Analysis of Correlation between ICI and Desired Carrier Power in OFDM Systems over Frequency-Selective Ricean Fading Channels under the Influence of Doppler Spread

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Abstract

For data detection in orthogonal frequency division multiplexing (OFDM) systems over mobile wireless channels, the inter-carrier interference (ICI) term arising from Doppler spread is correlated with the desired carrier term. In this paper, we carry out detailed analysis on the power correlation between the ICI and the desired signal term. Explicit closed-form expression for the covariance function of the ICI and the desired carrier power is derived for frequency-selective Ricean fading channels. This expression shows that, in the normal Doppler spread range corresponding to practical vehicular speeds, the ICI power is appreciably correlated with the desired carrier power.

Key Words: Orthogonal Frequency Division Multiplexing (OFDM), Inter-Channel Interference (ICI), Ricean Fading Channels, Doppler Spread, Frequency-Selective Fading Channels

1. Introduction

It is well known that the orthogonal frequency division multiplexing (OFDM) system will loose subcarrier orthogonality when the Doppler spread is a significant portion of the subcarrier spacing (fast fading) [1,2]. Destruction of subcarrier orthogonality will result in inter-carrier or inter-channel interference (ICI) thus degrading error rate performance. ICI can also be caused by timing and carrier frequency offsets [3,4]. However, we shall here only address the ICI inflicted by Doppler spread. Unlike the additive white Gaussian noise (AWGN), the ICI noise is correlated with the desired carrier term hence also is the ICI power correlated with the desired carrier power [5]. However, in analyzing QAM signaling in uncoded OFDM systems over frequency-selective fast Rayleigh fading channels (Doppler spread is significant), Wang et al. [6] model the ICI term as uncorrelated with the desired carrier term. They show that, by using a linear approximation for the time-varying subcarrier frequency responses, the probability density function (PDF) for the ICI can be represented by a two-dimensional Gram-Charlier series. From this representation, they prove that the uncorrelated ICI approximation is justified. The uncorrelated ICI model has also been used earlier in [7] for 16-QAM OFDM error rate evaluation over frequency-selective fast Rayleigh fading channels and in [8] for BPSK OFDM error rate calculation over frequency-flat fast Ricean fading. In both [7] and [8], for the carrier-to-ICI-plus-noise ratio (CINR), the authors separately averages the desired carrier power and the ICI-plus-noise power over fading channel realizations, then takes the ratio of the former to the latter (eqs. (17), (18) in [7] and eq. (11) in [8]). This is equivalent to assuming uncorrelated ICI. However, in [5], simulation results for OFDM error rate over fast fading channels show that the
uncorrelated ICI model can only be approximated for Rayleigh fading with acceptable accuracy but not for Ricean fading especially when the specular component or line-of-sight component is large. But, [5] has not given the exact behavior of the ICI correlation or how the correlation varies with Doppler spread. This prompts a motivation for a detailed analysis on the correlation between the ICI term and the desired carrier term under the influence of Doppler spread in OFDM systems. In the analysis, we shall use the Ricean fading model (which can be readily reduced to Rayleigh fading) for generality. For the frequency-selective Ricean fading model, the first channel tap consists of a specular component plus scattered components (zero-mean complex Gaussian components), then the rest channel taps are zero-mean complex Gaussian taps (Rayleigh fading taps). Such a frequency-selective Ricean model is also adopted by the International Telecommunication Union (ITU) [9]. This model applies for communication channels between satellite and terrestrial components where satellite propagation normally includes a line-of-sight component and diffused/reflected multipath components and hence tends to be Ricean distributed with fading rates set by user and satellite motions. However, Rayleigh fading will result when the line-of-sight path is obstructed [9]. We shall derive a closed-form expression for the covariance function of the ICI and desired carrier power. This expression shows that, in the normal Doppler spread range corresponding to practical vehicular speeds, the ICI power is appreciably correlated with the desired carrier power.

