ANALYSIS OF BROADBAND WIRELESS COMMUNICATION USING PARALLEL FDTD SIMULATIONS ON URBAN AREAS

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Abstract: Systems for broadband, high-speed wireless communication have been studied using a big parallel FDTD method applied to large 2-D areas. The effect of multipath interference on signal evaluation can be calculated accurately using FDTD.

Key words: FDTD method, parallel processing, linear equation, broadband wireless communications, TDMA, CDMA.

1. Introduction

In numerical study of high-speed communications, good accuracy is required to evaluate the systems and the effect of multipath interference, path loss, etc. Because the simulation of multipath propagation on urban areas requires large memory, we have developed a special parallel FDTD algorithm. Then, using this algorithm, signal propagation and evaluation of TDMA communication systems was done, and the effect of multipath interference on CDMA was studied.

2. Parallel FDTD Algorithm

2.1 Basis of the Method

For isotropic, frequency independent, linear materials, the 2D TM FDTD formulation can be represented as a set of linear equations, with variables E_z, H_x and H_y, as follows:

$$E_z^{n+1}(i,j) = C_a(m)E_z^n(i,j) + C_x(m)[H_y^n(i+1,j) - H_y^n(i,j)] - C_y(m)[H_x^n(i,j+1) - H_x^n(i,j)] \quad (1)$$

$$H_x^{n+1}(i,j) = H_x^{n+1}(i,j) - D_y(m)[E_z^{n+1}(i,j) - E_z^{n+1}(i,j-1)] \quad (2)$$

$$H_y^{n+1}(i,j) = H_y^{n+1}(i,j) - D_x(m)[E_z^{n+1}(i,j) - E_z^{n+1}(i-1,j)] \quad (3)$$

Where C_a, C_x, C_y, D_x and D_y are constants depending only on the properties of m -th material, and the values for the half time steps n+1/2 are represented as time n. If we define X_i as a vector containing all the variables for time step i, the relation between variables in time step i and i+1 would be $X_{i+1} = A_i X_i + \phi_i$, where ϕ_i represents source of currents and A_i is a constant matrix of coefficients. By arranging the equations respect to time steps, we can represent all the FDTD simulation as the next system of linear equations [1]-[2]:

$$\begin{bmatrix} I & 0 & 0 & \cdots & 0 & 0 \\ -A_1 & I & 0 & \cdots & 0 & 0 \\ 0 & -A_1 & I & \cdots & 0 & 0 \\ \vdots & \vdots & \vdots & \ddots & \vdots & \vdots \\ 0 & 0 & 0 & \cdots & I & 0 \\ 0 & 0 & 0 & \cdots & -A_1 & I \end{bmatrix} \begin{bmatrix} X_1 \\ X_2 \\ X_3 \\ \vdots \\ X_{T-1} \\ X_T \end{bmatrix} = \begin{bmatrix} \phi_1 \\ \phi_2 \\ \phi_3 \\ \vdots \\ \phi_{T-1} \\ \phi_T \end{bmatrix} \quad (4)$$

2.2 Parallelization and use of Green's Function

The solution of FDTD simulation becomes equivalent to solution of this triangular system; and it can be done by Gaussian substitution. Normally, substitution would require a lot of memory, but we carry out the solution in partial stages, solving eq.(5):

$$x_L + c_1^1 x_1^1 + c_2^1 x_2^1 + \ldots + c_{z1}^1 x_{z1}^1 = b_L^1 \quad (5)$$

Where x_L is a wanted variable (E_z, H_x or H_y) at some time step t and position (i,j), depending on other variables x_k^1 of time step t -1. Substitution is done until eq.(5) becomes:

$$x_L + c_1^n x_1^n + c_2^n x_2^n + \ldots + c_{zn}^n x_{zn}^n = b_L^1 \quad (6)$$

Where the variables x_k^n are components of the fields in time step t -n. We repeat this process for solving each variable x_k^n as a new x_L. The old and new equations are stored in different arrays on many processors. The recursive algorithm is shown in fig.(1). Memory can be saved because not all the variables are required to be in memory simultaneously.

