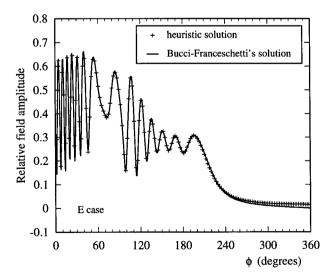
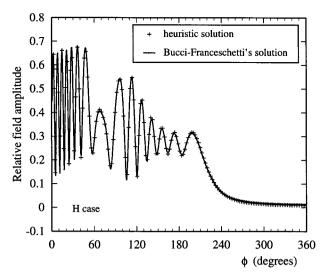


**Figure 8** Field scattered by an anisotropic half plane with  $|Z_x| = \infty$ ,  $Z_{y'} = 0$  illuminated at  $\beta_0 = 45^\circ$ ,  $\phi' = 60^\circ$  by a  $\beta'$ -polarized plane wave

tion coefficients for the scattering by a half plane with two isotropic face impedances [6]. The results shown here are relevant to the field scattered by an isotropically loaded  $(Z_{x'}/\zeta = Z_{y'}/\zeta = 0.5 - j1.4)$  half plane illuminated at normal incidence by a cylindrical wave radiated by an isotropic line source located at  $\rho' = 10\lambda$ ,  $\phi' = 45^{\circ}$ ,  $\rho'$  being the radial distance from the edge. The observation point is at a distance  $\rho = 5\lambda$  from the edge. Two different cases of incident cylindrical waves are considered: the E-case and the H-case. The first refers to an incident electric field parallel to the edge (TM with respect to the edge), whereas the latter regards an incident magnetic field parallel to the edge (TE with respect to the edge). Since depolarizing effects do not occur in the case of isotropically loaded surfaces, only the copolar components of the scattered electric field are reported in Figures 9 and 10. As can be seen, in both the cases, the results agree very well, thus validating the developed approach.



**Figure 9** Field scattered by an isotropically loaded  $(Z_{x'}\zeta = Z_{y'}/\zeta = 0.5 - j1.4)$  half plane illuminated at normal incidence by a cylindrical wave radiated by an isotropic line source located at  $\rho' = 10\lambda$ ,  $\phi' = 45^{\circ}$ 



**Figure 10** Field scattered by an isotropically loaded  $(Z_{x'}/\zeta = Z_{y'}/\zeta = 0.5 - j1.4)$  half plane illuminated at normal incidence by a cylindrical wave radiated by an isotropic line source located at  $\rho' = 10\lambda$ ,  $\phi' = 45^{\circ}$ 

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# AN EFFECTIVE THREE-DIMENSIONAL MODEL FOR UHF RADIO PROPAGATION INTO MULTISTORY OFFICE BUILDINGS

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ABSTRACT: An effective three-dimensional (3-D) direct-transmitted model combined with a patched-wall model is proposed to explore UHF radio propagation into multistory buildings in a microcellular environment. The model is verified by comparing the computed path loss with the measured one of 1.7 and 2.44 GHz radio waves propagating at different sites. The comparison demonstrates that the model has reasonably accurate prediction when the number of floor levels between the transmitting and the receiving antennas is no more than four.

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**Key words:** UHF radio propagation; field measurement; indoor radio propagation modeling

## I. INTRODUCTION

The increasing demand on system quality for future personal communication systems or wireless local area networks necessitates serving large numbers of users in buildings with similar good quality as on the streets. In this letter, radio signals at 1.7 and 2.44 GHz penetrating into buildings from a microcell base station installed outside have been investigated.

There have been a number of recent studies on ratio propagation into buildings [1–5]. In [1] and [3], statistical models for the received signal are developed based on the measured data. In [2, 4, 5], the building penetration loss is quantified by the experiments. In our approach, a 3-D direct-transmitted ray (DTR) model has been developed to explore radio wave propagation into a multistory office building. Our model will be ascertained by comparing the experimental results of 1700 MHz and 2.44 GHz radio waves propagating through the exterior wall, multiple floors, and multiple interior walls. Since the model is site specific, detailed geometric features and constitutive materials of these incidence boundaries are considered, and are described by a patched-wall model [6] which will enhance the prediction accuracy of the model.

