mitigate steady state error, the K.P. term is used.

The Design and tuning of a proportional integral derivative (P.I.D.) controller seem to be conceptually intuitive, but if several (and sometimes conflicting) goals such as short transient and high stability are to be achieved, it can be difficult in practice. Initial designs obtained by all means usually need to be modified Until the closed-loop system performs, or compromises as expected, repeatedly by computer simulations. This stimulates the development of "intelligent" instruments that can help engineers achieve the best overall control of P.I.D. for the entire operating envelope. This innovation has further contributed to the integration into P.I.D. hardware modules of certain sophisticated tuning algorithms. In line with these trends, this article presents the classic description of patent functionalities and methods of tuning, Packages of applications and modules for consumer hardware.

It's being seen in order to improve transient performance, several P.I.D. variants have been developed but standardizing and modularizing P.I.D. control is required, although challenging. The incorporation of device recognition and "intelligent" techniques in P.I.D. systems based on software helps to simplify to a useful degree the entire process of Design and tuning. This can also support the future production of widely available "plug-and-play" P.I.D. controllers that can be conveniently set up and function optimally for increased efficiency, improved quality, and reduced maintenance requirements.

In relation to the rise in plant production, the K.D. the coefficient is responsible for reducing output. To minimize settling time, the derivative gain is used. Any benefit has a grip on the system. To get the desired output, P.I.D. coefficients must be tuned. If the coefficients are not correctly tuned, the scheme would be unstable. The control system is required to set the best earnings for K_p, K_i and K_d in order to get the desired output. The method of setting optimum gains is tuning. For setting optimum gains, there are several tuning methods, such as Ziegler Nicholas. The steps for P.I.D. tuning is given below.

In the open loop, check the system's answer. Minimize the system's overflow answer by multiplying errors with K_p. To eliminate overshoots and stabilize the system, apply K_d Apply K_i to minimize steady state error. To get the desired answer, change the K_p, K_i and K_d values.

The equation for P.I.D. controllers is shown below:

$$PID = k_i \int_0^t e(t)\,d(t) + Kpe(t) + Kd\frac{d}{dt}e(t) \quad \ldots(19)$$

The P.I.D., the controller, is implemented in the SIMULINK environment, as shown in Fig 3.

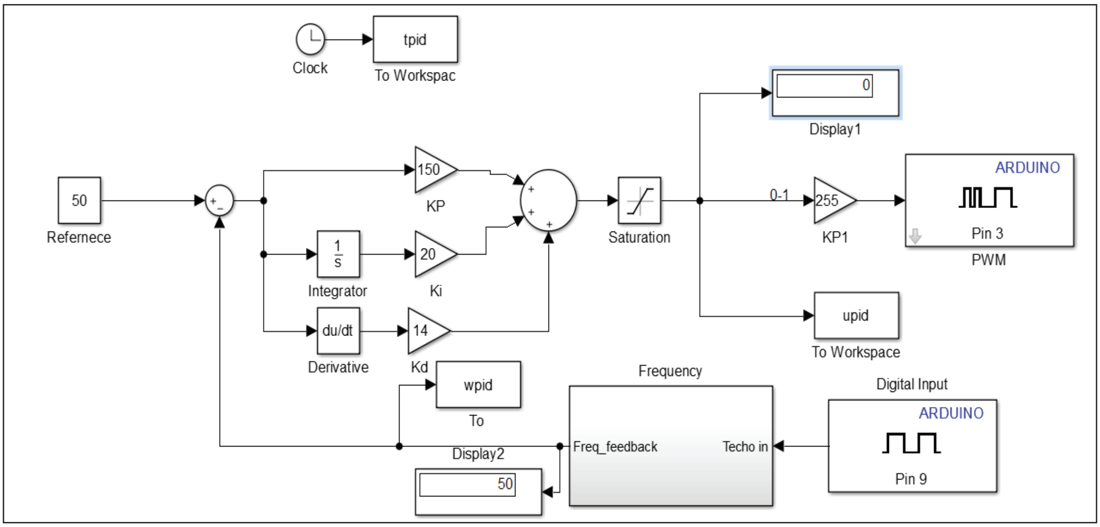

Fig. 3: P.I.D. Controller

4. Integral Sliding Mode Controller

Since no load or full load causes the plant parameters to shift, the results of the P.I.D. the controller is not satisfactory. Sliding mode control is suggested to address these issues. In the presence of disruptions, it provides optimum results. It is a technique of control that is non-linear and most commonly used. It has outstanding features such as precision, robustness and easy tuning and efficiency.

$$x_1 = v_w \quad \ldots(20)$$

$$u = v_a$$

$$\dot{x} \quad \dot{w}$$

$$= \frac{ukt}{J.L} - \frac{ktkbx_1}{J.L} - \frac{Rx_2}{L}$$

$$x_1 = x_2 + d_1(t)$$

$$x_2 = \frac{ukt}{J.L} - \frac{ktkbx_1}{J.L} - \frac{Rx_2}{L} + d_2(t)$$

$D_1(t)$ & $D_2(t)$ are disturbances. So, according to the definition of error,

$$e\_1 = x\_1 - x\_1r \qquad \ldots(21)$$

Taking derivative on both sides of Eq. 21

$$(e\_1)\dot{=}((x)\_1)\dot{-}(x\_1r)$$

$$(e\_1)\dot{=}((x)\_1)\ddot{-}((x)\_1r)^{\cdot\cdot} \qquad \ldots(22)$$

The sliding surface is chosen as,

$$S = C\_(1)\, e + e^{\cdot}$$

Taking (D^1) on both sides,

$$D^1 S = C\_(1)\, D^{(1)} e + e^{\cdot\cdot}$$

$$D^{(1)} S = CD^{(1)} e + ((x\_1)\ddot{-}(x\_1r))^{\cdot\cdot} \qquad \ldots(23)$$

((where x_1)) $\ddot{=} x\_2$ and $(x\_1)\ \ddot{=} (x\_2)^{\cdot\cdot}$

$$D^{(1)} S = C\_(1)\, D^{(1)} e + ((x\_2)\dot{-}(x\_1r))^{\cdot\cdot}$$

Put $(\dot{x}_2)$ equivalent value

$$D^1 S = C_1 D^1 e + \left(\frac{kt}{J.L} - \frac{ktkbx_1}{J.L} - \frac{Rx_2}{L} + d2(t) - x_{1r}\right)^{\cdot\cdot} \qquad \ldots(24)$$

As we know that, $U = U_{eq} + U_{ro}$

For (U_{eq}) put (D^1S) equal to zero so,

$$U_{eq} = -C_1 e + \frac{J.L}{kt}\left(\frac{ktkbx_1}{J.L} + \frac{Rx_2}{L}\right) + \ddot{x}_{1r} \qquad \ldots(25)$$

For (U_{ro}) put ($D^1S = -K_s$ sign(s))

$$(U_{ro}) = -K_s \text{ sign } (s)$$

As given above.

$$U = U_{eq} + U_{ro}$$

$$U = -K1e + \frac{J.L}{kt}\left(\frac{ktkbx_1}{J.L} + \frac{Rx_2}{L} + \ddot{x}_{1r}\right) - k_s sign(s) \qquad \ldots(26)$$

The SIMULINK design of the S.M.C. controller can be seen in Fig. 4.

