Using Low-Density Parity-Check Codes to Reduce the Effect of Laser Line width For Optical Communication

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Abstract-

The performance of coherent optical communication systems is degraded significantly by the phase noise of the semiconductor lasers. The phase noise is induced by spontaneous emission in the laser cavity and yields broadening in the laser linewidth. This paper addresses the application of the Low-Density Parity-Check (LDPC) codes as Forward Error Correcting (FEC) codes to relax the laser linewidth requirement. These codes are applied to three types of heterodyne optical receivers (BPSK, DPSK and QPSK) operating with finite laser linewidths.

1. Introduction

The more important reasons for using the coherent optical receiver is it's ability to amplify the received signal optically for better Signal-to-Noise Ratio (SNR) [1]. The semiconductor lasers used in coherent optical communication systems exhibit phase noise, that causes spectral broadening and center frequency instability. This can lead to sharp performance degradation [2]. The effect of the laser linewidth on the performance of coherent optical communications systems can be substantially reduced by using advanced laser sources having narrow linewidths or using FEC codes [3]. Recently there is

interest in block codes which exploits the advantage of iterative decoding constituted by the so called Low-Density Parity-Check (LDPC) codes. The LDPC decoding is based on a combination of simple and fast decoding of short linear block codes, such as Hamming codes, BCH codes or RS codes. When properly designed, the LDPC codes have large minimum Hamming distance [4]. The aim of this paper is to investigate the improvement gained by employing these advanced codes in reducing the effect of laser linewidth on the performance of heterodyne optical receivers.

Two feature parameters are used in this paper to calculate the receiver improvement due to FEC codes

- Coding Gain (*GC*) which is defined as the ratio between the received power without and with coding at a specific BER.
- Laser Relaxing Factor (*LRF*) which is defined as the ratio between laser linewidth without coding and with coding at a specific BER.

2. Analysis of Heterodyne Optical Receivers

Heterodyne coherent optical receivers with different modulation schemes are analyzed The analysis takes into account both local laser shot noise and transmitter and receiver lasers phase noise. When coherent Intermediate Frequency (IF) demodulation scheme is adopted, the parameters of the electrical Phase Locked Loop (PLL) are optimized to ensure minimum phase error.

2.1 Optical Bpsk Receivers

Figure 1 illustrates the heterodyne optical Binary Phase Shift Keying (BPSK) receiver [5]. The frequency of the received optical signal ω_R and the frequency of the local laser ω_{LO} differ by a radio frequency called IF and denoted by $\omega_{IF} = \omega_R - \omega_{LO}$. The photo current output is filtered by an IF band-pass filter. The output of the IF filter is used to drive an Automatic Frequency Control (AFC) device in the optical carrier recovery loop [6]. Let the received optical signal field be

$$E_R(t) = \sqrt{2P_R}\,a(t)\cos(\omega_R t + \phi_R(t)) \quad (1)$$

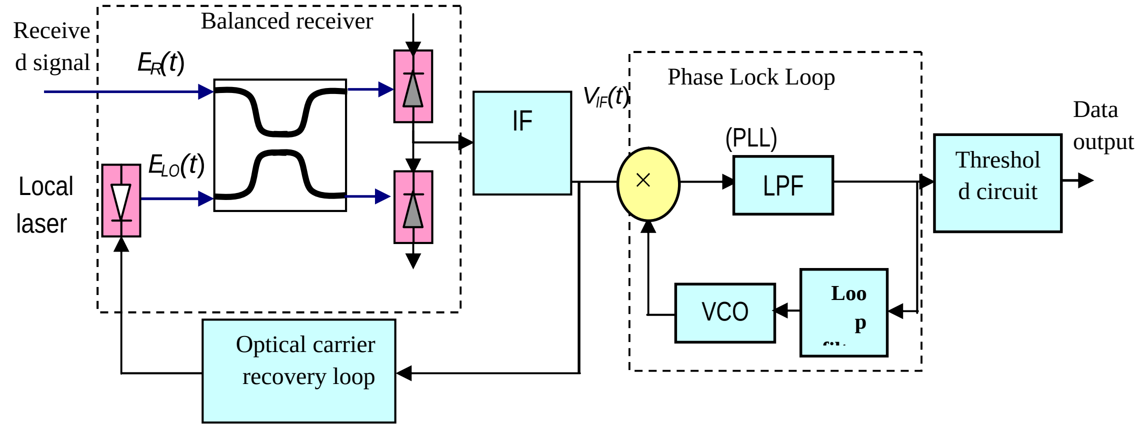

Fig.1 A heterodyne optical BPSK receiver.

where P_R is the power of the received optical signal, ω_R is the angular frequency of the received optical signal, $\phi_R(t)$ is the phase noise of the transmitter laser, and *a(t)* is the modulation signal {1,-1} using NRZ format. The optical Local Oscillator (LO), i.e. local laser field is given by

$$E_{LO}(t) = \sqrt{2P_{LO}}\cos(\omega_{LO}t + \phi_{LO}(t)) \qquad (2)$$

where P_{LO}, ω_{LO}, and $\phi_{LO}(t)$ are, respectively, power, angular frequency, and phase noise of the LO signal. The optical hybrid (3dB coupler) mixes the two optical signals (i.e. the received signal and LO electrical fields) [7]. This converts the received optical signal into a photocurrent corresponding to IF signal. The output of the IF amplifier for the upper branch is expressed as

$$V_{IF}(t) = \frac{1}{2}S_A\cos(\omega_{IF}t + \phi_{IF}(t)) + n(t) \qquad (3)$$

$$S_A = \Re\sqrt{P_R P_{LO}}$$

where $\Re$ is the photodiode responsitivity, ω_{LO} is the angular frequency, $\phi_{IF}(t)$ is the phase noise of the IF signal and equal to ($\phi_R(t)$ - $\phi_{LO}(t)$), and *n(t)* is the local laser shot noise. The phase noise is contained in the first term while the LO shot noise *n (t)* is shown in second term. The BPSK receiver output signal after decision circuit is $\pm S_A$.

