Exact Moments of Filtered Laser Phase Noise

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***Abstract*—In this paper, we derive an exact finite power series expression of the nth-order moment of a complex filtered phase noise random variable. This random variable is usually encountered in the error probability analysis of coherent heterodyne optical receivers. The result is then used to derive an infinite power series expression for the moment generating function of the same random variable. The two expressions represent a novel full statistical characterization of filtered phase noise. They also constitute an important step toward deriving optimal heterodyne receiver designs in the presence of phase noise. In a previous work by Banat (*J. Opt. Commun.*, vol. 5, no. 6, pp. 267–271, Dec. 2004), the author presented an approximate finite power series moment expression for filtered laser phase noise. The new results will be compared to those of Banat.**

***Index Terms*—Heterodyne optical receivers, laser phase noise, statistical characterization.**

I. Introduction

SEMICONDUCTOR laser phase noise is one of the major sources of performance degradation of heterodyne optical communication receivers [2]–[5]. A lot of literature exists on the design and analysis of heterodyne optical receivers. Most of this literature, however, assumes matched filter receiver models (or, equivalently, correlators or integrate-and-dump filters). These receiver designs are based on the principle of maximum likelihood and are known to be optimal in additive white Gaussian noise, i.e., in the absence of laser phase noise [6]. A brief literature review of optical receiver performance in the presence of laser phase noise can be found in [7].

Nicholson [2] and Jacobsen and Garrett [8] analyzed the differential phase-shift keying bit error rate (BER), including the effects of phase noise. In [4], the authors evaluated the phase-shift keying (PSK) BER degradation due to transmitter spectral spread caused by phase noise. Amplitude-shift keying and frequency-shift keying (FSK) phase noise performances were studied in [9] and [10], respectively. Accounting for phase noise effects in receiver performance evaluations leads to receiver decision variables that include multiplicative phase noise random factors [7], [11]. These factors are due to the application of the phase noise corrupted received signal to a matched filter. The name "filtered phase noise" will be used to refer to these random phase noise multiplicative factors. It is believed that an analytical expression for filtered phase noise moments represents a significant step toward a full analytical statistical representation of filtered phase noise. It is also believed that these can be very helpful in performance studies of heterodyne optical receivers.

Statistical characterization of filtered phase using a simulation technique that is based on a Brownian motion model was studied in [12]. A small-phase noise moment generating function (MGF) approximation was also derived in [12]. The probability density function (pdf) of the magnitude of filtered phase noise was found in [13] by numerically solving the Fokker–Planck differential equation. Analytical manipulations of Fokker–Planck-based partial differential equations were used in [14] to derive joint moments of the real and imaginary parts of filtered phase noise. Even ordered moments of the magnitude of filtered phase noise were also found in [14]. In [15] and [16], an exact analytical solution is provided for the Fokker–Plank equation. The solution provides the pdf of the phase noise-affected signal envelope in the form of an infinite series summation. A recursive formula for filtered phase noise moments was found in [17]. Other works on finding various moments of decision variables involving filtered phase noise quantities can be found in [18] and [19].

In a previous work [1], the author presented an approximate closed-form moment expression for filtered laser phase noise. In this paper, we present exact expressions of moments of filtered phase noise random variables in heterodyne optical receivers that use coherent demodulators. It is well known that knowledge of all moments of a random variable is equivalent to knowledge of its pdf. Hence, these expressions are presented as a novel full statistical characterization of filtered phase noise random variables.

The remainder of this paper is organized as follows: In Section II, we define the filtered phase noise random variables under study and give a sample application. In Section III, we derive the new moments and the MGF of filtered phase noise random variables. Some sample moments are listed in Section IV. Graphical results and comparisons are presented in Section V. Conclusions are given in Section VI.

II. Filtered Phase Noise Random Variables

Let us start by defining the complex filtered phase noise random variable

$$\eta = \frac{1}{T}\int_0^T e^{j\theta(t)}dt \tag{1}$$

where $\theta(t)$ is a laser phase noise process, and T is some integration interval. This random variable appears in the receiver error probability analysis of heterodyne optical fiber communication systems that use coherent demodulation. Specifically,

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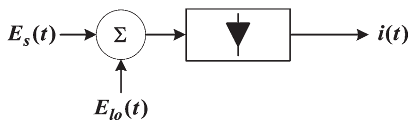

Fig. 1. Heterodyne optical receiver.

