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Self-Interference Cancellation in Full Duplex Communication using Steepest Descent Method

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Abstract: The present-day communication system uses Frequency Division Duplex (FDD) to emulate the benefits of Full Duplex Communication. But it requires more bandwidth as the cost of the spectrum is very high it becomes a major limitation. To overcome this problem implementation of Full Duplex Communication is the best solution. Implementation of full duplex communication is difficult because of a significant problem called self-interference. while transmitting and receiving signals on the same frequency band, receiving signal is interfered with the transmitted signal this phenomenon is called self-interference. The objective of this project is to minimize that self-interference signal from the received signal by using signal processing technique, LMS echo cancellation. Least Mean Square (LMS) echo canceller whose coefficients are updated iteratively is used to cancel the self-interference. An algorithm based on steepest descent method is used to obtain coefficients that change iteratively with varying step size to solve Weiner-Hopfs equation.

Keywords: Self-Interference Cancellation, Steepest Descent Method, Frequency Division Duplex, Full Duplex Communication, Least Means Square algorithm

I. INTRODUCTION

Traditional wireless communication systems contain both radio frequency (RF) transmitter (TX) and receiver (RX). Combination of both transmitter and receiver is known as a transceiver. There exist several traditional methods to successfully transmit and receive data at maximum efficiency. These methods exploit three different areas: time, frequency, and space. TDD architecture will sequentially transmit and receive a signal i.e., the TX and RX will never simultaneously be powered on. Although this method uses a single frequency, the major drawback lies in the fact that twice the time is needed to successfully transmit and receive a data packet of equal size.

Frequency domain method employs Frequency Division Duplexing (FDD). In FDD architecture, the TX and RX will be operating simultaneously but at different frequencies. Both paths will have Band Pass Filters (BPF) tuned to their specific frequency to block out any residual interference that may fall in each respective band. The major drawback of this architecture is that it requires twice the maximum efficient bandwidth. As the number of users and cost of bandwidth are increasing steadily, FDD technology may not be efficient.

A system employing spatial methods will use directional antennas, antenna diversity or cross- Duplex polarization techniques. Multiple Input Multiple Output (MIMO) systems use multiple antennas to increase efficiency by transmitting more packets of data to each antenna. Theoretically, this accommodates for the loss in spectral efficiency of time by effectively doubling throughput; however, maintenance of multiple antennas is required which is a difficult task.

Regarding the requirement of more spectrally efficient systems, it is clear that full duplex systems are a solution. Full duplex systems can exist using two methods – active or passive echo cancellation. Passive cancellation techniques typically rely on the attenuation of the transmitted signal before it reaches the receiver. This purpose can be served by Directional antennas or circulators. In order to achieve Full Duplex Communication and to maximize the echo signal cancellation both active and passive cancellation techniques should be used. Echo cancellation in a transceiver can be performed in three areas. These areas are at RF, analog baseband, and in the digital domain after quantization. It is preferred to perform the cancellation before the ADC otherwise the self-interference signal will saturate the dynamic range of the ADC and prevent the actual signal from being scaled up to the full-scale range. However, there exists a trade-off in this scenario. The cancellation performed in the digital aspect will be less effective if the received self- interference signal is weaker at the input of the ADC. The total cancellation (analog and digital combined) will ultimately end up being equal despite the amount of cancellation that happens in either the digital or baseband analog aspect.

This paper is organized as follows; Section II emphasizes on literature work. Section III describes about in-band full duplex communication, effect of self-interference and cancellation techniques. Section IV provides complete design approach of system. The results are discussed in Section V. This work is concluded in Section VI.

II. LITERATURE REVIEW

Many research papers have been published regarding implementation of full duplex systems theoretically and practically. Researchers at Stanford have shown that Wi-Fi 802.11ac systems at 2.45 GHz can be operated successfully in full duplex modes. Both analog RF and digital cancellation methods were employed to maximize the cancellation of the self-interference signal using the Wireless Open Access Research Platform (WARP) platform. A circulator was used to preserve real estimate and provide additional cancellation. The analog RF cancellation was implemented using fixed delay lines with variable attenuators per line that were tapped from the transmit path. This essentially emulates a Finite Impulse Response (FIR) filter in the analog domain. This allows the cancellation path to account for multi-path effects and the noise and harmonics created by the transmitter itself. Using delayed versions of the signal from before and after the main self-interference delay, the received signal can be recreated in a fashion similar to sinc interpolation. In the same way, by using previous and future samples known to the system the digital algorithm tries to predict the current value.

The major issue is the fact that the cancellation algorithms are not on-line. This means that a certain amount of time (900-1000 µs) will be elapsed before the circuits and digital algorithms adapt to the new environment. This might not be a problem in a slow fading channel; however, in a fast-fading channel with a short coherence time, the cancellation coefficients may not update fast enough. Moreover, to initialize the coefficients, Wi-Fi preamble information about the channel is required. Finally, the method they use to set the analog cancellation coefficients involves physically measuring the transfer function of each delay line. Mass producing devices with this implementation will be problematic because of the fact that this measurement only needs to be done once.

Stanford has also investigated a method of cancellation known as antenna cancellation. Two TX antennas are placed on either side of the RX antenna such that there exists a distance of $x+(\lambda/2)$ between one TX and RX pair and x between the other pair resulting in an offset spacing of one-half wavelength between the two TX antennas which causes destructive interference directly at the position of the RX antenna. The limitations of this design include bandwidth, power, and the channel centre frequency. This implementation will not be appreciated at lower frequencies due to the involvement of larger wavelengths. In addition, the powers of the two TX antennas are required to be equal to achieve maximum nullifying at the point of interest. The two TX antennas also cause nulls at other points in the environment that leads to dead spots. Moreover, the antenna placement needs to be manually implemented which might be difficult for an adaptive system in the field. Work in [4] explains that Orthogonal Frequency Division Multiplex (OFDM) systems will suffer from this architecture since each sub-carrier will experience a significant change in self-interference. Since the architecture needs three antennas, employing MIMO techniques may be used since a 3x3 system can theoretically triple the throughput. The advantage of this architecture, however, is that no knowledge of the channel is required, and no calibration or coefficients required to be updated. Noise and harmonics from the transmitter are also suppressed.

Researchers of Rice University successfully implemented a full duplex system using two separate antennas placed at a predefined distance in addition to analog RF and digital cancellation. They also used the WARP platform to perform cancellation measurements. Although they were able to show good cancellation results, the bandwidth used was only 625 kHz. Moreover, the antenna spacing used were of 20 cm and 40 cm; this will not be feasible for any sort of mobile device. The methods by which they implemented their algorithms and analog cancellation circuitry are not known. Further research from Rice University investigated the effects of directional antennas in a base station. Specifically, the half-power beam-width angle of the antennas was varied to observe how cancellation degrades. The results of this study proved that directionality certainly improves the cancellation of the self-interference signal. However, the issue with directionality is that the range of transmission is now limited to the beam-width angle and the base station design is constrained by the placement of the antennas to achieve an angle that attains maximum cancellation.

Another architecture uses a balun to invert the TX signal before injecting it into the RX path before the mixer. Self-Interference cancellation can work up to 100 MHz due to the relatively wide bandwidth of a balun. Gradient Descent Algorithm is used to correct the delay between the TX and RX path which controls a delay element between the balun and the injection point. However, this method fails to address the multi-path effects. The digital cancellation method used in [4] relies on the OFDM symbols. Inverse Fast Fourier Transform (IFFT) needs to be performed on the training symbols to recover the impulse response of the channel as they are initially defined in the frequency domain. To create a predicted self-interference signal, the transmitted data is convolved with this impulse response. Since the channel is a causal LTI system, the channel can be emulated using a FIR filter. The FIR filter coefficients can only be updated per every training pilot sent in an OFDM implying that this system is not truly on-line. In addition, [4] states that there must be an absence of interference during the training sequence. This creates an issue in an environment rich with EM signals at that particular band, however, all active echo cancellation schemes will suffer from this effect.

As stated earlier, coherence time will affect that rate that is needed to update the coefficients of the FIR filter [14]. A static environment will have a coherence time in the order of seconds and thus an estimation update rate of a few hundred milliseconds will be sufficient as stated in [5].

Digital cancellation and its implementation are investigated in [16]. The architecture is similar to that of [14], however, only a causal FIR filter is used. Based on the rms delay spread of the channel, algorithm estimates the number of taps needed. The coefficients of the FIR filter are updated based on pilot tones and thus there exists overhead in this scheme. The implementation is not true real-time. An experiment was performed on the impact of phase noise on active cancellation [15]. To mimic active echo cancellation, a test setup was created.

A known signal was first upconverted to RF and then fed to a splitter. The two splitter outputs were then fed to a vector signal analyzer (VSA) which down converted the signals using knowledge of the carrier frequency. The baseband signals were digitized with a sample rate of about 48 Msps. To emulate the active cancellation, the two signals were subtracted from each other after scaling and delaying.

It was found that, theoretically, if the signals are only corrupted by delay and noise, cancellation is independent of delay and only dependent on thermal noise. Practically, the upper bound of cancellation was found to be 55 dB – the dynamic range of the VSA is only 55-60 dB. This value of cancellation was constant across delay. However, when the WARP board was used to test the same experiment, the cancellation fell as a function of delay and floored at around 35 dB. Phase noise was considered as a suspect for this observation since the phase noise in the VSA receiver is much lower than that of the WARP board. When the same analysis was performed for residual error including phase noise, it was found that the cancellation is dependent on delay. Thus, phase noise acts as a bottleneck for active cancellation.

The cancellation operation is dependent on multiple factors such as precise modelling of RF cancellation delay lines, antenna placement, and training or preamble sequences that may not be common or standard amongst all communication protocols. Moreover, recalibration of the channel is required, and which is dependent on training sequences for active cancellation methods. Although this type of algorithm works for slow fading channels, it is not entirely a closed loop system in which the transceiver actively tracks the self-interference to adapt to any changes. With a closed loop system, processing is constantly performed, implying that such a system is real-time and on-line.

The aim of this work is to prove that the Least Means Squared (LMS) algorithm using steepest descent method works as a real time solution for echo cancellation. To show that adaptive cancellation is achievable with the LMS algorithm and that it can be used in other realms such as RF or analog baseband for cancellation. The main objectives of the work is as follows.

A. To design a full duplex system with a channel considering only scaling, delay and Additive White Gaussian Noise.
B. To design an LMS echo cancellation algorithm using steepest descent method.
C. To apply the designed algorithm on full duplex system.
D. To know how Bit Error Rate is varying with Signal to Noise Ratio.
E. To know how cancellation is varying with transmitted power.