The PLL allows the generation of variable output frequency by means of feedback. It generates an output signal that is proportional to the difference of the IF phase and the loop filter phase. Thus it maintains the phase difference between the electrical LO output and the IF signal. The PLL consists of three
basic blocks [8] phase detector, loop filter, and Voltage Controlled Oscillator (VCO).

The PLL shown in Fig. 2 incorporates an electronic VCO, whose output voltage is given by [9]

$$V_{VCO}(t) = V_O\cos(\omega_{IF}t + \phi_{VCO}) \qquad (6)$$

where V_O and ϕ_{VCO} are the amplitude and phase of VCO, respectively. The phase $\phi_{VCO}(t)$ is computed from

$$\phi_{VCO}(t) = K_{VCO}\int_{-\infty}^{t} V_C(t)\ dt \qquad (7)$$

where K_{VCO} is the VCO gain and $V_C(t)$ is the output of the loop filter.

The output of the IF filter $V_{IF}(t)$ and the output voltage of the VCO, $V_{VCO}(t)$ are mixed in the phase detector and then filtered by the Low-pass Filter (LPF) and smoothed by the loop filter

[10]. The output signal of the loop filter is expressed by

$$\begin{aligned} V_C(t) &= V(t) + n_C(t) \\ V(t) &= K_T \sin(\theta_d(t)) \\ n_C(t) &= n_{IF} V_{VCO}(t) \end{aligned} \quad (8)$$

where $V(t)$ is the output signal of the PLL branch, $n_C(t)$ is the noise associated with loop filter (consists of LO shot noise and loop filter noise), n_{IF} is the amplitude of the noise, $\theta_d(t)$ is the detector phase error, $K_T = K_{VCO}\ V_O\ A_{IF}$ is the total PLL gain, and A_{IF} is the IF amplifier gain.

The output signal $V_C(t)$ can be linearlized when the loop remains in lock with a small phase error, $\theta_d(t) << 1$ then $\sin\theta_d(t) \cong \theta_d(t)$. The output is given by

$$V_C(t) = K_T\ \theta_d(t) + n_C(t) \quad (9)$$

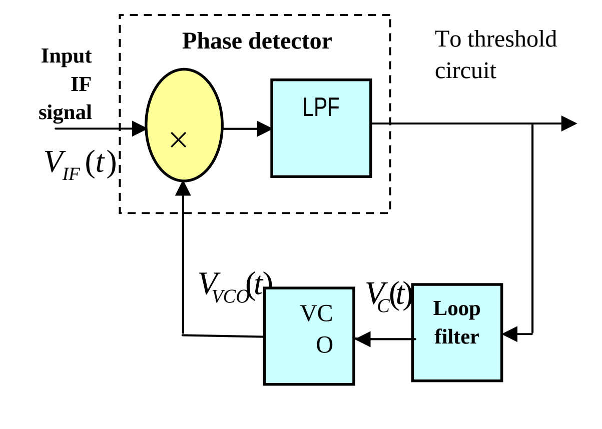

Fig. 2 Block diagram of electronical PLL.

The impact of local laser shot noise and laser phase noise on PLL phase error is analyzed now. The variance of the PLL phase error σ_T^2 can be splitted into two components

$$\sigma_T^2 = \sigma_{SN}^2 + \sigma_{PN}^2 \quad (10)$$

where σ_{SN}^2 and σ_{PN}^2 is, respectively, the variance of the PLL phase error due to shot noise and phase noise. Using the PLL linear model shown in Fig. 3a, σ_{SN}^2 and σ_{PN}^2 can be expressed as [11]

$$\begin{aligned} \sigma_{SN}^2 &= \int_{-\infty}^{\infty} \frac{S_{SN}(f)}{A^2} \left|H(j2\pi f)\right|^2 df \\ \sigma_{PN}^2 &= \int_{-\infty}^{\infty} S_{PN}(f) \left|1 - H(j2\pi f)\right|^2 df \end{aligned} \quad (11)$$

where $S_{SN}(f)$ is power spectral density of the shot noise, $S_{PN}(f)$ is the power spectral density of the phase noise $\phi_{IF}(t)$, A is the expected value of $S_A^2/2$, and $H(j2\pi f)$ is the closed loop transfer function of the PLL which is defined as the ratio of $\phi_{IF}(s)$ to $\phi_{VCO}(s)$ as expressed

$$H(s) = \frac{\phi_{VCO}(s)}{\phi_{IF}(s)} = \frac{AK_T G(s)}{s + AK_T G(s)} \quad (12)$$

where *G(s)* is the transfer function of the loop filter. For a first-order active filter (see Fig. 3b), *G(s)* is given by

$$G(s) = \frac{K_T(s\ \tau_2 + 1)}{s\ \tau_1} \quad (13)$$

τ_1, τ_2 = Time constants of the loop filter and expressed as $\tau_1 = C R_1$, $\tau_2 = C R_2$.

The local oscillator works as VCO [12]. The loop can be expressed as

$$\begin{aligned} H(s) &= \frac{AK_T(s\ \tau_2 + 1)/\tau_1}{s^2 + AK_T s\ (\tau_1/\tau_2) + K_T/\tau_1} \\ &= \frac{\omega_n^2 + 2\zeta\ \omega_n s}{s^2 + 2\zeta\ \omega_n s + \omega_n^2} \end{aligned} \quad (14)$$

where

$$\begin{aligned} \omega_n^2 &= K_T / \tau_1 \\ \zeta &= (\tau_2/2)\sqrt{K_T/\tau_1} \end{aligned} \quad (15)$$

Here ω_n and ζ is the natural angular frequency and damping factor of the second-order transfer function, respectively.

