The Indoor Radio Propagation Channel

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In this tutorial-survey paper the principles of radio propagation in indoor environments are reviewed. Modeling the channel as a linear time-varying filter at each location in the three-dimensional space, properties of the filter's impulse response are described. Theoretical distributions of the sequences of arrival times, amplitudes and phases are presented. Other relevant concepts such as spatial and temporal variations of the channel, large scale path losses, mean excess delay and RMS delay spread are explored. Propagation characteristics of the indoor and outdoor channels are compared and their major differences are outlined. Previous measurement and modeling efforts are surveyed and areas for future research are suggested.

I. Introduction

The invention of telephone in the 19th century was the first step toward shattering the barriers of space and time in communication between individuals. The second step was the successful deployment of radio communications. To date, however, the *location* barrier has not been surmounted; i.e., people are more or less tied to telephone sets or "fixed wireline" equipment for communication. The astonishing success of cellular radio in providing telecommunication services to the mobile and handheld portable units in the last decade has paved the way toward breaking the location barrier in telecommunications. The ultimate goal of personal communication services (PCS) is to provide instant communications between individuals located anywhere in the world, and at any time. Realization of futuristic pocket-size telephone units and subsequent Dick Tracy wrist-watch phones are major communication frontiers. Industry and research organizations worldwide are collectively facing great challenges in providing PCS [1]–[25].

An important consideration in successful implementation of the PCS is indoor radio communications; i.e., transmission of voice and data to people on the move inside buildings. Indoor radio communication covers a wide variety of situations ranging from communication with individuals walking in residential or office buildings, supermarkets or shopping malls, etc., to fixed stations sending messages to robots in motion in assembly lines and factory environments of the future.

Network architecture for in-building communications are evolving. The European-initiated systems such as the digital European cordless telecommunications (DECT), and the cordless telecommunications second and third generations (CT2 and CT3) are primarily in-building communication systems [7], [13], [21], while the universal portable digital communications (UPDC) in the United States calls for a unification of the indoor and outdoor portable radio communications into an overall integrated system [1]–[3]. Practical portable radio communication requires lightweight units with long operation time between battery recharges. Digital communication technology can meet this requirement, in addition to offering many other advantages. There is little doubt that future indoor radio communication systems will be digital.

In a typical indoor portable radiotelephone system a fixed antenna (base) installed in an elevated position communicates with a number of portable radios inside the building. Due to reflection, refraction and scattering of radio waves by structures inside a building, the transmitted signal most often reaches the receiver by more than one path, resulting in a phenomenon known as multipath fading. The signal components arriving from indirect paths and the direct path (if it exists) combine and produce a distorted version of the transmitted signal. In narrow-band transmission the multipath medium causes fluctuations in the received signal envelope and phase. In wide-band pulse transmission, on the other hand, the effect is to produce a series of delayed and attenuated pulses (echoes) for each transmitted pulse. This is illustrated in Fig. 1, where the channel's responses at two points in the three-dimensional space are displayed. Both analog and digital transmissions also suffer from severe attenuations by the intervening structure. The received signal is further corrupted by other unwanted random effects: noise and cochannel interference.

Multipath fading seriously degrades the performance of communication systems operating inside buildings. Unfortunately, one can do little to eliminate multipath disturbances. However, if the multipath medium is well characterized, transmitter and receiver can be designed to "match" the channel and to reduce the effect of these disturbances. Detailed characterization of radio propagation is therefore

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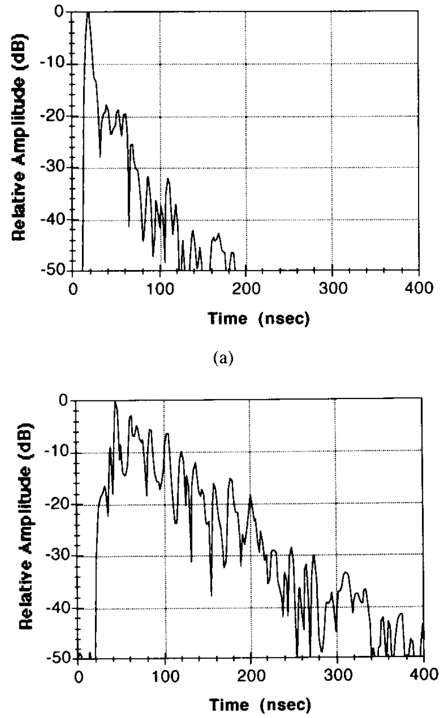

(b)

Fig. 1. The impulse responses for a medium-size office building. Antenna separation is 5 m. (a) Line of sight; (b) no line of sight. (Measurements and processing by David Tholl of TRLabs.)

a major requirement for successful design of indoor communication systems.

Although published work on the topic of indoor radio propagation channel dates back to 1959 [26], with a few exceptions, measurement and modeling efforts have all been carried out and reported in the past 10 years. This is in part due to the enormous worldwide success of cellular mobile radio systems, which resulted in an exponential growth in demand for wireless communications, and in part due to rapid advances in microelectronics, microprocessors, and software engineering in the past decade, which make the design and operation of sophisticated lightweight portable radio systems feasible.

A comprehensive list of measurement and modeling efforts for characterization of the analog and digital radio propagation within and into buildings are provided in references [26]–[208]. Extending the definition of indoor radio propagation to electromagnetic radiation within covered areas, mine and tunnel propagation modeling should also be included. These papers are listed in references [209]–[248]. (Reference [221] is a short review paper on the latter subject.)

The goal of this work is to provide a tutorial-survey coverage of the indoor radio propagation channel. Since the multipath medium can be fully described by its time and space varying impulse response, the tutorial aspect of this paper is based on characterization of the channel's impulse response. The general impulse response modeling of the multipath fading channel was first suggested by Turin [250]. It has been subsequently used in measurement, modeling, and simulation of the mobile radio channel by investigators following Turin's line of work [251]–[253], and by other researchers [254]–[259]. More recently, the impulse response approach has been used directly or indirectly in the indoor radio propagation channel modeling ([28]–[61], [64]–[71], [73], [74], [77]–[88], [97], [98], [117]–[124], [149], [188], [189], [191], [192], [196], [199], [200]).

After proper mathematical (the impulse response) formulation of the channel, other related topics such as channel's temporal variations, large-scale path losses, mean excess delay and rms delay spread, frequency dependence of statistics, etc., are addressed. The survey aspect of this paper reviews the literature. There are a number of important issues that either have not been addressed in the currently available measurement and modeling reports, or have received insufficient treatment. These areas are specified and directions for future research are provided. The survey covers papers published on the modeling of propagation as applied to portable radiotelephones or data services inside *conventional* buildings [26]–[208]. The mine and tunnel propagation papers are included for several reasons. The first reason is the similarities between some principles and applications. A good example is the leaky feeders ([99]–[101], [103], [104], [172], [173], for in-building, and [218]–[220], [227], [233], [235], [242], [244], for mine and tunnel propagation). The second reason is that a strong theoretical framework based on electromagnetic theory exists for mine and tunnel propagation (e.g., [212], [213], [225], [232], [234], [240], [248]) and not for the indoor office and residential building propagation. With a few exceptions, reported efforts on the latter subject are mainly directed toward measurements and statistical characterizations of the channel, with little emphasis on theoretical aspects. The interested researchers are encouraged to carry out a detailed comparison between the two types of propagation environments and bring out the points in common. Possible application of mine and tunnel propagation principles to some indoor environments is a challenging topic that will not be pursued in this report.

The main emphasis of this paper is on the tutorial aspect of the topic, although the survey aspect is also comprehensive. A general review of the indoor propagation measurement and models based on a totally different approach can be found in [27].

Finally, the indoor radio propagation modeling efforts can be divided in two categories. In the first category, transmission occurs between a unit located outside a building and a unit inside ([26], [89], [92]–[94], [112]–[114], [131], [132], [134], [136], [161], [164], [167], [168], [178], [183]). Expansion of current cellular mobile services to indoor applications and the unification of the two types of services has been the main thrust behind most of the measurements in this category. In the second category the transmitter and receiver are both located inside the building (the balance of references in [26]–[208]). Establishment of specialized indoor communication systems has motivated most of the researchers in this category. Although the impulse response approach is compatible with both, it has been mainly used for measurements and modeling efforts reported in the second category.

II. Mathematical Modeling of the Channel

A. The Impulse Response Approach

The complicated random and time-varying indoor radio propagation channel can be modeled in the following manner: for each point in the three- dimensional space the channel is a linear time-varying filter with the impulse response given by:

$$h(t,\tau) = \sum_{k=0}^{N(\tau)-1} a_k(t)\delta[\tau - \tau_k(t)]e^{j\theta_k(t)} \qquad (1)$$

where t and τ are the observation time and application time of the impulse, respectively, $N(\tau)$ is the number of multipath components, $\{a_k(t)\}, \{\tau_k(t)\}, \{\theta_k(t)\}$ are the random time-varying amplitude, arrival-time, and phase sequences, respectively, and δ is the delta function. The channel is completely characterized by these path variables. This mathematical model is illustrated in Fig. 2. It is a wide-band model which has the advantage that, because of its generality, it can be used to obtain the response of the channel to the transmission of *any* transmitted signal $s(t)$ by convolving $s(t)$ with $h(t)$ and adding noise.

The time-invariant version of this model, first suggested by Turin [250] to describe multipath fading channels, has been used successfully in mobile radio applications [251]–[253]. For the stationary (time-invariant) channel, (1) reduces to:

$$h(t) = \sum_{k=0}^{N-1} a_k\delta(t - t_k)e^{i\theta_k}. \qquad (2)$$

The output $y(t)$ of the channel to a transmitted signal $s(t)$ is therefore given by

$$y(t) = \int_{-\infty}^{\infty} s(\tau)h(t-\tau)\,d\tau + n(t) \qquad (3)$$

where $n(t)$ is the low-pass complex-valued additive Gaussian noise.

With the above mathematical model, if the signal $x(t) = \mathrm{Re}\{s(t)\exp[j\omega_0 t]\}$ is transmitted through this channel environment (where $s(t)$ is any low-pass signal and ω_0 is the carrier frequency), the signal $y(t) = \mathrm{Re}\{\rho(t)\exp[j\omega_0 t]\}$ is received where

$$\rho(t) = \sum_{k=0}^{N-1} a_k s(t - t_k)e^{j\theta_k} + n(t). \qquad (4)$$

In a real-life situation a portable receiver moving through the channel experiences a space-varying fading phenomenon. One can therefore associate an impulse response "profile" with each point in space, as illustrated in Fig. 3. It should be noted that profiles corresponding to points close in space are expected to be grossly similar because principle reflectors and scatterers which give rise to the multipath structures remain approximately the same over short distances. This is further illustrated in the empirical profiles of Fig. 4.

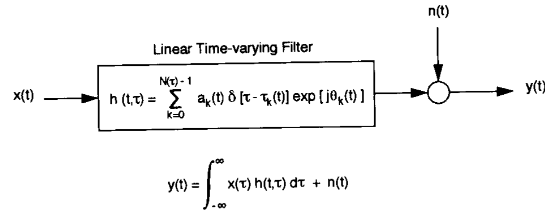

Fig. 2. Mathematical model of the channel.

B. The Discrete-Time Impulse Response

A convenient model for characterization of the indoor channel is the discrete-time impulse response model. In this model the time axis is divided into small time intervals called "bins." Each bin is assumed to contain either one multipath component, or no multipath component. Possibility of more than one path in a bin is excluded. A reasonable bin size is the resolution of the specific measurement since two paths arriving within a bin cannot be resolved as distinct paths. Using this model each impulse response can be described by a sequence of "0"s and "1"s (the path indicator sequence), where a "1" indicates presence of a path in a given bin and a "0" represents absence of a path in that bin. To each "1" an amplitude and a phase value are associated.

