# Theoretical Diversity Improvement in $\mathrm{GSC}(N, L)$ Receiver With Nonidentical Fading Statistics

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***Abstract*—The study on generalized selection combining ($\mathbf{GSC}(N, L)$) diversity systems that adaptively combines a subset of $N$ paths with the highest instantaneous signal-to-noise ratios (SNRs) out of $L$ available diversity paths has both theoretical and practical importance in the design of low-complexity receiver structures for cellular wideband CDMA, indoor millimeter-wave and ultra-wideband communications. This paper presents a novel mathematical framework to tackle the problem at hand by deriving a single integral expression for the moment generating function (mgf) of the $\mathbf{GSC}(N, L)$ output SNR when the $L$ resolvable multipaths are independent with nonidentical fading statistics. The mgf is then used to unify the performance evaluation of a broad range of digital modulation/detection schemes in practical wireless channels.**

***Index Terms*—Coherent receiver, diversity methods, noncoherent (quadratic) receiver, reduced-complexity receiver structures, ultra-wideband, wideband CDMA (WCDMA).**

## I. Introduction

THE ability to capture significant amount of transmitted signal energy present in the resolvable multipaths using only a modest number of rake fingers (correlators) is an important receiver design consideration for wideband CDMA (WCDMA) and ultra-wideband (UWB) communication systems. Even though a rake receiver structure that combines all the $L$ resolvable multipaths using maximal-ratio combining (MRC) technique is optimum from the performance viewpoint, it may not be desirable for practical implementations for a number of obvious reasons. For instance, its receiver complexity is dependent on the physical channel characteristics (i.e., the channel length $L$ may vary with operating environment as well as time), and therefore undesirable. When additional factors such as channel estimation errors and performance–complexity trade-off for MRC implementation with a large number of multipaths are taken into account, it seems reasonable to combine only a few "strongest" multipaths to achieve the desired level of performance. This is the main motivation behind the development of suboptimal coherent generalized selection combining ($\mathrm{GSC}(N, L)$) technique [1] in which a subset of $N \leq L$ paths with highest signal- to-noise ratios (SNRs) are optimally weighted and summed. For noncoherent and differentially coherent communications, the need for noncoherent $\mathrm{GSC}(N, L)$ is further emphasized owing to the noncoherent combining loss phenomenon inherent in the post-detection equal-gain combining (EGC) rake receiver [2]. In this case, there exists an optimum $N$ that minimizes the error probability performance for a given average SNR/bit and $L$.

The literature on $\mathrm{GSC}(N, L)$ is quite extensive (see [3]–[11] and the references therein). However, only a few of these contributions have examined the effects of independent but nonidentically distributed (i.n.d.) fading statistics on the receiver performance. In [5], a general formula for the moment generating function (mgf) of $\mathrm{GSC}(N, L)$ output signal-to-noise ratio (SNR) is derived for the i.n.d. Rayleigh fading case. The joint probability density function (pdf) of ordered instantaneous SNRs in descending order of magnitude, derived in [1], is used for studies of the average output SNR and average error rates for a number of different modulation/detection schemes in i.n.d. Rayleigh [6], [7] and Nakagami-$m$ [9]–[11] channels. In [8], a "virtual branch" transformation technique (which can be traced back to the "spacing" method [12]) is presented to tackle the $\mathrm{GSC}(N, L)$ receiver analysis in i.n.d. Rayleigh fading. In [10], the specific cases of $\mathrm{GSC}(2, 3)$ and $\mathrm{GSC}(2, 4)$ in conjunction with binary phase shift keying (BPSK) modulation scheme are treated for i.n.d. Nakagami-$m$ fading channels. In [9] and [11], the mgf of $\mathrm{GSC}(N, L)$ output SNR in i.n.d. Nakagami-$m$ fading channels is computed via an $N$-fold nested integral. This approach, however, is not desirable for numerical computation when $N$ gets large. This motivates us to derive a general yet simple-to-evaluate formula for the mgf similar to [3], [4], but for the i.n.d. case (including the mixed-fading scenario that generalizes [5] for fading environments other than a Rayleigh channel model). Such an analysis is important in view of the practical statistical channel models [13], [14] that have been developed from the empirical data (field measurements) indicate that different multipaths in the UWB and UMTS channels have i.n.d. fading statistics.

In this paper, we develop a novel mathematical framework for analyzing both coherent and noncoherent $\mathrm{GSC}(N, L)$ receiver performance over i.n.d. generalized fading channels. The key to our solution is the transformation of a multivariate nested integral that arise in the computation of mgf of SNR into a product

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form of univariate integrals. We exploit this result to obtain computationally efficient formulas for the mgf, pdf and the cumulative distribution function (cdf) of $\mathrm{GSC}(N, L)$ output SNR, which in turn are used to characterize the average symbol error probability (ASEP) of different modulation/detection schemes, average SNR and outage probability metrics over generalized fading channels. It should be highlighted that in [15], [16], the authors have independently derived the above mgf starting from [1, eq. (6)] and borrowing the idea of recursive substitution of the marginal mgf technique (to obtain the mgf of $\mathrm{GSC}(N, L)$ output SNR which entails only a single integral for arbitrary $N$ and $L$ and for different fading channels) developed in [3], [4] for the special case of i.i.d. fading statistics. Different from [15], in this paper (also in [17], [18]) we derive the desired mgf starting from the joint pdf of order statistics given in [19], which has led to the development of two alternative expressions for this mgf (each has merits in terms of computational complexity for different choices of $N$ and $L$). Numerical results for the noncoherent $\mathrm{GSC}(N, L)$ receiver are also provided.