The equivalent loop noise bandwidth B_L of the PLL is expressed as [13]

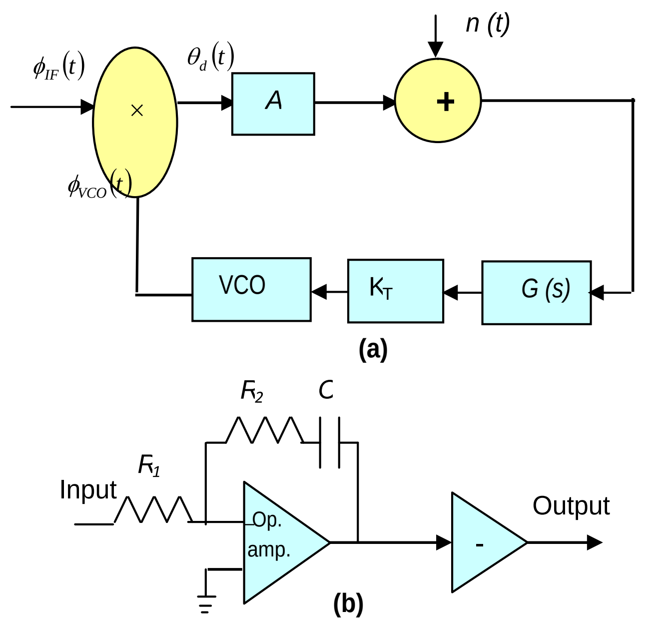

Fig. 3 (a) Linearlized model of the second-order PLL.
(b) Loop filter circuit (first-order active filter)

The total phase error variance is obtained by Eq. (10) an can be expressed as

$$\sigma_T^2 = \frac{\pi(1+4\zeta)^2 \ \Delta v}{8\zeta^2 B_L} + \frac{T_b}{SNR} B_L \qquad (17)$$

where Δv is the full width at half maximum laser linewidth, SNR is the signal-to-noise ratio, and T_b is the bit interval time.

In Eq.(17) the σ_T^2 depends on B_L . When B_L increases, then σ_{SN}^2 increases, but σ_{PN}^2 decreases. So there is an optimum value, for the loop bandwidth $(B_L)_{opt.}$, at which the total phase error variance σ_T^2 is minimum. This value is obtained by setting the first derivative of σ_T^2 in Eq. (17) with respect to B_L to zero.

For a BPSK system, the $SNR = \eta N_S$ where N_s is the number of photons per bit and η is the photodiode quantum efficiency. The expression of σ_T^2 given in Eq. (17) can be simplified at $\zeta = 1/\sqrt{2}$ to

$$\sigma_T^2 = \frac{0.53 \ \omega_n T_b}{SNR} + 2.22 \frac{\Delta v}{\omega_n} \qquad (18)$$

The optimum value of the natural angular frequency (ω_n)$_{opt}$, and the minimum value of total phase error variance $(\sigma_T^2)_{\min}$ are given , respectively, by

$$(\omega_n)_{opt} = 2.0466 \sqrt{\frac{\Delta v \ SNR}{T_b}}$$
$$(\sigma_T^2)_{\min} = 2.17 \sqrt{\frac{\Delta v \ T_b}{SNR}} \qquad (19)$$

In the presence of total phase error the probability of error is expressed by [14]

$$p = \frac{1}{2} - \exp^{(-SNR/2)} \sqrt{\frac{SNR}{\pi}} \sum_{m=0}^{\infty} \frac{(-1)^k}{2m+1} \left(I_m(SNR/2) + I_{m+1}(SNR/2)\right) \exp^{\left(-(2m+1)^2 \sigma_T^2/2\right)} \qquad (20)$$

where $I_m(.)$ is the first-kind Bessel function of order m. To minimize the effect of receiver phase error, the PLL bandwidth should be broadened. However the phase error variance caused by the local laser shot noise increases as the PLL bandwidth increases, see Eq. (17).

Therefore, the natural angular frequency (and hence the loop bandwidth B_L) is an important parameter to be used for decreasing the power penalty due to finite laser linewidth.

2.2 Optical DPSK Receiver

The heterodyne optical DPSK receiver uses a delay and multiplier demodulation schemes of T_b delay time, see Fig. 4. The total phase error due to laser phase noise for the consecutive symbol is given by [3]

$$\sigma_T^2 = 2\pi\Delta\nu T_b \qquad (21)$$

The error probability of DPSK receiver due to the total phase error can be expressed by [15]

$$p = \frac{1}{2} - \frac{SNR\,\exp^{(-SNR/2)}}{2}\sum_{m=0}^{\infty}\frac{(-1)^m}{2m+1}\left[I_m(SNR/2) + I_{m+1}(SNR/2)\right]^2 \exp^{\left(-(2m+1)^2\sigma_T^2/2\right)} \qquad (22)$$

2.3 Optical QPSK Receiver with Coherent Demodulation

The coherent QPSK receiver is shown in Fig. 5 and can be analyzed using the linear model of the PLL see in Fig. 3a. The total phase error variance due to phase noise and local laser shot noise is given in Eq. (19). The optimum value of the natural angular frequency (ω_n)$_{opt}$. and the minimum value of total phase error variance $\left(\sigma_T^2\right)_{\min}$ are expressed by Eq. (20).

The error probability for the coherent QPSK demodulation is express by [16]

$$p = \frac{1}{4\sqrt{2\pi\,\sigma_T^2}}\int_{-\infty}^{\infty}\exp^{\left(-\varphi^2/2\sigma_T^2\right)}\, erfc\left(-\sqrt{SNR/2}\,(\cos\varphi - \sin\varphi)\right) erfc\left(-\sqrt{SNR/2}\,(\cos\varphi + \sin\varphi)\right) d\varphi \qquad (23)$$

2.4 Optical QPSK Receiver with Differential Demodulation

The QPSK receiver differential demodulation shown in Fig. 6, has the advantage of less-complexity configuration than QPSK with coherent demodulation. The quadrature receiver can detect both in-phase and quadrature-phase components of the

