

Phase- and Polarization-Diversity Coherent Optical Techniques

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Abstract—Progress in phase- and polarization-diversity coherent optical techniques has led to impressive receivers able to tolerate wide laser linewidth and large polarization fluctuations. This paper reviews the advantages and the drawbacks of diversity receivers, and discusses recent experimental and theoretical research results.

I. INTRODUCTION

INTENSIVE RESEARCH in coherent optical communications [1] showed that several serious problems have to be resolved before coherent communications becomes practical. Four of these problems can be relieved via the diversity approach: limited receiver bandwidth; phase noise of the transmitter and local oscillator lasers; limited tuning range of local oscillators; and polarization fluctuations of the received signal.

Limited receiver bandwidth may become a limitation at high bit rates. In a coherent receiver, the frequency response of the photodetector and of the intermediate frequency (IF) electronics must extend to at least¹ $IF + R_b$, where R_b is the bit rate. Thus, at high bit rates it is desirable to keep IF as small as possible, and $IF = 0$ Hz (homodyning) is advantageous from this point of view. Unfortunately, if $IF = 0$ Hz, then the photocurrent is proportional to $\cos \phi$, where ϕ is the combined phase noise of the transmitter and local oscillator. If ϕ is left random (no phase locking), then the current is equal to zero when $\phi = 90^\circ$, rendering the system useless. If ϕ is kept small by means of phase locking, then the best receiver sensitivity is achieved: 9 photons/bit for phase shift keying (PSK) and 18 photons/bit for amplitude shift keying (ASK). However, phase-locked receivers require extremely narrow linewidth lasers: $\Delta\nu/R_b$ must be less than 3×10^{-4} for nonlinear loops [2] and 6×10^{-6} for balanced loops [3]. Since such linewidths are not currently available with semiconductor lasers, phase-locked homodyne receivers are not practical with present-day injection lasers.

The problem of phase noise/laser linewidth can be solved via several different approaches [4], [5]: a) an ASK heterodyne system, if suitably optimized (i.e., having a

receiver with a sufficiently large IF and sufficiently broad IF bandwidth), can tolerate extremely large amounts of phase noise; when the IF linewidth $\Delta\nu_{IF}$ is equal to the bit rate, the theoretical sensitivity penalty is less than 2 dB; b) a frequency shift keying (FSK) heterodyne system has the same property if the receiver is similarly optimized and the frequency deviation is sufficiently large; and c) a phase-diversity homodyne system has the same property, too, if the receiver is optimized and ASK or FSK formats are used. The advantage of phase-diversity homodyne systems with respect to heterodyne systems is that the phase-diversity receiver provides for phase-noise tolerance and baseband operation at the same time. To understand how phase-diversity receivers overcome phase noise, consider a three-branch phase-diversity receiver. A 3×3 coupler mixes the signal and the local oscillator (LO) fields in such a way that the detected currents are proportional to $\cos \phi$, $\cos(\phi + 120^\circ)$ and $\cos(\phi + 240^\circ)$, respectively, where ϕ is the phase noise. Simple trigonometry shows that if these currents are squared and added together, then the result is independent of ϕ .

Phase-diversity receivers have an important additional advantage when used for multichannel applications. The channel spacing in high-speed multichannel coherent systems increases linearly with the increase of the IF [6]. In turn, the required tuning range of the local oscillator is proportional to the channel spacing, and may present a difficult problem for multichannel systems. Phase-diversity receivers lead to a smaller required LO tuning range than conventional heterodyne receivers and are advantageous from this point of view.

Another problem with coherent receivers is associated with the single-mode fiber used to convey the transmitted optical signals. Large amounts of conventional single-mode fiber will be installed before coherent systems will be ready for deployment. Thus, it is likely that conventional (not single-polarization) single-mode fibers will be used to carry the signals to coherent receivers. One problem arising from this situation is that the polarization of the signal arriving at the receiver is unknown and is changing in time. As a result, the amplitude of the detected current in a single-detector receiver changes in time, too, introducing severe fading and even complete signal loss under certain conditions. Polarization-diversity receivers offer an elegant solution to this problem:

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¹More if the impact of the phase noise is taken into account.

they generate two demodulated signals stemming from two orthogonal polarizations of the received signal. The sum of the two is virtually independent of the state of polarization of the received signal.

