Signaling Overhead in Time-Invariant OFDM Systems with Adaptive Techniques

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Abstract

To realize "global information village (GIV)", the followed conditions must be fulfilled, high data rate wireless communication; high robustness; low cost. Orthogonal Frequency Division Multiplexing (OFDM) technique seems to be one of the best candidates to fulfill the above requirements. As a matter of fact, OFDM is the one candidate of 4G mobile communications. In this work one aspect of OFDM technique will be treated namely adaptive modulation techniques, where information is adaptively modulated on each subcarrier based on the current channel state. Emphasis will be put on the signaling overhead, which is necessary to inform the receiver which modulation method is applied on which subcarriers.

Keywords

GIV,robustness, OFDM, adaptive modulation techniques, signaling overhead.

1. Introduction

As one candidate for mobile communication systems of 4th generation OFDM comes more and more into the focus of research. Besides of its high spectral efficiency, which results from orthogonality and therefore overlapping of individual subcarriers, OFDM techniques can easily solve the problem of frequency-selectivity imposed on multipath radio channels without using sophisticated equalization algorithms. Because of these features, OFDM has been adopted as a standard for digital video broadcasting (DVB), digital audio broadcasting (DAB), wireless local area network (WLAN), etc.

However as well known from micro-wave theory, during signal propogation electromagnetic wave is scattered, diffracted and reflected on its way to the receiver via various signal paths. At the receiver the signal power level is therefore dependent on the frequency due to incoherent superposition of different signals. This leads to an interference pattern over the whole transmission bandwidth. If the receiver is mobile, this interference pattern is additionally depedent on time and location. Therefore the channel quality of each subcarrier can fluctuate very strongly. For subcarriers affected by deep fading a high-qualityesti data transmission is almost impossible, while for subcarriers with large amplitude high data rate can be applied. So for a multipath fading channel the bit error ratio (BER) would be dominated mostly by subcarriers with deep fading, while the channel capacity of good subcarriers are not exploited sufficiently. In order to eliminate this effect, adaptive modulation techniques are applied, where according to the current channel quality, data are adaptively modulated on each individual or grouped subcarriers [1]. In this way the subcarriers affected by deep fading are compensated by the subcarriers with high transmission quality and in total a significant gain can be achieved as will be shown in later sections.

Data can be adaptively modulated on each subcarriers to exploit the channel capacity. However, this will lead to another problem. The receiver has to know the transmission mode so that he can do demodulations. So an additional signaling overhead is necessary, which reduces bandwidth efficiency. We have to organize this signalization in an efficient way so that the overhead can be reduced to an acceptable amount without a significant system degradation. Furthermore through computer-based simulations we want to find out the theoretical lower and upper bounds of the signaling overhead under different channel conditions.

This work is divided into five sections. In section two a short review of OFDM techniques is given followed by mathematical descriptions of OFDM signal generation process and the system is described, where emphasis is put on the applied channel models. In the third section an adaptive technique is developed for time-invariant channels and the simulation results are presented under different channel condistions. In section four signaling overhead using Huffman coding scheme is introduced in time-invariant channel. In the last section this work is briefly conclution.

2. OFDM System with Adaptive Techniques

Now OFDM is considered as one candidate for 4G mobile communications. OFDM comes more and more into focus of researches. Efforts made on classic OFDM points like choosing suitable length of guard interval will be intensified and new research points will be increasingly exploited like adaptive techniques.

Frist DFT and IDFT play a central role in generation of OFDM signals. DFT is the discrete version of the Fourier transformation of discrete time signals and IDFT is the inverse transformation of DFT signals. Fig. 1 is a typical block diagram of an OFDM sysytem with pilot signal assisted. It begins with a binary source (information source), which is channel-coded using convolutional codes. These transmitted symbols s(k), including pilot symbols, go through the IDFT block. DFT or IDFT can be considered as the exchange from time domain to frequency domain or from frequency domain to time domain. So before going through IDFT block we can regard the transmitted symbols are in frequency domain, after going through IDFT block the transmitted symbols are transformed into time domain. The mathematical expression of the symbols in time domain is:

$$x_i(k) = IDFT\{s_n(k)\}$$

$$= \sum_{n=1}^{N} s_n(k) \cdot \mathrm{e}^{\mathrm{j}\frac{2\pi}{N}ni} \tag{1}$$