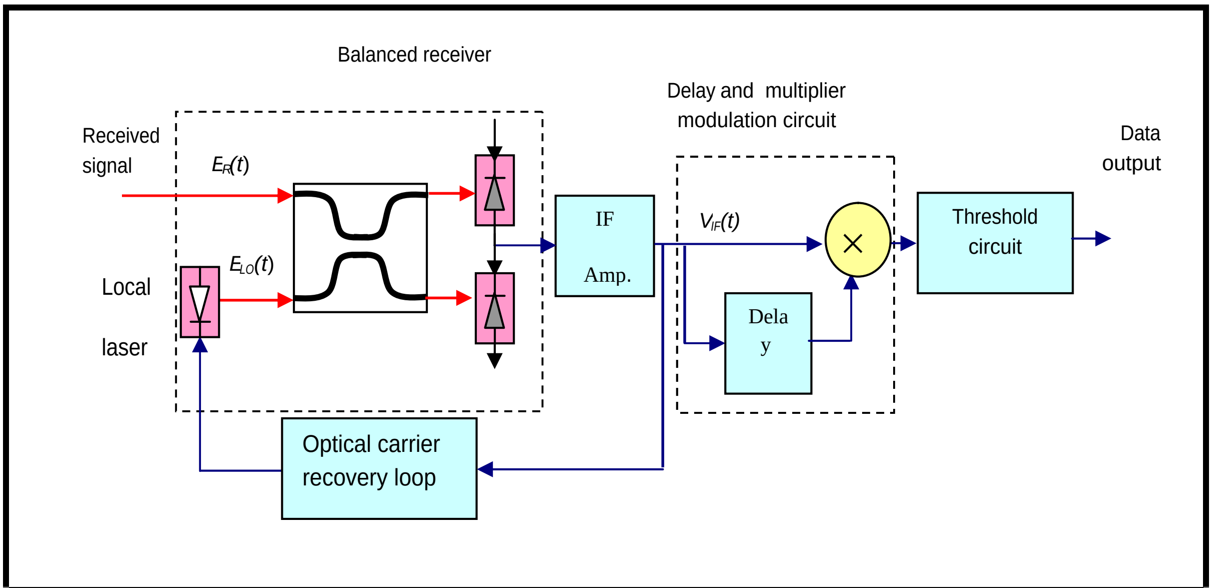

Fig. 4 A heterodyne optical DPSK receiver with delay and multiplier modulation circuit.

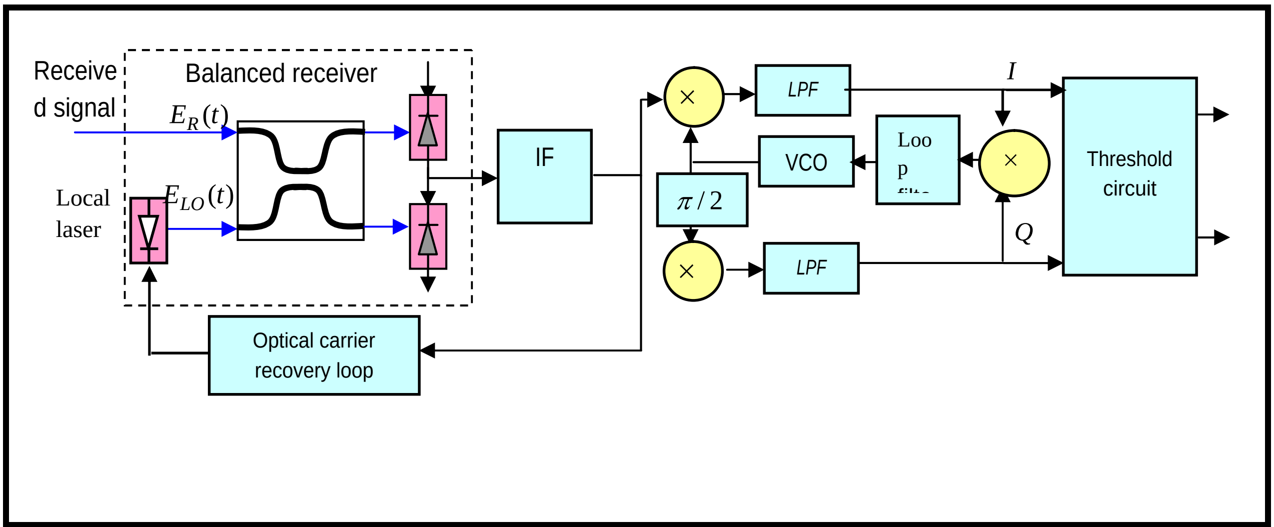

Fig. 5 A heterodyne optical QPSK receiver with coherent demodulation.

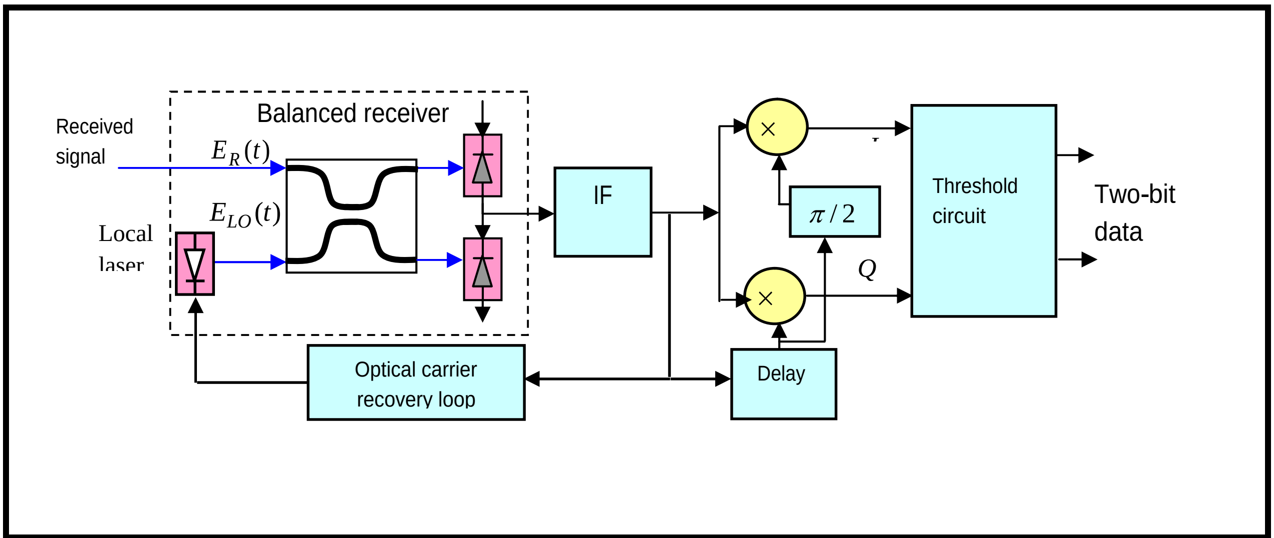

Fig. 6 A heterodyne optical QPSK receiver with differential demodulation.

