Antenna-Sector Time-Division Multiple Access for Broadband Indoor Wireless Systems

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Abstract—This paper investigates a hybrid space–time-division multiple access (S-TDMA) for broadband indoor wireless systems using sectored antennas. It is shown that portables which are located in different sectors of an indoor microcell may be able to reuse the same frequency and the same time slot. However, this requires careful scheduling of packet transmissions in order to avoid transmitting packets that would jam each other during the same time slot. It is proposed that the scheduling be performed in the base station, i.e., a central control architecture. The optimum scheduling algorithm, the one that maximizes the number of packets transmitted per frame, may be in the NP-complete class of problems, so it cannot be solved in real time. Therefore, a suboptimum algorithm, called the first fit algorithm (FFA), is proposed for frame scheduling. It was found that the capacity gain achieved by the FFA is dependent on the capture threshold, which is defined as the minimum signal-to-interference ratio required in order to achieve a given packet error rate goal. The capture threshold depends on the modulation and coding schemes. This paper investigates the performance of the FFA operating with multicarrier trellis-coded modulation. An alternative multicarrier modulation is analyzed, and the FFA performance is investigated when operating with the alternative and with the conventional multicarrier through computer simulations based on measured data which were obtained with a sectorization level of ten (using ten antenna sectors in the base station). The simulations have shown that the FFA can provide a large capacity gain when operating with multicarrier trellis-coded modulation and using differential detection. For example, while previous schemes can transmit only one packet at a time, the proposed scheme can transmit, on average, more than four packets per time slot in an open indoor location, or close to three packets per time slot in a closed indoor location with internal walls of concrete blocks.

Index Terms—Broadband communication, directive antennas, frequency division multiplexing, indoor radio communication, space division switching, time division multiaccess, trellis coded modulation, wireless LAN.

I. INTRODUCTION

SECTORED antennas have been used for some time in mobile cellular systems as a means of reducing interference, and consequently increasing system capacity. More recently, sectored antennas have also been applied to wireless local-area networks (WLAN) [1], [2]. In [1], the use of directional antennas is considered to be an excellent technique for dealing with multipath propagation problems. In [2], sectored antennas are proposed in order to combat the severe cochannel interference arising from a small frequency reuse factor. Although [1] and [2] propose different rules for the use of sectored antennas, the final goal in both cases is to achieve an increase in system capacity.

The multiple-access schemes described in [1] and [2] allow transmission/reception in only one of the antenna sectors at a time, usually in the one that provides the best communication channel between the base station and a given portable. Fig. 1(a) shows sectored antennas similar to those in [1] and [2]. The RF switch matrix used to select the desired antenna from the set can be fabricated compactly and inexpensively using either a gallium arsenide monolithic microwave circuit (GaAs MMIC) or microwave diode switching elements [4].

In [3], a new multiple-access scheme is proposed to achieve in-cell frequency reuse by allowing simultaneous transmissions over different antenna sectors. This new multiple-access scheme requires a multiport sectored antenna system such as the one illustrated in Fig. 1(b), which represents the same set of antennas as in Fig. 1(a), but with a switch matrix capable of simultaneously selecting two antennas and connecting each to a desired port that is connected to a transceiver. The sectored antenna system of Fig. 1(b) has the potential to double the capacity of the system by allowing simultaneous transmissions/receptions to occur in two antenna sectors sharing the same frequency spectrum. However, as we explain in Section II, this kind of spectrum sharing requires compatible portables.

The schemes proposed in [1] and [2] employ sectored antennas in both the base station and portable modules. The scheme proposed in [3] assumes that this approach causes the system to become excessively complex, and due to the size of the sectored antenna system, not suitable for a portable terminal. Therefore, the conservative approach of concentrating
the complexity of the system in the base station is adopted in order to be able to use simpler antennas in the portables: sectored antennas are used in the base station, and simple omnidirectional antennas are used in the portables.

The effect of using omnidirectional antennas in the portables is larger delay spread profiles in the channels. As a result, the system requires a modulation scheme that is capable of overcoming channel dispersion. At the same time, this scheme has to be bandwidth efficient, and it has to avoid the high-power consumption signal processors needed for equalization. An alternative multicarrier modulation that fits these requirements is proposed in [6].

In this paper, we investigate the performance of the multiple access proposed in [3] when operating with the alternative multicarrier scheme proposed in [6], and we compare the performance results with the results obtained when the system operates with the conventional multicarrier modulation.

It is important to state at this point that the results reported in this paper assume that the indoor channels are static since they are based on measurement data obtained with an experimental setup which does not account for channel dynamics. Intuitively, however, this technique would work well with quasi-static indoor networks where all of the portables are static and channel dynamics are caused by the movement of third elements. The third elements may be, for instance, the opening and closing of doors or people walking across the microcell. In this case, it is assumed that the base station is capable of updating the information about the compatibility relations among the portables. The concept of compatibility, which, in the multiple-access scheme, is the key to achieving capacity gain, is explained next.

II. COMPATIBILITY CONDITION

The concept of compatibility is explained with the help of Fig. 2, which shows a base station using a two-port sectored antenna system to communicate simultaneously with two portables $P_1$ and $P_2$. The signal strength values quoted in this figure were chosen for illustrative purposes. The reflectors represent reflecting structures, for example, walls or metallic doors. In order to verify the compatibility between portables $P_1$ and $P_2$, the minimum signal-to-interference ratio (SIR) acceptable for communication between the base station and a portable has to be specified. This parameter is the capture threshold, which we call $\Pi$. Its value is dependent on the coding and on the modulation scheme employed. For example, spread-spectrum modulation schemes can operate with a level of interference above the level of the signal, which results in a negative value of $\Pi$ (in decibels). However, the proposed multiple-access scheme is intended to operate with narrow-band (nonspread-spectrum) modulation schemes which can achieve high spectrum efficiency, around 1 bit/s/Hz, as required in the FCC proposal for U-NII wireless networks [5]. Therefore, we will be considering values of $\Pi$ in the range of $5–20$ dB.

It is assumed that an uplink transmission and a downlink transmission cannot occur simultaneously: either both transmissions are in the uplink direction (from portable to base station) or both are in the downlink direction (from base station to portable). This assumption is made because the level of interference that a downlink transmission would cause in an uplink reception would be unacceptable: in the base station, the power level of a transmitted signal is much higher than the power level of a received signal.

