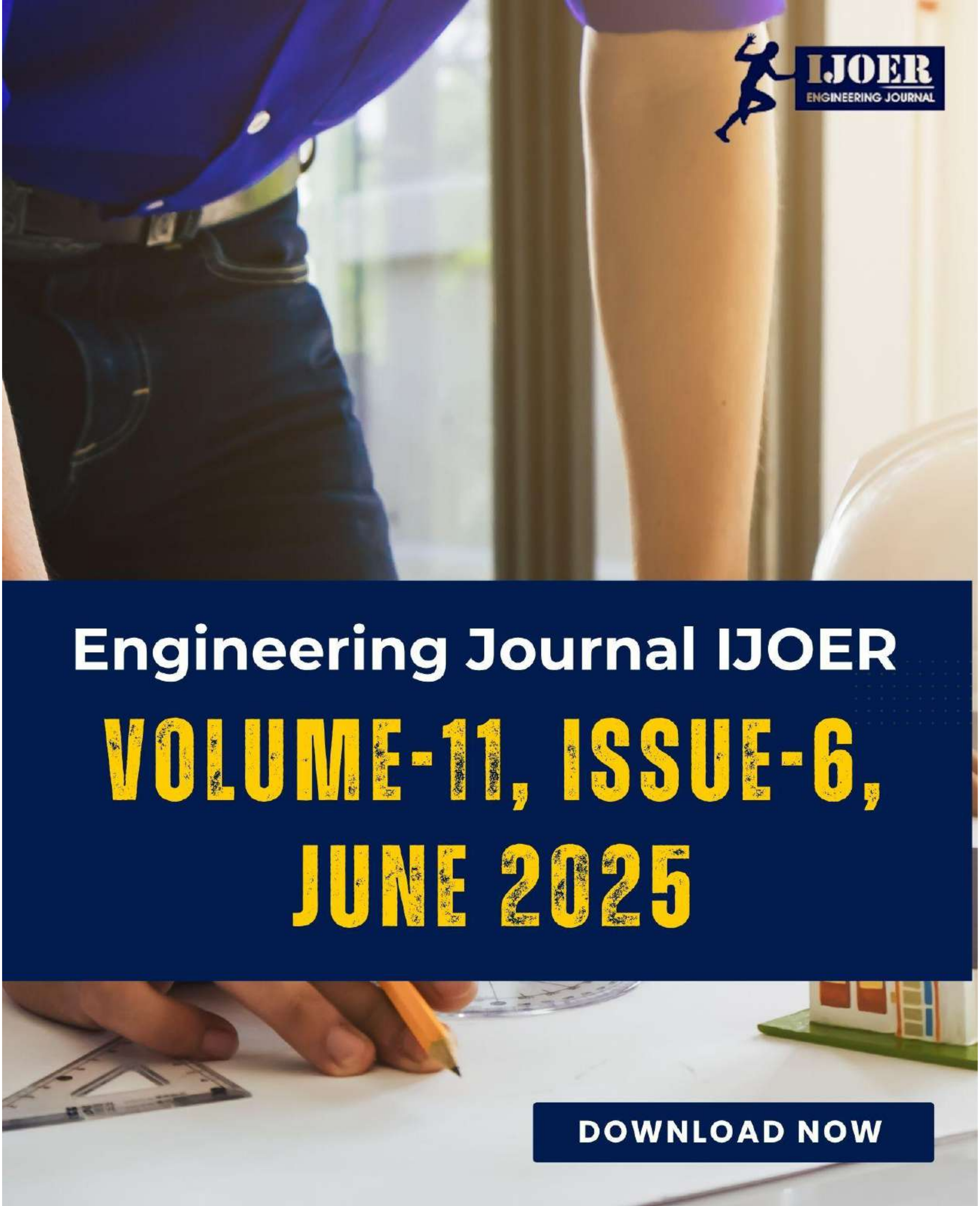

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VOLUME-11, ISSUE-6, JUNE 2025

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Preface

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Each article in this issue provides an example of a concrete industrial application or a case study of the presented methodology to amplify the impact of the contribution. We are very thankful to everybody within that community who supported the idea of creating a new Research with IJOER. We are certain that this issue will be followed by many others, reporting new developments in the Engineering and Science field. This issue would not have been possible without the great support of the Reviewer, Editorial Board members and also with our Advisory Board Members, and we would like to express our sincere thanks to all of them. We would also like to express our gratitude to the editorial staff of AD Publications, who supported us at every stage of the project. It is our hope that this fine collection of articles will be a valuable resource for *IJOER* readers and will stimulate further research into the vibrant area of Engineering and Science Research.

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Table of Contents

Volume-11, Issue-6, June 2025

A Novel Technique for Designing Multirate Filter Banks Exploiting Quasi-Newton Optimization

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Abstract— *This paper proposes a new technique for the design of multirate filter banks with linear phase in frequency domain. To match the ideal system response, low-pass analysis prototype filter response is optimized to minimize an objective function. The objective function is formulated as a weighted sum of pass-band error and stop-band residual energy of low-pass analysis filter, the square error of the overall transfer function at the quadrature frequency and amplitude distortion of the filter bank. Quasi-Newton optimization method is used to minimize the objective function by optimizing the filter tap weights of the prototype filter. Simulation results show that the proposed method is able to perform better than other existing methods.*

Keywords— Quadrature mirror filter bank, Quasi-Newton optimization method, Peak reconstruction error, Linear phase.

I. INTRODUCTION

Quadrature mirror filter (QMF) banks are multirate filter bank and have been extensively used for sub-band coding, where the signal is split into two or more sub-bands in the frequency domain, so that each sub-band signal can be processed in an independent manner and sufficient compression may be achieved [1]. At the receiver end the sub-band signals are recombined such that the original signal is properly reconstructed [2]. QMF banks find applications in many areas, such as analog to digital conversion [3], design of wavelet bases [4,5], image compression [6,7], digital trans-multiplexers [8], discrete multi-tone modulation systems [9], 2-D short-time spectral analysis [10], antenna systems [11], digital audio industry [12], biomedical signal processing [13,14,15].