The purpose of this paper is to review the state of the art in coherent diversity receivers; the material of this paper was partially presented in [7]. The paper is organized as follows. Principles of diversity receivers are reviewed in Section II. Phase- and polarization-diversity research results are discussed in Sections III and IV, respectively. Section V contains the conclusions of this study.

II. PRINCIPLES OF DIVERSITY RECEIVERS

A. General Receiver Structure

First, consider a conventional single-detector coherent receiver. Assuming the ASK modulation format and a linear state of polarization (SOP) of both the LO and the received signal, the photodetector current is given by

$$\text{Homodyne: } i = dA \cos \theta \cos \phi \quad (1)$$

$$\text{Heterodyne: } i = dA \cos \theta \cos (\omega t + \phi) \quad (2)$$

where d is the nonreturn-to-zero (NRZ) binary signal (it is equal to either zero or one); $A \equiv R \sqrt{P_S P_{LO}}$; R is the detector responsivity; P_S is the signal power on the detector surface; P_{LO} is the local oscillator power on the detector surface; θ is the angle between the SOP's of the signal and local oscillator; ϕ is the combined phase noise of the transmitter and local oscillator; and ω is the intermediate frequency in radians per second.

The phase noise ϕ is a nonstationary Wiener process:

$$\phi = \int_{-\infty}^t \dot{\phi}(t_1) dt_1, \quad \text{units: rad}$$

where $\dot{\phi}(t)$ is the instantaneous angular frequency noise in radians per second. The frequency noise $\dot{\phi}(t)$ can be modeled [1]–[5] as a white zero-mean Gaussian random process with the power spectral density (PSD) given by

$$S_{\dot{\phi}}(f) = 2\pi\Delta\nu, \quad \text{units: rad}^2 \cdot \text{Hz}, \quad -\infty < f < \infty \quad (3)$$

where $\Delta\nu$ is the full width half maximum (FWHM) linewidth at the "IF," i.e.,

$$\Delta\nu = \Delta\nu_T + \Delta\nu_{LO} \quad (4)$$

where $\Delta\nu_T$ and $\Delta\nu_{LO}$ are the linewidths of the transmitter and local oscillator, respectively. The PSD shape (3) implies the Lorentzian laser lineshape [1]–[5]. We note that even though the Lorentzian lineshape has been observed experimentally [8], $S_{\dot{\phi}}(f)$ may be different from (3) in two respects: first, the flicker noise at low frequencies; and second, a peak at the relaxation frequency of the laser—which is typically several gigahertz. The impact of both phenomena on the system performance will be small if

$$f_{FN} \ll R_b \ll f_p \quad (5)$$

where f_{FN} is the frequency range of the flicker noise, R_b is the system bit rate, and f_p is the frequency of the phase noise peak. In this paper, we assume that (5) is satisfied; further research will be needed to study the system performance if (5) is not satisfied.

Inspection of (1) reveals that if either θ or ϕ are equal to 90° , then the homodyne current is equal to zero, and all the data is lost. In case of heterodyning (2), the phase noise is not a problem if the intermediate frequency and the bandwidth of the IF filter are sufficiently large [5], [9], [10]: $\cos(\omega t + \phi)$ will go through a maximum at least once per bit, thus ensuring data recovery. However, polarization misalignment remains a problem.

Both phase-noise and polarization-fluctuation problems can be resolved with diversity receivers (Fig. 1). In all such receivers, the signal and the LO fields are combined to produce two or more² output fields $\{E_k\}$. The fields $\{E_k\}$ are linear combinations of E_S and E_{LO} where E_S is the signal field and E_{LO} is the local oscillator field; the specific relationship between $\{E_k\}$, E_S and E_{LO} is discussed in Section II-B. Following [11] and [12], we call an optical device used to generate the fields $\{E_k\}$ from the fields E_S and E_{LO} "a hybrid." The fields $\{E_k\}$ are detected separately, and the resulting currents $\{i_k\}$ are sent to the signal processing unit for data recovery. The structure of the signal processing unit is discussed in Section II-C.

