An Experimental Hardware Prototype for Fixed Wireless Broadband Access at 60-400 Mbit/sec in 4.6 MHz

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Abstract— **While the race for a ubiquitous, reliable and cost effective wireless broadband system has generated intense competition, the fundamental problem afflicting the wireless industry is cost of the transceiver hardware. We describe and discuss the results of some experimental outdoor field trials which demonstrate -for the first time- the hardware feasibility of fixed broadband wireless radio links with spectral efficiencies in excess of 50 bit/sec/Hz in severe multipath frequency selective environments (non line-of-sight). Custom cost-effective hardware has allowed data transmission in the range 60-200 Mbit/sec occupying a bandwidth of 2-4-5 MHz at MMDS frequencies ($\simeq$2.5 GHz). We also introduce a new modeling methodology inspired by basic theoretical physics which gives a different and practical perspective to the spatio-temporal transmission problem. We exploit the idea that extra-dimensionality can be created by properly sampling the wavefield space created by multipath propagation. Previously reported hardware experiments using the same concept have addressed the simpler problem of indoor communications in a 30 KHz cellular channel. The developed technology has the potential of making practical and cost effective the deployment of high capacity microwave data communications equipment for point to point and point to multipoint systems.**

I. Background

Broadband connectivity has been considered the future of the telecommunications industry for decades and it typically refers to those data services operating at data rates in excess of 1.544 Mbit/sec (known as T-1 rate). While the theoretical foundations of high data rate digital communications are appearently well-known, there has not been any practical solution to implement the massive amount of signal processing required to operate reliably over the poor quality "last-mile" channel. The principal obstacle to mass deployment of broadband data services is without any doubt cost. The wireless terrestrial transmission media presents probably the most difficult challenges. Several types of solutions are being proposed and initially deployed. The Multipoint Multichannel Distribution Service (MMDS) uses microwave transmission at frequency approximately around 2.5 GHz. It is currently envisioned by the wireless communication industry that the MMDS system can broadcast video, voice and data, allowing for interactive communications. One of the drawbacks of the system is that MMDS relies on traditional high capacity microwave technology which requires Line of Sight transmission and the development of high-performance RF [Radio Frequency] circuitry. The idea that multiple transmit/receive antennas have a substantial benefit on the achievable data rate in multipath fading environment has attracted remarkable interest in recent years at least in the research community [4], [5], [6]. Since it is possible to expect an increase in data throughput directly proportional to the number of sensors at the antenna arrays without any penalty in power and bandwidth, this technology appears to be a winning solution in many short and long-range wireless communications applications. However the cost increase of the transceiver is proportional to the number of antennas used by the system because the transceivers have multiple RF and DSP processing branches. Ultimately such a dilemma has slowed down many inititatives in industry. At Lucent, a system called BLAST (Bell-Labs Space-Time Architecture) was developed for a non-frequency selective, static (typically indoor) environment with extremely narrowband throughput [3]. Despite the undeniable value of that first study it is not difficult to see the restricted applicability of the investigation. In this paper we describe and discuss the results of some experimental outdoor field trials which have demonstrated -for the first time- the hardware feasibility of fixed broadband wireless radio links with spectral efficiencies in excess of 50 bit/sec/Hz in severe multipath environments. Wireless data transmission in the range 60-200 Mbit/sec occupying a bandwidth of 2-4 MHz at 2.5 GHz was achieved. We also introduce a new modeling methodology inspired by basic theoretical physics which gives a different and practical perspective to the spatio-temporal transmission problem and as a byproduct a new class of methodologies which we

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have defined STREAM$_{TM}$ (Spatial Transmission with Radio Enhanced Adaptive Modulation).

