Signal-Processing Strategy for Restoration of Cross-Channel Suppression in Hearing-Impaired Listeners

V.Purushothaman[1] , Naveen.M[2], Prem kumar.D[3], P.Manimaran[4], K.S.Sivaraman[5]
[1,2,3,4,5]Vel Tech engineering college
Chennai,India
[1]purushothamanece@gmail.com

Abstract—Because frequency components interact nonlinearly with each other inside the cochlea, the loudness growth of tones is relatively simple in comparison to the loudness growth of complex sounds. The term *suppression* refers to a reduction in the response growth of one tone in the presence of a second tone. Suppression is a salient feature of normal cochlear processing and contributes to psychophysical masking. Suppression is evident in many measurements of cochlear function in subjects with normal hearing, including distortion-product to acoustic emissions (DPOAEs). Suppression is also evident, to a lesser extent, in subjects with mild-to moderate hearing loss. This paper describes a hearing-aid signal processing strategy that aims to restore both loudness growth and two-tone suppression in hearing-impaired listeners. The prescription of gain for this strategy is based on measurements of loudness by a method known as categorical loudness scaling. The proposed signal-processing strategy reproduces measured DPOAE suppression tuning curves and generalizes to any number of frequency components. The restoration of both normal suppression and normal loudness has the potential to improve hearing-aid performance.

1.INTRODUCTION

The cochlea is a portion of the inner ear that looks like a snail shell (cochlea is Greek for snail.) The cochlea receives sound in the form of vibrations, which cause the stereocilia to move. The stereocilia then convert these vibrations into nerve impulses which are taken up to the brain to be interpreted. Two of the three fluid sections are canals and the third is a sensitive 'organ of Corti' which detects pressure impulses which travel along the auditory nerve to the brain. The two canals are called the vestibular canal and the tympanic canal.Maintaining the Integrity of the Specifications.

A cochlear implant is very different from a hearing aid. Hearing aids amplify sounds so they may be detected by damaged ears. Cochlear implants bypass damaged portions of the ear and directly stimulate the auditory nerve. Signals generated by the implant are sent by way of the auditory nerve to the brain, which recognizes the signals as sound. Hearing through a cochlear implant is different from normal hearing and takes time to learn or relearn. However, it allows many people to recognize warning signals, understand other sounds in the environment, and enjoy a conversation in person or by telephone.

Use of a cochlear implant requires both a surgical procedure and significant therapy to learn or relearn the sense of hearing. Not everyone performs at the same level with this device. The decision to receive an implant should involve discussions with medical specialists, including an experienced cochlear-implant surgeon. The process can be expensive. For example, a person's health insurance may cover the expense, but not always. Some individuals may choose not to have a cochlear implant for a variety of personal reasons. Surgical implantations are almost always safe, although complications are a risk factor, just as with any kind of surgery. An additional consideration is learning to interpret the sounds created by an implant. This process takes time and practice. Speech-language pathologists and audiologists are frequently involved in this learning process. Prior to implantation, all of these factors need to be considered.

In existing system, ear outer drum disabilities are using hearing machine it is to used to hear the sound signal. But inner drum disabilities persons cant able to hear the voice. The hearing machine only amplify input signal.

In this project we are going implement Signal-Processing Strategy for Restoration Of Cross-Channel Suppression in Hearing-Impaired Listeners. It is very useful for hearing disability person. Voice input given to mic after that we use some filters to compress the voice. Cleared voice stored in storage device of controller, after processing of controller give the another section depends upon the channel selection. In decompress section retrieve the given input voice but it has less amplitude so we have to amplify the signal. After amplification we use transducer for converting sound signal into vibrating signal.

The organization of this paper (after Section I) is as follows. Section II describes the gamma-tone filters how they generate time-frequency surfaces based on a gamma tone analysis. Section III describes about how OFD used as a digital multi-carrier modulation, section IV describe the multi-microphone sub-band adaptive processing. Section V describes about simulation of SNRvsBER. Finally, we discuss and summarize our contributions in Section VI and provide concluding remarks

II.GAMMA TONE FILTERS

Generally speaking, auditory features leverage the characteristics of the human auditory system, i.e., different forms of response to different frequency components of signals. It is generally accepted that the frequency analysis performed by the peripheral auditory system can be modeled to a reasonable degree of accuracy by a bank of linear bandpass filters. A particular novelty of our implementation is that the Gammatone filters are realized purely in the time domain. Specifically, the filters are applied directly on time series of speech signals by simple operations such as delay, summation and multiplication.

