Coherent Subcarrier Fiber-Optic Communication Systems with Phase-Noise Cancellation

Wing C. Kwong, Paul R. Prucnal, and Malvin C. Teich

***Abstract*— A phase-noise-cancelled coherent subcarrier fiber-optic communication system that uses integrated-optic Mach–Zehnder waveguide modulators for frequency-pair encoding is introduced. Optical double-sideband suppressed-carrier signals with the same random phase noise and information encoded in the frequency separation of the two sidebands are generated by properly biasing the modulator. Phase noise is eliminated at the receiver by a nonlinear operation on the two sidebands. System performance analysis with numerical examples, taking into account the modulation index and laser normalized linewidth, is presented.**

I. Introduction

COHERENT optical communication systems based on semiconductor lasers are currently receiving considerable attention, but their performance is degraded by the laser linewidth when phase information is used for data transmission [1]–[5]. For example, it has been shown that laser phase noise degrades system performance, imposes a bit error rate (BER) floor, and restricts the minimum allowable bit rate [2]–[5]. In conventional coherent frequency shift keyed (FSK) systems, frequency deviation $\geq \pm(2R_b+8\Delta\nu)$ or intermediate frequency (IF) linewidth $\leq 0.0025R_b$ is required to avoid the deleterious effects of laser phase noise where R_b is the bit rate and $\Delta\nu$ is the IF linewidth [4], [5]. A 100 channel optical frequency-division multiplexing experiment at 622 Mbit/s with laser linewidth of 20 MHz, frequency deviation of 2 GHz, and channel separation of 10 GHz has been demonstrated [5] where a large amount of optical bandwidth is wasted.

Optical bandwidth can be effectively utilized if phase-diversity configurations are employed [6]–[12]. Among the phase-diversity techniques, the "two-signal" encoding scheme is a promising candidate [9]–[12]. Two signals with the same random phase noise but different frequencies ("two-frequency" encoding) or polarizations ("two-polarization" encoding) are generated simultaneously at the transmitter. These two signals then propagate through the same optical channel and undergo the same random changes, for example, phase and polarization. Phase noise is then cancelled by performing a nonlinear operation on the two signals at the receiver. In general, the "two-frequency" encoding scheme [9] is more practical to employ in fiber-optic systems than the "two-polarization" encoding scheme [10] because the former does not require polarization-maintaining fibers and allows the use of polarization-diversity type receivers [13]. However, the "two-frequency" encoding scheme has only been demonstrated in free space using acousto-optic modulators which could be slow, bulky, and perform poorly if the system parameters change. In addition, nonsymmetric signal paths in the frequency-pair generation might result in incomplete phase-noise cancellation.

In this paper, we extend the previous results on free-space links to fiber-optic communication systems using coherent subcarrier [14] frequency-pair encoding techniques with integrated-optic Mach–Zehnder waveguide modulators which are compatible with fiber-optic systems. Data are modulated in a microwave subcarrier which, in turn, phase-modulates laser optical fields in the modulator. Frequency-modulated (FM) signals with harmonic frequencies related to the subcarrier are generated. Instead of reducing or avoiding the harmonics, they are used effectively by properly biasing the modulator to generate double-sideband suppressed-carrier signals with information encoded in the frequency separation of the two sidebands. Phase noise is eliminated at the receiver by a nonlinear operation on the two sidebands. As a result, the subcarrier frequency at the receiver is twice that of ordinary subcarrier systems where carriers are not suppressed. The electronic bandwidth requirements of the modulator and subcarrier generator are then half of that of ordinary subcarrier systems. The symmetric optical paths in the modulator can improve phase-noise cancellation at the receiver over acousto-optic devices. Since polarizations are not used for encoding, conventional polarization-diversity type receivers can be used [13]. In addition, some of the problems associated with pulsed laser sources, such as non-linearities, intensity noise, and pattern-dependent heating effects, are avoided as external modulators are used. Frequency stability within a channel is not a problem because information is encoded in the sideband difference, which is independent of laser instability.

