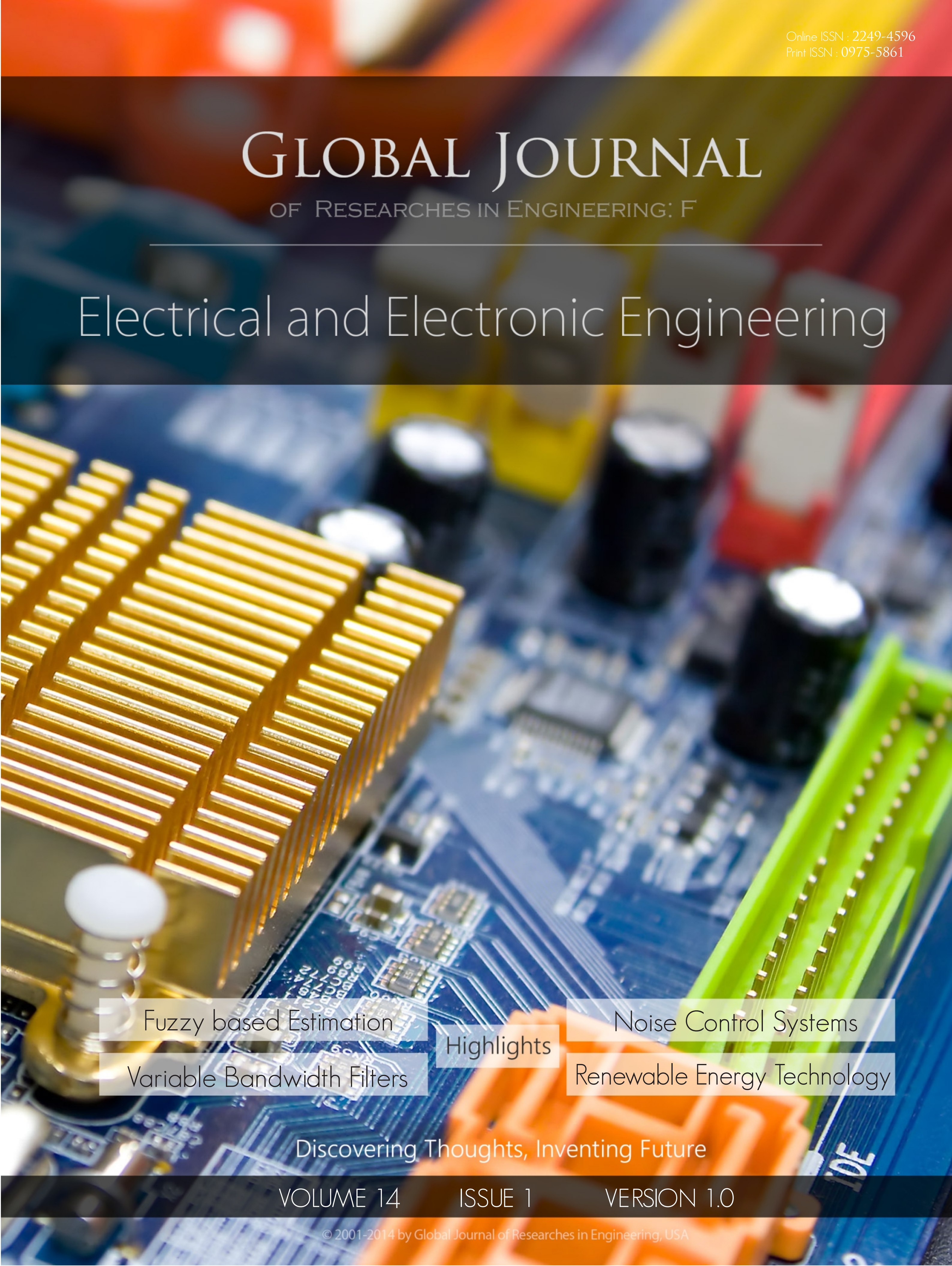

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Variable Bandwidth Filters

Highlights

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Discovering Thoughts, Inventing Future

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By James T. George & Elizabeth Elias
National Institute of Technology, India

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James T. George [α] & Elizabeth Elias [σ]

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I. Introduction

The main task of a hearing aid is to provide audibility, namely to amplify the required input signals to the levels that the hearing impaired can hear. An audiogram with frequencies on the X axis and hearing threshold on Y axis indicates the hearing capability of a person for various frequencies. An ideal hearing aid device should include features such as adjustable magnitude response on arbitrary frequencies, low power consumption for battery life, low processing delay to take advantage of lip-reading, linear phase to prevent the audio signal from distortion, small in size and weight for cosmetic appeal, noise reduction, programmability etc. (Yong and Ying, 2005). All these can be achieved by using the digital signal processing approach compared to the analog signal processing for implementing the hearing aid (Luo and Horst, 2002). Even though, the signal processing algorithms have dramatically improved in recent years, common hearing aid fitting procedures are still focusing on frequency bands and corresponding gain of the filters to match the audiogram of the specific user with individual hearing loss (Kuo et. al., 2008). So the basic requirement of a hearing aid is to provide precise adjustment of the gain versus frequency, which is also called audiogram matching/fitting. Sometimes, relatively minor changes may have a significant influence on performance, in particular with regard to speech intelligibility.

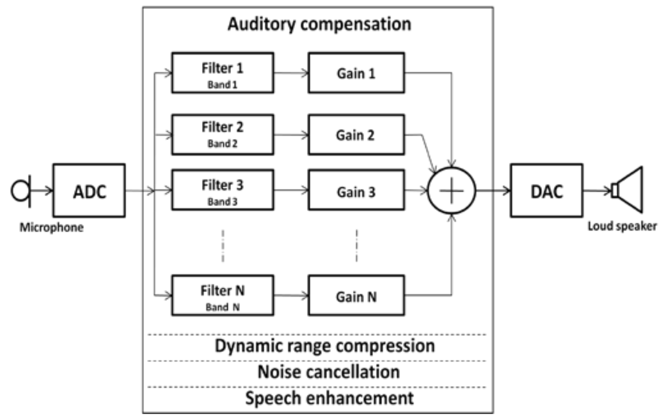

Figure 1 : Block diagram of digital hearing aid

Figure 1 shows the block diagram of generic hearing aid, including functional blocks such as auditory compensation, echo (feedback) cancellation, noise reduction/suppression, speech enhancement etc. (Kuo et. al., 2008). Auditory compensation is the main function in hearing aids, which performs frequency shaping to compensate for the hearing loss. An individual filter in the filter bank decomposes the input signal into (uniform or non-uniform) frequency bands. A low pass filter outputs the lowest frequencies, a high pass filter outputs the highest frequencies and a set of band pass filters output the remaining intermediate frequencies as shown in Figure 2, so that the prescribed gain based on the audiogram or hearing threshold can be applied to the different frequency bands to suit the needs of the hearing impaired people, or in other words the amplitude response of the filter bank should equalize or match the audiogram (McAllister et. al., 1994).

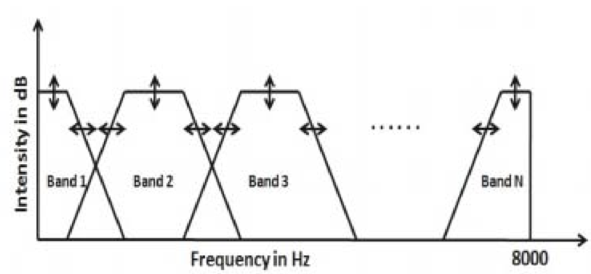

Figure 2 : General filter structure

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Most of the currently available hearing aid designs provide the filter bank with fixed bands (uniform or non-uniform). Thus the patients cannot take the full advantage to improve their specific auditive performance by using the hearing aid with limited number of fixed bands. This reduces the potential flexibility in fitting/matching of hearing loss with steeply sloping audiograms. One method of improving the same is to use an instrument with higher number of frequency bands for matching the audiogram with minimum matching error. But such an implementation would not only require a large amount of processing power but also increases the cost. Modern hearing aids with more than/up to 32 bands are available in the market. Therefore it would be very useful to design a filter structure with less number of bands and that can be easily customized with minimum change in parameters, for any hearing impaired person.

In this paper, a new method is proposed to realize a uniform as well as non-uniform filter structure for the hearing aid application with the facility for changing the band edges and gain to match the audiogram without any change in the number of bands and filter coefficients of the fixed filters associated with each band. Variable bandwidth filter with facility for the reduction and enhancement of the bandwidth without changing the filter order or filter coefficients, based on changing the sampling frequency of the input signal, is used for achieving the same (Fred, 2009, George and Elias, 2012). It is accomplished by interpolating the input series to obtain an intermediate time series at a new sample rate proportional to the required bandwidth. The interpolated series are processed by the fixed length FIR filter, now operating at the changed sample rate. The output of the filtered time series is then interpolated back to the original input sample rate, which effectively changes the bandwidth. The change in bandwidth of the particular band can be achieved by modifying only the bandwidth ratio given as input to the sample rate converter. All the circuits associated with each band can be implemented in fixed hardware also. With the help of the variable bandwidth filter approach, a wide range of hearing loss compensation can be achieved by using the same fixed hardware. In this case, the fixed filter and the associated sample rate converters can be implemented as fixed hardware in application specific integrated circuit (ASIC) and the programmable portion of the processing can be implemented in a digital signal processor/ micro-controller, thereby achieving small chip size and low power consumption (McAllister et. al., 1994).

The paper is organized as follows. Section 2 gives a brief overview about the audiogram and the common hearing loss patterns. Section 3 presents a review of the variable bandwidth filter. Section 4 gives the design example of the fixed filter and the structure of a 16 band non-uniform filter bank. The performance evaluation and comparison of the results are given in Section 5 and Section 6 concludes the paper.

II. Overview of Audiogram and Hearing Loss

An audiogram is a graph the hearing test is marked on, that shows the softest sound a person can hear at different pitches or frequencies. The Y axis represents the intensity or response of the ear measured in decibels (dB) corresponding to the frequency in hertz (Hz) marked on the X axis. An `O' often is used to represent the responses for the right ear and an `X' is used to represent the responses for the left ear. Curves displayed in decibels generally describe the individual hearing threshold of a person compared to the normal hearing average, which lies around 0 dB. Due to individual differences, all thresholds up to 20 dB are considered as normal. Figure 3 shows the audiogram of a normal hearing person. The threshold between 21 to 40 dB is considered as mild hearing loss, 41 to 55 dB is considered as moderate hearing loss, 56 to 70 dB as moderately severe, 71 to 90 dB is considered as severe hearing loss and greater than 90dB is considered as profound hearing loss.

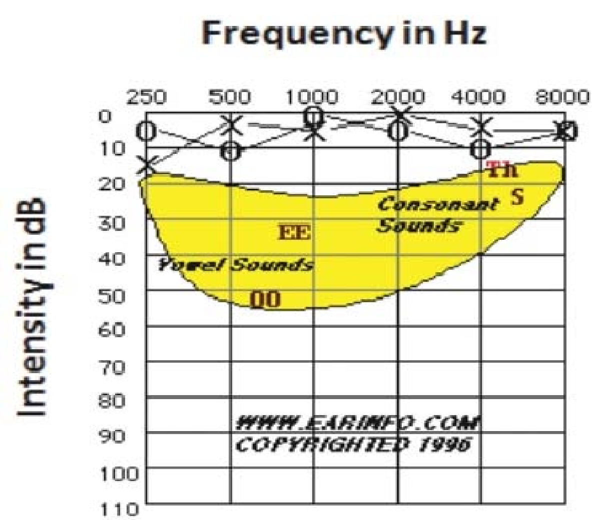

Figure 3 : Audiogram for normal hearing(www.earinfo.com)

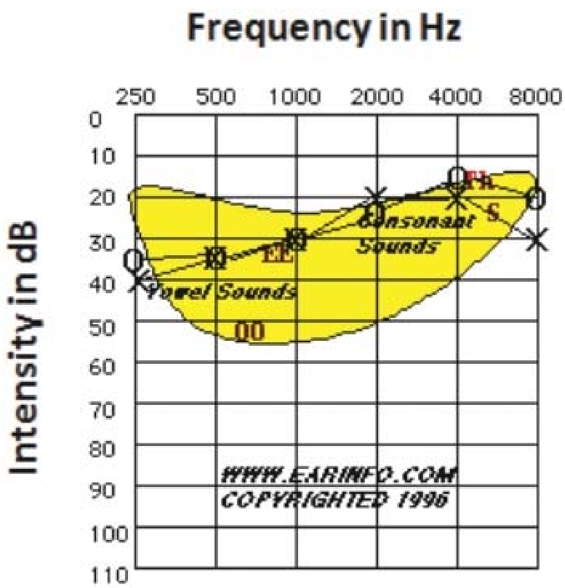

Figure 4 : Audiogram for mild to moderate hearing loss at low frequencies (www.earinfo.com)

Loudness level and frequencies of different speech sounds are also presented in audiograms. Because of the shape of the speech area, it is also referred to as `speech banana'. Vowels are low frequency sounds with a higher volume than consonants, which are soft high frequency sounds. The vowels carry the loudness impression of speech whereas consonants carry the meaning.

Hearing level measurements are done at each octave 250/500/1k/2k/4k/8k in a standard audiogram, which suggests that the uniform filter bank may face difficulties in matching the audiogram at all frequencies (Yong and Ying, 2005). A typical hearing loss shown in Figure 4, caused by aging, occurs at low frequencies. To achieve a better compensation, narrower bands with gain control need to be allocated at frequencies where

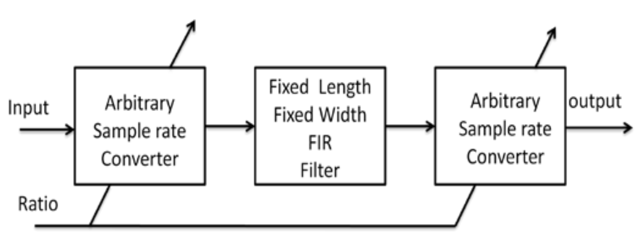

Figure 5 : Functional block diagram of variable bandwidth filter

the slope is steep. Therefore, a non-uniformly spaced digital filter bank structure with varying band edges using variable bandwidth filter becomes very useful. Several methods are described in the literature to realize the filter structure for hearing aid applications using combinations of techniques such as infinite impulse response (IIR), Finite impulse response (FIR), frequency response masking (FRM), frequency transformation, uniform and non-uniform filter banks etc. (Ito et. al., 2010, Deng, 2010, Brennan and Todd, 2001, Chong et. al., 2006, Lim, 1986, Ying and Yong, 2006, Ying and Debao, 2011), but all of them have fixed band edges with particular design. The next section explains the operation of the variable bandwidth filter for the hearing aid application with provision for re-configurability in band edges (George and Elias, 2012).

III. Review of Variable Bandwidth Filter

There are several methods (Oppenheim et. al., 1976, Roy and Ahuja, 1979, Hazra, 1984, Carson et. al., 2002, Johansson and Lowenborg, 2004, Mahesh and Vinod, 2011) available for the design of variable bandwidth filters for signal processing and communication application. Most of them require changes in the filter coefficients for achieving the revised frequency characteristics. For hearing aid application, the FRM based technique also requires modified structural design for getting the new frequency characteristics. But the fractional variation of the bandwidth (reduction as well as enhancement) of the fixed bandwidth prototype filter, without any change in the coefficients can be achieved by changing the sampling rate of the input signal (Fred, 2009, George and Elias, 2012).

In the case of constant form factor FIR filter, the length of the filter is inversely proportional to the transition bandwidth, which is proportional to the bandwidth of the filter. Thus if the bandwidth of the filter is changed by a factor, the length of the filter is also changed by the same factor.

In order to realize a variable bandwidth filter without changing the number of coefficients or coefficient values, we can utilize the relationship between the sampling rate and bandwidth (Fred, 2009). If the number of taps of the filter is fixed, one method of changing the bandwidth is to change the sampling rate. So we can change the absolute bandwidth of a filter by operating it at a different sampling rate. The technique proposed by us in George and Elias, 2012 is used for changing the bandwidth continuously in the decreasing and increasing fashion.

The implementation of the process is represented in Figure 5. An input signal which is initially oversampled is applied to an arbitrary sample rate converter (up or down). The modified signal is signal is processed by the fixed length, fixed bandwidth FIR filter and the output of the filter is then converted back to the original input sampling rate by using another arbitrary sample rate converter (down or up). This effectively changes the bandwidth.

Figure 6 shows the implementation of an arbitrary sample rate converter (Fred, 2004). Three M - path polyphase filters are used for calculating the sample values of the interpolant and the sample

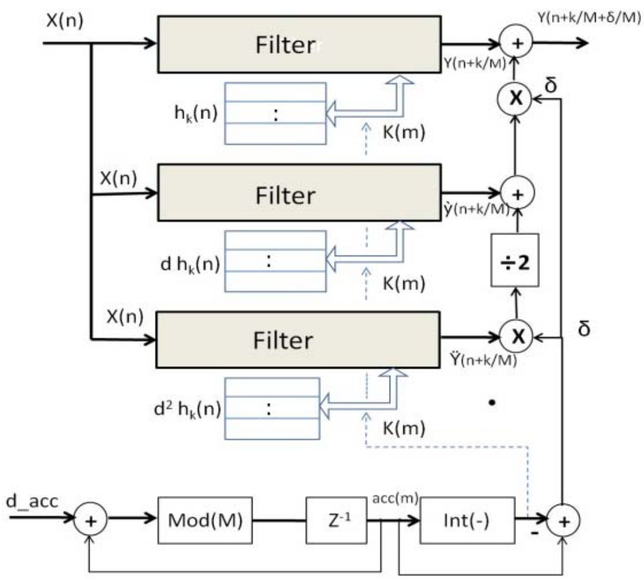

Figure 6 : Implementation of arbitrary sample rate converter

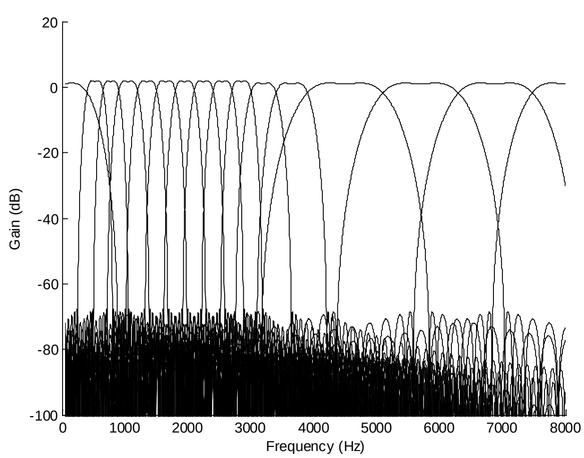

Figure 7 : Frequency response of the uniform filter bank

derivatives at the offset position k/M from the interpolating output phase centre. The computed output is formed from a local Taylor series as given by Equation (1).

$$y\left(n+\frac{k\delta}{M}+\frac{}{M}\right)\cong y\left(n+\frac{k}{M}\right)+\acute{y}\left(n+\frac{k}{M}+\frac{\delta}{M}\right)+\frac{2}{2!}\ddot{y}\left(n+\frac{k}{M}\right)+\ldots \quad (1)$$

The increment *d_acc* satisfies the following relation shown in Equation (2) and the required bandwidth can be achieved by just changing the parameter *d_acc* as shown in Equation (3).

$$\frac{d\_acc}{M}=\frac{T_{out}}{T_{in}}=\frac{fs_{in}}{fs_{out}}=\textit{Arbitrary sampling ratio} \quad (2)$$

$$d\_acc=M\ *sampling\left(bandwidth\right)ratio \quad (3)$$

Where

M - Number of polyphase filters

fs_{in} - Input sampling rate to interpolator

fs_{out} - Output sampling rate of interpolator

IV. Design Example

Finite impulse response (FIR) filters are widely accepted in many areas of signal processing, communication and bio-medical applications including hearing aid due to their exact linear phase and high stability under certain conditions. In the design of the variable bandwidth filter, we should preserve the finite length and linear phase for a desired set of frequency characteristics. Therefore priority should be given for the optimal design of the fixed bandwidth FIR filter.

Let us consider a typical example for designing a prototype fixed bandwidth FIR filter for a 16 band hearing aid to support up to 8 kHz. The design specifications of the low pass FIR filter are shown below.

Bandwidth : 269 Hz

Maximum Pass band ripple: 0.05dB

Minimum stop band attenuation: 65dB

Sampling frequency: 16kHz

Table 1 : Bandwidth, ratio and frequency shift of sub-bands

Band	Uniform			Non-uniform case-1([12])			Non-uniform case-2		
	Band-width (Hz)	Shift (Norma-lised)	Band-width Ratio	Band-width (Hz)	Shift (Norma-lised)	Band-width Ratio	Band-width (Hz)	Shift (Norma-lised)	Band-width Ratio
1	269	0	1	250	0	0.9375	400	0	1.5001
2	533	0.0335	1	250	0.0234	0.4688	250	0.0328	0.4688
3	533	0.0668	1	250	0.0391	0.4688	250	0.0484	0.4688
4	533	0.1001	1	250	0.0547	0.4688	300	0.0656	0.5625
5	533	0.1334	1	500	0.0781	0.9375	300	0.0844	0.5625
6	533	0.1667	1	500	0.1094	0.9375	300	0.1031	0.5625
7	533	0.2000	1	1000	0.1563	1.8750	300	0.1219	0.5625
8	533	0.2333	1	1000	0.2188	1.8750	300	0.1406	0.5625
9	533	0.2667	1	1000	0.2813	1.8750	300	0.1594	0.5625
10	533	0.3000	1	1000	0.3438	1.8750	300	0.1781	0.5625
11	533	0.3333	1	500	0.3906	0.9375	400	0.2000	0.75
12	533	0.3666	1	500	0.4219	0.9375	500	0.2281	0.9375
13	533	0.3999	1	250	0.4453	0.4688	1200	0.2813	2.25
14	533	0.4332	1	250	0.4659	0.4688	1200	0.3563	2.25
15	533	.4665	1	250	0.4766	0.4688	1175	0.4305	2.203
16	269	0.5	1	250	0.5	0.9375	525	0.5	1.968

Sampling frequency: 16kHz

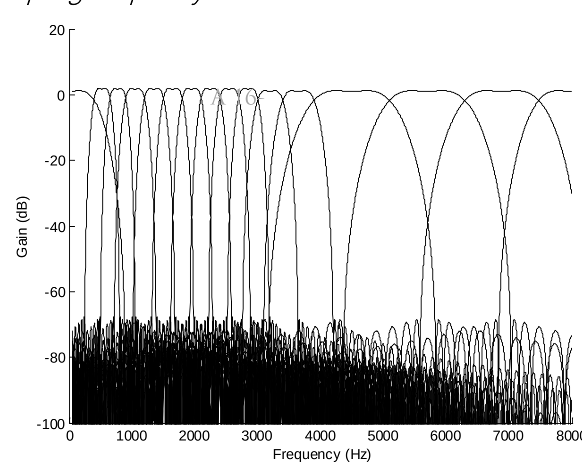

Figure 8 : Frequency response of the non-uniform case-2 filter bank

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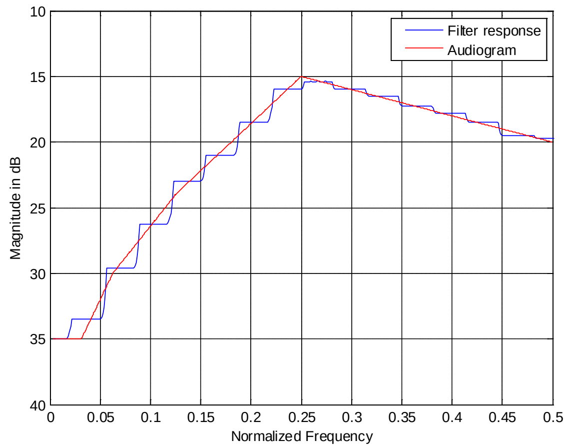

Figure 9 : Audiogram fitting for the uniform filter bank

The bandwidth, corresponding shift in centre frequency and bandwidth ratio for the various bands in the uniform and two examples/cases of the non-uniform bandwidth selection are given in Table 1. The frequency response of the filters in the uniform and non-uniform cases based on Table 1 are shown in Figure 7, 8 and 9 respectively.

The bandwidth of the particular band can be changed by modifying the bandwidth ratio *d_acc* and the corresponding centre frequency shift. The centre frequency shift can be achieved by using the spectrum shifting property of the Fourier Transform (Vaidyanathan, 1990).

V. Audiogram Matching/Performance Evaluation/Result and Discussions

In order to evaluate the performance of the 16 band filter bank for the hearing aid application, the structure is tested by using MATLAB 7.10.0 on a Toshiba satellite L750 laptop with Intel (R) core (TM) i5 2410M processor operating at 2.3 GHz. Various audiograms for the common types of hearing loss are

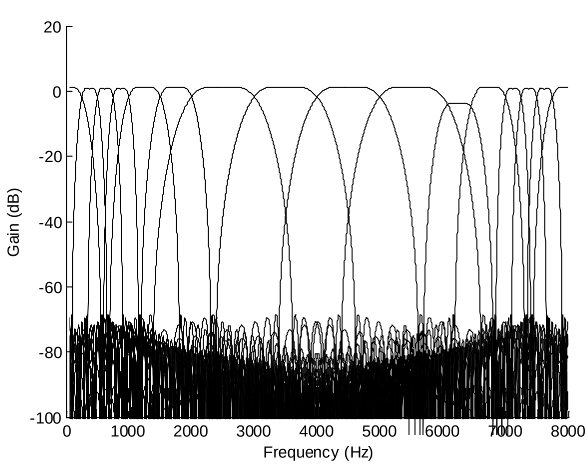

Figure 10 : Frequency response of the non-uniform case-1 filter bank

used for evaluating the effectiveness of the filter bank. It is observed that the matching error between the filter output and the audiogram is very small.