The advantage of this model is that it greatly simplifies any simulation process. It has been used successfully in the modeling [252] and the simulation [253] of the mobile radio propagation channel. Analysis of system performance is also easier with a discrete-time model, as compared to a continuous-time model.

C. Deduction of the Narrow-Band Model

When a single unmodulated carrier (constant envelope) is transmitted in a multipath environment, due to vector addition of the individual multipath components, a rapidly fluctuating CW envelope is experienced by a receiver in motion. To deduce this narrow-band result from the above wide-band model we let $s(t)$ of (4) equal to 1. Excluding noise, the resultant CW envelope R and phase $\emptyset$ *for a single point in space* are thus given by

$$\mathrm{R}e^{j\phi} = \sum_{k=0}^{\infty} a_k e^{j\theta_k}. \qquad (5)$$

Sampling the channel's impulse response frequently enough, one should be able to generate the narrow-band CW fading results for the receiver in motion, using the wide-band impulse response model.

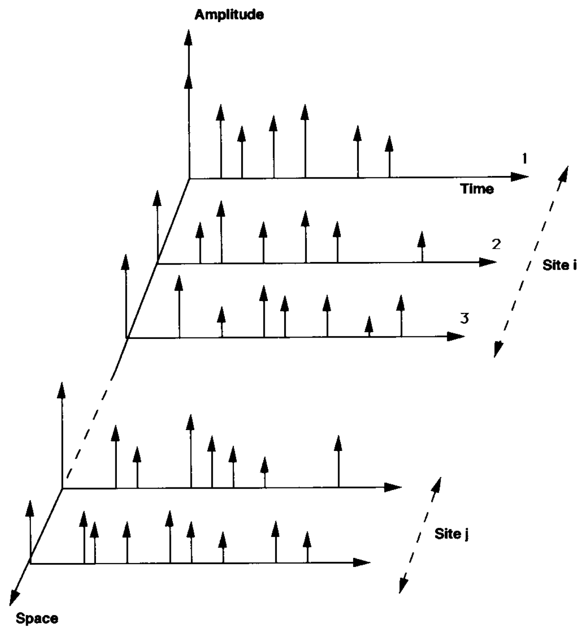

Fig. 3. Sequence of profiles for points adjacent in space.

III. Statistical Modeling of the Channel

A. A Model for Multipath Dispersion

The impulse response approach described in the previous section is supplemented with the geometrical model of Fig. 5. The signal transmitted from the base reaches the portable radio receivers via one or more *main waves*. These main waves consist of a line-of-sight (LOS) ray and several rays reflected or scattered by main structures such as outer walls, floors, ceilings, etc. The LOS wave may be attenuated by the intervening structure to an extent that makes it undetectable. The main waves are random upon arrival in the local area of the portable. They break up in the environment of the portable due to scattering by local structure and furniture. The resulting paths for each main wave arrive with very close delays, experience about the same attenuation, but have different phase values due to different path lengths. The individual multipath components are added according to their relative arrival times, amplitudes, and phases, and their random envelop sum is observed by the portable. The number of distinguished paths recorded in a given measurement, and at a given point in space depends on the shape and structure of the building, and on the resolution of the measurement setup.

The impulse response profiles collected in portable site i and portable site j of Fig. 3 are normally very different due to differences in the intervening (base to portable) structure, and differences in the local environment of the portables.

B. Variations in the Statistics

Let $X_{ijk}(i=1,2,\cdots,N; j=1,2,\cdots,M; k=1,2,\cdots,L)$ be a random variable representing a parameter of the channel at a fixed point in the three-dimensional space. For example, X_{ijk} may represent amplitude of a multipath component at a fixed delay in the wide-band model [a_k of (2)], amplitude of a narrow-band fading signal [R of (5)], the number of detectable multipath components in the impulse response [N of (2)], mean excess delay or delay spread (to be defined later), etc. The index k in X_{ijk} numbers spatially-adjacent points in a given portable site of radius 1–2 m. These points are very close (in the order of several centimeters or less). The index j numbers different sites with the same base-portable antenna separations, and the index i numbers groups of sites with different antenna separations. These are illustrated in Fig. 6.

With the above notations there are three types of variations in the channel. The degree of these variations depend on the type of environment, distance between samples, and on the specific parameter under consideration. For some parameters one or more of these variations may be negligible.

1) Small-Scale Variations: A number of impulse response profiles collected in the same "local area" or site are grossly similar since the channel's structure does not change appreciably over short distances. Therefore, impulse responses in the same site exhibit only variations in fine details (Figs. 3 and 4). With fixed i and j, $X_{ijk}(k = 1,2,\cdots,L)$ are correlated random variables for close values of k. This is equivalent to the correlated fading experienced in the mobile channel for close sampling distances.

2) Midscale Variations: This is a variation in the statistics for local areas with the same antenna separation (Fig. 6). As an example, two sets of data collected inside a room and in a hallway, both having the same antenna

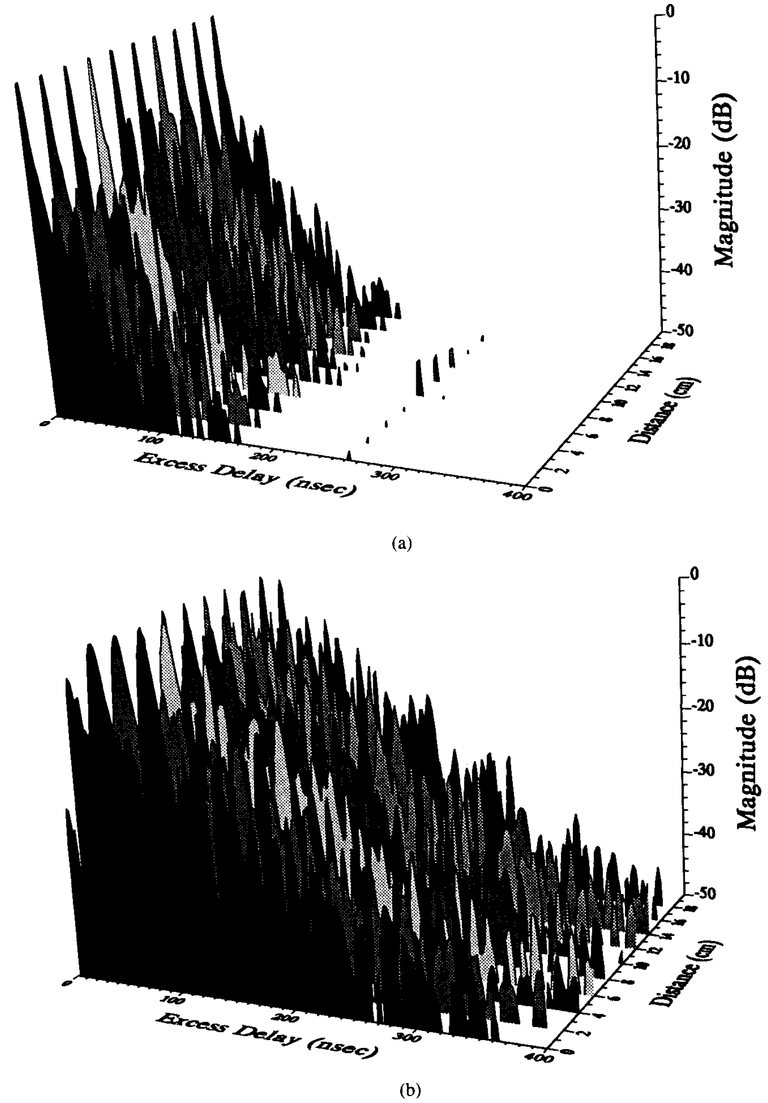

Fig. 4. Sequences of spatially adjacent impulse response profiles for a medium-size office building. Antenna separation is 5 m and center frequency is 1100 MHz. (a) Line of sight; (b) no line of sight. (Measurements by David Tholl of TRLabs, plotting by Daniel Lee of NovAtel.)

separation, may exhibit great differences. If μ_{ij} denotes the mathematical expectation of X_{ijk} (i.e., $\mu_{ij} = E_k(X_{ijk})$, where E_k denotes expectation with respect to k), then for fixed i, μ_{ij} is a random variable. For amplitude fading, this type of variation is equivalent to the shadowing effects experienced in the mobile environment. Different indoor sites correspond to intersections of streets, as compared to mid-blocks.

3) Large-Scale Variations: The channel's structure may change drastically, when the base-portable distance increases, among other reasons due to an increase in the number of intervening obstacles. As an example, for amplitude fading, increasing the antenna separation normally results in an increase in path loss. Using the previous terminology $\xi(d_i) = E_{jk}(X_{ijk}) = E_j(\mu_{ij})$ is different for different d_is (Fig. 6). If X_{ijk} denotes the amplitude, this type of variation is equivalent to the distance dependent path loss experienced in the mobile environment. For the mobile channel $\xi(d)$ is proportional to d^{-n}, where d is the base-mobile distance and n is a constant). Different path loss models for the indoor channel will be discussed in a subsequent section.

IV. Characterization of the Impulse Response

A. Distribution of the Arrival Time Sequence

1) General Comments: Although a number of investigators have adopted the impulse response approach to characterize the indoor radio propagation channel, with one

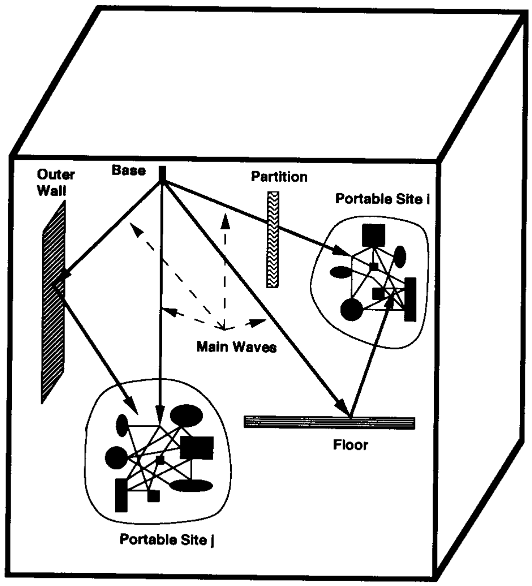

Fig. 5. A model for the radio propagation in indoor environments.

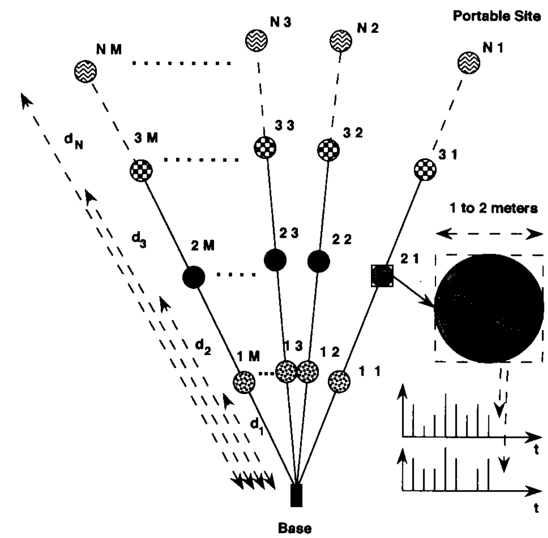

Fig. 6. Local and global variations.

exception (the work reported in [85]–[87]), distribution of the arrival time sequence has received insufficient attention.

