Air-interface of high data rate wireless access

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Abstract. High data rate, reliable wireless access has been the focus in the design of the future wireless broadband communications. The wireless access is increasingly becoming data centric, and has to meet the requirements of a wide array of traffic ranging from voice, multimedia, and other real-time applications.

The technology that work behind the scenes to make all this happen is constantly improving; improvements in the physical layer including new coding and modulation schemes that use spatial multiplexing (transmitting and receiving signals through multiple antennas), and improvements in medium access strategy (currently physical agnostic) is analyzed.

The role of MIMO MAC and inter-layer interactions in wireless open system interface are identified, to lay the framework for a cross-layer design, of a physical aware medium access layer.

Key words. Multiple Input Multiple Output(MIMO) wireless communication, Space-Time coding(STC), Orthogonal Frequency Division Multiplexing(OFDM), Wireless Local Area Network(WLAN), Cross-Layer design

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Contents

1 Introduction

The increasing popularity of wireless hot-spots and WLAN's can be attributed to features like mobility, high data rates and quality of service (QoS) that attracts the end user as much as the ease of deployment and scalability advantages that attracts the service providers. The technology alternatives that work behind the scenes to make it all possible are constantly improving. The improvements are across the layers of the open system interface (OSI), and work in tandem to ultimately realize the high data rate wireless access networks.

Improvements in the air interface or the physical layer, permit improved options like smart-multiple antennas, and a choice from the myriad coding and modulation schemes. Each scheme professes an edge over the others, to varying degrees, in terms of its effectiveness with respect to spectral usage, or its immunity to interference. As we shall see in section 4.2.2 Multiple input and Multiple output (MIMO) communications using space-time coding, and multi-carrier orthogonal frequency division multiplexing(OFDM) - section 4.2.3, are at the forefront in air-interface design for high data rate wireless access.

Improvements in medium access(MAC) strategies, have to take into account the performance limiting challenges of the wireless channel(time varying delay and Bit Error Rate(BER)). Additionally it would have to support a variety of services ranging from voice, short text, to video and audio streaming with their specified Quality of Service(QoS), in terms of delay, throughput and BER. A cross-layer design, for instance a physical(PHY) aware medium access(MAC) and data link layer can be used to better utilize radio resources(power and bandwidth), combat network congestion and facilitate handoff procedures.

Higher layers are likewise not left out in the changes designed to provide mobility and improved end to end throughput of the next generation wireless networks. Improved routing strategies can be developed for a multi-hop network with the knowledge of channel state information viz., received signal and interference level, number of active users etc. Thus new routes discovered based on additional link state information, can be more effective over those discovered merely using hop-distance information.

This report is organized into three major sections. The first section delves into the

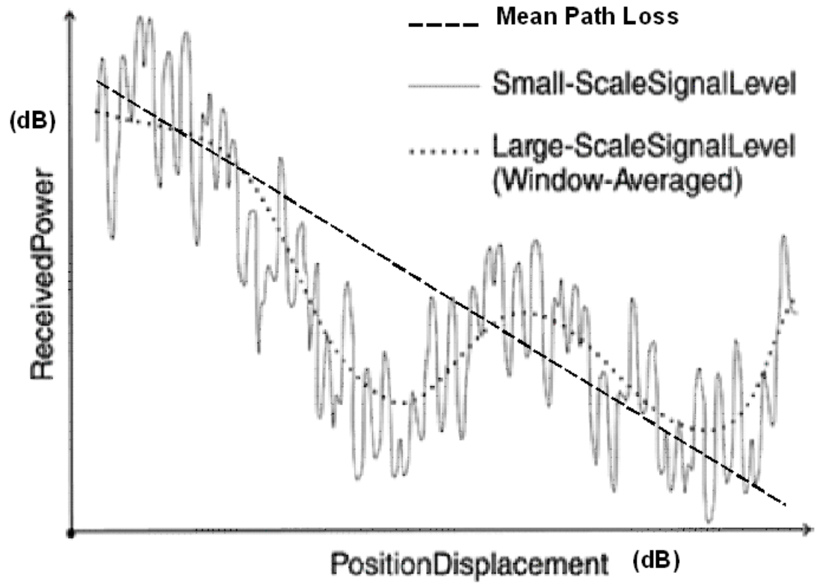

Figure 1: Small-scale and Large-scale spatial channel fading

characterization of the wireless channel in terms of mathematical models. The second part is about strategies to overcome the channel impairment, discussed in the first section. Emphasis is placed on space-time coding for MIMO communications as the strategy of preference, while mentioning the other traditional and newer alternatives, to combat the foes of the wireless channel. The final section deals with issues involved in the design of PHY aware MAC to meet the varied QoS requirements(in terms delay, throughput and BER), that the next generation wireless networks would have to support.

2 Fading in communication channels

Fading is the term used to describe the rapid fluctuations in the amplitude of the received radio signal over a short period of time. Fading is a common phenomenon in Mobile Communication Channels, and it is caused due to the constructive and destructive interference between two or more versions of the transmitted signals which arrive at the receiver at slightly different times. The resultant received signal can vary widely in *amplitude* and *phase*, depending on various factors such as the intensity, relative propagation time of the waves, bandwidth of the transmitted signal etc. The figure 1 summarizes the effects of fading, we will now look into mathematical models to characterize small-scale fluctuations in *amplitude* and *phase* of the received signal. See Annex A for the large-scale signal deterioration models.

2.1 Impulse response of channel, selectivity and distortion

The channel induced attenuation in amplitude, phase distortion and propagation delay, can be mathematically modeled as a time varying linear system, whose impulse response is:

$$\hat{h}(t) = a(t)e^{j\theta(t)}\delta(t-\tau) \tag{2.1}$$

The above expression for impulse response of a channel is just a scaled *impulse* in time which corresponds to channel loss/gain a(t) and channel induced propagation delay τ.

The output from this linear system model is the attenuated shifted response that the receiver sees, a result of convolution with the impulse response.

$$\begin{aligned} \hat{y}(t) &= \hat{h}(t) * \hat{x}(t), \qquad (2.2) \\ \hat{x}(t) &\text{ is the transmitted signal} \\ \hat{y}(t) &\text{ is the received signal} \end{aligned}$$

The multiple paths in a wireless system can be modeled as an aggregation of impulses due to the K-paths in the channel.

$$\hat{h}(t) = \sum_{i=1}^{K} a_i(t)e^{j\theta_i(t)}\delta(t-\tau_i) \tag{2.3}$$

$$\begin{aligned} \text{where } a_i(t) &: \text{ is the channel loss/gain of path i} \\ \text{and } \tau_i &: \text{ is the propagation delay of path i} \end{aligned}$$

The characterization of a channel into *flat* or *frequency-selective* is based on whether the the multi-path is resolvable or unresolvable, if T_m is the maximum excess delay amongst the multiple paths, and T_s is the transmission time of the symbol.

- A channel is considered **frequency-selective** when $T_m > T_s$, i.e the received multi-path components extend beyond the symbol duration and cause inter symbol interference(ISI).

- While the channel is considered to be **frequency-flat** when $T_m < T_s$, i.e the multi-path components arrive within the symbol duration and hence there is no channel induced ISI. There may sill be a loss in SNR due to the destructive interference of the multi-path symbols.

Thus while frequency-selective fading induces ISI distortion, flat fading may result in loss of SNR of the received signal [21].

2.2 Mathematical Model of Rayleigh Fading

Consider a transmitted signal $s(t) = Acos2\pi f_c t$ through a fading channel. The received signal can be expressed as (ignoring the effects of noise):

$$y(t) = A\sum_{i=1}^{N} a_i \cos(2\pi f_c t + \theta_i) \tag{2.4}$$

where
* a_i is the attenuation of the i^{th} multi path component
* θ_i is the phase-shift of the i^{th} multi path component
* a_i and θ_i are random variables

Thus expression 2.4 is re-written as:

$$y(t) = A\left\{\left(\sum_{i=1}^{N} a_i \cos(\theta_i)\right)\cos(2\pi f_c t) - \left(\sum_{i=1}^{N} a_i \sin(\theta_i)\right)\sin(2\pi f_c t)\right\} \tag{2.5}$$

We introduce two random processes $X_1(t)$ and $X_2(t)$, such that the above equation becomes:

$$y(t) = A\left\{X_1(t)\cos(2\pi f_c t) - X_2(t)\sin(2\pi f_c t)\right\} \tag{2.6}$$