II. System Description

As shown in Fig. 1, the laser source generates an optical continuous wave (CW) at a given center frequency f_c with phase noise $\theta_c(t)$. The electrodes of the integrated-optic Mach–Zehnder waveguide modulator are driven by a dc bias voltage V_{dc} added to a sinusoidal drive signal $V_{\mathrm{drive}}(t) =$

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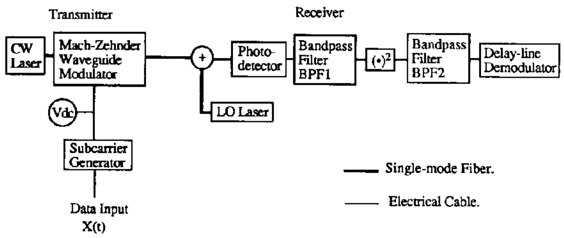

Fig. 1. Block diagram of the fiber-optic communication system using the frequency-pair encoding scheme with a Mach–Zehnder waveguide modulator.

$V\cos(2\pi x(t))$ generated by the subcarrier generator where V is the amplitude and $x(t) = \frac{1}{2}(f_s t + f_d \int_0^t X(t')dt')$ is the microwave subcarrier with continuous-phase FSK (CPFSK) modulated information. f_s is the microwave carrier frequency of the CPFSK modulation signal after phase-noise cancellation at the receiver. $X(t) \in \{-1, +1\}$ is a binary data sequence with a bit period T. f_d is the frequency deviation, defined by the modulation index $m \triangleq 2f_dT$.

The optical CW signal is split in the two waveguides of the modulator and then phase-modulated by the drive signal $V_{\text{drive}}(t)$. Because of the push-pull design of the electrode structure, two optical fields with the same optical parameters (i.e., f_c and $\theta_c(t)$), but opposite phase components are generated and later combined at the output with optical field [12]

$$E(t) \propto e^{j(2\pi f_c t + \theta_c(t) + \Theta + \Phi\cos(2\pi x(t)))} + e^{j(2\pi f_c t + \theta_c(t) - \Theta - \Phi\cos(2\pi x(t)))} \quad (1)$$

where $\Theta \propto V_{\text{dc}}$ is the phase difference between the fields imparted by the two waveguides and $\Phi \propto V$ is the phase modulation depth. Equation (1) simply represents a FM signal with harmonic frequencies related to the subcarrier $x(t)$. The harmonics are used effectively by setting $\Theta = \frac{1}{2}\pi$ to generate double-sideband suppressed-carrier signals where

$$E(t) \propto \sum_{\substack{n=-\infty \\ n=\text{odd}}}^{\infty} 2J_n(\Phi) e^{j(2\pi f_c t + 2\pi n x(t) + \theta_c(t) + \frac{1}{2}n\pi)}. \quad (2)$$

$J_n(\Phi)$ is the Bessel function of the first kind with integer order n. For Φ within the range of interest (i.e., $\Phi = [1.0, 1.5]$ rad), terms with $|n| > 3$ are negligible. Only those terms with $J_{\pm 1}(\Phi)$, which are the two sidebands, and $J_{\pm 3}(\Phi)$ contribute appreciably. The adverse effects caused by terms with $J_{\pm 3}(\Phi)$ can be minimized by carefully adjusting the system parameters.

The receiver consists of a LO laser, photodetector, bandpass filter BPF1, square-law device, narrow-bandpass filter BPF2, and delay-line demodulator [3]. Here, polarization fluctuations and intensity noise of the optical signal and LO laser are ignored for simplicity, otherwise polarization-diversity and dual-balanced detector configurations are required [9], [15]. The desired IF signal is selected by tuning the center frequency of the LO laser. The bandpass filter BFP1, which is used as a noise suppressor, is assumed to be ideal with a bandwidth $B_1 = (1 + 0.25\text{m})R_b + f_s + 12\Delta\nu_{\text{IF}}$ that is large enough to retain most of the energy in the information carrying signals with total IF linewidth $\Delta\nu_{\text{IF}}$ taken into consideration [16]. Therefore, the photocurrent at the square-law detector can be written as

$$i(t) \approx i \sum_{\substack{n=-3 \\ n=\text{odd}}}^{3} 2J_n(\Phi)\cos(2\pi f_{\text{IF}} t + 2\pi n x(t) + \theta_{\text{IF}}(t) + \frac{1}{2}n\pi) \quad (3)$$

where i is the photocurrent converted from the received optical signal and LO power [3] and $\theta_{\text{IF}}(t)$ is the total phase-noise process. The square-law detector, which is assumed to be ideal, performs the nonlinear operation on the two sidebands and results in a signal $a(t)$ with dc terms, double IF (i.e., $2f_{\text{IF}}$) terms with phase noise $\theta_{\text{IF}}(t)$, and frequency-difference terms without phase noise where

$$\begin{aligned} a(t) \approx\ & DC + \text{double IF} \\ & + 2i^2[(J_1^2(\Phi) - 2J_1(\Phi)J_3(\Phi))\cos(2\pi \cdot 2x(t)) \\ & - 2J_1(\Phi)J_3(\Phi)\cos(2\pi \cdot 4x(t)) \\ & + J_3^2(\Phi)\cos(2\pi \cdot 6x(t))]. \end{aligned} \quad (4)$$