The sequence of arrival times $\{t_k\}_0^\infty$ forms a point process on the positive time axis. Strictly speaking, the LOS path (if it exists) should be excluded from the sequence since its delay t_0 is not random. More appropriately, one should look at the distribution of $\{t_k - t_0\}_1^\infty$.

Several candidate point process models for the arrival time sequences are reviewed here.

2) Standard Poisson Model: As a preliminary model, one can postulate that the sequence of path arrival times $\{t_k - t_0\}_1^\infty$ follow a Poisson distribution. This distribution is encountered in practice when certain "events" occur with complete randomness (e.g., initiation of phone calls or occurrence of automobile accidents). In the indoor channel, if the obstacles which cause multipath fading are located with complete randomness in space, the Poisson hypothesis should be adequate to explain the path arrival times. If L denotes the number of paths occurring in a given interval of time of duration T, the Poisson distribution requires

$$\Pr(L = l) = \frac{\mu^l e^{-\mu l}}{l!} \tag{6}$$

where $\mu = \int_T \lambda(t)\,dt$ is the Poisson parameter. ($\lambda(t)$ is the mean arrival rate at time t.) For a stationary process (constant $\lambda(t))E[L] = \text{Var}[L] = \lambda$.

An important parameter in any point process is the distribution of interarrival times (defined as $x_i = t_i - t_{i-1}$, $i = 1, 2, \cdots$). For a standard (and stationary) Poisson process interarrival times are independent identically distributed (IID) random variables with an exponential distribution given by

$$f_X(x) = \lambda e^{-\lambda x} \qquad x > 0. \tag{7}$$

Analysis of measurement data collected in several indoor environments has established the inadequacy of the Poisson hypothesis to describe the arrival times [64], [65], [86], [87], [97], [98], [171]. It has also been observed that for insensitive receivers (i.e., for high threshold values) the Poisson fit to the data is relatively good. However, when threshold was lowered (weaker multipath components were included), deviations from the Poisson law were observed [32], [41].

Inadequacy of the Poisson distribution is probably due to the fact that scatterers inside a building causing the multipath dispersion are not located with total randomness. The pattern in location of these scatterers results in deviations from standard Poisson model, which is based on purely random arrival times.

3) Modified Poisson—The Δ-K Model: This model first suggested by Turin *et al.* [251] to describe the arrival time sequences in the mobile channel, has been fully developed by Suzuki [252]. It takes into account the clustering property of paths caused by the grouping property of scatterers (buildings in case of the mobile channel). The process is represented in Fig. 7. There are two states: $S-1$, where the mean arrival rate of paths is $\lambda_0(t)$, and $S-2$, where the mean arrival rate is $K\lambda_0(t)$. Initially, the process starts with $S-1$. If a path arrives at time t, a transition is made to $S-2$ for the interval $[t, t+\Delta)$. If no further paths arrive in this interval, a transition is made back to $S-1$ at the end of the interval. The model can therefore be explained by a series of transitions between the two states. Δ and K are constant parameters of the model, estimated using appropriate optimization techniques. For $K = 1$ or $\Delta = 0$, this process reverts to a standard Poisson process. For $K > 1$, incidence of a path at time t increases the probability of receiving another path in the interval $[t, t+\Delta)$, i.e., the process exhibits a clustering property. For $K < 1$, the incidence of a path decreases the probability

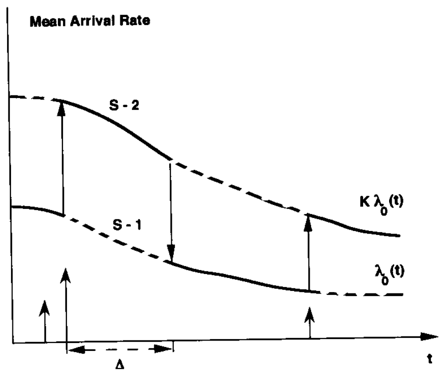

Fig. 7. The continuous-time modified Poisson process ($\Delta - K$ model).

of receiving another path, i.e., paths tend to arrive rather more evenly spaced than what a standard Poisson model would indicate.

A discrete version of this "$\Delta - K$" model has been successfully used to characterize and simulate path arrival times of the mobile channel [252], [253]. More recently, the model has been applied to limited indoor propagation data measured in several buildings [64], [65], and to a large data base consisting of 12 000 impulse response profiles obtained in two dissimilar office buildings [86], [87]. The fit has been very good. Most of the optimal K values, however, were observed to be less than 1, indicating that paths are more evenly distributed. Application of this model to impulse response data collected in several factory environments has not been successful [41].

The reported goodness of fit of the $\Delta - K$ model to the empirical data is due to one or both of two facts: 1) the phenomenological explanation given above, i.e., nonrandomness of the local structure; 2) the model uses more information from the data, as compared to the standard Poisson model. It should be noted that the $\Delta - K$ model uses empirical probabilities associated with individual small intervals of length Δ, while the standard Poisson model uses the total probability associated with a much larger interval T (normally, $T >> \Delta$). More details about the model can be found in [252].

4) Modified Poisson—Nonexponential Interarrivals: The IID exponential interarrivals give rise to a standard Poisson model. Other distributions can result in modifications of the Poisson process.

Extensive measurement data collected in several factory environments ([31], [34], [35]) were analyzed to construct a statistical model of the impulse response ([171], [179]). It was concluded that the Weibull interarrival distribution provides the best fit to the data, as compared to several other distributions. It should be noted, however, that there is no phenomenological explanation for choosing the Weibull interarrival distribution. A good Weibull fit is probably due to the fact that this distribution, in its most general form, has three parameters, increasing the flexibility to match the empirical data. For a specific choice of parameters the Weibull distribution reduces to an exponential distribution, and this model reverts to a standard Poisson model.

There seems to be no report in the literature investigating the independence of interarrivals.

5) The Neyman-Scott Clustering Model: The two-dimensional version of this model has been used in cosmology to study the distribution of galaxies [263], [264]. The process has cluster centers which follow a Poisson distribution, and elements in each cluster which also follow the Poisson law.

Empirical data collected in an office building has shown good fit to this double Poisson model [97], [98]. Clustering of paths was attributed to the building superstructure (such as large metal walls, doors, etc.), and multipath components inside each cluster were associated with multiple reflection from immediate environment of the portable [97], [98].

The above model is attractive and is consistent with the model of Fig. 5. Its available empirical verification mentioned above, however, is based on limited data. Its application to multipath data collected in several factory environments has also been unsuccessful [41].

6) Other Candidate Models: Using an extensive multipath propagation data base, validity of the $\Delta - K$ model for two office buildings has been established [86], [87]. It is recommended to investigate the distribution of the arrival times for other environments to determine the best fit model(s). Such an effort may reasonably include two other point process models, which have not been previously applied to the indoor propagation data. Both models are concerned with correlated events.

The first model is Gilbert's burst noise model, used to describe nonindependent error occurrences in digital transmission [265], [266]. The model has two states. State 1 corresponds to error free transmission. In state 2, errors occur with a preassigned probability. There is transition between these states. Such transitions, however, are independent of the events.

The second model is a pseudo-Markov model used to describe spike trains from nerve cells [267], [268]. This model also has two states $S - 1$ and $S - 2$. Interarrival distributions are assigned to events occurring in each state; the distributions may be different. A transition is made from $S-1$ to $S-2$ after occurrence of N_1 events, and from $S-2$ to $S - 1$ after N_2 events, N_1 and N_2 themselves being random variables with given distributions.

In both of the above models transition between the states is independent of the events, as opposed to the $\Delta - K$ model, in which an event causes a transition.

B. Distribution of Path Amplitudes

1) General Comments: In this section the distribution of path amplitudes is investigated. In a multipath environment if the difference in time delay of a number of "paths" (echoes) is much less than the reciprocal of the transmission bandwidth, the paths can not be resolved as distinct pulses. These unresolvable "subpaths" add vectorially (according to their relative strengths and phases), and the envelope of their sum is observed. The envelope value is therefore a

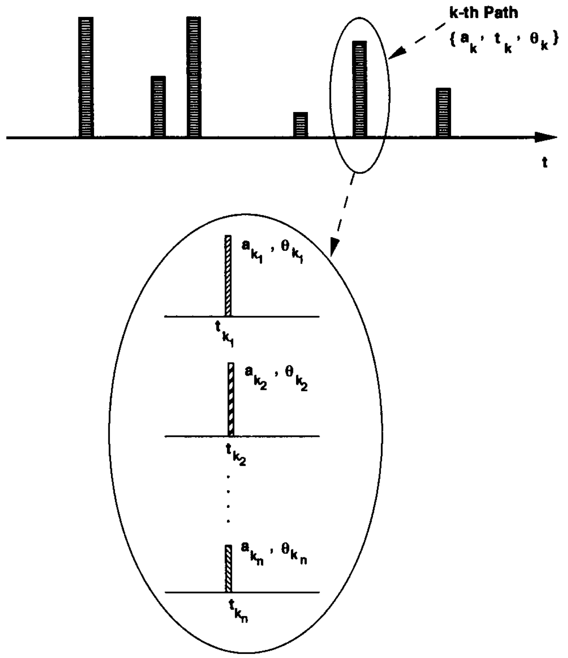

Fig. 8. A multipath component and its associated subpaths.

random variable. This is illustrated in Fig. 8. Mathematically, if $t_{k_i} - t_{k_j} < 1/W$, $i, j = 1, 2, \cdots, n$, where W is the transmission bandwidth, then

$$a_k e^{j\theta_k} = \sum_{i=1}^{n} a_{k_i} e^{j\theta_{k_i}} \tag{8}$$

is the resolved multipath component.

For ease of notation let $r = a_k$ for any k. In what follows r can also denote the CW fading envelope [R of (5)]. With proper interpretation, the definition of r may be extended to the narrow-band or wide-band temporal fading (i.e., variations in the signal amplitude when both antennas are fixed; such variations are due to the motion of people and equipment in the environment). If the latter definition is adopted, "spatial" separation between data points in this section should be replaced with "temporal" separations.

Amplitude fading in a multipath environment may follow different distributions depending on the area covered by measurements, presence or absence of a dominating strong component, and some other conditions. Major candidate distributions are described below.

2) The Rayleigh Distribution: A well-accepted model for small-scale rapid amplitude fluctuations in absence of a strong received component is the Rayleigh fading with a probability density function (pdf) given by

$$\Pr(r) = \frac{r}{\sigma^2} \exp\{-\frac{r^2}{2\sigma^2}\}, \qquad r \geq 0 \tag{9}$$

where σ is the Rayleigh parameter (the most probable value). The mean and variance of this distribution is $\sqrt{\pi/2}\,\sigma$ and $(2 - \pi/2)\sigma^2$, respectively.

The Rayleigh distribution is widely used to describe multipath fading because of its elegant theoretical explanation and occasional empirical justifications. To describe it theoretically, one can use the celebrated model of Clarke for the mobile channel [269]. In this model it is assumed that the transmitted signal reaches the receiver via N directions, the ith path having a complex strength $r_i e^{j\theta_i}$ that can be described by a phasor with an envelope r_i and a phase θ_i. At the receiver these signals are added vectorially and the resultant phasor is given by:

$$re^{j\theta} = \sum_i r_i e^{j\theta i}. \tag{10}$$

Clarke assumes that over small areas and in absence of a line-of-sight path, the r_is are approximately equal ($r_i = r'$, $i = 1, 2, \cdots, N$), and hence,

$$re^{j\theta} = r' \sum_i e^{j\theta i}. \tag{11}$$

The path phase θ_i is very sensitive to the path length, changing by 2π when the path length changes by a wavelength (which is fraction of a meter at UHF frequencies). Therefore, phases are uniformly distributed over $[0, 2\pi)$ and the problem reduces to obtaining distribution of the envelope sum of a large number of sinusoids with constant amplitude and uniformly-distributed random phases. Quadrature components I and Q of the received signal are independent, and, by the central limit theorem, Gaussianly distributed random variables. The joint distribution of $r(= \sqrt{I^2 + Q^2})$ and $\theta(= \arctan[Q/I])$ was first investigated by Lord Rayleigh [270]. The result is that r and θ are independent, r being Rayleigh-distributed and θ having a uniform $[0, 2\pi)$ distribution. A short derivation can be found in [271].

