Further Analysis on Selective Diversity Reception for Detection of *M*-ary Signals Over Nakagami Fading Channels

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ABSTRACT

The symbol error probability of *M*-ary PSK (MPSK) and QAM (MQAM) systems using the branch with the largest signal-to-noise ratio (SNR) at the output of *L*-branch selection combining (SC) in frequency-nonselective slow Nakagami fading channels with an additive white Gaussian noise (AWGN) is derived theoretically. For integer values of the Nakagami fading parameter *m*, the general formula for evaluating symbol error rate (SER) of MPSK signals in the independent branch diversity system comprises numerical analyses with the integral-form expressions. An exact closed-form SER performance of MQAM signals under the effect of SC diversity via numerical integration is presented. These performance evaluations allow designers to determine *M*-ary modulation methods for Nakagami fading channels.

Key Words : SC, Nakagami fading, MPSK, MQAM

Ⅰ. Introduction

In recent years, there have been increased interests in personal communication systems and wireless communications. The statistics for various fading channel models and the resulting communication evaluation have been considerably studied as summarized in [1]. The statistical properties of mobile radio environments can be often specified by the following propagation effects: 1) short-term fading 2) long-term fading [2]. In short-term fading, the scattering mechanism only results in numerous reflected components [3]. The Rayleigh model is used to characterize this fading in small geographical areas and sometimes does not account for large scale effects like shadowing by building and hills. When the fading index in the Rician model, *K*, goes to 0, the error performances lead to those of Rayleigh fading model. In long-term fading, the change of effective height for mobile communication antenna exists due to the nature of the terrain. Its statistics follow the log-normal distribution. But the Rician model can be obtained from the direct wave and its scattering components, and both waves carry information [4]. In a celluar system, Rayleigh fading is often the feature of large cells, whereas for the cells of smaller field, the envelope fluctuations of a received signal are closer to the Rician fading that is bounded by AWGN perturbations and Rayleigh fading [5]. When *M*-ary signals experience the fading channels, diversity schemes can minimize the effects of these fadings since deep fades simultaneously occur during the same time intervals on two or more paths.

In this paper, assuming that diversity branches are statistically independent with each other over Nakagami fading channels, we analyze the performance of SC diversity reception of MPSK and MQAM signals, respectively. With reference to a

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traditional SC, whereby the signal with the largest SNR is selected from the original diversity branches, the order statistics are applied [6]. Until now, the limited expressions for the error rate of *M*-ary signals have been represented using the probability density function (PDF) of the dual SC output for Rayleigh and Nakagami fading channels [7]. But we can represent the error performance of *M*-ary modulated signals in the independent, not dual branch, general *L*-branch diversity system by a finite-series formula which has been published previously. Recently, submissions of QPSK have been made to 3GPP (Third Generation Partnership Projects), whereas those of 16-QAM have been made to 3GPP2 (Third Generation Partnership Projects 2) [8], [9].

Ⅱ. System Model with SC Diversity Reception

The optimal combination of the received signals is obtained by using maximum ratio combining (MRC) which calls for the increased complexity with respect to other diversity schemes. Anyway, it is frequently considered since its performance can be assumed as the upper bound to compare suboptimal combining rules. On the other hand, among the suboptimal techniques, SC is attractive due to implementation simplicity and low cost.

The interest for SC application has been recently increased for the high-capacity mobile radio system. With reference to SC, applied here are the order statistics to select μ branches from the branch with the largest amplitude, or to choose μ branches from the branch with the largest SNR at the L diversity branches, for a data recovery while assuming that the noise power is constant across all branches when the PDF for this combining is analyzed to evaluate the error rate performance of *M*-ary signals on Nakagami fading channels [6][10].

Then the statistics of an instantaneous SNR γ are as follows:

$$\gamma=\gamma_L+\gamma_{L-1}+\gamma_{L-2}+\cdots+\gamma_{L-\mu+1},\quad L\geq\mu. \tag{1}$$

It is assumed that the statistical characteristics of diversity branches are independent each other over the Nakagami fading channels.