#### II. PROPAGATION MODELS

A. Ray-Fixed Coordinate Systems. To trace the rays in a three-dimensional space and to include the polarization of the field, the transmission or reflection coefficient of the propagating field has to be represented by a  $3 \times 3$  matrix in a global coordinate system in the case of oblique incidence at boundaries. The components of the matrix sometimes are difficult and time consuming to determine. With a ray-fixed coordinate system being introduced, the matrix is reduced to a 2 × 2 matrix whose elements are much more easily determined. The ray-fixed coordinate is a local coordinate system whose coordinate axes are determined by the direction of the incident (reflected) ray and the normal unit of the reflecting boundary. In Figure 1, the incident and reflected ray-fixed coordinates, represented by  $(\hat{s}_1, \hat{\alpha}_1, \hat{\beta}_1)$  and  $(\hat{s}_2, \hat{\alpha}_2, \hat{\beta}_2)$ , respectively, are the local coordinate systems.  $\hat{s}_1$  and  $\hat{s}_2$  are the unit vectors representing the directions of incident and reflected rays, respectively. The incident or reflected electric field at position R can always be resolved into components perpendicular  $(\hat{\alpha}_i)$  and parallel  $(\hat{\beta}_i)$  to the plane of inci-

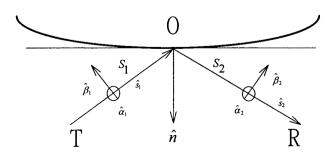


Figure 1 Ray-fixed coordinate system associated with the reflection at the boundary

dence. Detailed definitions of the both coordinate systems can be found in [6, 7]. In the ray-fixed coordinate systems, the matrices of the reflection and transmission coefficients become, respectively,

$$\overline{\overline{R_c}} = \begin{bmatrix} R_s & 0 \\ 0 & R_p \end{bmatrix} \quad \text{and} \quad \overline{\overline{T}} = \begin{bmatrix} T_s & 0 \\ 0 & T_p \end{bmatrix}. \tag{1}$$

Here,  $R_s(T_s)$  and  $R_p(T_p)$  are the reflection (transmission) coefficients of perpendicular and parallel polarizations, respectively, and they are determined by the *ABCD* matrix [8] since the boundaries may have layered structures. In this approach, the propagation problem is formulated as a transmission line problem by treating each layer as a segment of a transmission line with a specific impedance.

B. 3-D Model. To examine the propagation mechanism in detail, a three-dimensional (3-D) direct-transmitted ray (DTR) model is proposed. The model has been simplified to solve the two-dimensional NLOS (nonline-of-sight) indoor propagation problem effectively [9]. The easy-to-use DTR model is proposed here not only because of its computational efficiency, but also because of its ability to check whether the direct-transmitted ray is the dominant propagation mode in evaluating the median path loss. In the model, the received complex field is computed by

$$\vec{E}_R = \vec{E}_o \cdot G_t \cdot G_r \cdot L(d) \cdot \prod_i \overline{\bar{T}}(\theta_i)$$
 (3)

where  $G_t$  and  $G_r$  are the field-amplitude radiation gains at the direction of the direct-transmitted ray of the transmitting and the receiving antennas, respectively. L(d) is the free-space path loss with distance d between both antennas.  $\overline{T}(\theta_i)$  is the transmission coefficient for the ith incidence boundary with an incident angle  $\theta_i$ .

# III. MEASUREMENT SETUP AND EXPERIMENTAL SITES

A. Measurement Setup and Procedure. Narrowband (CW) signal strength measurements were made at 1.7 GHz. A halfwave length dipole antenna at a height of 4.5 m above the ground transmitted a 13 dBm CW signal. The transmitting system, including a signal generator, a section of cable, and the transmitting antenna, has been calibrated in an anechoic chamber to measure the 1 m transmitting field strength in free space. A receiving antenna connected to a preamplifier is also a half-wave length dipole antenna (Anritsu MP651B and MP663A) with the same height. Both the transmitting and the receiving antennas are vertically polarized during the measurement. The receiver (Advantest R3261A) can instantaneously measure the signal strength between -30 and -100 dBm over a 100 kHz interval. The received data are acquired automatically by a personal computer with a GPIB card. At each measured spot, the measured field strength is obtained by a spatial sector averaged over nine grid subpoints with a quarter-wavelength spacing between neighboring points. To assure that the propagation channel is time stationary during the measurement, the measured data have been averaged on screen over ten instantaneous sampled values.