Alias free efficient design of two-channel QMF banks while keeping minimum dimensions is a tough task. Therefore, various constrained and unconstrained optimization based techniques [16−31] have been developed for the design of linear phase QMF banks. Iterative methods [22−27] and genetic algorithms [28−31] have been proposed for the design problem of QMF based on multi-objective or single objective nonlinear optimization.

Fig. 1(a) shows the analysis and synthesis section of a popular multirate filter bank known as two-channel QMF bank. The discrete input signal x(n) is divided into two sub-band signals having equal band width, using the low-pass and high-pass analysis filters H0(z) and H1(z), respectively. Typical frequency responses of these filters are depicted in Fig. 1(b). The outputs of the synthesis filters are combined to obtain the reconstructed signal $\hat{x}(n)$. The reconstructed signal $\hat{x}(n)$ suffers from three types of errors: aliasing distortion (ALD), phase distortion (PHD), and amplitude distortion (AMD), due to the fact that the filters H0(z), H1(z), F0(z), and F1(z) are not ideal [32]. Therefore, the main stress of most of the researchers while designing the prototype filter for two-channel QMF bank has been on the elimination or minimization of these three distortions to obtain a perfect reconstruction (PR) or nearly perfect reconstruction (NPR) system [2, 16-21].

The overall transfer function of such an alias and phase distortion free system turns out to be a function of the filter tap coefficients of the low-pass analysis filter only [23]. Then, the AMD can only be minimized by optimizing the filter tap weights of the low-pass analysis filter using computer assistance techniques [2].

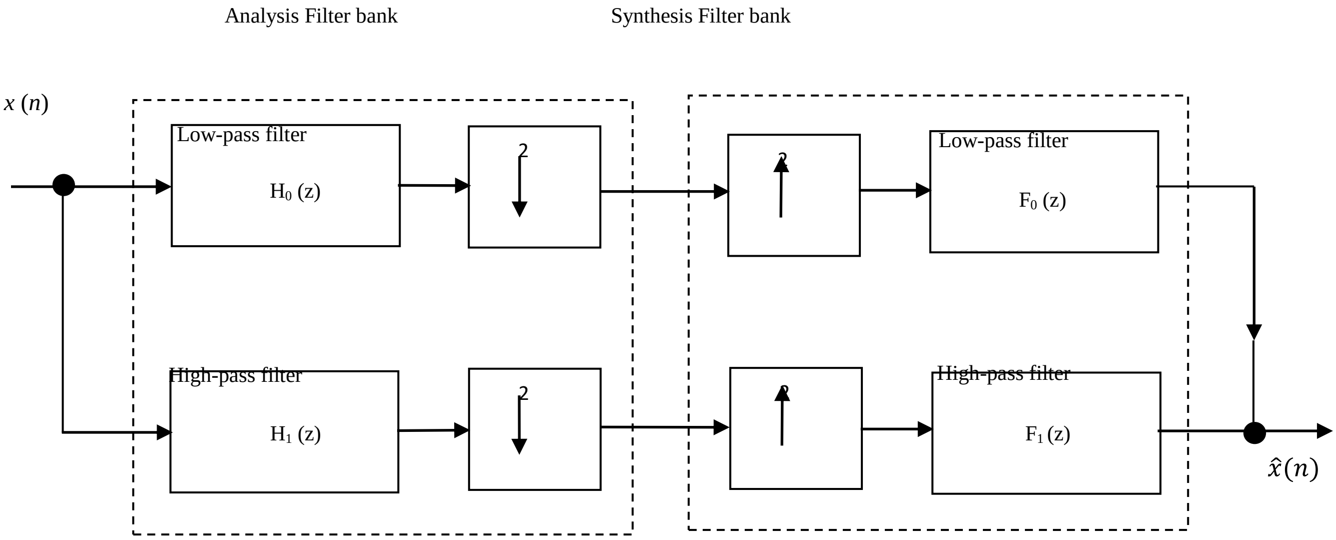

FIGURE 1 (a): Two-channel quadrature mirror filter bank

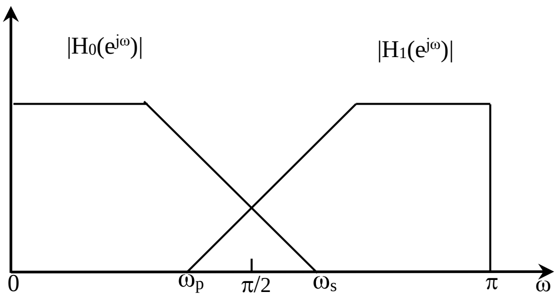

FIGURE 1 (b): Typical frequency responses of the analysis filters $H_0(z)$ and $H_1(z)$

It is well known [23] that the relation between output and input of an alias free two-channel QMF bank can be expressed. According to [23], the condition for perfect reconstruction can be written as:

$$|T(\omega)| = |H_r(\omega)|^2 + |H_r(\pi - \omega)|^2 = c, \text{ for all } \omega. \quad (1)$$

If the amplitude response of QMF bank in transition band is optimized, then the overall performance can be improved. Therefore, the aim is to optimize the coefficients of H0(z) by systematic computer-aided optimization technique, such that the filters satisfy the perfect reconstruction condition of Eq. (1) approximately.

In the above context, this paper presents a method to design a multirate filter bank. The objective function to be minimized [23] is formulated as a weighted sum of pass-band error and stop-band residual energy of low-pass prototype filter, the square error of the overall transfer function at the quadrature frequency $\omega = \pi/2$ and amplitude distortion of the QMF bank. Quasi-Newton optimization method is used to minimize the objective function by optimizing the filter tap weights of the prototype filter. The organization of rest of paper is as follows. Section 2 describes formulation of the design problem. Section 3 presents the proposed algorithm for design of prototype filter. Section 4 discusses the design results of the filter bank and comparison with already existing methods. Finally, conclusions are drawn in section 5.

II. DESIGN PROBLEM FORMULATION

To satisfy the perfect reconstruction condition of Eq. (1), an overall error function 'E' to be minimized is formulated as a weighted sum of four terms shown below:

$$E = \alpha_1.E_p + \alpha_2.E_s + \alpha_3.E_t \quad (2)$$

where α_1, α_2 , and α_3 re real constants, and E_p , E_s , and E_t are the measure of pass-band error, stop-band residual energy, square error of the overall transfer function at $\pi/2$, in transition band, and amplitude distortion, respectively. The objective function E is to be minimized iteratively using BFGS method by optimizing the coefficients of prototype filter.