Now, to derive the PDF for the instantaneous SNR, it is worthwhile to note that we introduce the following joint PDF of the statistics γ_L, $\gamma_{L-1},\cdots$, and γ_1, where $0\leq\gamma_1\leq\gamma_2\leq\cdots\leq\gamma_L$ [6][10]:

$$f(\gamma_L,\gamma_{L-1},\cdots,\gamma_1)=f(\gamma_L)f(\gamma_{L-1})\cdots f(\gamma_1). \tag{2}$$

Given each integration interval for the unnecessary random variables (rvs) $\gamma_{L-\mu}$, $\gamma_{L-\mu-1},\cdots$, and γ_1, we can perform the integration of those statistics and yield the marginal joint PDF of γ_L, $\gamma_{L-1},\cdots$, and $\gamma_{L-\mu+1}$.

Next, we can write, after the transform of rvs, the PDF of γ, $\gamma_{L-1},\cdots,\gamma_{L-\mu+2}$, and $\gamma_{L-\mu+1}$ as

$$f(\gamma_L,\cdots,\gamma_{L-\mu+2},\ \gamma_{L-\mu+1})\frac{1}{|J|}= f(\gamma,\ \gamma_{L-1},\cdots,\gamma_{L-\mu+2},\ \gamma_{L-\mu+1}) \tag{3}$$

where the Jacobian of the transformation, $|J|=1$ and $0\leq\gamma_{L-\mu+1}\leq\gamma_{L-\mu+2}\leq\cdots\leq\gamma$.

It follows that upon performing the integrations with respect to $\gamma_{L-1},\cdots,\gamma_{L-\mu+2}$, and $\gamma_{L-\mu+1}$, we can obtain the PDF of γ. In a traditional SC to select the branch with the largest SNR from the original diversity branches, it can be shown that $\mu=1$.

The probability density function (PDF) of the received instantaneous SNR γ at the output of *L*-branch SC in a Nakagami fading channel is given by [11]

$$f(\gamma)_{SC}=\frac{L}{\Gamma(m)}\left(\frac{m}{\gamma_0}\right)^m\gamma^{m-1}\exp\left(-\frac{m}{\gamma_0}\gamma\right)\left[1-\exp\left(-\frac{m}{\gamma_0}\gamma\right)\sum_{i=0}^{m-1}\left(\frac{m}{\gamma_0}\gamma\right)^i\frac{1}{i!}\right]^{L-1} \tag{4}$$

with assumption that γ_0 is the average received SNR per branch. In (4), we use the SC technique to select the signal with the largest SNR from *L* diversity branches for equal fading severity.

The parameter m is defined as the fading index. By setting $m=1$, we observe that (4) reduces to a Rayleigh PDF. For values of m in the range $\frac{1}{2} \le m \le 1$, we obtain PDFs that have larger tails than a Rayleigh-distributed random variable. For values of $m>1$, the tail of the PDF decays faster than that of the Rayleigh. As $m \to \infty$, we have a channel that becomes nonfading. At the other extremes, for $m=\frac{1}{2}$ we have a one-sided Gaussian distribution [4].

Now, we can express the last term of (4) by using the binomial theorem as follows:

$$\left[1-\exp\left(-\frac{m}{\gamma_0}\gamma\right)\sum_{i=0}^{m-1}\left(\frac{m}{\gamma_0}\gamma\right)\frac{1}{i!}\right]^{L-1} = \sum_{n_0=0}^{L-1}(-1)^{n_0}\binom{L-1}{n_0}\left[\sum_{i=0}^{m-1}\left(\frac{m}{\gamma_0}\gamma\right)^i\frac{1}{i!}\right]^{n_0} \exp\left(-\frac{m}{\gamma_0}n_0\gamma\right). \tag{5}$$

Also, the square bracketed term in (5) is

$$\left[\sum_{j=0}^{m-1}\left(\frac{m}{\gamma_0}\gamma\right)^i\frac{1}{i!}\right]^{n_0} = \sum_{n_1=0}^{n_0}\sum_{n_2=0}^{n_1}\cdots\sum_{n_{m-1}=0}^{n_{m-2}}\binom{n_0}{n_1}\binom{n_1}{n_2}\cdots\binom{n_{m-2}}{n_{m-1}} \left(\frac{1}{1!}\right)^{n_1-n_2}\left(\frac{1}{2!}\right)^{n_2-n_3}\cdots\left(\frac{1}{(m-2)!}\right)^{n_{m-2}-n_{m-1}} \cdot\left(\frac{1}{(m-1)!}\right)^{n_{m-1}}\left(\frac{m}{\gamma_0}\gamma\right)^{n_1+n_2+\cdots+n_{m-1}} \tag{6}$$