If the value of N is large (i.e, a large number of scattered waves are present), invoking the Central- Limit Theorem, we get approximate $X_1(t)$ and $X_2(t)$ to be Gaussian random variables with zero-mean and variance σ^2. The expression 2.6 can be re-written as:

$$y(t) = AR(t)\cos(2\pi f_c t + \theta(t)) \tag{2.7}$$

where, **The amplitude of the received waveform** $R(t)$ is given by:

$$R(t) = \sqrt{X_1(t)^2 + X_2(t)^2} \tag{2.8}$$

Since the processes $X_1(t)$ and $X_2(t)$ are Gaussian, it can be shown that $R(t)$ has a *Rayleigh Distribution* with a probability density function(pdf) given by: see p181,p190 Papoulis [17]

$$f_R(r) = \frac{r}{2\sigma^2} \exp\left\{\frac{-r^2}{2\sigma^2}\right\}, \quad r > 0 \tag{2.9}$$

and, **The phase of the received waveform** $\theta(t)$ is given by:

$$\theta(t) = \tan^{-1}\left(\frac{X_2(t)}{X_1(t)}\right) \tag{2.10}$$

Since the processes $X_1(t)$ and $X_2(t)$ are Gaussian, it can be shown that $\theta(t)$ has a *Uniform Distribution* with a probability distribution function(pdf) given by:

$$f_\theta(\theta) = \frac{1}{2\pi}, \quad -\pi \leq \theta \leq \pi \tag{2.11}$$

The distortion in the phase can be easily overcome if differential modulation in employed. It is the amplitude distortion $R(t)$ that severely degrades performance of digital communication systems over fading channels. It is usually reasonable to assume that the fading stays essentially constant for at least one signaling interval.

2.3 Instantaneous SNR per bit and Probability of Error

Probability of Error in AWGN: The probability of error in symbols transmitted can be determined from the minimum distance of the modulation used and the number of nearest neighbors in the signal constellation, for the AWGN channel the probability of error(P_e) for *Binary PSK* is given by:

$$P_e(\gamma_b) = 2Q\left(\sqrt{2\gamma_b}\right) \text{ where } \gamma_b = \frac{E_b}{N_o} \text{ for AWGN} \tag{2.12}$$

Probability of Error in AWGN+Rayleigh Fading: Since it is assumed that the fading stays constant in a signaling interval, we can represent the fading phenomenon using a random variable R. Since only amplitude distortion is considered, the instantaneous SNR per bit γ_b is now a random variable given by:

$$\gamma_b = R^2 \frac{E_b}{N_o} \tag{2.13}$$

Since R is *Rayleigh distributed*, R^2 and in turn γ_b has a *chi-squared distribution* with two degrees of freedom, which is an *exponential distribution* whose pdf is given by:

$$f_{\gamma_b}(\gamma_b) = \frac{1}{\bar{\gamma_b}} \exp\left\{\frac{-\gamma_b}{\bar{\gamma_b}}\right\}, \quad \gamma_b \geq 0 \tag{2.14}$$

Where $\bar{\gamma_b}$ is the average SNR per bit, and is given by:

$$\bar{\gamma_b} = \frac{E_b}{N_o} E[R^2] \tag{2.15}$$

So, the average probability of error, given the pdf of the random variable γ_b can be found directly as:

$$P_e = \int_0^{\infty} P_e(\gamma_b) f_{\gamma_b}(\gamma_b) d\gamma_b \tag{2.16}$$

for *Binary PSK* the above expression reduces to: from Rappaport [19],

$$P_e = \frac{1}{2}\left(1 - \sqrt{\frac{\bar{\gamma_b}}{1+\bar{\gamma_b}}}\right), \quad for\ \bar{\gamma_b} \gg 1,$$

where $\bar{\gamma_b}$ is defined in 2.15, P_e reduces to: (2.17)

$$P_e \approx \frac{1}{4\bar{\gamma_b}} \tag{2.18}$$

The destructive effect of fading is evident from the above equation. The value of P_e decreases linearly with increasing SNR per bit, rather than either a Q-function type or an exponential decrease. This will severely degrade the power-efficiency of a digital communication system if preventive measures are not taken to mitigate this problem.

The figure 2 illustrates this steep deterioration of the channel due to *Rayleigh Fading* when compared with additive white Gaussian noise(AWGN) channel. A 3dB SNR is adequate to support Bit Error Rates(BER) of 0.02 when there is no channel fading in AWGN channel, in contrast a 10dB SNR is required to support the same 0.02 BER using *Binary PSK* modulation scheme - which is a loss of 7dB due to fading.

Note: Though we have nice closed form expressions for BER or P_e for BPSK, we can arrive at similar performance curves using *Monte-Carlo* simulation by generating Raleigh and Gaussian random variates, and arrive at the statistical averages, by averaging over a large number of trials (restricted by the random number generator's capability to generate iid variates). Figure 2 performance curves, comparing AWGN and Raleigh Fading were obtained using the above mentioned closed form expressions, while figure 12 performance

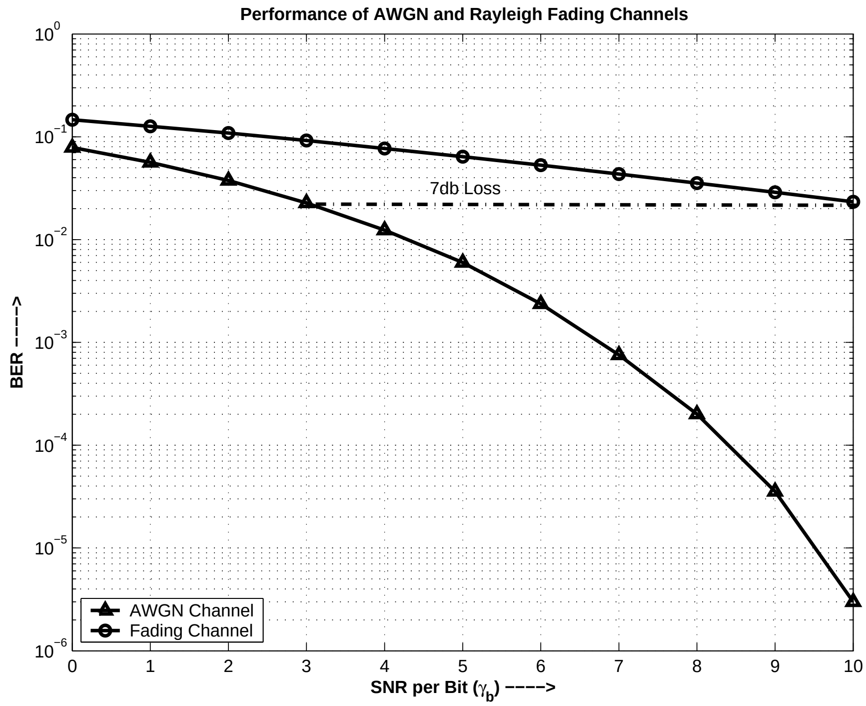

Figure 2: AWGN and Rayleigh Fading Performance

curves, for improvements due to transmit and receiver diversity, were obtained using *Monte-Carlo* simulation.

3 Mitigation of channel impairment

The previous sections detailed the physics of radio propagation and the theory behind channel fading and inter symbol interference(ISI) distortion. Additionally, the non-engineerable and non-clonable radio space is also affected by interference from multiple users sharing the same medium. Listed below are strategies summarized by [22] to alleviate the hostile radio environment, by carefully designed coding and modulation schemes.

3.1 Combat frequency-selective fading distortion

- Adaptive equalization, decision feedback Viterbi equalizer
- Spread Spectrum(SS) -Direct Sequence(DS) or Frequency Hopping(FH)
- Orthogonal Frequency Division Multiplexing(OFDM)

- Pilot Signal (Channel Estimation)
- Robust Modulation, Coding and Interleaving

3.2 Combat loss in SNR due to frequency-flat fading

Traditionally, some type of diversity has been employed to get additional uncorrelated estimates of the signal to combat the deterioration in SNR due to fading. The different types of diversity are:

- Temporal-diversity, redundancy in time e.g. coding, interleaving.
- Frequency redundancy or bandwidth expansion e.g. Spread Spectrum, DS and FH. Both time and frequency diversity trade bandwidth for performance viz. bit error-rate.
- Spatial diversity uses multiple transmit and/or receive antenna, to improve performance without loss in bandwidth.
- Polarization of E-field is also used as additional dimensions to improve diversity. It is either horizontal or vertical depending on the relative orientation of the field to the direction of propagation.

Though traditional schemes like adaptive decision feedback equalization and spread spectrum, and other new coding and modulation strategies like OFDM and Time Reversed STBC counter distortion and interference, emphasis is placed on orthogonal space-time coding proposed by Alamouti in this report.