III. SELF-INTERFERENCE IN FULL-DUPLEX COMMUNICATION

The wireless revolution has resulted in increased demands on the limited wireless spectrum, driving the quest for systems with higher spectral efficiency. In-band full-duplex (IBFD) operation has recently gained attention among the various ways to increase spectral efficiency. The main thought behind in-band full-duplex is as follows. Most contemporary communication systems contain terminals (e.g., base stations, relays, or mobiles) that function as both transmitters and receivers. Conventionally, these terminals operate in out of-band full-duplex or half-duplex, meaning that they transmit and receive either at over different frequency bands or different times. To enable wireless terminals to transmit and receive simultaneously over the same frequency band (i.e., IBFD operation) which offers the potential to double their spectral efficiency, as measured by the number of information bits reliably communicated per second per Hz. Hence it will be of great interest for next-generation wireless networks.

Beyond spectral efficiency, full-duplex concepts can also be advantageously used beyond the physical layer, such as at the access layer. From the access-layer point of view, enabling frame level in-band fullduplex, where a terminal is able to reliably receive an incoming frame while simultaneously transmitting an outgoing frame, could provide terminals with new capabilities. For example, terminals could detect collisions while transmitting in a contention-based network or receive instantaneous feedback from other terminal.

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A. Self-interference

Till now IBFD has not seen widespread use due to the potential debilitating effects of self-interference. Self-interference is the interference that a transmitting IBFD terminal causes to itself, which interferes with the desired signal being received by that terminal.

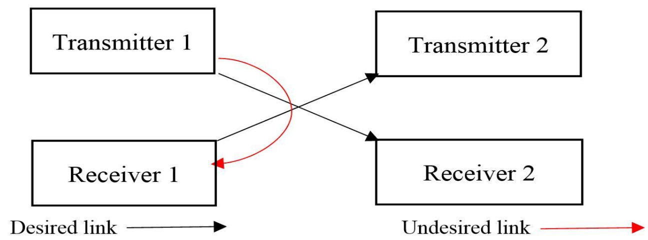

Fig. 1 Self-interference phenomenon

The fig. 1 illustrates the self-interference phenomenon which adversely affect the in-band full duplex communication. A full-duplex base-station to achieve the link SNR equal to that of a half-duplex counterpart, any full duplex design needs to provide 110dB of linear cancellation, 80dB of non-linear cancellation, and 60dB of analog cancellation.

B. Cancellation Techniques

There are number of self-interference cancellation technique that can be found in various research papers. They are generally categorized in two types, namely active and passive cancellation. Some previous work in this field [4,6,7,13,16,17,18] mainly uses passive, active or cascade of both cancellation technique so as to reduce self- interference signal level.

As mentioned in [13], complete self-interference cancellation is not possible because of the impairments of the radio circuits. Impairments like transmitter and receiver non-linearity, transmitter and receiver phase noise, ADC quantization noise can have severe impact on the SI cancellation level. Generally, power amplifier in transmitter chain introduces non-linearity in a circuit and this can be detrimental with increase in transmission power [11]. Therefore, a single cancellation scheme might not be sufficient to reduce the SI signal below noise floor due to which cascade of different cancellation schemes is suggested in various research papers [11].

C. Active Cancellation

Active SI cancellation technique is a process in which an inverse of interfering signal is generated and added to the self-interference signal in order to remove the interference. Active cancellation can be done in both baseband and band pass signal. It is generally classified as follows.

1) Antenna Cancellation
2) Analog Cancellation
3) Digital Cancellation

D. Antenna Cancellation

This cancellation technique is based on the fact that the two signals adds up in the space resulting in either constructive or destructive signal. Transmission signal (self- interfering signal) is divided and transmitted using two antennas TX1 and TX2 as shown in Fig. 2. The receiving antenna RX1 is placed at null point of two transmitting antennas TX1 and TX2, in such a way that the signal transmitted from the two antennas results in destructive combination at receiving end, thus mitigating some of the self-interference signal. After reducing self-interfering signal, the RX1 will be able to hear weaker signal of interest.

Antenna cancellation technique, as proposed in [1] utilizes three antennas, two for transmitting and one for receiving in a single node. Two transmitting antennas are kept at a distance of x and x+(λ/2) from the receiving antenna as shown in Fig. 2. [1]. The self-interfering signal from the two transmitting antennas adds up destructively at the receiving point.

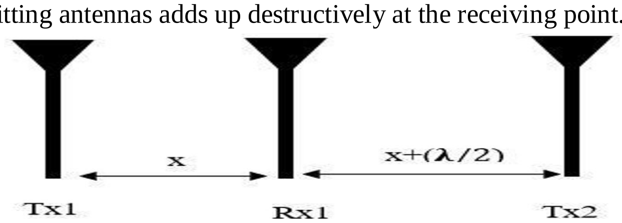

Fig. 2 Antenna setup for Antenna cancellation.

Additionally, antenna cancellation can only be achieved, if signal level from the transmitting antenna (TX1) which is close to the receiving antenna is attenuated to match with the signal level from another transmitting antenna (TX2) which is far away from the receiving antenna [1].

As mentioned in [1], maximum cancellation that can be achieved for a 5 MHz with centre frequency of 2.48GHz was 60.7dB when there is a perfect matching. In case of 1dB mismatch, the cancellation was observed to be 20dB.

Large SI signal voltage level covers most of dynamic range of the ADC which causes the low level SOI to suffer from large quantization noise [12]. Antenna cancellation reduces self-interference signal so that the ADC can accurately represent signal of interest. The amplitude or power of signal of interest is quite low with respect to interference signal and dynamic range of ADC is the limiting factor when representing the signal of interest. So, it is important to cancel out the interference in RF, so that the ADC can represent the SOI with enough precision [10].

Antenna separation method is not quite useful for a wideband signal where phase shifting is not uniform over entire bandwidth reducing the SI attenuation level [2]. It is mentioned in [2] that 40dB attenuation was achieved for RF signal having 2MHz bandwidth, whereas 10dB attenuation was observed for 20MHz bandwidth both centred at 2.4GHz carrier frequency.

E. Analog Cancellation

Analog cancellation technique implies cancellation of SI signal at RF level by adding up a phase inverted signal [3]. In [3], analog cancellation technique is proposed in which an additional auxiliary transmitter chain has been used for self-interference signal cancellation. In Fig. 2.3, original baseband signal u(n) has been up converted and transmitted using TX1 antenna resulting in band pass signal u(t). The self- interference channel h_{si} couples power in the receiving antenna RX thus interfering with signal of interest.

$$y(t) = h_{si} \otimes u(t) + h_{wire} \otimes d(t) + n \qquad (1)$$

The auxiliary transmitter chain TX2 uses a wired channel having response wire h with a canceller signal (t). It can be observed from the Fig. 3, that the received signal y(t) is the result of addition of three different signal including AWGN signal n and can be defined by the equation (1)

It can be seen from equation (1), in order to cancel out term $h_{si} \otimes u(t)$ and $h_{wire} \otimes d(t)$, h_{wire} must be equal to $-\hat{h}_{si}\, u \,/\, \hat{h}_{wire}$ where $\hat{h}_{si}$ and $\hat{h}_{wire}$ are the noisy estimation of SI h and wire h because of the estimation error, respectively.

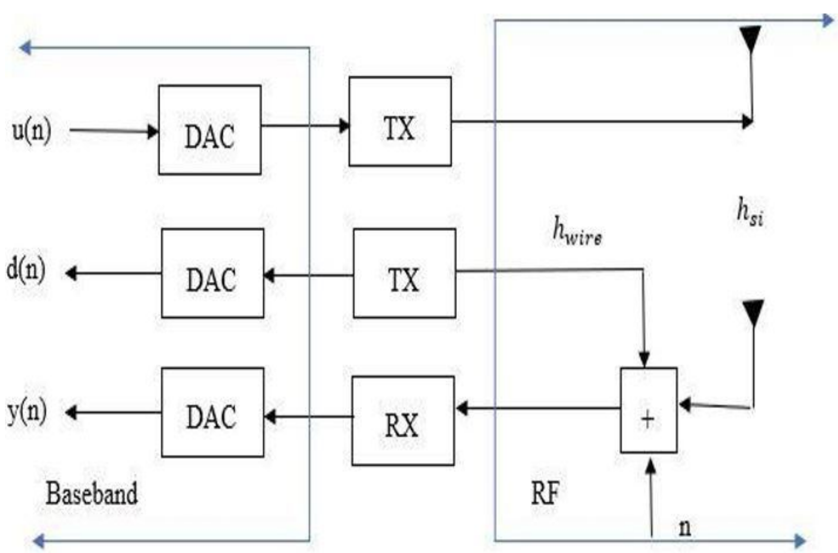

Fig. 3 Analog Cancellation Scheme utilizing two transmitting antenna and one receiving antenna.

The experiment in [3] was conducted with antenna separation of 20cm and 40 cm with a fixed distance of 6.5 cm between two nodes. The SI cancellation was observed to be 33 dB when separation was 20 cm where as 31dB cancellation was observed when separation was 40 cm. This is because greater the separation between the transmitting and receiving antenna, greater will be the estimation error because of the lower power SI signal coupling back to the receiving antenna. This will decrease the SI cancellation level.

F. Digital Cancellation

Digital cancellation is a technique in which self-interference is mitigated at the digital baseband level [1]. A prediction of received SI is formed, provided that transmitted samples are already known inside receiver where cancellation is performed.

In [11], A n on-linear digital cancellation technique is proposed along with RF cancellation which can be readily modified for the linear digital cancellation as shown in Fig. 4.

In [11] Fig. 4, x_n is original digital baseband transmit signal. The multipath channel between the transmitting antenna and receiving antenna is defined by the impulse response h_n. The estimated SI channel can be defined with an impulse response of $\hat{h}_n$. The estimated channel is modelled as FIR filter with L number of delays and the weight w_0 to w_{L-1}. The length of the estimation filter and tapping point can be different according to channel condition or multipath component. s_n is signal of interest which gets superimposed with self-interference signal and w_n is additive white Gaussian noise. So, from Fig. 2.4, total self-interference signal can be defined as shown in equation 2.

$$x_n^{si} = x_n \otimes h_n + S_n + w_n \qquad (2)$$

Similarly, the reference signal x_n is passed through the estimated channel to generate signal $\hat{x}_n^{si}$ which is used for cancelling out the self-interference signal and can be defined as shown in equation 3.

$$\hat{s}_n = x_n^{si} - \hat{x}_n^{si} \qquad (3)$$

The self-interference estimate is $\hat{x}_n^{si}$ defined as

$$\hat{x}_n^{si} = x_n \otimes \hat{h}_n = \sum_{k=0}^{L-1} w_k x_{n-k} \qquad (4)$$

After RF cancellation there is still some residual SI, which can be mitigated using digital cancellation at baseband [10]. Any adaptive estimation algorithm for example, LMS or RLS algorithm can be implemented to estimate a channel from residual SI signal.