## II. Derivation of the mgf of $\mathrm{GSC}(N, L)$ Output SNR

Let $\alpha_k (k \in \{1, 2, \ldots, L\})$ denote the set of i.n.d. random fading amplitudes associated with $\mathrm{GSC}(N, L)$ receiver inputs. For rake reception with matched filter receiver for each diversity path, we define the instantaneous SNR/symbol of the $k$th resolvable multipath as $\gamma_k = \alpha_k^2 E_s/N_0$ where $E_s/N_0$ is the symbol energy-to-Gaussian noise spectral density ratio. The corresponding average SNR/symbol is given by $\overline{\gamma}_k = \Omega_k E_s/N_0$ where $\Omega_k = E\left[\alpha_k^2\right]$ and $E[\cdot]$ represents the expected value (statistical average) of its argument. The pdf $f_k(\cdot)$, cdf $F_k(\cdot)$ and the marginal mgf $\phi_k(s, x) = \int_x^\infty e^{-st} f_k(t) dt$ of $\gamma_k$ for several common fading channel models are summarized in [3]. These formulas will be used in the computation of the mgf of $\mathrm{GSC}(N, L)$ output SNR over generalized fading channels. Note also that $\phi_k(s, 0) = \phi_k(s)$ and $\phi_k(0, x) = 1 - F_k(x)$.

Suppose $\gamma_{1:L} \leq \gamma_{2:L} \leq \cdots \leq \gamma_{L:L}$ represent the order statistics obtained by arranging the instantaneous SNRs $\gamma_1, \gamma_2, \ldots, \gamma_L$ in increasing order of magnitude, we have $\gamma_{gsc} = \sum_{k=L-N+1}^{L} \gamma_{k:L}$ as the $\mathrm{GSC}(N, L)$ output SNR. We are particularly interested in evaluating the mgf and the cdf of $\gamma_{gsc}$ because the mean of $\mathrm{GSC}(N, L)$ output SNR, outage probability and ASEP performance metrics of a variety of digital modulation/detection schemes can be computed using these quantities alone.

In [19], it is shown that the joint pdf of order statistics $\gamma_{1:L} \leq \gamma_{2:L} \leq \cdots \leq \gamma_{L:L}$ at $x_1, x_2, \ldots, x_L$ is given by

$$f_{\gamma_{1:L}, \gamma_{2:L}, \ldots, \gamma_{L:L}}(x_1, x_2, \ldots, x_L) = \mathrm{per}\left(\begin{bmatrix} f_1(x_1) & f_2(x_1) & \ldots & f_L(x_1) \\ f_1(x_2) & f_2(x_2) & \cdots & f_L(x_2) \\ \vdots & \vdots & \vdots & \vdots \\ f_1(x_L) & f_2(x_L) & \cdots & f_L(x_L) \end{bmatrix}\right) \quad (1)$$

where $0 \leq x_1 < x_2 < \cdots < x_L < \infty$, and $\mathrm{per}(A)$ denotes the permanent of a square matrix $A$. The permanent, also known as plus determinant in the statistical literature, is computed similar to the determinant except that all signs are positive. For example, the permanent of a $2 \times 2$ matrix is given by $\mathrm{per}\left(\begin{bmatrix} a_{11} & a_{12} \\ a_{21} & a_{22} \end{bmatrix}\right) = a_{11}a_{22} + a_{12}a_{21}$.

More generally, for $n \times n$ matrix $A = [a_{ij}]$, we have [20]

$$\mathrm{per}(A) = \sum_{\sigma \in S_n} \prod_{i=1}^{n} a_{i\sigma(i)} \quad (2)$$

where $S_n$ is the set of all permutations of integers $\{1, 2, \ldots, n\}$ and $\sigma \in S_n$ denotes the specific function $\sigma = (\sigma(1), \sigma(2), \ldots, \sigma(n))$ which permutes the integers $\{1, 2, \ldots, n\}$. For example, $S_2 = \{(1, 2), (2, 1)\}$, $S_3 = \{(1, 2, 3), (1, 3, 2), (2, 1, 3), (2, 3, 1), (3, 1, 2), (3, 2, 1)\}$ and so on. The cardinality of $S_n$ is equal to $n!$. The process of constructing all members of $S_n$ is recursive and most mathematical packages have explicit commands for this purpose. For instance, $S_n$ is constructed by the command perms $([1, 2, \ldots, n])$ in MATLAB.

Combining (1) and (2), we obtain a very compact representation for the joint pdf of $\gamma_{1:L}, \gamma_{2:L}, \ldots, \gamma_{L:L}$ as

$$f_{\gamma_{1:L}, \gamma_{2:L}, \ldots, \gamma_{L:L}}(x_1, x_2, \ldots, x_L) = \sum_{\sigma \in S_L} \prod_{i=1}^{L} f_{\sigma(i)}(x_i), \; 0 \leq x_1 < x_2 < \ldots < x_L < \infty. \quad (3)$$

Recognizing that the mgf of $\mathrm{GSC}(N, L)$ output SNR, $\phi_{\gamma_{gsc}}(\cdot)$, is the key to unified analysis of many modulation/detection schemes over fading channels [21]–[23], our immediate intention will be to derive the desired mgf first. Therefore, we are interested in computing

$$\phi_{\gamma_{gsc}}(s) = E\left[\exp\left(-s \sum_{i=L-N+1}^{L} \gamma_{i:L}\right)\right], \quad s \geq 0 \quad (4)$$

which can be written as

$$\phi_{\gamma_{gsc}}(s) = \sum_{\sigma \in S_L} \int \cdots \int_{0 \leq x_1 < x_2 < \cdots < x_L < \infty} e^{-s \sum_{i=L-N+1}^{L} x_i} \times \prod_{i=1}^{L} f_{\sigma(i)}(x_i) \, dx_L dx_{L-1} \ldots dx_1. \quad (5)$$

While the unordered $\gamma_i$'s are independent, the ordered $\gamma_{i:L}$'s are not, as can be seen from (1). Therefore, the expectation operation in (4) generally involves complexity of $L!$ computations of $L$-fold nested integral as depicted in (5). However, it is possible to simplify and speed-up the computation of (5) by exploiting the integral identities (A.2) and (A.9). This directly leads to the development of two generic formulas for the mgf of $\gamma_{gsc}$ over generalized fading channels (including the mixed-fading scenario):

$$\phi_{\gamma_{gsc}}(s) = \sum_{\sigma \in T_{L,N}} \int_0^\infty e^{-sx} f_{\sigma(L-N+1)}(x) \left[\prod_{i=1}^{L-N} F_{\sigma(i)}(x)\right] \times \left[\prod_{k=L-N+2}^{L} \phi_{\sigma(k)}(s, x)\right] dx, \quad 1 \leq N \leq L \quad (6)$$

