A Simple Indoor Microwave Channel Model With Realistic Motion Effects

A Progress Report and Draft for Discussion

John McKown, Motorola, and Larry Van Der Jagt, Knowledge Implementations, Inc.

INTRODUCTION

Standards committees currently addressing portable indoor digital radio systems have an immediate need for multipath channel simulation to investigate modulation, demodulation and multipath mitigation methods. Multipath is highly variable and notoriously laborious to characterize statistically in all the situations of interest. For the restricted purpose of preliminary design competition, however, it may be that accurate statistical characterization of multipath is not necessary as long as correct physical behavior is enforced. After all, even if enough data is collected to accurately characterize variations within factories, say, no one is likely to optimize against such detail.

This paper offers a rough-and-ready approach that eliminates measured data but preserves important motion effects. The model here is more appropriate for answering the question "which of two designs is better" than for "how will any particular system fare in the real world." We intend it to be useful to system designers using commercial digital simulation software.

Reference 1 presents the case that indoor multipath in the centimeter and millimeter bands is highly discrete (wide band power delay profiles are spikey) because walls and floors act as mirrors. In such cases, coherent ray tracing is a practical way to construct channel impulse responses and has the advantage that it ensures proper path phase changes due to motion.

Other channel models, crafted for analytic or computational simplicity, do not necessarily have this property. In particular, those which exploit the so-called *widesense stationary uncorrelated scattering assumptions* explicitly dispense with deterministic phase relations between the various echoes. This occurs through spatial averaging which is inappropriate for adaptive portable radios.

Below, we construct impulse responses by randomly positioning images of the transmitter. In effect, we are randomly positioning mirrors, i.e., walls, to generate the images. We do not, however, average the positions of these walls prior to using the channel.

MULTIPATH CHANNELS AS FIR FILTERS

The Friis transmission formula specifies the received power P_R at distance r from a transmitter of power P_T:

$$P_R = \left(\frac{\lambda}{4\pi r}\right)^2 g^T g^R P_T. \tag{1}$$

The g's are the antenna power gains at the appropriate angles. Therefore we consider channel impulse responses of form

$$h(t) = \sum_{n=0}^{N} \beta_n \delta(t - t_n) \tag{2}$$

where

$$\beta_n = \frac{\lambda \alpha_n \sqrt{g_n^T g_n^R}}{4\pi r_n}, \quad t_n = \frac{r_n}{c} \tag{3}$$

This expresses N+1 paths, one of which may be considered direct and the rest echoes from reflectors of intrinsic brightness, i.e., amplitude gain, α. This real channel filter is very wide-band. To use the complex baseband (or lowpass equivalent) representation of passband systems, we first convolve with some real passband system B of band width w. [2] The result HB can be brought to baseband by dropping the negative frequency components and then shifting by the carrier frequency f_c --- see Figure 1. If we take B(f) to be ideally flat, then the baseband transfer function C(f) is

$$\begin{aligned} C(f - f_c) &= 2H(f), f_c - \frac{w}{2} < f < f_c + \frac{w}{2} \\ &= 0, otherwise \end{aligned} \tag{4}$$

From Equation 2 we obtain

$$H(f)=\int_{-\infty}^{+\infty} e^{-2\pi jft}h(t)dt=\sum_{n=0}^{N}\beta_n e^{-2\pi jft_n} \tag{5}$$

From Equations 4 and 5,

$$C(f)=2\sum_{n=0}^{N}\beta_n e^{-2\pi j(f+f_c)t_n},\ -\frac{w}{2}<f<+\frac{w}{2}$$
$$=0, otherwise \tag{6}$$

The time-domain version of this filter is obtained with the inverse Fourier transform:

$$c(t)=\int_{-\infty}^{+\infty} e^{+2\pi jft}C(f)df$$
$$=2\sum_{n=0}^{N}\beta_n e^{-2\pi jf_c t_n}\frac{\sin(\pi w(t-t_n))}{\pi(t-t_n)} \tag{7}$$

We wish to apply this channel model in a discrete-time simulation. We choose the impulse-invariant method to convert equation 7 from continuous time to discrete time. We require that the sample interval T satisfies the Nyquist criterion T<1/w. This leads to the following FIR filter coefficients.

$$c_m=Tc(mT)$$
$$=2T\sum_{n=0}^{N}\beta_n e^{-2\pi jf_c t_n}\frac{\sin(\pi w(mT-t_n))}{\pi(mT-t_n)}$$
$$0\le m\le M-1 \tag{8}$$