It was observed in [12] that the amount of SI cancellation is directly proportional to the power of SI signal which gives the better estimation of SI channel by lowering the channel estimation error.

Linear signal processing method for digital cancellation cannot mitigate the effect of n on-linearity introduce by the transceiver chain. This is due to the fact, that the reference sample for the digital cancellation exist only in the digital state of the transmitter chain and does not include any non-linearity introduced by the component of transmitter chain, for example, power amplifier. This decreases the SI cancellation level by the digital cancellation method.

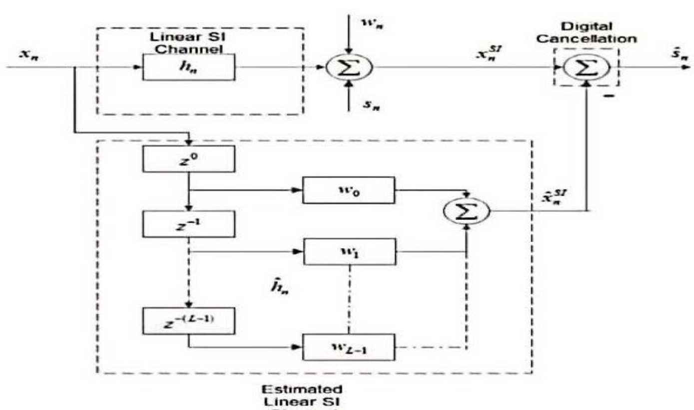

Fig. 4 Digital cancellation technique where the SI signal is regenerated and subtracted from the overall received signal at digital baseband.

The level of SI attenuation also depends on the number of training samples used for digital cancellation. Similarly, the length of the channel estimation filter should be long enough to produce good estimation.

G. Passive Cancellation

Passive cancellation is a technique, in which energy of a transmitted signal from a node is directed to a receiver at a different node, by using a directional antenna in order to suppress self-interference signal [10]. Similarly, path loss between the transmitting and receiving antenna also provides passive SI suppression.

In [10], it is mentioned that the directional diversity as one of the methods of the passive cancellation. This cancellation technique uses antenna isolation between the separate antennas or circulator isolation between shared antennas [11]. In directional diversity, node operating at full duplex completely relies on directional antenna for transmitting and receiving. The performance of this link depends on angular separation between these two antennas.

The larger the angle between the two antennas, more isolation is achieved between transmitting and receiving ends. A regular 2.4GHz patch antenna for both transmitter and receiver was used in [10]. It has been observed that the passive cancellation shows better performance in FD radio over HD radio even in the absence of extra hardware for cancellation.

Similarly, in [10] when using omnidirectional antenna there is not much angular separation due to which there is very low SIR when compared with the case using omnidirectional antenna. So, the choice of antenna type is also important, when there is requirement for passive cancellation.

In [3] antenna separation technique is used, in which path loss between transmitting and receiving antenna provides certain level of SI suppression. Since the distance between these two antennas is not high enough and does not contribute enough path loss, in this active cancellation technique is also used.

This work concentrates only on digital cancellation technique using LMS echo canceller which is implemented using Steepest descent method.

IV. DESIGN METHODOLOGY

A. Transmitter

The transmitter transmits modulated signal through antenna. Receiver receives the signal and demodulates it to get the actual signal. The transmitted signal is tapped and fed to LMS echo canceller which consists of FIR filter whose weights are varied iteratively. Algorithm calculates weights of FIR filter iteratively. The difference of altered transmitted signal from received signal is error signal which is fed to the receiver.

Transmitters are signal generators modulated with random bit pattern using BPSK modulation. b(n) be the bit pattern, NRZ encoding is performed on it by using the equation 5. After NRZ encoding $\hat{b}_n$ is multiplied with carrier signal to get BPSK modulated output $M(t)$ as shown in equations 6 & 7. x(n) shown in Fig. 5 is the sampled version of $M(t)$. The Fig. 6 is a block diagram of BPSK transmitter.

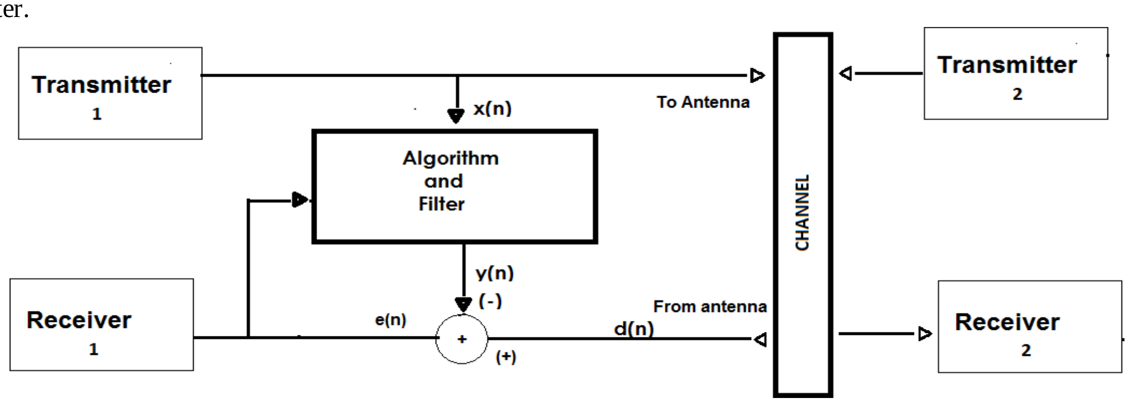

Fig. 5 Block view of Full duplex communication model.

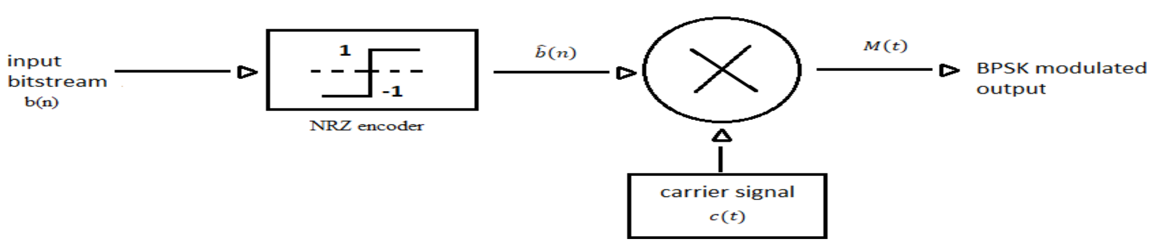

Fig. 6 BPSK Transmitter.

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$$\hat{b}(n) = \begin{cases} 1 & \Leftarrow \; b(n)=1 \\ -1 & \Leftarrow \; b(n)=0 \end{cases} \qquad (5)$$

$$c(t) = A_c \cos 2\pi f_c t \qquad (6)$$

$$M(t) = \hat{b}\, c(t) \qquad (7)$$

B. Receiver

Receivers are the BPSK demodulators. Carrier signal is multiplied with received signal be $M(t)$, as shown in equation 8. Output of multiplier $a(t)$ is fed to low pass filter whose cut-off frequency is greater than bitrate. Output of the filter be $\hat{a}(t)$. A threshold detector is used to recover bits $b_R(n)$ whose threshold level is 0 volts, as shown in equation 9. d(n) shown in Fig. 5 is a sample version of $M(t)$. The Fig. 7 is block diagram of BPSK receiver.

$$\alpha(t) = \hat{M}(t)c(t) \qquad (8)$$

$$b_R(n) = \begin{cases} 1 & \Leftarrow \; \hat{a}(t) > 0 \\ 0 & \Leftarrow \; \hat{a}(t) \le 0 \end{cases} \qquad (9)$$

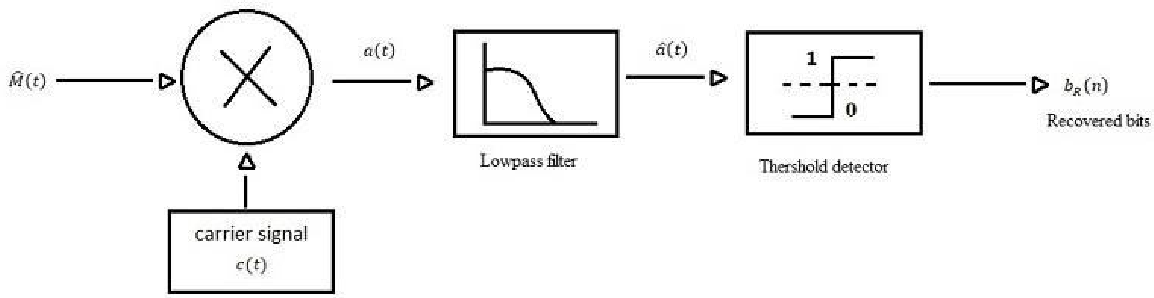

Fig.7 BPSK Receiver.

C. Algorithm and Filter

This is the main part of the work which enables the IBFD communication in a system. To reduce the self-interference, isolation of receiver from the transmitted signal must be achieved. This is impossible because both are operating at same frequency. However, the transmitted signal is known, it can be removed from the received distorted signal to get only desired signal i.e., TX2 signal. TX1 signal cannot be removed from received signal directly as TX1 and TX2 signals are mixed in the wireless channel. TX1 signal component present in the received distorted signal may undergo changes like scaling and delay in addition to the changes caused by the noise. Now there is need of emulating channel to get the TX1 signal component present in the received signal. A FIR filter is used to emulate the channel. The filter weights are calculated by an algorithm iteratively. A statistical signal processing algorithm known as LMS algorithm is used which uses Steepest Descent Method. The Fig. 8 shows a FIR filter with varying weights in accordance with the error signal.

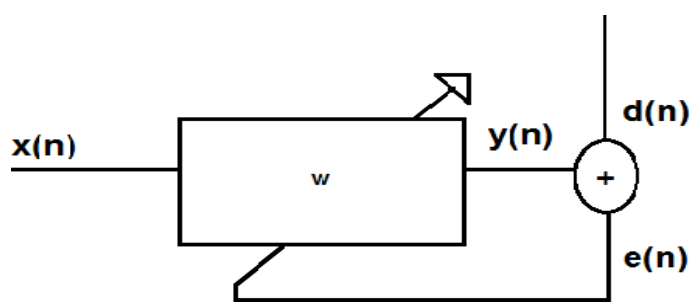

Fig.8 FIR filter adapting weights using error signal.