The results corresponding to the audiogram shown in Fig 4 are used for presenting the output. The purpose in selecting the mild to moderate hearing loss at low frequencies is that the audiogram contains steep slope up to the middle of the audio range. The audiogram fitting/ matching with the 16 band uniform and non-uniform cases based on the data given in Table 1. along with the selected audiogram are given in Fig 10, 11 and Fig 12 respectively. The corresponding errors between the audiogram and the magnitude responses of the filter banks in all the three cases are given in Fig 13. The advantage of the proposed method is that, the same hardware setup can be configured as uniform or non-uniform frequency bands based on the selection of the bandwidth ratio and frequency shift. The proper selection of the bandwidth or bandwidth ratio and frequency shift will decide the quality of the output.

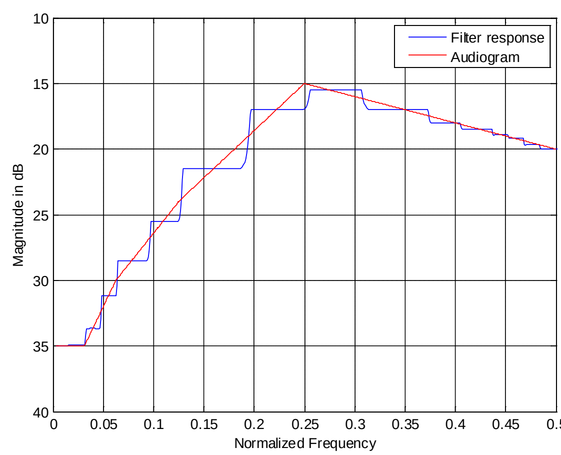

Figure 11 : Audiogram fitting for the non-uniform filter bank (case-1)

For verifying/analyzing the performance of the proposed method, the number of bands and bandwidth corresponding to each band are selected as given in Ying and Yong, 2006 for the non-uniform case -1. A modified selection of bandwidth for reducing the matching error is proposed as case-2 in Table 1. The matching error corresponding to each case is given in Table 2. In the case of uniform bands, the only possibility for reducing the matching error is to adjust the gain of the bands. Because of the wider bandwidth provided in the mid frequency region having large slope, further reduction in the matching error is not possible with the band allocation as specified in Ying and Yong, 2006 and given as case-1 in Table 1. In the case of non-uniform case-2, the control of band gain and selection of smaller bandwidth in those positions where the slope is steep, can reduce the matching error considerably as shown in Table 2.

Table 2 : Matching errors of the uniform and non-uniform filter banks

Maximum matching error in dB		
Uniform	Non-uniform (case-1)	Non-uniform (case-2)
1.59	2.1	1.08

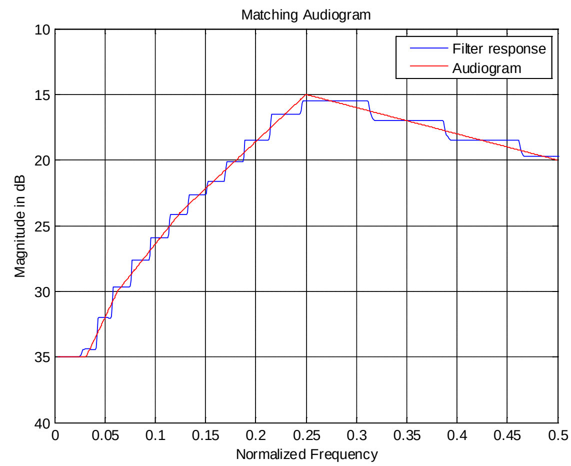

Figure 12 : Audiogram fitting for the non-uniform filter bank (case-2)

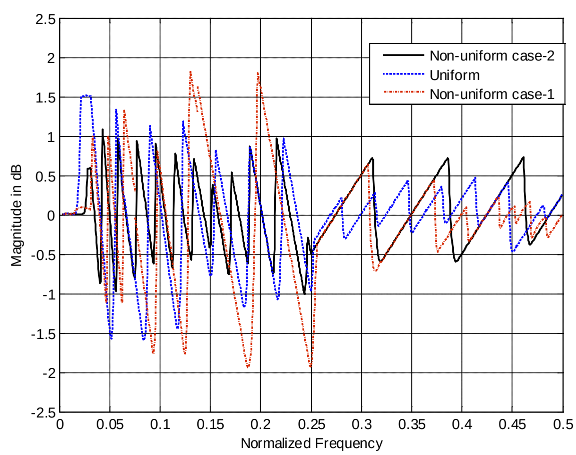

Figure 13 : Comparison of matching errors

VI. Conclusion

In this paper a technique is presented to obtain a 16 band uniform or non-uniform spaced filter bank using variable bandwidth filter to suit the fitting of audiogram for the hearing aid application. Each band is implemented as a combination of two arbitrary sample rate converters and a fixed bandwidth FIR low pass filter, which can be implemented in a fixed hardware. The advantage of the proposed scheme is that, the bandwidth and the position of the band can be configured by modifying the bandwidth ratio and frequency shift, based on the audiogram of the hearing impaired, at the time of hearing aid fitting, by an audiologist. This part of the hardware can be implemented in a FPGA/DSP/micro-controller part associated with the programmable hearing aid. The performance of the filter bank is evaluated with various audiograms for common types of hearing loss and the results are comparable with available/existing ones in the literature. Miniaturization, reduction in power consumption, programmability and affordable delay can be achieved by the implementation of the major part in the fixed hardware. Hearing aid fitting becomes easier and re-configuring the bands is also possible with minimum modification in the programmable part. Hence the same hearing aid can be easily re-programmed for the changes which can happen in the audiogram due to the aging of the user. Thus the re-usability of the hearing aid can be improved by the re-configurability.

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Fuzzy based Estimation of Low Cost Sensor- Less Control of Brushless DC Motor

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Abstract- This paper proposes a design for position control of sensor-less Brushless DC (BLDC) motor drive by means of the back electromotive force (EMF) method. A fuzzy controller based on regenerative observer is employed to control the BLDC motor drive. Most of the existing sensor-less methods have low performance at transients and low speed range. The controller is designed to overcome this problem. The whippings are avoided by the proposed controller using fuzzy switching gain adjustment. Additionally, the model for BLDC motor is also derived. Simulation results confirm the better performance and higher efficiency of the proposed model.

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FUZZYBASEDESTIMATIONOFLOWCOSTSENSORLESSCONTROLOFBRUSHLESSDCMOTOR

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Fuzzy based Estimation of Low Cost Sensor-Less Control of Brushless DC Motor

Mohammed S. Al-Numay [α] & NM. Adamali Shah [σ]

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I. Introduction

Recently, DC motors have been gradually replaced by BLDC motors due to their attractive features of high starting torque, high efficiency, low maintenance cost and compactness. The efficiency is likely to be higher than DC motor of equal size and more reliable due to the absence of commutator and brushes [1]. Hence the lateral stiffness of the motor is increased, allowing for high speed. When compared to the permanent magnet DC servo motor, BLDC motor has low inertia, large power to volume ratio, and low noise for the same output rating [2]. Therefore, due to high performance, BLDC motor drive has vast applications such as computers, robotics, automation, electric vehicles etc. The maximum speed of the BLDC motor is limited by the retention of the magnet against the centrifugal force alone. The power electronic converters required with BLDC motor are similar in topology to the PWM inverters used in induction motor drives. The device rating may be lower, if only a constant torque characteristic is required. The features of adjustable speed BLDC drive include energy saving, velocity or position control and amelioration of transients. However, BLDC motor still suffers from the extra mechanical position sensor for proper commutation. As a result, when a disturbance occurs on the position sensor, BLDC motor will run unsteady, and noise is produced. Additionally, the position sensor is easily damaged and poses difficulty in repair. Thus the cost of BLDC motor also increases due to the presence of the position sensor. Therefore, research on position sensor-less control for BLDC motor has become focus in the recent years [3]- [5]. In order to eliminate the position sensor, many position sensors-less control methods of BLDC motor with trapezoidal back EMF have been proposed in the literature over the last two decades [6]-[9]. The existing sensor-less drive methods of BLDC motor which are being widely used now have low performance in a transient state or low speed range and occasionally require additional circuits.

To overcome this drawback, fuzzy logic technique is employed to estimate the back EMF in order to improve the performance of the system. A BLDC motor is highly coupled nonlinear multivariable system. Since it is difficult to obtain an accurate mathematical model, fuzzy controller is used rather than the classical Proportional-Integral-Differential (PID) controller. The classical PID controller need accurate mathematical model and perform well under linear conditions.

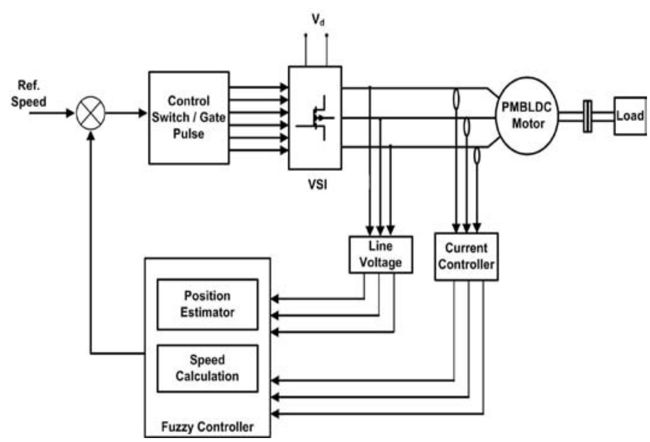

Figure 1 : Block diagram of proposed system

The fuzzy logic controller (FLC) is indeed capable of providing the high accuracy required by high performance drive system without the need of mathematical model. FLC accommodates nonlinearity without utilization of mathematical model. The FLC uses fuzzy logic as a design methodology which can be applied in developing nonlinear system for embedded control. Simplicity and less intensive mathematical design requirements are the most important features of the FLC. These features allow expeditious implementation of the controller, using inexpensive hardware technology. Fuzzy control is a real time controller using fuzzy rule base. Figure 1 shows the block diagram of proposed system. This system

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consists of BLDC motor, six-step inverter, gate drive of inverter, fuzzy controller and switching logic. In a fuzzy control of BLDC motor, the control accuracy is high and the response time is short. So, it is effective to control the speed of the motor.

In this paper, a fuzzy control is employed to improve the dynamic response and reduce the steady state error of the system. Due to the presence of parameter variation and load disturbance in a BLDC motor, closed loop control is necessary to obtain a desirable behavior. BLDC motor has three phase windings on stator and permanent magnet on rotor. In order to control the speed of the motor, rotor position is estimated by the proposed fuzzy control technique. Thus the exact back EMF estimation senses the position and speed of the BLDC motor. The estimated back EMF can measure the error at low speed which is the main drawback of existing system. Additionally, the proposed control technique can estimate the speed of the rotor continuously at transients as well as steady state even with changes in the external conditions.

Based on establishing the mathematical model for the BLDC motor and control system, the fuzzy controller for the position sensors-less motor drive was developed and simulated for different conditions. The simulation results confirm the better performance and higher efficiency of the proposed model.

II. System Design for Position Sensors-Less BLDC Motor

The control system of position sensors-less BLDC motor is shown in figure 1. The PWM based inverter topology is designed with six-switch voltage source configuration with constant dc-link voltage V_d.

For analysis and simplification, the following assumptions are made:

- The motor magnetic saturation is neglected.
- Stator resistances of all the windings are equal and self and mutual inductances are constant.
- Iron losses are negligible.
- The power switches are ideal.

The BLDC motor employed in this study is designed to generate trapezoidal back EMF in the stator terminal. The equivalent circuit topology of BLDC motor is shown in figure 2.

The model of BLDC motor involves solving many simultaneous differential equations; each one depends on the inputs to the motor and the constant parameters. In addition the model provides for dialogue boxes that can be used to vary the values of these constants. The state space equation for the BLDC motor model is derived as follows.

$$\begin{bmatrix} Va \\ Vb \\ Vc \end{bmatrix} = \begin{bmatrix} Ra & 0 & 0 \\ 0 & Rb & 0 \\ 0 & 0 & Rc \end{bmatrix} \begin{bmatrix} ia \\ ib \\ ic \end{bmatrix} + p \begin{bmatrix} La & 0 & 0 \\ 0 & Lb & 0 \\ 0 & 0 & Lc \end{bmatrix} \begin{bmatrix} ia \\ ib \\ ic \end{bmatrix} + \begin{bmatrix} ea \\ eb \\ ec \end{bmatrix} \qquad (1)$$

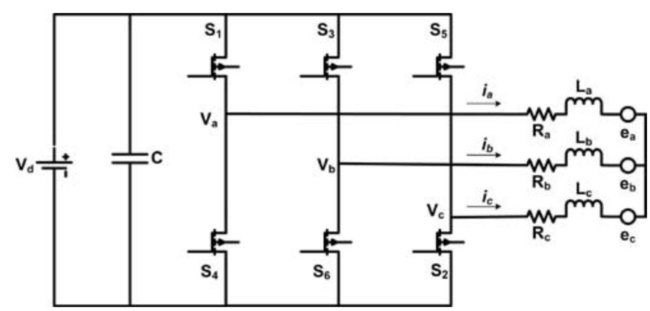

Figure 2 : BLDC motor configuration

The stator resistance per phase is assumed to be equal for all the three phases, therefore $R_a = R_b = R_c = R_s$

The induced back EMF's are all assumed to be trapezoidal, whose peak value is given by

$$E_P = (BLv)N = N(BLr\omega) = N\phi\omega = \lambda\omega \qquad (2)$$

Where λ is the flux linkage and ω is the angular velocity.

$$\begin{bmatrix} Va \\ Vb \\ Vc \end{bmatrix} = \begin{bmatrix} Rs & 0 & 0 \\ 0 & Rs & 0 \\ 0 & 0 & Rs \end{bmatrix} \begin{bmatrix} ia \\ ib \\ ic \end{bmatrix} + p \begin{bmatrix} La & 0 & 0 \\ 0 & Lb & 0 \\ 0 & 0 & Lc \end{bmatrix} \begin{bmatrix} ia \\ ib \\ ic \end{bmatrix} + \begin{bmatrix} ea \\ eb \\ ec \end{bmatrix} \qquad (3)$$

Where V_a, V_b and V_c are phase voltages. If there is no change in rotor reluctance with angle because of non-salient rotor and assuming three symmetric phases, inductances and mutual inductances M are assumed to be symmetric for all phases, i.e. (3) becomes:

$$\begin{bmatrix} V_a \\ V_b \\ V_c \end{bmatrix} = R_s * \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} I_a \\ I_b \\ I_c \end{bmatrix} + \mathbf{p} \begin{bmatrix} L & M & M \\ M & L & M \\ M & M & L \end{bmatrix} \begin{bmatrix} I_a \\ I_b \\ I_c \end{bmatrix} + \begin{bmatrix} E_a \\ E_b \\ E_c \end{bmatrix}$$

$$\begin{bmatrix} Va \\ Vb \\ Vc \end{bmatrix} = R_s \begin{bmatrix} 1 & 0 & 0 \\ 0 & 1 & 0 \\ 0 & 0 & 1 \end{bmatrix} \begin{bmatrix} ia \\ ib \\ ic \end{bmatrix} + p \begin{bmatrix} L & M & M \\ M & L & M \\ M & M & L \end{bmatrix} \begin{bmatrix} ia \\ ib \\ ic \end{bmatrix} + \begin{bmatrix} ea \\ eb \\ ec \end{bmatrix} \qquad (4)$$

The generated electromagnetic torque is given by

$$T_e = \left[e_a i_a + e_b i_b + e_c i_c \right] \frac{1}{\omega} \qquad (5)$$

The torque runs into computational difficulty at zero speed as the induced EMF is zero and hence a reformulation independent of the speed is desirable. As the induced EMF is proportional to the product of rotor speed and airgap flux linkage which is a function of rotor position θ, the induced EMF can be written as :

$$e_a = f_a(\theta)\ \lambda\ \omega$$
$$e_b = f_b(\theta)\ \lambda\ \omega$$
$$e_c = f_c(\theta)\ \lambda\ \omega$$

Electrical rotor speed and position are related by

$$\frac{d\theta}{dt} = \left(\frac{p}{2}\right) * \omega$$

The equation of motion for a simple system with inertia constant J, friction coefficient B, load torque T_l, electromagnetic torque T_e and mechanical speed ω is given by

$$J\frac{d\omega}{dt} + B\omega = T_e - T_l \qquad (6)$$

The state space model of BLDC motor is given by

$$X' = AX + BU \qquad (7)$$

Where $X = \begin{bmatrix} i_a & i_b & i_c & \omega & \theta \end{bmatrix}^T$

$$A = \begin{bmatrix} \frac{-R_s}{L} & 0 & 0 & 0 & 0 \\ 0 & \frac{-R_s}{L} & 0 & 0 & 0 \\ 0 & 0 & \frac{-R_s}{L} & 0 & 0 \\ \frac{\lambda p * f_a(\theta)}{J} & \frac{\lambda p * f_b(\theta)}{J} & \frac{\lambda p * f_c(\theta)}{J} & \frac{-B}{J} & 0 \\ 0 & 0 & 0 & \frac{p}{2} & 0 \end{bmatrix};$$

$$B = \begin{bmatrix} \frac{1}{L} & 0 & 0 & 0 \\ 0 & \frac{1}{L} & 0 & 0 \\ 0 & 0 & \frac{1}{L} & 0 \\ 0 & 0 & 0 & \frac{-1}{J} \\ 0 & 0 & 0 & 0 \end{bmatrix}; \quad U = \begin{bmatrix} V_a & V_b & V_c & T_L \end{bmatrix}^T$$

III. Estimation of Speed and Position Based on Fuzzy Logic Controller

The proposed fuzzy back EMF is divided into two parts. One is the stator current observed in terms of state equation and the other is the fuzzy function. The fuzzy membership functions for error, change in error and control output are shown in figure 3.

Fuzzy logic controller contains four main parts, out of which two perform transformations. They are fuzzifier (transformation 1), knowledge base, inference engine and defuzzifier (transformation 2).

a) Fuzzification

Fuzzification measures the values of input variables and converts them into suitable linguistic values. Knowledge base consists of a database and provides necessary definitions, which are used to define linguistic control rules. This rule base characterizes the control goals and control policy of the domain experts by means of a set of linguistic control rules. The input and output has five sets associated with seven linguistic labels: (NB) Negative Big, (NS) Negative Small, (Z) Zero, (PS) Positive Small and (PB) Positive Big as shown in figure 3.

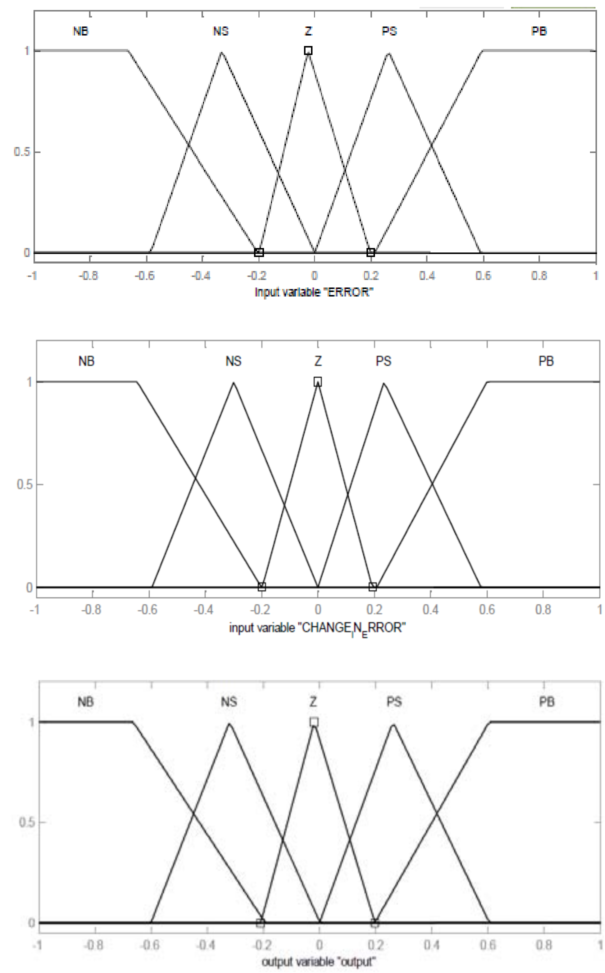

Figure 3 : Membership function

b) Inference Engine

Decision making logic or inference mechanism is a main part of fuzzy controller. It has the capability of simulating human decision making based on fuzzy concepts and of inferring fuzzy control actions employing fuzzy implication and the rules of inference in fuzzy logic. A typical rule is described as *IF (condition 1) AND (condition 2) THEN (conclusion).* Fuzzy inference consists of two processing methods namely, Mamdani's method and Sugeno or Takagi-Sugeno-Kang method to calculate fuzzy output [9]. Out of it Mamdani's method is more suitable for DC machine and induction machine control. The Table 1 shows the fuzzy rule-base.

c) Defuzzification

Defuzzification is a scale mapping, which converts the range of values of output variables into corresponding universe of discourse and also yields a non-fuzzy control action from an inferred fuzzy control action. The fuzzy function converts its internal fuzzy output variables into crisp values so that the actual system can use these variables. One of the most common ways is the center of area method, and will be used here.

Table 1 : Fuzzy Rules

Change in error	NB	NS	Z	PS	PB
NB	NB	NB	NB	NS	Z
NS	NB	NB	NS	Z	PS
Z	NB	NS	Z	PS	PB
PS	NS	Z	PS	PB	PB
PB	Z	PS	PB	PB	PB

IV. Simulation Results

The mathematical model is simulated by MATLAB / SIMULINK block. The closed loop model is shown in figure 4.

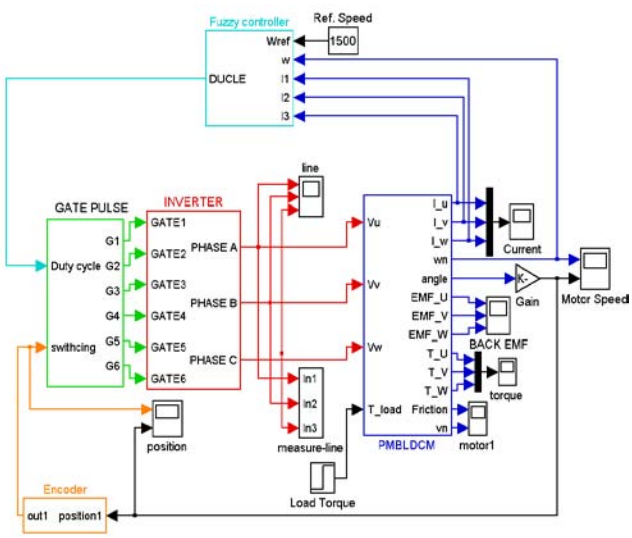

Figure 4 : Simulink model of proposed system

The block diagram of the proposed drive system is given in figure 1. The line voltage is measured from the DC-link. The line current of the BLDC motor is also calculated. Using these calculated values, back EMF is estimated with the help of Fuzzy Block. The speed and the rotor position are calculated by the estimated back EMF. The estimated speed is fed to the error detector which finds the difference between the actual and desired values. The error output is finally fed to the control switch where the pluses for the inverter are decided. Thus the inverter produces the exact voltage required for the BLDC motor.

The speed measured by fuzzy technique is almost same as the sensor value. The fuzzy based speed estimation is reliable and cost of the sensor is also eliminated. Thus the Fuzzy logic is found to be somewhat superior. That is, it doesn't need any physical component for the measurement of speed. Therefore the overall system cost is reduced and the maintenance problem of the sensor is also eliminated.

The results shown below reveal that the proposed fuzzy based control of BLDC motor is efficient. The back EMF of the BLDC motor has been shown in figure 5. Similarly the rotor position angle of the motor is also determined by the speed of the motor which is shown in figure 6. The speed of the motor determined by the estimated back EMF is shown in figure 7.

The speed is also varied from one point to another; from zero to full rated speed and from half rated to full. In all above aspects the simulations are done, performance is good for fuzzy based estimation of sensor-less control of BLDC motor. The maximum overshoot and ripples are reduced effectively after adding the Fuzzy controller. The simulation results confirm the better performance and higher efficiency of the proposed model. To realize the result, some of the waveforms taken after simulation for different speed range are given here for the purpose of reference.

The change in speed of the BLDC motor from 1000 rpm to 1200 rpm is shown in figure 8. Its corresponding back EMF and rotor position angle are shown in figure 9 and figure 10 respectively.

The results due to change in speed of the BLDC motor from 1000 rpm to 800 rpm are shown in figure 11. Its corresponding rotor angle is shown in figure 12.

