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Channel Capacity Analysis
Under Various Adaptation Policies and
Diversity Techniques over Fading Channels

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1. Introduction

The lack of available spectrum for expansion of wireless services requires more spectrally efficient communication in order to meet the consumer demand. Since the demand for wireless communication services have been growing in recent years at a rapid pace, conserving, sharing and using bandwidth efficiently is of primary concern in future wireless communications systems. Therefore, channel capacity is one of the most important concerns in the design of wireless systems, as it determines the maximum attainable throughput of the system [1]. It can be defined as the average transmitted data rate per unit bandwidth, for a specified average transmit power, and specified level of received outage or bit-error rate [2]. Skilful combination of bandwidth efficient coding and modulation schemes can be used to achieve higher channel capacities per unit bandwidth. However, mobile radio links are, due to the combination of randomly delayed reflected, scattered, and diffracted signal components, subjected to severe multipath fading, which leads to serious degradation in the link signal-to-noise ratio (SNR). An effective scheme that can be used to overcome fading influence is adaptive transmission. The performance of adaptation schemes is further improved by combining them with space diversity, since diversity combining is a powerful technique that can be used to combat fading in wireless systems resulting in improving link performance [3].

1.1 Channel and system model

Diversity combining is a powerful technique that can be used to combat fading in wireless systems [4]. The optimal diversity combining technique is maximum ratio combining (MRC). This combining technique involves co-phasing of the useful signal in all branches, multiplication of the received signal in each branch by a weight factor that is proportional to the estimated ratio of the envelope and the power of that particular signal and summing of the received signals from all antennas. By co-phasing, all the random phase fluctuations of the signal that emerged during the transmission are eliminated. For this

process it is necessary to estimate the phase of the received signal, so this technique requires the entire amount of the channel state information (CSI) of the received signal, and separate receiver chain for each branch of the diversity system, which increases the complexity of the system [5].

One of the least complicated combining methods is selection combining (SC). While other combining techniques require all or some of the amount of the CSI of received signal and separate receiver chain for each branch of the diversity system which increase its complexity, selection combining (SC) receiver process only one of the diversity branches, and is much simpler for practical realization, in opposition to these combining techniques [4-7]. Generally, SC selects the branch with the highest SNR, that is the branch with the strongest signal, assuming that noise power is equally distributed over branches. Since receiver diversity mitigates the impact of fading, the question is whether it also increases the capacity of a fading channel.

Another effective scheme that can be used to overcome fading influence is adaptive transmission. Adaptive transmission is based on the receiver's estimation of the channel and feedback of the CSI to the transmitter. The transmitter then adapts the transmit power level, symbol/bit rate, constellation size, coding rate/scheme or any combination of these parameters in response to the changing channel conditions [8]. Adapting certain parameters of the transmitted signal to the fading channel can help better utilization of the channel capacity. These transmissions provide a much higher channel capacities per unit bandwidth by taking advantage of favorable propagation conditions: transmitting at high speeds under favorable channel conditions and responding to channel degradation through a smooth reduction of their data throughput. The source may transmit faster and/or at a higher power under good channel conditions and slower and/or at a reduced power under poor conditions. A reliable feedback path between that estimator and the transmitter and accurate channel estimation at the receiver is required for achieving good performances of adaptive transmission. Widely accepted adaptation policies include optimal power and rate adaptation (OPRA), constant power with optimal rate adaptation (ORA), channel inversion with fixed rate (CIFR), and truncated CIFR (TIFR). Results obtained for this protocols show the trade-off between capacity and complexity. The adaptive policy with transmitter and receiver side information requires more complexity in the transmitter (and it typically also requires a feedback path between the receiver and transmitter to obtain the side information). However, the decoder in the receiver is relatively simple. The non-adaptive policy has a relatively simple transmission scheme, but its code design must use the channel correlation statistics (often unknown), and the decoder complexity is proportional to the channel decorrelation time. The channel inversion and truncated inversion policies use codes designed for additive white Gaussian noise (AWGN) channels, and are therefore the least complex to implement, but in severe fading conditions they exhibit large capacity losses relative to the other techniques.

The performance of adaptation schemes is further improved by combining them with space diversity. The hypothesis that the variation of the combiner output SNR is tracked perfectly by the receiver and that the variation in SNR is sent back to the transmitter via an error-free feedback path will be assumed in the ongoing analysis [8]. Also, it is assumed that time delay in this feedback path is negligible compared to the rate of the channel variation.

Following these assumptions, transmitter could adapt its power and/or rate relative to the actual channel state.

There are numerous published papers based on study of channel capacity evaluation. In [9], the capacity of Rayleigh fading channels under four adaptation policies and multibranch system with variable correlation is investigated. The capacity of Rayleigh fading channels under different adaptive transmission and different diversity combining techniques is also studied in [7], [10]. In [11], channel capacity of MRC over exponentially correlated Nakagami-*m* fading channels under adaptive transmission is analyzed. Channel capacity of adaptive transmission schemes using equal gain combining (EGC) receiver over Hoyt fading channels is presented in [12]. In [13], dual-branch SC receivers operating over correlative Weibull fading under three adaptation policies are analyzed.

In this chapter we will focus on more general and nonlinear fading distributions. We will perform an analytical study of the *κ-μ* fading channel capacity, e.g., under the OPRA, ORA, CIFR and TIFR adaptation policies and MRC and SC diversity techniques. To the best of authors' knowledge, such a study has not been previously considered in the open technical literature. The expressions for the proposed adaptation policies and diversity techniques will be derived. Capitalizing on them, numerically obtained results will be graphically presented, in order to show the effects of various system parameters, such as diversity order and fading severity on observed performances. In the similar manner an analytical study of the Weibull fading channel capacity, under the OPRA, ORA, CIFR and TIFR adaptation policies and MRC diversity technique will be performed.

1.1.1 *κ-μ* channel and system model

The multipath fading in wireless communications is modelled by several distributions including Nakagami-*m*, Hoyt, Rayleigh, and Rice. By considering important phenomena inherent to radio propagation, *κ-μ* fading model was recently proposed in [14] as a fading model which describes the short-term signal variation in the presence of line-of-sight (LOS) components. This distribution is more realistic than other special distributions, since its derivation is completely based on a non-homogeneous scattering environment. Also *κ-μ* as general physical fading model includes Rayleigh, Rician, and Nakagami-*m* fading models, as special cases [14]. It is written in terms of two physical parameters, *κ* and *μ*. The parameter *κ* is related to the multipath clustering and the parameter *μ* is the ratio between the total power of the dominant components and the total power of the scattered waves. In the case of *κ*=0, the *κ-μ* distribution is equivalent to the Nakagami-*m* distribution. When *μ*=1, the *κ-μ* distribution becomes the Rician distribution with *κ* as the Rice factor. Moreover, the *κ-μ* distribution fully describes the characteristics of the fading signal in terms of measurable physical parameters.

The SNR in a *κ-μ* fading channel follows the probability density function (pdf) given by [15]:

$$p_\gamma(\gamma) = \frac{\mu}{k^{(\mu-1)/2} e^{\mu k}} \left(\frac{1+k}{\bar{\gamma}} \right)^{(\mu+1)/2} \gamma^{(\mu-1)/2} \, e^{-\mu(1+k)\gamma/\bar{\gamma}} I_{\mu-1}\left(2\mu\sqrt{\frac{(1+k)k\gamma}{\bar{\gamma}}} \right). \qquad (1.1)$$

In the previous equation, $\overline{\gamma}$ is the corresponding average SNR, while $I_n(x)$ denotes the n-th order modified Bessel function of first kind [16], and κ and μ are well-known κ-μ fading parameters. Using the series representation of Bessel function [16, eq. 8.445]:

$$I_n(x) = \sum_{k=0}^{+\infty} \frac{x^{2k+n}}{2^{2k+n}\Gamma(k+n+1)k!},$$ (1.2)

the cumulative distribution function (cdf) of γ can be written in the form of:

$$F_\gamma(\gamma) = \sum_{p=0}^{+\infty} \frac{\mu^p \kappa^p}{e^{\mu\kappa}\Gamma(p+\mu)} \Lambda\left(p+\mu, \frac{\mu(1+\kappa)\gamma}{\overline{\gamma}}\right)$$ (1.3)

with $\Gamma(x)$ and $\Lambda(a,x)$ denoting Gamma and lower incomplete Gamma function, respectively [16, eqs. 8.310.1, 8.350.1].