II. Spatio-Temporal Wavefield Modeling

In the context of spatio-temporal processing one is typically interested in wavefields $e(t, \mathbf{r})$ propagating according to the wave equation

$$\left(\nabla^2 - \frac{1}{c^2}\frac{\partial^2}{\partial t^2}\right) e(t, \mathbf{r}) = 0 \qquad (1)$$

where c is the velocity of propagation of the medium, $\mathbf{r}$ identifies the spatial location (2-D $[\mathbf{r} = (r, \phi)]$ or 3-D $[\mathbf{r} = (r, \phi, \theta)~]$) of the propagating wave, t identifies the temporal location of the propagating wave. Assuming that the wavefield has a Fourier representation $e(t, \mathbf{r}) = \int d\omega E(\omega, \mathbf{r}) e^{j\omega t}$, each Fourier component satisfies the source-free Helmholtz equation [1]

$$\left(\nabla^2 + \mathcal{K}^2\right) E(\omega, \mathbf{r}) = 0, \qquad (2)$$

where $\mathcal{K} = \frac{\omega^2}{c^2}$ is the wavenumber associated with c. In particular we are interested in wavefields generated by linear space-time models expressed as

$$s(t, \mathbf{r}) = \int_T \int_\Gamma h(t, \tau, \mathbf{r}, \mathbf{r}') x(\tau, \mathbf{r}') d\mathbf{r}' d\tau \qquad (3)$$

where $h(t, \tau, \mathbf{r}, \mathbf{r}')$ is the time-variant/space-variant wavefield response of the channel (assumed space-time selective), T is the time interval of interest and Γ is the spatial volume of interest. Observe that $\mathbf{r}$ is a 3-D or 2-D parametrization of the spatial domain. Roughly speaking $h(t, \tau, \mathbf{r}, \mathbf{r}')$ describes the space-time response of the system to an impulse generated at time t [as measured at delay τ] and spatial location $\mathbf{r}'$ [as measured at location $\mathbf{r}$]. If the channel is time-invariant $[h(t, \tau, \mathbf{r}, \mathbf{r}') = h(t - \tau, \mathbf{r}, \mathbf{r}')~]$ we obtain a simple model for the wavefield of interest

$$S(\omega, \mathbf{r}) = \int_\Gamma H(\omega, \mathbf{r}, \mathbf{r}') X(\omega, \mathbf{r}') d\mathbf{r}', \qquad (4)$$

where $x(t, \mathbf{r}) = \int d\omega X(\omega, \mathbf{r}, \mathbf{r}') e^{j\omega t}$, $h(t, \mathbf{r}, \mathbf{r}') = \int d\omega H(\omega, \mathbf{r}, \mathbf{r}') e^{j\omega t}$.

Assuming a farfield scattering model we can write $H(\omega, \mathbf{r}, \mathbf{r}')$ as a a superposition of plane waves

$$H(\omega, \mathbf{r}, \mathbf{r}') = \int_\Theta \int_\Theta d\gamma d\gamma' \rho_h(\gamma, \gamma', \omega) e^{j\mathbf{r}\cdot\hat{\gamma}\mathcal{K}} e^{j\mathbf{r}'\cdot\hat{\gamma}'\mathcal{K}}$$

where Θ is the set of possibles directions of arrival, $\Theta = \begin{cases} \phi \in [-\pi, \pi] & \text{if 2-D} \\ \theta \in [0, \pi],~\phi \in [-\pi, \pi] & \text{if 3-D} \end{cases}$ γ and γ' are points in Θ, $\hat{\gamma}$ and $\hat{\gamma}'$ are unit vectors pointed in the directions γ and γ', $\rho_h(\gamma, \gamma', \omega)$ is a *scattering radiation density*. $X(\omega, \mathbf{r})$ is expressed as $X(\omega, \mathbf{r}) = \sum_{n=1}^{N_{tx}} q_n(\omega)\delta(\mathbf{r} - \mathbf{r}_n)$ where N_{tx} is the number of elements at the transmit array, $q_n(\omega)$ are the N_{tx} information bearing signals transmitted at the N_{tx} elements of the transmit array, $\mathbf{r}_n$, $n = 1, 2, ..., N_{tx}$ are the locations of the elements in the transmit array. Since a plane wave $e^{j\mathbf{r}\cdot\hat{\gamma}\mathcal{K}}$ is solution of the Helmoltz equation, $\rho_h(\gamma, \gamma', \omega)$ is an arbitrary complex function fully specified by the spatio-temporal propagation modes of the channel. For simplicity of notation from now on we will drop the explicit dependence on ω. Expanding with a complete and orthogonal (2-dimensional for 2-D propagation or 4-dimensional for 3-D propagation) basis [denoted $f_{n,m}(\gamma, \gamma')$] the scattering radiation density