TABLE I
COMPARISON OF COMPUTATIONAL COMPLEXITY BETWEEN (6) AND (7) FOR EVALUATING THE MGF OF GSC($N, L$) OUTPUT SNR OVER I.N.D. GENERALIZED FADING CHANNELS WHEN $L = 8$ AND $L = 5$. THE METRIC OF EFFECTIVENESS IS THE NUMBER OF SUMMATION TERMS (ONE-DIMENSION INTEGRAL EVALUATIONS)

| | Total Number of Summation Terms | | | | | | | |
|---|---|---|---|---|---|---|---|---|
| | GSC(1,8) | GSC(2,8) | GSC(3,8) | GSC(4,8) | GSC(5,8) | GSC(6,8) | GSC(7,8) | GSC(8,8) |
| Eq. (6) | 8 | 56 | 168 | 280 | 280 | 168 | 56 | 8 |
| Eq. (7) | 56 | 168 | 280 | 280 | 168 | 56 | 8 | n/a |

| | Total Number of Summation Terms | | | | |
|---|---|---|---|---|---|
| | GSC(1,5) | GSC(2,5) | GSC(3,5) | GSC(4,5) | GSC(5,5) |
| Eq. (6) | 5 | 20 | 30 | 20 | 5 |
| Eq. (7) | 20 | 30 | 20 | 5 | n/a |

$$\phi_{\gamma_{gsc}}(s) = \sum_{\sigma \in T_{L,N+1}} \int_0^\infty f_{\sigma(L-N)}(x) \left[\prod_{i=1}^{L-N-1} F_{\sigma(i)}(x)\right] \times \left[\prod_{k=L-N+1}^{L} \phi_{\sigma(k)}(s,x)\right] dx, \quad 1 \le N < L \quad (7)$$

where

$$\sum_{\sigma \in T_{L,N}} = \sum_{\sigma \in S_L, \sigma(1)<\sigma(2)<\ldots<\sigma(L-N), \sigma(L-N+2)<\ldots<\sigma(L)}.$$

The construction of all permutations in the group $T_{L,N}$ is also not difficult (i.e., it can be implemented using only four command lines in MATLAB). It should be emphasized that both (6) and (7) involve only one-dimension integration (instead of $L$-dimension integration as in (5)) because the integrand can be evaluated term by term for different $x$. For the special case of $N = L$, (5) can be evaluated in closed-form using identity (A.2) as

$$\begin{aligned}\phi_{\gamma_{gsc}(s)} &= \sum_{\sigma \in S_L} \int_0^\infty e^{-sx_1} f_{\sigma(1)}(x_1) \ldots \int_{x_{L-1}}^\infty e^{-sx_L} \\ &\quad \times f_{\sigma(L)}(x_L) dx_L \ldots dx_1 \\ &= \prod_{k=1}^{L} \phi_k(s) \end{aligned} \quad (8)$$

since $\phi_k(s,0) = \phi_k(s)$. Equation (8) is in fact a well-known expression for the mgf of MRC output SNR with i.n.d. fading statistics.

Since we have two distinct general formulas for computing the mgf of $\gamma_{gsc}$, it is instructive to compare the computational complexity (or speed) associated with (6) and (7) for different combinations of $L$ and $N$ values. The complexity of (6) is mainly dictated by the time required to compute $N\binom{L}{N}$ one-dimensional integrals because the cardinality of set $T_{L,N}$ is equal to $L!/[(L-N)!(N-1)!]$. On the other hand, (7) involves the complexity of evaluation of $(L-N)\binom{L}{N}$ one-dimensional integrals (i.e., cardinality of set $T_{L,N+1}$ is equal to $L!/[(L-N-1)!N!]$).

Table I compares the computational complexity of (6) and (7) in terms of the number of summands (i.e., number of one-dimension integrals) for two different values of $L$. It is also observed that the use of (6) is most attractive for $N \le \lfloor L/2 \rfloor$ while (7) outperforms (6) when $N \ge \lceil L/2 \rceil$. Based on these findings, we suggest the following efficient implementation of $\phi_{\gamma_{gsc}}(\cdot)$ in generalized fading channels:

$$\phi_{\gamma_{gsc}}(s) = \begin{cases} \text{eq.}(6), & \text{if } 1 \le N \le \lfloor \frac{L}{2} \rfloor \\ \text{eq.}(7), & \text{if } \lceil \frac{L}{2} \rceil \le N < L \\ \text{eq.}(8), & \text{if } N = L. \end{cases} \quad (9)$$

Although (6) and (7) can be evaluated in closed-form for i.n.d. Nakagami-$m$ channels (expressed in terms of Lauricella's hypergeometric function for real $m \ge 0.5$ but can be simplified into finite polynomials for positive integer $m$), their development are omitted here for brevity and also because the resulting formulas do not provide any computational speed advantage over (9) for any $L > 3$.

It should also be pointed out that for the special case of independent and identically distributed (i.i.d) diversity paths, we have $f_{\sigma(k)}(x) = f(x)$, $F_{\sigma(k)}(x) = F(x)$ and $\phi_{\sigma(k)}(s,x) = \phi(s,x)$ for $k = 1, 2, \ldots, L$, and there are $N\binom{L}{N}$ equal terms in the sum of (6). Thus, (6) immediately reduces to

$$\phi_{\gamma_{gsc}(s)} = N\binom{L}{N} \int_0^\infty e^{-sx} f(x) \times [F(x)]^{L-N} [\phi(s,x)]^{N-1} dx, \quad 1 \le N \le L \quad (10)$$

which is indeed the previous result derived in [3, eq. (3)]. From (7), we also get a new expression for the mgf of $\gamma_{gsc}$ with i.i.d. diversity paths:

$$\phi_{\gamma_{gsc}(s)} = (L-N)\binom{L}{N} \times \int_0^\infty f(x)[F(x)]^{L-N-1}[\phi(s,x)]^N dx, \quad 1 \le N < L. \quad (11)$$

The application of (9) for outage probability analysis is discussed in Section III. In Section IV, (9) is used to facilitate ASEP analysis of both coherent and noncoherent GSC($N, L$) diversity systems over generalized fading channels.