optical signal (I and Q). The total phase error variance due to the laser phase noise in this modulator is expressed in Eq. (21). The error probability p depends on the total phase error is given by [17]

$$p=\frac{SNR}{\pi\sqrt{2\pi\sigma_T^2}}\int_0^{\infty}\int_{-\infty}^{\infty}\int_{-\infty}^{\infty}\exp\left(-\varphi^2/\sigma_T^2\right)\exp\left(-(x-1)^2-(y-1)^2 SNR/2\right)\cdot \left[1-\frac{1}{2}erfc\left(\frac{-x\cos\varphi+y\sin\varphi}{\sqrt{x^2+y^2}}\sqrt{SNR}\right)\right]dx\,dy\,d\varphi \quad (24)$$

3. Construction And Decoding of Ldpc Codes

To construct an LDPC code, one has to replace each single parity-check equation of an LDPC code by the parity-check matrix of

simple linear block code, known as the constituent or local code. A part from the parity check matrix of the local code, the construction also depends on the codeword length, the number of super-codes, and a permutation matrix.

The parity–check matrix of a LDPC code is a sparse matrix ***H*** constructed in the

following manner. The matrix ***H*** is partitioned into m sub-matrices as [14]

$\boldsymbol{H_1}, \boldsymbol{H_2}, \boldsymbol{H_3}, \ldots, \boldsymbol{H_m}$	25

where $\boldsymbol{H_1}$ is a block-diagonal matrix obtained from an identity matrix, with ones on the main diagonal replaced by the parity-check matrix $\boldsymbol{H_0}$ of a local code $C_0(n, k)$ as shown [14]

$$\boldsymbol{H}_1 = \begin{bmatrix} \boldsymbol{H}_0 & 0 & .. & 0 \\ 0 & \boldsymbol{H}_0 & .. & 0 \\ . & & .. & . \\ 0 & 0 & . & \boldsymbol{H}_0 \end{bmatrix} \quad (26)$$

Each of the sub-matrices is derived from $\boldsymbol{H_1}$ through a random column permutation that denoted by

$$\boldsymbol{H_j} = \pi_{j-1}(\boldsymbol{H_1}) \quad j = 2, 3, \ldots, m \quad (27)$$

where π_{j-1} is the permutation of j-1 column.

into sub-words of length each $m \times k$, where $\boldsymbol{u}$ = [$\boldsymbol{u_1}$, $\boldsymbol{u_2}$, ….., $\boldsymbol{u_m}$].

In general, one seeks LDPC codes for which the local codes $C_0(n, k)$ has large minimum distance ($d_{\min}$) and a code rate as high as possible. The lower bound on the minimum distance of a LDPC code is expressed as [18].

The LDPC code matrix $\boldsymbol{H}$ is expressed as

$$\boldsymbol{H} = \left[\boldsymbol{H}_1^T \ \boldsymbol{H}_2^T \ \ldots.. \ \boldsymbol{H}_m^T\right]^T \quad (28)$$

The code rate of LDPC code is denoted by

$$R = \frac{K}{N} \geq 1 - m\left(1 - \frac{k}{n}\right) \quad (29)$$

where K and N is the information and codeword of the LDPC code, respectively. Therefore, this type of LDPC code, denoted as LDPC (N, m, n), is the intersection of m super-codes C_1, C_2, …., C_m whose parity check matrices are $\boldsymbol{H_1}$, $\boldsymbol{H_2}$,….., $\boldsymbol{H_m}$, respectively. The information $\boldsymbol{u}$ be encoded by using LDPC code parsed information is spitted [4]

$$D \geq d_{\min} \frac{\left[(d_{\min} - 1)(J - 1)\right]^{(m/4)} - 1}{(d_{\min} - 1)(J - 1) - 1} \quad (30)$$

where m and J denote the girth i.e., length of the shortest cycle of the global code graph and the column weight of the global code, respectively. Obviously, large girth leads to an exponential increase in the minimum distance, while large values of d_{min} lead to an increase of the basis of this exponential function. The code calls regular LDPC codes if the number of ones in all columns (column weight) is fixed and irregular LDPC codes other wise.

The LDPC codes can effectively be decoded using iterative decoding scheme. The LDPC decoder belongs to the super-code $\boldsymbol{C_1}$ with parity-check matrix $\boldsymbol{H_1}$. The N/n series-in series-out (SISO) decoders work in parallel on independent N/n constituent codes $\boldsymbol{C_0}$ of the super-code $\boldsymbol{C_1}$. For every coded bit a posterior probability and an extrinsic probability are computed. The extrinsic probabilities from super-code $\boldsymbol{C_1}$ are fed to N/n constituent codes $\boldsymbol{C_0}$ of the super-code $\boldsymbol{C_2}$. The procedure is repeated for every super code, extrinsic probabilities of super-code $\boldsymbol{C_2}$ are fed to N/n constituent codes $\boldsymbol{C_0}$ of the super-code $\boldsymbol{C_3}$, …. , while the extrinsic probabilities of super-code $\boldsymbol{C_{m-1}}$ are fed to N/n constituent codes $\boldsymbol{C_0}$ of the super-code $\boldsymbol{C_m}$ (these steps define one iteration). The procedure is terminated either when a pre-determined number of iterations is reached or when a valid codeword is generated [6].

4. Simulation Results

The simulation results are presented for 1Gb/s receivers operating at 1550nm wavelength with 80% photodiode quantum efficiency using MATLAB-7 environment. Table 1 lists the receiver sensitivity at BER = 10^{-12} for different modulation schemes and in the absence of laser linewidth (i.e. linewidth (Δv) =0). The table also contains the maximum allowable value of $(\Delta v T_b)$ which ensures a 1dB power penalty at BER = 10^{12}.

Table 1 Summary of the results related to uncoded optical receivers operating at 1Gb/s data rate and BER =10^{-12}.