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D. Steepest Descent Method

This method pivots on the point that the slope at any point on the surface provides the best direction to move in. It is a feedback approach to finding the minimum of the error performance surface. The greatest change in elevation of the surface of the cost function is given by the steepest descent direction for a given step laterally. The knowledge of this direction is used by the steepest descent procedure to move to a lower point on the surface and find the bottom of the surface in an iterative manner. From the Fig. 9, the following facts become evident that the slope of the function is zero at the optimum value associated with the minimum of the cost function. There is only one global minimum of the bowl- shaped curve and no local minima. The slope of the cost function is always positive at points located to the right of the optimum parameter value as shown in Fig. 9. The larger the distance from this point to the optimum value, the larger is the magnitude of the slope of the cost function at any given point.

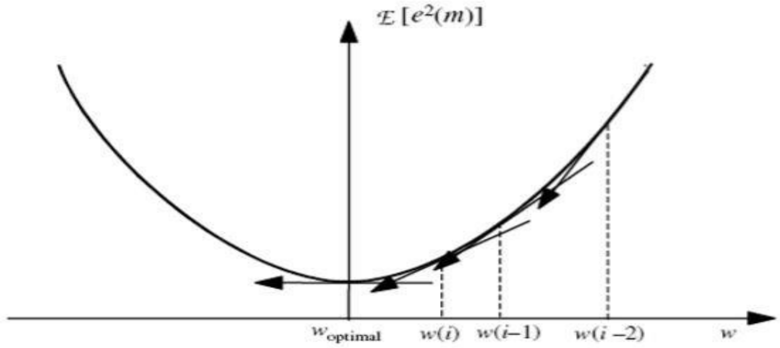

Fig.9 Cost function of MSE.

The aim is to iteratively descend to the bottom of the cost function surface, so that weight approaches optimum point i.e., the coefficients are updated repetitively using a strategy analogous to that of the ball rolling in a bowl. Convergence time is defined as the time taken by the algorithm to adapt and reach the optimum weight. It must be as less as possible because it directly reflects on the mean circuit delay. As the system is a communication system the mean circuit delay must be less, it must be almost on line. This convergence time dependent on step-size of the algorithm.

E. Mathematical Representation of Steepest Descent Method

To find the weights of the filter, Weiner-Hopfs equation need to be solved,

$$W=R^{-1}P \qquad (10)$$

Where W is the weight array of the filter, R is the auto correlation of received signal sample sequence, P is the cross correlation of received signal with transmission signal sample sequences.

Let X be the array of TX1 signal samples, D be the array of incoming received signal samples, μ be the step size, *g* be the gradient.

Therefore,

$$R = X.X^T \qquad (11)$$

$$P = D.X \qquad (12)$$

$$Gradient\ (g) = P - R.W \qquad (13)$$

Weight update equation of steepest descent method,

$$W = W + \mu.g \qquad (14)$$

Substituting *g* from equation 13,

$$W = W + \mu.(P - R.W) \qquad (15)$$

Substituting P and R from equation 11 and equation 12,

$$W = W + \mu.(D.X - X.X^T W) \qquad (16)$$

$$W = W + \mu.X.(D - X^T W) \qquad (17)$$

Since,

$$Y = X^T W \qquad (18)$$

$$W = W + \mu.X(D - Y) \qquad (19)$$

$$\text{Since } E = D - Y \qquad (20)$$

$$W = W + \mu.X.E \qquad (21)$$

Now, converting array form of equation to instantaneous equation

$$w(n+1) = w(n) + \mu.x(n).e(n) \qquad (22)$$

x(n), e(n) are signal samples and w is weight of the filter as shown in Fig. 9.

The step size in the steepest descent method is constant. An appropriate step size must be chosen which affects the convergence time of the algorithm. The equation 23 shows the constraints of step size (μ).

$$0 < \mu < \frac{2}{\|\lambda_{max}\|^2} \qquad (23)$$

Where, λ_{max} is the maximum eigen value of auto correlation matrix R.

$$\mu(n) = \mu(n-1) + \rho.e(n-1).e(n).x(n-1).x(n-1) \qquad (24)$$

But at an instant of time λ values cannot be found. So, an appropriate μ can be found using adaptive step size which is used in GASS (Gradient Adaptive Step Size) method proposed in [11]. Shown in equation 24. The ρ in equation 24 is a constant chosen as 0.0000001.

F. Channel

The channel should be wireless but only scaling, delay and additive white gaussian noise effects are considered as shown in Fig. 10. The degraded signal is a signal which is affected by channel impairments and self-interference. This signal is fed to the algorithm as a received signal. Algorithm reduces both self-interference and channel effects.

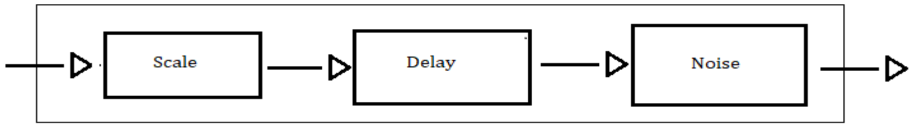

Fig.10 Wireless Channel.

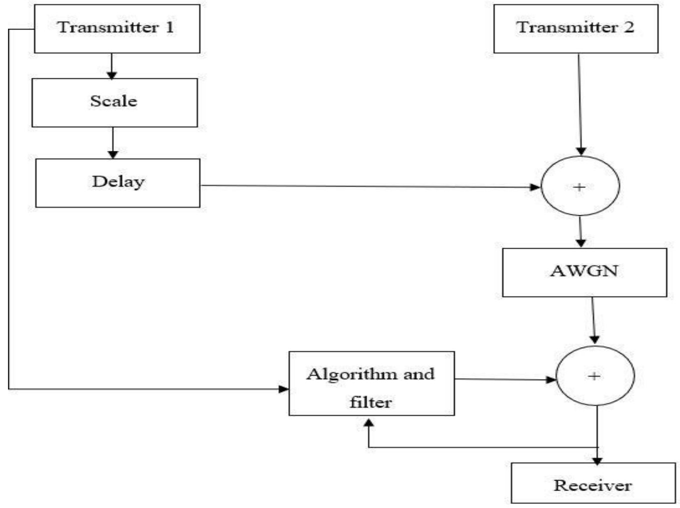

Fig.11 Simulation flow model.

The Fig. 11 shows the simulation flow model. Two blocks of same type of transmitter blocks are created, i.e., TX1 and TX2. Output of the TX1 signal is scaled, delayed, and added with TX2 Signal in AWGN effected channel. Received signal affected by self-interference and channel is fed to algorithm and filter.

The error signal generated is the desired signal which is fed to the receiver. The receiver detects and recovers bits. These recovered bits are compared with TX2 bits to compute BER.

G. Bit Error Rate Calculation

Bits recovered in receiver are compared with TX2 bits to calculate BER using equation 25. BER is calculated at different SNR values.

$$BER = \frac{Number\ of\ error\ bits}{Total\ number\ of\ bits} \quad (25)$$

H. Suppression Ratio/Cancellation Calculation

Calculating a parameter called cancellation, also known as suppression ratio, helps to know how much amount of self-interference is cancelled or suppressed. Cancellation is calculated using equation 26.

$$Cancellation\ in\ dB = 10\log_{10}\frac{Energy\ of\ received\ SI\ signal}{Energy\ of\ residual\ noise\ present\ in\ the\ output}\ \text{dB} \quad (26)$$

V. RESULTS AND DISCUSSION

Self-interference is generated using unmodulated sinusoidal signals. An LMS echo canceller is designed using algorithm based on steepest descent method in MATLAB. Firstly, the LMS echo cancellation is performed on unmodulated sinusoidal signals with constant step-size later with adaptive step-size. Full duplex channel is created using BPSK transceivers and channel considering only scaling, delay and additive white gaussian noise. LMS echo cancellation is performed on BPSK signals with adaptive step-size. Characteristics of LMS echo canceller like convergence time, weights, step-size, amount of cancellation are obtained.

A. LMS echo Cancellation on Unmodulated sine Signals

Convergence time must be as less as possible because it directly reflects on the mean circuit delay. As the system is a communication system the mean circuit delay must be less, it must be almost on-line. This convergence time dependent on step-size of the algorithm. Convergence time at different constant step-sizes is noted and plotted.

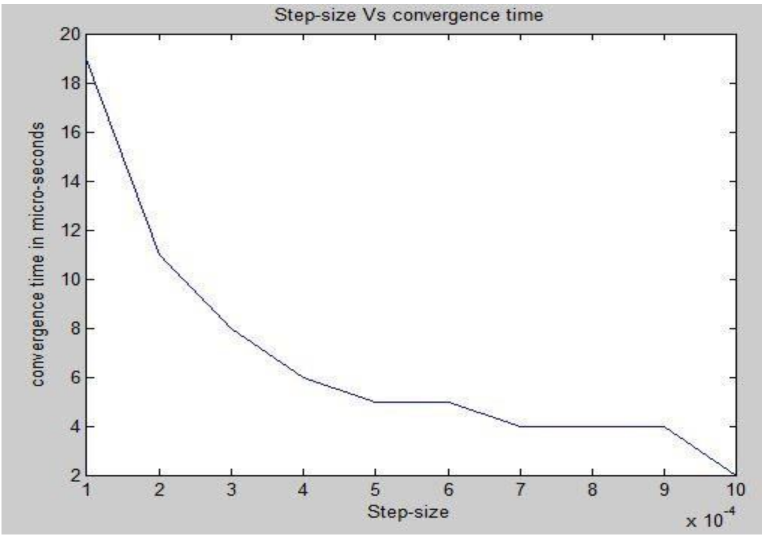

Fig. 12 Step-size vs Convergence time at 1MHz Carrier

The Fig. 12 shows the variation of convergence time with respect to step- size. Therefore, choosing an appropriate value for step-size plays a vital role in decreasing the convergence time without effecting the output of the system. This drawback can be resolved by using Adaptive step-size, using adaptive step-size LMS echo cancellation is performed on 1MHz and 5MHz unmodulated sinusoidal signals. Estimated and error signals with TX1 and TX2 signals are shown in the Fig. s 13 and 14.