B. Optical Hybrids: Function and Implementation

1. Phase-Diversity Homodyne Receivers:

a. *Definition of optical hybrids:* In phase-diversity homodyne receivers, the SOP's of the received signal and of the LO must be the same ($\theta = 0$); this is achieved by means of polarization control devices. Assuming that the SOP's of both input fields are linear, the output fields of symmetric hybrids should, by definition, be equal to the following:

In the case of two branches:

$$E_1 = \sqrt{0.5L} (E_S + E_{LO}) \quad (6a)$$

$$E_2 = \sqrt{0.5L} (E_S + jE_{LO}). \quad (6b)$$

In case of three or more branches:

$$E_k = \sqrt{L} [E_S + E_{LO} \exp(j360^\circ(k-1)/N)] / \sqrt{N} \quad (7)$$

$$1 \leq k \leq N$$

where E_k is the output field of the k th branch, $-10 \log L$ is the loss in decibels, and $N \geq 3$ is the number of branches. An optical device satisfying (6) is called the 90° hybrid; an optical device satisfying (7) with $N = 3$ is called the 120° hybrid. When the hybrid output fields are detected, the following signal currents appear (we assume, for simplicity, the ASK modulation format):

In case of two branches:

$$i_1 = LdA \cos \phi \quad (8a)$$

$$i_2 = LdA \sin \phi. \quad (8b)$$

²An exception is described in [35], see discussion in Section IV of this paper.

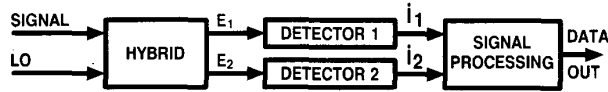


Fig. 1. A general block diagram of diversity coherent receivers. The case of two detectors is illustrated; many useful configurations use two, three, or four photodetectors.

In case of three or more branches:

$$i_k = \frac{2L}{N} dA \cos [\phi + 360^\circ (k - 1)/N] \quad (9)$$

where $A \equiv R \sqrt{P_S P_{LO}}$, and the powers are measured at the hybrid inputs. Inspection of (8) and (9) reveals that if the currents $\{i_k\}$ are squared and added together, the result is completely independent of the phase noise and is equal to

$$i_{TOT} = \sum i_k^2 = \frac{2L^2}{N} R^2 d^2 P_S P_{LO}. \quad (10)$$

b. Number of branches: Expression (10) shows that the signal i_{TOT} is inversely proportional to the number of branches N , so that the larger N , the larger the local oscillator power needed to suppress the receiver noise. Therefore, it is important to keep the number of branches small, and $N = 2$ is most advantageous from this point of view; the $N = 2$ diversity receiver is analogous to the optimum receiver structure for a signal with random phase in Gaussian noise [13].

Unfortunately, two-branch receivers have two serious disadvantages:

- a) 90° hybrids needed for two-branch receivers are inherently lossy devices: the loss of a symmetric 90° hybrid cannot be smaller than 2.3 dB [14], and all known implementations (see the next subsection) have a 3-dB loss; and
- b) receiver imperfections (such as polarization misalignment [15], local oscillator intensity noise [16], and reflections [17]) have stronger effect on two-branch receivers than on receivers with $N > 2$, and lead to larger sensitivity penalties.

Three-branch receivers ($N = 3$) require more electronic signal processing hardware and a larger local oscillator power than two-branch receivers. In return, three-branch receivers offer two advantages:

- a) 120° hybrids needed for three-branch receivers have no intrinsic loss and can be implemented as low-loss 3×3 couplers; and
- b) three-branch receivers are more tolerant to receiver imperfections than two-branch receivers.