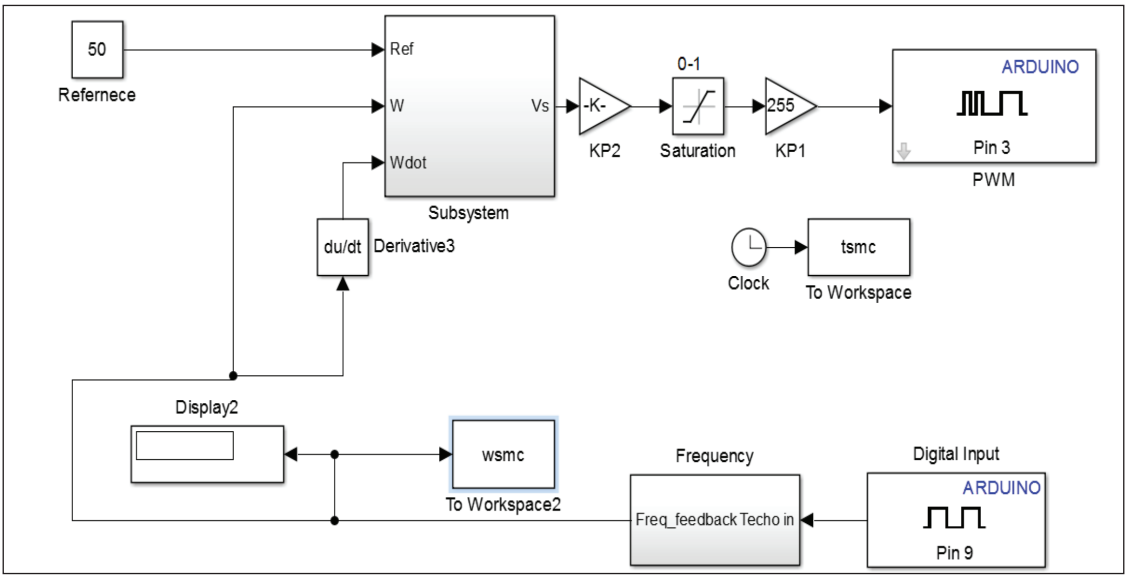

Fig. 4: S.M.C. Controller design

Design of IOFL for Speed Control

Induction motors have lately become the primary option for industrial load driving, especially where applications require variable speed, easy to operate, physically durable, more powerful, and cost-effective than D.C. motors. Nowadays, the trend is to substitute induction machines for D.C. machines. It is not easy to control the induction motor as the D.C. motors because it is a multiple-input multiple-output (MIMO), non-linear problem, and its operating conditions are away from the balance points whereby methods of linear approximation are not applicable. Various research studies have been performed by different researchers around the world to enhance the efficiency of induction motor drives for variable speed drive applications. In general, control of A.C. machines can be divided into two groups. The scalar control is the first and the oldest one. While this A.C. machine control system is simple to implement and can provide a better steady-state response, a less stable dynamic response is found. Its response to parameter variation is awkward. Further development has led to another control scheme called the Vector control to achieve better variable speed efficiency on those devices. The collective name was given to field-oriented control (F.O.C.), direct torque control (D.T.C.), non-linear control and predictive machine control is vector control.

A summary of the studies being performed on induction motor control shows that the focus is primarily on improving the D.T.C. system of these motors. This is because of their superior success over F.O.C. system. Different research experiments have been suggested by many researchers to enhance traditional D.T.C. using hysteresis controllers. The hysteresis controllers are replaced by proportional-integral (P.I.) controllers to produce the reference voltages. Although this increases performance but considers ripples for traditional direct torque controllers, the application of these controllers involves an exact model of the system, which is ideal for practical considerations. Besides, it is not an easy task to pick the controller gains for this controller, and the P.I. controllers are highly susceptible to disturbances, model uncertainties and variations of parameters. fuzzy logic, artificial predictive control methods for the neural network and the model are provided. The downside of these methods in solving complicated and complex problems is their complex existence. The researchers began concentrating on the non-linear control of induction motors as a solution to the above, in particular the input-output feedback linearization (IOFL), which was favored for its ultimate performance on robustness to varying parameters, simpler structure and fast reference tracking functionality.

IOFL is a technique that allows the designer/user to apply non-linear structures such as the induction motor to linear control strategies (I.M.). In order to be able to use the linear control technique, this linearization approach converts the non-linear system model into a linear equivalent one. In applying IOFL to P.I. flux controllers, a study introduces a design methodology. This P.I. controller, however, is still seen as a weakness in this technique. The solution to this problem is found by choosing to use the robust controller family type called the controller sliding mode (S.M.C.). But, for fast monitoring and improved robustness, the controller sliding mode fractional order (FOSMC) has opted for the velocity control loop. For different researchers, fractional equivalents of traditional controllers are nowadays a major concern.

In this paper, we addressed non-linear induction motor control based on linearizing input-output feedback technique using the space vector pulse width modulator (SVPWM). The benefit of this approach is that by compensating for the nonlinearities present in the machine and thereby ensuring that flux and electromagnetic torque are perfectly decoupled. In the (alpha β) frame, we used an engine model. This model needs no Further conversion to the direct and quadrature axis would lead to a delay in developing the controller as a result of this conversion. The major drawback of this system of controllers is that all states are necessary. Since it is a tough job.

We have also designed a sliding mode flux observer from stator current and voltage information and a P.I. type load observer to test rotor flux and load torque disturbance. The speed controller is another factor that we are going to research in this article. Using traditional controllers such as P.I.D., which is the most common controller for electrical drives, speed was controlled. However, the variations present in the system could not guarantee such controllers. The FOSMC, which is a well-known controller that controls systems with uncertainties robustly, may be the better choice. FOSMC's implementation not only addresses robustness, but also monitors the desired direction faster. The chattering (undesirable oscillations) present is the main downside of using S.M.C. families. This issue is alleviated by the use of continuous smooth approximation of the signum function.

1. Design of space vector P.W.M.

AC motor control is mostly achieved by regulating the DC/AC inverter output voltage, which is in turn fed to the system stator by producing an effective switching gate pulse driving each switch of power electronics. The first step is to establish the reference voltage and locate where it is located in the field. For the induction motor in (d, q) reference frame, a step-by-step SVPWM design procedure is provided.

$$V_s = V_{s\alpha} + jV_{s\beta} \qquad \ldots(27)$$

$$\theta = \tan^{-1}\left(\frac{V_{s\beta}}{V_{s\alpha}}\right) \qquad \ldots(28)$$

The sector S is decided by the value of θ and is categorised in one of the six regions. After deciding dwell times or application times, the next step is to determine the service cycle to be modulated with a frequency ramp signal equal to the frequency of the electronic power switching unit. The ideal output of the inverter phase voltages can be determined from the inverter's D.C. input and its simple gate pulses as follows:

$$V_\alpha = \frac{V_{dc}}{3}\left(2p_\alpha - p_b - p_c\right) \qquad \ldots(29)$$

$$V_b = \frac{V_{dc}}{3}\left(2p_b - p_a - p_c\right) \qquad \ldots(30)$$

$$V_b = \frac{V_{dc}}{3}\left(2p_c - p_a - p_c\right) \qquad \ldots(31)$$

2. IOFL Design

For a nonlinear system, the generalized form is,

$$\dot{x} = f(x) + g(x)u$$
$$y = h(x)$$

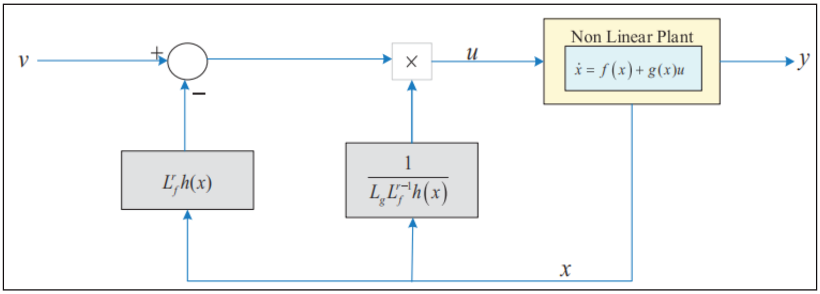

Fig. 5: Block diagram of IOFL

One which use the input-state linearization technique to linearize a non-linear system. However, this linearization strategy would not always lead to a linearized output, so the IOFL is used to achieve a linearized output.