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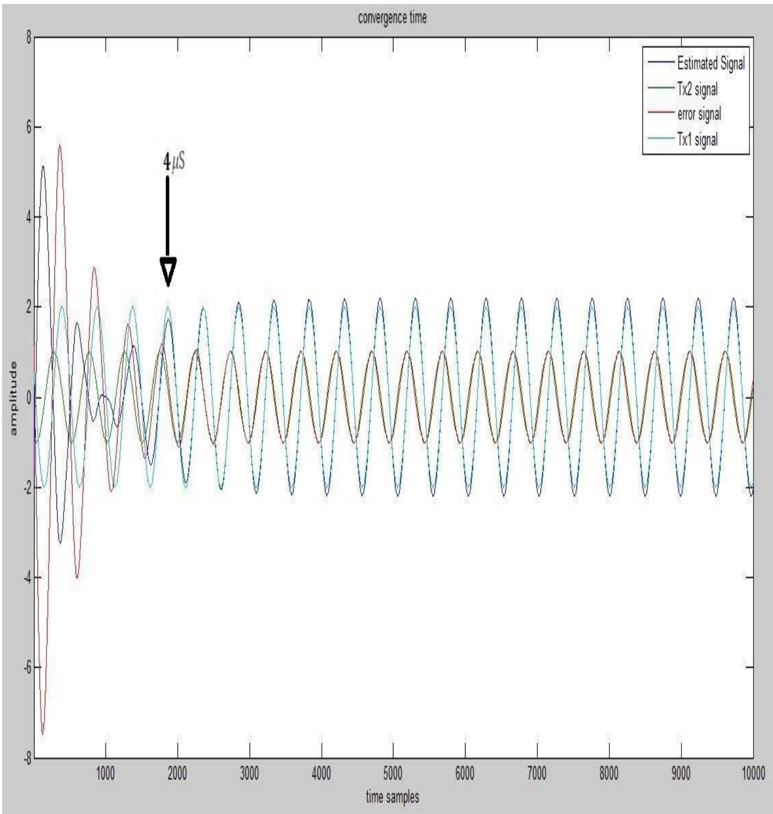

Fig. 13 Convergence time at 1MHz

The proposed method almost meets the requirement. The convergence time is 4µS for 1MHz sine wave, shown in Fig. 13. The convergence time is 1.7µS for 5MHz sine wave, shown in Fig. 14. The error signal is converging to match TX2 signal at convergence time.

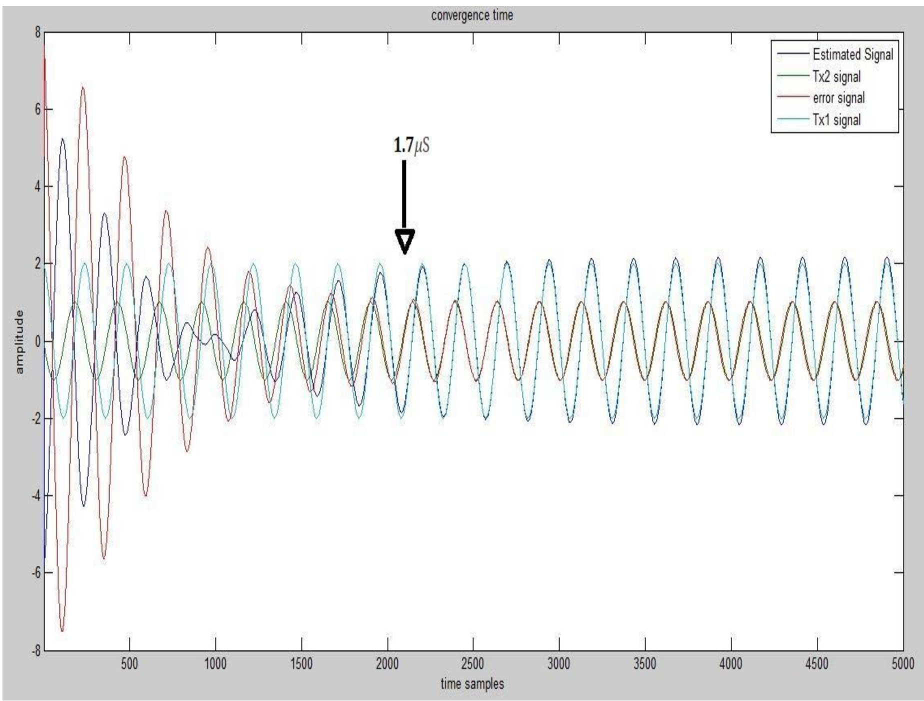

Fig. 14 Convergence time at 5MHz

The plots shown in the Fig. s 15 and 16 show the variation of weights with respect to iterations. It is clear that weights are converging to an optimum value after some iterations. Here as shown in Fig. 15 and 16 weight of the filter converging to 0.375 after 2500 iterations.

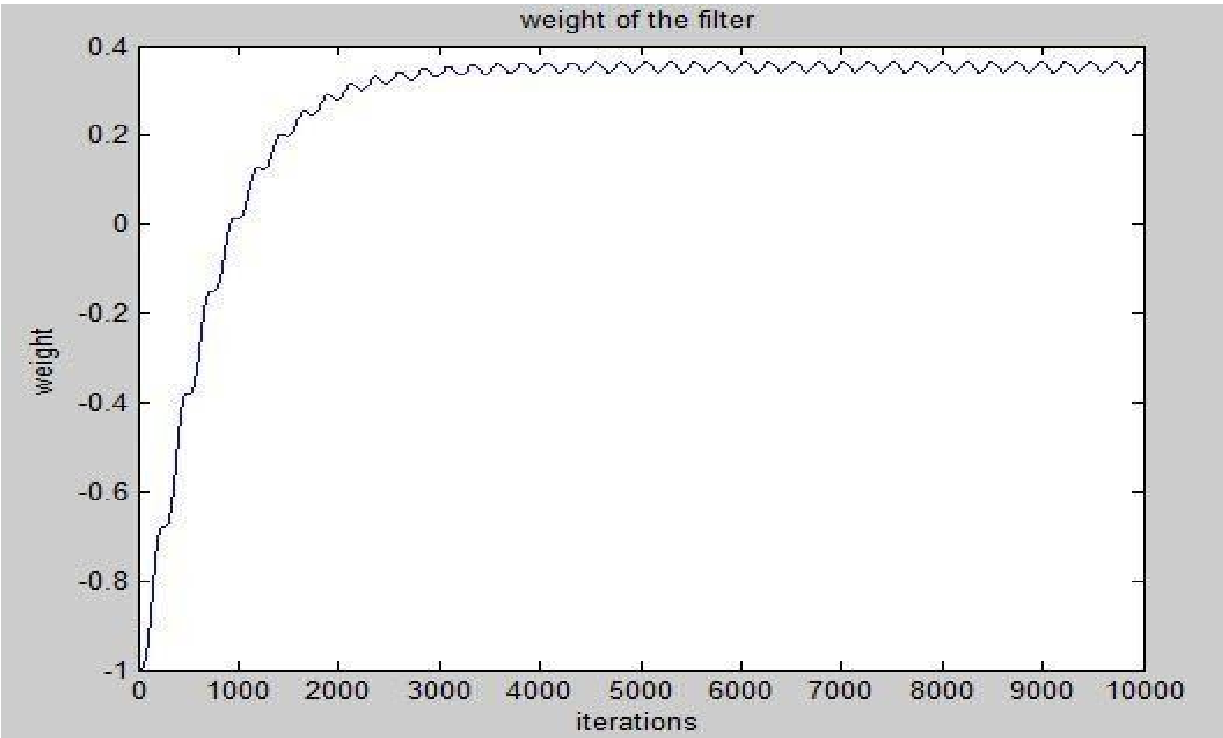

Fig. 15 Weight of the filter at 1MHz

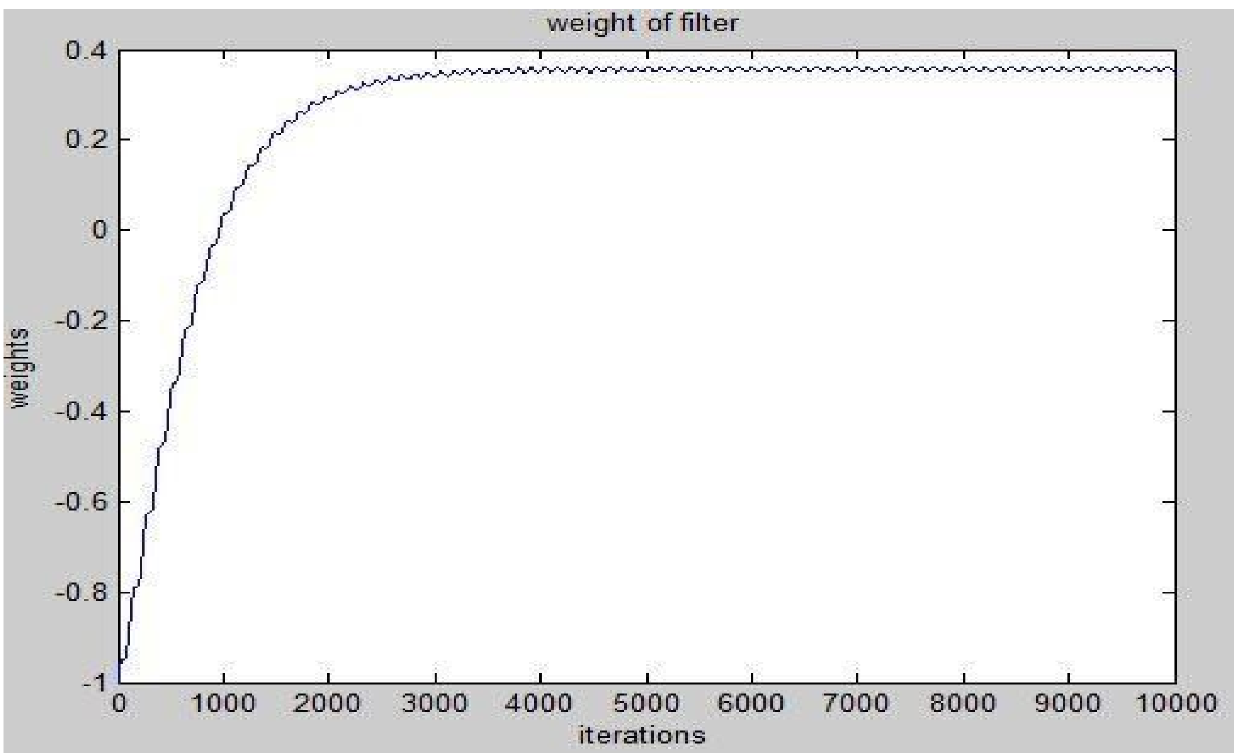

Fig. 16 Weight of the filter at 5MHz

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LMS echo cancellation using adaptive step-size is performed on unmodulated sinusoidal signals. It is observed that the step-size is converging to optimal value with respect to iterations. The step size (μ) is converging to 0.00019 after 2500 iterations as shown in Fig. 17 and 18.