Finally, from (5) and (6), the further expression for (4) can be written as [12]

$$f(\gamma)_{SC}=\frac{L}{\Gamma(m)}\left(\frac{m}{\gamma_0}\right)^m\sum_{n_0=0}^{L-1}(-1)^{n_0}\binom{L-1}{n_0} \sum_{n_1=0}^{n_0}\sum_{n_2=0}^{n_1}\sum_{n_{m-1}=0}^{n_{m-2}}\left[\prod_{j=1}^{m-1}\binom{n_{j-1}}{n_j} \cdot\left(\frac{1}{j!}\right)^{n_j-n_{j+1}}\left(\frac{m}{\gamma_0}\right)^{n_j}\right] \exp\left[-\frac{m}{\gamma_0}(1+n_0)\gamma\right]\gamma^{n+m-1} \tag{7}$$

where

$$n=n_1+n_2+\cdots+n_{m-1} \tag{8}$$

and

$$n_m=0. \tag{9}$$

Furthermore, this PDF in Nakagami fading channel can be expressed under the condition that the Nakagami fading parameter m is integer.

Ⅲ. Performance Analysis for MPSK Signals

The reliable phase information is still available by the transmittal of a pilot tone or by an data-aided estimate. Thus, it may be still valuable to evaluate the performance of MPSK with SC diversity receiver over Nakagami fading channels.

We find the symbol error probability for MPSK with SC diversity reception to be

$$P_{e,MPSK,SC} = \int_0^\infty P_{e,MPSK}\, f(\gamma)_{SC}\, d\gamma \tag{10}$$

where $P_{e,MPSK}$ is the conditional PDF with respect to γ when MPSK signals experience no fading, given by [13]

$$P_{e,MPSK}=\frac{1}{\pi}\int_{-\frac{\pi}{2}}^{\frac{\pi}{2}-\frac{\pi}{M}}\exp\left[-\gamma\sin^2\left(\frac{\pi}{M}\right)\sec^2\theta\right]d\theta. \tag{11}$$

Next, given that v is real number, substituting (7) and (11) into (10) and using the identity [14, p. 310, Eq. (3.351)]

$$\int_0^\infty x^n e^{-vx}dx=n!\,v^{-n-1}, \qquad Re\ v>0, \tag{12}$$

we find the symbol error probability under the Nakagami fading model to be

$$P_{e,MPSK,SC}=\frac{1}{\pi}\frac{L}{\Gamma(m)}\left(\frac{m}{\gamma_0}\right)^m\sum_{n_0=0}^{L-1}(-1)^{n_0}\binom{L-1}{n_0} \sum_{n_1=0}^{n_0}\sum_{n_2=0}^{n_1}\sum_{n_{m-1}=0}^{n_{m-2}}\left[\prod_{j=1}^{m-1}\binom{n_{j-1}}{n_j}\left(\frac{1}{j!}\right)^{n_j-n_{j-1}}\left(\frac{m}{\gamma_0}\right)^{n_j}\right] \cdot\int_{-\frac{\pi}{2}}^{\frac{\pi}{2}-\frac{\pi}{M}}\frac{\Gamma(n+m)}{\left[\frac{m}{\gamma_0}(1+n_0)+\sin^2\left(\frac{\pi}{M}\right)\sec^2\theta\right]^{n+m}}d\theta. \tag{13}$$

It is noted that SER for MPSK without diversity (L=1) under the Rayleigh fading(m=1) reduces to [13, Eq. (12)] in Fig.3.

Ⅳ. Performance Analysis for MQAM Signals

When MQAM signals with SC reception experience the Nakagami-m fading channel, we can derive the closed-form performance exact for $M=2^j$, when j is even, with the previous formula, (7). MQAM is, in practice, frequently used technique which requires less average transmitted power to achieve the same performance as MPSK signals.

Now, to derive the closed-form performance in a Nakagami fading channel, we introduce the exact SER in the presence of additive white Gaussian noise (AWGN) channel, represented as [15]

$$P_{e,MQAM}=\frac{2(\sqrt{M}-1)}{\sqrt{M}}erfc\left(\sqrt{\frac{\gamma}{c}}\right)-\frac{(\sqrt{M}-1)^2}{M}erfc^2\left(\sqrt{\frac{\gamma}{c}}\right) \tag{14}$$

where $c=\frac{2(M-1)}{3\log_2 M}$, given that j is even.