It is shown in [15], that the sum of κ-μ squares is κ-μ square as well (but with different parameters), which is an ideal choice for MRC analysis. Then the expression for the pdf of the outputs of MRC diversity systems follows [15, eq.11]:

$$p_\gamma^{MRC}(\gamma) = \frac{L\mu}{k^{(L\mu-1)/2}e^{L\mu k}} \left(\frac{1+k}{L\overline{\gamma}}\right)^{(L\mu+1)/2} \gamma^{(L\mu-1)/2} e^{-\mu(1+k)\gamma/\Omega} I_{L\mu-1}\left(2\mu L\sqrt{\frac{(1+k)k\gamma}{L\overline{\gamma}}}\right)$$ (1.4)

with L denoting the number of diversity branches.

The expression for the pdf of the outputs of SC diversity systems can be obtained by substituting expressions (1.1) and (1.3) into:

$$p_\gamma^{SC}(\gamma) = \sum_{i=1}^{L} p_{\gamma_i}(\gamma) \prod_{\substack{j=1 \\ j \neq i}}^{L} F_{\gamma_j}(\gamma)$$ (1.5)

where $p_{\gamma i}(\gamma)$ and $F_{\gamma i}(\gamma)$ define pdf and cdf of SNR at input branches respectively and L denotes the number of diversity branches.

1.1.2 Weibull channel and system model

The above mentioned well-known fading distributions are derived assuming a homogeneous diffuse scattering field, resulting from randomly distributed point scatterers. The assumption of a homogeneous diffuse scattering field is certainly an approximation, because the surfaces are spatially correlated characterizing a nonlinear environment. With the aim to explore the nonlinearity of the propagation medium, a general fading distribution, the Weibull distribution, was proposed. The nonlinearity is manifested in terms of a power parameter $\beta > 0$, such that the resulting signal intensity is obtained not simply as the modulus of the multipath component, but as the modulus to a certain given power. As β increases, the fading severity decreases, while for the special case of $\beta = 2$ reduces to the

well-known Rayleigh distribution. Weibull distribution seems to exhibit good fit to experimental fading channel measurements, for both indoor and outdoor environments.

The SNR in a Weibull fading channel follows the pdf given by [17, eq.14]:

$$p(\gamma) = \frac{\beta}{2a\overline{\gamma}} \left(\frac{\gamma}{a\overline{\gamma}} \right)^{\frac{\beta}{2}-1} e^{-\left(\frac{\gamma}{a\overline{\gamma}}\right)^{\beta/2}} \tag{1.6}$$

In the previous equation, $\overline{\gamma}$ is the corresponding average SNR, β is well-known Weibull fading parameter, and $a = 1/\Gamma(1+2/\beta)$.

It is shown in [18,19], that the expression for the pdf of the outputs of MRC diversity systems follows [19, eq.1]:

$$p_\gamma^{MRC}(\gamma) = \frac{\beta\gamma^{L\beta/2-1}}{2\Gamma(L)\left(\Xi\overline{\gamma}\right)^{L\beta/2}} e^{-\left(\frac{\gamma}{\Xi\overline{\gamma}}\right)^{\beta/2}} ; \qquad \Xi = \frac{\Gamma(L)}{\Gamma(L+2/\beta)} \tag{1.7}$$

with L denoting the number of diversity branches.

Similary, expression for the pdf of the outputs of SC diversity systems can be obtained as (1.5)

2. Optimal power and rate adaptation

In the OPRA protocol the power level and rate parameters vary in response to the changing channel conditions. It achieves the ergodic capacity of the system, i. e. the maximum achievable average rate by use of adaptive transmission. However, OPRA is not suitable for all applications because for some of them it requires fixed rate.

During our analysis it is assumed that the variation in the combined output SNR over κ-μ fading channels γ is tracked perfectly by the receiver and that variation of γ is sent back to the transmitter via an error-fee feedback path. Comparing to the rate of channel variation, the time delay in this feedback is negligible. These assumptions allow the transmitter to adopt its power and rate correspondingly to the actual channel state. Channel capacity of the fading channel with received SNR distribution, $p_\gamma(\gamma)$, under optimal power and rate adaptation policy, for the case of constant average transmit power is given by [8]:

$$<C>_{pra} = B \int_{\gamma_0}^{\infty} \log_2 \left(\frac{\gamma}{\gamma_0} \right) p_\gamma(\gamma) d\gamma, \tag{1.8}$$

where B (Hz) denotes the channel bandwidth and γ_0 is the SNR cut-off level bellow which transmission of data is suspended. This cut-off level must satisfy the following equation:

$$\int_{\gamma_0}^{\infty} \left(\frac{1}{\gamma_0} - \frac{1}{\gamma} \right) p_\gamma(\gamma) d\gamma = 1, \tag{1.9}$$

Since no data is sent when $\gamma < \gamma_0$, the optimal policy suffers a probability of outage P_{out} equal to the probability of no transmission, given by:

$$P_{out} = \int_{0}^{\gamma_0} p_\gamma(\gamma)d\gamma = 1 - \int_{\gamma_0}^{\infty} p_\gamma(\gamma)d\gamma \qquad (1.10)$$

2.1 *κ-µ* fading channels

To achieve the capacity in (1.8), the channel fading level must be attended at the receiver as well as at the transmitter. The transmitter has to adapt its power and rate to the actual channel state; when γ is large, high power levels and rates are allocated for good channel conditions and lower power levels and rates for unfavourable channel conditions when γ is small. Substituting (1.1) into (1.9), we found that the cut-off level must satisfy:

$$\sum_{i=0}^{\infty} \frac{(kL\mu)^i}{e^{L\mu k}\Gamma(i+L\mu)i!}\left(\frac{1}{\gamma_0}\Lambda\left(L\mu+i, \frac{\mu(1+k)\gamma_0}{\bar{\gamma}}\right) - \right.$$
$$\left. \frac{\mu(1+k)}{\bar{\gamma}}\Lambda\left(L\mu+i-1, \frac{\mu(1+k)\gamma_0}{\bar{\gamma}}\right)\right) - 1 = 0 \qquad (1.11)$$

Substituting (1.1) into (1.8), we obtain the capacity per unit bandwidth, $<C>_{opra}/B$, as:

$$\frac{\langle C\rangle_{opra}^{MRC}}{B} = \sum_{i=0}^{\infty} \frac{L\mu}{k^{(L\mu-1)/2}e^{L\mu k}}\left(\frac{1+k}{L\bar{\gamma}}\right)^{(L\mu+1)/2} \int_{\gamma_0}^{\infty} \log_2\left(\frac{\gamma}{\gamma_0}\right)\gamma^{L\mu+i-1}e^{-\mu(1+k)\gamma/\bar{\gamma}}d\gamma \qquad (1.12)$$

