CHIP LEVEL SIMULATION OF THE DOWNLINK IN UTRA-FDD

J.J. Olmos and S. Ruiz
Signal Theory and Communications Department (UPC)
Campus Nord-D4, Jordi Girona Salgado 1-3, 08034 Barcelona, Spain
E-mail: [olmos, silvia] @tsc.upc.es, fax +34 93 4017200

ABSTRACT

The specifications of UMTS Terrestrial Radio Access (UTRA) for the physical layer of the downlink make use of Orthogonal Variable Spreading Factor (OVSF) codes to preserve the orthogonality between downlink channels of different rates and spreading factors. This technique minimises the downlink intra-cell interference. In order to control the inter-cell interference, every base station multiplies the global downlink signal with a cell specific Gold code (scrambling code). Then, while the inter-cell interference may be modelled using the Gaussian hypothesis (that is: replacing the real interference with a Gaussian noise of the same power), the intra-cell interference requires detailed chip level simulations. In this paper we present results of a chip level simulation of the downlink UTRA physical layer. The objective is to evaluate the raw (uncoded) mean bit error rate (BER) of the system in a realistic environment and conditions. Then, by knowing the BER requirements of the different services, one can easily obtain the maximum capacity in terms of simultaneous connections at any combination of bit rates.

I. INTRODUCTION

In the UTRA-FDD mode, the base station allocates a different orthogonal code (channelisation code) to every connection of every user. Every connection uses a Dedicated Physical Data Channel (DPDCH). Since the modulation is QPSK at the bit level, before spreading every DPDCH can be represented by a flow of complex modulation symbols from the set $\{\pm 1 \pm j\}$. To achieve the chip rate, the real and imaginary parts of the modulation symbols are multiplied by the same connection specific channelisation code. After adding together all downlink contributions, the base station multiplies the global signal by a cell specific Gold code.

In the mobile receiver, and due to the multipath propagation, the original orthogonality is partially lost. This effect is more remarkable in the outdoor environment, where the delay spread of the mobile channel is equivalent to several chip intervals. So the intra-cell interference is no longer negligible, but as we shall see, its effects are still limited to values lower than what is predicted by the Gaussian hypothesis.

This work was supported by CICYT project TIC 98-684

0-7803-6465-5/00 $10.00 © 2000 IEEE

A good estimation of the effects of multipath in the UTRA downlink intra-cell interference requires detailed chip level simulations in a realistic environment. This means generating the exact UTRA downlink signal with the OVSF and Gold codes, filtering the chips with a root raised cosine filter, passing this signal through a environment specific (outdoor or indoor) mobile channel simulator, adding Gaussian noise, filtering again and finally the simulation of a coherent Rake receiver. In this paper we present the results of such a simulation. We assume an isolated UMTS single cell with quasi-ideal channel estimation (means ideal estimation on a slot by slot basis), perfect frequency and time synchronism and perfect fast power control (means no quantisation and no dynamic margin restrictions). We also assume that all services require the same mean BER, thus all connections, regardless of their spreading factor, have the same bit energy. All parts of the simulation program have been written in C++ language.

II. SOURCES OF NOISE

Usually, the loss of orthogonality is quantified using a parameter called the "orthogonality factor", [1], which measures the fraction of intra-cell interferent power that is perceived by the receiver as noise. Another degradation is due to the multipath-induced ISI, which obviously depends on the duration of one modulation symbol in comparison to the delay spread of the channel, on the autocorrelation of the spreading sequences and on the velocity of the mobile, since the higher this velocity the less accurate will be the channel estimation used in the Rake receiver, which can only be updated at the slot boundaries. Since we focus on the effects of multipath in a UTRA-FDD link, only one time-variant mobile channel is simulated, this means only one mobile station with multiple simultaneous connections.

It is illustrative to write all the terms that degrade the BER in a UTRA-FDD connection. The mean BER is a function of the global E_b/η, which can written as:

$$\frac{E_b}{\eta} = \frac{E_b}{N_0 + \chi + \rho \cdot I_0 + \eta_{ISI}} = \left(\left(\frac{E_b}{N_0}\right)^{-1} + \left(\frac{E_b}{\chi}\right)^{-1} + \rho \left(\frac{E_b}{I_0}\right)^{-1} + \left(\frac{E_b}{\eta_{ISI}}\right)^{-1} \right)^{-1} \quad (1)$$

Where E_b is the mean bit energy of the useful signal, N_0 is the thermal noise power spectral density, χ is the inter-cell interferent power density, ρ is the orthogonality factor, I_0 is the intra-cell interferent power density and η_{ISI} is the power density of the ISI present in the received signal.