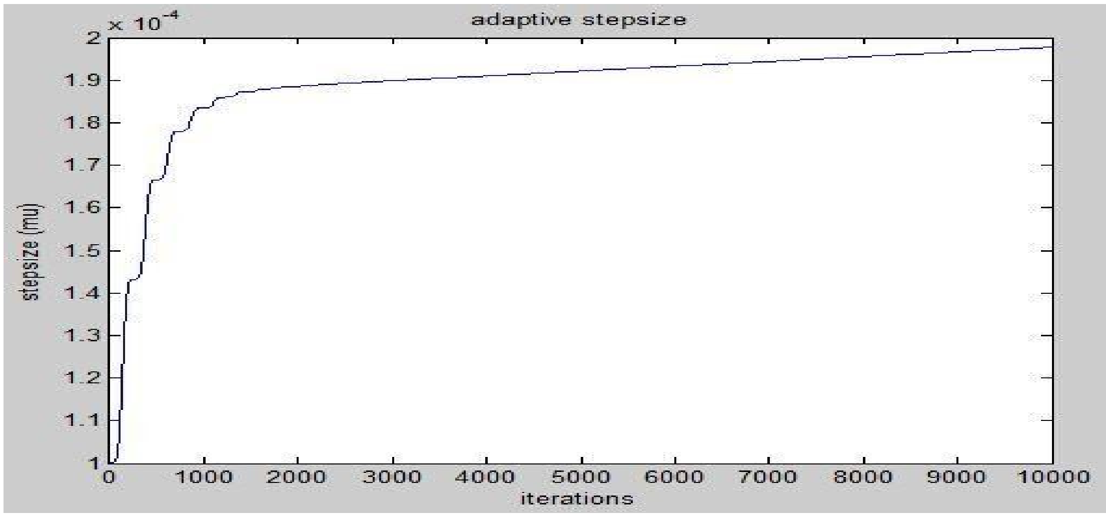

Fig. 17 Adaptive step size of algorithm at 1MHz

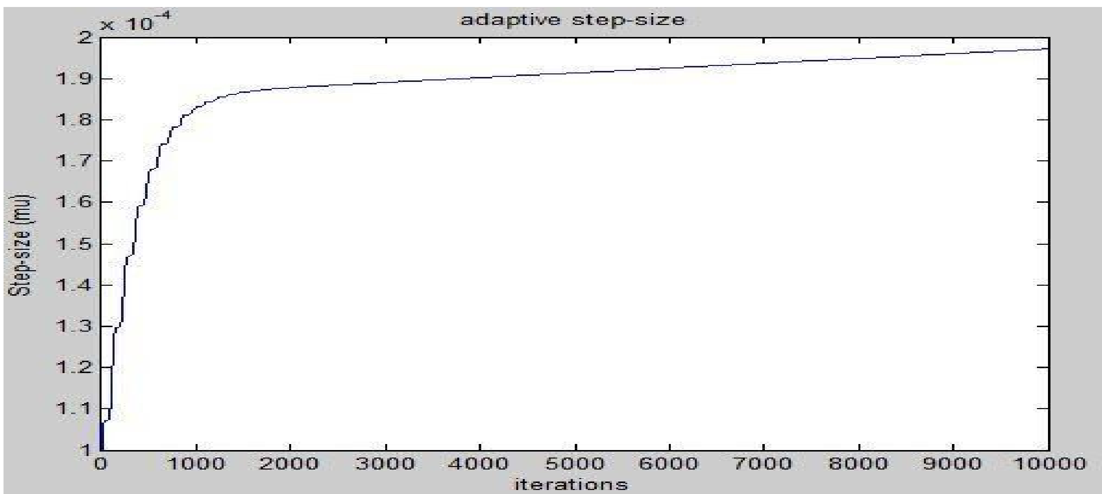

Fig. 18 Adaptive step size of algorithm at 5MHz

A full duplex channel is emulated using two BPSK transceivers. The simulation parameters are tabulated in the Table 1.

TABLE I
Simulation Parameters

Parameters	Value
Number of bits transmitted	100
Carrier Frequency	5 MHz
Sampling Frequency	100 Msps
Number of samples per bit	1000
Channel Type	AWGN
Transmitted Power	1 dBm
Modulation Type	BPSK

The NRZ encoded bits are multiplied with sinusoidal signal to generate BPSK signals. The Fig. 19 shows the PSK signal generated by two transmitters at the same point of time.

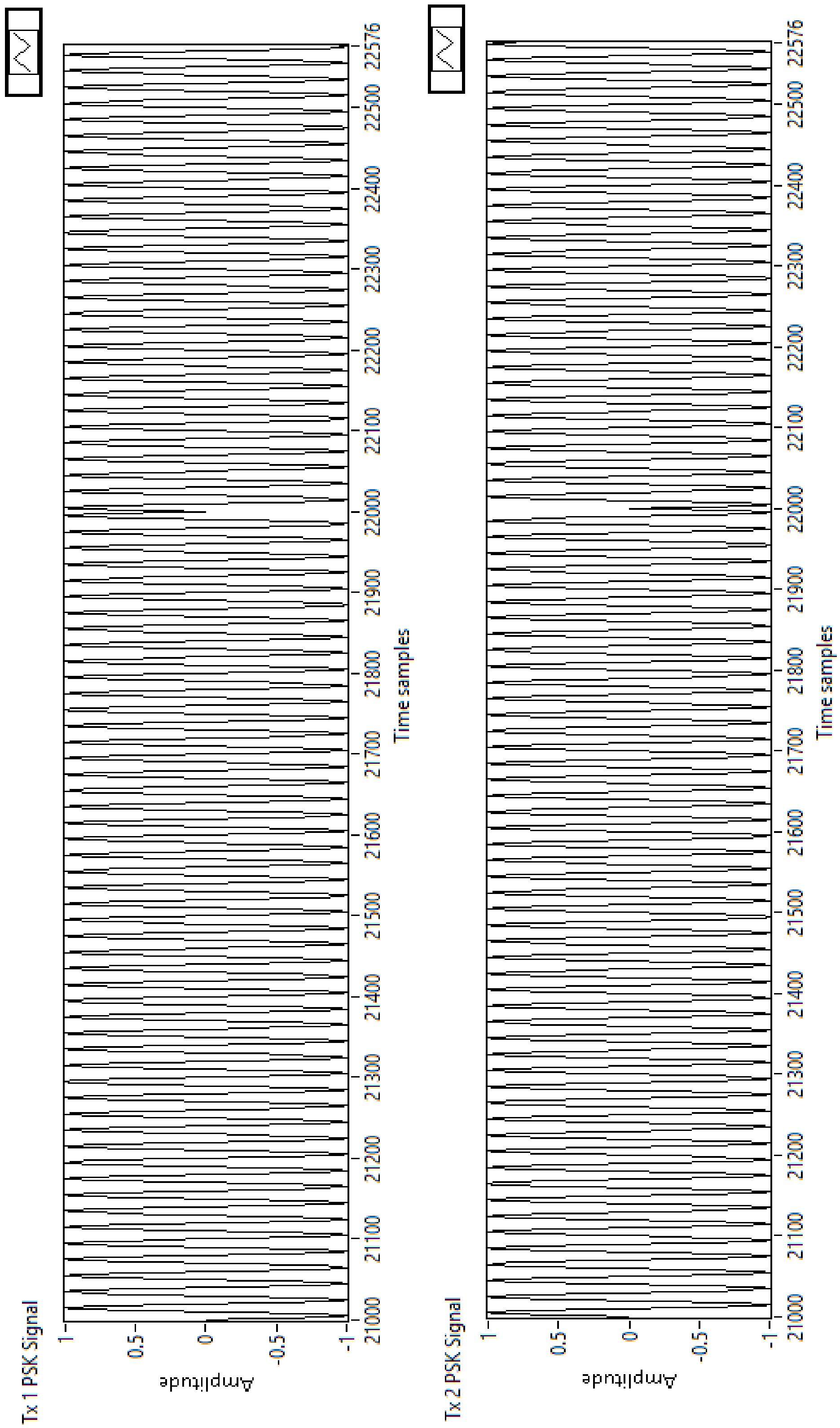

Fig. 19 PSK signals generated by TX1 and TX2.

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The generated BPSK signals are affected by SI as they are combined in channel. The SI affected signal is shown in Fig. 20. Signal after SI cancellation using LMS echo canceller is shown in Fig. 20.

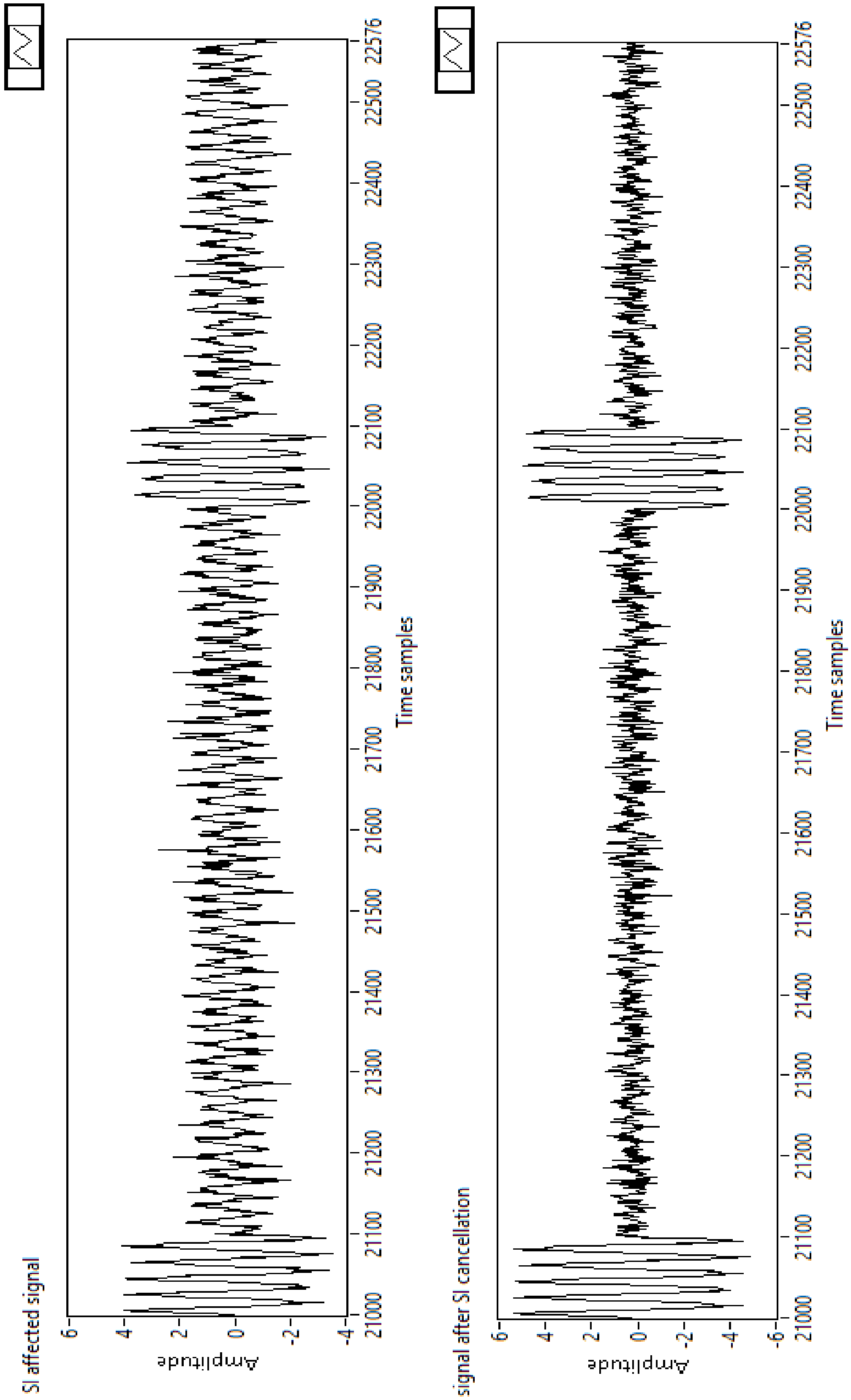

Fig. 20 SI affected signal and Signal after SI cancellation.