For perfect reconstruction, the overall amplitude response of QMF bank, for all ω, must be equal to square of amplitude response of prototype filter at zero frequency. Consequently, the PR condition of Eq. (1) at quadrature frequency $\omega = \pi/2$, is reduced to:

$$H_r\left(\frac{\pi}{2}\right) = \frac{1}{\sqrt{2}} H_r(0) \quad (3)$$

where $H_r(0)$ and $H_r(\pi/2)$ are the amplitude response of prototype filter at zero frequency and quadrature frequency, respectively. The square error E_t is given by:

$$E_t = [H_r\left(\frac{\pi}{2}\right) - \frac{1}{\sqrt{2}} H_r(0)]^2 \quad (4)$$

Further, E_p , E_s and E_{am} are defined as follows:

$$E_p = \frac{1}{\pi}\int_0^{\omega_p} \left[|H_r(0)| - |H_r(\omega)|\right]^2 d\omega \quad (5)$$

$$E_s = \frac{1}{\pi}\int_{\omega_s}^{\pi} \left|H_0(e^{j\omega})\right|^2 d\omega \quad (6)$$

$$E_{am} = \max_{\omega}|T(e^{j\omega})| - \min_{\omega}|T(e^{j\omega})| \quad (7)$$

For even filter length (N+1), the frequency response of low-pass prototype filter $H_0(e^{j\omega})$ is given by [2]

$$H_0(e^{j\omega}) = \left[\sum_{n=0}^{(N-1)/2} 2h_0(n)\cos\omega\left((N/2)-n\right)\right].e^{-j\omega N/2} \quad (8)$$

$$= H_r(\omega)\, e^{-j\omega N/2} \quad (9)$$

where $b(n) = 2h_0(n)$ and $H_r(\omega)$ is the amplitude function written as

$$H_r(\omega) = \mathbf{b}^{\mathbf{T}}\mathbf{c}(\boldsymbol{\omega}) \quad (10)$$

where vectors **b** and **c(ω)** are:

$$\boldsymbol{b} = [b_0\ b_1\ b_2\ \ldots\ldots b_{(N-1)/2}]^T,\ \mathbf{c}(\boldsymbol{\omega}) = [\cos\omega(N/2)\ \cos\omega((N/2)-1)\ \ldots\ldots\ldots\ \cos(\omega/2)]^T \quad (11)$$

The amplitude function at zero frequency $\omega = 0$ is calculated as

$$H_r(0) = \mathbf{b}^{\mathbf{T}}\mathbf{c}\ (\mathbf{0}) = \mathbf{b}^{\mathbf{T}}\mathbf{1}, \quad (12)$$

where **1** is the vector with all (N+1)/2 elements equal to unity.

Now, E_t , E_p and E_s can be realized as:

$$E_t = [\mathbf{b}^{\mathbf{T}}\mathbf{c}(\boldsymbol{\pi}/\mathbf{2}) - \frac{1}{\sqrt{2}}\,\mathbf{b}^{\mathbf{T}}\mathbf{c}(\mathbf{0})]^2 = [\mathbf{b}^{\mathbf{T}}\mathbf{d} - H_{r1}]^2 \quad (13)$$

where vector **d** and **c(0)** are equal to vector **c(ω),** when it is evaluated at $\omega = \pi/2$ and $\omega = 0$, respectively, and $H_{r1} = 0.707\ \mathbf{b}^{\mathbf{T}}\mathbf{c}(\mathbf{0})$.

Similarly E_p can be realized

$$E_p = \mathbf{b}^{\mathbf{T}}\mathbf{F}\ \mathbf{b} \quad (14)$$

where **F** is a real, symmetric and positive definite matrix, given by

$$\mathbf{F} = \frac{1}{\pi}\int_0^{\omega_p} (\mathbf{c}(\mathbf{0}) - \mathbf{c}(\boldsymbol{\omega}))(\mathbf{c}(\mathbf{0}) - \mathbf{c}(\boldsymbol{\omega}))^{\mathbf{T}} d\omega \quad (15)$$

Stop band error E_s is given as:

$$E_s = \mathbf{b}^{\mathbf{T}}\ \mathbf{G}\ \mathbf{b} \tag{16}$$

where **G** is a real, symmetric and positive definite matrix, calculated as

$$\mathbf{G} = \frac{1}{\pi}\int_{\omega_s}^{\pi} \mathbf{C}(\boldsymbol{\omega})\mathbf{C}(\boldsymbol{\omega})^{\mathbf{T}}\, d\omega \tag{17}$$

III. PROPOSED METHOD FOR DESIGN OF LOW- PASS PROTOTYPE FILTER

By using Eqs. (13), (14) and (16) the objective E can be rewritten as:

$$\mathrm{E} = \alpha_1\, \mathbf{b}^{\mathbf{T}}\mathbf{F}\ \mathbf{b} + \alpha_2\, \mathbf{b}^{\mathbf{T}}\ \mathbf{G}\ \mathbf{b} + \alpha_3[\mathbf{b}^{\mathbf{T}}\mathbf{d} - H_{r1}]^2 \tag{18}$$

$$= \alpha_1\, \mathbf{b}^{\mathbf{T}}\mathbf{F}\ \mathbf{b} + \alpha_2\, \mathbf{b}^{\mathbf{T}}\ \mathbf{G}\ \mathbf{b} + \alpha_3[\mathbf{b}^{\mathbf{T}}\mathbf{D}\mathbf{b} - 2H_{r1}\mathbf{b}^{\mathbf{T}}\mathbf{d} + H_{r1}{}^{\mathbf{2}}]$$

$$= \mathbf{b}^{\mathbf{T}}\mathbf{R}\ \mathbf{b} + \alpha_3[-2H_{r1}\mathbf{b}^{\mathbf{T}}\mathbf{d} + H_{r1}{}^{\mathbf{2}}] \tag{19}$$

where matrix **R** and **D** are

$$\mathbf{R} = \alpha_1\ \mathbf{F} + \alpha_2\ \mathbf{G} + \alpha_3\mathbf{D} \text{ and } \mathbf{D} = \mathbf{d}\mathbf{d}^{\mathbf{T}}$$