This is quite different from the widely adopted frequency-domain design, for instance where signals are transformed to frequency spectra first and the Gammatone filers then applied upon them. The time domain implementation avoids unnecessary approximation introduced by short-time spectral analysis, and saves a considerable proportion of computation involved in FFT. The bandwidth of the filters are set by a critical band function and so filter bandwidth increases with center frequency. If the energy at the output of each filter is calculated at a given point in time, and the values are plotted as a function of filter center frequency.

TIME DOMAIN GAMMA TONE FILTERS

The gammatone auditory filter can be described by its impulse response:

$$g(t) = at^{n-1}\, e^{-2\pi bt}\cos(2\pi f_c t + \varphi)$$

where fc is the central frequency of the filter, and φ is the phase which is usually set to be 0. The constant a controls the gain, and n is the order of the filter which is usually set to be equal or less then 4 [6]. Finally b is the decay factor which is related to fc and is given by:

$$b = 1.019 * 24.7 * (4.37 * f_c/1000 + 1)$$

A set of GFs with different fc form a gammatone filterbank, which can be applied to obtain the signal characteristics at various frequency, resulting in a temporal-frequency presentation similar to the FFT-based short-time spectral analysis. In order to simulate human auditory behavior, the central frequencies of the filterbank are often equally distributed on the Bark scale.We now derive the time domain GF implementation. First notice that (1) consists of two components: the filter envelope $at^{n-1}e^{-2\pi bt}$ and the amplitude modulator $\cos(2\pi fct + \varphi)$ at frequency fc.

Applying the Fourier analysis, the frequency domain representation of g(t) is obtained:the Laplace transform on the original continuous signal corresponds to the following Z transformation on the sampled discrete series:

$$G(z) = \frac{a(n-1)!}{2}(1 - e^{j2\pi f_c - 2\pi b})^{-n}.$$

Letting A(z) be an element transform:

$$A(z) = \frac{1}{1 - e^{j2\pi f_c/f_s - 2\pi b/f_s} z^{-1}},$$

G(z) can be regarded as a cascade of n recursive application of A(z). Note that A(z) is dependent on the central frequency fc, and so is the entire transform G(z). This can be simplified by a cascaded application of a series of filters which first remove the fc component and then apply a base filter ^ G(z) that is independent of fc, and finally compensate for fc. This process is shown in Fig.1, where x(t) is the input signal and y(t; fc) is the filtered signal dependent on fc.

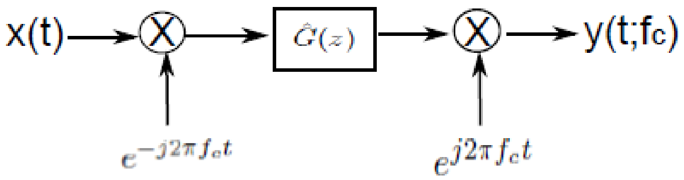


Fig Time Domain Gammatone Filtering

Pre-emphasis

It is well known that a pre-emphasis is helpful in reducing the dynamic range of spectrum and intensifying the low frequency components which usually involve more information of speech signals. Following the same idea, we implement the pre-emphasis as a 2-order low-pass filter given by:

$$H(z) = 1 + 4e^{-2\pi b/f_s} z^{-1} + e^{-2\pi b/f_s} z^{-2}$$

where b and fs are as defined in (2) and (4) respectively.

Average-based framing

The GFCC implementation proposed in [2] extracts frames from the GF outputs by down-sampling y(t; fm) to 100 Hz where fm is the central frequency of the m-th GF. This approach tends to result in high variation even with a low-pass filter applied. We use an average approach which uses a window covering K points and shifting every L points to frame y(t; fm). For the n-th frame, the average value of y(t; fm) within the window t ∈ [nL, nL + K) is computed as the m-th component:

$$\bar{y}(n;m) = \frac{1}{K}\sum_{i=0}^{K-1}\gamma(f_m)|y(nL+i; f_m)|$$

where | · | represents the magnitude of complex numbers, γ(fm) is a center frequency-dependent factor, and m is the index of the channel whose central frequency is fm.