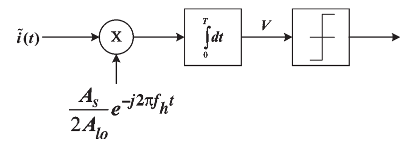

Fig. 2. OOK low-pass equivalent demodulator.

this random variable appears when the low-pass equivalent methods are applied in the derivation of receiver error probability.

As a sample application where a random variable like η can be encountered, consider a binary on–off keying (OOK) system in which the transmitted signal is given by

$$E_s(t) = \begin{cases} A_s \cos\left[2\pi f_s t + \theta_s(t)\right], & \text{“1”} \\ 0, & \text{“0”} \end{cases} \tag{2}$$

where A_s, f_s, and $\theta_s(t)$ are, respectively, the amplitude, the center frequency, and the laser phase noise process of the transmitted optical field. The received optical field is applied to the OOK heterodyne receiver shown in Fig. 1. The local laser field is given by

$$E_{\text{lo}}(t) = A_{\text{lo}} \cos\left[2\pi f_{\text{lo}} t + \theta_{\text{lo}}(t)\right] \tag{3}$$

where A_{lo}, f_{lo}, and $\theta_{\text{lo}}(t)$ are, respectively, the amplitude, the center frequency, and the laser phase noise process of this field. The detected current is usually given by

$$i(t) = LP\left\{\left|E_s(t) + E_{\text{lo}}(t)\right|^2\right\} + i_n(t) \tag{4}$$

where $LP\{\cdot\}$ denotes the low-pass equivalent, and $i_n(t)$ is a zero-mean white Gaussian noise process that is generated by the photodetector and other circuits that may be present in the receiver. Note that $i_n(t)$ will, generally, consist of shot noise and thermal noise components. However, in heterodyne systems, the amplitude of the local laser field A_{lo} is usually large enough for thermal noise to be ignorable. In other words, it can be assumed that shot noise is dominant. Substituting for the fields and simplifying, $i(t)$ can be written in the form

$$i(t) = \begin{cases} i_1(t) = \frac{A_s^2 + A_{\text{lo}}^2}{2} + A_s A_{\text{lo}} & \\ \qquad \times \cos\left[2\pi f_h t + \theta_h(t)\right] + i_{n1}(t), & \text{“1”} \\ i_0(t) = \frac{A_{\text{lo}}^2}{2} + i_{n0}(t), & \text{“0”} \end{cases} \tag{5}$$

where the heterodyne intermediate frequency $f_h = f_{\text{lo}} - f_s$, the heterodyne phase noise process $\theta_h(t) = \theta_{\text{lo}}(t) - \theta_s(t)$, and the quantities $i_{n1}(t)$ and $i_{n0}(t)$ are zero-mean white Gaussian receiver noise currents. Due to the dominance of shot noise, both $i_{n1}(t)$ and $i_{n0}(t)$ have the same variance [5]

$$\sigma_{i_{n1}}^2 = \sigma_{i_{n0}}^2 = \sigma_n^2 = \frac{A_{\text{lo}}^2}{2}. \tag{6}$$

Laser phase noise processes $\theta_s(t)$, $\theta_{\text{lo}}(t)$, and $\theta_h(t)$ are usually modeled as Wiener–Levy random processes with linewidths β_s, β_{lo}, and β_h, respectively. Therefore, any one of the three processes can be written in the form [5]

$$\theta(t) = 2\pi \int_0^t \zeta(\tau) d\tau \tag{7}$$

where $\zeta(t)$ is a zero-mean white Gaussian random process with double-sided power spectral density $\beta/2\pi$. The quantity β is the phase noise full-width at half-maximum linewidth and is equal to β_s, β_{lo}, or β_h, depending on the specific phase noise process. It can be easily found from (7) that $\theta(t)$ is a zero-mean nonstationary Gaussian random process with variance $\sigma^2 = 2\pi\beta t$. It can also be shown that $\sigma_h^2 = \sigma_s^2 + \sigma_{\text{lo}}^2$, and hence, $\beta_h^2 = \beta_s^2 + \beta_{\text{lo}}^2$. Note that subscripts h, s, and lo refer to $\theta_h(t)$, $\theta_s(t)$, and $\theta_{\text{lo}}(t)$, respectively.

