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A New Whale Optimization Algorithm-Based Fault Location Method by Focusing on Dispersed Model of the Transmission Line

Hossein Faramarzi¹, Navid Ghaffarzadeh²†

¹,² Faculty of Technical and Engineering, Imam Khomeini International University, Qazvin, Iran.

ABSTRACT

In this paper, a fault location approach is presented by using the Whale Optimization Algorithm (WOA) strategy in two terminal transmission feeders. Also, the Grey Wolf Optimization (GWO) method is discussed. Voltage and current are measured in both ends to collect the data required for the proposed strategy. The paper considers several types of faults and simulations, and the objective function identifies the fault location with a high accuracy in a short time. In addition, based on distributed model of the line, the fault location is defined and the optimization algorithm does not utilize the compressed model of the line, and the calculations are highly accurate. The WOA-based optimization method results in a notable reduction in the computational time. As the benefit of the proposed technique, accurate and timely location of the source of the fault is highly helpful to the repair crew. Almost in all cases, the accuracy of the proposed procedure is very high, and the error is kept below 1%.

Article Info

Keywords: Bergeron model in time domain, Dispersed model of the line, Fault location technique, Grey wolf optimization algorithm, Whale optimization algorithm.

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I. INTRODUCTION

Basic goal of power system is to continuously provide electrical energy to the users. Like any other system, failures may occur in a power system. After detecting fault condition and location, it is critical to apply correct remedial actions. Accurate determination of fault location and its condition are important. Since the visual search of faulted lines are costly and sometimes inconclusive, fast and precise designation of fault location is necessary for prompt restoration of power, particularly on transmission lines, which would lead to saving on time and resources for the electric utilities. This will help field personnel to figure out the fault locations from transmission line maps and drawings. As a result, power system operators can identify and isolate faulted sections by opening circuit breakers or switches correctly and timely. Subsequently, the power is restored to healthy transmission sections, and the customers are supplied reliably because long outages on the customer side are prevented and quick detection of faults helps beneficiaries to tackle them as fast as possible. It is worth noting that to increase the accuracy of calculations, line modeling plays a key role in the fault location process. Using distributed model of the line is one of the prevalent approaches in fault location optimization problems. Various methods have been proposed in the literature and have extensively been used to improve the solution of an optimal fault location problem. Linear programming (LP), nonlinear programming (NLP) mixed nonlinear techniques are just some examples. These methods have been classified and their advantages, limitations and requirements have been discussed in detail. One essential aspect of these methods is how to model the transmission line because it changes the accuracy of the calculations. Some
researches have focused on modeling lines with a compressed model, which reduces the accuracy of the calculations. So, to improve the precision of calculations, it is inevitable to use more comprehensive models of the transmission line. One of these accurate models is the distributed model of lines. According to the distributed time domain model of the lines, Ghiafeh Davoudi and Sadeh [1] proposed a method to detect the fault location using post-fault voltage and current samples at both terminals in which the objective was minimized by the genetic algorithm. Kezunovic and Mrkic [2] introduced a new fault location method based on the information of two end systems. It considers two different algorithm constructions utilizing two line models, and simulation performance is achieved by using EMTP software. It uses time sampling method for the proposed approach and it is concluded as the proposed method is cost-effective. Also, an algorithm is introduced by Sadeh [3] that uses time domain line model which is compensated with series connected FACTS device. It improves accuracy. An application of this technique can be developed to any series FACTS compensated line. Due to the calculation of the location and resistance of the fault, samples of voltage and current at both ends of the line are used synchronously. Ghazizadeh [4] suggested a novel arcing faults location technique for multi-segment combined transmission lines, which uses unsynchronized measurements from two terminals of the line. Also, Junzhang and Zhonghui [5] showed that fault location accuracy has been reduced by ignoring the line distributed capacitance. In document Ahmed and Attia [6], some optimization algorithms like teaching-learning-based optimization and harmony search algorithms are used. Also, some methods such as genetic algorithm, artificial bee colony, artificial neural networks and cause and effect are discussed along with the advantages and disadvantages of all methods. In Youssef [7], an accurate method is discussed about the calculations needed to find the exact fault location on transmission lines. It used travelling wave-based method. The method is unaffected by noise or spurious changes in line. Salehi [8] presents a closed-form solution for fault location that does not need the GPS-synchronized sampling of wide-area measurements. Sparse intelligence is used to record the unsynchronized measurement. Also, the Schur-Banachiewicz inversion formula is utilized to obtain a solution for fault location.

From the previous efforts, we can conclude that these approaches have many contrasts, especially as to points such as the model of transmission line, measured variables from one side or both sides, and the optimization strategies employed to get the best results. Since the fault location is not derived directly from equations related to the currents and voltages, more smart methods are used to detect fault location in transmission systems.

Based on the goal, the present paper focuses on the operational use and application of a pre-existing approach, which has been already described in the literature. In this paper, some structures like the model of the transmission line, optimization algorithm, and cost function are updated and all information about the model used in locating the fault are fully described. Here, our approach to find the fault location is as follows. First, the objective function along time interval is defined. Second, to perform the fault location (FL) calculations, samples of currents and voltages are taken. Then, the optimal solution of the problem is found. We also vary the accurate fault location to measure the advantages of the proposed approach. The results are compared with the results obtained by the GWO algorithm. The main objective of this paper is to find out the best solution for the problem using WOA. The results show that by using a distributed model of the line and information which are sent from PMUs, under the dynamic operation, online employment of the approach is feasible. In this paper, we calculate the location of different types of faults in the category shown in Fig. 1.

![Fig. 1. Classification of faults.](image-url)
A distributed model of transmission line (segments S to F) is shown in Figs. 4-5. A set of equations depict the relevance between voltages and currents to send and receive end buses as follows (Bergeron’s equations) [9, 10]:

\[ i_s(t) = \frac{1}{Z'_{ssS}} u_s(t) + I_s(t - \tau_{xs}) \]  
\[ i_x(t) = \frac{1}{Z'_{ssS}} u_x(t) + I_x(t - \tau_{xs}) \]

where the dependent source currents can be described by Eq. (3)- (4).

\[ I_s(t - \tau_{xs}) = \frac{4}{Z'_{ssS}} u_s(t - \tau_{xs}) + Z'_{ssS} i_s(t - \tau_{xs}) \]  
\[ I_x(t - \tau_{xs}) = \frac{4}{Z'_{ssS}} u_x(t - \tau_{xs}) + Z'_{ssS} i_x(t - \tau_{xs}) \]

where \( Z_s \) is the surge impedance of the line and \( R_{xs} \) is the resistance of the SF section.

\[ Z'_{ssS} = Z_s + \frac{R_{xs}}{4} \]  
\[ Z'_{ssS} = Z_s - \frac{R_{xs}}{4} \]

Where \( \tau_{xs} \) is the time required for the wave to traverse from S to F. \( T \) is the time required for the wave to traverse from S to R. In a single-phase Bergeron transmission line of two buses, if the fault takes place at point F, from above equation we can conclude that as a function of measured quantities at terminals S and R, the voltage at fault point can be extracted where \( R_{XR} \) is the resistance of the RF section.

\[ Z'_{ssR} = Z_s + \frac{R_{XR}}{4} \]  
\[ Z'_{ssR} = Z_s - \frac{R_{XR}}{4} \]

The part of transmission line from receive end to fault (F) point is named RF section. At the fault point, the voltage can be calculated using the quantities measured at terminals S and R. The following equations describe it.
\[ u_{\text{xs}}(t) = \left\{ \frac{1}{2Z_{\text{xs}}}Z_{\text{xs}}^2\left[u_{\text{s}}(t + \tau_{\text{xs}}) - Z_{\text{xs}}'i_{\text{s}}(t + \tau_{\text{xs}})\right] + \frac{1}{2Z_{\text{xs}}}Z_{\text{xs}}^2\left[u_{\text{s}}(t - \tau_{\text{xs}}) - Z_{\text{xs}}'i_{\text{s}}(t - \tau_{\text{xs}})\right] \right\} \]

\[ u_{\text{xs}}(t) = \left\{ \frac{1}{2Z_{\text{xs}}}Z_{\text{xs}}^2\left[u_{\text{s}}(t + (T - \tau_{\text{xs}})) - Z_{\text{xs}}'i_{\text{s}}(t + (T - \tau_{\text{xs}}))\right] + \frac{1}{2Z_{\text{xs}}}Z_{\text{xs}}^2\left[u_{\text{s}}(t - (T - \tau_{\text{xs}})) + Z_{\text{xs}}'i_{\text{s}}(t - (T - \tau_{\text{xs}}))\right] \right\} \]

Based on the Bergeron model, the distributed parameters are characterized by the surge impedance and phase velocity. To study this, we can define the objective function based on the fact that, at fault point, voltage is the same as whatever calculated from sending or receiving end.

### III. Optimal Fault Location Approach

This section formulates an optimization model to determine the solution of fault location problem. To get the solution of the optimization problem, it is necessary to determine an objective function which deals with decision variable. The goal is to obtain the distance between fault point and sending line. It is necessary to use an optimization technique to solve the fault allocation problem in such a way that the objective function is minimized. The formulation of the objective function is based on the voltage difference, considering \( n_{\text{xs}}, k \) as discrete variable.

#### A. Objective Function

The objective function is based on partial differential equations of the transmission line model which has two variables: position and time. By placing the measured voltage and current as boundary conditions in those functions, fault location can be calculated. To find the fault location, the key idea of this work is based on minimizing the voltage differential at fault point.

**1) Minimization of Voltage differential at Fault Point**

With the measured quantities at S and R terminals and using Eq. (1) and (2), we can conclude the following equation. Since the voltage of the fault point should be singular, the voltage differential at point F should be held at zero.

\[ F(u_{\text{xs}}, i_{\text{xs}}, u_{\text{ir}}, i_{\text{ir}}, t, \tau_{\text{xs}}) = \left| u_{\text{xs}}(t, \tau_{\text{xs}}) - u_{\text{ir}}(t, \tau_{\text{xs}}) \right| \]

Eq. (11) must be reached to its least value. It should be noted that the discretization of the measured voltage and current at discrete moments must be considered.

\[ F(k, x) = \left| u_{\text{xs}}(k, x) - u_{\text{ir}}(k, x) \right| \]

The fault point \( X_F \) is estimated by scanning the minimum value of the absolute difference between the fault voltage seen by both end. The scanning of \( F(k, x) \) for each \( \Delta t \) starts at \( t_{\text{Start}} = K_0 \Delta t \) and proceeds until the protection system pick-up at \( t_{\text{Stop}} = K_1 \Delta t \).

The mean value over time of \( F(k, x) \) leads to a better understanding of fault voltage behavior during the whole fault period. Its calculation is as follows:

\[ \text{Min (obj)} = \min \left( \frac{1}{\Delta t(k_1 - k_0)} \sum_{k=k_0}^{k-k_1} F(k, x) \right) \]

\[ k = \frac{1}{\Delta t} \]

\[ n_{\text{xs}} = \frac{\tau_{\text{xs}}}{\Delta t} \]

\[ x = C \times \tau_{\text{xs}} \]

\[ C = 3 \times 10^8 \text{ m/s} \]

\[ \Delta t: \text{Sampling step.} \]

\[ n_{\text{xs}}, k \text{ arbitrary integers.} \]

In this paper, the objective function is set at its minimum value using the WOA and GWO algorithms and the results are compared with each other. These algorithms are described in Section 5.

**2) Constraint**

In this paper, the inequality constraint for distance, \( x \), is given by Eq. (17) in which \( x \) is the distance from sending end to the fault point.

\[ 0 \leq x \leq L \]

### IV. Modal Transformation

The coupled direct equations in the three-phase transmission line are as below:

\[ \frac{\partial v(x, t)}{\partial x} = -L \frac{\partial i(x, t)}{\partial t} \]

\[ \frac{\partial i(x, t)}{\partial t} = -C \frac{\partial v(x, t)}{\partial t} \]

To get rid of the mutual effects, a transformation should be used to eliminate the foresaid mutual effects. Modal transformation can decompose the coupled equations into decoupled ones. By determining the equation in modal domain that is similar to the equation for a single-phase transmission line, decoupling process is terminated. A commonly used modal transformation matrix is defined as:

\[ M = \begin{bmatrix} 1 & 1 & 1 \\ 1 & -2 & 1 \\ 1 & 1 & -2 \end{bmatrix} \]

The phase quantities will be transformed to each other by using \( M \) matrix.

\[ I_{\text{ph}} = M \times I_{\text{m}} \]

\[ I_{\text{m}} = M^{-1} \times I_{\text{ph}} \]

Eq. (23) can be extracted by Eq. (22).
\[
\begin{bmatrix}
I_0 \\
I_1 \\
I_2
\end{bmatrix} = \frac{1}{3}
\begin{bmatrix}
1 & 1 & 1 \\
1 & -1 & 0 \\
1 & 0 & -1
\end{bmatrix}
\begin{bmatrix}
I_a \\
I_b \\
I_c
\end{bmatrix}
\]

(23)

in which \(I_0\) is ground mode current and \(I_1\) and \(I_2\) are the aerial mode. Aerial mode 1 will be used in the fault location procedure if the aerial modes have non-zero values in all fault types. Similarly, the voltages can be derived in modal domain. The equations governing the system were derived based on these equations, and finally, the first aerial mode is used to determine the fault location in three phase transmission line. In this paper, since fault location is determined by minimizing the objective function, two different optimization algorithms are considered. The objective function is defined in Eq. (13). To assess fault location, there is a need to use an optimization algorithm with some features such as having fewer control parameters and shorter computational time, being simple and fast to converge, and having an ability to explore wider search area. In the present work, the WOA and GWO algorithms are initially used to optimally locate the fault. The use of these algorithms is based on the following reasons. First, the algorithm is universal for problem-solving and does not need to transform the problem as a linear and mix-integer model. Second, we define some constraints to the problem. These algorithms are swarm-based methods which are driven from the collective behavior of social creatures. The implementation of swarm-based algorithms is easier than the evolutionary-based algorithms because they include fewer operators (i.e., selection, crossover, mutation). Also, these algorithms have lower input parameters as compared to evolutionary-based algorithms, like genetic algorithm. These methods use time domain representation of the signal and distributed-parameter model of the transmission line. Two approaches solve partial equations using optimization algorithm methods. They require the solution of partial differential equations.

A. Whale Optimization Algorithm (WOA)

In 2016, Mirjalili and Lewis developed an optimization algorithm which named Whale Optimization Algorithm (WOA). It is inspired by the hunting behavior of humpback whales in response to the search for food in the nature [11]. The main interesting point of these whales is how they hunt humpbacks. In WOA, each solution is thought to be a whale. In this solution, a whale tries to replete a new place in the search space considered as a reference the best element of the group. Two mechanisms are used by the whales to locate their prey and attack it. In the first one, the preys are encircled and the second creates bubble nets. Regarding optimization, when the whales look for a prey, the search space is explored and the exploitation occurs during the attack behavior. Random search and local search are two main characteristics of WOA. They play an important role to get the highest capability in solving the optimization problem. WOA has some good features, like simplicity, reliability, robustness, and flexibility. As already mentioned, the operational steps of this algorithm include the following ones: encircling the target, bubble-net attacking method, and searching for the target. Ashraf Darwish [12] and Xiaofei Wang and Hui Zhao [13] have described it in details.

B. Grey wolf Optimization (GWO)

The GWO algorithm is one of the recent meta-heuristic algorithms which is based on hunting and social leadership of grey wolves (Ashraf Darwish [12]). It was proposed by Iranian scholar Mirjalili in 2014. Gray wolves usually live in groups, and under the leadership of a head grey wolf, the wolves capture the prey through a series of processes, such as surrounding, hunting and attacking. In this algorithm, attaining the results is centered on three best grey wolves. The leader of the group is called alpha and is responsible for some activities such as making decisions about sleeping place and hunting [12, 13]. The second wolf is called beta, and he helps the wolf alpha in making decisions. The third grey wolf is called omega and is responsible for providing information to all the other wolves. All the other remaining gray wolves are called delta. They are responsible for dominating the omega. The main phases of the GWO algorithm are based on the following steps [12]:

- Tracking, chasing and approaching the prey.
- Pursuing, encircling and harassing the prey.
- Attacking the prey.

More details on this algorithm are available in [13]. In this paper the performance of our approach is evaluated by the GWO algorithm.

C. Flowchart

More details about the proposed approach is depicted in Fig. 6. In this paper, Eq. (13) is selected as the fitness function.
It is shown that in the optimization procedure, whenever the initial solutions are generated, the fitness function is evaluated for each solution. The solution that has the best fitness function (lower value of Eq. (13)) is used to update the current solutions. This step is repeated until the breaking rule is met. Finally, among all the solutions, the one with the best value of the fitness function is selected as the optimum solution of the problem.

V. SIMULATED CASE STUDY

In order to get the exact fault location, two different optimization algorithms are considered in this paper. It should be noted that there is an effective factor for any method to attain the location of the fault. This is mis-locality in the consequences, which is defined by Eq. (24).

\[
E_{FL} = \frac{X_{\text{measured}} - X_{\text{real}}}{L_{sr}} \times 100
\]

where \(X_{\text{measured}}\) is the measured location of a fault, \(X_{\text{real}}\) is the real location of the fault, and \(L_{sr}\) is the totality of line length. This equation provides a metric to analyze the accuracy of the proposed fault location methods.

VI. RESULTS AND DISCUSSION

A series of simulation studies were conducted to evaluate the performance of the approach using MATLAB/Simulink. The test system is shown in Fig. 7. The voltage of the system was 400 kV and the length of line was 120 km. The variables of this line are presented in Table 1. A fault occurs at point F with an distance of X km from the end bus (S) after the simulating results are obtained. Optimization would be done by using the WOA and GWO algorithms. The maximum number of iterations and initial population size were set at 500 and 100, respectively. In order to avoid the fluctuations of the performance, we repeated the algorithm for 20 times. After finding the voltage and current waveforms in the modal domain, optimization algorithm was utilized to locate the fault. The results were illustrated for comparison. A three-phase fault to ground has occurred on section SR without any resistance to ground.

The fault occurrence time was 0.02 s and its clearance time was .04 s. The voltage and current waveforms were achieved at buses S and R as shown in Figs. 9-12, respectively. Also, their adaptive waveforms in the modal domain are depicted in Figs. 13-16. These figures illustrate the voltage and current change during the time interval between 0-2 s. The average operation time of the WOA and GWO algorithms were 10.125 min and 1.957 min, respectively. So, based on the time value to achieve the result for all types, WOA took more time to optimize the objective function. It may be regarded as a bad attainment. A comparison between WOA and GWO was done to verifying which algorithm was better than the other.
Fig. 9. Three-phase voltage waveforms of terminal S.

Fig. 10. Three-phase voltage waveforms of terminal R.
Fig. 11. Line current waveforms with measurement at terminal S.

Fig. 12. Line current waveforms with measurement at terminal R.
Fig. 13. Three-phase voltage waveforms in modal domain for terminal S.

Fig. 14. Three-phase voltage waveforms in modal domain for terminal R.
Fig. 15 Line current waveforms in modal domain with measurement at terminal S.

Fig. 16 Line current waveforms in modal domain with measurement at terminal R
### TABLE II
Attained solutions for WOA at different locations and resistances of fault = 0 ohm.

<table>
<thead>
<tr>
<th>Fault type</th>
<th>Fault real location</th>
<th>Fault location measured by WOA</th>
<th>Error % in WOA</th>
<th>Error % in GWO</th>
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Fig. 17. Error profile at different fault location with resistance of fault = 0 ohm.

### TABLE III
Attained solutions for WOA at different locations and resistances of fault = 5 ohm.

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<tr>
<th>Fault type</th>
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<th>Fault location measured by WOA</th>
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<td>ABG</td>
<td>5</td>
<td>5.1837</td>
<td>0.1581</td>
<td>0.1685</td>
</tr>
<tr>
<td></td>
<td>50</td>
<td>50.615</td>
<td>0.5125</td>
<td>0.7192</td>
</tr>
<tr>
<td></td>
<td>100</td>
<td>99.981</td>
<td>-0.0158</td>
<td>-0.9208</td>
</tr>
<tr>
<td></td>
<td>115</td>
<td>115.967</td>
<td>0.8058</td>
<td>0.8175</td>
</tr>
<tr>
<td>AB</td>
<td>5</td>
<td>4.951</td>
<td>-0.0408</td>
<td>-0.1092</td>
</tr>
<tr>
<td></td>
<td>50</td>
<td>50.361</td>
<td>0.3008</td>
<td>0.5084</td>
</tr>
<tr>
<td></td>
<td>100</td>
<td>100.83</td>
<td>0.6917</td>
<td>0.8142</td>
</tr>
<tr>
<td></td>
<td>115</td>
<td>114.583</td>
<td>-0.3475</td>
<td>-0.4192</td>
</tr>
</tbody>
</table>

Fig. 18. Error profile at different fault location with resistance of fault = 5 ohm.
TABLE IV
ATTAINED SOLUTIONS FOR WOA AT DIFFERENT LOCATIONS AND RESISTANCES OF FAULT = 15 OHM.

<table>
<thead>
<tr>
<th>Fault type</th>
<th>Fault real location</th>
<th>Fault measured location by WOA</th>
<th>Fault measured location by GWO</th>
<th>Error % in WOA</th>
<th>Error % in GWO</th>
</tr>
</thead>
<tbody>
<tr>
<td>ABCG</td>
<td>5</td>
<td>5.2641</td>
<td>5.2798</td>
<td>0.2201</td>
<td>0.2332</td>
</tr>
<tr>
<td></td>
<td>50</td>
<td>50.135</td>
<td>50.261</td>
<td>0.1125</td>
<td>0.2175</td>
</tr>
<tr>
<td></td>
<td>100</td>
<td>99.893</td>
<td>99.6029</td>
<td>-0.0892</td>
<td>-0.3309</td>
</tr>
<tr>
<td></td>
<td>115</td>
<td>115.14</td>
<td>115.301</td>
<td>0.1167</td>
<td>0.2508</td>
</tr>
<tr>
<td>AG</td>
<td>5</td>
<td>5.0370</td>
<td>5.1470</td>
<td>0.0308</td>
<td>0.1225</td>
</tr>
<tr>
<td></td>
<td>50</td>
<td>50.218</td>
<td>50.4011</td>
<td>0.1817</td>
<td>0.3342</td>
</tr>
<tr>
<td></td>
<td>100</td>
<td>100.063</td>
<td>100.264</td>
<td>0.0525</td>
<td>-0.2300</td>
</tr>
<tr>
<td></td>
<td>115</td>
<td>115.271</td>
<td>115.740</td>
<td>0.2258</td>
<td>0.6167</td>
</tr>
<tr>
<td>ABG</td>
<td>5</td>
<td>5.1837</td>
<td>5.1989</td>
<td>0.1531</td>
<td>0.1658</td>
</tr>
<tr>
<td></td>
<td>50</td>
<td>50.615</td>
<td>50.863</td>
<td>0.5125</td>
<td>0.7192</td>
</tr>
<tr>
<td></td>
<td>100</td>
<td>99.981</td>
<td>98.795</td>
<td>-0.0158</td>
<td>-1.0042</td>
</tr>
<tr>
<td></td>
<td>115</td>
<td>115.567</td>
<td>115.605</td>
<td>0.4725</td>
<td>0.5042</td>
</tr>
<tr>
<td>AB</td>
<td>5</td>
<td>4.6821</td>
<td>4.3979</td>
<td>-0.2649</td>
<td>-0.5018</td>
</tr>
<tr>
<td></td>
<td>50</td>
<td>50.539</td>
<td>50.581</td>
<td>0.4492</td>
<td>0.4842</td>
</tr>
<tr>
<td></td>
<td>100</td>
<td>100.795</td>
<td>100.890</td>
<td>0.6625</td>
<td>0.7417</td>
</tr>
<tr>
<td></td>
<td>115</td>
<td>114.476</td>
<td>114.364</td>
<td>-0.4367</td>
<td>-0.5300</td>
</tr>
</tbody>
</table>

Tables II-IV show the comparison between the WOA and GWO algorithms. Based on different locations and fault resistance, the results of all fault types are illustrated in these tables. It is clear that WOA and GWO show the results accurately in spite of the fault type and fault resistance.

Figs. 17-19 show the error profile of the used algorithms at different fault locations. The analysis revealed a strong capability of the proposed approach (%error<1). It can be seen that the WOA algorithm has a higher capability than the other one. Also, Figs. 20-23 illustrate the effect of fault resistance on error value in different fault locations. As is shown, the fault resistance affects detection accuracy and %error. Although it is true for all fault types wherever the fault is, fault location is detected with minimal error values. The same has been included in Tables I-IV.

VII. CONCLUSIONS

A new correct fault location algorithm was used to compute the correct location of fault based on the Bergeron model of transmission line. In order to clarify the idea about the proposed method used in this paper, some factors such as simulation time, number of inputs, and rules are considered. Based on these factors, the complexity level is determined. In this article, some simulations were performed in different conditions, and the performance of the proposed method was compared with the GWO method. The aforesaid conditions are as follows: symmetrical and unsymmetrical faults, different fault locations, and different fault resistance. The results indicated the superiority of the proposed method in all cases. The main advantages of the proposed method are as follows:

- High accuracy of calculations
- High operating speed and low computational time
- Suitable to search for the solutions
- Low complexity of implementation
- Correct operation in different conditions of fault

Based on these advantages, the concerns on utilities about the service interruptions and down times are minimized by the proposed method. In this work, it is supposed that the location of PMU is known. In future works, the allocation of PMUs is recommended in the process of the problem definition.

Fig. 19. Error profile at different fault locations with resistance of fault = 15 ohm.
Fig. 20. Error value at different fault resistance for ABCG fault

Fig. 21. Error value at different fault resistance for AG fault

Fig. 22. Error value at different fault resistance for ABG fault.

Fig. 23. Error value at different fault resistance for AB fault.
REFERENCES


Hossein Faramarzi received his B.Sc. degree in Electrical Engineering from Shahed University, Tehran, Iran in 2009 and his M.Sc. degree in Electrical Engineering from Imam Khomeini International University, Qazvin, Iran in 2012. He is currently a Ph.D. candidate in Electrical Engineering at Imam Khomeini International University. His research interests cover power system optimization, energy hub management, power system protection, and renewable energy systems.

Navid Ghaffarzadeh is an associate professor in Department of Electrical Engineering at Imam Khomeini International University, Qazvin, Iran. His main areas of research interest are smart grid, intelligent systems, optimization applications in power systems, power system protection, renewable energy, and power quality. He has over 100 technical publications and has written seven books on power systems.
A Semi-Analytic Method for Solving a Class of Non-Linear Optimal Control Problems
Pooneh Omidiniya\textsuperscript{1} and Maryam Alipour\textsuperscript{2,†}

\textsuperscript{1,2} Department of Mathematics, Statistics and Computer Science, University of Sistan and Baluchestan, Zahedan, Iran.

Abstract
This paper, proposes an approximate analytical method to solve a class of optimal control problems. This method is an enhancement of the variational iteration method (VIM) that named modified variational iteration method (MVIM) and eliminates all additional calculations in VIM, thus requires less time to do the calculations. In this approach, first, the optimal control problem is converted into a nonlinear two-point boundary value problem via the Pontryagin's maximum principle, and then we applied the MVIM method to solve this boundary value problem. This suggested method is suitable for a large class of nonlinear optimal control problems that for the non-linear part of the problem, we used the Taylor series expansion. In the end, three examples are provided to demonstrate the simplicity and efficiency of the method. Numerical results of the proposed method versus other methods is presented in tables. All calculations were carried out using Mathematica software.

Article Info

Keywords:
Differential equations, Modified variational iteration method (MVIM), Numerical solution, Optimal control problems

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I. Introduction
Recently, non-linear optimal control problems (NOCPs) have been applied in different fields such as aircraft systems [1, 2], biomedicine [3] and robotics [4]. Because of the importance of this type of problem and its impact on science and engineering, researchers have shown considerable interest in this issue. Indeed, numerical approaches are usually applied to solve the problems. In direct methods, the problem can be converted into a linear or non-linear programming by using the discretization or parametrization techniques [5]. Indirect methods, on the other hand, are based on the Pontryagin’s maximum principle [6]. Various kinds of techniques have been proposed to solve NOCP’s. Shirazian et al. [7] suggested the application of the variational iteration method along with a shooting method in order to solve the extreme conditions resulting from the Pontryagin’s maximum principle. Kafash et al. [8] offered a numerical approach for solving optimal control problems (OCP) using the Boubaker polynomials expansion scheme. This approach is based on state parametrization. Indeed, the state variable is approximated by Boubaker polynomials with unknown coefficients and then performance index and boundary conditions are transformed into some algebraic equations. A new analytic technique based on VIM and some modifications was suggested to solve NOCPs in [9]. Jafari et al. [10] has been shown that the method proposed in [9] is exactly the same iterative formula as the ADM and HPM for solving NOCPs. Alipour et al. [11] introduced an approach for NOCPs which makes use of homotopy analysis and parametrization methods. Actually an appropriate parametrization of control is applied and state variables are computed using homotopy analysis method. Jajarmi et al. [12] came up with a novel analytical technique, called OHPM, to solve a class of NOCPs. In this paper, the authors argue that the proposed algorithm method has low computational complexity and fast convergence rate. A numerical technique based on the linear B-spline polynomials offered to solve OCP [13]. In this process, state and control functions are approximated in terms of B-spline functions. Also, in [14], OCPs were solved through the spectral homotopy analysis method (SHAM). SHAM is combination of
the hybrid spectral collocation technique and the homotopy analysis method. This article points out that SHAM is stronger than HAM due to it removes restrictions of the HAM such as the requirement for the solution to conform to the so-called rule of solution expression and the rule of coefficient ergodicity. In addition, more numerical methods in this area can be found in [15, 16, 17, 18, 19].

Ji-Huan He advised a method to solve non-linear differential equations using VIM [20, 21]. VIM also features a number of disadvantages which decrease its power. These are mainly associated with repetitive calculations and the introduction of excessive unnecessary terms. To overcome these downsides, Abassy et al. [22] proposed the modified variational iteration method (MVIM). Besides, MVIM improved the rate of convergence. This paper uses MVIM as a semi-analytical method to solve NOCPs. The proposed method is an indirect method and does not require discretization, linearization or transformation. In this approach, first, the optimal control problem is converted into a nonlinear two-point boundary value problem via the Pontryagin’s maximum principle, and then we applied the MVIM method to solve this boundary value problem. The examples show that our proposed method is rewarding thanks to its simplicity and small computation costs. The paper is organized as follows: Section 2. introduces NOCPs; MVIM for NOCPs is discussed in Section 3; Section 4 simulates the numerical examples to check the efficiency of the method; and Section 5 draws a number of conclusions based on the results.

II. NONLINEAR OPTIMAL CONTROL PROBLEMS

Consider the following nonlinear dynamical system:
\[ x'(t) = f(t, x(t)) + g(t, x(t))u(t), t \in [t_0, t_f] \]
\[ x(t_0) = x_0, \ x(t_f) = x_f, \]
where \( x(t) \in \mathbb{R}^n \) denotes the state variable, \( u(t) \in \mathbb{R}^m \) is the control variable, and \( x_0 \) and \( x_f \) are the given initial and final states at \( t_0 \) and \( t_f \) respectively. Moreover, \( f(t, x(t)) \in \mathbb{R}^n \) and \( g(t, x(t))u(t) \in \mathbb{R}^{n \times m} \) are two continuously differentiable functions in all arguments. Our purpose is to minimize the quadratic objective function.
\[ J(x, u) = \frac{1}{2} \int_{t_0}^{t_f} (x'(t)Qx(t) + u'(t)Ru(t)) \, dt \]
subject to the non-linear system (1), where \( Q \in \mathbb{R}^{n \times n} \) and \( R \in \mathbb{R}^{m \times m} \) are positive semi-definite and positive definite matrices, respectively. Since the performance index (2) is convex, the following extreme necessary conditions are also sufficient for optimality:
\[ x' = f(t, x) + g(t, x)u^* \]
\[ \lambda' = -H_x(x, u^*, \lambda) \]
\[ u^* = \arg \min_u H(x, u, \lambda) \]
\[ x(t_0) = x_0, \ x(t_f) = x_f, \]
where \( H(x, u, \lambda) = \frac{1}{2} [x'^TQx + u'^TRu] + \lambda^T [f(t, x) + g(t, x)u] \)
is the Hamiltonian function. Similarly (3) can be written in the form of:
\[ x' = f(t, x) + g(t, x)[-R^{-1} g'(t, x)\lambda], \]
\[ \lambda' = -(Qx + (\frac{\partial f(t, x)}{\partial x})^T \lambda + \sum_{i=1}^{n} \lambda_i [-R^{-1} g'(t, x)\lambda_i] \frac{\partial g(t, x)}{\partial x}) \]
\[ x(t_0) = x_0, \ x(t_f) = x_f \]
(4)
where \( \lambda(t) \in \mathbb{R}^n \) is the co-state vector with the \( i \)th component \( \lambda_i(t), i = 1, ..., n \), and \( g(t, x) = [g_1(t, x), ..., g_n(t, x)]^T \) with \( g_i(t, x) \in \mathbb{R}^n, \ i = 1, ..., n \). Also, the optimal control law is obtained by
\[ u^* = -R^{-1} g'(t, x)\lambda \]
(5)
There is no analytical solution for solving such a two-point boundary-value problem (TPBVP) in (4). Therefore, it is highly recommended to calculate analytic approximate or numerical solutions for it. We shall apply MVIM to solve the following initial value problem. Taylor series expansion is used here for the non-linear part of the problem.

III. THE MODIFIED VARIATIONAL ITERATION METHOD

In this section, consider the following differential equations,
\[ LV(t) + RV(t) + NV(t) = g(t) \]
\[ V(0) = f(t), \]
where \( L = \frac{d}{dt} \), \( R \) is a linear operator, \( N \) is a non-linear term and \( g(t) \) is an inhomogeneous term. Using VIM [20, 21] to solve the non-linear differential equation (6), the following variational iteration formula can be obtained:
\[ V_{n+1}(t) = V_n(t) - \int_0^t [L(V_n(\tau)) + R(V_n(\tau)) + N(V_n(\tau)) - g(\tau)] \, d\tau \]
(7)
It has been shown [22] that equation (7) is equivalent to the following equation:
\[ V_{n+1}(t) = V_n(t) - \int_0^t [R(V_n(\tau) - V_{n-1}(\tau)) + G_n(\tau) - G_{n-1}(\tau)] \, d\tau \]
(8)
where \( V_{-1} = 0, V_0 = f(t), V_1 = V_0 - \int_0^t [R(V_0 - V_{-1}) + (G_0 - G_{-1}) - g] \, d\tau \) and \( G_n(\tau) \) is obtained from \( NV_n(t) = G_n(t) + O(t^{n+1}) \). The Maclaurin series expansion is employed here for the non-linear part of the problem. Eq. (8) can be solved iteratively to obtain an approximate solution that takes the form \( V(t) \approx V_n(t) \), where \( n \) is the final iteration step.

Theorem 1. Suppose that \( V_0(t) = V_0 \) and the iterative sequence \( \{V_n(t)\} \) obtained from (7) converges to \( V(t) \); then \( V(t) \) is the solution of Eq. (6).

Proof. Considering the limits in the iterative formula (7), it follows that
\[ \lim_{n \to \infty} V_{n+1} = \lim_{n \to \infty} V_n - \int_0^t \left[ L(V_n(\tau)) + R(V_n(\tau)) + N(V_n(\tau)) \right] \, d\tau. \]
By considering $\lim_{n \to \infty} V_n = V$ and the continuity of $N$ operator, we will have
\[
\int_0^1 [L(V(t)) + R(V(t)) + N(V(t))]dt = 0.
\]
Then, differentiation of both sides concerning $t$ yields
\[
LV(t) + RV(t) + NV(t) = 0.
\]
(9)
Clearly, $V(t)$ satisfies (6). Moreover, if $t = 0$, then form (7), $V_{n+1}(0) = V_n(0)$, for every $n \geq 0$. Thus $V_0(0) = V_n(0) = V$. Hence, $V(t)$ is the solution of Eq. (6) and the proof is complete. Since the Maclaurin series is convergent, equation (8) also converges.

**The Modified Variational Iteration Method for Noep's**

We consider equation (4) as follows:
\[
x'(t) + Lx(t) + N\lambda(t) = 0,
\]
\[
\lambda'(t) + L\lambda(t) + N\lambda(t) = 0,
\]
\[
x(t_0) = x_0, \quad \lambda(t_0) = \alpha.
\]
(10)
where $L$ is a linear operator and $N$ is a non-linear operator.

To solve system (10) with MVIM, we should answer the following system:
\[
x'(t) + Lx(t) + N\lambda(t) = 0, \quad \lambda'(t) + L\lambda(t) + N\lambda(t) = 0, \quad x(t_0) = x_0, \quad \lambda(t_0) = \alpha.
\]
(11)
In equation (11)
\[
L(x(t) + N\lambda(t)) = -(f(t, x) + g(t, x)[-(R^{-1} g^T(t, x)\lambda)],
\]
\[
L(\lambda(t) + N\lambda(t)) = (Qx + \frac{\partial f(t, x)}{\partial x})^T \lambda + \sum_{i=1}^{n} \lambda_i [-\sum_{j=1}^{n} g^T(t, x)\lambda] \frac{\partial g_i(t, x)}{\partial x}.
\]
(12)
To solve equation (11) with MVIM, we construct the below iterations formula according to equation (8):
\[
x_{n+1}(t) = x_n(t) - \int_0^t \left[ R(x_n(t) - x_{n-1}(t)) + (G_n(t) - G_{n-1}(t)) \right] dt,
\]
(13)
\[
\lambda_{n+1}(t) = \lambda_n(t) - \int_0^t \left[ R(\lambda_n(t) - \lambda_{n-1}(t)) + (K_n(t) - K_{n-1}(t)) \right] dt,
\]
(14)
where $x_{-1}(t) = 0$, $\lambda_{-1}(t) = 0$, $x(0) = x_0$ and $\lambda(0) = \alpha$. We have:
\[
x_1(t) = x_0(t) - \int_0^t \left[ R(x_0(t) - x_{-1}(t)) + (G_0(t) - G_{-1}(t)) \right] dt,
\]
\[
\lambda_1(t) = \lambda_0(t) - \int_0^t \left[ R(\lambda_0(t) - \lambda_{-1}(t)) + (K_0(t) - K_{-1}(t)) \right] dt,
\]
and $G_n(t)$ and $K_n(t)$ are obtained from
\[
N\lambda_n(t) = G_n(t) + O(t^{n+1}) \quad \text{and} \quad N\lambda_n(t) = k_n(t) + O(t^{n+1}).
\]
Eqs. (13) and (14) can be solved iteratively to obtain an approximate solution that takes the form $x(t) \approx x_n(t)$ and $\lambda(t) \approx \lambda_n(t)$, where $n$ is the final iteration step.

The optimal control law is obtained by
\[
u^* = -R^{-1}g^T(t, x)\lambda.
\]
(15)
For stopping criterion, we consider the following criterion:
\[
\frac{|V_n - V_{n+1}|}{|V_{n+1}|} < \epsilon, \quad \text{where} \quad \epsilon > 0 \text{ should be chosen according to the desirable accuracy}.
\]

**Numerically Example**

In this section, we have solved three examples to illustrate the simplicity and efficiency of the proposed method.