$$\rho_h(\gamma, \gamma') = \sum_n \sum_m \psi_{n,m} f_{n,m}(\gamma, \gamma') \qquad (5)$$

where $\psi_{n,m}$ are coefficients of the expansion given by $\psi_{n,m} = \int_\Theta \int_\Theta d\gamma d\gamma' \rho_h(\gamma, \gamma') f^*_{n,m}(\gamma, \gamma')$. Using (5) we obtain $H(\mathbf{r}, \mathbf{r}') = \sum_{n,m} \psi_{n,m} \mathcal{H}_{n,m}(\mathbf{r}, \mathbf{r}')$ where

$$\mathcal{H}_{n,m}(\mathbf{r}, \mathbf{r}') = \int_\Theta \int_\Theta d\gamma d\gamma' f_{n,m}(\gamma, \gamma') e^{j\mathbf{r}\cdot\hat{\gamma}\mathcal{K}} e^{j\mathbf{r}'\cdot\hat{\gamma}'\mathcal{K}}.$$

So we can express

$$S(\mathbf{r}) = \sum_{k=1}^{N_{tx}} q_k \sum_{n,m} \psi_{n,m} \mathcal{H}_{n,m}(\mathbf{r}, \mathbf{r}_k). \qquad (6)$$

Now consider the receive array as a *sampling operator*: from $S(\omega, \mathbf{r})$ it returns a vector $\mathbf{s}$ of measurements (the array output)

$$\mathbf{s} = A_R \circ S(\mathbf{r}) \qquad (7)$$

where A_R is the *array sampling operator* [a vector valued linear and continuous functional]. Define $\mathbf{a}_{R,n,m,k}$ as the response of the receive array to a wavefield of the form $\mathcal{H}_{n,m}(\mathbf{r}, \mathbf{r}_k)$, $\mathbf{a}_{R,n,m,k} = A_R \circ \mathcal{H}_{n,m}(\mathbf{r}, \mathbf{r}_k)$. From the linearity of the sampling operator we get

$$\begin{aligned} \mathbf{s} &= A_R \circ S(\mathbf{r}) = \sum_k q_k \sum_{n,m} \psi_{n,m} A_R \circ \mathcal{H}_{n,m}(\mathbf{r}, \mathbf{r}_k) \\ &= \sum_k q_k \sum_{n,m} \psi_{n,m} \mathbf{a}_{R,n,m,k} = \sum_k q_k \mathbf{A}_{R,k} \mathbf{\Psi}, \qquad (8) \end{aligned}$$

where $\mathbf{\Psi}$ is a vector of coefficients $\psi_{n,m}$ and $\mathbf{A}_{R,k}$ is a matrix of vectors $\mathbf{a}_{R,n,m,k}$ ordered as to properly represent $\sum_{n,m} \psi_{n,m} \mathbf{a}_{R,n,m,k} = \mathbf{A}_{R,k}\mathbf{\Psi}$. The value of (8) is that it **decouples completely** the effect of the spatio-temporal channel and the parameters of the transmit/receive array. The vector $\mathbf{\Psi}$ fully captures the spatio-temporal channel. An array of isotropic elements in a 2-D spatial representation gives $[\mathbf{a}_{R,n,m,k}]_l = \mathcal{H}_{n,m}(\mathbf{r}_l = (r_l, \phi_l), \mathbf{r}_k = (r_k, \phi_k))$ where $\mathbf{r}_l = (r_l, \phi_l)$ is the location of the lth element in the receive array and $\mathbf{r}_k = (r_k, \phi_k)$ is the location of the kth element in the transmit array.