## III. GSC($N, L$) OUTPUT QUALITY INDICATORS

Outage probability criterion and the higher-order statistics of output SNR are often used as comparative performance measures of diversity systems in wireless (fading) channels. Therefore, in this section we shall derive analytical expressions for computing the outage probability performance of GSC($N, L$)

diversity systems as well as the mean and variance of combiner output SNR. A procedure for computing several other higher-order statistics of output SNR is also presented.

### A. *Outage Probability*

The outage probability $P_{\text{out}}$ is defined as the probability that the instantaneous output SNR falls below a certain specified threshold SNR $\gamma^*$. Thus, the knowledge of the cdf of $\gamma_{gsc}$ is of interest because the outage probability can be expressed in terms of this metric alone, *viz.*,

$$P_{\text{out}} = F_{\gamma_{gsc}}(\gamma^*). \tag{12}$$

However, recognizing that only the knowledge of $\phi_{\gamma_{gsc}}(\cdot)$ is readily available (see (9)), we exploit the Laplace inversion method suggested in [24] to compute the desired cdf as

$$F_{\gamma_{gsc}}(x) \cong 2^{1-C} e^{A/2} \sum_{c=0}^{C} \binom{C}{c} \sum_{b=0}^{c+B} (-1)^b \alpha_b \times \Re\left[\frac{\phi_{\gamma_{gsc}}\left(\frac{A+j2\pi b}{2x}\right)}{(A+j2\pi b)}\right] \tag{13}$$

where $\alpha_0 = 0.5$, $\alpha_b = 1$ for any $b \geq 1$, $\Re[\cdot]$ denotes the real part of its argument and the constants $A$, $B$ and $C$ are arbitrarily chosen to be 30, 18, and 24, respectively, to yield an accuracy of at least $10^{-4}$. From (13), it is evident that the evaluation of outage probability can be performed based entirely on $\phi_{\gamma_{gsc}}(\cdot)$.

### B. *Higher-Order Statistics of* $\mathrm{GSC}(N, L)$ *Output SNR*

If the marginal density of the $k$th-order statistic $(f_{\gamma_{k:L}}(\cdot))$ is available, then one can easily derive several higher-order statistics of $\gamma_{gsc}$ (such as the mean, variance and other performance measures that are of indication of the shape and dispersion properties of the output SNR). For example, the $n$th moment of $\gamma_{gsc}$ may be evaluated as

$$\Delta_n = E\left[\gamma_{gsc}^n\right] = \sum_{k=L-N+1}^{L} \int_0^\infty x^n f_{\gamma_{k:L}}(x)dx, \quad \text{integer } n \geq 0. \tag{14}$$

Thus combining [19, eq. (7)] and (2), the marginal density the $k$th-order statistic $\gamma_{k:L}$ can be expressed neatly as

$$f_{\gamma_{k:L}}(x) = \frac{1}{(L-k)!(k-1)!} \times \sum_{\sigma \in S_L} f_{\sigma(k)}(x) \times \left[\prod_{i=1}^{k-1} F_{\sigma(i)}(x)\right] \left[\prod_{j=k+1}^{L} [1 - F_{\sigma(j)}(x)]\right]. \tag{15}$$

The above expression can be further simplified into (16) by noting that the actual order in which the subscripts occur in the two bracketed sets of (15) is irrelevant (i.e., all selections of subscripts are invariant in each of the bracketed sets):

$$f_{\gamma_{k:L}}(x) = \sum_{\substack{\sigma \in S_L, \sigma(1) < \cdots < \sigma(k-1) \\ \sigma(k+1) < \cdots < \sigma(L)}} f_{\sigma(k)}(x) \left[\prod_{i=1}^{k-1} F_{\sigma(i)}(x)\right] \times \left[\prod_{j=k+1}^{L} [1 - F_{\sigma(j)}(x)]\right]. \tag{16}$$

For the special case of i.i.d. fading statistics, (16) simplifies into the compact formula given in [25]. Substituting (16) into (14), we immediately obtain a simple-to-evaluate yet general expression for the $n$th-order moment of $\gamma_{gsc}$ over generalized fading channels, *viz.*,

$$\Delta_n = \sum_{k=L-N+1}^{L} \sum_{\sigma \in T_{L,L-k+1}} \int_0^\infty x^n f_{\sigma(k)}(x) \times \left[\prod_{i=1}^{k-1} F_{\sigma(i)}(x)\right] \left[\prod_{j=k+1}^{L} [1 - F_{\sigma(j)}(x)]\right] dx. \tag{17}$$

Now the average SNR can be evaluated as $\overline{\gamma}_{gsc} = \Delta_1$ using (17). Also the variance of output SNR is given by $\mu_2 = \Delta_2 - (\Delta_1)^2$. More generally, the central moments may be computed via relation

$$\mu_n = E[(\gamma_{gsc} - \overline{\gamma}_{gsc})^n] = \sum_{k=0}^{n} \binom{n}{k} \Delta_k (-\Delta_1)^{n-k}. \tag{18}$$

Several quality indicators of $\mathrm{GSC}(N, L)$ combiner output SNR can be readily computed with the aid of (17) and (18). For example, the dispersion of the combiner output SNR about its mean is given by its variance $\mu_2$. The skewness, which is defined as $\nu = \mu_3 / \left(\mu_2^{3/2}\right)$, is a measure of the symmetry of a distribution. For symmetric distributions, $\nu = 0$. If $\nu > 0$, the distribution is skewed to the right. Kurtosis $\kappa = \mu_4 / \left(\mu_2^2\right)$ is a measure of the "tail weight" of a distribution. Finally, the coefficient of variation is defined as $cv = (\sqrt{\mu_2})/\Delta_1$.