**Example 1.** Consider the following nonlinear optimal control problem
\[
\min \int_{0}^{1} u^2(t) dt,
\]
Subject to:
\[
x'(t) = \frac{1}{2} x^2(t) \sin x(t) + u(t), \quad t \in [0, 1],
\]
\[
x(0) = 0, \quad x(1) = 0.5.
\]
(16)
The necessary equations for the optimal control are given as:
\[
x'(t) = \frac{1}{2} x^2(t) \sin x(t) - \frac{1}{2} \lambda(t), \quad t \in [0, 1],
\]
\[
\lambda' = -\lambda(t)x(t)\sin x(t) - \frac{1}{2} \lambda(t)x^2(t) \cos x(t),
\]
\[
x(0) = 0, \quad \lambda(0) = \alpha,
\]
(17)
that
\[
u(t) = -\frac{1}{2} \lambda(t).
\]
In this example, applying the following iteration formula
\[
x_{n+1} = x_n - \int_0^t \left[ R(x_n - x_{n-1}) + (G_n - G_{n-1}) \right] dt,
\]
\[
\lambda_{n+1} = \lambda_n - \int_0^t \left[ R(\lambda_n - \lambda_{n-1}) + (G_n - G_{n-1}) \right] dt,
\]
we consider
\[
R(\lambda(t)) = \frac{1}{2} \lambda(t),
\]
\[
G(\lambda(t)) = -\frac{1}{2} x^2(t) \sin x(t),
\]
\[
R(\lambda(t)) = 0,
\]
\[
G(\lambda(t)) = \lambda(t)x(t)\sin x(t) + \frac{1}{2} \lambda(t)x^2(t) \cos x(t),
\]
(18)
by applying Mathematica software, five-term approximations for $x$ and $\lambda$ were obtained as follows:
\[
x_5(t) = -\frac{\alpha t^2}{2},
\]
\[
\lambda_5(t) = \alpha - \frac{t^3 \alpha^3}{8} + \frac{t^5 \alpha^5}{192},
\]
In this case, we should have
\[
x_5(1) = -\frac{\alpha}{2},
\]
where $\alpha$ is an unknown parameter which will be obtained from the final state condition $x(t_f) = x_f$. Here, the value of $\alpha$ is derived from
\[
x_5(1) = -\frac{\alpha}{2} = 0.5,
\]
that is $\alpha = -1$. The optimal control is as follows:
\[ u(t) = u_\varepsilon(t) = -\frac{1}{2} \lambda \varepsilon(t) = \frac{1}{2} \left( 1 - \frac{t^3}{8} + \frac{t^5}{192} \right). \]

We consider \( \varepsilon = 10^{-3} \). Once the proposed method is applied, the numerical results for \( f_i \) and stopping criterion are as given in Table I. The maximum absolute error of the proposed method, modal series method \([23]\), and measure theory method \([24]\) are presented in Table II, which shows the proposed method has achieved similar results with modal series method. In addition, it should be noted that the basic VIM can not be calculated more than two iterations for example above. Also, the obtained numerical solution for \( x(t) \) and \( u(t) \) in five iterative are depicted in Fig 1.

**Table I:** Numerical results for different iteration, Example 1

| \( i \) | \( f_i \) | \( \frac{|f_i-f_{i-1}|}{f_i} \) |
|---|---|---|
| 1 | 0.25000 | - |
| 2 | 0.25000 | 0 |
| 3 | 0.23493 | 6.4146767 \times 10^{-2} |
| 4 | 0.23493 | 0 |
| 5 | 0.23533 | 1.6997408 \times 10^{-3} |

**Table II:** Numerical results for the proposed method versus other methods, Example 1

<table>
<thead>
<tr>
<th>Method</th>
<th>Objective value</th>
<th>Max state error</th>
<th>CPU time</th>
</tr>
</thead>
<tbody>
<tr>
<td>Proposed method (N=5)</td>
<td>0.2353</td>
<td>3.2 \times 10^{-5}</td>
<td>0.01562</td>
</tr>
<tr>
<td>Modal series method (n=5)</td>
<td>0.2353</td>
<td>2.8 \times 10^{-5}</td>
<td>-</td>
</tr>
<tr>
<td>Measure theory method</td>
<td>0.2425</td>
<td>4.3 \times 10^{-3}</td>
<td>-</td>
</tr>
</tbody>
</table>

**Example 2.** We consider the optimal maneuvers of a rigid asymmetric space craft \([2]\). The Euler’s equations for the angular velocities of the spacecraft are given by:

\[
\begin{bmatrix}
\dot{x}_1(t) \\
\dot{x}_2(t) \\
\dot{x}_3(t)
\end{bmatrix} =
\begin{bmatrix}
\frac{I_3-I_2}{I_1} x_1(t)x_3(t) \\
\frac{I_1-I_3}{I_2} x_1(t)x_2(t) \\
\frac{I_2-I_1}{I_3} x_1(t)x_2(t)
\end{bmatrix} +
\begin{bmatrix}
\frac{1}{I_1} \\
0 \\
0
\end{bmatrix} u_1(t) +
\begin{bmatrix}
\frac{1}{I_2} \\
0 \\
0
\end{bmatrix} u_2(t) +
\begin{bmatrix}
\frac{1}{I_3} \\
0 \\
0
\end{bmatrix} u_3(t),
\]

where \( x_1, x_2, \) and \( x_3 \) are the angular velocities of the spacecraft, \( u_1, u_2, \) and \( u_3 \) are the control torques, and \( I_1 = 86.24, I_2 = 85.07, \) and \( I_3 = 113.59 \text{ kg m}^2 \) are the spacecraft principle inertia. The optimal control is to find the control \( u(t) \) \((t \in [0,T]) \) that minimize the cost function

\[
J[x,u] = \frac{1}{2} \int_0^T (x^T(t)Qx(t) + u^T(t)Ru(t)) \, dt,
\]

Where: \( Q = \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \end{bmatrix}, R = \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \end{bmatrix} \).

In addition, the following boundary conditions should be satisfied:

\[ x_1(0) = 0.01 r/s, x_2(0) = 0.005 r/s, x_3(0) = 0.001 r/s, \]

\[ x_1(100) = x_2(100) = x_3(100) = 0 r/s. \]

According to the Pontryagin’s maximum principle, the following non-linear TPBVP should be solved:

\[
\begin{align*}
x_1'(t) &= -\frac{\lambda_1(t)}{I_1} x_2(t)x_3(t), \\
x_2'(t) &= -\frac{\lambda_2(t)}{I_2} x_1(t)x_3(t), \\
x_3'(t) &= -\frac{\lambda_3(t)}{I_3} x_1(t)x_2(t),
\end{align*}
\]

**Fig. 1(a):** Suboptimal state and control \( x(t) \), Example 1.

**Fig. 1(b):** Suboptimal state and control \( u(t) \), Example 1.
\[ \lambda_1'(t) = \frac{l_1 - l_2}{l_2} x_3(t) \lambda_1(t) + \frac{l_2 - l_1}{l_3} x_2(t) \lambda_3(t), \]
\[ \lambda_2'(t) = \frac{l_1 - l_2}{l_1} x_3(t) \lambda_1(t) + \frac{l_2 - l_1}{l_3} x_1(t) \lambda_3(t), \]
\[ \lambda_3'(t) = \frac{l_2 - l_1}{l_1} x_2(t) \lambda_1(t) + \frac{l_1 - l_2}{l_2} x_1(t) \lambda_2(t), \]
\[ x_1(0) = 0.01 \frac{\tau}{s}, x_2(0) = 0.005 \frac{\tau}{s}, x_3(0) = 0.001 \frac{\tau}{s}. \]

By applying the proposed method, the numerical results for \( J_i \) and stopping criterion are as given in Table III. The maximum absolute error of the proposed method, SHAM Chebyshev [14], SHAM Legendre [14] and HPM are as given in Table IV. It is noteworthy that the given method improves the maximum absolute error which indicates the efficiency of the method. Also, the obtained numerical solution for \( x(t) \) and \( u(t) \) in four iterative are depicted in Figures 2, 3 and 4.

Table III: Numerical results for different iteration, Example 2

<table>
<thead>
<tr>
<th>( i )</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
</tr>
</thead>
<tbody>
<tr>
<td>( J_i )</td>
<td>0.00468052</td>
<td>0.00467797</td>
<td>0.00467903</td>
<td>0.00467886</td>
</tr>
<tr>
<td>(</td>
<td>J_i - J_{i-1}</td>
<td>)</td>
<td>5.451082 × 10^{-4}</td>
<td>2.265426 × 10^{-4}</td>
</tr>
</tbody>
</table>
Fig. 2: Suboptimal state $x_2(t)$, Example 2.

Fig. 3: Suboptimal control $x_3(t)$ and $u_1(t)$, Example 2.

Fig. 4: Suboptimal control $u_2(t)$ and $u_3(t)$, Example 2.

Table IV: Numerical results for the proposed method versus other methods, Example 2

<table>
<thead>
<tr>
<th>METHOD</th>
<th>Objective value</th>
<th>Max state error</th>
<th>CPU time</th>
</tr>
</thead>
<tbody>
<tr>
<td>PROPOSED METHOD</td>
<td>0.004678</td>
<td>$2.40484 \times 10^{-14}$</td>
<td>0.046875</td>
</tr>
<tr>
<td>SHAM CHEBICHEV (M=6, N=50, H= -1.2)</td>
<td>0.004687</td>
<td>$1.0586 \times 10^{10}$</td>
<td>-</td>
</tr>
<tr>
<td>SHAM LEGENDER (M=6, N=50, H= -1.2)</td>
<td>0.004687</td>
<td>$1.0589 \times 10^{-9}$</td>
<td>-</td>
</tr>
<tr>
<td>HPM (M=6)</td>
<td>0.004687795533</td>
<td>$3.1420 \times 10^{-8}$</td>
<td>-</td>
</tr>
</tbody>
</table>

Example 3. Consider the non-linear system described by

$x_1' = x_2 + x_1 x_2$,

$x_2' = -x_1 + x_2 + x_2^2 + u$,

$x_1(0) = -0.8$, $x_2(0) = 0$, and the functional
minimize \( J = \frac{1}{2} \int_0^1 (x_1^2 + x_2^2 + u^2) \, dt \).

The extreme conditions are
\[
\begin{align*}
\lambda_1' &= - (x_1 + \lambda_1 x_2 - \lambda_2), \\
\lambda_2' &= - (x_2 + \lambda_1 (1 + x_1) + \lambda_2 (1 + 2x_2)), \\
x_1' &= x_2 + x_1 x_2, \\
x_2' &= -x_1 + x_2 + x_2^2 - \lambda_2,
\end{align*}
\]
that
\[
x_1(0) = -0.8, x_2(0) = 0, \quad \lambda_1(1) = \lambda_2(1) = 0,
\]
and the optimal control is \( u = -\lambda_2 \). By using Mathematica software, two-term approximations for \( \lambda_1, \lambda_2, x_1 \) and \( x_2 \), were obtained as follows:
\[
\begin{align*}
x_{12}(t) &= -0.8 + 0.08 \ t^2 - 0.1 \ t^2 \alpha_1, \\
x_{22}(t) &= 0.4 \ t^2 + 0.1 \ t^2 \alpha_1 - t(-0.8 + \alpha_2), \\
\lambda_{12}(t) &= \alpha_1 - 0.5t^2 \alpha_1 - t(-0.8 - \alpha_2) - \frac{t^2 \alpha_2}{2} + \frac{1}{2} t^2 \alpha_1 \alpha_2, \\
\lambda_{22}(t) &= -0.48t^2 + 0.1t^2 \alpha_1 + \alpha_2 + 0.1t^2 \alpha_2 + t^2 \alpha_2^2 - t(0.2\alpha_1 + \alpha_2),
\end{align*}
\]
That
\[
u(t) = u_{22}(t) = -\lambda_{22} = -0.536036 + 0.257913 t + 0.278123 t^2.
\]
We consider \( \varepsilon = 3 \times 10^{-2} \). Once the proposed method is applied, the numerical results for \( J_i \) and stopping criterion are as given in Table 6. The obtained numerical solution for \( x(t) \) and \( u(t) \) in three iterative are depicted in Fig 5 and 6.

![Fig 5: Suboptimal state \( x_1(t) \) and \( x_2(t) \), Example 3.](image)

Table V: Calculation of \( \alpha \) for \( N = 2 \), Example 3

<table>
<thead>
<tr>
<th>( \alpha_1 )</th>
<th>( \alpha_2 )</th>
<th>( j )</th>
</tr>
</thead>
<tbody>
<tr>
<td>-1.39061579</td>
<td>0.536036193</td>
<td>0.410274</td>
</tr>
</tbody>
</table>

Table VI: Numerical results for different iteration, Example 3

| \( i \) | \( J_i \) | \( \frac{\left| J_i - J_{i-1} \right|}{J_i} \) |
|-----|--------|-----------------|
| 1   | 0.417241 | -               |
| 2   | 0.410274 | 1.6981 \times 10^{-2} |
| 3   | 0.427145 | 3.9497 \times 10^{-2} |
| 4   | 0.431156 | 9.3028 \times 10^{-3} |

Table VII: Minimum of objective value for the proposed method, Example 3

<table>
<thead>
<tr>
<th>Method</th>
<th>Objective value</th>
<th>CPU time (sec)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Proposed method (N=3)</td>
<td>0.410274</td>
<td>0.015625</td>
</tr>
<tr>
<td>VIM (N=3) [18]</td>
<td>0.4102</td>
<td>-</td>
</tr>
</tbody>
</table>

Conclusions

Due to its high computing demands, VIM cannot solve some non-linear optimal control problems. Hence, we proposed the modified variational iteration method to find a solution for this type of optimal control problems. The suggested method eliminates all additional calculations in VIM thus requires less time to do the computations. As stated, in Example 1, the VIM cannot be counted more than two iterations due to additional calculations, but the proposed method is applicable in the high iteration. In addition, as seen in Table IV, the better max state error is obtained compared to other methods. As a future research direction, we can apply this method for solving optimal control problems ruled by partial differential equations and integral equations.
REFERENCES


Pooneh Omidinyia received her MSc degree from Sistan and Baluchestan University. Her research interests include optimal control, iterative method.

Maryam Alipour received her MSc degree from Mazandaran University and Ph.D. degree from Shahid Bahonar University of Kerman in applied mathematics. She is currently an Assistant Professor with university of Sistan and Baluchestan, Zahedan, Iran. Her research interests include optimal control, iterative method and fractional calculus.
Application of Improved Salp Swarm Algorithm Based on MPPT for PV Systems Under Partial Shading Conditions

Javad Farzaneh¹, Ali Karsaz²,†

¹ Department of Electrical and Computer Engineering Faculty, Semnan University, Semnan, Iran
²,† Department of Electrical and Electronic Engineering, Khorasan Institute of Higher Education. Mashhad, Iran

Abstract

Maximum Power Point Tracking (MPPT) is an important concept for both uniform solar irradiance and Partial Shading Conditions (PSCs). The paper presents an Improved Salp Swarm Algorithm (ISSA) for MPPT under PSCs. The proposed method benefits a fast convergence speed in tracking the Maximum Power Point (MPP), in addition to overcoming the problems of conventional MPPT methods, such as failure to detect the Global MPP (GMPP) under PSCs, getting trapped in the local optima, and oscillations around the MPP. The proposed method is compared with original algorithms such as Perturbation and Observation (P&O) method (which is widely employed in MPPT applications), Differential Evolutionary (DE) algorithm, Particle Swarm Optimization (PSO), and Firefly Algorithm (FA). The obtained results show that the proposed method can detect and track the MPP in a very short time, and its accuracy outperforms the other methods in terms of detecting the GMPP. The proposed ISSA algorithm has a higher speed and the convergence rate than the other traditional algorithms.

Article Info

Keywords: Improved Salp swarm algorithm, Photovoltaic systems, Maximum power point tracking, Partial shading condition.

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I. INTRODUCTION

On account of restrictions on fuel resources, increased fossil fuel price, effects of fossil fuel sources on the environment and increased energy consumption, employing Renewable Energy Sources (RESSs) has significantly increased during the last decades [1]. Among RESSs, solar energy and Photovoltaic (PV) systems have greatly been utilized as they generate reliable electrical energy without producing emissions [2-3].

PV systems are considered as a reliable source of electrical energy generation thanks to the following reasons: they can be utilized widely all over the globe, they do not require fuel for energy generation, and their maintenance cost is very low [4]. Specific features, such as no depreciation and the minimum number of moving parts, have resulted in exclusive advantages for PV systems, including durability and increased lifelong up to roughly 20 years. Another advantage of PV systems is that they can be used in different capacities. Therefore, it is possible to supply the required energy of a residential home in small dimensions or even supply the needs of a town. Furthermore, the application of PV systems is economically feasible for supplying electrical energy in places where it is either impossible to transfer the electricity from the grid or when high amounts of time and cost should be invested. One major hurdle in employing PV systems is the high installation cost and low energy transformation efficiency, where the latter is somehow due to the nonlinear characteristics and
dependency of I-V and P-V characteristics on temperature and solar irradiance. Solar irradiance and temperature have direct bearings on the generated current and voltage of the PV system, respectively. Therefore, it is incumbent on us to employ MPPT technology in PV systems to extract the maximum power under different irradiance and temperature conditions [5]. Up to date, different methods have been introduced for MPPT under uniform insolation conditions, such as Perturbation and Observation (P&O) [6], Incremental Conductance (IC) [7], Fuzzy Logic-based methods [8], Artificial Neural Network (ANN) [9], Fractional Open-Circuit Voltage or Short-Circuit Current methods [10], and Hill Climbing (HC) [11]. The mentioned methods have great ability in MPPT applications under uniform insolation conditions. Among these methods, P&O and IC have the most extensive applications in comparison with the other mentioned methods [12]. In addition to temperature and irradiance, the other effective factor on the generated power of PV systems is the Partial Shading Condition (PSC). When PSC occurs due to the presence of clouds, shadings by trees, and or other objects in the neighborhood of the PV system, the output power of a PV system dramatically decreases. Normally, bypass diodes are used for reducing the PSC effect. If the bypass diodes are used in the case of PSC, the P-V characteristics curve of a PV system will have several peaks instead of one single peak. In this situation, conventional methods are unable to detect the GMPP and normally trap in the local optima [13].

Many different methods have already been introduced for MPPT under PSCs, all of which have their advantages and disadvantages. Among the available MPP tracking methods, optimization algorithm-based methods such as Firefly Algorithm (FA) [20], Particle Swarm Optimization (PSO) algorithm [14], Cuckoo Search (CS) algorithm [15], Ant Colony Optimization (ACO) algorithm [16], Grey Wolf Optimization (GWO) algorithm [17], Artificial Bee Colony (ABC) [18], Bat Algorithm (BA) [19], evolutionary algorithms [20], and Gravitational Search Algorithm (GSA) [21] are widely used. The reason behind this is that these types of techniques have many features, including the ability to detect the GMPP under any conditions, very small probability of trapping in the local optima, no oscillation around the MPP, and easy implementation in comparison to other methods.

Ref. [22] used the FA method for MPPT under PSCs, where a boost converter was utilized for connecting the load to the PV system. The method was compared with PSO and P&O methods, and it has been observed that the FA method has faster speed and higher accuracy in detecting the GMPP compared to the other two methods. A combined method was introduced in [16] for the MPPT application. In the first step, using an ACO algorithm, the proposed method reaches the neighborhood of the MPP and then tracks it by using the P&O method. GWO algorithm was employed in [23] for MPPT under PSCs. The author has compared the proposed method with P&O and other improved PSO methods and showed that the proposed method has a higher convergence speed without oscillations around the MPP. The average efficiency of the method is 99.85%. Fibonacci series was employed as a method of tracking the GMPP [24], where it was utilized as a mathematical basis for dividing the P-V curve. Since the method works based on a mathematical basis, it has very suitable accuracy, although its implementation cost is high. In [25], a method similar to the two-step search approach was used for MPPT under PSC. In this method, at first, the P-V curve is completely swept and the local optimum points are stored. Then, by perturbing the voltage and observing the power variations, the GMPP is tracked. The speed of this method is acceptable and has suitable tracking accuracy. An Adaptive Neuro-Fuzzy Inference System (ANFIS) was used for MPPT under uniform solar irradiance [26]. The suggested method has a very high convergence speed, and the accuracy of the MPPT depends on the learning quality of the ANFIS system such that if it is trained with enough data, it will have good accuracy. A new MPPT method based on the PSO algorithm was presented in [27], where extra coefficients are added to the PSO equations to reduce the computational effort of the algorithm. Nonetheless, it cannot be surely said whether the algorithm succeeds in continuously tracking the MPP. The reason is that when the particles are close to the MPP, the speed of the algorithm becomes significantly low or even is zero. One of the most common problems with the PSO algorithm is that when there is a small difference in the solar irradiances, the changes in the duty cycle should be small enough to be able to perform MPPT more accurately.

Ref. [28] makes use of the PSO algorithm for MPPT under PSC. In this paper, a boost converter is utilized for increasing the efficiency of the system for each solar panel, but this leads to the increased capital cost. Additionally, [29] used PSO for MPPT under PSCs. The author defined linear equations for parameters of the PSO algorithm to increase the speed and accuracy of the MPPT. The equations were tuned so that the system would have higher exploration and diversity at the beginning of running the algorithm, and the exploration value is decreased through running the algorithm so that the solutions converge. In [30], the ACO algorithm was used for MPPT. ACO is an inspiration by ant behaviors in the colony and the foraging path. The population size was assumed to be four. The proposed method was compared with PSO, P&O, and fractional open-circuit voltage methods. The simulation results were obtained for four different case studies of PSCs, highlighting the superiority of the ACO-based method over the other three methods. The authors in [31] combined the P&O algorithm with the genetic algorithm (GA) structure. This resulted in a reduction in the population size of the algorithm. Due to the reduction in the number of iterations (NOI), the MPPT’s speed increased significantly. In [32], a combined method was used for MPPT under PSCs. Noting that the P&O
method is unable to detect the global optimum and the probability of trapping in the local optima is high, the author used it in conjunction with the PSO algorithm. In this method, at first, the P&O method starts the search process and once it reaches the local optimum point, the PSO method takes action. Since the P&O method is used in the first stage, it reduces the search space for the PSO method; thus, the system converges in a short time. A combination of PSO and P&O methods was employed in [33] for MPPT applications, where the P&O method is used under uniform solar irradiance and the PSO algorithm for tracking the MPP only at the beginning of PSCs. In ref. [34], by mixing the GWO and P&O methods, the author made an effort to extract the MPP under PSCs. Initially, the range of GMPP is determined by the GWO algorithm, and then the MPP is tracked using the P&O method. The speed of this approach is acceptable. In [35], the author has made some modifications in the FA method to reduce the convergence time while increasing the tracking speed of the MPPT. In this paper, using the average coordination of all fireflies as a representative point, the desired firefly moves only towards their average, instead of moving towards each of the brighter fireflies. In this method, by reducing the number of firefly movements, the tracking speed is increased. Nonetheless, the probability of not detecting the global optimum by the system is increased because the variety of fireflies’ movements has dramatically reduced.

Li, Hong, et al. have proposed an overall distribution PSO and MPPT algorithm for a PV power system to track the GMPP under PSCs [36]. The main difference between our methodology and this new published paper is that in our algorithm the new proposed metaheuristic method SSA is used. The SSA is able to explore the most promising regions of the search space, move salps abruptly in the initial steps along with move gradually in the final steps of iterations compared to the standard and improved versions of the PSO methods. Moreover our proposed algorithm based on SSA method ensures and improves the average fitness of all salps, and enhance that best solution found so far over the course of optimization.

Recently, Yang, Bo, et al. proposed a new hybrid algorithm based on memetic algorithm and salp swarm optimization, named as MSSA [37]. The Memetic algorithm has developed into a broad class of algorithm and can properly hybrid a population based global search and heuristic local search. So the new ideal in this paper is to adopt the memetic computing framework to enhance the search ability of SSA, which mainly contains two operations, “local search in each chain” and “global coordination in virtual population”. Therefore, this paper is also proposed a new method to improve the standard SSA as well as our methodology. In our proposed algorithm with a simple manner, contribution of previous position of salp to update new positions have increased. So our proposed algorithm is an alternative algorithm to Yang, Bo’s idea in a simple manner.

The present work aims to use an improved salp swarm algorithm (ISSA) for the MPPT under PSCs in PV systems. The proposed method is compared with its counterparts, including the SSA method; the P&O method, which is a common method for MPPT; the DE method, which is one of the mostly-used evolutionary algorithms in engineering sciences and MPPT applications; and FA, and PSO methods, which work based on swarm intelligence and have suitable efficacy in tracking the MPP. It is worth mentioning that the optimization methods at the start of the algorithm are initialized randomly and the probability of trapping in the local optima is very high. Another challenge that should be investigated is the ability to detect the global optimum, which cannot be perceived or detected in only one or two times of executing of the simulation software. Thus, the suggested method is compared with DE, FA, PSO, and SSA methods by running the software for many times and the obtained results verify the superiority of the proposed method over its counterparts. The advantages of our proposed scheme make it different from some other works are:

1- The proposed ISSA benefits from many interesting features, making it a highly reliable method for accurate track of the MPP within a short time.
2- The ability to detect the global optimum, which cannot be perceived or detected in only one or two times of executing of the simulation software.
3- The proposed ISSA algorithm has a higher speed and converges very fast.

In the rest of the paper, the model of the solar cell, the effects of temperature and solar irradiance, as well as the effect of shading conditions on PV systems are given in Section II. In Section III, the objective function and the P&O method are described. SSA and ISSA methods are explained in Section IV, and their applications in MPPT are described. The simulation results are thoroughly presented in Section V and finally, conclusions are included in Section VI.

II. CHARACTERISTIC OF THE PV SYSTEM AND THE EFFECTS OF PSCS

A. Modeling of the PV module

A current source connected in shunt to a diode can be used to represent the equivalent circuit of an ideal PV cell. A resistor ($R_p$) is connected in parallel to the equivalent circuit to limit the leakage current flow. Similarly, a series resistor ($R_s$) is employed to measure the losses. There are two types of modeling for the PV cell: (i) single-diode model, and (ii) double-diode mode [38]. Between these two models, the later is more accurate, but more parameters are required to accurately model the PV cell. As a result, a single-diode model is used in this paper for simplification. Fig. 1 illustrates the schematic of the single-diode model.
In general, equations of the output voltage and current of a solar cell are expressed as (1).

\[ I = I_{PV} - I_0 \left( \exp \left( \frac{V + R_S I}{V_{oc}} \right) - 1 \right) - \frac{V + R_S I}{R_P} \]

(1)

Where \( I_{PV} \) is the PV current and \( I_0 \) is the reverse saturation current. \( I_{PV} \) is defined based on the following equation:

\[ I_{PV} = (I_{PV,n} - K_I \Delta T) \frac{G}{G_n} \]

(2)

The generated current due to solar irradiance at standard conditions (temperature of 25°C and irradiance of 1000W/m²) is \( I_{PV,n} \). In addition, \( \Delta T \) is the temperature difference between the actual temperature (T) and the ambient standard temperature (\( T_n \)). \( K_I \) indicates the temperature coefficient of the short-circuit current. \( G \) is the irradiance and \( G_n \) denotes the irradiance at nominal conditions. Furthermore, for the inverse saturation current, we have:

\[ I_0 = \frac{I_{ren} + K_I \Delta T}{\exp\left( \frac{V_{oc,n} + K_V \Delta T}{q} \right)} \]

(3)

The short-circuit current and open-circuit voltage at nominal conditions are expressed by \( I_{ren} \), \( V_{oc,n} \), and the voltage factor of the open-circuit is defined as \( K_V \). The heating voltage of a panel with \( N_s \) number of solar cells connected in series is stated as (4).

\[ V_S = \left( \frac{N_s k T}{q} \right) \]

(4)

q is the electron charge (\( q = 1.6 \exp(-19)c \)), \( k \) is the Boltzmann factor (\( k = 1.3805 \exp(-23) \frac{J}{Kelvin} \)), and T is cell’s temperature in Kelvin.

B. Effects of temperature and irradiance on the PV panel

Characteristics of the assumed PV panel for this study at standard conditions (temperature = 25°C, air mass = 1.5, irradiance =1000 W/m²) are given in Table.1. One of the effective factors of the produced power by PV systems is irradiance. By increasing solar irradiance, the produced power is proportionally increased, and vice versa. Fig. 2(a) depicts I-V and P-V curves of the PV panel under-study (MSX-60). As it is obvious from the figure, in the temperature of 25°C, when the irradiance is increased from 400 W/m² to 1000 W/m² with steps of 200, the generated current is increased and this entails the increase in the produced power. Besides, the temperature is also effective on the generated power by PV systems. By increasing temperature the output power is decreased, and vice versa. Fig. 2(b) illustrates that when the temperature is increased from 25°C to 55°C in the steps of 10°C, the open-circuit voltage is decreased, and as a result, the output power of the solar panel will decrease. Solar irradiance in graphs of Fig. 2(b) is assumed to be constant, equal to 1000 W/m².

<table>
<thead>
<tr>
<th>Table. 1 Parameters of the MSX-60 PV module at STC</th>
</tr>
</thead>
<tbody>
<tr>
<td>Maximum power (W), ( P_{app} )</td>
</tr>
<tr>
<td>Nominal open-circuit voltage (V), ( V_{ocn} )</td>
</tr>
<tr>
<td>Maximum power voltage (V), ( V_{MPP} )</td>
</tr>
<tr>
<td>Nominal short-circuit current (A), ( I_{scn} )</td>
</tr>
<tr>
<td>Maximum power current (A), ( I_{MPP} )</td>
</tr>
<tr>
<td>Number of series cells, ( N_s )</td>
</tr>
</tbody>
</table>

C. Effects of PSCs on PV systems

The produced power by a PV system depends on temperature and irradiance, and these two factors have direct impacts on the output power produced by the PV system. Another effective event on PV systems is PSCs. When a PV system is used in urban environments, PSC is not negligible. In most cases, the reason for the occurrence of PSC is the presence of trees, buildings, and other objects around PV systems. Hence, the impact of PSCs on the PV system should be investigated. When a part of a module or even the whole module of a PV system is under shading conditions, the output power reduces. In Fig. 3, the PV module has n cells. And a cell separated from the other cells is shown, which has current I and voltage V.
Fig. 2. The P-V and I-V curves under varying irradiation (b) The P-V and I-V curves under variable temperature.

In Fig. 3 (a), all cells are under direct solar irradiation and are connected in series. Thus, the same current $I$ is flowing through them. In Fig. 3(b), the upper cell is under shading condition and its produced current is zero. The remaining $n-1$ cells are under solar irradiation. As a result, the whole current produced by $n-1$ cells must flow through resistors $R_p$ and $R_s$, and this reduces the output power. In the cases the PV modules are connected in series or shunt to increase the output power, bypass diodes are used to reduce the impact of PSCs. For instance, when PSC occurs and a module is under shading condition, diodes prevent the loss of power produced by other modules and impede the creation of a hotspot. Consequently, only the module under shading condition produces less power. P-V characteristics of PV systems differ in the presence and absence of the bypass diodes. Under shading conditions, different currents flow through the modules, therefore the P-V curve has several peaks. In Fig. 4, four PV panels are connected in series, wherein Fig. 4(a) the irradiance is the same for all panels and is equal to 1000 W/m$^2$. In Fig. 4(b), it is assumed that the PSC has occurred and irradiance is 1000 W/m$^2$, 850 W/m$^2$, 700 W/m$^2$, and 300 W/m$^2$, respectively. Fig. 5 depicts P-V and P-I curves in the presence of bypass diodes in the PV system of Fig. 4. As it is evident from Fig. 5, the uniform irradiance is equal to 1000 W/m$^2$, the produced power is 240 W. Moreover, when the PSC occurs, there are several peaks in the P-V curve and the produced power is decreased as well.
III. OBJECTIVE FUNCTION AND P&O METHOD

The formulation of the MPPT as an optimization problem is as follows:

\[
\text{Maximize } P_{\text{pv}} (d) \\
\text{Subject to } \begin{cases} 
  d_{\text{min}} \leq d \leq d_{\text{max}} \\
  V_{\text{min}} < V_{\text{pv}} < V_{\text{max}}
\end{cases}
\]

In Eq. (5), \(P_{\text{pv}}\) is the output power, \(d\) is the duty cycle of the DC-DC buck-boost converter, and \(d_{\text{min}}\) and \(d_{\text{max}}\) define the minimum and maximum duty cycles of the DC-DC converter. In this work, the latter two parameters are set 0.1 and 0.9, respectively.

A. PERTURBATION AND OBSERVATION METHOD

The P&O method is a widely-used conventional method for MPPT under uniform irradiance as it has easy implementation and low cost compared with other MPPT methods [39]. In some papers, the P&O method is combined with other MPPT methods to increase the tracking speed [40]. The performance of the P&O method is such that, at first, a small controlled disturbance is established in the initial operating voltage of the PV system based on the tracking criteria. This is performed in such a manner that its value increases in case the produced power of the PV system is increased, i.e., the operating point moves toward the MPP. As a result, in the next disturbance, the operation voltage is established in the same direction exactly similar to the previous disturbance. This is continued until the MPP is reached. However, if the produced power by the PV system is reduced, it means that it has moved further away from the MPP. Therefore, the direction of the disturbance must be changed [41]. The efficacy of the P&O method is reduced by quick changes in environmental conditions and under PSCs because it is unable to track the GMPP under PSCs [42].

IV. SALP SWARM ALGORITHM

A. Inspiration and the proposed mathematical model

Salp swarm algorithm was introduced by Seyed Ali Mirjalli et al. in 2017 [42]. A swarm formed by a group of salps is known as a salp chain. The population of salps is divided into two groups to mathematically model salp chains: leader salp and follower salps. The leader salp is the one at the front end of the salp chain, and other salps are considered as followers. It is clear from the naming that the leader has the task of leading and guiding the followers. Like other population-based algorithms, an n-dimensional search space is assigned to define the positions of the salps, where \(n\) indicates the number of variables for a specific problem. A two-dimensional matrix, \(x\), is used for storing the positions of the salps. Moreover, it is taken for granted that there is a food source, \(F\), in the search space and the salp is trying to reach that. The following equation is used for updating the position of the leader salp:

\[
x^l_j = \begin{cases} 
  F_j + c_1 \left( \left( u_j - l_j \right) c_2 + l_j \right) & c_3 \geq 0 \\
  F_j - c_1 \left( \left( u_j - l_j \right) c_2 + l_j \right) & c_3 < 0
\end{cases}
\]

It can be understood that the position of the leader is updated only corresponding to \(F\) as the food source. In the above equation, \(x^l_j\) is the position of the leader in the \(j^{th}\) dimension, \(F_j\) represents the food source in the \(j^{th}\) dimension,
and \(ub_i\) and \(lb_i\) are the upper and lower boundaries of the \(j^{th}\) dimension. The most important parameter to be considered in the SSA is coefficient \(c_1\) because this coefficient makes a random balance between exploration and exploitation. Furthermore, \(c_2\) and \(c_3\) are two random numbers generated uniformly in the interval \([0, 1]\). These two numbers determine the movement of the next position in the \(j^{th}\) dimension, either towards positive or negative infinity. Moreover, the size of the steps is defined by \(c_2\) and \(c_3\). The equation to update the position of the followers is:

\[
\epsilon_j^m = \frac{1}{2} \epsilon_j^m + \epsilon_j^{m-1} \quad (7)
\]

where \(i \geq 2\) and \(x_{ji}\) indicates the position of the \(i^{th}\) follower salp in the \(j^{th}\) dimension.

### B. Improved SSA algorithm

In the SSA algorithm, the positions of the follower salps are updated based on (7). For the system to have a high convergence speed and converge within a shorter time, equation (7) is changed so that the previous positions of the follower salps have more contribution to updating new positions of the salps. This leads the salps to have smaller displacements with respect to their previous positions. According to (7) and assuming that \(\epsilon_j^m\) and \(\epsilon_j^{m-1}\) are positive values, (7) can be rewritten as:

\[
x_j^m = \frac{1}{2}(x_j^m + x_j^{m-1}) = \frac{1}{2}((x_j^m + x_j^{m-1})^2)^\frac{1}{2}
\]

\[
= \frac{1}{2}(x_j^m)^2 + (x_j^{m-1})^2 + 2x_j^m x_j^{m-1})^\frac{1}{2} \quad (8)
\]

In the above equation, for the contributions of previous positions of the salps have more effects on updating new positions, the value of \(2(x_j^{m-1})^2\) is taken into account, not \(2(x_j^m x_j^{m-1})\). As a result of this, the salps move less than their previous positions and finally, this increases the convergence speed. By simplifying (8), we have:

\[
x_j^m = \frac{1}{2}(x_j^m)^2 + 3(x_j^{m-1})^2)^\frac{1}{2} \quad (9)
\]

As it is clear from the above equation, contributions of previous positions of the salp to updating new positions have increased. Due to this, when the system will not experience large variations and have more inertia when reaches the optimal point, leading to increase convergence speed.

### C. Application of the ISSA method to MPPT

In this section, the ISSA method is used for MPPT Fig. 6 shows the block diagram of the proposed system. The system under-study consists of four PV modules connected in series. Also, a buck-boost DC-DC converter is used as a link between the load and the PV system. In the following, the steps of the ISSA method for MPP tracking are illustrated in detail. Fig. 7 exhibits the flowchart of the proposed method.

Step 1: Adjusting the parameter \(c_1\) of the ISSA and determining the population size, \(N\). In this algorithm, the position of each salp is considered as the duty cycle of the buck-boost DC-DC converter. The power generated by the PV system is assumed as the fitness of each salp, which is corresponding to the position of each salp.

Step 2: Initialization of the salps. This step initializes the population of the salps in an acceptable solution space respect to the lower bound, \(d_{min}\), and the upper bound, \(d_{max}\). The position of a salp shows the duty cycle of the DC-DC converter. As was mentioned in the previous section, if the size of the population size large, the calculation time will increase; otherwise, the algorithm might be trapped in local optima. Therefore, the population size is assumed four, meaning that one salp for each panel.

Step 3: Fitness evaluation. In this step, the buck-boost converter operates sequentially corresponding to the position of each salp. For each duty cycle, the output power of the PV system is taken into account as the fitness of the corresponding salp. This step is repeated for positions of all of the salps available in the population.

Step 4: Updating salps' positions. The position of the leader salp is updated according to (6), and positions of the follower salps are updated based on (9).

Step 5: The termination criterion, similar to other optimization methods, is the number of iterations. Once it is reached, the algorithm stops, and the system operates based on the optimal duty cycle.
To analyze the global MPP tracking using the ISSA method, a thorough comparison is made between SSA; FA; and PSO methods; which work based on swarm intelligence; DE; and P&O methods, respectively. Extensive studies have been performed on MATLAB/Simulink software under different patterns of PSCs. The input inductance of the buck-boost DC-DC converter is $1\, \text{mH}$, the output capacitor is $220\, \mu\text{F}$, the input capacitor is $470\, \mu\text{F}$, and the switching frequency is $50\, \text{kHz}$. To ensure that the system will experience steady-state conditions before the start of the next particle to track the MPP, the sampling time interval is set $0.04\, \text{s}$. For the PSO method, $C_1=1.2$, $C_2=1.6$, and $W=0.4$ are taken into account, as in [28]. The number of iterations (NOI) is also set 10. The values of NOIs in these two methods are set 10. The above-mentioned parameters are summarized in Table 2.