Now, by making change of variables, , $<C>_{opra}/B$ can be obtained as:

$$\frac{\langle C\rangle_{opra}^{MRC}}{B} = \sum_{i=0}^{\infty} \frac{(L\mu k)^i}{\Gamma(i+L\mu)i!e^{L\mu k}}\left(\int_{0}^{\infty} \log_2\left(\frac{t\bar{\gamma}}{\mu(1+k)\gamma_0}\right)t^{L\mu+i-1}e^{-t}dt - \right.$$
$$\left. \int_{0}^{\gamma_0\mu(1+k)/\bar{\gamma}} \log_2\left(\frac{t\bar{\gamma}}{\mu(1+k)\gamma_0}\right)t^{L\mu+i-1}e^{-t}dt\right) = \sum_{i=0}^{\infty} \frac{(L\mu k)^i}{\Gamma(i+L\mu)i!e^{L\mu k}}(I_1 - I_2) \qquad (1.13)$$

Integral I_1 can be solved by applying Gauss-Laguerre quadrature formulae:

$$I_1 = \int_{0}^{\infty} f_1(t)e^{-t}dt \cong \sum_{k=1}^{R} A_k f_1(t_k); \quad f_1(t) = \log_2\left(\frac{t\bar{\gamma}}{\mu(1+k)\gamma_0}\right)t^{L\mu+i-1} \qquad (1.14)$$

In the previous equation A_k and t_k, $k=1,2,\ldots,R$, are respectively weights and nodes of Laguerre polynomials [20, pp. 875-924].

Similarly, integral I_2 can be solved by applying Gauss-Legendre quadrature formulae:

$$I_2 = \left(\frac{\gamma_0 \mu (1+k)}{2\bar{\gamma}} \right)^{L\mu+i} \int_{-1}^{1} f_2(u)du \cong \left(\frac{\gamma_0 \mu (1+k)}{2\bar{\gamma}} \right)^{L\mu+i} \sum_{k=1}^{R} B_k f_2(u_k)$$

(1.15)

where B_k and u_k, $k=1,2,\ldots,R$, are respectively weights and nodes of Legendre polynomials.

Convergence of infinite series expressions in (1.13) is rapid since we need about 10 terms to be summed in order to achieve accuracy at the 5th significant digit for corresponding values of system parameters.

2.2 Weibull fading channels

Substituting (1.7) in (1.8) integral of the following form need to be solved

$$I = \frac{1}{\ln 2} \int_{\gamma_0}^{\infty} \gamma^{L\beta/2-1} \ln\left(\frac{\gamma}{\gamma_0} \right) e^{-\left(\frac{\gamma}{\Xi\bar{\gamma}} \right)^{\beta/2}} d\gamma .$$

(1.16)

After making a change of variables $t = (\gamma / \gamma_0)^{\beta/2}$ and some simple mathematical manipulations, we get:

$$I = \frac{4\gamma_0^{L\beta/2}}{\beta^2 \ln 2} \int_{1}^{\infty} t^{L-1} \ln(t) e^{-\left(\frac{\gamma_0}{\Xi\bar{\gamma}} \right)^{\beta/2} t} dt$$

(1.17)

Furthermore, this integral can be evaluated using partial integration:

$$\int_{1}^{\infty} u d\upsilon = \lim_{t\to\infty}(u\upsilon) - \lim_{t\to 1}(u\upsilon) - \int_{1}^{+\infty} \upsilon du$$

(1.18)

with respect to:

$$u = \ln t; \quad d\upsilon = t^{L-1} e^{-\left(\frac{\gamma_0}{\Xi\bar{\gamma}} \right)^{\beta/2} t} dt .$$

(1.19)

Performing L-1 successive integration by parts [16, eq. 2.321.2], we get

$$\upsilon = -e^{-mt} \sum_{p=1}^{L} \frac{(L-1)!}{(L-p)!} \frac{t^{L-p}}{m^p}$$

(1.20)

denoting $m = (\gamma_0 / \Xi\bar{\gamma})^{\beta/2}$. Substituting (1.20) in (1.18), we see that first two terms tend to zero. Hence, the integral in (1.17) can be solved in closed form using [16, eq 3.381.3]

$$I = \frac{(L-1)!}{m^L} \sum_{p=0}^{L-1} \frac{\Gamma(p,m)}{p!} \quad (1.21)$$

with $\Gamma(a, x)$ higher incomplete Gamma function [16]. Finaly, $<C>_{opra}/B$ using L-branch MRC diversity receiver over Weibull fading channels has this form

$$\frac{\langle C \rangle_{opra}^{MRC}}{B} = \frac{2}{\beta \ln 2} \sum_{p=0}^{L-1} \frac{\Gamma(p,m)}{p!} \cdot \quad (1.22)$$

3. Constant power with optimal rate adaptation

With ORA protocol, the transmitter adapts its rate only while maintaining a fixed power level. Thus, this protocol can be implemented at reduced complexity and is more practical than that of optimal simultaneous power and rate adaptation.

The channel capacity, $<C>_{ora}$ (bits/s) with constant transmit power policy is given by [1]:

$$\langle C \rangle_{ora} = B \int_0^\infty \log_2(1+\gamma) p_\gamma(\gamma) d\gamma \quad (1.22)$$

3.1 κ-μ fading channels

To achieve the capacity in (1.22), the channel fading level must be attended at the receiver as well as at the transmitter.

After substituting (1.1) into (1.22), by using partial integration:

$$\int_0^\infty u dv = \lim_{\gamma \to \infty}(uv) - \lim_{\gamma \to 0}(uv) - \int_0^\infty v du \quad (1.23)$$

with respect to:

$$u = \ln(1+\gamma); \quad du = \frac{d\gamma}{1+\gamma}; \quad dv = \gamma^{p+\mu-1} e^{-\frac{\mu(1+k)\gamma}{\bar{\gamma}}}; \quad (1.24)$$

and performing successive integration by parts [16 , eq. 2.321.2], we get

$$v = e^{-\frac{\mu(1+k)\gamma}{\bar{\gamma}}} \sum_{q=1}^{p+\mu} \frac{(p+\mu-1)! \gamma^{p+\mu-k}}{(p+\mu-q)!} \left(\frac{\bar{\gamma}}{\mu(1+k)} \right)^q \quad (1.25)$$

By substituting (1.25) in (1.23), we see that first two terms tend to zero. Hence, the integral in (1.23) can be solved in closed form using [16, eq. 3.381.3]. Finaly, $<C>_{ora}/B$ over κ-μ fading channels has this form:

$$\langle C \rangle_{ora} = \frac{B}{\ln 2} \sum_{p=0}^{\infty} \sum_{q=1}^{p+\mu} \frac{\mu^{2p+\mu-q} \kappa^p (1+\kappa)^{p+\mu-q} (n-1)!}{e^{\mu\kappa} \overline{\gamma}^{-p+\mu-q} \Gamma(p+\mu) p!} e^{\frac{\mu(1+\kappa)}{\overline{\gamma}}}$$

$$\Gamma\left(-n+p+\mu, \frac{\mu(1+\kappa)}{\overline{\gamma}}\right) \tag{1.26}$$

On the other hand, substituting (1.4) into (1.22) and applying similar procedure, the expression for the $\langle C \rangle_{ora}/B$ with MRC diversity receiver is derived as:

$$\langle C \rangle_{ora}^{MRC} = \frac{B}{\ln 2} \sum_{p=0}^{\infty} \sum_{q=1}^{p+\mu} \frac{\mu^{2p+\mu-q} \kappa^p (1+\kappa)^{p+\mu-q} (n-1)! L^p}{e^{L\mu\kappa} \overline{\gamma}^{-p+\mu-q} \Gamma(p+\mu) p!} e^{\frac{\mu(1+\kappa)}{\overline{\gamma}}}$$

$$\Gamma\left(-n+p+\mu L, \frac{\mu(1+\kappa)}{\overline{\gamma}}\right) \tag{1.27}$$

Convergence of infinite series expressions in (1.26) and (1.27) is rapid, since we need 5-10 terms to be summed in order to achieve accuracy at the 5th significant digit for corresponding values of system parameters.