Log-based cosine transform

With Cochleagrams, the discrete cosine transform (DCT) is requested to obtain component-uncorrelated cepstral features2. While the DCT can be applied on Cochleagrams directly (as in [2]), we place a logarithm on Cochleagrams as usually adopted in the MFCC processing. Though there isnot much theoretical advantage with the logarithm, we find that it leads to more stability in numerical processing. The following equation presents the exact cepstral form:

$$F(n,u) = (\frac{2}{M})^{0.5}\sum_{i=1}^{M}\{\frac{1}{3}log(\bar{y}(n,i))cos[\frac{\pi u}{2N}(2i-1)]\}$$

where M is the total number of channels which is 32 is our case, and u ranges from 0 to 31 accordingly. Based on the observation that most values of F(., u) are close to zero when u >= 13, we choose the first 12 components of F(n, u) as the feature vector. This results in the static GFCC features, as illustrated in Fig.

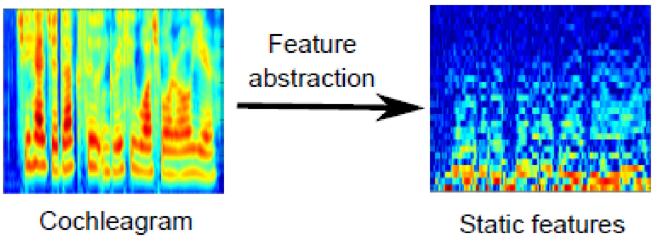


Fig COCHLEAGRAM and GFCC

Dynamic feature generation

Dynamic features are generally helpful in capturing temporal information. We produce the first and second order dynamic features as follows:

$$\Delta F(n,u) = \frac{\sum_{k=1}^{K} k(F(n+k,u) - F(n-k,u))}{2\sum_{k=1}^{K} k^2}$$

$$\Delta\Delta F(n,u) = \frac{\sum_{k=1}^{K} k(\Delta F(n+k,u) - \Delta F(n-k,u))}{2\sum_{k=1}^{K} k^2}$$

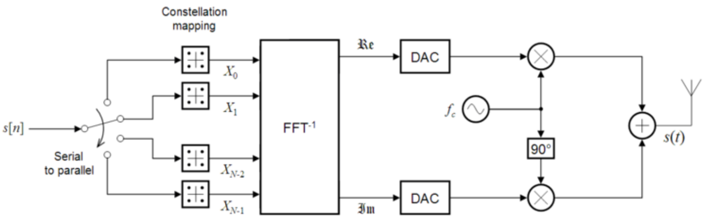
where K is set to 2, which means the length of the context window is 5.

III. Orthogonal Frequency-Division Multiplexing (OFDM)

OFDM is a frequency-division multiplexing (FDM) scheme used as a digital multi-carrier modulation method. A large number of closely spaced orthogonal sub-carrier signals are used to carry data on several parallel data streams or channels. Each sub-carrier is modulated with a conventional modulation scheme (such as quadrature amplitude modulation or phase-shift keying) at a low symbol rate, maintaining total data rates similar to conventional *single-carrier* modulation schemes in the same bandwidth.

The primary advantage of OFDM over single-carrier schemes is its ability to cope with severe channel conditions (for example, attenuation of high frequencies in a long copper wire, narrowband interference and frequency-selective fading due to multipath) without complex equalization filters. Channel equalization is simplified because OFDM may be viewed as using many slowly modulated narrowband signals rather than one rapidly modulated wideband signal. The low symbol rate makes the use of a guard interval between symbols affordable, making it possible to eliminate inter symbol interference (ISI) and utilize echoes and time-spreading (on analogue TV these are visible as ghosting and blurring, respectively) to achieve a diversity gain, i.e. a signal-to-noise ratio improvement. This mechanism also facilitates the design of single frequency networks (SFNs), where several adjacent transmitters send the same signal simultaneously at the same frequency, as the signals from multiple distant transmitters may be combined constructively, rather than interfering as would typically occur in a traditional single-carrier system.

