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ARTICLE

Genetic Algorithm Optimization Model for Determining the Probability of Failure on Demand of the Safety Instrumented System

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ABSTRACT

A more accurate determination for the Probability of Failure on Demand (PFD) of the Safety Instrumented System (SIS) contributes to more SIS reliability, thereby ensuring more safety and lower cost. IEC 61508 and ISA TR.84.02 provide the PFD determination formulas. However, these formulas suffer from an uncertainty issue due to the inclusion of uncertainty sources, which, including high redundant systems architectures, cannot be assessed, have perfect proof test assumption, and are neglected in partial stroke testing (PST) of impact on the system PFD. On the other hand, determining the values of PFD variables to achieve the target risk reduction involves daunting efforts and consumes time. This paper proposes a new approach for system PFD determination and PFD variables optimization that contributes to reduce the uncertainty problem. A higher redundant system can be assessed by generalizing the PFD formula into KooN architecture without neglecting the diagnostic coverage factor (DC) and common cause failures (CCF). In order to simulate the proof test effectiveness, the Proof Test Coverage (PTC) factor has been incorporated into the formula. Additionally, the system PFD value has been improved by incorporating PST for the final control element into the formula. The new developed formula is modelled using the Genetic Algorithm (GA) artificial technique. The GA model saves time and effort to examine system PFD and estimate near optimal values for PFD variables. The proposed model has been applied on SIS design for crude oil test separator using MATLAB. The comparison between the proposed model and PFD formulas provided by IEC 61508 and ISA TR.84.02 showed that the proposed GA model can assess any system structure and simulate industrial reality. Furthermore, the cost and associated implementation testing activities are reduced.

1. Introduction

Standards and regulations began being issued following catastrophic industrial accidents that occurred during the second half of the last century.

The target of these standards and regulations has been to obtain safeguarding methods for the prevention and mitigation of risks associated with the industrial process. In 1974, 28 people died and 36 were injured in Flixborough, UK.

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^[1]. In 1976, 17000 people were affected by dioxin release, 200 were poisoned, and 600 were evacuated ^[2]. In 1987, 167 people died in Piper Alpha, UK ^[3]. Consequently, the standards IEC 61508 ^[4], IEC 615011 ^[5], and ISA-TR.84.00.02 ^[6] were issued, which are considered the most recent and widely accepted standards. Moreover, SIS is defined as a protection layer in the mentioned standards, which contributes to reduce the risk posed by industries including potential hazards such as those in the oil and gas industry.

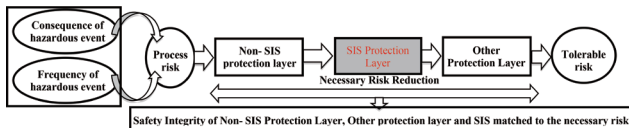


Figure 1. Flow chart shows SIS contribution to risk reduction

The function of SIS is to continuously monitor the process parameters, and in case any abnormal deviation occurs, it performs a predetermined action to return the process to its safe state. As shown in Figure 1, the SIS layer is considered as one of the most critical protection layers due to its ability to reduce the overall risk. Moreover, SIS consists of several safety-instrumented functions (SIFs). Each SIF consists of various subsystems (Sensor, Final Control Element, and Logic solver). Each SIF protects against an identified hazard and contributes to reduce the overall risk by a risk reduction Factor (RRF) or its inverse probability of failure on demand (PFD). Further, RRF identifies SIF associated safety integrity level (SIL).

The standards ^[4,5] provide a framework for establishing the overall requirements and technical activities related to the safety life cycle of SIS. As shown in Figure 2, the safety life cycle consists of three phases (analysis, realization, and operation), each of which consists of several steps. This paper is concerned with the PFD formula used for the SIS verification step included in the realization phase.

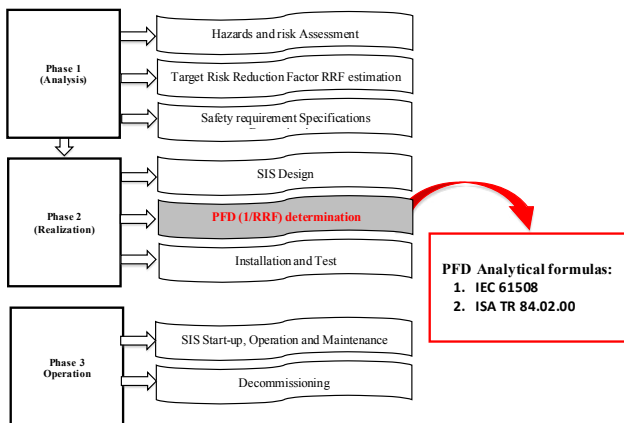


Figure 2. Phases of safety Life Cycle

The conventional PFD determination formulas provided by IEC 61508 and ISA-TR.84.00.02 suffer from the uncertainty problem as demonstrated in ^[7]. This problem is due to assumptions, approximation and limitation contaminations. As an example of the formula limitations, only 1oo1, 1oo2, 2oo2, 2oo3, and 2oo4 architectures can be examined. Therefore, there is no capability for a higher redundant system structure examination ^[8]. Additionally, the failure rates (λ) are assumed as being constant. Another impractical assumption is the perfectness of the proof test as it is assumed to reveal all undetected dangerous failures neglecting the effectiveness of the test procedures ^[9]. Further, the CCF data estimation methodology is based on the checklist questions, which involve operational, environmental, and human factors that may be difficult to be accurately answered ^[7]. However, parameters such as PST and online channel output comparison can reasonably improve the PFD value, which are not included in the formulas ^[10]. Finally, high efforts and significant time are consumed to examine the PFD and determine the values of design parameters (such as proof test interval, main repair time, and main time to restore, etc.). As a result, research contributed to identify, describe the uncertainty sources in conventional PFD formulas, and develop new formulas capable of minimizing the problem. As discussed in ^[7,11], the uncertainty sources in SIS PFD examination are completeness, model, and parameter sources. The authors identified and described the uncertainty sources and ranked the highest importance for completeness uncertainty followed by parameter and model uncertainty. They stressed the uncertainty assessment importance during any SIS reliability examination and demonstrated that analysis results should include the performed uncertainty assessment.

In ^[12], authors reduced one source of uncertainty by developing a new PFD formula considering non-constant failure rate through incorporating the degradation effect within different subsequent proof testing intervals, but the formula is limited to FCE, where PFD examination formulas for LS and SE are not included.

In ^[10,13], authors reduced another source of uncertainty by deriving the PFD formula while incorporating PST. As they presented the impact of PST incorporation on PFD value, the derived formula in ^[10] can examine 1oo1 architecture only and does not include other important variables influencing the PFD such as the β factors.

A Koon generalized formula for PFD presented is in ^[14], where the PFD formula for different system architectures has been derived based on IEC 61508.

Also, Authors in ^[15] proposed a generalized KooN formula for PFD determination based on ISA TR84.00.02. Both formulas have the capability of examining the re-

liability of SIS with high redundancy architecture. But the formula derived in ^[15] includes the contributions from DU failure, DD failures, DD and DU failures combination, and CCF for DU failure and DD failures with binary functions representing the independent failure coefficient or CCFs. The two proposed formulas assumed the perfect tests where the PTC factor was not included.

Another Generalized KooN formula presented in ^[16] considered whether or not the DD failures are repaired. The proposed equation counts for CCF for DU failure and DD failures, along with the perfect proof tests being assumed.

In ^[17], the classification of the non-perfect test and its effect on four IEC61508 formulas was discussed. Accordingly, two models from scientific papers and the PDS method in addition to the PTC impact were examined using the PTC factor, but the human error effect has not been deeply discussed.

The practicality of perfect test assumption has been discussed in ^[18], which involved investigating the reasons of non-perfectness. The investigation resulted in five reasons known as the ‘five Ms’ (Method, Machine, Manpower, Milieu, and Material).

Further, in ^[19,20], the authors discussed the non-perfectness of the proof test, PTC determination, and the factors that may affect it; they demonstrated that perfect PTC assumption is not practical. In addition, different practical considerations can influence PTC. In ^[9], The authors proposed procedures tables for PTC determination; that based on the test procedures.

The above-discussed researches have some merits and provided contributions. However, some drawbacks will be explained in more detail in Section 2.

This paper introduces the optimization model for PFD determination using the GA artificial technique. The GA is a stochastic search technique that guides a population of solutions towards optimum values using the principles of evolution and natural genetics after searching a small portion of the search space ^[21]. Further, the GA model can accurately determine the PFD value and identify the best values for PFD variables in order to achieve the target RRF. Another credit for using GA is time and effort saving. In order to add the capability of higher redundant system examination, a generalized formula has been developed. Moreover, CCF and DD failures have been considered. The PTC factor has been incorporated into the formula to simulate the proof test effectiveness and the PFD value has been improved by incorporating PST for the final control element into the formula. The proposed model has been implemented on the SIS design for crude oil test separator using MATLAB. The model results were

compared with conventional method results for interpretation.

The rest of this paper has been organized as follows:

Section 2 shows the problem areas of conventional PFD determination formulas through the formulas uncertainty assessment.

Section 3 shows the development for the proposed model and implementation procedures.

Section 4 explains the results for the practical case study where the proposed model is implemented.

Finally, the conclusion is given in Section 5.

2. Traditional PFD Determination Formulas and Main Contribution of the Paper

Conventional PFD determination formulas provided in IEC 61508 ^[4], ISA-TR.84.00.02 ^[6] and proposed alternatives still need improvements as they suffer from limitations and drawbacks, such as uncertainty contamination, that can affect SIS design. Further, uncertainty can be defined as something ‘not definitely ascertainable or fixed’ ^[11], which is caused by assumptions, approximations, limitations, lack of understanding, and time and effort consumptions ^[7]. Moreover, PFD cannot be perfectly described since the knowledge of its phenomena is not complete, so uncertainty can be reduced when knowledge about the system increases or when new technology such as Artificial Intelligence can be used.

In this section, the thoroughness of PFD conventional determination methods will be assessed through the uncertainty assessment. Thus, the general sources for uncertainty should be primarily be defined, followed by PFD formula uncertainty assessment being applied.

The general uncertainty sources are as follows:

- (1) Model Uncertainty
- (2) Parameter Uncertainty
- (3) Completeness Uncertainty

Model Uncertainty: is due to simplification, assumption or approximation included in the model, which simulate the system or have different results for the same system when assessed using different models. Another concern is that reducing one uncertainty source may influence another source.

Parameter Uncertainty: is usually, due to the probabilistic and non-probabilistic distributions for available data described the parameter values.

Completeness Uncertainty: is usually due to the inclusion of assumptions, simplifications, and approximation; it can be known or unknown since each type has different causes. The cause of known completeness uncertainty is the non or improper inclusion for known factors that

have influenced the model due to difficulty of estimation, insufficient data availability or lack of understanding of the model. Therefore, it can be reduced by specific and conceptual incorporation for known factors that contribute to the model output; the cause of unknown completeness uncertainty are identified factors with marked contribution for their reduction through the expansion of the searching space, including the indirect factors that may impact the model or by incorporating new intelligent methods or algorithms into the model.

Table 1 illustrates the uncertainty sources for conventional PFD formulas.

Table1. Uncertainty assessment in the conventional PFD formula

1. Model Uncertainty	1.1 Formula Implementation Procedures.
	1.2 Perfect Proof test assumption.
	1.3 Different results obtained for the same system.
2. Parameter Uncertainty	2.1 Constant λ_D assumption.
	2.2 Relative Unavailability of Failure rate data.
	2.3 CCF estimation methodology.
3. Completeness Uncertainty	3.1 Lack of system architectures inclusion.
	3.2 Lack of testing and maintenance strategies' effect on mission time inclusion.
	3.3 Non-inclusion of human error variable.
	3.4 Non-inclusion of the PST effect on the formula.
	3.5 Channel output comparison limited to the 1oo2 structure (1oo2D).

From Table 1:

(1.1) Implementation procedures can increase the model complexity. Further, the complexity increases with completeness uncertainty reduction through incorporating new parameters, and parameter uncertainty may increase by the procedures of estimating the value of input parameters.

(1.2) Perfect proof test assumption does not simulate industrial reality, and the non-perfect proof test must be considered due to the effect of the implementation quality of the test procedures, errors committed by the maintenance crew during the test, test equipment quality, and some inherent conditions for the tests.

(1.3) Different results can be obtained for the same system using different formulas, especially with high DC or long MRT.

(2.1) λ_D is a determination methodology usually determined from general data bases; such data bases are built based on the data from components that were installed several years before the data collection, and at different operational and environmental conditions. The collected failure rate data are based on recorded maintenance strategies, which may not be accurately applied.

(2.2) Failure rate data for new technology devices are often not available.

(2.3) CCF estimation is based on checklist questions that involve operational, environmental, and human fac-

tors that may be difficult to be accurately answered.

(3.1) The formulas are limited to 1oo1, 1oo2, 2oo2, 2oo3, and 2oo4 (only the ISA formula) system architectures where higher redundant architectures such as 3oo5 cannot be examined.

(3.2) Conventional formulas do not include the effect of testing and maintenance strategies on mission time of the subsystems.

(3.3) Conventional formulas do not include any variable that simulates human error although human error is a well-known variable that may increase the probability of failure.

(3.4) PST is not included in the formula, although it can reasonably improve the PFD value.

(3.5) Channel output comparison is limited to the 1oo2 structure named 1oo2D, while applying the online comparison between output channels can reasonably improve the DC factor as it can be applied for any system architecture with redundancy.

Main contributions of this paper

This research contributes to reducing the uncertainty associated with the PFD determination criterion as described below:

Reference to point (1.1) in order to overcome the model complexity, GA will be used to model the formula in addition to widely decreasing the time and efforts required to determine the PFD value and the values of PFD variables. Moreover, it solves the problem of obtaining different results for same system: point (1.3).

Reference to point (1.2) the proof test PTC factor (PTC) will be incorporated into the formula to simulate the effectiveness of the proof tests.

Reference to point (3.1) The newly developed formula will be generalized into the KooN system that can examine high redundant system architectures.

Reference to point (3.4) The PST variable for the final control element will be incorporated into the formula to simulate the effect of PST on system PFD value.

Points (2.1), (2.2), (2.3), (3.3) and 3.5 are out of this research scope and will be considered in the future work.

3. Procedures of the Developed GA-based Optimization Model for the Probability of Failure on Demand

In order to overcome the drawbacks associated with conventional PFD determination methods and to reduce the uncertainty sources in the PFD formula discussed in Section 2, we must develop a new formula that can assess any KooN structure, incorporate all variables that can influence PFD, and finally model the formula using GA to find the optimal solutions for PFD variables as shown in Figure 3.

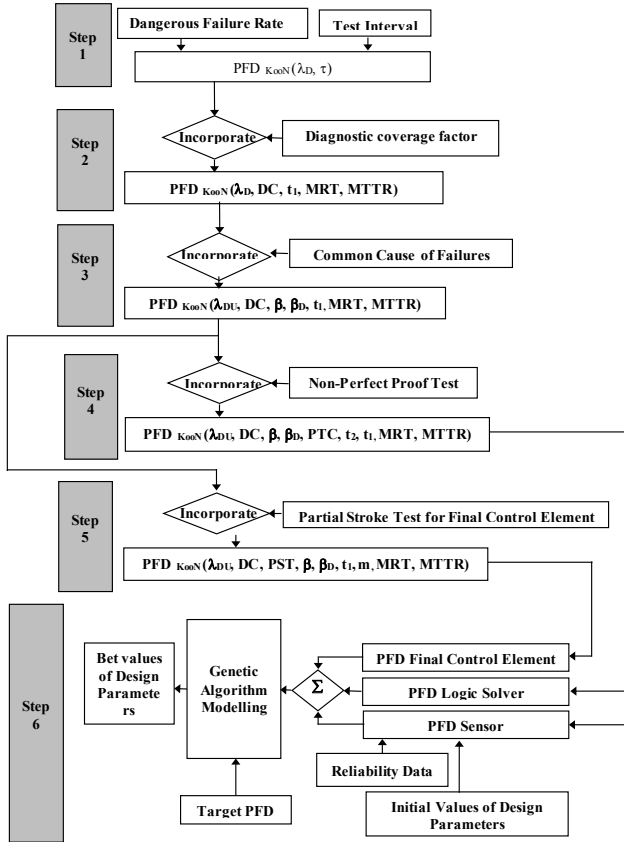


Figure 3. Procedures of the developed GA optimization model for PFD

Step 1: Generalized the formula for KooN structure

KooN expresses the system structure, where N is the number of channels/equipment sets, and K is the number needed to initiate the trip as it describes the connection between subsystem channels/equipment.