In the downlink case, the base station uses antenna sector 5 to transmit to portable $P_1$, and antenna sector 8 to transmit to portable $P_2$. Portable $P_1$ receives a signal from antenna sector 5 at an average power level of 0 dBm, and interference from antenna sector 8 through a reflected path at an average power level of $-20$ dBm. Simultaneously, portable $P_2$ receives a signal from antenna sector 8 at an average power level of $-10$ dBm, and interference from antenna sector 5 through a reflected path at an average power level of $-100$ dBm. Therefore, the SIR for the signal received by portable $P_1$ is 20 dB, and the SIR for the signal received by portable $P_2$ is 90 dB. Then we say that portables $P_1$ and $P_2$ are compatible in the downlink case if $\Pi \leq 20$ dB.

In the uplink case, the base station receives, through antenna sector 5, the signal transmitted by $P_1$ at an average power level of 0 dBm and interference of $-100$ dBm from the reflected signal transmitted by $P_2$, yielding an SIR of 100 dB for the $P_1$ transmitted signal. Simultaneously, the base station receives, through antenna sector 8, the signal transmitted by $P_2$ at an average power level of $-10$ dBm and interference of $-20$ dBm from the reflected signal transmitted by $P_1$, yielding an SIR of 10 dB for the $P_2$ transmitted signal. Therefore, we say that portables $P_1$ and $P_2$ are compatible in the uplink case if $\Pi \leq 10$ dB.

In this example, which does not consider the power control mechanism that we describe below, compatibility in the downlink case does not necessarily correspond to compatibility in the uplink case. In fact, we observed through measurement results that the average number of compatibilities among
portables in an indoor microcell is, in general, larger for the
downlink case than for the uplink case. This downlink/uplink
average compatibility asymmetry arises because, in the uplink
case, portables located closer to the base station cause stronger
interference with the signals of portables located on the
periphery of the indoor microcell. In a sense, this problem is
similar to the near–far effect that decreases the uplink capacity
of CDMA systems. Therefore, it is expected that power control
can be used to improve the average number of compatibilities
in the uplink case.

Consider now the use of power control in Fig. 2. The power
control mechanism assumes that the portables adjust their
transmitting power so that the base station receives, through
the best sector for communication with a given portable, the
same average power level from any portable. In this case,
portable $P_2$ increases its transmitted power so that the base
station can receive its signal with the same power level of the
signal received from portable $P_1$. Then $P_2$ has to add 10 dB
to its transmitted signal power. By doing so, it also adds 10
dB to the interference caused in the $P_1$ signal. This is shown
in Fig. 2(c). If the uplink SIR values are recalculated with the
power control values of Fig. 2(c), the SIR values obtained are
90 dB for the $P_1$ signal and 20 dB for the $P_2$ signal. Therefore,
the minimum of the two SIR values, 20 dB, is 10 dB better
than the minimum SIR value obtained without power control.

In order to express the compatibility conditions mathemati-
cally, we define the following parameters.

- $G_{i,s}$ (portable-sector power gain): represents the average
  power received through the channel between portable $P_i$
  and antenna sector $s$. It has the same value for both uplink
  and downlink channels (provided that the same power
  level is transmitted in both cases). This is true because
  we assume that uplink and downlink transmissions occur
  in the same frequency band (but not at the same time);
  therefore, uplink and downlink channels are reciprocal.
- $B(i)$: represents the best antenna sector for communicat-
ing with $P_i$.

The $G_{i,s}$ parameters are illustrated in Fig. 3.

Using these definitions, we can express the conditions for
the existence of compatibility in the downlink case between
two portables $P_i$ and $P_j$ as

$$\begin{align}
\{ & \quad \text{10 log} \frac{G_{i,B(i)}}{G_{j,B(j)}} \geq \Pi \\
& \text{10 log} \frac{G_{j,B(j)}}{G_{i,B(i)}} \geq \Pi 
\}.
\end{align}$$

In the uplink case, the conditions for the existence of
compatibility between two portables $P_i$ and $P_j$ are

$$\begin{align}
\{ & \quad \text{10 log} \frac{G_{i,B(i)}}{G_{j,B(i)}} \geq \Pi \\
& \text{10 log} \frac{G_{j,B(i)}}{G_{i,B(i)}} \geq \Pi 
\}.
\end{align}$$

III. COMPATIBILITY CONDITION FOR THE $N$-PORT CASE

In the previous section, we considered compatibility for
the two-port case, which implies using a system of sectored
antennas with two ports as shown in Fig. 1(b). Now, we
consider a system of sectored antennas with $N$ ports as shown
in Fig. 4.

An $N$-port antenna system would allow a maximum of $N$
simultaneous transmissions in $N$ different antenna sectors. As
such, it would have a switching matrix capable of selecting
$N$ of the antenna sectors and connecting each of them to one
of $N$ ports (each port is connected to a transceiver). In this
case, the conditions for compatibility among $N$ portables $P_1,
P_2, \cdots, P_N$ in the downlink case are given in (3) and, in the
uplink case, the compatibility conditions are given in (4), both
shown at the bottom of the next page.

If the $N$ conditions of (3) are satisfied, then there may be $N$
simultaneous packet transmissions during a time slot to these
$N$ portables (downlink case). This means that, during this time
slot, the downlink system capacity is multiplied by $N$, i.e., $N$
packets are transmitted through $N$ sectors instead of the single
packet that could be transmitted in this time slot if a single port sectored antenna system were used in the base station. Similarly, if the $N$ conditions of (4) are satisfied, then there may be $N$ simultaneous packet transmissions during a time slot from these $N$ portables (uplink case), which corresponds to the uplink system capacity being multiplied by $N$. In general, however, an increase in the number of ports does not correspond to a proportional increase in system capacity, as can be observed from measurement and simulation results presented later in this paper. The reason is that the number of subsets of $N$ portables that correspond to compatible portables decreases as $N$ increases. For example, the probability of two portables being compatible is larger than the probability of three portables being compatible. Therefore, as the number of ports $N$ increases, it is more likely that some of them will not be used during a given time slot due to a lack of compatible portables. The capacity gain eventually saturates at a given number of ports.

IV. FRAME SCHEDULING

This section discusses how to take advantage of the existing compatibilities among portables in an indoor microcell in order to increase system capacity. This can be done by properly scheduling simultaneous transmissions of packets that belong to compatible portables. Fig. 5 depicts the scenario where scheduling is performed. It shows that the base station has a buffer where packets collected from a high-speed wireline backbone network are temporarily stored while waiting to be transmitted to their destination portables during downlink time slots. It also shows that each portable has a buffer that temporarily stores packets generated in the end-user terminal equipment and are waiting to be transmitted to the base station during granted uplink time slots. The scheduling problem consists of finding among the awaiting packets those that can share a time slot in the next frame, namely, those that belong to compatible portables.