It has been shown that even when as few as six sine waves with uniformly distributed and independently fluctuating phases are combined, the resulting amplitude and phase follow very closely the Rayleigh and uniform distributions, respectively [272].

The assumption that r_is are equal is unrealistic since it implies the same attenuation for each path. It has been shown, however, that if the magnitudes are not equal but any single one of them does not contribute a major fraction of the received power (i.e., if $r_i^2 << \Sigma r_i^2, i = 1, 2, \cdots, N$), then the Rayleigh distribution can still be used to describe variations of the resulting amplitude.

There are several reported empirical justifications for application of the Rayleigh distribution to the indoor propagation data. Extensive CW measurements in five factory environments has shown that small scale fading is primarily Rayleigh, although the Rician fading also described some LOS paths [31]. However, when only signal levels below the median were considered, the distribution appeared to be lognormal. Analysis of the wide-band data collected in the same factory environments indicate that for heavy clutter situations amplitude of the multipath components are Rayleigh distributed [171]. Wide-band propagation data in an office building has shown better Rayleigh than lognormal fit [97], [98]. The data, however, were limited, and the Rayleigh fit was observed only after "inflating" (i.e., increasing the number of) weaker components. Wide-band [178], [183], and narrow-band CW [89], [92], [183] measurements with either the transmitter or the receiver located outside the building and the other located inside, has shown a good Rayleigh fit to the collected data. CW measurements with both antennas inside buildings have shown Rayleigh characteristics [106], [107], and Rice or Rayleigh distributions depending on the presence or absence of a LOS path [175]. CW measurements in an office building at 900 MHz by one investigator [197], and at 21.6 GHz and 37.2 GHz in a university campus building by another [143] showed good Rayleigh fit to the fast fading component. CW measurements at 1.75 GHz showed that when the transmission path was obstructed by human body, the fading statistics was Rayleigh, and for the LOS cases it was Rician [148]. Finally, CW measurements at 900 MHz, 1800 MHz, and 2.3 GHz showed that the small scale variations were Rayleigh distributed [134].

3) The Rician Distribution: The Rician distribution occurs when a strong path exists in addition to the low level scattered paths. This strong component may be a line-of-sight path or a path that goes through much less attenuation compared to other arriving components. Turin calls this a "fixed path" [250]. When such a strong path exists, the received signal vector can be considered to be the sum of two vectors: a scattered Rayleigh vector with random amplitude and phase, and a vector which is deterministic in amplitude and phase, representing the fixed path. If $ue^{j\alpha}$ is the random component, with u being Rayleigh and α uniformly distributed, and $ve^{j\beta}$ is the fixed component (v and β are not random), then the received signal vector $re^{j\theta}$ is the phasor sum of the above two signals. Rice [273] has shown the joint pdf of r and θ to be

$$\Pr(r,\theta) = \frac{r}{2\pi\sigma^2}\exp\{-\frac{r^2+v^2-2rv\,\cos(\theta-\beta)}{2\sigma^2}\}$$
$$r \geq 0,\quad -\pi \leq (\theta\,-\beta) \leq \pi. \quad (12)$$

Furthermore, since the length and phase of the fixed path usually changes, β is itself a random variable uniformly distributed on $[0, 2\pi)$. Randomizing β causes r and θ to become independent, θ having a uniform distribution and r having a Rician distribution given by the pdf

$$\Pr(r) = \frac{r}{\sigma^2}\exp\{\frac{-r^2+v^2}{2\sigma^2}\}I_o(\frac{rv}{\sigma^2}), r \geq 0 \qquad (13)$$

where I_0 is the zeroth-order modified Bessel function of the first kind, v is the magnitude (envelope) of the strong component and σ^2 is proportional to the power of the "scatter" Rayleigh component.

In the above equation if v goes to zero (or if $v^2/2\sigma^2 << r^2/2\sigma^2$), the strong path is eliminated and the amplitude distribution becomes Rayleigh, as expected. Therefore, the Rician distribution contains the Rayleigh distribution as a special case. On the other hand, if the fixed path vector has a length considerably longer than the Rayleigh vector (power in the stable path is considerably higher than the combined random paths), r and θ are both approximately Gaussian, r having a mean equal to v and θ having zero mean. That is, in this case, the Rician distribution is well approximated around its mode by a Gaussian distribution.

Analysis of local wide-band data in several factory environments has shown that over "certain range of signal amplitudes" the Rician distribution shows good fit [171]. Extensive temporal fading data (i.e., measurements with both antennas stationary) collected by one investigator indicates that even in the absence of a LOS path, the Rician distribution shows much better fit to the data, as compared to the Rayleigh distribution [77]–[79]. Similar results have been reported for CW temporal fading measurements at several factory environments by another investigator [31], [34]. Analysis of CW data over a number of buildings using both leaky feeders and dipole antennas has shown the signal envelope to be "weakly Rician" [172]. Also, CW measurements inside a university building [175], and in an office environment [148] has indicated that when a LOS path between the transmitter and receiver exists, the envelope data follow the Rician distribution. Finally, CW measurements at 21.6 GHz and 37.2 GHz with directional transmitting antennas indicated that amplitude fading was "close to Rician" [143].

4) The Nakagami Distribution: This distribution (also called the m-distribution), which contains many other distributions as special cases, has been generally neglected, perhaps because the Nakagami's works are mostly written in Japanese.

To describe the Rayleigh distribution, the length of the scatter vectors were assumed to be equal and their phases to be random. A more realistic model, proposed by Nakagami [274], also permits the length of the scatter vectors to be random. Using the same notation we have $r = |\Sigma r_i e^{j\theta_i}|$. The Nakagami-derived formula for the pdf of r is

$$\Pr(r) = \frac{2m^m r^{2m-1}}{\Gamma(m)\Omega^m}\exp\{-\frac{mr^2}{\Omega}\}, r \geq 0 \qquad (14)$$

where $\Gamma(m)$ is the Gamma function, $\Omega = E\{r^2\}$ and $m = \{E[r^2]\}^2/\text{Var}[r^2]$, with the constraint $m \geq 1/2$. Nakagami is a general fading distribution that reduces to Rayleigh for $m = 1$ and to the one-sided Gaussian distribution for $m = 1/2$. It also approximates, with high accuracy, the Rician distribution, and approaches the lognormal distribution under certain conditions.

A search of the literature indicates that application of this distribution to indoor radio propagation data has been generally neglected. One investigator has applied it to the analysis of global (large area) data, with the conclusion that the other distributions tested (Suzuki and lognormal) show better fit [112]–[114]. Simulations of the CW envelope fading based on analytical ray tracing techniques (i.e., no measurements) showed that the fast fading component was Nakagami-distributed [202].

5) The Weibull Distribution: This fading distribution has a pdf given by

$$\Pr(r) = \frac{\alpha b}{r_0}\left(\frac{br}{r_0}\right)^{\alpha-1}\exp\left[-\left(\frac{br}{r_0}\right)^{\alpha}\right], \qquad r \geq 0 \qquad (15)$$

where α is a shape parameter, r_0 is the rms value of r, and $b = [(2/\alpha)\Gamma(2/\alpha)]^{1/2}$ is a normalization factor [275].

There is no theoretical explanation for encountering this type of distribution. However, it contains the Rayleigh distribution as a special case (for $\alpha = 1/2$). It also reduces to the exponential distribution for $\alpha = 1$. The Weibull distribution has provided good fit to some mobile radio fading data [276]. Narrow-band measurements at 910 MHz in several laboratories with both antennas stationary showed that the Weibull distribution accurately described fading during the periods of movement [72]. A search of the literature shows no other direct empirical justification for application of this distribution to the indoor data.

6) The Lognormal Distribution: This distribution has often been used to explain large scale variations of the signal amplitudes in a multipath fading environment. The pdf is given by

$$\Pr(r) = \frac{1}{\sqrt{2\pi}\sigma r}\exp\{-(ln\, r - \mu)^2/2\sigma^2\}, \qquad r \geq 0. \qquad (16)$$

With this distribution, log r has a normal (Gaussian) distribution. There is overwhelming empirical justification for this distribution in urban and ionospheric propagation. A heuristic theoretical explanation for encountering this distribution is as follows: due to multiple reflections in a multipath environment, the fading phenomenon can be characterized as a multiplicative process. Multiplication of the signal amplitude gives rise to a lognormal distribution, in the same manner that an additive process results in a normal distribution (the central limit theorem).

A key assumption in the theoretical explanation of the Rayleigh and Rician distribution was that the statistics of the channel do not change over the small (local) area under consideration. This implies that the channel must have spatial homogeneity for Rayleigh and Rician distributions to apply. Measurements over large areas, however, are subject to another random effect: changes of the parameters of the distributions. This spatial inhomogeneity of the channel seems to be directly related to the transition from Rayleigh distribution in local areas to lognormal distribution in global areas [252].

A study of the indoor radio propagation modeling reports reveals that with one exception ([86], [87]) there is no direct reference to the global statistics of path amplitudes. The fact that the *mean* of local data are lognormal, however, seems to be well established in the literature (impulse response data collected in several factory environments [41], and CW data recorded inside several buildings with the transmitter placed outside the building [26], [92], [93], [131], [132]). The good lognormal fit has also been observed for some local data (small number of profiles in each location at several factory environments [33], [36], [41], [179], CW fading data for obstructed factory paths [31], and limited wide-band data at several college buildings [65]). In one set of CW measurements "local short time fluctuations" of the signals were measured (with transmitter and receiver stationary during the measurements) [70]. The results showed better lognormal fit than Rayleigh fit to the "local" temporal fading data. CW measurements in a modern office building at 900 MHz showed that the "room-related slow fading" was lognormally distributed [197]. Large scale variations of data collected at 900 MHz, 1800 MHz, and 2.3 GHz for transmission into and within buildings were found to be lognormal [134].

The strongest empirical justification for applicability of the lognormal distribution to indoor data have been reported in [85]–[87]. The data base for these measurements consists of 6000 impulse response profiles collected at each one of two office buildings. Four transmitter-receiver antenna separations of 5, 10, 20, and 30 m were considered, twenty locations per antenna separation were visited, and for each location 75 profiles at sampling distances of 2 cm were recorded. (The measurement plan was based on the channel variations depicted in Fig. 6., with $N = 4$, $M = 20$, $L = 75$, $d_1 = 5$ m, $d_2 = 10$ m, $d_3 = 20$ m, and $d_4 = 30$ m.) Analysis of this extensive data base indicated that distribution of the multipath components' amplitude is lognormal for both local and global data [86], [87]. Local data consist of all profiles recorded in one location (i.e., 75 profiles), and global data consist of all profiles for one antenna separation, i.e., 1500 profiles.

