Impact of Interferometric Noise on the Remote Delivery of Optically Generated Millimeter-Wave Signals

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***Abstract*— In this paper, we report for the first time results relating to reflections/multipath induced interferometric noise in millimeter-wave fiber-radio systems for the broadcast of very narrow linewidth wave signals. We use a rigorous formulation based on the modified Chernoff bound which provides an accurate upper bound on the bit-error rate (BER) and power penalty (PP). Simulated results show good agreement with the analytical findings. We conclude that interferometric noise (IN) can be a significant impairment in systems of this type.**

***Index Terms*— Microwave generation, millimeter-wave generation, millimeter-wave radio communication, noise, optical fiber communication.**

I. Introduction

RECENTLY, the applicability of a novel optical method for generation and transport of millimeter-wave signals for distribution to more than 1000 millimeter-wave radio stations via optical fiber has been demonstrated [1]. This new method is based on the mixing of two optical carriers, generated from a single laser using a Mach–Zhender modulator on a pin-diode [2]. The frequencies of the two carriers are displaced by the required millimeter-wave frequency. The two carriers are separated by an optical filter following their generation; one of them is modulated with the subcarrier and then both signals are sent via a fiber optical-distribution network to the receiver antenna units. There, optical heterodyning takes place to produce the wanted millimeter-leave frequency, which after amplification and filtering, is radiated. Due to the high degree of coherence between the two carriers, this method has proven to provide millimeter-wave signals with linewidths of much less than 1 kHz [2].

Interferometric noise (IN) can, however, be a major impairment constituting a severe limitation in the scaling capability of a network and, therefore, imposes very strict component requirements [3].

IN can appear in an optical system when the received signal is accompanied by weak delayed replicas of itself or other lightwave components, which are displaced in frequency from the signal by less than the receiver bandwidth. Several factors, such as reflections in splices, and leakage signals in routers can create crosstalk which gives rise to IN [4], [5].

In this paper, we analyze incoherent IN arising in fiber/radio networks in terms of bit-error rate (BER) and power penalty (PP), which enables us to assess the required crosstalk isolation for the various system components in such networks. We emphasize the novelty of this analysis which differs from other discussions of IN in optical networks because of the coherent heterodyne method of millimeter-wave signal generation. We also present results from simulations of a fiber-radio system which characterize the IN in terms of eye closure and the variance of the noise power. The simulation tool is modular and reconfigurable to give various combinations of reflections and/or multiple signal paths in the network.

II. Theoretical Analysis

The analysis considers the characteristics of the wanted optical carriers received together with unwanted delayed replicas originating from reflections. The polarization states of all the interfering terms are taken to be aligned with the carrier polarization state. This is a worse-case assumption, which has been proven to be dominant in optical systems [4].

The case of a carrier directly modulated with random digital-binary data is considered with the reception of the millimeter-wave signal following transmission being based on coherent detection.

The envelope of the total optical-electrical field (signal and crosstalk) arriving at the photodetector can be described by

$$\begin{aligned}E(t) = &Ee^{j\left(\Omega-\frac{\omega}{2}\right)t+j\phi(t)} + m(t)Ee^{j\left(\Omega+\frac{\omega}{2}\right)t+j\phi(t)}\\ &+ E\sum_{k=1}^{N}\sqrt{\epsilon_k}m(t-\tau_k)e^{j\left(\Omega+\frac{\omega}{2}\right)(t-\tau_k)+j\phi(t-\tau_k)}\\ &+ E\sum_{k=1}^{N}\sqrt{\epsilon_k}e^{j\left(\Omega-\frac{\omega}{2}\right)(t-\tau_k)+j\phi(t-\tau_k)}\end{aligned} \quad (1)$$

where

$$m(t) = \sum_{l=-\infty}^{\infty} a_l p(t-lT) \quad (2)$$

where E is the electrical optical-signal amplitude, ω is the millimeter-wave angular frequency, a_l is the random binary data; $a_l \in \{a, 0\}$ with both events having equal probability of occurrence, $p(t)$ is the digital data-pulse shape, Ω is the

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angular frequency of the laser diode, $\phi(t)$ is the laser phase noise, τ_k represent the time delays of the interferers, N is the number of interferers, and ϵ_k is the crosstalk isolation as a ratio between the power of the kth interferer and the power of the wanted signal.