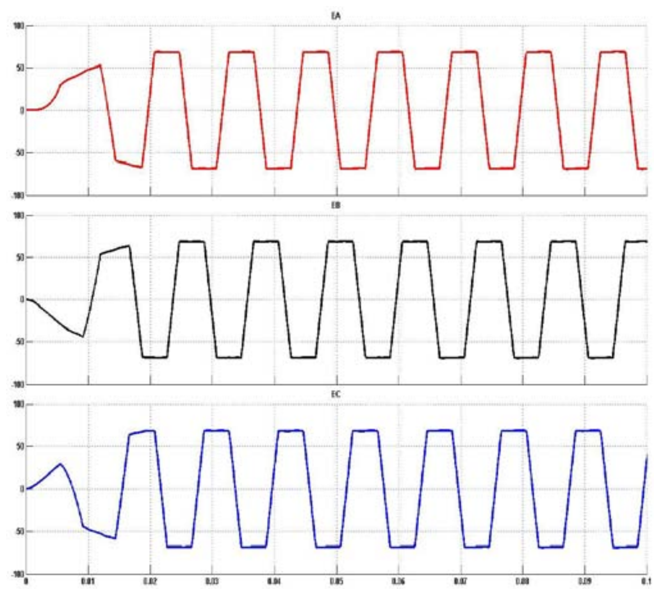

Figure 5 : Trapezoidal Back EMF

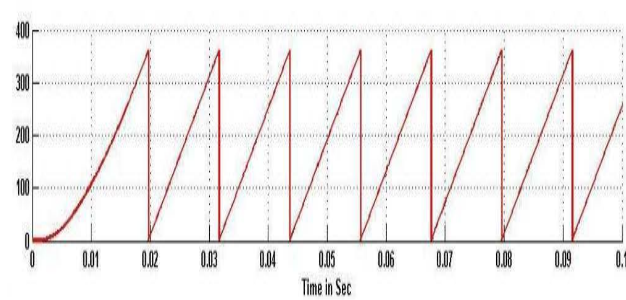

Figure 6 : Rotor angle

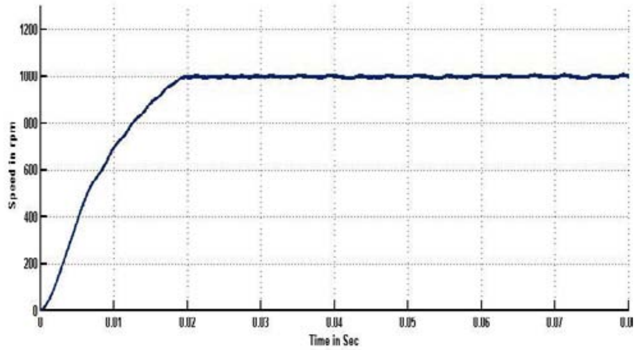

Figure 7 : Speed response of BLDC motor

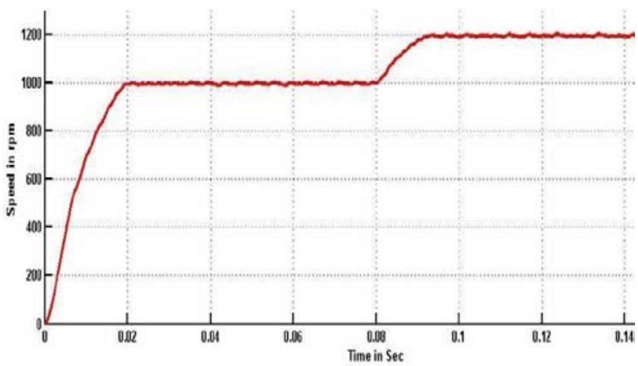

Figure 8 : Speed response of BLDC motor

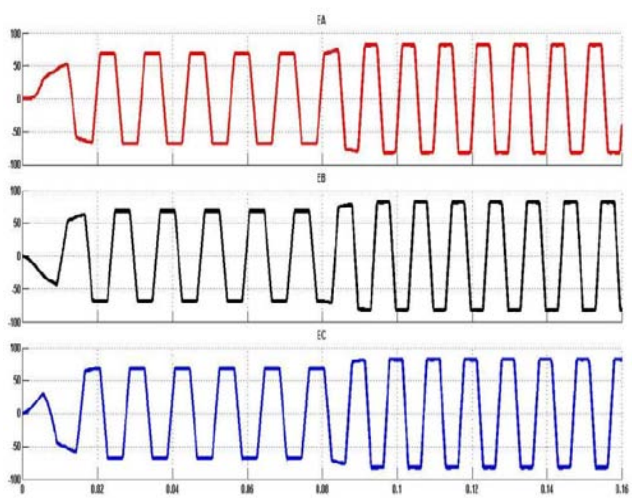

Figure 9 : Trapezoidal back EMF

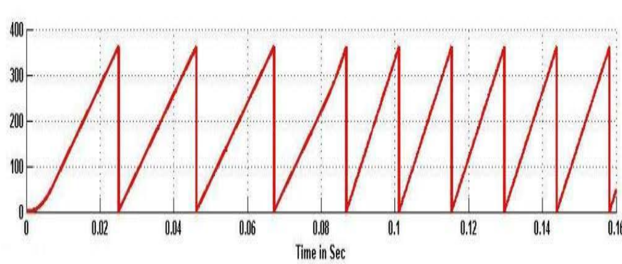

Figure 10 : Rotor angle

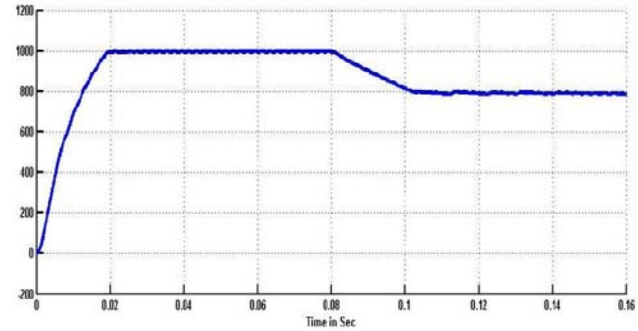

Figure 11 : Speed response of BLDC motor

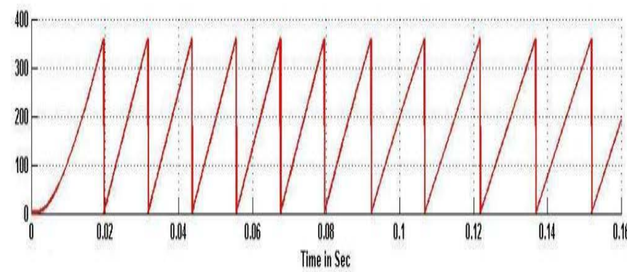

Figure 12 : Rotor angle

Motor Parameter Used

Phase Voltage	300 V
No. of Poles	4
Number of turns per phase	800
Rated Speed	1000 RPM
Resistance per phase (Rs)	10 Ohms
Self Inductance (La)	10 mH
Mutal Inductance (M)	1.5 mH
Maximum flux density	0.8167 web/m2
Moment of Inertia (J)	0.0021 Kg-m2
Friction Co-efficient (B)	0.089 Nm/(rad/sec)

V. Conclusion

A control system to estimate the speed and rotor position based on fuzzy back EMF observer is developed with the help of fuzzy logic technique for BLDC motor without position sensors. The proposed model is used to estimate the speed of BLDC motor under variable and fixed condition of back EMF. The proposed method has higher performance than the conventional sensors-less method without any additional circuitries. Simulation results confirm the better performance and higher efficiency of the proposed model.

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Proximity Coupled Rectangular Microstrip Antenna with X-slot for WLAN Application

By I.V.S. Rama Sastry & Dr. K. Jaya Sankar
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Abstract- One of the major limitations of microstrip antenna is narrow bandwidth. Various techniques can be used to improve it's BW. In this paper, Proximity coupling technique and X slot are employed to enhance the BW of Rectangular Microstrip antenna (RMSA). A RMSA with X slot has been designed at a frequency of 2.45GHz on FR4, one substrate and it is proximity coupled with microstrip feed, which is on the other substrate. Various parameters viz. Return loss, VSWR, input impedance, Gain, are obtained from HFSS simulation and Network analyzer test is performed to obtain VSWR. Percentage BW of VSWR from network analyzer is obtained as 7.14, which is more than that of a microstrip feed RMSA.

Keywords: rectangular microstrip antenna, proximity coupling, microstrip feed, X-slot, BW, return loss.

GJRE-F Classification : FOR Code: 090699, 090609

PROXIMITYCOUPLEDRECTANGULARMICROSTRIPANTENNAWITHX-SLOTFORWLANAPPLICATION

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Proximity Coupled Rectangular Microstrip Antenna with X-slot for WLAN Application

I.V.S. Rama Sastry [α] & Dr. K. Jaya Sankar [σ]

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Keywords: rectangular microstrip antenna, proximity coupling, microstrip feed, X-slot, BW, return loss.

I. Introduction

The microstrip antenna in its simple form cannot satisfy the bandwidth requirements for most wireless communication systems. Antenna plays a vital component in wireless application systems. The microstrip antenna can be used for wireless applications as it has features such as light weight, easily mounted and it is easy to mass production. Although there are many features that suits well for microstrip antenna to be deployed for wireless applications, there is a very serious limitation where it has a very narrow bandwidth. The typical bandwidth of the microstrip antennas is between 1% to 3%.By overcoming this limitation the microstrip antenna can be used to its full potential. An alternative bandwidth enhancement technique is studied and then proposed in order to broaden the bandwidth of the microstrip antenna [1]. The wireless application that is selected to be studied is the Wireless Local Area Network (WLAN) based on the IEEE 802.11b standard. This WLAN band spans from 2.4GHz to 2.48GHz. Inserting slot on ground plane and stacked patch supported by wall, the bandwidth can improve up to 25% without significant change in the frequency [1].

In this paper the bandwidth of RMSA is enhanced by using combination of two techniques, Proximity coupled feed line and adding X slot in the patch.

a) *Proximity Coupled Feed*

The microstrip antenna consists of a grounded substrate where a microstrip feed line is located. Above this material, there is another dielectric laminate with a rectangular microstrip patch etched on its top surface. There is no ground plane separating the two dielectric layers. The power from the feed network is coupled to the patch electromagnetically, as opposed to a direct contact. This form of microstrip patch is sometimes referred to as an electromagnetically coupled patch antenna.

A key attribute of the proximity-coupled patch is that its coupling mechanism is capacitive in nature. This is in contrast to the direct contact methods, which are predominantly inductive. The difference in coupling significantly affects the obtainable impedance bandwidth, because the inductive coupling of the edge- and probe-fed geometries limits the thickness of the material useable. Thus, bandwidth of a proximity-coupled patch is inherently greater than the direct contact feed patches.

b) *Slots in The Patch*

Recently, most applications need larger bandwidths, so in these application areas, microstrip antennas' biggest handicap is the narrow bandwidth. Because of this, there are lots of studies going on and various bandwidth enhancement techniques are reported [2]. These techniques to increase bandwidth include introducing multiple resonances into the structure [3]. This may take the form of stacked patches, coplanar parasitic patches, or patches that have novel shapes such as the U-shaped slot patch antenna.Using special feed networks or feeding techniques [4] to compensate for the natural impedance variation of the patch is another method. Etching a slot on the patch is a simple design. This design avoids the use of stacked or coplanar parasitic patches, either of which increases the thickness or the lateral size of the antenna. So, while changing the current distribution on the microstrip patch, by enhancing the impedance bandwidth sometimes more than one resonant frequency is obtained [5]. In this study, slots with various dimensions etched on present rectangular patch antennas in literature are tested and it is seen that their bandwidths can be enhanced by sacrificing a bit from the initial resonant frequencies. The simulation measurement without slots give near resonant frequencies to the experimental results; but all the bandwidths are under 5% and by using the slots with direct contact feeding techniques we can enhance the bandwidth up to 15%.

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II. Design of Proximity Coupled Fed Microstrip Antenna with X-Slot

This section describes the design of Rectangular Microstrip patch antenna (RMSA) to obtain the results at frequency of 2.45 GHz

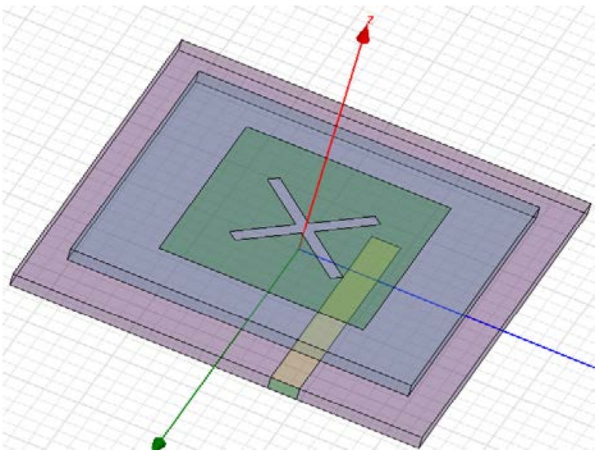

Figure 2.1 : Geometry of Proximity Coupled Fed Antenna with X-Slot

The geometry of proximity coupled microstrip line fed RMSA with X-Slot is shown in figure 1. In the design of RMSA the substrate material Fr4 (Glass-epoxy) with dielectric constant (ε_r) of 4.4 and thickness (h) of 1.60 mm, has been used. The dimensions of RMSA at frequency of 2.45 GHz have been calculated by using Transmission line model [6]. The length (L) and width (W) are calculated as 28.8 mm. A slot of X shape is made on RMSA. The slot length (L_S) is considered as less than $\lambda/4$ and width (W_S) is less than $\lambda/10$.

A 50 Ω. Impedance matching line of $\lambda/4$ length is used as microstrip feed line to couple the power, to RMSA. It is etched on top portion of the bottom substrate material (Fr4). The length (L_f) and width (W_f) of feed line are calculated.

The design of Proximity coupled RMSA with X-slot is simulated with HFSS simulator [7]. To get the satisfied simulation results, the design parameters are optimised. The optimised design parameters of proximity coupled RMSA with X-slot are shown in table 2.1.

Table 2.1 : Optimised design parameters

f_r	ε_r	h	W	L	F_L	F_W	L_W	L_S
2.45 GHz	4.4	1.6 mm	26 mm	24 mm	30 mm	3.4 mm	1.75 mm	17.32 mm

With the above design parameters, the RMSA has been fabricated and tested using network analyzer. From measurement VSWR characteristic is obtained which is shown in figure 3.6 VSWR

a) Photo copies of fabricated RMSA

The fabricated design of proximity coupled RMSA photos are shown in figures

Figure 2.2 : Proximity coupled RMSA with X-slot

Figure 2.3 : RMSA with X-slot (on first substrate)

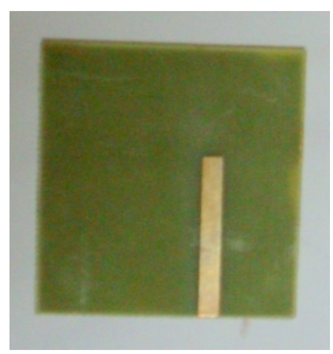

Figure 2.4 : Microstrip feed line (on Second Substrate)

III. Simulation Results

The High Frequency Simulation Software (HFSS) is used to model and simulate the Proximity coupled RMSA. HFSS software [8] is the industry-standard simulation tool for 3-D full-wave electromagnetic field simulation. Using this software, input port impedances, s-parameters, return loss, radiation pattern, etc. are obtained and the results are shown in figures from.

a) Return Loss

Return loss is the difference between forward and reflected power, in dB, generally measured at the input to the coaxial cable connected to the antenna. If the power transmitted from the source is P_T and the power reflected back to the source is P_R, then the return loss is given by P_R/P_T. For maximum power transfer, the return loss should be as small as possible. This means that the ratio should be as small as possible, or expressed in d B, the return loss should be as large a negative number as possible. For example a return loss of -40dB is better than one of -20dB.

The designed antenna resonates at 2.45GHz with return losses of -26 dB which is shown in figure3.1 and percentage BW is calculated as 8.17%

- *Bandwidth Calculation*

$$\%BW = \frac{f_2 - f_1}{f_r} \times 100\%$$

$$= \frac{2.4715 - 2.2709}{2.45} \times 100 = 8.17\%$$

Where f_1 and f_2 lower and upper frequencies

b) Voltage Standing Wave Ratio (VSWR)

The most common case for measuring and examining VSWR is when installing and tuning transmitting antennas. When a transmitter is connected to an antenna by a feed line, the impedance of the antenna and feed line must match exactly for maximum energy transfer from the feed line to the antenna to be possible. When an antenna and feed line do not have matching impedances, some of the electrical energy cannot be transferred from the feed line to the antenna. Ideally, VSWR must lie in the range of 1-2 which is achieved in figure 4.4 for the frequency 2.45 GHz, near the operating frequency value.

c) Radiation Pattern

The radiation pattern is a graphical depiction of the relative field strength transmitted from or received by the antenna. Antenna radiation patterns are taken at one frequency, one polarization, and one plane cut. The patterns are usually presented in polar or rectilinear form with a dB strength scale.

Since a Microstrip patch antenna radiates normal to its patch surface, the elevation pattern for $\theta = 0^o$ and $\theta = 90^o$ would be important. Figure 3.10 below shows the gain of the antenna at 2.45GHz for $\theta = 0^0$ and $\theta = 90^o$.

d) Input Impedance

We expect pure real impedance at frequencies where the patch resonates, that is, where the patch is designed to radiate. As a result, the input impedance plot in Fig 3.2 shows that around the desired radiating frequency, sufficient reactance cancellation can only occur inside a narrow bandwidth. In addition, one needs to match the resonant resistance with the characteristic impedance of the feed line. A small antenna can be tuned to resonate with an appropriate addition of reactance, or it can be made to self-resonate so that the reactance cancellation at resonance happens naturally in the antenna structure. Since adding external reactance for this purpose increases the power loss and it also requires extra space, it is advisable to follow the second alternative.

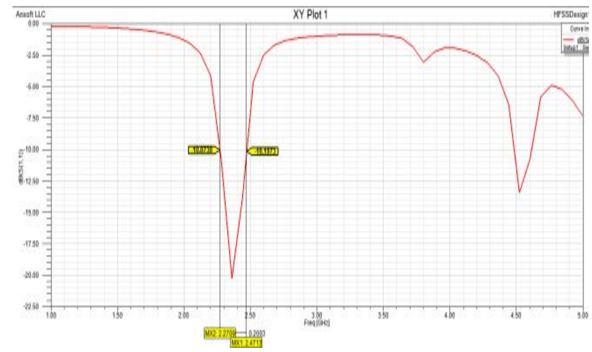

Figure 3.1 : Return loss

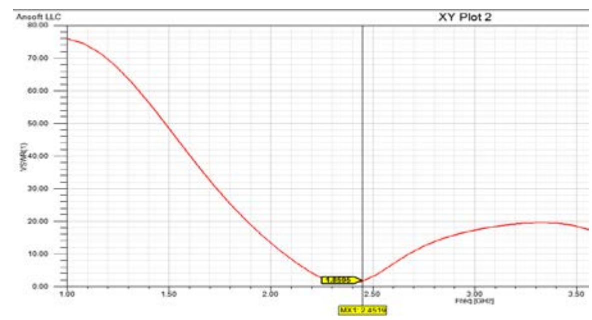

Figure 4.5 : VSWR plot

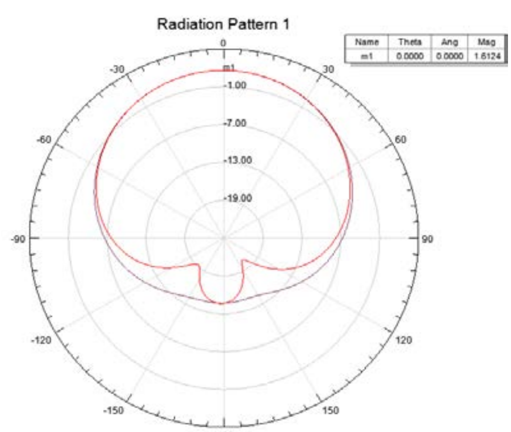

Figure 4.6 : Radiation Pattern

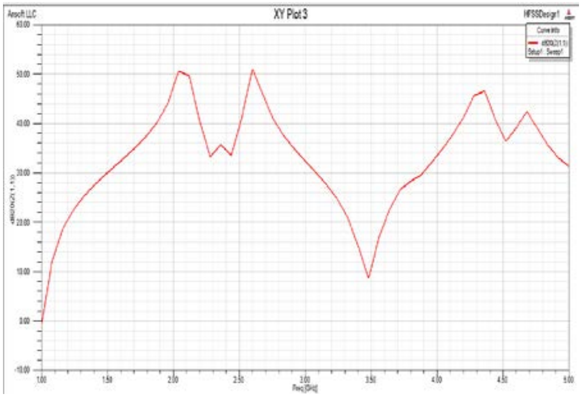

Figure 4.7 : Z-parameter

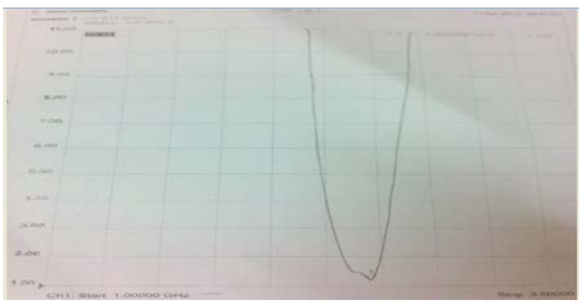

Figure 3.8 : Network Analyzer test result –VSWR Plot

e) Fabricated Antenna Results

The fabricated Proximity coupled RMSA with X-slot is tested with HP Make network analyser. From this test,The VSWR characteristics are obtained. The percenge Bandwidth from VSWR plot is calculated as 7.14% ,

- *Practical Bandwidth Calculation*

$$= \frac{2.5 - 2.325}{2.45} \times 100 = 7.14\%$$

f) BW Comparison

The percentage BWs of Proximity coupled RMSA with X-slot is compared, those values are calculated from HFSS simulation results and Network analyser test.Table 4.1shows comparison of % BW between the simulation and practical results.

Table 4.1 : Simulation & Practical Test Results

Results	**Rectangular MSA(Without X-Slot)**	**Proximity coupled fed Rectangular MSA with X-slot**
Simulation Results	5.68%	8.17%
Practical Results (Network Analyzer Test)	2.4%	7.14%

IV. Conclusion

Proximity coupling and X slot techniques are introduced for improving the BW of RMSA at 2.45GHz. The proposed design of RMSA has been simulated and practically tested with network analyzer. For proposed antenna design, BW of 8.17% is obtained from HFSS simulation, where as from network analyser test the BW of 7.14% is calculated .The bandwidth before adding the slot and the stacked patch was 2.5% whereas after adding the slot and the stacked patch the bandwidth increased. Thus the proposed design is useful for BW enhancement for RMSA, which is required for Wireless LAN applications.

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Community based Micro Off-Grid Power System using Renewable Energy Technology (Ret): Investment Analysis, Cost Benefit and Main Factors

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Abstract- Bangladesh is the eighth populous country in the world with the highest population density and a vast majority of its population, who are living in villages and small towns, are deprived of power. Recently some steps are being taken by the Government of Bangladesh (GoB) to bring those people under electrification who are in remote inaccessible un-electrified area where grid expansion is expensive and it is being done by encouraging them to use renewable energy such as: Solar photovoltaic, Solar thermal power, Wind power, Biogas, Mini-Hydro etc for electricity generation in stand-alone system. But the high installation and maintenance cost of these renewable energy based power generation in stand-alone system are becoming the main hindrances to the village people to afford it. Though some financial helps are provided from government and some NGOs, but those are not enough and only a few people get these privileges.

Keywords: renewable energy technology; developing country; community; off-grid; biogas.

GJRE-F Classification : FOR Code: 090608

COMMUNITYBASEDMICROOFFGRIDPOWERSYSTEMUSINGRENEWABLEENERGYTECHNOLOGYRETINVESTMENTANALYSISCOSTBENEFITANDMAINFACTORS

Community based Micro Off-Grid Power System using Renewable Energy Technology (Ret): Investment Analysis, Cost Benefit and Main Factors

Md. Tariqul Islam [α] & Md. Imtiazul Haque [σ]

Abstract- Bangladesh is the eighth populous country in the world with the highest population density and a vast majority of its population, who are living in villages and small towns, are deprived of power. Recently some steps are being taken by the Government of Bangladesh (GoB) to bring those people under electrification who are in remote inaccessible un-electrified area where grid expansion is expensive and it is being done by encouraging them to use renewable energy such as: Solar photovoltaic, Solar thermal power, Wind power, Biogas, Mini- Hydro etc for electricity generation in stand-alone system. But the high installation and maintenance cost of these renewable energy based power generation in stand-alone system are becoming the main hindrances to the village people to afford it. Though some financial helps are provided from government and some NGOs, but those are not enough and only a few people get these privileges. That's why the potentiality of renewable energy couldn't be used in an effective way. The objective of this paper is to present an alternative and - from our point of view – more realistic aspect of using renewable energy effectively in the reliable energy supply. This paper will discuss about an idea of cost effective community based micro off-grid power system which will emphasize on using Renewable Energy Technologies (RETs) like utilizing Biogas for power generation. Furthermore, this paper will also visualize as to how this micro off-grid system can be implemented in those villages and small towns which are currently detached from electricity and it can be done by dividing each village or town into small communities and bringing them under electrification. Many factors, such as technology costs and investment analysis, benefit-cost ratio, payback period and available potentials have been incorporated in order to fulfill this task.

Keywords: renewable energy technology; developing country; community; off-grid; biogas.

I. Introduction

Power stands as a judging criterion for indicating a strong socio-economic development of any country. Industrial growth is solely dependent on power. Without power, the development efforts of a country cease to exist. The growing need for power is intensifying day by day. With the limited resources available, developed countries have been satisfying their ongoing demand for power by applying proper technologies and hence maximizing their industrial growth. But developing country like Bangladesh is far away from this race due to its weak administrative structure, poor human resource management, high population density and lack of knowledge regarding application of proper scientific methods to utilize their available resources. Power crisis has become so acute that the gap between total generation and total demand is getting larger and larger. The Government of Bangladesh (GoB) has been failed to mitigate this power crisis in many aspects due to their imprudence in policy making regarding power crisis. The pervasive corruption and irregularities in power sector have made this problem more acute. In Bangladesh, about less than 10% of the rural people are connected with the national grid of electricity supply and about two-third of the country's 86000 villages are still outside the reach of the national grid [1]. The main reason for that is, the GoB has failed to give them full access to electricity leaving them detached from the national grid. In this situation, they can be given access to electricity by building community based micro power plants which will not be connected to the national grid rather they will be operated only those areas which are not electrified yet. Those micro power plants may be designated as off-grid power plants whose generating fuel will be renewable energy such as: Biogas. In this community based micro off-grid power system, a village or small town will be segmented into several communities where each community will consist of a certain number of consumers and they will be supplied electricity from one common micro off-grid power plant. Here total installation and maintenance cost will be shared by the consumers of each community which will result in less financial burden for a single family as compared to stand-alone RETs based power generation system.