4 Spatial diversity

The basic idea in using multiple antennas in the transmission and reception of uncorrelated copies of the signal is the increased probability in MIMO systems that at least one of them is of good quality, despite deep-fade in the channel.

In order to tap the above mentioned processing gain in MIMO systems, an optimal signal combining strategy at the receiver, and a handcrafted orthogonal signal design at the transmitter should be employed. Orthogonal Space-Time coding proposed by Alamouti and further generalized by Tarokh et.al [12] is needed to prevent multiple copies of the signal, arriving at the receiver, from interfering with one another.

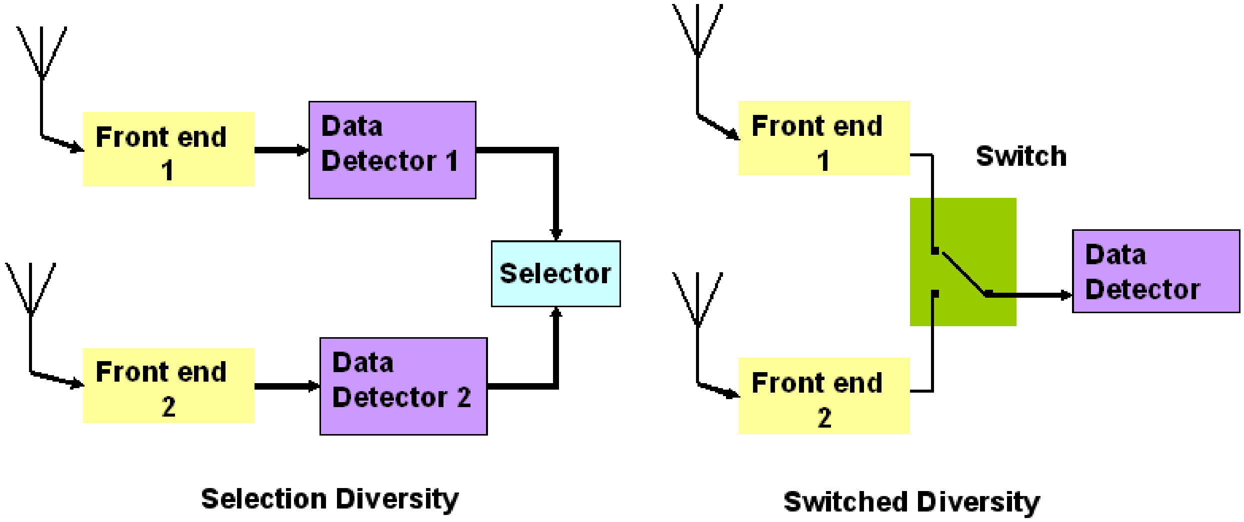

Figure 3: Selection and Switched Diversity

4.1 Receiver Diversity

The first system to incorporate spatial diversity was at the access point or the base station. The access point had no weight, size or power limitations (unlike the mobile users), to implement the multi-antenna systems. The economics of sharing the cost of hardware upgrade amongst the multiple users also motivated the development of the multiple antenna base stations. The receiver diversity at the access point leads to an increase in the up-link capacity(from mobile user to the access point) and was incorporated into GSM and IS-136 standards.

4.1.1 Selection and Switched Diversity

The switched antenna diversity was amongst the earliest diversity techniques used at the access point, the antenna's are separated by distances greater than $\frac{\lambda}{2}$ to ensure a certain degree of independence in their reception. When the SNR falls beneath a threshold for useful detection, the RF path was switched to the alternate path to receive the transmission despite the signal fade in the earlier path, see figure 3.

The hard switch when replaced by a soft decision; results in the selection combining receiver. It works on similar lines, and utilizes the fact that not all of the independent paths will fade simultaneously. Thus when the signal SNR falls beneath the threshold for correct decoding; the alternate antennas are scanned to choose the one with highest SNR.

4.1.2 Linear Combination Receiver, Equal Gain and Maximal Ratio

Maximal-ratio combining(MRC), and equal gain combining are linear combining strategies at the receiver. Assuming a flat fading channel that precludes ISI, a linear combination strategy involves co-phasing the received signals by appropriate choice of weighting coefficients that would ultimately maximize the SNR of reception, the MMSE receiver also optimizes the weighting coefficients to this end. See figure 4 for the structure of the maximal combining receiver.

The received signals are:

$$y_0[k] = h_0 x[k] + n_0[k] \tag{4.1}$$
$$y_1[k] = h_1 x[k] + n_1[k] \tag{4.2}$$

where $x[k] = s_n$ is the transmitted signal during symbol interval k, and $h_{0,1} = \alpha_{0,1} e^{j\phi_{0,1}}$ is the quasi static channel response that is assumed to remain unchanged during a symbol interval. $n_{0,1}[k]$ are the additive noise terms.

The MRC chooses weighting coeffs a_i to combine the signals received in all the branches:

$$\hat{y}[k] = \sum_{i=0}^{N_r-1} a_i y_i[k] \tag{4.3}$$
$$= x[k] \sum_{i=0}^{N_r-1} a_i \alpha_i e^{j\phi_i} + \sum_{i=0}^{N_r-1} a_i n_i[k] \tag{4.4}$$

substituting $x[k] = s_n$ in 4.4, the SNR of the MRC, γ_{mrc} is:

$$\gamma_{mrc} = \frac{\mathbf{E}\left[|s_n \sum_{i=0}^{N_r-1} a_i \alpha_i e^{j\phi_i}|^2\right]}{\mathbf{E}\left[|\sum_{i=0}^{N_r-1} a_i n_i[k]|^2\right]} \tag{4.5}$$

using Cauchy-Schwartz the upper bound on the SNR obtained in MRC is found by reducing 4.5 to:

$$\gamma_{mrc} \leq \left(\frac{E_x}{N_0}\right) \sum_{i=0}^{N_r-1} \alpha_i^2, \text{ when } a_i \propto \alpha_i e^{-j\phi_i} \tag{4.6}$$
$$\gamma_{mrc} = \left(\frac{E_x}{N_0}\right) \sum_{i=0}^{N_r-1} \alpha_i^2, \text{ i.e. meets the upper bound} \tag{4.7}$$

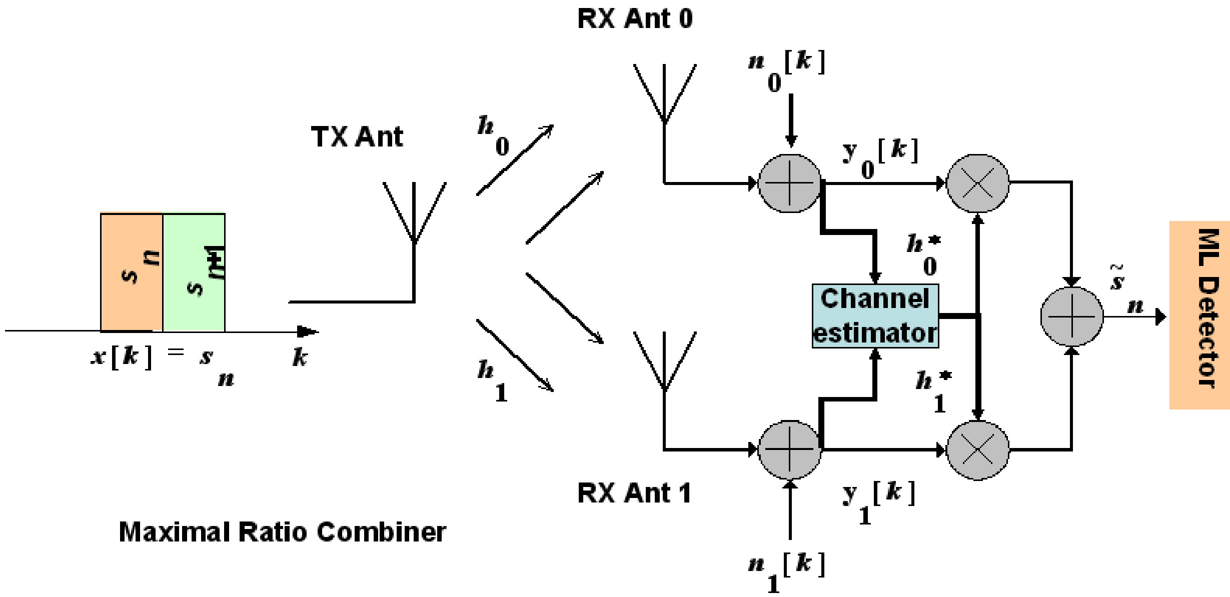

Figure 4: Maximal Ratio Combining Receiver

The MRC produces an estimate of the transmitted wave form $\hat{s}_n$:

$$\hat{s}_n = h_0^* y_0[k] + h_1^* y_1[k] \tag{4.8}$$

$$= (\alpha_0^2 + \alpha_1^2)s_n + h_0^* n_0[k] + h_1^* n_1[k] \tag{4.9}$$

Performance of MRC receiver: A closed form expression of the performance of MRC reciever when using *BPSK* modulation is provided in [26], which are obtained in similar lines as the probability of error calculation for Rayleigh fading in section 2.3, approximate P_e for $\bar{\gamma}_b \gg 1$, where nR is the number of receivers:

$$P_e \approx \left(\frac{1}{4\bar{\gamma}_b}\right)^{nR} \binom{2nR-1}{nR} \tag{4.10}$$

Note: The complexity of the maximal ratio combiner in estimating the channel gain/loss and conjugating it to obtain the weighting coefficients a_i is avoided in the equal gain combiner, where $a_i = e^{-j\phi_i}$ is chosen to be equal for all the branches. The simplicity of this receiver is preferred over the slight loss in performance due to sub optimality.