7) The Suzuki Distribution: This is a mixture of the Rayleigh and Lognormal distributions, first proposed by Suzuki [252] to describe the mobile channel. It has the pdf

$$\Pr(r) = \int_0^{\infty} \frac{r}{\sigma^2}\exp\left(-\frac{r^2}{2\sigma^2}\right)\frac{1}{\sqrt{2\pi}\sigma\lambda} \cdot \exp\left[-\frac{(ln\sigma - \mu)^2}{2\lambda^2}\right] d\sigma. \qquad (17)$$

This distribution, although complicated in form, has an elegant theoretical explanation: one or more relatively strong signals arrive at the general location of the portable. The main wave, which has a lognormal distribution due to multiple reflections or refractions, is broken up into subpaths at the portable site due to scattering by local objects. Each subpath is assumed to have approximately equal amplitudes and random uniformly distributed phases. Furthermore, they arrive at the portable with approximately the same delay. The envelope sum of these components has a Rayleigh distribution with a lognormally distributed parameter σ, giving rise to the mixture distribution of (17).

The Suzuki distribution phenomenologically explains the transition between the local Rayleigh distribution to the global lognormal distribution. It is consistent with the model depicted in Fig. 5. It is, however, complicated for data reduction since its pdf is given in an integral form. A successful application of this distribution for the mobile channel has been reported in [277]. A review of indoor propagation papers indicate that it has been generally neglected (probably because of the complexity of data reductions). Its only reported application is by one investigator for CW data collected with the transmitter located outside and the receiver placed inside different floors of a building [112]- [114]. Application of the Rayleigh, Weibull, Nakagami, lognormal, and Suzuki distributions to the large area data showed good Suzuki and lognormal fits (Suzuki better). For one set of data, the optimum μ and λ [(17)] were found to be 6.7 dB and 1.4 dB, respectively [113], [114].

C. Distribution of Path Phases

1) General Comments: Performance of digital indoor communication systems is very sensitive to statistical properties of the phase sequence $\{\theta_k\}_0^\infty$. Although the importance of this issue has been recognized by most investigators, a comprehensive search of the literature shows that, to date, no *empirically driven* model for the phase sequence has been reported. This is probably due to difficulties associated with measuring the phase of individual multipath components. (Recording of signal phase is incompatible with some measurement techniques.)

The signal phase is critically sensitive to path length and changes by the order of 2π as the path length changes by a wavelength (30 cm at 1 GHz). Considering the geometry of the paths, moderate changes (in the order of meters) in the position of the portable results in a great change in phase. When one considers an ensemble of points, therefore, it is reasonable to expect a Uniform $[0, 2\pi)$ distribution; i.e., on a global basis, θ_k has a $U[0, 2\pi)$ distribution. This phenomenologically reasonable assumption can be taken as a fact with no need for empirical verification. For small sampling distances, however, great deviations from uniformity may occur. Furthermore, phase values are strongly correlated if the channel's response is sampled at the symboling rates (tens to hundreds of kilobits per second). Phase values at a fixed delay for a given site (Figs. 3-6) are, therefore, correlated. Adjacent detectable multipath components of the same profile, on the other hand, have independent phases since their excess range (excess delay multiplied by the speed of light) is larger than a wavelength, even for very high resolution (a few nanoseconds) measurements.

Taking the above into consideration, it is accurate to say that the absolute phase value of a multipath component at a fixed point in space is not important; emphasis of the modeling should be placed on *changes* in phase as the portable moves through the channel. Let $\theta^{(m)}$ denote the phase of a multipath component at a fixed delay for profile number m, where $m = 1, 2, 3, \cdots$ numbers adjacent points in space at a given site. Equivalently $\theta^{(m)}$ may denote the phase of a multipath component occupying a given bin (the discrete-time impulse response model) at spatial point m. For the first profile in a sequence ($m = 1$), $\theta^{(m)}$ is assumed to have a $U[0, 2\pi)$ distribution. Subsequent phase values are assumed to follow the following relation:

$$\theta^{(m)} = \theta^{(m-1)} + \emptyset(s_m/\lambda), \qquad m = 2, 3, \cdots \tag{18}$$

where s_m is the spatial separation between the $(m-1)$st and mth profiles, λ is the wavelength, and $\emptyset(s_m/\lambda)$ is a phase increment. On a sequence of spatially separated profiles, the chain of values defined by (18) is interrupted when a path at a given excess delay time (or at a given bin) ceases to exist. A new chain of values (with uniformly distributed first component) starts if a path with the same excess delay appears at a later profile.

Appropriate choices for $\emptyset(s_m/\lambda)$ will impose the necessary spatial correlation on phase values. Using this approach two models for this phase increment may be considered.

2) The Random Phase Increment Model: $\emptyset(s_m/\lambda)$ is a random variable; i.e., starting with a $U[0, 2\pi)$ initial phase, each subsequent value is obtained by adding a random phase increment to the previous phase value. Parameters of the probability distribution of this increment are functions of s_m/λ. As an example, one can assume $\emptyset(s_m/\lambda)$ to be a Gaussianly distributed random variable with zero mean and standard deviation $\sigma_{s/\lambda}$. By making $\sigma_{s/\lambda}$ an increasing function of s/λ (or s, for a fixed λ) one can control the degree of correlation between $\theta^{(m-1)}$ and $\theta^{(m)}$. For $s = 0$, $\sigma_{s/\lambda} = 0$, and $\theta^{(m-1)} = \theta^{(m)}$ (assuming a time-invariant channel). The correlation between $\theta^{(m-1)}$ and $\theta^{(m)}$ decreases as $\sigma_{s/\lambda}$ increases, until they become uncorrelated.

The functional form of $\sigma_{s/\lambda}$ (or for that matter, the probability density function of $\emptyset(s_m/\lambda)$) is unknown. Using a large data base for phase, it may be possible to determine these unknowns. A possible method is to simulate the above model using various choices for the probability distribution of $\emptyset(s_m/\lambda)$ and the functional form of its moments, until the small-scale and large-scale statistics of the experimental data are reproduced.

The above model (with Gaussian phase increment and exponential $\sigma_{s/\lambda}$) has been previously applied in simulation of the phase components of a wide-band mobile radio channel model [253]. To date, however, there is no report on its application to the indoor channel.

3) The Deterministic Phase Increment Model: In this model changes in the phase value of a multipath component at a fixed delay when the portable moves through space is not random; i.e., knowing $\theta^{(1)}$ and $\emptyset(s_m/\lambda)$, $\theta^{(m)} (m = 2, 3, \cdots)$ can be calculated deterministically.

To use the above model some simplifying assumptions should be made in order to reduce the degree of randomness of the channel. In one such application, it was *assumed* that in a length of one meter in space, all multipath components with the same delay are caused by reflection from the same fixed (but randomly located) scatterer [37], [41]. In this simulation application the initial phase was generated according to a $U[0, 2\pi)$ distribution. Other spatially separated phases (with the same delay) were obtained by adding $\emptyset(s_m/\lambda)$ to the previous phase. The phase increment was calculated using the single scatterer and the local geometry [37], [41]. This is a one-hop model which excludes multiple reflections. This phase model was used in a simulation package for predicting the impulse response of open plant and factory environments. Since the phase of individual multipath components was not measured in the original experimentation, justification for the above approach was provided by generating the narrow-band CW fading data from the wide-band channel model [(5)], and comparing the results with the measured data reported in [31]. The general characteristics (i.e., the periodic spacing of the nulls) were found to be similar [37], [41].

In two other simulation applications, the deterministic phase model has been used for the indoor channel [118] and the mobile channel [278]. In both applications, it was *assumed* that the angle of arrival of the kth multipath component with respect to the direction of motion of the vehicle (portable) Ψ_k remains the same for small spatial separations. Therefore:

$$\emptyset(s_m, \lambda) = \frac{2\pi s_m}{\lambda} \cos \Psi_k. \tag{19}$$

For the mobile channel, $\Psi_k(k = 1, 2, 3, \cdots)$ were generated according to a uniform distribution. The power spectra of simulated CW data generated using a wide-band channel simulator (reported in [253]) showed better agreement with theory [278], when compared with the spectra obtained using the random phase increment model. For the indoor channel, $\Psi_k(k = 1, 2, 3, \cdots)$ were estimated with a 5° resolution based on measurements reported in [117] and by using the Fourier transform method (details are reported in [118]).

It is important to note that neither one of the above two models are derived from empirical data. Justifications for each have , therefore, been provided indirectly. There is no theoretical basis for choosing the Gaussian phase increment in the first model. Assumption of a one-hop reflection for each multipath component caused by a single scatterer (which remains the same with the motion of the portable) is also too simplistic for the complicated indoor channel, and violates the realistic multipath dispersion model of Fig. 5. Derivation of a phase model from actual measurement data is therefore strongly recommended.

D. Interdependences within Path Variables

1) Correlations Within a Profile: Adjacent multipath components of the same impulse response profile are likely to be correlated. However, with exceptions to be noted below, the existence and the degree of this type of correlation has not been established yet.

Correlation between the arrival times, if it exists, is due to the grouping property of local structures. The $\Delta - K$ model described before is an example of a correlated arrival time model. Furthermore, the echo pattern dies out with time; i.e., the probability of receiving echoes decreases with increasing the excess delay for large delays since multipath components go through higher path losses and become less detectable at larger delays.

For high resolution measurements amplitudes of adjacent multipath components of the same profile are likely to be correlated since a number of scattering objects that produce them may be the same. Analysis of impulse response data collected in several factory environments has shown that for the LOS topography correlation on amplitudes of initial paths is small. More specifically, amplitude of the LOS path is uncorrelated with amplitude of the path at 8 ns, anticorrelated (negative correlation coefficient) with the component at 16 ns, and uncorrelated with subsequent multipath components [33], [41]. For later echoes (excess delays greater than zero), two components have uncorrelated amplitudes if their time difference is greater than 25 ns. For the obstructed (i.e., non-LOS) topography, amplitudes become uncorrelated for excess delay differences greater than 25 ns, independent of the actual excess delay. The overall conclusion is that path component amplitudes are correlated only if they arrive within 100 ns of each other [41].

It should be noted that in the above-mentioned work the arrival times are modeled as independent events; i.e., paths at each bin arrive with different probabilities, but independent of each other [41]. This seems to be inconsistent with correlated amplitude fading since independent arrival times implies independent scattering and hence uncorrelated amplitudes at all excess delays. Analysis of extensive multipath propagation data collected in two office environments has shown small correlation between the amplitude of adjacent multipath components of the same profile. Typical correlation coefficients were between 0.2 and 0.3 [85]–[87].

It is reasonable to assume that at frequencies of interest (above 1 GHz) phase components of the same impulse response profile are uncorrelated since their relative excess range (excess delay multiplied by the speed of light) is larger than a wavelength, even for high-resolution measurements.

The amplitude sequence is correlated with the arrival time sequence because later paths of a profile experience higher attenuations due to greater path lengths and possibility of multiple reflections.

There is no reason to believe that the phase sequence is correlated with the arrival-time sequence, or with the amplitude sequence.

2) Correlations between Spatially Separated Profiles: A number of impulse response profiles collected in the same "local" area are grossly similar since the channel's structure does not change appreciably over short distances. This can be observed in Fig. 4.

"Spatial" correlations (i.e., correlation between impulse response profiles at points close in space) govern the amplitudes, arrival times, and phases. The degree of these correlations, however, are likely to be different. A study of the published works shows that with two exceptions to be noted below, spatial correlations have not been quantified.