The ratios E_b/N_0, E_b/χ and E_b/I_0 can be obtained through system level simulations or analytical calculations based on simplifying hypothesis. They depend on the number of active connections per cell, on their SF, on the spatial distribution of the mobile stations, on the cell site configuration (number of sectors), behaviour of power control, type of environment, voice activity factor, etc...In turn, the estimation of ρ and E_b/η_{ISI}, must be done by chip-level simulations of the UTRA radio link.

The estimation of the orthogonality factor is interesting because it allows us to model the intra-cell interference using the Gaussian hypothesis even if the OVSF codes are used in the link. We simply have to add at the receiver an equivalent Gaussian noise with power equal to ρ times the received intra-cell interferent power. Since the inter-cell interference can also be modelled with the Gaussian hypothesis, it is sufficient to characterize the link for one single connection of each of all the possible bit rates (or SF).

The estimation of E_b/η_{ISI} is interesting if we can find and adequate theoretical expression for the function that relates the BER to the global E_b/η. If so, and if we know the values of E_b/N_0, E_b/χ and E_b/I_0, we could calculate E_b/η using expression (1) and then the BER. We could also solve the inverse problem, that is: which are the minimum signal to inter-cell (intra-cell) power ratios, SIR_X (SIR_I) that we must design for a given BER?, where the SIR_X and SIR_I in the downlink are given by:

$$SIR_X = \frac{2}{SF}\cdot\left(\frac{E_b}{\chi}\right) \quad ; \quad SIR_I = \frac{2}{SF}\cdot\left(\frac{E_b}{I_0}\right) \qquad (2)$$

Assuming that the inter-cell interference is a fraction of the intra-cell interference we could solve equation (1) and obtain SIR_X and SIR_I.

Unfortunately and in general, the function that relates the BER to E_b/η must be found by computer simulation. This is due to the intervention of the fast power control and to the use of a Rake receiver with (possibly) less arms than independent propagation paths. The improvement due to these features is difficult to evaluate without simulations.

III. SIMULATION DESCRIPTION

The program accepts as input a list where, for every spreading factor (SF), we specify how many independent downlink connections using that SF will be simulated. The possible SF are: 4, 8, 16, 32, 64, 128, 256 and 512. Since the chip rate is 3840 kchip/s, the possible bit rates are: 15, 30, 60, 120, 240, 480, 960 and 1920 kbit/s. The program allocates first the OVSF codes for the connections with higher bit rate, and then proceeds with the lower bit rates by applying the restrictions of the OVSF code tree [2]. In summary, these restrictions mean that the higher the bit rate the more resources are consumed, to the point that a base station can accept 512 simultaneous independent connections at 15 kbit/s but only 4 simultaneous independent connections at 1920 kbit/s. We must keep in mind that high bit rate users are also power demanding users, so the OVSF code tree restrictions do not place any additional limit in a power limited base station. The program guarantees that the simulated combination of rates and services is a valid one in the UTRA context. The global downlink complex signal, with only one sample per chip, is multiplied by the Gold code of period 38400 chips. For simplicity, the transmitted signal is generated in blocks of 512 chips. This means from two bits of a connection at 15 kbit/s to 256 consecutive bits of a connection at 1920 kbit/s.

The mobile channel simulator can simulate any number of independent complex propagation paths. Each path has Rayleigh or Rice statistics (selectable) and "classic"[3] Doppler spectrum. The power, delay and Doppler spread of every propagation path can be adjusted independently, so the simulator can produce any power delay profile. The delays of the propagation paths are kept constant during all the simulation. To simulate a time-variant channel, the impulse response is updated 100 times for every coherence time interval of the channel. All the paths have the same Doppler spread, which depends on the carrier frequency (2 GHz) and on the vehicle speed that we are simulating. We don't take into account shadowing fading because the fast power control in the downlink will compensate it. The fast fading is implicit in the mobile channel impulse response.

The Rake receiver has 4 arms whose delays are dynamically adapted to track the points of maximum energy in the instantaneous global impulse response. The precision in the setting of the Rake delays is 1/8 of a chip interval, but only one sample per chip is processed in the whole system simulator. The Rake arms lock to delays that are at least one chip interval apart from each other and the signal energy at any arm is within 10 dB from the maximum signal energy. If this can not be accomplished less than 4 arms are used.