II. Potential of Renewable Energy Technologies (RETs) and its Present Scenario in Bangladesh

Bangladesh is heavily dependent upon conventional energy sources e.g. Coal, Gas, Diesel,

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Furnace oil etc and a large portion of total energy demand of the country is satisfied by these traditional biomass fuels. At present, Bangladesh receives energy supply both from renewable and non-renewable sources. In 2009, natural gas accounted for 50 percent of total energy supply, which declined to 46 percent in 2010. Contribution of bio-mass to total energy supply increased from 33.3 percent to 34.6 percent during this period. It may be mentioned that, use of oil as energy has increased significantly during this time. In 2009, oil represented 11.1 percent of total energy supply, which increased to 18.3 percent in 2010 [2]. Whereas Renewable energy contributes only a few percent of the total energy consumption in the country, mainly through biomass, e.g. agricultural residues contribute almost half the national total, with cow dung, bagasse and fuel wood making up the rest. The trend of annual energy consumption is presented below in the “Figure-I”. The potential for renewable energy other than biomass is quite high, but current utilization is minimal. These sources are biomass (including biogas and solid waste), solar energy, tidal and wave. The hundred plus miles long coastal areas and hilly sections provide ample wind for wind turbines. Waterways of varied forms and speed provide ample wave and gravity driven water flow for ecologically balanced hydroelectric generators. The lush vegetation provides ample photosynthesis and biomass for fuel for a variety of purposes. Also, more than two-thirds of the land area is grid free where decentralized applications of various RETs have been proven to be the most cost-effective options for generating electricity and heat [3].

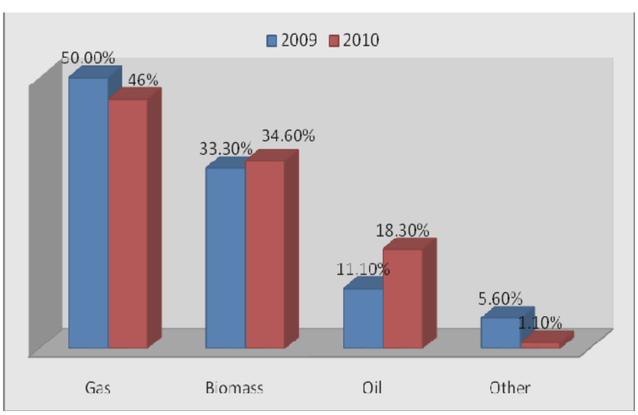

Figure 1 : The trend of annual energy consumption in Bangladesh [2]

The size and economic potential of the renewable energy resources (e.g., solar photovoltaic, solar thermal power, wind power, biogas, hydro etc.) in Bangladesh are yet to be determined and the capacity of renewable energy development is presently low. An estimated potential of Renewable Energy Technology (RET) in Bangladesh is shown in “Table-I” [4].

Table 1 : Potential of Renewable Energy Technology (Ret) In Bangladesh

Sources	**Unit size**	**Potential number**	**Total conventional unit**
Solar LED based lantern	5W	11 million (below poverty line)	55 MW
Solar LED based lantern + 10W CFL	15W	11million (below poverty line)	165 MW
Solar Home System	30W	12 million	360 MW
Mini grids	12.5W	40011	30000 MW
Mini grid of moderate size			10000 MW
Solar water pumping		225000 (around)	1200 MW
Grid connected PV system			600 MW
Solar PV System		1% area of Bangladesh with 10% efficiency	40000 MW
Rice husk gasifier	200kW	500	100 MW
Wind electricity			1000 MW
Micro hydro			1.2 MW (according to BPDB)
Biogas power plants		0.202 million (from poultry waste)	Potential 400 MW Possible generation 100 MW

Although investment cost of renewable energy based power generation is generally higher compared to fossil fuel alternatives, this option becomes economically viable when all externalities (e.g. environmental cost, health hazards etc.) and lower operating cost are taken into consideration. In Bangladesh, currently renewable energy based power generation is being implemented using the following methodologies [5]:

- Solar power generation using solar rays
- Wind-mill power generation using wind power
- Production of Biogas using waste
- Electricity produced by Biomass Gasification Method using wood, rice husk, etc.

a) Solar Photovoltaic

The largest amount of sun light is available in between two broad bands encircling the earth between 15″ and 35″ latitude north and south. Fortunately, Bangladesh is located between 20° 34′ to 26° 38′ north latitude and 88°01′ to 92°42′ east longitude which is a good location for solar energy utilization. Daily average solar radiation varies between 4 to 6.5 KWh per square meter [6]. Maximum amount of radiation is available on the month of March-April and minimum on December-

January. Despite of having large potential, utilization of solar energy has been limited to traditional uses such as crop and fish drying in the open sun. Solar photovoltaic (PV) is gaining acceptance for providing electricity to households and small businesses in rural areas. In 1988, Bangladesh Atomic Energy Commission (BAEC) installed several pilot PV systems. The first significant PV-based rural electrification programme was the Norshingdi project initiated with financial support from France. Three Battery charging stations with a total capacity of 29.4 kWp and a number of standalone solar home systems (SHS) with a total capacity of 32.586 kWp were installed [7]. Rural electrification Board (REB) owned the systems and the users paid a monthly fee for the services. Solar photovoltaic (PV) systems are in use throughout the country with over 200,000 household-level installations having capacity of about 50 MW (June 2011) [2]. Scaling-up of solar PV systems assisted by the development partners are being implemented through the Rural Electrification Board (REB), Local Government Engineering Department (LGED), Bangladesh Power Development Board (BPDB) and other agencies implementing solar energy program. Renewable Energy Research Centre of the University of Dhaka has installed a model 1.1KW grid connected photovoltaic system. There is a strong potential for solar energy within the country.

b) Solar Thermal Power/Concentrating Solar Power (CSP)

The technology involves harnessing solar radiation for generation of electricity through a number of steps finally generating mechanical energy to run a generator. This technology needs to be disseminated in the country to supplement the power supply.

c) Biomass

Bangladesh has strong potential for biomass gasification based electricity. More common biomass resources available in the country are rice husk, crop residue, wood, jute stick, animal waste, municipal waste, sugarcane bagasse etc. This technology can be disseminated on a larger scale for electricity generation.

d) Biogas

Biogas mainly from animal and municipal wastes may be one of the promising renewable energy resources for Bangladesh. Presently there are tens of thousands of households and village-level biogas plants in place throughout the country. It is a potential source to harness basic biogas technology for cooking, and rural and peri-urban electrification to provide electricity during periods of power shortfalls. According to an estimate "29.7 billion cubic meter of biogas can be obtained from the livestock of the country which is equivalent to 1.5 million tons of kerosene (which is the principal fuel in the rural areas) [7]. Apart from this, it is also possible to get biogas from human excreta, poultry dropping, waste, marine plants etc. If each family of Bangladesh can be associated with a biogas plant, then only human excreta will give about 10 billion cubic meter biogas". According to Institute of Fuel Research & Development (IFRD) - there is potential of about four million biogas plants in our country [8].

e) Hydro

Micro-hydro and mini-hydro have limited potential in Bangladesh, with the exception of Chittagong and the Chittagong Hill tracts. Hydropower assessments have identified some possible sites from 10 kW to 5 MW but no appreciable capacity has yet been installed. There is one hydro power plant at Kaptai established in the 1960s with installed capacity of 230 MW.

f) Wind Energy

Wind Energy has also made some inroads but its potential is mainly limited to coastal areas, and offshore islands with strong wind regimes. These coastal settings afford good opportunities for wind-powered pumping and electricity generation. The long term wind flow in Bangladesh, especially in islands and in southern coastal belt of the country indicate that the average wind speed remains between 3 to 4.5 m/s for the month of march to September and 1.7 to 2.3 m/s for the remaining period of the year [9]. There is a good opportunity in island and coastal areas for the application of windmills for pumping and electricity generation. Presently there are 2 MW of installed wind turbines at Feni and Kutubdia. A number of small wind generators have been recently installed by Grameen Shakti at its Chakaria shrimps farm, BRAC and GTZ (a German NGO). BRAC alone has installed 11 wind turbines at various coastal areas. These are small low cutting, DC operation type systems, supplying power to the target group to improve their quality of life.

g) Others

Other renewable energy sources include bio-fuels, gasohol, geothermal, river current, wave and tidal energy. Potentialities of these sources are yet to be explored.

At present it is estimated that renewable sources of power generation is about 55 MW. As per approved renewable energy policy 5% of the total generation (500 MW) would be added by 2015 and 10% of the total generation (1600 MW) would be added by 2020 from renewable sources [14]. IDCOL has supported NGOs in installation of SHSs in more than 380,000 households. Under the new initiative, BPDB is in process of installation of 100 MW Wind Power and 9-14 MW Grid connected Solar Power through PPP. Targets of power generation from renewable energy sources as fixed by the GoB are presented in the "Table-II" [2].

Table 2 : Targets of Electricity Generation by 2015 Utilizing Renewable Energy Technologies (RETs) and Achievements Till Date

Sources	Achievement (MW) by 2011	Estimated Production (MW) by 2015
Solar PV	50	200
Wind Power	2	200
Biomass	< 1	45
Biogas	< 1	45
Others	< 1	15
Total	55	500

III. Barriers for Sustainability of Renewable Energy Technologies (RETs) in Bangladesh

There are plenty of barriers hindering widespread deployment of potential Renewable Energy Technology (RET) in Bangladesh [10].

- Rural people have lack of idea about renewable energy resources, technical/economic information about RETs, equipment suppliers, and potential financiers.
- High initial capital costs and higher perceived risks of the renewable energy technology.
- Availability and access to existing renewable energy resource information is limited. A central information point does not exist, instead information is scattered among various sectors.
- There is not much campaign or awareness programs for the renewable energy consumption.
- The decision makers, who are urban dwellers, don't feel the necessity of renewable energy.
- GoB budgets for subsidizing RETs projects are limited as the demand for financing the various national priority areas (health, education, disaster management etc.) is great.
- The currently small and dispersed size of the renewable energy market in Bangladesh does not facilitate benefits such as economies of scale.
- NGOs working in Bangladesh are not sufficient or they do not have enough financial backup to promote the use of sustainable energy in the extreme rural areas
- NGOs have lack of technical know how people related to renewable energy technology.
- There are not much training materials and trained persons for technical backup support.
- Natural disasters are one of the barriers for promotion of sustainable energy.
- In our country financing sources are not interested in sustainable energy technology.
- Installation and maintenance cost of SHS are high due to bad communication and scattered localities.
- Lack of expertise and services in resource assessment, system design, installation, operation and maintenance of renewable energy technologies.

IV. Design of our Proposed Community based Micro Off-Grid Biogas Plant

Due to several reasons mentioned earlier, RETs based power generation has not gained widespread implementation. As a result of this, a vast majority of the population living in the rural areas have limited access or in some cases, no access to electricity at all. To help ease the problem, we have come up with the idea of Community Based Micro Off-Grid Power System using Renewable Energy Technology (RET). As Bangladesh is an agricultural country and most of the villagers earn their livelihood by farming, cows/bullocks/buffalos are part and parcel of most of the farmers' households. So, the dung egested by these livestock can be used as a potential source for generating biogas which then can be used to produce electricity. According to our proposed community based micro off-grid power system, what we will do is dividing a village into small communities where each community will consist of at least five families, each family having a member of five persons. These individual communities will meet their electricity demand by installing Biogas Plant which will be independent of National Grid, that's why the term Off-grid Power System has been coined. Here we will show the design and necessary calculation of a simple biogas plant which can meet the proposed communities' electricity demand. The system design includes the estimation of total gas required, amount of feedstock (or dung) required and the number of animals required having feedstock of a given amount.

a) Calculation of Net Electricity Demand of a Typical Community

Biogas system design for supplying the required demand for five families each having five members is considered here. Here, we have assumed a typical electrical load profile of a single family consisting five members and it is shown in the "Table-III".

Table 3 : A Typical Electrical Load Profile of A Signle Family In A Village

Load	Quantity	Wattage rating	Hours/day	Units (Wh)
1. Fan	2	80	4	640
2. Light (Energy saving/CFS bulb)	3	23	4	276
3. TV (19''color)	1	160	4	640
4. Mobile charger	1	10	4	40
		Total demand = 399 W		Total = 1.596 KWh

Therefore, a single community of five families will have a net demand of 399W×5≈2KW approximately. We have assumed that daily at most 4 hours these loads will be operated. Therefore, the total units needed will be 2KW×4Hr = 8KWh per day.

b) Amount of Gas Required Per Day

Here, 1KWh Electrical energy output is equivalent of 0.7 m^3 gas [11]. And 1000 *liters* of gas is equivalent to 1 m^3 of gas.

Therefore, 8KWh Electrical energy output ≡5.6 m^3 of gas

∴Total gas required = 5.6 m^3 / day or 5600 Litters/ day.

c) Number of Livestocks Required to Fulfill Daily Gas Requirement

Amount of gas produced from 1 Kg of fresh dung = 40 Litters

∴Total amount of dung

$$\text{required} = \frac{Total\ gas\ required}{Gas\ per\ kg\ of\ dung} = \frac{5600}{40} = 140\ kg$$

As most of the people in rural villages are farmer, so it is quite possible for every family to have cows for agricultural purpose. So we will consider here cow dung to fulfill daily gas requirement. 10 Kg dung/day/cow is an approximate; it may vary with breed of the cow [11].

Thus, in order to have 140 *kg* of dung, total no. of cows required$= \frac{140}{10} = 14$ cows

In our case, we have considered a total of five families. Therefore, if one family possesses 3 cows on average, a total of 15 cows will be owned by five families. Thus, the total demand of dung needed per day can easily be met by these families all by themselves.

d) Design of Digester and Gas Holder

In order to make slurry, water should be added to equal amount of dung i.e. dung : water = 1 : 1 [11]

Total mass of slurry = dung + water = 140 + 140 = 280 Kg Here, Specific gravity of slurry is about 1090 Kg/m^2

So, Volume of slurry per day

$$= \frac{Total\ mass\ of\ slurry}{Specific\ gravity\ of\ slurry} = \frac{280}{1090} m^3 = 0.2568 m^3$$

We have assumed the retention period of slurry will be 45 days.

So, Total Volume of the Digester = per day volume of slurry×retention period = 0.2568×45 = 11.556 m^3 ≈12 m^3.

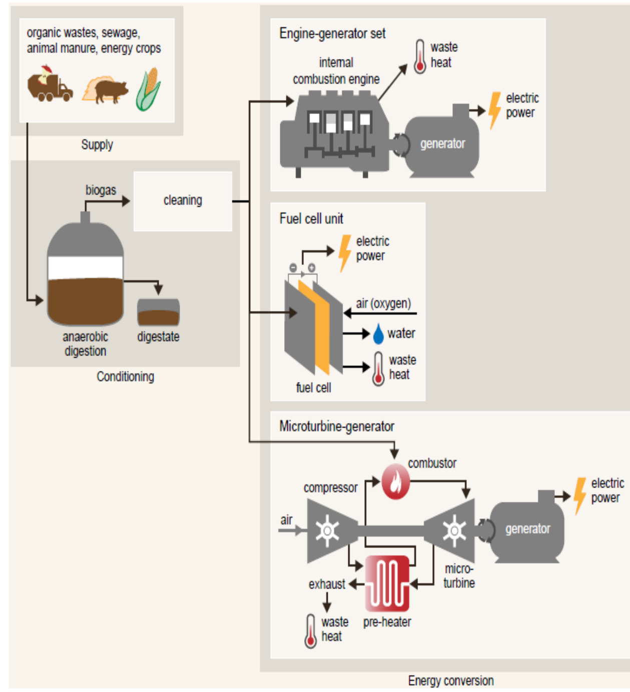

Figure 2 : Schematic view of our proposed micro off-grid Biogas plant

e) Dimension of the Digester and Gas Holder Tank

Depth to diameter ratio should be between 1 and 1.3. The volume of the gas holder tank should be about 60%of the per day gas volume. It should be kept in mind that, the temperature of the slurry in the tank plays an important role in the amount of gas production. Production yield is maximum between 45 and 55°C [11].

f) Generation of Electricity

The average calorific value of biogas is about 21-23.5 MJ/m^3, so that 1 m^3 biogas corresponds to 0.5-0.6 litter of diesel fuel or about 4.70 KWh (FNR, 2009). Theoretically, biogas can be converted directly into electricity using a fuel cell. However, very clean gas and an expensive fuel cell is necessary for this process. In most cases, biogas is used as fuel for combustion engines, which convert it to mechanical energy, powering an electric generator to produce electricity. Technologically far more challenging is the first stage of the generator set: the combustion engine using the biogas as fuel. In theory, biogas can be used as fuel in nearly all types of combustion engines, such as gas engines (Otto motor), diesel engines, gas turbines and Stirling motors etc.

For using Biogas in gas or diesel engines, the Biogas must fulfill certain requirements [12]:

- The methane content should be as high as possible as this is the main combustible part of the gas;
- The water vapor and CO_2 content should be as low as possible, mainly because they lead to a low calorific value of the gas;

- The Sulphur content in particular, mainly in form of H_2S, must be low, as it is converted to corrosion-causing acids by condensation and combustion.

Appropriate electric generators are available in virtually all countries and in all sizes. In most commercially run biogas power plants today, internal combustion motors have become the standard technology either as gas or diesel motors.

Figure 3 : Biogas Generator (Brand name: BETTER & Model no: BG350)

Table 4 : Main Technical Data of Biogas Generator (Brand Name: Better & Model No: Bg350)

Model	Power (KW)	Frequency (HZ)	Voltage (V)	Ignition System
BG350	2-3	50/60	230	T.C.I

V. Financial Analysis

a) Investment Cost

Table 5 : A Rough Cost Estimate of $12m^3$ Fixed Dom Biogas Plant

Material	Quantity	Unit	Unit Cost in Local Market (in BDT)	Total Cost (in BDT)
Bricks	4000	Piece	5	20,000
Sand	220	Cubic Feet (cft)	30	6,600
Cement	45	Bag (50Kg)	450	20,250
Brick Chips	75	Cubic Feet (cft)	50	3,750
PVC pipe (6 in Diameter)	15	Feet (ft)	100	1,500
MS rod	75	Kg	70	5,250
Earth Cutting	3000	Cubic Feet (cft)	5	15,000
Plastic Emulation paint				5,000
Gas pipe/ valve/ Gi pipe				5,000
Mixing Device				1,500
Kitchen Waste Bin (prefarmantation busket)	10	Piece	500	5,000
Mason				10,000
Other cost				3,000
Technical Supervisor fee				10,000
Biogas Generator	1	Piece		25,000
Total Cost				136,850

Each community collectively will contribute BDT 136,850 to the capital cost of this project. In addition, the communities will provide unskilled labor for the construction work. So, here each family in a community will have to pay = 136,850/5 = 27,370 BDT which is a very less amount and affordable in comparing with stand-alone cost. The monthly operation and maintenance costs for running this Biogas plant are estimated at BDT 2,000. To manage the Biogas power plant in the future, training modules will develop to train local community leaders in responsibilities, technical capabilities, staff and financial management, record keeping, accounting and leadership qualities.

b) Revenue

Here, From the Biogas plant electricity generated per day is equal to 8 KWh.

So, Electricity generated per year = 8KWh× 365 = 292020 KWh.

In Bangladesh, Rate of quick rental power is BDT 16 per KWh.

So, net revenue will be earned from electricity generated by the Biogas plant per year = 2,920×16 BDT = 46,720 BDT

Again, Slurry generated per month = 1,150 Kg

So, Slurry generated per year = 1,150×12 = 13,800 Kg

In Bangladesh, Rate of compost fertilizer is BDT 75 per 40 kg. Net revenue will be earned from fertilizer per year = 345×75 BDT = 25,875 BDT

So, Total Revenue will be earned from the Biogas plant per year = 46,720+25,875 BDT= 72,595 BDT

c) Payback (Payout) Period

The payback method, which is often called the simple payout method, mainly indicates a project's liquidity rather than its profitability [13]. The simple payback and discounted payback period methods tells us how long it takes cash inflows from our community based off-grid Biogas power plant project to accumulate to equal (or exceed) the project's cash outflows, which is an indicator of our project risk. Here calculation of the Simple Payback Period (θ) and the Discounted Payback Period (θ') at MARR = 13% is given in the "Table-VI".

Here, I = Initial investment for the project = 136,850 BDT

S = Salvage (market) value at the end of the study

period = 0 BDT (We have assumed that the biogas plant will have no value after its projected study period)

N = Project study period = 15 Years

O&M = Operation and management cost of the project per year = 2000×12 BDT = 24,000 BDT

B = Benefit from the project per year = 72,959 BDT

MARR = Minimum Attractive Rate of Return = 13% (As, the local Govt. bank and private bank in Bangladesh has MARR on average about 12%)

So, Annual worth of Investment = I (A/P, i%, N)

$$= 136{,}850 \times \left[\frac{i*(1+i)^n}{(1+i)^n - 1}\right]$$

$$= 136{,}850 \times 0.155 \text{ BDT}$$

$$= 21{,}212 \text{ BDT}$$

Here, Net profit from the proposed project per year = 72,595-24,000 BDT = 48,959 BDT

Table 6 : Calculation of The Simple Payback Period and the Discounted Payback Period at Marr = 13%

End of Year, k	Net Cash Flow (in BDT)	Cumulative PW at i=0%/yr through Year k (in BDT)	Present Worth of Cash Flow at i=13%/yr (in BDT)	Cumulative PW at MARR=20%/yr through Year k (in BDT)
0	-136,850	-136,850	-136,850	-136,850
1	48,595	-88,255	43,005	-93,845
2	48,595	-39,660	38,057	-55,788
3	48,595	**+8,935**	33,679	-22,109
4	48,595	+57,530	29,804	**+7,695**
5	48,595	+106,125	26,375	+34,070

So, here the Simple Payback Period is $\theta = 3yrs$ because the cumulative balance turns positive at EOY 3. And the Discounted Payback Period is $\theta' = 4yrs$ because the cumulative discounted balance turns positive at EOY 4.

d) Benefit-Cost Ratio

Conventional B/C ratio with Annual worth (AW) [13]:

$$B/C = \frac{AW\,(Benefits\ of\ the\ proposed\ project)}{AW\,(Total\ costs\ of\ the\ proposed\ project)}$$

$$= \frac{AW(B)}{CR + AW(O\,\&\,M)}$$

$$= \frac{68885}{21212 + 42000}$$

$$= 1.089$$

Here, B/C ratio is greater than 1. So, our proposed project is acceptable.

VI. Conclusion

In this paper we have presented the idea of physically implementing smart micro off-grid community based power generation using RETs which can solve the existing power crisis especially faced by the vast majority of general mass who are living in the developing countries. In Bangladesh, people living in many rural areas have no access to electricity at all. Growing demand for power has already put an immense pressure on fossil fuels and with limited resources available in Bangladesh, it will be almost impossible to meet this huge increasing demand. The current reserve of fossil fuels is depleting in an alarming rate which will create an agonizing situation for the inhabitants of this country. Because, GoB has already failed to provide access to electricity in the rural areas; what is going to happen in near future when this fossil fuel reserve runs out completely can easily be imagined. Our proposed idea of smart micro off-grid community based power generation can be used as a helping tool for the solution of this problem. Detailed investment and benefit-cost ratio analysis, payback period calculation have been carried out in a realistic way to show that our proposed idea can be successfully implemented as well as keeping the poor economic condition of the general rural people in contrast. Technical assistance and economic support from GoB as well as from other NGOs should also be incorporated to implement our proposed idea thereby reducing intense pressure on national grid and conventional limited fossil fuel reserve.

Appendix

GoB-Government of Bangladesh

NGO-Non Government Organization

RET-Renewable Energy Technology

BPDB-Bangladesh Power Development Board

REB-Rural Electrification Board

IFRD-Institute of Fuel Research & Development

LGED-Local Government Engineering Department

CSP-Concentrating Solar Power

SHS-Solar Home System

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Reducing the Vulnerability of Digital Protective Relays to Intentional Remote Destructive Impacts: Technical-and-Economic Aspects

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I. Introduction

Some of my previous publications [1-3] illustrated the vulnerability of digital protective relays (DPR) to intentional remote destructive impacts (IRDI) such as electromagnetic and cybernetic attacks. In [4-6] rationalizes the necessity of DPR protection and [6] explains a specific method of protection based on an in-series combination of DPR and reed switch-based starting element, which unlocks the DPR only if at least one of the controlled parameters (current, voltage, angle between them, etc.) approaches the DPR operating threshold, Fig. 1.

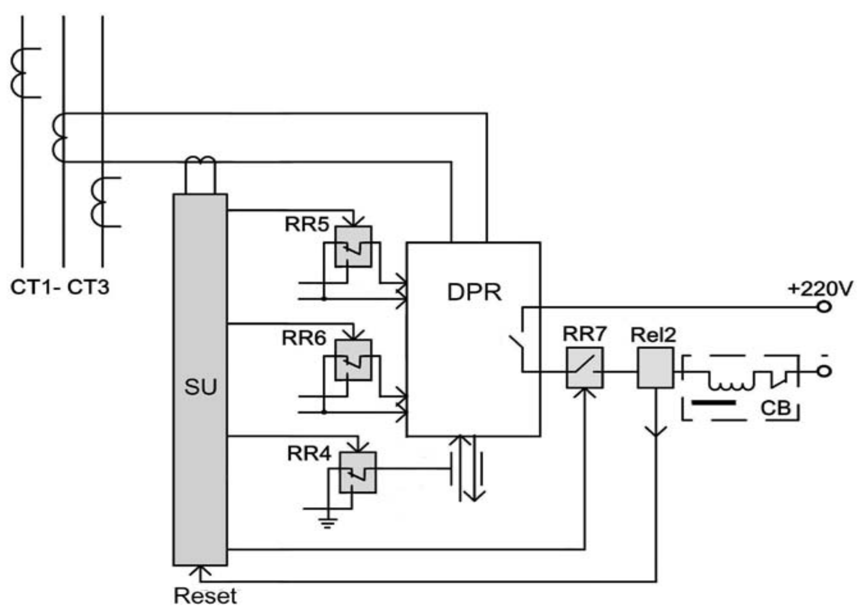

Figure 1 : Block diagram of the proposed method of DPR protection against IRDI

The statement itself and the proposed method of DPR protection against IRDI is so unusual and so different from anything that has been known before, that experts inevitably have raised a barrage of questions and a tornado of emotions (alas, they are not always positive). Lack of the answers to many of the emerging questions in previously published articles often has led to misunderstandings and hence to complete rejection of the proposed method. Therefore, we will try to formulate the most frequently asked questions on this matter and answer them.