B. Measurement Sites. The measurement sites are on different floors of Engineering Building Four, a typical office

building, at the National Chiao-Tung University in Hsin-Chu. Basically, the exterior walls and floorboards are made of reinforced concrete. Figure 2 illustrates a floor layout of the south wing, and the transmitting antenna is situated outside the building. Measurements along the hallway will be executed at each floor repeatedly from the second to the eighth floors. The measured spots along the hallway for the third floor are also illustrated in Figure 2. On the second and third floors, measurements of 1.7 and 2.44 GHz radio waves are carried out in several classrooms such as rooms 301–303 and 201–203 shown in Figure 2. The measured spots in these rooms on each floor are arranged in three rows, with the spacing between the neighboring rows equal to 2.7 m.

One more measurement site, the third floor of Engineering Building Three, is chosen. The building is similar to Engineering Building Four in terms of main construction features and floor layout. Figure 3 illustrates the top view of the measured site in Engineering Building Three. The path loss of the 1.7 GHz radio wave is measured along the corridor of the third floor of Engineering Building Three.

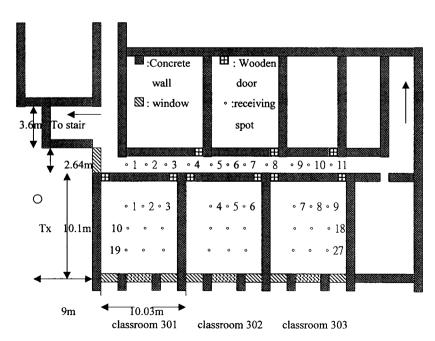
#### IV. COMPARISONS

In the numerical simulation made with the DTR model, the dielectric constants of the concrete wall, plasterboard, window glass, and floorboard are chosen equal to 7-0.7j, 3-0.005j, 3-j0.05, and 10-0.5j for 1.7 GHz, and 7-0.5j, 3-0.0037j, 3-0.037j, and 10-0.37j for 2.44 GHz, respectively [9, 10]. Plasterboards mainly form most of the interior walls above the fourth floor. To examine the prediction accuracy of the DTR model, the path loss computed by the model is compared with the measured one. In Figure 4, a comparison is made for the 1.7 GHz case of measurement along the hallway on the second and sixth floors of Engineering Building Four. In the case of 2.44 GHz, the measured path losses and the computed result at the second and sixth

floors are illustrated in Figure 5. It is found that the model gives good prediction accuracy for both frequencies. To make the paragraph more concise, comparison results for other floors at frequencies of 1.7 and 2.44 GHz are computed and given in Table 1. It is found that the DTR model still yields good prediction accuracy, except at the seventh and the eighth floors. This is because the DTR model may have neglected some significant multiply transmitted-reflected rays, and yields an overestimation of the path loss when the receiver is at higher floors such as at the seventh or eighth floors. When the receiving positions are in the classrooms of Engineering Building Four, the comparison results for 1.7 and 2.44 GHz on the second and third floors are illustrated in Figures 6-9. It is found that the model shows a similar good performance. It is noted that values of  $\sigma_e$  for the measuring spots located in the classrooms are larger than those on the different floors. This is due to the fact that a large fading depth will be expected in the classrooms. The comparisons above show that the direct-transmitted ray is the dominant propagation mode when the number of floor levels between the transmitting and receiving antennas is no more than four. Figure 10 shows the comparison result of the measurement site on the third floor of Engineering Building Three, and the model again yields consistent prediction accuracy.

## V. CONCLUSION

The mechanism of UHF radiowave propagation into multistory office buildings is explored by using an effective three-dimensional DTR model. It is found that: 1) the direct transmitted wave is the dominant mode when the number of floor levels between the transmitting and receiving antennas is not so large such that the level of the direct wave is still higher than those multiply-transmitted signals at the receiver, and 2) the path loss neither increases nor decreases monotonically as a function of increasing floor level. This is be-



**Figure 2** Third floor layout of the south wing of Engineering Building Four. Measurement along the hallway will be executed repeatedly at each floor from the second to the eighth floors. Only at the second and third floors are the measured spots planned in the classrooms