This paper is organized as follows. Section 2 gives the OFDM signal model. Section 3 describes the Ricean channel model. Section 4 presents ICI analysis under the Doppler spread influence. Derivation will be given for the expression of covariance function between the ICI and desired carrier power. Section 5 gives a discussion on the difference made by the correlated and uncorrelated ICI models. Then, section 6 presents simulation results. Finally, Section 7 draws the conclusion.

2. OFDM Signal Model

For an OFDM system, N complex data symbols \( X_k \) over a time interval \( T \) constitute an OFDM symbol block, \( k = 0, 1, \ldots, N - 1 \). Thus each data symbol occupies a symbol interval \( \Delta t = T / N \). The signal bandwidth is \( 1 / \Delta t \). Data are transmitted one block at a time. Before a block is transmitted, the \( N \) symbols \( \{X_k\} \) in that block are first converted from serial to parallel form and then passed through an \( N \)-point inverse discrete Fourier transformer (IDFT) to produce \( N \) parallel complex outputs given as

\[
X_n = \frac{1}{N} \sum_{k=0}^{N-1} X_k e^{j2\pi nk/N}, \quad n = 0, 1, \ldots, N - 1
\] (1)

Thus we are viewing \( X_k \) as being in the frequency domain and \( x_n \) in the time domain. The parallel \( \{x_n\} \) are then converted back to a serial sequence over the block. Next, the serial sequence is transformed into analog form by digital-to-analog (D/A) conversion. The D/A conversion is equivalent to letting \( f_k = k / T \), and \( t = n \Delta t = nT / N \). The analog form is

\[
x(t) = \frac{1}{N} \sum_{k=0}^{N-1} X_k e^{j2\pi f_k t}, \quad 0 < t < T
\] (2)

Equation (2) only represents the initial block or zeroth block. If the \( i \)th block is spoken of, we should replace \( t \) by \( t - iT \) and the range of \( t \) should be \( iT < t < (i + 1)T \). In view of (2), \( x(t) \) can be viewed as a sum of complex subcarriers respectively at frequencies \( \{f_k = k / T\} \) with amplitudes \( X_k \), each subcarrier occupies a subcarrier width \( \Delta f = 1 / T \). With this subcarrier spacing, it is readily shown that all subcarriers are orthogonal to one another. Then, the complex baseband signal \( x(t) \) is frequency up-shifted to \( f_c \) for transmission in the channel passband [10] in the real passband form as \( s(t) = \text{Re}[x(t)e^{j2\pi f_c t}] \).

3. Frequency-Selective Ricean Fading Channel Model

Consider a frequency-selective fading channel whose equivalent lowpass discrete-time channel impulse response can be modeled by a tapped delay line [10,11]. Thus, the equivalent baseband discrete frequency response of the channel is given by

\[
H_k = \sum_{m=0}^{N-1} h_m e^{-j2\pi nk/N}, \quad k = 0, 1, \ldots, N - 1
\] (3)

where \( h_m \) is the discrete-time channel impulse response.
or tap gain and v is the channel dispersion length. If \( h_m = 0 \) for \( m \neq 0 \), we have a frequency-nonsel ective or flat fading channel. For most fading media, \( \{ h_m \} \) can be assumed spatially uncorrelated [10,12], i.e., \( E[(h_m - \overline{h}_m)(h_{m'} - \overline{h}_{m'})] = 0 \), \( m \neq m' \), where \( E[\cdot] \) denotes expectation and \( \overline{h}_m = E[h_m] \). For Ricean fading, the tap gain \( h_0 \) corresponds to the shortest path delay and contains scattered paths (zero-mean Gaussian components) plus a specular component. Thus, \( h_0 \) is a non-zero mean Gaussian random variable (RV) and the amplitude \( |h_0| \) is a Rice RV. The rest of the taps \( \{ h_m, m \neq 0 \} \) are zero mean Gaussian RVs and \( \{|h_m|, m \neq 0\} \) are Rayleigh RVs. By virtue of (3), we easily find that \( h_k \) is a complex Gaussian RV with mean \( E[h_k] = \overline{h}_k \) and variance \( V[h_k] \). For Ricean fading, the tap gain \( h_0 \) is a non-zero mean Gaussian RV and the amplitude \( |h_0| \) is a Rayleigh RV. Thus, \( h_0 \) is a non-zero mean Gaussian RV and the amplitude \( |h_0| \) is a Rayleigh RV. Therefore, the correlation between the desired carrier and the ICI is given by