The $2x(t)$ term represents the desired signal, while other terms can be filtered out by BPF2 with a bandwidth $B_2 = (1 + 0.5m)R_b$. The filtered phase-noise-cancelled signal can be written as $a(t) \approx 2i^2(J_1^2(\Phi) - 2J_1(\Phi)J_3(\Phi))\cos(2\pi f_s t + 2\pi f_d \int_0^t X(t')\,dt')$, which simply represents a conventional binary orthogonal CPFSK modulation signal with microwave carrier f_s and frequency deviation f_d. Data are then recovered by the delay-line demodulator.

III. System Performance Analysis

In coherent systems, the noise in optical heterodyning arises from the LO and from its interaction with the beat signal [17]. For sufficiently large LO power P_{LO}, we assume that all receiver noise sources can be ignored and that shot-noise limited behavior is a satisfactory approximation. Under these conditions, the single-sided power spectral density (PSD) of the noise process within the passband of the bandpass filter BPF1 can be approximated as $N \approx 2qRP_{LO}$ where q is the electron charge and R is the responsivity. The noise process after the nonlinear operation at the square-law device is non-Gaussian. The computation of the statistics can be simplified by assuming a Gaussian. Although this provides an overestimation of the error probability, the nonlinear noise process is expected to give a statistic well approximated by a Gaussian statistic for the case of a wide passband B_1, large signal-to-noise rate (SNR) and narrow-passband B_2. Thus, the noise PSD at the output of the demodulator can be written as

$$N_a(f) \approx 2i^2(J_1^2(\Phi) + J_3^2(\Phi))N + \frac{1}{4}N^2 B_1 \qquad f_s - \frac{1}{2}B_2 \le f \le f_s + \frac{1}{2}B_2. \quad (5)$$

The probability of error for demodulating the phase-noise-cancelled CPFSK signal using a delay-line demodulator with

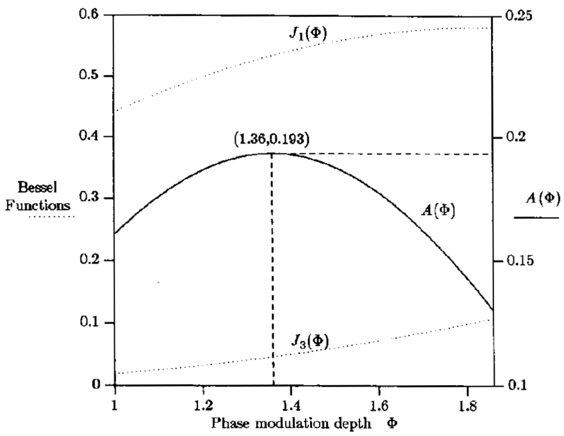

Fig. 2. The first-order $J_1(\Phi)$ and third-order $J_3(\Phi)$ Bessel functions of the first kind (dotted curves) and the ratio of Bessel functions $A(\Phi)$ (solid curve) versus the phase modulation depth Φ, over the range $\Phi = [1.0, 1.85]$.

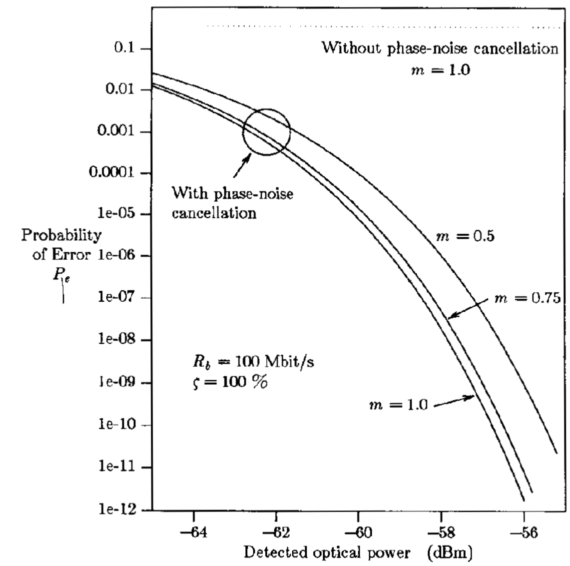

Fig. 3. Probability of error P_e versus detected optical power for various modulation indices with (solid) and without (dotted) phase-noise cancellation. A bit rate $R_b = 100$ Mbit/s and normalized linewidth $\zeta = 100\%$ are used in the calculation.