The factor T scales such that if the input to the discrete-time filter consists of samples of the input to the continuous-time filter then the discrete output consists of samples of the continuous output. The number of taps, M, must be chosen large enough to avoid significantly truncating the last sinc function:

$$M\ge\frac{\max\{t_n\}}{T}+(\text{safety margin}) \tag{9}$$

It may be desirable to effect the discrete-time filter of equatior in the (discrete) frequency domain. However, this blockwise processing assumes that the time evolution of the channel, i.e., the time-dependence of t_n, can be ignored during the block. To check that t_n can be evaluated block-by-block rather than sample-by-sample, we first choose a phase error, θ, allowable between processing blocks of length N_{FFT}. Then, if v is the receiver's linear speed, we require

$$2\pi\frac{vN_{FFT}T}{\lambda}=2\pi\left(\frac{v}{c}\right)\left(\frac{f_c}{w}\right)N_{FFT}\le\theta \tag{10}$$

where the sample rate is assumed to be at the Nyquist limit. Note that as the band width w decreases, the processing blocks must get shorter. As an example, we take $\theta=2\pi/100$, f_c=3 GHz and w=1 MHz. Then at the typical pedestrian velocity of 1m/sec, relation 10 allows block lengths of up to about 1000 samples.

A PROBABILISTIC CHARACTERIZATION OF MULTIPATH CHANNELS.

The deterministic model above can be made representative of classes of buildings by choosing its parameters randomly fron appropriately defined ensembles. Our viewpoint will be that particular channels are deterministic (except for the effects of random motion) once their initial parameters are chosen. These initial values are to be chosen from probability densities that characterize the scenarios of interest.

The variables whose initial values are to be chosen randomly are

(a) the number of echoes N,
(b) the delays t_n and
(c) the brightnesses α_n.

We also provide for random motion of the receiver under the constraints that translations are within a horizontal plane and rotations are about a vertical axis. Our model will thus be two-dimensional; transmitter, receiver and all scatterers will be in a plane. Antenna patterns will be specified only as horizontal slices. The random variables which characterize the motion are

(d) the linear acceleration and
(e) the angular acceleration.

There are thus five kinds of random variables to be chosen from five ensembles. The first three (a through c above) characterize sites. They are randomly initialized and subsequently update by calculations which involve repeated draws from the last twc ensembles, which characterize motion. These updates will be performed sample by sample or block by block, depending on the implementation and the phase-continuity requirements discussed above.

Below, we rationalize (rather than derive) the five probability density functions. It is preferable, no doubt, to extract these densities from suitable data if it is available. In reality, of course, the brightnesses, delays and numbers of reflectors are probably not independent.

THE NUMBER OF REFLECTORS

N should depend on the density of reflectors in the environment. Experience and intuition agree that metal-walled factories are worse than plaster-walled homes and many offices are in between. We choose to trivialize the density for N to a delta function and assign names as follows:

residential	N=4
office	N=8
factory	N=16

DELAYS

Delays are bounded from below by the direct path. We take a typical maximum building dimension as an upper bound (thus ignoring full-length reverberations). We take the delay probability density to be uniform between these bounds. The upper bounds for the three cases are

residential	15m/c
office	50m/c
factory	100m/c

BRIGHTNESSES

The quantity α varies between -1 and zero as the virtual reflector varies from very shiny, e.g., a metal wall, to perfectly absorbent. We expect that more metal walls should make larger α's more likely. To express this expectation we adopt the blatantly nonphysical densities shown as Figure 2. When applied to the direct path, α other than unity can be interpreted as expressing a blocked line of sight.

ACCELERATIONS

We take the accelerations from zero-mean densities so that initial velocities are preserved (as averages) during the simulation. We choose Gaussians. Although the long tails occasionally permit impossibly large accelerations, we expect our sample rates will generally be high enough, i.e., individual accelerations will be applied briefly enough, that velocities will remain smooth and reasonable. The variances are set to provide mean acceleration magnitudes characteristic of an adult's whole-body motion:

	mean magnitude	variance
linear	$.38 m/sec^2$	$.32 m^2/sec^4$
rotational	$1 rad/sec^2$	$2.23 rad^2/sec^4$

SOFTWARE IMPLEMENTATION

We expect our channel model to be used more or less as illustrated in Figure 3, most likely within a commercial system simulation tool of the time-driven, block diagram-oriented, complex baseband, DSP variety (Acolade, Boss, Hypersignal, SPW, etc.). s1 is a signal of interest and s2 is an interferer. The two FIR filters are coupled by use of the same geometrical database (they use the same scatters in the same positions). Since the transmitters are in different locations, however, the two FIR filters are different.