### C. *Computational Results and Remarks*

In the following, results for the outage probability and the mean output SNR statistics are presented to illustrate the utility of the analytical expressions derived in the preceding subsections. To perform a comparative study of diversity receiver performance over i.n.d. channels, it is plausible to introduce an exponential power decay model because a single parameter $\delta > 0$ can be used to represent the mean SNR imbalances across the diversity paths. Suppose the mean SNR of the $k$th diversity path is $\overline{\gamma}_k = Ce^{-k\delta}\overline{\gamma}_s$ where $\overline{\gamma}_s$ denotes the average SNR/symbol and the parameter $C$ is chosen such that the constraint $\sum_{k=1}^{L} \overline{\gamma}_k = \overline{\gamma}_s$ is satisfied, solving for $C$ yields

$$\overline{\gamma}_k = \frac{(1 - e^{-\delta})e^{-\delta(k-1)}}{1 - e^{-L\delta}} \overline{\gamma}_s. \tag{19}$$

The above model has also been widely used in the literature for modeling of multipath intensity profile (MIP) in frequency selective fading channels [2]. Note also that the receiver performance for i.i.d. case may be evaluated from the analytical framework for i.n.d. case by setting $\delta \to 0$ (i.e., $\delta = 10^{-5}$).

It should be emphasized, however, that our mathematical framework and analysis is applicable to any i.n.d. channel models such as those described in [13] and [14]. The results obtained for mixed fading scenarios assumes more significance by noting that measurements in UWB and UMTS/WCDMA propagation environments have indicated that the each of the resolvable multipaths have different fading statistics. Similarly for millimeter-wave communication systems where antenna arrays are employed, some of the antenna elements may receive line-of-sight signals with varying Rice factors, while others

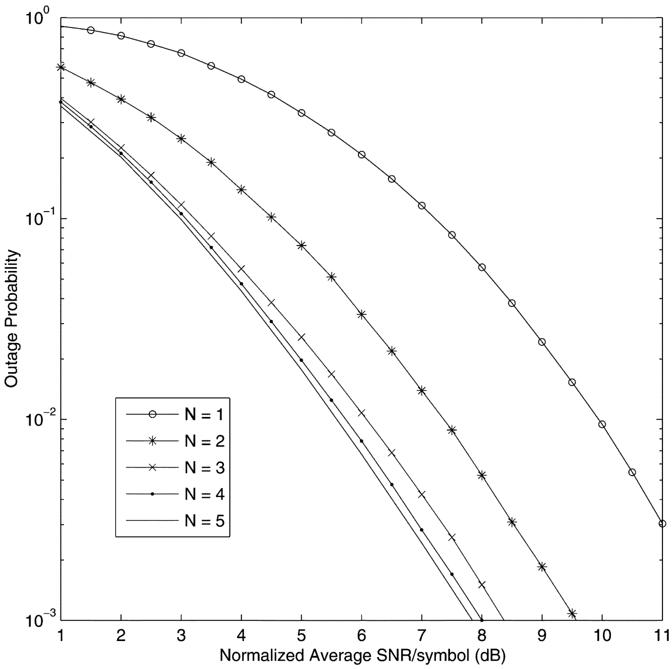


Fig. 1. Outage probability $F_{\gamma_{gsc}}(\gamma^*)$ versus the normalized average SNR/symbol $\overline{\gamma}_s/\gamma^*$ for GSC$(N, 5)$ receiver in a mixed fading $[m_1 = 0.75, m_2 = 1, K_3 = 2.5, K_4 = 1, K_5 = 0]$ and exponentially decaying MIP $(\delta = 0.1)$ environment.

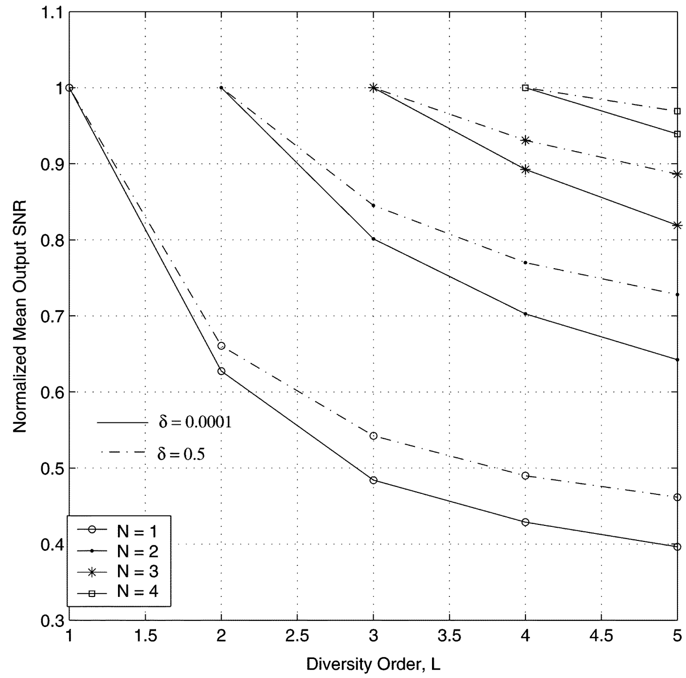


Fig. 2. Comparison of normalized average GSC$(N, L)$ output SNR versus diversity order $L$ in i.n.d. Nakagami-$m$ channels $[m_1 = 5.5, m_2 = 4, m_3 = 2.5, m_4 = 1, m_5 = 0.75]$ for two different multipath intensity profiles ($\delta = 10^{-4}$ and 0.5).

may subject to Rayleigh or Nakagami-$m$ fading because these signals may take completely different (independent) propagation paths before arriving at the receiver. Fig. 1 illustrates the outage probability metric plotted as a function of the normalized average SNR/symbol $\overline{\gamma}_s/\gamma^*$ for a $\mathrm{GSC}(N, 5)$ receiver in a mixed fading scenario (the amplitudes of the first and the second multipath are subject to Nakagami-$m$ fading with $m_1 = 0.75$ and $m_2 = 1$ while the amplitudes of the third, fourth and the fifth multipath are subject to Rice fading with Rice factors $K_3 = 2.5$, $K_4 = 1$ and $K_5 = 0$, respectively) with exponentially decaying MIP $(\delta = 0.1)$. As expected, the relative outage improvement declines with increasing $N$ and thus, the law of diminishing returns prevails. In Fig. 2, the normalized mean output SNR $\overline{\gamma}_{gsc}/\overline{\gamma}_s$ is plotted against the diversity order $L$ in i.n.d. Nakagami-$m$ channels for two different MIP. Similar analysis is not found in [11]. For a fixed $L$, the total energy captured increases with increasing values of $N$, as anticipated. For a fixed value of $N$, however, the normalized average output SNR declines with increasing $L$. The rate at which $\overline{\gamma}_{gsc}/\overline{\gamma}_s$ decreases with increasing $L$ declines as $\delta$ increases because an increase in diversity order does not significantly increase the diversity gain and/or the percentage of energy captured.