Four-branch receivers ($N = 4$) require four photodetectors which generate four photocurrents: $i_1 = A \cos \phi$; $i_2 = A \sin \phi$; $i_3 = -A \cos \phi$; and $i_4 = -A \sin \phi$. Consider the quantities $i_{13} \equiv i_1 - i_3 = 2A \cos \phi$ and $i_{24} \equiv i_2 - i_4 = 2A \sin \phi$, which can be easily generated by the back-to-back connection of the photodiodes (1 with 3 and 2

with 4, respectively), the so-called balanced configuration. If i_{13} and i_{24} are squared and added together, the result is independent from the phase noise, as in other phase-diversity receivers. Thus, a balanced four-branch receiver³ combines the best features of both phase-diversity and balanced receivers: it protects the system against the phase noise, the intensity noise of the local oscillator, and reflections in the local oscillator path (see Section III-D). Such a receiver needs four photodetectors, but only two sets of RF hardware (amplifiers, filters, and squarers); in other words, it has four optical branches and two electronic branches. A hybrid for such a receiver has no intrinsic loss, so that both signal and LO powers are used efficiently. However, a four-branch receiver requires a larger LO power than either two-branch or three-branch receivers.

The foregoing discussion indicates that two-, three-, and four-branch receivers have serious advantages and warrant further investigation. The currently available knowledge suggests that receivers with larger number of branches ($N \geq 5$) do not offer any additional advantages; judging from (10), their performance is expected to be worse than that of their counterparts with $N < 5$.

c. Implementation of optical hybrids for two-branch receivers: To implement a hybrid for two-branch receivers (expression (6)), the following four techniques have been proposed.

1) In the bulk optics implementation shown in Fig. 2(a) (after [11]), the output fields E_1 and E_2 have orthogonal polarizations; hence, the phase relationship between the signal and the LO components of E_1 and E_2 can be controlled independently by the input polarization controllers.

2) In the all-single-mode-fiber implementation shown in Fig. 2(b) (after [12]), the two fiber-optic polarization analyzers are orthogonal to each other, and the principle of operation is similar to 1), except for two differences (see discussion below); a more detailed explanation of the principle of operation of this hybrid is contained in Appendix A.

3) In an integrated-optics implementation shown in Fig. 2(c) (after [2]), differential time delay is employed to generate the necessary phase shift.

4) In an overmoded fiber implementation shown in Fig. 2(d) (after [18]), the necessary phase shift is obtained using the phase shift between the two light modes excited in the device.

From the four above implementations, only the first two have been used in system experiments until now. The main advantage of the first approach is that the losses can be allocated at will between the signal and the LO paths; if a 3-dB coupler is used, then both signal and LO suffer a 3-dB loss. The main advantage of the second approach is that the entire processing occurs in the single-mode-fiber domain, so that insertion losses and alignment prob-

³Balanced processing can, in principle, be used in three-branch receivers, too.

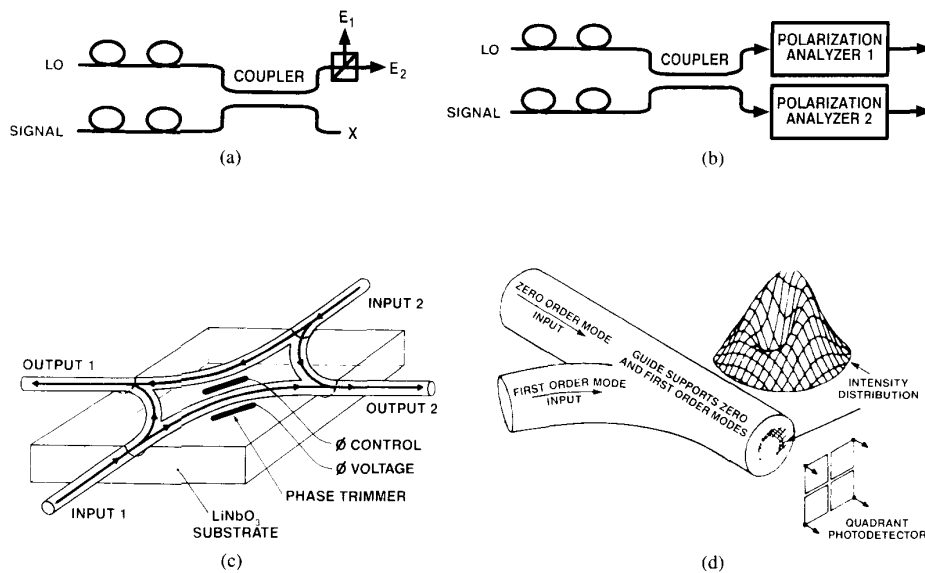


Fig. 2. Implementation of 90° hybrids. (a) A bulk optics implementation, after [11]. (b) An all-fiber implementation, after [12]. (c) An integrated-optics implementation, after [2]. (d) An all-fiber implementation, after [18].