On the other hand, when j is odd, there is no equivalent $\sqrt{M}$-ary pulse amplitude modulation (PAM) signaling system which can be to equate a performance. In this case, if the detector bases its decisions on the optimum distance metric (maximum-likelihood criterion), the symbol error probability is tightly upper bounded by [4], [15]

$$P_{e,MQAM}\le\frac{2(\sqrt{M}-1)}{\sqrt{M}}erfc\left(\sqrt{\frac{\gamma}{c}}\right)-\frac{(\sqrt{M}-1)^2}{M}erfc^2\left(\sqrt{\frac{\gamma}{c}}\right) \tag{15}$$

where $erfc(\cdot)$ is the error function defined as

$$erfc(x)=\frac{2}{\sqrt{\pi}}\int_x^\infty e^{-t^2}dt. \tag{16}$$

Average SER for MQAM under the effect of SC diversity can be shown to be given by

$$P_{e,MQAM,SC}=\int_0^\infty P_{e,MQAM}\,f(\gamma)_{SC}\,d\gamma. \tag{17}$$

Given that j is even, it can be, from (17), assumed that

$$\frac{2(\sqrt{M}-1)}{\sqrt{M}}\int_0^\infty erfc\left(\sqrt{\frac{\gamma}{c}}\right)\exp\left[-\frac{m}{\gamma_0}(1+n_0)\gamma\right]\gamma^{n+m-1}d\gamma=J_1 \tag{18}$$

and

$$\frac{(\sqrt{M}-1)^2}{M}\int_0^\infty erfc^2\left(\sqrt{\frac{\gamma}{c}}\right)\exp\left[-\frac{m}{\gamma_0}(1+n_0)\gamma\right]\gamma^{n+m-1}d\gamma=J_2. \tag{19}$$

Given that μ, ν, and β are real numbers, using the identity [14, p. 649, Eq. (6.286.1)]

$$\int_0^\infty erfc(\beta x)e^{\mu^2x^2}x^{\nu-1}dx=\frac{\Gamma\left(\frac{\nu+1}{2}\right)}{\sqrt{\pi}\nu\beta^\nu}\,{}_2F_1\left(\frac{\nu}{2},\frac{\nu+1}{2};\frac{\nu}{2}+1;\frac{\mu^2}{\beta^2}\right),$$
$$Re\,\beta^2>Re\,\mu^2, Re\,\nu>0, \tag{20}$$

where

$${}_2F_1(a,b;c;z)=\frac{\Gamma(c)}{\Gamma(a)\Gamma(b)}\sum_{n=0}^\infty\frac{\Gamma(a+n)\Gamma(b+n)}{\Gamma(c+n)}\frac{x^n}{n!}, \tag{21}$$

it is straightforward to show that (18) is [14, p. 649, Eq. (6.286.1)] (See Appendix.)

$$J_1=\frac{2(\sqrt{M}-1)}{\sqrt{M}}\frac{c^{n+m}}{\sqrt{\pi}(n+m)}\Gamma\left(n+m+\frac{1}{2}\right){}_2F_1\left(n+m,n+m+\frac{1}{2};n+m+1;-\frac{m(1+n_0)c}{\gamma_0}\right). \tag{22}$$

By making the change of variable, integrating by parts and using identities [14, p.198, Eq. (2.667.4) and p. 649, Eq. (6.286.1)], (19) thus becomes

$$J_2=\frac{(\sqrt{M}-1)^2}{M}\Gamma(m+n)\left[\frac{\gamma_0}{m(1+n_0)}\right]^{m+n}-\frac{4(\sqrt{M}-1)^2}{\pi M}\sum_{i=0}^{m+n-1}\Gamma(m+n)c^{n+m-i-1}\cdot\left[\frac{m}{\gamma_0}(1+n_0)\right]^{-1-i}\frac{1}{2n+2m-2i-1}$$

$$\cdot\ {}_2F_1\left(n+m-i-\frac{1}{2},\ n+m-i\ ; n+m-i+\frac{1}{2}; -\frac{\gamma_0+cm(1+n_0)}{\gamma_0} \right) \tag{23}$$