Also, an equation Taking consecutive differentiation of the above output equation results in the following:

$$y = \frac{d(L_f h)}{dt}\left[f(x) + g(x)u\right] = L_f^2 h(x) + L_g L_f h(x)$$

While in the generalize context,

$$Y^r = L_f^r h(x) + L_g L_f^{r-1} h(x) u \qquad \ldots(32)$$

Here, the outputs are viewed as the produced electromagnetic torque of the motor and the rotor flux square. The further derivation of the controller can be simplified by this. The method, looking from the built model, is MIMO is considered because it has two inputs, namely Vs alpha and Vsβ, and two outputs, namely y_1 and y_2. Successive differentiation of the output before the effect occurs in at least one of the inputs gives us the following:

$$y_1 = V_\alpha = -k_{\alpha 1}\left(T_{em} - T_{emref}\right) + T_{emref}$$
$$y_2 = V_\beta = -k_{\beta 1}\left(\lambda_r^2 - \lambda_{ref}^2\right) - k_{\beta 2}\left(\lambda_r^2 - \lambda_{ref}^2\right) + \lambda_{ref}^2$$

The dynamics above will be stable if the roots of polynomials lie on the left side of the complex plane. This requires for the characteristic equation involving constants $k_{\alpha 1}$, $k_{\alpha 2}$ and k_{b2} to satisfy the Hutwith criterion.

3. Proportional, Integral and Derivative Speed Controller

Earlier, we developed a non-linear controller based on external references using IOFL to independently control electromagnetic torque and flux. The consumer sets the flux reference, but the torque reference is obtained from the velocity controller block. The P.I. controller, which is the most common controller for electric drives, has been used to control speed. But the speed controller type used matters.

The governing equation can be defined as,

$$\frac{d\omega_m}{dt} = \frac{1}{J_{eq}}\left(T_{em} - T_1\right) - \frac{f_r}{J_{eq}}\omega_m \qquad \ldots(33)$$

Using a suitable tuning mechanism, the gains of P.I.D. controllers should be tuned.

4. Design of FOSMC for speed loop

The dynamics of the speed loop can be rearranged from the model equations derived earlier to simplify our design approach without altering the dynamic structure.

$$\omega = \phi\omega_m + l_o\left(T_{em} - T_1\right) + d \qquad \ldots(34)$$

where, $\phi = \Delta\phi\omega_m + \Delta l\left(T_{em} - T_1\right)$

"Δ" Introduces the ambiguity of the parameter and the subscript 'o' is used to denote the parameter's nominal value.

Using the exponential law,

$$S_w = -k_r S_w - k_s sign\left(S_w\right) \qquad \ldots(35)$$

$$T_{emref} = \frac{1}{l_o}\left[\omega_{mref} - \phi_o\omega_m + l_o T_1 + \lambda_{10} D_t^{1-\alpha} e_\omega + k_r S_w + k_s sign\left(S_w\right)\right]$$

The chattering present in this controller's control input makes its implementation troublesome as it can cause shaft damage and linked loads. Here, the enhancement of this controller can be accomplished by substituting a smoother sigmoid function for the signum function.

5. Sliding mode rotor flux observer

The sliding-mode observer is designed to monitor the stator currents and the rotor flux is measured using the observer's control signals. The suggested estimation scheme for rotor flux uses stator currents, but the stator voltage is not needed. In practical applications, it has significant advantages, particularly for sensor-less field-oriented control (F.O.C.) of induction motors. Measuring the rotor fluxes and feeding them back to the controller is difficult. We are therefore obliged to design an observer for the states of rotor flux.

The earlier model can be re-written as,

$$\frac{d\hat{i}_s}{dt} = \beta U - \frac{R_s}{\sigma}\hat{i}_s + \frac{1}{\sigma}\overline{V}_s$$

Therefore, the observer expression is,

$$\frac{d\hat{i}_s}{dt} = \beta U - \frac{R_s}{\sigma}\hat{i}_s + \frac{1}{\sigma}\overline{V}_s \quad \ldots(36).$$

$$\frac{d\hat{\lambda}_r}{dt} = -U$$

6. Load Torque Observer

In different industrial, civil, and military sectors, PMSM has been commonly used. Fields due to their high-power factor benefits, high performance, quick dynamic reaction, high reliability, etc. The conventional high-power asynchronous machines or hydraulic systems used in some large and high-end equipment have been steadily replaced in recent years, such as servo broaching machines, all-electric injection molding machines, crank servo presses, etc. However, special efforts are needed to suppress the velocity fluctuation under operating conditions with large load and inertia variations when the PMSM drive is used in the large high-end equipment described above. On-line load torque identification is needed to achieve this.

The attractive advantages of robustness are sliding mode observers (S.M.O.), Low sensitivity to the parameter variations of the device, quick response, and easy implementation against disturbances. Many SMO-based methods of parameter recognition have therefore been proposed. SMO-based control schemes to manage PMSM drive systems with time-varying parameters and disturbances. Extending the S.M.O. by the inclusion of the Mechanical parameter errors/disturbances in the state space equations to estimate the mechanical parameters and device disturbances of the PMSM drive system. The sign feature can, however, trigger a high frequency chattering/buffeting problem in the aforementioned traditional S.M.O.s, which could generate system oscillation, performance degradation, or even system instability. A Compromise solution is to add a low-pass filter (L.P.F.) to obtain useful estimated parameter information. Unfortunately, the implementation of the L.P.F. will lead to a phase delay that will affect the accuracy of the calculation and the drive system's output. Proper compensation is therefore needed to mitigate the impact of the L.P.F., especially when A.C. signals are calculated, where there are continuous phase delay displays in the above-mentioned S.M.C.s, one can note that the word T1 is present in the expression of the control rule. However, this mechanical shaft load disturbance is an uncertain concept or, using calculation methods, it is difficult to determine its meaning. This includes the formulation of an observer of load torque that can derive its value from flux estimates and current values.

$$T_1 = \widehat{K_{p1}}\left(\widehat{\omega_m} - \omega_m\right) + K_{i1}\int\left(\widehat{\omega_m} - \omega_m\right)dt \qquad \ldots(37)$$

Where, $\left(\widehat{\omega_m} = \dfrac{1}{j_{eq}s + f_r}\left(T_e - \hat{T}_1\right)\right)$

Fig. 6 shows the representation of load torque observer,

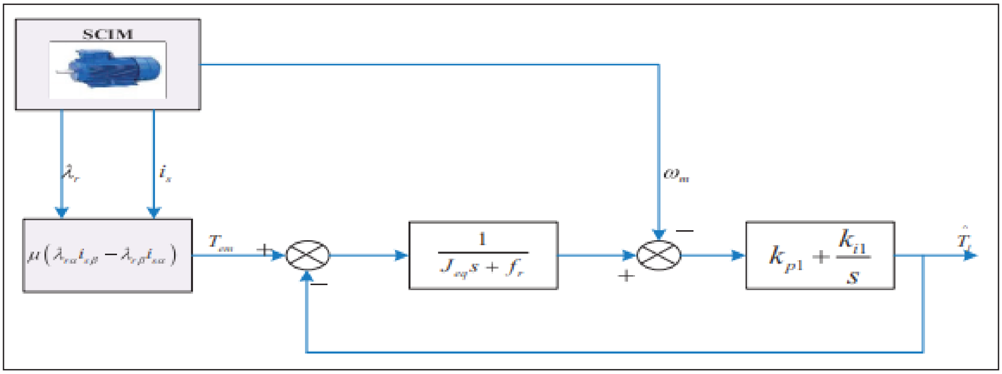

Fig. 6: Load Torque observer

D.C. motor speed control using FOSMC Controller

Among the most commonly used components in both academic and industrial applications are D.C. motors. An electric motor with a certain associated power supply and amplifier stage to control the power input to the motor in response to a lower-level control signal is an important feature of any position or speed control device. The transfer role of the studied D.C. motor needs to be determined in order to adequately tune the FO-PI controller as indicated earlier. The applied voltage V_a, which is the manipulated variable, will control the position $\theta(t)$, which is the controlled variable. For the speed control, the controlled variable is the angular velocity $\omega(t)$.