## IV. ASEP Analysis

The ASEP or the average bit error probability (ABEP) is generally obtained by averaging the conditional error probability $P_s(\gamma)$ over the pdf of $\mathrm{GSC}(N, L)$ output SNR. Since $f_{\gamma_{gsc}}(\cdot)$ is not readily available, we may apply Parseval's theorem (frequency convolution theorem) to transform the product integral into the frequency domain. Thus, we obtain [23]

$$\begin{aligned}\overline{P}_s &= \frac{1}{2\pi}\int_{-\infty}^{\infty} G_\gamma(\omega)\phi_{\gamma_{gsc}}(-j\omega)d\omega \\ &= \frac{1}{\pi}\int_0^{\infty} \Re[G_\gamma(\omega)\phi_{\gamma_{gsc}}(-j\omega)]d\omega \qquad (20)\end{aligned}$$

where $G_\gamma(\omega) = \int_0^\infty P_s(\gamma)e^{-j\omega\gamma}d\gamma$ can be evaluated in closed-form for a broad range of digital modulation/detection schemes, and they are summarized in [23, Table 2]. Moreover, if the CEP can be expressed in a "desirable" exponential form (if such representation exists), then (20) simplifies into a finite-range integral whose integrand is composed of only the mgf of $\gamma_{gsc}$ (by applying Cauchy's theorem). Two examples are provided next to highlight the utility of $\phi_{\gamma_{gsc}}(\cdot)$ in the ASEP analysis of both coherent and noncoherent $\mathrm{GSC}(N, L)$ receiver structures with nonidentical fading statistics.

### A. Coherent GSC Receiver

The ABEP of $M$-ary PSK and $M$-ary DPSK with coherent GSC receiver may be computed using [26, eqs. (42), (38), (40)] and $\phi_{\gamma_{gsc}}(\cdot)$ as defined in (9). Using this mgf approach [26], it is possible to write down ASEP and/or ABEP formulas for other digital modulation schemes, but they are omitted here for brevity. We would also like to point out that the analysis of coherent GSC is applicable to differentially coherent and noncoherent detection schemes. In determining the optimum processing of the received waveforms, an assumed lack of phase knowledge of the transmitted signal is different from lack of phase knowledge of the diversity channels. Therefore, the processed outputs of several paths can be added coherently even though the final detection-decision process is necessarily noncoherent [27]. Such an analysis is useful because it provides a lower bound on the error rates with noncoherent GSC receiver (which may be difficult to analyze for certain modulation schemes such as $M$-ary DPSK).

### B. Noncoherent GSC Receiver

Since square-law detection (also known as post-detection EGC) circumvents the need to co-phase and weight the diversity branches, the multichannel quadratic receiver has a simple implementation and suitable for use in noncoherent and differentially coherent communication systems. The ABEP of several binary and quaternary modulation schemes with noncoherent GSC receiver may be computed using [28]

$$\overline{P}_b = \frac{1}{(1+\eta)^{2N-1}2\pi} \times \int_0^{2\pi} \frac{g(\theta)}{1-2\beta\cos\theta+\beta^2}\phi_{\gamma_{gsc}} \times \left[\frac{b^2}{2}(1-2\beta\cos\theta+\beta^2)\right] d\theta \quad (21)$$

where $0^+ < \beta = a/b < 1$,

$$g(\theta) = \sum_{k=0}^{2N-1} \binom{2N-1}{k} \beta^{k+1-N}\eta^k \times \{\cos[(k-N+1)\theta] - \beta\cos[(k-N)\theta]\} \quad (22)$$

and the values for constants $(a, b)$ for binary orthogonal frequency shift keying (BFSK), binary DPSK (BDPSK) and differentially detected 4-PSK (DQPSK with Gray coding) are given by (0,1), $(0, \sqrt{2})$ and $(\sqrt{2-\sqrt{2}}, \sqrt{2+\sqrt{2}})$, respectively, and $\eta = 1$. Note that (21) assumes an indeterminate form when $\beta = 0$ for any $N > 1$ but the limit $\beta \to 0$) converge smoothly to the exact ABEP. Thus, the ABEP of BDPSK and BFSK may be computed using (21) with good accuracy by letting $\beta = 10^{-4}$ instead of zero. Alternatively, one may utilize (20) in conjunction with entry 4 in [23, Table 2].

### C. Numerical Examples

Fig. 3 illustrates the ABEP performance of BPSK with a coherent $\mathrm{GSC}(3,5)$ receiver in a mixed fading environment (which includes Rayleigh, Rice and Nakagami-$m$ fading statistics) for varying $\delta$ values. The relative diversity improvement declines with increasing $\delta$ because the channel becomes less dispersive. When $\delta$ is small, the inclusion of the third "strongest" multipath reduces the ABEP appreciably compared to a heavily decayed MIP scenario where most of the signal energy is contained only in the first or second path. Although not shown in this figure, we also found that the gap between the curves for $\mathrm{GSC}(3,5)$ and $\mathrm{GSC}(5,5)$ gets closer as $\delta$ increases. This trend further highlights the benefits of $\mathrm{GSC}(N,L)$ design in practical wireless channels.

In Fig. 4, the efficacy of $\mathrm{GSC}(N,6)$ receiver in the UMTS Vehicular A channel model is examined. It is observed that significant performance improvement over classical selection diversity can be realized by combining a few additional multipaths. In fact, $\mathrm{GSC}(4,6)$ provides performance almost identical to the MRC receiver. However, increasing $N$ beyond 4 does not significantly improve the ABEP performance of BPSK owing to the large power imbalance across the multipaths. It should be noted that the performance improvement in i.n.d. fading also depends on the fading parameter (i.e., amount of fading) for each diversity path in addition to a larger choice of $N$.