The KooN system has minimal cut sets of order $(n - k + 1)$; the average PFD of the kooN system is expressed as:

$$\begin{aligned}
 PFD_{avg} &\approx \int_0^\tau \binom{n}{n-k+1} \cdot (\lambda_D \tau)^{n-k+1} [22,23] \\
 &= \binom{n}{n-k+1} \frac{(\lambda_D \tau)^{n-K+1}}{n-K+2} \\
 &= \frac{N!}{(k-1)! \times (N-K+1)!} \times \frac{(\lambda_D (\lambda_D \tau))^{n-K+1}}{n-K+2} \\
 &= \frac{N(N-1)(N-2)\dots(K+2)(K+1)K}{(N-K+2)!} \times (\lambda_D (\lambda_D \tau))^{n-K+1} \\
 &= \prod_{i=K}^N \left[\frac{(i \lambda_D (\lambda_D \tau))^{n-K+1}}{N-i+2} \right] \\
 PFD_{KooN} &= \prod_{i=1}^{N-K+1} \left[\frac{(N-i+1) \lambda_D \tau}{i+1} \right] \quad (1)
 \end{aligned}$$

The system structure is identified as per the fault tolerance requirement based on the following factors:

(1) The determined safety fractional factor

$$(SFF) = \frac{\sum \lambda_S + \sum \lambda_{DD}}{\sum \lambda_S + \sum \lambda_{DD} + \sum \lambda_{DU}} [4]$$

(2) Device type (A or B)

(3) Determined target SIL

Step 2: Incorporating the DC factor into the formula.

Many dangerous failures of modern safety devices may be revealed by diagnostic self-testing of the portion of failures detected by a diagnostic called Detected Dangerous (DD) failures with a dangerous detected failure rate λ_{DD} ; the diagnostic testing is assumed to be carried out frequently enough for the failures to be revealed immediately. In subsystems with redundant elements, a failure may sometimes be repaired while the subsystem is online and can perform its safety function. In other cases, the subsystem must taken off-line to repair the failure; the mean downtime of the subsystem to repair the failure of an item in the subsystem that has been revealed by diagnostic self-testing is known as mean time to restoration (MTTR). The remaining portion of dangerous failure rate is considered as undetected dangerous (UD) failure rate, λ_{DU} which is supposed to be revealed by a periodic manual proof test with test interval t_1 . When a failure is detected by manual proof test, the subsystem has to be taken offline to be repaired with the mean repair time MRT .

Thus, the dangerous failure rate λ_D is the sum of the undetected failure rate λ_{DU} revealed by a periodic manual test with test interval t_1 and mean repair time MRT in addition to detected failure rate λ_{DD} revealed by diagnostic self-testing with mean time to restoration $MTTR$,

$$\lambda_D = \lambda_{DD} + \lambda_{DU} [4] \quad (2)$$

DC is the measure of the effectiveness of the self-diagnostic test expressed by the fraction of dangerous failures detected by self-dangerous to total dangerous failures,

$$DC = \frac{\sum \lambda_{DD}}{\sum \lambda_D} [4, 6]$$

$$\lambda_{DD} = DC \lambda_D \quad (3)$$

$$\lambda_{DU} = (1-DC) \lambda_D \quad (4)$$

From Eqs. 2, 3 and 4 in Eq.1;

$$PFD_{KooN} = \prod_{i=1}^{N-K+1} \left[(N-i+1) \left((1-DC) \lambda_D \left(\frac{t_1}{i+1} + MRT \right) + DC \lambda_D (MTTR) \right) \right] \quad (5)$$

Step 3: Incorporating CCF into the formula

An additional PFD arises where more failures may occur due to a common cause (CC). Thus, the total dangerous failure rate equals the sum of independent dangerous failure rate in addition to the common cause failure rate;

$$\lambda_D = \lambda_{Ind,D} + \lambda_{CC}^{[24,25,26]} \quad (6)$$

Where:

$\lambda_{Ind,D}$ is the independent dangerous failure rate, and

λ_{CC} is the common cause failure rate.

The fractional of CCF rate is defined by the beta factor (β),

$$\beta = \lambda_{CC} / \lambda_D \quad (7)$$

The β factor provides the fraction of undetected dangerous failures with a CC while the β_D provides the fraction of detected dangerous failures that have a CC; thus:

$$PFD_{KooN-total} = (1-\beta) PFD_{KooN} + \beta PFD_{IooI}^{[22]} \quad (8)$$

From eqs. 6, 7 and 8 in Eq. 5;

$$PFD_{KooN} = \prod_{i=1}^{N-K+1} \left[(N-i+1) \left((1-\beta)(1-DC)\lambda_D \left(\frac{t_1}{i+1} + MRT \right) + (1-\beta_D)(DC)\lambda_D(MTTR) \right) + \beta(1-DC)\lambda_D \left(\frac{t_1}{2} + MRT \right) + \beta_D(DC)\lambda_D(MTTR) \right] \quad (9)$$

Step 4: Incorporating the PTC factor into the formula.

In Eq. 9, the proof test was considered to be perfect for revealing all undetected dangerous failures with the device being regarded new in condition at the end of the test interval. However, practically, the proof test can never be perfect as it depends on the quality of the procedure(s), errors committed by the maintenance crew during the test, test equipment quality, some inherent conditions for the tests, and the inherent features of the system itself. Moreover, device condition can be considered as a new condition only when major maintenance overhaul has been carried out or when a demand is made at demand period t_2 . PTC is the percentage effectiveness of proof tests to check the existence of undetected danger failures expressed by the fraction of revealed failures during proof test to the non-revealed failures during proof test

$$PTC = \lambda_{DU_{rr}} / \lambda_{DU_{nrr}} \quad (10)$$

$$PFD_{DUKooN} = PTC PFD_{DUKooN}(\lambda_{DU}, t_1, MRT) + (1-PTC) PFD_{DUKooN}(\lambda_{DU}, t_2, MRT) \quad [27, 28, 29, 30, 31, 32] \quad (11)$$

From eqs. 10 and 11 in Eq. 9;

$$PFD_{KooN} = \prod_{i=1}^{N-K+1} \left[(N-i+1) \left((1-\beta)(1-DC)\lambda_D \left(PTC \left(\frac{t_1}{i+1} + MRT \right) + (1-PTC) \left(\frac{t_2}{i+1} + MRT \right) \right) + (1-\beta_D)(DC)\lambda_D(MTTR) \right) + \beta(1-DC)\lambda_D \left(PTC \left(\frac{t_1}{2} + MRT \right) + (1-PTC) \left(\frac{t_2}{2} + MRT \right) \right) + \beta_D(DC)\lambda_D(MTTR) \right] \quad (12)$$

Step 5: Incorporating PST for the final control element

PST for final elements has the advantage of reducing the frequency of full tests to save deferment; thus, now there are two test intervals:

(1) More frequently performed partial closure test (Partial Stroke test) at t_{PST}

Where $t_{PST} = t_1 / m$ and m is the number of partial test intervals per full test interval with lower efficiency than FPT (Low coverage factor PST).

(2) Less frequent full test period ($t_1 = t_{PST} * m$) with higher efficiency than partial test (high coverage factor PTC).

$$PFD_{DUKooN} = PST PFD_{DUKooN}(\lambda_{DU}, \frac{t_1}{m}, MRT) + (1-PST) PFD_{DUKooN}(\lambda_{DU}, t_1, MRT) \quad [33, 34, 35] \quad (13)$$

From Eq. 13 in Eq. 12:

$$PFD_{KooN} = \prod_{i=1}^{N-K+1} \left[(N-i+1) \left((1-\beta)(1-DC)\lambda_D \left(PST \left(\frac{t_1}{m(i+1)} + MRT \right) + (1-PST) \left(\frac{t_1}{i+1} + MRT \right) \right) + (1-\beta_D)(DC)\lambda_D(MTTR) \right) + \beta(1-DC)\lambda_D \left(PST \left(\frac{t_1}{2m} + MRT \right) + (1-PST) \left(\frac{t_1}{2} + MRT \right) \right) + \beta_D(DC)\lambda_D(MTTR) \right] \quad (14)$$

In Eq. 14, failures detected by PST are considered as dangerous undetected failures, knowing that the sum of the test interval and the time to perform the repair of a detected failure is more than the mean time to restoration (MTTR) used to determine the achieved safety integrity for that safety function. Thus, PST does not affect SFF and consequently, it does not affect the architecture constrains.

The final formula to be modelled using GA is the total system probability of failure on demand expressed by the sum of the probability of failure on demand for the three subsystems (sensor, logic solver, and final control element).

$$PFD_{sys} = PFD_s + PFD_{LS} + PFD_{FE}^{[4]} \quad (15)$$

Step 6: Modelling the formula using the GA technique

The target of modelling the formula using GA is to find the best value for variables of PFD developed formula achieving target PFD, satisfying design constraints, and reflecting industrial realities;

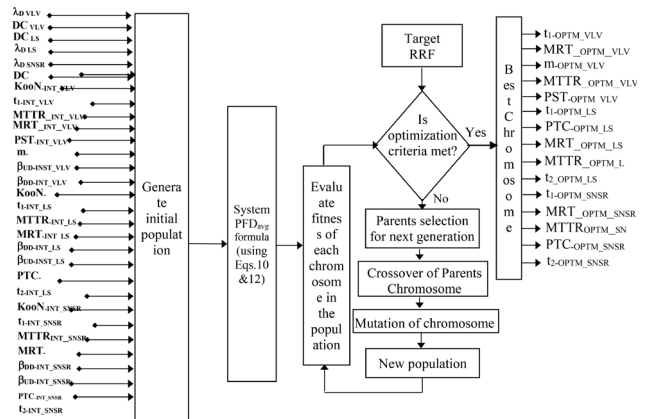


Figure 4. GA Model flowchart

GA uses populations with allowed solutions called 'individuals' that count in the group of parallel algorithms

with constraints, implying that it is necessary to set limits at least for the values of the optimized parameters as shown in Figure 4.

Here, we use the MATLAB GA Toolbox for simulation, where the first step is to define the cost function, the next step is to encode the problem into suitable GA chromosomes, and then construct the population; some works recommend 20 to 100 chromosomes in one population since a higher number of chromosomes will give a better chance to find optimal results. However, due to execution time limitation, 80 chromosomes are used for each generation; each chromosome comprises the model parameters with varied value bounds depending on the cost functions. The population in each generation is represented by the 80 x 31 matrix, with the initial values of parameters as follows:

NL=3; KL=1; TPPTLS=1460; MLS=6; LLS=10; MTTRLS=8; CPTLS=0.8; NV=3; KV=1; TPPTV=356; MV=6; LV=10; MTTRV=8; CPTV=1; CPPTV=0.7; NS1=2; KS1=1; TPPTSN-SR1=514; MSNSR1=6; LSNSR1=10; MTTRSN-SR1=8; CPTSNSR1=0.8; NS2=2; KS2=1; TPPTSN-SR2=730; MSNSR2=6; LSNSR2=10; MTTRSN-SR2=8; CPTSNSR2=0.8.

Each row is one chromosome that comprises the parameter values, and the last column is added to accommodate fitness values (F) of corresponding chromosomes; the final values of parameters are determined by minimizing a certain cost function.

The cost function (CF1) as shown in Eq. (16) minimizes the integrated square error $e(t)$ and improves the overall performance.

$$CF_1 = \int_0^{\infty} (e(t))^2 dt \quad (16)$$

Figures 5A and 5B show the structure of the GA tuning system for cost function. It can be noted that the input for the GA tuning system is the error signal only, which is sometimes not enough to obtain good results.

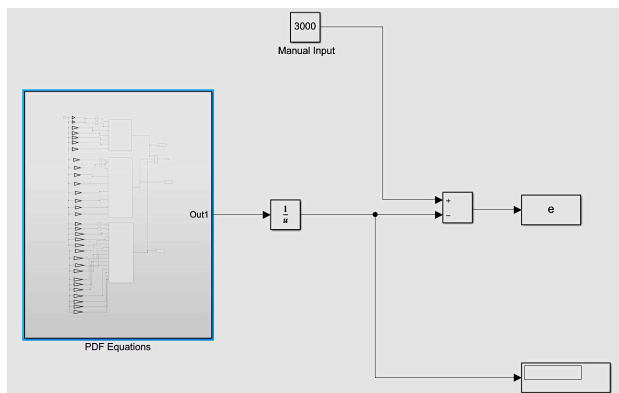


Figure 5A. GA based optimization model using MATLAB

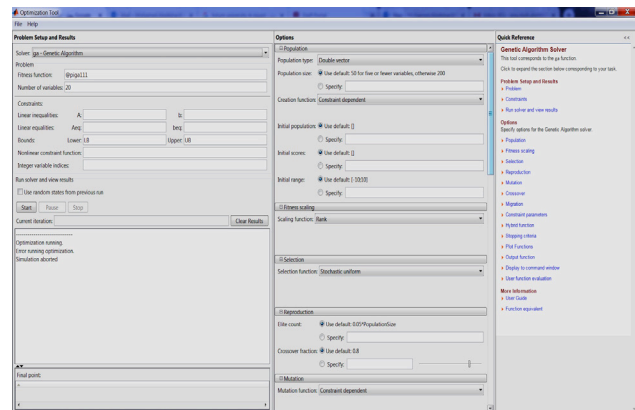


Figure 5B. GA based optimization model configuration on MATLAB

4. Model Implementation and Result Interpretation for Practical Case Study from the Oil and Gas field.

The Mansoura Petroleum company/West Khelala Gas processing plant includes the test separator (V-304) shown in Figure 6 below, where V-304 is a horizontally mounted cylindrical and 2 phase; the gravity separation and size is 300 BBL/D for liquids and 38 MMscfd for gas with a 9m3 surge volume. The produced liquid from the test separator is discharged under level control, LIC-011 using LCV-011 [36].

The Hazard and Operability (HAZOP) Study Report [36] resulted from the HAZOP session prepared for the mentioned unit.

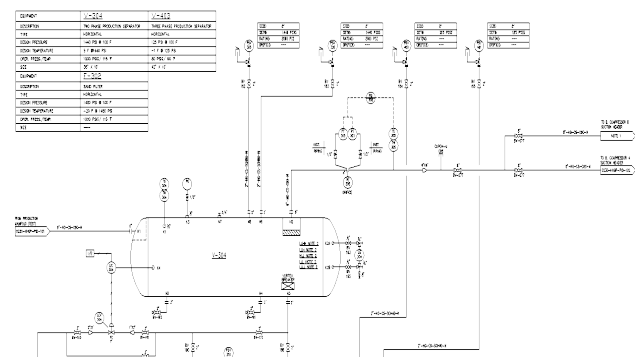


Figure 6. Test Separator P&ID (Model under study) [38]

Further, HAZOP concluded that high-level and high-pressure parable hazards exist and a safely-instrumented function is required for protection with target SIL 3 ($10^{-4} \leq PFD_{avg} \leq 10^{-3}$).

The initial designed SIFs is composed of three subsystems: level transmitter (LT), pressure transmitter (PT), logic solver (LS), and emergency shutdown valve (ESDV). When high pressure or level hazards are detected by the

transmitter, the system should shut the supply source to the separator in order to prevent the event^[37].

The structure of the system is shown in Figure 7 below. Each subsystem has a parall structure. The sensor layer is made up of two identical pressure and level transmitters structured in the 1oo2 architecture.

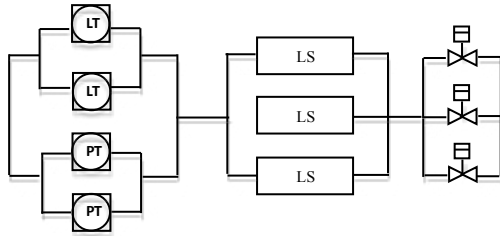


Figure 7: case study initial SIS structure

The logic solver layer (LS) is structured in the 1oo3 architecture and the shutdown valves are structured in the 1oo3 architecture.

Tables 2 and 3 illustrate the variables and corresponding PFD on demand calculations using the conventional and developed formulas.

From Table 3:

(1) The determined system PFD and all subsystem PFDs are almost the same without taking PTC and PST into account.

(2) The determined system PFD and all subsystem

PFDs derived from IEC are almost equal to results derived by the developed formula, as well as the resulting PFDs with PTC being higher than the PFDs without considering the non-perfectness of the proof tests. Moreover, ISA does not have the capability of considering the PTC variable.

(3) The derived formula is the only approach that can count for the PST. Further, PST could decrease the PFD again after increasing due to the PTC consideration.

4.1 Effect of GA:

Table 4 shows the PFD variables and corresponding PFD derived by the GA based model.

From Table 4:

(1) The GA optimization model tended to increase the PST variable for the final control element from 50% to 60% with monthly interval, including what extended the FPT interval T_1 from annual to two years and what decreased the cost, process disturbance, human error, and material degradation associated with manual full FPT.