The cascade of the raised cosine transmission-reception filter, mobile channel, and the Rake combiner (means the entire Rake except the correlator with the local codes, see figure 1) is a linear system that can be simulated with a single FIR filter with time-variant coefficients. The only problem is that the Gaussian noise, to take into account thermal noise plus inter-cell interference, can not be added at the input of this filter. Since the Rake delays are separated at least one chip interval, the noise contributions of the different arms are uncorrelated and its variances can be added. So we can calculate the noise variance at the output of the Rake combiner and add the complex Gaussian noise at that point. The noise variance depends on the Rake coefficients, which in turn depend on the channel impulse response, and so it must be re-calculated every time the mobile channel is updated. Since the received signal power depends on the power delay profile, the noise variance also must be accordingly modified to obtain the desired E_b/N_0.

After the global FIR filtering, we multiply the complex samples by the local codes (OVSF plus Gold), integrate over 512 chips and detect the bits one by one using minimum distance criterion. The program simultaneously calculates the output bits of all the active connections in the mobile station. So the multiplication by the local OVSF codes and the 512 chips integral is repeated as many times as different connections are in use. Every integral is divided into multiple partial integrals depending on the SF of the connection, since the number of bits to detect per connection is equal to $2\cdot 512/SF$.

In order to guarantee that the channel has gone through all its possible states, every simulation processes a number of frames equivalent to at least 1000 coherence time intervals of the mobile channel. Furthermore, to have a good estimation of the BER, every simulation runs until at least 100 errors have occurred. Only when both of these two conditions are met the program stops or changes to the next simulation point (next E_b/N_0).

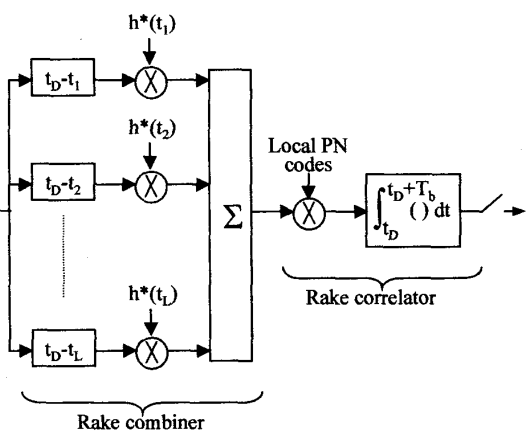

Figure 1.- Block diagram of the Rake receiver

IV. PROGRAM VALIDATION AND RESULTS

In general, even if we can calculate the global E_b/η present in the link with expression (1), there is not an analytical expression to calculate the BER as a function of E_b/η. On the other hand, and in order to validate the simulation program, it is important to check its results versus some theoretical prediction. There are some assumptions under which we can find a theoretical formula for the BER in an UTRA radio link.

First, if we assume perfect channel estimation, the Rake receiver behaves as an order L MRC diversity combiner (with L being the number of arms in the Rake). Due to the limited bandwidth of the receiver, the propagation paths with delay differences lower than a chip interval merge into a single contribution at the output of the receiving raised cosine filter and are not individually resolvable. So, only the propagation paths that are at least one chip interval apart from each other are resolvable. If the number of resolvable propagation paths in the power delay profile (PDP) that models the propagation environment is higher than L, then the Rake performs first a "selection" diversity (by selecting the paths with higher energy) and then an MRC combination, so in this case it is not a pure MRC processing. But if the number of resolvable propagation paths is lower or equal than L, then there is a closed expression for the bit error probability, [4]:

$$P_b = \tfrac{1}{2}\sum_{k=1}^{L}\pi_k\left[1-\sqrt{\frac{\gamma_k}{1+\gamma_k}}\right] \tag{3}$$

In (3), P_b is the mean BER, L is the number of diversity paths, γ_k is the mean E_b/η ratio in the kth diversity path (exponentially distributed) and:

$$\pi_k = \prod_{\substack{i=1\\ i\neq k}}^{L}\frac{\gamma_k}{\gamma_k-\gamma_i} \tag{4}$$

This expression assumes that there is no ISI and that each one of the propagation paths has a Rayleigh envelope. Due to the intervention of the power control loop, this assumption could not hold in slowly time variant channels, where the fast power control would change the statistics of the received signal. In turn, in an outdoor environment with a vehicle moving at 120 km/h or more, we can assume that only the mean signal power is affected by the power control. That is, the slow fading will be removed and the fast fading will remain. To obtain γ_k from the global E_b/η, we simply multiply it by the fraction of power carried by the kth path.

The first power delay profile that we will simulate corresponds to an outdoor environment with 4 propagation paths separated one chip interval from each other and relative powers of 0,-3,-6 and –9 dB (respectively) and a mobile speed of 120km/h. This environment is proposed as case number 3 (Annex B) in the UTRA specifications, [5], for performance measurements.