Let us suppose that Fig. 6(a) represents the values of the $G_{p,s}$ parameters, as defined in Section II, related to the portables and the base station of Fig. 5. The values quoted in this figure were chosen for illustrative purposes. With these values, the two-by-two compatibility relations among the portables can be obtained, as shown in Fig. 6(b) for a capture threshold of 10 dB, where “1” means compatibility and “0” means incompatibility. Let us also suppose that a two-port frame has been scheduled using these compatibility relations. Fig. 7 represents this frame with its downlink and uplink portions. Here, the word “port” is used to denote a port in the sectored antenna system which can be switched to any one of the antenna sectors. Therefore, in this figure, the base station operates with a two-port sectored antenna system as shown in Fig. 1(b). The frame is composed of a number of equal time slots, each capable of carrying one packet per port. Thus, with the two-port antenna system, a maximum of two packets can be transmitted per time slot. For example, each packet could transport one ATM (asynchronous transfer mode) cell [8] plus some wireless overhead. The packet owners and the sectors being used for the transmission of each packet are also shown in Fig. 7. The first time slot of the downlink subframe carries two packets, one for $P_1$ and the other for $P_3$.

\[
\begin{align*}
10\log \left[ \frac{G_{1,1}}{G_{1,1} + G_{1,2} + \cdots + G_{1,N}} \right] & \geq \Pi \\
10\log \left[ \frac{G_{1,2}}{G_{1,1} + G_{1,2} + \cdots + G_{1,N}} \right] & \geq \Pi \\
\vdots \\
10\log \left[ \frac{G_{N,1}}{G_{N,1} + G_{N,2} + \cdots + G_{N,N}} \right] & \geq \Pi \\
10\log \left[ \frac{G_{1,1}}{G_{1,1} + G_{1,2} + \cdots + G_{1,N}} \right] & \geq \Pi \\
10\log \left[ \frac{G_{1,2}}{G_{1,1} + G_{1,2} + \cdots + G_{1,N}} \right] & \geq \Pi \\
\vdots \\
10\log \left[ \frac{G_{N,1}}{G_{N,1} + G_{N,2} + \cdots + G_{N,N}} \right] & \geq \Pi \\
\end{align*}
\]
For each packet, the base station uses the antenna sectors that provide the best signal level, namely, sectors 2 and 6. The time slot and frequency spectrum sharing between a packet and a packet is possible because of the compatibility between their owners.

We propose that uplink and downlink modes use the same frequency spectrum (time-division duplex). This guarantees reciprocity between uplink and downlink channels, allowing the parameters to be used for both downlink and uplink compatibility verification with the conditions defined in (1) and (2), respectively. Uplink/downlink traffic does not need to be symmetrical. In fact, it is expected that the uplink traffic will be a small fraction of the downlink traffic. This is the case, for example, for the current traffic in the Internet where the downloading of files generates highly asymmetrical traffic. Accordingly, the downlink subframe in Fig. 7 was made intentionally larger than the uplink subframe for illustrative purposes.

It is assumed that scheduling is performed in the base station (central control architecture). For this purpose, the base receives updated information about the compatibility relations among the portables and about the number of packets waiting transmission in each buffer. For variable bit-rate (VBR) services, the number of packets that arrive in the portables’ buffers and in the base station’s buffer during one frame duration varies from frame to frame. This means that each new frame requires its own scheduling. Therefore, the time available for scheduling each frame is one frame duration: during the transmission of the $i$th frame, the base station is working out the scheduling of the $(i+1)$th frame, which has to be ready for transmission before the end of the $i$th frame transmission.

V. FIRST FIT ALGORITHM

The necessity of solving the scheduling problem in less than one frame duration was stated in the previous section. However, it turns out that for the $N$-port case ($N \geq 3$), the optimum scheduling problem, which maximize the number of packets transmitted per frame, is in the NP-complete class of problems (problems that cannot be solved in polynomial time, or nonpolynomial problems) [7]. The practical result is that, for a large number of packets, solving it in real time becomes impractical.

The NP-completeness of this problem is not proved here, but it is indicated that it can be proved by mapping the $N$-dimensional matching problem [7] to the $N$-channel ($N \geq 3$) frame-scheduling problem. By knowing that the $N$-dimensional ($N \geq 3$) matching problem is in the class of NP-complete problems, it is established that the $N$-channel frame-scheduling problem ($N \geq 3$) is also in this class. The two-port optimum frame-scheduling problem is a special case that requires further investigation to determine if it belongs to the NP-complete class or if it belongs to the $P$ class (problems that can be solved in polynomial time [7]). However, we prefer to concentrate on finding a suboptimum frame-scheduling solution that fits the generic $N$-port case, and that can be solved in real time. This suboptimum solution is the first fit algorithm (FFA) that we describe below.

The FFA can be better explained with the help of Fig. 8. This algorithm places the packets into the time slots one at a time in the order that they arrive in the buffer. It does so according to the following simple rules.

1) First, the algorithm occupies the time slots of the first port. If, after performing this step, there are packets left in the buffer, then it performs the following steps.

2) For allocating packets in ports 2 and up, the algorithm always places the next packet of the buffer into the lowest indexed time slot containing only packets that are compatible with the packet being allocated. The conditions of compatibility are given by (3) for the downlink case and by (4) for the uplink case, where $N$ is such that $N - 1$ is the number of packets already allocated in the time slot. When searching for the lowest indexed time slot, the port index has priority over the time slot index. Therefore, the algorithm looks first for a time slot in port 2, and if it cannot find a valid time slot, then it looks for one in port 3, and so on. This helps to distribute packet transmissions evenly in the time axis.

3) If a packet does not fit in any of the time slots of this frame, then the algorithm leaves it to be transported in
Fig. 8. First fit algorithm.

the next frame, and tries to allocate the next packet of
the buffer by going back to step 2).

4) The algorithm repeats steps 2) and 3) until all of the
packets of the buffer are considered or until the frame
is full.

In the scheduling of downlink packets, the packets are stored
in the base station’s buffer while awaiting transmission, so the
FFA operates in a local buffer. However, in the scheduling of
uplink packets, the buffer is distributed in the sense that it is
composed of all of the portables’ buffers. In this case, the base
station has to receive updated information about the number
of packets awaiting transmission in each portable buffer, so
that the distributed buffers can be operated in conjunction as
if they were a single buffer.

The FFA serves the packets according to their order of
arrival in the buffer. However, this algorithm could be mod-
ified in order to serve packets according to some policy that
privileges real-time services. See [9], for example.

It has been assumed that the base station keeps updated
information about the compatibility condition among the porta-
bles and about the number of packets awaiting transmission
in each buffer. In [3], a frame structure with overheads that
allow the base station to acquire and update this information
is proposed.