### Table 2

<table>
<thead>
<tr>
<th>Method</th>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>PSO</td>
<td>$C_1$</td>
<td>1.2</td>
</tr>
<tr>
<td></td>
<td>$C_2$</td>
<td>1.6</td>
</tr>
<tr>
<td></td>
<td>$W$</td>
<td>0.4</td>
</tr>
<tr>
<td></td>
<td>NOI</td>
<td>10</td>
</tr>
<tr>
<td>FA</td>
<td>$\beta_0$</td>
<td>2</td>
</tr>
<tr>
<td></td>
<td>$\alpha$</td>
<td>0.6</td>
</tr>
<tr>
<td></td>
<td>$\gamma$</td>
<td>1</td>
</tr>
<tr>
<td></td>
<td>NOI</td>
<td>6</td>
</tr>
<tr>
<td>DE</td>
<td>$F_{\text{Max}}$</td>
<td>0.2</td>
</tr>
<tr>
<td></td>
<td>$F_{\text{Min}}$</td>
<td>0.8</td>
</tr>
<tr>
<td></td>
<td>CR</td>
<td>0.3</td>
</tr>
<tr>
<td></td>
<td>NOI</td>
<td>10</td>
</tr>
<tr>
<td>P&amp;O</td>
<td>$T_a$</td>
<td>0.005</td>
</tr>
<tr>
<td></td>
<td>$\Delta d$</td>
<td>0.005</td>
</tr>
<tr>
<td>SSA</td>
<td>$c_1$</td>
<td>$3.2e^{-3.5l^2}$</td>
</tr>
<tr>
<td></td>
<td>NOI</td>
<td>10</td>
</tr>
<tr>
<td>ISSA</td>
<td>$c_1$</td>
<td>$2.3e^{-3.2l^2}$</td>
</tr>
<tr>
<td></td>
<td>NOI</td>
<td>10</td>
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</tbody>
</table>

The values of NOIs in these two methods are set 10. The above-mentioned parameters are summarized in Table 2.

MPPT techniques have been used as controllers to supply the buck-boost converter with a desired duty cycle to study and compare dynamic responses of the PV system under PSCs. MPPT methods were investigated from the following perspectives: tracking time, convergence speed, oscillations around the MPP, and the efficiency of MPP tracking under PSCs. Two PSC patterns were utilized to evaluate the efficacy of the methods used in this paper in tracking the MPP. In the first PSC pattern, the irradiance on each module is 1000 W/m², 800 W/m², 600 W/m², and 400 W/m², respectively, and the temperature is set 25°C. P-I and P-V curves for the first PSC pattern and uniform irradiance (1000 W/m²) are given in Fig. 8(a) and (b), respectively. In the first PSC pattern, there are four peaks, in which the GMPP is 116.468 W and is located on the P-I curve on the second peak. Details of the simulation results obtained for the PV system (power and duty cycle of the buck-boost converter) are provided in Fig. 9 for different MPPT techniques under the first PSC pattern.
According to Fig. 9, the P&O method has reached a power of 108.5 W in less than 0.7 s. In terms of speed, the P&O method has acceptable speed but cannot detect the global MPP and is trapped in local optima. Further, power and duty cycle curves given in the corresponding curves of the P&O method are presented in smaller intervals to clearly show the changes in power and duty cycle. As is observed from the figure, duty cycle continuously changes in the P&O method and the output power has limited oscillation, which results in power loss. The FA method has succeeded to track the global MPP in less than 3 s and reach a power of 115.52 W. As is seen in the figure, the algorithm has tracked the global point in the last iterations and had changes when searching for the accurate point in the neighborhood of the optimal duty cycle. The algorithm continues its operation with the optimal duty cycle after the iterations are finished. Power and duty cycle curves in smaller intervals clearly show the convergence of the FA method. As is seen, the particles in the last iterations are seeking for the accurate MPP in the duty cycle range of 0.55. PSO, DE, and SSA methods track the GMPP within 1.76 s, 1.6 s, and 1.56 s, respectively. However, the ISSA method does it in a shorter time, within almost 1.2 s. Compared to other methods, as is seen in Fig. 9, the ISSA method has successfully detected the GMPP in a very short time and extracted the maximum power. As the system tends to maintain its position in the proposed system, it converges with a high speed after detecting the considered point. Moreover, the suggested method benefits a significantly high accuracy. Simulation results are summarized in Table 3. Also, to illustrate the impact of data uncertainties on the obtained solution, this table shows the results under 15 percent fluctuations in input data including temperature and irradiance for each panel.

<table>
<thead>
<tr>
<th>Shading pattern</th>
<th>Technique</th>
<th>Power (W)</th>
<th>Tracking speed (s)</th>
<th>Power at the GMPP (W)</th>
<th>Voltage at the GMPP (V)</th>
<th>Current at the GMPP (A)</th>
<th>Tracking efficiency (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>First case</td>
<td>P&amp;O</td>
<td>108.5</td>
<td>0.76</td>
<td>116.46</td>
<td>53.95</td>
<td>2.16</td>
<td>92.44</td>
</tr>
<tr>
<td></td>
<td>FA</td>
<td>115.52</td>
<td>2.92</td>
<td></td>
<td></td>
<td></td>
<td>99.01</td>
</tr>
<tr>
<td></td>
<td>PSO</td>
<td>115.56</td>
<td>1.53</td>
<td></td>
<td></td>
<td></td>
<td>99.04</td>
</tr>
<tr>
<td></td>
<td>DE</td>
<td>115.45</td>
<td>1.76</td>
<td></td>
<td></td>
<td></td>
<td>98.89</td>
</tr>
<tr>
<td></td>
<td>SSA</td>
<td>115.58</td>
<td>1.35</td>
<td></td>
<td></td>
<td></td>
<td>99.12</td>
</tr>
<tr>
<td></td>
<td>ISSA</td>
<td>115.59</td>
<td>1.22</td>
<td></td>
<td></td>
<td></td>
<td>99.13</td>
</tr>
</tbody>
</table>
P&O

FA

PSO
Fig. 9. Simulation results for the PV system under the first PSC, a) power and duty cycle curves, b) a with higher magnification

For further study, the second PSC pattern was also studied, in which the power difference between the global optimal point and one of the local optima is small to better show the efficiency of the algorithms. The irradiance for each panel is considered 1000 W/m², 800 W/m², 500 W/m², and 300 W/m², respectively, and the temperature is set 25°C. Another point that is worth noting here is that the efficacy of meta-heuristic algorithms cannot be proved only by one single run. Therefore, for the second PSC pattern, each algorithm was run 40 times to find the most efficient and superior MPP tracking method under PSCs. P-I and P-V curves for the second PSC case are given in Figs. 10 (a) and (b),
respectively. As observed in the P-V curve, the difference between the GMPP and one of the local optima is significantly small, making it difficult to detect the GMPP. The GMPP value is 98.21 W and the local optimal value with a small difference to the GMPP is 96.14 W. It can be concluded from the obtained results of the first PSC case that the P&O method fails to detect the GMPP, and that is why it was not compared with other methods in the second PSC case. Details of simulation results obtained for the PV system from different MPPT methods under the second PSC case are shown in Fig. 11.

![P-I curve](image1)

![P-V curve](image2)

**Fig. 10.** (a) P–I curve (b) P–V curve under the second PSC

![DE](image3)

![SSA](image4)

![ISSA](image5)

**Fig. 11.** Details of the PV system simulation results under the second PSC case.

In Fig. 11, meta-heuristic methods were compared from the MPPT ability point of view. As is evident in Fig. 11, the FA method was successful in tracking the GMPP for 22 out of 40 times, was trapped in a local optimum with a small difference to the GMPP for 16 times, and was trapped in other local optima for 2 times. These numbers for the PSO, DE, SSA, and ISSA methods were 23, 15, 2; 15, 20, 5; 24, 15, 1; and 24, 15, 1, respectively. Finally, from the first pattern of the PSC, it can be concluded that the ISSA method has a high convergence speed and is a suitable method for tracking the GMPP. According to the results obtained from the second PSC run for a high number of repetitions, notable conclusions can be drawn. The results of the second pattern show that the DE method is trapped in the local optima, so it cannot be considered as a reliable method. FA and PSO methods are more reliable than the DE method, and in most cases, they can detect the global optimum with satisfactory accuracy. Nevertheless, their convergence times are longer. Additionally, simulation results illustrate that SSA and ISSA methods have acceptable accuracy in detecting the GMPP. Table 4 describes a quality comparison between different proposed methods.
<table>
<thead>
<tr>
<th>Criteria</th>
<th>ISSA</th>
<th>SSA</th>
<th>PSO</th>
<th>FA</th>
<th>DE</th>
<th>P&amp;O</th>
</tr>
</thead>
<tbody>
<tr>
<td>Periodic Tuning</td>
<td>Not Required</td>
<td>Not Required</td>
<td>Not Required</td>
<td>Not Required</td>
<td>Not Required</td>
<td>Not Required</td>
</tr>
<tr>
<td>Tracking Accuracy</td>
<td>Highly Accurate</td>
<td>Highly Accurate</td>
<td>Highly Accurate</td>
<td>Highly Accurate</td>
<td>Accurate</td>
<td>Low (may locate local peak)</td>
</tr>
<tr>
<td>Steady state oscillation</td>
<td>Zero</td>
<td>Zero</td>
<td>Zero</td>
<td>Zero</td>
<td>Zero</td>
<td>High</td>
</tr>
<tr>
<td>Tracking Speed (without PSC condition)</td>
<td>Fast</td>
<td>Medium</td>
<td>Medium</td>
<td>Medium</td>
<td>Medium</td>
<td>Medium</td>
</tr>
<tr>
<td>Ability to track under PSCs</td>
<td>High</td>
<td>High</td>
<td>High</td>
<td>High</td>
<td>Average</td>
<td>Poor</td>
</tr>
<tr>
<td>Algorithm complexity</td>
<td>Medium</td>
<td>Medium</td>
<td>Medium</td>
<td>Medium</td>
<td>Medium</td>
<td>Simple</td>
</tr>
<tr>
<td>Efficiency</td>
<td>High</td>
<td>High</td>
<td>High</td>
<td>High</td>
<td>Average</td>
<td>Poor in PSCs</td>
</tr>
</tbody>
</table>

VI. CONCLUSION

The original SSA algorithm is amongst the most recently introduced algorithms. In this paper, a new method based on the improved SSA is proposed for the MPPT under PSCs in PV systems. In this study, the method of updating the follower salps' positions is improved to increase the tracking speed of the MPP in the SSA algorithm. The ISSA method was compared with the original algorithm; the P&O method as a conventional method for MPPT; the DE method, which is included in evolutionary algorithms; and FA, and PSO methods as swarm intelligence algorithms, under PSCs. The simulation results obtained from MATLAB/Simulink software show that the proposed method, SSA, FA, PSO, and DE can track the GMPP under PSCs. In addition, ISSA and SSA methods have a noticeable capability in detecting the global MPP, even when the difference between the global optimum and the local optima is very small. Moreover, the speed of the ISSA method in tracking the GMPP under PSCs is higher than other methods. High efficiency and speed are the main advantages of the proposed method.

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Javad Farzaneh was born in Mashhad, Iran, in 1991. He received the B.Sc. degree from the Bahar Institute of Higher Education, Mashhad, Iran, in 2014, the M.Sc. degree in electrical engineering from the Semnan University, Semnan, Iran, in 2017. His research interests are Artificial Neural Networks, Adaptive Neuro-Fuzzy Inference Systems (ANFIS), Stochastic Modeling and Estimation, System Identification, Metaheuristic Optimization, Renewable Energy, Power Electronics.

Ali Karsaz received his B.S. degree in Electrical Engineering from the Amirkabir University of Technology (Tehran Polytechnic), Tehran, Iran, in 1999. He received his M.Sc. and Ph.D. degrees in Control Engineering both form Ferdowsi University of Mashhad, Iran, in 2004 and 2008, respectively. Since 2008, he has been Assistant Professor of control and biomedical engineering at Khorasan Institute of Higher Education and he was Chair of the Division of Control Department from 2012 until now. His current research interests include the development of mathematical models for analysis and control of Biological Systems, Pharmacodynamics, System Biology Mathematical Modeling, Artificial Neural Networks, Stochastic Modeling and Estimation, System Identification, Time Series Analysis and Prediction, Renewable Energy, Power Electronics, Inertial Navigation Systems, Multi-sensory Multi-target Tracking. He has published over 130 peer-reviewed articles in these and related research fields.
Model-Free Tracking Control via Adaptive Dynamic Sliding Mode Control With Application To Robotic Systems

Mohammad Reza Shokoohinia¹,†, Mohammad Mehdi Fateh²

¹,² Department of Electrical and Robotic Engineering, Shahrood University of Technology, 3619995161 Shahrood, Iran.

Abstract

In this paper, a novel model-free control scheme is developed to enhance the tracking performance of robotic systems based on an adaptive dynamic sliding mode control and voltage control strategy. In the voltage control strategy, actuator dynamics have not been excluded. In other words, instead of the applied torques to the robot joints, motor voltages are computed by the control law. First, a dynamic sliding mode control is designed for the robotic system. Then, to enhance the tracking performance of the system, an adaptive mechanism is developed and integrated with the dynamic sliding mode control. Since the lumped uncertainty is unknown in practical applications, the uncertainty upper bound is necessary in the design of the dynamic sliding mode controller. Hence, the lumped uncertainty is estimated by an adaptive law. The stability of the closed-loop system is proved based on the Lyapunov stability theorem. The simulation results demonstrate the superior performance of the proposed adaptive dynamic sliding mode control strategy.

Article Info

Keywords: Adaptive Dynamic Sliding Mode Control, Model-Free Tracking Control, Robotic Systems, Voltage Control Strategy.

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I. INTRODUCTION

Since robotic systems are always affected by the environmental disturbances, it is necessary to develop robust and adaptive controllers to suppress the effects of them [1-6]. Sliding mode control (SMC) is a well-known robust control technique that has good tracking performance due to its robustness against the uncertainties and disturbances.

However, this control strategy suffers from the chattering problem, due to the discontinuous control law [7-9]. A common method adopted to improve the chattering is to replace the switching function by the saturation function. Since the chattering is reduced by this method, an indefinite steady-state error is also caused depending on the selection of the boundary layer. Thus, the chattering and accuracy become a tradeoff problem in this design [10,11]. Another technique is to reduce the switching gain in the controller. However, if the controller is not powerful enough to confront the uncertainties, the robustness of SMC becomes poor.

Dynamic sliding mode control (DSMC) is an effective scheme to reduce the control chattering. The time derivative of the control signal is considered as the new control variable for the augmented system in which the augmented system includes the original system and the integrator. In DSMC, the chattering problem can be effectively reduced due to the integration method in obtaining the control signal [12-14]. However, similar to SMC, the uncertainty bound should be known in the design of DSMC. In this paper, to overcome this problem, the uncertainty is estimated using an adaptive mechanism. In [15-19], the Fourier series expansion is used for the controller design. These methods require more computation, whereas the proposed method is simpler and
can alleviate the computation burden. Many studies focused on the torque control strategy (TCS) of robotic systems. In this strategy, the control law computes the torques which should be produced by the motors. The system actuators should be excited, so that they produce the desired torques. However, the actuator dynamics are not considered in the TCS and its input is not calculated in this strategy. To solve the aforementioned problem, voltage control strategy (VCS) has been developed which is more effective and requires less computation [20-23]. As a result, voltage-based approaches are superior. Therefore, in this paper, using VCS, an adaptive DSMC is developed for robust control of robotic systems.

The purpose of this paper is developing an adaptive dynamic sliding mode controller for tracking control of robotic systems. The proposed method is model-free and does not require the robot dynamics. The proposed method does not need the uncertainty upper bound. In fact, the lumped uncertainty is estimated using an adaptive rule. Using Lyapunov direct method, it is guaranteed that the tracking errors converge to zero.

The remainder of the paper is organized as follows. Section 2 introduces the robotic problem formulation. In Section 3, the proposed adaptive dynamic sliding mode controller is designed. Stability analysis is presented in Section 4. Simulation results are discussed in Section 5. Our conclusions are given in Section 6.

II. Problem Formulation

The dynamics of a robotic system can be described as [24]

\[ D(q)\ddot{q} + C(q, \dot{q})\dot{q} + G(q) = \tau_t \]  

(1)

\[ J_m r^{-1}\dot{\ddot{q}} + B_m r^{-1}\dot{\dot{q}} + r\tau_t = K_m I_a \]  

(2)

\[ R I_a + K_m r^{-1}\dot{\dot{q}} + \ddot{\xi} = v(t) \]  

(3)

Where \( \ddot{\xi} = LI_a + d \), \( Q \) is the vector of joint positions, \( V \) is the vector of motor voltages, \( I_a \) is the vector of motor currents and \( d \) is a vector of external disturbances. The details are completely explained in [15].

Substituting \( \tau_t \) from (1) into (2) results in

\[ J_m r^{-1}\dot{\ddot{q}} + B_m r^{-1}\dot{\dot{q}} + r(D(q)\ddot{q} + C(q, \dot{q})\dot{q} + G(q)) = K_m I_a \]  

(4)

Using (4), one can calculate \( I_a \) as

\[ I_a = k_m^{-1} \left( (J_m r^{-1} + rD(q)\ddot{q}) + (B_m r^{-1} + rC(q, \dot{q})\dot{q} + rG(q)) \right) \]  

(5)

Substitution of (5) into (3) yields

\[ v = R k_m^{-1} \left( (J_m r^{-1} + rD(q)\ddot{q}) + (B_m r^{-1} + rC(q, \dot{q})\dot{q} + rG(q)) + k_m r^{-1}\dot{\dot{q}} + \ddot{\xi} \right) \]  

(6)

Now, (6) can be rewritten as

\[ v = \ddot{\xi} + C_i \dot{\dot{q}} + G_i + \ddot{\xi} \]

\[ \ddot{\xi} = R k_m^{-1}(J_m r^{-1} + rD(q)\ddot{q} + B_m r^{-1} + rC(q, \dot{q})\dot{q} + rG(q)) \]

Using (8) and (9), we can rewrite (7) as

\[ v = \ddot{\xi} + g \]

(8)

III. The proposed Adaptive Dynamic Sliding Mode Control

The structure of the proposed adaptive controller is shown in Fig. 1. In this block-diagram, error, the first and second sliding surfaces, adaptive rule, and integrating the control signal have been clearly illustrated.

The tracking error \( e \) and its time derivatives \( \dot{e}, \ddot{e} \) are defined as follows:

\[ e = q_a - q \]

(9)

\[ \dot{e} = \dot{q}_a - \dot{q} \]

\[ \ddot{e} = \ddot{q}_a - \ddot{q} \]

Now, define a sliding surface as follows:

\[ s(t) = \dot{e}(t) + a_e e(t) + a_2 \int_0^t e(\tau)d\tau \]

(10)

The time derivative of (10) becomes

\[ \dot{s}(t) = \dot{\dot{e}} + a_e \dot{e} + a_2 e \]

(11)

Substituting \( \dot{\dot{e}} \) from (9) into (11) results in

\[ \dot{s}(t) = \dot{\ddot{q}}_a - \dot{\dot{q}} + a_e \dot{e} + a_2 e \]

(12)

Using (8), (12) can be written as

\[ \dot{s}(t) = \dot{\ddot{q}}_a - \dot{\dot{q}}_a + a_e \dot{e} + a_2 e \]

(13)

The secondary sliding surface is considered as

\[ \sigma(t) = \dot{s}(t) + b_s \dot{s}(t) + b_2 \int_0^t s(\tau)d\tau \]

(14)

The time derivative of (14) becomes

\[ \dot{\sigma}(t) = \ddot{s}(t) + b_s \ddot{s}(t) + b_2 \int_0^t \dot{s}(\tau)d\tau \]

(15)

Using (10) and (13), (15) can be written as

\[ \dot{\sigma}(t) = \ddot{\ddot{q}}_a - \ddot{\dot{q}}_a - v + g \]

(16)

Using (8) and (9), we have

\[ \dot{e} = \ddot{q}_a - v + g \]

(17)

Substitution of (17) into (16) yields
\[ \sigma(t) = \ddot{q}_d - \dot{v} + g(t) + a_1(\ddot{q}_d - v) + a_2 \hat{\nu} + b_1 \ddot{q}_d \\
- b_2 \dot{v} + b_1 g(t) + a_2 \ddot{\nu} + a_2 \dot{v} e(t) + a_2 \ddot{e}(t) \]

\[ + a_2 b_2 \int_0^t e(\tau) d\tau \]

(18)

we can rewrite (18) as

\[ \dot{\sigma}(t) = \ddot{q}_d - \dot{v} + \mu_1 \dot{Y}(t) + f(t) + \mu_2 \dot{e} + \mu_3 \dot{e} + \mu_4 \int_0^t e(\tau) d\tau \]

(19)

Where \( Y(t) = \ddot{q}_d - v \), \( f(t) = \mu_4 g + \dot{\nu} \), \( \mu_1 = a_1 + b_2 \), \( \mu_2 = a_2 + a_2 \dot{\nu} + b_2 \), \( \mu_3 = \mu_1 + a_2 \dot{\nu} + b_2 \) and \( \mu_4 = a_2 \dot{\nu} + b_2 \). The control law in adaptive DSMC is proposed by

\[ v_{ADSMC} = \ddot{q}_d + \mu_1 \dot{Y}(t) + \mu_2 \dot{e} + \mu_3 \dot{e} + \mu_4 \int_0^t e(\tau) d\tau + v_r \]

(20)

where \( \dot{f} \) is the estimation of \( f \) and \( v_r \) is the robust control term which will be calculated in the next section. It follows from (19) and (20) that

\[ \dot{\sigma} = \ddot{f} - v_r \]

\[ \dot{f} = f - \dot{f} \]

(21)

The sampling interval in the experiment is short enough as compared with the variation of \( f \), thus, the term \( f \) is also assumed to be a constant during the estimation (i.e. \( \dot{f} = f - \dot{f} \rightarrow \dot{f} = -\dot{f} \)) [25-28].

\[ V = \sigma \sigma - \frac{1}{y} \dot{f} \dot{f} \]

(23)

Using \( \dot{\sigma} \) defined in (21) we have

\[ V = \sigma (\dot{f} - v_r) - \frac{1}{y} \dot{f} \dot{f} \]

(24)

The adaptive law can be proposed as follows:

\[ \dot{f} = \gamma \sigma \]

(25)

Using (25), one can easily conclude that

\[ V = -\sigma v_r \]

(26)

where the robust control term is selected as follows:

\[ v_r = k \sigma \]

(27)

Substituting (27) into (26), we have

\[ V = -k \sigma^2 \]

(28)

Therefore, it has been guaranteed that \( V \leq 0 \). Using Barbalat’s lemma [29], it can be found the tracking error asymptotically converges to zero.

Remark: The final Lyapunov function for the total robotic system is the sum of Lyapunov function as defined in (22).

\[ V = \frac{1}{2} \sigma^2 + \frac{1}{2y} \dot{f}^2 \]

(22)

Differentiating the Lyapunov function, we have

\[ V = \frac{1}{2} \sigma^2 + \frac{1}{2y} \dot{f}^2 \]

\[ V = \sigma \sigma - \frac{1}{y} \dot{f} \dot{f} \]

IV. STABILITY ANALYSIS

Consider a Lyapunov function candidate:

\[ V = \frac{1}{2} \sigma^2 + \frac{1}{2y} \dot{f}^2 \]

(22)

V. SIMULATION RESULTS

In order to demonstrate the performance of the proposed controller, an articulated robot is considered, with a symbolic representation given in Fig. 2. The parameters of the robotic system are presented in [30]. The Denavit–Hartenberg (DH) parameters of the robot and the parameters of permanent
magnet dc motors are given in Table 1 and Table 2, respectively. The external disturbance $d$ is a step function with the amplitude of 2 volts which is inserted into the system at $t = 6$ sec. The maximum voltage of each motor is set to $v_{\text{max}} = 40\text{V}$. The desired position for each joint is formulated by

$$q_d = 1 - \cos(\pi t / 7)$$

(29)

This desired trajectory is shown in Fig. 3. The sliding surface parameters have been selected as $a_1 = 10$, $a_2 = 0.1$, $b_1 = 20$, $b_2 = 0.2$. The parameters $\gamma$ and $k$ have been set to 3000 and 10, respectively.

<table>
<thead>
<tr>
<th>Link</th>
<th>$\theta$</th>
<th>$d$</th>
<th>$a$</th>
<th>$\alpha$</th>
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<tr>
<td>1</td>
<td>$\theta_1$</td>
<td>$0.28$</td>
<td>$0$</td>
<td>$\pi/2$</td>
</tr>
<tr>
<td>2</td>
<td>$\theta_2$</td>
<td>$0$</td>
<td>$a_2 = 0.76$</td>
<td>$0$</td>
</tr>
<tr>
<td>3</td>
<td>$\theta_3$</td>
<td>$0$</td>
<td>$a_3 = 0.93$</td>
<td>$0$</td>
</tr>
</tbody>
</table>

**TABLE I**

The Denavit–Hartenberg parameters

<table>
<thead>
<tr>
<th>Motor</th>
<th>$R$</th>
<th>$J_m$</th>
<th>$B_m$</th>
<th>$K_m$</th>
<th>$r$</th>
<th>$L$</th>
</tr>
</thead>
<tbody>
<tr>
<td>1,2,3</td>
<td>1.26</td>
<td>0.0002</td>
<td>0.001</td>
<td>0.26</td>
<td>0.01</td>
<td>0.001</td>
</tr>
</tbody>
</table>

**TABLE II**

The motor parameters

| FIG. 2. The symbolic representation of the articulated robot |

Fig. 4 shows the tracking performance of ADSMC. According to Fig. 4, it can be found that the proposed controller design procedure performs well and has fast tracking performance. Also, the control inputs are depicted in Fig. 5. As shown in this figure, these signals are smooth and do not exceed the allowable values. From Figs. 4 and 5, the simulation results indicate that the favorable tracking performance is obtained.

Now, the proposed method is compared with the DSMC method. In the DSMC, the control signal is as follows:

$$v_{\text{DSMC}} = \ddot{q}_d + \mu_1 \dot{q}(t) + \mu_2 \dot{e} + \mu_3 e + 
\mu_4 \int_0^t e(\tau) d\tau + \rho \text{sgn}(\sigma), \quad |\sigma| < \rho$$

(30)

$$v_{\text{DSMC}}(t) = \int_0^t v_{\text{DSMC}}(\tau) d\tau$$

The value of $\rho$ is selected as 20. In the DSMC method, the sliding surface parameters are chosen as the proposed method, i.e. $a_1 = 100$, $a_2 = 400$, $b_1 = 200$, $b_2 = 600$ in the DSMC. The corresponding tracking errors and control efforts are shown in Figs. 8 and 9, respectively. As illustrated in Fig. 8, the tracking performance of the DSMC is improved, but it is still weak comparing to the proposed ADSMC method. Consequently, according to Figs. 4-9, it is concluded that the suggested ADSMC method has a significant advantage over the DSMC method regarding the tracking performance.

**FIG. 3. The desired trajectory**

![Fig. 3. The desired trajectory](image-url)
Remark: Increasing the parameter $k$ will increase the amplitude of the control signal and actuator saturation will occur. For example if $k = 100$, the control signal will exceed the permitted range. For the sliding surface parameters $a_1$ and $b_1$, small values such as 1 or 2 will result in poor tracking performance and the tracking errors cannot
converge to zero. For the sliding surface parameters $a_2$ and $b_2$, large values will increase the overshoot in the tracking error profile. Due to these issues, the controller parameters have been adjusted using the trial and error process. Also, optimization algorithms such as PSO, BA, ILCOA, GA, OSA, WOA or GPEA [31-44] can be used for tuning the controller parameters.

Finally, in order to show the efficiency of using the second sliding surface, the performance of the proposed method is compared with that of the conventional sliding mode control [40] in which the control signal is designed as:

$$v_{SMC} = \ddot{q} + a_1 \dot{\sigma} + a_2 \sigma + \rho \text{sgn}(\sigma), \quad |\dot{\sigma}| < \rho_1$$ (31)

Where $\rho_1 = 10$, $a_1 = 100$ and $a_2 = 400$.

The tracking errors and control signals related to the conventional sliding mode control are illustrated in Fig. 10 and Fig. 11, respectively. As it can be seen the control signals are affected by the chattering phenomenon which is not desirable. In fact, in conventional sliding mode control, we are faced with the chattering problem. In order to overcome this problem, dynamic sliding mode control can be used. Moreover, in order to reduce the tracking error, adaptive dynamic sliding mode control is proposed in this paper. These comparisons reveal the superiority of the proposed method.

**VI. Conclusion**

A new approach based on an adaptive dynamic sliding mode control and VCS has been developed for tracking control of robotic systems. First, a Dynamic sliding mode control scheme has been designed properly. Then, in order to reduce the chattering, an adaptive mechanism has also been developed. The control signal is designed using voltage control strategy. The stability of the robotic system has been proven based on the Lyapunov criteria. The computer simulation results show that the voltage-based adaptive dynamic sliding mode controller can perform successful control and achieve desired tracking performance.

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Mohammad Reza Shokoohinia received his M. Sc. degree in Control engineering at Garmsar University, Iran in 2012. He is now a Ph. D. candidate in Control engineering at Shahrood University of Technology. His research interests include Dynamic Sliding Mode Control, Robotics and Intelligent Systems.
E-mail: mshokoohinia@yahoo.com (Corresponding author).

Mohammad Mehdi Fateh received his M. Sc. degree in electrical engineering from Tar-biat Modares University, Iran in 1991. He received his Ph. D. degree in robotic Engineering from Southampton University, UK in 2001. He is a full professor with the Department of Electrical and Robotic Engineering at Shahrood University of Technology, Iran. His research interests include non-linear, robust and fuzzy control, robotics and intelligent systems.
E-mail: mmfateh@shahroodut.ac.ir
Quantization Noise Reduction Using Random Dither in Direct Torque Control of an Induction Motor

Amir Ghasemian and Asghar Taheri

1, 2Department of Electrical and Computer Engineering, University of Zanjan, Zanjan, Iran

A direct torque control (DTC) induction motor drive is presented in this paper. Quantization errors of current and voltage measurements are simulated and considered. To reduce the average quantization error and other offset errors of current and voltage measurement and eliminating the increasing integrator errors, a random dither signal is added to the truncating analog to digital converter (ADC) outputs. In this method, the ADC mean error is reduced to zero and therefore, integrator output error is mitigated. The proposed quantization method can improve the digital converter result; thus, this method can decrease the current measurement result. Thus, the torque and flux ripples were decreased. The proposed dither injection method can be used in Digital signal processor (DSP) or FPGA implemented applications. Experimental results show the performance of the proposed method.

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I. INTRODUCTION

Digital control systems need analog to digital converter (ADC) to measure physical data. ADC operation can be modeled by two sub-processes called sampling and quantization [1-4]. The sampling operation incurs no loss of information as long as the input is band-limited in accordance with the sampling theorem [5]. But during quantization, truncating or rounding of the analog signal to a digital signal level, always result in signal degradation [5]. Also, offset error is important where the digital signal must be integrated. Integrating a signal with a small offset error causes the output error to increase linearly by time. In direct torque control (DTC) of an induction motor, estimation of the flux needs integration of the motor current [1, 6, 7]. Therefore, if a small offset error flaws in the current measurement then the flux estimation error will increase linearly.

The signal dithering is a method to overcome these errors; this method has many applications in digital audio and image processing [5, 8]. In control systems, the dithering has not been used commonly as a method to reduce input noises, but it is used as a method by which high-frequency disturbances is introduced to a slow dynamic system to suppress problems like static friction and squeal [9-11]. In this paper, the dithering is used to reduce the mean error of measuring stator currents to zero and suppress the integration of the torque and the flux calculation errors. The DTC method controls both the stator flux and the electrical torque decoupled without using any current feedback, PWM algorithm, or rotary coordinate conversion module. Therefore, compared with the field-oriented control method, the DTC technique has a
simpler structure, faster torque response, and better robustness against parameters changes [12]. However, the basic switching-table-based DTC method has some outstanding drawbacks such as variable switching frequency due to the presence of the hysteresis controllers. Also, this method has high torque and flux ripples due to the low number of inverter switching states, and high sampling requirement [13-15].

A suitable look-up table for DTC of three-level dual voltage source inverter fed open-ended winding IM drive is proposed [15], where the VVs selection for lower hysteresis boundary conditions of torque and flux are restructured with null voltage states.

Two simple control methods with a fixed frequency are proposed for improving DTC of induction machines, which is named as CFTC-DTC in [16]. The CFTC-DTC was initially introduced to reduce torque ripple and achieve constant switching frequency in inverters. However, when compared to the original DTC, the CFTC-DTC algorithm suffers from slow torque dynamic response owing to the selection of zero-voltage vectors during torque transient [16].

A suitable duty cycle control technique to reduce torque, flux ripples, and harmonic loss in DTC of six-phase induction machine is proposed in [4]. Three different methods are proposed to improve the performance of DTC Switching Table of machine. The quantization noise cancellation method is used in other applications [17 and 18]. Using of the dither injection method can be used any application with ADC to decrease the error of digital converter. According to these technical points, the quantization noise reduction of the DTC of an induction motor is very important.

### II. DIRECT TORQUE CONTROL

In a two-level inverter, a DC voltage is inverted to a three-phase output, by means of two switches for each output phase. These six switches can be commanded in six different ways. For each value of the motor torque and the vector form of the stator flux, a suitable switching state can be selected to excite the motor to simultaneously control the flux and torque. This is the basic principle of the DTC scheme.

Figure (1) shows the DTC diagram. The torque and flux vectors must be estimated or calculated from measurable motor parameters. Estimated torque and flux are compared with their desired values and the error signals go to hysteresis blocks to determine the desired changes of the flux and torque. For torque error signal, the hysteresis block is a tree-level hysteresis block. But for flux error, it is a two-level block. Then the suitable switching states are determined by a switching table. Also, the appropriate switching states depend on the angle of the stator flux in the vector space.

![Used DTC method diagram](image1)

If the vector space is divided into six different sectors, as shown in figure (2), each sector has its own suitable switching states depending on the hysteresis block outputs.

![Two-level inverter switching vectors and stator flux sector](image2)

Table (1) shows a switching table for a two-level inverter DTC system. In this way, the torque and flux of the induction motor can remain with a suitable hysteresis band around their desired values, and therefore can be controlled directly by changing the switches.

<table>
<thead>
<tr>
<th>Sector</th>
<th>1</th>
<th>2</th>
<th>3</th>
<th>4</th>
<th>5</th>
<th>6</th>
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</thead>
<tbody>
<tr>
<td>Flux</td>
<td>Increase</td>
<td>V2</td>
<td>V3</td>
<td>V4</td>
<td>V5</td>
<td>V6</td>
</tr>
<tr>
<td>Torque</td>
<td>No Change</td>
<td>V0</td>
<td>V7</td>
<td>V0</td>
<td>V7</td>
<td>V0</td>
</tr>
<tr>
<td></td>
<td>Decrease</td>
<td>V6</td>
<td>V1</td>
<td>V2</td>
<td>V3</td>
<td>V4</td>
</tr>
<tr>
<td>Increase</td>
<td>V3</td>
<td>V4</td>
<td>V5</td>
<td>V6</td>
<td>V1</td>
<td>V2</td>
</tr>
<tr>
<td>No Change</td>
<td>V7</td>
<td>V0</td>
<td>V7</td>
<td>V0</td>
<td>V7</td>
<td>V0</td>
</tr>
<tr>
<td>Decrease</td>
<td>V5</td>
<td>V6</td>
<td>V1</td>
<td>V2</td>
<td>V3</td>
<td>V4</td>
</tr>
</tbody>
</table>

### III. QUANTIZATION AND OFFSET ERRORS

The process of converting an analog signal to a digital value in ADCs consists of two stages; one is the sampling, and the other is the quantization (figure 3). During the sampling stage, the input analog signal periodically is held to give enough time to the next stage to measure and convert the
analog signal. During the quantization stage, the held analog signal is truncated or rounded to a digital neighbor level and then converted to a binary digital number.

**Fig. 3.** The quantization model

An ideal ADC uses a rounding method; therefore, the mean output error will be zero. If the ADC uses the truncation method, a mean offset error of value $1/2$ LSB, will be added to the actual mean value of the original signal. Today many ADCs use the rounding method to eliminate this offset error because if this digitized signal needs to be integrated, this small offset error will produce serious problems.

Earlier airplanes with mechanical computers were working more accurately, during fly, because of high-frequency vibrations. This high-frequency vibration was called dither [12]. After that, dither as a technique found many applications in mechanical and electromechanical systems and digital signal processing [5]. In the mechanical and the electromechanical systems, dither means superposition of high-frequency vibration to stabilize a low-frequency vibration [10]. Suppressing squeal in car wiper system [10] or brake system, [11], are two examples of these applications. However, in digital signal processing, it has a little different meaning. In this area, dithering is an intentionally applied form of noise to reduce quantization errors [5].

### IV. DTC SIMULATION

The DTC of an induction motor is simulated in MATLAB Simulink. In this simulation, the induction motor is modeled in $d$-$q$ coordinate system. A simple PID controller measures the motor speed and determines a suitable torque setpoint for the DTC. The DTC measures the motor currents and voltages and estimates its torque and flux vectors. Then, the appropriate switching state is calculated to keep the motor torque and flux in their determined boundaries.

**Fig. 4.** DTC system simulation in MATLAB Simulink

In figure 7(a), averaged voltages are shown. This figure shows that the averaged voltages which are applied to the motor are three-phase sinusoidal voltages. The motor current in phase (a) is shown in figure 7(b). In the experimental setup, two phases of these currents are measured with ADCs and the third phase current is calculated.

**Fig. 5.** (a) Stator flux magnitude, (b) Stator flux vector in $d$-$q$ coordinate
V. SIMULATION OF THE QUANTIZATION ERROR

To consider ADC error, small ADC blocks designed in the model. To model ADC, firstly, the transfer function of an ideal ADC [5] is used. In the used DTC system, the controller measures the currents of two phases of the motor using hall effect sensors, and then the third phase current is calculated using these two measurements. Therefore two current sensors are sufficient for this controller. Figure 8(a) shows the current measurement of phase a, using this ideal ADC that uses rounding method for the quantization. Since three-phase currents are dependent on each other, Ia and Ib are measured by two ADCs and Ic is calculated. The quantization error of measurement Ia is shown in figure 8(b). As it is shown in this figures, the average quantization error of ideal ADCs over a sufficient period of time will be zero.

Zero average ADC errors help the integration carried out properly. Also, integrator blocks do not increase errors. The resulting motor direct component (d-axis) of flux estimation error, and the total flux magnitude error, using ideal ADC models are shown in figures 9 (a) and (b), respectively.

These flux errors are small enough and do not increase with time. Therefore, the motor flux is estimated in a suitable range around its actual value.