Receiver types	**Receiver sensitivity when (Δv =0) P_R** (dBm)	**$\Delta v T_b$ for 1dB power penalty**
BPSK	-49.62	2.86×10^{-3}
DPSK	-48.58	8.69×10^{-3}
QPSK with coherent demodulation	-49.28	3.78×10^{-4}
QPSK with differential demodulation	-47.93	6.00×10^{-4}

receivers having significant laser phase noise. The design of these codes depends on BCH codes having different code rates. These codes are LDPC (6393, 3591), LDPC (6393, 3213), and LDPC (6393, 2835) which give code rates equal to 0.9051, 0.8951 and 0.7143, respectively. The local codes used here are BCH (63, 57), BCH (63, 51) and BCH (63, 45), respectively. The column weight for all codes is fixed at J = 3 and the codes are extended by *m* = 8. Table 2 shows the main parameters of the LDPC codes used to improve optical receivers sensitivity.

Table 2 A main parameters of the three LDPC codes used in the simulation.

Coding systems C(N, K)	**Local code** $C_o(n, k)$	**Local code** d_{min}	**Shortest cyclic**	**Column weight**	**LDPC code** D_{min}	**Code Rate** =K/N
LDPC (3969, 3591)	BCH (63, 57)	3	8	3	15	0.9051
LDPC (3969, 3213)	BCH (63, 51)	5	18	3	45	0.8951
LDPC (3969, 2835)	BCH (63, 45)	7	8	3	91	0.7143

Simulation results for coded heterodyne optical receivers are given in Figs. 7-10 for BPSK, DPSK, QPSK with coherent demodulation , and QPSK with differential demodulation, respectively. The figures show the BER characteristics when the receivers incorporate the three LDPC codes shown in Table 2. The results are depicted for $\Delta v T_b$ listed in Table 1 and compared against the BCH (255, 223) code and RS(255,239) + RS(255, 239) concatenation code.

A second test is conducted to examine the systems performance under fixed received power, and different values of $\Delta v T_b$. The aim of the test is to find the receivers response for a change in a laser linewidth. Figs. 11-14 show the variation of *BER* with normalized linewidth $\Delta v T_b$ for BPSK, DPSK, QPSK with coherent demodulation, and QPSK with differential demodulation, respectively when the LDPC codes are used, these results are compared against the BCH(255, 223) code and RS(255, 239)+RS(255,239) concatenation code.

Table 3. summarizes the main results related to a 1Gb/s heterodyne optical receivers operating under LDPC coding schemes at BER = 10^{-12}. The results in this table highlight the following facts for BPSK receiver. The LDPC code gives higher *CG* and *LRF* than the BCH code operating at the same code rate. For example, the LDPC (3969, 3213) code gives *CG* = 10.38 dB and *LRF* = 3.35. These values are to be compared with *CG* = 2.78 dB and *LRF* = 2.74 for BCH (255,223) code having the same code rate. The LDPC code gives higher *CG* and *LRF* than the concatenated code operating at the same code rate. For example, the LDPC (3969, 3213) code gives *CG* = 10.38 dB and *LRF* = 3.35. These values are to be compared with *CG* = 4.9 dB and *LRF* = 3.24 for RS(255,239) + RS(255, 239) code have the same code rate. Decrease code rate of LDPC code gives improvement in *CG* and *LRF*. For Example, moving from LDPC (3969, 3591) code to LDPC (3969, 2835) code gives 1.3 dB improvement in *CG* and 3.38% enhancement in *LRF*. The results indicate clearly that LDPC code offers higher *CG* and *LRF* when implemented with QPSK system compared with BPSK and DPSK systems.

Similar conclusions can be deduced for other optical receives discussed in section 2

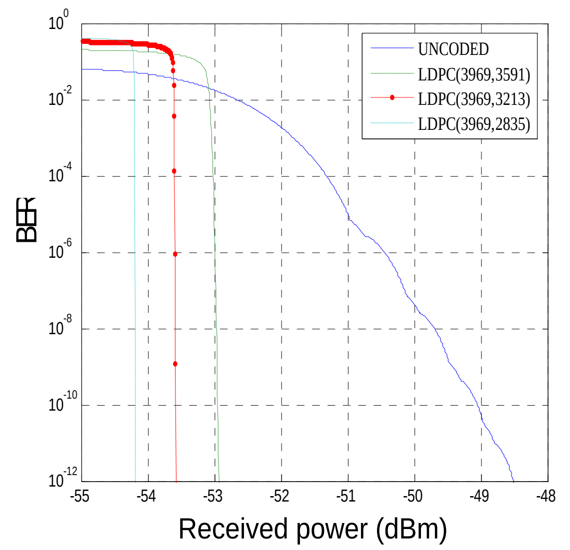

Fig. 7 BER versus received power for heterodyne optical BPSK receiver operating at 1Gb/s rate with different LDPC code and $\Delta v T_b = 2.86 \times 10^{-3}$.

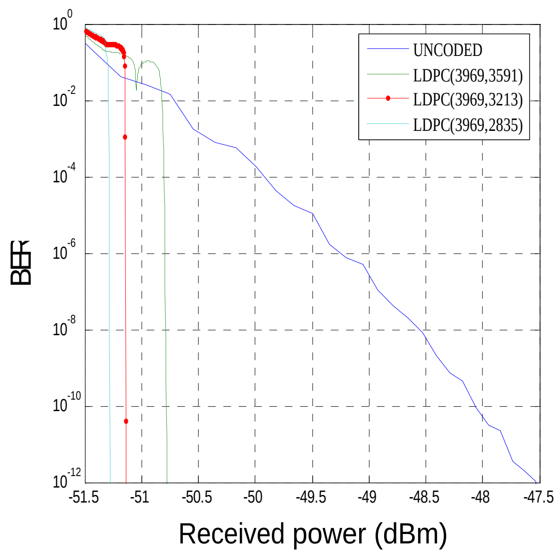

Fig.8 BER versus received power for heterodyne optical DPSK receiver operating at 1Gb/s rate under different LDPC with $\Delta v T_b = 8.69\times 10^{-3}$.