In order to improve the speed of the algorithm, numerical Green's function for free space were used. In a linear system, we have that the values of Green's function g(**r**,t,**r'**,t') for any component of the fields is equivalent to the coefficients c_k^n for t'=t-n. These values are always the same for any time and position as long as the surrounding space is the same and they can be calculated beforehand and stored for posterior use. The suitability of this approach has been established already in [3].

Instead of doing the substitution for all points, the improved algorithm evaluates the surrounding cells for the position (i,j) and if there is only free space and no sources (antennas), the stored values can be used directly for c_k^n. In the other case, Gaussian substitution is done.

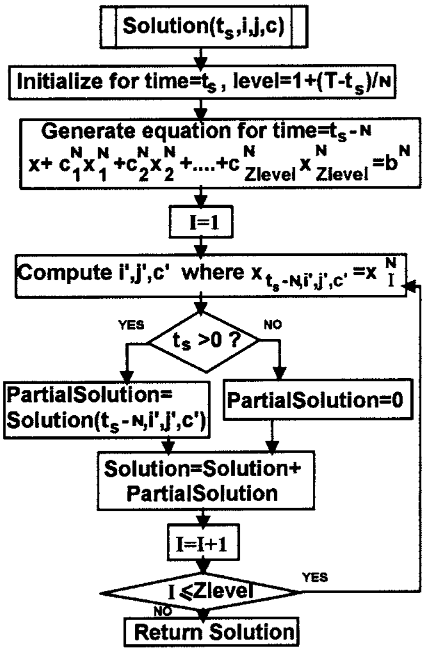

Fig.1: Flowchart of the recursive algorithm

3. Analysis for TDMA communications

For the calculation of Bit Error Rate (BER) of 1 Mbps digital wireless communication systems for TDMA, a model using concrete buildings was used, as shown in fig.(2) and parameters:

- Cell size: 0.02 m
- Time increment Δt: 37.7 ps
- Relative permittivity (concrete): 3.0
- Conductivity of concrete, σ: 0.005 S/m
- Current amplitude: 1000.0 A/m^2

We assumed an antenna of point type, generating a Gaussian pulse transmitted with a carrier f_c= 900 MHz as shown in eq.(7), which is the signal for symbol of bit 1. Symbol of bit 0 has no signal. Basic signal also can be considered as a periodic set of pulses c(t) for symbol 1, affected by the sequence of bits. For any sequence of bits b(t), the modulated signal d(t) to be transmitted can be obtained from the addition of the basic signal for all bits with value 1 adjusted with time shift, as in eq.(7). For any point in the model, if the dynamic response of the signal in eq.(7) is E_z*, then the dynamic response for the sequence of bit represented in eq.(8) is shown in eq.(9). Multiple channels are created in time by assigning a time slot with a limited number of bits to channel 1, then transmitting the data of channel 2 and so on. In real TDMA systems, the standards divide each frequency space in 3 or 8 channels.

$$J_z^{i,n} = J_{\max} \cos(2\pi f_c (t - \varsigma\Delta t)) e^{-\alpha(t-\varsigma\Delta t)^2} \quad (7)$$

$$d(t) = \sum d_i(t) = \sum J_{\max} \cos(2\pi f_c (t - \varsigma\Delta t - \tau_i)) e^{-\alpha(t-\varsigma\Delta t-\tau_i)^2} \quad (8)$$

$$E_z(t) = \sum_{i-th\ bit=1} E_z^*(t - \tau_i) \quad (9)$$

Where J_{max} is the current density amplitude of the source, and $\alpha = -16/(\varsigma\Delta t)^2$, $\varsigma = 6631$. The period between signals is T=1 μs, bit rate is 1 Mbps, and τ_i=(i-1)T. For a symbol 1, the modulated pulse is transmitted in the first half T/2, for the symbol 0, no signal is transmitted on all T. The signal is mostly limited to bandwidth of 300 MHz; components outside 750 MHz - 1050 MHz are negligible.