We start by evaluating the orthogonality factor. To this purpose we run simulations without power control, with a single active connection ($I_0=\chi=0$) and ideal channel estimation (update the Rake coefficients every time the channel is updated). In this way we have the minimum ISI. We compare the BER of that simulations with the BER of another kind of simulations where we have multiple active connections ($I_0>0$) but no other noise ($N_0+\chi=0$) except the minimum ISI. When the two BER are equal, we can interpret that the decision device is experiencing the same noise variance, an so:

$$N_0 + \eta_{ISI} = \rho \cdot I_0 + \eta_{ISI} \quad \Rightarrow \quad \rho = N_0 / I_0 \qquad (5)$$

In figure 2 we show results for the BER in the above described environment. The solid lines are the BER curves for one single connection, while the dashed curves are BER vs. E_b/I_0 for multiple active connections (each with the same SF). The dashed line without symbols is the theoretically predicted BER using expression (3) for a 4 arm MRC diversity combiner with unequal powers on each arm. It can be seen how the BER curves for one single connection and high SF tend to coincide with the theoretical formula. This is because at low bit rate (high SF) the bit duration is much higher than the delay spread of the channel and there is almost no ISI. There is also a plot, obtained with a bit level simulation with no ISI, that gives almost exactly the same result as the theoretical formula. From the observation of the plots in figure 2, we obtain an orthogonality factor of about 6 dB (ρ=0.25) for this environment.

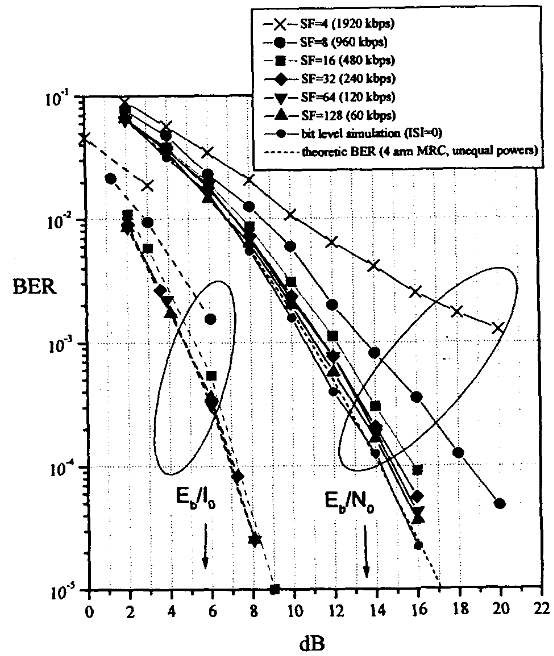

Figure 2.- BER for UTRA-FDD downlink with 120 km/h outdoor channel and ideal channel estimation

In order to estimate the ISI, we can run simulations with one single active connection but with quasi-ideal channel estimation, that is, updating the Rake coefficients only once every slot with the exact conjugate of the global impulse response. For high E_b/N_0, the errors are only due to the ISI, and thus we can estimate E_b/η_{ISI} by comparing the asymptotic BER with the theoretical (no ISI) curve. The argument is the same as before: when the two BER are equal the decision device sees the same noise power. In figure 3 we present this kind of results. We can see how, the lower the SF, the higher is the asymptotic BER, and also how the BER for high SF is now considerably higher than the theoretical curve. Since it is difficult to reach to the asymptotic BER in all the plots, we have evaluated the BER for 100dB of E_b/N_0, and built table 1.

As an example of the use of table 1 we can predict the BER for a SF=128 and E_b/N_0=18dB: first calculate the parallel (in linear) of 18dB // 16.2 dB, this gives approximately 14dB, so the BER will be the corresponding to 14dB in the theoretic curve (about 10^{-4}) and we can check in the plotted BER for SF=128 that the result of our prediction is correct. Strictly speaking, this way of estimating E_b/η_{ISI} is an approximation, since the theoretical curve that we use for comparison assumes a constant noise density independent of the signal power, while the power of the ISI fluctuates with the useful signal power. So it must not be expected a perfect fit in all the cases.