To facilitate the use of the low-pass equivalent method in determining the error probability, let us consider the OOK demodulator block diagram shown in Fig. 2, where the input low-pass equivalent detected current is given by

$$\tilde{i}(t) = \begin{cases} \tilde{i}_1(t) = A_s A_{\text{lo}} e^{j[2\pi f_h t + \theta_h(t)]} + \tilde{i}_{n1}(t), & \text{“1”} \\ \tilde{i}_0(t) = \tilde{i}_{n0}(t), & \text{“0”} \end{cases} \tag{8}$$

where $\tilde{i}_{n1}(t)$ and $\tilde{i}_{n0}(t)$ are the low-pass equivalents of the bandpass parts of $i_{n1}(t)$ and $i_{n0}(t)$, respectively. Both $\tilde{i}_{n1}(t)$ and $\tilde{i}_{n0}(t)$ are zero-mean Gaussian with variances equal to A_{lo}^2. Note that the constant terms in $i(t)$ do not have a bandpass component, and hence, they do not contribute to $\tilde{i}(t)$. It can be easily shown that the decision variable V in Fig. 2 is given by

$$V = \begin{cases} m\eta_h + I_{n1}, & \text{“1”} \\ I_{n0}, & \text{“0”} \end{cases} \tag{9}$$

where I_{n1} and I_{n0} are zero-mean Gaussian random variable with variances

$$\sigma_{I_{n1}}^2 = \sigma_{I_{n0}}^2 = \sigma_{I_n}^2 = \frac{m}{2}. \tag{10}$$

The filtered phase noise random variable η_h is identical to η in (1), when $\theta_h(t)$ is substituted for $\theta(t)$. This completes our example of a typical situation where the random variables under study in this paper can appear. Examples based on PSK, quadratic-amplitude modulation, FSK, and other modulation techniques can be constructed very similarly.

III. Moments of Filtered Phase Noise Random Variables

The derivation we present in this section improves on the results of [1] in the sense that here, we use only exact relations. As will be seen in Section V, there is a visible difference between the results obtained through the exact procedure in this paper and those obtained through the approximate procedure in [1].

It was shown in [5], [7], [20], and [21] that the statistics of filtered phase noise random variables similar to the one in (1) are dependent on βT, which is the laser linewidth duration product (LDP). Hence, such statistics must also be dependent on ρ, which is the linewidth duration factor given by

$$\rho = \pi\beta T. \tag{11}$$

The modified symbols $\eta(\rho)$ and $\theta_\rho(t)$ will be used from now on to signify this dependence. The nth-order moment of $\eta(\rho)$ is given by

$$\mu_n(\rho) = E\left[\eta^n(\rho)\right] = E\left[\left\{\frac{1}{T}\int_0^T e^{j\theta_\rho(t)}dt\right\}^n\right]. \tag{12}$$

Note that the nth power of the integral in (12) can be written in the form of an n-fold integral as follows:

$$\mu_n(\rho)=E\left[\frac{1}{T^n}\int_0^T e^{j\theta_\rho(t_1)}\int_0^T e^{j\theta_\rho(t_2)}\cdots\int_0^T e^{j\theta_\rho(t_n)}dt_n\cdots dt_2dt_1\right]. \tag{13}$$

The integration in (13) is performed over the hypervolume spanned by $0 \le t_1, t_2, \ldots, t_n \le T$. Note that this hypervolume can be partitioned into $n!$ subhypervolumes, where a typical one of which is defined by $0 \le t_n \le t_{n-1}, 0 \le t_{n-1} \le t_{n-2}, \ldots, 0 \le t_1 \le T$. All such $n!$ subhypervolumes can be obtained by reordering the variables $t_1, t_2, \ldots, t_n$. Since the integrand has the same form in all variables, it should be possible to evaluate the integration over only one of the subhypervolumes and multiply the result by $n!$. This results in

$$\mu_n(\rho)=E\left[\frac{n!}{T^n}\int_0^T e^{j\theta_\rho(t_1)}\int_0^{t_1} e^{j\theta_\rho(t_2)}\cdots\int_0^{t_{n-1}} e^{j\theta_\rho(t_n)}dt_n\cdots dt_2dt_1\right]. \tag{14}$$