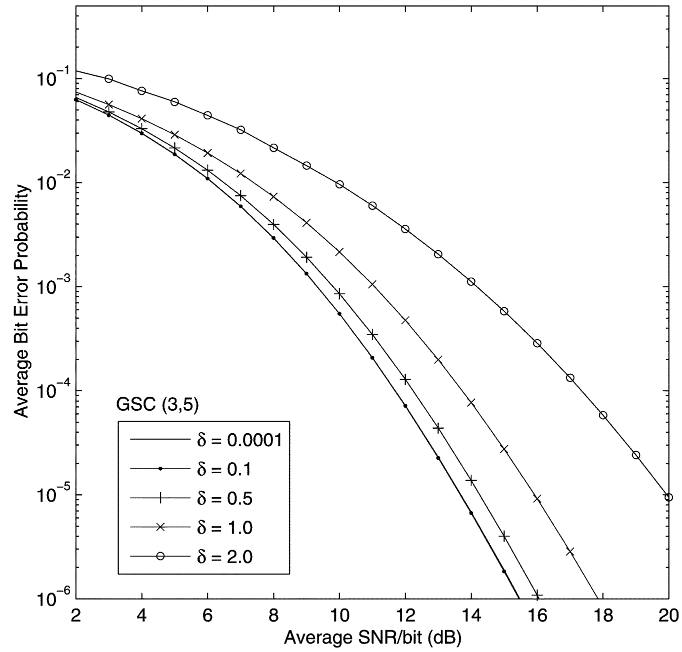


Fig. 3. ABEP performance of BPSK versus average SNR/bit for coherent $\mathrm{GSC}(3,5)$ receiver in different multipath intensity profiles. The fading parameters of the resolvable multipath signals are given as $[m_1 = 0.75,\ m_2 = 3,\ K_3 = 2.5,\ K_4 = 1,\ m_5 = 1]$.

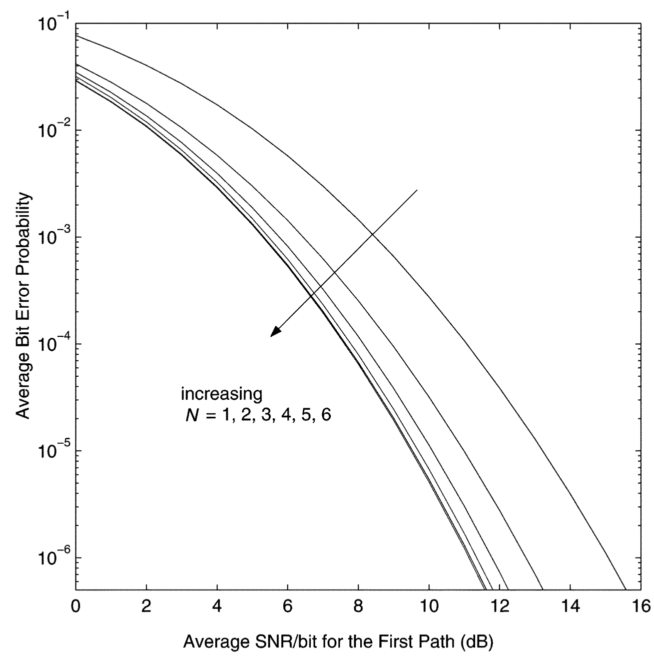


Fig. 4. ABEP performance of BPSK with a coherent $\mathrm{GSC}(N,6)$ receiver in a UMTS vehicular A channel model with relative mean powers (in dB) for the first six resolvable multipaths are given by $[0, -1, -9, -10, -15, -20]$. The fading parameters of these multipaths are assumed to be $[K_1 = 3,\ m_2 = 4,\ m_3 = 3.5,\ K_4 = 1.5,\ K_5 = 0,\ m_6 = 0.6]$.

Finally in Fig. 5, we investigate the trade-off between the diversity gain and noncoherent combining loss for DQPSK in a

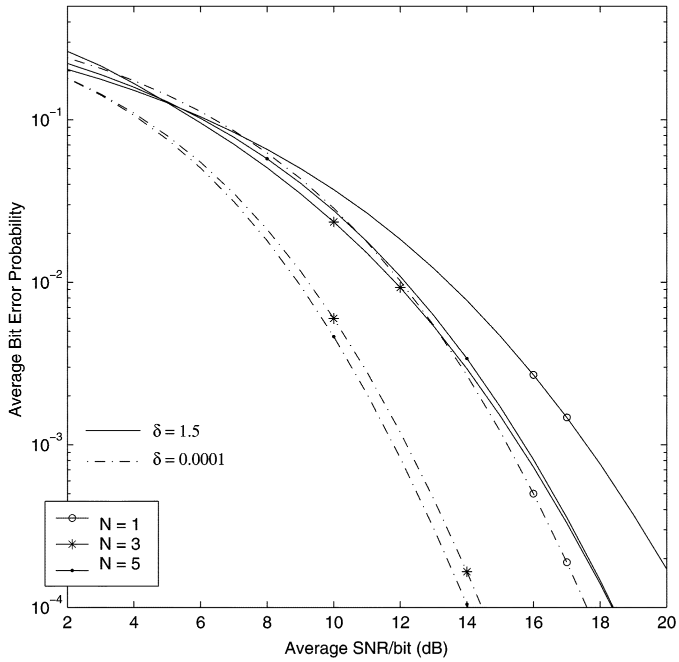


Fig. 5. Investigation into the trade-off between diversity gain and noncoherent combining loss for DQPSK in conjunction with noncoherent $\mathrm{GSC}(N, 5)$ receiver in a generalized fading channel $[m_1 = 0.75,\ m_2 = 3,\ K_3 = 2.5,\ K_4 = 1,\ m_5 = 1]$.

mixed fading environment. When $\delta = 1.5$, we observe that noncoherent $\mathrm{GSC}(3, 5)$ provides better performance than post-detection EGC $\mathrm{GSC}(5, 5)$ for average SNR/bit between 5 and 19 dB owing to the noncoherent combination loss phenomenon. Comparison between the curves corresponding to $\mathrm{GSC}(1, 5)$ reveals that statistical diversity gain has a stronger influence on the ABEP performance of DQPSK at higher average SNR/bit but the amount of energy captured is more critical at the lower average SNR/bit region.