The photodetected signal for a normalized receiver responsivity is $i(t) = |E(t)|^2$. Following some algebra and taking into account that the remote-antenna unit-band pass filter rejects all signals but those centered around the millimeter-wave frequency, ω, and neglecting beat terms between reflected components, we get

$$i(t) = E^2 m(t)\left\{\cos[\omega \mathrm{t}] + \sum_{\mathrm{k=1}}^{\mathrm{N}} \sqrt{\epsilon_{\mathrm{k}}}\cos[\omega \mathrm{t} + \Phi_{\mathrm{k}}]\right\} + E^2 \sum_{m=1}^{N} \sqrt{\epsilon_m} m(t-\tau_m)\cos[\omega \mathrm{t} + \Phi_{\mathrm{m}}] \quad (3)$$

where $\Phi_k = \left(\frac{\omega}{2} + \Omega\right)\tau_k + \phi(t) - \phi(t-\tau_k)$ and $\Phi_m = -\left(\frac{\omega}{2} + \Omega\right)\tau_m - \phi(t) + \phi(t-\tau_m)$, and both are independent identically distributed random variables described by a uniform distribution since we consider the delays τ_k much greater than the laser coherence time [4]. The first term of (3) is the desired signal, the second is an IN term resulting from the beating between the modulated carrier and the reflections of the nonmodulated carrier, and the third term is an IN term resulting from the beating between the nonmodulated carrier and reflections of the modulated carrier. From (3), we can observe that the beating noise terms convert the phase noise into amplitude IN.

At the input of the decision circuit of the receiver we get

$$\begin{aligned} i_o(t_s) &= E^2 a_0 + \mathrm{IN} + n(t_s) \\ \mathrm{IN} &= E^2 a_0 \sum_{k=1}^{N} \sqrt{\epsilon_k}\cos(\Phi_{\mathrm{k}}) \\ &= E^2 \sum_{k=1}^{N} \sqrt{\epsilon_k} a_k \cos(\Phi_{\mathrm{k}}) \end{aligned} \quad (4)$$

where t_s is the sampling time, a_0 is the binary random variable of the wanted signal and a_k, $k \in \{1, 2, \cdots, N\}$ are the binary random variables of the noise terms arising from the beating between the reflections of the modulated carrier with the nonmodulated carrier. $n(t_s)$ is the additive Gaussian receiver noise.

Here we adopt the moment generating function (MGF) approach to capture the statistics of the random variable which describes the IN. It can shown that the MGF's for the signal plus IN, for a transmitted "1" and a transmitted "0," can be expressed as follows (see also Appendix):

$$M_0(s) = \prod_{k=1}^{N}\left[\frac{1}{2} I_0(E^2 \sqrt{\epsilon_k} as) + \frac{1}{2}\right] \quad (5)$$

$$M_1(s) = M_0(s) \prod_{k=1}^{N} I_0(E^2 \sqrt{\epsilon_k} as) e^{E^2 as}. \quad (6)$$

$I_0(\cdot)$ is the modified Bessel function of the first kind of order zero. The modified Chernoff bound (MCB) can then be used to provide a tight upper bound on the BER [6] as follows:

$$\begin{aligned} \mathrm{BER} &\leq \mathrm{MCB(s)} \\ &= \frac{M_n(s)}{2\sqrt{2\pi} s \sigma_n}[M_0(s)e^{-sD} + M_1(-s)e^{sD}], \qquad s > 0 \end{aligned} \quad (7)$$

where D is the decision threshold, $M_n(s)$ is the MGF of the additive Gaussian receiver noise [6]. For the remainder of this paper, we assume D is 0.5. It should be noted that this is not necessarily the optimum threshold value [3]. σ_n^2 represents the power of the Gaussian receiver noise.