III. Detection

For simplicity we consider the uniform linear array configuration, both at the transmitter and at the receiver. We then have

$$[\mathbf{a}_T(\gamma)]_k = e^{j\frac{2\pi}{\lambda}(k-1)d_T sin\theta} \quad k=1,2,...,N_{tx}$$

and

$$[\mathbf{a}_R(\gamma)]_k = e^{j\frac{2\pi}{\lambda}(k-1)d_R sin\theta} \quad k=1,2,...,N_{rx}$$

where d_T and d_R are element separations, respectively at the transmitter and at the receiver arrays and λ is the carrier wavelength. These expressions can be used in (??) and numerical computation of $\mathbf{a}_{R,n,m,k}$ can be performed. The vector $\mathbf{\Psi}(\omega)$ can be truncated to $4N+2$ terms $[\psi_{-N,-N}(\omega),...,\psi_{N,N}(\omega)]^T$. According to the Maximum Likelihood (ML) principle the optimal estimation of the channel parameters can be obtained maximizing

$$\begin{aligned} L_N(\mathbf{\Psi}) &= -\sum_{i=1}^{N_{rx}} \left| [\mathbf{r}]_i - \left[\sum_k q_k \mathbf{A}_{R,k}\mathbf{\Psi} \right]_i \right|^2 \\ &= -||\mathbf{r} - \sum_k q_k \mathbf{A}_{R,k}\mathbf{\Psi}||^2 \\ &= -||\mathbf{r} - \mathbf{S}\mathbf{\Psi}||^2, \end{aligned} \tag{9}$$

which is obtained as $\hat{\mathbf{\Psi}} = \left[\mathbf{S}^H\mathbf{S}\right]^{-1}\mathbf{S}^H\mathbf{r}$. The matrix $\mathbf{S}$ is formed from appropriate training data.

The transmitted signals can be expressed in the time domain as

$$\begin{aligned} \tilde{q}_k(t) &= \sum_l a_k(l) p_k(t-lT_s) \\ &= \sum_l a_k(l) p_{k,l}(t) = \int d\omega q_k(\omega) e^{j\omega t}, \end{aligned} \tag{10}$$

where $a_k(l)$ is a QAM symbol, T_s is the symbol period and $p_k(t)$ a pulse shaping filter. Denoting convolutions as $*$, we have

$$r_i(t) = \sum_{k,l} a_k(l) \sum_{n,m} \tilde{a}_{R,k,n,m,i}(t) * \tilde{\psi}_{n,m}(t) * p_{k,l}(t) + \tilde{n}_i(t)$$

where

$$\tilde{a}_{R,k,n,m,i}(t) = \int d\omega \left[\mathbf{a}_{R,n,m,k}(\omega)\right]_i e^{j\omega t}$$

$r_i(t) = \int d\omega \left[\mathbf{r}(\omega)\right]_i e^{j\omega t}$, $\tilde{n}_i(t) = \int d\omega \left[\mathbf{n}(\omega)\right]_i e^{j\omega t}$ and $\tilde{\psi}_{n,m}(t) = \int d\omega \psi_{n,m}(\omega) e^{j\omega t}$.

The connection with the commonly assumed MIMO (Multiple Input Multiple Output) model (see for example [8], [9], [11], [10])

$$r_i(t) = \sum_{k=1}^{N_{tx}} \sum_{l=-\infty}^{+\infty} a_k(l) h_{k,i}(t-lT_s) + \tilde{n}_i(t), \quad i=1,...,N_{rx}$$

is evidently obtained as

$$h_{k,i}(t) = \sum_{n,m} \tilde{a}_{R,k,n,m,i}(t) * \tilde{\psi}_{n,m}(t) * p_k(t),$$