1) Transmitter

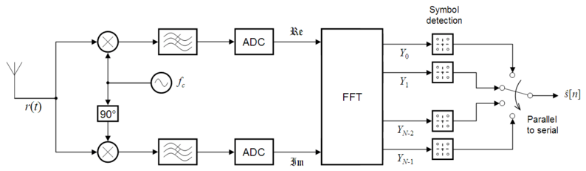


An OFDM carrier signal is the sum of a number of orthogonal sub-carriers, with baseband data on each sub-carrier being independently modulated commonly using some type of quadrature amplitude modulation (QAM) or phase-shift keying (PSK). This composite baseband signal is typically used to modulate a main RF carrier.

$s[n]$ is a serial stream of binary digits. By inverse multiplexing, these are first demultiplexed into N parallel streams, and each one mapped to a (possibly complex) symbol stream using some modulation constellation (QAM, PSK, etc.). Note that the constellations may be different, so some streams may carry a higher bit-rate than others.

An inverse FFT is computed on each set of symbols, giving a set of complex time-domain samples. These samples are then quadrature-mixed to passband in the standard way. The real and imaginary components are first converted to the analogue domain using digital-to-analogue converters (DACs); the analogue signals are then used to modulate cosine and sine waves at the carrier frequency, f_c, respectively. These signals are then summed to give the transmission signal, $s(t)$.

2) Receiver



The receiver picks up the signal $r(t)$, which is then quadrature-mixed down to baseband using cosine and sine waves at the carrier frequency. This also creates signals centered on $2f_c$, so low-pass filters are used to reject these. The baseband signals are then sampled and digitised using analog-to-digital converters (ADCs), and a forward FFT is used to convert back to the frequency domain.

This returns Nparallel streams, each of which is converted to a binary stream using an appropriate symbol detector. These streams are then re-combined into a serial stream, $\hat{s}[n]$, which is an estimate of the original binary stream at the transmitter.

If Nsub-carriers are used, and each sub-carrier is modulated using Malternative symbols, the OFDM symbol alphabet consists of M^Ncombined symbols.

The low-pass equivalent OFDM signal is expressed as:

$$\nu(t) = \sum_{k=0}^{N-1} X_k e^{j2\pi kt/T}, \quad 0 \le t < T,$$

where $\{X_k\}$are the data symbols, $N$is the number of sub-carriers, and $T$is the OFDM symbol time. The sub-carrier spacing of $\frac{1}{T}$makes them orthogonal over each symbol period; this property is expressed as:

$$\frac{1}{T}\int_0^T \left(e^{j2\pi k_1 t/T}\right)^* \left(e^{j2\pi k_2 t/T}\right) dt$$
$$= \frac{1}{T}\int_0^T e^{j2\pi (k_2 - k_1)t/T} dt = \delta_{k_1 k_2}$$

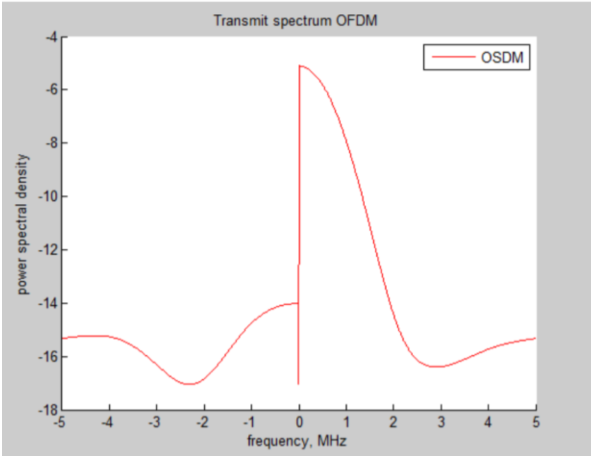
where $(\cdot)^*$denotes the complex conjucate operator and $\delta$is the Kronecker delta.

To avoid intersymbol interference in multipath fading channels, a guard interval of length $T_g$is inserted prior to the OFDM block. During this interval, a *cyclic prefix* is transmitted such that the signal in the interval $-T_g \le t < 0$ equals the signal in the interval $(T - T_g) \le t < T$. The OFDM signal with cyclic prefix is thus:

$$\nu(t) = \sum_{k=0}^{N-1} X_k e^{j2\pi kt/T}, \quad -T_g \le t < T$$

The low-pass signal above can be either real or complex-valued. Real-valued low-pass equivalent signals are typically transmitted at baseband—wireline applications such as DSL use this approach. For wireless applications, the low-pass signal is typically complex-valued; in which case, the transmitted signal is up-converted to a carrier frequency f_c. In general, the transmitted signal can be represented as:

$$s(t) = \Re\left\{\nu(t) e^{j2\pi f_c t}\right\}$$
$$= \sum_{k=0}^{N-1} |X_k| \cos\left(2\pi[f_c + k/T]t + \arg[X_k]\right)$$