An alternative simplified representation treats each IN random variable as Gaussian, adding the IN variance to that of the receiver noise to provide a Gaussian approximation (GA) to the BER as follows:

$$\mathrm{BER} = \frac{1}{2}\mathrm{Q}\left[\frac{\mathrm{E^2 a - D}}{\sqrt{\sigma_{\mathrm{n}}^2 + \sigma_1^2}}\right] + \frac{1}{2}\mathrm{Q}\left[\frac{\mathrm{D}}{\sqrt{\sigma_{\mathrm{n}}^2 + \sigma_0^2}}\right] \quad (8)$$

where

$$\sigma_1^2 = \frac{3a^2E^4}{4}\sum_{k=1}^{N}\epsilon_k \quad \sigma_0^2 = \frac{a^2E^4}{4}\sum_{k=1}^{N}\epsilon_k \quad (9)$$

and σ_1^2 and σ_0^2 are the power of the IN on "1"'s and "0"'s, respectively. $Q(\cdot)$ represents the error function. Note that the GA uses only minimal information concerning IN statistics (just the variance). We can, by considering the central limit theorem, reasonably expect this to provide a good approximation for N sufficiently large, but for a small number of interfering terms this model is obviously inadequate since the arcsine probability density function (PDF) is bounded, whereas the Gaussian is not. In support of the analysis, simulations of a fiber-radio network were carried out with a view to replicating the predicted noise characteristics. A block-diagram representation of the system is shown in Fig. 1.

III. Results and Discussion

Results from the simulations are shown in Figs. 2–4. Fig. 2 is the normalized received-eye pattern for a signal detected along with a single-26-dB interferer, illustrating the interferometric conversion of phase noise to amplitude fluctuations. Fig. 3 shows the IN standard deviation as a function of number of interferers for received "1"'s and "0"'s. The plot shows simulated results and the behavior predicted by the analysis (9). Equal distribution of power between the interferers is assumed, that is, $\epsilon_k = -26$ dB $\forall k \in \{1, \cdots, N\}$. We note good agreement between the simulated and analyzed cases.

Fig. 4 shows the BER variation with crosstalk isolation given by the two approaches described above. Again, we assume equal distribution of power between the interferers. We consider an optical receiver which provides a $\mathrm{BER} = 10^{-9}$ in the absence of IN. From this figure, it is clear that for small N the GA overestimates the BER and the crosstalk isolation tolerance as expected. For example, for $N = 1$ the GA can overestimate the crosstalk isolation tolerance by up