$$P_{DC_{motor}}(s) = \frac{w(s)}{Va(s)} = \frac{L_m}{(L_\alpha s + R_\alpha)(J_s + b) + K_b K_m} \qquad \ldots(38)$$

The time constant $T_a = \dfrac{L_a}{R_a}$, *is negligible*, so Eq 38 is simplified to,

$$P_{DC_{motor}}(s) = \frac{K_m}{(R_a)(J_s + b) + K_b K_m} = \frac{\dfrac{K_m}{R_a b K_b K_m}}{T_s + 1}$$

$$= K_{DC-MOTOR}$$

Where, $T = \dfrac{R_a J}{R_a b + K_b K_m}$ and $K_{DC-motor} = \dfrac{K_m}{R_a b + K_b K_m}$

The transfer function from position $\theta(t)$ as output (controlled variable) to armature voltage V_a as input (manipulated variable) is evident in Eq 39:

$$P_{DC-motor} = \frac{\theta(s)}{V_a(s)} = \frac{K_{DC-motor}}{s(Ts+1)} \qquad \ldots(39)$$

Fig. 7 represents the Design for the FOSMC speed controller.

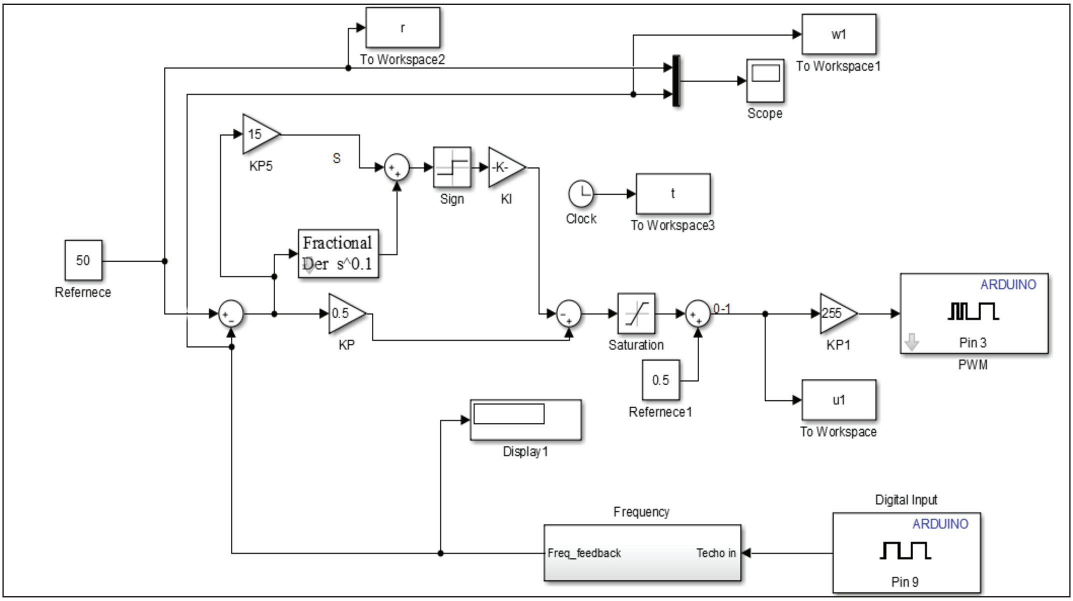

Fig. 7: FOSMC controller representation

RESULTS AND ANALYSIS

It implements and checks the controllers that are proposed. Considering various cases, the findings are taken. Various cases for checking the controllers are open loop responses, error signals, no load response, under load response and control efforts. The uncertainties and disruptions of the parameters are introduced by varying load, so in order to get an excellent output response, the controller must reduce it. In order to decide how to eradicate system disruptions, control signals from controllers are very important.

No Load Response

In order to verify the device response, P.I.D, S.M.C. and FOSMC are introduced when no load is present. The statistics display various controllers' responses. Finally, the joint response of no-load controllers is addressed. Figure 8 shows the no load response of a P.I.D designed controller.

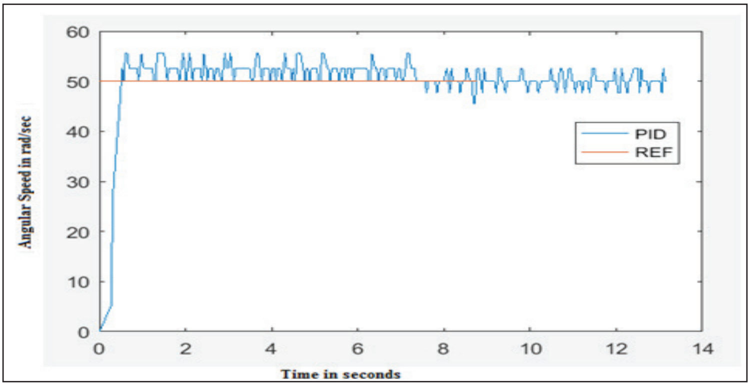

Fig. 8: No load Response for a P.I.D.

The no load output response of S.M.C. controller for D.C. motor speed control is shown in Fig. 9. Red line indicates reference speed and blue line represents the controller response.

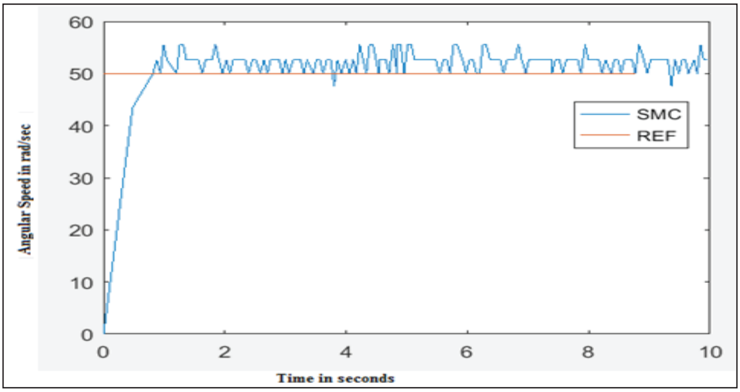

Fig. 9: No load Response for S.M.C.

Fig. 10 shows the output of FOSMC for D.C. motor speed control when there is no load connected.

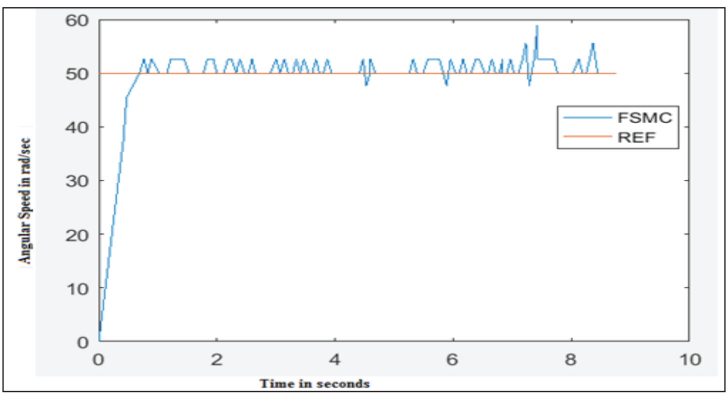

Fig. 10: No load response for FOSMC

To observe the robustness of FOSMC controller more clearly, Fig. 11 shows a comparison response for the no load scenario.