The objective function given by Eq. (18) is a quadratic function and matrix **R** is a Hermitian positive definite matrix, therefore, E can be minimized by Quasi Newton optimization method [33]. If $\mathbf{b_i}$ is the approximation of the minimum point at the ith stage of iteration and λ_i is the optimal step length in the search direction, then the new or improved approximation in the (i+1)th stage of iteration using BFGS method can be calculated as

$$\mathbf{b}_{i+1} = \mathbf{b}_i - \lambda_i[\mathbf{B}_i]\nabla E_i = \mathbf{b}_i + \lambda_i\mathbf{s}_i \tag{20}$$

where ∇E_i is the gradient of the objective function E and $\mathbf{s}_i$ is the search direction, when evaluated at the design vector $\mathbf{b}_i$, both are given by

$$\nabla E_i = 2\mathbf{R}\mathbf{b} + \alpha_3[-2H_{r1}\mathbf{d}] \text{ and } \mathbf{s}_i = -[\mathbf{B}_i]\nabla E_i \tag{21}$$

and matrix $[\mathbf{B_i}]$ is the estimate of inverse of Hessian matrix. Initially the matrix $[\mathbf{B_i}]$ is taken as the identity matrix [**I**] and up-dation of this matrix is done using BFGS formula [33].

$$[\mathbf{B}_{i+1}] = [\mathbf{B}_i] + \left[1 + \frac{\mathbf{g}_i^T[\mathbf{B}_i]\mathbf{g}_i}{\mathbf{d}_i^T\mathbf{g}_i}\right]\frac{\mathbf{d}_i\mathbf{d}_i^T}{\mathbf{d}_i^T\mathbf{g}_i} - \frac{\mathbf{d}_i\mathbf{g}_i^T[\mathbf{B}_i]}{\mathbf{d}_i^T\mathbf{g}_i} - \frac{[\mathbf{B}_i]\mathbf{g}_i\mathbf{d}_i^T}{\mathbf{d}_i^T\mathbf{g}_i} \tag{22}$$

where

$$\mathbf{d}_i = \mathbf{b}_{i+1} - \mathbf{b}_i = -\lambda_i[\mathbf{B}_i]\ \nabla E_i \text{ and } \mathbf{g}_i = \nabla E_{i+1} - \nabla E_i \tag{23}$$

The optimum step length λ_i in the direction of $\mathbf{s}_i$ can be obtained by equating the derivate of objective function $E(\mathbf{b}_i + \lambda_i\mathbf{S}_i)$ with respect to λ, to zero. The derivate $dE/d\lambda = 0$, gives following expression for optimum step length

$$\lambda_i = \{\mathbf{b}_i^T\mathbf{R}\mathbf{s_i} - \alpha_3 H_{r1}\mathbf{d}^{\mathrm{T}}\mathbf{s_i}\} / \{\mathbf{s}_i^T\mathbf{R}\mathbf{s_i}\} \tag{24}$$

The unit energy constraint on the filter coefficients is also imposed within some prespecified limit as proposed in [17,27]. The design algorithm that minimizes the objective function in step by step manner proceeds through following steps:

(1) Specify filter length (N+1), stop band edge frequency (ω_s) and pass band edge frequency (ω_p).

(2) Assume initial values of $\alpha_1, \alpha_2, \alpha_3$ and α_4.

(3) Start with an initial design vector $\mathbf{h_0}$ = [0 0 0 0 0 …………$\frac{1}{\sqrt{2}}$], $\mathbf{h_0}$ is zero except $\mathbf{h_0}$ ((N-1)/2) = $\frac{1}{\sqrt{2}}$; satisfying the unit energy constraint within a pre-specified tolerance given by

$$g_1 = \left|1 - 2\sum_{k=0}^{(N-1)/2} h_0^2(k)\right| \le \varepsilon_1$$

(4) Set the iteration number, i = 1, and $\mathbf{b_i} = \mathbf{2h_0}$.

(5) Compute the objective function E_i, by using Eq. (18), at the design vector $\mathbf{b_i}$.

(6) Compute ∇E_i and the search direction $\mathbf{s}_i$,, at the design vector $\mathbf{b}_i$.

(7) Determine the optimum step length λ_i, by using Eq. (24).

(8) Compute the new approximation

$$\mathbf{b}_{i+1} = \mathbf{b}_i - \lambda_i[\mathbf{B}_i]\nabla E_i \; = \; \mathbf{b}_i + \lambda_i \mathbf{s}_i$$

(9) Obtain the constraint g_1, at the point $\mathbf{b_{i+1}}$, if it is violated then choose the optimum point as $\mathbf{b_i}$, stop the algorithm and go to step (13).

(10) Compute the amplitude distortion E_{am} using Eq. (13) at the design vector $\mathbf{b_{i+1}}$.

(11) Compute the objective function E_{i+1} and ∇E_{i+1} at the design vector $\mathbf{b_{i+1}}$. Also compute matrix $[B_{i+1}]$. If $E_{i+1} \geq E_i$,, choose the optimum point as $\mathbf{b_i}$, stop the procedure and go to step (13). If $E_{i+1} < E_i$, set $E_i = E_{i+1}$, and $\mathbf{b_i} = \mathbf{b_{i+1}}$.

(12) Set the new iteration number as $i = i + 1$, and go to step (5).

(13) The optimum solution is $\mathbf{h_0 = (1/2)\, b_i}$, and stop the procedure.

IV. RESULTS AND DISCUSSION

Design example is presented in this section to illustrate and examine the effectiveness of the proposed algorithm. The performance of the algorithm is evaluated in terms of following important parameters:

Mean square error in the pass band (E_p);

Stop band error (E_s);

stop-band first lobe attenuation (A_L);

stop-band edge attenuation (A_s) = $-20 \log_{10}(H_0(\omega_s))$;

maximum reconstruction error (ϵ) in dB = $\max_{\omega} |10\log|T(e^{j\omega})||$

A MATLAB program has been written which implements the design technique described above.