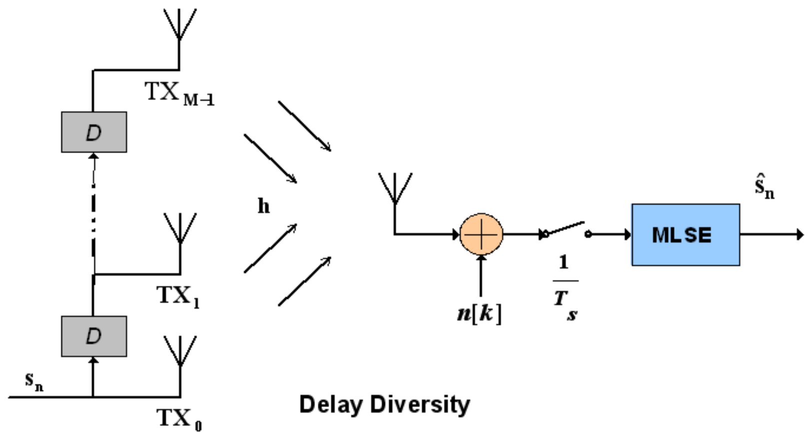

Figure 5: Transmit Delay Diversity Scheme

4.2 Transmitter Diversity

The 3G Partnership Project (3GPP) is looking into the open loop and closed loop variants of transmitter diversity to increase the down-link capacity (from the access point to the mobile user)[7]. Transmitter diversity as mentioned earlier has to have specific orthogonality requirements. In the open loop design, the transmitter is not aware of the channel state, while in the closed loop design the receiver reports the Rayleigh attenuation coefficients and the delays associated with the multiple paths, by decoding special pilot signals this process is called *Channel Estimation*.

4.2.1 Delay Diversity

Delay diversity transmit scheme first proposed by Wittneben and simultaneously by Winters, the scheme involves the creation of a multi-path environment intentionally. It incorporates delay elements at the multiple Tx antenna to create the multi-path environment, see figure 5. The processing gain achieved by the delayed diversity is realized by using an equalizer at the receiver.

4.2.2 Alamouti Orthogonal Space-Time Block Coding

Space-time codes(STC), are orthogonal signal design technique that utilize the added spatial dimension of multi-antenna systems in coding. There are two classes of STC viz. state-time block code(STBC) first proposed by Alamouti and the state-time trellis codes(STTC) proposed by Tarokh e.t.al. The simplicity and linear processing at the receiver when using

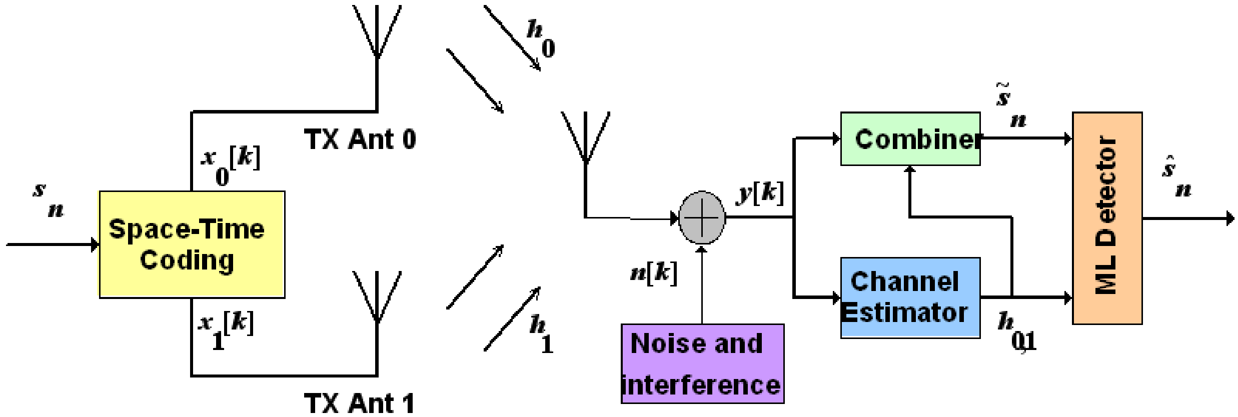

Figure 6: Alamouti Transmitter Diversity

STBC is preferred over the exponentially increasing complexity of the STTC, which involves convolution encoding and Viterbi decoding. We will only look into STBC in this report .

2Tx-1Rx Alamouti encoder The figure 6 gives the structure of STBC with two transmit antennas. The essence of Space-Time coding for trasmit diversity lies in hand-crafting signals which are orthogonal in the chosen dimensions, such that they do not interfere with one other, when they arrive simultaneously at the receiver.

The transmission of two symbols occurs simultaneously through the two Tx antenna's over two time intervals(hence full-rate). The first(even) interval two of the incoming symbols is transmitted, the following(odd) interval complex conjugates of the two symbols are transmitted as shown in figure 7:

Let S_n be the sequence of symbols to be transmitted. Let $x_0[k], x_1[k]$ be the signals transmitted by antenna zero and antenna one respectively at time instant k. Thus the transmissions at k and k+1 are:

$$x_0[k] = s_n, \; x_1[k] = s_{n+1} \tag{4.11}$$

$$x_0[k+1] = -s^*_{n+1}, \; x_1[k+1] = s^*_n \tag{4.12}$$

Channel is assumed to fade slowly, i.e. remains constant over two symbol durations:

$$h_0[k+1] = h_0[k] = h_0 = \alpha_0 e^{j\phi_0} \tag{4.13}$$

$$h_1[k+1] = h_1[k] = h_1 = \alpha_1 e^{j\phi_1} \tag{4.14}$$

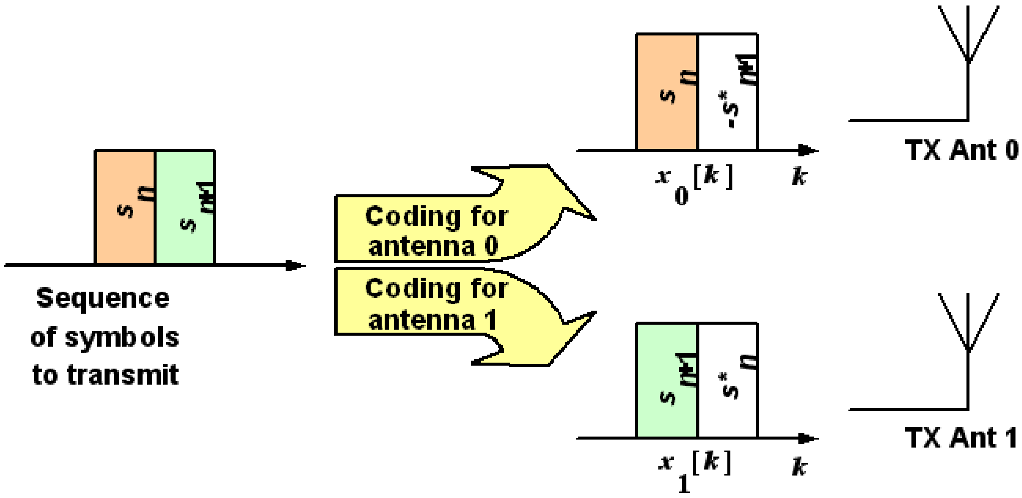

Figure 7: Transmitter Diversity,Space-Time Coding

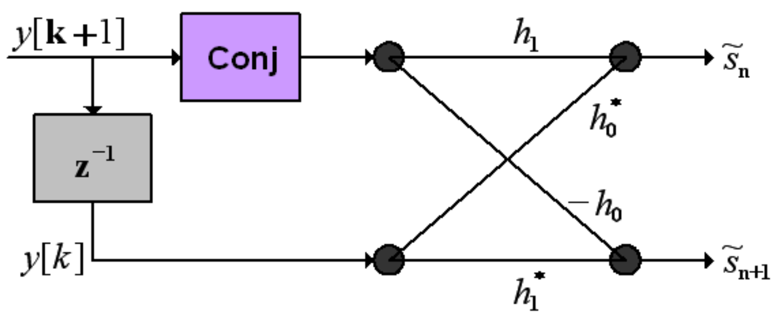

Figure 8: Alamouti-Receiver-Combiner

thus the received signal is:

$$y[k] = h_0 x_0[k] + h_1 x_1[k] + n[k] \tag{4.15}$$

2Tx-1Rx Alamouti decoder The decoder utilizes the specific transmission sequence to decode the orthogonal signals iteratively. ML decoding in the AWGN case with no ISI reduces to minimum distance decoding. Since the receiver estimates the channel from the pilot transmissions, the received estimates can be divided by the channel gain to obtain the transmitted signal estimate, when the receiver combines the received signals as follows:

$$\hat{s}_n = h_0^* y[k] + h_1 y^*[k+1] \tag{4.16}$$

$$\hat{s}_{n+1} = h_1^* y[k] - h_0 y^*[k+1], \quad \text{reduces to:} \tag{4.17}$$

$$\hat{s}_n = (\alpha_0^2 + \alpha_1^2) s_n + h_0^* n[k] + h_1 n^*[k+1] \tag{4.18}$$

$$\hat{s}_{n+1} = (\alpha_0^2 + \alpha_1^2) s_{n+1} + h_1^* n[k] - h_0 n^*[k+1] \tag{4.19}$$

A general technique for constructing STBC for any number of transmitters was formulated by Tarokh et.al. These codes still retain the ML decoding and it was shown that it is possible to construct full rate STBC for real signal constellations. For example the STBC mapper for four transmit antenna with rate 4/8 is given below:

$$\begin{bmatrix} s_1 \\ s_2 \\ s_3 \\ s_4 \end{bmatrix} \rightarrow \begin{bmatrix} s_1 & -s_2 & -s_3 & -s_4 & s_1^* & -s_2^* & -s_3^* & -s_4^* \\ s_2 & s_1 & s_4 & -s_3 & s_2^* & s_1^* & s_4^* & -s_3^* \\ s_3 & -s_4 & s_1 & s_2 & s_3^* & -s_4^* & s_1^* & s_2^* \\ s_4 & s_3 & -s_2 & s_1 & s_4^* & s_3^* & -s_2^* & s_1^* \end{bmatrix} \tag{4.20}$$

4.2.3 Space-Frequency Coding: Orthogonal Frequency Division Multiplexing(OFDM)

The *flat-frequency* channel model assumed by Alamouti space-time coding fails in the case of wideband transmissions. At high data rates the symbol duration is much smaller than the *excess delay-spread* due to fading. Space-Frequency coding(SFC) is a strategy that uses the transmitter diversity technique proposed by Alamouti for *freq-selective* fading channels. SFC uses the adjacent subbands of the channel in an OFDM structure instead of adjacent times in the STBC proposed by Baraniuk e.t.al [1]. See figure 9 for the structure of SF-OFDM scheme.

OFDM is a multi-carrier communication technique wherein, a single stream of input data is split into multiple streams using a serial to parallel converter at the transmitter. Each substream is modulated onto *orthogonal subbands* using Inverse Fast Fourier Transform(IFFT). The orthogonal waveforms despite its overlap in the frequency domain can be successfully demodulated at the receiver using the inverse transformation (FFT) provided there is no Inter symbol Interference(ISI) or Inter subband Interference. Guard bands and cyclic prefixes are introduced in the modulation scheme to prevent Inter symbol Interference(ISI) and inter subband interference[9]. See figure 11 to see how the guard interval precludes inter symbol interference(ISI) while cyclically extending the data into the guard intervals prevents

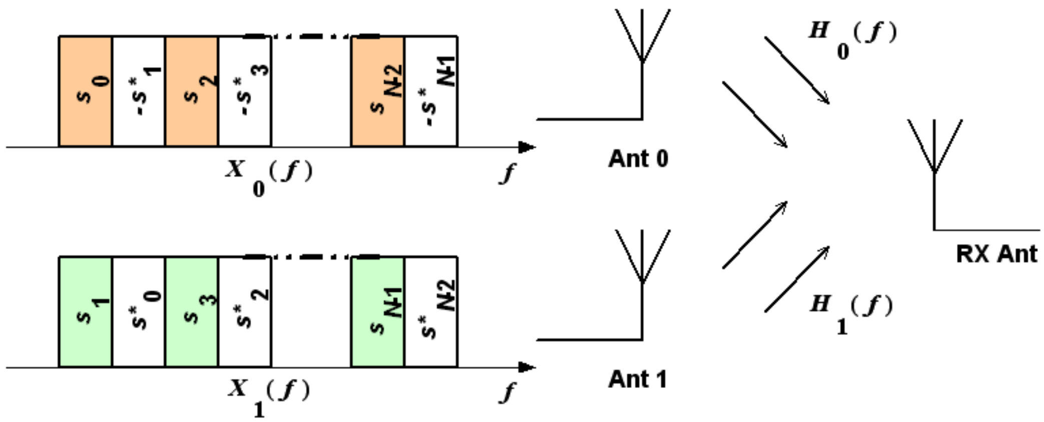

Figure 9: Transmitter Diversity for Frequency Selective Channels using SF-OFDM

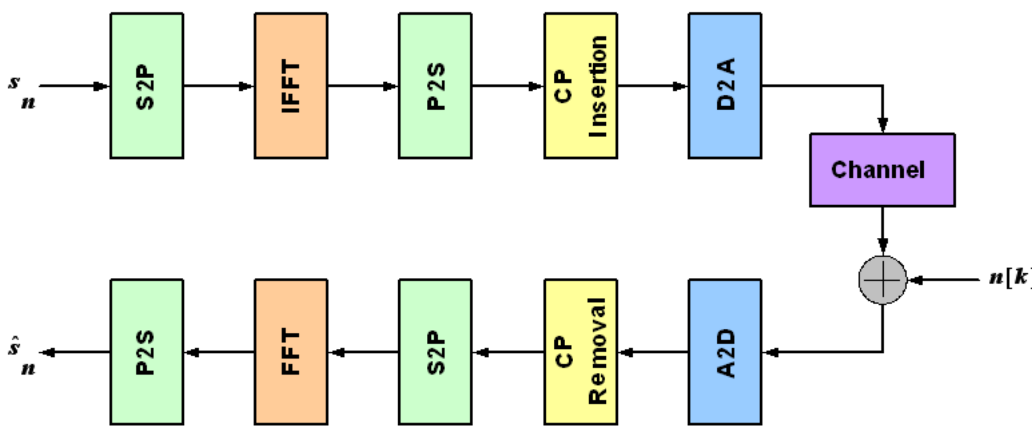

Figure 10: OFDM Transmission Reception

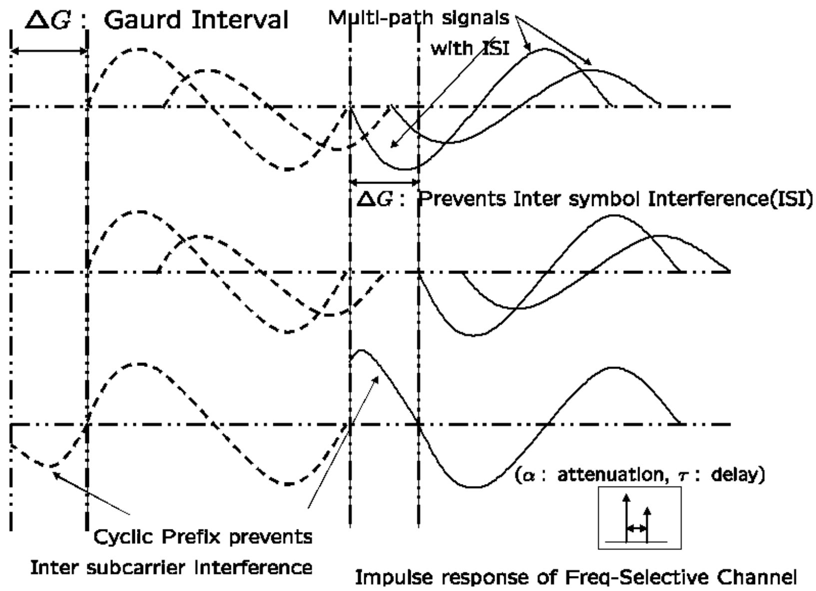

Figure 11: Guard Interval precludes ISI while Cyclic Prefix Prevents Inter Subband Interference

spurious high frequency components (also called inter subband interference, due to sharp discontinuities) from being introduced in the signal. The cyclic prefix that was added in the transmission is removed, before inverse orthogonal transformation (FFT) and ultimate parallel to serial conversion at the receiver, to recover the original data stream. See figure 11 for the overall structure of the OFDM transmission and reception. Further details on the space-frequency OFDM, or on other techniques like time-reversed STBC for *freq-selective* channels is beyond the scope of this report.