The power spectrum of a time-series x(t)describes how the variance of the data x(t) is distributed over the frequency components into which x(t) may be decomposed. This distribution of the variance may be described either by a measure μor by a statistical cumulative distribution function s(f)= the power contributed by frequencies from 0 up to f. Given a band of frequencies [a, b), the amount of variance contributed to x(t) by frequencies lying within the interval [a, b) is given by S(b)-S(a). Then S is called the spectral distribution function ofx. Provided S is an absolutely continuous function, then there exists a spectral density function S'. In this case, the data or signal is said to possess an absolutely continuous spectrum. The spectral density at a frequency f gives the rate of variance contributed by frequencies in the immediate neighbor hood of f to the variance of x per unit frequency.

The nature of the spectrum of a function x gives useful information about the nature of x , for example, whether it is pperodic or not. The study of the power spectrum is a kind of generalisation of Fourier analysis and applies to functions which do not possess Fourier transforms.

IV. Multi-Microphone Sub-band Adaptive Processing

Speech enhancement combining multi-microphone methods with intermittent adaptive processing and diversity of processing in sub-bands (Fig.1) has been suggested by Toner and Campbell It allows noise features within sub-bands, such as the noise power, the correlation or coherence between signals from multiple sensors, and the behavior of the adaptive algorithm (Fig. 2), to influence the subsequent processing during the "noisy speech" period. Close spacing of the sensors reduces the required order (complexity) of the adaptive filter and thus the sub-band computational load both for adaptive and intermittent fixed processing. It also reduces the adjustment noise when using continuously adapting schemes. Sub-band operation gives faster adaption through the freedom to use different adaptive step-sizes in each band. Separate decisions can be made on the appropriate form of processing for each sub-band. The inherent parallelism of the approach allows for future parallel processor implementation.

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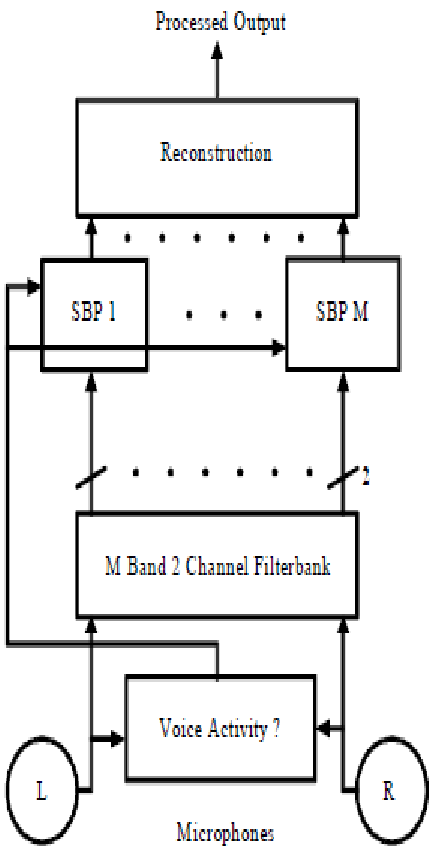


Figure 1: Diverse Sub-band adaptive processing

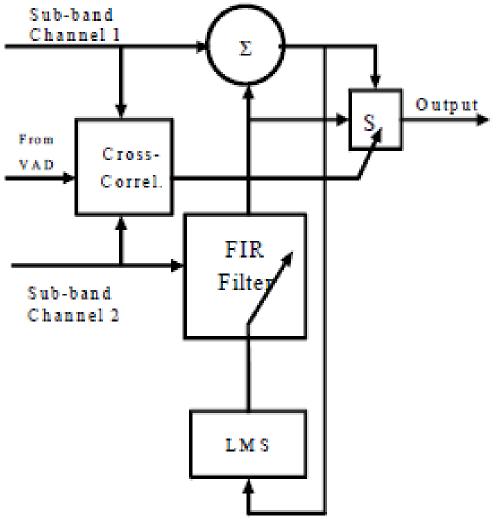


Figure 2: Sub-band processor (SBP)

V. SNR vs BER

E_b/N_0 (the **energy per bit to noise power spectral density ratio**) is an important parameter in digital communication or dat transmission. It is a normalized signal-to-noise ratio (SNR) measure, also known as the "SNR per bit". It is especially useful when comparing the bit error rate (BER) performance of different digital modulation schemes without taking bandwidth into account.