\[
\sigma_v^2 = \sum_{n=0}^{\infty} \left| h(n) \right|^2
\]

where \( \sigma_v^2 \) is the variance of \( h_v \).

4. Inter-Carrier Interference Analysis

In this section, we will present an analysis of ICI for frequency-selective fast Ricean fading channels. The analysis results can be easily reduced to the Rayleigh fading results by eliminating the specular component.

4.1 Average Carrier and ICI Powers

Suppose we append a cyclic prefix of length \( G \geq v - 1 \) to each block, then the time-domain sequence should be given as

\[
x_n = \frac{1}{N} \sum_{k=0}^{N-1} x_k e^{j2\pi nk/N}, \quad -G \leq n \leq N - 1
\]

where the prefix symbols are \( x_n = x_{N+n}, n = -G, \ldots, -1 \).

Then, after frequency down-shift, analog-to-digital (A/D) conversion, and removal of the prefixed symbols at the receiver, the noiseless received discrete-time signal is given by

\[
y_n = \sum_{m=0}^{\infty} h_m(n) x_{n-m}, \quad n = 0, 1, \ldots, N - 1
\]

Note that we have now used \( h_v(n) \) to include a time-index \( n \) to account for fast fading. Also, after the removal of the prefix, only \( N \) samples of \( y_n \) are taken [10]. The convolution sum in (5) is circular, i.e., the index \( n - m \) for \( x_{n-m} \) is of modulo \( N \).

Then, the serial sequence of \( y_n \) is converted to parallel form and discrete Fourier transformed (DFT) to yield

\[
Y_k = \sum_{n=0}^{N-1} y_n e^{-j2\pi nk/N}, \quad k = 0, 1, \ldots, N - 1
\]

Using (4) and (5), we can rearrange (6) as

\[
Y_k = \sum_{m=0}^{N-1} \sum_{n=0}^{N-1} h_v(n) x_{n-m} e^{-j2\pi nk/N}
\]

\[
= \frac{1}{N} \sum_{n=0}^{N-1} \sum_{m=0}^{N-1} h_v(n) X_k e^{j2\pi nk/N} e^{-j2\pi nk/N}
\]

\[
= \frac{1}{N} \sum_{n=0}^{N-1} \sum_{m=0}^{N-1} h_v(n) X_k e^{j2\pi nk/N}
\]

The first term in (7) is the contribution from the desired \( k \)th subcarrier. The second term is ICI.

By comparing (6) and the 4th equality of (7), we see that

\[
y_n = \frac{1}{N} \sum_{k=0}^{N-1} H_k(n) X_k e^{j2\pi nk/N}
\]

\[
= \frac{1}{N} \sum_{k=0}^{N-1} \sum_{m=0}^{N-1} h_v(n) X_k e^{j2\pi nk/N}
\]

We now wish to calculate the desired carrier power and the ICI power in (7). The first tap gain of our Ricean model can be decomposed as

\[
h_v(n) = h_v'(n) + \overline{h}_v
\]