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II. Answers on Most Important Expert Questions

Question 1. According to the diagram the reed switches are hung on DPR all-around like fairy lights on the Christmas tree

It is clear that the reed switches are not hung on the inputs and outputs of DPR like “fairy lights”, rather they are located inside the single shielded enclosure

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together with all other elements of the proposed protection device, and the design of the enclosure is similar to that of the DPR with the only difference being that there is no need for a screen but there is an access to the operating threshold control units of the threshold element reed relays. This separate module has the same terminal blocks for connection to external circuits, as the DPR.

Question 2. There is a common perception that reed switches are not reliable (they "stick"). Is it reasonable to use them in devices which should be characterized with enhanced reliability?

Reed switches, or rather reed relays used in the starting unit (SU) of protection devices, have a wide range of benefits compared to conventional electromagnetic relays. First, the contact points of dry reed switches are enclosed in a sealed cylinder filled with a mixture of pressurized inert gases or vacuum and so they are not exposed to adverse environmental factors (moisture, dust, gases). These contacts require no adjustment or cleaning during the whole lifetime.

Second, the reed relay is 3-5 times, or more, faster than the conventional electromechanical relays. Third, under alternating current such a relay has a reset ratio of 0.9 - 0.95, which is much higher than for conventional relays. Fourth, the reed relay allows for easy approach to the galvanic isolation level between input and output (between coils and contacts) of tens of kilovolts, which is unattainable for conventional electromechanical relays [7]. Fifth, unlike the conventional relays, the reed relays have clear and stable pick-up thresholds under gradual increase of control coil current/voltage, thereby enabling the development of sensitive reed-based measuring units for protection purposes. In addition to the above, it should be noted that the dry reed switches are insensitive to the position in space and can be easily combined with electronic, electromagnetic and magnetic components to develop a number of different functional modules and devices on their basis [8].

High-quality vacuum and gas-filled reed switches manufactured by leading companies specializing in this industry (such as proposed for use in the device [9]) are not cheap ($15 - $30 each), but they are highly reliable and widely used not only in industry and communications, but also in military and aerospace. Thanks to a number of advantageous options, the reed switches occupy an intermediate position between semiconductors and electromechanical switching elements. Therefore, the automatic telephone exchanges, such as reed-based ATX ("Quant", etc.) are called "quasi-electronic". According to the specifications the lifetime of such ATX is 40 years and the number of failed reeds should not exceed 0.3 % within the period. So the figures speak for themselves.

However, the reed relays have one fundamental difference from the conventional electromechanical relays: their magnetic system is not isolated from the contacts, rather, it is formed by them. This difference causes low overcurrent capacity of the reed switches. Unlike the conventional relays, the reed switches do not withstand even the short-term contact overcurrent. This is due to fact that the magnetic field of the current flowing through the closed contacts of the reed switch is directed opposite to that of the coil magnetic field, holding the contacts closed, and weakens it, thus decreasing the contact force up to the appearance of the gap. This leads to the increased erosion and sometimes to welding of the reed contacts even under the short-term currents exceeding the maximum allowable value for particular type of the unit. Poor awareness of this aspect and ignorance of the reed differences from the conventional relays (with regard to the overload capacity) often leads to the equipment failures and, as a consequence, to the distrust of the reed switches. Under properly selected operating mode the reed switches provide reliable circuit switching throughout millions of operation cycles.

When using reed switches for switching external circuits with a wide variation of the current, no one wants to monitor the reed current operating mode. It is much easier not to use them at all, which often happens in practice. In the proposed construction some reed switches are included only in the device's internal circuits, where the current load is tenfold less than the maximum allowable reed current. Other reed switches turn off the digital input circuits with the maximum current of several milliamps, which is two orders less than the limit. And the current of several Amperes can flow only though the reed switch connected in-series with the output terminals of the DPR intended to switch-on the circuit breaker trip coil. However, 1) these reed switches do not switch these currents directly (they only assemble the circuit under no-current condition), and 2) their type (Bestact R15U, produced by the Japanese company Yaskawa) provides high current margin.

Question 3. Modern DPRs combine 10-20 and more different functions in a single module. Does it mean that the proposed protection device must contain the same number of input relays?

No, it does not. The point is that the variety of DPR functions embedded in a single terminal is based on the measurements of current, voltage and angle between them. Accordingly, the input relays of the proposed protection device must contain threshold elements for current, voltage and angle between them. Thresholds of pick-up of all these elements must be less than the minimum values selected as the DPR set points.

Question 4. Why do we need to use expensive DPRs together with some new and also expensive protection devices, if we can just go back to the cheap and resistant to IRDI electromechanical relays?

Indeed, the electromechanical protection relays (EPR) have been operated for over a hundred years and still provide reliable protection against emergency operation for all types of electrical equipment. Suffice it to say that some large and diversified national power systems (for example, in Russian) are even today nearly 70-90% equipped with EPRs. However, despite the fact that EPRs have proven their high reliability, about 30-40 years ago all the world's leading manufacturers of protective relays stopped developing and improving EPRs and began to intensively develop first the solid state relays completely duplicating the functions and characteristics of EPRs, and then the microprocessor-based relays with advanced features and improved performance. About 20-25 years ago, most of the world's leading manufacturers of protective relays stopped producing EPRs and concentrated all their efforts on the DPRs. The main reason for this phenomenon was that it was much more profitable to produce and test printed circuit boards with electronic components on the automatic equipment, than to produce miniature mechanical elements on high-precision turning and milling machines and manually assemble them into the rather complex mechanical design, make manual tests and customization.

Due to the large difference in production costs between the EPR and the DPR, the consumer stands to gain too as today the cost of DPRs produced by world's leading relay producers is much less than the cost of EPRs with similar characteristics. The statement that today the EPRs are much cheaper than MPDs is not correct and is not supported by the analysis of world market prices. For example, if electromechanical relay for three-stage line distance protection type LZ31 (made by ABB) could cost about 30-35 thousand USD (according to current prices), its microprocessor-based analogue with improved characteristics, such as the relay type D30 (made by General Electric) costs only 7,500 USD and Chinese analogue of the relay (type GTL- 823 made by Guatong Electric) costs even less – only 5,000 USD today.

In addition, a powerful advertising campaign pursued by the developers of DPRs, universities and research organizations interested in financing of new projects did the trick. Today, to raise the question about returning to EPRs means to become an outcast in the community of experts and to gain a character of retrograde who is trying to stop technological progress. None of the experts or decision-making officials would take such a responsibility. And even if they take it, it is safe to say that they will be inundated with charges of obscurantism and incompetence. In addition, for the sake of objectivity, it should be noted that DPRs do have some features and functionalities unattainable by EPRs.

With all of these factors in mind it can be stated that the question of returning to the EPR is not relevant.

Question 5. Suppose that the return to EPRs is really not possible today. But why not to use DPRs completed with these EPRs instead of inventing some new reed switch-based devices?

In fact, the combination of EPR and DPR has long been used in practice, Fig. 2.

Figure 2 : Section of Distance Protection panel for critical 160 kV lines consisting of electromechanical relays type LZ31 (above) connected in-parallel with microprocessor-based protection MiCOM P437 (bottom)

However, they are not connected in-series (as suggested) – they are connected in-parallel to duplicate each other in order to increase the reliability. As explained previously [5], such a method of DPR and EPR combination (i.e., in-parallel) is essentially not correct. In parallel connections the EPR really must completely duplicate the DPR functions and have the same set points. In any combination of multifunction DPRs and EPRs the whole set of expensive EPRs should be used, thus making this project very doubtful because of its high cost and availability of large areas for installation of a large number of different EPRs. The suggested protective reed-based device should be much more simple, smaller and cheaper than a set of EPRs necessary to protect one DPR. And this is the only way to make its use promising.

Question 6. To provide the versatility and full functionality the functional capabilities of the suggested protective device should be the same as of the set of EPRs. Hence, its cost should be about the same. Why will it be cheaper?

Let's look how the EPR works. For example, let's consider current-dependent time delay relay - Inverse Definite Minimum Time (IDMT). It is an electromechanical induction disc type relay where an aluminum disk begins to turn slowly and the movable contact associated with this disk starts to approach the fixed contact under the certain threshold current. After some time determined by the speed of disk rotation (based on the current flowing through the relay coil) the contact closes (via the intermediate relay) the circuit of the circuit breaker trip coil.

The starting element of the proposed DPR protection device requires no current-dependent time delay. This starting element should trip only under the certain current, somewhat smaller than the disk pickup current. That's all. Other functions are not required, since all other functions will be done by the activated DPR. That is, in this case, instead of a complex and expensive IDMT relay we need only the simplest relay consisting of a coil and a reed switch.

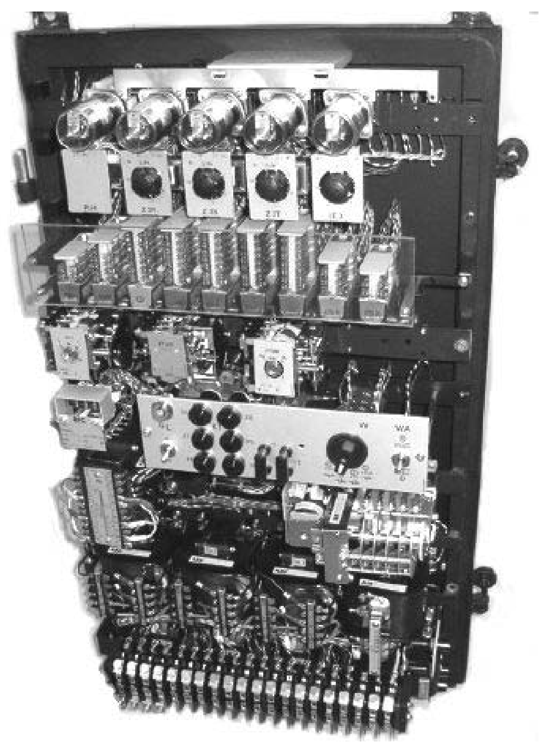

Figure 3 : Line Distance Protection relay LZ31 type

As another example, let's consider several types of Line Distance Protection relays. The electromechanical alternative of this relay (for example, type LZ31, Fig. 3) contains many complex and interconnected electromechanical assemblies providing three or four stages of line impedance measurement to the short circuit point corresponding to these delay stages, special form of characteristics, etc. As noted above, the cost of such relays is 30-35 thousand USD. However, the whole complex is actuated by the simplest starting element controlling the balance between line current and voltage, Fig. 4. The element is actuated by an imbalance between preset current and voltage values.

Large and complex distance protection relay types RYZKB, RYZOE, RYZFB, manufactured by ASEA in the 70s, Fig. 5, implement several protection features. However, all these relays are equipped with very simple starting element (see the diagram in Fig. 5).

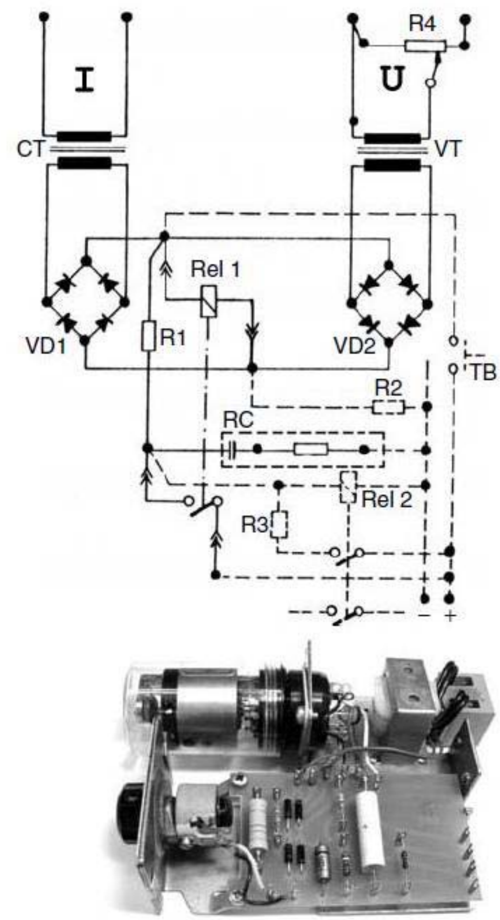

Figure 4 : Principle of operation and design of the starting element of the distance protection relay LZ31

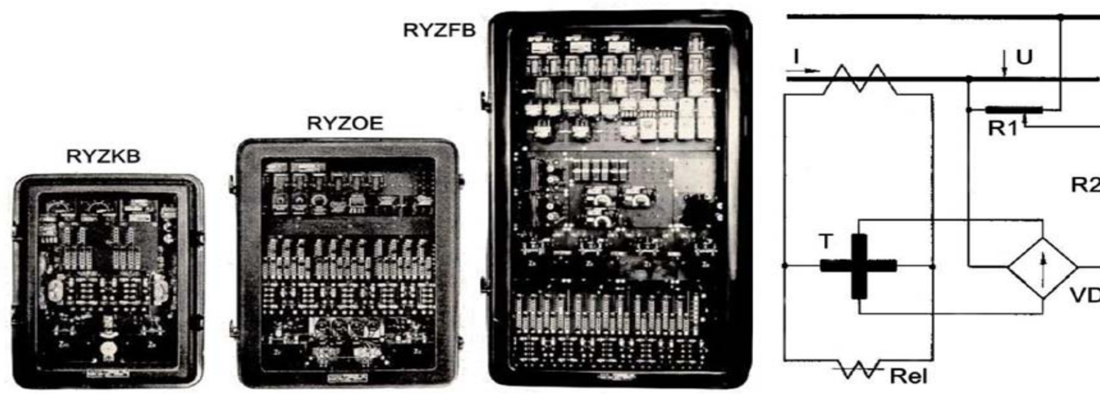

Figure 5 : Line Distance Protection electromechanical relays of various types made by ASEA and the diagram of their starting element (produced in 70s)

These starting elements were an integral part of complex structures and were not sold separately. The exceptions were some types of relays, which were produced in Russia, for example, relay type KPC - 112, Fig. 6, containing special chokes and four-pole inductors with rotor. In essence, this relay is a separate starting element of distance protection. However, it is too complex, expensive and large. Anyway, the combination of such obsolete designs with the up-to-date DPR technology is hardly a good idea.

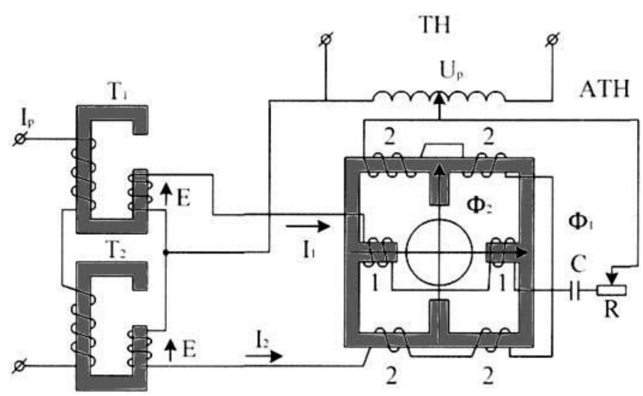

Figure 6 : Relay KPC-112 (ChEAZ, Russia) with the induction mechanism

In this respect, starting element of distance protection of type HZM (Westinghouse) could be much more appropriate, Fig. 7.

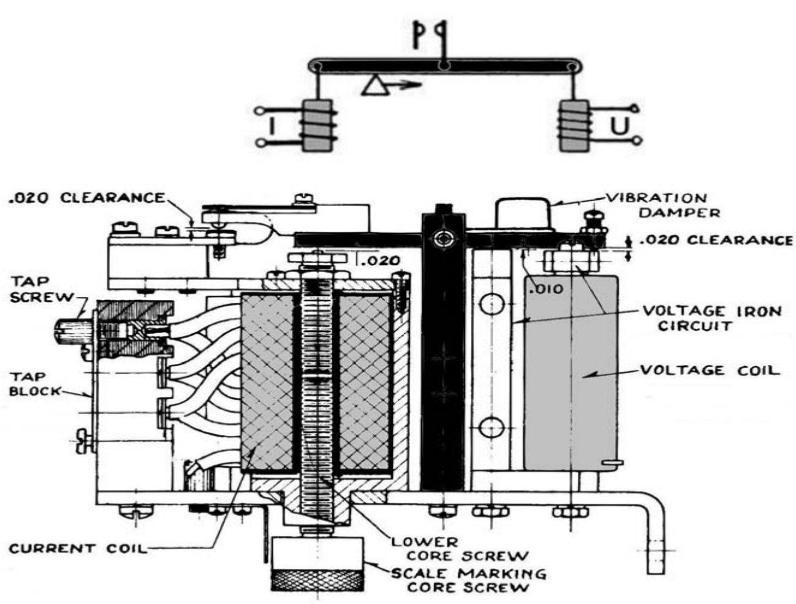

Figure 7 : Balance electromagnetic starting element used in distance protection relay of type HZM (Westinghouse)

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This is a very simple device comprising of T-shaped core with swinging rocker (upper part of the letter "T") and two coils: voltage coil and current coil acting on the ends of the rocker. The position of the rocker with attached contact depends on the balance of the magnetic fields generated by the current and voltage coils. This assembly is an internal part of HZM relay design and has never been sold separately.

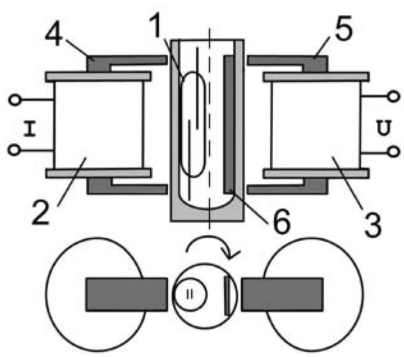

Figure 8 : The simplest starting element of distance protection with adjustable threshold

1 – reed switch; 2 and 3 - coils with control windings; 4 and 5 - U-shaped flat ferromagnetic cores; 6 - magnetic shunt.

The reed relay, built on the same principle of balance between current and voltage (Fig. 8), is much simpler and more reliable [9]. This relay responds to the difference between the magnetic fields generated by the current and voltage coils and its threshold can be adjusted within a wide range by turning the reed capsule. Such a starting element can be successfully used in the SU.

Thus, the proposed device with a small number of reed-based elements of current, voltage and the difference between them is much simpler and cheaper than a full-featured set of EPRs. In addition, reed-based starting elements do not require maintenance during operation, have significantly less delay within the overall relay response time and provide a high level of insulation between input and output unattainable for old EPRs.

Question 7. In some cases, the circuit breaker trip command is issued directly by the protective relays (such as transformer gas protection relay) and is simultaneously duplicated by the signals sent to the logical inputs of DPR, thus triggering the fault recorder. How then will the proposed device (which blocks logical inputs of DPR) work?

This is easy to resolve: it is only necessary to send a signal from trigger relay contacts (in this case, gas relay) also to one of the inputs of SU of protection device. In this case the DPR is unblocked and the fault recorder starts operation and records gas relay trip information.

Question 8. There is a requirement of inadmissibility to include any additional locking elements into the circuit of breaker trip coil, and in the proposed device this circuit is switched by the contact of the addition auxiliary relay. Is this acceptable?

In fact, normally the open contact of an auxiliary relay is not connected into the trip coil circuit breaker. It is connected into the circuit connecting the output DPR relay contact to the trip coil of the circuit breaker. That is, this additional contact does not block the control circuit of the circuit breaker trip coil; it only blocks the output circuit of DPR. The control circuit of the breaker trip coil remains free to connect any external contacts or manually operated keys.

Question 9. How to deal with complex protection units, for example, with the protection units providing the offset from the transformer excitation current inrush and containing filters of 2 and 5 harmonics? Should the proposed device also contain such filters? Or another example: the differential protection. How to ensure the device operation if the emergency mode exists only in the protected area?

No, the SU does not need such filters or excitation current inrush offset for operation. The tripping of the SU under the transformer excitation current inrush only unblocks the DPR for about 10 seconds and nothing else. The DPR will block against unnecessary pick-ups with its own algorithm. After 10 seconds the SU reverts to the original state and blocks the DPR again. The same applies to the differential protection. The SU device does not care where the fault is, within or outside the protected area. The only important thing is the presence of short circuit current, while the damage zone will be determined by the DPR after the SU unlocks it. The SU response time is about 6 ms and almost does not affect the total response time of relay protection since the DPR proper time is 30-40 ms.

Question 10. If the DPR and the EPR are connected in-series, the capabilities of relay protection will actually be limited by the capabilities of the EPR, as it has reduced capabilities and characteristics. Is this good?

No, it is not. The proposed device in no way defines any properties or characteristics of relay protection. It only unblocks the DPR at the moment when at least one parameter of the entire set of monitored parameters approaches the DPR set point. The subsequent behavior of the relay protection and its response to emergency mode will be completely determined by the properties and characteristics of the DPR.

It is obviously that in practice there are more complex modes of DPR operation, which are not discussed in the article, and such modes will require special starting elements to be developed. This is possible. However, even if this will require the development of such a special starting element, then based on a combination of reed switches and magnetic circuits it is possible to create simpler, cheaper and faster units compared to the traditional electromechanical relays. For example, the device shown in Fig. 9 may be well used to control the angle between the current and the voltage or as power measuring element.

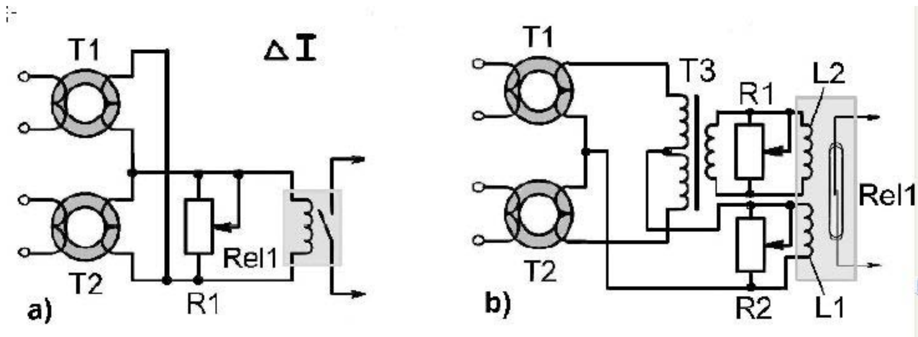

Figure 9 : Variants of quasi-electronic starting elements of differential protection

The combination of magnetic, reed switch-based and high-voltage discrete semiconductor elements provides additional opportunities. For example, Fig. 9a illustrates the simplest device responding to the current difference, and Fig. 9b illustrates the device with differential current desensitization by the passing current value (offset).

Thus, the above analysis clearly shows that the practical implementation of the proposed method of DPR protection is quite feasible from the technical and economic point of view. Certainly, such an implementation should be done by the manufacturers of DPRs, which can offer consumers a quasi-electronic SU as an option for improving the safety and reliability of the relay protection for critical objects.

III. Interest in the Proposed DPR Protection Device and the Experts Opinion

Naturally, the production may be interesting for manufacturers of DPRs, while the question of interest in application of this device is much more complicated. Some relay experts believe that their professional duties are limited only to the operation of relay protection. They openly declare that they absolutely don't care about the problems of relay protection against IRDI. In their opinion, these issues must be resolved by "special services" (Intelligence).

These experts believe that they have enough problems with relay protection, so they are totally against any additional remedies that may complicate their already difficult lives.

Many experts enthusiastically accept any newfangled ideas and trends in relay protection, no matter what these innovations may lead to. Whether it is "proactive relay protection", "artificial intelligence" relaying, "digital substation" or "smart grid", they accept anything … to get the money for development and implementation. Why not? Why not to make some money, as long as the money is green?! Moreover, such projects are well paid and even covered by special national programs. So why not come up with some other "smart" toys for relay protection? The main thing is to provide attractive groundings and use scientific words for designation.

There are experts who believe that if the computational capabilities of modern microprocessors used in the DPR allow providing relay protection and a lot of other options and functions, such options should be fully used. Why not to connect the set of different sensors to the DPR and use it to monitor the status of electrical equipment? Why not to use the same DPR as an information-measurement system or ACS substation controller in addition to its basic functions? That is, they support the principle: the more the better and you always "need" if you "can".

Some officials responsible for making strategic decisions on directions of relay protection development act on the principle: if any developed nation began to work in a certain direction, we cannot keep up, let's catch up and overtake! This policy reminds me an anecdote narrated by the Englishman, a good specialist in relay protection:

A large tribe of Indians lived near one of the big American cities. They were dissatisfied with their weather predictor, who made too many mistakes. So their forward-minded chief decided to call the Weather Forecast Office, as the weather played a very important role in the life of the Indians. He was told that an accurate forecast was not available yet, but it was assumed that the coming winter should be cold. So, the Indians began to gather firewood just in case. After some time, the chief decided to check the forecast and called the Weather Forecast Office again. The answer was not long in coming: "Yes, of course, the winter will be very cold, now we know it! Look how actively the Indians are gathering firewood for the winter; they never make a mistake!"

Many experts think: we want to move forward in the direction of technological progress, so you must provide us with protection against all these IRDI! But it is time to realize that the current trends of relay protection development in the direction of "digital substations", "smart grids", relay protection with artificial intelligence and so on, where the DPRs are not only used as a protection relays, but also receive commands and communicate with many external objects across multiple communication channels, do not enable proving reliable protection of the DPR against existing dangers.