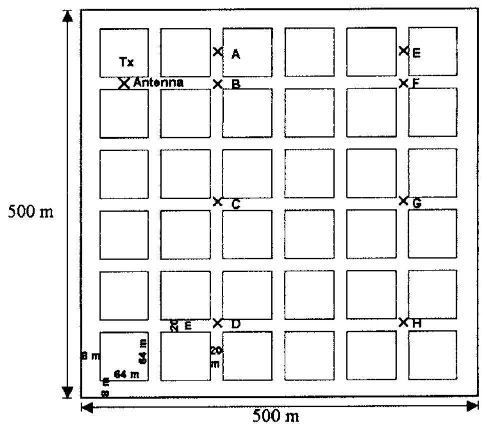

Fig.2: Urban model, size 500m x 500m

For the generation of the received signal, at first, we used the result of the FDTD simulation for only one modulated pulse (symbol 1). Then we generated the signal for a sequence of bits, using time shift and linear composition (addition), that is, we carried out eq.(9). To recover the data from the received signal, first we used envelope detection to eliminate the carrier and to get only the baseband pulses. Then we used a signal correlator and a detector for getting the data. To calculate a value for BER, we used eq.(10), where P(0|1) is the probability of error when the original bit is 1, and P(1|0) is the probability of error when the original bit is 0, $E_D = (E_0 + E_1)/2$, E_0 and E_1 are the average energy of each symbol. This formula is convenient when we have full knowledge of the probability distribution of the errors.

$$BER = \frac{1}{2}[P(1|0) + P(0|1)] \quad (10)$$

$$P(1|0) = \frac{1}{\sigma\sqrt{2\pi}} \int_{E_D}^{\infty} \exp(-\frac{(E-E_0)^2}{2\sigma^2}) dE \quad (11)$$

$$P(1|0) = \frac{1}{\sigma\sqrt{2\pi}} \int_{-\infty}^{E_D} \exp(-\frac{(E-E_1)^2}{2\sigma^2}) dE \quad (12)$$

For a model without multipath propagation, the BER depends only on the SNR, because the pulse does not change shape or has undesired reflections. But for wireless data communication on urban areas, the effect of reflections is important. So, basically, we can study 3 cases: (a) Noise, no multipath interference (b) Multipath interference, no noise, (c) Multipath interference and noise.

Detection of the signal is done using a threshold E_D. Case (a) is textbook case and it is well known. Case (b) is applicable only for very strong signals (noise is negligible). But in the real world, case (c) is the most common. The effect of multipath interference is the displacement of the center of the probability curves from E_0 and E_1 towards E_0' and E_1'. This displacement is unknown beforehand and it can increase BER.

We have decided that both receptor and emitter use a similar dipole antenna (same impedance) with power = 1 W. The value of J_{max} was scaled according to this output power.

Thermal noise is the main candidate as source of errors for eq.(11) and (12). The power and the expected value of the voltage's variance are shown in eq.(13) and eq.(14). Here, $K=1.38 \times 10^{-23}$ J/°K is Boltzmann's constant, T is temperature in Kelvin (290°K= 17°C), R=73 Ω is the resistance, B=80 MHz is the bandwidth in Hz.

$$N = KTB \ (in\ Watts) \quad (13)$$

$$<\sigma_V^2> = 4KTBR \quad (14)$$

The results of the calculation of BER without noise (case (b)) in the model are a perfect communication (BER=0 for all points). The effect of the multipath scattering is never too strong, and it is always proportional to the signal. For the case (c), with noise, we have generated all the possible sequences of 6 bits (2^6=64 sequences), and evaluated the probability of error of the last 5 bits, because the first bit does not suffer multipath interference and therefore it does not belong to case (c). The total number of bits evaluated per position is 5 bits/sequence x 64 sequences= 320 bits. Then we can calculated the probability of error using the integrals on eq.(11) and eq.(12) for all 320 bits, and get the expected BER as shown in eq.(15), for each j-th sequence and i-th bit. This is done for a total of 121 points in the model.

$$BER = \frac{1}{320}\left(\sum_{j=1}^{64} \sum_{i=1, i-th\ bit\ of\ j=0}^{5} P(1\,|\,0) + \sum_{j=1}^{64} \sum_{i=1, i-th\ bit\ of\ j=1}^{5} P(0\,|\,1) \right) \quad (15)$$

The calculated BER is shown in fig.(3). The dotted squares are the buildings. From the figure, we can see that the reception of the data is good on most of the model, except on the opposite area from the emitting antenna. And for nearby points, the best reception generally is in the corners. Along the streets, the BER appears to be related to the street distance from the emitter (that is, the distance moving only along streets).