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After SI cancellation, demodulation is performed and the signal after SI cancellation is multiplied by the carrier and passed through low pass filter to obtain demodulated signal which is shown in Fig. 21 after demodulation.

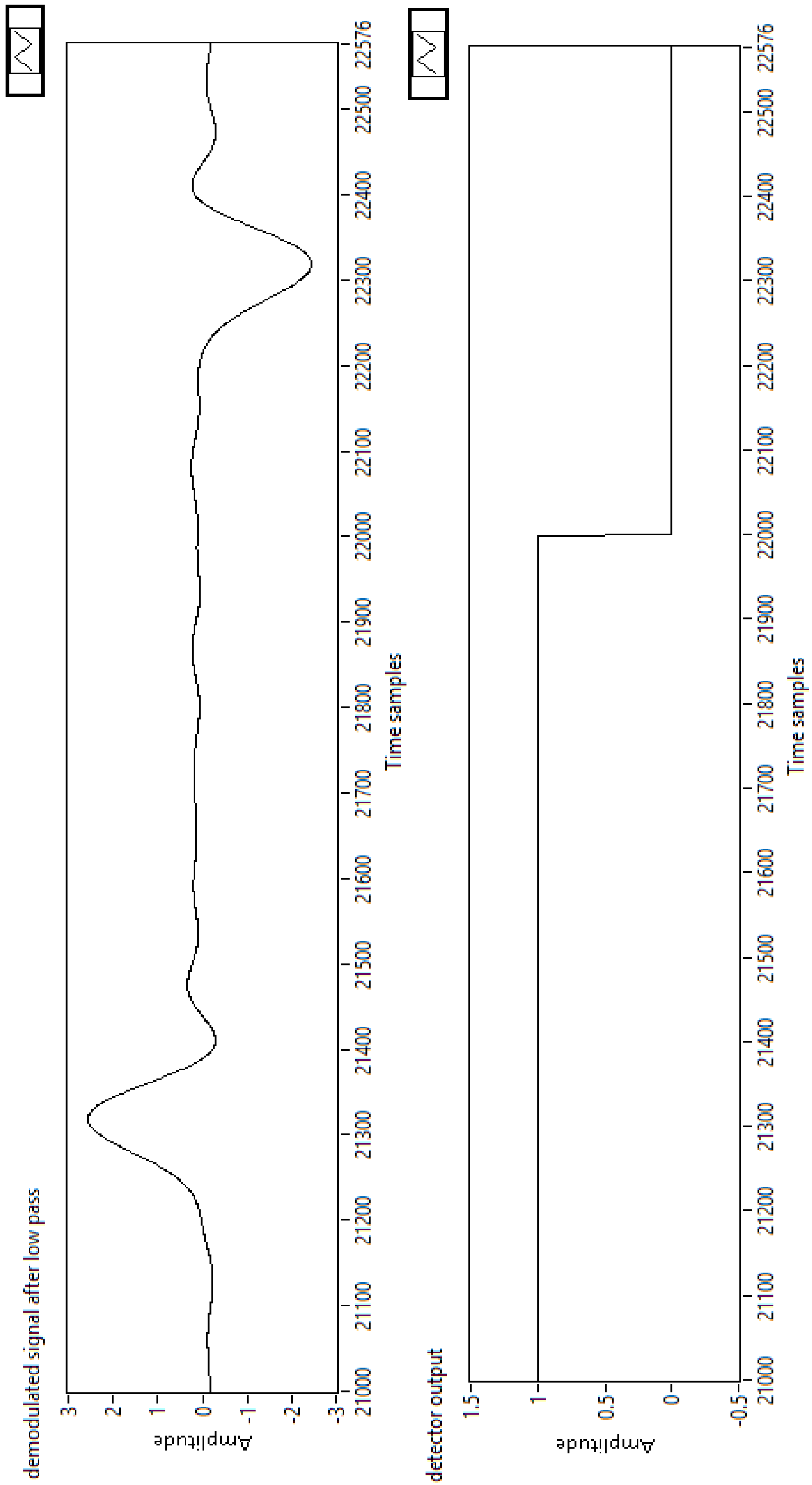

Fig. 21 Output of the threshold detector.

BER is standard measure to know the efficiency of a digital communication system. It is a ratio of number of error bits to the total number of bits transmitted. It can be calculated theoretically for AWGN channel using equation 27. The plot between SNR and theoretical BER is shown in the Fig. 22.

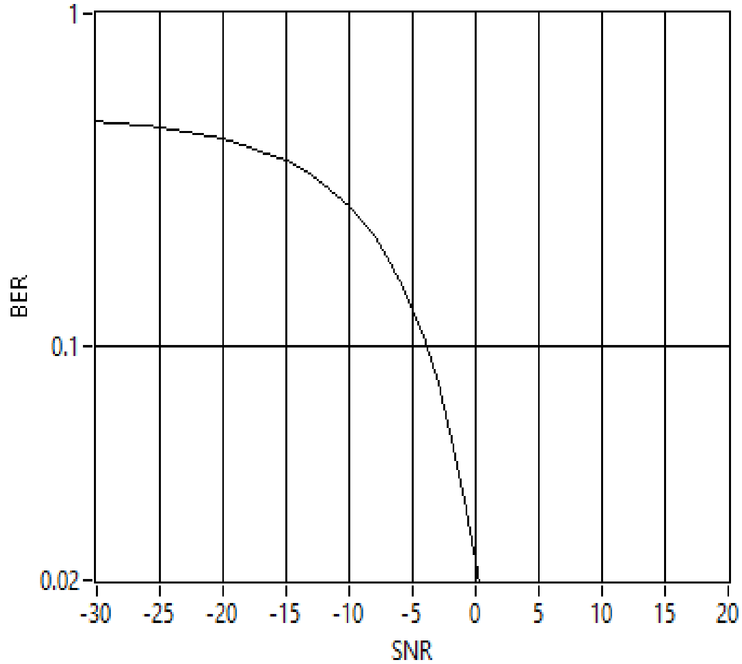

Fig. 22 SNR vs theoretical BER.

The measured BER varies slightly with different bit patterns, four of them are shown in Fig. 23. BER measured at different bit patterns are averaged to plot a SNR Vs BER graph, as shown in Fig. 24. In plots of Fig. s 22, 23, 24 BER is logarithmic. SNR is varied from -30dB to 20dB, BER calculated is varying from 0.4 to 0.04 respectively as shown in Fig. 24.

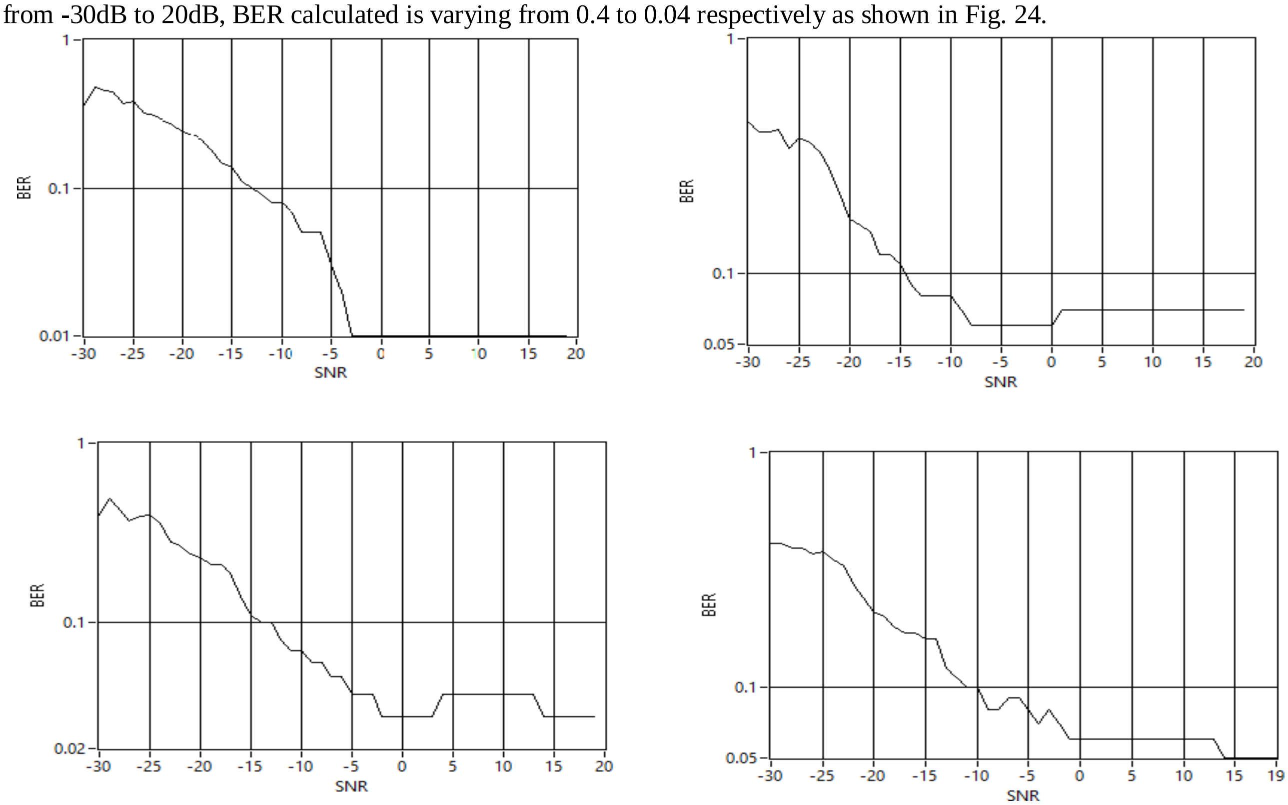

Fig. 23 SNR vs BER for different patterns.