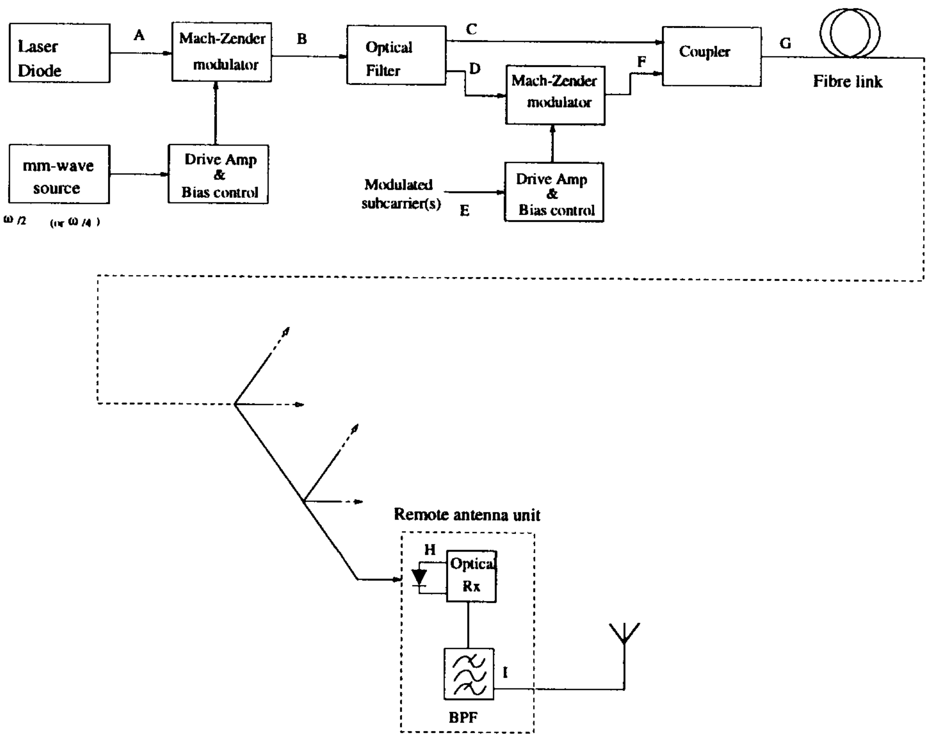

Fig. 1. Block diagram of the simulated fiber/radio network.

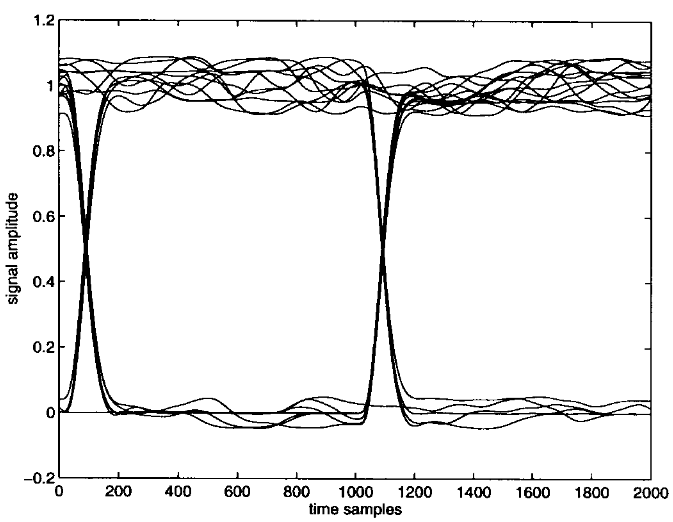

Fig. 2. Simulated eye-diagram for $N = 1$, $\varepsilon = -26$ dB.

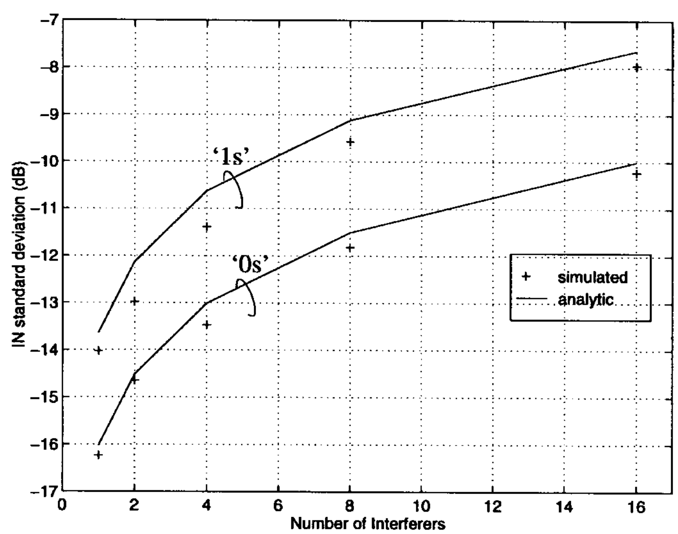

Fig. 3. Standard deviation of the IN versus number of interferers for $\varepsilon = -26$ dB.

to 2 dB. As we decrease the number of interferers the whole process tends to Gaussian (by the central limit theorem) and the GA improves, ultimately agreeing closely with the MCB. However, it is a very common situation in optical networks for just one or two IN terms to be dominant.