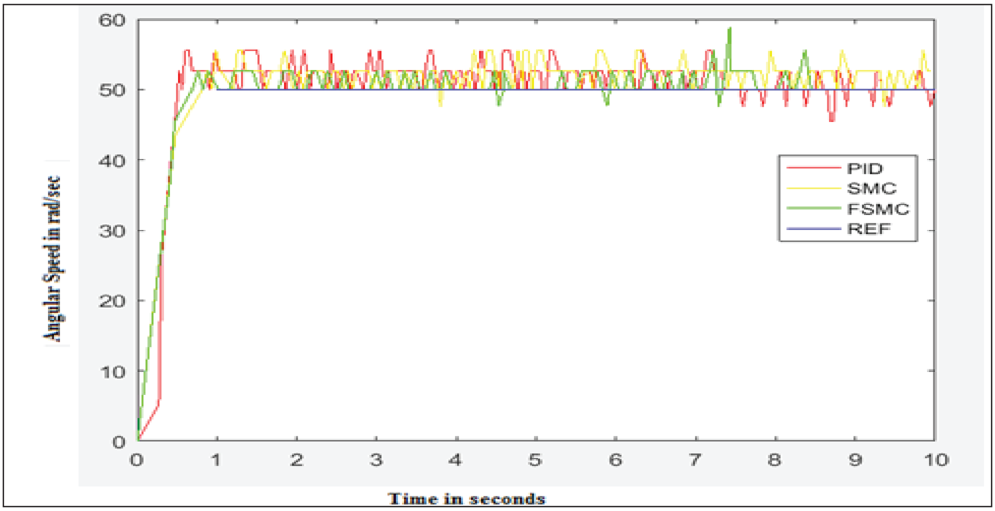

Fig. 11: Response comparison for no load in P.I.D., S.M.C. and FOSMC

Parametric Uncertainty Response

The uncertainties and disruptions of the parameters are introduced by varying load, so in order to get an excellent output response, the controller must reduce it. In order to decide how to eradicate system disruptions, control signals from controllers are very important. The control efforts of P.I.D is shown in Fig. 12.

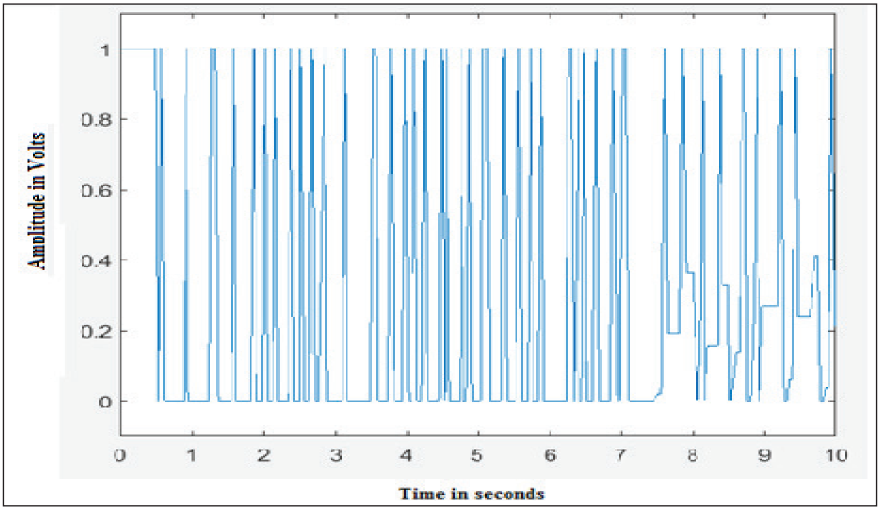

Fig. 12: P.I.D. response to parametric uncertainties

Fig. 13 shows the response of S.M.C. to parametric uncertainties.

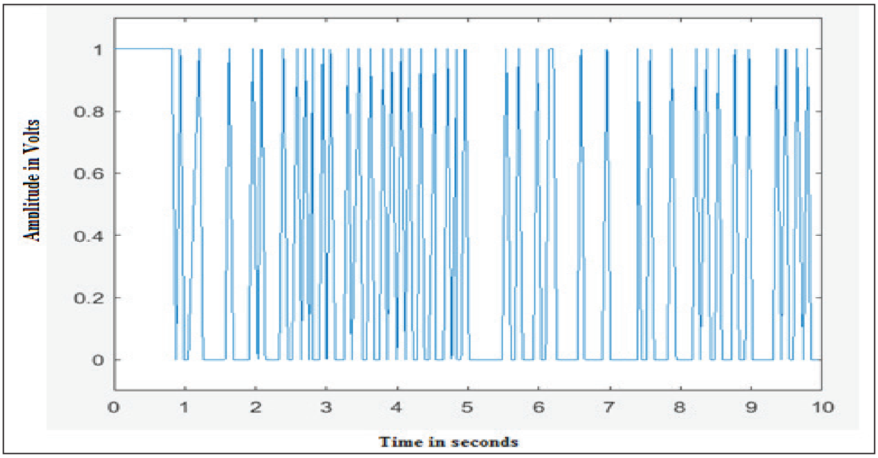

Fig. 13: S.M.C. response to parametric Uncertainties

Fig. 14 shows the response of FOSMC to parametric uncertainties.

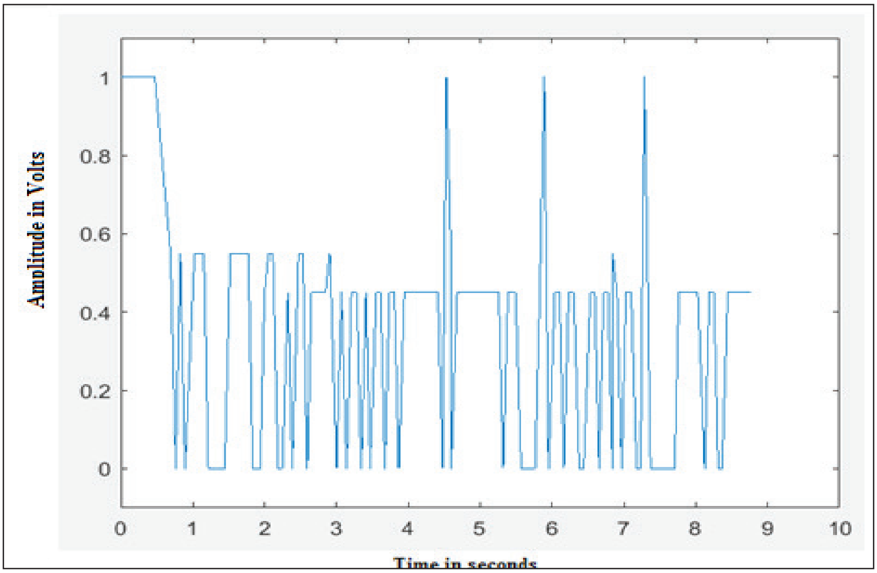

Fig. 14: FOSMC response to parametric uncertainties

9.3.Error Signal Plots

The error signal is determined from the reference & feedback differences. The error shows us how the real signal deviates from the fixed point. If the error signal is strong, then the output is not completely tracked. To get an excellent answer, the error signal needs to be reduced. The error signal of a P.I.D. is shown in Fig 15.