Inspection of impulse response profiles collected in several factory environments over one meter (4.3 λ) tracks show very small variations between the amplitudes of consecutive profiles [35]. More specifically, the LOS path varied by no more than one or two dB over the entire track for most cases. It was observed that for the LOS topography and for excess delays less than 100 ns, amplitudes are uncorrelated at spatial separation of $\lambda/2$ and "slightly anticorrelated" for 2.5λ–3λ (λ = 23 cm). For excess delays beyond 100 ns, amplitudes were uncorrelated for spatial separations greater than $\lambda/2$. The obstructed topography, on the other hand, showed average correlation coefficients close to zero for all spatial separations and excess delays [36], [41]. On the basis of these observations a two-dimensional Gaussian distribution has been proposed for simulating spatially correlated amplitudes in dB [41]. It has been assumed that a multipath component that exist at a particular excess delay, exists at the same excess delay over the entire 1-m track due to the high spatial correlation on the arrival times [33], [41].

Correlation on log-amplitude of impulse response profiles of spatially- adjacent points were investigated using an extensive multipath propagation data base of 12 000 impulse response profiles collected at two office environments [85]-[87]. The average correlation coefficient was between 0.7 and 0.9 for the portable antenna displacement of 2 cm, but it dropped fairly rapidly for larger displacements. Spatial correlations were higher for initial paths than for later paths [86], [87].

The phase components are expected to be correlated only at very small spatial separations, and become uncorrelated much faster than amplitudes, although there is no empirical verifications.

V. Other Channel Related Issues

A. Temporal Variations of the Channel

Due to the motion of people and equipment in most indoor environments, the channel is nonstationary in time; i.e., the channel's statistics change, even when the transmitter and receiver are fixed. This is reflected in the time-varying filter model of (1). Analysis of this time-varying filter model, however, is very difficult. Most digital propagation measurements have therefore assumed some form of stationarity while collecting the impulse response profiles. The data were later supplemented by CW temporal fading measurements. Examples of CW temporal envelope fading are shown in Fig. 9. In Fig. 9(a) the immediate environment of the portable was clear of motion for the first 20 s, while motion occured after 20 s. Fig. 9(b) corresponds to a measurement during which there was constant motion in the vicinity of the portable throughout the 30-s measurement period. Examination of these figures reveal great variations in the signal level, even though both antennas are stationary. Deep fades of up to 20 dB below the mean value can be observed in these figures.

A review of the literature shows that in a number of measurements temporal stationarity or quasi-stationarity of the channel have either been observed or assumed in advance. Other experiments have shown that the channel is "quasistatic" or "widesense-stationary," only if data is collected over short intervals of time [78], [80]. The assumption of stationary or quasistatic channel in a time span of a few seconds may be reasonable for residential buildings or office environments in which one does not expect a large degree of movement. The situation may be different in crowded shopping malls, supermarkets, etc., where great number of people are always in motion. To avoid distortions caused by the motion of people and equipment, a number of indoor measurements have been carried out at night or during the weekends.

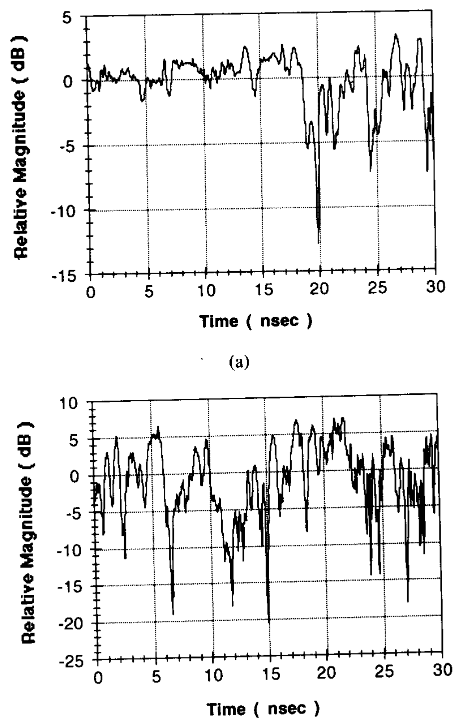

Fig. 9 Temporal CW envelop fading for a medium-size office building. Carrier frequency is 915 MHz and both antennas were stationary during the measurements. (a) Antenna separation 10 m; (b) antenna separation 20 m. (Measurements and processing by David Tholl of TRLabs.)

Indoor channel's temporal variations has been studied extensively by one investigator [77]–[80]. A major conclusion is that for office buildings where the environment is divided into separate rooms fading normally occurs in "bursts" lasting tens of seconds with a dynamic range of about 30 dB. For open office environments, however, fading is rather continuous with a dynamic range of 17 dB [78], [79].

Extensive CW measurements around 1 GHz in five factory environments [31], [34] and office buildings [77]–[79] have shown that even in absence of a direct line-of-sight path between the transmitter and receiver, the temporal fading data show a good fit to the Rician distribution. Another work reporting measurements at 60 GHz, however, indicates that with no LOS path the CW envelop distribution is nearly Rayleigh.

A measure of the channel's temporal variation is the width of its spectrum when a single sinusoid (constant envelop) is transmitted. This has been estimated to be about 4 Hz [78], [79] for an office building. A maximum value of 6.1 Hz has also been reported [66], [72].

The "local" and "global" short time variations of the channel have been reported in [70]. This report also includes temporal variations of the rms delay spread and the received power.

B. Large-Scale Path Losses

1) General Comments: While the impulse response approach is useful in characterization of the channel at a microscopic level, path loss models describe the channel at a macroscopic level. Path loss information in indoor environments are essential in determination of the size of the coverage area for radio communication systems, and in selecting optimum locations for base antennas. Obtaining three-dimensional propagation contour plots using a building's blueprint and the knowledge of its construction material is a challenging job which requires detailed and reliable path loss models.

Path loss measurement and modeling has been reported by many investigators [31], [34], [35], [40], [42], [45], [46], [50], [59]–[61], [63], [67], [74], [78], [81], [89], [96]–[98], [104]–[110], [113], [117], [131]–[136], [142], [143], [160], [161], [167], [168], [170], [172], [174]–[176], [183], [185], [196], [197], [201]. The indoor channel exhibits much larger path losses as compared to the mobile channel. Furthermore, large variations in the path loss are possible over very short distances. The propagation environment is very complicated and a universally accepted path loss model is not yet available. A review of indoor propagation measurements, however, indicate that there are four distinct path loss models. These models are briefly reviewed in this section. Although in principle one can obtain path losses using the wide-band pulse transmission techniques, a number of available models were derived from narrow-band CW measurements.

2) Model 1: According to this model the received signal power follows an inverse exponent law with the distance between antennas; i.e., $P(d) = P_0 d^{-n}$, where $P(d)$ is the power received at a distance d from the transmitter, and P_0 is the power at d =1 m. P_0 depends on the transmitted power, frequency, antenna heights and gains, etc. Path loss is therefore proportional to d^{-n}, where n depends on the environment. On a logarithmic scale this corresponds to a straight line path loss with a slope of $10n \log d$. A number of investigators have used this model [31], [34], [35], [40], [42], [45], [63], [67], [74], [78], [81], [96]–[98], [106]–[110], [117], [134], [143], [196], [197]. The reason is simplicity of the model and its previous successful application to the mobile channel.

The reported values of n are 1.5–1.8 for LOS and 2.4–2.8 for the obstructed factory channels [31], [34]; 1.81–5.22 for four different buildings [40], [42]; 1.4–3.3 for several manufacturing floors [63], [67]; 3.9 within a commercial building when transmitter and receiver are on adjacent floors [96]; less than 2 for hallway and equal to 3 for room measurements in an office building [97], [98]; equal to unity at 1700 MHz, and ranging from 1 to 3 at 433 and 861 MHz for LOS measurements in the same corridor of an office building [81]; between 1.2 and 6.5 for 900 MHz measurements in different indoor environments; 1.6 for a LOS, 2.1 and 4.5 for two non-LOS locations in an office building [74]; between 5 and 6 for three frequencies at university campus buildings [134]; 2 for LOS, 3.3, 3.8, and 4.5 for non-LOS at three other frequencies in rooms of an office environment [196]; 2 for open-plan buildings such as grocery store and retail store [45]; 1.1 and 1.2 for 37.2 GHz and 21.6 GHz, respectively, at a university building [143]; and between 1 to 2 for LOS measurements in an office building [117]. A comprehensive table of n values for various topographies and different choices of construction materials is provided in [108]. It should be noted that path loss in free-space follows a d^{-2} law. Smaller reported path losses (relative to free space) can be interpreted by taking the waveguide nature of walls in hallways and rooms into account; i.e., reflections from parallel walls can contribute constructively to the received power if their phase differences are small. The higher values of n are probably due to large attenuations encountered when transmitted signals have to penetrate walls, ceilings, floors, etc.

The model described above provides the mean value of random path loss. A standard deviation σ is also associated with each measurement. Reported σs are 7.1 dB for five factory environments [31], 16.3 dB for all data combined in four buildings, 2.5 dB for a LOS, 3.1 dB and 4.75 dB for two non-LOS locations in an office building [74], and ranging from 4.3 to 13.3 dB when data was classified according to building type and number of floors between transmitter and receiver [40], [42], etc. It has also been shown in one set of measurements at 900 MHz in several buildings [106]–[110] that the best fit to a straight-line (on a log scale) path loss model (i.e., model 1) occurs where "rooms are of similar size, linearly arranged, and with one wall of uniform attenuation between each room" [106]. In these measurements it was observed that departure (+ or -) from the straight line path loss is between 1 and 16 dB, depending on topography and construction material, but it is most often less than 10 dB (+ or -).

3) Model 2: This model has been suggested by one investigator [176] and has been quoted in the literature on many occasions. In this model the received power follows a d^{-n} law as in model 1. The exponent n, however, changes with distance. A distance-dependent exponent that increases from 2 to 12 with increasing d has been reported for indoor measurements carried out in a multistory office building [176]. In the measurements the fixed transmitter was located in the middle of a corridor and the portable receiver was placed inside rooms or along other corridors, on the same floor and on other floors. The value of n estimated from the data was 2 (for $1 < d < 10$ m), 3 (for $10 < d < 20$ m), 6 (for $20 < d < 40$ m), and 12 (for $d > 40$ m). The large values of n are probably due to an increase in the number of walls and partitions between the transmitter and receiver when d increases.

4) Model 3: This model associates logarithmic attenuations with various types of structure between the transmitter and receiver antennas. Adding these individual attenuations, the total path loss in dB can be calculated.

One set of measurements have shown an attenuation of 7 dB for 8-in. concrete block walls, 3 dB for wood and brick sliding, 2 dB for aluminum sliding, 12 dB for metal walls, and 1 dB per meter for office furniture [160] (also reported in [174]). In a comprehensive path loss measurement campaign floor attenuations were estimated at 12.9 dB, 18.7 dB, 24.4 dB, and 27 dB for 1, 2, 3, and 4 floors, respectively, in one building; and 16.2 dB, 27.5 dB, and 31.6 dB for 1, 2, and 3 floors, respectively, in another building [40]–[42], [45]. Attenuations were measured at 13 dB for a concrete block wall and 40–50 dB for one floor and one wall [42]. In the same set of measurements, attenuation factors were estimated at 1.4 dB for each cloth-covered office partition and 2.4 dB for each concrete wall [45]. Detailed path loss values versus obstruction type for a set of measurements where transmitter and receiver were placed on two sides of the obstruction have been reported in [42]. Another investigator has found path loss per a wall or a floor in a concrete building to be about 8.5–10 dB, regardless of whether transmitter was located inside the building or outside it [170]. Path loss through a floor with corrugated steel structure has been reported to be approximately 30 dB [136]. "Wall losses" were measured at 3.8 dB per wall for double plasterboard walls [142]. Finally, 60-GHz measurements have shown the loss relative to the LOS path to be 7 dB for penetrating a wooden door, and 27 dB for passing through a concrete wall [175].