VI. SIMULATING WITH OFFSET ERRORS

Although many of today’s ADCs removed truncation offset errors, they still have other sources of offset errors. These offset errors are calculated during the fabrication process, and they are mentioned in the ADC datasheets. Therefore compensating these offset errors is a problem when long interval integrations are carried out. To model ADCs with the truncation method, a constant offset is added to the ideal ADC model prepared in the previous section. This way the truncating ADC offset error is simulated by setting the offset equal to 0.5 LSB. Also, other offset errors can be simulated by changing this value. Measured phase a current is shown in figure 10 (a). As mentioned above, the mean quantization error is not zero in this case. The average quantization error for this ADC has a negative value as shown in figure 10 (b).
According to figure 11 (a), the current measurement errors resulted in increasing errors in the direct flux component. Figure 11(b) shows the increasing fluctuation in the flux magnitude. As it is shown in figure 12 (a), the actual flux magnitude has gone away from its estimated value; and it is not close to the estimated flux magnitude. The circular graph of the actual stator flux vector is shown in figure 12 (b). According to that, the controller controls its estimated flux around its desired values, but the actual flux goes away from its desired value, as the error increases.

VII. DITHER INJECTION

To eliminate the current measurement offset error, we cannot add a constant to the ADC output. Because the ADC offset error is 0.5 LSB and this error will be vital if it is integrated.
As it is shown, adding dither changed the mean error back to zero. Again, by decreasing the mean errors to zero, the increasing flux estimation errors are disappeared. The resulting motor direct component of flux estimation error, and the total flux magnitude error, using dither injection are shown in figures 14 (a) and (b), respectively.

Experimental results were added to the manuscript. The experimental results are shown below. The proposed methods is developed and tested in the experimental setup. The proposed techniques in DTC of IM are implemented on a Tms320f28337 board. Experimental results of the proposed techniques in DTC IM are shown in Figure 15. Figure 15 shows experimental results of load torque, motor speed, stator flux of d axis according to q axis, phase current, and flux amplitude.

Figure 15. Experimental results of DTC method, a) motor load torque and speed, b) stator flux of d axis according to q axis, c) phase current, and d) flux amplitude
VIII. CONCLUSION

A random digital dither signal is used to reduce the mean offset error of the motor current measurement to zero. Simulation results show that the increasing integration errors eliminated, and the motor torque and flux controlled properly as the case where an ideal ADC without offset errors is used. According to results, the proposed method decreases the ADC error in DTC of induction motor and improve motor performance.

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Amir Ghaseemian was born in Mazandaran, Iran, in 1984. He received the B.Sc. degree from Iran University of Science and Technology, Tehran, Iran, in 2007 and the M.Sc. degree in 2010, and Ph.D. degree with the Faculty of Electrical and Computer Engineering, University of Zanjan, Zanjan, Iran, all in electrical engineering. His current research interests include modeling, analysis, and control of power converters, digital control of dc–dc converters, switched and hybrid dynamical systems, and power electronic systems for renewable energy sources.

Asghar Taheri was born in Zanjan, Iran, in 1977. He received the B.S., M.S., and Ph.D. degrees in electrical and electronics engineering from Amirkabir University of Technology, Tehran, Iran, and Iran University of Science and Technology, Tehran, in 1999, 2001, and 2011, respectively. Since 2010, he has been a Faculty Member with the University of Zanjan, where he was an Assistant Professor from 2011 to 2016 and has been an Associate Professor since 2016. His current research interests include modeling, analysis, and control of power converters, motor drives and control, and multiphase machine drives, multilevel inverter, Z-source inverter, power electronic systems for renewable energy sources, process control, DSP and FPGA based system designs, hardware in the loop, and computer aided control.
A Novel Complete Dynamic and Static Model of 48-Pulse VSC-based GUPFC for Parallel Transmission Lines

Mahyar Abasi1, Mahmood Joorabian2,†, Alireza Saffarian3, and Seyyed Ghodratollah Seifossadat4

1,2,3,4 Department of Electrical Engineering, Faculty of Engineering, Shahid Chamran University of Ahvaz, Ahvaz, Iran

A Generalized Unified Power Flow Controller (GUPFC) is a member of the flexible alternating current transmission system (FACTS) devices family that operates based on voltage source converter (VSC) and is known as the most efficient FACTS device. GUPFC can control the voltage of one bus and the active and reactive power flows on at least two transmission lines with equal voltage levels. This paper presents the mathematical modeling of power injection by GUPFC for the first time. Besides, the accurate design and details of the control system for series and shunt converters of a GUPFC, along with a new mathematical function for pulse generation based on a 48-pulse VSC when the GUPFC is placed in the middle point of a parallel transmission line, are presented in this study. The power injection modeling introduced in this paper is very useful and efficient in Newton-Raphson power flow studies and in modeling different parts of the control system and power electronics converter in dynamic and transient studies. The modeling is implemented in MATLAB/Simulink for a 400 km, 230 kV, and 60 Hz nominal frequency double-circuit transmission line. The satisfactory results provided in the simulations section of the paper verify the validity and accurate performance of the proposed model.

Article Info

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I. INTRODUCTION

Various solutions have so far been proposed for solving the problems caused by electrical energy transmission. One of the most significant and efficient solutions is the application of FACTS devices. Advances in the manufacturing of power electronics converters with high-speed switching [1] and the improved technology of control processors have led to a dramatic growth of novel ideas in designing FACTS devices [2]. To date, 10 types of compensators from the FACTS family have virtually been introduced in terms of theory as well as the experimental and practical manufacturing processes [3]. In 2000, Fardanesh et al. introduced, for the first time, a new concept of the most complete and advanced type of multi-converter FACTS, called GUPFC. Fig. 1 shows the various topologies of GUPFC in the power system. Generally speaking, this device consists of n (n ≥ 2) series and shunt converters [4].

†Corresponding Author: mjoorabian@scu.ac.ir
Tel: +98-9161183017, Department of Electrical Engineering, Faculty of Engineering, Shahid Chamran University of Ahvaz, Ahvaz, Iran

Once the GUPFC was introduced, similar FACTS devices including GIPFC and MUPFC were also presented [5-6]. These devices have two major common features. First, they can be placed on more than one single transmission line; second, regarding their special topology, each of them contains both series and shunt converters. The main difference between GUPFC and GIPFC is that GUPFC can be used on at least two parallel transmission lines with the same voltage level. However, GIPFC is placed on at least two transmission lines with different voltage levels. The major difference between GUPFC and MUPFC is in the number of shunt converters. GUPFC has only one shunt converter while the number of shunt converters in MUPFC is equal to the number of its series converters. Furthermore, MUPFC can be employed in transmission lines with equal or different voltage levels. After the publication of the first paper in the field of GUPFC, building and installing a GUPFC device was practically proposed in Xinjiang, China, due to several reasons including numerous problems in power transmission to long distances,
huge economic loss, and inappropriate power flow distribution which posed a risk to the system safety [7].

Basically, we can classify the literature on GUPFC into several main categories. A number of papers associated with GUPFC have focused specifically on the GUPFC modeling problem in the realm of power flow studies [8-14]. The mathematical modeling of GUPFC using the rating method, and a specific device for Newton-Raphson power flow studies have been presented for power systems [8]. The GUPFC modeling is discussed in the optimal power flow in the nonlinear internal point [9]. The authors in [10] provide the results obtained from utilizing GUPFC in a 28-bus, 330 kV in Nigeria for optimal power flow and voltage drop reduction. Ref. [11] argues the optimal location of GUPFC, the optimal power flow, and the modification of the Jacobian matrix in the presence of GUPFC. The developed model of GUPFC in the analysis of the Newton-Raphson power flow is given in [12] using the combined non-conformity method. Also, the development of GUPFC in Newton-Raphson method without modifying the Jacobian matrix and considering the power flow constraints is discussed in [13]. In another study, an impedance compensation method is used to solve numerical instability problems of power flow analysis in power systems equipped with the GUPFC [14].

The second category includes the dynamic studies and the analysis of various types of stabilities in the presence of GUPFC [15-20]. A control algorithm based on a fuzzy-neural network controller is presented in [15] to improve transient stability. A new algorithm for calculating the power quality index under steady-state and transient conditions is introduced in [16] using the harmonics technique in the presence of GUPFC. To study small-signal stability, a dynamic modeling based on the Heffron-Phillips model of a power system equipped with a GUPFC along with a PSS is proposed in [17]. An algorithm for enhancing the performance of GUPFC using a fuzzy logic-based controller is presented in [18]. Ref. [19] deals with the effect of GUPFC on small-signal stability of power systems. Also, the authors in [20] discuss the sub-synchronous resonance phenomenon in transmission lines compensated by GUPFC.

Still another group of studies concentrate on the operation and economic analyses of the presence of GUPFC in power systems [21-26]. A sensitivity-based approach for deciding on the proper location of GUPFC to perform an optimal power flow and its effect on power generation and transmission costs is suggested in [21]. Furthermore, a novel optimization algorithm for solving the economic problems of emissions and loss reduction in a power system equipped with GUPFC is presented in [22]. A new optimization method based on the PSO algorithm is given in [23] to optimize the generation cost objectives subject to various practical constraints in a power system equipped with GUPFC. The authors in [24] use a nonlinear internal point optimization method to calculate the transmission capability of a power system equipped with GUPFC. The increase in the loading capability of power system through the optimal placement of GUPFC based on PSO is presented in [25]. Line congestion management algorithm based on demand management for a combination of bilateral pool electricity market in a power system equipped with GUPFC is suggested in [26].

A different body of research deals with power quality assessment in power systems with GUPFC [27-28]. Ref. [27] studies the effect of GUPFC on the improvement of power quality and the total harmonic distortion (THD) level. Also, the concept of using GUPFC for reducing voltage flicker phenomenon and improving power quality is proposed in [28]. Finally, in the last group, Ref. [29] analyzes the positive effects of GUPFC on the distance relaying operation in transmission lines.

A thorough scrutiny of the literature shows that all power flow modeling studies performed in the presence of GUPFC for models given in Figs. 1 (a) and (b). Moreover, a major part of literature deals with the impact of GUPFC on power systems and discusses the accurate modeling of different control parts while related power electronics have rarely been investigated.

The topology given in Fig. 1(a) is used when the goal is to control the near-end bus voltage of the double-circuit line and to control power at a double-circuit line with distinctive remote-end buses with different voltages. The two circuits in the topology may differ in terms of distance. The topology given in Fig. 1(b) is employed when the goal is to regulate the near-end bus voltage and control power on one of the double-circuit lines with a common remote-end bus. The distances of both circuits in this topology are equal. The topology given in Fig. 1(c) is utilized when the aim is to regulate voltage along the path on the double-circuit line, especially in the middle of the line. The near and remote-end buses of both circuits in this topology are the same. According to the mentioned differences in different topologies provided in Fig. 1, the GUPFC modeling is different in Newton-Raphson power flow equations and in building the Jacobian matrix. Furthermore, control parameters in the series and shunt converters of the GUPFC in different topologies given in Fig. 1 are different because of power system stabilization in the presence of GUPFC, satisfying different constraints of power flow, and bus voltage control.

The main contributions of the present study are: 1) the accurate mathematical modeling of active and reactive power injections by GUPFC based on the topology given in Fig. 1 (c), which is exclusive for Newton-Raphson power flow studies, 2) the accurate modeling of different parts of the control system of the 48-pulse VSC-based GUPFC, and 3) the presentation of a new mathematical function to generate a 48-pulse signal required in the GUPFC, instead of using the classic 48-pulse signal generator system. The logic used in this block generates the firing angle signals of the 48-pulse VSC’s gate turn-off
Using harmonic filters along with FACTS devices to remove harmonics produced by voltage source inverter embedded within the filters leads to increased installation and operation costs of FACTS devices in the power system. Hence, to solve the power quality problem and reduce the harmonic distortions level in output voltage of the converters used in the FACTS devices, the paper presents 48-pulse voltage source inverters. The application of 48-pulse converters within the GUPFC helps reduce the harmonic level of the output voltage so that there is no need to use harmonic elimination filters, making it really cost-effective.

The paper is organized as follows. The mathematical modeling of the active and reactive power injection by the GUPFC is presented in Section 2 of the paper, which is exclusive to Newton-Raphson power flow studies. Section 3 describes the accurate design of the GUPFC's controller system. Section 4 discusses the inverter modeling of the 48-pulse VSC-based GUPFC using a new mathematical function for the pulse generator. Software simulations and the obtained results are provided in Section 5. Section 6 presents conclusions, and finally, the Appendixes include the test system data.

For mathematical modeling of the power equations of GUPF, Thevenin and Norton equivalent circuits are used. Fig. 2 shows such circuits for a GUPFC installed on a double-circuit line.

In addition, the complex power lost in such circuits for a GUPFC installed on a double-circuit line.
\[ S_{\text{imp}12} = \frac{V_1}{2} (I_{12})^* = \frac{V_{\text{sa}}^2}{|Z_{12}|^2} (R_{12} + j X_{12}) \quad (11-a) \]
\[ S_{\text{imp}34} = \frac{V_3}{2} (I_{34})^* = \frac{V_{\text{sh}}^2}{|Z_{34}|^2} (R_{34} + j X_{34}) \quad (11-b) \]

The exchanged active power of the shunt compensator is calculated in Eq. (12) with regard to the active power balance condition of the GUPFC:
\[ P_{\text{sh}} = -(P_{12}+P_{34}) \quad (12) \]
where \(P_{12}\) and \(P_{34}\) are found from (13-a) and (13-b):
\[ P_{12} = P_1 + P_2 - \frac{V_{\text{sa}}^2}{|Z_{12}|^2} R_{12} \quad (13-a) \]
\[ P_{34} = P_3 + P_4 - \frac{V_{\text{sh}}^2}{|Z_{34}|^2} R_{34} \quad (13-b) \]

Moreover, to calculate reactive power of the shunt converter, the same calculations as used for finding reactive power of the series converter is utilized in this step. Finally, (14) helps obtain the reactive power of the shunt converter.
\[ Q_{\text{sh}} = \frac{V_{\text{sa}}}{|Z_{\text{sh}}|^2} (X_{\text{sh}} \cos(\alpha_1-\alpha_{\text{sh}}) + R_{\text{sh}} \sin(\alpha_1-\alpha_{\text{sh}})) \quad (14) \]

The complex powers injected to buses 1 to 4 are calculated using (15) to (18).
\[ S_{1,\text{GUPFC}} = P_{1,\text{GUPFC}} + jQ_{1,\text{GUPFC}} = (P_1 + P_{\text{sh}}) + j(Q_1 + Q_{\text{sh}}) \quad (15) \]
\[ S_{2,\text{GUPFC}} = P_{2,\text{GUPFC}} + jQ_{2,\text{GUPFC}} = P_2 + jQ_2 \quad (16) \]
\[ S_{3,\text{GUPFC}} = P_{3,\text{GUPFC}} + jQ_{3,\text{GUPFC}} = P_3 + jQ_3 \quad (17) \]
\[ S_{4,\text{GUPFC}} = P_{4,\text{GUPFC}} + jQ_{4,\text{GUPFC}} = P_4 + jQ_4 \quad (18) \]

As a result, the equivalent circuit of the double-circuit line compensated by GUPFC is represented in terms of complex powers injected to buses 1 to 4, as shown in Fig. 3.

![Fig. 3. Equivalent circuit of the double-circuit line compensated by GUPFC in terms of the injected complex power.](image)

Once the steady-state model of GUPFC shown in Fig. 3 is extracted, a modification is applied to the Jacobian matrix and Newton-Raphson power flow equations based on the extracted model and power flow studies are carried out regarding the new model.

III. ACCURATE DESIGN OF THE GUPFC'S CONTROLLER SYSTEM

Referring to Fig. 4, the shunt converter controls the voltage of bus 1. This is performed through exchanging reactive power with the grid, i.e. via controlling the voltage of the DC-link. To realize this, at first, the proper angle \(\alpha\) for synchronization of voltage and current is calculated using a phase-locked loop (PLL). Then, using the abc-dq0 transform, the three-phase voltage and current components fed to the shunt converter are transformed into active and reactive components. Next, the measured voltage of the bus connected to the shunt converter is instantaneously compared with the reference voltage and is injected to the PI controller so that the error signal is calculated to generate the reference reactive current. In the next step, a comparison is made between the reference reactive current calculated in the previous step and the reactive power of the shunt converter-connected bus. The difference between these two quantities is used as the input in the proportional-integral (PI) controller to generate the phase angle alpha of the inverter voltage. D-alpha is another component used in this converter which is generated by applying the average DC-link voltage values and the reactive power to the PI controller. The last component required for pulse generation by the firing pulse generation system is sigma. Put differently, Sigma is the same modulation index that impacts the amplitude of the converter’s output voltage and its value is 172.5. Based on Fig. 1 (c), a GUPFC has two series converters with similar control systems. In this section, we investigate only one series converter knowing that the second one is identical. As per Fig. 5, the series converter is employed to automatically control power flow and adjust the current flowing from the transmission line. In this converter, similar to the shunt converter, the three-phase voltage measured from the bus is connected to one side of the series converter and the PLL calculates the proper angle, \(\alpha\), to synchronize voltage and current components. Then, the active and reactive components related to three-phase currents and voltages of the buses are calculated using the abc-dq0 transform. According to Fig. 5, the reference active and reactive currents can be found using the reference active power, the reference reactive power, the active voltage, and the reactive voltage. The calculated reference active and reactive currents are compared with their counterparts from the bus. After that, the error signal is fed to the proportional-integral-differential (PID) controller to calculate the reference active and reactive voltages. Next, a limiter controls the reference active and reactive voltages to prevent any possible system instability. Finally, the output of the real and imaginary voltage limiters is employed for calculating Alpha and Sigma. As the value of D_Alpha is obtained from the shunt converter, it is set zero in this section of the paper.

Coefficients \(k_i\) and \(k_p\) used in the series and shunt controllers of the GUPFC are calculated first by using the closed-loop Ziegler-Nichols method and then by trial and error to stabilize the test power system, achieve the balance among constraints, reach ideal operation conditions, and achieve power and power quality control.
IV. MODELING OF THE 48-PULSE VSC-BASED GUPFC’S INVERTER

Fig. 6 illustrates the inverter system considered for this compensator, which is a 48-pulse GTO voltage source converter and includes four three-level inverters and four phase-shifting transformers. All inverters considered for this converter can generate a three-phase square semi-sine voltage waveform using the three-level GTO bridge inverter. The secondary winding of the phase-shifting transformer uses three phase sequences of the series voltage wave series with the primary winding to generate a virtually sine voltage waveform. The amplitude of the voltage waveform can adopt one of these values: the DC-link voltage, the minus value of the DC-link voltage, or zero. Period of the zero voltage at each quarter-cycle can be defined as a dead-time with a value between 0 to 90°.

A. Analysis of 48-pulse VSC inverters

Two 24-pulse converters with a 7.5° phase shift with respect to each other form a 48-pulse converter. Using symmetry, a -3.75° phase-shift on the transformer connections of one of the 24-pulse converters and a +3.75° phase-shift on the other transformer create the required 7.5° phase-shift. Hence, the firing pulse needs +3.75° and -3.75° phase-shifts, respectively. Four 12-pulse converters that are interconnected using four 12-pulse transformers with phase-shifting windings are used to form a 48-pulse converter. Thanks to its perfect performance and low harmonic rate on the AC side, this converter, needless of AC filters, can be used for applications that demand high power. The output voltage contains harmonic orders of \( n = 48r \pm 1 \), where \( r = 0, 1, 2, \ldots \), meaning that harmonics of orders 47th, 49th, 95th, 97th, … have amplitudes of 0.47, 0.49, 0.95, … per unit on the base of the fundamental component. Also on the DC side, the lower flowing harmonic current will be considered as the 48th order harmonic component. The phase-shifting pattern for individual 12-pulse converters will be as follows [30]:

- Analysis of the first converter

The output voltage produced by the first converter is:

\[
v_{ab1}(t) = 2\left[V_{ab1}\sin(\omega t + 30°) + V_{ab1}\sin(11\omega t + 195°) + V_{ab13}\sin(13\omega t + 255°) + V_{ab2}\sin(23\omega t + 60°) + V_{ab25}\sin(25\omega t + 120°) + \ldots \right]
\]  

(19)

- Analysis of the second converter

The output voltage produced by the second converter is:

\[
v_{ab2}(t) = 2\left[V_{ab2}\sin(\omega t + 30°) + V_{ab2}\sin(11\omega t + 15°) + V_{ab2}\sin(13\omega t + 75°) + V_{ab23}\sin(23\omega t + 60°) + V_{ab25}\sin(25\omega t + 120°) + \ldots \right]
\]  

(20)

- Analysis of the third converter

The output voltage produced by the third converter is:

\[
v_{ab3}(t) = 2\left[V_{ab3}\sin(\omega t + 30°) + V_{ab3}\sin(11\omega t + 285°) + V_{ab3}\sin(13\omega t + 345°) + V_{ab2}\sin(23\omega t + 240°) + \ldots \right]
\]
\[ V_{ab25} \sin((25\omega t + 300^\circ) + \ldots) \]  
(21)

- Analysis of the fourth converter

The output voltage produced by the fourth converter is:

\[
v_{ab12}(t) = 2V_{ab} \sin(\omega t + 30^\circ) + V_{ab13} \sin(11\omega t + 105^\circ) + V_{ab14} \sin(13\omega t + 165^\circ) + V_{ab12} \sin(23\omega t + 240^\circ) + V_{ab25} \sin((25\omega t + 300^\circ) + \ldots)
\]  
(22)

These four AC output voltages, given by (19) to (22), are added to the series connection of secondary windings of the transformers. The following expresses the 48-pulse output AC voltage:

\[
v_{ab12}(t) = \sum_{n=1}^{\infty} V_{ab} \sin((n\omega t + 30^\circ) + \ldots) \quad \forall t
\]  
(25)

Applying an 11.25° phase-shift to (25) using a PST:

\[
v_{ab12}(t) = 2\sum_{n=1}^{\infty} V_{ab} \sin((n\omega t + 30^\circ) + 11/25^\circ) \quad \forall t
\]  
(26)

In the above equation, \(i = 1\) for the positive-sequence harmonics is the abc sequence and \(i = -1\) for negative-sequence harmonics is the cba sequence. The line-to-neutral voltage of phase a is as follows.

Now, by lagging the firing pulse by 11.25°, we have:

\[
v_{ab12}(t) = 2\sum_{n=1}^{\infty} V_{ab} \sin((n\omega t + 18/75^\circ + 11/25^\circ)) \quad \forall t
\]  
(27)

As a result, the general term for the line-to-line voltage of a 48-pulse converter is given by (28):

\[
v_{ab12}(t) = 8\sum_{n=1}^{\infty} V_{ab} \sin((n\omega t + 18/75^\circ + 11/25^\circ)) \quad \forall t
\]  
(28)

The line-to-neutral voltage is:

\[
v_{an_{4a}}(t) = \frac{8}{\sqrt{3}} V_{ab} \sin(\alpha t) - V_{ab12} \sin(47\omega t) + V_{ab13} \sin(95\omega t) + V_{ab14} \sin(97\omega t) + \ldots
\]  
(29)

Rewriting (29) gives:

\[
v_{an_{4a}}(t) = \frac{8}{\sqrt{3}} \sum_{n=1}^{\infty} V_{ab} \sin((n\omega t + 18/75^\circ + 18/75^\circ)) \quad \forall t
\]  
(30)

Voltages \(v_{an_{4a}}(t), v_{bn_{4a}}(t)\) have similar patterns except with \(120^\circ\) and \(240^\circ\) phase-shift with respect to \(v_{an_{4a}}(t)\).

### V. SOFTWARE SIMULATIONS AND RESULTS

Fig. 1 (c) provides the system under study in this paper. The system consists of one coupled double-circuit line with characteristics of 400 km, 230 kV, and 60 Hz representing the distributed model. Both ends of the system are modeled using a Thevenin equivalent circuit. The Appendix provides the detailed data related to the simulated system. The GUPFC employed in this study is placed in the middle of the coupled double-circuit line and its main task is to control active and reactive power flows between the two lines. The shunt converter considered for the GUPFC has a rated power of 200 MVA so that it can supply the power required by each series converter up to 100 MVA.

### B. Modeling the 48-pulse generator mathematical function

Various methods have so far been introduced for modulations in the power electronics industry, the most popular one of which is pulse width modulation (PWM). PWM methods are characterized by fixed-amplitude pulses where the pulse widths are modulated for controlling the modulator output voltage and reducing its harmonic content. Different methods of PWM include single-pulse modulation, multi-pulse modulation, and sine modulation, each of which differs from others in terms of the harmonic content in the output voltage. For PWM implementation, a digital IC such as a microcontroller or timer 555 can be used. The generated signal is a train of pulses forming a square waveform. In the related literature, various schemes have been introduced for the design of PWM's switching algorithm. The most well-known scheme is a design that compares wave and triangular waveforms. Fig. 7 shows a new mathematical function for describing pulse generation operations in a 48-pulse VSC. There are four switches at each leg of the inverter, with an overall number of 12 switches embedded for each inverter.

As mentioned in the control system design of the GUPFC, the inputs to the pulse generator function are \(\delta, \text{D}_\text{Alpha}, \text{Alpha}, \text{and} \omega t = \theta\). Function mod is used with angular values of \(2\pi\) and \(\beta\) for pulse generation. Angle \(\beta\) is influenced by half of the switch-off duration of switch Q1 in the half cycle \((-\pi/2)\). \(\text{D}_\text{Alpha}, \omega t = \theta\), and the angle of the beginning of each half-cycle, which can be zero or one. Then the output of the mod function is fed to the step function with a magnitude of one, the output of which is zero or one. Next, the values produced by step functions are sent as switching signals of GTOs. Finally, the operation of the two sets of switches (Q1 and Q3) and (Q2 and Q4) at each leg reverse each other. In this modeling, with no need for complex comparators of various waveforms and complicated pulse generation hardware and using only one simple mathematical function, the signals required for switching GTOs on and off are generated.
A. Simulation results for the output of the GUPFC’s shunt converter

Fig. 8 (a) illustrates the diagram of phase A’s secondary voltage for a shunt converter during one cycle. As is seen, the output voltage is a pure 48-pulse signal. Also, Fig. 8 (b) shows the FFT diagram of the output voltage and its THD value (= 4.35%). As expected, the values of the 47th and 49th harmonics are the greatest after the fundamental harmonic value.

Moreover, Fig. 9 shows the voltage of the secondary side and voltage and current of the primary side of the transformer connected to the shunt converter.

Fig. 10 illustrates the active and reactive powers measured at the shunt converter bus. The depicted powers in these diagrams are calculated using three-phase currents and voltages measured from the bus connected to the shunt converter. As is clear, the active and reactive power exchanges between the shunt converter and the grid are positive and negative, respectively. This means that the shunt converter receives active power from the grid to supply the internal loss power and charge the DC-link capacitor, and injects reactive power to the line to maintain the voltage of the bus connected to the shunt converter.

Fig. 7. New mathematical function for generating 48 pulses.

Fig. 8. (a) The output voltage of the 48-pulse VSC-based inverter for the parallel converter within a single periodic cycle. (b) Harmonic analysis diagram of the output voltage of the 48-pulse VSC-based inverter for the parallel converter within a
single periodic cycle.

Fig. 9. The voltage of the secondary side and voltage and current of the primary side of the transformer connected to the shunt converter.

Fig. 10. The active and reactive power diagram for the shunt converter.

Fig. 11 shows the voltage of the bus connected to the shunt converter. Based on the control constraints of the GUPFC, a task of GUPFC is to control the voltage of the bus. In other words, the shunt converter controls the voltage of the middle point of the line. As from Fig. 11 illustrates, the shunt converter suitably maintains the voltage of the middle point of the line close to the reference voltage.

Fig. 11. Diagram of the voltage of the bus connected to the shunt converter.

As is observed in Figs. 10 and 11, an oscillation is seen in the active and reactive power and shunt converter bus voltage 1 s after the start of the simulation. The reason behind this is the great variations in the value of the reference active power at both series converters (converter 1 and converter 2). This change in the power for series converter 1 from a value of 1 to 1.2 and for series converter 2 from 0.9 to -1.2 within 1 s causes changes in the operating mode of the GUPFC besides creating oscillation and transients in the active and reactive powers and voltage of the shunt converter because of satisfying power flow and voltage balance constraints. The established transient and oscillatory state, when the operating mode of the GUPFC changes is utmost two cycles, as shown in Figs. 10 and 11. It is rapidly damped and the system returns to the steady state.

Fig. 12 presents the voltage of the DC-link connected to three converters. As is observed, the voltage magnitude is maintained almost fixed. As discussed in the section related to the design of the control system, the main aim of controlling D_Alpha is to adjust and balance the capacitor voltages of the DC-link common among all converters. According to Fig. 12, the control of D_Alpha in the control system is performed acceptably.

Fig. 12. Voltage of the DC-link common among GUPFC’s converters.

B. Simulation results for the output of the GUPFC’s series converter 1

Figs. 13 and 14 present the results of active and reactive powers calculated by the control system and the dq reference frame for series converter 1. In these diagrams, active and reactive powers are calculated using the block diagrams provided in Fig. 5.

It is clearly seen that the values of reference active power for the series converter 1 during the time intervals of 0-0.5 s, 0.5-1 s, and 1-1.5 are 1 p.u., 1.1 p.u., and 1.2 p.u., respectively. Additionally, the reference reactive power for the series converter 1 during the whole period is considered -0.3 p.u. As seen in Figs. 13 and 14, series converter 1 is successful in controlling the active and reactive powers.

Fig. 13. Active power diagram of series converter 1.

Fig. 14. Reactive power diagram of series converter 1.

C. Simulation results for the output of the GUPFC’s series converter 2

Figs. 15 and 16 present the results of active and reactive powers calculated by the control system and the dq reference frame for
series converter 2. In these diagrams, active and reactive powers are calculated using the block diagrams provided in Fig. 5.

It is clearly seen that the values of reference active power for the series converter 2 during the time intervals of 0-0.5 s, 0.5-1 s, and 1-1.5 are 1 p.u., 0.9 p.u., and 1.2 p.u., respectively. Moreover, the reference reactive power for the series converter 2 during the entire period is considered −0.4 p.u. As seen in Figs. 15 and 16, series converter 2 is successful in controlling the active and reactive powers.

![Fig. 15. Active power diagram of series converter 2.](image)

![Fig. 16. Reactive power diagram of series converter 2.](image)

D. Simulation results for the active power balance of the GUPFC’s three converters

Figs. 17 and 18 present active power diagrams calculated by three-phase voltage and currents measured from buses located at the two ends of series converters 1 and 2. Fig. 19 shows the active power diagram of the shunt converter (Fig. 12) and the sum of active powers of series converters 1 and 2 (Figs. 17 and 18). As is seen from Fig. 19, the active power balance equation for the GUPFC is met.

![Fig. 17. The measured active power from the transformer of series converter 1.](image)

![Fig. 18. The measured active power from the transformer of series converter 2.](image)

![Fig. 19. Active power of the shunt converter along with the sum of active powers of the GUPFC’s two series converters.](image)

VI. CONCLUSION

This paper presents the accurate modeling of power injection by a GUPFC placed in the middle of a double-circuit line for the first time. The proposed model is very efficient in static studies such as the power flow. Moreover, accurate modeling of the control system and inverter of the 48-pulse VSC-based GUPFC, specific to dynamic studies, is provided. The control circuit utilized in the current study relies on the $dq$ theory. A new mathematical function that operates based on the PWM theory is presented for the pulse generation system to turn on/off the GTOs of the GUPFC’s inverter, where 48 pulses required for turning the GTOs on/off are produced. The suggested technique is robust against noise and can be implemented in various programmable ICs. Furthermore, according to the simulation results, the designed GUPFC can successfully satisfy control constraints and is very effective in terms of improving power quality and reducing harmonic levels.

APPENDIX

The detailed data of the double-circuit transmission line, Thevenin equivalent circuit of both sides of the line, the control system of the GUPFC, power electronic inverter of the GUPFC, and transformers placed between the GUPFC and the transmission line are listed in Table 1.
### Table I
Detailed data of the test system.

<table>
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<th>Line and source data</th>
<th>Source 1</th>
<th>Source 2</th>
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<tbody>
<tr>
<td>V</td>
<td>230 kV</td>
<td>230 &lt; 20.8 kV</td>
</tr>
<tr>
<td>X/R</td>
<td>10</td>
<td>10</td>
</tr>
<tr>
<td>r1</td>
<td>0.02546 Ohm/km</td>
<td>0.8716 Ohm</td>
</tr>
<tr>
<td>r2</td>
<td>0.3864 Ohm/km</td>
<td>1.3074 Ohm</td>
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<td>0 Ohm</td>
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<td>GUPFC data</td>
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<td>Current regulator2</td>
<td>K_r 15</td>
</tr>
<tr>
<td>Current regulator</td>
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<td>Current regulator2</td>
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<tr>
<td>Low pass filter</td>
<td>I_L 100</td>
<td>Mag Limit</td>
</tr>
<tr>
<td>D_Alpha Bpk</td>
<td>K_p 0.001</td>
<td>D_Alpha</td>
</tr>
<tr>
<td>Three level bridge</td>
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<td>Snubber resistance</td>
<td>Rs (ohm)</td>
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<tr>
<td>Snubber capacitance</td>
<td>Cs (F)</td>
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<tr>
<td>Internal resistance</td>
<td>Rm (ohm)</td>
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<tr>
<td>Power electronic device</td>
<td>GTO</td>
<td></td>
</tr>
<tr>
<td>Zigzag transformer phase shifting</td>
<td>Nominal power (VA)</td>
<td>25e6</td>
</tr>
<tr>
<td>Nominal frequency (Hz)</td>
<td>60</td>
<td></td>
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<tr>
<td>Primary nominal voltage (Volt-rms)</td>
<td>12.5e3</td>
<td></td>
</tr>
<tr>
<td>Secondary nominal voltage phase shift (Volt-rms)</td>
<td>12.5e3</td>
<td></td>
</tr>
<tr>
<td>Resistance winding 1 zig-zig (pu)</td>
<td>0.05/30</td>
<td></td>
</tr>
<tr>
<td>Inductance winding 1 zig-zig (pu)</td>
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</tr>
<tr>
<td>Resistance winding 2 zig-zig (pu)</td>
<td>0.05/30</td>
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Mahyar Abasi was born in 1989 in Malayer, Iran. He received the B.Sc. degree in electrical power engineering from the Islamic Azad University of Arak Branch, Arak, Iran, in 2012, and the M.Sc. degree in electrical power engineering from Shahid Chamran University of Ahvaz, Ahvaz, Iran, in 2015. He is currently working towards his PhD degree in the Department of Electrical Engineering, Shahid Chamran University of Ahvaz, Ahvaz, Iran. His research interests are fault location, power system protection, FACTS devices, and power quality assessment.

Mahmood Joorabian was born in Shooshtar, Iran, in 1961. He received his B.E.E degree from University of New Haven, CT, USA, M.Sc degree in Electrical Power Engineering from Rensselaer Polytechnic Institute, NY, USA and Ph.D. degree in Electrical Engineering from University of Bath, Bath, UK in 1983, 1985 and 1996, respectively. He has been with the Department of Electrical Engineering at Shahid Chamran University of Ahvaz as Senior Lecturer (1985), Assistant Professor (1996), Associate Professor (2004) and Professor (2009). His main research interests are fault location, FACTS devices, power system protection, applications of intelligence technique in power systems, operation and planning of deregulated power system, load and energy management, renewable energy, microgrid, smart grids, and power quality assessment.

Alireza Saffarian was born in 1981 in Ahvaz, Iran. He received the B.Sc. and M.Sc. degrees in electrical engineering from Amirkabir University of Technology, Tehran, Iran, in 2003 and 2005, respectively, and the Ph.D. degree from University of Tehran, Tehran, Iran, in 2011. Currently, he is an Assistant Professor with the Department of Electrical Engineering, Shahid Chamran University of Ahvaz, Ahvaz, Iran. His research interests include power system protection, power system stability, and power quality assessment.

Seyyed Ghodratollah Seifossadat was born in Ahvaz, Iran, on August 28, 1963 and received the B.Sc. degree in electrical engineering from the Iran University of Science and Technology (IUST), Tehran, in 1989 and the M.Sc. degree in electrical engineering from the Ferdosi University of Mashhad, Mashhad, Iran, in 1992 and the Ph.D. degree from IUST in 2006. Currently, he is a Professor with the Department of Electrical Engineering, Shahid Chamran University of Ahvaz, Iran, where he has been since 1992. His research interests are power system protection, fault location, power electronics, electric machinery, FACTS devices, and power quality assessment.
Improving Quality of Movement in a Linear Switched Reluctance Motor Using a Fuzzy Logic System

Allahverdi Azadrou1, Siamak Masoudi †, Reza Ghanizadeh3 and Payam Alemi4

Department of Electrical Engineering, Urmia Branch, Islamic Azad University, Urmia, Iran.

ABSTRACT

This work deals with minimizing fluctuations of propulsion force and improving the motion quality in a linear switched reluctance motor. In order to minimize the jerks in the moving part of the motor, a new profile has been used to generate an appropriate reference speed profile. The results indicate that at speed 0.5 m/s, the motor reaches its command speed at the proposed time while, using conventional speed profile it takes almost 1.4 times the desired time. In order to control the speed and increase the motion quality, a simple fuzzy logic system has been used which is able to overcome the uncertainties problem in nonlinear systems. The fuzzy control system can regulate the motor performance so that it tracks the reference speed with minimum error and fluctuation. To illustrate the performance of the fuzzy method, a conventional PI method along with a model reference adaptive control (MRAC) strategy have been applied to the motor and the obtained results for three control methods have been compared. Speed overshoot using conventional PI method is about 20 percent of the final speed while this is about 6 percent for fuzzy and MRAC methods. The system is designed and its efficiency is shown through simulation and experimental tests in different performance situations. The obtained results confirm that the fuzzy strategy outperforms other methods.

Article Info

Keywords: Force Ripple, Fuzzy Logic System, Linear Motor, Speed Control, Switched Reluctance Motor.

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I. INTRODUCTION

Switched reluctance motors (SRMs) have very interesting characteristics over electrical motors. They are very robust, low price, simple in structure, and fault tolerant with fewer thermal problems compared to other electrical motors. The linear type of these motors has all the above-mentioned benefits along with the ability of direct motion. They do not need any mechanical rotational to linear motion converters making them more suitable in special direct motion applications such as elevators, transport systems, etc. Different types of linear switched reluctance motors (LSRMs) have been presented by researchers previously which can be used in different industries [1]. These motors have inherently nonlinear dynamics with some uncertainties cause to be complicated their analysis and control.