lems associated with bulk optics are eliminated; since the device is symmetric, both signal and LO suffer a 3-dB loss. Note that the loss of a symmetric 90° hybrid cannot be smaller than 2.3 dB [14].

d. Implementation of optical hybrids for three-branch receivers: For three-branch receivers, ideal symmetric lossless 3×3 couplers can, in principle, be used as 120° hybrids. Unfortunately, practical 3×3 couplers can suffer from several imperfections, including loss, nonuniform distribution of input power among the output ports, and polarization dependence. As a result of these imperfections, practical 3×3 couplers are generally not guaranteed to act as 120° hybrids. One can, however, find selected 3×3 couplers that are sufficiently close to the ideal performance to be used as 120° hybrids [19].

2. Polarization-Diversity Heterodyne Receivers: A bulk-optics hybrid for a polarization-diversity receiver can be implemented as shown in Fig. 2(a). The polarization controller on the LO path is adjusted to produce equal LO powers at the outputs E_1 and E_2 ; it does not have to be readjusted as the signal polarization varies in time. The polarization controller on the signal path is not needed and can be omitted. An all-fiber hybrid can be implemented as shown in Fig. 2(b). The two fiber-optic polarization analyzers are orthogonal to each other, and the polarization controller on the LO path is adjusted to produce equal LO powers at the outputs E_1 and E_2 . The polarization controller on the signal path is not needed and can be omitted. The main advantage of the first approach is that the losses can be allocated at will between the signal and the LO paths; if a 3-dB coupler is used, then both signal and LO suffer a 3-dB loss.

The main advantage of the second approach is that the entire processing occurs in the single-mode-fiber domain, so that insertion losses and alignment problems associated with bulk optics are eliminated; if a 3-dB coupler and polarization analyzers are used, then both signal and LO suffer a 3-dB loss. The output currents generated by the two detectors are identical in both cases; for the ASK format, the currents are

$$i_1 = \sqrt{2} L R d \sqrt{P_S P_{LO}} \cos \theta \cos (\omega t + \phi) \quad (11a)$$

$$i_2 = \sqrt{2} L R d \sqrt{P_S P_{LO}} \sin \theta \cos (\omega t + \phi + \psi) \quad (11b)$$

where the signal was for simplicity assumed to be linearly polarized with an angle θ with respect to polarization of E_1 , and ψ is an arbitrary angle. The currents i_1 and i_2 are squared and added together. If the intermediate frequency and the IF bandwidth are much larger than the signal bandwidth stemming from the modulation, phase noise, and polarization fluctuations, then the lowpass part of the result is (see Appendix B for a detailed derivation):

$$i_{TOT} = \text{lowpass} \{i_1^2 + i_2^2\} = L^2 R^2 d P_S P_{LO} \quad (12)$$

where $\text{lowpass} \{x\}$ denotes the lowpass part of x . Note that the result is independent both of polarization (of θ) and of the phase noise (of ϕ); it can be shown that (12) is valid for an arbitrary polarization of the signal.

We emphasize that a hybrid for polarization-diversity receivers does not have to have any intrinsic loss whatsoever, and L can, in principle, be equal to unity. Indeed, some implementations [20] have a 0-dB intrinsic loss (their actual implementation loss is, of course, larger than 0 dB); see Section IV-B for more detailed discussion.

C. Signal Processing for Diversity Receivers

For most diversity receivers the signal processing consists of demodulation, summation, filtering, and threshold comparison, as shown in Fig. 3. The signal processing configuration of Fig. 3 is sufficient for receivers implementing either phase diversity only or polarization diversity only algorithms; more sophisticated configurations are needed if both these algorithms are implemented in a single receiver, or if either of them is used in conjunction with such functions as LO intensity noise cancellation via balanced detection.

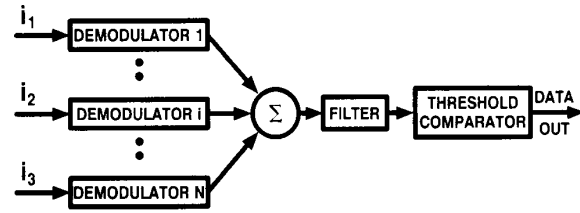


Fig. 3. General block diagram of signal processing for diversity receivers.