Then, \( h_v'(n) \) is a zero mean complex Gaussian RV or \( h_v'(n) \) is a Rayleigh RV. Using the classical correlation model for Rayleigh fading channels [12,13], we have

\[
E[h_v(n)h_v'(l)] = E[h_v'(n)] = |\overline{h_v'}|^2
\]

\[
= \sigma_v^2 V_0(2\pi f_M T(n-l)/N) + |\overline{h_v'}|^2
\]

where \( \sigma_v^2 = E[|h_v'|^2] \), \( f_M \) is the maximum Doppler frequency, and \( J_0(x) \) is the zero order Bessel function of the first kind. Also,
Now, for a fixed channel realization, the desired carrier power can be readily obtained from the first term of (7) as

$$P_{S,k} = \sigma_x^2 \sum_{n_1=0}^{N-1} \sum_{n_2=0}^{N-1} \mathbb{E}[|h_n(n)|^2]$$

and the ICI power from the second term of (7) as

$$P_{ICI,k} = \sigma_x^2 \sum_{n_1=0}^{N-1} \sum_{n_2=0}^{N-1} \mathbb{E}[|h_n(n)|^2]$$

where $\sigma_x^2 = \mathbb{E}[|X_0|^2]$ is the transmitted signal power. We must note here that the carrier power $P_{S,k}$ and the ICI power $P_{ICI,k}$ are correlated since they are both related to $\{h_m(n)\}$. Now, using (9) through (12) and averaging $P_{S,k}$ over all channel realizations, the $k$th average carrier power can be obtained as

$$E[P_{S,k}] = \sigma_x^2 \sum_{n_1=0}^{N-1} \sum_{n_2=0}^{N-1} \rho_{n_1-n_2}$$

$$E[P_{S,k}] = \sigma_x^2 \sum_{n_1=0}^{N-1} \sum_{n_2=0}^{N-1} \rho_{n_1-n_2}$$

where we have denoted the correlation coefficient $\rho_i = E[h_m(n)h_m((n+i))]/\sigma_m^2$ for $m = 1, 2, \ldots, N-1$ and $\rho_0 = E[h_m(n)h_m((n+i))]/\sigma_m^2$ for $m = 0$, while $\rho_0 = 1$. In arriving at the 3rd equality in (14), we have used the fact

$$N \sum_{n_1=0}^{N-1} \sum_{n_2=0}^{N-1} \rho_{n_1-n_2} = N + 2 \sum_{n=1}^{N-1} (N-n) \rho_n$$

Then, using (13) and the facts

$$\sum_{n_1=0}^{N-1} \sum_{n_2=0}^{N-1} e^{j2\pi n_1 N/n_2} = 0, \ \alpha \neq kN, \ \alpha, k : \text{integer}$$

the average ICI power can be obtained as

$$E[P_{ICI}] = E[P_{R_1}] + \sigma_x^2 \sum_{n_1=0}^{N-1} \sum_{n_2=0}^{N-1} \rho_{n_1-n_2}$$

It is interesting to find that

$$E[P_{S,k}] + E[P_{R_1}] = \sigma_x^2 (\bar{h}_0^2 + \sum_{n_1=0}^{N-1} \sigma_n^2)$$

Note that all subcarriers have the same average carrier power and average ICI power as expressions (14) and (18) are all independent of $k$. Equation (19) makes perfect sense which states that the sum of the average desired carrier power and the average ICI power equals the total average received signal power. The total received power is split between the two in accordance with (14) and (18). Note also that, when $h_0 = 0$, the above results reduce to the Rayleigh fading results. Moreover, when $\rho_i = 1$ (slow fading or zero Doppler), (18) becomes zero implying zero ICI. Then, as $\rho_i \rightarrow 0, \ i \neq 0$, (very fast fading or $f_d \rightarrow \infty$), the average carrier power and the average ICI power will both level to constants as

$$E[P_{S,k}] \rightarrow \sigma_x^2 (\bar{h}_0^2 + 1/N \sum_{n=0}^{N-1} \sigma_n^2)$$

$$E[P_{ICI,k}] \rightarrow \sigma_x^2 (\bar{h}_0^2 + 1/N \sum_{n=0}^{N-1} \sigma_n^2)$$

Note that for Rayleigh fading channels, the average carrier-to-average ICI power ratio is given by $E[P_{S,k}] / E[P_{ICI}] = 1/(N-1)$ which only depends on the OFDM block length $N$.