a pulse-shaping filter can then be written as [3]

$$P_e \approx \Psi(-\sqrt{\text{SNR}}) \tag{6}$$

where

$$\sqrt{\text{SNR}} = \frac{a^2(\Phi)}{2\sqrt{a^2(\Phi)N_a H + N_a^2 B_2 H}}. \tag{7}$$

$a(\Phi) = 2i^2(J_1^2(\Phi) - 2J_1(\Phi)J_3(\Phi))$, $\Psi(\cdot)$ is the unit normal cumulative distribution function, and

$$H = \int_0^{R_b} \left| \frac{1+\cos(\pi f T)}{2\,\text{sinc}\,(fT)} \cdot \frac{1-(fT/m)^2}{\cos(\pi f T/2m)} \right|^2 df \quad fT < 1, \tag{8}$$

is the single-sided noise bandwidth of the pulse-shaping filter which is assumed to have a raised-cosine shape. The delay τ in the delay-line demodulator is designed to be $T/2m$ to optimize the demodulation process [3].

For the case of high SNR, (7) can be simplified and approximated to be SNR $\propto A(\Phi)$ where

$$A(\Phi) = \frac{(J_1^2(\Phi) - 2J_1(\Phi)J_3(\Phi))^2}{J_1^2(\Phi) + J_3^2(\Phi)} \tag{9}$$

for a fixed optical detected power and modulation index. The behavior of $J_1(\Phi)$, $J_3(\Phi)$, and $A(\Phi)$ with Φ are summarized in Fig. 2 where these three functions are plotted against the phase modulation depth over the range $\Phi = [1.0, 1.85]$. As shown in the figure, $\Phi_{\text{opt}} = 1.36$ rad yields the best SNR in the proposed system and is therefore used in the following calculations.

Shown in Fig. 3 are the BER curves for different modulation indices with (solid) and without (dotted) phase-noise cancellation in an environment with an IF linewidth comparable with the bit rate. A bit rate $R_b = 100$ Mbit/s and normalized linewidth $\zeta = \Delta\nu/R_b = 100\%$ are used for the calculations. The system without phase-noise cancellation suffers severe degradation and a BER floor is observed, while the proposed system performs far better and has no BER floor. Based on the calculation, the receiver sensitivity for BER$\leq 10^{-9}$ can be as low as -57.3 dBm for $m = 1.0$. As also shown in Fig. 3, the BER depends on the modulation index and improves as m increases.

Shown in Fig. 4 are the BER curves for various modulation indices with different detected signal powers and $R_b = 100$ Mbit/s. With phase-noise cancellation, the solid and dashed curves represent the case with $\zeta = 100\%$ and 0%, respectively. The BER of the proposed system is seen to be a weak function of the laser linewidth. As a whole, the BER improves with increasing modulation index. The improvement is rapid for small values of m, but it slows down as m increases. This can be explained by the relationship between the single-sided noise bandwidth H and the modulation index. A larger modulation index allows a smaller delay τ, and hence shorter rise and fall times for pulses at the delay-line demodulator output. More filtering of the pulse-shaping filter is achieved. As a result, the value of H becomes smaller for larger modulation indices and this, in turn, decreases the noise power and increases the SNR. The dotted curve represents the case without phase-noise cancellation, a detected optical power $P_r = -57.5$ dBm, and a normalized linewidth $\zeta = 2.0\%$; the system suffers severe degradation even for such a small linewidth.