4.1 Design example:

Example: For filter length $(N+1) = 48$, $\omega_s = 0.6\pi$, $\omega_p = 0.4\pi$, $\alpha_1 = 0.95$, $\alpha_2 = 0.04$, $\alpha_3 = 1$ and $\alpha_4 = 10^{-4}$, the following 24-filter coefficients for the FIR low-pass prototype filter are obtained:

$h_0(0) = 0.000208516673605$ $h_0(1) = -0.000277085005004$

$h_0(2) = -0.000485925710620$ $h_0(3) = 0.000939830506683$

$h_0(4) = 0.000828146074712$ $h_0(5) = -0.002254669530186$

$h_0(6) = -0.001114175660144$ $h_0(7) = 0.004518779506026$

$h_0(8) = 0.001102671159255$ $h_0(9) = -0.008074502198911$

$h_0(10) = -0.000416487855403$ $h_0(11) = 0.013299020942940$

$h_0(12) = -0.001522982128226$ $h_0(13) = -0.020679775981083$

$h_0(14) = 0.005591974419610$ $h_0(15) = 0.030998856797507$

$h_0(16) = -0.013310060758246$ $h_0(17) = -0.045983099224892$

$h_0(18) = 0.027954420876814$ $h_0(19) = 0.070646028734991$

$h_0(20) = -0.059744465465400$ $h_0(21) = -0.126933649477169$

$h_0(22) = 0.172673420719003$ $h_0(23) = 0.593622292543939$

The corresponding normalized magnitude plots of analysis filters $H_0(z)$ and $H_1(z)$ are displayed in Fig. 2a. Figures 2b and 2c, respectively, depict the amplitude of distortion function and attenuation characteristics of low-pass filter $H_0(z)$. The reconstruction error (in dB) of QMF bank is plotted in Fig. 2d. The corresponding important parameters are $E_p = 3.144\times10^{-10}$, $E_s = 2.96\times10^{-8}$, A_s = 55.06 dB, A_L = 64.09 dB, and PRE (ϵ) in dB = 0.0092 dB.

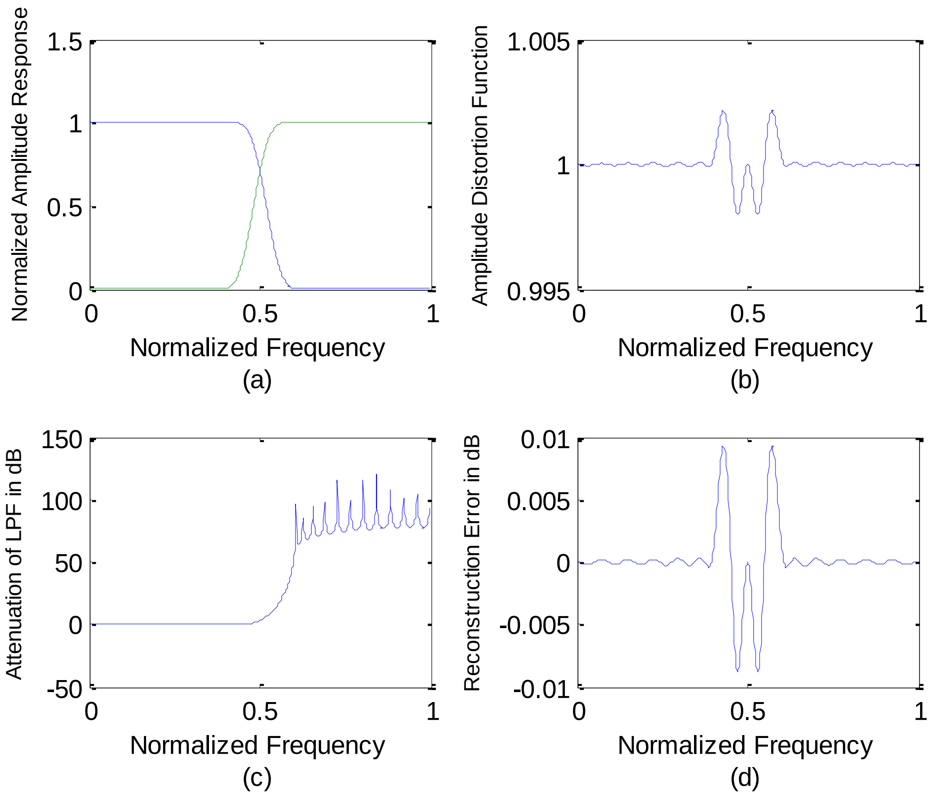

FIGURE 2 (a) Amplitude response of analysis filters for filter length = 48, (b) Amplitude of distortion function, (c) Attenuation characteristics of low-pass analysis filter, (d) Reconstruction error in dB.

From the above results, it is clearly indicate that the performance of the proposed method significantly improved when compared than to earlier known techniques in terms of PRE, E_p and A_s with similar design specifications.

V. CONCLUSION

A new iterative method for the design of multirate filter banks has been developed by formulating the perfect reconstruction condition in the frequency domain. The quadratic objective function is minimized without any matrix inversion which generally affects the effectiveness of some methods. Design example shows that the proposed technique is very effective in designing the quadrature mirror filters. The peak reconstruction error is minimum by the proposed method that makes it suitable for real time applications. Further, it is possible to extend this approach for the design of QMF banks with more than two bands.

CONFLICT OF INTEREST

The authors declare no conflict of interest.

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Enhanced DV-Hop Self-Localization Procedure for Wireless Sensor Networks

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Abstract— *In this paper, a novel self-localization procedure for wireless sensor networks is presented. Due to errors in calculating universal coordinates and inappropriate relative pairwise distance assessment, the errors in the appraisal of localization may occur. Minimization of these errors is necessary for effective localization arrangements. In irregular or sparse networks, the earlier proposed DV-Hop positioning procedure shows poor accuracy and is not effective. To avoid the weaknesses of this procedure, in this work an enhanced DV-Hop procedure for self-localization (EDVHPSL) is proposed. The anticipated EDVHPSL method delivers improvement in results in terms of localization error. MATLAB simulations have been performed to implement the proposed EDVHPSL by changing the benchmark parameters during simulation process. The parameters are number of beacon nodes, communication range, number of sensor nodes and hop count.*

Keywords— Communication range, Wireless Sensor Network (WSN), Sensor Node, Beacon nodes, Localization error minimization.