(2) Increasing the MTTR from 8 Hrs to 48 Hrs ensures enough time and permeability for practical repair activities in case of failure detection.

(3) Decreasing the PTC from 80% to 60%, consequently decreases the associated test procedures and related activities, and the associated cost facilitates the implementation.

Table 2. Design and reliability data

Subsystem	λ_p	K	N	DC	t_1	t_2	m	$\frac{MRT}{MTTR}$	B	β_0	% PTC	% PST
PT	1.90E-06	1	2	51.1	4380	43800	N/A	8	0.1	0.05	80	N/A
LT	7.60E-06	1	2	10	4380	43800	N/A	8	0.1	0.05	80	N/A
LS	3.20E-08	1	3	81.25	8760	87600	N/A	8	0.1	0.05	80	N/A
FE	3.35E-06	1	3	25	4380	N/A	6	8	0.1	0.05	80	50

Table 3. Comparison between resulted PFD using conventional formulas and developed formula

PFD using IEC formula			PFD using ISA formula			PFD using developed formula		
Without PTC	With PTC	With PST	Without PTC	With PTC	With PST	Without PTC	With PTC	With PST
2.09E-04	6.09E-04	N/A	2.09E-04	N/A	N/A	2.09E-04	6.85E-04	6.85E-04
1.87E-03	6.13E-03	N/A	1.80E-03	N/A	N/A	1.75E-03	6.12E-03	6.12E-03
2.68E-06	7.37E-06	N/A	2.66E-06	N/A	N/A	2.64E-06	7.39E-06	2.64E-06
5.53E-04	1.55E-03	N/A	5.60E-04	N/A	N/A	5.53E-04	1.55E-03	5.53E-04
5.56E-04	1.56E-03	N/A	5.63E-04	N/A	N/A	5.56E-04	1.56E-03	3.30E-04

Table 4. Variables and corresponding PFD derived by the GA model

Subsystem	λ_p	K	N	% DC	t_1	t_2	m	$\frac{MRT}{MTTR}$	B	β_0	% PTC	% PST	Target PFD	PFD
PT	1.90E-06	1	2	51.1	8760	43800	N/A	48	0.1	0.05	0.6	N/A	5.60E-04	5.50E-04
LT	7.60E-06	1	2	10	8760	43800	N/A	48	0.1	0.05	0.6	N/A		
LS	3.20E-08	1	3	81.25	17520	87600	N/A	48	0.1	0.05	0.6	N/A		
FE	3.35E-06	1	3	25	8760	N/A	12	48	0.1	0.05	N/A	60		

(1) The credit of the GA's ability to keep the PFD below the target PFD with the above mentioned optimization is due to the incorporated PST and its contribution to improve the PFD and generalized formula for any KooN with implementation practicality credit because of the PTC incorporation.

(2) The presented generalized analytical formula for PFD determination has the capability of assessing any KooN architecture; thus, it can be used in PFD where higher redundancy architectures need to be assessed.

(3) PFD is obtained using the presented formula (excluding PST) is $5.56\text{E-}03$, which is very close to the values obtained using the conventional formula $5.63\text{E-}04$, as it could keep the system PFD below the target PFD despite the above improvement for PFD variables.

4.2 Effect of Generalizing the PFD Formula into KooN:

Table 5 and Figure 8 show the PFD values for the logic solver structured in 1oo1, 1oo2, 2oo2, 2oo3, 1oo3, 2oo4, 2oo5, and 3oo5 system architectures with a variety of test intervals without CC.

From Table 5 and Figure 8:

(1) For 1oo1 system architecture, when the logic solver receives the initiation signal from the sensor, it de-energizes the final control element to trip with $2.64\text{E-}06$ PFD when it is being tested annually.

(2) For the same test interval, 1oo2 is used, and only one of the two logic solvers set needed to initiate the SIF with $9.43\text{E-}10$ PFD. Resultantly, a great drop occurred at the PFD.

(3) For 2oo2, both logic solvers must initiate the SIF when it's important to keep the plant running. Further, it was required to be sure of the event having occurred to initiate the SIF with $5.31\text{E-}05$ PFD. A great change resulted in the PFD as it increased to the double again, due to which this architecture gave an advantage of keeping the plant running and ensured the occurrence of the event in order to trip. However, it increased the PFD.

(4) For 2oo3 (triple marginal redundancy), two logic solvers out of the three set must initiate the SIF with $2.83\text{E-}09$ PFD. A great change resulted in the PFD as it decreased again.

(5) 2oo3 architecture had an advantage over the 2oo2 architecture for the resulting PFD to be much lower than the resulted PFD from 2oo2.

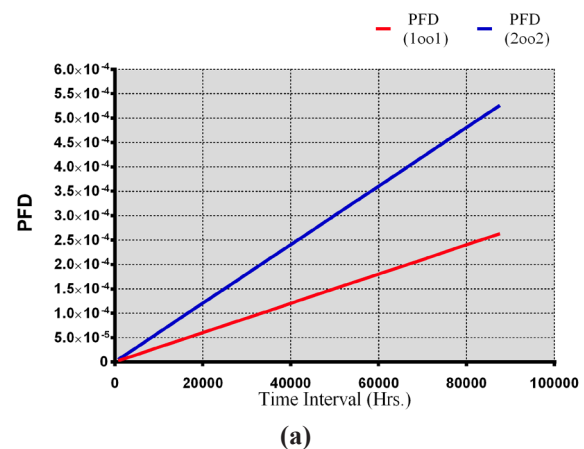
(6) However, the resulting PFD from 1oo2 architecture was lower than the resulting counterpart from 2oo3. Contrastingly, 2oo3 architecture had an advantage over the 1oo2 architecture that kept the plant running and ensured the occurrence of the event in order to trip.

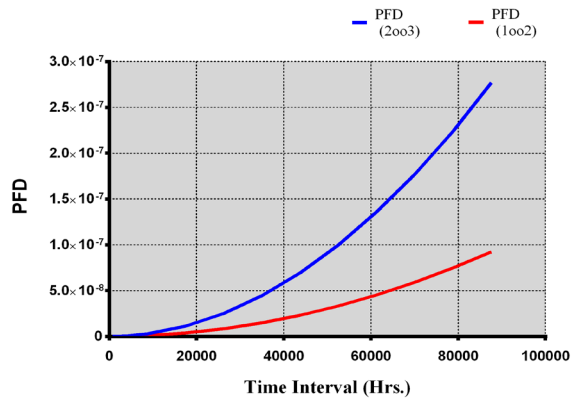
(7) The higher the redundant architecture, the lower the PFD. Therefore, when we decrease the PFD by increasing the redundancy, the PFD examination is not available using the conventional PFD formulas. However, here, we could determine the PFD of 2oo4 ($1.52\text{E-}13$ PFD), 2oo5 ($8.17\text{E-}18$ PFD), and 3oo5 ($3.79\text{E-}13$ PFD) system architectures using the developed formula.

(8) On the other hand, it can be noticed from the curves that for all system architectures, the PFD increases with the increase of test interval.

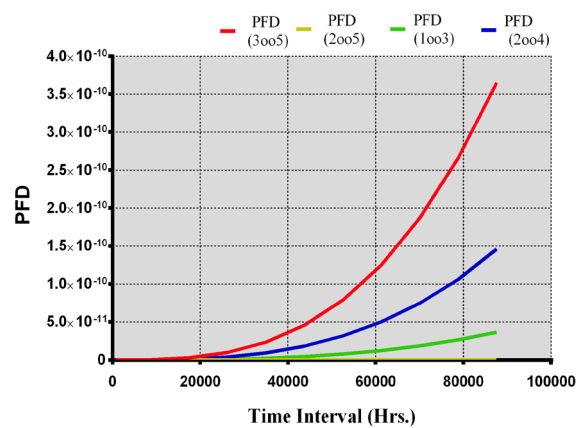
Table 5. Effect of generalizing PFD formula into KooN

T1	PFD (1oo1)	PFD (1oo2)	PFD (2oo2)	PFD (2oo3)	PFD (1oo3)	PFD (2oo4)	PFD (2oo5)	PFD (3oo5)
Month-ly	2.42E-06	8.20E-12	4.83E-06	2.46E-11	3.28E-17	1.31E-16	7.36E-22	3.28E-16
3 months	6.74E-06	6.16E-11	1.35E-05	1.85E-10	6.47E-16	2.59E-15	3.68E-20	6.47E-15
6 months	1.32E-05	2.35E-10	2.64E-05	7.05E-10	4.75E-15	1.90E-14	5.17E-19	4.75E-14
1 year	2.65E-05	9.43E-10	5.31E-05	2.83E-09	3.79E-14	1.52E-13	8.17E-18	3.79E-13
2 years	5.28E-05	3.73E-09	1.06E-04	1.12E-08	2.97E-13	1.19E-12	1.26E-16	2.97E-12
3 years	7.91E-05	8.36E-09	1.58E-04	2.51E-08	9.94E-13	3.98E-12	6.32E-16	9.94E-12
4 years	1.05E-04	1.48E-08	2.11E-04	4.45E-08	2.35E-12	9.39E-12	1.99E-15	2.35E-11
5 years	1.32E-04	2.31E-08	2.63E-04	6.94E-08	4.58E-12	1.83E-11	4.84E-15	4.58E-11
6 years	1.58E-04	3.33E-08	3.16E-04	9.99E-08	7.90E-12	3.16E-11	1.00E-14	7.90E-11
7 years	1.84E-04	4.53E-08	3.68E-04	1.36E-07	1.25E-11	5.01E-11	1.85E-14	1.25E-10
8 years	2.10E-04	5.91E-08	4.21E-04	1.77E-07	1.87E-11	7.48E-11	3.15E-14	1.87E-10
9 years	2.37E-04	7.48E-08	4.74E-04	2.24E-07	2.66E-11	1.06E-10	5.05E-14	2.66E-10
10 years	2.63E-04	9.23E-08	5.26E-04	2.77E-07	3.65E-11	1.46E-10	7.68E-14	3.65E-10





(b)



(c)

Figure 8. Effect of generalizing PFD formula into KooN on PFD.

Figure 8 (a). no fault tolerance; Figure 8 (b). one fault tolerance; Figure 8 (c). two fault tolerance

4.3 The Effect of Incorporating PTC into the Formula

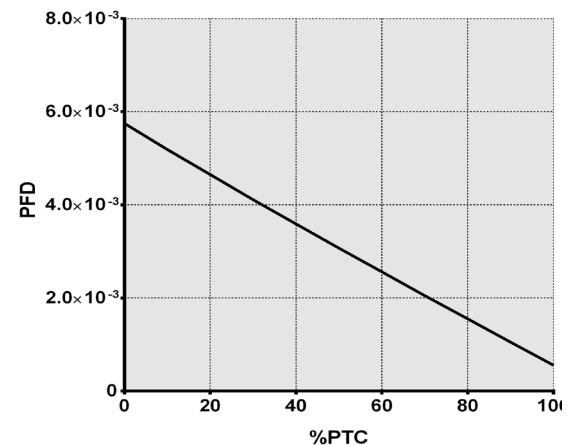
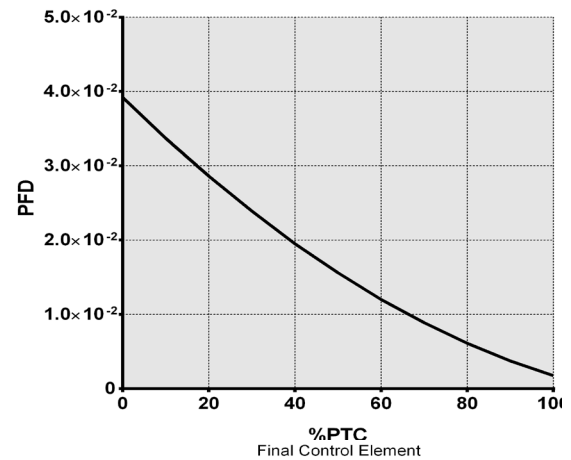
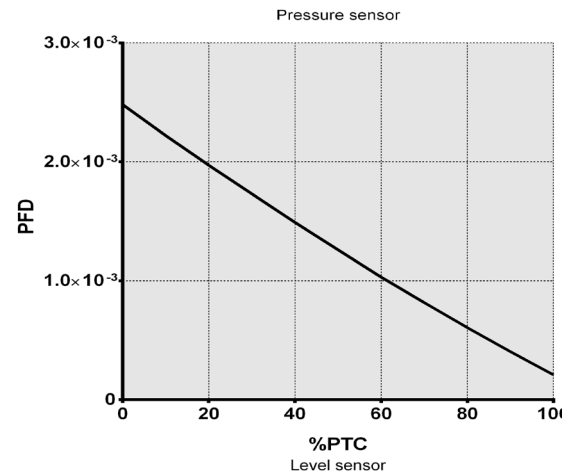
Table 6 and Figure 9 show the impact of the PTC variable on the system PFD and each subsystem PFD.

In Table 6 and Figure 9, when the PTC increases, the PFD decreases while increasing the PTC improves safety by reducing PFD value. However, this consumes time, effort, and money as sticking seals during the full proof test (FPT) is relatively high since it requires flow bypasses, capital, and installation costs.

Table 6. Impact of PTC variable on system PFD and Subsystems PFDs

%PTC	Logic solver PFD	Pressure sensor PFD	Level sensor PFD	Final Control Element PFD	System PFD
100	2.64E-06	2.09E-04	1.75E-03	5.53E-04	5.56E-04
90	5.01E-06	4.04E-04	3.73E-03	1.05E-03	1.06E-03

80	7.37E-06	6.06E-04	6.11E-03	1.55E-03	1.56E-03
70	9.74E-06	8.16E-04	8.88E-03	2.05E-03	2.07E-03
60	1.21E-05	1.03E-03	1.20E-02	2.56E-03	2.58E-03
50	1.45E-05	1.26E-03	1.56E-02	3.07E-03	3.10E-03
40	1.68E-05	1.49E-03	1.95E-02	3.59E-03	3.63E-03
30	1.92E-05	1.73E-03	2.39E-02	4.11E-03	4.17E-03
20	2.16E-05	1.97E-03	2.86E-02	4.65E-03	4.73E-03
10	2.39E-05	2.22E-03	3.37E-02	5.19E-03	5.29E-03
0	2.63E-05	2.48E-03	3.92E-02	5.75E-03	5.87E-03



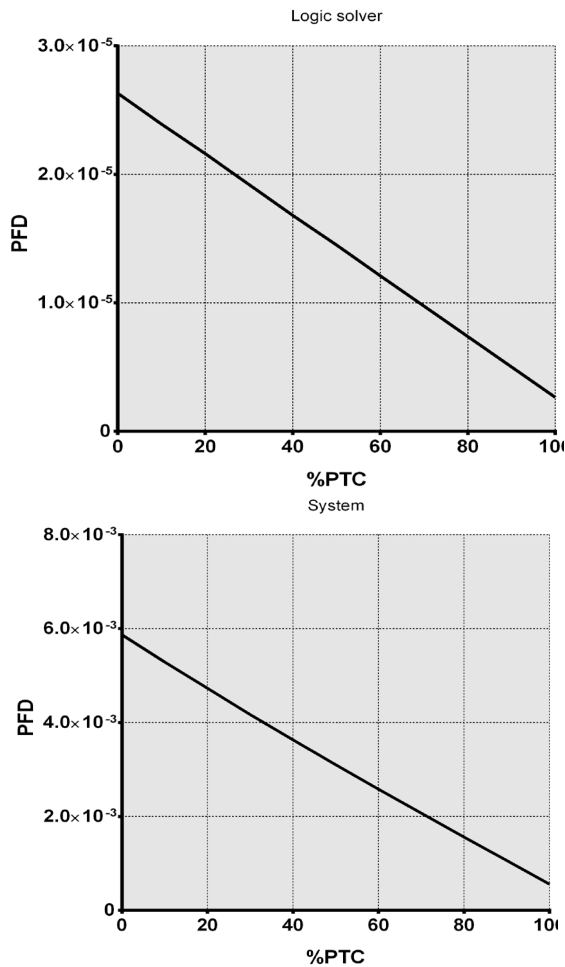


Figure 9. The impact of the PTC variable on system PFD and subsystems PFD

From Table 7 and Figure 10, the behaviour of the curves shows the system PFD with variety test intervals demonstrated through 80% PTC system PFD, which is higher than the system PFD with perfect proof test assumption. However, it reflects the industrial reality and must be considered due to the dependency on the quality of the procedure(s), errors committed by the maintenance crew during the test or repairs required, test equipment quality, some inherent conditions for the tests, and inherent features of the system itself.