Some apologists of the DPR perceive the above statement as rejection of everything new and progressive. One prominent and important functionary in the field of relay protection wrote "the author is trying to slow down the technological progress" in the review on the article devoted to hazards of modern tendencies in relay protection development. Such guardians of technological progress either mistakenly or deliberately garble my position, since I don't advocate abandoning new technologies in the field of diagnosis of electrical equipment or information-measuring systems, but only separating their relaying.

IV. Summing Up

I firmly believe that we can ensure effective protection of the DPR against IRDI, as well as improve the reliability of its operation under normal operating conditions, only if the microprocessor-based relay protection will be used solely to meet the challenges of relay protection. I also believe that after the stormy period of euphoria associated with the new opportunities in relay protection provided by microprocessor technology, it is the time to come down to earth and re-evaluate the situation.

However, I have to say with much regret that the majority of experts and officials engaged in resolving the pressing problems of today have little interest in the potential problems and hazards associated with current trends in relay protection development. This is understandable, because none of them will take the responsibility for the collapse of the power system under the IRDI, since no instructions or regulations on this subject exist yet. On the other hand, the implementation of new techniques and new technologies brings honors and awards.

Probably, the developers of electromagnetic and cybernetic weapons give today's tendencies in protective relaying an ovation, like the snake who is satisfyingly watching a frog trying to jump into its mouth.

We can only trust that the wisdom of experts and officials will take precedence over the short-term mercantile interests and hope that as expressed in the well known proverb (a peasant needs thunder to cross himself and wonder) the "peasant" will not wait for the "thunder".

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New Delay Less Sub Band Adaptive Filtering Algorithm for Active Noise Control Systems

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Abstract- The delay less SAF scheme in an ANC system involves the decomposition of input noise (i.e., the reference signal) and error signals into sub bands using analysis filter banks, and combining the sub band weights into a full-band noise canceling filter by a synthesis filter bank called weight stacking. Typically, a linear-phase finite-impulse response (FIR) low-pass filter (i.e., prototype filter) is designed and modulated for the design of such filter banks. The filter must be designed so that the side-lobe effect and spectral leakage are minimized. The delay in filter bank is reduced by prototype filter design and the side-lobe distortion is compensated for by oversampling and appropriate stacking of sub band weights. Experimental results show the improvement of performance and computational complexity of the proposed method in comparison to two commonly used sub band and block adaptive filtering algorithms.

Keywords: DFT, dolph-linear-phase finite-impulse response, sub band adaptive filtering, SAF, UDFM.

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NEWDELAYLESSSUBBANDADAPTIVEFILTERINGALGORITHMFORACTIVENOISECONTROLSYSTEMS

Strictly as per the compliance and regulations of :

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Abstract- The delay less SAF scheme in an ANC system involves the decomposition of input noise (i.e., the reference signal) and error signals into sub bands using analysis filter banks, and combining the sub band weights into a full-band noise canceling filter by a synthesis filter bank called weight stacking. Typically, a linear-phase finite-impulse response (FIR) low-pass filter (i.e., prototype filter) is designed and modulated for the design of such filter banks. The filter must be designed so that the side-lobe effect and spectral leakage are minimized. The delay in filter bank is reduced by prototype filter design and the side-lobe distortion is compensated for by oversampling and appropriate stacking of sub band weights. Experimental results show the improvement of performance and computational complexity of the proposed method in comparison to two commonly used sub band and block adaptive filtering algorithms. Sub band adaptive filtering (SAF) techniques play a prominent role in designing active noise control (ANC) systems. They reduce the computational complexity of ANC algorithms, particularly, when the acoustic noise is a broadband signal and the system models have long impulse responses. In the commonly used uniform-discrete Fourier transform (DFT) -modulated (UDFTM) filter banks, increasing the number of sub bands decreases the computational burden but can introduce excessive distortion, degrading performance of the ANC system. In this paper, we propose a new UDFTM-based adaptive sub band filtering method that alleviates the degrading effects of the delay and side-lobe distortion introduced by the prototype filter on the system performance.

Keywords: DFT, dolph-linear-phase finite-impulse response, sub band adaptive filtering, SAF, UDFM.

I. Introduction

Subband adaptive filtering (SAF) techniques play a prominent role in designing active noise control (ANC) systems. They reduce the computational complexity of ANC algorithms, particularly, when the acoustic noise is a broadband signal and the system models have long impulse responses. Active noise control (ANC) is a method of canceling a noise signal in an acoustic cavity by generating an appropriate anti-noise signal via canceling loudspeakers.

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In general, the SAF methods offer a good alternative approach to meet ANC system requirements, due to their inherent spectral decomposition and down sampling operations. Since the spectral dynamic range and eigen value spread of the covariance matrix of noise signal decrease in each sub band, the performance, i.e., convergence rate, noise attenuation level, and stability of the ANC system, improves using SAF techniques. Hence, one expects that increasing the number of sub bands (or block length) M should improve the performance.

The delay less SAF scheme in an ANC system involves the decomposition of input noise (i.e., the reference signal) and error signals into sub bands using analysis filter banks, and combining the sub band weights into a full-band noise canceling filter by a synthesis filter bank called weight stacking. Typically, a linear-phase finite-impulse response (FIR) low-pass filter (i.e., prototype filter) is designed and modulated for the design of such filter banks. The filter must be designed so that the side-lobe effect and spectral leakage are minimized. The latter requires a high-order FIR filter, introducing a long delay, which increases with M as the bandwidth shrinks. The long delay and side-lobe interference introduced by the prototype filter degrade the performance of SAF algorithms for large M, limiting the computational saving that can be obtained by increasing the number of sub bands. Improving the system performance and reducing the computational burden by increasing M has inspired the work presented herein. The focus of this project is to design a high-performance SAF algorithm. The performance limiting factors of existing SAF structures were found to be due to the inherent delay and side-lobes of the prototype filter in the analysis filter banks.

A delay less structure targeted for low-resource implementation is proposed to eliminate filter bank processing delays in sub band adaptive filters (SAFs). Rather than using direct IFFT or poly phase filter banks to transform the SAFs back into the time-domain, the proposed method utilizes a weighted overlap-add (WOLA) synthesis. Low-resource real-time implementations are targeted and as such do not involve long (as long as the echo plant) FFT or IFFT operations. Also, the proposed approach facilitates time distribution of the adaptive filter reconstruction calculations crucial for efficient real-time and hardware implementation. The method is implemented on an

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oversampled WOLA filter bank employed as part of an echo cancellation application. Evaluation results demonstrate that the proposed implementation outperforms conventional SAF systems since the signals used in actual adaptive filtering are not distorted by filter bank aliasing. The method is a good match for partial update adaptive algorithms since segments of the time-domain adaptive filter are sequentially reconstructed and updated.

A delay less method for adaptive filtering through SAF systems is proposed. The method, based on WOLA synthesis of the SAFs, is very efficient and is well mapped to a low-resource hardware implementation. The performance of an open-loop version of the system was compared against a conventional SAF system employing the same WOLA analysis/synthesis filter banks, with the proposed delay less system offering superior performance but at greater computational cost. The performance is identical to the DFT-FIR delay less SAF system that employs straightforward poly phase filter banks.

However, the proposed WOLA-based TAF synthesis offers a superior mapping to low-resource hardware with limited-precision arithmetic. Also the WOLA adaptive filter reconstruction may easily be spread out in time simplifying the necessary hardware. This time-spreading may be easily combined with partial update adaptive algorithms to reduce the computation cost for low-resource real time platforms.

II. Generic Invertibility of Multidimensional Fir Multirate Systems and Filter Banks

We study the inevitability of M-variety polynomial (respectively: Laurent polynomial) matrices of size N by P. Such matrices represent multidimensional systems in various settings including filter banks, multiple-input multiple-output systems, and MultiMate systems. The main result of this paper is to prove that when $N - P \geq M$, then $H(z)$ is generically invertible; whereas when $N - P < M$, then $H(z)$ is generically noninvertible. As a result, we can have an alternative approach in design of the multidimensional systems. During the last two decades, one dimensional multiage systems in digital signal processing were thoroughly developed. In recent years, due to the high demand in multidimensional processing including image and video processing, volumetric data analysis and spectroscopic imaging, multidimensional multirate systems have been studied more extensively.

One key property of a multidimensional multirate system is its perfect reconstruction, which guarantees that an original input can be perfectly reconstructed from the outputs. We show that there is a sharp phase transition on the invariability depending on the size and dimension of a given Laurent polynomial matrix. Specifically when $N - P \geq M$, the $N \times P$ polynomial (resp. : Laurent polynomial) of M-variety matrix is generically invertible; whereas when $N - P < M$, the matrix is generically noninvertible. Using this sharp phase transition property, we develop a fast algorithm to compute a particular left inverse for a given Laurent polynomial matrix. These results suggest an alternative approach in designing multidimensional filter banks by freely generating filters for the analysis side first. If we allow an amount of over sampling then we can almost surely find a perfect reconstruction inverse for the synthesis poly phase matrix. These results also have potential applications in multidimensional signal reconstruction from multi-channel filtering and sampling. Speech signals from the uncontrolled environments may contain degradation components along with the required speech components. The degradation components include background noise, reverberation and speech from other speakers. The degraded speech gives poor performance in automatic speech processing tasks like speech recognition and speaker recognition and is also uncomfortable for human listening [1]. The degraded speech therefore needs to be processed for the enhancement of speech components. Several methods have been proposed in the literature for this purpose, majority them can be grouped into spectral processing and temporal processing methods. In spectral processing methods, the degraded speech is processed in the transform domain, where as, in temporal processing methods, the processing is done in the time domain, for enhancing the speech components. Each of them has their own merits and demerits. These two approaches may be effectively combined by exploiting their merits and aiming to minimize the demerits. This may lead to speech enhancement methods which are more effective and robust compared to only spectral or temporal processing.

Frequency-domain and sub band implementations improve the computational efficiency and the convergence rate of adaptive schemes. The well-known multi delay adaptive filter (MDF) belongs to this class of block adaptive structures and is a DFT-based algorithm. In this paper, we develop adaptive structures that are based on the trigonometric transforms DCT and DST and on the discrete Hartley transform (DHT). As a result, these structures involve only real arithmetic and are attractive alternatives in cases where the traditional DFT-based scheme exhibits poor performance. The filters are derived by first presenting a derivation for the classical DFT-based filter that allows us to pursue these extensions immediately. The approach used in this paper also provides further insights into sub band adaptive filtering.

III. The Implementation of Delay Less Sub Band Active Noise Control Algorithms

Wideband active noise control systems usually have hundreds of taps for control filters and the cancellation path models, which results in high computational complexity and low convergence speed. Several active noise control algorithms based on sub band adaptive filtering have been developed to reduce the computational complexity and to increase the convergence speed. The sub band structure is similar to the frequency domain structure but differs in the time domain processing of the sub band signals. This paper discusses several issues associated with implementing the delay less sub band active noise control algorithms on a DSP Platform, such as the modeling of the cancellation path in sub bands and the partial Update of different sub bands.

Single channel ANC systems often use sub band techniques to overcome the difficulties of high computational complexity and low convergence speed associated with a wideband control filter containing thousands of taps.

This paper will discuss various method of noise reduction for wireless communication network. Noise is an, unwanted and inevitable interference, in any form of communication. It is non-informative and plays the role of sucking the intelligence of the original signal. Any kind of processing of the signal contributes to the noise addition. A signal traveling through the channel also gathers lots of noise. It degrades the quality of the information signal. The effect of noise could be reduced only at the cost of the bandwidth of the channel, which is again undesired, as bandwidth is a precious resource. Hence to regenerate original signal, it is tried to reduce the power of the noise signal, or in the other way, raise the power level of the Informative signal, at the receiver end this leads to improvement in the signal to noise ratio(SNR).

Adaptive algorithms that allow neighboring nodes to communicate with each other at every iteration. At each node, estimates exchanged with neighboring nodes are fused and promptly fed into the local adaptation rules. In this way, an adaptive network is obtained where the structure as a whole is able to respond in real-time to the temporal and spatial variations in the statistical profile of the data. Different adaptation or learning rules at the nodes, allied with different cooperation protocols, give rise to adaptive networks of various complexities and potential. Obviously, the effectiveness of any distributed implementation depends on the modes of cooperation that are allowed among the nodes. Figure 1 illustrates three such modes of cooperation. In an incremental mode of cooperation information flows in sequential manner from one node to the adjacent node. This mode of operation requires a cyclic pattern of collaboration among the nodes, and has the advantage that for the last node in the cycle, the data from the entire network are used to update the desired parameter estimate, thereby offering excellent estimation performance.

Moreover, for every measurement, every node needs to communicate with only one neighbor. However, incremental cooperation has the disadvantage of requiring the definition of a cycle, and network processing has to be faster than the measurement process, since a full communication cycle is needed for every measurement. This may become prohibitive for large networks. Incremental networks are also less robust to node and link failures. An alternative protocol is the diffusion implementation where every node communicates with all of its neighbors as dictated by the network topology. This approach has no topology constraints and is more robust to node and link failure. It will have some performance degradation compared to an incremental solution, and also every node will need to communicate with its neighbors for every measurement, possibly requiring more energy than the incremental case.

The mainstay of the proposed model is improving the system performance and reducing the computational burden. In this paper, we first demonstrate that the increased delay degrades the system performance more than that of the spectral leakage (or side-lobe effects) in a uniform sub-band filtering method. It is shown how the spectral leakage can be reduced by choosing a proper decimation factor and weight stacking methodology. We then present a new SAF (Sub-Band Adaptive Filtering) algorithm that reduces computational complexity by increasing the number of subbands M without degrading the performance of the ANC (Active Noise Control) system. The performance of the proposed method is compared with those of MT (Moragan and Thi) and DFT-MDF (Discrete Fourier Transform and Multi-Delay Adaptive Filter) methods. The results show that the maximum noise attenuation level (NAL) of the proposed method is higher than that of MT and comparable to that of the DFT-MDF method. However, the new method achieves the maximum NAL with much lower computational complexity and higher robustness than the other two methods.

IV. Methodology

The gradient based adaptation starts with an old optimization technique known as the method of steepest descent. This has been discussed in the next chapter in detail. It is recursive in the sense that starting from some initial arbitrary value for tap weight vector, it improves with increasing number of iterations. The final value so computed for tap weight vector converges to Wiener solution. The fixed step size least mean square (FSS LMS) algorithm is an important member of the

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family of stochastic gradient algorithms. The term stochastic gradient is intended to distinguish it from the method of steepest descent that uses deterministic gradient in a recursive computation of the Wiener filter for stochastic inputs. This algorithm does not require measurements of the pertinent correlation functions, nor does it require matrix inversion. Subsequent works have discussed issue of optimization of step size or methods of varying step size to improve performance. There are different types of adaptive filtering algorithms, they are 1. Least mean square (LMS) algorithm. 2. Normalized least mean square (NLMS) algorithm. 3. Variable step size LMS (VSLMS) algorithm. 4. Variable step size Normalized LMS (VSNLMS) algorithm. 5. Recursive least squares (RLS) algorithm.

a) Normalized least mean Square (NLMS) Algorithm

One of the primary disadvantages of the LMS algorithm is having a fixed step size parameter for every iteration. This requires an understanding of the statistics of the input signal prior to commencing the adaptive filtering operation. In practice this is rarely achievable. Even if we assume the only signal to be input to the adaptive echo cancellation system is speech, there are still many factors such as signal input power and amplitude which will affect its performance.

The normalized least mean square algorithm (NLMS) is an extension of the LMS algorithm which bypasses this issue by selecting a different step size value, μ(n), for each iteration of the algorithm. This step size is proportional to the inverse of the total expected energy of the instantaneous values of the coefficients of the input vector **x**(n). This sum of the expected energies of the input samples is also equivalent to the dot product of the input vector with itself, and the trace of input vectors auto-correlation matrix, **R**.

$$tr(\boldsymbol{R}) = \sum_{i=0}^{N-1} E[x^2(n-i)]$$

$$= E[\sum_{i=0}^{N-1} x^2(n-i)]$$

The recursion formula for the NLMS algorithm is stated in equation.

$$w(n+1)=w(n)+\frac{1}{x^T(n)x(n)}e(n)x(n)$$

b) Derivation of the NLMS algorithm

To derive the NLMS algorithm consider the standard LMS recursion, for which we select a variable step size parameter, μ(n). This parameter is selected so that the error value, $e^+(n)$, will be minimized using the updated filter tap weights, w(n+1), and the current input vector, **x**(n). $w(n+1) = w(n) + 2\mu(n)e(n)x(n)$, $e^+(n) = d(n) - w^T(n+1)x(n)$, $=(1- 2\mu(n)x^T(n)x(n))e(n)$

Next we minimize $(e^+(n))^2$, with respect to μ(n). Using this we can then find a value for μ(n) which forces $e^+(n)$ to zero.

$$\mu(n) = \frac{1}{2x^T(n)x(n)}$$

This μ(n) is then substituted into the standard LMS recursion replacing μ, resulting in the following NLMS equation.

$$w(n+1) = w(n) + 2\mu(n)e(n)x(n) \ ,$$

$$w(n+1)=w(n)+\frac{1}{x^T(n)x(n)}e(n)x(n)$$

c) Implementation of the NLMS algorithm

The NLMS algorithm has been implemented in Matlab and in a real time application using the Texas Instruments TMS320C6711 Development Kit. As the step size parameter is chosen based on the current input values, the NLMS algorithm shows far greater stability with unknown signals. This combined with good convergence speed and relative computational simplicity makes the NLMS algorithm ideal for the real time adaptive echo cancellation system. As the NLMS is an extension of the standard LMS algorithm, the NLMS algorithms practical implementation is very similar to that of the LMS algorithm. Each iteration of the NLMS algorithm requires these steps in the following order

(a) The output of the adaptive filter is calculated.

$$y(n) = \sum_{i=0}^{N-1} w(n)x(n-i) = w^T(n)x(n)$$

(b) An error signal is calculated as the difference between the desired signal and the filter output

$$e(n)=d(n)-y(n)$$

(c) The step size value for the input vector is calculated.

$$\mu(n) = \frac{1}{2x^T(n)x(n)}$$

(d) The filter tap weights are updated in preparation for the next iteration.

$$w(n+1) = w(n) + \mu(n)e(n)x(n)$$

Each iteration of the NLMS algorithm requires 3N+1 multiplications, this is only N more than the standard LMS algorithm, this is an acceptable increase considering the gains in stability and echo attenuation achieved.

V. Active Noise Control System

Active noise control (ANC) is a method of canceling a noise signal in an acoustic cavity by generating an appropriate anti-noise signal via canceling loudspeakers. Due to recent advances in wireless technology, new applications of ANC have emerged, e.g., incorporating ANC in cell phones, Bluetooth headphones, and MP3 players, to mitigate the environmental acoustic noise and therefore improve the speech and music quality. For practical purposes, ANC as a real-time adaptive signal processing method should meet the following requirements: 1) minimum computational complexity (lower computational delay and power consumption), 2) stability and robustness to input noise dynamics, and 3) maximum noise attenuation.

Acoustical noise can sometimes disturb or even harm nearby people. Hence, it is necessary to find ways to reduce such unwanted noise. Traditionally, passive means (i.e., physical barriers) to attenuate the noises have been employed. Unfortunately, the barriers are not effective to isolate lower frequency noises; and to achieve significant reduction the barriers have to be rather bulky. In effect, the passive barrier is not a cost-effective solution to reducing low-frequency noises (for example, noises that come from industrial blowers, diesel engines, transformers, earth-moving machines, and propeller-driven aircraft.) Because of that shortcoming of the physical barriers, active means to reduce low frequency noise (less than 500-1000 Hertz) have been investigated by researchers in the field of adaptive acoustic control. Active noise control (ANC) promises a good reduction of the noises in the form of a small package of a DSP controller, microphone(s), and loudspeaker(s). For the better or the worse, the ANC systems are effective only when the intended noise is periodic, and so random noises like the white noise will not be reduced.

There are different ANC schemes that have been developed. My project is involved with the implementation of one of the schemes that is called single-channel adaptive feedback ANC. The implementation was on a Texas Instruments TMS320C54 evaluation module (EVM) board; in addition to this, I used a microphone and a loudspeaker. Two types of noise exist in the environment, broadband noise, where its energy is more or less evenly distributed across the frequency spectrum, or narrowband noise, where the energy is mostly concentrated around specific frequencies. In ANC roughly two types of control strategies can be distinguished as shown by Fuller, their use strongly depends on the deterministic behavior of the disturbance:

Feedback ANC: A controller is used to modificate the response of a system, for example by adding artificial damping. In this way vibration levels can be reduced even for a broadband random disturbance.

Feedforward ANC: When the disturbance is deterministic, or in particular harmonic, a controller can be used to adaptively calculate a signal that cancels the disturbance. When vibrations are induced by rotating machinery this often results in harmonicvibrations and the amount of noise reduction achieved by feedforward ANC systems is far superior to that of feedback ANC systems as shown by Hansen & Snyder. The basic idea of feedforward ANC is to generate a signal (secondary noise), that is equal to a disturbance signal (primary noise) in amplitude and frequency, but has opposite phase. Combination of these signals results in cancellation of the primary (unwanted) noise. This ANC technique is well-known for its use in cancelling unwanted sound as shown by Nelson & Elliott [6], but it is used for the control of vibration. A block diagram of an adaptive digital filter is shown in fig.2.1, where n is a time index. This filter forms the basis for feedforward ANC, based on the FXLMS algorithm. The adaptive filter actually consists of two parts. The digital filter, W(z) calculates its output by using a reference x(n) and adjustable filter coefficients, or weights. The filter coefficients are updated by an adaptive algorithm, using x(n) and an error signal e(n) in such a way that the squared error $e^2(n)$ is minimized.

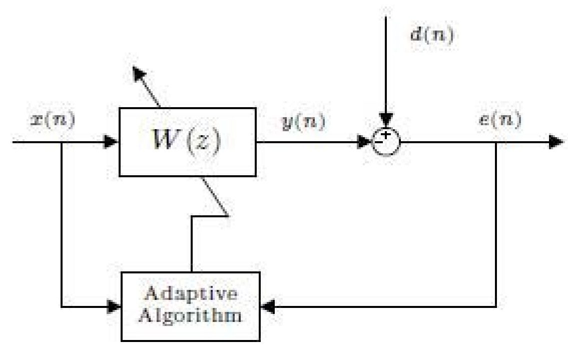

Figure 5.1 : ANC using an adaptive filter

We can define the error e(n) as:

$$e(n) = d(n) - y(n) \qquad (5.1)$$

where d(n) is an unwanted disturbance. The adaptive filter will try to calculate an output y(n) that is equal to the unwanted disturbance d(n), so this disturbance will be cancelled.

a) Concept of an ANC system

The basic concept of the feedforward ANC system that is used with the experimental setup can be found in Figure 2.2, where the grey part represents the controller and the white part represents the physical world. This is a very general concept, in this report vibrations are considered, but it can also be applied to acoustic applications as shown by Nelson and Elliott [6] or more specific to sound cancellation in ducts as shown by Kuo and Morgan [5].

b) System Description

The harmonic noise is produced at the noise source (e.g. an engine or a shaker).

Through the transfer function P(z) of the primary path this results in a vibration d(n) somewhere in the construction. This vibration will be reduced, by generating the appropriate controller output y(n) and sending it trough the transfer function S(z) of the Secondary Path to the construction. The remaining vibration e(n) can then be measured by a sensor. The adaptive filter looks similar to that of Figure 2.1 but is slightly more complicated. That is to compensate for the effects of the Secondary Path, which will be explained later.

c) Conventional versus Indirect Feedforward ANC

In conventional feedforward ANC systems, the disturbance frequency information is available or can be derived from the noise source, for example from the engine velocity. When the disturbance frequency is exactly known, the reduction that can be achieved by a conventional feedforward ANC system has its limit at infinity for the ideal case with a pure harmonic noise-free disturbance and linear Secondary Path. In other applications the disturbance frequency information may not be available, because the disturbance frequencies are unknown or slowly varying. In that case indirect feedforward ANC can be used as shown in this report, where the reference signal x(n) is generated from the error e(n), instead of from the frequency information of the noise source. Conventional feedforward ANC with a single frequency disturbance was imple- mented on the experimental setup by H.J. van der Veen. This report focuses on different kinds of indirect feedforward ANC methods, where if possible harmonic disturbances with two frequencies are used. They are tested at the experimental setup and will be compared with each other.

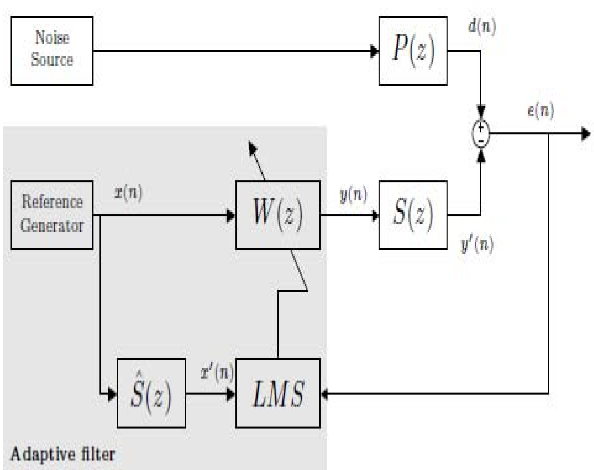

Figure 5.1 : Block diagram of an ANC system

In practical applications there is a transfer function S(z) between the digital controller signal and the physical world, which contains the D/A converter, power amplifier, actuator element and construction. In general, this Secondary Path transfer function S(z) gives a change in amplitude and a phase shift, so the adaptive filter should compensate for the effects of S(z) to ensure convergence. A straightforward solution would be to place the inverse $S(z)^{-1}$ in series with S(z), but because this inverse does not necessarily exists, the so-called Filtered-x LMS (FXLMS) algorithm is more generally used. This algorithm places an estimate of S(z) in the reference signal to the weight update.