Generally, these motors generate a force with high ripples and so, low quality which again complicates their application. Various speed regulation strategies have been suggested by researchers to improve the motor performance and to solve the torque (or force) fluctuation problem in SRMs. Generated force in LSRMs can be controlled by an appropriate force distribution function (FDF) [2]–[4]. In order to minimize the torque ripple in a rotational switched reluctance motor, an advanced torque sharing function has been proposed in [5] which has used combination of the genetic algorithm and fuzzy PSO methods. An appropriate FDF has been presented in [6] which could noticeably reduce the force fluctuations in a LSRM. We know that considerable force fluctuation can occur

† Siamak Masoudi: s.masoudi@tabrizu.ac.ir
Tel: +98-9141246770, Fax: +98-9141246770, Islamic Azad University. Department of Electrical Engineering, Urmia Branch, Islamic Azad University, Urmia, Iran.
In the phase commutation period. In recent papers, many efforts have been made to decrease the force (or torque) pulsations. In conventional control methods, it is possible to optimize the conduction angles in order to reach a flat force, but it has been shown that under higher loading conditions, noticeable force fluctuation is unavoidable [7]. Some interesting techniques for force (or torque) fluctuation reduction have been presented by researchers in recent years. In [8], dynamic equations of a SRM have been shown and the generated torque has been studied considering the contribution of the mutual inductances. Another current profile analysis has been done in [9] which cause to minimization of the torque pulsation over a wide range of speeds. Other current profile-based optimization methods have been suggested in [10]. In the methods, the current profile has been optimized to produce less ripple in the output current and torque. Destructive interference is the other method which has been used in [11] for improving the torque fluctuation in a SRM. Optimization of the SRM is another method to lower the force or torque pulsation. We have used a seeker optimization algorithm to optimize the LSRM construction in [12]. The obtained results indicated that the optimized structure generates better force with fewer ripples. In [13] and [14], new structures of SRMs have been proposed generating a torque with significantly low ripples. It is obvious that the fluctuated force will affect the speed of the translator; consequently, the actual speed would be different from the command value leading to disturbance in the LSRM performance. Accordingly, an appropriate speed controller can regulate the motor performance accurately. Fuzzy control is a practical alternative for a variety of challenging control applications as it provides a convenient method for constructing nonlinear controllers via heuristic information [15] – [18]. An interesting approach to the analysis of continuous time Takagi-Sugeno fuzzy systems using fuzzy Lyapunov functions has been presented in [16] which is independent of normalized fuzzy weighting functions. Some special systems such as nontriangular structural system have some problems complicating their analysis. In [17], an adaptive fuzzy switched control strategy has been investigated which solved the system problems. Another combined fuzzy control strategy has been proposed for a nonlinear system in [18]. Using Nussbaum function method, an adaptive output feedback control strategy has been proposed in [19] which is appropriate for nonlinear systems with output constraints. Fuzzy control strategy has the advantage of independence of the system parameters; thus, the method can be used for systems with unknown variables [20] and [21].

In this work, we have been used a fuzzy control strategy to smooth the output actual speed of the moving part of the LSRM. The controller has been implemented to the motor as a speed regulator. The proposed linear motor structure and its dynamic equations have been discussed in section 2. In section 3, block diagram of the control system along with the fuzzy control method have been investigated. Also, a new speed profile to reach a smooth motion quality has been introduced in section 3. In order to have a comparison between the proposed control methods, a new model reference adaptive control strategy has been demonstrated in section 4 and finally, simulation results and experimental results for three control strategies have been demonstrated and discussed in section 5 and 6.

II. PROPOSED SYSTEM

A. The linear motor and its characteristics

Control of a four-phase double-sided linear switched reluctance motor (DLSRM) is proposed in this work. The motor has a moving part with eight winding sets located on it along with two stators on both sides of it [6]. Because of two stators located on both sides of a translator, two same electromagnetic forces are applied to the common translator between them. So, this double sided construction yields high propulsion force density thus making the motor appropriate for applications requiring strong drivers. It is obvious that DLSRMs inherently have complete nonlinear characteristics complicating their performance control over electrical motors. Achieving an appropriate model for the DLSRM with a complete and accurate data of the motor inductance and force is very useful in the motor analysis. This can be done by a precise 3-D finite element analysis. The obtained data of the selected DLSRM are shown in Fig. 1.

---

**Fig. 1.** Selected DLSRM and data from FEM, (a) Selected DLSRM, (b) inductance, (c) force

The dimensions of the proposed DLSRM are shown in
B. Electromagnetic Equation

Voltage balancing equation for any one phase can be represented as:

\[ v_k = R_i + \frac{d\lambda_k}{dt} + \lambda_k = L_k(i_k, x), i_k \]  

\[ \frac{d\lambda_k}{dt} = L_k(i_k, x) + g_k \cdot v_x - i_k \]  

where, \( v_k \), \( i_k \), \( \lambda_k \), \( R \), and \( L_k \) are voltage, current, linkage flux, winding resistance, and inductance of the \( k \)-th phase, respectively where \( k \) denotes the phases a, b, c, and d. In addition, \( v_x = dx/dt \) and \( g_k(x) = dL_k/dx \) are linear motion velocity and the force function respectively, in which \( g_k(x) \) is defined as the rate of change of phase inductance according to position. Force of each phase \( F_k \) can be obtained from the phase current and translator position. The resultant propulsion force, \( F_e \) is the sum of the phase forces. SRM mathematical analysis has been carried out in [18]. The generated propulsion force in the LSRM can be written as:

\[ F_e = M \frac{dv}{dt} + Cv + F_L \]  

\[ F_e = F_a + F_b + F_c + F_d \]  

\[ F_L = k_pv \]  

where \( F_a \) and \( F_L \) show the propulsion and output load forces, respectively while \( F_k \) with \( k = a, b, c, \) and \( d \) denotes force of each phase. \( v, M \) and \( C \) are the speed of the moving part (m/s), moving part total mass and frictional coefficient, respectively. \( k_p \) is a constant factor.

Phase currents can be used to express an output force equation. To express the output force in terms of phase currents and translator position, the output force equation is derived from the coenergy \( W_e \). From the definition of the coenergy and assuming that two “\( x \)” and “\( y \)” phases are on simultaneously, then according to \( \lambda_x = L_x i_x \) and \( \lambda_y = L_y i_y \), the differential coenergy is expressed as:

\[ dW_e(i_x, i_y, x) = \lambda_x d_i_x + \lambda_y d_i_y + F_e dx \]

\[ = (L_x i_x) dx + (L_y i_y) dx + F_e dx \]  

(6)

The coenergy can be found by integrating (6) along a path of integration. The most convenient integration path is to integrate over \( x \) holding \( i_x \) and \( i_y \) fixed at zero, integrate over \( i_y \) by holding \( i_x \) fixed at zero, and finally integrate over \( i_x \). In the first part of the integration, the integral is zero because \( F_e \) is zero when both \( i_x \) and \( i_y \) are zero. Thus, the coenergy is calculated as:

\[ W_e(i_x, i_y, x) = \int_0^y \lambda_x(0, \epsilon, x) d\epsilon + \int_0^y \lambda_y(\epsilon, i_y, x) d\epsilon = \int_0^y (L_x i_x) d\epsilon + \int_0^y (L_y i_y) d\epsilon = \frac{1}{2} L_x i_x^2 + \frac{1}{2} L_y i_y^2 \]  

(7)

where \( \epsilon \) is an integration variable. Then, the propulsion force \( F_e \) is calculated as:

\[ F_e = \frac{\partial W_e(i_x, i_y, x)}{\partial x} |_{i_x, i_y, \epsilon} = \frac{1}{2} g_x i_x^2 + \frac{1}{2} g_y i_y^2 \]  

(8)

Now \( F_e \) is expressed in terms of variables, \( i_x \) and \( i_y \), as the rate of change of phase inductance, \( g_x \) and \( g_y \). After calculating the propulsion force, using (3) and (4), we can write:

\[ \dot{v} = -k_pv + \frac{1}{M}[Cv + F_e] \]  

(9)

III. FUZZY CONTROL SYSTEM DESIGN

The block diagram of the system composed of the LSRM and its control system is demonstrated in Fig. 2. In this study, the speed of the motor has been controlled by the fuzzy logic control approach. The fuzzy system has two inputs: the speed error \( e \) and change in the speed error \( \Delta e \). The output of the controller is the reference current of each phase.

![Fig. 2. Block diagram of the proposed control system](image)

A. Jerk Minimization-Based Speed Profile

At the first step and after defining the desired position of the translator, it is necessary to produce a speed profile. An appropriate speed profile can reduce the jerks of the motion noticeably, and so it can enhance the motion quality [22]. Assuming that \( x(t) \) indicates the position of the translator and \( t_0 \) is the final time of motion, we can write the objective function as follows which should be minimized.

<table>
<thead>
<tr>
<th>Table 1.</th>
<th>SELECTED LSRM DIMENSIONS</th>
</tr>
</thead>
<tbody>
<tr>
<td>Parameter</td>
<td>Value</td>
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<tr>
<td>Air gap length</td>
<td>2</td>
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<td>Stator pole width</td>
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<tr>
<td>Stator pole height</td>
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<tr>
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</tr>
<tr>
<td>Translator pole height</td>
<td>26</td>
</tr>
<tr>
<td>Translator height</td>
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</tr>
<tr>
<td>Turns per phase</td>
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<tr>
<td>Rated current</td>
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</tr>
<tr>
<td>Rated voltage</td>
<td>150</td>
</tr>
<tr>
<td>Friction coefficient (C)</td>
<td>10</td>
</tr>
<tr>
<td>Translator mass (M)</td>
<td>5</td>
</tr>
<tr>
<td>Phase resistance</td>
<td>2.5</td>
</tr>
</tbody>
</table>
\[ j = \int_0^t \left( \frac{d^3x(t)}{dt^3} \right)^2 \, dt \]  
(10)

If \( x(t) \) be the response of the Euler-Poisson problem, then the function \( j \) will have an extremum value. According to this, we can write the obtained equation as:

\[ \frac{d^6x(t)}{dt^6} = 0 \]  
(11)

This means the minimum jerk profile \( x(t) \) is a polynomial like (12).

\[ x(t) = \alpha_5 t^5 + \alpha_4 t^4 + \alpha_3 t^3 + \alpha_2 t^2 + \alpha_1 t + \alpha_0 \quad (12) \]

Consequently, the reference speed profile of the system is obtained as:

\[ v_{ref}(t) = 5\alpha_5 t^4 + 4\alpha_4 t^3 + 3\alpha_3 t^2 + 2\alpha_2 t + \alpha_1 \quad (13) \]

In order to solve the equation (13), five boundary conditions are required which are \( x(0) = 0, \, \dot{x}(0) = v_0 = 0, \, \ddot{x}(0) = 0, \, \dot{x}(t_d) = 0 \), and \( \ddot{x}(t_d) = 0 \). Subscript \( \cdot d \) indicates the parameters at the end of the motion period. Finally, the reference speed is obtained as:

\[
v_{ref}(t) = \left( -\frac{3a_0}{2t_d^2} - 3\frac{v_0}{t_d^3} - 3\frac{v_d}{t_d^4} + 6\frac{x_0}{t_d^5} + 6\frac{x_d}{t_d^6} \right) t^4 + \left( 3\frac{a_0}{2t_d^2} + 8\frac{v_0}{t_d^3} + 7\frac{v_d}{t_d^4} + 15\frac{x_0}{t_d^5} - 15\frac{x_d}{t_d^6} \right) t^3 + \left( -3\frac{a_0}{2t_d^2} - 6\frac{v_0}{t_d^3} - 4\frac{v_d}{t_d^4} - 10\frac{x_0}{t_d^5} + 10\frac{x_d}{t_d^6} \right) t^2 + a_0 t + v_0 \]
(14)

The proposed speed profile was applied to the system and the obtained results were compared with the conventional trapezoidal approach findings. At speed 0.5 m/s and using a conventional trapezoidal profile, speed of the motor reaches its command value at \( t=0.17 \) s while this time is less than \( t=0.12 \) s using the proposed speed profile. This can be seen at speed 2.5 m/s in Fig. 3 (b). It can be seen from Fig. 3 that the proposed method has a better performance in reference speed tracking.

**B. Fuzzy Control Strategy**

A zero order Takagi-Sugeno fuzzy system is selected to be applied to the DLSRM. The system is able to approximate the nonlinear characteristics of the motor. The proposed fuzzy system elements are shown in Fig. 2. The inference part consists of inference mechanism unit, database unit, and fuzzy rule base unit. The database unit includes all required data such as linguistic variables of input signals, membership function of each linguistic variable, and a set of fuzzy rules. The system consists of a mapping from an input vector \( z = [z_1, ..., z_m]^T \in \Omega_z \subset R^m \) to a scalar output variable \( y_f \in R \), where \( \Omega_z = \Omega_{z_1} \times ... \times \Omega_{z_m} \) and \( \Omega_{z_i} \subset R \). Assuming that each variable \( z_k \) has fuzzy sets \( G^k_1, j = 1, ..., M_k \), the TS fuzzy system can be introduced by a set of if-then rules such as the following:

\[ R^{(i)}: \text{If } z_1 = G^1_k \text{ and } z_2 = G^2_k, \text{ Then } y_f = y_f^i = q_0^i + q_1^iz_1 + q_2^iz_2 = \theta^T[1 \, z^T]^T \]

where \( G^k_1 \) indicate fuzzy sets, \( k=1, 2, \quad i = 1, ..., N_r \) and \( z^T = [z_1, z_2] \) is the input vector, \( \theta^T = [q_0^i \, q_1^i \, q_2^i]^T \) shows a vector of the adjustable parameters, \( y_f^i \) is the scalar output of the i-th rule, and \( N_r \) is the total number of fuzzy rules. By using the central average defuzzifier and product inference, the output of the TS fuzzy system can be expressed as:

\[ y_f(z) = \frac{\sum y_{f}^{i} \, \mu_{i} G_{1}^{k}(z)}{\sum \mu_{i} G_{1}^{k}(z)} \quad (16) \]

\[ \mu_{i} = \prod_{k=1}^{K} G_{1}^{k}(z) \quad (17) \]

where \( \mu_{G_{1}^{k}} \) denotes the membership function of the fuzzy set \( G_{1}^{k} \). We can rewrite the output of the fuzzy system expressed in (9) as follows:

\[ y_f(z) = w^T(z) \theta \]

where \( \theta = [y_{f}^{1}, ..., y_{f}^{N_r}]^T \) is a vector grouping all consequent parameters, and \( w(z) = [w_{1}(z), ..., w_{N_r}(z)]^T \) indicates a set of fuzzy basis functions defined as

\[ w_{i}(z) = \frac{\mu_{i}(z)}{\sum_{j=1}^{N_r} \mu_{j}}, \quad i = 1, ..., N_r \]

(19)

Defining enough fuzzy rules, the fuzzy logic system (19) can be used for approximating any real continuous function. The speed error and change of the speed error represent the inputs of the fuzzy system. Each input and output variable have fuzzy sets as shown in Fig. 4. According to Fig. 4 (a), seven triangular fuzzy sets were used to introduce each input variable with fuzzy sets having 50% overlap with each other. For the output variable, a fuzzy inference mechanism was calculated using a singleton membership function shown in Fig. 4 (b). In this work, we have selected singleton membership function for the output of the system. This cause to fast response of the controller.
The main advantage of the fuzzy control system is that it does not need lots of data to control the LSRM performance. Totally the used fuzzy controller has the following advantages over the other control strategies:

- Simplicity and flexibility
- Can handle problems with imprecise and incomplete data
- Can model nonlinear structure of the LSRM
- Cheaper and need very low memory in implementation
- Cover a wide range of operation conditions.

The fuzzy rules are listed in TABLE II.

### TABLE II

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<td>PB</td>
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### IV. Model Reference Adaptive Control (MRAC)

A new MRAC system has been proposed in [23] which can be used for speed control of the electrical motors. Implementation of this control strategy needs state equation of the system same as [23]:

$$\dot{x} = f(x) + G(x)u$$  \hspace{1cm} (20)

where $x \in \mathbb{R}^n$ are state variables and $u \in \mathbb{R}^m$ is the input of the system. This equation can be written as

$$\dot{x} = A(x)x + B(x)u$$  \hspace{1cm} (21)

where $f(x) = A(x)x$ and $G(x) = B(x)$. The speed is the time derivative of the position i.e:

$$\dot{v} = \dot{v}$$  \hspace{1cm} (22)

Using (6), we can write:

$$\dot{v} = -\frac{C+k_p}{M}v + \frac{1}{M}F_e$$  \hspace{1cm} (23)

The matrix form of (22) and (23) is written as follows:

$$\begin{bmatrix} \dot{x}_1 \\ \dot{x}_2 \end{bmatrix} = \begin{bmatrix} 0 & 1 \\ 0 & -\frac{C+k_p}{M} \end{bmatrix} \begin{bmatrix} x_1 \\ x_2 \end{bmatrix} + \begin{bmatrix} 0 \\ 1/M \end{bmatrix} [F_e]$$  \hspace{1cm} (24)

$$u(t) = F_e$$  \hspace{1cm} (25)

In the proposed system, position speed $x_1 = v$ and derivative of the speed $x_2 = \dot{v}$ are state variables while, according to (25), propulsion force $F_e$ is the output. In (10), $A(x)$ and $B(x)$ are the state-dependent coefficient matrices that represent the nonlinear system in the form (20) in a pseudo-linear form. In this work we examine the controllability of the LSRM system using the equation (3) discussed in the [23]. Using the state-dependent Riccati equation, all state variables should reach zero by minimizing the following integral quadratic cost function:

$$J = \frac{1}{2} \int_0^\infty (x^T Q(x) + u^T R(x) u) dt$$  \hspace{1cm} (26)

where $Q(x) \in \mathbb{R}^{n \times n}$ and $R(x) \in \mathbb{R}^{m \times m}$ are the state-dependent weight matrices. For any $x$, it is assumed that $Q(x) \geq 0$, $R(x) > 0$  \hspace{1cm} (27)

Briefly description of the method has been demonstrated in [23]. This control strategy was implemented to the LSRM system and simulation results along with experiments results were obtained.

### V. Simulation Results

We use Matlab/Simulink to simulate the DLSRM along with the proposed control strategy. After generating the reference speed by the presented speed profile, the reference speed was compared with the actual value. Then, the error was introduced into the proposed fuzzy speed controller. Four phase currents were the output of the speed controller, which entered the PI current controller after comparison with the corresponding actual values. The proportional and integral parameters of the applied PI controller extracted by a trial and error approach were set to 0.05 and 2.2, respectively. In order to compare the performance of the proposed fuzzy system, a conventional PI controller along with a MRAC strategy discussed in [23] were...
applied to the motor and the obtained results have been shown under all conditions. The proportional and integral gains of the PI speed controller were set to 1.3 and 3.2, respectively.

The simulation results including the speed, speed error, and total propulsion force for three control strategies are shown in Fig. 5. According to Fig. 5 (a), the profiles of the speed and position are same for all control methods, so only one curve is shown in the figure. In Fig. 5 (b), propulsion forces obtained from three control strategies are demonstrated. Generally, the figures are same and reach their final value in a same time. Pick value of force in MRAC and PI controllers are about 20 N.m while this is lower in the proposed fuzzy control method. Current of phase “a” is depicted in Fig. 5 (c). According to these figures, the current is high at starting while it reduces to a lower value in steady state. According to figures, proposed fuzzy control method has better performance than the conventional PI method while has the same performance as modern MRAC strategy discussed in [23].

The advantage of the fuzzy method is that it does not require accurate information about the nonlinear LSRM structure and is easier to implement. Comparison of the speed between three control methods obtained by simulation and experimental have been shown in Fig. 9. There are some effective methods for evaluating the performance of control strategies which can be useful in comparing different systems. A useful method discussed in [12] introduces the critical parameters of the results including the percentage of overshoot, rising time, settling time, the criteria encompassing integral of absolute error (IAE) along with integral of time absolute error (ITAE). The parameters for the simulation test have been obtained and reported in Table 3. In order to study the performance of the system under load variations, an 8 kg load was added to the translator when it was moved at a constant speed. Then, the above-mentioned parameters were measured again and written in Table 3. The data confirm that the proposed control strategy has outperformed the conventional PI method in both no load and full load conditions, while it is approximately same as MRAC method. Considering the simple structure and implementation of the proposed fuzzy control method, it seems to be more appropriate than two other control strategies.

![Fig. 5. Comparison of simulation results, (a) speed and position, (b) force (N), (c) phase current (A)](image-url)
TABLE III. Critical Parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>No load</th>
<th>Full load</th>
</tr>
</thead>
<tbody>
<tr>
<td>Overshoot (%)</td>
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<td>0.88</td>
</tr>
<tr>
<td>Rising time</td>
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<td>0.013</td>
</tr>
<tr>
<td>Settling time</td>
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<td>0.017</td>
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</tr>
<tr>
<td>ITAE</td>
<td>0.001</td>
<td>0.001</td>
</tr>
</tbody>
</table>

VI. EXPERIMENTAL RESULTS

In this study, a DLSRM with dimensions and parameters outlined in Table 1 was selected. Then the proposed control strategy along with a conventional PI controller was applied to regulate the motor performance. The motor and its experimental drive are shown in Fig. 6. Steel sheets with 0.5 mm thickness were used to construct the stator poles and yoke. For creating the translator, nonlaminated iron was used which coils of AWG#15 wire placed on translator poles. In order to implement the control system, an ARM Cortex-M4 microcontroller STM32f407 was used with 72 MHz processing frequency. Four Hall sensors sensed the actual phase currents and sent them to the current control unit. All signals were stored in a flash memory. Also, the propulsion force was calculated by the corresponding equations and finite element analysis information. A magnetic sensor strip with 10 μm resolution runs alongside the stator giving the position feedback signal. The speed information was also obtained through a capture channel of the timer inside the microcontroller.

The results measured in the experimental test for three control strategies are shown in Fig. 7. The test conduct under the same conditions as the simulation in Fig. 5. According Fig. 7, force in the steady state condition is 12 N. At start and to overcome the moment of inertia, the generated force is higher. The pick start force is 17 N, 20 N, and 20.5 N for proposed fuzzy method, MRAC, and PI method, respectively. This also happens during braking.

![Fig. 6. Experimental setup](image)

![Fig. 7. Force comparison of three methods](image)
Fig. 9. The speed error in proposed fuzzy control strategy and MRAC method has same overshoot about 0.03 m/s while this exceeds 0.1 m/s in the conventional PI controller. So, the proposed fuzzy control method and MRAC strategy have almost the same performance but fuzzy control method has some benefits explained in section 3.

Fig. 9. Speed error comparison, (a) simulation, (b) experimental

REFERENCES


Allahverdi Azadrou was born in Salmas, Iran, in 1983. He received his B.S. degree in electrical engineering from Islamic Azad University, Abhar Branch, Abhar, Iran, and M.S. degree from Dezful Branch, Dezful, Iran. He is a member of science in Islamic Azad University, Salmas Branch, Iran since 2012. Currently he is Ph.D student in Islamic Azad University, Urmia Branch, Urmia, Iran.

Siamak Masoudi was born in Miyaneh, Iran. He received his B.S. degree in power engineering from Tarbiat Moallem Tabriz University, Tabriz, Iran, in 2006, and his M.S. and Ph.D. degrees in Electrical Machines and Drives from University of Tabriz, Tabriz, Iran, in 2009 and 2016, respectively. From 2010, he is a member of science in Islamic Azad University, Abhar Branch, Zarjan, Iran. His current research interests include Electrical machines design and control and linear motors/generators.

Reza Ghanizadeh received the B.S. degree in electrical engineering from the Ardabil Branch-Islamic Azad University (Ardabil), Iran, in 2009. He received the M.S. and the Ph.D. degrees from University of Birjand (Birjand), Iran, in 2013 and 2017, respectively. He is currently an assistance professor in faculty of electrical
engineering at Urmia Branch, Islamic Azad University (Urmia), Iran. His research interests include power quality, microgrids control, power electronics, high voltage studies, power systems stability and FACT devices.

Payam Alemi was born in Tabriz, Iran, in 1982. He received his B.S. degree from the University of Tabriz, Tabriz, Iran, in 2005, and his M.S. degree from the Science and Research Branch, Islamic Azad University, Tehran, Iran, in 2008, and Ph.D. degree in Electrical Engineering, from Yeungnam University, Gyeongsan, Korea, in 2014. Then he joined Simon Fraser University, BC, Canada for his Post-doctoral program until 2016. Currently he is an assistant professor in the department of electrical engineering, Islamic Azad University, Urmia Branch, Urmia, Iran. His current research interests include the control of multilevel power converters, power loss analysis for converters, LCL filters and machine drives, DC-DC converters.
Solving Economic Load Dispatch by a New Hybrid Optimization Method
Hossein Sharifzadeh

1 Department of Electrical and Computer Engineering, Hakim Sabzevari University, Sabzevar, Iran.

ABSTRACT
This paper presents a new solution method to efficiently handle non-convexity stemmed from valve points in the economic load dispatch problem. The proposed solution technique integrates both the advantage of fast solution algorithms of linear programming and powerful solution techniques of nonlinear programming to find the global solution. In the first step of the proposed solution framework, non-convex terms are replaced by some linear segments and the new linear model solved by modern fasts algorithms. In the second step, a nonlinear programming algorithm as a powerful local search algorithm solves the original non-convex model to improve the solution obtained in the previous step. By exploiting the main strength of linear and nonlinear programming algorithms, the proposed solution approach can quickly converge to nearly the global solution method. By experimental results on three test cases with different sizes, we show that the presented method outperforms the other algorithms published in the literature in the quality of the solution.

I. INTRODUCTION
Among different operational tools provided to efficiently operate power systems in dispatch centers, economic load dispatch (ELD) has special importance as it determines final generation levels of generators. Economic dispatch optimally assigns output power of generators while satisfies power system demand. To this end, an optimization problem with a cost function, as the objective function, and a set of practical constraints is defined [1]. The cost function shows the operational or just fuel cost of power plants and usually models as a quadratic term.

Nonetheless, it is shown that the cost function can better be represented by taking a rectified sine into account as it more precisely models the effect of opening several valves, in a power plant, on cost function [2]. However, from an optimization viewpoint, the considered rectified sinusoidal term changes the ELD problem into a non-smooth non-convex optimization problem. A wide range of solution algorithms has been published in the literature to tackle the optimization problem and can be categorized into two gradient and artificial intelligence (AI) methods. The gradient-based methods can include Lagrangian relaxation [3], modified lambda-iteration [4], linear programming [5], and quadratic programming [6]. The traditional gradient-based methods rely on derivatives of objective and constraints in optimization problems and as a result, they can not effectively manage the non-differentiable absolute function induced by rectified sinusoidal term. Moreover, they are local optimizer in nature and their performances dramatically depend on the starting point. Therefore, they have poor performances in the no-convex problem with multiple local minimums [7].

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Applications of AI algorithms have demonstrated in some studies [8]-[10]. AI based algorithms particularly in the ELD subject may consist of bat algorithm [11], gravitational search algorithm [12], symbiotic organisms search algorithm [13], firefly algorithm [14], cuckoo search algorithm [15], cooperative search algorithm [16], phasor particle swarm optimization [17], hybrid optimization framework [18], hybrid PSO–SQP [19], and hybrid GA–PS–SQP [20]. The algorithms do not require a differentiable search space. Moreover, they work with a set of solutions, so-called population, swarm, etc., compared to the gradient-based methods use a single solution. Hence, they may find the global solution using the parallel search. However, the computation burdens of these algorithms are high. Moreover, the performances of these methods severely depend on algorithm parameters limiting their application in practice.

In this paper, a new hybrid solution method is employed to tackle with the non-convexity of economic load dispatch accommodating the valve point effects. In the first step, the proposed method decomposes the non-convex problem to the set of linear problem method. The best solution among these linear problem methods can be found by a strong branch and bound algorithm. In the second step of the presented approach, the original problem is solved using a nonlinear programming (NLP) algorithm considering non-convex terms using the starting point, obtained in the first step, most likely located near the global solution depending on the quality of the linear segments approximation. Thus, NLP may converge to the global solution by providing an effective starting point.

The rest of the paper is organized as follows. The economic load dispatch problem formulation accounting for valve loading points is modeled in section II. Section III presents the proposed solution framework for solving non-convex ELD. Experimental results to show the effectiveness of the proposed solution technique is provided in section IV. Section V concludes the paper.

II. PROBLEM STATEMENT

The economic load dispatch tool aims to minimize total operation cost considering the physical and operational limits of power systems. From a formulation point of view, the aims and the limits can be represented by an objective function and set of equality and inequality constraints in an optimization model as follows [2]:

\[
\begin{align*}
& \text{Minimize} \sum_{j} \left( a_j P_j^2 + b_j P_j + c_j + e_j \sin \left( f_j \left( P_j^{\text{min}} - P_j \right) \right) \right) \\
& \text{subject to:} \\
& P_j^{\text{min}} \leq P_j \leq P_j^{\text{max}}, \quad j = 1, 2, ..., n \\
& \sum_{j=1}^{n} P_j = D
\end{align*}
\]

In the formulation, \( j \) shows an index of generators from 1 to \( n \) (number of generators). \( a_j, b_j, c_j, e_j, f_j \) are the coefficient cost of generator \( j \). The decision variable \( P_j \) stands for the generation level of generator \( j \). The minimum and maximum power limits of generator \( j \) are represented by \( P_j^{\text{min}} \) and \( P_j^{\text{max}} \) respectively. The power system demand is shown by \( D \).

The objective function in (1) indicating the total cost of generators includes two convex and non-convex components. The quadratic term \( a_j P_j^2 \) is a convex function while the term \( e_j \sin \left( f_j \left( P_j^{\text{min}} - P_j \right) \right) \) is a non-convex and non-smooth function added due to valve loading effects. The non-nonconvex term changes the ELD problem to non-convex one with many local optimal solutions.

In the next section, a two layers method is proposed to cope with the non-convex and non-smooth space of the presented ELD problem.

III. THE PROPOSED ELD SOLUTION METHOD

As mentioned before, the ELD accommodating valve point effects is a non-convex problem. Here, to tackle with the non-convexity, the non-convex cost functions replaced with some of the linear segments finally producing a mixed integer linear problem. Subsequently, a mixed integer programming (MIP) solution technique is used to solve the new model. As the mixed inter algorithms are very powerful at present, the new model can be solved very fast. The obtained solution of mixed inter then used as the starting solution of the original non-convex ELD problem solved by a nonlinear programming (NLP) approach. The NLP can remove the approximation error and converge to the nearly global solution as it currently uses the high-quality starting point obtained in the first step. The mixed integer problems usually solved by powerful branch and bound (B&B) algorithms [21]. In the B&B algorithm, firstly, all integer values relaxed and they can adopt fractional values. The resulting relaxed LP model is solved. In the B&B algorithm, a so-called tree is generated composed of a root (the first LP model) with some branches, nodes, and leaves as shown in figure 1. Nodes correspond to some integer variables that have fractional values in the solution of the LP model. In each node, a decision is made about rounding up or down the fractional values of the integer variable shown by corresponding branches. Subsequently, two new nodes in two branches ending of the node show the new models with fixing the fractional to integer values. The nodes that have been not branched yet called leaves. The two new models may be solved in two new nodes and other new nodes can be generated. The procedure continues until some stopping criteria satisfied. Consequently, the solution space of the original non-convex model is approximated by some probably
smaller linear models with a more efficient solution algorithm.

A mixed integer linear model of a nonlinear function can be found as described below.

Consider that the $B+1$ breakpoints or equivalently $B$ segments are used to linearly approximate a nonlinear function and $x^\text{min} = x_0 < x_1 < \ldots < x_B = x^\text{max}$ stand for the breakpoints.

Let us show the non-convex function with its $k$th linear segment approximation as follows:

$$y_k = m_k \hat{x}_k + d_k , \quad x_{k-1} \leq \hat{x}_k \leq x_k$$  \hspace{1cm} (4)

Where $x_{k-1}$ and $x_k$ show the beginning and ending points of the segments respectively. Here, in the ELD problem, $\hat{x}_k$ and $y_k$ can be interpreted as the generation level and associated variable with its segmented variable $\hat{x}_k$. Based on (7) and (8) only one continuous variable $\hat{x}_k$ can be non-zero that is determined based on the corresponding non-zero binary variable $z_k$.

Relying on the technique, the MIP model for ELD problem can be described as follows:

Minimize $\sum_{j=1}^{m} F_j$  \hspace{1cm} (10)

$$P_j^{\text{min}} \leq P_j \leq P_j^{\text{max}}, \quad j = 1, 2, \ldots, n$$  \hspace{1cm} (11)

$$\sum_{j=1}^{m} P_j = D$$  \hspace{1cm} (12)

$$F_j = \sum_{k=1}^{B} (m_{j,k} \hat{p}_{j,k} + d_{j,k}) , \quad j = 1, 2, \ldots, n$$  \hspace{1cm} (13)

$$P_j = \sum_{k=1}^{B} \hat{p}_{j,k} , \quad j = 1, 2, \ldots, n$$  \hspace{1cm} (14)

$$p_{j,k}^* \leq \hat{p}_{j,k} \leq p_{j,k}^d, \quad k = 1, 2, \ldots, B, \text{ and } j = 1, 2, \ldots, n$$  \hspace{1cm} (15)

$$\sum_{k=1}^{B} z_{j,k} = 1, \quad j = 1, 2, \ldots, n$$  \hspace{1cm} (16)

$$z_{j,k} \in \{0,1\}, \quad \forall k = 1, 2, \ldots, B, \text{ and } j = 1, 2, \ldots, n$$  \hspace{1cm} (17)

Where $\hat{p}_{j,k}$ is the generation level of generator $j$ in segment $k$ and limited to up and down breakpoints of the segment i.e. $p_{j,k}^*$ and $p_{j,k}^d$ respectively. The linear cost of generator $j$ is shown by $F_j$. Other symbol denotations are straightforward as they are a formulation extension of single function approximation in (5)-(9) to multi-function cost approximation of $n$ generators in ELD problem.

The solution obtained from the MIP model of ELD (10)-(17) can be used as the starting point of nonlinear ELD model (1)-(3) to converge to the nearly global solution.

**IV. NUMERICAL RESULTS**

This section deals with performance evaluation of the proposed solution technique for solving non-convex non-smooth economic load dispatch. To show the efficiency of the presented method three case studies with different size are selected as follows:

**Case I:** A small scale 3 unit test system.

**Case II:** A medium scale 13 unit test system.

**Case III:** A large scale 40 unit test system.

The proposed solution technique is implemented using
GAMS software. We used a laptop with CPU Core i3, 2.4GHZ clock frequency and 4 GB RAM in all simulations. The mixed integer model is solved by CPLEX and the nonlinear model using CONOPT. It is noted that the optimality of obtained results is not sensitive to the selected solver.

Application results of the proposed hybrid approach on these case studies are reported in the following. Moreover, the performance of the presented technique is compared with other solution algorithms in the literature to evaluate the capabilities of the proposed method more precisely.

**Case I: 3 unit case study**

The load demand of the three units test case is 850MW. All of its units have valve loading effects in their cost function. The required data for the test case can be found in [23].

For Case I, result of the presented algorithm and other ELD solution techniques, which are adopted from their published paper, are provided in Table I. As the table shows, different algorithms nearly have the same optimal solutions due to relatively a simple ELD problem in this small test case.

Nonetheless, most of the solution techniques present a noticeable fluctuation in results among different algorithms runs shown by best, average and worst costs. However, the proposed solution method always converges to a unique solution showing its robustness in solving the nonconvex ELD problem. Note that the optimal cost obtained by the proposed method is lower than the mean cost and worst costs of all the reported algorithms in Table I illustrating its merit in solving the ELD problem in practice. Table II demonstrates the generation levels of units of case I for the optimal solution of the proposed method.

The outcome of MIP, i.e. the approximate objective function as (9), for the case I is 8233.64$. The solution of MIP is employed as the initial point of the nonlinear programming solver to solve the original non-convex model, here CONOPT solver. The final solution achieved by the CONOPT is 8234.07$ showing its robustness in solving the nonconvex ELD problem. Table III shows the outcome pertaining to the application of the proposed hybrid solution method and results taken from other published approaches in the literature to solve the economic load dispatch problem for case II. The presented method outperforms most of the other methods in finding the better optimal solution, as the table III shows. Again, while the solution of the proposed piecewise technique is unique, dissimilar final solutions of other techniques challenge their applicability in real-word problems. Table IV illustrates the unit dispatch as the solution of the suggested method in solving ELD in test system II. For the case II, the output of the MIP solver is 17959.455$ finally reaches to 17960.366$ by the nonlinear solver.

**TABLE I: Comparison Of Solution Algorithms For Solving ELD In Case I**

<table>
<thead>
<tr>
<th>Methods</th>
<th>Best cost ($)</th>
<th>Mean cost ($)</th>
<th>Worst cost ($)</th>
</tr>
</thead>
<tbody>
<tr>
<td>MFEP[23]</td>
<td>8234.08</td>
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<td>8234.24</td>
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<td>IFEP[23]</td>
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<td>FA[24]</td>
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<td>PS*[25]</td>
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<td>8352.41</td>
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<td>Proposed hybrid algorithm</td>
<td>8234.07</td>
<td>8234.07</td>
<td>8234.07</td>
</tr>
</tbody>
</table>

*Violation of demand constraint

**TABLE II: Generation Levels, And Associated Cost In Optimal Solution Of Case I**

<table>
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<tr>
<th>Unit</th>
<th>Generation (MW)</th>
<th>Cost ($)</th>
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<tr>
<td>2</td>
<td>149.73310</td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>400.0000</td>
<td></td>
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</tbody>
</table>

**Case II: 13 unit case study**

The generator characteristics of thirteen test system are adopted from [23]. The demand for this test case is D=1800MW. The cost functions of all 13 units test case have valve loading effects. Table III shows the outcome of the MIP, i.e. the approximate objective function as (9), for the case II is 17960.366$. The solution of the proposed method outperforms most of the other methods in finding the better optimal solution, as the table III shows.

**TABLE III: Generation Levels, And Associated Cost In Optimal Solution Of Case II**

<table>
<thead>
<tr>
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<tr>
<td>2</td>
<td>149.6</td>
<td></td>
</tr>
<tr>
<td>3</td>
<td>222.749</td>
<td></td>
</tr>
<tr>
<td>4</td>
<td>109.867</td>
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</tr>
<tr>
<td>5</td>
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<td>6</td>
<td>60</td>
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<tr>
<td>7</td>
<td>109.867</td>
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</tr>
</tbody>
</table>

**TABLE IV: Comparison Of Solution Algorithms For Solving ELD In Case I**

<table>
<thead>
<tr>
<th>Methods</th>
<th>Best cost ($)</th>
<th>Mean cost ($)</th>
<th>Worst cost ($)</th>
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<tr>
<td>CEP[23]</td>
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<td>CSOMA[26]</td>
<td>17960.36.6</td>
<td>17967.87</td>
<td>17970.83</td>
</tr>
</tbody>
</table>

*Proposed hybrid algorithm | 17960.36.6 | 17960.366 | 17960.366 |
Case III: 40 unit case study

The forty unit test case demand is 10500MW. The valve loading effects appear in the cost function of all of its units. The data for the test case are taken from [23].

Table V compares the obtained results of the proposed solution technique and reported outcomes of other methods in the literature in solution quality of case III. The first point that should be highlighted is the considerable difference between reported solutions due to the large scale nonconvex ELD problem in this test case. For the same reason, more solution oscillations can be seen in multiple runs of the reported algorithms. Nevertheless, the proposed method converges to the best solution and without any change showing its ability to solve the large scale nonconvex ELD problem. Optimal dispatch for test case III found by the proposed algorithm is illustrated in Table VI. While the returned cost by the MIP solver in the case III is 121413.57$, the final optimal solution is 121412.53$ given by the CONOPT nonlinear solver.

<table>
<thead>
<tr>
<th>Methods</th>
<th>Best cost ($)</th>
<th>Mean cost ($)</th>
<th>Worst cost ($)</th>
</tr>
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<td>FEP [23]</td>
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<td>Θ-MBA [29]</td>
<td>121412.53</td>
<td>121412.7</td>
<td>121412.95</td>
</tr>
<tr>
<td>Proposed hybrid algorithm</td>
<td>121412.53</td>
<td>121412.5</td>
<td>121412.5</td>
</tr>
</tbody>
</table>

Although the presented hybrid technique establishes a nearly global solution and virtually outperforms the other ELD solution methods published in the literature, its computational burden also should be evaluated for the practical application. Because of the different software and hardware used in this paper compared with other techniques presented in the literature, comparisons of computation times are fairly invalid. However, to illuminate the application of the hybrid method, the elapsed time of the suggested technique for each experiment, for all the three test cases, are shown in Table VII. As can be seen from Table VII, the computation times of the presented hybrid algorithm in all the three experiments are less than 2 seconds showing the fast convergence of the method and its application for practical test cases.