III. PHASE-DIVERSITY RECEIVERS: RECENT RESEARCH RESULTS

A. Theory

Fig. 4 shows a recently proposed signal processing configuration for phase-diversity receivers [4] which consists of wide-band filters, rectifiers, and a narrow-band filter. The most critical components of this structure are the wide-band filters. If the bandwidth of these filters is too narrow, then the input voltage is distorted by the filters, and the input phase noise is converted to intensity noise at the decision gate, leading to a bit error rate (BER) floor or a severe sensitivity penalty. If the bandwidth of the wide-band filters is too large, then too much noise is collected; much of the additional noise is rejected by the narrow-band filter, leading to a moderate sensitivity penalty. The optimum bandwidth is approximately $\sqrt{R_b^2 + (6\Delta\nu)^2}$, as shown in [4], [5], [9], [10].

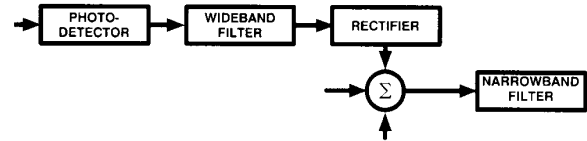


Fig. 4. A phase-diversity receiver employing wide-band IF filters, rectifiers, and a narrow-band baseband filter, after [4].

Fig. 5 shows sensitivity penalty versus $\Delta\nu T$, where $\Delta\nu$ is the IF linewidth and T is the bit duration [4]. Receiver bandwidth is assumed to be optimum as discussed above, so that the bandwidth increases as we move along the horizontal axis from small linewidth to large linewidth. Note that the penalty increases rapidly for $0 < \Delta\nu T < 10$ percent; in this region, postdetection filtering does not help, so that this segment of the penalty curve corresponds to receivers with predetection filtering only. For $\Delta\nu T > 10$ percent, the curve corresponds to receivers employing both predetection and postdetection filtering since they have better performance in this region. Inspection of Fig. 5 reveals that the optimized structure is extremely tolerant with respect to phase noise: the sensitivity penalty is less than 2 dB even when the linewidth is of the same order of magnitude as the bit rate. It is interesting to note that only a short time ago it was widely believed that the linewidth must be much smaller than the bit rate in coherent systems.

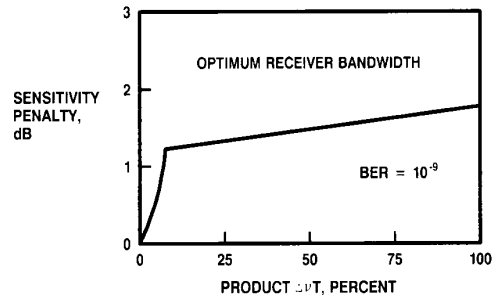


Fig. 5. Sensitivity penalty versus $\Delta\nu T$ for optimized phase-diversity receivers with wide-band IF filters, after [4]. $\Delta\nu$ is the IF linewidth, and T is the bit duration.

TABLE I
PROPERTIES OF VARIOUS COHERENT RECEIVERS¹

Receiver Type	Sensitivity	Linewidth ²	Approximate Receiver Bandwidth ³
Heterodyne	42 photons/bit	One bit-rate	$\sim 2\sqrt{R_b^2 + (6\Delta\nu)^2}$
Phase-locked Homodyne	18 photons/bit	$3 \cdot 10^{-4} \times$ bit rate	$\sim R_b$
Phase Diversity Homodyne	42 photons/bit	One bit-rate	$\sim \sqrt{R_b^2 + (6\Delta\nu)^2}$

¹ The table assumes that the ASK modulation format with non-synchronous demodulation is used; many of the results are applicable to wide-deviation FSK, too.

² For 2 dB penalty.

³ Notation: R_b = bit rate, $\Delta\nu$ = IF linewidth.