or in the frequency domain

$$H_{k,i}(\omega) = \sum_{n,m} a_{R,k,n,m,i}(\omega)\psi_{n,m}(\omega)P_k(\omega),$$

where $p_k(t) = \int d\omega P_k(\omega) e^{j\omega t}$, and $[\mathbf{a}_{R,n,m,k}(\omega)]_i = a_{R,k,n,m,i}(\omega)$. At this point we can look for the optimum MMSE (Minimum Mean Square Error) linear filter $W_{k,i}(\omega)$, whose T_s-spaced sampled output can be expressed as

$$Z_k(\omega) = \sum_{m=-\infty}^{+\infty} \sum_{i=1}^{N_{rx}} W_{k,i}(\omega - 2\pi m/T_s) R_i(\omega - 2\pi m/T_s).$$

We have considered spectrum folding due symbol-rate sampling and used the notation $R_i(\omega) = [\mathbf{r}(\omega)]_i$. The Mean Squared Error is written as

$$MSE_k = T_s \int_{-2\pi/2T_s}^{+2\pi/2T_s} E\left\{ |Z_k(\omega) - A_k(\omega)|^2 \right\} d\omega,$$

where $A_k(\omega)$ is the Fourier domain representation of $\sum_{m=-\infty}^{+\infty} a_k(m)\delta(t-mT_s)$. For simplicity we assume that there is no excess bandwidth, that is that $P_k(\omega)$ is exactly bandlimited to the Nyquist bandwidth. Since we have

$$H_{k,i}(\omega) = \mathbf{\Psi}(\omega)^T \mathbf{a}_{R,k,i}(\omega) P_k(\omega) = \mathbf{a}_{R,k,i}(\omega)^T \mathbf{\Psi}(\omega) P_k(\omega)$$

we can compact in vector notation

$$\mathbf{H}_k(\omega) = \mathbf{A}_{R,k}(\omega)\mathbf{\Psi}(\omega)P_k(\omega) \tag{11}$$

where

$$\mathbf{H}_k(\omega)^T = \left[H_{k,1}(\omega), H_{k,2}(\omega), ..., H_{k,N_{rx}}(\omega)\right].$$

Basically the expression (11) states that $\mathbf{H}_k(\omega)$ is in the subspace spanned by the columns of $\mathbf{A}_{R,k}(\omega)$ which forces the channel vector to lie in the appropriate *a-priori* known subspace: the subspace of the baseband channels generated by plane waves obeying the Helmoltz equation. Compared to unstructured methods [8] this channel parametrization in practice reduces the Mean Square Error of an MMSE detector computed using (11). On the other hand the structure of the model is such that no estimation of angle of arrivals or channel gains is necessary.

The optimum linear MMSE filter is obtained from

$$\begin{aligned} \mathbf{W}_k(\omega) &= \left[W_{k,1}(\omega), W_{k,2}(\omega), ..., W_{k,N_{rx}}(\omega)\right]^T \\ &= \left[\mathbf{R}_H(\omega) + \mathbf{R}_N(\omega)\right]^{-1} \mathbf{H}_k(\omega) \\ &= \mathbf{R}(\omega)^{-1} \mathbf{A}_{R,k}(\omega)\mathbf{\Psi}(\omega)P_k(\omega) \end{aligned} \tag{12}$$

where $[\mathbf{R}_N(\omega)]_{m,n} = E\{N_n(\omega)N_m^*(\omega)\}$ with $[\mathbf{n}(\omega)]_m = N_m(\omega)$ [1] and

$$\begin{aligned}\mathbf{R}_H(\omega) &= \sum_{k=1}^{N_{tx}} \mathbf{H}_k(\omega)^* \mathbf{H}_k(\omega)^T \\ &= \sum_{k=1}^{N_{tx}} \mathbf{A}_{R,k}(\omega)^* \mathbf{\Psi}(\omega)^* \mathbf{\Psi}(\omega)^T \mathbf{A}_{R,k}(\omega)^T |P_k(\omega)|^2 .\end{aligned} \tag{13}$$

The evident problem is that the expansion of $\rho_h(\gamma, \gamma')$ is exact only with an infinite number of components in $\mathbf{\Psi}(\omega)$. Reducing the number of components in $\mathbf{\Psi}(\omega)$ one could trade off bias and variance [7] of the reconstruction error. The MMSE filters can be implemented in the time-domain or in the frequency domain. The preferred implementation used in the experimental prototype is a multirate architecture [13]. The length of the expansion in (11) used in the hardware experiments is $N = 64$.