V. CONCLUSIONS

To handle the nonconvexity appears in the ELD problem with valve point effects, we propose a hybrid solution method of powerful local and global searching ability. We describe how the solution method exploits both the branch and bound algorithm as the global search technique and NLP solvers as the local search algorithm to converge to unique high-quality solutions. Comparison results of solution algorithms of the nonconvex ELD problem show the considerable advantage of the presented hybrid technique concerning the other algorithms published in the literature both in the solution quality and robustness of the solution in the multiple
algorithms runs. It is noticed that the performance of the AI-based algorithms highly depends on their parameters. However, the proposed technique outperforms the other methods without a trial and error parameter tuning mechanism. Another crucial factor the presented solution method is the low computational burden allowing to be used in practical applications.

REFERENCES


Hossein Sharifzadeh is an assistant professor of electrical engineering at Hakim Sabzevari University (HSU). He mainly works on applications of modern optimization techniques and uncertainty handling on power systems. His current research interest is application of global optimization techniques to power system problems.
Novel Stability Criteria for Piecewise Affine Systems with Time-Varying Delay

Salehe Afsharian\textsuperscript{1}, Hussein Eliasi\textsuperscript{2,†}

\textsuperscript{1,2} Faculty of Electrical and Computer Engineering, University of Birjand, Birjand, Iran

Abstract

This article aims to derive new sufficient conditions to guarantee the stability of piecewise affine systems with time-varying delay (PWA-TVD). The set of delay-dependent linear matrix inequality (LMI) describes the novel stability criteria. This approach considers the PWA-TVD system with a time-delayed state-dependent switching signal. The newly suggested Lyapunov-Krasovskii functional (L-K-F) and improved estimation of its derivative have a crucial role in decreasing the complexity and conservativeness of the proposed stability results. The suggested L-K-F belongs to the current and time-delayed states, the integral of the states over the time-varying delay, and time derivation of the states. A new inequality was used to obtain an upper bound (UB) for the time derivation of the Lyapunov functional. Then based on this UB, less conservative results are achieved. The theoretical results are applied to the numerical examples. The results confirm the effectiveness of the presented method. The conservative index is the maximum admissible UB of time delay.

Keywords:
Linear matrix inequality (LMI), Lyapunov-Krasovskii functional, PWA systems, Stability analysis, Time-varying delay

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I. Introduction

A piecewise affine (PWA) system is a dynamical system that contains multiple affine subsystems and a switching signal that allows switching between various subsystems. These systems have broad applications in a range of science and engineering. This type of system constructs a robust structure for to model a wide variety of hybrid and nonlinear systems in that nonlinearities are estimated by a set of linear-affine models around different operating points.\cite{1}.

A time delay is inevitable in the real world. It exists in industrial processes, control systems, energy systems, and so on. The time-delay may cause unsatisfactory efficiency, undesired oscillation, and also system instability. So, it is imperative to analyze the stability criterion of systems with time-delay \cite{2}.

The PWA systems with time-delay have a key role in the modeling of such systems as communication networks, highway transportation systems, automotive clutch systems, etc. Delay is an outstanding design and performance characteristic of these systems and cannot be ignored.

To study the stability of switching systems and PWA systems, the primary approach is to obtain a common Lyapunov function for the system under arbitrary switching \cite{3}. For less conservative results, multiple Lyapunov functions have been proposed \cite{4} - \cite{7}. Lyapunov-Krasovskii functional and Lyapunov-Razumikhin functional techniques are used for time-delay systems to obtain stability criteria \cite{8}. A critical goal in the delay-dependent stability analyses is to improve conditions in which one guarantees the stability for the allowable UB of time delay as large as possible. The conservativeness of the results depends on two points. The first one is the choice of proper Lyapunov functional. In order to improve the results, an augmented vector has been used to make a Lyapunov functional. Augmented state variables contain information about current and delayed states, their integral and their derivative. The stability of time-delay systems depends on the current states and their history, so the...
use more states information in Lyapunov functional can be a less conservative way for stability analysis [9-12]. In all of these work, the novel augmented L-K-F are proposed, and according to this selection, the results will be less conservative. The other one is to estimate the bound of integral that appears in the time derivation of the Lyapunov functional. Several remarkable approaches have been reported to estimate a tighter bound of integral terms such as Wirtinger’s inequality [13], free-weighting matrix [14], Jensen’s inequality [2], double integral inequality [15], delay-dependent-matrix based (DDMB) reciprocally convex inequality [16], and etc.

The first time the PWA with time-delay systems to stability analysis has been discussed in [17]. The sufficient conditions that guarantee the stability of the PWA-TVD system have been investigated in [18-19]. In which, switching law is based on only the current states, and the results are more conservative. In [20-21], the state-dependent switching condition depends on the delayed states that are considered, but the delay is constant. In [22], stability conditions are derived for PWA-TVD systems with the switching law, which is dependent on the states dependent delay and the delayed-states. Then a switching signal that is independent of the time-delayed states for the Lyapunov function is designed. In all of these works, LMIs depend only on the bound of delay and bound of its derivative, and therefore, the results are more conservative.

In the present work, the PWA-TVD systems with switching law, which is based on both the states and delayed states, are considered. We aim at improving the existing results by proposing a proper L-K-F and developing a new inequality used to derive a more accurate estimation of the lower bound of the integral, which appears in the time derivation of the L-K-F. The proposed L-K-F contains more information about the states. This Lyapunov functional depends on the current and delayed states, the integral of the states across the time-delay, and the time derivative of the states. Integral terms with time-varying delay intervals are considered in the suggested Lyapunov functional, and new stability conditions are presented, which depend on time-varying delay, and its derivative does not depend only on the bound of these. Also, we use new inequality to find the bound for the derivative of the Lyapunov functional, and we then propose stability criteria, which enhance the feasible region of stability, for the TVD-PWA system. The new stability results for the TVD-PWA system are formulated in terms of LMIs. These LMIs depend on time-varying delay and derivative of delay.

The paper is organized as below. The preliminaries and problem formulation are defined in in Section II. The main stability theorem is proposed in Section III. The paper is terminated with a conclusion paragraph in Section V.

Notation: $\mathbb{R}^{r}$ is the set of all qxr real matrices. $S^r$ denotes a set of symmetric rxr matrices. In addition Q > 0, for $Q \in \mathbb{R}^{n \times n}$, means that Q is positive definite. We describe Sym(X) = X + X^T, for any square matrix X $\in \mathbb{R}^{n \times n}$. The notation $I_r$ represents the appropriate unit matrix. The symbol * represents the symmetric structure. K^T denotes the matrix basis for the null space of $K \in \mathbb{R}^{n \times m}$.

II. Problem Formulation

The following formulation represents the PWA-TVD system:

$$S_1: \begin{cases} \dot{x} = A_1 x(t) + \sum_{i} A_i^d x_d(t-a_i); \ x(t) = \sigma(t), \ t \notin [-h_d,0] \end{cases}$$

where $x(t) \in \mathbb{R}^{n+1}$ is the system current states, $x_d(t) = x(t-d_0(t))$ is the delayed-states, $A_i, A_i^d \in \mathbb{R}^{n \times n}$ are the current and time-delayed state matrices respectively, $a_i \in \mathbb{R}$ is the affine terms, and $\sigma(t)$ is an initial condition. The state space is partitioned to a number of polyhedral cell that are shown with $\chi_i = (a_i,b_i) \in \mathbb{R}^2$, $E_i [E_i^d]$ are the cell boundary matrices. $l=\{1, \ldots , \Xi \}$ shows the index of the subsystems. $d_0(t)$ is a time-varying delay, which value and its rate are bounded:

$$0 \leq d_0(t) \leq d_0, -\mu \leq d_0'(t) \leq \mu$$

The state space can be partitioned as below:

$$[E_i^d] [\bar{x}(t)] \succeq 0; \ \forall [x_d; x_d^T] E_i \chi_i , i \in l$$

The continuity matrices $F_i = [F_i F_i^d]$ satisfy:

$$[F_i F_i^d] [\bar{x}(t) \ x_d(t)] = [F_j F_j^d] [\bar{x}(t) \ x_d(t)] \ \forall [x_d; x_d^T] E_i \chi_i , i, j \in l$$

The following in [23] can construct these matrices.

Define the following augmented matrices: $\bar{A}_i = \begin{bmatrix} A_i & a_i \\ 0 & 0 \end{bmatrix}$.

$$\bar{A}^d_i = \begin{bmatrix} A_i^d & 0 \\ 0 & 0 \end{bmatrix}$$

The subsystems that contain the origin is expressed by $I_0$, and other subsystems are denoted by $I_1$.

III. Main Result

In this section, the set of delay-dependent LMI-based sufficient conditions is extracted to analyze the stability of the PWA-TVD. We introduce the following lemmas that will be used in the extraction of the stability criteria. Then, the main theorem is presented.
**Lemma 1** [24]: If $y$ is a differentiable function: $[\theta_1, \theta_2] \rightarrow \mathbb{R}^n$, for any matrices $M_1 \in \mathbb{R}^{3 \times 3n}, V_1, V_2 \in \mathbb{R}^{3 \times n}$, and symmetric matrices $M_1, M_2 \in \mathbb{R}^{3 \times 3n}$, and $Q \in \mathbb{R}^{n \times n}$, and satisfying:

$$
\begin{bmatrix}
M_1 & M_2 & V_1 \\
* & M_3 & V_2 \\
* & * & Q
\end{bmatrix} \geq 0
$$

(6)

the following inequality holds:

$$
- \int_{\theta_1}^{\theta_2} y^T(s)Qy(s)ds \leq \omega_1(\theta_1, \theta_2)\psi_1(\theta_1, \theta_2)
$$

(7)

Where

$$
\omega_1(\theta_1, \theta_2) = \left[\begin{array}{cc}
y^T(\theta_1) & y^T(\theta_2)
\end{array}\right] \left[\begin{array}{cc}
M_1 + \frac{1}{3}M_2
\end{array}\right] + \text{Sym}(V_1[-I_4 \ 0] + V_2[-I_4 \ -I_4])
$$

$$
\psi_1(\theta_1, \theta_2) = \left(\left[\begin{array}{cc}
y^T(\theta_2) & y^T(\theta_1)
\end{array}\right] \right)^T
$$

Theorem 1: (Finsler's Lemma) Let $W \in \mathbb{R}^n$, $S \in \mathbb{S}^n$, and $K \in \mathbb{R}^{m \times n}$ so that $\text{Rank}(K) < n$. The statements (1) and (2) are equivalent to:

(1) $W^T S W < 0, \forall KW = 0, x \neq 0$

(2) $K^T S K < 0$

Finsler's lemma can be used to give novel linear matrix inequality characterizations to stability and control problems.

We now present novel stability criteria for the PWA-TVD system, which conditions are dependent of time-delay.

**For $V_i (i \in \mathcal{I}_0)$**

$$
\begin{bmatrix}
Z_1 & Y_1 & N_1 \\
* & Z_2 & N_2 \\
* & * & R
\end{bmatrix} \geq 0, \quad
\begin{bmatrix}
Z_3 & Y_2 & N_3 \\
* & Z_4 & N_4 \\
* & * & R
\end{bmatrix} \geq 0
$$

(8)

$$
Z_{1,2,3,4} \in \mathbb{R}^{3(n+1) \times (n+1)}, \quad N_{1,2,3,4} \in \mathbb{R}^{3n \times 3n}, \quad Y_i \in \mathbb{R}^{3n \times 3n}
$$

$$
\begin{bmatrix}
E_{i}^T H_i E_i & E_{i}^T H_i E_i \\
E_{i}^T H_i E_i & E_{i}^T H_i E_i
\end{bmatrix} \geq 0
$$

(9)

For $V_i (i \in \mathcal{I}_1)$

$$
\begin{bmatrix}
Z_1 & Y_1 & N_1 \\
* & Z_2 & N_2 \\
* & * & R
\end{bmatrix} \geq 0, \quad
\begin{bmatrix}
Z_3 & Y_2 & N_3 \\
* & Z_4 & N_4 \\
* & * & R
\end{bmatrix} \geq 0
$$

(10)

$$
\begin{bmatrix}
B_i^T & \Psi_i(B_i^T)
\end{bmatrix} \leq 0
$$

where

$$
B_i = [A_i \ A_i^d \ 0 \ -I_d \ 0 \ 0 \ 0 \ 0 \ 0], \quad h_i(-1) = -d_i(t),
$$

$$\psi_i = \text{Sym}(E_i P_i E_i^T) + \text{Sym}(E_i S_i E_i^T) + \Omega_1 + \Omega_2 + h_i E_i R_i E_i^T + \Phi_1 + \Phi_2
$$

$$\Omega_1 = E_i Q_i E_i^T - h_i E_i Q_i E_i^T + \text{Sym}(E_i Q_i E_i^T)
$$

$$\Omega_2 = h_i E_i Z_i E_i^T - h_i E_i Z_i E_i^T + \text{Sym}(E_i Z_i E_i^T)
$$

(11)

$$
\begin{bmatrix}
E_{i}^T H_i E_i & E_{i}^T H_i E_i \\
E_{i}^T H_i E_i & E_{i}^T H_i E_i
\end{bmatrix} > 0
$$

(12)

$$
\begin{bmatrix}
B_i^T & \Psi_i(B_i^T)
\end{bmatrix} \leq 0
$$

(13)
\(\tilde{\omega}_1 = E_2 Q E_1^T - h E_2 Q E_2^T + \text{Sym}\left\{E_2 Q E_3^T\right\}\)
\(\tilde{\omega}_2 = h E_4 Z E_4^T + E_1 t Z E_1^T + \text{Sym}\left\{E_1 Z E_2^T\right\}\)
\(E_1 = [\bar{b}_1, \bar{b}_2], E_2 = [\bar{b}_4, h \bar{b}_5], E_3 = [\bar{b}_3, d_0(t) b_7 (h_d - d_0(t)) \bar{b}_3], E_4 = [\bar{b}_6, \bar{b}_1, \bar{b}_2, h \bar{b}_2 - b_3], E_5 = [\bar{b}_5, \bar{b}_6, \bar{b}_3, \bar{b}_4, \bar{b}_5, \bar{b}_6, \bar{b}_2 - b_3], E_7 = [d_0(t) b_7, \bar{b}_1, \bar{b}_2, d_0(t) \bar{b}_1 - b_7].\)
\(E_8 = [\bar{b}_0, b_0, \bar{b}_4, \bar{b}_3, \bar{b}_2, b_5, \bar{b}_6, E_10 = [\bar{b}_3, \bar{b}_6, \bar{b}_2 - b_3], E_{11} = [h_d - d_0(t)] b_2 - b_3 (h_d - d_0(t)) \bar{b}_2 - b_3], E_{12} = [\bar{b}_0, b_0, \bar{b}_6, \bar{b}_5].\)
\(\bar{\omega}_1 = d_0(t) E_{13} \left( \frac{1}{3} Z \right) E_{13}^T + \text{Sym}\{E_{13} \tilde{N}_1 (\bar{b}_1 - b_2) \bar{b}_2^T + \tilde{N}_2 (2 \bar{b}_2 - b_2 - b_2) \bar{b}_2^T\}\)
\(\bar{\omega}_2 = (h_d - d_0(t)) E_{14} \left( \frac{1}{3} Z \right) E_{14}^T + \text{Sym}\{E_{14} \tilde{N}_1 (\bar{b}_2 - b_2) \bar{b}_2^T + \tilde{N}_2 (2 \bar{b}_2 - b_2 - b_2) \bar{b}_2^T\}\)
\(\bar{R} = \begin{bmatrix} R & R_{12} \\ R_{12} & R_{22} \end{bmatrix} > 0, \bar{S} = \begin{bmatrix} S_{11} & 0 & S_{12} & 0 & S_{13} & 0 \\ 0 & * & S_{22} & 0 & S_{23} & 0 \\ * & * & S_{32} & 0 & S_{33} & 0 \\ * & * & * & S_{44} & 0 & 0 \\ * & * & * & * & S_{55} & 0 \\ * & * & * & * & * & 0 \end{bmatrix} > 0\)
\(Q = \begin{bmatrix} Q_{11} & 0 & Q_{12} & Q_{13} & Q_{14} & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ * & * & Q_{22} & Q_{23} & Q_{24} & 0 \\ * & * & * & Q_{33} & Q_{34} & 0 \\ * & * & * & * & Q_{44} & 0 \\ * & * & * & * & * & 0 \end{bmatrix},\)
\(Z = \begin{bmatrix} Z_{11} & 0 & Z_{12} & Z_{13} & Z_{14} & 0 \\ 0 & 0 & 0 & 0 & 0 & 0 \\ * & * & Z_{22} & Z_{23} & Z_{24} & 0 \\ * & * & * & Z_{33} & Z_{34} & 0 \\ * & * & * & * & Z_{44} & 0 \\ * & * & * & * & * & 0 \end{bmatrix}\)

\(\eta_i(y) = \begin{bmatrix} x^T(y) \\ \bar{x}^T(y) \end{bmatrix} \int_y^{t(t)} \bar{x}^T(w) dw\)

According to Schur complement and positive definiteness of \(S, Q, Z, R\), the proposed L-K-F is positive definite from LMI (9)

Define:
\(\xi (t) = [x^T(t) \ x^T(t - t_d) \ \bar{x}^T(t) \ \bar{x}^T(t - t_d)]\)
\(\frac{1}{d_0(t)} \int_{t_d(h)}^{t} x^T(w) dw \ \int_{t_d(h)}^{t} \bar{x}^T(w) dw\)

Calculating the time derivative of \(V_i\) gives:
\(\frac{\partial V_i}{\partial t} = \sum_{i=2}^{n} \frac{\partial V_i}{\partial t}\)

where
\(\frac{\partial V_i}{\partial t} = \xi (t) \text{Sym} \{E_i P E_i^T\} \xi (t),\)
\(\frac{\partial V_i}{\partial t} = \xi (t) \text{Sym} \{E_i SE_i^T\} \xi (t),\)
\(\frac{\partial V_i}{\partial t} = \xi (t) \text{Sym} \{E_i SE_i^T\} \xi (t),\)
\(\frac{\partial V_i}{\partial t} = \xi (t) \text{Sym} \{E_i SE_i^T\} \xi (t),\)

By applying Lemma (1) with LMI (8), the UB of the derivative can be estimated as:
\(\frac{\partial V_i}{\partial t} = \xi (t) \eta (t)\)

According to Lemma (2), it is clear that if LMI (10) holds, then inequality (14) is satisfied.

For \(i \in I_1\) consider the L-K-F as bellow:
\(V_i(t, x_1) = V_i(t, x_1) + \sum_{g=2}^{n} V_g(t, x_1)\)
\(V_i(t, x_1) = \eta_i(t) P, V_2(t, x_1) = \eta_2(t) S_n_2(t),\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Q n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
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\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)
\(V_i(t, x_1) = \int_{t_d(h)}^{t} n_i(t) Z n_i(t) dy,\)

According to LMI (12) and positive definiteness of \( R, S, Q, \) \( Z \), the Lyapunov functional is positive definite. LMI (13) with respect to LMI (11) guarantees that the derivative of Lyapunov functional decreases over time.
Note that $V_2 = V_2$, $V_4 = V_4$, $V_5 = V_5$, and the condition that guarantees the continuity of $V_1$ and $V_1$ at the boundaries can be obtained by using the appropriate continuity matrices.

The suggested Lyapunov functional is continuous, but its derivative is not continuous at all points of the state space, so Theorem 1 is confirmed if sliding behavior does not occur at the boundaries. It is necessary to study of the appearance of the charming sliding modes at the boundaries for the PWA systems. An analysis to recognize this sliding mode according to the Filippov solution is discussed in [23].

IV. NUMERICAL EXAMPLES

This section presents three numerical examples to illustrate the effectiveness of the proposed stability conditions. In the first example, it is considered a switched linear time-delay system. In this system, the delay is constant and switching law depends on the current states. In the second example, it is considered the equation of the water level changes in a tank. In this example, the delay is time-varying and switching law depends on the current states. The third example is about the equation of motion of a simple pendulum. In this example, the delay is constant and switching law depends on the delayed states as well as the current states. The characteristics of these examples are given in Table 1. The conservativeness of delay-dependent stability conditions is checked by computing the maximum allowable UB of time delay.

<table>
<thead>
<tr>
<th>Type of delay</th>
<th>Dependence type of switching law</th>
</tr>
</thead>
<tbody>
<tr>
<td>Example 1</td>
<td>Constant</td>
</tr>
<tr>
<td>Example 2</td>
<td>Time-varying</td>
</tr>
<tr>
<td>Example 3</td>
<td>Constant</td>
</tr>
</tbody>
</table>

**Example 1**: Suppose the switched linear time-delay system

$$\dot{x} = A_1x(t) + A_2x(t-d_h)$$

with the system matrices given by:

$$A_1 = \begin{bmatrix} -0.1 & 0 \\ 0 & -0.1 \end{bmatrix},$$
$$A_2 = \begin{bmatrix} -0.1 & 0 \\ 0 & -0.1 \end{bmatrix},$$
$$A_3 = \begin{bmatrix} 0 & 5 \\ -1 & 0 \end{bmatrix},$$
$$A_4 = \begin{bmatrix} 0 & 1 \\ 0 & 0 \end{bmatrix}.$$

and the cell partition:

$$E_1 = \begin{bmatrix} 1 \\ -1 \end{bmatrix}, E_2 = \begin{bmatrix} 1 \\ 1 \end{bmatrix}, E_3 = \begin{bmatrix} 1 \\ -1 \end{bmatrix}, E_4 = \begin{bmatrix} 1 \\ 1 \end{bmatrix}.$$

In this system, the time-delay is constant ($d_h$), and the switching law depends on the current states. Table 2 shows the maximum allowable time delay obtained by Theorem 1 ($h_d$) and existing methods [17-21]. As shown in this table, the results obtained by Theorem 1 are less conservative than the currently existing ones.

**Table II**

<table>
<thead>
<tr>
<th>Methods</th>
<th>Feasible UB for delay ($h_d$)</th>
</tr>
</thead>
<tbody>
<tr>
<td>[17]</td>
<td>0.0142</td>
</tr>
<tr>
<td>[18,19]</td>
<td>0.0142</td>
</tr>
<tr>
<td>[20,21]</td>
<td>0.0168</td>
</tr>
<tr>
<td>Theorem 1</td>
<td>0.0193</td>
</tr>
</tbody>
</table>

The state trajectories of the system with initial conditions [-2,0]T are shown in Fig.1. As shown in this figure, the system is stable for $d_h = 0.02$ and unstable for $d_h = 0.021$. According to this, the theoretical maximum UB for time-delay in the system is between 0.02 and 0.021. Therefore, the simulation results confirm less conservativeness in Theorem 1.

**Example 2**: Fig.2 shows a water tank and a pipe with length of $L$. In this example, a water tank system with a nonlinear model has the following statement [18]:

$$\dot{x}(t) = \frac{1}{A_0} \left( -\frac{1}{H} \sqrt{\varphi g x(t)} + u_{in}(t-d_h(t)) \right)$$

(15)
where \( g = 9.8 \text{ ms}^{-2} \), \( \varphi = 1000 \text{ kgm}^{-1} \), \( H = 11.3882 \text{ m}^{3/2} \text{kg}^{1/2} \) and \( A = 10 \text{ m}^2 \). The pipe length \( L \) makes a time-varying delay \( (d_h(t)) \) in the water inflow to the tank. The goal is to keep the water level at \( x = 0.5 \text{ cm} \). We obtained a PWA model of the system with linearization around two operating points \( x_0 = 0.25 \text{ cm}, \ x_{0.75} = 0.75 \text{ cm} \), as follow as:

\[
\dot{x}(t) = \frac{-g x(t)}{2AH\sqrt{0.25\varphi g}} + \frac{1}{A\varphi} u_{in}(t-d(t)) \\
+ \left( \frac{0.25g}{2AH\sqrt{0.25\varphi g}} \right) \sqrt{0.25\varphi g} \varphi AH; \quad 0 \leq x(t) < 0.5
\]

\[
\dot{x}(t) = \frac{-g x(t)}{2AH\sqrt{0.75\varphi g}} + \frac{1}{A\varphi} u_{in}(t-d(t)) \\
+ \left( \frac{0.75g}{2AH\sqrt{0.75\varphi g}} \right) \sqrt{0.75\varphi g} \varphi AH; \quad 0.5 \leq x(t) < 1
\]

and the cell partition is obtained:

\[ E_1 = [ -1 \ 0.5 ], \ E_2 = [ 1 \ -0.5 ] \]

The switching law is based on the current states.

Suppose a control input denoted as below:

\[ u_{in}(t) = 0.2A\varphi x(t), \quad 0 \leq x(t) < 0.5 \]

\[ u_{in}(t) = -0.1A\varphi x(t), \quad 0.5 \leq x(t) < 1 \]

Fig 3 shows the level of water for the system without delay. As shown in this figure, the system without delay is stable. But in practice the system should be considered with delay. Increasing the delay can cause closed-loop instability. Therefore, it is necessary to find the maximum allowable upper bound of the time-delay for which the stability of the closed loop system is guaranteed.

Table 3 shows the results for the maximum allowable time-varying delay for various bounds of the time derivative of the time delay obtained by Theorem 1. In Table 3, \( \mu \) is the UB of the derivative of time delay, as shown in (2). It can be verified with the simulation that the nonlinear system is marginal stable for constant delay \( d_h = 6.53 \) as shown in Fig. 4 and the system is unstable for \( d_h = 6.54 \). The maximum allowable constant delay obtained with Theorem 1 for the equivalence PWA system is \( h_c = 6.265 \).

### Example 3

In this example, a simple pendulum motion model with constant time delay \( (d_h) \) between the sensor and the processor is considered as below [20]:

\[
M_p L_p \ddot{\theta}(t) = -M_p g L_p \sin(\theta(t)) + T(t-d_h)
\]  

(16)

where \( M_p = 1 \text{ kg} \) is the mass of pendulum, \( L_p = 9.8 \text{ m} \) is the length of pendulum, \( g \) is the gravity of earth, and \( T \) is the input torque. The general model of this system with time delay and the switching conditions, which depend on both the states and the delayed states, is obtained in [20]:

\[
[p(\dot{\theta}(t))] = \begin{bmatrix} 0 & 1 & 0 & -0.825 \end{bmatrix} \begin{bmatrix} \dot{\theta}(t) \\ \dot{\theta}(t-d_h) \end{bmatrix} + \begin{bmatrix} -0.825 \\ 0 \\ 0 \\ 0 \end{bmatrix} \begin{bmatrix} \dot{\theta}(t-d_h) \\ \dot{\theta}(t-d_h) \end{bmatrix}; \quad \forall \begin{bmatrix} \dot{\theta}(t-d_h) \\ \dot{\theta}(t-d_h) \end{bmatrix} \in \Psi_1
\]

\[
[p(\ddot{\theta}(t))] = \begin{bmatrix} 0 & 1 & 0 & -0.825 \end{bmatrix} \begin{bmatrix} \ddot{\theta}(t) \\ \ddot{\theta}(t-d_h) \end{bmatrix} + \begin{bmatrix} -0.825 \\ 0 \\ 0 \\ 0 \end{bmatrix} \begin{bmatrix} \ddot{\theta}(t-d_h) \\ \ddot{\theta}(t-d_h) \end{bmatrix}; \quad \forall \begin{bmatrix} \ddot{\theta}(t-d_h) \\ \ddot{\theta}(t-d_h) \end{bmatrix} \in \Psi_2
\]

\[
[p(\dddot{\theta}(t))] = \begin{bmatrix} 0 & 1 & 0 & -0.825 \end{bmatrix} \begin{bmatrix} \dddot{\theta}(t) \\ \dddot{\theta}(t-d_h) \end{bmatrix} + \begin{bmatrix} -0.825 \\ 0 \\ 0 \\ 0 \end{bmatrix} \begin{bmatrix} \dddot{\theta}(t-d_h) \\ \dddot{\theta}(t-d_h) \end{bmatrix}; \quad \forall \begin{bmatrix} \dddot{\theta}(t-d_h) \\ \dddot{\theta}(t-d_h) \end{bmatrix} \in \Psi_3
\]

\[
[p(\dddot{\theta}(t))] = \begin{bmatrix} 0 & 1 & 0 & -0.825 \end{bmatrix} \begin{bmatrix} \dddot{\theta}(t) \\ \dddot{\theta}(t-d_h) \end{bmatrix} + \begin{bmatrix} -0.825 \\ 0 \\ 0 \\ 0 \end{bmatrix} \begin{bmatrix} \dddot{\theta}(t-d_h) \\ \dddot{\theta}(t-d_h) \end{bmatrix}; \quad \forall \begin{bmatrix} \dddot{\theta}(t-d_h) \\ \dddot{\theta}(t-d_h) \end{bmatrix} \in \Psi_4
\]

with partitions:

\( \Psi_1 = \{ -0.7854 \leq \dot{\theta}(t) \leq 0, -0.7854 \leq \dot{\theta}(t-d_h) \leq 0 \} \);

\( \Psi_2 = \{ -0.7854 \leq \dot{\theta}(t) \leq 0, 0 \leq \dot{\theta}(t-d_h) \leq 0.7854 \} \);

\( \Psi_3 = \{ 0 \leq \dot{\theta}(t) \leq 0.7854, -0.7854 \leq \dot{\theta}(t-d_h) \leq 0 \} \);

\( \Psi_4 = \{ 0 \leq \dot{\theta}(t) \leq 0.7854, 0 \leq \dot{\theta}(t-d_h) \leq 0.7854 \} \);

The simulation results are given in Fig. 5 and 6 and show that the system is stable with \( d_h = 0.19 \), but becomes unstable when \( d_h = 0.196 \), which implies that the theoretical UB of the delay is between 0.19 and 0.196. The maximum allowable delay obtained with Theorem 1 for this system is \( h_c = 0.18 \). Due to switching based on delayed-states, the methods proposed in
[17-19] cannot deal with this example and the maximum upper bound that is obtained in [20], is \( h_u = 0.161 \). As a result, the proposed approach can achieve less conservative estimate of the UB for the delay in this system.

![Figure 5: The pendulum angle for \( d_h = 0.19 \)](image)

![Figure 6: The pendulum angle for \( d_h = 0.196 \)](image)

V. CONCLUSIONS

This paper introduced novel stability conditions for the PWA-TVD system. It is considered that the system switches are based on both the states and the delayed states. The stability criteria for the PWA-TVD system have been established in terms of LMIs. By choosing the improved L-K-F and using new inequality for the estimation of its derivative, the derived results are less complex and conservative. Future research can focus on several extensions such as stability condition for uncertain PWA systems, the use of different Lyapunov functional to obtain less conservative results, and the use of the presented theorem to stabilize PWA systems with time delay and uncertainty.

APPENDIX

The following algorithm is used to find the maximum allowable upper bound for delay:

**Step 1:** Select the positive constants \( h_d \) and \( \mu \).

**Step 2:** Solve LMIs (8)-(13) and then check the feasibility of LMIs for \( \forall d_h(t) \in [0, h_d], \quad d_h(t) \in (-\mu, \mu) \).

**Step 3:** If LMIs will be feasible, increase \( h_d \) and then solve LMIs (8)-(13) again (step 2), else if LMIs will be infeasible, exit the algorithm.

REFERENCES


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**Salehe Afsharian** received the B.S. and M.S. degree in control engineering from the Ferdowsi University of Mashhad (FUM), Iran, in 2009 and 2012, respectively. Since 2014, she is doing a PhD dissertation in Control Engineering at Department of Electrical and Computer Engineering, University of Birjand, Iran. Her research interests include Hybrid System, Time-delay Systems, Stability Analysis, Model Predictive Control, Modeling and Simulation, Renewable and Sustainable Energy, and Control.

**Hussein Eliasi** received the B.S. degree in electrical engineering from the Ferdowsi University of Mashhad (FUM), Iran, in 2001. He received the M.S. and the Ph.D. degrees from AmirKabir University of Technology (Tehran), Iran, in 2005 and 2011, respectively. He is currently an assistance professor in faculty of electrical engineering at University of Birjand, Iran. His research interests include Constrained Nonlinear Dynamics Systems, Robust Control, Parameter Optimization of Model Predictive Control, Transient Stability in Power Systems and Load Following by Nuclear Power Plants.
Optimal Impedance Voltage-Controller for Electrically Driven Robots

Mohsen Jalaeian F.¹, Mohammad Mehdi Fateh², and Morteza Rahimiyan³,
¹,²,³ Department of Electrical and Robotic Engineering, Shahrood University of Technology, Shahrood, Iran.

This paper presents a novel optimal impedance voltage-controller for Electrically Driven Lower Limb Rehabilitation Robots (EDLR). To overcome the dynamical complexities, and handle the uncertainties, the proposed method employs an expected forward model of the actuator. The existing value of lumped uncertainty is represented by the output difference of the model and system. A voltage-controller, which compensates for the uncertainties, is designed based on this uncertainty estimator. Parameters of the controller are optimized using genetic algorithms. Key contributions of this paper are I) uncertainty estimation through the expected model’s output, II) overcoming the changes in motors’ parameters, III) introducing a class of closed-loop system termed as “Repeatable”, and IV) designing an optimal impedance voltage-controller that is non-sensitive to the parameter variations. Significant merits of the approach are swift calculations, efficiency, robustness, and guaranteed stability. Furthermore, the simplicity of design, ease of implementation and model-free independent joint structure of the approach are noticeable. The method is compared with an adaptive robust sub-controller and a Taylor-series-based adaptive robust controller, through simulations in passive range of motion and active assistive rehabilitation exercises. The results show the superiority of the proposed method in tracking performance and the time of calculations.

I. INTRODUCTION

The human lower limb’s dysfunctionality has significant effects on daily life quality [1]. Nowadays, patients suffering from such diseases may improve their lower limb motor function through robotic rehabilitation programs [2]. Robots provide several benefits for the therapeutic process, namely reducing the costs [3], speeding up the process [4], reducing pain [5], as well as being user-friendly [6]. However, control of rehabilitation robots is a difficult complex problem, due to the lack of an accurate model for the human body, and the high sensitivity to interconnection forces [7]. In addition, the Electrically Driven Robots (EDR), has coupled third-order nonlinear differential equations [8]. Moreover, since the robot may be powered by batteries, energy consumption is another important issue.

Over the last decade, in the field of therapeutic robots’ control, several scientific articles such as a PID [9], adaptive robust controller [10], as well as voltage-based adaptive control [11] are presented.

In terms of using system equations, robot control methods are divided into four categories: (i) Model Free Controllers (MFC) [9], (ii) Computed Torque Control (CTC) [12], (iii) Integrated Dynamics Methods (IDM) [8], and (iv) Voltage Control Strategy (VCS) [11,13,14]. Although the robot control signal is applied through the actuators, they are not considered in the MFC and CTC methods. The IDM’s have heavy time-consuming calculations, whereas the dynamic effects of the complex coupled nonlinear dynamics of the robot, environment, and the mechanical part of the motor, altogether are observable in the motor current. Thus, the VCS methods design the controller using the electrical portion of
the actuator’s dynamics [13]. However, the electrical current’s derivative is hard to be measured [15,16], and actuator parameters may vary due to heat [17]. The VCS strategies try to approximate, estimate, or predict the unmeasurable parts of the equation using different types of observers or approximators. To do so, an adaptive uncertainty estimator [8], an adaptive impedance approach [11], fuzzy approximators [14,18,19], an estimator based on Fourier series expansion [16], a FAT-based robust adaptive approximator [20], and a Taylor series approximator [21] are introduced.

To the best of our knowledge, the variation of motor parameters is neglected in the literature. To overcome these challenges, a novel voltage-controller is proposed in this research that uses the known part of the system to estimate the model imperfection. In [22], a forward internal model is used to control a lower extremity exoskeleton; however, since it utilizes the dynamical model of the robot, the final control law is a complex highly nonlinear coupled system. In this paper, we employ the VCS to reach a simple final controller. The advantages of the approach are swift calculations, fast response, efficiency, guaranteed closed-loop stability, simplicity of design, ease of implementation, and independent joint structure.

This paper includes the following sections: Section 2 is allocated to detail the dynamical model of the EDLR. The proposed method is described in Section 3 while the analysis of its stability is expressed in Section 4. Then, the description and analysis of the simulations and comparisons are detailed in Section 5. Finally, the paper is concluded in the last section.

II. MODEL OF AN EDLR

The EDLR (Fig. 1) is a robot, with two degrees of freedom, which is used to provide rehabilitation exercises for the human lower limb in the sagittal plane [2,4].

![Fig. 1. The EDLR, assembly drawing.](image)

The equations of motion of an EDLR can be written as Equ. (1), [10]. In this paper, the matrices and vectors are written in bold characters while the scalars are written in narrow fonts.

\[
\begin{align*}
\mathbf{M}(\mathbf{q})\ddot{\mathbf{q}} + \mathbf{H}(\mathbf{q}, \dot{\mathbf{q}}) + \mathbf{f}(\mathbf{q})^{\top}\mathbf{F}_e + d &= \mathbf{T}_r \\
\text{in which, } \mathbf{q} &= [q_1, q_2, \ldots, q_n]^T \text{ is the vector of joints angles, where } n \text{ is the robot’s degree of freedom, } \\
\mathbf{M}(\mathbf{q}) &\in \mathbb{R}^{n \times n} \text{ is the inertia matrix, } \\
\mathbf{H}(\mathbf{q}, \dot{\mathbf{q}}) &\in \mathbb{R}^{n \times n} \text{ is a vector containing the centripetal, Coriolis, and gravity forces, } \\
\mathbf{f}(\mathbf{q}) &\in \mathbb{R}^{n \times 1} \text{ is the Jacobian matrix of the robot, } \\
\mathbf{F}_e &\in \mathbb{R}^{n \times 1} \text{ is a vector representing the patient-exerted-forces, } \\
\mathbf{T}_r &\text{ are the joints torques vector, and } d \text{ is the lumped bounded uncertainty. The torque of } i^{th} \text{ joint is produced by an electrical DC motor through a gearbox with the speed transmission ratio of } r_i, \text{ that means: }
\end{align*}
\]

\[
\dot{\mathbf{q}}_i = r_i\dot{\theta}_i
\]

in which, \( \dot{\theta}_i \) is the angular speed of the motor shaft. The dynamic equations of each DC motor consist of two parts, namely a mechanical portion and an electrical equation.

\[
\begin{align*}
J_{m,i}\ddot{\theta}_i + B_{m,i}\dot{\theta}_i + r_iT_{r,i} &= K_{m,i}\dot{\theta}_i \\
L_i\ddot{I}_i + R_iI_i + K_{b,i}\dot{I}_i &= V_i
\end{align*}
\]

where, \( J_{m,i} \) and \( B_{m,i} \) are the inertia and friction of the motor. \( V_i, I_i \), and \( T_{r,i} \) are the motor’s terminal voltage, armature current, and joint torque, respectively. \( K_{m,i} \) and \( K_{b,i} \) are the motor torque transmission and back-EMF constants. \( L_i \) and \( R_i \) are the actuator inductance and resistance.