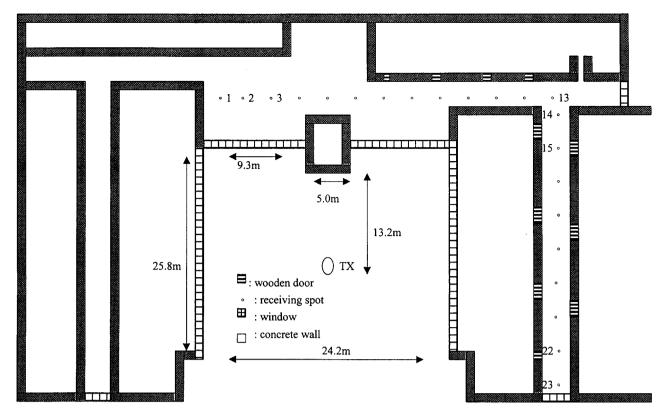
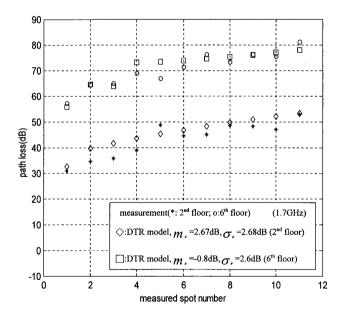
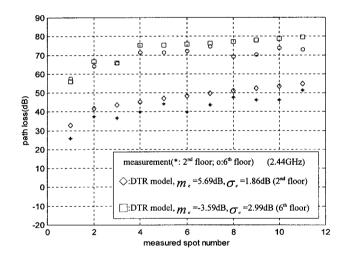


Figure 3 Top view of the third floor of Engineering Building Three. Measurement spots are located along the corridor



**Figure 4** Measured and computed path losses of 1.7 GHz radio wave as a function of measured spot number on the second and sixth floors of Engineering Building Four

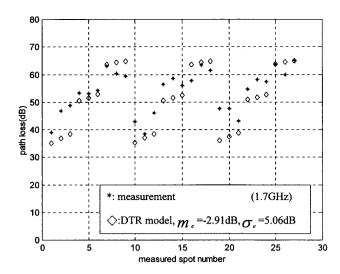
cause the increasing number of multiply transmitted-reflected or -scattered received paths compensates for the increasing path loss when the number of floor levels is increased. After comparison with the measured path loss, the DTR model combined with a patched-wall model yields consistent prediction accuracy for different propagation scenarios and frequencies.



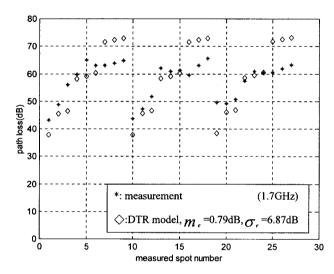
**Figure 5** Measured and computed path losses of 2.44 GHz radio wave as a function of measured spot number on the second and sixth floors of Engineering Building Four

TABLE 1 Values of  $m_{\rm e}$  and  $\sigma_{\rm e}$  Illustrated from the Third to Eighth Floors in Engineering Building Four for 1.7 and 2.44 GHz

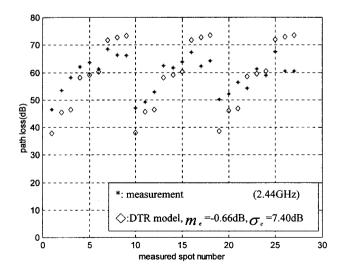
Floor No.		3rd	4th	5th	7th	8th
1.7 GHz	$m_e$ (dB)	1.21	-1.70	-0.89	7.13	10.65
	$\sigma_{e}$ (dB)	2.4	2.9	3.95	2.58	4.02
2.44 GHz	$m_e$ (dB)	2.24	3.94	4.53	13.59	13.1
	$\sigma_{e}$ (dB)	3.98	2.31	4.64	3.08	4.98



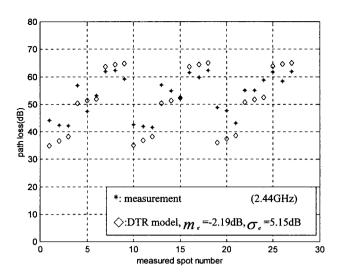
**Figure 6** Measured and computed path losses of 1.7 GHz radio wave for the measured spots located in classrooms 201–203 in Engineering Building Four