In a number of measurements the transmitter was placed outside and receiver inside a building to estimate penetration losses. The overall median attenuation for 900-MHz measurements in a steel building has been estimated to be 29 dB [89]. The value of penetration loss at 900 MHz, 1800 MHz, and 2.3 GHz were reported to be 14.2 dB, 13.4 dB, and 12.8 dB, respectively, with the loss decreasing at 1.4 dB per floor as height was increased [134]. The range of penetration loss values for 937 MHz measurements by another investigator has been found to be between 10 to 25 dB [161]. The median path loss relative to free space on floors with a LOS path to the base station was found to be on the average 20.8 dB for one building, and 12.9 dB for another, with standard deviations of 3.2 dB and 2.3 dB, respectively [113]. Penetration loss measurements for 14 buildings has shown the signal level to be on the average 14 dB less inside the buildings (first floor), as compared to the adjacent street [167]. The mean penetration loss has been found to be approximately 11 dB at the first floor, with a loss reduction of about 0.67 dB per floor as the floor height was increased [168].

5) Model 4: This model associates a dB per unit of distance (antenna separations) path loss within a building.

Measurements were carried out in a large and a medium size office building at 850 MHz, 1.7 GHz, and 4 GHz [59]. Three models were tested: 1) A piecewise linear model (i.e., model 2 above), 2) A simple two-path model with an attenuation term, and 3) a simple free space rule plus a linear (on a log scale) attenuation. The third approach provided results as good or better than the first two. It was concluded that linear attenuation per unit distance is more appropriate since it conforms better with the physical reality of cluttered environments [59]. Another set of measurements at 850 MHz, 1.9 GHz, 4 GHz, and 5.8 GHz in large office buildings showed that a free-space attenuation plus a linear loss (0.3–0.6 dB/m) is sufficient to describe the data for different frequencies and different areas, with no clear dependence on frequency [60], [61].

Propagation measurements at 850 MHz in a large manufacturing environment has shown that an overall attenuation of 0.39 dB/ft describes path loss data for the case where both transmitter and receiver were located on the same floor [136]. The same investigators have carried out measurements at 150, 450, and 850 MHz in a large office building. Their best fit path loss model has α dB/ft loss for distances of up to d_1 ft and β dB/ft after that where α, β, and d_1 were 0.45, 0.22, and 150, respectively for the 150 MHz data; 0.38, 0.24, and 130 for 450 MHz; and 0.42, 0.27, and 70 for the 850-MHz signals [135].

C. Mean Excess Delay and rms Delay Spread

A one-number representation of an impulse response profile is the rms delay spread $\tau_{\rm rms}$ defined as:

$$\tau_{\rm rms} = \{[\sum_k (t_k - \tau_m - t_A)^2 a_k^2] / \sum_k a_k^2\}^{1/2} \qquad (20)$$

where τ_A is the arrival time of the first path in a profile and τ_m is the mean excess delay defined as:

$$\tau_m = [\sum_k (t_k - t_A) a_k^2] / [\sum_k a_k^2]. \qquad (21)$$

The above expressions show that τ_m is the first moment of the power delay profile ($|h(t)|^2$) with respect to the first arriving path, and the delay spread $\tau_{\rm rms}$ is the square root of the second central moment of a power delay profile. $\tau_{\rm rms}$ is a good measure of multipath spread; it gives an indication of the potential for intersymbol interference. Strong echoes (relative to the LOS path) with long delays contribute significantly to $\tau_{\rm rms}$. It has been shown that the performance of communication systems operating in a multipath environment are very sensitive to the value of $\tau_{\rm rms}$ [279], [280].

The mean excess delay and the rms delay spread have been estimated in a number of indoor propagation measurements. Numerical values depend on the size and type of the building, existence or absence of a clear LOS path, etc. Reported values are between 20 and 50 ns for small and medium-size office buildings [98], [117], between 30 and 300 ns for various factory environments [35], under 100 ns at several university buildings [200], less than 160 ns over 90% of the area in a shielded building [199], less than 80 ns in an office building, under 120 ns in a large office building [60], [61], and up to 200 ns in other large office buildings [55]. Large delay spreads of up to 200 ns have also been reported for smaller office buildings, although such a large delay spread seems to be due to external geographical features [55]. The median rms delay spreads reported are 96 ns for LOS and 105 ns for the obstructed factory channels [30], [35], 15.3 ns to 52.6 ns for several other factory environments [63], between 70 to 90 ns at three dissimilar office buildings [46], 8.3 ns for a LOS , 8.3 ns and 14.1 ns for two non-LOS situations in an office building [74], and 25 ns for a medium-size office building [98]. Much smaller median $\tau_{\rm rms}$ of 3–13 ns have also been reported for 60–GHz measurements in several buildings [189].

Measurements in a laboratory building have shown that τ_m and $\tau_{\rm rms}$ depend on the size of the rooms [187]. Measurements at an office building and a university building, however, indicate that the median $\tau_{\rm rms}$ is about the same for both buildings, but standard deviation of $\tau_{\rm rms}$ depends on the building [81]. Measurements in a large and in a small office building have also shown about the same $\tau_{\rm rms}$ [55]. However, the large delay spreads in the small building were attributed to external geographical features. Another investigator has found that $\tau_{\rm rms}$ depends on factory inventory, building construction and location of walls [35]. Average delay spread of 130 ns was obtained in a shielded building [199].

Dependence of $\tau_{\rm rms}$ on the transmitter-receiver antenna separation has been reported by several investigators. A number of measurements have shown correlation between $\tau_{\rm rms}$ and $T-R$ separation [54], [81], [82], [117], [199], [200]. There are also reports indicating uncorrelatedness [35], [98].

High linear correlation between $\tau_{\rm rms}$ and large-scale path losses was observed in a shielded building [199]. This is not supported by another investigator who showed $\tau_{\rm rms}$ and path loss to be uncorrelated [35].

Analysis of a large data base of 12 000 impulse response profiles collected at two office buildings [85]–[88] has shown that: 1) $\tau_{\rm rms}$ is typically between 10 and 50 ns, with mean values between 20 and 30 ns, and standard deviations between 3 and 5 ns; 2) $\tau_{\rm rms}$ over large areas (1500 profiles collected at each one of four transmitter-receiver antenna separations) can be well described with normal distributions; 3) mean $\tau_{\rm rms}$ increases with increasing antenna separation; 4) $\tau_{\rm rms}$ for spatially adjacent profiles are highly correlated; and 5) average $\tau_{\rm rms}$ at each location has a high linear correlation with the average path loss for that location.

D. Frequency Dependence of Statistics

The reported indoor radio propagation measurements have been carried out at frequencies as low as 35 MHz [26], and as high as 60 GHz [107], [126], [175], [189]. However, most of the measurement and modeling reports are for the 800-1000 MHz band ([40], [45], [46], [51]–[61], [63]–[65], [72]–[74], [77]–[80], [89], [92]–[95], [101], [103]–[109], [113], [117], [131], [132], [134]–[136], [167], [172], [173], [178], [182], [183], [197], [200], [201]), the 1.2-1.4 GHz band ([28]–[39], [41], [44], [131], [132], [170], [191]), the 1.6-1.8 GHz band ([58], [59], [80], [104], [117], [134], [142], [144], [147], [148], [164], [188], [189]), and for the 2.3-2.6 GHz band ([134], [164], [196], [199], [200]). Measurements have also been reported at 150 MHz [26], 400 MHz [131], [132], 1.9 GHz [46], [60], [61], 4 GHz [38], [59]–[61], 4.75 GHz [196], 5.8 GHz [60], [61], [200], 11.5 GHz [196], 21.6 GHz [143], and 37.2 GHz [143]. Some differences in the channel's behavior at different frequencies are reviewed below.

Narrow-band temporal CW envelope fading measurements at 910 MHz and 1.7 GHz in an office building and a university building have shown less severe fading in the 910-MHz data, with a dynamic range difference of 10 dB [80]. The overall conclusion, however, is that channel's statistics (both narrow-band and wide-band) are nearly the same at the two frequencies, with greater variations with the type of environment than with the frequency. This conclusion has been supported by another investigator on the basis of measurements at 850 MHz and 1.7 GHz in a multistory office building [58], at 850 MHz, 1.7 GHz, and 4 GHz in two dissimilar office buildings [59], and at 850 MHz, 1.9 GHz, 4 GHz, and 5.8 GHz in a large commercial building [61]. In these measurements, the rms delay spread and the attenuation appeared to be about the same at all frequencies. These results are not supported by the results of another measurement in a multistory office building, which has shown that path loss experienced between floors, or through a substantial wall is 11 dB greater at 1.7 GHz, as compared to the 900 MHz [104].

CW penetration loss measurements (with the transmitter outside and the receiver inside a building) at 35 MHz and 150 MHz by one investigator [26], at 441 MHz, 900 MHz, and 1.4 GHz by a second [131], [132], and at 900 MHz, 1800 MHz, and 2.3 GHz by a third investigator [134] indicate that the penetration loss is slightly smaller at higher frequencies. Path loss measurements inside office buildings at 900 MHz and 1650 MHz, however, indicate that loss through floors is greater at the higher frequency [182]. Measurement of penetration loss into houses at 860 MHz, 1550 MHz, and 2569 MHz (with a geosynchronous satellite transmitter) has also shown a path loss increase with increasing the frequency [164]. Path loss measurements in three dissimilar office buildings at two frequencies (915 MHz and 1.9 GHz) indicate that "different frequencies behave differently in different buildings" [46]. In one building floor attenuation was 13 dB higher at 1.9 GHz, while in the other two attenuations were similar at the two frequencies [46].

Measurements at 2.4 GHz, 4.75 GHz, and 11.5 GHz in rooms in an office environment showed that mean $\tau_{\rm rms}$ was about the same for 2.4 GHz and 4.75 GHz, and, on the average, 30% lower for 11.5 GHz [196]. Path loss exponent n was 2 at all frequencies for LOS cases, and increased with frequency for non LOS situations (n=3.3, 3.8, and 4.5, at 2.4 GHz, 4.75 GHz, and 11.5 GHz, respectively).

There has been recent interest in 60 GHz for radio communication inside buildings. This frequency coincides with a peak in atmospheric oxygen absorption, which results in an additional attenuation of 15 dB/km (compared to free space attenuation) [189]. It is therefore, unusable for long-range radio communication applications. Delay spread measurements at 1.7 GHz and 60 GHz have shown that the rms delay spread is considerably smaller at 60 GHz (median values for the three environments were 8, 22, and 24 ns at 1.7 GHz, and 3, 6, and 13 ns at 60 GHz) [189]. The reduction in rms delay spread at 60 GHz was attributed to a reduction in the power received from distant reflectors. Another investigator comparing path loss at 900 MHz and 60 GHz concludes that 60-GHz signals penetrate walls less easily, and, therefore, this frequency is usable for transmission in confined rooms, while the 900-MHz signal attenuation through walls is less severe and thus 900 MHz can provide coverage to several rooms [107].

It should be noted that a comprehensive comparison between the indoor channel's statistics at different frequencies, addressing issues such as amplitude and arrival-time distributions is unavailable. Such comparisons for the mobile channel at 488 MHz, 1280 MHz, and 2920 MHz, however, have shown insignificant differences [251], [252].