Table I compares properties of several types of coherent receivers. Inspection of Table I reveals that phase-diversity homodyne receivers have the same sensitivity and linewidth requirements as heterodyne receivers, but require a much smaller bandwidth. Phase-locked homodyne receivers [2], [3] have better sensitivity, but impose much more stringent requirements on the laser linewidth. The main disadvantages of phase-diversity receivers are

their complexity and sensitivity to implementation imperfections.

In many phase-diversity experiments, the IF is not equal to zero; instead, the IF of several megahertz [14] or even $0.5 R_b$ [15] may be employed. This approach is called quasi-homodyne or intradyne. Intradyne phase-diversity receivers have the same sensitivity and linewidth requirements as homodyne phase-diversity receivers, but require a somewhat larger bandwidth. The design of automatic frequency control (AFC) loops is simpler for intradyne receivers than for homodyne receivers.

Reflections of light at discontinuities in connectors, splices, and other components are a major potential source of degradation in phase-diversity systems. A theory of this phenomenon is only beginning to evolve [17]; however, it is already clear that the following three mechanisms are important [17], [23], [28], [30]:

a) Reflection of the incoming power backwards into the laser [30] can cause linewidth broadening, frequency hopping, reflection-induced noise, and other degradations; this effect can be minimized by incorporating an optical isolator in the laser package.

b) Refraction index discontinuities in the signal transmission path can form Fabry-Perot cavities which convert the signal phase noise into high levels of intensity noise downstream of the isolator [23]; this effect leads to roughly the same penalty in both homodyne and in heterodyne systems; to ensure reliable system performance, it is desirable to suppress all reflections in the signal path to well below -20 dB for both types of systems [17], [28].

c) Refraction index discontinuities in the LO path can also form Fabry-Perot cavities which convert the LO phase noise into high levels of intensity noise, similar to item b) above; however, the impact of this phenomenon on homodyne phase-diversity receivers is vastly different from its impact on heterodyne receivers [17]. The reason is that the spectrum of the dominant component of the resulting intensity noise has a low-pass shape, i.e., is concentrated around zero frequency. Thus, it is substantially attenuated in heterodyne systems if the IF is sufficiently large. Homodyne systems, however, operate in the baseband, and have no mechanism for protection against this noise. As a result, a well-designed heterodyne system can tolerate reflections of up to -20 dB [28], whereas a phase-diversity homodyne system requires suppression of reflections to well below -35 dB [17].

B. Experimental Research

1. *Two-Branch Receivers:* The first phase-diversity experiment [15] used a two-branch receiver. Both signal and local oscillator fields were derived from the same HeNe laser, leading to a very narrow IF linewidth. The differential phase shift keying (DPSK) modulation format was used at 140 Mbit/s; the hybrid was implemented using a bulk optics polarizing beamsplitter; and the demodulators were implemented using a standard delay-and-

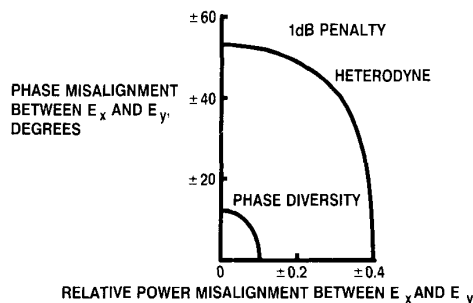


Fig. 6. Tolerance of two types of coherent receivers to polarization fluctuations, after [15].

multiply approach. In later experiments, the same group demonstrated a two-laser system with a separate LO, and demonstrated the feasibility of ASK reception. One interesting result is shown in Fig. 6 illustrating the tolerance to polarization misalignment of conventional heterodyne and two-branch phase-diversity homodyne receivers. Inspection of Fig. 6 reveals that two-branch phase-diversity receivers are less tolerant than conventional heterodyne receivers: they incur a 1-dB penalty when polarization deviates by $\sim 10^\circ$ from the optimum value, while conventional receivers can tolerate much larger polarization fluctuations. The reason for this phenomenon is that the 90° hybrid used relies on the correct SOP for proper operation (see Section II-B), so that once the SOP deviates from the design value, the phase relationship between the branches is disturbed, too.