For the ANC system of Figure 2.2, containing a Secondary Path transfer function S(z), the residual error can be expressed as:

$$e(n)=d(n)-y`(n); \qquad (5.2)$$

where y`(n) is the output of the Secondary Path S(z). If S(z) is assumed as an IIR filter with denominator coefficients $[a_1,\ldots\ldots\ldots,a_N]$ and numerator coefficients $[b_0,\ldots\ldots.b_{M-1}]$, then the filter output y`(n) can be written as the sum of the filter input y(n) and the past filter output:

$$y'(n)=\sum_{i=1}^{N}a_i y'(n-i)+\sum_{j=0}^{M-1}b_j y(n-j). \qquad (5.3)$$

It can be achieved in a similar way that the gradient estimate becomes:

$$\nabla\hat{\xi}(n)=-2\mathbf{x}'(n)e(n), \qquad (5.4)$$

where:

$$\mathbf{x}'(n)=\sum_{i=1}^{N}a_i\mathbf{x}'(n-i)+\sum_{j=0}^{M-1}b_j\mathbf{x}(n-j). \qquad (5.5)$$

Note that in practical applications, S(z) is not exactly known, therefore the parameters a_i and b_j are the parameters of the Secondary Path Estimate S^ (z). The weight update equation of the FXLMS algorithm is:

$$w(n+1)=w(n)+\mu x`(n)e(n) \qquad (5.6)$$

and x`(n) can be calculated from Equation 5.5.

The FXLMS algorithm is very tolerant to modelling errors in the Secondary Path Estimate S^ (z) as shown by Kuo & Morgan [5]. The algorithm will converge when the phase error between S(z) and S^ (z) is smaller than 90^0. Convergence will be slowed down though, when the phase error increases. From the weight update Equation 2.6 can be seen that a step size μ has to be chosen. This step size affects important properties such as performance, stability and error after convergence. A more in-depth analysis can be found in Kuo & Morgan [5] and Elliott & Nelson. Furthermore, a modification of the standard FXLMS is presented to make the choice of μ independent of the power of x`(n).

VI. System Analysis

a) Adaptive Filter Theory

In the past three decades, interest in adaptive systems has increased, leading to widespread use of adaptive techniques in fields such as Communications, Signal Processing, Sonar and Biomedical Engineering.

Adaptive systems adapt to the environment changes and search for the optimal system parameters based on a reference signal. In the case of a filter, the system parameters are the tap weights of the filter. The performance of an adaptive algorithm is highly dependent on the reference input and additive noise statistics. In the context of Wiener filter theory, there are assumptions of time invariance, linearity and Gaussian statistics such that the mean square error criteria will be the optimum cost function. These assumptions are often for the ease of mathematical analysis, but do not take into account of the broader problems of signals with non-Gaussian statistics. In the digital communication systems, efficient bandwidth utilization is economically important to maximizing profits, while at the same time maintaining performance and reliability. More importantly, the adaptive filter solution has to be relatively simple, which often leads to the use of the conventional Least Mean Square (LMS) algorithm. However, the performance of the LMS algorithm is often sub-optimal and the convergence rate is small. This, therefore, provides the motivation to explore and study variable step size LMS adaptive algorithms for various applications.

b) The Wiener Filter

These are a class of linear optimum discrete time filters known collectively as Wiener filters. Wiener filters are a special class of transversal Finite impulse response (FIR) filters that build upon the Mean Square Error (MSE) cost function to arrive at an optimal filter tap weight vector, which reduces the MSE signal to a minimum. Theory for a Wiener filter is formulated for general case of complex valued time series with filter specified in terms of its impulse response because baseband signal appears in complex form under many practical situations.

c) Mean Square Error Criterion

The linear filter with the aim of estimating the desired signal d(n) from input x(n). Assume that d(n) and x(n) are samples of infinite length, random processes illustrates in Fig 3.1 . In 'optimum filter design', signal and noise are viewed as stochastic processes. The filter is based on minimization of the mean square value of the difference between the actual filter output and some desired output, as shown in fig.3.1.

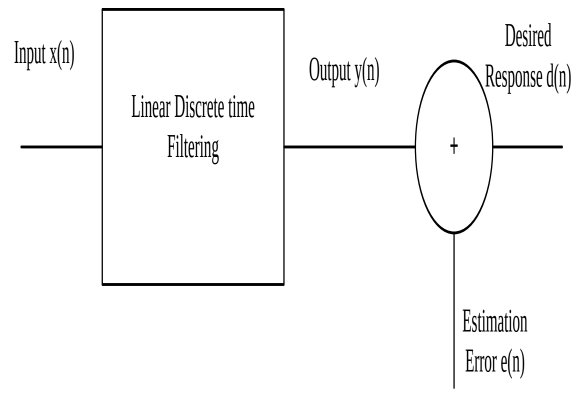

Figure 6.1 : Prototype Wiener Filtering Scheme

The requirement is to make the estimation error as small as possible in some Statistical sense by controlling impulse response coefficients w_0, w_1, ……., w_{N-1} . Two basic restrictions are: 1. The filter is linear, which makes mathematical analysis easy to handle.2. The filter is an FIR (symmetrical and odd ordered) filter.

The filter output is y(n) and the estimation error is given by e(n). The performance of the filter is determined by the size of the estimation error, that is, a smaller estimation error indicates a better filter performance. As the estimation error approaches zero, the filter output y(n) approaches the desired signal d(n). Clearly, the estimation error is required to be as small as possible. In simple words, in the design of the filter parameters, an appropriate function of this estimation error as performance or cost function is chosen and the set of filter parameters is selected, which optimizes the cost function. In Wiener filters, the cost function is chosen to be

$$x = E[e(n)^2] \qquad (6.1)$$

Where E[.] denotes the expectation or ensemble average since both d(n) and x(n) are random processes.

d) Wiener Filter: Transversal, Real valued case

Consider an adaptive transversal filter as shown in Fig 3.2. Assume that the filter input x(n) and the desired response d(n) are real valued stationary processes. The filter tap weights w_0, w_1,…………w_{N-1} are also assumed to be real valued , where N equals the number of delay units or tap weights.

The filter input x(n) and tap weight vectors, w, can be defined as column vectors,

$$\begin{aligned} x(n) &= [\, x(n)\;\; x(n-1)\ldots\ldots\, x(n-N+1)] \\ w &= [w_0\;\; w1 \ldots\ldots\ldots\, w_{N-1}]^{T} \end{aligned} \qquad (6.2)$$

The filter output is defined as

$$y(n) = \sum_{i=0}^{n} = w_i x(n-i) = w^T x(n) = x^T(n) w(n) \qquad (6.3)$$

Subsequently, the error signal can be written as

$e(n) = d(n)\text{-}y(n) = d(n) - w^T x(n) = d(n) - x^T(n)w$ (6.4)

Substituting (3.5) into (3.1) , the cost function is obtained as,

$E[(e(n)^2] = E[(d(n)\text{-}\ w^T x(n))\ (d(n)\text{-}x^T (n)w)]$ ((6.6)

Expanding the last expression of (6.6) we obtain,

$$E[\ d(n)^2] - E[\ d(n)x^T(n)w] - E[d(n)\ w^T x(n)] + E[w^T x(n)\ x^T (n)w] \quad (6.7)$$

Since w is not a random variable,

$$E[d(n)^2] - E[\ d(n)x^T (n)]w - w^T E[d(n)\ x(n)] + w^T E[\ x(n)\ x^T (n)]w \quad (6.8)$$

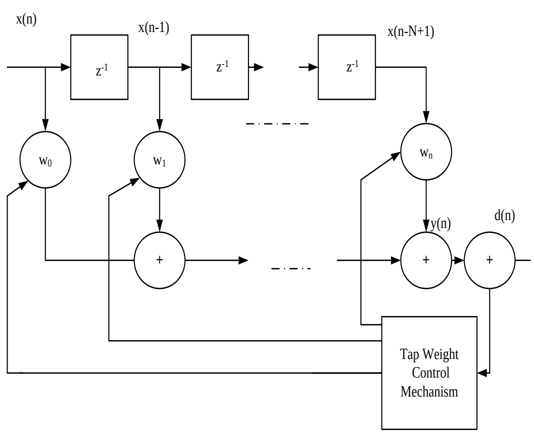

Figure 6.1 : Structure of an Adaptive Transversal Filter

Next, we can express E[d(n) x(n)] as an N x 1 cross correlation vector

$$p = E[d(n)x(n)] = [p_0, p_1, \ldots\ldots\ldots\ldots p_{N-1}]^T \quad (6.9)$$

And E[x(n) x^T (n)] as a N x N autocorrelation matrix R

$$R = E[x(n)x^T(n)] =$$

$$\begin{bmatrix} r_{00} & r_{01} & r_{02} & \cdots & r_{0,N-1} \\ & \vdots & & \ddots & \vdots \\ r_{N-1,0} & r_{N-1,1} & & \cdots & r_{N-1,N-1} \end{bmatrix} \quad (6.10)$$

From (6.9), p^T□$= E[d(n)\ x(n)]$ and hence $p^T w = w^T p$ This implies that $E[d(n)\ x^T$□$(n)]w = E[d(n)x(n)]w^T$□. Subsequently, we get

$$□□ = E\,[d(n)^2] - E[\ d(n)x^T(n)]w - w^T□E[d(n)\ x(n)] + w^T□E[\ x(n)\ x^T□(n)]w, \quad = E[d(n)^2] - 2p^T w + w^T□R\,w \quad (6.11)$$

This is a quadratic function of tap weight vector 'w' with a single global minimum. To obtain the set of filter tap weights that minimizes the cost function, □ solve the system of equations that results from setting the partial derivatives of □□□with respect to every tap weight of the filter i.e. the gradient vector to zero. That is

$$\frac{\partial \xi}{\partial w_i} = 0 \quad (6.12)$$

For i = 0, 1, ...N-1 where N = number of tap weights The gradient vector in (3.12) can also be expressed as□□ $= 0$ (6.13)

Where Ñ is the gradient operator defined as column vector

$$\nabla = \left[\frac{\partial \xi}{\partial w_1} \quad \frac{\partial \xi}{\partial w_2} \quad \frac{\partial \xi}{\partial w_{N-1}}\right] \quad (6.14)$$

and 0 on the right hand side of (3.13) denotes the column vector consisting of N zero. It has been further proved that the partial derivatives of x with respect to the filter tap weights can be solved such that

$$□□ = 2Rw2p \quad (6.15)$$

By letting □□ = 0, the following equation is obtained, in which the optimum set of Wiener filter tap weights can be obtained, *Rw = p* This implies that

$$w = R^{-1}p = w_0 \quad (6.16)$$

Where w_0 indicates the optimum tap weight vector. This equation is known as the Wiener Hopf equation and can be solved to obtain the tap weight vector, which corresponds to the minimum point of the cost function.

e) Iterative Search Algorithm

It has been shown in the previous section that the Wiener Hopf equation can be solved to obtain the optimum filter tap weights by minimizing a cost function, if the required statistics of the underlying signals 'R' and 'p' are available. Although this method is straightforward, it presents serious computational difficulties, especially when the filter contains a large number of tap weights and the input data rate is high. An alternative is to use an iterative search algorithm that starts at some arbitrary initial point in the tap weight vector space and moves progressively towards the optimum filter tap weight vector in steps. Each step is chosen with the aim of reducing the cost function. The principle of finding the optimum filter tap weight vector by progressive minimization of the underlying cost function by means of an iterative algorithm is central to the development of adaptive algorithms (e.g. LMS). In simplified terms, adaptive algorithms are actually iterative search algorithms derived for minimizing the cost function by replacing the true statistics with estimates obtained.

f) Method of Steepest Descent

Assume that the cost function to be minimized is convex (If the cost function corresponds to a convex quadratic surface, it has a unique minimum point. In other words, when the cost function is convex, the iterative search algorithm is guaranteed to converge to the optimum solution), we may start with an arbitrary point on the performance surface and take a small step in the direction in which the cost function decreases fastest. This corresponds to a step along the steepest descent slope of the performance at that point. Repeating this successively, convergence towards the bottom of the performance surface (corresponding to the set of parameters that minimize the cost function) is guaranteed.

The method of steepest descent is an alternate iterative search method to find w_0 (in contrast to solving the Wiener Hopf equation directly). The method of steepest descent algorithm belongs to a family of iterative methods of optimization. It is a general scheme that performs an iterative search for a minimum point of any convex function of a set of parameters. Here, this method is implemented in transversal filter with the convex function referring to the cost function and the set of parameters referring to the filter tap weights. It uses the following procedures to search the minimum point of the cost function of a set of filter tap weights.

a) Begin with an initial guess of the filter tap weights whose optimum values are to be found for minimizing the cost function. Unless some prior knowledge is available, the search can be initiated by setting all the filter tap weights to zero, i.e. w(0).
b) Use this initial guess to compute the gradient vector of the cost function with respect to the tap weights at the present point.
c) Update the tap weights by taking step in the opposite direction (sign change) of the gradient vector obtained in step 2. This corresponds to step in the direction of the steepest descent in the cost function at the present input. Furthermore, the size of the step is chosen proportional to the size of the gradient vector.
d) Go back to Step 2, and iterate the process until no further significant change is observed in the tap weights i.e. the search has converged to an optimal point.

According to the above procedures, if w(n) is the tap weight vector at the nth iteration, then the following recursive equation may be used to update w(n).

$$w(n+1) = w(n) - \mu\Delta_n\xi \qquad (6.17)$$

Where μ is the positive scalar called step size, and $\Delta_n\xi$ denotes the gradient vector evaluated at the point w = w(n).

g) Error Performance Surface

The estimation error e(n) can be given as:

$$E(n) = d(n) - \sum_{i=0}^{N-1} w_i x(n-i) \qquad (6.18)$$

The cost function can be written as

$$\xi = E[d(n)^2] - \sum_{i=0}^{N-1} w_i^* E[x(n-i)d^*(n)] \qquad (6.19)$$

The cost function or the mean squared error is precisely a second order function of the tap weights in the filter. Since 'w' can assume a continuum of values in the N dimensional w-plane, the dependence of the cost function depends on the tap weights w_0, w_1, ...w_{N-1} may be visualized as a bowl shaped (N+1)-dimensional surface with N degrees of freedom represented by the tap weights of the filter. The surface so described is called the error performance surface of the transversal filter. The surface is characterized by a unique minimum, where the cost function ξ attains its minimum value. At this point, the gradient vectorΔ ξ is identically zero. The height ξ corresponds to the physical description of filtering the signal x(n-i) with the fixed filter weight w, from which a prediction error signal e(n) with power of ξ is generated. Some filter setting $w_0=(w_{o0},w_{o1})$ will produce the minimum MSE (w_o is the optimum filter tap weight vector). This theory is the base of basic adaptive algorithms of adaptive signal processing. The gradient based adaptation starts with an old optimization technique known as the method of steepest descent. It is recursive in the sense that starting from some initial arbitrary value for tap weight vector, it improves with increasing number of iterations. The final value so computed for tap weight vector converges to Wiener solution.

The LMS algorithm has been extensively analyzed in literature and a large number of results on its steady state misadjustment and tracking performance have been obtained. The fixed step size least mean square (FSS LMS) algorithm is an important member of the family of stochastic gradient algorithms. The term 'stochastic gradient' is intended to distinguish it from the method of steepest descent that uses deterministic gradient in a recursive computation of the Wiener filter for stochastic inputs. This algorithm does not require measurements of the pertinent correlation functions, nor does it require matrix inversion. Subsequent works have discussed issue of optimization of step size or methods of varying step size to improve performance.

h) Performance of an Adaptive Algorithm

The factors that determine the performance of an algorithm are clearly stated below. Essentially, the most important factors as described here *1. Rate of Convergence:* This is defined as the number of iterations required for the algorithm to converge to its steady state mean square error. The steady state MSE is also known as the Mean asymptotic square error or MASE. *2. Misadjustment:* This quantity describes steady-state behavior of the algorithm. This is a quantitative measure of the amount by which the ensemble averaged final value of the mean-squared error exceeds the minimum mean-squared error produced by the optimal Wiener filter. The smaller the misadjustment, the better the asymptotic performance of the algorithm. *3. Numerical Robustness:* The implementation of adaptive filtering algorithms on a digital computer, which inevitably operates using finite word-lengths, results in quantization errors. These errors sometimes can cause numerical instability of the adaptation algorithm. An adaptive filtering algorithm is said to be numerically robust when its digital implementation using finite-word-length operations is stable. *4. Computational Requirements:* This is an important parameter from a practical point of view. The parameters of interest include the number of operations required for one complete iteration of the algorithm and the amount of memory needed to store the required data and also the program. These quantities influence the price of the computer needed to implement the adaptive filter.

5. Stability: An algorithm is said to be stable if the mean-squared error converges to a final (finite) value. Ideally, one would like to have a computationally simple and numerically robust adaptive filter with high rate of convergence and small misadjustment that can be implemented easily on a computer. In the applications of digital signal processing e.g. adaptive echo cancellation, the above factors play an important role.

There are different types of adaptive filtering algorithms, they are

- Least mean square (LMS) algorithm
- Normalized least mean square (NLMS) algorithm
- Variable step size LMS (VSLMS) algorithm
- Variable step size Normalized LMS (VSNLMS) algorithm
- Recursive least squares (RLS) algorithm.

i) The structure of Adaptive filter

The block diagram for the adaptive filter method utilized in this section. Here w represents the coefficients of the FIR filter tap weight vector, x(n) is the input vector samples, z^{-1} is a delay of one sample periods, y(n) is the adaptive filter output, d(n) is the desired echoed signal and e(n) is the estimation error at time n. The aim of an adaptive filter is to calculate the difference between the desired signal and the adaptive filter output, e(n). This error signal is fed back into the adaptive filter and its coefficients are changed algorithmically in order to minimize a function of this difference, known as the cost function. In the case of acoustic echo cancellation, the optimal output of the adaptive filter is equal in value to the unwanted echoed signal.

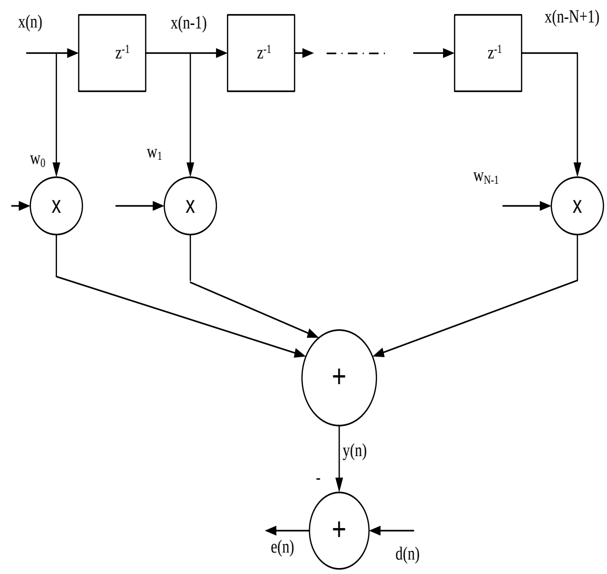

Figure 6.2 : Adaptive filter block diagram

When the adaptive filter output is equal to desired signal the error signal goes to zero. In this situation the echoed signal would be completely cancelled and the far user would not hear any of their original speech returned to them.

VII. Least Mean Square (LMS) Algorithm

The LMS algorithm is a type of adaptive filter known as stochastic gradient-based algorithms as it utilizes the gradient vector of the filter tap weights to converge on the optimal wiener solution. It is well known and widely used due to its computational simplicity. With each iteration of the LMS algorithm, the filter tap weights of the adaptive filter are updated according to the following formula

$$w(n+1) = w(n) + 2\mu e(n)x(n) \qquad (7.1)$$

Here **x**(n) is the input vector of time delayed input values

$$x(n) = [x(n)\ x(n-1)\ x(n-2) \ldots\ x(n-N+1)]^T \qquad (7.2)$$

The vector w(n) represents the coefficients of the adaptive FIR filter tap weight vector at time n.

$$w(n) = [w_0(n)\ w_1(n)\ w_2(n) \ldots\ w_{N-1}(n)]^T \qquad (7.3)$$

The parameter μ is known as the step size parameter and is a small positive constant. This step size parameter controls the influence of the updating factor. Selection of a suitable value for μ is imperative to the performance of the LMS algorithm, if the value is too small the time the adaptive filter takes to converge on the optimal solution will be too long; if μ is too large the adaptive filter becomes unstable and its output diverges.

a) Derivation of the LMS algorithm

The derivation of the LMS algorithm builds upon the theory of the wiener solution for the optimal filter tap weights, w_o, as outlined in section 3.2.2. It also depends on the steepest descent algorithm as stated in equation 3.23, this is a formula which updates the filter coefficients using the current tap weight vector and the current gradient of the cost function with respect to the filter tap weight coefficient vector, ξ(n)

$$w(n+1)=w(n)-\mu\, \xi(n)$$

Where $$\xi\,(n)=E[e(n)^2] \quad (7.4)$$

As the negative gradient vector points in the direction of steepest descent for the N-dimensional quadratic cost function, each recursion shifts the value of the filter coefficients closer toward their optimum value, which corresponds to the minimum achievable value of the cost function, ξ(n).

The LMS algorithm is a random process implementation of the steepest descent algorithm, from equation 3.23. Here the expectation for the error signal is not known so the instantaneous value is used as an estimate. The steepest descent algorithm then becomes equation 3.24.

$$w(n+1)=w(n)-\mu\, \xi(n)$$

Where $$\xi\,(n) = e(n)^2 \quad (7.5)$$

The gradient of the cost function, ▼ξ(n), can alternatively be expressed in the following form.

$\xi(n)=(e^2(n))$

$$=\frac{\partial e^2\,(n)}{\partial w}$$

$$= 2e(n)\frac{\partial e\,(n)}{\partial w} = -\,2e(n)\frac{\partial\,(d(n)-y(n))}{\partial w} = (7.6)$$

$$-\,2e(n)\frac{\partial w^T\,(n)x(n)}{\partial w} = -\,2e(n)x(n)$$

Substituting this into the steepest descent algorithm of equation 3.8, we arrive at the recursion for the LMS adaptive algorithm.

$$w(n+1) = w(n) + 2\mu e(n)x(n) \quad (7.7)$$

b) Implementation of the LMS algorithm

Each iteration of the LMS algorithm requires 3 distinct steps in this order:

i. *The output of the FIR filter, y(n) is calculated using equation 3.27*

$$y(n)=\sum_{i=0}^{N-1} w(n)x(n-i)=w^T(n)x(n) \quad (7.8)$$

ii. *The value of the error estimation is calculated using equation 3.28.*

$$e(n)=d(n)-y(n) \quad (7.9)$$

iii. *The tap weights of the FIR vector are updated in preparation for the next iteration, by equation 7.9*

$$w(n+1) = w(n) + 2\mu e(n)x(n) \quad (7.10)$$

The main reason for the LMS algorithms popularity in adaptive filtering is its computational simplicity, making it easier to implement than all other commonly used adaptive algorithms. For each iteration the LMS algorithm requires 2N additions and 2N+1 multiplications (N for calculating the output, y(n), one for 2μe(n) and an additional N for the scalar by vector multiplication). One of the primary disadvantages of the LMS algorithm is having a fixed step size parameter for every iteration. This requires an understanding of the statistics of the input signal prior to commencing the adaptive filtering operation. In practice this is rarely achievable. Even if we assume the only signal to be input to the adaptive echo cancellation system is speech, there are still many factors such as signal input power and amplitude which will affect its performance. The normalized least mean square algorithm (NLMS) is an extension of the LMS algorithm which bypasses this issue by selecting a different step size value, μ(n), for each iteration of the algorithm. This step size is proportional to the inverse of the total expected energy of the instantaneous values of the coefficients of the input vector **x**(n). This sum of the expected energies of the input samples is also equivalent to the dot product of the input vector with itself, and the trace of input vectors auto-correlation matrix, **R**.

$$tr(\boldsymbol{R}) = \sum_{i=0}^{N-1} E[x^2(n-]i)$$
$$= E[\sum_{i=0}^{N-1} x^2(n-i)] \quad (7.11)$$

The recursion formula for the NLMS algorithm is stated in equation 3.31.

$$w(n+1)=w(n)+\frac{1}{x^T(n)x(n)}e(n)x(n) \quad (7.12)$$

c) Derivation of the NLMS algorithm

To derive the NLMS algorithm we consider the standard LMS recursion, for which we select a variable step size parameter, μ(n). This parameter is selected so that the error value, $e^+(n)$, will be minimized using the updated filter tap weights, w(n+1), and the current input vector, **x**(n). $w(n+1) = w(n) + 2\mu(n)e(n)x(n)$, $e^+(n) = d(n) - w^T(n+1)x(n), =(1-2\mu(n)x^T(n)x(n))e(n)$ (7.13)

Next we minimize $(e^+(n))^2$, with respect to μ(n). Using this we can then find a value for μ(n) which forces $e^+(n)$ to

$$\text{zero.}\mu(n)=\frac{1}{2x^T(n)x(n)} \quad (7.14)$$

This $\mu(n)$ is then substituted into the standard LMS recursion replacing μ, resulting in the following NLMS equation.

$$w(n+1) = w(n) + 2\mu(n)e(n)x(n)$$

$$w(n+1)=w(n)+ \frac{1}{x^T(n)x(n)} e(n)x(n) \quad (7.15)$$

d) Implementation of the NLMS algorithm

The NLMS algorithm has been implemented in Matlab and in a real time application using the Texas Instruments TMS320C6711 Development Kit. As the step size parameter is chosen based on the current input values, the NLMS algorithm shows far greater stability with unknown signals. This combined with good convergence speed and relative computational simplicity makes the NLMS algorithm ideal for the real time adaptive echo cancellation system.

As the NLMS is an extension of the standard LMS algorithm, the NLMS algorithms practical implementation is very similar to that of the LMS algorithm. Each iteration of the NLMS algorithm requires these steps in the following order.