### 4.2 Correlation between Carrier Power and ICI Power

We wish to investigate the correlation between the ICI power and the desired carrier power. It suffices to find the covariance defined as
To obtain the covariance of (22), we must find $E[P_{SA}P_{ICL}].$ Using (12) and (13), we can obtain

$$E[P_{SA}P_{ICL}]=\frac{\sigma_x^4}{N^4}\sum_{k=0}^{N-1}\sum_{m_1=0}^{N-1}\sum_{m_2=0}^{N-1}\sum_{m_3=0}^{N-1}\sum_{m_4=0}^{N-1} e^{-j2\pi (m_1-m_2)N/N} e^{-j2\pi (m_3-m_4)N/N} \sum_{n_1=0}^{N-1}\sum_{n_2=0}^{N-1}\sum_{n_3=0}^{N-1}\sum_{n_4=0}^{N-1} E[h_{m_1}(n_1)h_{m_2}^*(n_2)h_{m_3}(n_3)h_{m_4}^*(n_4)]$$

The term $E[h_{m_1}(n_1)h_{m_2}^*(n_2)h_{m_3}(n_3)h_{m_4}^*(n_4)]$ in (23) is non-zero only when $(m_1 = m_2) \neq (m_3 = m_4), (m_1 = m_3) \neq (m_2 = m_4), (m_1 = m_4) \neq (m_2 = m_3), (m_1 = m_2 = m_3 = m_4) \neq 0,$ or $m_1 = m_2 = m_3 = m_4 = 0.$ Using two formulas derived from the moment theorem for complex Gaussian processes [14] given as

$$E[h_{m_1}(n_1)h_{m_2}^*(n_2)h_{m_3}(n_3)h_{m_4}^*(n_4)] = 0$$

then with great patience, (23) can be rearranged as

$$E[P_{SA}P_{ICL}]=\left\{\frac{\sigma_x^4}{N^2} |h_0|^2 + \frac{\sigma_x^4}{N^2} \sum_{n_1=0}^{N-1}\sum_{n_2=0}^{N-1}\sum_{m=0}^{N-1} \sum_{n_3=0}^{N-1} |\sum_{m_1=0}^{N-1} \sum_{m_2=0}^{N-1} e^{-j2\pi m_1 p /N} \sum_{m_3=0}^{N-1} \sum_{m_4=0}^{N-1} e^{-j2\pi m_2 p /N} \rho_{m_1-m_2-n_1-n_2} |^2 \right\}$$

$$=\frac{\sigma_x^4}{N^4} \sum_{n_1=0}^{N-1}\sum_{n_2=0}^{N-1}\sum_{m=0}^{N-1} e^{-j2\pi m p /N} \sum_{n_3=0}^{N-1}\sum_{n_4=0}^{N-1} e^{-j2\pi m p /N} \rho_{m_1-m_2-n_1-n_2}$$

$$=\frac{\sigma_x^4}{N^4} \sum_{n_1=0}^{N-1}\sum_{n_2=0}^{N-1}\sum_{m=0}^{N-1} e^{-j2\pi m p /N} \sum_{n_3=0}^{N-1}\sum_{n_4=0}^{N-1} e^{-j2\pi m p /N} \rho_{m_1-m_2-n_1-n_2}$$

Substituting (14), (18), and (25) into (22), we can show that

$$\text{Cov}[P_{SA}, P_{ICL}] = E[(P_{SA} - E[P_{SA}])(P_{ICL} - E[P_{ICL}])]$$

$$= E[P_{SA}P_{ICL}] - E[P_{SA}]E[P_{ICL}]$$

$$= \frac{\sigma_x^4}{N^4} \sum_{n_1=0}^{N-1}\sum_{n_2=0}^{N-1}\sum_{m=0}^{N-1} e^{-j2\pi m p /N} \sum_{n_3=0}^{N-1}\sum_{n_4=0}^{N-1} e^{-j2\pi m p /N} \rho_{m_1-m_2-n_1-n_2}$$