Table 7. System PFD with 80% PTC to system PFD with perfect proof test assumption

T1	System PFD With Perfect proof test	System PFD with 80% PTC
6 months	5.55E-04	1.56E-03
1 Year	1.11E-03	3.15E-03
2 years	2.23E-03	6.66E-03
3 years	3.39E-03	1.09E-02
4 years	4.59E-03	1.61E-02

5 years	5.86E-03	2.29E-02
6 years	7.20E-03	3.16E-02
7 years	8.65E-03	4.27E-02
8 years	1.02E-02	5.66E-02
9 years	1.19E-02	7.39E-02
10 years	1.37E-02	9.50E-02

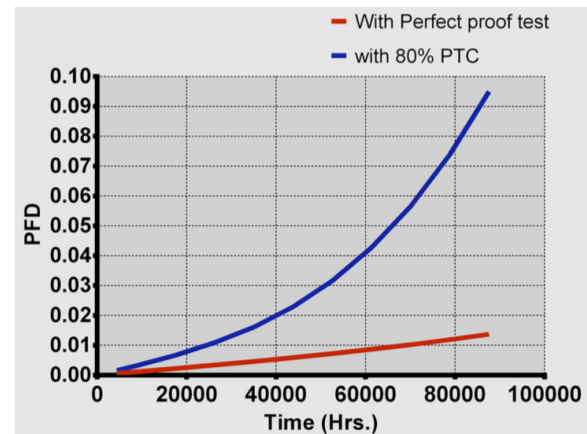


Figure 10. System PFD with 80% PTC to system PFD with perfect proof test assumption.

4.4 The Impact of Incorporating PST Into the Formula

Table 8 and Figure 11 show the effect of PST on the system PFD and the final control element PFD.

Further, Table 9 and Figure 12 show the effect of the number of partial stroke tests m on the system PFD and the final control element PFD.

From tables 8 and 9 as well as figures 11 and 12, when the PST increases, the PFD decreases. The incorporation of PST into the formula for the final element only reasonably decreased the PFD as well as increased the PST and/or decreased the partial test interval m , which can achieve further reduction of PFD.

Table 8. The impact of PST variable on system PFD and The final control element PFD

%PST	Final Control Element PFD	System PFD
0	5.53E-04	5.64E-04
10	5.07E-04	5.18E-04
20	4.66E-04	4.77E-04
30	4.15E-04	4.26E-04
40	3.69E-04	3.80E-04
50	3.23E-04	3.34E-04
60	2.77E-04	2.89E-04
70	2.36E-04	2.47E-04
80	1.97E-04	2.09E-04

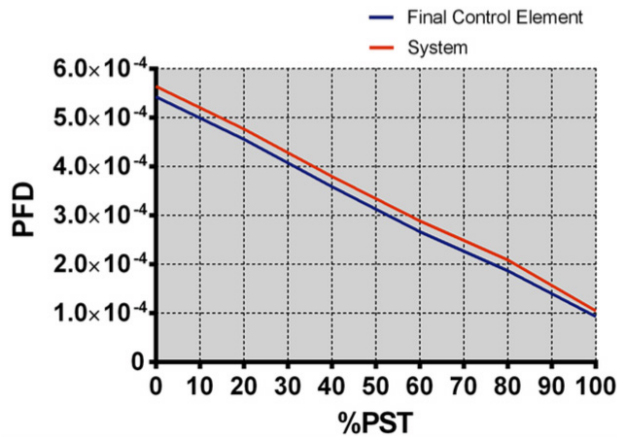


Figure 11. The impact of the PST variable on system PFD and the final control element PFD

Table 9. The impact of partial stroke tests number on system PFD and on Final control element PFD

M	Final Control Element PFD	System PFD
10	1.56E-04	1.67E-04
9	1.61E-04	1.72E-04
8	1.67E-04	1.78E-04
7	1.75E-04	1.86E-04
6	1.86E-04	1.97E-04
5	2.00E-04	2.12E-04
4	2.22E-04	2.34E-04
3	2.59E-04	2.70E-04
2	3.33E-04	3.44E-04
1	5.53E-04	5.64E-04

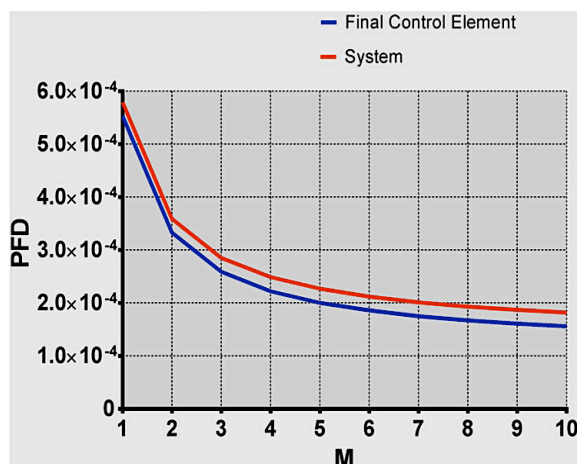


Figure 12. The impact of the m variable on system PFD and the final control element PFD.alue

Table 10 and Figure 13 show the system PFD curve without partial stroke testing and PFD curve for the system when 80% effective partial stroke tests were applied 6

times per full test interval.

Table 10. System PFD with 80% PST and 6 M to system versus PFD without partial stroke testing

T1	System PFD without PST	System PFD with 80% PST & 6M
6 months	1.56E-03	1.97E-04
1 year	3.15E-03	3.80E-04
2 years	6.66E-03	7.48E-04
3 years	1.09E-02	1.12E-03
4 years	1.61E-02	1.49E-03
5 years	2.29E-02	1.86E-03
6 years	3.16E-02	2.23E-03
7 years	4.27E-02	2.61E-03
8 years	5.66E-02	2.99E-03
9 years	7.39E-02	3.37E-03
10 years	9.50E-02	1.97E-04

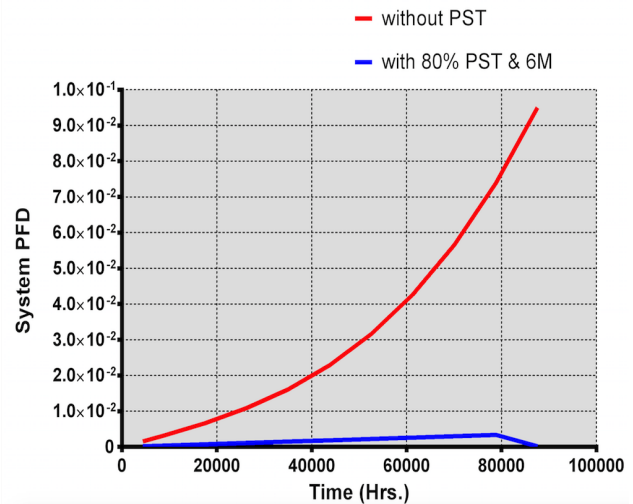


Figure 13. system PFD with 80% PST and 6 M to system PFD without partial stroke testing

From Table 10 and Figure 13, the system PFD was reasonably reduced when partial stroke tests were applied for the final control element with 80% PST 6 times per full test interval. Thus, the system safety increased and cost reduced as well due to the extension of the full proof test interval.

5. Conclusion

The proposed model in this paper can assess any KooN architecture, which is not limited to (1oo1, 1oo2, 2oo2, 2oo3, 1oo3, and 2oo4) architectures contributing to reduce the model uncertainty. Moreover, all known terms that influence the PFD values have been included, investigated, and explained; the presented formula has the capability

of determining PFD for SIS operating in the low demand mode such that it is used in the oil and gas industry but not capable of the same task at a high complexity time, which is dependent on safety systems. Moreover, the presented PFD formula provides a wider interpretation of systems with non-perfect proof tests, as the results showed reasonable reduction in PFD with the increase of PTC and/or test frequency. Further, incorporating PST into the PFD formula for the final control element improves safety by reducing the PFD value since part of DU failures detected and repaired within a shorter time interval than the full test interval. In addition, increasing the PST and/or decreasing the stroke test interval can achieve further reduction in PFD. Moreover, cost is reduced by extending the full test interval; sticking seals during PST are less than those during FPT, thereby decreasing full flow bypasses and reducing engineering, capital, and installation costs. Consequently, the failures detected by PST are considered as dangerous undetected failures, knowing that practically, t_1 +MRT of a detected failure cannot be eliminated to few minutes or hours may exceed MTTR used to determine the achieved safety integrity for that safety function. Therefore, PST does not affect SFF and consequently, it does not affect the architecture constraints as it also contributes to reducing the completeness uncertainty.

The presented PFD formula has been incorporated into the GA model for formulating optimal design of SIS in order to achieve the required RRF. The efficiency has been realized numerically in the practical case study. Moreover, it saved effort, time and cost and facilitated assessing systems with higher complexity that contributed to reducing the model uncertainty.

Data Availability

The data used to support the findings of this study are included within the article.

Conflicts of Interest

The author declares that there is no conflict of interest regarding the publication of this paper.

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Abbreviations

CCF: Common Cause Failures
FCE: Final Control Element
GA: Genetic Algorithm
HAZOP: Hazard and Operability Study

IEC: International Electrotechnical Commission
ISA: The Instrumentation, Systems, and Automation Society
LS: Logic Solver
LT: Level Transmitter
PFD: Probability of Failure on Demand
PT: Pressure Transmitter
PST: Partial Stroke Testing
RRF: Risk Reduction Factor
SIF: Safety Instrumented Function
SIL: Safety Integrity Level
SIS: Safety Instrumented System

Symbols

λ_D : Dangerous Failure Rate (Per Hr)
 λ_{DD} : Detected Dangerous Failure Rate (Per Hr)
 λ_{DU} : Undetected Dangerous Failure Rate (Per Hr)
 λ_{CC} : Common Cause Failure Rate (Per Hr)
 β : Undetected Failures Common Cause Factor
 β_D : Detected Failures Common Cause Factor
DC: Diagnostic Coverage Factor
K: Number needed to initiate the trip
N: Number of channels/equipment sets
m: Number of Partial Stroke Tests Per Full Proof Test Interval
MRT: Mean Repair Time (Hrs)
MTTR: Mean Time to Restoration (Hrs)
 t_1 : Proof Test Interval (Hrs)
 t_2 : Intervals Between Demands (Hrs)
PTS: Partial Stroke Test Coverage Factor
PTC: Proof Test Coverage

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ARTICLE

Adaptive Noise Cancellation Algorithms Implemented onto FPGA-Based Electrical Impedance Tomography System

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ABSTRACT

Electrical Impedance Tomography (EIT) as a non-invasive of electrical conductivity imaging method commonly employs the stationary-coefficient based filters (such as FFT) in order to remove the noise signal. In the practical applications, the stationary-coefficient based filters fail to remove the time-varying random noise which leads to the lack of impedance measurement sensitivity. In this paper, the implementation of adaptive noise cancellation (ANC) algorithms which are Least Mean Square (LMS) and Normalized Least Mean Square (NLMS) filters onto Field Programmable Gate Array (FPGA)-based EIT system is proposed in order to eliminate the time-varying random noise signal. The proposed method was evaluated through experimental studies with biomaterial phantom. The reconstructed EIT images with NLMS is better than the images with LMS by amplitude response $AR = 12.5\%$, position error $PE = 200\%$, resolution $RES = 33\%$, and shape deformation $SD = 66\%$. Moreover, the Analog-to-Digital Converter (ADC) performances of power spectral density (PSD) and the effective number of bit $ENOB$ with NLMS is higher than the performances with LMS by $S_i = 5.7\%$ and $ENOB = 15.4\%$. The results showed that implementing ANC algorithms onto FPGA-based EIT system shows significantly more accurate image reconstruction as compared without ANC algorithms implementation.

1. Introduction

Electrical Impedance Tomography (EIT) employs multi-frequency impedance measurement within several electrodes that attached on the periphery of the dielectric object in order to reconstruct the conductivity distribution. This conductivity distribution represents a useful meaning that varies with the application of EIT such as an anomaly functional biological tissues in the medical imaging applications or a multiphase flow

visualization in the process imaging applications. In terms of EIT hardware design type, Field Programmable Gate Array (FPGA)-based systems are continuously gaining interest as compared to programmable digital signal processor (PDSP)-based systems due to their flexibility and higher bandwidth. Moreover, the FPGA-based EIT system allows us to take high-speed impedance measurement in a multi-electrode sensor with a sweep frequency which suitable with the required performance

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of the EIT system for medical imaging such as cell sensing in micro-channel^[1], thrombus formation in blood flow^[2], and for process imaging such as multi-phase flow^[3], characterizations of Lithium-Ion battery^[4].

In practical EIT applications, the noise signal sources mainly behave as a time-varying random noise, which is difficult to overcome by the stationary-coefficient based filters^[5-9]. In FPGA-based EIT system, the readout chain in the architecture of noise cancellation, the electronic circuit components, and the electronic circuit noise signal levels synergies each other to achieve the SNR of EIT system^[10,11]. Consequently, the architecture of noise cancellation of FPGA-based EIT system should be considered in advance, because of insufficiently high signal-to-noise SNR in EIT system leads to the lack of impedance measurement sensitivity due to the time-varying random noise^[12,13]. The challenge in the development of FPGA-based EIT system for medical imaging is to achieve an acceptable SNR level of as low as 90 dB^[14] by finding a matching the architecture of noise cancellation and the high sensitivity detection.

In order to eliminate the noise signal caused by a time-varying random noise in the FPGA-based EIT system, one of the alternatives is by employing the adaptive noise cancellation (ANC) that can be integrated directly onto FPGA^[15]. As compared with stationary-coefficient based filters such as FFT filters, using ANC provides a more appropriate application in the case of statistical parameters of signals change over time^[16]. Previous investigators of FPGA-based EIT system^[17-23] did not consider the ANC and only applied with a specific architecture of noise cancellation based-on stationary-coefficient based filters by assuming there is no temporal variant environment (leads to the time-varying random noise).

The implementation of ANC onto FPGA-based EIT system is used to discriminate between the sensing signal D_i and the noise signal h which depends on the electronics circuit design and frequently leads to a different architecture of noise cancellation^[24-26]. The ANC automatically adjusts the filter transfer function based on an optimization algorithm. In digital signal processing fields, two different ANC algorithms are popular in the FPGA-based system, i.e., the Least Mean Square (LMS) and the Normalized the Least Mean Square (NLMS). The LMS finds the transfer function coefficients by calculating the least mean squares of the error signal between the output signal Y_i and the sensing signal D_i itself^[27]. On the other hand, the NLMS is developed to overcome the drawbacks of LMS due to the instability of the algorithm in calculating the transfer function coefficient^[28].

In this paper, two different ANC algorithms, i.e., LMS and NLMS are analyzed and implemented onto an FPGA-based EIT system in order to obtain higher accuracy of image reconstruction and the ADC performance parameters (the power spectral density (PSD) and the effective number of bits (ENOB)). The 2D reconstructed images quantitatively under experimental conditions in terms of amplitude response AR , position error PE , resolution RES , and shape deformation SD are also compared.

2. Implementation of Adaptive Noise Cancellation Algorithms

2.1 Architecture on FPGA-Based EIT System

Figure. 1 shows the proposed implementation of ANC algorithms on a field-programmable gate array (FPGA)-based electrical impedance tomography (EIT) system. FPGA-based EIT system is composed of a circular-shape sensor, multiplexer (MUX), signal conditioning, and FPGA. The circular-shape sensor composed of 32 electrodes is attached to a MUX in order to interchange the boundary condition of electrodes, whether as transmitter electrodes (V_{hc} and V_{lc}) or receiver electrodes (V_{hp} and V_{lp}). FPGA consists of a voltage generator, implemented ANC algorithms, and voltmeter. The transmitter electrodes

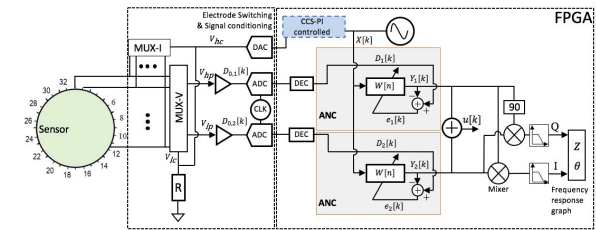


Figure 1. Adaptive filter implementation in FPGA-based EIT system.

(V_{hc} and V_{lc}) are connected to the constant current source (CCS-PI controlled) in order to maintain the constant current inside of the sensor. Meanwhile, the receiver electrodes are connected to ANC algorithms and voltmeter through an analog-to-digital converter (ADC). The ANC algorithms are connected to each receiver electrodes readout.