3.2 Weibull fading channels

After substituting (1.6) into (1.22), when MRC reception is applied over Weibull fading channel, we can obtain expression for the ORA channel capacity, in the form of:

$$\frac{\langle C \rangle_{ora}}{B} = \frac{\beta}{2\Gamma(L)(\Xi\overline{\gamma})^{L\beta/2} \ln 2} \int_0^{\infty} \gamma^{L\beta/2} \ln(1+\gamma) e^{-\left(\frac{\gamma}{\Xi\overline{\gamma}}\right)^{\beta/2}} d\gamma. \tag{1.28}$$

By expressing the logarithmic and exponential integrands as Meijer's G- functions [21, eqs. 11] and using [22, eq. 07.34.21.0012.01], integral in (1.28) is solved in closed-form:

$$\frac{\langle C \rangle_{ora}}{B} = \frac{\beta}{2\Gamma(L)(\Xi\overline{\gamma})^{L\beta/2} \ln 2} H_{2,3}^{3,1}\left((\Xi\overline{\gamma})^{-\beta/2} \left|\begin{array}{l}(-L\beta/2, \beta/2), (1-L\beta/2, \beta/2) \\ (0,1), (-L\beta/2, \beta/2), (-L\beta/2, \beta/2)\end{array}\right.\right) \tag{1.29}$$

with:

$$H_{p,q}^{m,n}\left(x \left|\begin{array}{l}(a_1, \alpha_1)....(a_p, \alpha_p) \\ (b_1, \beta_q)....(b_p, \beta_q)\end{array}\right.\right) \tag{1.30}$$

denoting the Fox H function [23].

4. Channel inversion with fixed rate

Channel inversion with fixed rate policy (CIFR protocol) is quite different than the first two protocols as it maintains constant rate and adapts its power to the inverse of the channels fading. CIFR protocol achieves what is known as the outage capacity of the system; that is the maximum constant data rate that can be supported for all channel conditions with some probability of outage. However, the capacity of channel inversion is always less than the capacity of the previous two protocols as the transmission rate is fixed. On the other hand, constant rate transmission is required in some applications and is worth the loss in achievable capacity. CIFR is adaptation technique based on inverting the channel fading. It is the least complex technique to implement assuming that the transmitter on this way adapts its power to maintain a constant SNR at the receiver. Since a large amount of the transmitted power is required to compensate for the deep channel fades, channel inversion with fixed rate suffers a certain capacity penalty compared to the other techniques.

The channel capacity with this technique is derived from the capacity of an AWGN channel and is given in [8]:

$$\langle C \rangle_{cifr} = B \log_2 \left(1 + 1 \Big/ \int_0^\infty \left(p_\gamma \left(\gamma \right) / \gamma \right) d\gamma \right). \tag{1.31}$$

4.1 *κ-μ* fading channels

After substituting (1.1) into (1.31), and by using [16, eq. 6.643.2]:

$$\int_0^\infty x^{\mu - \frac{1}{2}} e^{-\alpha x} I_{2\nu} \left(2\beta \sqrt{x} \right) dx = \frac{\Gamma \left(\mu + \nu + \frac{1}{2} \right)}{\Gamma \left(2\nu + 1 \right)} \beta^{-1} e^{-\frac{\beta^2}{2\alpha}} \alpha^{-\nu} M_{-\mu,\nu} \left(\frac{\beta^2}{\alpha} \right) \tag{1.32}$$

where $M_{k,m}(z)$ is the Wittaker's function, we can obtained expression for the CIFR channel capacity in the form of:

$$\langle C \rangle_{cifr} = B \log_2 \left(1 + \frac{(\mu - 1)}{e^{-\frac{\mu k}{2}} \left(\frac{1 + k}{k \overline{\gamma}} \right)^{\frac{\mu}{2}} M_{1 - \frac{\mu}{2}, \frac{\mu - 1}{2}} \left(\mu k \right)} \right). \tag{1.33}$$

Case when MRC diversity is applied can be modelled by:

$$\langle C \rangle_{cifr}^{MRC} = B \log_2 \left(1 + \frac{(L\mu - 1)}{e^{-\frac{\mu k L}{2}} \left(\frac{1 + k}{k L \overline{\gamma}} \right)^{\frac{L\mu}{2}} M_{1 - \frac{L\mu}{2}, \frac{L\mu - 1}{2}} \left(\mu k L \right)} \right). \tag{1.34}$$

Similarly, after substituting (1.5) into (1.31), with respect to [16, eqs. 8.531, 7.552.5, 9.14]:

$$\Lambda\left(a, x\right) = \frac{x^a}{a} e^{-x} {}_1F_1\left(1; 1 + a; x\right)$$ (1.35)

$$\int_0^\infty e^{-x} x^{s-1} {}_pF_q(a_1, ..., a_p; b_1, ..., b_q; \alpha x) dx = \Gamma(s) {}_{p+1}F_q(s, a_1, ..., a_p; b_1, ..., b_q; \alpha x)$$ (1.36)

$${}_1F_1\left(a; b; x\right) = \sum_{k=0}^\infty \frac{(a)_k x^k}{(b)_k k!}$$ (1.37)

expressions for the CIFR channel capacity over κ-μ fading with SC diversity applied for dual and triple branch combining at the receiver can be obtained in the form of:

$$\left\langle C\right\rangle_{cifr}^{SC-2} = B\log_2\left(1 + 1\Bigg/\sum_{p=0}^\infty\sum_{q=0}^\infty f_1\right)$$

$$f_1 = \frac{\mu^{p+q+1}\kappa^{p+q}(1+\kappa)\Gamma(p+q+2\mu-1)}{2^{p+q+2\mu-2}e^{2\mu\kappa}\,\overline{\gamma}\,\Gamma(p+\mu)p!\Gamma(q+\mu)q!(q+\mu)}$$ (1.38)

$${}_2F_1\left(p+q+2\mu-1, 1; 1+q+\mu; \frac{1}{2}\right)$$

$$\left\langle C\right\rangle_{cifr}^{SC-3} = B\log_2\left(1 + 1\Bigg/\sum_{p=0}^\infty\sum_{q=0}^\infty\sum_{r=0}^\infty\sum_{s=0}^\infty f_2\right)$$

$$f_2 = \frac{\mu^{p+q+r+1}\kappa^{p+q+r}(1+\kappa)\Gamma(p+q+r+s+3\mu-1)}{3^{p+q+r+s+3\mu-3}e^{3\mu\kappa}\,\overline{\gamma}\,\Gamma(p+\mu)p!\Gamma(q+\mu)q!(q+\mu)\Gamma(r+\mu)r!(r+\mu)(1+r+\mu)_s}$$ (1.39)

$$\times\,{}_2F_1\left(p+q+r+s+3\mu-1, 1; 1+q+\mu; \frac{1}{3}\right)$$

Number of terms that need to be summed in (1.38) and (1.39) to achieve accuracy at 5th significant digit for some values of system parameters is presented in Table 1 in the section Numerical results.

4.2 Weibull fading channels

After substituting (1.6) into (1.31) we can obtain expression for the CIFR channel capacity when MRC diversity is applied in the form of:

$$\frac{\left\langle C\right\rangle_{cifr}^{MRC}}{B} = \log_2\left(1 + \frac{(\Xi\overline{\gamma})\Gamma(L)}{\Gamma(L-2/\beta)}\right).$$ (1.40)

5. Truncated channel inversion with fixed rate

The channel inversion and truncated inversion policies use codes designed for AWGN channels, and are therefore the least complex to implement, but in severe fading conditions they exhibit large capacity losses relative to the other techniques.