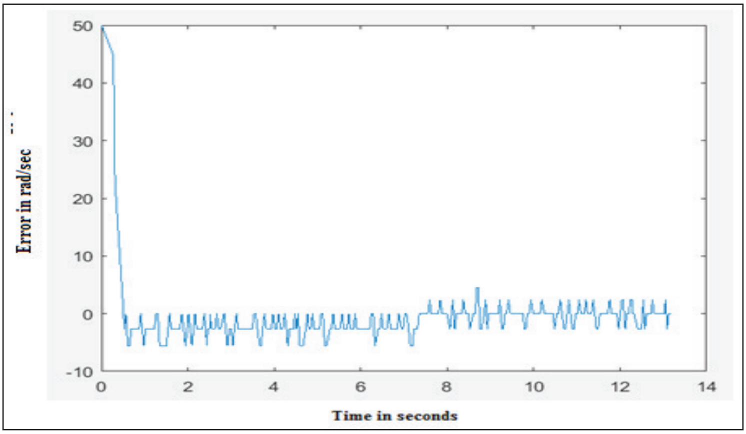

Fig. 15: Error response of a P.I.D.

The error response of a S.C.M. is shown in Fig 16.

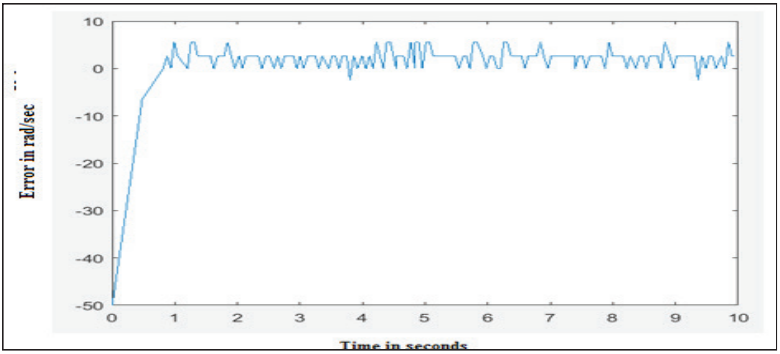

Fig. 16: Error signal of S.M.C.

Fig. 17 shows the error signal produced by FOSMC. The error is closer to zero in this case. The system response is excellent as compared to P.I.D. and S.M.C.

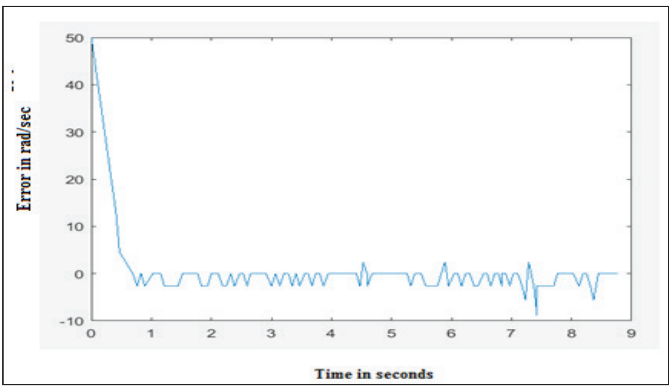

Fig. 17: Error response of FOSMC

9.4 Response under Load

To evaluate the system response, P.I.D, S.M.C. and FOSMC are introduced when the load is put on. The statistics indicate numerous responses from controllers. Figure 18 shows the response for a P.I.D. when the load is inserted after 3 seconds.

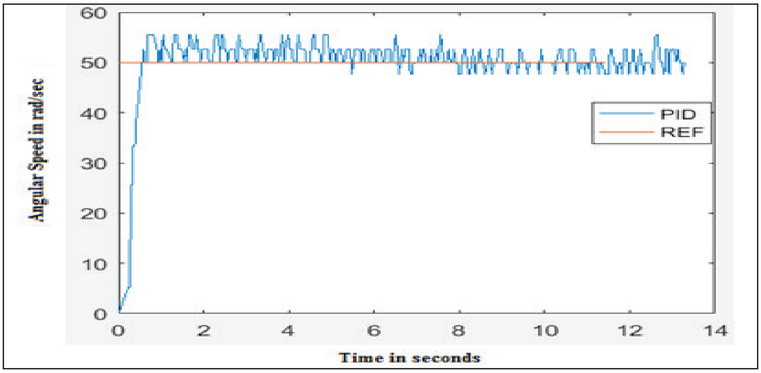

Fig. 18: P.I.D. response after load insertion at 3 seconds

Fig. 19 shows the response for a P.I.D. when the load is inserted after 3 seconds.

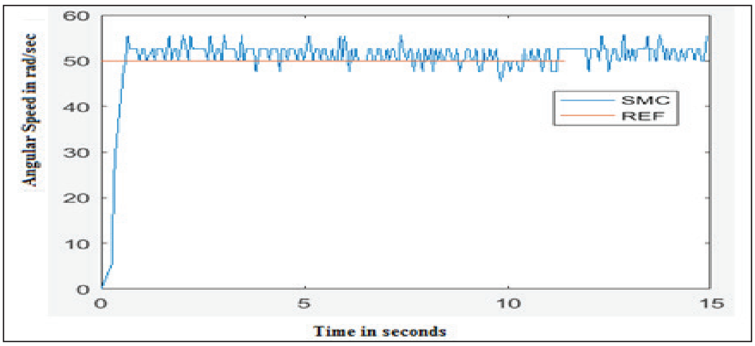

Fig. 19: S.M.C. response after load insertion at 3 seconds

Fig. 20 shows the response for a P.I.D. when the load is inserted after 3 seconds.

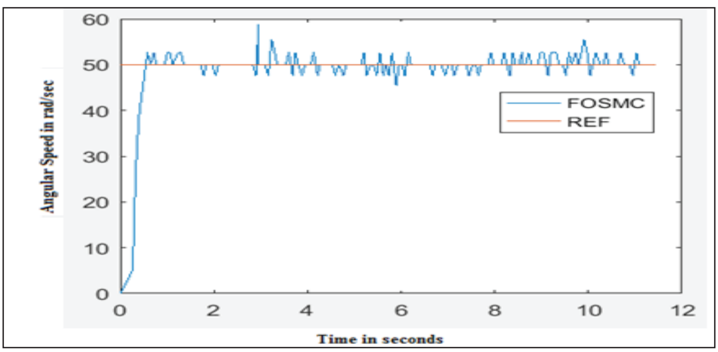

Fig. 20: FOSMC SMC response after load insertion at 3 seconds

Joint response of P.I.D., S.M.C. and FOSMC with load is presented in Fig. 21.

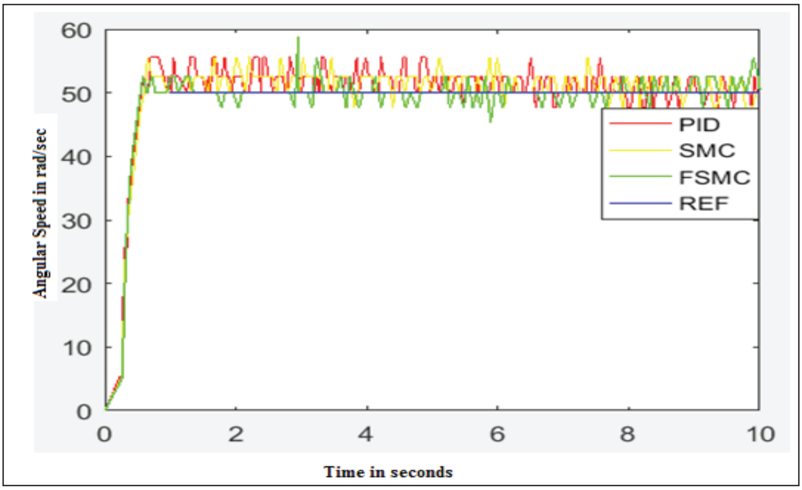

Fig. 21: Response comparison of P.I.D., S.M.C. and FOSMC

CONCLUSION

The speed of the D.C. motor is controlled and implemented using the fractional order controller. The speed controller has non-linear behaviour and is a time variant. Its parameters are altered due to the alteration of load on the D.C. motor. So, the main goal of this research is to build a control system that is not influenced by these conditions. Many controllers have been used for D.C. motor speed control, but some issues have been introduced in the past with each controller. For controlling a system, high performance controllers are often in demand.