Both the GA and the MCB formulations described above can be used to provide an estimate of power penalty. We define the power penalty as the extra signal power required to obtain a BER of 10^{-9}. In Fig. 5, we compare the power penalty given by the GA and the MCB. Again, as with the BER comparison, we find that for a small number of interferers the GA significantly overestimates the crosstalk isolation tolerance when compared to that given by the very tight MCB. It can be seen that for a power penalty of 2 dB, the GA overestimates the crosstalk isolation tolerance by 4 and 2 dB for $N = 1$ and $N = 2$, respectively. This is because, as Fig. 6 clearly illustrates, the GA erroneously predicts error floors for a small number of interferers which leads to the significant overestimation of the PP. This, as noted earlier, is due to the unbounded nature of the GA.

IV. Conclusion

In this paper, we have addressed the effect of IN on fiber/radio networks and we have presented a rigorous for-

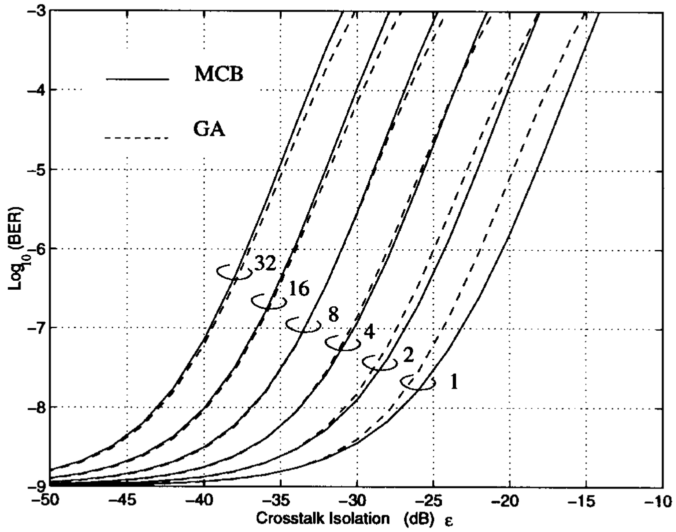

Fig. 4. BER versus the crosstalk isolation for various numbers of interferers.

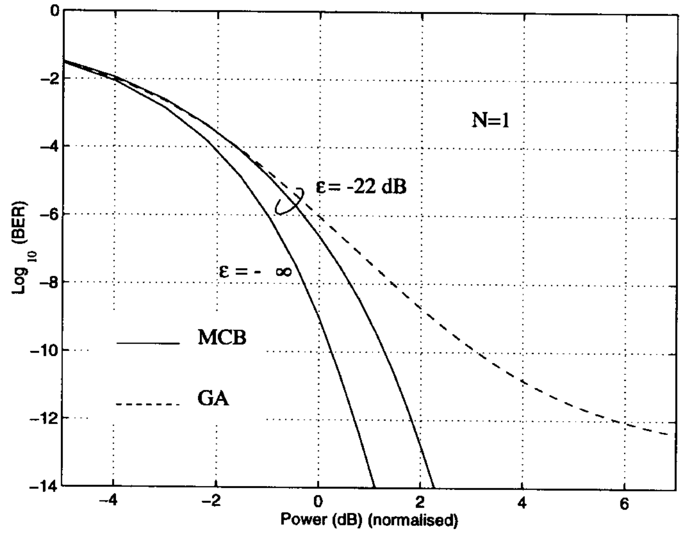

Fig. 6. BER versus optical power.