For each receiver electrodes readout, the converted analog data of $V_{hp}[k]$ or $V_{lp}[k]$ are the sensing signal at receiver electrodes $D_1[k]$ and $D_2[k]$. Practically, the sensing signal D_i contains noise signal h . Where i is the order number of signal. By using the filter tap weights vector W , the output signal Y_i can be calculated, which is the free-noise signal of sensing signal D_i . On this FP-

GA-Based EIT System, the input signal X is the output of the voltage generator. In order to calculate filter tap weights vector W_i , the error signal e_i should be maintained as low as possible which is given by the following relationship

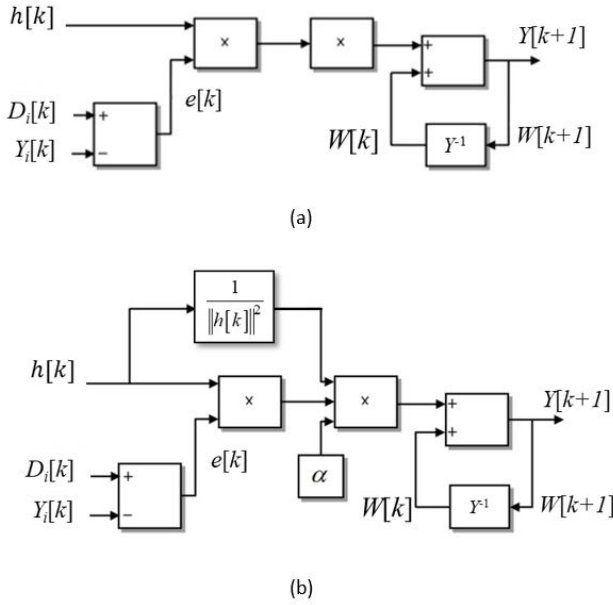


Figure 2. Scheme of an ANC algorithm: (a) block diagram of LMS algorithm, and (b) block diagram of NLMS algorithm

$$\mathbf{e}_i = \mathbf{D}_i - \mathbf{Y}_i \quad (1)$$

The calculation of the filter tap weights vector W itself varies among the proposed ANC algorithms. In this study, the two ANC algorithms are explained in the following sub-section.

2.2 Adaptive Noise Cancellation Algorithms

2.2.1 Least Mean Squares (LMS) Algorithm

In order to minimize the error signal e_i , LMS algorithm is based on a stochastic gradient algorithm that uses a gradient vector of the filter tap weight to converge on the optimal Wiener solution [29]. The block diagram of LMS algorithm within an ANC scheme is shown in Figure. 2(a). It should be noted that the sensing signal D_i induced by excitation voltage has a similar statistical property with the input signal X . Thus, generally, the sensing signal D_i can be expressed by a weighting transformation which applied on the input signal X . With the filter tap weights vector W , it is feasible to train the input signal X and finally obtain the best filtering result of the estimation of the desired signal Y_i . The estimation of the desired signal Y_i contains the noise

signal h as low as possible by implementing the filter tap weights vector W .

In order to achieve the optimal filtering, LMS adjust its filter parameters iteratively when the statistical properties of the sensing signal changes with the following equation [30]:

$$W[k+1] = W[k] + 2\mu e[k] X[k] \quad (2)$$

where, μ [31] is the step size parameter to control the speed of the LMS filter by using weight update for each iteration of ANC, and k is the number of iteration. In order to apply eq. (2) onto eq. (1), a structure of filter within the adaptive system employs a digital finite impulse response (FIR) filter in a time instant n [30].

$$Y_i[n] = \sum_{k=0}^{L-1} W_k[n] * X[n-k] = W^T[n] X[n] \quad (3)$$

$$X[n] = [x(n), x(n-1), \dots, x(n-L+1)]^T \quad (4)$$

$$W[n] = [w_n(0), w_n(1), \dots, w_n(L-1)]^T \quad (5)$$

Where, L_j is the filter length, and j is the value determined from the sequence 2^n of 2, 4, 8, 16, 32, 64, 128, 256, 512, 1024, 2048.

2.2.2 Normalized Least Mean Squares (NLMS) Algorithm

The NLMS algorithm, as shown in Figure 2(b) derived from LMS revises the characteristics of the step size parameter μ [31] in order to solve the LMS filter drawback [31,32]. In NLMS filter, the filter tap weights $W[k+1]$ is normalized with the squared Euclidian norm of the noise signal $h[k]$ as the following equation:

$$W[k+1] = W[k] + \left(\frac{\alpha}{c + \|h[k]\|^2} \right) e[k] X[k] \quad (6)$$

where, α is the NLMS adaptation constant that optimizes the convergence rate of the NLMS filter, and its possible range is $0 < \alpha < 2$. c [31] is time-dependent filter coefficients which are the positive constant term and is always less than 1. In order to compute the norm of the noise signal $h[k]$, eq. (7) is used as follows

$$\|h[k]\| = \sqrt{\sum |h[k]|^2} \quad (7)$$

In the case of the LMS algorithm, a proper value selection of the step size parameter $m^{[-]}$ is important to the performance of ANC. The lower value of step size parameter $m^{[-]}$, the longer time of LMS filter to converge on the optimal Wiener solution. However, the higher value of the step size parameter $m^{[-]}$, the LMS filter becomes unsteady and diverge. Moreover, the noise signal h amplitude is not steady during the change of time due to the instability of the input signal X that leads to the step size parameter $m^{[-]}$ varies in time and affects to the convergence rate ^[31,32]. In this regards, finding the best of the step size parameter $m^{[-]}$ of LMS is troublesome.

3. Experiments

3.1 Experimental Setup and Conditions

As shown in Figure.3, the experimental setup of an FPGA-based EIT system is composed of a high resolution and high speed ADC by implementing the Red-Pitaya platform which has a dual-port (14 bit DAC, 125 MSps) DAC1401D125 from NXP Semiconductors, a dual-ADC's (14 bit ADC, 125 MSps) LTC2145 from Linear Technology, and a FPGA-ZYNQ 7010 System on Chip (SoC) from Xilinx Co.Ltd. This FPGA is chosen because it has potentially sufficient for the implementation of a complete instrument with an ANC and a spectrum analyzer with a portable system. The input stage of FPGA-ZYNQ 7010 has been modified by inserting 1:1 RF balun as an isolation transformer, enabling the acquisition of signals up to 125 MHz in order to increase the bandwidth of measurement and to characterize the jitter effect ^[33]. This FPGA-based EIT system has 32-channels electrode with vessel diameter $d = 125$ mm. The 32-channel electrodes are attached on the active electrode circuit that converts the analog signal transmission between electrode and FPGA system into the digital signal.

In order to eliminate the need of constant current circuit, an artificial constant current source proportional-integral controlled (CCS-PI controlled) is used by implementing a PID controller inside of FPGA (see Figure.1 and Figure. 4(a)). This method could be possible because of the intensity of input signal X and the range resistance of R_{load} which is the electrical properties of an object interest are known priori. Figure. 4(b) shows the comparison of current I_{out} that injected onto the object of interest in the case of variant R_{load} . The result shows that constant current is stable until frequency 35 kHz.

3.2 Phantom Conditions

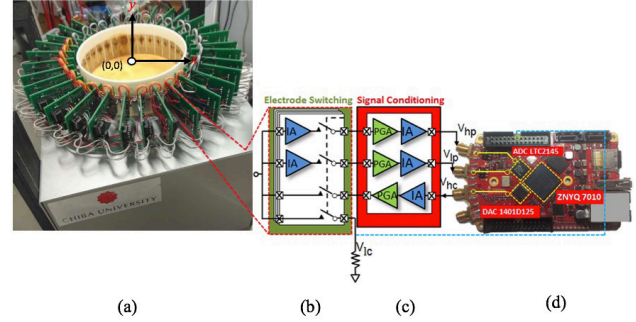


Figure 3. Manufactured of FPGA-based EIT system which consists of (a) 32 active electrodes attached to circular-shaped vessel, (b) MUX-I or MUX-V (electrode switching), (c) Signal conditioning, and (d) FPGA system based on ZYNQ-7010

Two agar phantoms to demonstrate the feasibility of the proposed method were built based on NaCl powder (Wako, Osaka, Japan) and agar powder (Wako). Agar phantom-1 was composed of chicken meat with a diameter of $d_1 = 50$ mm and immersed in agar phantom background with conductivity $\sigma_b = 0.3$ S/m.

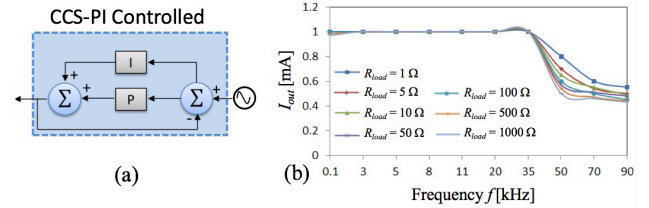


Figure 4. (a) Block diagram of constant current source PI controlled, and (b) constant current at frequency variant.

Agar phantom-2 was composed of one cylinder agar block inclusion with a diameter of $d_2 = 20$ mm and conductivity $\sigma_i = 0.7$ S/m was immersed in agar phantom background with conductivity $\sigma_b = 0.3$ S/m. Image noise analysis is evaluated with input signal $X = 2 V_{p-p}$ and five different positions of cylinder agar inclusions along the x -axis from right to left until the center of the conductive medium.

3.3 Image Reconstruction Evaluation Method

Five different positions of cylinder agar inclusions are investigated by using the image reconstruction performance parameters in EIDORS, which are amplitude response AR , position error PE , resolution RES , and shape deformation SD ^[34]. $AR^{[-]}$ indicates how the inclusion's amplitude contribute to the overall reconstructed image's amplitude by measuring the ratio of image pixel amplitudes between the normalized inclu-

sion's amplitude A_m [1] and the normalized area of inclusion A_{inN} [1]. m is the pixel matrix of the reconstructed image with resolution 64×64 .

$$AR = \frac{\sum_m A_m}{A_{inN}} \quad (8)$$

PE [1] indicates the accuracy of the reconstructed image in detecting the location of inclusion. r_t is the predicted center of the inclusion in the reconstructed image, while r_q is the actual location of the center of the inclusion in the phantoms.

$$PE = r_t - r_q \quad (9)$$

RES [1] indicates the smallest visible object by calculating the ratio of inclusion's area A_{in} [mm²] and total reconstructed image's area A_T [mm²].

$$RES = \sqrt{A_{in} / A_T} \quad (10)$$

SD [1] indicates the fraction of the outside area of inclusion A_c [mm²] in the reconstructed image, which is not covered by the inclusion area A_{in} [mm²]. A_{in} [mm²] is the area of inclusion in the reconstructed image.

$$SD = \frac{A_c}{A_{in}} \quad (11)$$

3.4 Experimental Results and Evaluation

Figure 5 shows the comparison of reconstructed images of agar phantom-1 by using and without adaptive filter algorithms in the case of different input signal amplitude $X = [X, X/2, X/3, X/5]$. The ANC algorithms used Figure 5 was NLMS. Without using the ANC algorithm on FPGA-based EIT system shows the inaccuracy of the reconstructed image as compared with using the ANC.

Figure 6 shows the comparison of reconstructed images of agar phantom-2 based on LMS and NLMS algorithm implementation. It can be seen that the LMS filter algorithm is dominantly contributed by image noise from near the boundary position to the center position of object inclusion. Reconstructed images of NLMS is preferable from the inclusion, which is placed close to the electrodes to the center position. Although NLMS reconstructed images pattern has less tolerant to noise and a lower dynamic range, it performs better when the target is placed close to the electrodes. The image artifact on LMS and NLMS reconstructed images has increased slightly among the different inclusion.

The evaluation of reconstructed images is shown in Figure 7. The reconstructed EIT images with NLMS is better than the images with LMS by amplitude response $AR = 12.5\%$, position error $PE = 200\%$, resolution $RES = 33\%$, and shape deformation $SD = 66\%$. The desired behavior of amplitude response AR is constant in the whole position inside the sensor. The reconstructed images of LMS and NLMS showed similar behavior in the case of AR . In the case of position error PE , reconstructed images of LMS have bigger PE as compared with NLMS. The instability of the sensing signal causes the higher PE of LMS, and it is indicated by high image noise as shown in Figure 6. Meanwhile, in the case of resolution RES , the desired behavior is small and uniform. Both reconstructed images of LMS and NLMS showed similar behavior in the case of RES . Furthermore, this instability sensing signal can be quantified by using the shape deformation SD . As can be seen in the image reconstruction of LMS, it has high shape deformation. The value of shape deformation SD also indicates that the LMS filter algorithm is not properly filtered and still create image artifact in the reconstructed images.

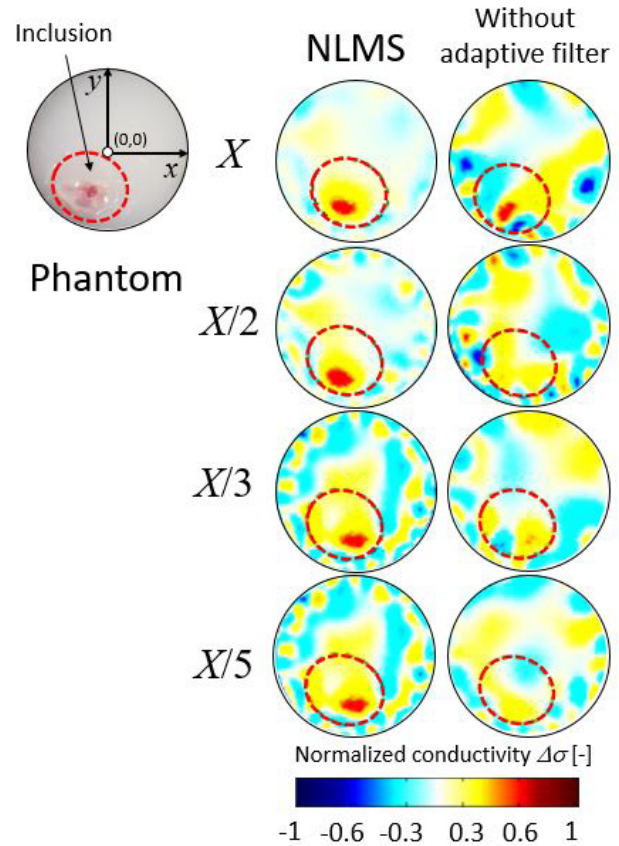


Figure 5. Comparison of reconstructed images between using and without adaptive filter.

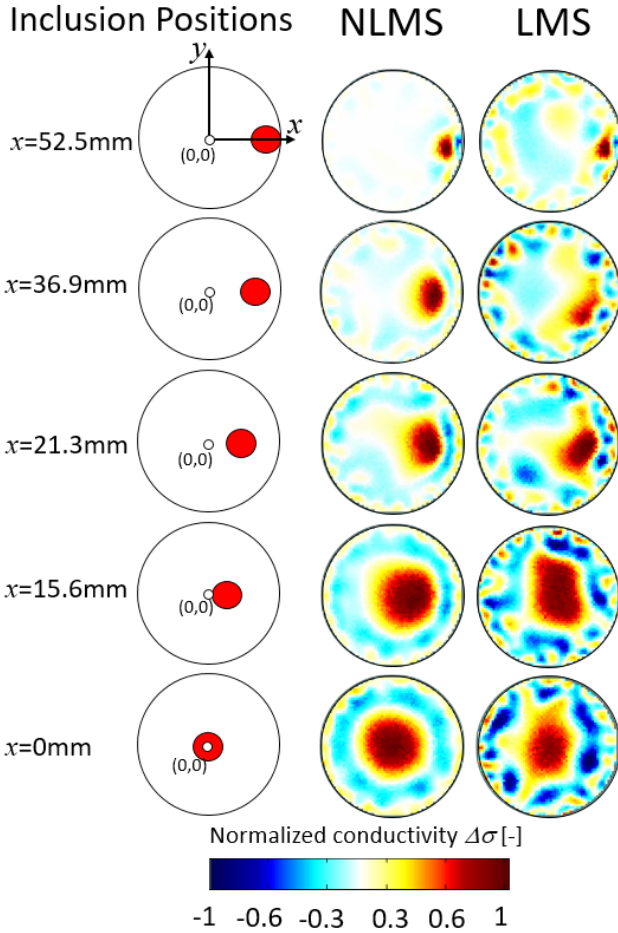


Figure 6. Comparison of reconstructed images.

4. Discussion

4.1 Analysis Method and Condition

We discuss the analysis of ANC algorithms implemented onto FPGA-based EIT system in terms of the capability to reduce noise signal. Both phase and amplitude of sensing signals entering ADC $D_{0,i}[k]$ as shown in Figure.1 have affected by the noise signal as well as input signal leaving DAC $X[k]$. The noise signal sources in FPGA-based EIT system as shown in Figure. 8 consists of the input signal noise $g[k]$, the sample-and-hold (S/H) noise $r[k]$, the voltage reference noise $b[k]$, and the quantizer noise $q[k]$. Based on the dependent response of the input signal variance, the performance of ANC algorithms was investigated.