To solve the problem with S.M.C., a fractional S.M.C. is suggested. FOSMC uses both the fractional sliding surface and the fractional plant model. There are several uses for the fractional calculus that can be applied. Implementation of PID, SMC & FOSMC controllers. In terms of control signals, output responses and error signals, comparisons are made between these controllers. FOSMC's findings are higher than S.M.C. & P.I.D.An illustrative instance of a fractional order in this paper. The controller for speed control of a load-changing D.C. motor is Introduced. The fractional-order controller's robustness has It was assessed by altering the gains and time constants of the D.C. motor as a result of the brake unit shift. Using a fractional-order P.I. controller, the experimental experiments are carried out and the output performance is compared to the integer-order P.I. controller. Using the same tuning algorithm, both controllers were designed. Thanks to the versatility of the parameters of the fractional order, the device can meet more requirements and thus become more robust for dynamic changes. The experimental results show that the controller of the fractional order outperforms the controller of the classical integer order. On the other hand, computational complexity means the Design of a fractional-order controller and additional effort may also be needed for the implementation itself.

REFERENCES

1. Yulin, D. 2010. The analysis and implement of PLC-based PI control for the permanent magnet D.C. motor. Proceedings of Second International Conference on Communication Systems, Networks and Applications.

2. Khalid Khan, M. 2003. Design and Application of Second Order Sliding Mode Control Algorithms. Doctor of Philosophy, University of Leicester, pp. 11-26.

3. Kumar, D. and Mija, S. 2014. Design and performance evaluation of robust S.M.C. schemes for speed control of D.C. motor. Proceedings of IEEE International Conference on Advanced Communication Control and Computing Technologies (ICACCCT).

4. Nouri, K., Dhaouadi, R. and Braiek, N. 2008. Adaptive control of a non-linear D.C. motor drive using recurrent neural networks. Applied Soft Computing. Adaptive control of a non-linear D.C. motor drive using recurrent neural networks. Applied Soft Computing.

5. Huang, G. 2008. PC-based P.I.D. speed control in D.C. motor. Proceedings of International Conference on Audio, Language, and Image Processing, pp. 400-407.

6. Ananthababu, P. and Reddy, B. 2009. Control of PMDC motor using fuzzy P.I. controller. Proceedings of International Conference on Control, Automation, Communication and Energy Conservation, pp. 1-4.

7. Gowthaman, E. and Balaji, C. D. 2013. Self-tuned P.I.D. based speed control of PMDC drive. Proceedings of International Multi-conference on Automation, Computing, Communication, Control and Compressed Sensing, pp. 686-692.

8. Velagic, J. and Galijasevic, A. 2009. Design of fuzzy logic control of permanent magnet D.C. motor under real constraints and disturbances. Proceedings of International Conference on Control Applications & Intelligent Control, pp. 461-466.

9. Kumar, U. and Dohera, D. 2015. Separately excited D.C. motor speed control of using various tuning conventional controllers. *International Research Journal of Engineering and Technology,* **2**(8): 2015.

10. Jin, Y., Luo, Y., Wang, C. and Chen, Y. 2009. LabVIEW based experimental validation of fractional order motion controllers, Chinese Control and Decision Conference (CCDC 2009), pp. 323–328.

11. Luo, Y., Quan Chen, Y. and Pi, Y. 2011. Experimental study of fractional order proportional derivative controller synthesis for fractional order systems, *Mechatronics,* **21**: 204–214.

12. Ruszewski, Sobolewski, A. 2012. Comparative studies of control systems with fractional controllers, Przeglazd Elektrotechniczny, *Electr. Rev.,* pp. 204–208.

13. Kaneda, Y., Sadahiro, T. and Yamakita, M. 2011. FPGA Implementation of digital differentiator using Richardson extrapolation and high sampling rate acting like fractional delay, SICE Annual Conference, Waseda University, Tokyo, Japan, September 13–18, pp. 2378–2383.

14. Caponetto, R., Dongola, G. and Gallo, A. 2009. Fractional integrative operator and its FPGA implementation, in: Proceedings of the ASME 2009 International Design Engineering Technical Conferences & Computers and Information in Engineering Conference (IDETC/CIE 2009), San Diego, California, U.S.A., August 30–September 2, pp. 1–7.

15. Qu, L., Hu, H. and Huang, Y. 2010. Fractional order P.I.D. controller based on particle swarm optimization implemented with FPGA, *In:* Proceedings of the International Conference on Artificial Intelligence and Computational Intelligence (AICI-2010), IEEE Computer Society, 23–24 Oct, pp. 165–169.

16. Podlubny, I. 1999. Fractional-order systems and PI Dλ µ controllers. *IEEE Transactions on Automatic Control,* **44**: 208–214.

17. Domingues, J., Valerio, D. and da Costa, J.S. 2009. Rulebased fractional control of an irrigation canal. In Proceedings 35th annual conference of IEEE industrial electronics IECON '09 (pp. 1712–1717).

18. Ahn, H.S., Bhambhani, V. and Chen, Y.Q. 2009. Fractional-order integral and derivative controller for temperature profile tracking. *Sadhana,* **34**: 833–850.

19. Xue, D.Y., Zhao, C.N. and Chen, Y.Q. 2006. Fractional order P.I.D. control of a DC-motor with elastic shaft: A case study. In Proceedings American control conference.

20. Tehrani, K., Amirahmadi, A., Rafiei, S., Griva, G., Barrandon, L., Hamzaoui, M., Rasoanarivo, I. and Sargos, F. 2010. Design of fractional order P.I.D. controller for boost converter based on multi-objective optimization. In 14th International power electronics and motion control conference (EPE/PEMC) (pp.179–185).

21. Changmao, Q., Naiming, Q. and Zhiguo, S. 2010. Fractional P.I.D. controller design of hypersonic flight vehicle. In international conference on computer, mechatronics, control and electronic engineering (CMCE), pp. 466–469.

22. Fan, H., Sun, Y. and Zhang, X. 2007. Research on fractional order controller in servo press control system. In International Conference on Mechatronics and Automation, ICMA, pp. 2934 –2938.

23. Bouafoura, M.K. and Braiek, N.B. 2010. PI Dλ µ controller design for integer and fractional plants using piecewise orthogonal functions. *Communications in Nonlinear Science and Numerical Simulation,* **15**: 1267– 1278.

24. Podlubny, I. 1999. Fractional-order systems and PI Dλ µ controllers. *IEEE Transactions on Automatic Control,* **44**: 208–214.

25. Valéio, D. and da Costa, J.S. 2006. Tuning of fractional P.I.D. controllers with Ziegler– Nichols-type rules. Signal Processing, 86, 2771–2784. Special Section: Fractional Calculus Applications in Signals and Systems.

26. Zamani, M., Karimi-Ghartemani, M., Sadati, N. and Parniani, M. 2009. Design of a fractional order P.I.D. controller for an A.V.R. using particle swarm optimization. Control Engineering Practice: Vol. 17. In Special Section: The 2007 IFAC symposium on advances in automotive control, pp. 1380–1387.