Where N is the number of subcarriers. Following IDFT block, the guard interval is inserted into these symbols. In order to efficiently overcome ISI (intersymbol interference), the length of the guard interval has to be larger than the maximum time spread in the channel, so that the data block will not be interfered by the previous OFDM symbols. Additionally, in order to eliminate inter-carrier interference (ICI), cyclic extention is adopted. Through the cyclic extention, the exact integer number inside the FFT integration time is recovered, so that the contribution from other subcarriers is canceled again as desired [3]. Therefore, the OFDM symbol is the following expression:

$$x_f(k) = \begin{cases} x(N+k), & k = -N_g, -N_g+1, ..., -1 \\ x(k), & k = 0, 1, ..., N-1 \end{cases} \tag{2}$$

where N_g is the length of the guard interval. The transmitted signals are sent into a multipath fading channel. The received signals can be represented by:

$$y(t) = h(t) * x(t) + v(t) \tag{3}$$

where $v(t)$ is the Additive White Gaussian Noise (AWGN) and $h(t)$ is the channel impulse response. So after going through DFT block, as the above explained,

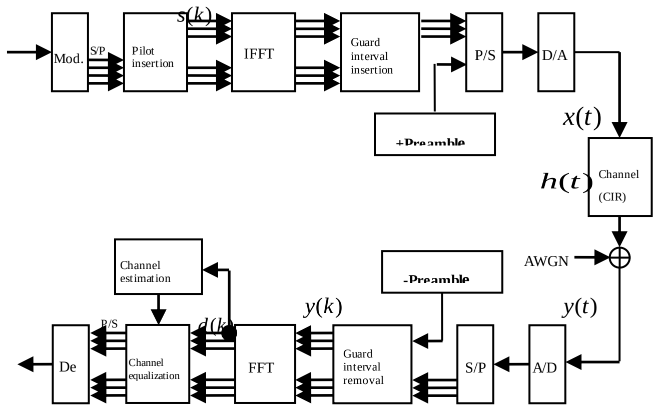

Fig.1 OFDM system block diagram

the received signal is converted from time domain to frequency domain again. Assume that the received symbols are called d(k).

Its mathematical expression is:

$$d_n(k) = \frac{1}{N}\sum_{i=1}^{N} y_i(k)\cdot \mathrm{e}^{-\mathrm{j}\frac{2\pi}{N}ni} \tag{4}$$

The outputof the DFT operation is input into the equilazator, where according to the estimated channel transfer function the transformed data symbols are equalized. The output of the equalizator is the detected transmit symbols, which are made up of the signaling field and data blocks. The signaling field is separated from the data blocks and then decoded and demodulated to obtain the signaling information. According to the estimated result the data blocks are demodulated.

In this work we consider a time-invariant channel model with exponentially decreasing power delay profile (PDP), which describes the reality very well [4]. So we have:

$$A_c(\tau) = e^{-\frac{\tau}{3T}} \tag{5}$$

where the factor 3 is the correction factor and T is the delay spread that takes the value of 50ns for typical indoor applications. As mentioned before the investigated system has a bandwidth $\Delta f = 20\text{MHz}$ corresponding to the inverse of T , further we choose a guard time of 800ns corresponding to 16 samples of the PDP. So we restrict the discrete channel transfer function to 16 samples only so that no ISI occurs:

$$A_c[i] = e^{-\frac{i}{3}}, \quad i = 0,1,2,3,...,15 \tag{6}$$

For comparison purposes we normalize the PDP to unit power:

$$\bar{A}_c[i] = \frac{1}{\sum_{i=0}^{15} e^{-\frac{i}{3}}} e^{-\frac{i}{3}}, \quad i = 0,1,2,3,...,15 \tag{7}$$