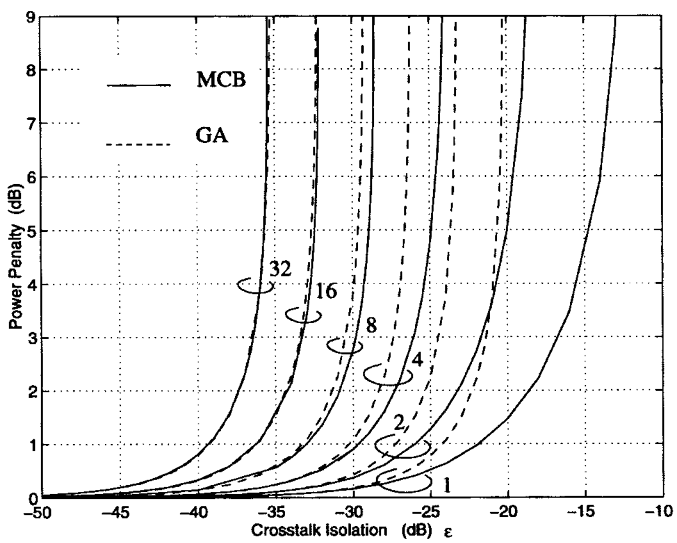

Fig. 5. Power penalty versus the crosstalk isolation for various numbers of interferers.

mulation based on the MCB/MGF approach for calculating the BER and PP. This paper also enables the assessment of the crosstalk isolation requirements for the various optical components in such networks.

The appropriateness of the GA for the characterization of the impact of IN on the system performance was also investigated, and from this we conclude that the GA is appropriate provided the number of significant interferers, N, is greater than about 4. However, for very small N, the GA can significantly overestimate the BER and the power penalty, and it was observed that such an approach can overestimate the crosstalk isolation tolerance by up to 5 dB for a PP of 2 dB. In such a case, recourse has to be made to the more accurate MCB formulation if the impact of IN is to be correctly assessed.

We have also reported simulation results which illustrate the interferometric conversion of phase noise to amplitude noise. The simulated results for the standard deviation of the IN on the detected signal are in good agreement with the theory.

Our results clearly show that IN can be considered a significant impairment in fiber-radio systems.

APPENDIX

In this Appendix, we derive the MGF's of the random variables (RV's) which characterize the signal and IN at the input of the decision device.

At the sampling time, the RV's which describe the IN for a received "0" and a received "1" can be written as follows:

$$\mathrm{IN}_0 = E^2 \sum_{k=1}^{N} \sqrt{\epsilon_k} a_k \cos(\Phi_\mathrm{k}) \tag{10}$$

$$\mathrm{IN}_1 = E^2 \sum_{k=1}^{N} \sqrt{\epsilon_k} a_k \cos(\Phi_\mathrm{k}) + \mathrm{E}^2 \mathrm{a}_0 \sum_{\mathrm{n}=1}^{\mathrm{N}} \sqrt{\epsilon_\mathrm{n}} \cos(\Phi_\mathrm{n}).^1 \tag{11}$$

The MGF of a random variable X is defined as $E[e^{sX}]$ [6] with $E[\cdot]$ denoting statistical average and s representing the Laplace domain variable. Hence, it can be shown that the MGF of IN_0, conditioned on the random variable a_k can be written as:

$$M_{\mathrm{IN}_0}(a|a_k) = \prod_{k=1}^{N} I_0(s\sqrt{\epsilon_k} a_k E^2). \tag{12}$$

$I_0(\cdot)$ denotes the modified Bessel function of order 0. Removing the conditioning of (12) on a_k we get

$$M_{\mathrm{IN}_0}(s) = \prod_{k=1}^{N} \left[\frac{1}{2} I_0(s\sqrt{\epsilon_k} a E^2) + \frac{1}{2}\right]. \tag{13}$$

Similarly, it could be shown that the MGF of IN_1 can be written as

$$M_{\mathrm{IN}_1}(s) = M_{\mathrm{IN}_0}(s) \prod_{k=1}^{N} I_0(s\sqrt{\epsilon_k} a E^2). \tag{14}$$

[1]We note that the RV's Φ_k and Φ_n have the same distribution but are distinct RV's even for $k = n$. See (3) in this paper.

Therefore, the MGF's of the random variable which characterize the signal plus IN for the received "0" 's and "1" 's, can be expressed as $M_0(s) = M_{\mathrm{IN}_0}$ and $M_1(s) = M_{\mathrm{IN}_1} e^{E^2 as}$, respectively.

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