IV. Description of the Hardware Architecture

WJ Communications Inc. (WJCI) has developed a hardware prototype which enabled STREAM_{TM} with up to 6 transmit antennas and 12 receive antennas. The bandwidth of the radio is selectable up to 5 MHz. We present results for the 2.5 GHz band. Fig. 1 show the architecture of the transceiver (receiver section). The transmit and receiving antennas (Gain = 7dBi, 70^o azimuthal 3-dB beamwidth, 60^o vertical beamwidth) are connected to the site RF distribution (not shown for simplicity). The Wideband RF Modules (one per antenna, see Fig. 1) are in charge of filtering, amplifying, downconverting and digitizing the RF spectrum of interest. The wideband receiver is designed to be driven from a site RF distribution network and minimizes this interface requiring a single, low gain RF feed per antenna element. High dynamic range amplifiers and mixers are utilized to obtain a large instantaneous dynamic range preserving signal fidelity. The signal as collected by the antenna is filtered and amplified. It is then mixed by a first Local Oscillator and filtered, amplified and mixed down to zero IF (Intermediate Frequency) by an analog quadrature downconverter. The transceivers are locked to a common frequency reference. The channel coding strategy is inspired to the concatenated approach. Encoders and interleavers are specified in [2]. Data formatting refers to the organization of the information bitstream in a way that is compatible with the particular application of the communication system. The platform used for the hardware trials was completed in July 2000. Integration progress for the hardware modules with relative spectral efficiencies is reported in Fig. 2. The signal modulation employed at each antenna is 64-QAM and roll-off factor of the baseband filters is 0.15. The Peak to Average Ratio remains within 6-9 dB which makes the technique significantly more advantageous than OFDM. We show

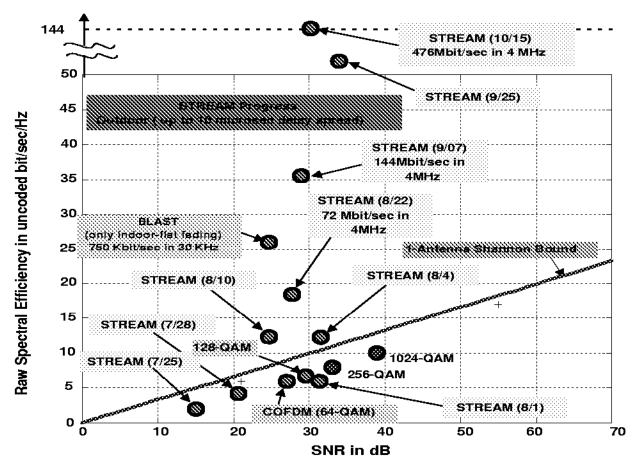

Fig. 2. Progress of STREAM_{TM} in terms of achieved spectral efficiency in Non Line of Sight.

hardware results of a typical fixed wireless broadband access scenario (about 0.8 mile range) in Non Line Of Sight. The delay spread is about 1 μsec and the bandwidth used by the radio is 4 MHz. The antennas are arranged in a uniform linear array configuration (1 wavelength spacing). The Peak to Average Ratio remains within 6-9 dB which makes the technique significantly more advantageous than OFDM. In Fig. 3 we summarize the achieved data rates for BER less than 10^{-6} in 4 MHz averaged over different locations. Observe that we report results for a number of antennas larger than the ones available in hardware. To accomplish virtual antenna measurements, we exploited the channel time invariance and simulated signal collections for 9-12 and 20 transmit antennas. Such measurements realized the achievement of 54 bit/sec/Hz, 60 bit/sec/Hz and approximately 144 bit/sec/Hz in a 4 MHz channel. The data rates of Fig. 3 account for 28% overhead due to channel coding and frame overhead for training and time-frequency synchronization.