Making use of

$$E\left[e^{j\theta_\rho(t)}\right] = e^{-\pi\beta t} \tag{15}$$

and successively carrying out the integration process in (14), we obtain

$$\mu_n(\rho) = \frac{n!}{T^n}\int_0^T\int_0^{t_1}\cdots\int_0^{t_{n-1}} e^{-\pi\beta\sum_{k=1}^{n} a_k t_k} dt_n\cdots dt_2dt_1 \tag{16}$$

where

$$a_k = 2k - 1. \tag{17}$$

The structure in (16) allows for the integration to be evaluated as a recursion by noting that we can write

$$\mu_n(\rho) = g_n(\rho, T) \tag{18}$$

where

$$g_n(\rho, t_0) = \frac{n}{T}\int_0^{t_0} e^{-\pi\beta a_1 t_1} g_{n-1}(\rho, t_1)dt_1 \tag{19}$$

and, in general

$$g_i(\rho, t_{n-i}) = \frac{i}{T}\int_0^{t_{n-i}} e^{-\pi\beta a_{n-i+1} t_{n-i+1}} g_{i-1}(\rho, t_{n-i+1})dt_{n-i+1}. \tag{20}$$

Note that

$$g_0(\rho, t_n) = 1. \tag{21}$$

Laplace-transforming (20) over the variable t_{n-i}, we obtain

$$G_i(\rho, s) = \frac{i}{sT}G_{i-1}(\rho, s + \pi\beta a_{n-i+1}). \tag{22}$$

This can be repeated for all values of i until $G_n(\rho, s)$ is obtained to be equal to

$$G_n(\rho, s) = \frac{n!}{T^n}\prod_{m=0}^{n}\frac{1}{s + b_m} \tag{23}$$

where

$$b_m = \pi\beta m^2. \tag{24}$$

Applying the inverse Laplace transform to (23) with respect to the variable s, we obtain

$$g_n(\rho, t) = \frac{n!}{T^n}\sum_{m=0}^{n} c_{n,m} e^{-m^2\pi\beta t} \tag{25}$$

where

$$c_{n,m} = \frac{1}{(\pi\beta)^n \prod_{\substack{k=0\\k\ne m}}^{n}(k^2 - m^2)}. \tag{26}$$

Finally, using (18) to substitute T for t, we can write an expression for the nth-order moment of $\eta(\rho)$ in the form

$$\mu_n(\rho) = \frac{n!}{\rho^n}\sum_{m=0}^{n} d_{n,m} e^{-m^2\rho} \tag{27}$$

where

$$d_{n,m} = \frac{1}{\prod_{\substack{k=0\\k\ne m}}^{n}(k^2 - m^2)} = \frac{(-1)^m 2}{(n-m)!(n+m)!}. \tag{28}$$

Based on this result, the MGF of $\eta(\rho)$ can be written as an infinite power series in terms of the moments of $\eta(\rho)$. The MGF may, therefore, be expressed as

$$\Psi_\eta(s) = E\left[e^{s\eta(\rho)}\right] = \sum_{n=0}^{\infty} \frac{\mu_n(\rho)}{n!} s^n. \tag{29}$$

Now, substituting for $\mu_n(\rho)$ using (27), we obtain

$$\Psi_\rho(s) = \sum_{n=0}^{\infty} \left(\frac{s}{\rho}\right)^n \sum_{m=0}^{n} d_{n,m} e^{-m^2\rho}. \tag{30}$$