VI. CHANNEL MEASUREMENTS

An experiment, which is described with details in [3],
has been devised to measured the $G_{ps}$ parameters which
are illustrated in Fig. 3. It can measure with a sectorization
level of 10, i.e., it assumes that the base station operates
with ten antenna sectors, each covering 36° of the horizontal
plane. Three typical indoor locations were measured with this
experiment. The first location was the southwest section of the
fourth floor of the Galbraith Building located on the University
of Toronto Campus. This location is depicted in Fig. 9. It
represents a closed indoor environment where the walls are
made of a hard material, in this case concrete blocks. For this
location, 50 uniformly distributed positions were measured.

The second location was a 10 $\times$ 112 room (room
GB402 in the Galbraith Building) divided into cubicle offices with
soft material partitions. This room was chosen to represent a
semiopen indoor environment. For this location, 20 positions
were measured, each corresponding to the interior of one
cubicle office.

The third location was a 10 $\times$ 11 m2 classroom (room
GB404 in the Galbraith Building) furnished mainly with desks
and chairs. This room was chosen to represent an open
indoor environment. For this location, 36 uniformly distributed
positions were measured.

In [3], the measurement results were analyzed in terms of
the FFA maximum throughput. For example, Fig. 10 shows
the uplink maximum throughput in terms of packets per time
slot that can be achieved in location 1 with the FFA. It is clear
in Fig. 10 that a small capture threshold is desirable since the
capacity gain is inversely proportional to the capture threshold.
However, the capture threshold depends on modulation scheme
and on coding. In the following section, a coded multicarrier
modulation is described, and making use of the measured data
($G_{ps}$ parameters), the FFA is simulated in conjunction with
the coded multicarrier modulation.
VII. FFA MAXIMUM THROUGHPUT WITH MULTICARRIER MODULATION

In [3], the FFA maximum throughput was investigated as a function of the capture threshold. However, the question may arise as to with which value of capture threshold a system can operate. Of course, the answer depends on the type of modulation and coding being used by the system.

The aim of this section is to investigate the performance of the proposed multiple-access scheme assuming the use of multicarrier modulation. This type of modulation [12]–[15] was chosen because it is naturally resistant to channel dispersion which occurs in indoor channels due to multipath propagation. This phenomenon affects signals when they operate at a symbol rate such that the symbol period is comparable to the channel time dispersion. In this case, intersymbol interference (ISI) causes an intolerable degradation in the bit-error rate (BER) performance. Therefore, we need to obtain means to transmit reliably at the required high bit rates. One of the possible solutions is to use adaptive equalization [16]. However, according to Mitzlaff [4], there are practical difficulties created by the sheer size, expense, and power consumption of the hardware needed to equalize a 10 Mbit/s data signal. Bit rates larger than 10 Mbits/s are expected for multimedia services, which makes equalizing these channels a challenging proposition. Therefore, we adopted multicarrier modulation as the solution to combat ISI.

A. Orthogonal Frequency-Division Multiplex

The specific type of multicarrier modulation explored in this paper is orthogonal frequency-division multiplex (OFDM) which is popular because it can be generated and demodulated using fast Fourier transforms (FFT’s). The baseband OFDM signal can be expressed as

\[ S(t) = \sum_{\nu=-\infty}^{\infty} \sum_{n=0}^{N-1} A_{\nu,n} e^{j2\pi f_n t} u(t) \]  \hspace{1cm} (5)

where

\[ f_n = \frac{n}{T_O} \quad n = 0, \ldots, N-1, \quad u(t) = \begin{cases} 1, & 0 \leq t \leq T_S \\ 0, & \text{otherwise} \end{cases} \]

Here, \( N \) is the number of carriers, \( T_O \) is the observation period, and \( T_S \) is the symbol duration (\( T_S = T_O + \) cyclic prefix). A cyclic prefix can be added to eliminate the ISI [15]. Therefore, there is a guard period between two consecutive observation periods which is given by \( \Delta = T_S = T_O \). The term \( A_{\nu,n} \) represents the information symbol transmitted by the \( n \)th carrier during the \( \nu \)th symbol period.

Fig. 11(a) and (b) shows the block diagrams of an OFDM transmitter and receiver, respectively. In the transmitter, the information sequence is applied to a serial-to-parallel converter (S/P), where \( N \) symbols are assembled. Each assembled vector \( \{A_{\nu,n}\}_{n=0,1,2,\ldots,N-1} \) is padded with \((J-1)N\) zeros. Then the \( JN \)-point inverse discrete Fourier transform (IDFT) is performed on the resulting vector. It is easy to see that the output of the IDFT corresponds to \( JN \) samples of the baseband OFDM signal of (5) taken during the \( \nu \)th symbol period at a sampling rate \( f_{\text{samp}} = JN/T_O \). These samples are applied to a parallel-to-serial (P/S) converter, and the real and imaginary parts are applied to digital-to-analog converters (D/A). The output signals of the D/A converters are low-pass filtered to generate the complex signal \( S(t) \) of (5). Then, real and imaginary parts of \( S(t) \) are used to modulate the in-phase and quadrature components of the RF carrier, respectively.

In the receiver, the received signal \( R(t) \) is down-converted, and \( JN \) samples are taken during the \( \nu \)th observation period. These samples are applied to a discrete Fourier transform (DFT) to obtain the Fourier coefficients of the signals in the observation period \([T_O, T_O + T_S] \). The output \( K_{\nu,n} \) of the \( n \)th carrier at time \( \nu T_S \) is represented by

\[ K_{\nu,n} = \frac{1}{T_O} \int_{T_O}^{T_O+T_S} R(t)e^{-j2\pi f_n (t-\nu T_S)} dt. \]  \hspace{1cm} (6)

The demodulated vector \( \{K_{\nu,n}\} \) is applied to a parallel-to-serial converter, and the decision block of the receiver decides which symbol was transmitted in the \( n \)th carrier during the \( \nu \)th symbol period.

B. Alternative OFDM

We can see in Fig. 11 that a \( JN \)-point IDFT is used in the modulation of an \( N \)-carrier OFDM signal, and a \( JN \)-point DFT is used in the demodulation. With the \( JN \)-point IDFT, samples of the transmit signal are generated at a sampling rate of \( f_{\text{samp}} = JN/T_O \). Bingham [13] stated that if \( f_{\text{sam}} = 2N_{\text{tot}}/T_O \), then \( N_{\text{tot}} \) carriers are available for modulation, which makes \( J = 2 \). Now we start analyzing an alternative OFDM signal proposed by Macedo and Sousa [6] where \( J = 1 \). The main merit of this alternative OFDM signal is that it can be generated using an \( N \)-point IDFT, and demodulated using an \( N \)-point DFT, while the conventional OFDM requires at least a \( 2N \)-point IDFT in the generation and a \( 2N \)-point DFT in the demodulation. Therefore, with this alternative OFDM, there is the potential to cut the amount of processing by \((1/2)((1/1 + 1/\log_2 N))\).