Table 1.- Estimation of E_b/η_{ISI} for downlink UTRA-FDD in an outdoor environment at 120 km/h

SF	Asymptotic BER	E_b/η_{ISI} (dB)
4	$1.58 \cdot 10^{-3}$	10.2
8	$1.33 \cdot 10^{-4}$	13.8
16	$4.63 \cdot 10^{-5}$	15.2
32	$2.69 \cdot 10^{-5}$	15.9
64	$2.30 \cdot 10^{-5}$	16
128	$1.98 \cdot 10^{-5}$	16.2

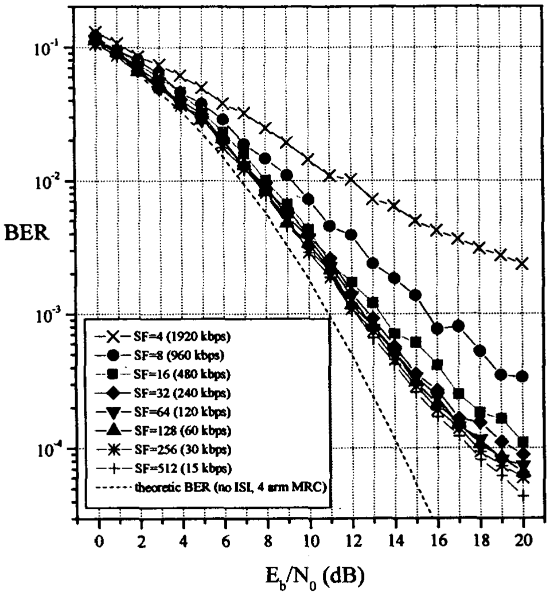

Figure 3.- BER for UTRA-FDD downlink with 120 km/h outdoor channel and quasi-ideal channel estimation

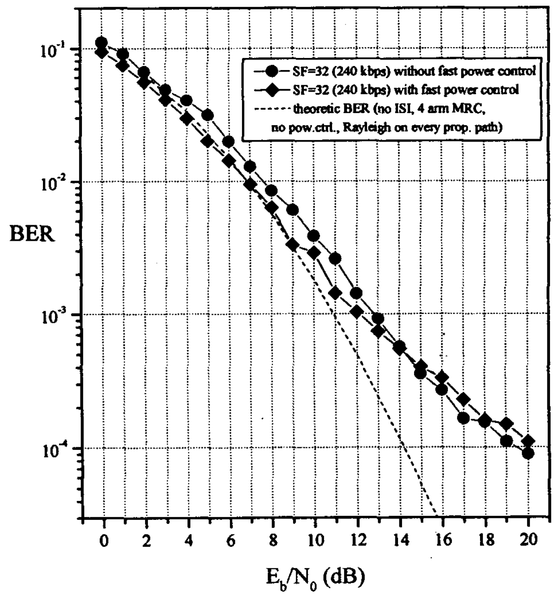

Figure 4.- Comparison of BER in UTRA-FDD downlink with and without fast power control in outdoor environment at 120 km/h

We have verified that, at 120km/h of mobile speed, the ISI is very sensitive to the updating rate of the Rake coefficients. Really, in our simulations we update the Rake every half slot. This is because the performance with one slot updates was very poor and, in fact, the receiver can always be designed to process the received samples half a slot forward and backwards from the pilot bits, so the effective updating error is equivalent to half slot, even if the pilots are transmitted only once per slot.

In order to prove that the fast power control has almost no effect at 120km/h, in figure 4 we compare the BER for a single connection of SF=32 with and without fast power control. We can see that, at this speed, there is only a slight gain in using fast power control and in some cases it can even degrade the performance.

V. CONCLUSIONS

In order to study the UTRA-FDD link, a fully ad-hoc program has been written in C++ language. In contrast with a commercially available simulation package, this allows total control of the model as well as faster execution of the program.

We have simulated the downlink for the UTRA-FDD mode with different number of simultaneous connections and bit rates in a quite realistic outdoor environment. We have seen that the Gaussian hypothesis leads to pessimistic results and that it is required to simulate at chip level to measure the degradation of orthogonality due to the mobile channel and the system imperfections. We have presented BER curves for an outdoor environment and obtained the orthogonality factor in that case.

REFERENCES

[1] Harri Holma, Antti Toskala (Editors), "WCDMA for UMTS. Radio Access for Third Generation Mobile Communications", Ed. John Wiley & Sons, 2000

[2] 3GPP. Spreading and Modulation (FDD), 3G TS 25.213 version 3.1.1 (1999-12)

[3] W.C. Jakes, "Microwave Mobile Communications", pag. 21, John Wiley & Sons, 1974

[4] J.G.Proakis, "Digital Communications", Ed. Mc.Graw Hill, 1983, pp. 486

[5] 3GPP. UE Radio transmission and reception (FDD), 3G TS 25.101 version 3.1.0 (1999-12)