4.3 Advantages of Spatial Diversity

The processing gain achieved in MIMO systems, is realized into the following reliability and capacity gains:

1. Increased reliability of the wireless link. The array gain of MIMO systems and SNR increase due to it, increases the robustness of the wireless link by decreasing the outage probability and the associated Bit Error Rate(BER) in the transmission. The outage probability has been shown by Laneman[11] to be proportional to $\frac{1}{SNR^N}$ for N=2, where N is the diversity order of the system.

2. Increased capacity/range for given transmitter power and bandwidth constraint. The spatial diversity utilizes the multiple data pipes and thereby increases the spectral

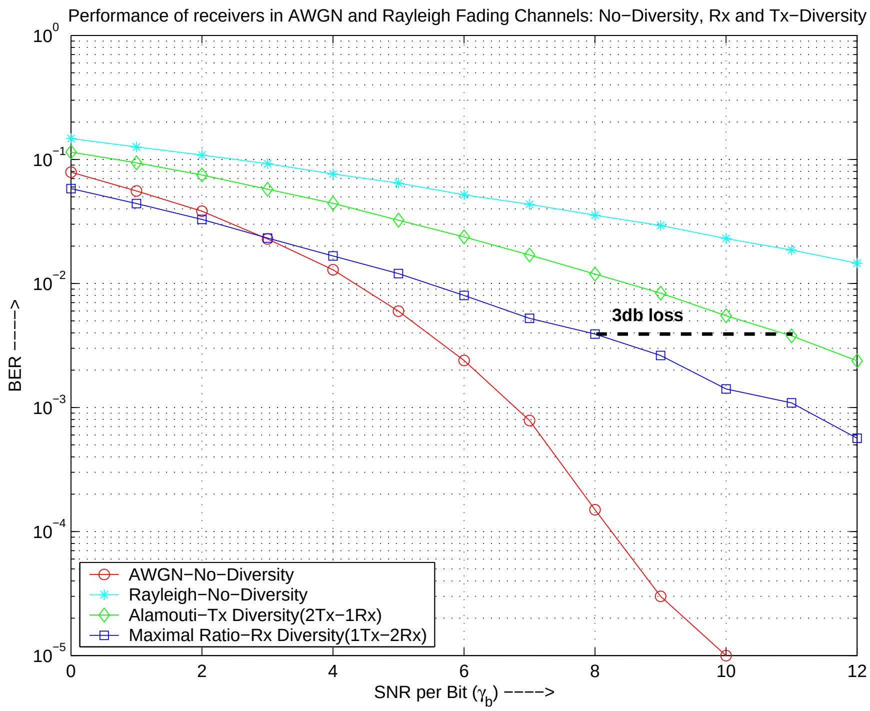

Figure 12: BER improvement with diversity

efficiency of the system, without increasing the power and bandwidth requirements.

4.3.1 Increased Reliability a MIMO Link

The increased reliability of a MIMO link can be seen in the figure 12. The figure shows that the maximal-ratio combiner performs 3dB better over the transmit diversity scheme with an equivalent *diversity-order* (N) of two. The loss in performance of the transmit diversity can be attributed to the fact that, only half of the power is radiated from each transmit antenna(when N = 2), this normalization is done to keep the received power constant, and independent of the number of transmitters.

4.3.2 Increased Range of a MIMO link

Smart antennas provide enhanced coverage through range extension, hole filling and better building penetration [14]. Given the same transmitter power output at the base station and subscriber unit, smart antennas can increase range by increasing the gain of the base station antenna. From the path loss models discussed in Annex A, the uplink power received from a mobile unit at a base station is:

$$P_r = P_t + G_s + G_b - PL \tag{4.21}$$

where P_r is the power received at the base station, P_t is the power transmitted by the subscriber, G_s is the gain of the subscriber unit antenna and G_b is the gain of the base station antenna. On the uplink, if a certain power $P_{r,min}$, is required at the base station, by increasing the gain of the base station - G_b, the link can tolerate greater path loss PL. Using expressions for path loss from AnnexA:

$$PL(d) = PL(d_0) + 10nlog\left(\frac{d}{d_0}\right) + X_\sigma \tag{4.22}$$

Thus by increasing the tolerable path loss, one can increase the reception range, d of the base station.

4.3.3 Increased Capacity of MIMO systems

The capacity bound on communication systems developed my Shannon as $C = Blog_2(1 + SNR)$, can be extended to analyze MIMO systems. The Shannon's transmission rate bound C, is limited by the bandwidth B of the communication system, and the transmission signal to noise ratio(SNR).

The characterization of the multi-path channel as a time varying linear system with a complex impulse response h is used to model the received observations(refer section 2.1 on *flat-frequency* scaled impulse model):

$$y = h^H s + n, \quad \text{s : signal, n : AWGN} \tag{4.23}$$

Capacity of a SISO Link: The received signal strength from equation 2.3 is thus hs and the power would be $h^2E[s^2]$, from which the capacity of a **SISO** system(single transmit and receive antenna) is found as:

$$C = \log_2(1 + \rho|h|^2) \quad \text{(bits/s/Hz)} \tag{4.24}$$

where ρ is the SNR at the receive antenna, and h is the complex gain of the single channel.

Capacity of a SIMO Link: The channel model 4.23 can also be used to model multiple paths of SIMO or receiver diversity system. The Maximum Likelihood(ML) receiver statistic reduces to minimizing the exponent of the Gaussian random variable :

$$||y - hs||^2 = ||y||^2 + |s|^2||h||^2 - 2Re[s^* h^H y] \tag{4.25}$$

$$= ||h||^2 \left[s - \frac{h^H y}{||h||^2} \right]^2 + const, \quad \text{(completing the square)} \tag{4.26}$$

$$= ||h||^2 (s - \hat{s})^2, \quad \text{where } \hat{s} \sim N_C \left(s, \frac{\sigma^2}{||h||^2} \right) \tag{4.27}$$

since there are 'M' uncorrelated AWGN paths, the exponent of the ML function is additive, from which we get $\sum_{i=1}^{M} |h_i|^2$ as the processing gain for the M path uncorrelated Gaussian channels. Using this processing gain, the capacity for a **SIMO** system (single transmit and M receive antenna's) is given by:

$$C = \log_2 \left(1 + \rho \sum_{i=1}^{M} |h_i|^2 \right) \quad \text{(bits/s/Hz)} \tag{4.28}$$

where h_i is the complex gain for the receive antenna i.

Capacity of a MISO Link: Similarly the capacity for a **MISO** system (N transmit and single receive antenna) is given by:

$$C = \log_2 \left(1 + \frac{\rho}{N} \sum_{i=1}^{N} |h_i|^2 \right) \quad \text{(bits/s/Hz)} \tag{4.29}$$

where the normalization of SNR-ρ by N ensures capacity is independent of the number of transmitters.

Capacity of a MIMO Link: In contrast to the limited diversity design viz. *receiver diversity* -capacity given by equation 4.28 and *transmitter diversity* -capacity given by equation 4.29, a complete **MIMO** system with N transmit and M receive antennas, offers much more potential. Its capacity given by:

$$C = \log_2 \left[det \left(I_M + \frac{\rho}{N} \mathbf{H}\mathbf{H}^* \right) \right] \quad \text{(bits/s/Hz)}, \tag{4.30}$$

grows linearly with $min(M, N)$ see [8] for further details .