We expect that the initial software which (eventually) expresses this channel model will consist of various C subroutines, ASCII input files and a demonstration program which exercises them. To use the model, a would-be simulator would modify the source code as necessary for use as a primitive in his preferred simulation package.

Since the channel filter is generally time dependent, it will probably be executed separately from static, band limiting filters in the simulation, rather than being convolved with them and executed as a composite (it will be a separate block in the block diagram). In many cases, users will want to insert this channel filter bock into an existing simulation which carries a single, global, sample rate several times higher than the Nyquist rate in order to generate smooth waveforms for a slicer (symbol identity decider). In such cases, w in Equation 8 can be chosen anywhere between the bandwidth provided by the band limiting filters and the bandwidth appropriate for the actual sample rate.

EDITABLE INPUT FILES

As presently envisaged, the main input file contains the following.

transmitter antenna pattern file name
receiver antenna pattern file name

transmitter antenna boresight direction
initial receiver antenna boresight direction
initial receiver location
initial receiver linear velocity
initial receiver angular velocity
receiver linear acceleration variance
receiver angular acceleration variance
carrier frequency
bandwidth
sample frequency
N1 (number of reflectors to randomly position)
maximum path length
brightness density hump
specified-reflector file name

The transmitter location is taken to be the coordinate origin. There are 3 other input files of 2 types. The antenna pattern files have 181 lines of form

degrees off boresight gain in dBi

which specify half of a symmetric pattern in 1 degree increments.

The specified-reflector file allows manual placement of reflectors and has the form

N2 (the number of reflectors in this file)
$x1, y1, \alpha 1$
$x2, y2, \alpha 2$
.
.
.
$xN2, yN2, \alpha N2$

SUBROUTINES

The channel can be expressed in a conventional way as various subroutines. An initialization routine reads the input files, checks the parameters for validity, calculates derivative quantities like the number of taps M and the FFT length. The outermost filtering routine would accept one complex input sample and return one complex output sample each time it was called. Perhaps on the very first call it would call the initialization routine. Output samples would all be zero until enough input samples to fill a processing block had arrived. Then the coefficients of equation 8 would be evaluated and FFTs called to perform the convolution. The block of output values would then be released one sample at a time.

THE ROLE OF CHANNEL MODELS IN STANDARDS

Channel models can play an important role in both the generation of a wireless LAN standard and in the use of that standard in a commercial environment. The role of the model varies as the standard progresses through its life cycle. In the standard writing stage, a medium model serves to benchmark proposed solutions and to predict the performance of those solutions prior to the availability of experimental data.

Once the standard is written, channel models can be used to test conformance to the standard. They provide an ensemble of test scenarios in which implementations must exhibit acceptable performance to be considered conformant. That is, channel models define the range of environmental characteristics over which interoperability of conformant implementations will be assured.

The problem of developing a standard for interoperable, multivendor, wireless LANs is substantially more difficult than the problem of developing a standard for use on wired media. The bulk of this difficulty arises from the fact that methods traditionally used to characterize and test wired media can not be applied to the time varying, unconfined media of wireless LANs.

One approach is summarized by the following statement. A conformant physical layer entity provides communication of minimum quality by exchanging conformant signals with a standard reference entity through a conformant medium. Conformant media, of course, must be defined in terms of channel models --- generally, time dependent ones.

This approach requires careful definition of the system used to verify conformance. In particular, signals must be exchanged with the antenna system which will be used in actual operation. Obviously, the room (or open field) in which this test is performed will affect the electric field in the vicinity of the antennas.

Generally, objects in the vicinity will affect both the delays and the angles of arrival of incident radio waves. It is certain that all implementors will not use the same antenna structure. Indeed, it is highly likely that intelligent antennas which adapt to the direction of arriving signals will be implemented. It is important to impose a repeatable, three dimensional, time varying electric field, which does not artificially favor particular designs, upon the equipment under test.

In light of this problem, we propose that conformance testing for wireless LANS and similar equipment be conducted as follows.