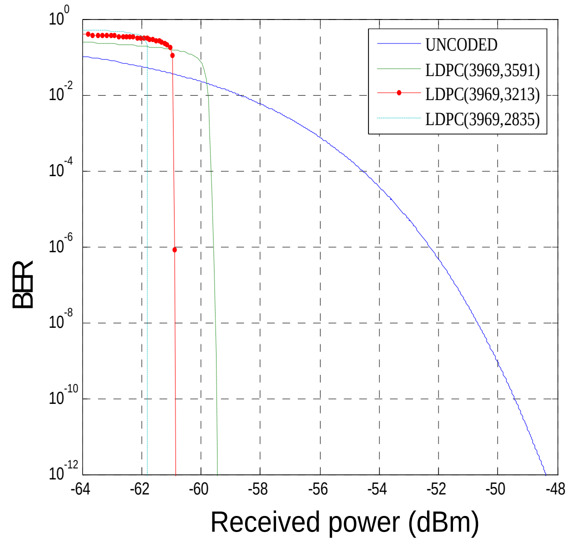

Fig. 9 BER versus received power for coherent QPSK receiver operating at 1Gb/s rate and incorporating coherent demodulation scheme with different LDPC codes and $\Delta v T_b = 3.78\times 10^{-4}$.

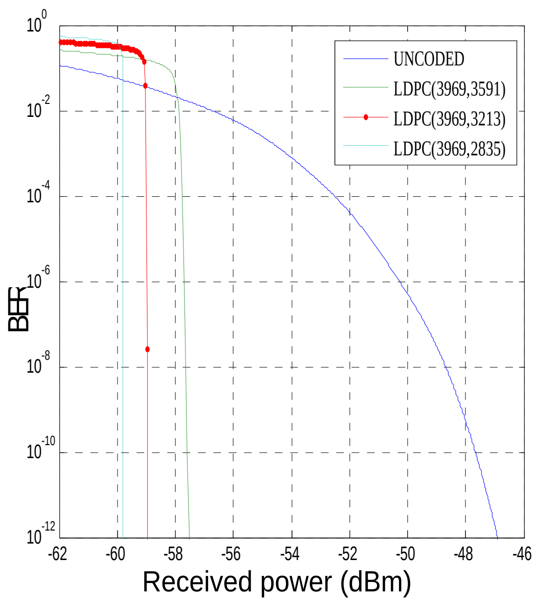

Fig. 10 BER versus received power for differential QPSK receiver operating at 1Gb/s rate and incorporating coherent demodulation scheme with different LDPC codes and $\Delta v T_b = 6\times 10^{-4}$.

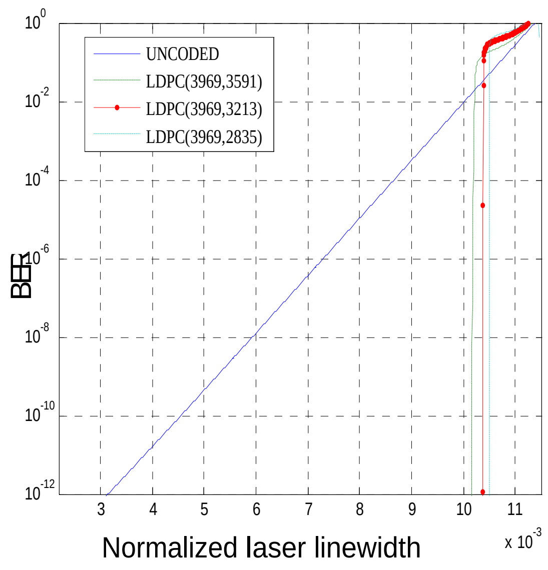

Fig. 11 BER versus $\Delta v T_b$ for heterodyne optical BPSK receiver operating at 1Gb/s rate and P_R = -48.65 dBm with different LDPC codes.

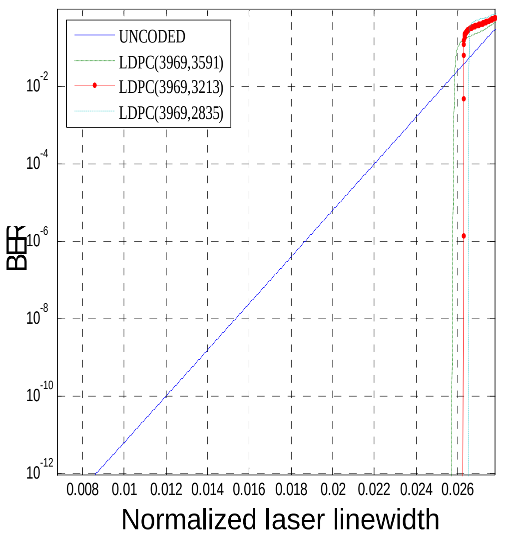

Fig. 12 BER versus $\Delta v T_b$ for heterodyne optical DPSK receiver operating at 1Gb/s rate and P_R = -47.45 dBm with different LDPC codes

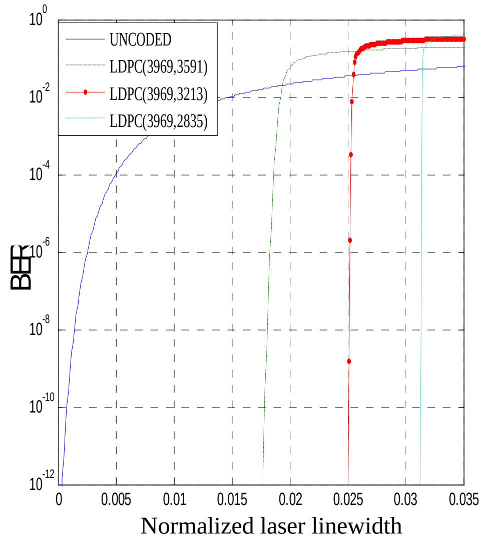

Fig. 13 BER versus $\Delta v T_b$ for heterodyne optical QPSK receiver operating at 1Gb/s rate and P_R = -48.39 dBm incorporating coherent demodulation scheme with different LDPC codes.