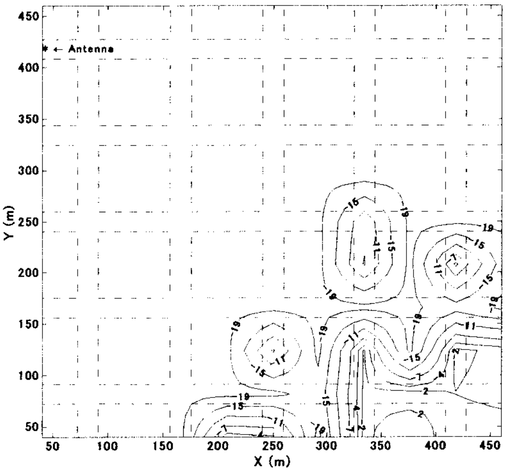

Fig.3: Calculated log(BER) for 1 Mbps TDMA

4. Analysis for CDMA Communications

The original digital data signal for one channel is $b_k(t)$; it is modulated by the carrier $\cos(w_c t)$, $w_c=2\pi f_c$ and by a higher bit rate code sequence or chip sequence $c_k(t)$, which increases the bandwidth of the baseband data signal, as shown in eq.(16) for QPSK, and eq.(17) for D -QPSK. Then this signal $d_k(t)$ is transmitted in the urban model, when it can be affected by noise and multipath interference. As the band of frequencies is shared (multiple access), at any time the total transmitted signal d(t) is the union of many signals d_k, each one with a different data transmission b_k and different code sequence c_k.

In QPSK and QAM case, the value of $\phi(c_k(t))$ is 0, $\pi/2$, π and $3\pi/2$ for the bits 00, 01, 10 and 11 respectively (for the 2 first bits in QAM case). In differential QPSK (D-QPSK), the shift is relative to the last chip, that is, the shift is cumulative, not absolute, with $\psi_k(t)=\psi_k(t-T_c)+\phi(c_k(t))$, T_c is period between chips. If CDMA uses a BPSK, QPSK or D-QPSK to modulate the data with the chip sequence, then $\rho(c(t))=1$. If it uses QAM, $\rho(c(t))$ is variable. In this paper, we use D-QPSK.

$$d_k(t) = b_k(t)\cos\left(w_c t + \phi(c_k(t))\right)\rho(c_k(t)) \quad (16)$$

$$d_k(t) = b_k(t)\cos\left(w_c t + \psi_k(t)\right)\rho(c_k(t)) \quad (17)$$

$$d(t) = \sum_{k=1}^{120} d_k(t) \quad (18)$$

For the simulation of 2 Mbps wireless systems for CDMA, we assumed a signal with a carrier of 900 MHz. The chip rate is 8 Mcps. The simulation is the downlink (base station to user) for simplicity, because it is not affected for the near-far problem. Differential QPSK is also used because it does not need synchronization. The code-set to be used is a Gold code of length 2^7-1=127.

The base station is sending data to 120 channels on different positions at the same time. We generate random data sequences $b_k(t)$ of 8 bits for each point, then modulated it using the Gold code $c_k(t)$ of each receiver and the carrier. The summation of all channel signals $d_k(t)$ is the signal d(t) to be transmitted, see eq.(18). The signal d(t) (and therefore all $d_k(t)$ too) is modified by the channel into the received signal $d^{(r)}(t)$ (signal received by k-th channel is $d_k^{(r)}(t)$). The bit sequence $b_k(t)$ and transmitted signal $d_k(t)$ for one channel are shown in fig.(4); the received signal $d_k^{(r)}(t)$ for the same channel (at point D) is shown in fig.(5). The total transmitted signal d(t) is shown at fig.(6), the total received signal for points F (LoS) and D (NLoS) is shown in fig.(7) and fig.(8) respectively.