Equation (26) is for frequency-selective channels where $v > 1.$ For frequency-nonselective channels where $v = 1,$ the first term of (26) must be modified to be conditioned only on $m_1 - m_2 - n_4 + n_3 = 0$ as this term cannot possibly equal $\pm N.$ We also can see that (26) is independent of $k.$ If $f_{fs} \to \infty,$ thus $\rho_k \to 0,$ $i = 1, 2, \ldots, v$ (Note that $\rho_0 = 1$ always), (26) would reduce to

$$\text{Cov}[P_{SA}, P_{ICL}] = \frac{\sigma_x^4}{N^4} \sum_{n_1=0}^{N-1}\sum_{n_2=0}^{N-1}\sum_{m=0}^{N-1} e^{-j2\pi m p /N} \sum_{n_3=0}^{N-1}\sum_{n_4=0}^{N-1} e^{-j2\pi m p /N} \rho_{m_1-m_2-n_1-n_2}$$

Thus, as Doppler spread approaches infinity, the ICI power becomes uncorrelated with the desired carrier power. When Doppler spread approaches infinity, fading becomes extremely fast. As a result, each channel impulse response $h_m(n)$ becomes a white process. Hence, the $\{h_m(n)\}$ are both spatially and temporally uncorrelated.

Finally, we wish to note here that, for slow or quasi-static fading where the channel fading remains constant over at least one OFDM symbol block, then $H_k(n) \to H_k,$ and it can be readily shown that (7) simplifies to $Y_k = H_kX_k$ (by using (16)). Thus, the ICI term vanishes and hence no correlation or covariance between ICI and the carrier power will exist.

### 5. Correlated ICI Model vs. Uncorrelated ICI Model

A word on the difference between correlated ICI model and uncorrelated ICI model is in order. When including in (7) an independent, identically distributed (i.i.d.) AWGN term $Z_k$ with zero mean and variance $\sigma_Z^2,$ the received CINR for the $k$th sub-carrier under a fixed channel realization is given by

$$Y_k = \frac{P_{c,k}}{P_{ICL} + \sigma_Z^2}$$
Averaging over channel realizations, the average CINR is given by

\[ \overline{\gamma}_k = E \left[ \frac{P_{CA}}{P_{ICL} + \sigma^2_Z} \right] \]  

(29)

In [7] and [8], the ICI is assumed uncorrelated with the carrier term and is approximated by a Gaussian RV (by invoking the central limit theory for large \( N \)). The \( P_{ICI} \) in (28) is thus replaced by \( E[P_{ICI}] \). The average received CINR becomes

\[ \overline{\gamma}_k = \frac{E[P_{CA}]}{E[P_{ICL}]} + \sigma^2_Z \]  

(30)

In the next section, simulation results will show that the CINR based on correlated ICI and uncorrelated ICI as respectively given by (29) and (30) do make difference.

### 6. Simulation Results

In our simulations, we normalize the transmitted signal power \( \sigma^2_X = 1 \) and the channel power \( |h_0|^2 + \sum_{m=1}^{L} \sigma^2_m = 1 \), and assume an exponential channel power profile \( \sigma^2_m = \sigma_0^2 \exp(-\pi m T / 2\sqrt{N}) \) thus \( \sigma_0^2 = \frac{1}{K_R + \sum_{m=0}^{L} e^{-\pi m T / 2\sqrt{N}}} \),