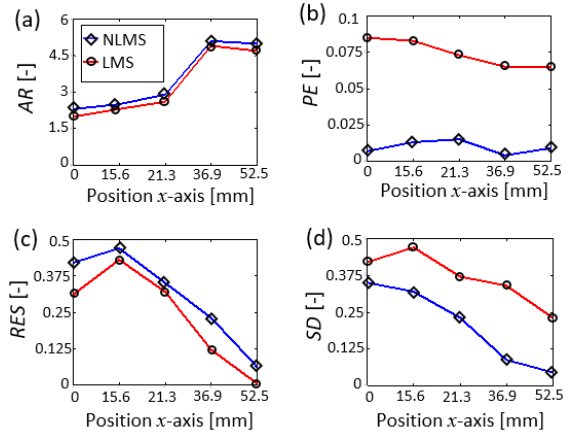


Figure 7. Evaluation of image reconstruction performances: (a) amplitude response AR , (b) position error PE , (c) resolution RES , and (d) shape deformation SD .

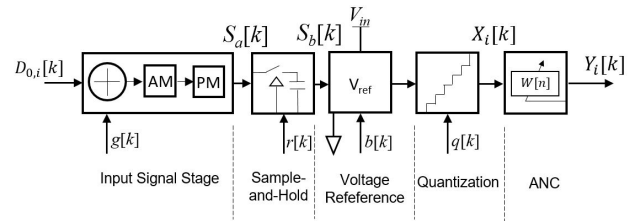


Figure 8. Noise signal sources model on FPGA-based EIT system.

The input signal noise $g[k]$ of an ADC behaves as an additive white noise^{[35], [36]} which mainly affects the amplitude sensing signal. The additive white noise is caused by such as a thermal, and circuit node noise. The contribution of additive white noise can be expressed by

$$s_a[k] = D_{0,i}[k] + g[k] \quad (12)$$

The sample-and-hold (S/H) noise $r[k]$ or aperture jitter affects mainly on the phase sensing signal. Time fluctuations occur in ADC's sample-and-hold which generate a parametric noise. Time fluctuations are also known as aperture jitter which is defined as the variation of the sampling instant at a time kt . Hence, the effect of aperture jitter can be described as follows

$$s_b[k] = s_a(kt + r[kt]) \quad (13)$$

At this stage, $s_b[k]$ is converted into digital form, as described in eq. (15), where m is the resolution of the ADC, V_{ref} is the voltage reference, and $q[k]$ is the quantization noise:

$$s_c[k] = s_b[k] \frac{2^m}{V_{ref}} + q[k] \quad (14)$$

At the voltage reference stage, the voltage reference noise $b[k]$ caused by a non-ideal source from instability voltage over temperature and time. The voltage reference noise $b[k]$ is described as a non-ideal source, $V_{ref} = V_{rb} (1+b[k])$. Here, V_{rb} is the nominal value. This noise behaves as a parametric noise that the character noise depends on the voltage reference topology^[37]. Differential amplifier equation is to solve the constant values of m and V_{rb} . The voltage reference noise as a reference in input ADC can divide the ADC output by $2^m/V_{ref}$. Hence, ADC output is expressed as

$$s_d[k] = s_c[k] \frac{V_{rb}}{2^m} \quad (15)$$

Lastly, the ADC output is stated in terms of the additive and parametric noise are described by

$$X_i[k] = (g[k] + s_b(kt + (r[k]))) (1 + b[k]) \quad (16)$$

The voltage reference noise is considered $b[k] \ll 1$.

Considering the noise signal sources as aforementioned, thus we can analyze how this noise signal sources affect the amplitude and phase sensing signal by the following mathematical relationship

$$PN = \frac{1}{V_0} g[k] + 2\pi V_0 r[k] \quad (17)$$

Where $1/V_0$ is normalization of amplitude fluctuation generated by input signal noise and summing using the conversion of aperture jitter $2\pi V_0 r[k]$ false from analog to digital converter. The unity of PN is radian. Meanwhile, the amplitude signal noise AN is also generated by input signal noise, but with a difference summation that is induced by voltage reference noise $b[k]$:

$$AN = \frac{1}{V_0} g[k] + b[k] \quad (18)$$

The AN is a non-dimensional. Then, the phase and amplitude signal noise can be fitted by polynomial law to compute the power spectrum density:

$$S_Q[f] = \frac{1}{V_0^2} S_g[f] + 4\pi^2 V_0^2 s_r[f] \quad (19)$$

$$S_I[f] = \frac{1}{V_0^2} S_g[f] + S_b[f] \quad (20)$$

Thus, the PSD of phase signal noise $S_Q[f]$ and ampli-

tude signal noise $S_I[f]$ can be calculated by Eq. (19) and Eq. (20) respectively. In this PSD of amplitude and phase signal noise, the contribution of each noise signal sources is identified as a complete description of the device limitations.

4.2 Analysis Results

We realize that the readout of the estimation of the output signal Y_i is available at a single frequency signal. Because the impedance measurements are a sweep frequency measurement, we need to confirm the free noise of sensing signal D_i at the readout of the voltmeter in the spectral region. In this regards, we need to apply the power spectral density PSD analysis. The ANC algorithms of LMS and NLMS are supposed to suppress the noise signal h_i in order to obtain the estimation of the output signal Y_i without the drawbacks from the noise signal. The LMS and NLMS are evaluated through PSD analysis by comparing with PSD of noise signal floor NF . As shown in Figure. 9, the PSD of NF is $NF = -164 \text{ dBV}^2/\text{Hz}$ in the case of ADC 14 bit, primary signal $X = 2 \text{ V}_{pp}$, and the step size parameter $\mu = 0.002$ ^[1]. As closer the PSD of the desired signal to PSD of NF , it indicates as better ANC algorithms performance.

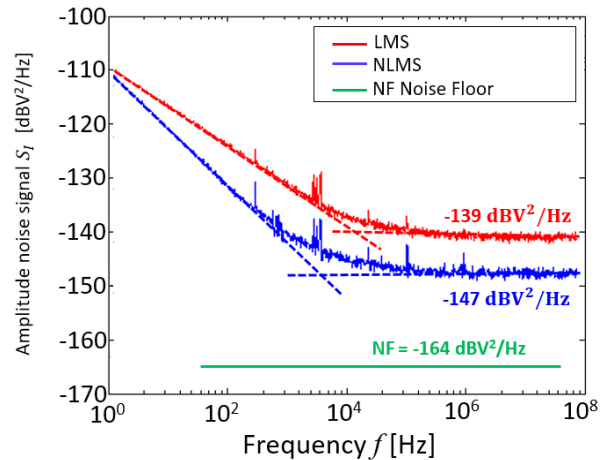


Figure 9. Comparison of PSD of amplitude sensing signal S_I between NLMS, LMS, and Noise floor in the case of phantom as shown in Figure 5

The spectrum analysis of four noise sources has been considered as a common noise of two ADC of receiver electrodes, as shown in Figure. 8. The output of ADC was synchronously sampled by one sample every one signal period in order to measure the influence of noise signals. The sensing signal D_i which is the output of ADC has affected by the noise signal can be analyzed in terms of amplitude and phase signal which were detected by using two points of a cosine wave of $u_i[k]$. These points consist

of zero-crossing detection and voltage peak detection. Zero crossing detection is related to phase signal noise PN , the measurement of PN results in the prediction of the phase noise generated by quadrature noise. Voltage peak detection is related to amplitude signal noise AN , the measurement of AN result in the prediction of the amplitude noise generated by in-phase noise.

Some of the important ADC performance parameters are signal-to-noise ratio SNR , signal-to-noise and distortion ratio $SNDR$, effective number of bits $ENOB$, spur free dynamic range $SFDR$, total harmonic distortion THD , inter modulation distortion IMD and effective resolution bandwidth $ERBW$ [38]. Among these ADC performance parameters, $ENOB$ reflects the resolution and the accuracy of an ideal ADC circuit under consideration in dynamic measurement that is suitable for high sensitivity detection analysis in most EIT applications [39,40].

The contribution of ANC performance increases $ENOB$ performance, and it directly related to SNR measurement. The Analog-to-Digital Converter (ADC) performances of PSD and the effective number of bit $ENOB$ with NLMS is higher than the performances with LMS by $S_I = 5.7\%$ and $ENOB = 15.4\%$. Therefore, high image noise in the LMS adaptive filter indicates the voltage signal with low SNR . Figure. 8 shows the PSD of the voltage noise with LMS and NLMS algorithm that it can be obtained by connecting the combination of two ADC inputs to ground through a $50\ \Omega$ resistor ($X = 2\ V_{p-p}$, $f_s = 125\ \text{MHz}$). This noise corresponds to the amplitude noise induced by the input stage that contains signal instability. As shown in Table I when LMS is applied to the system, it presents an additive white noise of $S_I = -139\ \text{dBV}^2/\text{Hz}$. Meanwhile, NLMS shows better performance in term of the additive white noise value. The additive white noise value of NLMS is $S_I = -147\ \text{dBV}^2/\text{Hz}$. This additive white noise close to the quantization of noise floor $NF = -164\ \text{dBV}^2/\text{Hz}$. From PSD plot of additive white noise, the level of additive white noise is also described for actual $ENOB$ value of ADC. The actual $ENOB$ of LMS algorithm is $ENOB = 9.7\ \text{bit}$. Meanwhile, the actual $ENOB$ of NLMS algorithm increases to $ENOB = 11.2\ \text{bit}$.

Table 1. Comparison of ANC algorithms performance

ANC	$ENOB$	S_I
LMS	9.7 bit	-139 dBV ² /Hz
NLMS	11.2 bit	-147 dBV ² /Hz

Figure. 10(a) and (b) show the comparison of the capability to reduce noise signal between LMS and NLMS in terms of PSD analysis of phase signal noise $PN = S_\phi$ [dBrad²/Hz]. While Figure. 11(a) and (b) show the com-

parison of PSD of amplitude signal noise $AN = S_I$ [dB/Hz]. Four voltage levels of input signal X , $X = 2\ V_{p-p}$ to $X/5 = 0.4\ V_{p-p}$, were used to investigate the PSD level in order to compare the performance between LMS and NLMS.

As shown in Figure. 10(a) and (b), phase signal noise PN are dominated by the additive noise of the input stage, although the aperture jitter also occurs. Furthermore, it increases proportionally to the input amplitude variance $1/V_0^2$. As it can be seen that under this condition, PN is dependent on the input amplitude variance $1/V_0^2$. Additive flicker and additive white noise of NLMS at each input amplitude level are higher than LMS.

The PSD of amplitude signal noise S_I produces the same result that NLMS has higher PSD level as compared with LMS as shown in Figure. 10(a) and (b). In terms AN , the influence of additive noise by harmonic contributions of the signal generator is dominant. The PSD analysis results of PN and AN are appropriate with the noise model, as shown in Eq. (19) and (20). Finally, the capability of reducing the noise of NLMS contributes to increasing the accuracy of the reconstructed image is shown in Figure. 6 and increasing the $ENOB$ that reflects the high sensitivity detection of FPGA-based EIT system as shown in Table 1.

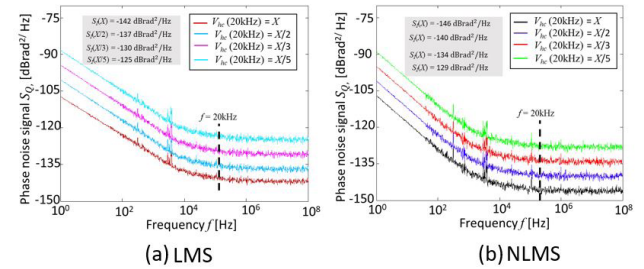


Figure 10. PSD of phase sensing signal of (a) LMS and (b) NLMS in the case homogenous agar condition.

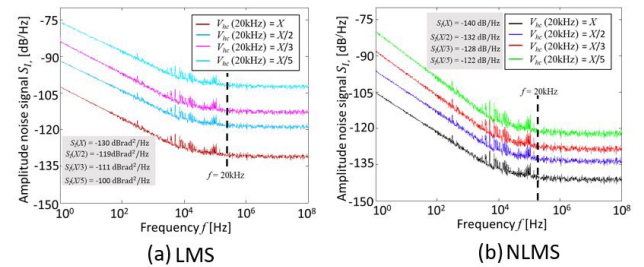


Figure 11. PSD of amplitude sensing signal of (a) LMS and (b) NLMS in the case homogenous agar condition.

5. Conclusion

In this paper, we proposed the implementation of adaptive noise cancellation (ANC) algorithms, i.e. least means square (LMS) and normalized least means square (NLMS) algorithm, for FPGA-based EIT system in order to elimi-

nate the noise signal and to increase the sensitivity detection by increasing the effective number of bit (*ENOB*) of ADC parameter. This study evaluates the benefit of ANC algorithms application on the new field which is the Electrical Impedance Tomography (EIT) system. The ANC of LMS and NLMS algorithm work by calculating the transfer function coefficients in order to minimize the error signal between the sensing signal D_i and the noise signal h itself. The ANC of LMS and NLMS algorithm was compared and evaluated through experimental studies.

The experimental results showed that the implementation of ANC algorithms onto FPGA-based EIT system:

(1) Implementing ANC algorithms on FPGA-based EIT system shows significantly more accurate image reconstruction to show the inclusion as compared without ANC algorithms implementation.

(2) ADC performances of NLMS has $S_f = -147 \text{ dBV}^2/\text{Hz}$ and effective number of bit (*ENOB*) = 11.2 bit.

(3) ADC performances of LMS has $S_f = -139 \text{ dBV}^2/\text{Hz}$ and *ENOB* = 9 bit.

Higher ADC performances of NLMS as compared with LMS leads to the better image reconstruction performances in terms of amplitude response (*AR*), position error (*PE*), resolution (*RES*), and shape deformation (*SD*).

We realize that several ANC algorithms are already proposed, not only LMS or NLMS. On this regards, this consideration opens a new opportunity to apply the variant of ANC algorithms that could be suitable with different EIT applications.

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ARTICLE

Robust Power System Stabilizer Design Using Kharitonov's Theorem: A Case Study

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ABSTRACT

This paper proposes a robust power system stabilizer(PSS) for a steam synchronous generator in Barka II power station. The PSS should be capable of damping small-disturbance oscillations (inherently existing in power systems due to e.g. load changes, lines switching...etc.) within a certain settling time for different load conditions. Also, the proposed PSS must have the conventional structure and its parameters must not be violated. To achieve this goal, robust control provides many advantages. The suggested controller is tuned by the Kharitonov's theorem and uses the standard structure employed in industry. The problem is cast into a nonlinear constrained optimization problem to achieve the desired settling time without violating the practical values of the controller parameters. Performance of the robust PSS is evaluated by several simulations in the presence of system uncertainty due to load changes.

Abbreviations

AVR: Automatic voltage regulator, CPSS: Conventional power system stabilizer, LTI: Linear time-invariant systems, PSS: Power system stabilizer

Nomenclature

All units are in per unit (p.u) unless otherwise stated.

Δ : deviation

δ : Torque angle of machine, rad.

ω : Speed .

x'_d : Generator direct-axis transient reactance.

x_d, x_q : Direct and quadrature-axis synchronous reactance, respectively.

I_d, I_q :D-axis, Q-axis armature and field winding cur-

rents, respectively.

P, Q :Real and reactive power loading, respectively.

V : Terminal voltage of the generator.

V_∞ : Infinite bus voltage.

x_e : Transmission line reactance.

H : machine inertia constant , sec

M : machine momentum $=2H$, sec.

E'_q : The quadrature-axis transient voltage of machine.

E'_f : The field voltage of machine .

T'_{d0} : D-axis open circuit field time constant, sec.

K_E : Gain of the excitation system.

T_E : Time constant of the excitation system, sec.

x : The vector of the state variables.

u : The vector of input variables.

A, B, C : State, input, and output matrices, respectively

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G_{PSS} : Transfer function of PSS.

1. Introduction

Power systems have to provide good service to consumers in terms of constant value of voltage and frequency. Automatic voltage regulators (AVR) are installed to Synchronous generators to adjust its terminal voltage. For tight control of terminal voltage, the AVR uses high gain in the loop (which may result in instability). Power systems are subject to small disturbances as a result of line switching, changing on the value of the load and other reasons. These small disturbances that happen on the system may grow to cause system separation; consequently, a great loss on the national economy. In order to decrease the swings that happen during transient rotor angle instability and fix the voltage value to constant value, a stabilizing signal is added to the excitation system. The power system stabilizers are developed to support and help in damping these low-frequency oscillations (1-3 Hz) by adding a signal to the excitation system^[1-5].