IV. Some Sample Moments

The first moment of $\eta(\rho)$ can be found by substituting $n = 1$ in (27) to obtain

$$\mu_1(\rho) = \frac{1}{\rho}(1 - e^{-\rho}). \tag{31}$$

Note that this is exactly the same expression that can be found by direct evaluation of the mean value of $\eta(\rho)$ using (1) [1], [5], [20], [21]. Substituting $n = 2$ in (27) shows that the second moment of $\eta(\rho)$ is equal to

$$\mu_2(\rho) = \frac{1}{6\rho^2}(3 - 4e^{-\rho} + e^{-4\rho}). \tag{32}$$

This is identical to the expression in [1]. The third moment of $\eta(\rho)$, which is obtained from (27), is

$$\mu_3(\rho) = \frac{1}{60\rho^3}(10 - 15e^{-\rho} + 6e^{-4\rho} - e^{-9\rho}). \tag{33}$$

This is somehow different from the same moment that was derived in [1], which is given by

$$\mu_{3,\text{approx}}(\rho) = \frac{1}{72\rho^3}(16 - 27e^{-\rho} + 12e^{-3\rho} - e^{-9\rho}). \tag{34}$$

The difference between the two results is not quite substantial. This will be illustrated graphically in Section V.

V. Results

The nth-order moment of $\eta(\rho)$ has been plotted in Fig. 3 as function of n for several values of the LDP. Note that the curves have been presented as connected lines in order to make visible the general trend in these curves as n is varied. The data points are indicated using markers. Obviously, the moments get smaller as the LDP gets larger. This observation is extremely important because it is well known that $\eta(\rho)$ (or a related quantity) usually multiplies the signal term at the input of the symbol (or bit) decision circuitry in the receiver; see (9) for an example. Hence, a larger LDP results in smaller multipliers, leading to more frequent errors in the decision process. Furthermore, and because of this trend, if the moments are to be used in calculating the MGF of $\eta(\rho)$, then fewer moments are needed as the LDP gets larger. Finally, all curves in Fig. 3 are approximately linear curves. Noting that the vertical axis is logarithmic leads to a general conclusion that the dependence of the moments on n has a decaying exponential nature.

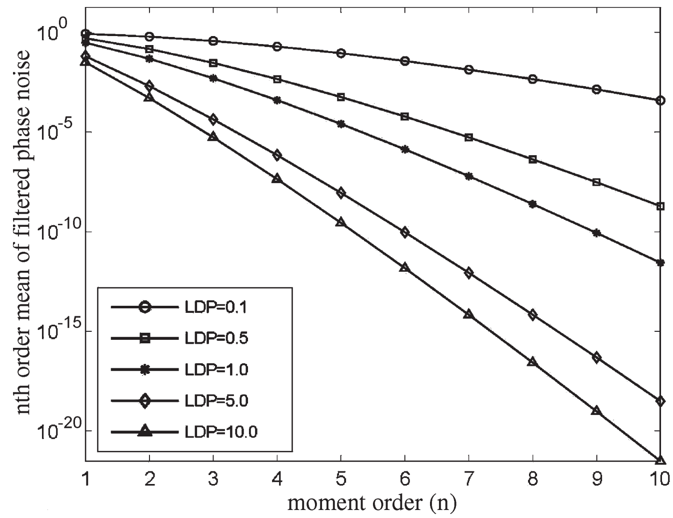

Fig. 3. nth-order mean of filtered phase noise for several values of the LDP.

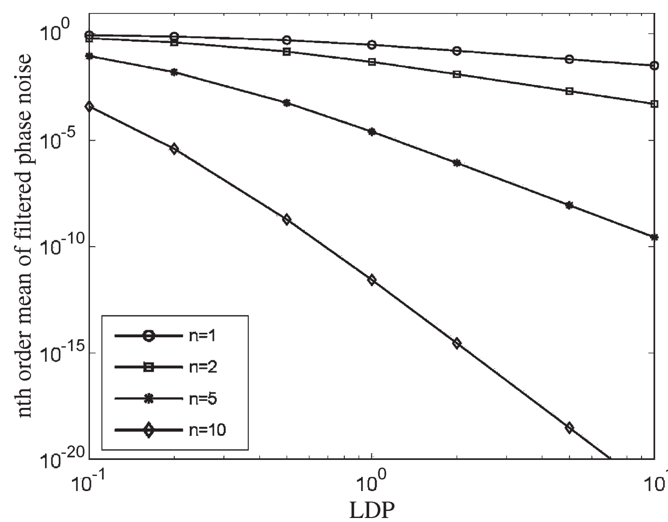

Fig. 4. nth-order mean of filtered phase noise as function of the LDP for several values of n.