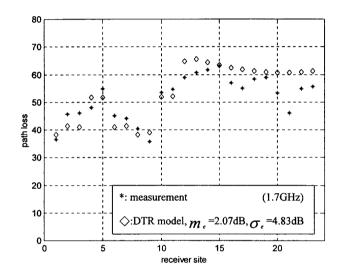
**Figure 7** Measured and computed path losses of 1.7 GHz radio wave for the measured spots located in classrooms 301–303 in Engineering Building Four



**Figure 8** Measured and computed path losses of 2.44 GHz radio wave for the measured spots located in classrooms 201–203 in Engineering Building Four



**Figure 9** Measured and computed path losses of 2.44 GHz radio wave for the measured spots located in classrooms 301–303 in Engineering Building Four



**Figure 10** Measured and computed path losses of 1.7 GHz radio wave as a function of measured spot number on the third floor of Engineering Building Three

## **ACKNOWLEDGMENT**

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# APERTURE-COUPLED CYLINDRICAL DIELECTRIC RESONATOR ANTENNAS FORMING FOUR-ELEMENT LINEAR ARRAYS

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ABSTRACT: Two four-element linear aperture-coupled cylindrical dielectric resonator antenna arrays have been theoretically modeled and experimentally implemented. The arrays produced broadside radiation shaping the E- and H-plane, respectively. The resonant frequency, return loss, radiation pattern, gain, unloaded Q-factor, and impedance bandwidth of the arrays have been measured. Overall, good agreement was obtained between theory and experiment. © 1999 John Wiley & Sons, Inc. Microwave Opt Technol Lett 20: 151–153, 1999.

**Key words:** dielectric resonator antennas; linear antenna arrays; aperture coupling

# 1. INTRODUCTION

Cylindrical dielectric resonator antennas (CDRAs) offer wide impedance bandwidth, low Q-factor, and high radiation efficiency as well as small size, low cost, and ease of excitation [1, 2]. Among the various excitation methods for the CDRAs, the aperture coupling has been widely used as it allows the CDRAs to be integrated with MMICs [3, 4]. A two-element linear CDRA array was studied experimentally by Chow et al. [5]. The CDRAs were excited by aperture-coupled sources and operated at the fundamental  $TM_{11\delta}$ -mode. Both the E-and H-plane configurations were considered. It was found that the antenna gain at resonance ( $\sim 5~\mathrm{GHz}$ ) was  $\sim 9~\mathrm{dBi}$ 

( $\sim$  7.9) for both array configurations, which was 2.3 dB higher than that of the single element.

Continuing Chow's work, two four-element linear broadside arrays using aperture-coupled CDRAs are reported in this letter, demonstrating the feasibility of this array configuration. The arrays shape the E- and H-plane, respectively. The arrays are analytically formulated using the perfect magnetic wall model (PMWM), numerically simulated with a software package based on the finite-element method (FEM) and experimentally implemented. The resonant frequency, radiation pattern, gain, unloaded Q-factor, and impedance bandwidth of each array configuration are reported.

### 2. MODELING OF ARRAYS

The E- and H-plane configurations of the CDRA arrays are presented in Figure 1(a) and (b), respectively. According to the principle of pattern multiplication, the total electric field  $\overline{E}$  for a four-element linear array is determined by the expression

$$\overline{E} = 4 \times \overline{E}_o \times AF \tag{1}$$

Top View

where  $\overline{E}_o$  is the electric field due to a single element placed at the center of the array and AF is the array factor,

 $(\phi=0^{\circ})$ Substrate S S S S Aperture  $W_{a}$   $W_{a}$   $W_{a}$ 

feed

(a)

Ground Plane

Microstrip Feedline

Top View

Substrate Ground Plane Microstrip Feedline

Substrate Ground Plane Microstrip Feedline

Substrate Ground Plane Microstrip Feedline

**Figure 1** Four-element linear aperture-coupled CDRA arrays. (a) Top view of Configuration A shaping  $E_{\theta}$  at  $\phi=0^{\circ}$  (E-plane). (b) Top view of Configuration B shaping  $E_{\phi}$  at  $\phi=90^{\circ}$  (H-plane)