Fig. 12 compares the four-carrier conventional OFDM signal with the four-carrier alternative OFDM signal. In Fig. 12(a), an eight-point IDFT is used to generate eight
samples per observation period $T_o$ of a four-carrier OFDM signal. As seen in this figure, padding with zeros is required in order to obtain the appropriate sampling rate which has to be at least $(2N/T_o)$. Fig. 12(b) illustrates how the alternative OFDM signal can be generated in the case of four carriers. During the $i$th symbol period, the vector $\{S_i,k\}$ is generated by taking the $N$-point IDFT of the information vector $\{A_i,n\}$ (without padding with zeros). The elements of the vector $\{S_i,k\}$ are then serially transmitted by Nyquist pulses with a signaling rate of $1/T$, where $T = T_o/N$. Here, a cyclic prefix can also be added to combat ISI. The resulting signal has a symbol period of $T_S = T_o/pT$, where $pT$ is the cyclic prefix and $p$ is an integer. Fig. 13 illustrates one symbol period of the alternative OFDM signal with and without the addition of a cyclic prefix of size $p = 2$.

The baseband transmitted signal can be expressed as

$$S(t) = \sum_{i=-\infty}^{\infty} \sum_{k=-p}^{N-1} S_{i,k} h(t - kT - iT_S)$$  \hspace{1cm} (7)

where

$$S_{i,k} = \sum_{n=0}^{N-1} A_{i,n} e^{j(2 \pi n k / N)}, \quad (S_{i,k} = S_{i,N-k}) | k = 1, 2, \ldots, p$$  \hspace{1cm} (8)

and $h(t)$ is the Nyquist pulse waveform. More precisely, $h(t)$ is assumed to be the raised-cosine pulse, which can be expressed as

$$h(t) = \left[ \frac{\sin(\pi t/T)}{\pi t/T} \right] \left[ \frac{\cos(\alpha \pi t/T)}{1 - (2\alpha \pi t/T)^2} \right]$$  \hspace{1cm} (9)

where $\alpha$ is the rolloff factor. The real and imaginary parts of $S(t)$ are then used to modulate the in-phase and quadrature components of the RF carrier, respectively.

### C. Alternative OFDM Under Frequency-Selective Fading

This subsection analyzes the BER performance of the alternative OFDM in a frequency-selective channel. This type of channel can be modeled as a tapped delay line where each tap is an independent complex Gaussian random process. Assuming a static channel, the impulse response of the channel can be expressed as

$$g(t) = \sum_{l=1}^{P} \beta_l \delta(t - \tau_l)$$  \hspace{1cm} (10)

where $\delta(t)$ represents the Dirac delta function, $\beta_l$ is the complex envelope of the signal received on the $l$th path which is assumed to be a complex Gaussian random process with zero mean and variance $\sigma_l$, and $\tau_l$ is the propagation delay for the $l$th path.

When transmitting the alternative OFDM signal through the channel expressed by (10), the received signal can be expressed as

$$R(t) = S(t) \ast g(t) + Z(t)$$  \hspace{1cm} (11)

where $Z(t)$ represents added complex Gaussian noise. By substituting (7) and (10) into (11), we obtain

$$R(t) = \sum_{i=-\infty}^{\infty} \sum_{k=-p}^{N-1} \sum_{l=1}^{P} \beta_l S_{i,k} h(t - kT - iT_S - \tau_l) + Z(t).$$  \hspace{1cm} (12)

In order to demodulate the $i$th information vector, $N$ samples of $R(t)$ are taken at times $t = uT + iT_S$. The generated samples are

$$R_{i,u} = R(t)|_{t = uT + iT_S}.$$  \hspace{1cm} (13)

For simplification purposes, let us express $R_{i,u}$ as a function of $G_m$, where $G_m$ is defined as

$$G_m \stackrel{\text{def}}{=} \sum_{l=1}^{P} \beta_l h(mT - \tau_l).$$  \hspace{1cm} (14)

The values of $R_{i,u}$ can be approximated considering that $G_m$ has significant values only for $-L \leq m \leq L$. In this case, $R_{i,u}$ can be expressed as

$$R_{i,u} = \Gamma_{i,u} + \Lambda_{i,u} + \Omega_{i,u} + Z_{i,u}$$  \hspace{1cm} (15)

where

$$\Gamma_{i,u} = \sum_{m=0}^{L} S_{i,u} (N+m+u) \mod N | G_m,$$

$$0 \leq u \leq N - 1$$  \hspace{1cm} (16)

$$\Lambda_{i,u} = \sum_{m=0}^{L} (S_{i,-u-N-m} - S_{i,1-N-m}) G_m$$

$$0 \leq u < L$$  \hspace{1cm} (17)

$$\Omega_{i,u} = \sum_{m=0}^{L} (S_{i,m+u-N} - S_{i,-m+u-N}) G_m$$

$$N - L \leq u \leq N - 1$$  \hspace{1cm} (18)

and $Z_{i,u}$ represents the samples of the added Gaussian noise.
The next step in the demodulation of the \(i\)th symbol, which carries the \(i\)th information vector \(\{A_i,n\}\), is to take the DFT of the vector \(\{R_{i,n}\}\) to obtain
\[
K_{i,n} = \gamma_{i,n} + \lambda_{i,n} + \omega_{i,n} + z_{i,n} \quad (19)
\]
where \(\gamma_{i,n}, \lambda_{i,n}, \omega_{i,n},\) and \(z_{i,n}\) are the DFT of the terms \(\Gamma_{i,u}, \Lambda_{i,u}, \Omega_{i,u},\) and \(Z_{i,u}\), respectively.

By taking the DFT of the term \(\Gamma_{i,u}\), we obtain
\[
\gamma_{i,n} = A_{i,n} \sum_{m=-L}^{L} G_m e^{-j2\pi mn/N} = A_{i,n} C_n. \quad (20)
\]

Therefore, \(\gamma_{i,n}\) represents the information symbol transmitted by the \(n\)th carrier multiplied by a distortion factor \(C_n\) which can be expressed as
\[
C_n = \sum_{m=-L}^{L} G_m e^{-j2\pi mn/N} = \sum_{l=1}^{P} \beta_l \sum_{m=-L}^{L} h(mT - \tau_l) e^{-j2\pi mn/N}. \quad (21)
\]

We can see that \(C_n\) is the sum of \(P\) zero-mean complex Gaussian random variables; therefore, its envelope is a Rayleigh random variable.