The truncated channel inversion policy inverts the channel fading only above a fixed cutoff fade depth γ_0. The capacity with this truncated channel inversion and fixed rate policy $<C>_{tifr}/B$ is derived in [8]:

$$\left\langle C \right\rangle_{tifr} = B \log_2 \left(1 + 1 \Big/ \int_{\gamma_0}^{\infty} \left(p_\gamma \left(\gamma \right) / \gamma \right) d\gamma \right) \left(1 - P_{out} \right). \tag{1.41}$$

5.1 κ-µ fading channels

After substituting (1.2) into (1.40) we can obtain expression for the CIFR channel capacity over κ-µ fading channel in the following form:

$$\left\langle C \right\rangle_{tifr} = B \log_2 \left(1 + 1 \Big/ \sum_{p=0}^{\infty} f_3 \right) \left(1 - \sum_{i=0}^{\infty} \frac{\left(k\mu \right)^i}{e^{\mu k} \Gamma \left(i + \mu \right) i!} \Lambda \left(\mu + i, \frac{\mu \left(1 + k \right) \gamma_0}{\overline{\gamma}} \right) \right) \tag{1.42}$$

$$f_3 = \sum_{p=0}^{\infty} \frac{\mu^{p+1} \kappa^p \left(1 + \kappa \right) \Lambda \left(p + \mu - 1, \frac{\mu \left(1 + k \right) \gamma_0}{\overline{\gamma}} \right)}{e^{\mu \kappa} \overline{\gamma} \Gamma \left(p + \mu \right) p!}$$

Case when MRC diversity is applied can be modelled by:

$$\left\langle C \right\rangle_{tifr}^{MRC} = B \log_2 \left(1 + 1 \Big/ \sum_{p=0}^{\infty} f_4 \right) \left(1 - \sum_{i=0}^{\infty} \frac{\left(kL\mu \right)^i}{e^{L\mu k} \Gamma \left(i + L\mu \right) i!} \Lambda \left(\mu L + i, \frac{\mu L \left(1 + k \right) \gamma_0}{\overline{\gamma}} \right) \right) \tag{1.43}$$

$$f_4 = \sum_{p=0}^{\infty} \frac{\mu^{p+1} \kappa^p \left(1 + \kappa \right) L^p \Lambda \left(p + \mu L - 1, \frac{\mu L \left(1 + k \right) \gamma_0}{\overline{\gamma}} \right)}{e^{\mu \kappa L} \overline{\gamma} \Gamma \left(p + \mu L \right) p!}$$

Convergence of infinite series expressions in (1.42) and (1.43) is rapid, since we need about 10-15 terms to be summed in order to achieve accuracy at the 5th significant digit.

5.2 Weibull fading channels

After substituting (1.6) into (1.41) we can obtained expression for the CIFR channel capacity over Weibull fading channels when MRC diversity is applied in the form of:

$$\frac{\langle C \rangle_{tifr}^{MRC}}{B} = \log_2 \left(1 + \frac{\Xi \overline{\gamma} \Gamma(L)}{\Gamma\left(L - 2/\beta, \left(\gamma_0/\Xi\overline{\gamma}\right)^{\beta/2}\right)} \right) \frac{\Gamma\left(L, \left(\gamma_0/\Xi\overline{\gamma}\right)^{\beta/2}\right)}{\Gamma(L)}. \qquad (1.44)$$

6. Numerical results

In order to discuss usage of diversity techniques and adaptation policies and to show the effects of various system parameters on obtained channel capacity, numerically obtained results are graphically presented.

In Figs. 1.1 and 1.8 channel capacity without diversity, $<C>_{ora}$ given by (1.22), for the cases when κ-μ and Weibull fading are affecting channels, for various system parameters are plotted against $\overline{\gamma}$. These figures also display the capacity per unit bandwidth of an AWGN channel, C_{AWGN} given by:

$$C_{AWGN} = B \log_2(1 + \gamma). \qquad (1.45)$$

Considering obtained results, with respect that C_{AWGN} = 3.46 dB for average received SNR of 10dB we find that depending of fading parameters of κ-μ and Weibull distribution, channel capacity could be reduced up to 30 %. From Fig. 1.1 we can see that channel capacity is less reduced for the cases when fading severity parameter μ, and dominant/scattered components power ratio κ, have higher values, since for smaller κ and μ values the dynamics in the channel is larger. Also from Fig. 1.8 we can observe that channel capacity is less reduced in the areas where Weibull fading parameter β has higher values.

Figures 1.2-1.4,1.6 show the channel capacity per unit bandwidth as a function of $\overline{\gamma}$ for the different adaptation policies with MRC diversity over κ-μ fading channels. It can be seen that as the number of combining branches increases the fading influence is progressively reduced, so the channel capacity improves remarkably. However, as L increases, all capacities of the various policies converge to the capacity of an array of L independent AWGN channels, given by:

$$C_{AWGN}^{MRC} = B \log_2(1 + L\gamma) \qquad (1.46)$$

Thus, in practice it is not possible to entirely eliminate the effects of fading through space diversity since the number of diversity branches is limited. Also considering downlink (base station to mobile) implementation, we found that mobile receivers are generally constrained in size and power.

In Fig. 1.5 comparison of the channel capacity per unit bandwidth with CIFR adaptation policy, when SC and MRC diversity techniques are applied at the reception is shown. As expected, better performances are obtained when MRC reception over κ-μ fading channels is applied.

Figure 1.7 shows the calculated channel capacity per unit bandwidth as a function of $\overline{\gamma}$ for different adaptation policies. From this figure we can see that the OPRA protocol yields a small increase in capacity over constant transmit power adaptation and this small increase

in capacity diminishes as $\overline{\gamma}$ increases. However, greater improvement is obtained in going from complete to truncated channel inversion policy. Truncated channel inversion policy provides better diversity gain compared to complete channel inversion varying any of parameters.
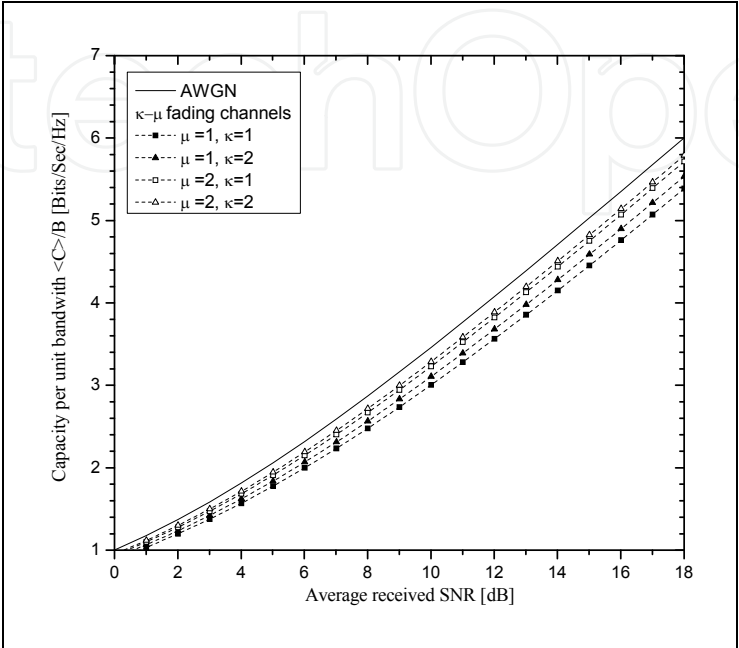