Finally, the closed-form SER of MQAM with SC diversity over a Nakagami fading channel is

$$P_{e,MQAM,SC}=\frac{L}{\Gamma(m)}\left(\frac{m}{\gamma_0}\right)^m \sum_{n_0=0}^{L-1}(-1)^{n_0}\binom{L-1}{n_0}\sum_{n_1=0}^{n_0}\sum_{n_2=0}^{n_1}\cdots\sum_{n_{m-1}=0}^{n_{m-2}}\left[\prod_{j=1}^{m-1}\binom{n_{j-1}}{n_j}\left(\frac{1}{j!}\right)^{n_j-n_{j+1}}\left(\frac{m}{\gamma_0}\right)^{n_j}\right]$$
$$\cdot\left\{\frac{2(\sqrt{M}-1)}{\sqrt{M}}\frac{c^{n+m}}{\sqrt{\pi}(n+m)}\Gamma\left(n+m+\frac{1}{2}\right){}_2F_1\left(n+m,n+m+\frac{1}{2};n+m+1;-\frac{m(1+n_0)c}{\gamma_0}\right)\right.$$
$$-\frac{(\sqrt{M}-1)^2}{M}\Gamma(m+n)\left[\frac{\gamma_0}{m(1+n_0)}\right]^{m+n}+\frac{4(\sqrt{M}-1)^2}{\pi M}\sum_{i=0}^{m+n-1}\Gamma(m+n)c^{n+m-i-1}$$
$$\cdot\left[\frac{m}{\gamma_0}(1+n_0)\right]^{-1-i}\frac{1}{2n+2m-2i-1}$$
$$\left.\cdot\ {}_2F_1\left(n+m-i-\frac{1}{2},n+m-i\ ;n+m-i+\frac{1}{2};-\frac{\gamma_0+cm(1+n_0)}{\gamma_0}\right)\right\}. \tag{24}$$

When L goes to 1, which corresponds that there is no diversity channel, we can find that the evaluation for (24) is perfectly equivalent to [16, Eq. (3.15)] in Fig.3..

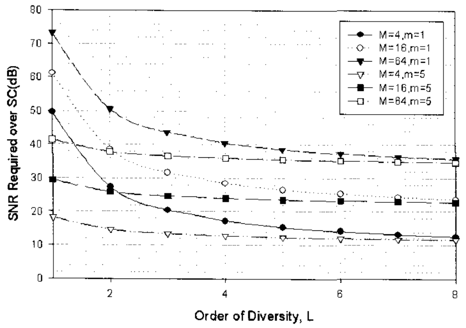

Fig. 1. Average SNR versus order of diversity for MPSK.

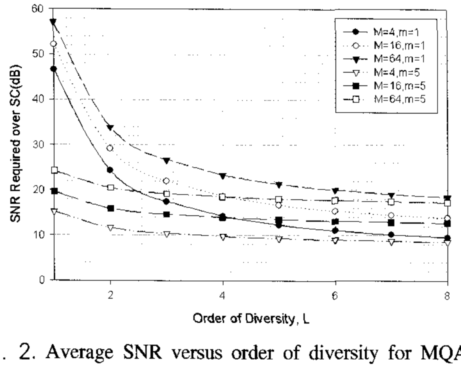

Fig. 2. Average SNR versus order of diversity for MQAM.

V. Numerical Results

Figs. 1-2 show the required SNR of a selection diversity system for the number of diversity branches, respectively. To produce these figures, the error performance of 10^{-5} is chosen. A substantial gain in the SNR per symbol required over SC for MQAM is more achieved than that for MPSK, for equal alphabet sizes *M*. It is noted that as the number of diversity branches decreases, the SNR required over SC is more deviated. But for larger the order of diversity, these curves become more horizontal approximately, for equal alphabet sizes *M*. Also, the SNR for *m*=1 is fitted more closely to that for *m*=5. So, we can show that, by decreasing the received SNR per branch, the performance of SC for *m*=1 is a little more improved than that for m=5 and otherwise, the performance in *m*=1, i.e., the Rayleigh fading channels is always worse than