[1] The modification of these expressions for the excess-bandwidth case can be easily determined considering the folded spectra for the channels $H_{k,i}(\omega - 2\pi m/T_s)$ and noise $N_i(\omega - 2\pi m/T_s)$, for $m = -J, -J+1, ..., J$.

$$\mathbf{H}_k(\omega)^T = \left[H_{k,1}(\omega - 2\pi J/T_s), ..., H_{k,N_{rx}}(\omega + 2\pi J/T_s)\right],$$

where $2\pi J/T_s$ is large enough to cover for the excess bandwidth of $P_k(\omega)$.

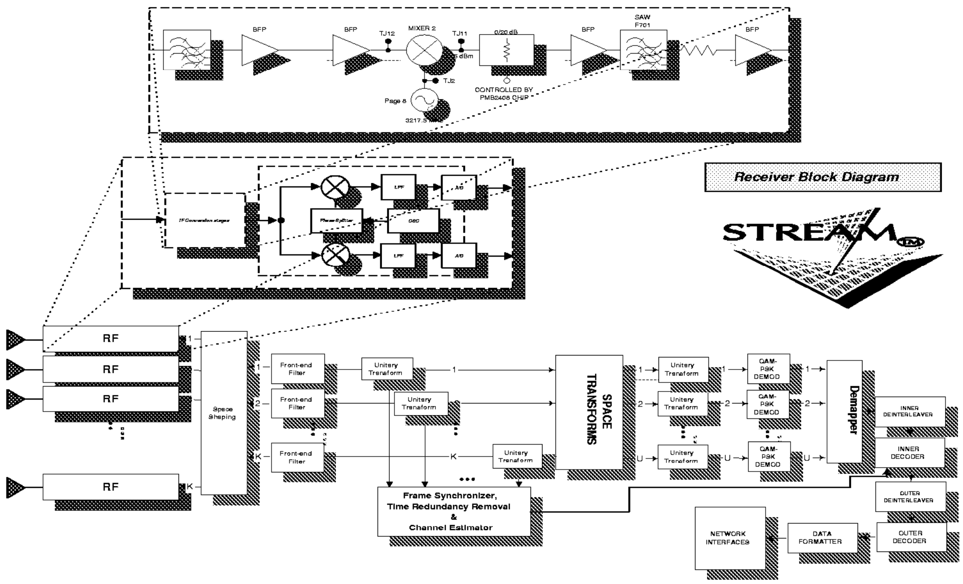

Fig. 1. The Hardware architecture of the Transceiver (receiver section) developed by WJ Communications. DSP data flow is only shown at the functional level.

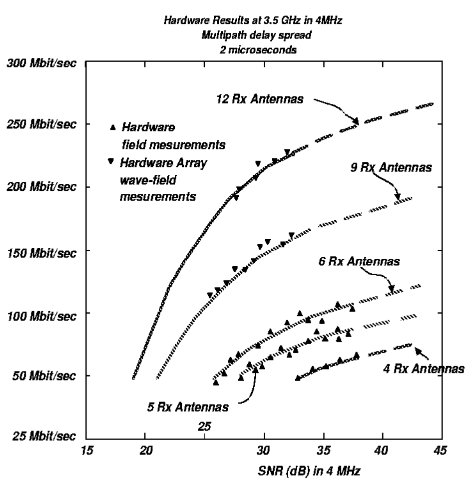

Fig. 3. Bit Error Rate results averaged over 80 different locations in 4 MHz 4 - 5 - 9 - 12 receive antennas. Curves are for convolutionally coded 64-QAM.

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