For PSS tuning, different techniques such as classical^[6,7], intelligent^[8], adaptive^[9], robust^[10-15], switching control^[16] methods are reported. However, considering the variable nature of the loads (causing system uncertainty), the robust control is the best choice. The main advantages of robust approach in comparison of other methods is in its simplicity (one controller with fixed parameters facing the system uncertainty due to load changes).

In the present paper, a robust control method based on Kharitonov's theorem is proposed to design a robust conventional PSS which guarantees satisfactory operation under uncertain operating conditions. In addition to not violating the practical ranges of the controller parameters. The Kharitonov's theorem addresses few polynomials which are obtained from closed loop characteristic equation. Unlike^[12] which uses the root locus method to stabilize many polynomials and does not consider the practical limits of the controller parameters, the suggested design avoids these difficulties by solving the problem using constrained nonlinear optimization approach.

This paper is organized as follows: the case study system, its structure and mathematical model for performance evaluation of the proposed robust control method, in sections 2 and 3, are described. In section 4, control design strategy based on Kharitonov's theorem, is presented. Simulation results and conclusion are presented in sections 5 and 6 respectively.

2. Case Study System

The case study system is Barka II power station. It is

composed of 3x200 MVA steam +2x175 MVA gas turbines, Figure 1a. The design of PSS installed at one of the steam turbines is presented in^[18]. In that study, the problem is solved as a single machine infinite bus (SMIB) system. In which, the rest of the system (the other 4 machines+ the oman grid) is represented by Thevenin's equivalent using the short circuit fault level at the 11 kv bus. This is shown in Figure 1b by V_{∞}, x_e . The PSS of^[18] is based on the heaviest load in summer and it has the notch filter structure to eliminate the conjugate complex poles near the imaginary axis of the system.

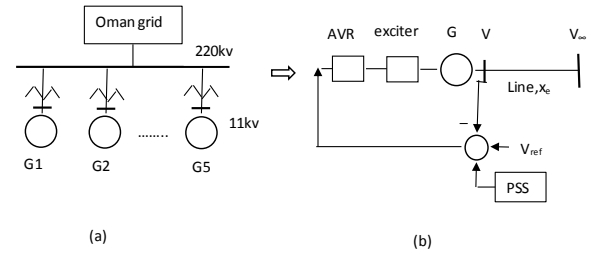


Figure 1 The case study system , Barka II power station

In present manuscript, a conventional PSS is designed for one of the gas turbines which achieves: (1) robustness against load variations, and (2) practical ranges of controller parameters.

3. Mathematical Model

The proposed robust saturated controller is tested using a single machine infinite bus system (SMIB). The mathematical nonlinear model describing the dynamics of SMIB is given by:

$$\begin{aligned} \dot{\delta} &= \omega_0(\omega - 1) \\ 2H\dot{\omega} &= P_m - P_e = P_m - [E'_q + (x_q - x'_d)I_d]I_q \\ T'_{d0}\dot{E}'_q &= E_q - [E'_q + (x_d - x'_d)I_d] \\ T_E\dot{E}_f &= -E_f + K_E(V_{ref} - V_t + u) \end{aligned} \quad (1)$$

where I_d, I_q , and V_d are given by, Figure2:

$$\begin{aligned} I_d &= \frac{E'_q - V \cos \delta}{x'_d + x_e} \\ I_q &= \frac{V \sin \delta}{x_q + x_e} \\ V &= \sqrt{[(I_q x_q)^2 + (E'_q - I_d x'_d)^2]} \end{aligned} \quad (2)$$

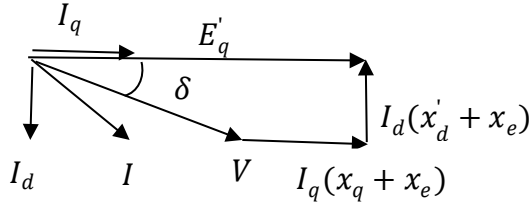


Figure 2. Phasor diagram

The SMIB data are in per unit (p.u.) unless otherwise stated and have their usual meaning ^[1] and are listed in Table 1. Three operating conditions (light, nominal, and heavy) are considered to cover the whole range of load changes all over the year.

Table 1. Data for Barka power station II, gas turbine

x_d	2.11
x'_d	0.207
x_q	1.97
ω_o	314 rad/s
T'_{do}	10.8 s
H	1.256 s
K_E	25
T_E	0.05 s
x_e	0.15
P _{Light}	0.4
P _{Nominal}	0.7
P _{Heavy}	1
Q _{Light}	0.35
Q _{Nominal}	0.525
Q _{Heavy}	0.875
V	1

Note that H in the above table is calculated as follows:

$$H = \text{kinetic energy / machine rating} = \frac{1}{2} J \omega^2 / S$$

$$= 0.5 * 5091 \text{ kg.m}^2 * (3000 \text{ rpm} \times 2\pi / 60)^2 / (175 \text{ MVA} * 10^6)$$

Substituting (2) into (1), and linearizing the resulting mathematical model for small disturbance around the three operating conditions (using the matlab commands trim, and linmod) to get the state equation:

$$\dot{x} = Ax + Bu, y = Cx \quad (3)$$

Where $x = [\Delta\delta, \Delta\omega, \Delta E'_q, \Delta E'_f]^T$ and the entries of the matrix A are functions of the active and reactive machine load (P, Q). The input u is generated by the PSS. Since the most commonly used input to the PSS is the change in speed $\Delta\omega$, the matrix C is selected as

$$C = [0, 1, 0, 0] \quad (4)$$

The plant (synchronous machine) transfer function model for the three operating conditions are

$$G_{\text{Light}} = \frac{21.71s}{s^4 + 20.22s^3 + 233.8s^2 + 4247s + 3900}$$

$$G_{\text{Nominal}} = \frac{28.87s}{s^4 + 20.22s^3 + 326.3s^2 + 6115s + 5511}$$

$$G_{\text{Heavy}} = \frac{30.18s}{s^4 + 20.22s^3 + 425.3s^2 + 8108s + 8056}$$

By using the transfer functions of the nominal, heavy, and light loads, the plant uncertain transfer function is obtained by taking the minimum and the maximum coefficients value of the parameters.

$$G_{\text{plant}} = \frac{[21.71 \ 30.18]s}{s^4 + 20.22s^3 + [233.8 \ 425.3]s^2 + [4247 \ 8108]s + [3900 \ 8056]} \quad (5)$$

4. Control system strategy

The control target is to design a conventional PSS to robustly stabilize the uncertain plant (5) i.e. to damp the system oscillations in less than 10-15 sec and not to violate the controller's parameters limitation imposed in industry. The conventional PSS can be a single lead or double lead controller

$$G_{\text{PSS,Single lead}} = K \frac{T_1s + 1}{T_2s + 1}$$

$$G_{\text{PSS,Double lead}} = K \frac{[T_1s + 1][T_3s + 1]}{[T_2s + 1][T_4s + 1]} \quad (6)$$

The practical ranges of PSS parameters (all positive scalars) are ^[3]:

$$0.1 \leq K \leq 50, \quad 0.2 \leq T_1 \leq 1.5 \text{ sec}, \quad 0.02 \leq T_2 \leq 0.15 \text{ sec},$$

$$0.2 \leq T_3 \leq 1.5 \text{ sec}, \quad 0.02 \leq T_4 \leq 0.15 \text{ sec}.$$

The closed loop system is shown in Figure 3

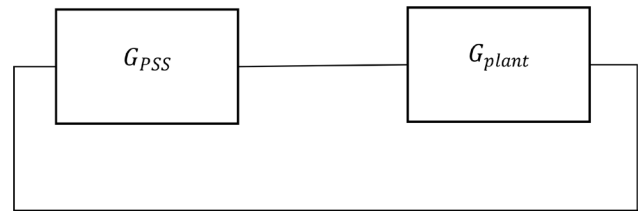


Figure 3. The closed loop system

The closed loop characteristic polynomial is

$$1 + G_{\text{PSS}}G_{\text{plant}} = 0 \quad (7)$$

Note that (7) is an uncertain polynomial function of the PSS parameters.

4.1 Kharitonov's Theorem

When the coefficients of the characteristic polynomial are fixed, the best method is to use the Routh-Hurwitz to test

if the system is stable or not. When the exact value parameters of the system are not fixed due to changing loads of the SMIB, an infinite number of polynomials have to be checked for stability using the Routh method. Kharitonov's theorem avoid such difficulty by checking the stability only 4 polynomials^[17]. It can be used to estimate the stability of uncertain dynamical systems. When the range of the polynomial coefficients is known, the Kharitonov's theorem can be used to find the stability of the dynamical system. The polynomial

$$p(s) = a_n s^n + a_{n-1} s^{n-1} + \dots + a_0 \quad (8)$$

is called an interval polynomial when the coefficients a_i is independent of each other and changes in an interval having minimum and maximum limit. An interval polynomial is represented as

$$p(s) = \sum_{i=0}^n [a_i^-, a_i^+] s^i \quad (9)$$

The system is robustly stable if and only if the following four Kharitonov polynomials are stable^[17].

$$p_1 = a_0^- + a_1^- s + a_2^+ s^2 + a_3^+ s^3 + a_4^- s^4 + \dots \quad (10)$$

$$p_2 = a_0^+ + a_1^+ s + a_2^- s^2 + a_3^- s^3 + a_4^+ s^4 + \dots \quad (11)$$

$$p_3 = a_0^+ + a_1^- s + a_2^- s^2 + a_3^+ s^3 + a_4^+ s^4 + \dots \quad (12)$$

$$p_4 = a_0^- + a_1^+ s + a_2^+ s^2 + a_3^- s^3 + a_4^- s^4 + \dots \quad (13)$$

Note that (9) is a polynomial of descending power of s . Whereas, (10-13) are polynomials of ascending powers of s . The “-” and “+” show the minimum and maximum bounds of the polynomial coefficients. In the Kharitonov's theorem, the $p(s)$ is considered as the closed loop polynomial. In our PSS problem, the polynomial coefficients are dependent. In this case, the Kharitonov's theorem represents only a sufficient condition for the stability of uncertain polynomial if the coefficients are assumed independent. Kharitonov's theorem provides a simple and powerful tool for robust adjusting of practical controllers.

4.2 Application of Kharitonov's Theorem to Robust PSS Design

It is very difficult to try to find an analytical expression for the closed loop uncertain polynomial as a function of the PSS parameters (all are positive scalars), extracting the 4 Kharitonov's polynomials, arranged an ascending powers of s . This can be easily done numerically using the matlab

commands: conv, flip. Trying to stabilize the Kharitonov's polynomials by adjusting the PSS parameters using the root locus method as presented in^[12] is also difficult. The easiest way is to cast the robust PSS problem in the following constrained nonlinear optimization to shift the rightmost closed loop pole (among the 4 kharitonov's polynomials) to the left in the complex plane as much as possible.

$$\text{minimize}_{K,T_1,T_4} J = \max \{\text{real part of closed loop pole}\} \forall p_1..p_4 \quad (14)$$

Subject to the following constraints

$$0.1 \leq K \leq 50, 0.2 \leq T_1 \leq 1.5, 0.02 \leq T_2 \leq 0.15, 0.2 \leq T_3 \leq 1.5, 0.02 \leq T_4 \leq 0.15.$$

The optimization problem (14) is highly nonlinear. To avoid trapping into a local minimum solution, it is best solved using a probabilistic approach e.g. the particle swarm optimization (matlab command, “particleswarm”).

Solving (14), the obtained single lead and double lead robust PSS (providing the max possible left shift to the rightmost closed loop poles, thus achieving the fastest oscillation damping) are

$$G_{\text{Single lead}} = 13.0602 \frac{1.5s + 1}{0.02s + 1}$$

$$G_{\text{Double lead}} = 1.8635 \frac{[1.5s + 1]}{[0.02s + 1]} \frac{[1.5s + 1]}{[0.02s + 1]}$$

As seen there is no violation to the practical ranges of the PSS parameters.

5. Performance Evaluation

For a cleared three phase short circuit fault on the generator terminal which causes 0.1 rad angle disturbance, the performance at different loads is given in Figure 4,a,b,c, for single lead robust PSS.

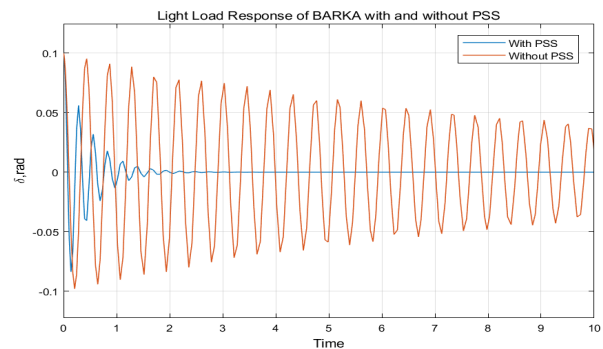


Figure 4 (a). Light Load Response of Barka Power II station with and without PSS

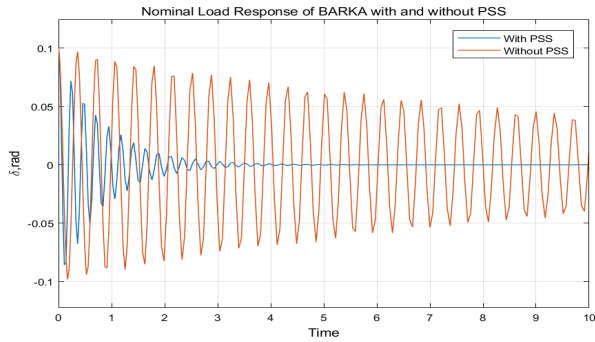


Figure 4 (b). Nominal Load Response of Barka Power II station with and without PSS

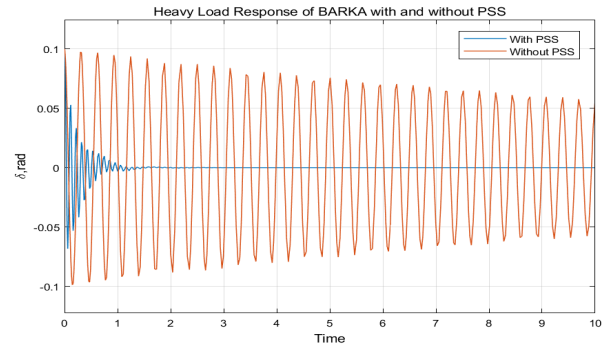


Figure 5 (b) (c). Heavy Load Response of Barka Power II station with and without PSS

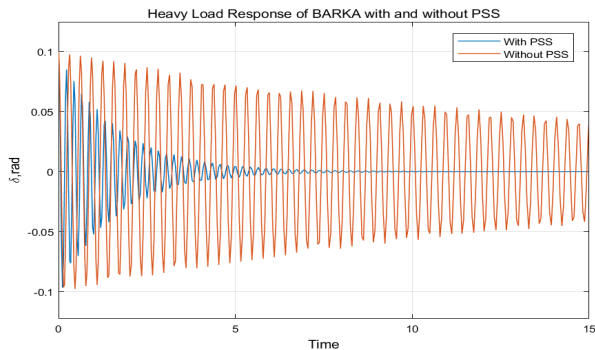


Figure 4 (c). Heavy Load Response of Barka Power II station with and without PSS

Similarly for double lead robust PSS, the response at different loads is shown in Figure 5a,b,c.

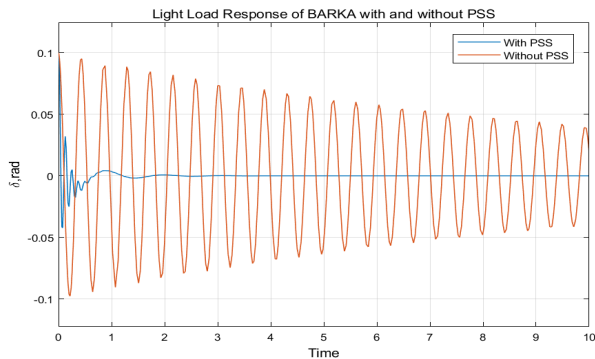
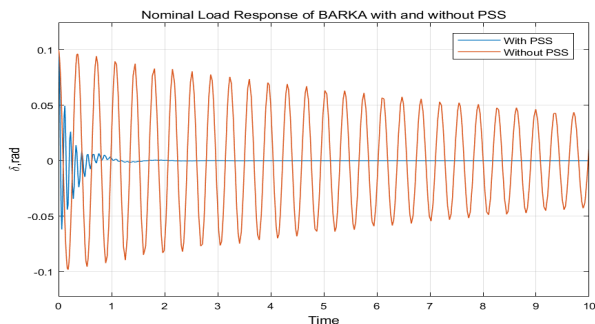


Figure 5 (a). Light Load Response of Barka Power II station with and without PSS



It is evident that when the load increases (the system becomes nearer to instability), the settling time increases.

