Experimental Results and Transceiver Hardware Design for Fixed Broadband Wireless Non-Line-Of-Sight Transmission: up to 1 Gbit/sec in 4.6 MHz

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Abstract— **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 60-960 Mbit/sec occupying a bandwidth of 2-4-5 MHz at MMDS frequencies (≃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. 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. Spatio-Temporal Wavefield Modeling

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 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 have defined STREAM$_{TM}$ (Spatial Transmission with Radio Enhanced Adaptive Modulation). Due to space limitations our treatment is brief and concise. We direct the interested reader to the upcoming book [1]. 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 [2]

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

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 , is the spatial v olume 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

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

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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

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

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

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

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)

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

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

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

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.

II. Wavefield Representations Based on Fourier Spatial Expansion

The selection of a basis that parsimoniously represents the radiation density is a critical component of the technology we introduce. One straightforward representation for the "channel" wavefield can be used employing Fourier bases. Consider the 2-D case with

$$f_{n,m}(\gamma,\gamma') = \frac{1}{2\pi} e^{jn\phi} e^{jm\phi'}. \tag{9}$$

The basis set $\mathcal{H}_{n,m}(\mathbf{r},\mathbf{r}')$ can be written [2]

$$\begin{aligned} \mathcal{H}_{n,m}(\mathbf{r} = (r,\phi), \mathbf{r}' = (r',\phi')) &= \int_0^{2\pi}\int_0^{2\pi} d\theta d\theta' \\ & e^{j\mathcal{K}r\cos(\theta-\phi) - jn\theta} e^{j\mathcal{K}r\cos(\theta'-\phi') - jm\theta'} \\ &= 2\pi j^{n+m} J_n(\mathcal{K}r) J_m(\mathcal{K}r') e^{-j(n\phi + m\phi')}, \end{aligned} \tag{10}$$

where $J_n(\mathcal{K}r)$ is the Bessel function of the first kind. So we obtain

$$[\mathbf{a}_{R,n,m,k}]_l = 2\pi j^{n+m} J_n(\mathcal{K}r_l) J_m(\mathcal{K}r_k) e^{-j(n\phi_l + m\phi_k)}$$
$$l = 1,2,...,N_{rx}, k = 1,2,...,N_{tx}. \tag{11}$$

It is worth to examine in more detail the relationship of the wavefield model with the traditional model based on manifolfds. Define $\mathbf{a}_R(\gamma)$ and $\mathbf{a}_T(\gamma)$ as the manifolds of the receive and transmit array [in other words $[\mathbf{a}_R(\gamma)]_m$ and $[\mathbf{a}_T(\gamma)]_m$ are the responses of the m-th sensor to a plane wave propagating in the direction γ]. Since

$$\mathbf{a}_R(\gamma) = A_R \circ e^{j\mathcal{K}\mathbf{r}\cdot\hat{\gamma}}$$

and

$$\mathbf{a}_T(\gamma) = A_T \circ e^{j\mathcal{K}\mathbf{r}'\cdot\hat{\gamma}'}$$

where A_R and A_T are the *array sampling operators* [receiver and transmitter respectively], we get

$$\begin{aligned}\mathbf{a}_{R,n,m,k} &= A_R \circ \int_\Theta \int_\Theta d\gamma d\gamma' \\ &\quad f_{n,m}(\gamma,\gamma') e^{j\mathbf{r}\cdot\hat{\gamma}\mathcal{K}} e^{j\mathbf{r}_k\cdot\hat{\gamma}'\mathcal{K}} \\ &= \int_\Theta \int_\Theta d\gamma d\gamma' f_{n,m}(\gamma,\gamma') \\ &\quad \mathbf{a}_R(\gamma)\,[\mathbf{a}_T(\gamma')]_k \,. \qquad (12)\end{aligned}$$

This says that $[\mathbf{a}_{R,n,m,k}]_l$ is the Fourier decomposition of the product between the manifolds of the transmit and receive arrays which implies

$$\mathbf{A}_{R,k}\mathbf{w}_{\gamma,\gamma'} = \mathbf{a}_R(\gamma)\,[\mathbf{a}_T(\gamma')]_k$$

where $\mathbf{w}_{\gamma,\gamma'}$ is a vector organization of the basis functions $f^*_{n,m}(\gamma,\gamma')$.

III. 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. The transmit and receiving antennas (Gain = 7dBi, 70° azimuthal 3-dB beamwidth, 60° vertical beamwidth) are connected to the site RF distribution (not shown for simplicity). The Wideband RF Modules 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 [3]. 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. 1. The signal modulation employed at each antenna is 64-QAM and roll-off factor of the baseband filters is 0.15. We show 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. 2 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, approximately 144 bit/sec/Hz and approximately 208 bit/sec/Hz in a 4MHz channel. The data rates of Fig. 2 account for 28% overhead due to channel coding and frame overhead for training and time-frequency synchronization.