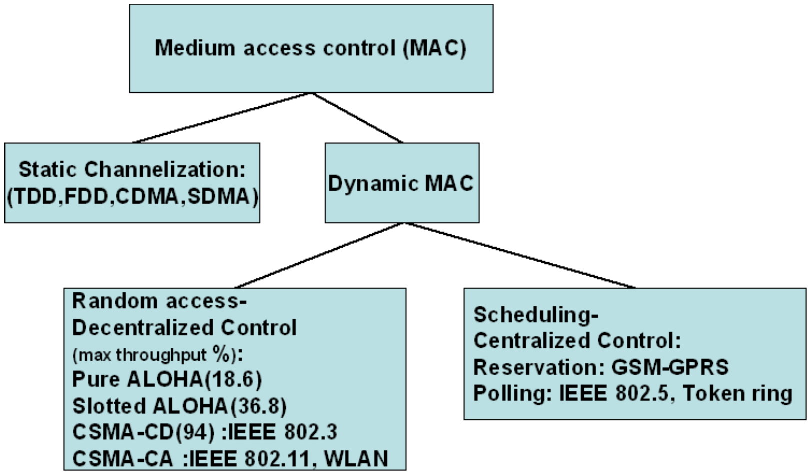

Figure 13: Medium access strategies

5 Medium Access strategies in wireless access networks

Why does one need to modify medium access(MAC) at all? Isn't getting high throughput largely a physical(PHY) issue?[3]. The answer to this is, while PHY data rates are increasing, throughput is increasingly dominated by overheads(preamble, PLCP headers), radio round-trip time, pipeline delay etc. Thus in order to efficiently utilize the wireless medium a system view and design of the MAC is needed to manage the the services of the PHY while providing an unmodified interface to the higher layers.

The classical theme in medium access strategy carries through to even the latest standards like 802.11e, the figure 13 gives the underlying framework of this medium access strategy. MAC buffers data till it reaches a threshold length or age, and schedules or accesses the radio channel to transmit the burst of data. In conjunction with the Link Control entity, MAC is responsible for the segmentation and fragmentation of data, error-checking and re-transmission of packets in error, and sequencing of the fragments/segments at the peer end.

The means by which the bursty channel access is acquired may be static, with a specific time slot, frequency channel, code or space being allocated priori, to the individual users, as seen in time division duplexing(TDD), frequency division duplexing(FDD), code division multiple access(CDMA) and space division multiple access(SDMA) systems.

The dynamic access strategy is differentiated into centralized and decentralized control strategies. The *centralized scheduling* scheme performs well under heavy traffic loads and guarantees the specified service level agreement(SLA), by incorporating *admission control*.

The *reservation* scheme used in GSM-GPRS and *polling* schemes used in satellite communication and 802.5 token ring are part of the scheduled dynamic MAC.

The decentralized control in the dynamic MAC dates back to the pure ALOHA radio networks developed at the University of Hawaii. Improvements in the capacity of this *random-access* scheme came about by reducing the collision detection time from two frames to one in the slotted ALOHA system. Proceeding on similar lines, the carrier sensing multiple access(CSMA) reduces the collision detection to the propagation time of one symbol. Decentralized random access performs well under light traffic, with little overhead in handshake or coordination, which are essential ingredients in the centralized scheduling mechanisms. The Ethernet technology uses the CSMA-collision detection(CD), while the wireless LAN's use the CSMA-collision avoidance(CA) random access strategy [13].

5.1 QoS Support in wireless LAN's

Mobility, ease of deployment and scalability of WLAN makes it the dominant solution for local wireless networking. Despite the possible use of MIMO communications at the physical layer, and the use of the mandatory distributed coordination function(DCF) and optional point coordination function(PCF) of 802.11 standard, for medium access control. The service is only a *best effort* scheme, with no guarantee on throughput, delay and jitter(variance in packet delay). The DCF is a contention based channel access mechanism, like the CSMA-CA, while the PCF is a contention free centralized scheduling scheme, see figure 13.

Extensive research in bringing the QoS framework into wireless LAN's has culminated in 802.11e, which proposes enhanced distributed channel access(EDCA) and HCF controlled channel access(HCCA). The EDCA is a distributed traffic engineering and service differentiation mechanism, that aims in providing fair throughput amongst different applications by tuning the four WLAN parameters: 1) Inter-Frame Space(IFS), 2) Contention Window(CW), 3) Backoff Interval(BI) and 4) Persistence Factor(PF).

Inter-Frame Space(IFS) is the period of time a mobile station(MS) is required to wait after it senses an idle channel, before actual transmission. *Contention Window(CW)* is the window frame bounded by CW_{max} and CW_{min} from which the MS chooses a random backoff interval(BI)(uniformly distributed within CW), when it senses a busy channel. *Backoff Interval(BI)* is used in a backoff timer, the MS waits for the backoff timer to expire before persisting on the channel access attempt, BI = Random(CW_{min}, CW_{max}) x Slot Time.

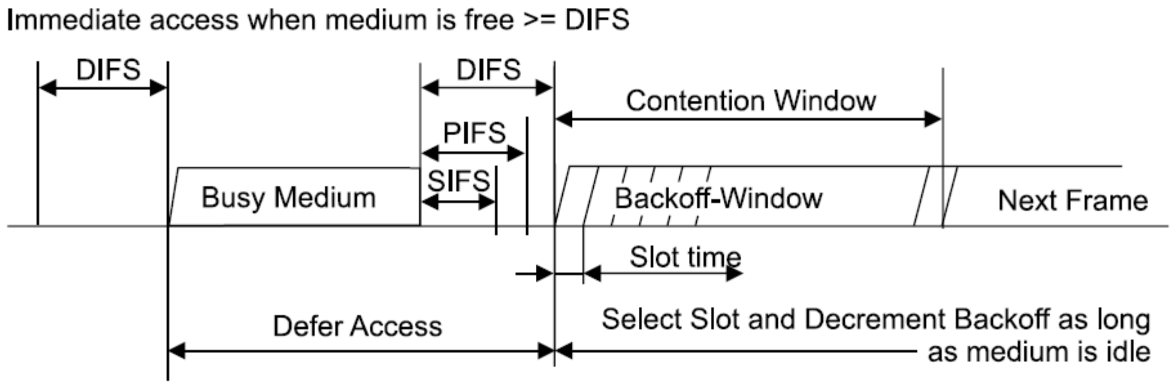

Figure 14: EDCA-QoS-Parameters: IFS,CW and BI

Persistence Factor(PF) is the probability $p < 1$, with which a MS that encounters a busy channel would persist on the channel access attempt. See figure 14 to visualize the EDCA QoS parameters.

HCAA uses HCF, which achieves prioritization and service differentiation by polling the wireless stations in the infra-structured centralized framework. Since the above two mechanisms of centralized and distributed control are complimentary, the key to success is in identifying the correct trade-off between EDCA and HCAA in meeting the service level agreement(SLA) for real-time applications like voice over IP(VOIP), and videoconferencing with stringent bandwidth, delay and jitter requirements [4, 18].

5.2 Cross-Layer design perspective

Having gone through PHY layer details of coding and modulation in MIMO systems and MAC QoS support in wireless LAN's, we will now look at the interaction between the layer entities for improved QoS in terms of delay, throughput and bit error rate (BER). The introduction of knowledge between the functionalities in different layers is bound to realize benefits in throughput increase, network latency reduction and energy savings in mobile nodes, when higher entities such as MAC (data link) and NWK are aware of the PHY layer state [2].

For instance the adaptation of MAC and Radio resource management (RRM) algorithms for re-use of radio resources at the transmitter would benefit from the channel state information as seen by the receiver. The other higher layer functions like handoff, routing algorithms, scheduling algorithms in addition to link adaptation strategies, would very much benefit from being aware of the channel state information including: estimates of channel impulse response, location information, vehicle speed, signal strength, interference level and etc. The above mentioned optimizations in power control, routing and mobility management, and interoperability in an heterogeneous wireless network with a mix of coexisting technologies(GSM,GPRS,UMTS and WLAN), would not be feasible unless some sort of cross-layer approach is adopted in the OSI stack.

6 Role of MIMO aware MAC

The current 802.11 MAC strategy, takes care of interference by two methods 1) physical carrier sensing and 2) virtual carrier sensing by using RTS/CTS and network allocation vector(NAV). Additionally many methods have been suggested including those of [18], who have proposed methods to take care of service differentiability and QoS in terms of throughput and delay.

In light of the current developments to move to multiple antenna elements(MEA)'s as the defacto physical layer alternative for broadband wireless access. It is clear that the current MAC schemes would under-utilize the interference canceling capability of MEA's, which would facilitate simultaneous transmissions in the carrier sensing range. Some groups have initiated a MAC re-design for both infrastructure and ad-hoc wireless networks using MEA's. This report is focused on the work by P Ramanathan, B V Veen and et.al [24, 25], who have proposed a MAC protocol, that exploits multiuser diversity to enable concurrent communications, they show this is better than multiple streaming. Other relevant work in this area are by R Sivakumar, M A Ingram and et.al [5, 6, 20, 23], who have proposed a controlled multiple streaming, spatial multiplexing MAC, with service differentiability provided by choosing different number of simultaneous streams, and the proposed MIMA/AS MAC by M Park, S M Nettles et.al [15, 16], who propose an antenna selection MAC that uses Alamouti coding for transmission and reception of control packets.