Out of Figure 4,5, the following settling time comparison can be done between the single lead and double lead robust PSS, Table 2.

Table 2. Settling time, s

	Single lead robust PSS	Double lead robust PSS
Light load	3	2
Nominal load	5	2.1
Heavy load	10	2.5

If the designer is not satisfied with a single lead PSS at heavy load (the settling time is 10 s), he can use the double lead PSS. The price is more complex hardware.

6. Conclusion

A simple method is proposed for the design of the PSS. The method is based on Kharitonov's theorem. The suggested design retains the conventional PSS structure with controller parameters in the practical ranges. Robustness of the proposed controller is evaluated via several simulations in the MATLAB/SIMULINK environment. The double lead robust PSS is more effective in quenching system oscillations than the single lead. The price is more hardware complexity.

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ARTICLE

Optimal PV Allocation & Minimal tap-Changing Transformers Achieving Best Distribution Voltage Profile & Minimum Losses in active distribution networks

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ABSTRACT

In distribution systems, voltage levels of the various buses should be maintained within the permissible limits for satisfactory operation of all electrical installations and equipment. The task of voltage control is closely associated with fluctuating load conditions and corresponding requirements of reactive power compensation. The problem of load bus voltage optimization in distribution systems that have distributed generation (DG) has recently become an issue. In Oman, the distribution code limits the load bus voltage variations within $\pm 6\%$ of the nominal value. Several voltage control methods are employed in active distribution systems with a high share of photovoltaic systems (PV) to keep the voltage levels within the desirable limits. In addition to the constraint of targeting the best voltage profile, another constraint has to be achieved which is the minimum loss in the distribution network. An optimised solution for voltage of load buses with on-load tap-changing (OLTC) transformers and PV sources is presented in this paper. This study addresses the problem of optimizing the injected power from PV systems associated with the facilities of tap-changing transformers, as it is an important means of controlling voltage throughout the system. To avoid violating tap-changing constraints, a method is depicted for determining the minimal changes in transformer taps to control voltage levels with distributed PV sources. The taps of a range $+5$ to -15% , can be achieved by tap-changing transformers. The OLTC operation was designed to keep the secondary bus within the voltage standard for MV networks.

1. Introduction

Recently, the integration of distributed generation gives rise to many challenges to distribution operators. A distribution network integrated with PV is termed active distribution network. The problem of voltage control in radial networks has been studied in several literatures^[1]. Different methods for allocating PV sources in distribution

systems including transformers with tap-changing facilities^[2-6]. Although these methods are efficient, it needs more steps and computational time than the presented method. Most of transformers have a motorized on-load tap-changers (OLTC) that adjusts the transformer turns ratio, typically in steps of 1.25% and are utilized to improve the voltage profile of high and medium voltage (HV&MV) grids^[7]. Some transformers have no-load tap

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changers (NLTC). These transformers need to be taken out of service to change the tap, which is not advisable in modern smart grid operations. The OLTC is normally controlled by an automatic voltage regulator AVR relay that increases or decreases the tap positions of the transformer. The transformer OLTC are usually at its HV side as the current of this side is lower and there are more turns are available. This makes the voltage regulation more precise. The switching signals to the OLTC transformers can be automated using an automatic voltage controller either local or at a control centre.

The increased constraints, number of decision variable, and nonlinearity of the DG optimization problems, especially in the case of distribution networks with DGs, made the exact methods incapable of solving such problems without a number of simplifying assumptions that significantly reduce the solution accuracy [9].

This paper investigates the optimum voltage control strategy in busses equipped with load tap-changing transformers and PV sources in radial networks. Several voltage control methods are employed in power systems to keep the voltage levels within the desirable limits. Some of these methods of voltage control in power systems are summarized. The excitation control and voltage regulators at the generating stations provide the basic means of voltage control to maintain the voltage at its scheduled level. There are other additional techniques to control the voltage using reactive power injection or absorption by reactors, synchronous condensers, shunt capacitors, static var compensators SVCs, Flexible AC transmission (FACT) devices and tap changing transformers. These techniques can be applied at sending or receiving end of transmission and distribution lines in industries, substations, and distribution substations to maintain the voltage levels within the standard limits. The users' load connected to the distribution network are devices, which operate at a voltage of a nominal value. If that value drops or increases it will affect the operation of that device and it may damage it, therefore it was needed to have a voltage control method to keep the voltage at its normal nominal value [5].

The voltage control of distribution systems is obtained basically by changing tap position of on-load or off-load tap changers. Off-load tap changing adjustments are usually for seasonal load variations of the special operational requirement of local substations and adjusting the voltage in distribution transformer at the consumer end. On the other hand, on-load tap changing is employed for changing the turn-ratio of the transformer to regulate the system voltage while the transformer is delivering the load. By changing the turns ratio of the transformer both the voltage ratio and the secondary voltage are changed. Tap

changing is widely used voltage control method employed at every voltage level. The disadvantage of using OLTC is the voltage collapse due to discrete OLTC switching at intervals of tens of seconds because of the interaction of OLTC dynamics, system loading, and generator reactive power limits. In this case, OLTC blocking can forestall voltage collapse.

This paper investigates the optimum PV allocation strategy in active distribution systems in conjunction with minimal taps of OLTC transformers compared with the case if no tap changing transformers are available. The adopted range for the OLTC has been decided based on datasheets of manufacturers of distribution transformers with tap-changing capabilities.

2. Problem Formulation

Most types of optimization functions in the Engineering applications are as constrained optimization problem. The voltage control by minimum tap changes with PV connected to transformer buses can be cast into a nonlinear constrained optimization. The constrained multivariable optimization problem with an objective function that needs to be maximized or minimized, is followed with constrains functions. The objective function to be minimized is to: (1) minimize the bus voltage deviations from 1 p.u. to achieve the best voltage profile, (2) minimize the transformers tap deviation from 1 p.u so as not to violate the taps limits, and (3) minimize system losses. The function minimization is subject to the constraints of load flows in the active distribution network.

The problem is stated mathematically as finding the tap settings T and PV active power injections so that the 'objective function' for the n -bus radial system is minimized for the radial system shown in figure 1.

$$\min J = (V-I)^T A (V-I) + (T-I)^T B (T-I) + \sum P_{Loss} \quad (1)$$

Subject to the load flow constraints.

$$T_n V_n - T_{n-1} V_{n-1} = \frac{R_{n-1}(P_1 + \sum P_{Loss(n-1)} + P_{n-1}) + X_{n-1}(Q_1 + \sum Q_{Loss(n-1)} + Q_{n-1})}{T_{n-1} V_{n-1}} \quad (2)$$

Where:

A and B are positive definite diagonal element matrices will determine the relative importance of bus voltage deviation and changes in taps to be as close as possible to one p.u. Matrix A is a diagonal unity matrix multiplied by an accelerating α factor.

$$V = [V_1 \angle 0, V_2 \angle \delta_1, \dots, V_n \angle \delta_{n-1}]^T, \text{ The vector } 1 = [1, 1, \dots, 1]^T, T = [T_1, T_2, \dots, T_n]^T$$

For any section in distribution networks, the phase

angle between voltages at that section can be neglected. Hence, we get

$$\sum P_{\text{Loss}} = \frac{(T_2 V_2)^2 - (T_1 V_1)^2}{|Z_1|^2} R_1 + \frac{(T_3 V_3)^2 - (T_2 V_2)^2}{|Z_2|^2} R_2 + \dots$$

$$\sum Q_{\text{Loss}} = \frac{(T_2 V_2)^2 - (T_1 V_1)^2}{|Z_1|^2} X_1 + \frac{(T_3 V_3)^2 - (T_2 V_2)^2}{|Z_2|^2} X_2 + \dots$$

The resulting equations can be put in a compact form as follows:

$$\mathbf{g} = [g_1, g_2, \dots, g_{n-1}]^T = [0, 0, \dots, 0]^T \quad (3)$$

where

$$g_{n-1} = T_n V_n - T_{n-1} V_{n-1} - \frac{R_{n-1}(P_1 + \dots + P_{n-1}) + X_{n-1}(Q_1 + \dots + Q_{n-1})}{T_{n-1} V_{n-1}} = 0 \quad (4)$$

In this paper, a method is introduced to solve a constrained optimization problem of allocating PV sources in electrical networks with bus voltage constraints. The method of minimal taps of OLTC transformer is used taking into account the line losses and percentage of PV penetration.

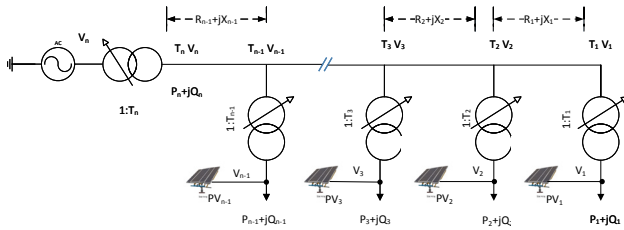


Figure 1. Radial System under study with PV penetration

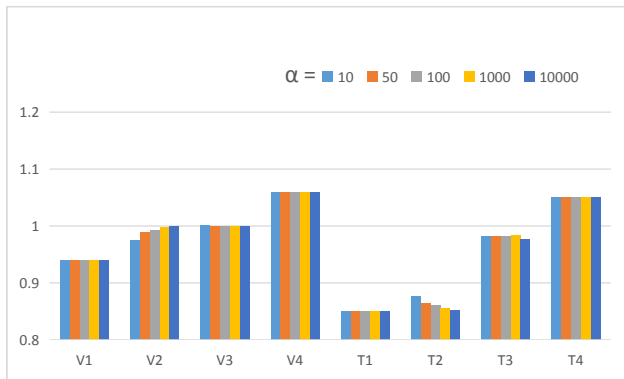


Figure 2. Optimized Values of PV penetration Transformer taps in per unit at variable α factor

Under normal loading conditions, voltage control requirements will be maintained within transformer tap limits. Hence, the weighting matrices A and B could

be chosen with equal weight. On the other hand, under extreme heavy loading conditions adjusting the voltage levels to their desired levels could call up taps outside their physical limits. To avoid tap violation outside their physical limits, the control algorithm is penalized by choosing the elements of the matrix B much larger than those of matrix A. The PV penetration will improve the voltage level as well with a constrained level of maximum 20 % of the load level at that particular bus. This will result in relaxed voltage profile. If the losses are only minimized, the voltage profile may be violated during the optimization of the PV system allocation [9]. This is why both voltage and transformer tap constraints should be taken into account.

3. Problem Solution

There are two approaches to solve the above constrained nonlinear optimization problem: (1) to transform it to unconstrained one using the Lagrangian multiplier method, or (2) to solve it directly as a nonlinear optimization problem.

The voltage control problem is cast into minimizing the objective function J subject to the load flow equality constraints. The problem is that we wish to minimize J subject to the constraint $\mathbf{g}=0$.

A solution can be found using the method of Lagrangian multipliers. There will be a penalty vector λ for too big constraints called Lagrange multiplier vector. Therefore, the Lagrangian of the problem is

$$L = J + \lambda^T \cdot \mathbf{g} \quad (5)$$

L is the Lagrangian which will be the unconstrained function, J is the objective function in the constrained optimization problem, λ is the Lagrange multiplier vector, g are the constrain functions of constrained optimization problem.

The constrained nonlinear optimization problem will be solved by the direct method approach. Therefore, it is required to minimize J (1), subject to the power flow constraints $\mathbf{g}=0$ in (4). The above method is applied to a typical radial networks similar to those in Oman's power system as shown in Figure 1.

4. Improved Optimization Technique

In this section, the simulation results relevant to part of a distribution network in Oman are presented. Minimizing the objective function J (1) subject to the constraints g (3) can be solved directly as a constrained nonlinear optimization problem using the function "fmincon" in MATLAB®

for different values of α coefficient ($\alpha = 10, 10^2, 10^3, 10^4, \dots$).

Applying an increasingly active and reactive load by 7% each year for four years for system of figure 1, there was no voltage violation and the results are within the acceptable limits ($\pm 6\%$). The results can be summarized by the following table 1A for different values of the accelerating factor and an increasing load up to year 2022.

Therefore, this method can be used to produce a good voltage profile by just optimizing the tap-changer of transformers in the presence of PV distributed generation. Results recorded in table A1 and figure 2 show a tight control grip will be carried out over the bus voltage profile for higher α values. However, PV supports the voltage within standard range.

The methods of this paper could be applied successfully in a real network. To make the operation in the network more smarter, sensors and measuring IED devices send the required signals to a central computer^[10]. The optimized values of transformers taps and PV output power are then sent back to each bus to control the bus voltages hence the line losses. During partial or no solar insolation, a trade-off has to be done between energy storage or other renewable energy sources.

5. Conclusions

The results here presented show interesting applications of an innovative methodology for voltage control analysis using tap-changing transformers and PV sources, which can become a viable way to achieve consistent and robust voltage control in distribution networks with PV sources. Results show the superiority of the proposed algorithm to the existing ones. This can improve considerably the performance of electrical transmission and distribution systems. The algorithm can be further improved by using online data measurements at each bus with adaptive PV-level penetration.

Appendixes

Table A1. Optimised results for 7% load growth at $\alpha=100$

$P_{1,2,3}$	2018	2019	2020	2021	2022
V_1	0.9456	0.9400	0.9400	0.9400	0.9400
V_2	0.9988	0.9922	0.9922	0.9925	0.9875
V_3	0.9998	1.0004	1.0004	1.0006	0.9915
V_4	1.0600	1.0600	1.0600	1.0600	1.0600
T_1	0.8500	0.8500	0.8500	0.8500	0.8500
T_2	0.8580	0.8614	0.8617	0.8621	0.8668
T_3	0.9810	0.9820	0.9831	0.9843	0.9942

T_4	1.0500	1.0500	1.0500	1.0500	1.0500
PV_1	0.0250	0.0350	0.0380	0.0410	0.0437
PV_2	0.0300	0.0400	0.0430	0.0460	0.0493
PV_3	0.0270	0.0370	0.0400	0.0424	0.0454
V_{Norm}	0.0810	0.0852	0.0852	0.0852	0.0862
T_{Norm}	0.2134	0.2111	0.2108	0.2104	0.2068
P_{loss1}	0.0404	0.0442	0.0448	0.0459	0.0463
P_{loss2}	0.1705	0.1814	0.1836	0.1856	0.1872
P_{loss3}	0.2681	0.2641	0.2595	0.2548	0.2172

Table A2. Line data

From Bus	To Bus	R [pu]	X [pu]
1	2	0.07	0.19
2	3	0.09	0.195
3	4	0.067	0.184

Table A3. Load data

Bus No.	P [pu]	Q [pu]
1	0.35	0.16
2	0.04	0.19
3	0.37	0.15

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- A brief description of the novelty and importance of the findings detailed in the paper

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- Project sponsors
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- Other financial relationships/support
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Main Manuscript Outline

V . Introduction

The introduction should highlight the significance of the research conducted, in particular, in relation to current state of research in the field. A clear research objective should be conveyed within a single sentence.

VI . Methodology/Methods

In this section, the methods used to obtain the results in the paper should be clearly elucidated. This allows readers to be able to replicate the study in the future. Authors should ensure that any references made to other research or experiments should be clearly cited.

VII . Results

In this section, the results of experiments conducted should be detailed. The results should not be discussed at length in

this section. Alternatively, Results and Discussion can also be combined to a single section.

VIII. Discussion

In this section, the results of the experiments conducted can be discussed in detail. Authors should discuss the direct and indirect implications of their findings, and also discuss if the results obtain reflect the current state of research in the field. Applications for the research should be discussed in this section. Suggestions for future research can also be discussed in this section.

IX. Conclusion

This section offers closure for the paper. An effective conclusion will need to sum up the principal findings of the papers, and its implications for further research.

X. References

References should be included as a separate page from the main manuscript. For parts of the manuscript that have referenced a particular source, a superscript (ie. [x]) should be included next to the referenced text.

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XI. Glossary of Publication Type

J = Journal/Magazine

M = Monograph/Book

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D = Dissertation/Thesis

P = Patent

S = Standards

N = Newspapers

R = Reports

Kindly note that the order of appearance of the referenced source should follow its order of appearance in the main manuscript.

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Graphs, figures and tables should be labelled closely below it and aligned to the center. Each data presentation type should be labelled as Graph, Figure, or Table, and its sequence should be in running order, separate from each other.

Equations should be aligned to the left, and numbered with in running order with its number in parenthesis (aligned right).

XII. Others

Conflicts of interest, acknowledgements, and publication ethics should also be declared in the final version of the manuscript. Instructions have been provided as its counterpart under Cover Letter.



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