I. INTRODUCTION

In WSNs, the sensor nodes are deployed in an unplanned infrastructure where there is no prior knowledge of location. The problem of estimating the spatial coordinates of a sensor node (relative or absolute) is referred to as localization [1]. It is a term used to define the process of finding the geographic location of the sensor nodes in a coordinate system. The sensor nodes must be localized in space in order to identify the location of an event. The positioning of sensor nodes is accomplished using a localization system which is a key part of WSNs. Localization systems not only help to locate events but can also be used as a

base for routing, density control, tracking, and network protocols. The straightforward localization approach gathers the information (e.g. connectivity or pair-wise distance measurement) about the entire network into one place, where the collected information is processed to estimate the location of sensor nodes [2]. Localization is an unavoidable challenge when dealing with WSNs and an important problem because many of the sensor network protocols and applications simply suppose that all sensor nodes in the network are aware of their individual locations. Secondly, if a sensor node is reporting a critical event or data by means of geographical routing technique, the location of individual sensor nodes must be known in prior [3].

Existing localization algorithms [4] estimate the locations of sensor nodes either by using knowledge of the positions of a few sensor nodes or their inter-sensor node measurements such as distance and angle. Sensor nodes with known location information are called anchors or beacons and their locations are obtained by using a GPS, or by their manual placement at points with known coordinates. While the GPS is one of the most popular technique and is widely accessible, the high cost, high energy consumption and restricted indoor usage makes it difficult for WSNs. Limitation of size, battery and hardware resources of sensor nodes prohibits the use of GPS hardware in every sensor node. The various localization methods have their own merits and demerits and their performance also depend on many other factors like, accuracy, coverage, complexity, scalability, robustness, fault tolerance, cost and energy.

1.1 Self Localization:

Localization is a mechanism to establish spatial relationships between sensors in a wireless or wired sensor network. Many of these sensor network systems are embedded to monitor or control the behavior of physical systems (as compared with strictly virtual information systems), and therefore sensor nodes often need to determine their action based on their physical location. Networked applications are often implemented in the form of a layered network protocol stack and localization benefits span several layers of the protocol stack. Several issues render the localization problem more challenging for large scale, densely distributed sensor networks than in many other domains. Sensor networks must satisfy several physical constraints. Wireless sensor network consists of a large number of various mini integrated self powered sensor nodes, these nodes communicate wirelessly with neighboring nodes to form a wireless sensor network. The process of computing the physical locations of nodes in a wireless sensor network is known as localization, self localization becomes critical for large scale sensor networks because manual deployment is often impractical due to time requirements, economic constraints or inherent limitations. The spatial relationship of these sensors is typically an important factor and introduces a new challenge for the researchers working in this area.

1.2 Existing DV-Hop Positioning Algorithm Error Analysis:

Errors in localization estimates can result from incorrect estimates of relative pairwise distances and errors in global coordinate calculations. These errors must be minimized for effective localization planning. Existing DV-Hop positioning algorithm is one of the representative algorithms of band-less localization technology [5, 6]. The main idea is to express the distance between the unknown sensor node and the beacon node as the product of the hop distance and the number of hops. Implementation of the algorithm consists of three steps.

In the first step, each beacon node transmits location information to all neighboring sensor nodes. At this time, the hop count is initialized to 0, and each receiving sensor node maintains the beacon node information and the minimum hop count value for each beacon. In the second step, once the sensor node has a relative hop count value to other beacons, it estimates the hop size for a single hop and then broadcasts it throughout the network. According the following equation (1), the hop size is estimated.

$$SizeHop_i = \frac{\sum \sqrt{(x_i - x_j)^2 + (y_i - y_j)^2}}{\sum s_{ij}} \tag{1}$$

wherever, s_{ij} is the total number of hops between the beacon nodes and (x_i, y_i) and (x_j, y_j) are the coordinates of beacon node i and j. In the last step, when the unknown sensor node receives 3 or more distance estimates from the beacon node, it calculates its position using trilateration.

II. ENHANCED DV-HOP PROCEDURE FOR SELF- LOCALIZATION (EDVHPSL)

The DV-Hop positioning algorithm has low accuracy and performs poorly in sparse or irregular networks. Considering the drawbacks of this algorithm, efforts were made to improve it and EDVHPSL is proposed. The assumptions taken into account for EDVHPSL are described in the following steps.

Step 1: Statistics Broadcasting

Each guidance node transmits position information to neighboring sensor nodes with a hop count initialized to one. The format of the broadcast information is (idi, xi, yi, sHopi), where idi is the ID of the beacon node and (xi, yi) are its coordinates. Each receiving sensor node maintains a record of the information received from the guidance node. If a sensor node receives a packet with the same ID, then it compares the sHopi of the packet and if the new sHopi is smaller than the already received sHopi table, it is considered to be otherwise dropped and the packet is no longer forwarded. Each sensor node will increase sHopi by one before transmitting it to other neighboring sensor nodes

Step 2: Calculation of Distance

As per equation (1), the hop size is calculated to estimate the distance between the beacon node and unknown sensor node.

The average hop size is calculated as per equation (2).

$$SizeHope_{avg} = \frac{\sum SizeHop_i}{Nos.of\ HopSize} \tag{2}$$

As per equation (3), the distance W_i between the beacon node and unknown sensor node is considered by multiplying by its hop count with the average hop size

$$W_i = SizeHop_{avg} \times Hopcount_i \tag{3}$$

Step 3: Calculation of Location

According to the distance information obtained with respect to the beacon nodes of the unknown sensor node, the unknown node calculates its coordinates using trilateration as depicted in Fig. 1.