(1) A reference transceiver is defined and implemented in su[illegible] manner that it includes a channel simulator. The simulator is

pplied such that signals transmitted from the reference transmitter to the receiver of the equipment under test have the time characteristics of multipath.

(2) The reference transceiver and the equipment under test are placed in an anechoic chamber at sufficient separation that the equipment under test is illuminated by a single (quasi-)plane wave.

(3) Passing the test requires successfully exchanging specified data at a specified rate. The test must be passed using each antenna available to the candidate equipment (unless the antennas are used as an array).

Motion might be handled in at least two ways. The test procedure might allow the candidate equipment to be physically manipulated at will during the test so as to properly orient individual antennas. Alternatively, the candidate transceiver might be subjected to specified motions on a robot arm.

Channel models used for design or conformance testing must characterize not just multipath but also interference. In the 2.4-2.483 GHz band the dominant interference is that of microwave ovens. The radiation patterns from these ovens have been extensively studied. As is the case with propagation measurements, however, directionality and polarization have typically not been examined. Impulse interference may also require explicit treatment.

Obviously, there are many details yet to fix in a test procedure of this type. Given this basic structure, the degree to which conformant devices interoperate in the real world will depend on the degree to which real channels resemble the ones used in the conformance test.

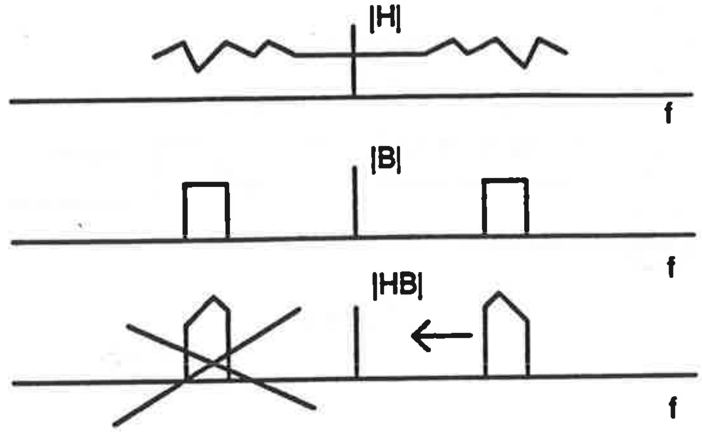

Figure 1

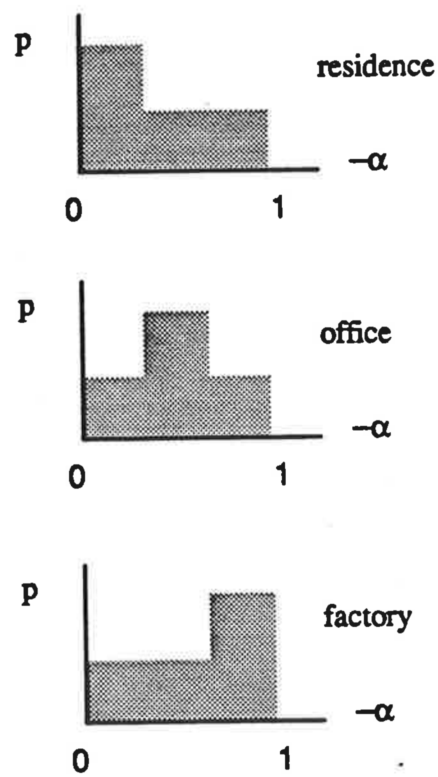

Figure 2.

REFERENCES

[1] J. W. McKown and R. L. Hamilton, "Ray Tracing as a Design Tool for Radio Networks," IEEE Network, vol. 5, no. 6, pp27-30, November 1991.

[2] M. C. Jeruchim, P. Balaban, K.C. Shanmugan, "Simulation of Communication Systems," ISBN 0-306-43989-1, Plenum Press, New York and London, 1992

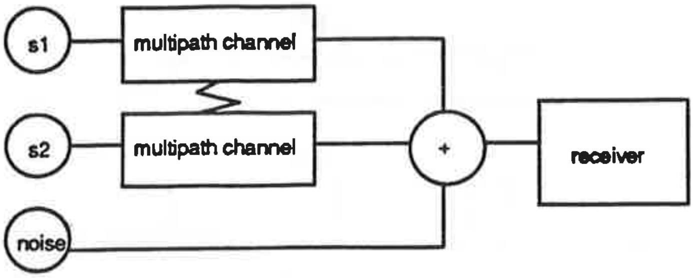

Figure 3