where \( K_R = |h_0|^2 / \sigma_0^2 \) is the Ricean K-factor. Using \( N = 64 \) and \( L = N / 4 = 16 \) for a frequency-selective Ricean fading channel, it is found that the covariance function of (26) has a maximum value \( \text{Cov}[PC, P_{ICL}]_{\text{max}} \) at the normalized Doppler shift \( f_d T = 0.7 \). We can normalize the covariance by dividing it by this maximum, i.e., \( \frac{\text{Cov}[PC, P_{ICL}]}{\text{Cov}[PC, P_{ICL}]_{\text{max}}} \). This normalized covariance is plotted against \( f_d T \) in Figure 1. Note that the normalized covariance is independent of \( K_R \). This can be easily seen by observing that \( \sigma^2_m = \sigma_0^2 \exp(-\pi m T / 2\sqrt{N}) \) and that \( \sigma_0^2 \) (the only quantity that is a function of \( K_R \) in the expression of the normalized covariance) will appear both in the numerator and denominator of the normalized covariance and hence will be cancelled out. After about \( f_d T = 2 \), the normalized covariance will go down well below 0.1 and eventually decays to zero as Doppler spread goes to infinity. Taking an IEEE 802.11a standard with \( f_c = 5 \) GHz and \( \Delta f = 1.25 \) MHz, this value of \( f_d T = 2 \) corresponds to a vehicular speed about 542 km/hr. Thus for practical vehicular speeds (usually less than 542 km/hr), the Doppler spread \( f_d T \) is within 2. From Figure 1, we see that ICI is appreciably correlated with the carrier term in the region \( f_d T < 2 \).

Next, we fix the transmitted signal-to-noise ratio\( \text{SNR} = \frac{\sigma^2_X}{\sigma^2_Z} = 25 \) dB. Using the same frequency-selective Ricean fading channel as simulated for Figure 1, we present in Figure 2 plots of \( \overline{\gamma}_k \) vs. normalized Doppler spread \( f_d T \) as given by (29) and (30) for \( K_R = -\infty, 9, \) and 17 dB. (Note that log scale is adopted for normalized Doppler spread in Figure 2 while linear scale is adopted in Figure 1, thus the point for \( f_d T \rightarrow 0 \) in Figure 2 is at \( -\infty \)). The dashed curves are the CINR based on the correlated ICI model given by (29) while the solid curves are the CINR based on the uncorrelated model given by (30). Taking \( K_R = 9 \) dB in Figure 2 for a specific numerical example, the error differences between the dashed and solid curves respectively at \( f_d T = 0.1, 0.7, \) and 2 are approximately 0.5, 2, and 1 dB. Then, comparing values of covariance corresponding to the same Doppler spread points as given in Figure 1, we can readily observe that a larger covariance would result in a larger error difference. This makes perfect sense and stresses the main

![Figure 1](image-url)  

**Figure 1.** Covariance function of ICI and desired carrier power vs. normalized Doppler spread. Exponential channel power profile, \( N = 64, L = 16. \)
point of our study. All curves are the results averaged over 20,000 channel realizations using the modified Jakes model given by [15]. Note that $K_R = -\infty$ corresponds to Rayleigh fading. From Figure 2, we see that, for a given CINR, the uncorrelated model gives a smaller Doppler spread (overly optimistic) than the correlated model. The behavior at both ends of Figure 2 warrants some explanations. As indicated earlier in the previous section, when Doppler spread $f_D T \to 0$ (corresponding to $\rho_i \to 1$), the ICI power will approach zero. Thus the CINR approaches the SNR $\sigma_C^2 / \sigma_I^2 = 25$ dB. This accounts for the curve merging at left. On the other hand, as Doppler spread approaches infinity, fading becomes so fast that $\rho_i \to 0, i \neq 0$ (note again that $\rho_0 = 1$ always), then both the carrier and ICI power will level to a constant as also indicated earlier. As a result, the CINR also levels to a constant (different constant values for different $K_R$ values). This accounts for the curve merging at the right end.

7. Conclusion

The power correlation between the ICI and the desired signal term is analyzed for OFDM application in fast frequency-selective Ricean fading channels under the influence of Doppler spread. It is found that, in the normal Doppler spread range corresponding to practical vehicular speeds, appreciable correlation exists between the ICI power and the desired carrier power.

References


Hill (1966).


*Manuscript Received: May 4, 2011
Accepted: Oct. 12, 2011*