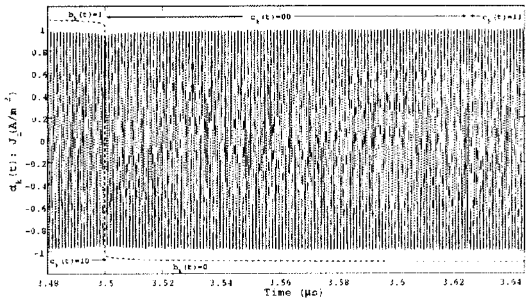

Fig.4: Transmitted signal $d_k(t)$, one channel (enlarged)

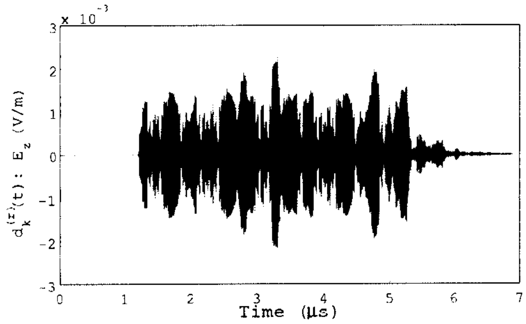

Fig.5: Received signal $d_k^{(r)}(t)$, same channel at point D

The multipath scattering distorts the signal more for point D than for point F. The original d(t) has two "gaps" or time intervals without signal, around 0.7 μs and 3.1 μs. In the case of point F, these gaps are visible, but in the case of point D, only the first gap is noticeable. The original d(t) has 2 main bursts of power between 1.5 μs and 2.8 μs. For point F, the relative size and shape of these bursts do not change much, but for point D, the second burst changes a lot.

5. Conclusion

Parallel FDTD method can be very useful in the modeling of big urban channels and in the design of better systems for high-speed data communications. The merit of FDTD is better accuracy than other methods, and for high data rates, accuracy becomes very important. This research will allow for optimal system design for applications such as multimedia wireless communications. By changing the method of modulation and the bit rate, the best modulation and the fastest Bit Rate with good BER can be found.

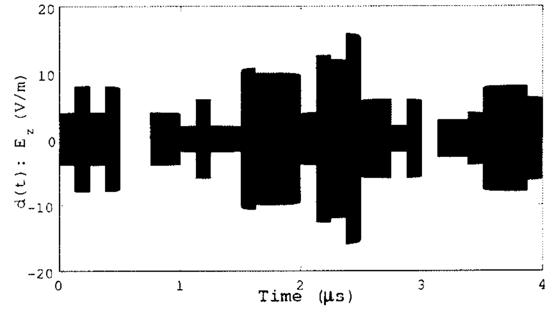

Fig.6: Total transmitted signal d(t)

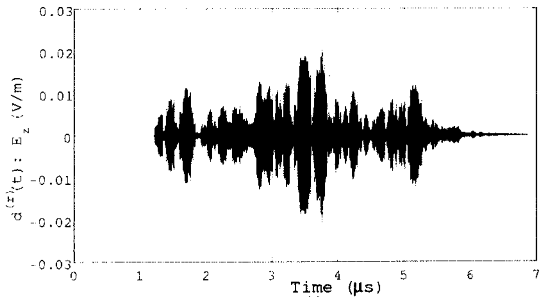

Fig.7: Total signal received $d^{(r)}(t)$ at point D (NLoS)

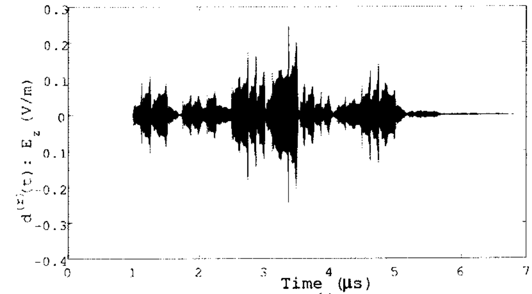

Fig.8: Total signal received $d^{(r)}(t)$ at point F (LoS)

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