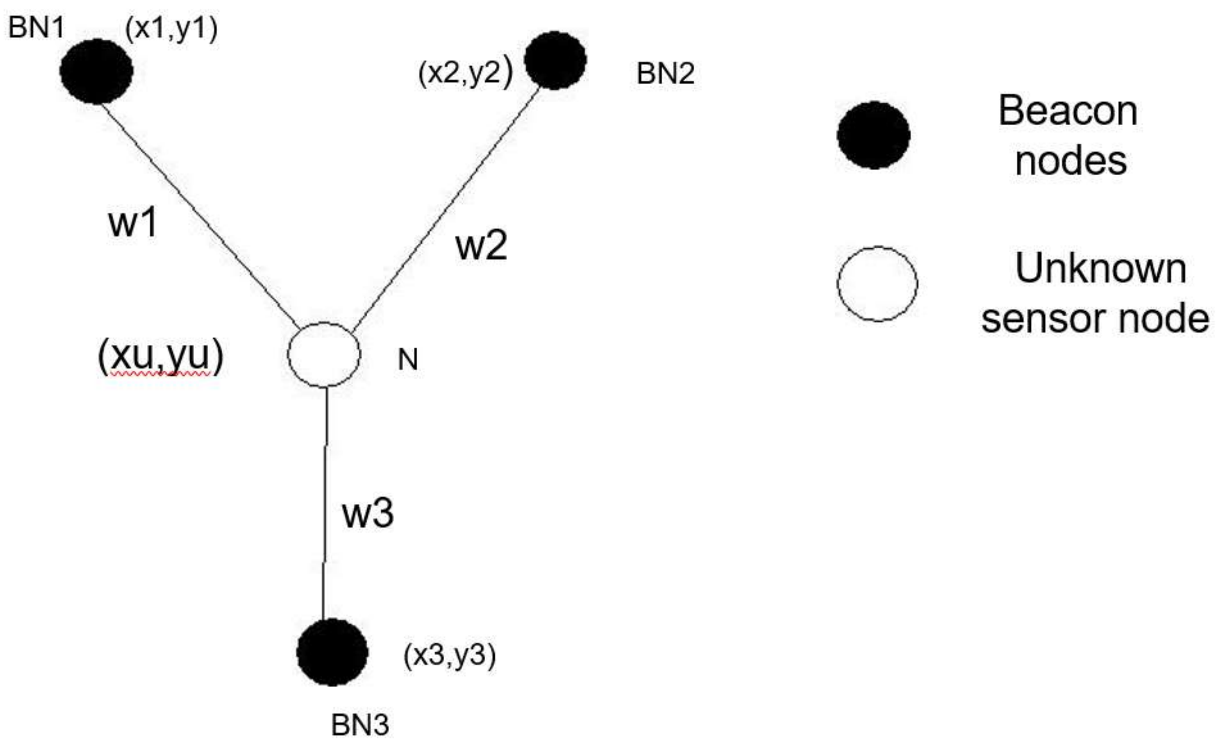

FIGURE 1: Trilateration method based calculation

Step 4: Assessment of localization error

Using equation (4), the localization error can be modestly predictable.

$$E_i = \frac{\sum \sqrt{(x_i - x_j)^2 + (y_i - y_j)^2}}{L} \tag{4}$$

Where, (x_i, y_i) and (x_j, y_j) are the coordinates of beacon node *i* and *j*. and L is the communication range.

The algorithm of the proposed EDVHPSL is as follows;

Steps: Algorithm: EDVHPSL

START

1: Area A is used to deploy sensor nodes

2: Nos. of beacon nodes = *BN* are defined

3: Position of beacon nodes acquires

4: BN coordinates are (x_i, y_i) then send this positions to all adjoining sensor nodes

5: Hop count will be compared

6: Case 1**:** hop count is less assent the position packet acknowledged from the beacon

7: Case2: Else, discard

8: Wait for more position packets, appraisal the hop size between beacon node *BN* and *N*

9: Determine $SizeHope_{avg} = \frac{\sum SizeHop_i}{Nos.of\ HopSize}$

10: Refer $SizeHope_{avg}$ to all neighbours

11: Estimate $W_j = s_j \times SizeHop_{avg}$

12: Unknown sensor node position is computed

13: Determine $E_i = \frac{\sum\sqrt{(x_i - x_j)^2 + (y_i - y_j)^2}}{L}$

END

III. EDVHPSL SIMULATION RESULTS

EDVHPSL is realized in MATLAB 7, and to verify the performance of the proposed algorithm, 500 nodes (beacon nodes = 100 and sensor nodes = 400) are randomly distributed in a 100m x 100m area. All nodes (beacons and sensor nodes) have an adjustable communication range L. The layout of beacon nodes and sensor nodes is shown in the fig. 2.