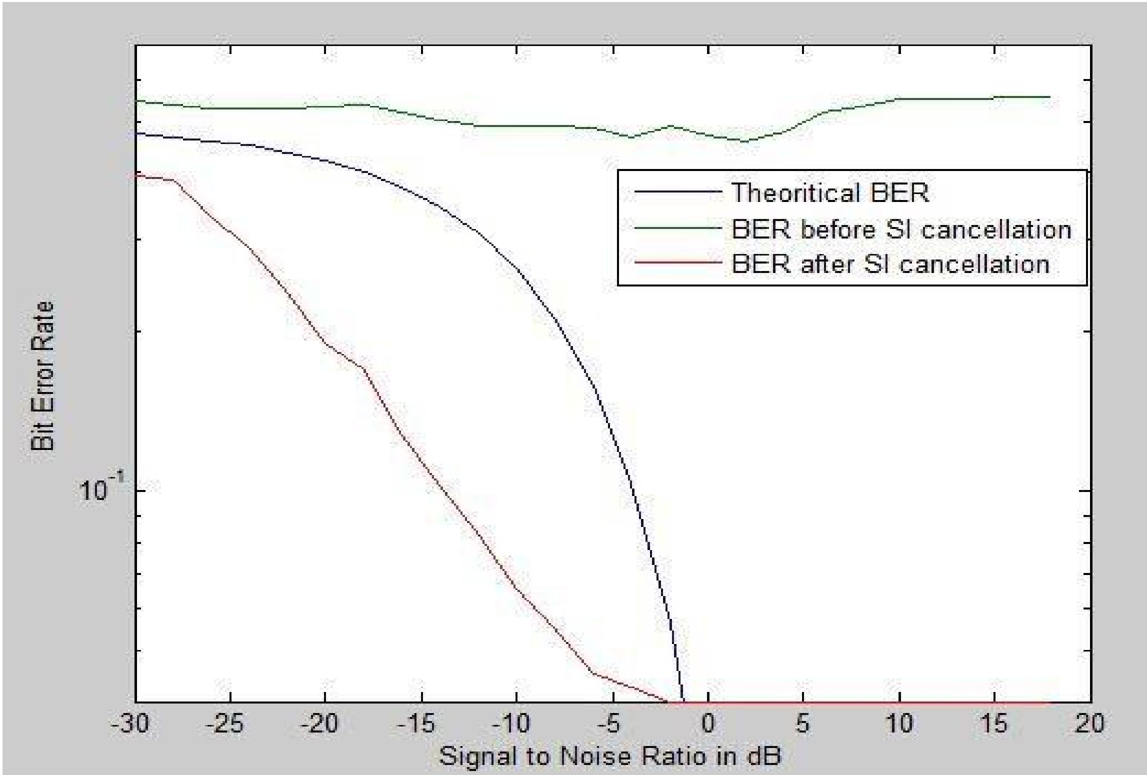

Fig. 24 SNR vs BER.

Cancellation (or suppression ratio) is parameter shows the amount of self-interference suppressed. It is calculated using equation (26). At different transmitted powers, cancellation is calculated and plotted which is shown in the Fig. 25. The amount of cancellation is increasing linearly with transmitted power till 75dB after 12dBm amount of cancellation decreases.

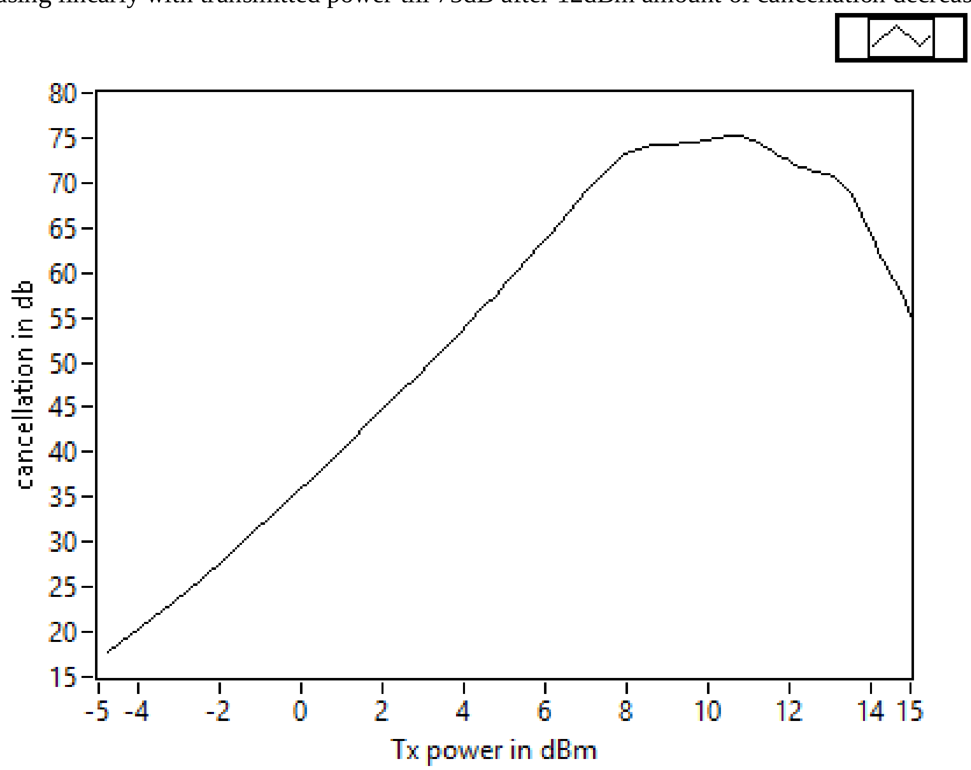

Fig. 25 Transmitted power Vs Cancellation.

VI. CONCLUSIONS

In this paper, the effect of LMS algorithm using Steepest Descent Method on SI cancellation in full duplex system was studied. A simple BPSK transceiver was used in LabVIEW environment to carry out this analysis. The performance of LMS algorithm was based on steady state error and convergence rate.

Increasing the step size to certain threshold has increased the amount of SI attenuation. It is important to choose the right value of step size because increasing the step size to beyond some threshold can result in divergence of error signal, resulting in the lower amount of SI attenuation. Hence, the adaptive step-size is used to mitigate the above effect.

By varying SNR from -30 dB to 20 dB, BER calculated is varying from 0.4 to 0.04 respectively. Bit errors from -30 dB to 0 dB are less than theoretical values calculated for AWGN after 0 dB bit error rate saturates to 0.04. It can be concluded that the Bit error due SI are reduced to the maximum extent.

By increasing the transmitted power, the amount of SI cancellation increases till 75 dB and saturates. At higher powers again amount of SI cancellation decreases. It can be concluded that this method of cancellation is limited to low powers i.e., below 12 dBm.

The length of adaptive filter must be equal to the number of taps used in SI coupling channel in order to get the better estimation of channel. So, it is important to know the characteristics of SI channel. A longer filter length increases the computational resources than a shorter filter length. However, increasing filter length can only attenuate the SI signal to some threshold. A small filter length can increase the convergence speed.

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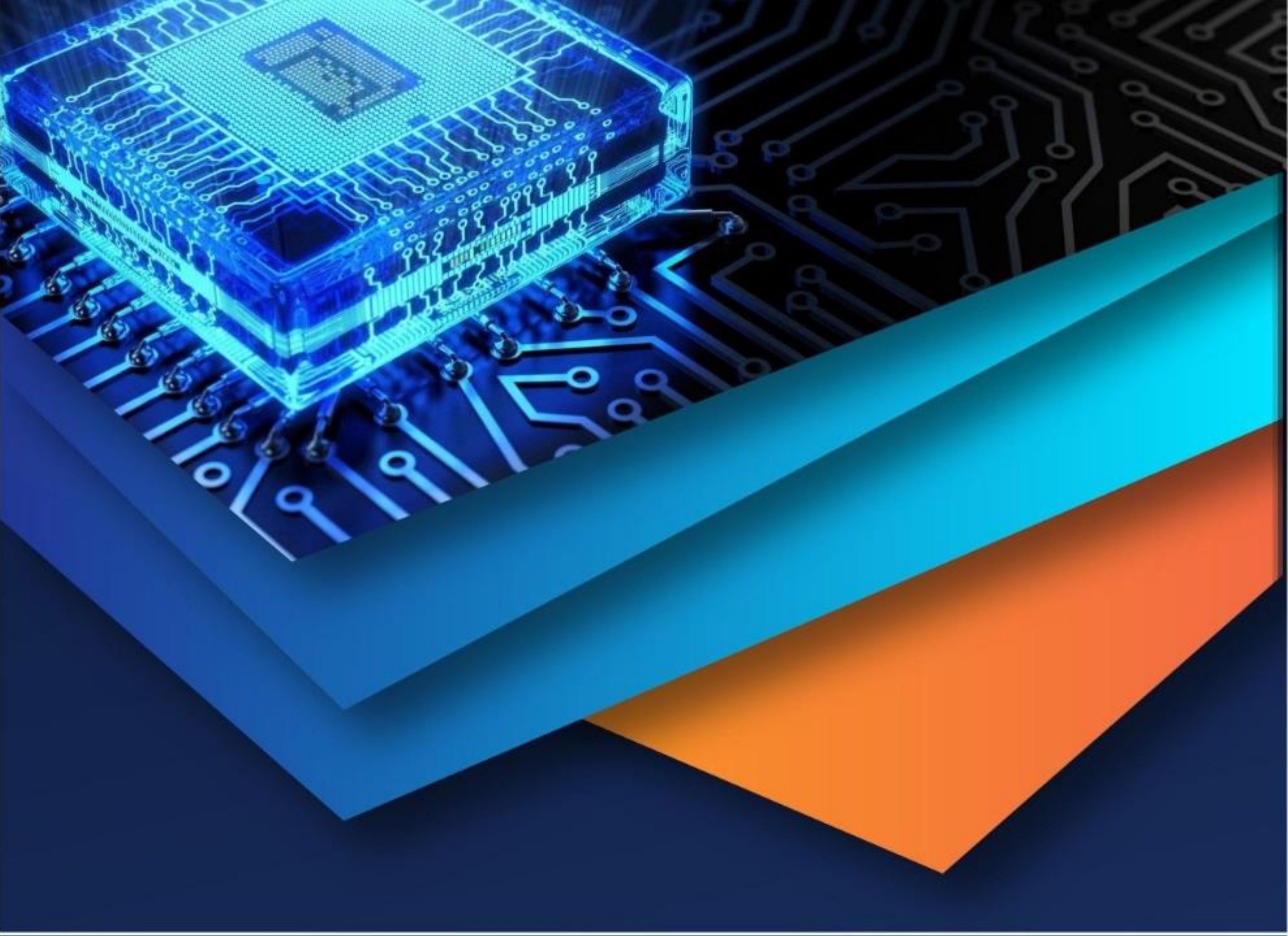

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