## V. Conclusion

This paper investigates the performance of both coherent and noncoherent $\mathrm{GSC}(N, L)$ receiver over generalized fading channels. The mgf of $\gamma_{gsc}$ is used to unify the performance evaluation of different modulation/detection schemes while the outage probability performance is predicted from the cdf expression. Concise analytical formulas for the higher order moments and central moments are also derived. Our mathematical framework can be applied to the design and analysis of several wireless systems of interest such as $\mathrm{GSC}(N, L)$ rake receiver design for wideband CDMA and UWB communications, and $\mathrm{GSC}(N, L)$ antenna array design for use in millimeter-wave indoor wireless communications.

## Appendix

In this appendix, we show (using the principles of mathematical induction) that it is possible to transform two different multivariate nested integrals into a product of univariate integrals. These nested integrals arise in the computation of the mgf of $\mathrm{GSC}(N, L)$ output SNR with nonidentical fading statistics. Define $I_{N-1}(x_N)$ as

$$I_{N-1}(x_N) = \sum_{\sigma \in S_{N-1}} \int_0^{x_N} g_{\sigma(N-1)}(x_{N-1}) \dots \int_0^{x_3} g_{\sigma(2)}(x_2) \times \int_0^{x_2} g_{\sigma(1)}(x_1) dx_1 \dots dx_{N-1}, \quad N \geq 2 \quad \text{(A.1)}$$

where $g_{\sigma(k)}(\cdot)$ is an arbitrary statistical function and $G_{\sigma(k)}(y) = \int_0^y g_{\sigma(k)}(x)dx$.

We will prove (using the principles of mathematical induction) that

$$I_{N-1}(x_N) = \prod_{k=1}^{N-1} G_{\sigma(k)}(x_N). \quad \text{(A.2)}$$

For $N = 2$, we have

$$I_1(x_2) = \int_0^{x_2} g_{\sigma(1)}(x_1)dx_1 = G_{\sigma(1)}(x_2) \quad \text{(A.3)}$$

implying that (A.2) holds for $N = 2$.

For $N = 3$, we obtain (using integration by parts)

$$\begin{aligned} I_2(x_3) &= \int_0^{x_3} g_{\sigma(2)}(x_2)G_{\sigma(1)}(x_2)dx_2 \\ &\quad + \int_0^{x_3} g_{\sigma(1)}(x_2)G_{\sigma(2)}(x_2)dx_2 \\ &= 2G_{\sigma(1)}(x_3)G_{\sigma(2)}(x_3) - I_2(x_3) \end{aligned} \quad \text{(A.4)}$$

because

$$\int_0^{x_3} g_{\sigma(i)}(x_2)G_{\sigma(j)}(x_2)dx_2 = G_{\sigma(j)}(x_3)G_{\sigma(i)}(x_3) - \int_0^{x_3} G_{\sigma(i)}(x_2)g_{\sigma(j)}(x_2)dx_2. \quad \text{(A.5)}$$

Equation (A.4) implies that (A.2) also holds for $N = 3$.

Assume (A.2) holds for $N = D - 1$. This implies

$$I_{D-2}(x_{D-1}) = \prod_{k=1}^{D-2} G_{\sigma(k)}(x_{D-1}). \quad \text{(A.6)}$$

Using the definition of (A.1) and the assumption of (A.6), we can write $I_{D-1}(x_D)$ as

$$\begin{aligned} I_{D-1}(x_D) &= \sum_{\substack{\sigma \in S_{D-1} \\ \sigma(1) < \dots < \sigma(D-2)}} \int_0^{x_D} g_{\sigma(D-1)}(x_{D-1}) \\ &\quad \times I_{D-2}(x_{D-1})dx_{D-1} \\ &= \sum_{i=1}^{D-1} \int_0^{x_D} g_i(x_{D-1}) \\ &\quad \times \prod_{k=1, k\neq i}^{D-1} G_k(x_{D-1})dx_{D-1}. \end{aligned} \quad \text{(A.7)}$$

Applying integration by parts on (A.7), we get

$$I_{D-1}(x_D) = (D-1) \prod_{k=1}^{D-1} G_{\sigma(k)}(x_D) - (D-2)I_{D-1}(x_D) \quad \text{(A.8)}$$

implying that (A.2) holds for $N = D$. Therefore, by mathematical induction, (A.2) holds for all $N \geq 2$. This completes our proof.

In fact, (A.2) may also be deduced from selection diversity combining (SDC) and MRC analyzes with i.n.d. fading statistics. Let $f_{\sigma(k)}(\cdot)$ and $F_{\sigma(k)}(\cdot)$ denote the pdf and cdf of SNR of the $\sigma(k)$th diversity path, respectively. If we substitute $g_{\sigma(k)}(x_i) = f_{\sigma(k)}(x_i)$ in (A.1), we have an identical problem formulation for computing the cdf of SDC combiner output SNR with $N-1$ diversity paths. In this case, (A.2) is in agreement with the well-known result for SDC. Alternatively, if we substitute $g_{\sigma(k)}(x_i) = e^{-sx_i} f_{\sigma(k)}(x_i)$ and $x_N = \infty$ in (A.1), the resulting expression resembles the general problem formulation for computing the mgf of MRC output SNR with $N-1$ diversity paths. Once again, the accuracy of (A.2) is validated by comparison with the well-known result for MRC (see (8)).

Using the above approach, we also get another interesting integral identity

$$\begin{aligned} J_{N-1}(x_1) &= \sum_{\sigma \in S_{N-1}} \int_{x_1}^{\infty} g_{\sigma(2)}(x_2) \int_{x_2}^{\infty} g_{\sigma(3)}(x_3) \\ &\quad \cdots \int_{x_{N-1}}^{\infty} g_{\sigma(N)}(x_N) dx_N \dots dx_2 \\ &= \prod_{k=1}^{N-1} H_{\sigma(N-k+1)}(x_1), \quad N \geq 2 \end{aligned} \tag{A.9}$$

where $H_{\sigma(k)}(y) = \int_y^{\infty} g_{\sigma(k)}(x) dx$.

The accuracy of (A.2) and (A.9) have been validated numerically by letting $g_{\sigma(k)}(\cdot) = f_{\sigma(k)}(\cdot)$. We also found that the product form of $I_{N-1}(x_N)$ and $J_{N-1}(x_1)$ yields a tremendous improvement in computation efficiency over their multivariate integral counterpart, specifically for large $N \geq 4$ values.

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