Therefore in this channel model we consider 16 MPCs, each of which is the sum of a number of micropath components. These micropath components are unresovable with respect to the inverse of the bandwidth Δf . As derived before each MPC can be described by a complex random variable. The in-phase and quadrature-phase of this complex random variable is Gaussian distributed (central limit theorem) and the absolute amplitude shows a Rayleigh distribution. So the complex amplitude of these MPCs can be described by the following equation:

$$h[i] = \sqrt{\frac{\bar{A}_c[i]}{2}}(\hat{n}_{\mathrm{re}} + j\hat{n}_{\mathrm{im}}),\ i = 0,1,2,3,...,15 \tag{8}$$

where $\hat{n}_{re}$ and $\hat{n}_{im}$ are gaussian distributed with zero mean and unit variance. Using 64 points FFT from eq.1, which corresponds to 64 subcarriers in the considered transmit bandwidth, $h[i]$ is Fourier transformed into $H[k]$:

$$H[\frac{2\pi}{64}k] = \sum_{n=0}^{15} x[n]e^{-j2\pi kn/N}, \ k = 0,1,2,...,63 \tag{9}$$

These 64 subcarriers are orthogonal to each other as derived in section 2. This transfer function is normalized to the average receive power. As can be seen due to MPCs the absolute amplitude is strongly frequency-selective. The amplitude difference between different subcarriers can ranges from serveral decibels to over 30 dB.

3. Adaptive Modulation in Time-Invariant System

Now we assume that the transmission of signaling bits is not error-free and apply very robust BPSK to modulate the signalization bits, while everything else including channel model (time-invariant channel model) remains (see the block diagram in sec.2). Note that in WLAN standards the physical layer convergence procedure (PLCP) header, which contains information about the data rate and length of the PLCP data service unit (PSDU) and therefore serves as a kind of signaling too, is also convolutionally coded with code rate 1/2 and then BPSK-modulated [5]. So we are faced with the following situation, for each subcarrier one of the five possible modulation orders (no data modulaton, BPSK, 4-QAM, 16-QAM, 64-QAM) is selected depending on the current channel state. If we ignore the occurrence probability of each modulation order for this moment, three bits pro subcarrier are required to encode these 5 possible modulation orders (ceil(ld(5)) = 3), where the ceil function rounds its argument towards plus infinity:

No data modulation $-\!\!\longrightarrow$ 000

BPSK $-\!\!\longrightarrow$ 001

4-QAM $-\!\!\longrightarrow$ 010

16-QAM $-\!\!\longrightarrow$ 100

64-QAM $-\!\!\longrightarrow$ 110

These three bits are convolutionally encoded with code rate of 1/2, so that totally 6 bits pro subcarrier are transmitted. These 6 bits are modulated using BPSK as mentioned before which leads to 6 symbols for each subcarrier. Totally we have 64 subcarriers (acually no signalization needed for the four pilot subcarriers and twevle free subcarriers, but they are set to 000 to fill the fest data structure). So the signaling field is made up of a symbol matrix with 64×6 symbols. This signaling field is the signaling overhead at this moment and is inserted between preamble and information blocks as shown in Fig. 1.

Now we are interested in how much the system performance is degradated due to imperfect signaling. The symbol error ratio is compared between two cases: perfect signaling and signaling with BPSK in the Fig. 2.

This comparison is performed for two data rates 24 Mbps and 12 Mbps respectively which correspond to 16 QAM and 4 QAM on each subcarrier in average. We expect that due to the additive noise there may be a small gap in each curve pair in the low SNR region althogh BPSK is applied. But in high SNR regions a coincidence of curves has to occur. This is the case for blue curves which result from the data rate of 24 Mbps (16-QAM) in average. However, in the case of 4-QAM, the curves drift and a kind of saturation behaviour is observed for the SER with BPSK signaling. This is caused by missing certain amount of termination bits in the signaling field. We remember that a viterbi decoder is used to decode the received bits and this decoder requires a predefined initial state and final state. These termination bits, normally filled with zeros, are aimed to bring the decoder to the predefined final state. A state, which is not predefined, will lead to an sharply increased decoding error probability[6]. These errors cause an incorrect signaling and therefore a higher SER. However this can be observed only when the SER already reaches a relative low value (10^{-4}), so that this effect can not be seen clearly in the blue curves where more robust modulation (4-QAM) was applied.