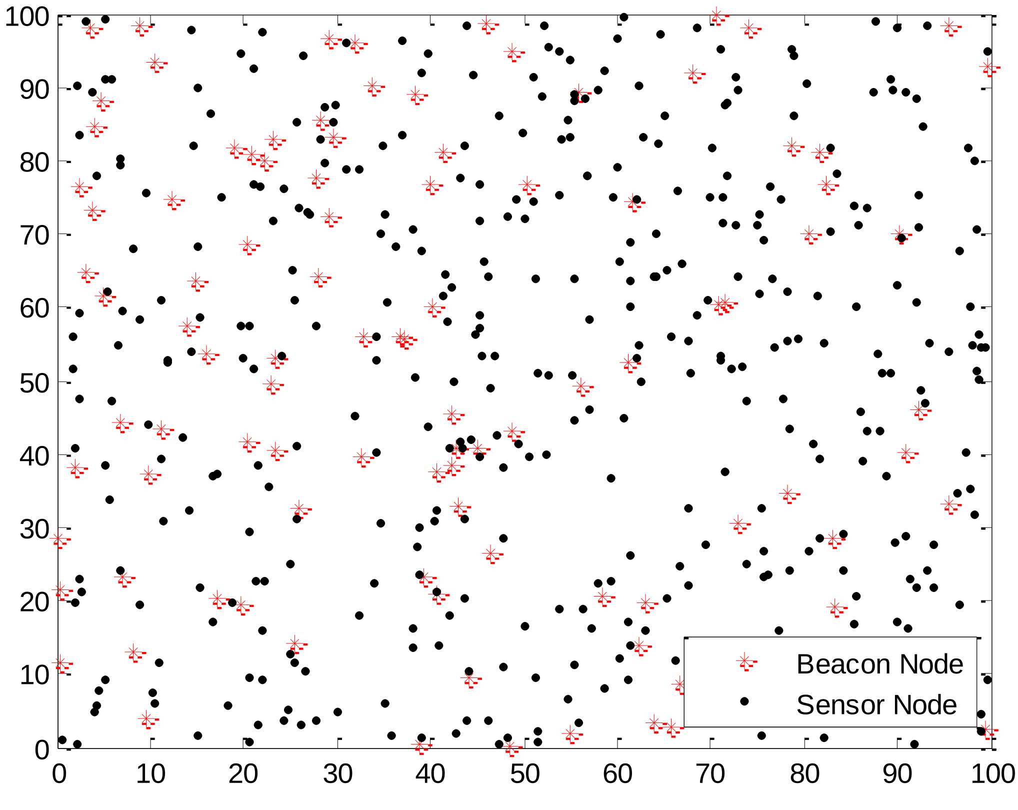

FIGURE 2: Deployment randomly sensor nodes and beacon nodes

Simulations has been done by variations in number of beacon nodes and communication range. The enactment of the DV-Hop positioning procedure and the proposed EDVHPSL are matched.

3.1 Communication Range Varying Effect:

Observance the number of beacon nodes = 100 and the number of sensor nodes = 400, the communication range is changed to 80m, 70 m, 60 m, 50 m, 40 m, 30 m,20 m, and 10 m, and the performances of DV-Hop and proposed EDVHPSL positioning algorithm are analysed. . It can be seen from Fig. 3 that the localization error of EDVHPSL is lower than that of the DV-Hop positioning algorithm. The reasons for the low errors are due to the fact that at the maximum communication range (80 m) most nodes are able to directly communicate with each other in a simpler way.

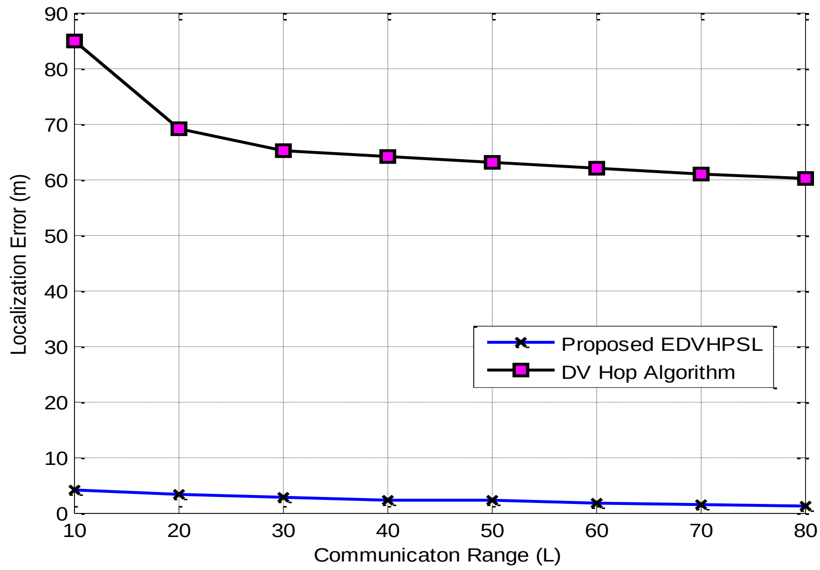

FIGURE 3: Localization error verses communication range

3.2 Beacon Nodes Varying Effect:

Number of beacon nodes are changed to 100, 90, 80, 70, 60, 50, 40, 30, 20 and 10 while keeping L = 50 m, simulations are performed for DV-Hop and EDVHPSL positioning algorithm. Fig. 4 depicts that the localization error of the proposed EDVHPSL is lower than that of the DV-Hop positioning algorithm, if the beacon nodes are increased.

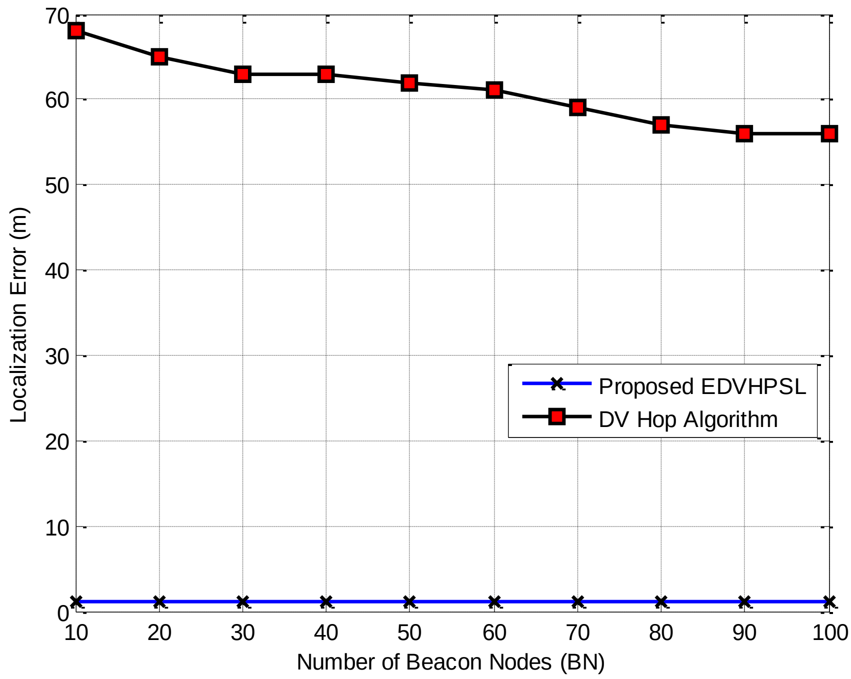

FIGURE 4: Localization error verses beacon nodes

IV. CONCLUSIONS

In this paper, the causes for the localization error in the existing DV-Hop positioning algorithm were described and an enhanced DV-Hop positioning algorithm is proposed. The proposed EDVHPSL technique, which is a amendment in the DV-

Hop positioning algorithm, delivers upgraded results in terms of localization error. During the simulation process, EDVHPSL is implemented by altering essential parameters named as hop count, range, number of beacon nodes, and number of sensor nodes.

CONFLICT OF INTEREST

The authors declare no conflict of interest.

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