Fig. 1.1 Average channel capacity per unit bandwidth for a κ-μ fading and an AWGN channel versus average received SNR.
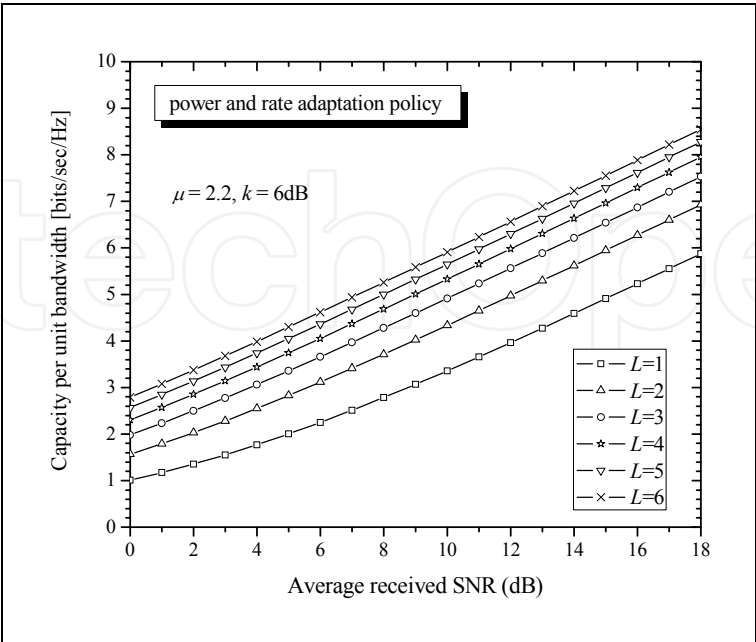

Fig. 1.2 Power and rate adaptation policy capacity per unit bandwidth over κ-μ fading channels, for various values of diversity order.
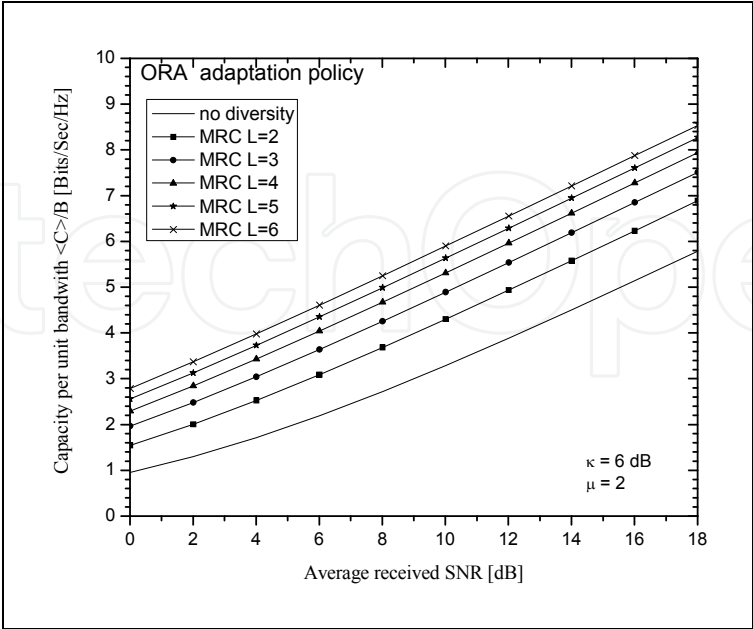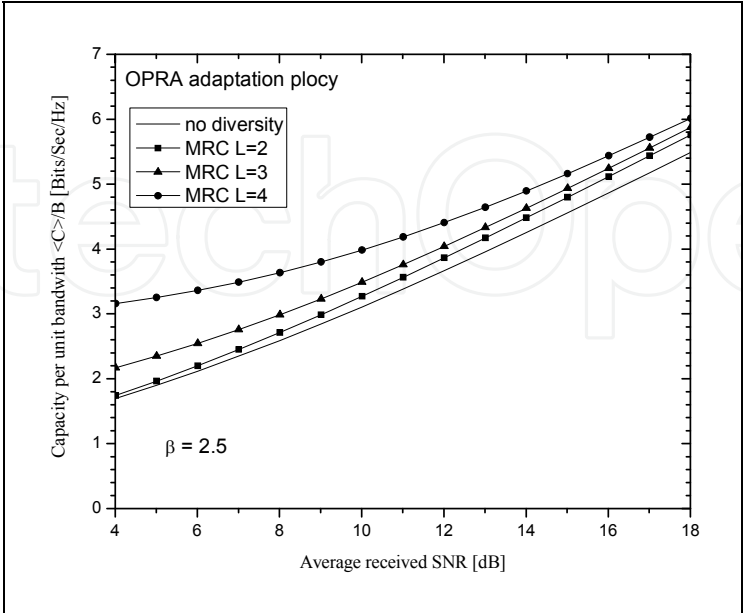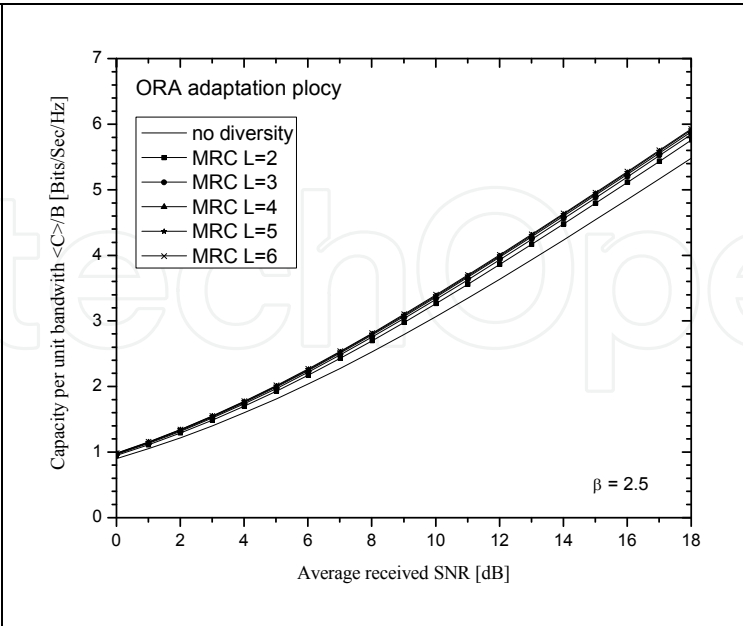
Fig. 1.3 ORA policy capacity per unit bandwidth over κ-μ fading channels, for various values of MRC diversity order.
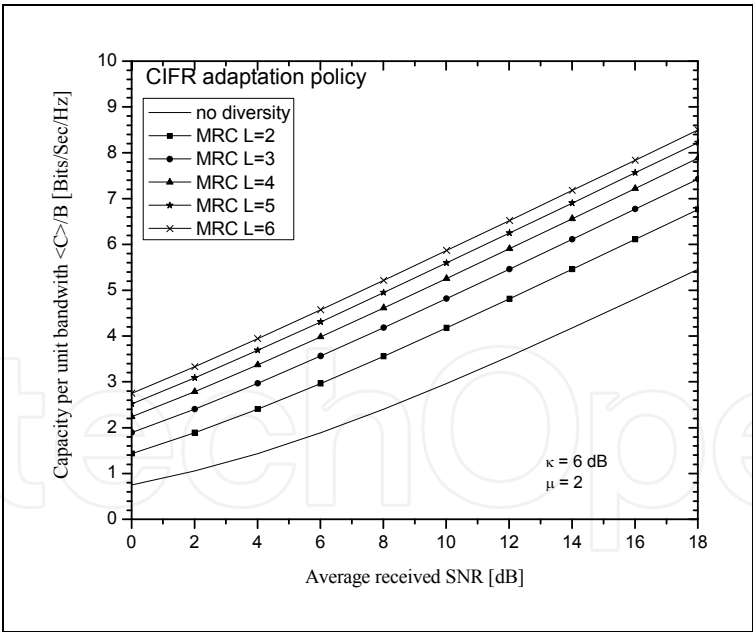
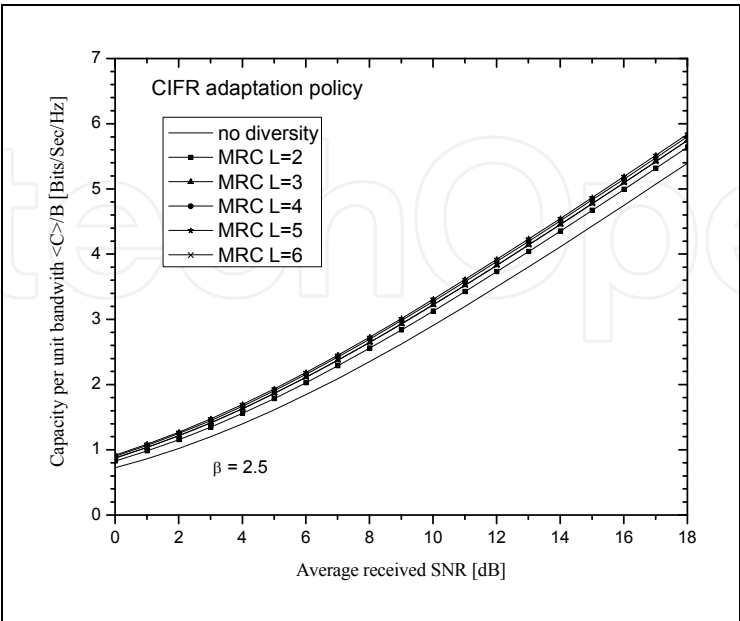Fig. 1.4 CIFR policy capacity per unit bandwidth over κ-μ fading channels, for various values of MRC diversity order.