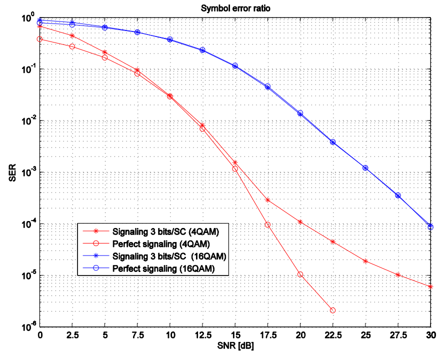

Fig. 2 Comparison of SER between perfect signaling and BPSK signaling

The effect of termination can also be confirmed by Fig. 3. The curve marked with star is the signaling BER which is convolutionally coded and BPSK modulated as mentioned before. Obviously the BER profile does not match the fact that the signaling bits must be well protected, since the curve shows a very low falling rate with increasing SNR and the slope keeps constant over almost 20dB from 10 dB to 30 dB. At high SNR values a remainding error probability (at 30 dB still 10^{-5}) is observed. This situation can not be improved through interleaving operation or even using perfect channel knowledge, which implies the remaining errors are not caused by any deeply fading subcarriers or additive noise. So obviously the remaining error lies in some certain fixed number of bits only. In order to confirm this expectation 9 arbitarily chosen addional data blocks are concatenated prior to the 6 signaling blocks, so the total number of bits considered is over 2 times more than the initial signaling bit number. The BER of all 15 blocks is evaluated. The result shows the weight of this remaining error is weakened by transmitting more number of bits. This leads to the conclusion that certain number of bits at the end of the signaling fields are disturbed because the termination bits are missing. If the signaling field is terminated, a signaling BER profile represented by the curve marked with black points is obtained which shows a much better performance compared to the no termination cases. This would also give a better SER performance. Now a way has to be found to terminate the signaling field. Since if using 3 bits for each subcarrier the whole data structure is completely filled and no empty places ar e available for termination bits.

4. 4. Signaling Overhead Using Huffman Coding Scheme in Time-Invariant Channels

In the third section signaling techniques with an overhead of 6 OFDM blocks were investigated in the case of time-invariant channels. It was shown that this signaling overhead is enormous compared to the data transmitted($L_{\mathrm{DBL}} = 10$). In order to improve

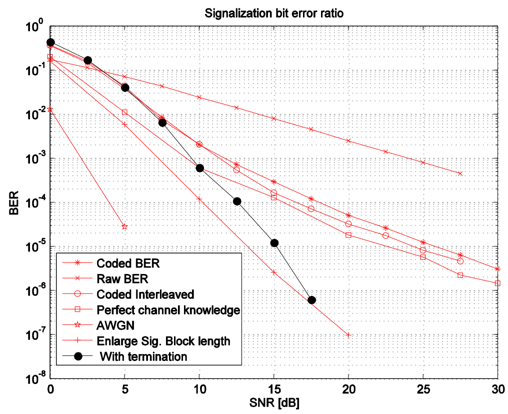

Fig.3 Signaling BER for different signaling methods

the bandwidth efficiency, we can increase L_{DBL}. However large L_{DBL} requires additional signaling overhead if the system performance has to be maintained at some acceptable level. So L_{DBL} can not be scaled up to an arbitrary value. As mentioned before in average 4 bits are transmitted on each subcarrier, so 16-QAM will be the modulation scheme which is most frequently applied over all subcarriers. Now we take the statistical properties of the source into account and assume for this moment that the transmitter has the perfect knwoledge about CSI. The coding procedure is performed using Huffman coding scheme which minimizes the average codeword length.

First the occurrence probabilities of each modulation order have to be known. For this purpose the time-invariant channel model is applied, since no correlation between successive BATs is desired. So the parameters set in previous section are reserved and the simulation is run for 500000 iterations which corresponds to 500000 independent channel realizations. For each realization the modulation order on each of all 48 data subcarriers is recorded. The total occurrence of the modulation orders is then counted over all subcarriers and iterations. Therefore the occurrence probabilities are given by the ratio of occurrence and the product of the iteration number and the number of subcarriers (500000 × 48 = 24000000). For example 16-QAM occurs totally 11469600times in the 500000 channel realizations, so the occurrence probability is then equal to 11469600/24000000 = 0.4779. Other four occurrence probabilities from "no data modulation" to "64-QAM" are obtained similarly and saved together in the following vector:

$$\bar{w}_1 = (0.0274, 0.0261, 0.1860, 0.4806, 0.2799) \tag{10}$$

We assign symbols $\mathrm{S_{M0}}$, $\mathrm{S_{M1}}$, $\mathrm{S_{M2}}$, $\mathrm{S_{M4}}$ and $\mathrm{S_{M6}}$ to the above modulation orders from "No modulation" to "64-QAM".