As the description implies, E_b is the signal energy associated with each user data bit; it is equal to the signal power divided by the user bit rate (*not* the channel symbol rate). If signal power is in watts and bit rate is in bits per second, E_b is in units of joules (watt-seconds). N_0 is the noise spectral density, the noise power in a 1 Hz bandwidth, measured in watts per hertz or joules. These are the same units as E_b so the ratio E_b/N_0 is dimensionless; it is frequently expressed in decibels. E_b/N_0 directly indicates the power efficiency of the system without regard to modulation type, error correction coding or signal bandwidth (including any use of spread spectrum). This also avoids any confusion as to *which* of several definitions of "bandwidth" to apply to the signal.

But when the signal bandwidth is well defined, E_b/N_0 is also equal to the signal-to-noise ratio (SNR) in that bandwidth divided by the "gross" link spectral efficiency in (bit/s)/Hz, where the bits in this context again refer to user data bits, irrespective of error correction information and modulation type.

E_b/N_0 must be used with care on interference-limited channels since additive white noise (with constant noise density N_0) is assumed, and interference is not always noise-like. In spread spectrum systems (e.g., CDMA), the interference *is* sufficiently noise-like that it can be represented as I_0 and added to the thermal noise N_0 to produce the overall ratio $E_b/(N_0+I_0)$.

B. Relation to carrier-to-noise ratio

E_b/N_0 is closely related to the carrier-to-noise ratio (CNR or C/N), i.e. the signal-to-noise ratio (SNR) of the received signal, after the receiver filter but before detection:

$$C/N = E_b/N_0 \cdot \frac{f_b}{B}$$

where

f_b is the channel data rate (net bitrate), and
B is the channel bandwidth

The equivalent expression in logarithmic form (dB):

$$\mathrm{CNR_{dB}} = 10\log_{10}(E_b/N_0) + 10\log_{10}\left(\frac{f_b}{B}\right),$$

Caution: Sometimes, the noise power is denoted by $N_0/2$ when negative frequencies and complex-valued equivalent baseband signals are considered rather than passband signals, and in that case, there will be a 3 dB difference.

C. Relation to E_s/N_0

E_b/N_0 can be seen as a normalized measure of the **energy per symbol to noise power spectral density** (E_s/N_0):

$$\frac{E_b}{N_0} = \frac{E_s}{\rho N_0},$$

where E_s is the energy per symbol in joules and ρ is the nominal spectral efficiency in (bit/s)/Hz. E_s/N_0 is also commonly used in the analysis of digital modulation schemes. The two quotients are related to each other according to the following:

$$\frac{E_s}{N_0} = \frac{E_b}{N_0}\log_2 M,$$

where M is the number of alternative modulation symbols.

Note that this is the energy per bit, not the energy per information bit.

E_s/N_0 can further be expressed as:

$$\frac{E_s}{N_0} = \frac{C}{N}\frac{B}{f_s},$$

where

C/N is the carrier-to-noise ratio or signal-to-noise ratio.
B is the channel bandwidth in hertz.
f_s is the symbol rate in baud or symbols per second.

D. Shannon limit

Main article: Shannon–Hartley theorem

The Shannon–Hartley theorem says that the limit of reliable information rate (data rate exclusive of error-correcting codes) of a channel depends on bandwidth and signal-to-noise ratio according to:

$$I < B\log_2\left(1+\frac{S}{N}\right)$$

where

I is the information rate in bits per second excluding error-correcting codes;
B is the bandwidth of the channel in hertz;
S is the total signal power (equivalent to the carrier power C); and
N is the total noise power in the bandwidth.