Similar results are presented considering channels affected by Weibull fading. Figures 1.9-1.12 show the channel capacity per unit bandwidth as a function of $\overline{\gamma}$ for the different adaptation policies with L-branch MRC diversity applied. Comparison of adaptation policies is presented at Fig. 1.13.
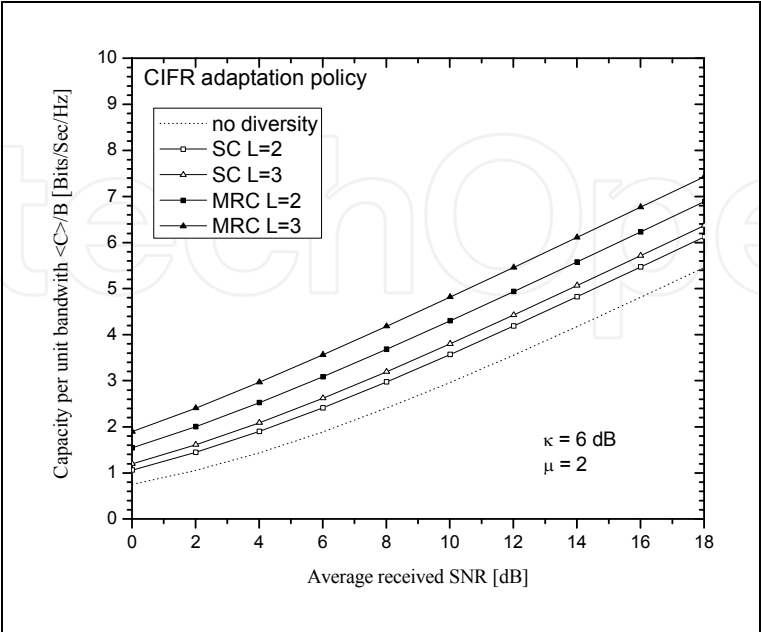
Fig. 1.5 CIFR policy capacity per unit bandwidth over κ-μ fading channels, for MRC and SC diversity techniques various orders.
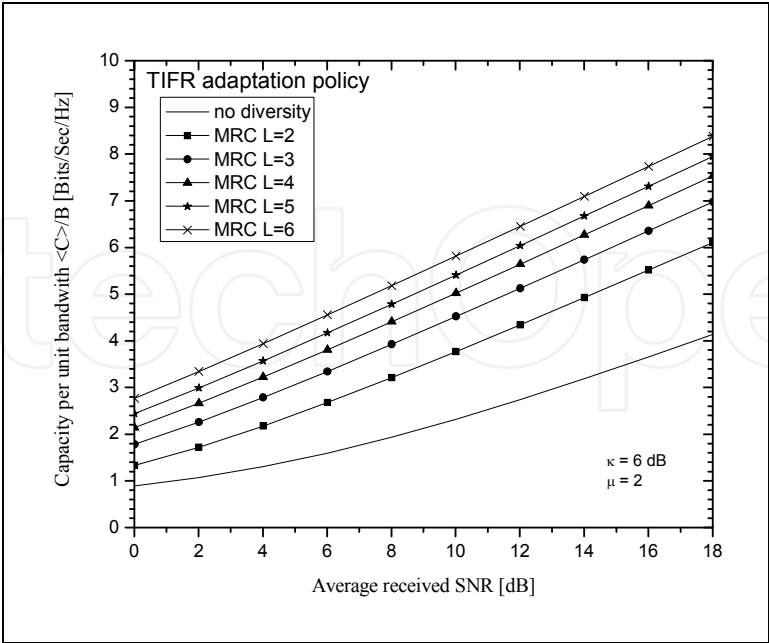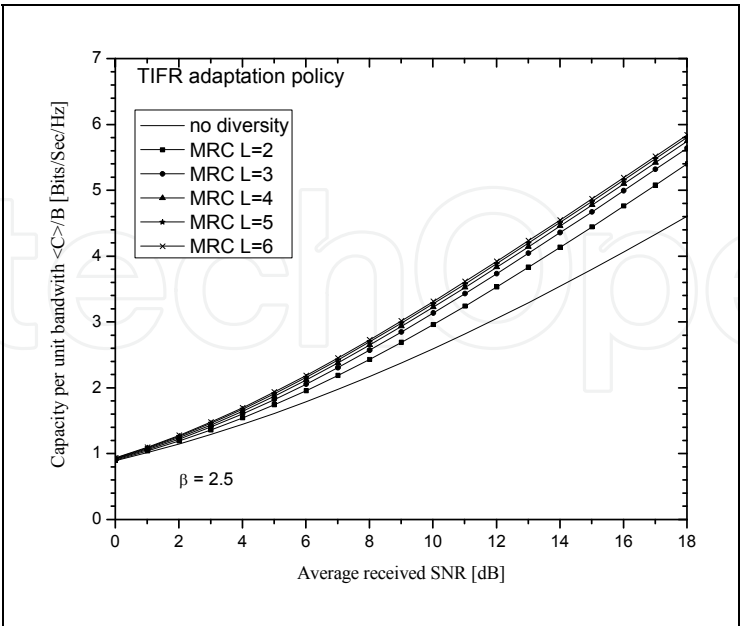