27. Lee, C.H. and Chang, F.K. 2010. Fractional-order P.I.D. controller optimization via improved electromagnetism-like algorithm. *Expert Systems with Applications,* **37**: 8871–8878.

28. Biswas, A., Das, S., Abraham, A. and Dasgupta, S. 2009. Design of fractional-order PI Dλ µ controllers with an improved differential evolution. Engine.

29. Cervera, J., Banos, A., Monje, C. and Vinagre, B. 2006. Tuning of fractional P.I.D. controllers by using QFT. In 32nd annual conference on IEEE industrial electronics, IECON, pp. 5402–5407.

30. Monje, C.A., Vinagre, B.M., Chen, Y.Q., Feliu, V., Lanusse, P. and Sabatier, J. 2004. Proposals for fractional PI Dλ μ tuning. In 1st IFAC workshop on fractional derivatives and applications, Bordeaux, France. 1st IFAC workshop on fractional derivatives and applications, Bordeaux, France.

31. Yeroglu, C. and Tan, N. 2011. Note on fractional-order proportional – integral – differential controller design. *I.E.T. Control Theory Appl.,* **5**(17): 1978 – 1989.

32. Chestnut, H. and Mayer, R.W. 1959. Servomechanisms and regulating system design. John Wiley & Sons, INC., 2" Edition.

33. Boiko, I., Fridman, L. and Castellanos, I.M. 2004. "Analysis of second-order sliding-mode algorithms in the frequency domain," at *IEEE Trans. Autom. Control,* 49(6): 946950.

34. Jiang Jing, 2007. Sliding-mode variable structure control for the position tracking servo system WSEAS *Transactions on Systems,* **6**(2): 294-297.

35. Jaferov, E.M., Parlakc, M.N.A. and Istcfanopulos, Y. 2005. "A new variable structure PID-controller design for robot manipulators," *IEEE Trans. Control Sys. Tech.,* **13**: 122-130.

36. White, B.A., Lipczynski, R.T. and Daniels, A.R. 1983. "A simple digital control scheme for a D.C. motor," *I.E.E. proc. B, Electr. Power Appl.*, **130**(2): 143-147.

37. Uma maheswararao. Ch, Y.S. Kishore Babu, K. Amaresh, 2011. "Sliding mode speed control of a D.C. motor," 2011 International Conference on Communication Systems and Network Technologies (CSNT), pp. 387-391, 3-4, June 2011, Katra, Jammu, India.

38. Utkin, V.I. 1977. "Variable structure systems with sliding modes," IEEE Trans., AC22, no. 2, pp. 212-222.

39. Bengiamin, N.N. and Kauffinann, B. 1984. "Variable structure position control," *IEEE Trans., Control System Magazine,* **4**(3): 3-8.

40. Chan, W.C. and Hsu, Y.Y. 1981. Automatic generation control of interconnected power system using variable-structure controllers. *I.E.E. proc, C,Gen. Trans. & Distrib.,* **128**(5): 269-279.

41. Chan, W.C. and Hsu, Y.Y. 1983. "An optimal variable structure stabilizer for power system stabilization, *" IEEE Trans., PAS-102, pp.* 1738-1746.

42. Bengiamin, N.N. and Chan, W.C. 1982. "Variable structure control of electric power generation, " IEEE Trans., PAS-101, pp.376-380.

43. Okyay Kaynak, Kemalettin Erbatur, and Meliksah Ertugrul, 2001. "The fusion of computationally intelligent methodologies and sliding-mode control—a survey," *IEEE Transactions on Industrial Electronics,* **48**(1): 4-17.

44. Young, K.D. 1978. "Controller design for a manipulator using theory of variable structure systems, "*IEEE Trans., MSC-8,* no.2, pp.101-109.

45. Zinober, A.S. (Ed.), 1993. Variable structure and Lyapunov control, Springer-Verlag, London, U.K. (1993)

46. Spurgeon, S. 1991. Hyperplane design techniques for discrete-time variable structure control system. *Int. J. Contr.,* **55**: 445-456.

47. Slotine, J.J., Sastry, S.S. 1983. Tracking control of non-linear systems using sliding surfaces with application to robot manipulator. *Int. J. Contr.,* **33**(2): 465-492.

48. Edwards, C. and Spurgeon, S.K. 1998. Compensator based output feedback sliding mode controller design. *Int. J. Contr.,* **71**: 601-614.

49. Itkis, U.1976. Control systems of variable structure, John Wilely & Sons.

50. Wang, S., Habibi, S. and Burton, R. 2006. "Sliding mode control for a model of an electrohydraulic actuator system with discontinuous non-linear friction", The Proceedings of 2006 American Control Conference, Minneapolis, Minnesota U.S.A., pp. 5898-5904.

51. Harifi, A., Aghagolzadeh, A., Alizadeh, G. and Sadeghi, M. 2005. "Designing a sliding mode controller for antilock brake system", International Conference on Computer as a Tool, Serbia and Montenegro, pp. 611-616.

52. Sadati, N. and Ghadami, R. 2008. "Adaptive multi-model sliding mode control of robotic manipulators using soft computing," *Neurocomputing,* **71**(2): 27022710.

53. Wai, R. and Chang, L. 2006. Adaptive stabilizing and tracking control for a non-linear inverted-pendulum system via sliding-mode technique. *IEEE Trans. Ind. Electron.,* **53**: 674692.

54. Liang, H. and Chong, K.T. 2006. "Variable parameter sliding controller design for vehicle brake withwheel slip," *J. Mech. Sci. Technol.,* **20**(11): 18011812.

55. Xie, W.F. 2007. "Sliding-mode-observer-based adaptive control for servo actuator with friction," *IEEE Trans. Ind. Electron.,* **54**(3): 1517-1527.

56. Hazim H. Tahir, Ali A. A. Al-Rawi, 2006. "Variable structure control design of process plant based on sliding mode approach," Proceedings of the 25 Chinese Control Conference, pp. 932-937, 7-11 August 2006, Harbin, Heilongjiang, China.

57. Utkin, V.I. 1977. Variable Structure systems with Sliding Modes. *IEEE Transaction on Automatic Control,* **22**(2): 212-222.

58. Raymond A. DeCarlo, Stanislaw H. Zak, and Gregory Mathews, 1988. "Variable Structure Control of Nonlinear Multivariable Systems: A Tutorial", *Proceedings of the IEEE,* **76**(3).

59. Hung, J.Y., Gao, W. and Hung, J.C. 1993. "Variable Structure Control: A Survey", *IEEE Trans. on Industrial Electronics,* **40**(1).

60. Slotine, J. and Li, W. 1991. Applied Nonlinear Control. New Jersey: Prentice-Hall; p. 461.

61. Shafiei, S.E. 2010. Sliding mode control of robot manipulators via intelligent approaches. In: Shafiei SE, editor. Advanced Strategies for Robot Manipulators. Rijeka: IntechOpen; pp. 135-172.

62. Yu, X. and Kaynak, O. 2009. Sliding-mode control with soft computing: A survey. *IEEE Transactions on Industrial Electronics,* **56**(9): 3275-3285.

63. Utkin, V.I. 1977. Variable structure systems with sliding modes. *IEEE Transactions on Automatic Control,* **22**(2): 212-222.

64. Tannuri, E.A., Agostinho, A.C., Morishita, H.M. and Moratelli, L. 2010. Dynamic partitioning systems: An experimental analysis of sliding mode control. *Control Engineering Practice,* **18**(10): 1121-1132.

65. Young, K.D., Utkin, V.I. and Ozguner, U. 1999. A control engineer's guide to sliding mode control. *IEEE Transactions on Control Systems Technology,* **7**(3): 328-342.