Let us assume that each element of the vector \(\{A_i,n\}\) is chosen out of \(\{e^{j2\pi m/M}|m = 0, 1, \cdots, M-1\}\). Therefore, the average power of the \(n\)th carrier can be expressed as
\[
b_n = E[|C_n|^2] = \sum_{l=1}^{P} \beta_l \sum_{m=-L}^{L} \sum_{k=-L}^{L} h(mT - \tau_l) h(kT - \tau_l) \cdot \cos[2\pi(m-k)n/N]. \quad (22)
\]

The second and third terms of (19), \(\lambda_{i,n}\) and \(\omega_{i,n}\), represent intersymbol interference (ISI). The term \(\lambda_{i,n}\) represents interference coming from past symbols, and \(\omega_{i,n}\) represents interference coming from future symbols. The third term \(z_{i,n}\) represents added Gaussian noise.

The interference power in the \(n\)th carrier, related to the term \(\lambda_{i,n}\), can be expressed as
\[
\sigma^2_{\lambda,n} = E[|\lambda_{i,n}|^2]. \quad (23)
\]

By taking the DFT of the term \(\Lambda_{i,u}\), and substituting into (23), we obtain
\[
\sigma^2_{\lambda,n} = \frac{1}{N} \sum_{l=1}^{P} \beta_l \sum_{m=-L}^{L} \sum_{k=-L}^{L} h(mT - \tau_l) h(kT - \tau_l) \cdot I(u, k, m, q) h(qT - \tau_l) \cdot \cos[2\pi(m-k)n/N] \quad (24)
\]

where
\[
I(u, k, m, q) = \begin{cases} 1, & u - m = k - q \\ 0, & \text{otherwise} \end{cases}
\]

\(\beta_l\) is the variance of the signal received on the \(l\)th path, and \(p\) is the size of the cyclic prefix. Similarly, the interference power in the \(n\)th carrier, related to the term \(\omega_{i,n}\), can be developed and expressed as
\[
\sigma^2_{\omega,n} = \frac{1}{N} \sum_{l=1}^{P} \beta_l \sum_{m=-L}^{L} \sum_{k=-L}^{L} h(mT - \tau_l) h(kT - \tau_l) \cdot I(u, k, m, q) h(-mT - \tau_l) h(-kT - \tau_l) \cdot \cos[2\pi(u-k)n/N] \quad (26)
\]
where

\[ I(u, k, m, q) = \begin{cases} 1, & u + m = k + q \\ 0, & \text{otherwise.} \end{cases} \]  

(27)

Observe that the term \( \sigma_{m,n}^2 \) can be attenuated or even eliminated by increasing the cyclic prefix \( p \). We could also have added a cyclic extension in order to attenuate the term \( \sigma_{m,n}^2 \). In practice, only the cyclic prefix is worth adding because, in general, \( \sigma_{m,n}^2 \ll \sigma_{0,n}^2 \).

The average SNIR in the \( n \)th carrier is given by

\[ \overline{P}_n = \frac{b_n}{\sigma_{X,n}^2 + \sigma_{z,n}^2 + \sigma_z^2} \]  

(28)

where \( \sigma_z^2 \) is the variance of the Gaussian random noise \( z(\hat{t}) \).

The BER of the QDPSK signal (see [16, p. 786]) in the \( n \)th carrier can be computed by

\[ P_{b,n} = \frac{1}{2} \left[ 1 - \frac{p_n}{\sqrt{2} \mu_n} \right] \]  

(29)

where

\[ \mu_n = \frac{\overline{P}_n}{1 + \overline{P}_n}. \]  

(30)

The term \( \overline{P}_n \) is the average received SNIR in the \( n \)th carrier, which can be computed by (28).

To confirm the validity of these equations, BER performance based on this theory was compared with the results of computer simulations. The channel model used was a 16-ray Rayleigh fading channel where the strength of each ray fluctuates independently with the Rayleigh distribution. Fig. 14 illustrates the power delay profile (PDP) of the proposed channel model. This PDP is based on indoor measurement results which are reported in [11]. Assuming that the number of carriers is \( N = 32 \), Fig. 15 shows the average bit-error rate performance over the 32 carriers, which can be computed by

\[ P_b = \frac{1}{N} \sum_{n=0}^{N-1} P_{b,n}. \]  

(31)

Each carrier was modulated by QDPSK at a symbol rate of \( 20/32 \) Msymbols/s, so the total symbol rate was 20 Msymbols/s. The pulse-energy-to-noise-density ratio \( Eb/N_0 \) was set to 30 dB. Simulations were run for five values of the cyclic prefix \( (p = 0, 1, 2, 3, \text{ and } 4) \). This means that a guard period of \( (p + p + N) \times T_s / 8 \) was wasted per symbol period. For \( p = 4 \), the wasted guard period corresponds to a power loss of only 0.51 dB. It can be seen that the theoretical results agreed well with the simulation results.

**D. Conventional Versus Alternative OFDM**

The bandwidth efficiency of the alternative OFDM depends on the rolloff factor \( \alpha \). A small \( \alpha \) is desired because bandwidth efficiency is inversely proportional to \( \alpha \). For example, \( \alpha = 0.1 \) corresponds to an excess bandwidth of 10%. In this case, a symbol rate of 20 Msymbols/s occupies a bandwidth of 22 MHz. This bandwidth efficiency is close to what can be achieved with the conventional OFDM. Fig. 16 plots...
the theoretical bit-error rate performance $P_b$ for QDPSK of conventional and alternative OFDM in the 16-path Rayleigh channel model of Fig. 14. Three values of the rolloff factor were considered, $\alpha = 0.1$, 0.5, and 1.0. The conventional OFDM curve was plotted using the theory developed in [15] for OFDM under frequency-selective fading. A cyclic prefix of $(4/36)T_S$ (or a 0.51 dB loss) was considered for both OFDM schemes. It can be seen that using a small rolloff factor causes degradation in the performance at high values of $E_b/N_0$. This is due mainly to the term $\omega_{\tau,n}$ of (19), which represents the ISI coming from future symbols, a kind of ISI that does not occur in the conventional OFDM. However, for $E_b/N_0$ below 20 dB, the range of practical interest, the performance of the alternative OFDM is close to the performance of the conventional OFDM, even for a rolloff factor as small as 0.1. Therefore, the alternative OFDM can match the bandwidth efficiency of the conventional OFDM without sacrificing performance.