Fig. 1.6 TIFR policy capacity per unit bandwidth over κ-μ fading channels, for various values of MRC diversity order.
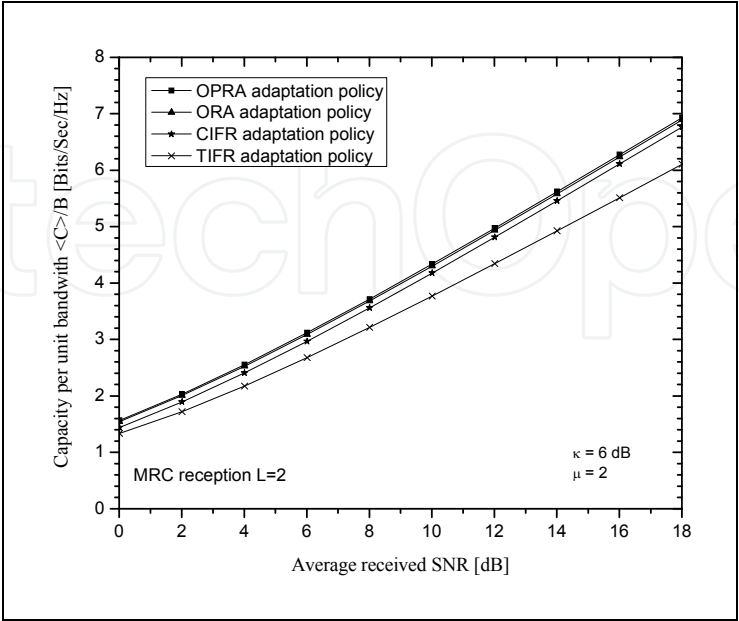
Fig. 1.7 Comparison of adaptation policies over MRC diversity reception in the presence of κ-μ fading.
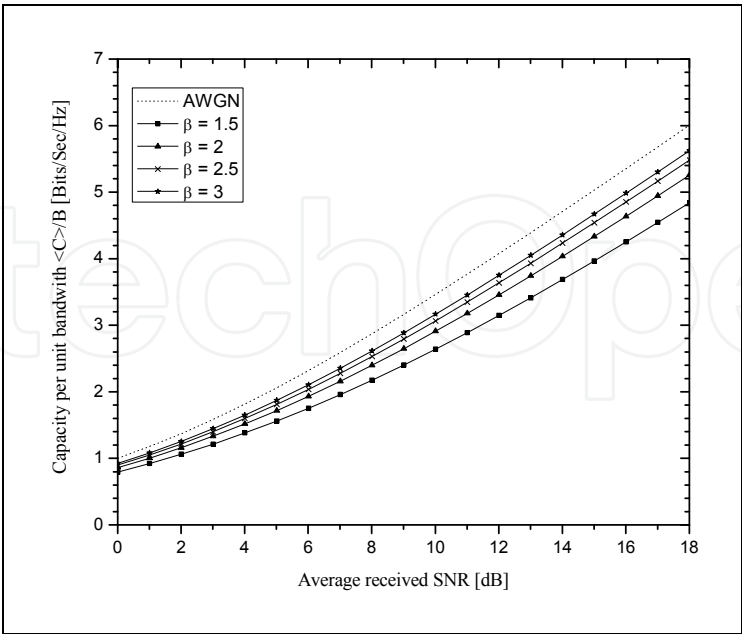
Fig. 1.8 Average channel capacity per unit bandwidth for a Weibull fading for various values of system parameters and an AWGN channel versus average received SNR [dB].

Fig. 1.9 ORPA policy capacity per unit bandwidth over Weibull fading channels, for various values of MRC diversity order.

Fig. 1.10 ORA policy capacity per unit bandwidth over Weibull fading channels, for various values of MRC diversity order.

Fig. 1.11 CIFR policy capacity per unit bandwidth over Weibull fading channels, for various values of MRC diversity order.
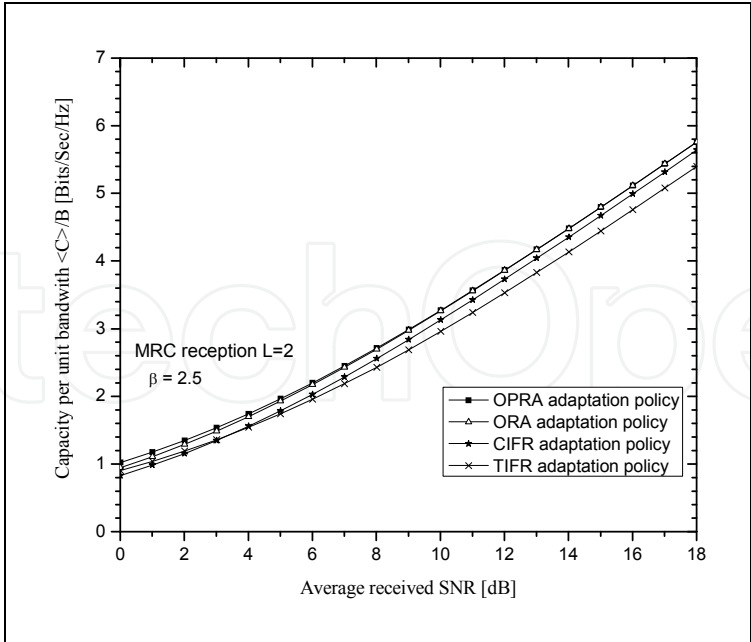

Fig. 1.12 TIFR policy capacity per unit bandwidth over Weibull fading channels, for various values of MRC diversity order.

Fig. 1.13 Comparison of adaptation policies over MRC diversity reception in the presence of Weibull fading.

The nested infinite sums in (1.38) and (1.39), as can be seen from Table 1, for dual and triple branch diversity case, converge for any value of the parameters κ, μ and $\overline{\gamma}$. As it is shown in this Table 1, the number of the terms need to be summed to achieve a desired accuracy, depends strongly on these parameters and it increases as these parameter values increase.

Expression (1.38) 6th significant digit		$\overline{\gamma} = 5$ dB	$\overline{\gamma} = 10$ dB	$\overline{\gamma} = 15$ dB
$\kappa = 1$	$\mu = 1$	8	9	10
$\kappa = 2$	$\mu = 2$	15	15	16
Expression (1.39) 6th significant digit		$\overline{\gamma} = 5$ dB	$\overline{\gamma} = 10$ dB	$\overline{\gamma} = 15$ dB
$\kappa = 1$	$\mu = 1$	19	21	24
$\kappa = 2$	$\mu = 2$	23	26	28

Table 1. Number of terms that need to be summed in (1.38) and (1.39) to achieve accuracy at the specified significant digit for some values of system parameters.

7. Conclusion

Cases when wireless channels are affected by general and nonlinear fading distributions are disscused in this chapter. The analytical study of the κ-μ fading channel capacity, e.g., under the OPRA, ORA, CIFR and TIFR adaptation policies and MRC and SC diversity techniques is performed. The main contribution are closed-form expressions derived for the proposed adaptation policies and diversity techniques. Based on them, numerically obtained results are graphically presented in order to show the effects of various system parameters. Since κ-μ model as general physical fading model includes Rayleigh, Rician, and Nakagami-m fading models, as special cases, the generality and applicability of this analysis are more than obvious. Nonlinear fading scenario is discussed in the similar manner, as an analytical

study of the Weibull fading channel capacity, under the OPRA, ORA, CIFR and TIFR adaptation policies and MRC diversity technique.

8. Acknowledgment

9. References

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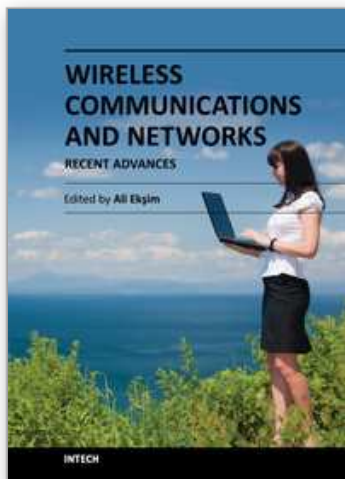
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