Shown in Fig. 5 are the BER curves for different normalized linewidths with (solid) and without (dotted) phase-noise cancellation for $m = 1.0$ and $R_b = 100$ Mbit/s. The system without phase-noise cancellation [3] suffers severe degradation and exhibits a BER floor for $\zeta > 1.0\%$, while the proposed system performs better, and has no BER floor, even with a large amount of phase noise. The BER does improve slightly because the passband B_1 is narrower and more noise filtering is achieved as ζ decreases. For a BER $\leq 10^{-8}$, the proposed

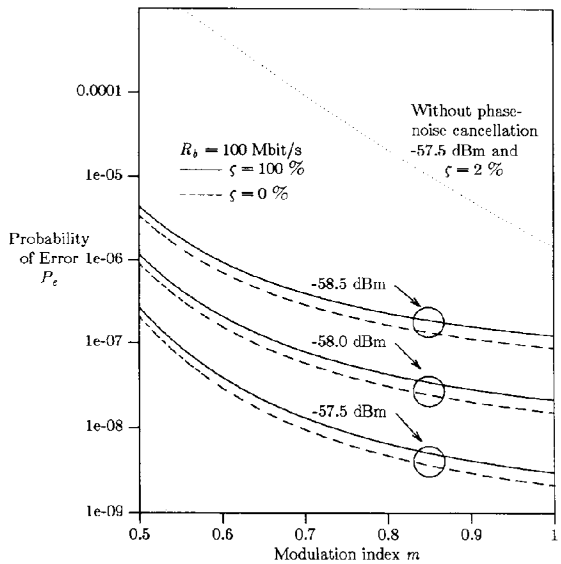

Fig. 4. Probability of error P_e versus modulation index m for various detected optical powers with (solid and dashed) and without (dotted) phase-noise cancellation. A bit rate R_b = 100 Mbit/s and normalized linewidth ζ = 100% are used in the calculation. Normalized linewidths of 100% and 0% are used in the solid and dashed curves, respectively. A detected optical power of −57.5 dBm and normalized linewidth ζ = 2% are used in the dotted curve.

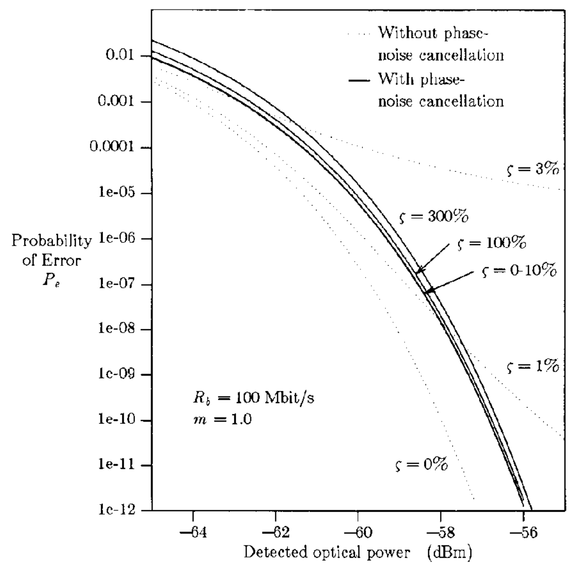

Fig. 5. Probability of error P_e versus detected optical power for various normalized linewidths with (solid) and without (dotted) phase-noise cancellation. A bit rate R_b = 100 Mbit/s and modulation index m = 1.0 are used in the calculations.

system performs better than the system without phase-noise cancellation as ζ increases beyond 1.0%. Comparing the dotted curve with $\zeta = 0\%$ with the corresponding solid curve, there is less than a 2 dB penalty in sensitivity resulting from use of the proposed system for BER = 10^{-9}. The penalty arises from additional noise generated from the nonlinear operations. Also, the overestimation of the noise power may be a cause. Therefore, the performance of the proposed system is inferior to that of conventional narrow-deviation FSK systems so that it should not be used when very small laser linewidths (i.e., $\zeta < 1\%$) are involved.

IV. Summary

A phase-noise-cancelled coherent subcarrier fiber-optic network utilizing the frequency-pair encoding technique with integrated-optic Mach–Zehnder waveguide modulators is studied. The electronic bandwidth requirements of the modulator and subcarrier generator are half of that of ordinary subcarrier systems. The receiver sensitivity is determined and related to the modulation index and normalized linewidth. With phase-noise cancellation, the proposed system only suffers a minor performance degradation, even for an IF linewidth comparable with the bit rate, while extensive performance degradation is found using conventional FSK systems without phase-noise cancellation. However for the case of normalized linewidth less than 1.0%, the proposed system performs worse than a conventional CPFSK system (by about 2 dB) and should therefore not be used. Finally, the use of external modulation is a limitation of this system, since it imposes additional power loss and cost. However, the proposed system will be very attractive if the modulator and laser are fabricated in a same package or chip. One can also see that an amplitude shift keyed (ASK) modulation format can be applied to this system with only minor modifications. This can be done simply by transmitting the two sidebands for data bit +1, but transmitting no light for data bit −1.

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