The block diagram of a robotic-assisted lower limb rehabilitation process is depicted in Fig. 2.

![Fig. 2. Block-diagram of robotic rehabilitation.](image)

Substituting Equs. (1)-(2) in Equ. (3), the integrated mechanical portion is obtained:

\[
\begin{align*}
(r_m\mathbf{M}(\mathbf{q}) + r_m^{-1}\mathbf{J}_m)\ddot{\mathbf{q}} + r_m^{-1}\mathbf{B}_m\dot{\mathbf{q}} + r_m\mathbf{H}(\mathbf{q}, \dot{\mathbf{q}}) + r_m\mathbf{f}(\mathbf{q})^{\top}\mathbf{F}_e + r_md &= \mathbf{K}_m\dot{\mathbf{q}} \\
\end{align*}
\]

in which, \( \mathbf{J}_m, \mathbf{B}_m, \mathbf{K}_m, \) and \( \mathbf{r}_m \) are the diagonal matrices of the mechanical parameters of all motors.

Property 1 (Inertia Matrix Invertibility):

The \( \mathbf{M}(\mathbf{q}) \), is a symmetric positive definite matrix, which is bounded even if \( \mathbf{q} \) is not bounded. Therefore, \( \mathbf{M}(\mathbf{q}) = (r_m\mathbf{M}(\mathbf{q}) + r_m^{-1}\mathbf{J}_m) \) is a positive definite bounded matrix, and its inverse exists and remains non-zero and bounded [23].

Considering property 1, one can rewrite the dynamics of mechanical portion as:
\( \ddot{q} = \ddot{M}(q)^{-1}(K_M I - (r_M B_M \dot{q} + r_M H(q, \dot{q}) + r_M f(q)^T F_e + r_M d) \)  

**Assumption 1 (Desired Trajectory):**

The desired angular position \((q_d)\), velocity \((\dot{q}_d)\), and acceleration \((\ddot{q}_d)\) are given bounded.

**Assumption 2 (Voltage Supplier):**

The input voltage, which is produced by a real supplier, is bounded because of factual limits. 

\[ |V| \leq V_{\text{max}} \]  

(7)

**Assumption 3 (Measurement Noises):**

It is assumed that each measured signal is passed through a low-pass filter; thus, the measurement noise is not considered in the course of this paper.

## III. THE PROPOSED POSITION CONTROLLER

The voltage equation of each actuator is considered independently without losing generality. The subscript \( i \) is omitted, and \( \alpha \pm K_\sigma r^{-1} \) is defined for the sake of simplicity, hence, Eqn. (4) can be rewritten as:

\[
\alpha \ddot{q} + \dot{R} \ddot{I} + R I = V
\]  

(8)

Based on Eqn. (5), the motor current represents the dynamical effect of the entire mechanical parts of the system. Therefore, Eqn. (8) can be used to design the controller in an independent joint structure.

### A. Case 1: Expected-Model Voltage Control with fixed known parameters:

Here, we consider a simple forward model of the system defined as:

\[
\ddot{q}(t) = \alpha^{-1}(V(t - \varepsilon) - R I(t))
\]  

(9)

The unknown portion of the system can be computed by attention to Eqns. (8) and (9) as:

\[
\psi(t) \triangleq \dot{q}(t) - \ddot{q}(t) \equiv \alpha^{-1}(\dot{R} I(t) - \ddot{R} I(t))
\]  

(10)

in which, \( \ddot{R} I(t) = (V(t) - V(t - \varepsilon)) \), and \( \varepsilon \) is a very small positive value as a time delay. Then, the control law can be designed as:

\[
V = R I + \alpha(\dot{q}_d + \lambda e + \psi)
\]  

(11)

where, \( \varepsilon \approx q_d - q \), is the error of the joint angle, and \( \lambda \) is a real positive scalar. The tracking error dynamics, which is discussed later in details, is equivalent to \( \dot{e} + \lambda e = \alpha^{-1}\ddot{q}(t) \).

### B. Case 2: Extended-Model Voltage Control with parameters variations:

Motor parameters may vary during the runtime owing to the motor heat and temperature changes [17]. Hence, the parameters \( R \) and \( \alpha \) are supposed to be unknown, and two suggested design parameters, \( \ddot{R} \) and \( \ddot{\alpha} \), are utilized, instead. The new model can be proposed as:

\[
\ddot{q}(t) = \ddot{\alpha}^{-1}(V(t - \varepsilon) - \ddot{R} I(t))
\]  

(12)

Using the new model, Eqn. (12), total imperfection can be measured as:

\[
\psi(t) \triangleq \dot{q}(t) - \ddot{q}(t)
\]  

(13)

Analytically, the imperfection term \( \psi \) equivalents:

\[
\psi \equiv \alpha^{-1}(\dot{R} I(t) - \ddot{R} I(t) - \xi)
\]  

(14)

in which, \( \xi = (\alpha^{-1}V(t) - \ddot{\alpha}^{-1}V(t - \varepsilon)) \). Accordingly, a new control law is proposed as:

\[
V(t) = \ddot{R} I(t) + \dot{\alpha}(\dot{q}_d + \lambda e + \psi)
\]  

(15)

Applying the proposed control law, Eqn. (15), to the electrical equation of the actuator, Eqn. (8), yields:

\[
\ddot{q} = \alpha^{-1}\left((\ddot{R} I(t) + \dot{\alpha}(\dot{q}_d + \lambda e + \psi) - \ddot{R} I(t))
\right)
\]  

(16)

For analyzing the closed-loop system, the imperfection term, \( \psi \), can be substituted with its analytical equivalence, Eqn. (14):

\[
\alpha \ddot{q} = \ddot{\alpha}(\dot{q}_d + \lambda e) + \ddot{R} I(t) - (\ddot{R} I(t) + \ddot{R} I(t) - \ddot{\alpha}^{-1}\ddot{R} I(t) - \xi)
\]  

(17)

That yields:

\[
(\dot{e} + \lambda e) \equiv (\ddot{\alpha}^{-1}\ddot{R} I(t) - \ddot{\alpha}^{-1}\ddot{R} I(t) - \xi)
\]  

(18)

Apply some manipulations and exchanging \( \xi \), we obtain:

\[
(\dot{e} + \lambda e) \equiv \ddot{\alpha}^{-1}(V(t) - V(t - \varepsilon))
\]  

(19)

Thus, the closed-loop system can be represented by:

\[
(\dot{e} + \lambda e) = \ddot{\alpha}^{-1}\delta(t)
\]  

(20)

Note that, although \( \ddot{\alpha} \) is noticed in Eqn. (21), the estimation errors, \( \ddot{R} \triangleq (R - \ddot{R}) \) and \( \ddot{\alpha} \triangleq (\alpha - \ddot{\alpha}) \), do not play any role in the closed-loop system.

The pseudo-code of the proposed algorithm is detailed in Fig. 3.

- Initialize the control parameters (\( \ddot{R} \) and \( \ddot{\alpha} \))
- Initialize the imperfection indicator (\( \psi = 0 \))
- While (time < final_time)
  - Compute the control signal (\( V \))
  - Apply recent control signal into the system
  - Measure the feedback variables (\( I, \dot{q} \))
  - Compute the forward model output (\( \dot{q} \))
  - Compute the imperfection indicator (\( \psi \))
- End While

Fig. 3. The proposed controller pseudo-code.

In a rehabilitation robot with revolute joints driven by DC motors with bounded terminal voltages, the electrical current, its derivative, and the angular velocity remain bounded [11]. The closed-loop system stability is studied here. According to assumption 2, \( V(t) \) is bounded. When voltage is saturated, \( |V| = V_{\text{max}} \), it can be concluded that \( \delta(t) = 0 \), and it leads the Eqn. (21) to become \( \dot{e} + \lambda e = 0 \). If the voltage is not saturated, \( |V| < V_{\text{max}} \), then \( |\delta(t)| < 2 V_{\text{max}} \), and consequently, the input value of Eqn. (21) is bounded.
Therefore, the boundedness of error, $e$, and its derivative, $\dot{e}$, are guaranteed [24]. Thus, according to assumption 1, the actual position and velocity of the joint, ($q$ and $\dot{q}$), and of the motor, ($\theta$ and $\dot{\theta}$), remain bounded as well. Having assumption 2 and equation (8), the following is gained:

$$LI + RI = \nu$$  (22)

in which $\nu = (V - a\dot{q})$ is a bounded value as proved. Since Eq. (22) is a first-order linear differential equation with positive-constant coefficients, it is stable based on the Routh-Hurwitz criterion. With bounded input, $\nu$, the responses ($I$ and $L$) are bounded [24]. Although the motor parameters vary during time, they are always positive and bounded due to physical constraints. Ergo, Equs. (12) and (13) claim that $\dot{q}$ and $\dot{\psi}$ are bounded. In addition, Eq. (3) shows that the produced joints torque, $T_r$, is bounded. So, the closed-loop system is stable. □

IV. PROPOSED IMPEDANCE CONTROL, DESIGN, AND OPTIMIZATION

The main idea of mechanical impedance control is to enhance the desirability of dynamical interactions between a robot and its environment [25,26]. The desired impedance law for each joint is defined as a second order equation [7].

$$M_d(\ddot{q}_d - \ddot{q}_r) + B_d(\dot{q}_d - \dot{q}_r) + K_d(q_d - q_r) = f(q)^T{T_e}$$  (23)

in which, $M_d, B_d, K_d$ are the desired inertia, damping, and stiffness, respectively, and $q_d$ is the desired trajectory. In addition, $q_r$ is a regenerated trajectory for the joint angle. The block diagram of the proposed impedance control is presented in Fig. 4.

$$\ddot{q}_r = \ddot{q}_d + M_d^{-1}[B_d(\dot{q}_d - \dot{q}_r) + K_d(q_d - q_r) - f(q)^T{T_e}]$$  (24)

Since $f_e, q_d, \dot{q}_d$ and $\ddot{q}_d$ are bounded, and the desired impedance coefficients are positive, the $q_r$ is bounded. The block-diagram of using impedance controller in rehabilitation.

Fig. 4. The block-diagram of using impedance controller in rehabilitation.

Successful uncertainty handling is one of the main advantages of the proposed method, which enables the designer to tune the controller parameters using an offline optimization algorithm, like Genetic Algorithms (GAs) or Particle Swarm Optimization (PSO) [15,23]. Here, this characteristic is termed as “repeatability” of the closed-loop control system.

Definition 1: Repeatable closed-loop system

A closed-loop system is termed “Repeatable” if a fixed set of controller-parameters leads to a repetitive closed-loop response, even in the presence of uncertainty and external disturbances. In other words, in a repeatable closed-loop system:

$$\int_0^t (e(t) - e(t)(\phi,d)) dt \preceq \int_0^t (e(t))^2 dt$$  (25)

where, $e(t)$ is the tracking error of the closed-loop system without uncertainty or external disturbances, and $e(t)(\phi,d)$ is the tracking error of the closed-loop system in the presence of model uncertainty, $(\phi)$, and external disturbances, $(d)$, and $(\cdot)$ is used to denote the expectation value of the variables.

Remark 1:

Offline techniques are not being able to optimize the controller-parameters, due to the erratic essence of the system response, in the case of the unrepeatable system. However, if a system is repeatable, the offline optimization algorithm can be used to optimize its parameters; while uncertainties and unknown external disturbances have neglectable impacts on the results.

Remark 2:

The EDR with the proposed controller is repeatable, thus, the offline techniques may be utilized to optimize its parameters, without any worries about the exposure to uncertainty and external disturbances.

V. SIMULATIONS

In this section, two simulation scenarios (comparative position control, and optimized impedance control) of the LLRR are studied, in which, MATLAB® 2018a software is utilized within an ASUS laptop powered by Windows 10, and an Intel® Core™ i7-2.6GH CPU.

A. Position Control:

The proposed controller is applied to a 2-DoF EDR for hip and knee therapeutic exercises. Human joints have a certain Range of Motion (ROM), which are considered inside the interval of (-30°, +120°) for hip and (-90°,0°) for knee, [10,28,29]. In addition, Maximum Angular Speeds (MAS) of hip and knee joints are considered equal to 30 degrees per second. The parameters of motor and robot are detailed in Table 1. Interested readers can find full details of the dynamic equations in [10]. The proposed controller is applied to the position control in comparison with two other controllers, namely an adaptive robust sub-controller (ARSC) [10], and a Taylor-series-based adaptive robust controller (TSARC) [8].
In order to perform a Passive ROM exercise for lower limb rehabilitation in the joint training practice, previous scientific works set the desired joint trajectory as a four-part piecewise linear periodic path, which is very simple but not smooth [9,10], or as a sinusoidal trajectory, which has not enough rest in each period for the patient. This study proposes an adjustable smooth and continuous trajectory whose derivative is computable analytically. Thus, the angular position and speed are desirably adjustable. Since the exercises are periodic, the time inside each period is defined as $\eta = \text{Rem} \left( \frac{\text{time}}{T_{\text{prd}}} \right)$, in which $T_{\text{prd}}$ is a pregiven cycle-time that is determined by the physiotherapist, and $\text{Rem}(\cdot)$ computes the reminder value of the division. Afterward, the desired trajectory is proposed based on the $\eta$.

$$a_d(\eta) = \left( \max\text{ (ROM)} - \min \text{ (ROM)} \right) \times \left( y(\eta, \eta_1) - y(\eta, \eta_2) \right) + \min \text{ (ROM)}$$

where $y(\eta, \eta_i) = \left( 1 + \exp \left( - \frac{\eta - \eta_i}{a} \right) \right)^{-1}$, and $a$ is a setting parameter for maximum angular speed.

The desired trajectory, which is introduced in Remark 3, is applied to the proposed position control method. The tracking performances of hip and knee joints are shown in Fig. 5.

![Fig. 5. PROM tracking performance of hip and knee joints.](image)

According to Fig. 5, the proposed controller shows better tracking performance due to the smaller error and zero overshoot. There is an initial angle error that the controller overcoming it by nearly two seconds, (First zoomed areas in Fig. 5). To study the control performance, an external disturbance ($d = 2N$) is applied to the robot between the 35sec and 45sec, which the controller succeeds to damp its effects completely (Second zoomed areas in Fig. 5). Tracking errors are depicted in Fig. 6, where its value is small with the proposed controller, and the external disturbances do not affect the closed-loop errors (repeatability).

![Fig. 6. PROM tracking error of hip and knee joints (comparison).](image)

Based on Fig. 6, the proposed method proves superiority since its error magnitude is smaller than the other two, while no oscillation is observed in its response. To have a numerical comparison, several indices are studied in Table 2.

<table>
<thead>
<tr>
<th>Joint</th>
<th>Proposed Method</th>
<th>TSARC</th>
<th>ARSC</th>
</tr>
</thead>
<tbody>
<tr>
<td>ACT</td>
<td>1.22e-5</td>
<td>3.78e-5</td>
<td>1.60e-4</td>
</tr>
<tr>
<td>MCT</td>
<td>2.21e-4</td>
<td>2.90e-3</td>
<td>4.2e-3</td>
</tr>
<tr>
<td>MAV</td>
<td>2.778</td>
<td>2.785</td>
<td>3.956</td>
</tr>
<tr>
<td>MAC</td>
<td>0.66</td>
<td>0.26</td>
<td>0.82</td>
</tr>
<tr>
<td>SAE</td>
<td>8.68e2</td>
<td>8.78e2</td>
<td>5.8e4</td>
</tr>
<tr>
<td>MSE</td>
<td>0.21</td>
<td>0.21</td>
<td>7.39</td>
</tr>
</tbody>
</table>

ACT=Average cycle time, MCT=Max cycle time, MAV=Mean Absolute Voltage, MAC=Mean Absolute Current, SAE=Sum Absolute Error, MSE=Mean Square Error.

From the performance point of view, the tracking error of the proposed controller in both indices (Sum Absolute Error and Mean Square Error) is smaller than the other two. In addition, from the efficiency point of view, Mean Absolute Voltage, Mean Absolute Current and Average Cycle Time of the controller computation are studied while the proposed controller is better in those features. The proposed controller is superior in efficiency and performance, in spite of its great simplicity in design and implementation. Furthermore, the proposed structure is not sensitive to parameters estimation error. The controller has a few design parameters which should be set once in the first run. In contrast with a majority of controllers that have many design parameters and are sensitive to their setting values, which must be readjusted for each new trajectory. Control signals, the terminal voltages of motors, are illustrated in Fig. 7. As seen, the voltages are saturated when the error is big; however, the controller

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Dimension</th>
<th>Joint 1</th>
<th>Joint 2</th>
</tr>
</thead>
<tbody>
<tr>
<td>$m_i$</td>
<td>kg</td>
<td>5.5</td>
<td>5</td>
</tr>
<tr>
<td>$I_i$</td>
<td>m</td>
<td>0.5</td>
<td>0.45</td>
</tr>
<tr>
<td>$d_i$</td>
<td>m</td>
<td>0.45</td>
<td>0.4</td>
</tr>
<tr>
<td>$I_1$</td>
<td>kg.m²</td>
<td>0.25</td>
<td>0.2</td>
</tr>
<tr>
<td>$F_s$</td>
<td>N.m</td>
<td>0.9</td>
<td>0.9</td>
</tr>
<tr>
<td>$F_e$</td>
<td>N.m</td>
<td>0.8</td>
<td>0.8</td>
</tr>
<tr>
<td>$\sigma_i$</td>
<td>N.m/ rad.s⁻¹</td>
<td>1.2</td>
<td>1.2</td>
</tr>
<tr>
<td>$K_m$</td>
<td>N.m/A</td>
<td>4.32</td>
<td>4.32</td>
</tr>
<tr>
<td>$R_i$</td>
<td>$\Omega$</td>
<td>4</td>
<td>4</td>
</tr>
<tr>
<td>$L_m$</td>
<td>H</td>
<td>2.5e-3</td>
<td>2.5e-3</td>
</tr>
<tr>
<td>$K_e$</td>
<td>v/ rad.s⁻¹</td>
<td>0.287</td>
<td>0.287</td>
</tr>
<tr>
<td>$K_g$</td>
<td>$r^{-1}$</td>
<td>-</td>
<td>16</td>
</tr>
</tbody>
</table>

**Remark 3:**

To study the control performance, an external disturbance ($d = 2N$) is applied to the robot between the 35sec and 45sec, which the controller succeeds to damp its effects completely (Second zoomed areas in Fig. 5). Tracking errors are depicted in Fig. 6, where its value is small with the proposed controller, and the external disturbances do not affect the closed-loop errors (repeatability).
overcomes and becomes unsaturated as fast as possible. The effect of external disturbance on voltages can be seen in the zoomed area in Fig. 7. Motors currents are displayed in Fig. 8 as well. Fig. 8 shows that the current is small and smooth. Based on the aforementioned principle, the dynamical behaviors of mechanical parts can be observed in the currents, as seen in Fig. 8.

Fig. 7. motors voltages (control signals) of hip and knee joints.

Fig. 8. DC motors currents.

B. Optimized Impedance Control:

The proposed scheme is utilized in impedance control of the LLRR in the presence of uncertainty and external disturbance. Based on remarks 1 and 2, GA is used to minimize the Mean Square Error (MSE) of tracking performance. Note that, the cost function for optimization can be selected based on error and energy [30], or one of them.

As expected, changes on $R$ do not affect system performance. Nevertheless, the effect of changing in $\hat{\alpha}$ should be taken into account. Fig. 9 illustrates the impact of changes in $\hat{\alpha}$. The optimized $\hat{\alpha}$ is used, to minimize the error. Fig. 10 illustrates the patient’s exerted force.

Fig. 9. The impact of changes in $\hat{\alpha}$ on impedance error.

The tracking performance is also depicted in Fig. 11, where the regenerated and actual trajectories are shown.

Fig. 10. Patient’s exerted force.

Fig. 11. Tracking performance (regenerated and actual trajectories).

Fig. 11 shows that the actual joints angles track the regenerated trajectories. However, the difference between them is shown in Fig. 12, as the impedance tracking error, with less than a few degrees which is neglectable in the context of rehabilitation.

Fig. 12. The difference between the measured and commanded errors.

Applied voltages to the actuators are shown in Fig. 13.

Fig. 13. Actuators’ voltages (control signal).

These voltages provide the joints torques, which are depicted in Fig. 14.
It can be seen that the actuators’ voltages are smooth and bounded. In other words, the closed-loop system that contains the robot, patient, and the environment, altogether behaves as the reference desired impedance rule.

In the future studies, authors going to utilize model predictive compensator or dynamic-growing fuzzy-neural acceleration-based compensator, which are first introduced by the authors in [31,32], to compensate for the effects of $\delta(t)$.

VI. CONCLUSIONS

The EDLRs can be controlled through the terminal voltage of their actuators in an independent joint manner. However, the parameters of actuators may vary over time and the value of the motor current derivative is unmeasurable. In this paper, a novel approach has been proposed to overcome these challenges. The proposed control method, which is applied to the position and optimal impedance control of an EDLR, has overcome the uncertainty and complexities in dynamics of the robot. Swift calculations, high performance and efficiency, robustness, and guaranteed stability are the main merits of the proposed method. Comparatively, the higher performance of the method is validated showing less tracking error. At the same time, greater efficiency is achieved through brief calculations, smaller control signal (voltage), and reduced power consumption leading to smaller motor sizing. In addition, the most significant advantages of the method are the independent joint structure, simplicity of design, and ease of implementation. The method has been compared with two others, namely an adaptive robust sub-controller, and a Taylor-series based robust controller through simulations. The results of simulations and comparisons have confirmed the aforementioned merits.

REFERENCES


Mohsen Jalaeian-F received the M.Sc. degree of control engineering from the Ferdowsi University of Mashhad, Iran, in 2012. Currently, he is a Ph.D. candidate at Shahrood University of Technology. He is also with the Center of Excellence on Soft Computing and Intelligent Information Processing (SCIP). His research interests include robotics, control, optimization, AI/ML (soft-computing), fuzzy logic, and deep reinforcement learning.

Mohammad Mehdi Fateh received his B. Sc. degree from Isfahan University of Technology in 1988. He received his M.Sc. degree in electrical engineering from Tarbiat Modares University, Iran in 1991, and his Ph.D. degree in robotic engineering from Southampton University, the UK in 2001. He is a full professor with the Department of Electrical and Robotic Engineering at Shahrood University of Technology, Iran. His research interests include nonlinear, robust and fuzzy control, robotics and intelligent systems, mechatronics, and automation.

Morteza Rahimiyan received the B.Sc. degree from the Isfahan University of Technology, Isfahan, Iran, in 2003, and the M.Sc. and Ph.D. degrees (with honors) from the Ferdowsi University of Mashhad, Mashhad, Iran in 2006 and 2011, respectively, all in electrical engineering. He is currently an associate professor at Shahrood University of Technology, Shahrood, Iran. His research interests include operation, planning, and economics of power systems, as well as optimization.
Linear Modelling of Six Pulse Rectifier and Designee of Model Predictive Controller with Stability Analysis

Hamid Radmanesh¹, Masood Saeidi ²*

¹Electrical Engineering Department, Shahid Sattari Aeronautical University of Science and Technology, Tehran, Iran
²Department of Electrical and Computer Engineering, Iran University of Science and Technology, Tehran, Iran

The AC/DC converter is one of the popular power electronic converters in industrial applications such as in the railway, power supply systems and electric vehicle. In this paper, a three-phase controllable rectifier is considered and its linear model is extracted. Because of MPC controllers benefits, the continuous control set model predictive controller (CCs-MPC) is designed for controlling this rectifier output DC voltage. By considering rectifier dynamic response, the suitable criteria to choice the model predictive controller parameters such as sampling time, prediction horizon and control horizon is proposed. In experimental implantation the computing burden of microcontroller is limit therefore the reaching to optimal and minimum complexity in algorithms implantation is vital problem. In other words by using these proposed criteria for selection of sample time, prediction and control horizon the tradeoff between computational burden, system performance and dynamic stability is made. When using designed MPC controller, the rectifier and grid performance such as total harmonic distribution (THD), power factor (PF) and output voltage ripple have acceptable value. This controller can eliminated the effect of heavy load change on rectifier performance which is very common problem in industrial system. Also, this controller stability guaranteed is checked by using the dual-mode method. The simulation results and controllers performance are validated in MATLAB software.

Keywords: Linear model, Power Conversion, Predictive control, Rectifier, Stability analysis

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I. INTRODUCTION

Once Todays, power electronic converters have become the most important controllable device in power systems [1]. With using of power electronics converter to merge all of the new energy sources, make possible many advantages such as reduced component counts, weight reduction and control simplicity in power utility and microgrids [2]. In the past, due to the low level of semiconductor manufacturing technology, the converters switching speed was low and making using them impossible or very limit [2]. Nowadays, thanks to semiconductor manufacturing improvements, these converters can be used in high switching frequency and high power rating, so today, in the power system and industrial application these converters are used at the scale of Gigawatt and with frequency up to 400 kHz [3]. After this progression, they are widely used in microgrids, hybrid/electric transportation vehicles or hybrid energy systems. Benefit of implementation control methods on power electronic converters have been investigated and different methods for control these converters have been proposed in [4]. The conventional Linear controller is used for controlling the output voltage by appropriate tuning the Kp and Ki values. However, this control method have some disadvantage as result researcher tried to develop new control method form past years. Due to the existence of uncertainties and nonlinear behavior in the power electronic system, using the linear...
controller theory is not recommended [5]. So, researchers have tried to use new control methods that solve the linear controller theory problems and use all capacity of this converter. To overcome these problems, new control methods such as neural networks, fuzzy, and sliding mode control as known intelligent control methods have been proposed in [5-6]. The nonlinear controller such as MPC, fuzzy and sliding mode controllers have some benefits and advantage with compression of linear controllers such as dynamic speed, low ripple and ability to consider system model nonlinearity effect and etc. Therefore in recent years especially after microprocessor computational power progression, most of the nonlinear controller are implemented in linear control system. In spite of creating computational burden and being complex in both design and implementation procedures, using these controllers can indeed overcome the disadvantage of ordinary linear controllers but linear controllers nowadays still the first priority in power electronic converters in compression with this intelligent methods. In recent years theory of the predictive control methods are presented and this control method has gradually been used in the all fields of industrial application. The initial implementation of model predictive control algorithm has been introduced in the chemical and thermal industry [7]. With the advancement of computer processors, the processing speed has been increased dramatically and providing the ability to implement new control method such as model predictive control algorithms in most industrial system [8].

According to the discrete nature of the power electronic converters, the use of predictive control algorithm in these systems has been suggested in [9, 10]. One of the main reasons for using this method in power electronic converters, is the possibility of considering constraints in the controller design. In [11, 12], designing and implementation of constrained model predictive controller methods in power electronic converters have been investigated. The application of predictive control in various types of power electronic converters are reviewed in [13], which has demonstrated the use of this control algorithm in the distributed generation sources become widespread in last years [13].

Generally, model predictive control can be divided into Continuous Control Set (CCS)-MPC and Finite Control Set (FCS)-MPC [14]. In FCS-MPC, searching and selecting the optimal switching mode from among the number of switching states and after applying it to the system, while in the CCS-MPC, the optimal control signal is produced by optimizing the process according to the system dynamic equations. CCS-MPC requires a modulator to generate control signals, which cause the fixed switching frequency, while in FCS-MPC the switching frequency is variable [15]. In [16], these two control methods in the induction motor drive are compared. In fact, for designing an appropriated controller, modeling the power electronic converters are necessary. In [17, 18], the modeling of these converters and electrical machines drives are discussed. In these references, the Taylor expansion method for small signal modeling without considering large signal modeling and parameter variation is used. Therefore, using small signal model leads to a significant error conditions such as in fault conditions and severe load change, that it may cause disturb control system. The model predictive controller can reduce uncertainties effects on the system performance hence the use of linear models for designing a prediction controller will lead to satisfactory results. But in some literature, such as [19, 20], designing and using of nonlinear model predictive controllers in power electronic converters is proposed, although it contributes to the better performance than linear MPC but leads to complex design and operation aren't suitable for industrial application.

Today, all controllers are implement digitized, but it is necessary to select the suitable sampling time for these systems. The importance of selecting the appropriate sampling time in the MPC algorithm is discussed in [21]. Some of the most important reasons for choosing the appropriate sampling time in the MPC controller are impacts on the calculations time and the dependence of the controller stability. Design and implementation of constrain MPC is another important topic which is discussed in [22, 23]. In most articles that refer to the use of MPC controller in power electronic converters and electrical drives, the issue of choosing an appropriate sampling time and the prediction and control horizon, aren't considered. In most MPC controller applications in power electronics, have not offer analytical method for designing of MPC parameters such as sampling time, prediction horizon and control horizon while in this article, attempt has been made to set these parameters using a dynamic model and system behavior.

In this paper, CCS-MPC for controlling a three-phase rectifier is designed. According to the system dynamic response, a criterion is proposed for selecting the prediction and control horizon in the MPC controller, and MPC controller stability proof is investigated by use of the dual-mode method. In the some related works in power electronics, the stability proof is not considered for MPC controller or other new control methods. While in this study, the stability of the proposed controller is investigate and proved. In the VI it is shown that the stability of the MPC controller is related to the prediction horizon and sampling time which indicates the importance of the criterion controller parameters designing procedure.

This paper is arranged as follows. The mathematical and linearized model of the three-phase controllable rectifier is presented in Section 2. Model predictive controller design of rectifier is presented in Section 3. Stability analysis of modified CCS-MPC controller with using dual-mode method
is illustrated in Section 4. The simulation results are describe in Section 5. Conclusions are given in Section 6.

II. MATHEMATICAL MODEL OF THREE-PHASE CONTROLLABLE RECTIFIER

In this section, the three-phase converter (rectifier), which uses the IGBT switches, is investigated. Rectifiers are the oldest and the most used power electronic converters. To use them in high power applications, the input voltage of these converters is usually three phases. To reduce the output voltage fluctuations, a capacitor is usually used as a voltage filter in rectifier structure. Also, a resistive load is used in rectifier structure. The three-phase rectifier topology is shown in Fig. 1.

![Fig. 1. Studied rectifier structure](image)

The state space method is used to model this converter. Dynamic elements are selected as system state variables; therefore, capacitor voltage and grid current are selected as rectifier system state variables [25]. With this assumption and with using Table 1 the state equations of the system are given as follows:

\[
\frac{di_a(t)}{dt} = -\frac{R}{L} i_a - \frac{1}{L} v_{za}(t) + \frac{1}{L} v_g(t) \\
\frac{di_b(t)}{dt} = -\frac{R}{L} i_b - \frac{1}{L} v_{zb}(t) + \frac{1}{L} v_g(t) \\
\frac{di_c(t)}{dt} = -\frac{R}{L} i_c - \frac{1}{L} v_{zc}(t) + \frac{1}{L} v_g(t) \\
I_{dc}(t) = I_{dca}(t) + I_{dcb}(t) + I_{dcd}(t) \\
\frac{dv_{dc}(t)}{dt} = \frac{1}{c} I_{dc}(t) - \frac{1}{cR_L} v_{dc}(t) \\
v_{rabc} = \frac{1}{2} m_i(t) v_{dc} \\
v_{rbc} = v_{abc} - v_z \\
i_{abc} = \frac{1}{2} m_i(t) i_{dc}
\]

Where \(V_{abc}, V_{bpc}, V_{cgc}\) are the amplitude of grid voltages and \(V'\) is amplitude voltage drop on grid impedance. By using the above equations, the model of the system is written as follows:

\[
\frac{di_a(t)}{dt} = -\frac{R}{L} i_a - \frac{1}{L} v_{ra}(t) + \frac{1}{L} v_r(t) \\
\frac{di_b(t)}{dt} = -\frac{R}{L} i_b - \frac{1}{L} v_{rb}(t) + \frac{1}{L} v_r(t) \\
\frac{di_c(t)}{dt} = -\frac{R}{L} i_c - \frac{1}{L} v_{rc}(t) + \frac{1}{L} v_r(t) \\
\frac{dv_{dc}(t)}{dt} = \frac{1}{c} I_{dc}(t) - \frac{1}{cR_L} v_{dc}(t) \\
\frac{dv_{dc}(t)}{dt} = \left(\frac{i_a + i_b + i_c}{cV_{dc}}\right) \left(\frac{v^2 + 2v v' \cos(\phi) + v'^2}{cV_{dc}}\right) - \frac{1}{cR_L} v_{dc}
\]

With sinusoidal grid voltage assumption and using (4) the above equations are rewritten as follows:

\[
\frac{di_a(t)}{dt} = -\frac{R}{L} i_a - \frac{m_i}{2L} v_{dc} - \frac{1}{L} V_{ag} \\
\frac{di_b(t)}{dt} = -\frac{R}{L} i_b - \frac{m_i}{2L} v_{dc} - \frac{1}{L} V_{bg} \\
\frac{di_c(t)}{dt} = -\frac{R}{L} i_c - \frac{m_i}{2L} v_{dc} - \frac{1}{L} V_{cg} \\
\frac{dV_{dc}(t)}{dt} = \frac{3m_i}{2c} I_{dc}(t) - \frac{V_{abg}}{cR_L} \sin(\omega t) - \frac{V_{abc}}{cR_L} \sin(\omega t + \phi) \\
\phi = \tan^{-1}\left(\frac{XL}{R}\right)
\]

### Table 1: Parameter definition of studied rectifier

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Definition</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>(I_{abc})</td>
<td>Grid three phase current</td>
<td>Amp</td>
</tr>
<tr>
<td>(V_{abc})</td>
<td>Grid three phase voltage</td>
<td>Volt</td>
</tr>
<tr>
<td>(V_{dc})</td>
<td>Dc link voltage</td>
<td>Volt</td>
</tr>
<tr>
<td>(C)</td>
<td>Capacitor</td>
<td>F</td>
</tr>
<tr>
<td>(m_i)</td>
<td>modulation index</td>
<td>-</td>
</tr>
<tr>
<td>(I_{dc})</td>
<td>output dc current</td>
<td>-</td>
</tr>
<tr>
<td>(R)</td>
<td>Grid resistance</td>
<td>(\Omega)</td>
</tr>
<tr>
<td>(L)</td>
<td>Grid inductance</td>
<td>H</td>
</tr>
</tbody>
</table>

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In (15-18) with considering that grid currents have dependency with together, therefore, one phase current and dc link capacitor voltage are selected as state variables and grid voltages are system input variables respectively. In the state equations, there are some nonlinear term such as \( \frac{i_a + i_b + i_c}{CV_{dc}} \) and \( m_l (t) \) therefore for linear analyses, with using 10 kW rectifier parameter (Table I) and using the Taylor expansion method can be linearize of (15-18) around the nominal operation point \((i_{abc} = 26.18, V_{dc} = 400, V' = 19.09)\), transfer function is calculated as:

\[
H(s) = \frac{v_{dc}(s)}{m_l(s)} = \frac{2.8286(s + 175100)}{(s + 51.17)(s + 125.7)}
\]  

(19)

### TABLE III

**RECTIFIER SYSTEM PARAMETER VALUES**

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Unit</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( P_{rectifier} )</td>
<td>Kw</td>
<td>10</td>
</tr>
<tr>
<td>( V_g )</td>
<td>Volt</td>
<td>180</td>
</tr>
<tr>
<td>( V_{dc} )</td>
<td>volt</td>
<td>400</td>
</tr>
<tr>
<td>( f_{grid} )</td>
<td>Hz</td>
<td>60</td>
</tr>
<tr>
<td>( L_{grid} )</td>
<td>mH</td>
<td>1.93</td>
</tr>
<tr>
<td>( R_{grid} )</td>
<td>( \Omega )</td>
<td>0.1</td>
</tr>
<tr>
<td>( R_{L} )</td>
<td>( \Omega )</td>
<td>16</td>
</tr>
<tr>
<td>( C )</td>
<td>( \mu F )</td>
<td>990</td>
</tr>
</tbody>
</table>

### III. MODEL PREDICTIVE CONTROL DESIGN

Because of nonlinear behavior and parameter uncertainties in power electronic converters, by using the new controller, the defects of linear controllers are eliminated [25]. The model predictive control algorithm is one of the best options for using in power electronic converter applications.

#### A. Model Predictive Control Algorithm

Model predictive control algorithm generates a proper control signal to stabilize the rectifier voltage. MPC controller is implemented in 4 forms which are: DMC, AMC, PFC, and GPC [26]. In [26], a full comparison between 4 forms of the MPC controller is taken and its result is that using GPC format for unstable, non-minimum phase and system with very small zero is appropriate. Therefore in this paper, the GPC form for MPC implementation is used. For nonlinear effects minimizing, Using of MPC is recommended [26], therefore in this paper, it is applied to the rectifier system.

The rectifier control system is shown in Fig. 2.

![Fig. 2. Rectifier control system.](image)

The MPC is a multivariable control algorithm and the calculation of the optimal control move is based on solving the optimization problem defined by a cost function and the control goal can be prescribed by the quadratic cost function. The MPC controller cost function can be formulated as (20), a control signal can be produced, with minimization of the cost function \( J \):

\[
J(N_p, N_c, N_i) = \sum_{j=1}^{N_c} \delta(j)[\hat{y}(t + j|t) - w(t + j)]^2 + \sum_{j=1}^{N_i} \lambda(t)\Delta u(t + j - 1)^2
\]  

(19)

where, \( N_p \) is prediction horizon, \( N_c \) control horizon, \( N_i \) model delayed, \( \Delta u \) control signal, \( y(t) \) model output(t) reference set point and \( \delta(t) \) and \( \lambda(t) \) are the weight factors [18].

#### B. Tuning of the Model Predictive Controller Parameter

Choosing the appropriate values for the controller parameters in the cost function can be effective in reducing the computation burden. Therefore, with proper selection of these parameters, optimal control signals can be generated which
improves the power electronic converters performance and efficiency. The prediction horizon and sample time should be selected according to the system specification [21]. If the cost function doesn't have any constraints, by applying optimization methods, the cost function can be optimized by conventional method but in case of constrained MPC controller design, complex minimization method such as active set and Gauss-Seidel method are applied [22, 23].

The MPC controllers are implemented in both schemes of finite control set (FSC-MPC) and continuous control set (CCS-MPC). The idea of using FCS-MPC is referred as to the natural discrete property of power electronic converter hence by applying the model prediction, the switches finite state is predictable [29]. This method has advantages, such as lack of need for modulators, the simple implementation, and the intuitive understanding algorithm. Having variable switching frequency, system response fluctuations and the steady-state error in the system output response are considered as its disadvantage. Another scheme for implementing the MPC controller is acquired via the CCS approach. In this method, the MPC controller generates an appropriate reference signal which is used in SPWM or SVM modulators. Some of the CCS-MPC advantages are given as constant switching frequency, the possibility of eliminating the steady-state error, less sampling time and designed controller with proof of the possibility of stability moreover using long horizon police and provide MPC with the high degree of robustness [9]. In comparison with FCS-MPC controller, CCS-MPC requires less time for computing and it has a clear design approach otherwise it is vulnerability to noise and external disturbances effects.