**E. Coding Requirement**

In dispersive channels, when the total bandwidth is subject to frequency-selective fading, the components of $\{A_{\tau,n}\}$ transported by carriers that are close to a spectral null are badly attenuated, while other components may have their power enhanced. This could cause errors to be concentrated in these badly attenuated carriers. This problem can be treated by using codes capable of spreading each information bit among the carriers as much as possible. In this case, recovery of the modulated data may still be possible provided that enough carriers are not badly attenuated. A coding scheme capable of fulfilling this purpose is trellis-coded modulation (TCM). Moreover, TCM does not require bandwidth expansion in order to achieve the information redundancy required for coding. Instead, TCM expands the constellation.

We used a simple TCM code combined with multicarrier modulation and interleaving to simulate the FFA in the three locations described in Section VI. The simulation is represented by the block diagram of Fig. 17, which can be used to represent both the conventional OFDM and the alternative OFDM simulations. However, in the alternative OFDM case, the IDFT and DFT blocks are $N$-point FFT’s; therefore, there is no need for padding with zeros. Moreover, the low-pass filters (LPF) used in the transmitter of the alternative OFDM signal are raised-cosine filters.

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The chosen TCM code expands the constellation from the two-point constellation of Fig. 18(a) to the four-point constellation of Fig. 18(b). Therefore, a BPSK signal is encoded into a QPSK signal.

In the design of the trellis code for fading channels, the main objective is to achieve as large a shortest error event path [16, p. 831] as possible since this parameter is equivalent to the amount of diversity in the received signal. In order to increase the shortest error event path, we have to increase the number of states in the trellis that is used to encode the BPSK signal into the QPSK signal. We used the four-state trellis depicted in Fig. 19 in the simulations. This figure also shows how the two-phase information signal is mapped into the four-phase signal. Fig. 20 shows the shortest error event path provided by this trellis code. As we can see, the shortest error event path spans over a period of three encoded symbols, and since three consecutive symbols are transmitted by different carriers, the TCM code provides diversity over three carriers.

This trellis-coding scheme can be made more effective by maintaining frequency separation, which can be accomplished by interleaving or sufficiently separating successive encoded...
symbols. To better understand this, we should note that if a given carrier is deeply attenuated for being close to a spectral null, then it is likely that the adjacent carriers will also be attenuated to some extent. Therefore, if interleaving is not used, three consecutive symbols that are transmitted by the three adjacent and attenuated carriers will suffer the fading effect. It is easy to see that the TCM coding, which can provide diversity over three carriers, will not be very effective if the three carriers happen to be attenuated. By interleaving the symbols, we avoid that consecutive symbols are transmitted by adjacent carriers, which makes the TCM coding scheme more effective.

In the simulations, the block interleaver in Fig. 21 was used.

F. Representing the Measured Channels in the Simulations

As seen in (10), a static frequency-selective channel can be modeled as a tapped delay line, each tap \( \beta_q \) being the complex envelope of the signal received on the \( q \)th path. Unfortunately, the noncoherent channel-measuring experiment described in [3] could only be used to obtain the amplitude frequency response across a 20 MHz channel. This information is not enough to determine the complex tap values of (10). However, the simulations can be run without knowing the actual tap values because, provided that the ISI is completely eliminated by a cyclic prefix, the exact tapped delay line of a given measured channel can be replaced by any tapped delay line whose amplitude frequency response satisfactorily approximates the actual amplitude frequency response. These alternative tapped delay lines were obtained under the assumption that any measured channels can be represented by the 16-path Rayleigh channel model in Fig. 14. Then, for each measured channel, a total of \( 10^4 \) tapped delay line outcomes were drawn, and the outcome which best approximated the measured amplitude frequency response was picked. As an example, Fig. 22 plots the experimentally measured amplitude frequency response for one of the measured channels, and the amplitude frequency response of the tapped delay line chosen to represent this channel in the simulations. It has been observed that, for this channel and for all of the other measured channels, the
amplitude frequency response of the chosen tapped delay line approached the respective experimentally measured amplitude frequency response reasonably well.

**G. Simulation Description**

The FFA maximum throughput algorithm described in [3] assumes that a single value of the capture threshold is used by all portables in the microcell. The simulation process described here uses a modified version of this algorithm, where each portable works with its particular capture threshold value. This makes more sense because portables transmit through channels with different characteristics, which causes different portables to be able to stand different amounts of interference. Therefore, in order to maximize throughput, each portable should operate with its minimum acceptable capture threshold value.

The simulation process starts with all of the portables operating at a low capture threshold value and communicating with the base station sector that provides highest average power. The simulation runs until each portable has transmitted at least 100 packets. At this time, the number of packets that arrived in the base station corrupted by error are counted for each portable, and any portable that transmitted more than five corrupted packets has its capture threshold increased by 1 dB, and this ends the first iteration. In the second iteration, the simulation restarts with the portables operating at their new capture threshold. Again, the simulation runs until each portable has transmitted at least another 100 packets. At this time, the number of packets that arrived corrupted by error are counted again for each portable, and any portable that transmitted more than five corrupted packets has its capture threshold increased by 1 dB. This process is repeated for a number of iterations until each portable transmits less than five corrupted packets. If a given portable could not transmit fewer than five corrupted packets per set of 100 packets after having increased its capture threshold to 20 dB, the base station switches the communication with this portable to the antenna sector that provides the second highest average power. This is done because, sometimes, the best sector for communicating with a given portable is not the one that provides highest average power; it also matters how the power is distributed among the carriers. The simulation ends when all of the portables transmit fewer than five corrupted packets per set of 100 transmitted packets. The throughput achieved in the last iteration is taken as the output of the simulation.

Note that five corrupted packets per set of 100 transmitted packets represents a reasonable packet error rate if the system operates with an automatic repeat request (ARQ) protocol. We observed that only a few portables were subjected to a packet error rate of (5/100), and most of the portables transmitted no corrupted packets during the last iteration of the simulation. Therefore, the average packet error rate was much smaller than (5/100).

In order to better understand the computations that were performed during the simulations, let us suppose that, in a given time slot, two portables \( P_1 \) and \( P_2 \) are allowed to transmit. Fig. 23 illustrates this scenario.