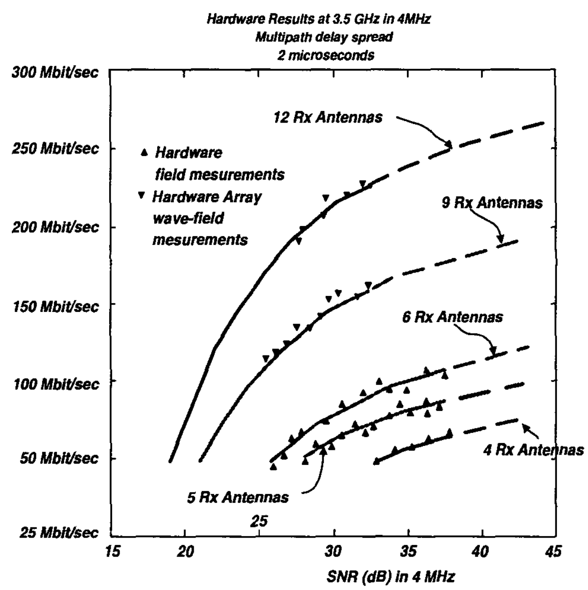

Fig. 2. Bit Error Rate results averaged over 80 different locations in 4 MHz 4 - 5 - 9 - 12 receive antennas. Curves are for convolutionally coded 64-QAM.

IV. Line of Sight Channels

The Line of Sight Channel is obviously of interest in long-haul microwave applications. We summarize in Fig. 3 the results of a typical 5 Km microwave

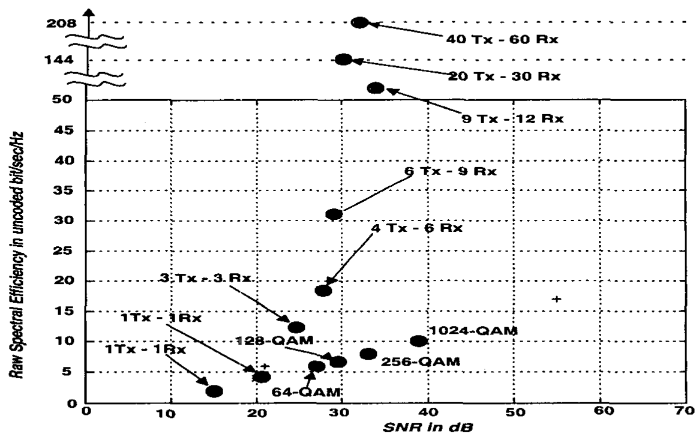

Fig. 1. Progress of $STREAM_{TM}$ in terms of achieved spectral efficiency in Non Line of Sight.

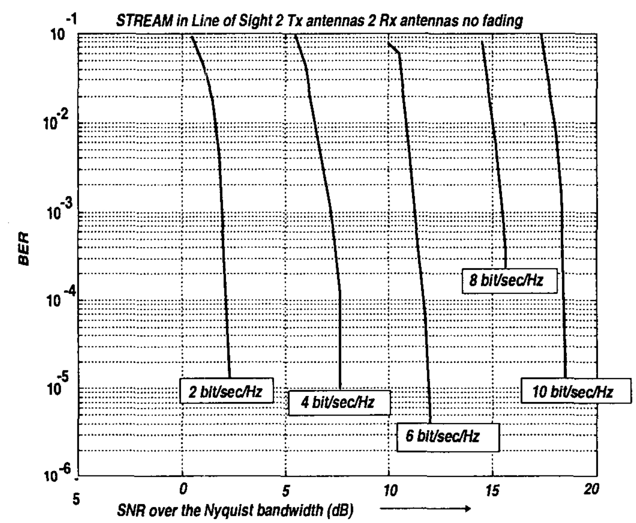

Fig. 3. $STREAM_{TM}$ results in Line of Sight for two transmitters and two receivers at 10 GHz.

link experiment. The bandwidth is set to 4 MHz and there are two antennas at the transmitter and two antennas at the receiver. The multipath environment is essentially Ricean and delay spread is less than 1 μsec. The specular component of the channel is 15 dB below the main signal. The results of Fig. 3 are impressive. At 2 and 4 bit/sec/Hz we are 2 and 2.5 dB from MIMO (Multiple-Input-Multiple-Output) capacity. At 6 and 8 dB we are 3-5 dB from MIMO capacity.

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