According to the Huffmann coding scheme, the average codeword length can be determined:

$$\bar{L}(\mathrm{S_M}) = \sum_i P(\mathrm{S_{Mi}}) L(\mathrm{S_{Mi}}) \quad i = 1, 2, 4, 6$$

$$\begin{aligned} \bar{L}(\mathrm{S_M}) &= 0.0274 \times 4 + 0.0261 \times 4 + 0.1860 \times 3 \\ &+ 0.4806 \times 1 + 0.2799 \times 2 = 1.8178(\mathrm{bits}) \end{aligned} \tag{11}$$

Now the 48 data subcarriers are only considered in flat channels. So the signaling overhead is reduced to $\bar{L}(\mathrm{S_M}) \times 48 = 87.25$ bits on average. However for some extreme channel a much higher signaling overhead than the average value may occur. Therefore we have to investigate how often this kind of extreme channel realizations can occur. For this purpose we use the codewords above and determine the total number of signaling bits required for each channel realization. If the channel quality distinguishes strongly from each other, more signaling overhead will be required as mentioned above. Therefore we have to determine the probability density function (PDF) of all possible signaling overheads which may occur. The PDF is depicted as a function of required signling bits required for one channel realization in Fig.4. The corresponding probability distribution function is depicted beside the PDF also in Fig. 4.

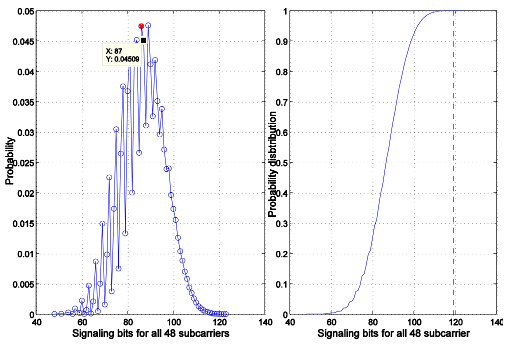

Fig.4 PDF and CDF of the signaling overhead for uncorrelated channels

The PDF begins with the signaling overhead of 48 bits. The associated probability is extremely low. Due to the fixed data rate and therefore the fixed number of transmitted bits pro OFDM symbol the combination of the modulation orders on the 48 data subcarriers is not arbitrary any more. This leads to the fact that not all integers greater than 48 is a possible signaling overhead as shown in Fig. 4. Actually we can describe possible combinations using the finite-state machine model. This machine has 48 containers, each container can have five possible values (0, 1, 2, 4, 6) corresponding to five different modulation orders. But the sum of the values in each container must give the total number of transmitted bits. Since in a finite state machine we can reach all possible state from any initial state, we start with the state where all containers have the value of 4. As mentioned before the sum remains always the same, so if one container take a lower value, at least one container should take a higher value. In these values only 6 is larger than 4. So we assume that any one of the 48 containers modifies its value to 6, the sum would take the value of 194 if any other containers were not modified. To obtain the same sum at least one of the rest containers has to modify its value so that this additional 2 is canceld. We note the only possibility is one container change its value from 4 to 2, since a change from 4 to 1 or 0 would give an overbalance. So we reach from the initial state to another state at which one container has 6 and one container has 2 besides the rest 46 containers. Furthermore we note that if we put the values into a 1×48 vector V_{con}

$$V_{\text{con}} = \overbrace{(4,4,\ldots,4)}^{48} \tag{12}$$

the order of each components is of no importance for our consideration as we are interested in the total signaling overhead only. But we have to bear in the mind that for the calculation of probabilities the order does count. Therefore we can take the container containing 6 to the very beginning and the container with 2 to the second place of this vector. So we have

$$V_{\text{con}} = \overbrace{(6,2,\ldots,4)}^{48} \tag{13}$$

Note the "6" can come from any "4" of all 48 "4"s and the "2" from any "4" of the rest 47 "4"s or vice versa. We call the above mentioned modification "basic combination". This constallation leads to 3 more signaling bits compared to all "4" constallation, since any "4" requires only 1 bit but "6" requires 2 bits and "2" requires 3 bits (1.5 bits pro subcarrier). Obviously any two "4"s in the vector can be replaced by this basic combination until we have the constallation:

$$V_{\text{con}} = \overbrace{(6,2,\ldots,6,2)}^{48} \tag{14}$$

This gives an additional overhead of 5 bits (1.67 bits pro subcarrier). The third basic combination is to use two "4→1" to balance three "4→6" modifications. Any five "4"s can be replaced by this basic combination. In this case the vector becomes:

$$V_{\text{con}} = \overbrace{(6,6,6,1,1,\ldots,4,4)}^{48} \tag{15}$$

It leads to an additional overhead of 9 bits (1.8 bits pro subcarrier). We conclude that using Huffman coding scheme the theoretical lower bound of the signaling overhead is 48 bits which corresponds to the all "4"s case, the theoretical upper bound is 134 bits corresponding to nine times the third basic combination and one times the second combination.

5. Conclusion

In this work a signaling method was developed for the time-invariant multipath channel. Almost no performance degradation was observed by using terminated signaling field with 5 signaling bits pro 2 subcarriers. In the practice, however, we have to deal with time-variant multipath channels. Here great efforts were made to find out the theoretical lower and upper bounds of the signaling overhead, which was necessary to make the receiver able to demodulate the received data. In the case of slowly time varying channels, the successive bit allocation tables were strongly correlated. It led to an enormous potential to save the signaling overhead by transmitting the modification of successive BATs instead of transmitting the whole BATs everytime. It was confirmed through simulations that in the case of perfect channel knowledge the signaling overhead could be reduced to the half, if only the modifications were transmitted. But in the reality the perfect CSI can not be provided. The transmitter has to estimate the channel based on the preamble symbols from the last reception. In this case the signaling overhead is expanded. If the estimation errors result from the additive noise only, the signaling overhead can still be maintained at a low level for high SNR. It increases the more, the lower the SNR is chosen, until the correlation of the channel is not available any more. Then we obtain a uncorrelated channel, which requires a comparable signaling overhead as that for the time-invariant channel.

Reference

[1] A. J. Paulraj, D. A. Gore and R. U. Nabar H. Bolsckei: An Overview of MIMO Communications - a Key to Gigabit Wireless In Proceedings of the IEEE, Vol. 92, pp. 198 - 218, Feb. 2004.

[2] A. Czylwik: Adaptive OFDM for Wideband Radio Channels In Proceedings of the IEEE Global Telecommunications Conference (GLOBECOM), London, pp. 713- 718, 1996.

[3] R.W. Chang:Synthesis of band-limited orthogonal signals for multi-channel data transmission Bell System Technical Journal 46, pp. 1775-1796, 1966.

[4] S. B. Weinstein and P. M. Ebert: Data Transmission by Frequency Division Multiplexing Using the Discrete Fourier Transform IEEE Transactions on Communications, Vol. 19, pp. 628-634 Oktober 1971.

[5] B. Hirosaki,: An Orthogonally Multiplexed QAM System Using the Discrete Fourier Transform IEEE Transactions on Communications, Vol. 29, Juli 1981.

[6] John. G. Proakis and Dimitris. G. Manolakis: Digital Signal Processing Pearson Education, Upper Saddle River, New Jersey, 2007.

[7] H. Schulze and C. Lüders: Theory and Applications of OFDM and CDMA John Wiley & Sons Ltd, The Atrium, Southern Gate, Chichester, West Sussex PO19 8SQ, England, 2005.

[8] R. Van Nee and R. Prasad: OFDM for Wireless Multimedia Communications Artech House Bosten and London, 2000.

[9] D. J. Clowes: Variable Data Rate Transmission for a Rayleigh Fading Channel Commun. Lab., Defence Res. Telecommun. Establishment, Ottawa, Ont., Canada, Tech. Memo. 22, Feb. 1969.

[10] A. Goldsmith and P. Varaiya: Capacity of Fading Channels with Channel Side Information IEEE Transaction on Information Theory, vol. 43, pp. 1986-1992, November 1997.

[11] A. J. Goldsmith and S. G. Chua : Adaptive Coded Modulation for Fading Channels in Proceedings of IEEE Transactions on Communications, vol. 48, pp. 367-369, March 2000.

[12] T. Suzuki, S. Sampei, and N. Morinaga: Space Path Diversity Combining Technique for 10 Mbits/s Adaptive Modulation/TDMA in Wireless Communication Systems in Procedings of IEEE VTC'96, pp. 1003-1007, IEEE, May 1996.