This equation can be used to establish a bound on E_b/N_0 for any system that achieves reliable communication, by considering a gross bit rate R equal to the net bit rate I and therefore an average energy per bit of $E_b = S/R$, with noise spectral density of $N_0 = N/B$. For this calculation, it is conventional to define a normalized rate $R_l = R/2B$, a bandwidth utilization parameter of bits per second per half hertz, or bits per dimension (a signal of bandwidth B can be encoded with $2B$ dimensions, according to the Nyquist–Shannon sampling theorem). Making appropriate substitutions, the Shannon limit is:

$$\frac{R}{B} = 2R_l < \log_2\left(1+2R_l\frac{E_b}{N_0}\right)$$

Which can be solved to get the Shannon-limit bound on E_b/N_0:

$$\frac{E_b}{N_0} > \frac{2^{2R_l}-1}{2R_l}$$

When the data rate is small compared to the bandwidth, so that R_l is near zero, the bound, sometimes called the *ultimate Shannon limit*,is:

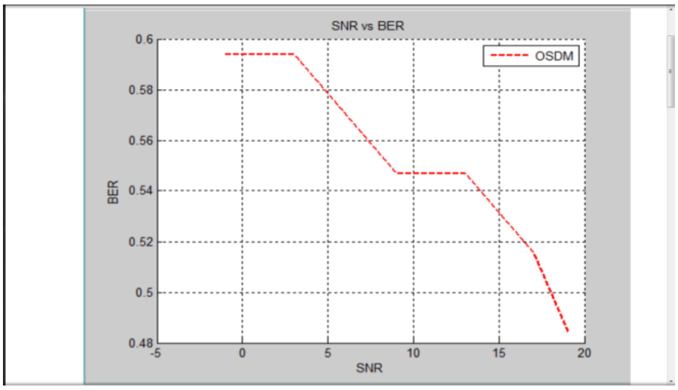
$$\frac{E_b}{N_0} > \ln(2)$$

which corresponds to –1.59 dB because:

$\ln(2) = 0.693$ and

$$10\log_{10}(0.693) = -1.59\ \mathrm{dB}$$

Note that this often-quoted limit of -1.59 dB applies *only* to the theoretical case of infinite bandwidth. The Shannon limit for finite-bandwidth signals is always higher



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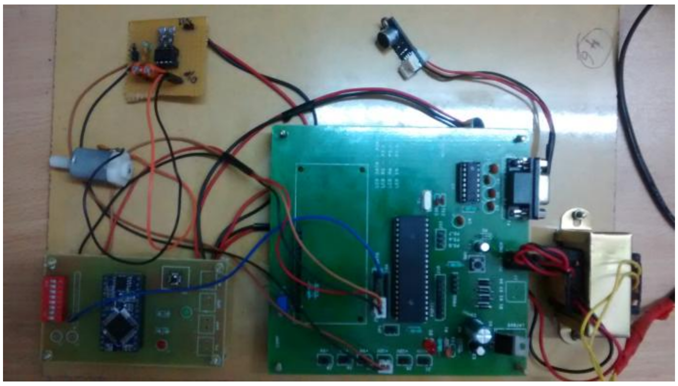

VI. Discussion

The idea of restoring normal two-tone suppression through a hearing aid has been proposed before. Turicchia and Sarpeshkar described a strategy for restoring effects of two-tone suppression in HI individuals that uses multiband compression followed by expansion [14]. The compressing-and-expanding (companding) can lead to two-tone suppression in the following manner. For a given band, a broadband filter was used for the compression stage and a narrowband filter for the expansion stage. An intense tone with a frequency outside the narrowband filter passband of the expander but within the passband of the broadband filter of the compressor results in a reduction of the level of a tone at the frequency of the expander but is then filtered out by the narrowband expander, producing two-tone suppression effects. They suggested that parameters for their system can be selected to mimic the auditory system; however, this was not demonstrated.
Subsequent evaluation of their strategy only resulted in small improvements in speech intelligibility.
Strelcyk *et al*. described an approach to restore loudness growth—restoration of normal loudness summation and differential loudness. Loudness summation is a phenomenon where a broadband sound is perceived as being louder than a narrowband sound when the two sounds have identical sound pressure level. Loudness summation is achieved in the system of Strelcyk *et al*. by widening the bandwidth of channel filters as level increases.

VII. CONCLUSION

Finally this project has been completed successfully to give for deaf peoples as their hearing aid machine. the aim of making speech more intelligible, and to correct impaired hearing as measured by audiometry has done successfully. This project has various advantages of immensely reducing noise. Reducing the errors. Reduces the cost of hearing aid machine. Reducing the cost for surgery for hearing aid. In case patient didn't hear properly, it helps to listen again by recording through various channels.

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