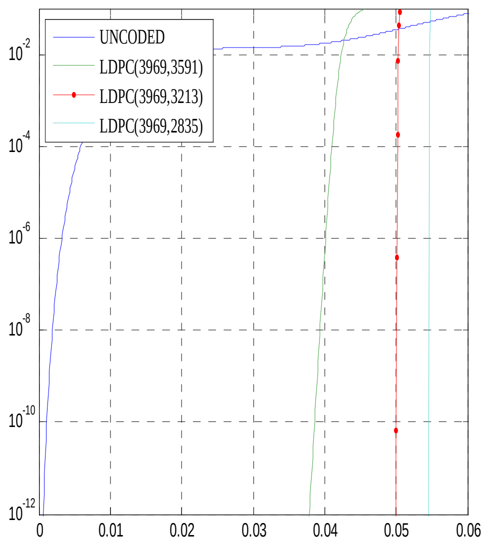

Fig. 14 BER versus $\Delta v T_b$ for heterodyne optical QPSK receiver operating at 1Gb/s rate and P_R = -46.85 dBm incorporating differential demodulation scheme with different LDPC codes.

requirement in heterodyne optical receivers. The results indicate clearly that LDPC code offers higher *CG* and *LRF* when implemented with QPSK system compared with BPSK and DPSK systems.

LRF of 83.28 and 67.14 can be obtained by employing LDPC (3969, 3213) code in QPSK system with differential demodulation and QPSK system with coherent demodulation, respectively, and this results are to be compared with 3.55 and 3.05 for BPSK and DPSK systems, respectively.

The result indicate that BPSK with coding gives high improvement in BER compare with other modulation scheme.

Table 3. Summarized results related to a 1Gb/s heterodyne optical receivers operating under LDPC codes at BER = 10 12

Receiver types	Coding schemes	Code rate	P_R (dBm)*	*CG* (dB)	(ΔvT_b) **	*LRF*
BPSK	Uncoded	-	-48.65	-	2.86×10^{-3}	-
	LDPC (3969, 3591)	0.9052	-55.32	6.67	9.81×10^{-3}	3.43
	LDPC (3969, 3213)	0.8095	-59.03	10.38	10.1×10^{-3}	3.53
	LDPC (3969, 2835)	0.7142	-60.21	11.56	10.15×10^{-3}	3.55
	BCH (255, 223)	0.8745	-51.43	2.78	8.59×10^{-3}	2.74
	RS(255,239)+ RS(255, 239)	0.8784	-54.15	5.50	10.2×10^{-3}	3.24
DPSK	Uncoded	-	-47.55	-	8.69×10^{-3}	-
	LDPC (3969, 3591)	0.9052	-50.77	3.22	25.71×10^{-3}	2.96
	LDPC (3969, 3213)	0.8095	-51.14	3.59	26.25×10^{-3}	3.02
	LDPC (3969, 2835)	0.7142	-51.28	3.73	26.51×10^{-3}	3.05
	BCH (255, 223)	0.8745	-50.62	2.06	20.6×10^{-3}	1.89
	RS(255,239)+ RS(255, 239)	0.8784	-51.08	2.52	23.1×10^{-3}	2.5
QPSK with coherent demodulation	Uncoded	-	-48.39	-	3.71×10^{-4}	-
	LDPC (3969, 3591)	0.9052	-59.44	11.09	17.9×10^{-3}	48.35
	LDPC (3969, 3213)	0.8095	-60.94	12.09	25×10^{-3}	67.14
	LDPC (3969, 2835)	0.7142	-62.09	13.40	31.1×10^{-3}	82.28
	BCH (255, 223)	0.8745	-54.43	6.04	48.3×10^{-4}	13.01
	RS(255,239)+ RS(255, 239)	0.8784	-59.51	11.12	182×10^{-4}	49.06
QPSK with differential demodulation	Uncoded	-	-46.93	-	6.11×10^{-4}	-
	LDPC (3969, 3591)	0.9 052	-57.50	10.57	380×10^{-4}	63.0
	LDPC (3969, 3213)	0.8095	-59.00	12.07	500×10^{-4}	83.3
	LDPC (3969, 2835)	0.7142	-59.81	12.89	546×10^{-4}	91.0
	BCH (255, 223)	0.8745	-52.77	5.84	57.3×10^{-4}	9.55
	RS(255,239)+ RS(255, 239)	0.8784	-57.75	10.82	401×10^{-4}	66.80

***** This value (ΔvT_b) is chosen to yield a 1 dB power penalty to the uncoded systems.

****** Maximum allowable value of (ΔvT_b) which ensures a 1dB power penalty at BER =10^{-12} for the coded systems

5. Conclusions

This paper addresses the possibility of using LDPC codes to relax laser linewidth requirement in heterodyne optical receivers. The results indicate clearly that LDPC code offers higher *CG* and *LRF* when implemented with QPSK system compared with BPSK and DPSK systems.

LRF of 83.28 and 67.14 can be obtained by employing LDPC (3969, 3213) code in QPSK system with differential demodulation and QPSK system with coherent demodulation, respectively, and this results are to be compared with 3.55 and 3.05 for BPSK and DPSK systems, respectively.

The result indicate that BPSK with coding gives high improvement in BER compare with other modulation scheme.

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استخدام نظام التصحيح نوع مراقب تعادلِ ذو كثافةِ منخفضةِ لتقليل تاثير عرض حزمة أليزر في أنظمة الاتصالات البصرية

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الخلاصة:

تعتمد خصائص أنظمة الاتصال بالألياف الضوئية المتزامنة وبشكل رئيسي على الضوضاء الزاوية المتولدة من الانبعاث الذاتي لليزر شبه الموصل و هذا ما يؤدي إلى تولد عرض في حزمة الترددات المتولدة من مصدر الليزر بدل من التردد الواحد. في هذه النشرية تم دراسة عملية تحسين أنظمة الاتصال البصري وذلك باستخدام نظام التصحيح الأمامي (forward error correcting code) نوع (Low- Density Parity-Check (LDPC) code) لتحسين المستلمات البصرية وذلك بتقليل تأثير عرض الحزمة الترددية لليزر. وتم تطبيق نظام التصحيح هذا على أنظمة الاتصالات للمستلمات من نوع (BPSK, DPSK, and QPSK) و الذي يعمل باستخدام مولد ليزري ذو عرض حزمه ترddيه محدد.

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