In order to calculate CCS-MPC algorithm, use of online and offline methods are recommended. An online method based on calculations of the control law in each sample interval [10]. The offline mode is based on obtaining explicit control signals with consideration of system operation points and keep these explicit signals. Specification all operating point in which the optimal control moves are determined by evaluating a linear function. Explicit MPC controllers require lower computation time than the conventional controller. Therefore it is useful for applications which require small sample time [31]. Principles of determining suitable operation points are under discussion been in [32], further investigation is beyond the scope of this paper. To implement the CCS-MPC controller, selection of the sample time, the predictive horizon and the control horizon is necessary. The volume of computations depends on the choice of these parameters so that they should be precisely selected until lower processing time needed for the algorithm computations. Some criteria for choosing sample time and prediction horizon are given in [33] but, in this reference, the first order system with constant time delay is under discussion, while rectifier system model is two order system with one zero. Therefore, this method will not provide any appropriate responses in the rectifier system. In this paper, it is suggested that the sampling time should be selected based on model dynamic response and theorem of Shannon which is using in continues model discretization process. Based on digital control theorem, for converting S-domain system to Z-domain model, in each oscillation cycle between 8-12 samples is require until the quantization error is ignorable. In this paper, this number is selected 10, and in selecting of the prediction horizon for oscillating systems, it should be noted that the prediction horizon should be able to cover at least one peak or one valley of the wave in order to provide sufficient information about the system model so that the prediction process can be done to produce the control signal properly still, the value of the control horizon should not be taken very high because it diminishes the calculation rate of system.

C. Model Predictive Controller Design

In this paper, the primary prediction and control horizon is given the initial value 11 and 2 respectively but the changes in the prediction and control horizon in the system response are still checked in following simulation and investigate this parameters effect on controller response. The sample time for the system transfer function discretization according to the system dynamic response is selected 0.00083(s). Sample time is selected 0.00083 (s) and using of zero-pole match method for discretization. The discrete transfer function of the system is given as follow:

$$h(Z^{-1}) = \frac{0.116Z^{-1} + 0.107Z^{-2}}{1 - 1.858Z^{-1} + 0.8622Z^{-2}}$$  \hspace{1cm} (21)

Commonly, in GPC technic, controlled autoregressive integrated moving average (CARIMA) model type is used. The equation of CARIMA model can be derived as follows:

$$A(Z^{-1})y(t) = B(Z^{-1})u(t-1) + C(Z^{-1})\xi(t)\frac{\Delta}{\Delta}$$  \hspace{1cm} (22)

Where for power electronics and drive application, d is considered 1, $\zeta(t)$ represent noise in system and $\Delta = 1 - Z^{-1}$, $\Delta$ is deviation operator. If $\zeta(t)$ is white noise, $C(Z^{-1})$ is set to 1 thus (22) can be simplified as:

$$A(Z^{-1})y(t) = B(Z^{-1})u(t-1) + \frac{\xi(t)}{\Delta}$$  \hspace{1cm} (23)

In order to calculate the prediction step, the following Diophantine equation is considered as following [33]:

$$1 = E_f(Z^{-1})A(Z^{-1}) + Z^{-1}F_f(Z^{'})$$  \hspace{1cm} (24)

Calculation of F and E terms are described in [33]. The best possible prediction for $y$ is:

$$y(t+j) = G_f(Z^{-1})\Delta U(t+j-1) + F_f(Z^{-1})y(t)$$  \hspace{1cm} (25)

In which: $G_f(t) = E_f(Z^{-1})B(Z^{-1})$. In (25) the term of $G_f(Z^{-1})\Delta U(T + j - 1)$ is divide into 2 terms, concerning past and future. Sum of the past output term with $F_f(Z^{-1})$ is
named free response ($\hat{f}$) and system response to future value is force response. System transfer function $h(Z^{-1})$ is expressed as following:

$$
A(Z^{-1})\psi_{dc}(t) = B(Z^{-1})m_{1}(t-1)
$$

(26)

At the first step, assuming that there is no constraint in the system, the control signal is obtained by minimizing as follow:

$$
\frac{\partial j}{\partial U} = 2(G^T G + \lambda I) U + 2G^T (f - w) = 0
$$

(27)

$$
U = (G^T G + \lambda I)^{-1} G^T (w - f)
$$

(28)

In MPC controller, receding horizon approach is used and in any optimization one term of control effort ($U$) is applied to the system [34]. In the above equation, $G$ is system dynamic matrix, $f$ denotes the free response of the system, $\lambda$ the weighting factor and $W$ the reference trajectory.

**IV. STABILITY ANALYSIS OF CCS-MPC CONTROLLER WITH USING DUAL MODE METHOD**

To ensure the stability of the controller, it should be demonstrated mathematically. In linear controllers, it is possible to prove the stability through conventional methods, such as an NYQUIST diagram, root locus curve or bode diagram analysis [35]. Stability analysis method of MPC controllers is not as same as linear controllers. In the references such as [36], [37], mathematical methods for proving the stability of MPC controllers have been discussed.

The method examined in this paper is to prove the stability of the MPC controller, known as the Dual Mode method. In (20), if the upper limit of the first summation approach to infinity, the optimization problem is feasible. But in practical and industrial applications, the computation power of microprocessors are limited, therefore, prediction and control horizon are selected on the basis of a specified criteria such as the defined method. Hence MPC controller requires stability investigation. The basis of the Dual Mode method is minimizing the cost function from one until prediction horizon and applying state feedback from the perdition horizon to the infinity. Accordingly, if the conditions of the Dual Mode problem are satisfied, it is strongly claimed that the MPC controller is stable. The rectifier system transfer function in (19) is rewritten as the state space model in below:

$$
\begin{bmatrix}
  x_1(k+1) \\
  x_2(k+2)
\end{bmatrix} = \begin{bmatrix}
  1.83 & -0.83 \\
  1 & 0
\end{bmatrix} \begin{bmatrix}
  x_1(k) \\
  x_2(k)
\end{bmatrix} + \begin{bmatrix}
  1 \\
  0
\end{bmatrix} v_i(k)
$$

(29)

$$
\nu(k) = \begin{bmatrix}
  0.2383 & 0.218
\end{bmatrix} \begin{bmatrix}
  x_1(k) \\
  x_2(k)
\end{bmatrix}
$$

The quadratic cost function with an infinite horizon is written as follows:

$$
J = \sum_{j=0}^{\infty} x^T(k+j)Qx(k+j) + u^T(k+j)Ru(k+j)
$$

(30)

As being as, the (30) until infinite prediction horizon is minimized, it will certainly be stable, however in practice the control horizon is limited, so the (30) is rewritten as following and this is done by choosing the terminal weighting matrix $\bar{Q}$ so that the term $x^T(k+N_p +1)Q\Delta x(k+N_p +1)$ is equal to the cost over the mode 2, which is equivalent to the state feedback effect [34].

$$
J = \sum_{j=0}^{N_p} x^T(k+j)Qx(k+j) + u^T(k+j)Ru(k+j) + x^T(k+N_p +1)\bar{Q}\Delta x(k+N_p +1)
$$

(31)

Terminal weighting matrix is achieved by a solution of the LYAPUNOV equation:

$$
\bar{Q} - (A + BK)^T Q (A + BK) \geq K^T RK
$$

(32)

Where $A, B$ are calculated from the state space model and denotes is stated feedback gain.

The LYAPUNOV equation (32) has a unique solution for $\bar{Q}$ if and only if: 1) the eigenvalues of $A + BK$ are located inside in the unit circle, 2) $Q + K^T RK$ where $Q = C^T C$ positive definite [34] is. For the controllable system, by designing an appropriate $K$, closed-loop poles can be placed at the specified locations [34]. Using terminal weighting matrix, LYAPUNOV function can be defined as $V(x) = x^T \bar{Q}x$ where $V(x) > 0$ and $\dot{V}(x) < 0$ which guarantee the MPC controller stability.

In summary, this paper suggests dual-mode receding horizon control method with state feedback controller which is applied inside the attractive region and a receding horizon controller applied outside the terminal region [33]. Dual mode method concept is shown in Fig. 3.

**Fig.3.** The schematic diagram of the dual-mode method

By using equations of (29) and (32), state feedback gain ($K$) and terminal weighting matrix ($\overline{Q}$) are defined as follow:
\[ K = \begin{bmatrix} 1.833 & 1.077 \end{bmatrix} \]
\[ \bar{Q} = \begin{bmatrix} 5.34 & 1.83 \\ 1.83 & 1 \end{bmatrix} \] (33)

After calculation of \( \bar{Q} \) and \( K \), we can claim that designed CCS-MPC controller is stable [38]. In case of power electronic converters and electrical drive applications, stability problem of MPC controller isn't a major issue because their models aren't complex, therefore, using this paper the proposed method for sampling time and prediction horizon selection, don't require for power electronic converters and electrical drive stability evaluate but without using appropriate criteria, stability analysis is needed.

V. SIMULATION RESULTS

Plotting of the time or frequency response is the first step in any controller design. The rectifier system step response is shown in Fig. 4(a). According to Fig. 4(a), the system output voltage without a controller is not appropriate and has 5% steady state error, as well as the system settling time, is not suitable. Therefore a controller should be designed to improve the system output voltage.

Fig. 4(b) depicts the simulation result of the system via the CCS-MPC controller. As can be seen from Fig. 4(b), the use of CCS-MPC controller in the rectifier system has made it possible to provide the fast response with high steady-state precision without having any oscillation or overshoot rectifier output voltage. As a result, the selection of criteria for MPC parameter is acceptable.

![Fig. 4](image)

**Fig. 4.** The Rectifier output voltage step response: without (a) and with (b) CCS-MPC controller with the prediction horizon 11, and the control horizon 2

The effect of sampling time, prediction and control horizon variation on system performance is discussed in following. Initially, the sampling time will change without altering other parameters. As shown in Fig. 5(a), by increasing the sample time, system dynamic response is also strongly affected and reducing the system speed response. In fact, by increasing the sampling time, a portion of the model is ignored and the signal is not properly recovered and ultimately it reduces system dynamic speed. The effect of reducing the sampling time on the performance of the system has been investigated in Fig. 5(b). The simulation result shown in Fig. 5(b) illustrates the influence on DC side voltage when sampling time is reduced. When the sampling time is set to 0.0002 seconds, the output voltage has a 550V peak in its response, therefore, if sampling time more reduced, it may lead to voltage instability. The system output oscillation reason is related to reduction its sampling time because when the sampling rate is smaller, the little amount of data from the model response is available then the controller cannot be able to generate the optimal signal, that it may even lead to system instability. In other words, in the selection of sampling time, should be established a tradeoff between output response speed, stability, and computational complexity, thus the criterion which is proposed in this paper is an effective solution. The effect of changing the prediction horizon on the system response is investigated in Fig. 5(c). As can be seen from Fig. 5(c), the effect of prediction horizon reduction is approximately equivalent to reducing the sampling time with this difference that decreasing the prediction horizon will reduce the burden of the complexity computations. So, in practice, there will be a relax tradeoff between the system response and facilitates calculation.

Control horizon is another tuning parameter in CCS-MPC controller. In contrast to the two parameters of prediction horizon and sampling time, this parameter has no significant effect on the system stability and it only affects the transient response of the system. The choice of control horizons is important only in constrained systems and its selection is not important for unconstrained systems so that the long control horizon result in more computational volume [12]. Fig. 5(d) shows the variation of the control horizon on the system response.

![Fig. 5](image)

**Fig. 5 (a)**

**Fig. 5 (b)**
The effect of (a) increasing the sampling time, (b) reducing the sampling time, (c) prediction horizon change, (d) control horizon change on the system response.

A. Analysis of Rectifier Performance with CCS-MPC Controller

In order to show the MPC controller benefits, converter performance when using MPC controller should be investigated. Grid current, PF and THD are very important factors in rectifier converter control, therefore in Fig. 6 is showing these parameters when using the MPC controller. As inferred from Fig. 6, THD and PF parameter in the rectifier system is controlled in the acceptable range, therefore, can be recommended for using this control algorithm in high power rectifier system however in traditional Thyristor rectifier, power quality problem is an open topic of research that requires further attention. In output voltage harmonic spectrum is shown that after the main DC component, harmonics in f=300 Hz and f=600 Hz are other harmonics which is less than IEC61000 3-2 standard values therefore from power quality aspect, this system output voltage is desirable. Following, controller performance has been investigated when the load is increased. Due to the fact, when using the linear model for system behavior representation, this model is valid on system operation point so the system load changes are should be changed around the operation point [28]. In this paper, a modified structure of the MPC controller is used and therefore we expect that this controller can reduce the effect of heavy load change in rectifier performance.

Fig. 8 shows rectifier output DC voltage when the load is increased 20%, 50%, 100% and 150% respectively. As can be noticed, Fig. 8 evidence that when using MPC in the rectifier control system heavy load variations do not distribute output voltage severely because, in the new structure, nonlinearity of system is reduced by selection of appropriate correction factor. From Fig. 7 can be inferred that when load changes, the rectifier output voltage have steady state error that depends on the value of the new load because when load is increased dramatically, system voltage droop in grid impedance is increased therefore rectifier input voltage is reduced and as a result, this effect cause voltage steady state error in output voltage.
In practical applications, overvoltage and lower voltage conditions events are very probable case. This events causes may be related to starting high load system, the transient overvoltate of capacitors or short circuit fault in the system, therefore in this conditions the investigation of the rectifier performance is very necessary. Fig. 8(a) shows the rectifier output voltage when a three-phase 30% overvoltage fault has occurred in $t = 0.25$ (s) until $t = 0.4$ (s). As shown in Fig. 9(a), at the fault duration event in the grid, the rectifier output voltage is stabilized. In online CCS-MPC, cost function optimization will be done in any sampling time so the differences between the linear model and real system are decreased as result CCS-MPC controller is robustness again external conditions. This MPC algorithm reduces the uncertain effects of the system on the output voltage and it robust the system in front of severe faults such as over or under voltage.

In the final step, should be tested rectifier performance in harmonic condition. In this case, 20% harmonic order 5 is applied to grid voltage and investigate rectifier performance. As shown in Fig. 8(b), with using MPC controller in rectifier voltage harmonic effects are eliminated in output DC voltage. In practical application, the input voltage of the rectifier converter has some harmonic such as 5 and 7 orders so as a result from Fig. 8(b) using of MPC controller is guaranteed rectifier output voltage in the normal range.

**Fig. 9 (a)**

![Figure 9(a)](image)

**Fig. 9: Rectifier output voltage, (a) with three phase overvoltage, (b) in the harmonic condition.**

**VI. CONCLUSION**

In this paper, the linear model of the controllable rectifier is obtained and the specified transfer function is calculated. The main goal of this paper is designing a CCS-MPC controller for regulating rectifier output DC voltage and investigating its performance in the rectifier converter. The criterion for selecting of CCS-MPC sampling time, prediction and control horizon is proposed. This criterion is the specified procedure for tuning the sampling time and prediction horizon with the consideration system dynamic response and the computing power of the processor. Using of CCS-MPC controller demonstrates that it has the appropriate result in some grid and converter characteristics such as grid PF, THD, and output voltage drop value. In particular, when sever changes or faults and disturbances affects the system, MPC controller has the ability to eliminate this outer disturbance. In this paper, MPC controller stability proof is investigated by using the dual model method. The main result from MPC controller stability procedure, is that, if a valid criterion is used in traditional power electronic converter, the stability analysis won’t be a vital problem but if don’t regard suitable tuning method in selection of MPC controller parameters, stability analyses and determination of stability margin are vital step in MPC controller design.

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Hamid Radmanesh was born in 1981. He received the B.Sc. Degree in electrical Engineering from Malek-Ashtar University of Technology, Tehran, Iran, in 2006, the M.Sc. Degree in electrical Engineering from Shahed University, Tehran, in 2009, and the Ph.D. degree in electrical engineering from Amirkabir University of Technology (AUT), Tehran, in 2015. He has authored more than 100 published technical papers and has been involved in several industrial projects and educational programs in the fields of power electronics and power systems. His research interests include transient in power system, renewable energy, power quality, HVdc transmission systems, and more-electrical aircraft.

Masood Saeidi was born in 1994. He received the B.Sc. degree in electrical engineering from Shahid Beheshti University, Tehran, Iran in 2016 and currently, is the M.Sc. Of electrical Engineering in Iran University of Science and Technology, Tehran, Iran since 2016. His main interesting activity is related to power electronic converter modeling, model predictive control applications in power electronics, electrical machine control, power system control and renewable energy.
Robust Congestion Control Using Sliding Mode Control In TCP/IP Computer Networks

Shoorangiz Shams S Farahani1, †, Narges Masoomabadi2, Mohammadreza Jahed-Motlagh3

1Department of Electrical Engineering, Islamshahr branch, Islamic Azad University, Islamshahr, Iran,
2Department of Electrical Engineering, Science and Research Branch, Islamic Azad University, Tehran, Iran,
3Department of Electrical Engineering, Iran university of Science and Technology, Tehran, Iran

Based on the recent Internet advances, congestion control is considered as an important issue and has spurred a significant amount of research. In this study, second-order sliding mode control is used to adjust the average queue length and maintain the closed-loop system performance. The control law is obtained in two steps. First, the nonlinear state-space form of the network is extracted based on state variables as the average queue length and congestion window size. Then, the proportional-Integrator-derivative and proportional- derivative sliding surface are defined according to the tracking error. Also, in order to avoid chattering, the derivative of the sliding surface is considered and the closed-loop system stability is investigated based on Lyapunov theory. The proposed scheme renders good tracking specifications and closed-loop system robustness. The simulation results show that the proposed methods outperform proportional integral (PI) and proportional integral derivative (PID) schemes. Also, robustness to disturbances increases and chattering and transient response degradation are avoided.

I. INTRODUCTION

A set of protocols is defined as TCP/IP reference model to enable end-to-end (E2E) communications over the Internet. The Transmission Control Protocol (TCP) is an Internet standard which is utilized by many applications[1]. In [2], the scalable TCP is established as a Multiplicative Increase Multiplicative Decrease (MIMD) protocol. It explains the conditions where the total multiplicative increase dynamics of window size is transformed to an additive increase one. In [3], a framework is developed for networked TCP applications which supports both congestion avoidance and slow start algorithms. The aforementioned model addresses the router network which supports any Active Queue Management (AQM) techniques.

Congestion control is an important function which motivates wide acceptance of TCP. Congestion of packets at the outgoing queues in routers renders low reliability and network performance degradation. So, more effective congestion control schemes are needed. Since 1990s control theory has been applied to solve congestion problem in communication networks [4].

Recently, there has been a vast amount of research on using sliding mode control in congestion control schemes [5]-[12]. A TCP congestion control mechanism based on a sliding window mechanism where an additive increase multiplicative decrease (AIMD) algorithm is employed to fit the transmission rate to available network resources is
introduced in [5], a TCP congestion control protocol which uses the congestion window growth function as an exponential function and introduces an adaptive increasing factor in the aforementioned function is established in [6]. SMC to control congestion in TCP networks where stability analysis is assessed using Lyapunov theorem is assessed in [7], an adaptive generalized minimum variance (AGMV) as a congestion controller for dynamically varying TCP/AQM networks is presented in [8], sliding mode variable structure control (SMVS) is presented as a congestion controller for AQM in [9], an SMC based on TCP input–delayed model for AQM routers supporting TCP data transfer is established in [10], SMC for uncertain time delay TCP/AQM network systems is presented in [11] and an SMC technique to control congestion in differentiated service communication networks is addressed in [12].

Time delay could be found in different fields. In [13], delay and data loss compensation are studied in Internet-based process control systems, and Internet transmission delay is overcome by considering a variable sampling time. In this regard, two compensation elements in feedforward and feedback channels are presented. Delay can lead to epidemic outbreak in complex networks and it must be considered in the analysis of congestion control methods [14]. In [15], decentralized LMI-based strategies are presented in a delayed nonlinear network to control congestion. The schemes are robust to queue size changes and the consequent delay changes. Also, the control problem is solved using linear matrix inequalities (LMIs). The stability of a TCP/RED congestion control model is studied in [16] where delays are state-dependent and Random early detection (RED) is considered for Internet congestion control. However, RED parameters are difficult to study because of model discontinuous terms and delay.

During the last two decades, great attention has been paid to improving TCP congestion performance by considering slow start, congestion avoidance, fast retransmit, and fast recovery as the congestion control scheme modules. In [17], AQM schemes are used to evaluate TCP performance to control congestion where different TCP and AQM variants are studied. The study is conducted by a practical setup and shows the importance of choosing proper TCP and AQM variants. In [18], the bibliography of TCP/IP congestion control schemes in the last two decades is classified and some main results are obtained. In [19], a scalable testing method is presented for congestion control schemes in real-world TCP implementations where the TCP congestion control is either tested at the interface level or it is tested using an equivalence class of test inputs simultaneously.

In this study, second-order sliding mode control is used to adjust the average queue length and maintain the closed-loop system performance. In this strategy, the control law is obtained in two steps. Also, in order to avoid chattering, the sliding surface derivative is considered and the closed-loop system stability is investigated based on Lyapunov theory.

The reminder of this article is organized as follows: Section II summarizes the TCP/IP dynamic model. Sections III and IV present the sliding mode congestion controller for TCP/IP dynamic model based on sliding surface proportional-integrator-derivative (SSPID) and sliding surface proportional-derivative (SSPD), afterwards, in section V, the performance of our approach is assessed by a simulation set up, and finally section VI provides the conclusion of this paper.

II. TCP/IP Dynamic Model

Among different dynamic models which are developed for TCP/IP, the model by [20] is one of the most widely used models which incorporates the fluid flow and stochastic differential equations to represent TCP/IP dynamics. In the aforementioned model, the queue size and the congestion window size are considered as state variables. The model is expressed as follows:

\[ w(t) = \frac{1}{R(t)} w(t - R(t)) - \frac{1}{2} \frac{R(t - R(t))}{R(t)} p(t - R(t)) \]

\[ q(t) = N \frac{w(t)}{R(t)} - C \]

where \( w(t) \) is the TCP congestion window size (packets), \( q(t) \) the queue size (packets), \( R(t) \) the round-trip time (RTT) (seconds), \( C \) the link capacity (packets / sec), \( N \) the traffic load (the TCP session number) and finally \( p(t) \) is the packet drop probability. In (1), the congestion window size and the queue size are considered as state variables. Also, the packet drop probability and the queue length are input and output signals, respectively. By considering \( x_1(t) = w(t) \) and \( x_2(t) = q(t) \) as state variables, \( u(t) = p(t - R(t)) \) as input signal and \( y(t) = q(t) \) as system output, the network nonlinear state-space model is as follows:

\[ x_1(t) = \frac{1}{R(t)} x_1(t) x_2(t - R(t)) - \frac{1}{2} \frac{R(t - R(t))}{R(t)} u(t) \]

\[ x_2(t) = N \frac{x_2(t)}{R(t)} - C \]

Also, RTT is the queue length function and we have:

\[ R(t) = \frac{q(t)}{C} + \tau_p \]

where \( \tau_p \) represents the propagation delay (seconds). So, the state-space model (2) can be written as:
\[ i_1(t) = \frac{1}{C + \tau_p} x_1(t) - \frac{x_1(t)}{2} \left( \frac{x_1(t)}{C + \tau_p} + \tau_p \right) u(t) \]  
(4)

\[ i_2(t) = N - \frac{x_2(t)}{C + \tau_p} \]

There exist some constraints on the input signal and system states which shall be considered in the controller as a saturation function. The constraints are as follows:

\[ 0 \leq q(t) \leq q_{\text{max}} \]
\[ 0 \leq w(t) \leq w_{\text{max}} \]
\[ 0 \leq p(t) \leq 1 \]

where \( q_{\text{max}} \) is the queue capacity (packets) and \( w_{\text{max}} \) is the maximum congestion window size (packets).

### III. Sliding Mode Congestion Controller for Tcp/Ip Dynamic Model

Sliding mode control (SMC) can be considered as an important control scheme which is robust to disturbances and modeling uncertainties. Since there exist uncertainties occurring in TCP/IP computer networks, SMC is suggested for congestion control in this paper. In this scheme, states should reach a sliding surface in a limited period of time and remain there. The queue length tracking error can be written as:

\[ e(t) = q(t) - q_d = x_1(t) - x_{2d} \]

where \( q_d \) is the optimal average queue size. In this section, the sliding surface is considered as a sliding surface proportional-integrator-derivative (SSPID):

\[ s(t) = \dot{e}(t) + \lambda_1 e(t) + \frac{1}{2} \lambda_2 e(t)^2 \]

The sliding mode control ensures that system states converges to the sliding surface asymptotically in a limited time. The Lyapunov function is considered as:

\[ v(t) = \frac{1}{2} s^2(t) \]

By deriving (8), we have

\[ \ddot{v}(t) = s(t) \ddot{s}(t) \]  
(9)

To ensure the stability of the closed loop system, (9) shall be negative. So, the inequality (10) is used:

\[ v(t) = s(t)s(t) \leq -\eta |s(t)| \rightarrow s(t) \leq -\eta \text{sign}(s(t)) \]

Using (7), the sliding surface derivative is as follows

\[ \dot{s}(t) = \dot{e}(t) + \lambda_1 \dot{e}(t) + \lambda_2 e(t) \]

Considering the tracking error (6), we have

\[ \dot{s}(t) = x_2(t) - x_{2d} + \lambda_1 (x_2(t) - x_{2d}) + \lambda_2 (x_2(t) - x_{2d}) \]

Since we have

\[ \dot{x}_2(t) = N \left( \frac{x_1(t)}{C + \tau_p} - \frac{x_1(t)}{2} \left( \frac{x_1(t)}{C + \tau_p} + \tau_p \right) u(t) \right), \]

(13)

By substituting (2), we have

\[ \dot{s}(t) = \frac{N}{C} \left( \frac{x_1(t)}{C + \tau_p} - \frac{x_1(t)}{2} \left( \frac{x_1(t)}{C + \tau_p} + \tau_p \right) u(t) \right) \]

(14)

\[ e(t) = q(t) - q_d = x_1(t) - x_{2d} \]

(5)

\[ N \left( \frac{x_1(t)}{C + \tau_p} - \frac{x_1(t)}{2} \left( \frac{x_1(t)}{C + \tau_p} + \tau_p \right) \right) \]

(15)

\[ \dot{s}(t) = \frac{2N}{C} \left( \frac{x_1(t)}{C + \tau_p} - \frac{x_1(t)}{2} \left( \frac{x_1(t)}{C + \tau_p} + \tau_p \right) u(t) \right) \]

(16)

where

\[ \dot{\Phi}(x,t) = \frac{2N}{C} \left( \frac{x_1(t)}{C + \tau_p} - \frac{x_1(t)}{2} \left( \frac{x_1(t)}{C + \tau_p} + \tau_p \right) u(t) \right) \]

(17)

\[ \dot{\Psi}(x,t) = -N - \frac{2}{C} \frac{x_1(t)}{C + \tau_p} + \frac{x_1(t)}{2} \left( \frac{x_1(t)}{C + \tau_p} + \tau_p \right) \]

(18)

So, the sliding surface derivative is:

\[ \dot{s}(t) = \dot{\Phi}(x,t) + \dot{\Psi}(x,t) u(t) - \dot{x}_2(t) + \dot{\lambda}_1 (x_2(t) - x_{2d}) + \dot{\lambda}_2 (x_2(t) - x_{2d}) \]

(19)

In order to satisfy (18), the control signal is considered as follows:
\( u(t) = \frac{1}{\Psi(x,t)} \)  
(20)

\[-\Phi(x,t) + \dot{x}_{md} - \lambda_1 (x(t) - x_{md}) - k_1 (x(t) - x_{md}) - k_2 \text{sign}(s(t)) \]

In (20), \( k \) is sliding mode gain and plays a key role in dealing with uncertainties.

**IV. Sliding Mode Congestion Controller for Tcp/Ip Dynamic MODEL**

In this scheme, the sliding surface is considered as a sliding surface proportional -derivative:

\[ s(t) = \dot{e}(t) + \lambda e(t) \]  
(21)

The sliding surface derivative is defined as:

\[ \dot{s}(t) = -k_1 s(t) - k_2 \text{sign}(s(t)) \]  
(22)

There exist two design parameters, namely, \( k_1, k_2 \) in (22). SMC suffers from the disadvantage of high frequency oscillations of the control signal which is considered as chattering. In case of using high order sliding mode control, increasing the sliding mode gain end in chattering decrease in the control signal. By considering \( k_1 \), a stable dynamic is obtained for sliding surface derivative where the derivative of the sliding surface tends to be zero. To extract the SMC law, first, the sliding surface derivative is calculated. So, we have:

\[ \dot{s}(t) = \dot{e}(t) + \lambda e(t) = \Phi(x,t) + \Psi(x,t)u(t) - x_{md} + \lambda (x(t) - x_{md}) \]  
(23)

Using Eq. (22), we have

\[ \Phi(x,t) + \Psi(x,t)u(t) - x_{md} + \lambda (x(t) - x_{md}) = -k_1 s(t) - k_2 \text{sign}(s(t)) \]  
(24)

So, the SMC law is as follows:

\[ u(t) = \frac{1}{\Psi(x,t)} \]

\[-k_1 s(t) - k_2 \text{sign}(s(t)) + x_{md} - \lambda (x(t) - x_{md}) - \Phi(x,t) \]

(25)

In the following, by choosing the sliding surface derivative (22), Lyapunov stability is established. The Lyapunov function is considered as:

\[ v(t) = \frac{1}{2} s^2(t) \]  
(26)

So, we have

\[ \dot{v}(t) = s(t) \dot{s}(t) = s(t) \left( -k_1 s(t) - k_2 \text{sign}(s(t)) \right) \]  
(27)

Since

\[ s(t) \geq 0 \rightarrow \text{sign}(s(t)) = 1 \rightarrow s(t) \text{sign}(s(t)) > 0 \]

\[ s(t) < 0 \rightarrow \text{sign}(s(t)) = -1 \rightarrow s(t) \text{sign}(s(t)) > 0 \]

so,

\[ \dot{v}(t) = -k_1 s^2(t) - k_2 s(t) \text{sign}(s(t)) \leq 0 \]  
(28)

The above inequality shows that if Eq. (22) is satisfied, the closed-loop system stability is guaranteed. In other words, (29) shows that the sliding surface converges to zero asymptotically. On the other hand, according to Eq. (6), if the sliding surface converges to zero, the tracking error will converge to zero asymptotically.

**V. Performance Analysis**

In this part, first the performance of SSPID and SSPD is studied. In this regard, changes in queue length, packet loss probability and the congestion window size and RTT are studied. Afterwards, the impact of disturbance is examined on the system performance. Finally, the results of the proposed controller are compared with PI and PID controllers.

**A. Performance Analysis of the Proposed Controllers**

In this part, the performance of SSPID and SSPD is studied in case of tracking of the desired queue length and packet drop probability. Also, the congestion window size, RTT, the sliding surface and its derivative are shown. The sliding mode controller sets the average queue length such that the system performance is guaranteed in the presence of external disturbances and model uncertainties. The simulations are run for the following network parameters:

\[ N = 60, C = 3750, \tau_d = 0.2 \]. The parameters set in SSPID are \( \lambda_1 = 16, \lambda_2 = 64, k = 5 \) and the parameters set in SSPD are \( k_1 = 15, k_2 = 10, \lambda = 20 \). It is worth noting that the state variables and control signal are constrained in TCP-based computer network systems. Fig.1 shows the tracking of the desired queue length in SSPID and SSPD.

![Fig. 1. Tracking of the desired queue size in in SSPD and SSPID](image)
The control signal implies that the packet drop probability is high in transient time, however, it is reduced after the transient time. The packet drop probability in SSPD is less than that of in SSPID, so a better control signal is gained in SSPD comparing with SSPID. Fig. 3 shows the congestion window size and RTT in SSPD and SSPID.

The congestion window size in SSPD and SSPID converges to a constant value after a certain period of time which shows the closed-loop system stability. Also, RTT converges to a constant value after the transient time. Fig. 4 shows the sliding surface and its derivative in SSPD and SSPID. It can be seen that the sliding surface and its derivative converge to zero in SSPD. However, the sliding surface in SSPID does not converge to zero.

**B. Performance Analysis of the Proposed Controllers in the Presence of Disturbance**

In this part, the performance of the proposed controller is studied in the presence of disturbance. A step disturbance (value=20, time=10) is applied to the average queue length. Table I shows the performance of the proposed controllers in case of average queue length error, average congestion window size error, average RTT error and average packet drop probability with and without disturbance.

<table>
<thead>
<tr>
<th>Scheme</th>
<th>Performance metrics</th>
<th>Average queue length error</th>
<th>Average congestion window size error</th>
<th>Average RTT error</th>
<th>Average packet drop probability</th>
</tr>
</thead>
<tbody>
<tr>
<td>SSPID (without disturbance)</td>
<td>6141.8</td>
<td>3895.1</td>
<td>5.480</td>
<td>0.0591</td>
<td>3.1167 x 10^{-4}</td>
</tr>
<tr>
<td>SSPID (without disturbance)</td>
<td>6.7812</td>
<td>5.8491</td>
<td>5.480</td>
<td>0.0591</td>
<td>3.1167 x 10^{-4}</td>
</tr>
<tr>
<td>SSPID (without disturbance)</td>
<td>0.0616</td>
<td>0.0596</td>
<td>5.1880</td>
<td>0.0591</td>
<td>3.1167 x 10^{-4}</td>
</tr>
<tr>
<td>SSPID (without disturbance)</td>
<td>5.9104 x 10^{-4}</td>
<td>5.3271 x 10^{-4}</td>
<td>3.8958 x 10^{-4}</td>
<td>0.0591</td>
<td>3.1167 x 10^{-4}</td>
</tr>
<tr>
<td>SSPID (without disturbance)</td>
<td>0.0616</td>
<td>0.0596</td>
<td>5.1880</td>
<td>0.0591</td>
<td>3.1167 x 10^{-4}</td>
</tr>
</tbody>
</table>

Table I shows that SSPD without disturbance outperforms SSPD with disturbance in terms of average queue length error, average congestion window size error, average RTT error and average packet drop probability. Also, SSPD without disturbance outperforms SSPID with disturbance in terms of average queue length error, average congestion window size error, average RTT error and average packet drop probability. Moreover, SSPD without disturbance ends in better performance comparing with SSPD with disturbance, and SSPID with/without disturbance. Also, SSPID with disturbance performs the worst comparing with SSPID without disturbance, and SSPID with/without disturbance.
C. Comparison of the Proposed Controller with PI and PID Controllers

In this part, the performance of the proposed controller (SSPD) is compared with PI [21] and PID [22] controllers. In [21], given that \( W_0 \), \( p_0 \) and \( q_0 \) are the equilibrium points of TCP nonlinear system, the transfer function of \( \delta p \) and \( \delta q \) is determined as follows:

\[
G_{p_0}(s) = \frac{R_0C_0K}{R_0s + 1} e^{-R_0s} \tag{30}
\]

where

\[
\delta p = p(t) - p_0, \quad \delta q = q(t) - q_0, \quad K = \frac{R_0C_0}{2N_0}, \quad R_0 = T_0 + \frac{q_0}{C_0} \tag{31}
\]

In [21], the performance of PI controllers is studied in TCP networks. In this paper, Skogestad’s method [23] is suggested where the system transfer function as the following first order approximation is taken into consideration:

\[
G_{p_0}(s) \approx \frac{R_0C_0K^2}{K + \frac{1}{2}} e^{-\frac{3}{2}R_0s} \tag{32}
\]

The controller parameters are derived as follows:

\[
k_p = \frac{1}{C_0K^2 \tau_c + 1.5R_0} \
\]

\[
k_i = \min \{ KR_0 + 0.5R_0, 4(\tau_c + 1.5R_0) \} \tag{33}
\]

The following values are obtained as the controller parameters:

\[
k_p = 4.2196, \quad k_i = 3.7056
\]

Based on the model, a PID controller is as follows [22]:

\[
G_r(s) = 2.9067 \times 10^{-5} \left( 1 + \frac{1}{5.45s} + 0.4157s \right) \tag{34}
\]

In this section, the desired average queue length is 120. Fig. 5 shows the tracking of the desired queue length in the proposed controller, PI and PID controllers.

Fig. 5 shows that the proposed scheme renders satisfactory tracking of the desired queue length comparing with PI and PID controllers. The PID controller ends in an undesired transient closed-loop response and a very high overshoot. Also, the PI controller renders unsatisfactory performance comparing with the proposed scheme since the desired queue length significantly fluctuates. Fig. 6 shows the packet drop probability in the proposed controller, PI and PID controllers. Also, the congestion window size is shown in Fig. 7.

Fig. 6. Packet drop probability in the proposed controller, PI and PID controllers

Fig. 7. Congestion window size in the proposed controller, PI and PID controllers
The packet round-trip time in the proposed controller is less than that of the PI and PID controllers. The congestion window size in Fig. 7 and RTT in Fig. 8 demonstrate that the proposed method outperforms PI and PID schemes. The performance of the proposed controller, PI and PID controllers are compared in Table II.

**Table II. Performance of The Proposed Controller, PI and PID Controllers**

<table>
<thead>
<tr>
<th>Scheme</th>
<th>Performance metrics</th>
</tr>
</thead>
<tbody>
<tr>
<td>PI</td>
<td>Average queue length error</td>
</tr>
<tr>
<td>5427.2</td>
<td>3395.3</td>
</tr>
<tr>
<td>PID</td>
<td>Average congestion window size error</td>
</tr>
<tr>
<td>6.2886</td>
<td>5.1880</td>
</tr>
<tr>
<td>0.0617</td>
<td>0.0591</td>
</tr>
<tr>
<td>3.6×10^-4</td>
<td>3.1167×10^-4</td>
</tr>
<tr>
<td>The proposed controller</td>
<td>Average packet drop probability</td>
</tr>
<tr>
<td>0.2360</td>
<td>0.594</td>
</tr>
</tbody>
</table>

Table II demonstrates that the proposed controller outperforms PI and PID controllers. The PI controller has the highest mean queue length error, however, the PID controller outperforms the PI controller with less packet drop probability and less congestion window size. Also, the PID controller has the worst packet round-trip time and the proposed scheme renders the best queue length tracking.

**VI. Conclusions**

In this study, the sliding mode control scheme is presented for desired queue length tracking in TCP-based computer network. In this regard, the proportional-integrator-derivative and proportional-derivative sliding surface is defined and the control signal is extracted. Also, the Lyapunov stability conditions are satisfied. The proposed scheme renders satisfactory performance in the presence of disturbance. Also, no chattering occurs in the control signal and satisfactory transient response is gained.

The comparison between the proposed method with PI and PID control schemes shows the superiority of sliding mode control comparing with the aforementioned controllers where congestion window size remains unchanged, satisfactory control signal is achieved and less overshoot is obtained.

**References**


Shoorangiz Shams Shamsabad Farahani obtained her B.Sc., M.Sc. degrees, and the PhD in Electrical engineering in 2001, 2005 and 2013, respectively. She is an Assistant Professor in Islamic Azad University, Islamshahr branch, Islamshahr, Iran. Her current research interests include switching systems, communication networks, wireless sensor networks, complex systems, stability analysis, fuzzy control, robust control and nonlinear control. She has supervised and co-supervised several graduate students in these areas.

Narges Masoumabadi obtained her B.Sc. and M.Sc. degrees in Electrical engineering in 2010 and 2015, respectively. Her current research interests include complex systems, communication networks, robust control and nonlinear control.

Mohammad Reza Jahed-Motlagh obtained his B.Sc., M.Sc. degrees, and the PhD in Electrical engineering in 1978, 1986 and 1990, respectively. He is a full Professor in Iran University of science and Technology, Tehran Iran. His current research interests include control engineering, complex systems, wireless sensor networks, robust control and nonlinear control. He has supervised and co-supervised several graduate students in these areas.