6.1 Review of MIMO MAC Design by P Ramanathan, B V Veen and et.al [24]

Fully Adaptive Antenna arrays allows multiple users, in a rich multipath environment, to couple to the best spatial channel, and to setup concurrent transmissions, while canceling interference from other users. The linearly independent channels created by multipath can be used to send multiple data streams between a transmit-receive pair. This being the underlying philosophy of smart-antennas in wireless networks, the above authors propose a MAC design with minimal changes to existing 802.11 MAC.

To facilitate concurrent multi-user communication by adaptive interference cancellation [24], propose an omni-directional RTS/CTS exchange followed by beamformed DATA/ACK.

- The RTS/CTS scheme would be incorporated with pilot symbols to obtain the channel state information (CSI)

- A modified NAV ensures accurate channel estimation, by preempting other nodes in the sensing range from initiating RTS/CTS for 'sifs+CTS' duration

- The transmitter adjusts its weights based on the CSI received in the CTS, while the receiver adjusts its weights based on the CSI received in the RTS, thus the transmitter-receiver pair are beamformed to the best spatial channel

- On successful packet transmission, reception of ACK, both the transmitter and receiver backoff for RTS+sifs+CTS duration to prevent collision with any ongoing omni directional control exchange.

More work needs to be done in looking at the tradeoff's of the designs suggested by different proponents, and a cross-layer perspective needs to be taken, to design our PHY aware MAC for the future MEA equipped devices.

7 Conclusion

MIMO systems would be the preferred physical layer alternative for broadband wireless access systems. Current medium access techniques do not account for the MIMO architecture. Some groups have initiated MAC redesign for both infrastructure and ad-hoc wireless networks using MEA's [5, 6, 20, 23–25]. More work needs to be done in understanding the node level and multi-node level, cross layer optimization issues, for the ultimate design of a PHY aware MAC, for multi-antenna systems in broadband wireless networks.

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A Large-scale Path Loss Models

The physics of the propagation of radio waves in any medium are dependent on the frequency of the signal, and its interaction with the obstacles in its journey from transmitter to the receiver. Detailed models have been summarized by Rappaport [19] based on reflection, diffraction and scattering of the transmitted waves by the obstacles in its path. These models are crucial in determining the coverage of a transmission and the capacity a link may support without undue errors.

A.1 Path loss in free space:

Free space propagation is the simplest propagation model of the radio waves in an isotropic medium. It determines the signal strength at the receiver when there is a clear line of sight(LOS) to the transmitter.

$$Frii's\ Equation:\ P_r(d) = \frac{P_t G_t G_r \lambda^2}{(4\pi)^2 d^2}, \quad G = \frac{4\pi A_e}{\lambda^2} \tag{A.1}$$

P_r, P_t are the receiver(Rx) and transmitter(Tx) power respectively, $\lambda = \frac{c}{\nu}$ is the wavelength of the carrier, and d is the distance between the Tx and Rx and G is the antenna gain which is related to the effective aperture A_e.

$$Path\ Loss(dB) = 10log\frac{P_t}{P_r} = -10log\left[\frac{G_t G_r \lambda^2}{(4\pi)^2 d^2}\right] \tag{A.2}$$

While the free space model is suitable for a rural environment with no scatterers and clear LOS, any urban terrestrial or indoor wireless communication is fraught with interference from multiple reflection paths due to local scatterers such as houses, buildings and other man-made structures or by natural objects such as foliage, mountains and clouds. There are many models proposed to explain the observed fading phenomenon let us see two of them:

A.2 Two Ray Reflection model:

In this model the two paths from the transmitter, line of sight(LOS) and the ray reflected from the earths surface($\Gamma = -1$ assuming perfect reflection), interfere with each other at the receiver. The difference in the distance traveled in the two paths introduces a phase difference $\Delta\phi = \frac{2\pi\Delta R}{\lambda}$ in the E-field. Using Pythagorean theorem and binomial expansion

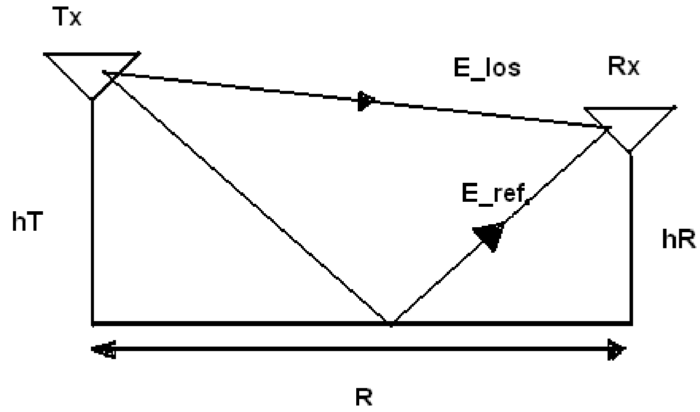

Figure 15: Reflection and the plane-earth model

approximation $\sqrt{1+x} = 1+\frac{x}{2}$ when $x < 1$, the path difference and hence the phase difference in the E-field is: see figure 15

$$R_d = \sqrt{R^2 + (h_T - h_R)^2}, \quad R_r = \sqrt{R^2 + (h_T + h_R)^2} \tag{A.3}$$

$$\Delta R = \frac{(h_T + h_R)^2 - (h_T - h_R)^2}{2R} = 2\frac{h_T h_R}{R} \tag{A.4}$$

$$\Delta\phi = \frac{2\pi\Delta R}{\lambda} = \frac{4\pi h_T h_R}{\lambda R} \tag{A.5}$$

The total E-Field at the receiver is the sum of the incident(LOS) and the reflected waves:

$$|E_{TOT}(d,t)| = |E_{LOS}(d,t) + E_{ref}(d,t)| \tag{A.6}$$

from the free-space propagation model, the E-field(units V/m) is given by:

$$E(d,t) = \frac{E_0 d_0}{R_d} cos\left(\omega_c\left(t - \frac{R_d}{c}\right)\right), (d > d_0) \tag{A.7}$$

thus E_{TOT} becomes

$$E_{TOT}\left(d, t=\frac{R_r}{c}\right) = \frac{E_0 d_0}{R_d} cos\left(\omega_c\left(\frac{R_r - R_d}{c}\right)\right) - \frac{E_0 d_0}{R_d} cos(0°) \tag{A.8}$$

$$= -\frac{E_0 d_0}{R_d}\left(1 - cos(\Delta\phi)\right) = \tilde{E}_d \mathrm{Re}\left\{1 - exp(-j\Delta\phi)\right\} \tag{A.9}$$

$$= |\tilde{E}_d|\left[1 + cos^2\Delta\phi - 2cos\Delta\phi + sin^2\Delta\phi\right]^{1/2} \tag{A.10}$$

$$= |\tilde{E}_d|\left[2 - 2cos\Delta\phi\right]^{1/2} = 2|\tilde{E}_d| sin\left(\frac{\Delta\phi}{2}\right) \tag{A.11}$$

The relationship between E-field and power received is:

$$P_r = \frac{|E_{TOT}|^2 A_e}{\eta_0}\ where\ \eta_0 = 120\pi = 377\Omega \text{ is the free space impedance} \tag{A.12}$$

$$\text{Thus } P_r = 4\frac{|\tilde{E}_d|^2 A_e}{\eta_0} sin^2\left(\frac{2\pi h_T h_R}{\lambda R}\right) \tag{A.13}$$

The factor $\frac{|\tilde{E}_d|^2 A_e}{\eta_0}$ is the power received via the direct path, using the free space propagation for this direct path and using $sin\theta \approx \theta$ for small θ we get:

$$P_r = 4P_t\left(\frac{\lambda}{4\pi R}\right)^2 G_T G_R sin^2\left(\frac{2\pi h_T h_R}{\lambda R}\right) = P_t G_T G_R\left(\frac{h_T h_R}{R^2}\right)^2 \tag{A.14}$$

$$Path\ Loss(dB) = 40logd - (10logG_T + 10logG_R + 20logh_T + 20logh_R) \tag{A.15}$$

Thus from the two ray model the received power is attenuated by a factor R^n where the exponent $n = 4$

A.3 Indoor Propagation: Attenuation Factor Model

The generalized attenuation factor model, falls right off the two ray model:

$$Path\ Loss(d)[dB] = PL(d_0) + 10nlog\left(\frac{d}{d_0}\right) \tag{A.16}$$

the path loss exponent n takes the following values:

Sample path-loss exponents from [10]

Environment	n
Free space:	2
Flat rural:	3
Rolling rural:	3.5
Suburban,low rise:	4
Dense urban,skyscrapers:	4.5