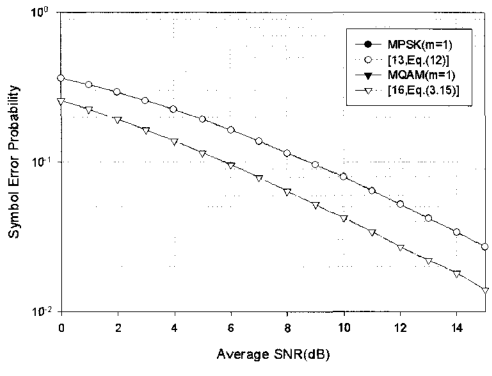

Fig. 3. Error performance comparisons of MPSK and MQAM signals adopting SC diversity technique for *M*=4 in Rayleigh fading

the performance in m=5, even when the higher order diversity is employed. We can find that the actual evaluation with increasing M does not have the important effect on the performance of M-ary signals in SC diversity system. These selected numerical results illustrate the performance of SC is improved very restrictedly in Nakagami fading conditions with increasing the diversity branches for the single value of M and m.

Ⅵ. Conclusion

An alternative solution to the problem of obtaining acceptable performances on a Nakagami fading channel is the diversity technique, which is widely used to combat the fading effects of time-variant channels. When M-ary signals experience the fading channels, diversity schemes can minimize the effects of these fadings since deep fades simultaneously occur during the same time intervals on two or more paths.

The performances for MPSK and MQAM signals under the effect of SC have been evaluated in Nakagami fading channels. Average SER formula of MPSK signals achievable by SC diversity systems has been derived in terms of integral expressions. Also an exact closed-form SER of MQAM signals under the effect of SC has been presented. These results show that the obvious performance gain is achievable with increasing the diversity branches, however the performance improvement saturates as L increases more than 4.

The results of the present works are sufficiently general in offering a convenient method to evaluate the performance of several current M-ary modulation systems using SC diversity technique on Nakagami fading channels.

Appendix: The Finite-series Representation of ${}_2F_1(\cdot)$ in (22)

Given that n is real number more than $1/2$, we assume that $G(z)$ is given by [16]

$$G(z) = z^n\,{}_2F_1(n, n+\tfrac{1}{2};\ n+1;\ -z) = \sum_{i=0}^{\infty} \frac{n}{n+i}\,\frac{(2n-1+2i)!!}{2^i(2n-1)!!}\,\frac{(-1)^i}{i!}\,z^{n+i},\quad Re\ n > \frac{1}{2} \tag{A.1}$$

where

$$(2n-1)!! = (2n-1)(2n-3)\cdots 3\cdot 1. \tag{A.2}$$

We note that since n is more than $1/2$, $G(0)$ is 0.

Then, (A.1) can be represented as follows:

$$G(z) = n\int_0^z x^{n-1}\left[\sum_{i=0}^{\infty}\frac{(2n-1+2i)!!}{2^i(2n-1)!!}\,\frac{(-x)^i}{i!}\right] dx + G(0) \tag{A.3}$$

The infinite-series formula of (A.3) becomes

$$\sum_{i=0}^{\infty}\frac{(2n-1+2i)!!}{2^i(2n-1)!!}\,\frac{(-x)^i}{i!} = (x+1)^{-n-\frac{1}{2}} \tag{A.4}$$

Hence, (A.3) can be expressed as

$$G(z) = \int_0^z n x^{n-1}(x+1)^{-\frac{2n+1}{2}}\,dx. \tag{A.5}$$

Consequently, (A.5) may be presented as follows [14, p. 73, Eq. (2.221)]:

$$G(z) = -n\sum_{i=1}^{n-1}\frac{(-1)^i}{i+\frac{1}{2}}\binom{n-1}{i}\left[(z+1)^{-i-\frac{1}{2}}-1\right] = -\sum_{i=0}^{n-1}\frac{(-1)^i}{i+\frac{1}{2}}\,\frac{\Gamma(n+1)}{\Gamma(i+1)\Gamma(n-i)}\left[(z+1)^{-i-\frac{1}{2}}-1\right] \tag{A.6}$$

Finally, ${}_2F_1(\cdot)$ in (A.1) can be written as

$${}_2F_1(n, n+\tfrac{1}{2};\ n+1;\ -z) = -z^{-n}\sum_{i=0}^{n-1}\frac{(-1)^i}{i+\frac{1}{2}}\,\frac{\Gamma(n+1)}{\Gamma(i+1)\Gamma(n-i)}\left[(z+1)^{-i-\frac{1}{2}}-1\right] \tag{A.7}$$

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