During the \( i \)th symbol period, the base station receives symbols \( \gamma_{i,n_1} \) and \( \gamma'_{i,n_2} \) through channels “P1S6” (Portable 1–Sector 6) and “P2S6” (Portable 2–Sector 6), respectively. Symbol \( \gamma_{i,n_1} \) carries the desired information transmitted by portable \( P_1 \) in the \( n \)th carrier during the \( i \)th symbol period. Symbol \( \gamma'_{i,n_2} \) represents the interference from \( P_2 \) received by the base station in the same carrier during the same symbol period. Therefore, the symbol received during the \( i \)th symbol period in the \( n \)th carrier from portable \( P_1 \) can be expressed as

\[
R_{i,n_1} = \gamma_{i,n_1} + \gamma'_{i,n_2} + z_{i,n}
\]  

(32)

where \( z_{i,n} \) is the Gaussian noise term. In the simulations, this term was such that the \( Eb/N0 \) averaged over the \( N \) carriers was made equal to 20 dB. The term \( \gamma_{i,n_1} \) can be calculated as

\[
\gamma_{i,n_1} = D_{i,n} C_{i,n}
\]  

(33)
where $C_n$ represents the channel response for the $n$th carrier. In the case of conventional OFDM, it is expressed by (see [15])

$$C_n = \sum_{l=1}^{P} \beta_{l} e^{-j2\pi f_l n}$$  \tag{34}$$

where $\beta_{l}$ is the complex envelope of the signal received on the $l$th path. In the case of alternative OFDM, it is expressed by (21). The term $D_{t,n,1}$ represents the differentially encoded information transmitted by portable $P_1$ during the $t$th symbol period in the $n$th carrier:

$$D_{t,n,1} = D_{t-1,n,1} A_{t,n,1}$$  \tag{35}$$

where $A_{t,n,1}$ represents the information symbol, encoded by the TCM scheme and interleaved, that must be recovered during the $t$th symbol period in the $n$th carrier.

In the simulations, demodulation assumed differential phase estimation. In this case, the demodulated symbol can be expressed as

$$K_{t,n,1} = R_{t,n,1} R_{t-1,n,1}^*$$

$$= (\gamma_{t,n,1} + \gamma_{t-1,n,2} + z_{t,n})$$

$$\cdot (\gamma_{t-1,n,1} + \gamma_{t-1,n,2} + z_{t-1,n})^*$$

$$= (\gamma_{t,n,1}^* \gamma_{t-1,n,1}^*)^* + (\text{interference and noise terms})$$

$$= (D_{t,n,1} C_n)(D_{t-1,n,1} C_n)^*$$

$$+ (\text{interference and noise terms})$$  \tag{36}$$

where $|C_n|$ represents the gain of the $n$th carrier in channel “P1S6” (see Fig. 23). The $K_{t,n,1}$ sequence is then deinterleaved and decoded by the TCM decoder (see Fig. 17). In the case of alternative OFDM, the ISI term $\lambda_{t,n}$ was disregarded based on the assumption that an adequate cyclic prefix can be used. The other ISI term, $\omega_{t,n}$, was also disregarded since, in the simulations, $\omega_{t,n} \ll z_{t,n}$. In the case of the conventional OFDM, ISI was disregarded based on the assumption that an adequate cyclic prefix can be used (see [15]).

The other features of the simulations are the following.

- The number of carriers used was $N = 32$.
- A rolloff factor of 0.1 was assumed in the case of the alternative OFDM.
- A packet size of only 64 bits was used in order to accelerate the simulations, so the number of symbols (bits) per packet transmitted on each carrier was two; one was used to provide phase reference for the differential detector, and the other was encoded by the trellis code. This packet size seems to be too small, but we can expect that larger packet sizes would provide similar results. This is due to the fact that, when a subset of $N$ portables share a given time slot, the signal and the $N-1$ interference terms related to the transmission of each of the $N$ packets can be assumed to remain unchanged during the time slot in the case of quasi-static channels. There is also a Gaussian noise term affecting each packet, but this random term is 20 dB below the power level of the signal, so the errors are caused primarily by the interference terms. Therefore, considering only the deterministic $N-1$ interference terms, if a one-symbol packet is received without errors, then a multisymbol packet will also be received without errors since the interference terms are the same for each symbol period of the packet.

- The Viterbi algorithm was used in the decoder.
- Only uplink traffic was simulated, and it considered the power control mechanism described in Section II.
- Simulations were run with the modified FFA operating with different numbers of ports, from two ports up to a number of ports where the maximum throughput saturated.

H. Simulation Results

The simulation results for locations 1, 2, and 3 are shown in Figs. 24–26, respectively. In location 1, the maximum throughput saturated with three ports, meaning that it is pointless to use more than three ports in this location with multicarrier modulation, the TCM coding scheme, and a sectorization level of 10. The maximum throughput in location 1 was 2.8 packets per time slot when using the alternative OFDM, and 2.5 packets per time slot when using the conventional OFDM. In location 2, the maximum throughput saturated at 3.9 packets per time slot when using the alternative OFDM, and at 4.2 packets per time slot when using the conventional OFDM. In location 3, the maximum throughput saturated with five ports at 4.3 packets per time slot when using the alternative OFDM, and at 4.2 packets per time slot when using the conventional OFDM. The main conclusion is that the alternative OFDM scheme provides performance similar to the performance achieved with the conventional scheme.

It was mentioned in Section VI that each one of these three locations is representative of an indoor environment, namely, closed, semiopen, and open. Therefore, the simulation results can provide a raw estimation of the capacity gain that can be obtained in these environments with the FFA operating...
with a sectorization level of ten, multicarrier trellis-coded modulation, and differential estimation.

VIII. CONCLUSIONS

A new technique has been proposed for reusing frequency in broadband indoor wireless systems. It consisted of using sectored antennas and a suboptimum frame-scheduling algorithm to allocate simultaneous packet transmissions to/from the base station.

Measurement results obtained in some typical indoor locations were used in conjunction with coded multicarrier modulation to simulate the proposed multiple-access scheme. The simulations showed that the proposed frequency reuse scheme can transmit, on average, close to three packets per time slot in a closed indoor location, and more than four packets per time slot in a semiopen or open indoor location.

Due to the limitations of the experiment, the results can be guaranteed only for the case of static channels. However, we expect to obtain similar results with quasi-static indoor scenarios, where the base station will be able to update the information regarding the compatibility relations among the portables in the microcell.

We suggest the following areas for further research on the topic.

- Investigate different levels of sectorization, and the use of antenna arrays to replace the sectored antennas.
- Study maximum throughput and portable blocking issues when the microcell operates with specific traffic compositions, for example, ATM traffic.
- Investigate how channel dynamics affect the amount of overhead required for allowing the base station to track the compatibility relations in the microcell, and investigate throughput loss when the tracking is not perfect.
- Study the effect of intercell interference.
- Investigate the performance of the proposed multiplexing scheme in outdoor fixed wireless system. For example, local multipoint distribution systems (LMDS) and wireless local loop systems (WLL).

REFERENCES

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