The output of the adaptive filter is calculated.

$$y(n)=\sum_{i=0}^{N-1} w(n)x(n-i) = w^T(n)x(n) \quad (3.35)$$

An error signal is calculated as the difference between the desired signal and the filter output

$$e(n)=d(n)-y(n) \quad (7.16)$$

The step size value for the input vector is calculated.

$$\mu(n) = \frac{1}{2x^T(n)x(n)} \quad (7.17)$$

The filter tap weights are updated in preparation for the next iteration.

$$w(n+1) = w(n) +\mu(n)e(n)x(n) \quad (7.18)$$

Each iteration of the NLMS algorithm requires 3N+1 multiplications, this is only N more than the standard LMS algorithm, this is an acceptable increase considering the gains in stability and echo attenuation achieved.

VIII. Comparison of Adaptive Filtering Algorithms

Algorithm: LMS Algorithm

Average attenuation: -18.2 dB

Multiplication operations: 2N+1

Comments: Is the simplest to implement and is stable when the step size parameter is selected appropriately. This requires prior knowledge of the input signal which is not feasible for the echo cancellation system.

Algorithm: NLMS Algorithm

Average attenuation: -27.9dB

Multiplication operations: 3N+1

Comments: Simple to implement and computationally efficient. Shows very good attenuation and variable step size allows stable performance with non-stationary signals. This was the obvious choice for real time implementation.

Algorithm: VSSLMS Algorithm

Average attenuation: -9.8 dB

Multiplication operations: 4N+1

Comments: Displays very poor performance, possibly due to non-stationary nature of speech signals. Only half the attenuation of the standard LMS algorithm. Not considered for real time implementation.

Algorithm: VSSNLMS Algorithm

Average attenuation: -9.9 dB

Multiplication operations: 5N+1

Comments: Increase in multiplications gives negligible improvement in performance over VSSLMS algorithm.

The real time acoustic echo cancellation system was successfully developed with the NLMS algorithm. The system is capable of cancelling echo with time delays of up to 75 ms, corresponding to reverberation off an object a maximum of 12 meters away. This proves quite satisfactory in emulating a medium to large size room.

The utility of SBC is perhaps best illustrated with a specific example. When used for audio compression, SBC exploits what might be considered a deficiency of the human auditory system. Human ears are normally sensitive to a wide range of frequencies, but when a sufficiently loud signal is present at one frequency, the ear will not hear weaker signals at nearby frequencies. We say that the louder signal masks the softer ones. The louder signal is called the masker, and the point at which masking occurs is known, appropriately enough, as the masking threshold. The basic idea of SBC is to enable a data reduction by discarding information about frequencies which are masked. The result differs from the original signal, but if the discarded information is chosen carefully, the difference will not be noticeable, or more importantly, objectionable.

IX. Encoding Audio Signals

The simplest way to digitally encode audio signals is pulse-code modulation (PCM), which is used on audio CDs, DAT recordings, and so on. Digitization transforms continuous signals into discrete ones by sampling a signal's amplitude at uniform intervals and rounding to the nearest value representable with the available number of bits. This process is fundamentally inexact, and involves two errors: discretization error, from sampling at intervals, and quantization error, from rounding.

The more bits used represent each sample, the finer the granularity in the digital representation, and thus the smaller the error. Such quantization errors may be thought of as a type of noise, because they are effectively the difference between the original source and its binary representation. With PCM, the only way to mitigate the audible effects of these errors is to use enough bits to ensure that the noise is low enough to be masked either by the signal itself or by other sources of noise. A high quality signal is possible, but at the cost of a high bitrate (e.g., over 700 kbit/s for one channel of CD audio). In effect, many bits are wasted in encoding masked portions of the signal because PCM makes no assumptions about how the human ear hears. More clever ways of digitizing an audio signal can reduce that waste by exploiting known characteristics of the auditory system. A classic method is nonlinear PCM, such as mu-law encoding (named after a perceptual curve in auditory perception research). Small signals are digitized with finer granularity than are large ones; the effect is to add noise that is proportional to the signal strength. Sun's Au file format for sound is a popular example of mu-law encoding. Using 8-bit mu-law encoding would cut the per-channel bit rate of CD audio down to about 350 kbit/s, or about half the standard rate. Because this simple method only minimally exploits masking effects, it produces results that are often audibly poorer than the original. Sub-band coding is used for example in G.722 codec. It uses sub-band adaptive differential pulse code modulation (SB-ADPCM) within a bit rate of 64 kbit/s. In the SB-ADPCM technique used, the frequency band is split into two sub-bands (higher and lower) and the signals in each sub-band are encoded using ADPCM.

As explained in Section 2, in the proposed algorithm the TAF is obtained with a delay relative to the input signal. The amount of delay depends on the method of filter reconstruction. For sequential synthesis the delay is (*La* □□*Ls*) / *2* samples while for batch synthesis it is *La* / *2* samples. All of the delayless SAF methods reviewed in Section 1 have to deal with a plant reconstruction delay. The delay leads to a "synchronization problem" between the input signals and the plant, causing problems in tracking a dynamic plant. The extent of the problem depends on the plant time-dynamics and the TAF reconstruction delay. To demonstrate the effects of the delay, we simulated the system with the same system set up and input signals as described in the previous section with the following changes. The echo plant was switched to a new plant after *30* seconds through the experiment. With the employed analysis/synthesis filters used in the experiments, tracking problems were barely observable due to the low reconstruction delay of the system. Thus, the analysis and synthesis window lengths were increased to *La* □*1024* and *Ls* □*256* samples to better observe the effects of the delay. To simplify the analysis, batch synthesis was used for TAF WOLA reconstruction. This leads to a filter reconstruction delay of *La* / *2* □*512* samples. Delaying the input signals by the same amount so that they are synchronized with the plant could compensate for the filter reconstruction delay. Of course this is counter productive as it creates delays in an otherwise delayless system. The ERLE drops at *30* seconds, and stays low for around *64* msecs (corresponding to *512* samples of delay) before it starts to rise again. This low-time of ERLE causes a drop in echo cancellation performance and creates artifacts in the output. Repeating the experiment with delay compensation, the ERLE drops later and start to rise right away as shown in the figure. The echo plant swap is unlikely to happen in practice; rather gradual plant variations might occur.

X. Simulation Results

The input file 'file1.wav' is read by the command waveread and impulse response of secondary path is plotted as shown in fig 5.1. This means the output of secondary path y`(n) initially. Then estimating the secondary propagation path S^(z) and identifying this path using NLMS adaptive filter. This is shown in fig 5.2 and also shows the required signal d(n),output signal y`(n) and error signal e(n).If the number of iterations are increased, the error signal e(n) is reduced. Here d(n) and y`(n) are having same signal value from $0.5*10^4$ to $3*10^4$.

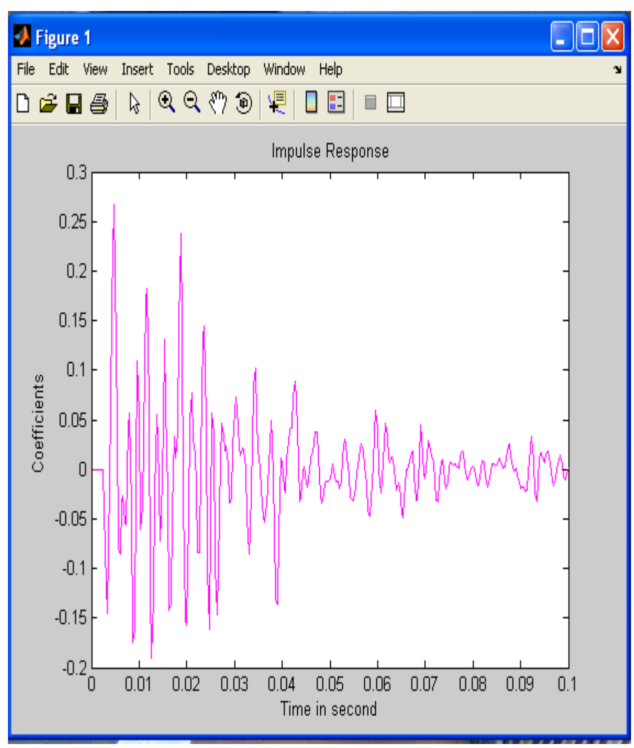

Figure 8.1 : Secondary path filter response

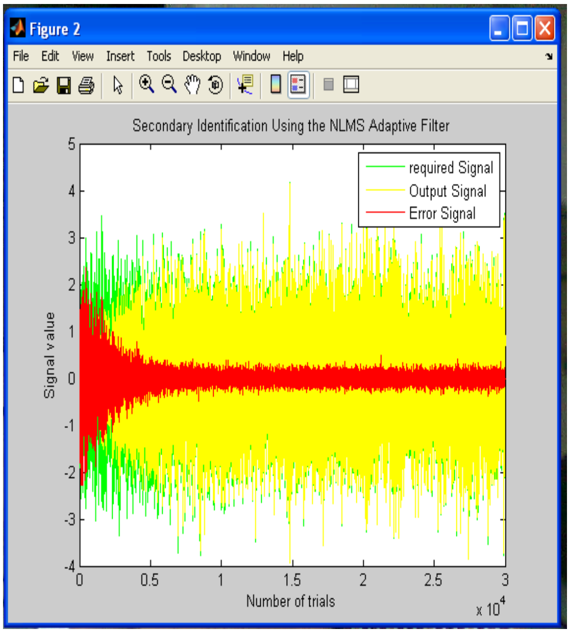

Figure 8.2 : Secondary path identification

a) Using NLMS algorithm

Fig 8.3 shows accuracy of the estimated secondary propagation path S^(z). Also the summer output e(n) in time-domain. Here the d(n) and y`(n) follows the same path and error signal e(n) is zero.

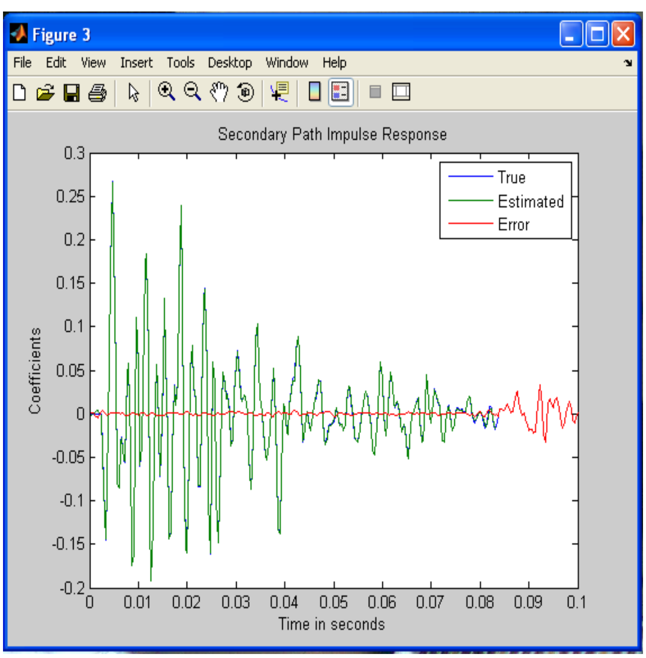

Figure 8.3 : Accuracy of Secondary path

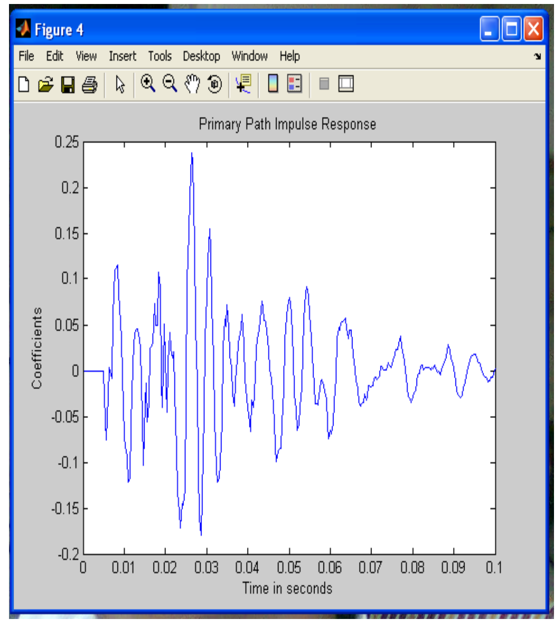

Figure 8.4 : Primary path filter response

Fig 8.4 shows the primary propagation path filter response P(z).The output of P(z) is d(n),this is the actual noise to be canceled by generating anti-noise through the S^(z).Then the noise in the system is to be canceled and fig 5.5 shows the power spectral density of the canceled noise d(n). Power spectral density means the distribution of d(n) over frequency-axis. The voice frequency range is considered as from (0.3-3.5) KHz.

Fig 8.6 shows residual error signal spectrum of e(n) or power spectral density of d(n) and y`(n).From (0-1300)Hz, there is small difference between d(n) and y`(n).After that both are equal. Fig 5.7 shows the power spectral density of d(n)-y`(n).At 200Hz, power/frequency is -50 db/Hz (0.0031) approximately zero.

The total complexity is plotted in Fig.5.8, number of real multiplications versus the number of subbands M. The plot is for the PFFT-2 method with L_{SAF} = 4N/M, as it results in better performance than that of PFFT-1. For comparison purposes, included the computational complexity of the MT and DFT-MDF algorithms. As shown, the computational complexities of all methods reduce almost exponentially with M. The proposed technique compared to the other methods for small values of M has higher computational complexity.

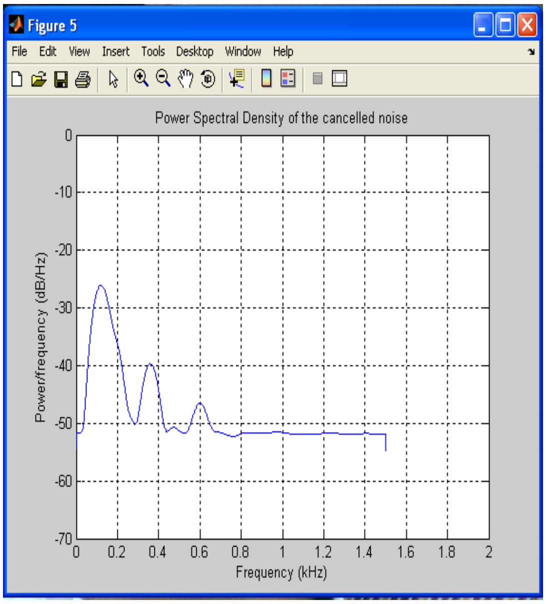

Figure 8.5 : Power spectral density of canceled

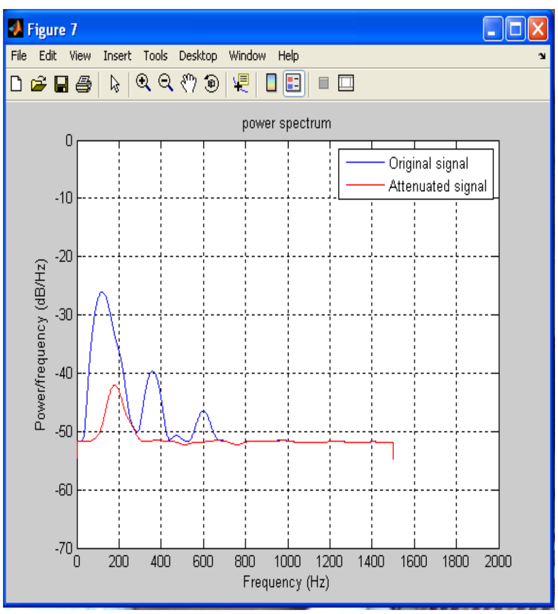

Figure 8.6 : Residual error signal spectrum Noise

The new technique works very well with a larger number subbands, improving the system performance and attaining lower complexity, whereas the MT method fails to converge and the performance of the DFT-MDF method deteriorates.

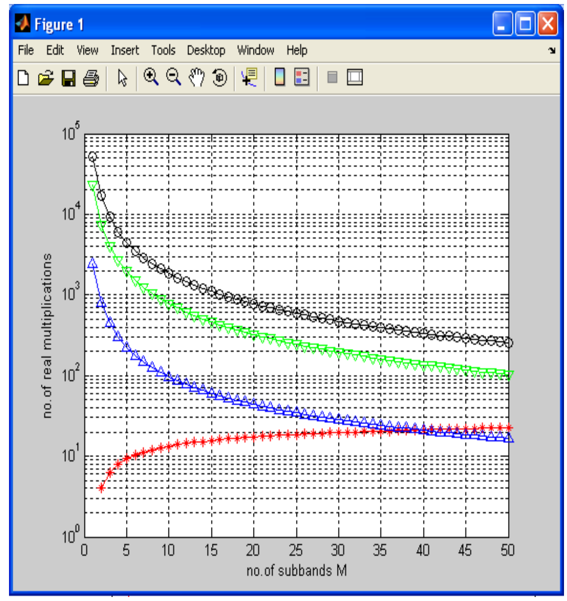

Figure 8.7 : Comparison of computational complexity per input sample versus number of subbands (M)

XI. Conclusion

Acoustic paths such as those encountered in ANC application usually have long impulse responses, which require longer adaptive filters for noise cancellation. Subband adaptive filters working with a large number of subbands have been shown to be a good solution to this problem. The focus of this project was to design such a high-performance SAF algorithm. The performance limiting factors of existing SAF structures were found to be due to the inherent delay and side-lobes of the prototype filter in the analysis filter banks. Hence, the analysis filter banks were modified to reduce the inherent delay. A new weight stacking transform was designed to alleviate the interference introduced by the side-lobes. The modifications resulted in a new subband method that, unlike existing methods, improves the performance and reduces the computational complexity for a large number of subbands.

Experimental results showed that the proposed method outperformed the two commonly used SAF and BAF methods. The proposed technique compared to the other methods for small values of M has higher computational complexity. The new technique works very well with a larger number of subbands, improving the system performance and attaining lower complexity, whereas the MT method fails to converge and the performance of the DFT-MDF method deteriorates.

XII. Future Scope

Adaptive digital signal processing is a rapidly growing branch of DSP and has great significance in the design of adaptive systems. The various signal processing applications demand for reduction in trade off between misadjustment and convergence rate, taking realization of algorithm into account. The modifications resulted in a new subband method that, unlike existing methods, improves the performance and reduces the computational complexity for a large number of subbands. There is a scope of improvement in replacing the existing time domain adaptive filters with frequency domain adaptive filters. There's a lot, which can be done in future for improvement on the methods for noise cancellation. The field of digital signal processing and in particular adaptive filtering is vast and further research and development in this area can result in some improvement on the methods studied in this paper.

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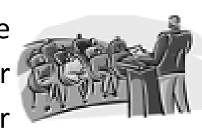

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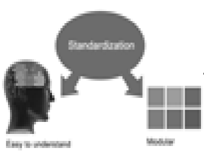

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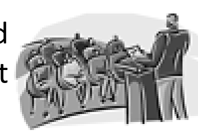

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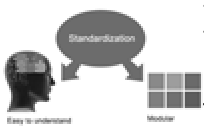

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1. General,

2. Ethical Guidelines,

3. Submission of Manuscripts,

4. Manuscript's Category,

5. Structure and Format of Manuscript,

6. After Acceptance.

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Approach:

- Use past tense except for when referring to recognized facts. After all, the manuscript will be submitted after the entire job is done.
- Sort out your thoughts; manufacture one key point with every section. If you make the four points listed above, you will need a least of four paragraphs.

- Present surroundings information only as desirable in order hold up a situation. The reviewer does not desire to read the whole thing you know about a topic.
- Shape the theory/purpose specifically - do not take a broad view.
- As always, give awareness to spelling, simplicity and correctness of sentences and phrases.

Procedures (Methods and Materials):

This part is supposed to be the easiest to carve if you have good skills. A sound written Procedures segment allows a capable scientist to replacement your results. Present precise information about your supplies. The suppliers and clarity of reagents can be helpful bits of information. Present methods in sequential order but linked methodologies can be grouped as a segment. Be concise when relating the protocols. Attempt for the least amount of information that would permit another capable scientist to spare your outcome but be cautious that vital information is integrated. The use of subheadings is suggested and ought to be synchronized with the results section. When a technique is used that has been well described in another object, mention the specific item describing a way but draw the basic principle while stating the situation. The purpose is to text all particular resources and broad procedures, so that another person may use some or all of the methods in one more study or referee the scientific value of your work. It is not to be a step by step report of the whole thing you did, nor is a methods section a set of orders.

Materials:

- Explain materials individually only if the study is so complex that it saves liberty this way.
- Embrace particular materials, and any tools or provisions that are not frequently found in laboratories.
- Do not take in frequently found.
- If use of a definite type of tools.
- Materials may be reported in a part section or else they may be recognized along with your measures.

Methods:

- Report the method (not particulars of each process that engaged the same methodology)
- Describe the method entirely
- To be succinct, present methods under headings dedicated to specific dealings or groups of measures
- Simplify - details how procedures were completed not how they were exclusively performed on a particular day.
- If well known procedures were used, account the procedure by name, possibly with reference, and that's all.

Approach:

- It is embarrassed or not possible to use vigorous voice when documenting methods with no using first person, which would focus the reviewer's interest on the researcher rather than the job. As a result when script up the methods most authors use third person passive voice.
- Use standard style in this and in every other part of the paper - avoid familiar lists, and use full sentences.

What to keep away from

- Resources and methods are not a set of information.
- Skip all descriptive information and surroundings - save it for the argument.
- Leave out information that is immaterial to a third party.

Results:

The principle of a results segment is to present and demonstrate your conclusion. Create this part a entirely objective details of the outcome, and save all understanding for the discussion.

The page length of this segment is set by the sum and types of data to be reported. Carry on to be to the point, by means of statistics and tables, if suitable, to present consequences most efficiently.You must obviously differentiate material that would usually be incorporated in a study editorial from any unprocessed data or additional appendix matter that would not be available. In fact, such matter should not be submitted at all except requested by the instructor.

Content

- Sum up your conclusion in text and demonstrate them, if suitable, with figures and tables.
- In manuscript, explain each of your consequences, point the reader to remarks that are most appropriate.
- Present a background, such as by describing the question that was addressed by creation an exacting study.
- Explain results of control experiments and comprise remarks that are not accessible in a prescribed figure or table, if appropriate.
- Examine your data, then prepare the analyzed (transformed) data in the form of a figure (graph), table, or in manuscript form.

What to stay away from

- Do not discuss or infer your outcome, report surroundings information, or try to explain anything.
- Not at all, take in raw data or intermediate calculations in a research manuscript.
- Do not present the similar data more than once.
- Manuscript should complement any figures or tables, not duplicate the identical information.
- Never confuse figures with tables - there is a difference.

Approach

- As forever, use past tense when you submit to your results, and put the whole thing in a reasonable order.
- Put figures and tables, appropriately numbered, in order at the end of the report
- If you desire, you may place your figures and tables properly within the text of your results part.

Figures and tables

- If you put figures and tables at the end of the details, make certain that they are visibly distinguished from any attach appendix materials, such as raw facts
- Despite of position, each figure must be numbered one after the other and complete with subtitle
- In spite of position, each table must be titled, numbered one after the other and complete with heading
- All figure and table must be adequately complete that it could situate on its own, divide from text

Discussion:

The Discussion is expected the trickiest segment to write and describe. A lot of papers submitted for journal are discarded based on problems with the Discussion. There is no head of state for how long a argument should be. Position your understanding of the outcome visibly to lead the reviewer through your conclusions, and then finish the paper with a summing up of the implication of the study. The purpose here is to offer an understanding of your results and hold up for all of your conclusions, using facts from your research and generally accepted information, if suitable. The implication of result should be visibly described. Infer your data in the conversation in suitable depth. This means that when you clarify an observable fact you must explain mechanisms that may account for the observation. If your results vary from your prospect, make clear why that may have happened. If your results agree, then explain the theory that the proof supported. It is never suitable to just state that the data approved with prospect, and let it drop at that.

- Make a decision if each premise is supported, discarded, or if you cannot make a conclusion with assurance. Do not just dismiss a study or part of a study as "uncertain."
- Research papers are not acknowledged if the work is imperfect. Draw what conclusions you can based upon the results that you have, and take care of the study as a finished work
- You may propose future guidelines, such as how the experiment might be personalized to accomplish a new idea.
- Give details all of your remarks as much as possible, focus on mechanisms.
- Make a decision if the tentative design sufficiently addressed the theory, and whether or not it was correctly restricted.
- Try to present substitute explanations if sensible alternatives be present.
- One research will not counter an overall question, so maintain the large picture in mind, where do you go next? The best studies unlock new avenues of study. What questions remain?
- Recommendations for detailed papers will offer supplementary suggestions.

Approach:

- When you refer to information, differentiate data generated by your own studies from available information
- Submit to work done by specific persons (including you) in past tense.
- Submit to generally acknowledged facts and main beliefs in present tense.

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Introduction	Containing all background details with clear goal and appropriate details, flow specification, no grammar and spelling mistake, well organized sentence and paragraph, reference cited	Unclear and confusing data, appropriate format, grammar and spelling errors with unorganized matter	Out of place depth and content, hazy format
Methods and Procedures	Clear and to the point with well arranged paragraph, precision and accuracy of facts and figures, well organized subheads	Difficult to comprehend with embarrassed text, too much explanation but completed	Incorrect and unorganized structure with hazy meaning
Result	Well organized, Clear and specific, Correct units with precision, correct data, well structuring of paragraph, no grammar and spelling mistake	Complete and embarrassed text, difficult to comprehend	Irregular format with wrong facts and figures
Discussion	Well organized, meaningful specification, sound conclusion, logical and concise explanation, highly structured paragraph reference cited	Wordy, unclear conclusion, spurious	Conclusion is not cited, unorganized, difficult to comprehend
References	Complete and correct format, well organized	Beside the point, Incomplete	Wrong format and structuring

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