E. Comparison between the Indoor and the Mobile Channels

The indoor and outdoor channels are similar in their basic features: they both experience multipath dispersions caused by a large number of reflectors and scatterers. They can both be described using the same mathematical model. However, there are also major differences, briefly described in this section.

The *conventional* mobile channel (with an elevated base antenna and low-level mobile antennas) is stationary in time and nonstationary in space. Temporal stationarity is due to the fact that signal dispersion is mainly caused by large fixed objects (buildings). In comparison, the effect of people and vehicles in motion are negligible. The indoor channel, on the other hand, is stationary neither in space nor in time. Temporal variations in the indoor channel's statistics are due to the motion of people and equipment around the low-level portable antennas.

The indoor channel is characterized by higher path losses and sharper changes in the mean signal level, as compared to the mobile channel. Furthermore, applicability of a simple negative-exponent distance-dependent path loss model, well established for the mobile channel, is not universally accepted for the indoor channel.

Rapid motions and high velocities typical of the mobile users are absent in the indoor environment. The indoor channel's Doppler shift is therefore negligible.

Maximum excess delay for the mobile channel is typically several microseconds if only the local environment of the mobile is considered [251], and more than 100 μs if reflection from distant objects such as hills, mountains, and city skylines are taken into account [259]. The outdoor rms delay spreads are of the order of several μs without distant reflectors [254], [255], and 10 to 20 μs with distant reflectors [259]. The indoor channel, on the other hand, is characterized by excess delays of less than one μs and rms delay spreads in the range of several tens to several hundreds of nanoseconds (most often less than 100 ns). As a result, for the same level of intersymbol interference, transmission rates can be much higher in indoor environments.

Finally, the relatively large outdoor mobile transceivers are powered by the vehicle's battery with an antenna located away from the mobile user. This is in contrast with lightweight portables normally operated close to the user's body. As a result, much higher transmitted powers are feasible in the mobile environment.

VI. Performance Analysis of Indoor Systems

A. A Summary of the Channel

Propagation of radio waves inside a building is a highly complicated process. The impulse response approach described in this paper can be used to fully characterize the channel. A study of the literature [26]-[208] and the author's recent measurement and modeling efforts in this area show that:

1) The number of multipath components in each impulse response profile, N, is a random variable. Mean value of N is different for different types of buildings. The path variable sequences $\{a_k\}$, $\{t_k\}$, $\{\theta_k\}$ for every point in space are random sequences. The mean and variance of the distribution of a_ks are also random variables due to large-scale inhomogeneities in the channel over large areas.
2) Adjacent multipath components of the same impulse response profile are not independent. A standard Poisson hypothesis is inadequate to describe the arrival-time sequences. Adjacent amplitudes are likely to have correlated fading for high resolution measurements since a number of scattering objects that produce them may be the same. (Such correlations, although small, have been reported by two investigators. Measurements with smaller resolutions at different types of environments are needed to prove or disprove this type of correlation.) Phase components for the same profile, however, are uncorrelated since at frequencies of interest their relative excess range is much larger than a wavelength. The amplitude sequence and the arrival-time sequence are correlated because later paths of a profile go through multiple reflections and hence experience higher attenuations.
3) The impulse response profiles for points that are close in space are correlated since the channel's structure does not change appreciably over very short distances. Spatial correlations govern the amplitudes, the arrival-times and the phases, as well as the mean and variance of the amplitudes. There are small scale local changes in the channel's statistics and large scale global variations due to shadowing effects and spatial nonstationarities.
4) Path loss in an indoor environment is very severe most of the time. It is also very dynamic, changing appreciably over short distances. Simple path loss rules are successful in describing the mobile channel, but not the indoor channel.

5) The channel's parameters have great dependence on the shape, size and construction of the building. Variations with frequency have also been reported.
6) In its most general form the channel is nonstationary in time. Temporal variations are due to the motion of people and equipment around both antennas.

Any realistic channel model should take the above factors into account. Furthermore, it should derive its parameters from actual field measurements rather than basing them on simplified theory.

B. Performance Analysis Methods

Performance analysis of communication systems operating in an indoor environment is complicated both by the harsh propagation environment and by the sophisticated processing needed at the transmitter and receiver ends. Systems designers can employ one or more of the following distinct approaches:

1) Analyze the performance using a mathematical model based on realistic assumptions.
2) Analyze the performance using simplified theory.
3) Analyze the performance by simulating the entire communication link using realistic simulated statistical models.

Because of the complexity of the channel, task 1) is virtually impossible to perform. To carry out any feasible mathematical modeling of the system, the realistic channel model described in the previous section has to be replaced by a simplified one [approach 2) above]. Mathematical analysis of systems reported in the literature for the mobile and indoor channels are often based on standard (but oversimplified) assumptions of uniformly distributed Poisson arrival items, independent and identically distributed (IID) amplitude fadings, undifferentiated local and global statistics, negligible spatial correlations, no interpath interference, etc. Because of these oversimplifications, the reliability of any system simulation result is questionable.

Approach 3) is feasible and reliable on the condition of availability of a realistic simulation model for the channel. Such a simulation model should take into account all the realistic conditions mentioned in the previous section. It should be based on extensive field measurements at different types of areas and at various frequencies.

To emphasize the superiority of approach 3), we cite a pioneering work in performance analysis of various spread spectrum communication systems operating in the mobile channel [281]. Two methods were used and compared:

1) a simplified mathematical analysis based on standard assumptions.
2) a realistic simulation using an elaborate channel simulator (reported in [253]).

Several receiver configurations (including RAKE) were considered. It was observed that analytical results, although based on standard and reasonable assumptions, were between 3 and 7 dB optimistic compared to the realistic simulation results.

It should be noted that while elaborate simulations are needed for accurate results, simpler theoretical models should not be entirely dismissed since they are useful for providing insight.

VII. Recommendation for Future Work

Although a number of investigators have taken the impulse response approach, with two exceptions (the works reported fully in [41] and [87]), the published measurement and modeling efforts fall short of fully characterizing the impulse response (and hence, the channel). More specifically, there are a number of uninvestigated issues and a number of inconclusive points that demand further measurement and modeling. Some of these issues are reviewed here:

1) As of yet, there is no empirically driven phase model for the indoor channel. The models reviewed in this paper are both heuristic.
2) With one exception (measurements carried out at two office buildings [85]–[87]) modeling of the arrival-time sequence has not received sufficient attention. Other available models are based on limited data.
3) Distribution of signal amplitudes is not conclusive. Different experimenters have arrived at different distributions. Although each result may be justifiable when interpreted with the knowledge of conditions governing the measurements, a consistent model presenting amplitude distribution under a diversified set of conditions is unavailable. Furthermore, distinction between "local" and "global" statistics has often been ignored. Exception to this paragraph is the work reported in [85]–[87] in which analysis of 12 000 impulse response profiles in two office buildings has shown very good lognormal fit to both local and global data.
4) Path loss models reported in the literature are very different. Even with a given model, the range of parameters is large, and the dependence on the environment is not well established. As an example, path loss experienced by signals passing through concrete walls has been reported at 7 dB [160], [174], 8.5 to 10 dB [170], 13 dB [42], and 27 dB [175] by different investigators.
5) Spatial and temporal correlations between the path variables have not been fully addressed. Although consecutive impulse response measurements show strong correlations at close separations, the correlations have not been quantified (exceptions are the works reported for the factory channel [41] and two office buildings [86], [87]). The same is true for interdependences within path variables.
6) Range of the *median* values of rms delay spread reported in the literature are between 3 ns [189] and 120 ns [55]. Although the channel's parameters may have great dependences on frequency and topography, such factors are probably insufficient to account for differences of this magnitude. Extensive treatment of

various properties of the rms delay spread for data collected at two office buildings has been reported in [87], [88].

7) The degree of the channel's dependence on frequency and type of environment is not clear. Different investigators have reported different results.

It is the author's belief that inconsistencies between the works surveyed in this paper are, to a great extent, due to insufficiency of data. As an example, a well-recognized impulse response model reported in the literature is based on only 200 profiles.

On the basis of the results of this survey the following comments and recommendations are made:

1) To carry out *extensive* measurements in a diversified set of conditions (building type, size and topography, frequency, antenna types and patterns, etc.) to arrive at a consistent impulse response model. Such measurements should be based on a reasonable mathematical model and a comprehensive measurement plan. The author's experience in one such measurement campaign (12 000 impulse response profiles collected at two office environments using the mathematical model of (2) and the variation model of Fig. 6 [85]–[88]) shows that most of the unresolved issues can be consistently and confidently resolved.

2) It is well established that the indoor channel is, in general, time varying. Therefore, there are two types of variations: 1) variation due to the motion of the portable unit, 2) variation due to the dynamics of the changing environments. Although 1) may have greater influence than 2), the two are different, and therefore most impulse response modeling reports (including [85]–[88]) were concerned with 1). Full characterization of $h(t, \tau)$ rather than $h(t)$ is a challenging job which requires major modeling efforts. One possibility is to modify the time-invariant impulse response models using the information available on temporal variations from the CW fading measurements. However, the mechanism for such modifications and improvements is not clear.

3) It is extremely useful to devise elaborate channel simulation models based on empirical data. One such model for the factory and open plan office buildings is SIRCIM reported in [41]. Another one is being developed by the author using the extensive multipath propagation data base reported in [85]–[88].

4) While 1)–3) above are useful to characterize the channel at a "microscopic" level, reliable path loss models are needed for characterization on the "macroscopic" level. Such models are needed to determine the coverage area and select optimum base locations. Extensive CW propagation measurements at a diversified set of conditions aimed at devising reliable path loss prediction models (similar to what is reported fully in [45]) are recommended.

5) Recently reports of the application of analytical ray tracing techniques to indoor radio propagation modeling has appeared in the literature [47], [50], [76], [149], [192], [193], [202]. This technique has been proposed to predict path loss, the time-invariant impulse response, and the RMS delay spread. With the computing powers currently available, this approach provides a challenging but feasible approach to propagation modeling. Reliable site-specific propagation prediction models for each building based on its detailed geometry and construction can be used as very effective tools in engineering indoor communication systems. Application of this approach, "while in its infancy," appears to be very promising [47]. It is recommended to follow this line of work; i.e., to devise elaborate ray tracing techniques and validate the theoretical models with actual field measurements.

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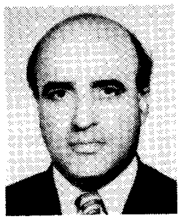

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In the summer of 1977, he served as a Consulting Engineer at Stanford Research Institute, Menlo Park, California. From 1977 to 1979 he was a member of the Technical Staff at Bell Telephone Laboratories, Holmdel, N.J., where he was involved in system design for high-capacity mobile telephone systems. Since 1979 he has been a Faculty Member at the Electrical Engineering Department of Sharif University of Technology, Teheran, Iran, where he is currently Associate Professor. He was on sabbatical leave at NovAtel Communications Ltd., Calgary, Alberta, Canada, in the 1990–91 academic year, where he worked on various projects related to mobile and personal communications. He spent the summer of 1992 at the Electrical Engineering Department of the University of Ottawa, where he was involved in propagation modeling at infrared optics for indoor wireless communications. He is on leave at TRLabs, Calgary, Alberta, Canada, in the summer of 1993, supervising several projects related to indoor radio propagation modeling and PCS system design. He has worked on different aspects of wireless communications, including general system architecture, channel assignment, propagation modeling, performance analysis, and capacity evaluations for digital cellular, microcellular, and indoor wireless communication systems. His channel simulation package SURP has been used internationally in the design of digital cellular radio communication systems. He has also served as a consultant to the Iranian Telecommunications Research Center and the Telecommunication Company of Iran.