In Fig. 4, we plot the nth-order moment of $\eta(\rho)$ as a function of the LDP for several values of n. As was the case above, moments are seen to decrease with increasing n or LDP. However, exponential dependence of the moments on the LDP is somewhat less obvious than that on n.

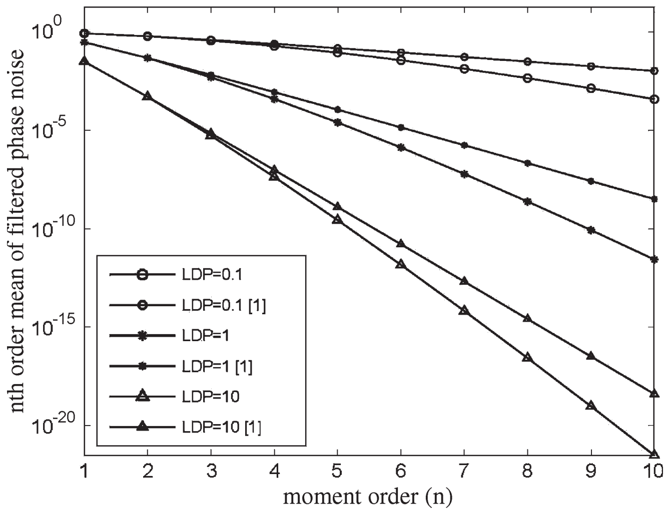

Fig. 5. nth-order mean of filtered phase noise for several values of the LDP. Comparison with the results of [1].

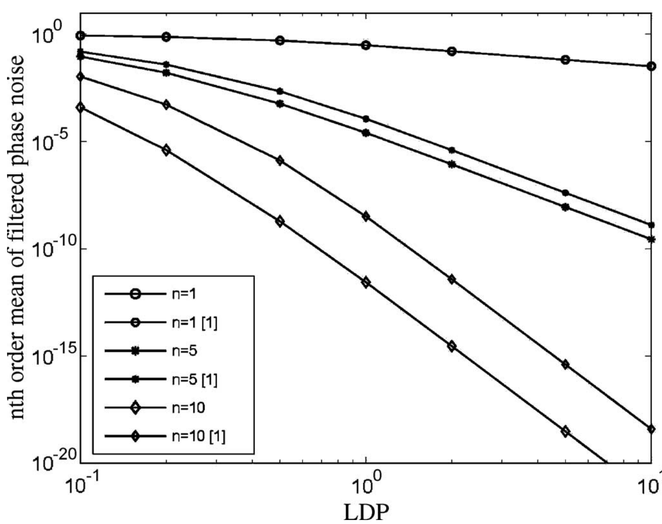

Fig. 6. nth-order mean of filtered phase noise as function of the LDP for several values of n. Comparison with the results of [1].

In Figs. 5 and 6, we compare the exact moments as given by (27) with those found through the approximate method in [1]. It can be easily seen from these two figures that there is a visible difference between the two methods for moments other than the first. Specifically, it is clear that the approximate method always produces higher moments than the exact method. Based on our discussions of Figs. 3 and 4 above, this is a form of underestimation of the effect of phase noise multiplying factors in optical receivers. This is certainly a good justification for the use of the new exact method rather than approximate methods in as many situations as possible.

VI. Conclusion

This paper has presented a new exact finite power series expression of the nth-order moment of a complex filtered phase noise random variable that is encountered in the error probability analysis of coherent heterodyne optical receivers. The result was used to derive an infinite power series expression for the MGF of the same random variable. The two expressions represent a novel full statistical characterization of filtered phase noise. Various moments calculated using the new expression were plotted and compared to corresponding moments calculated using an approximate method published earlier by the author. Comparison results demonstrate visible differences between the two methods, emphasizing the need for the new exact method.

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