A wide-linewidth two-branch phase-diversity receiver employing a distributed feedback (DFB) laser was reported in [14], [16], [21]. In initial experiments, both signal and LO fields were derived from the same DFB laser with a 40-MHz linewidth. The LO field was delayed by 7 km of single-mode fiber so that the signal and the LO fields were practically independent at the receiver, leading to an IF linewidth of 80 MHz. ASK modulation format was used at 150 Mbit/s; an all-fiber optical hybrid [12] was used; and the processor of Fig. 4 was used for data recovery. Fig. 7 shows experimental and simulated eye diagrams corresponding to this configuration. Initial experiments led to a noisy eye (Fig. 7 (a)). To understand the reasons for eye closure, the receiver was simulated using SYSTID [22]. The simulation program included shot noise, phase noise, and LO intensity noise; the resulting eye diagram (Fig. 7 (b)) closely resembles the experimental eye. Elimination of the noise sources one by one reveals that the dominant noise source is the LO intensity noise; eye diagram (d) shows the predicted improvement when the relative intensity noise (RIN) is improved. Experiments revealed [23] that most of the intensity noise was generated by the phase-noise-to-intensity-noise conversion due to reflections in single-mode fiber components. Suppression of reflections led to a clear eye (Fig. 7 (c)), in a good agreement with the simulation result (Fig. 7 (d)). A subsequent theoretical study [17] showed that this receiver is highly sensitive to reflections

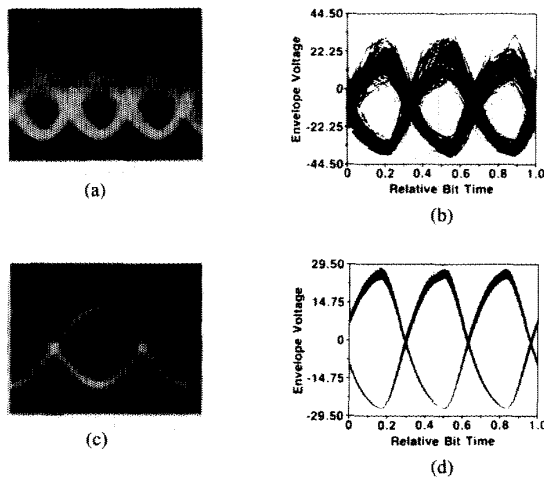


Fig. 7. Experimental and simulated eye diagrams of a wide-linewidth two-branch phase-diversity receiver, after [14], [16], [21]. (a) Relative intensity noise, $RIN = -125$ dB/Hz (experimental). (b) Relative intensity noise, $RIN = -125$ dB/Hz (simulation). (c) Relative intensity noise, $RIN \leq -145$ dB/Hz (experimental). (d) Relative intensity noise, $RIN \leq -145$ dB/Hz (simulation).

in the LO path; for small penalty, the reflections have to be suppressed to well below -35 dB [17].

Later, the same receiver was tested in a two-laser experiment: one DFB laser was used as the transmitter, and another DFB laser was used as the LO [14]. High sensitivity (-56.5 dBm) and a simple AFC loop were demonstrated.

Fig. 8 shows the sensitivity penalty versus LO power for two-branch and three-branch phase-diversity receivers [14]. The general behavior of the curves is the same for both receivers: for small LO powers, the penalty is large due to incomplete suppression of thermal noise; for large LO powers, the penalty is also large due to increasing impact of the LO intensity noise. An optimum is achieved for an intermediate value of the local oscillator power, P_{LO} . The optimum P_{LO} is larger for three-branch receivers than for two-branch receivers because three-branch receivers have to split P_{LO} between a larger number of p-i-n-FET modules. Note that for large P_{LO} three-branch receivers suffer less penalty than two-branch receivers. The reason is that the intensity noise at the decision gate in two-branch receivers is proportional to

$$n(t)[\cos \phi + \sin \phi] \quad (13)$$

whereas in three-branch receivers this noise is proportional to

$$n(t)[\cos \phi + \cos(\phi + 120) + \cos(\phi + 240^\circ)] \quad (14)$$

where $n(t)$ is the intensity noise at the output of the p-i-n-FET modules. It is easy to see that while the term in square bracket in (14) is identically equal to zero, the term in square brackets in (13) depends on ϕ , and can be as large as $\sqrt{2}$.

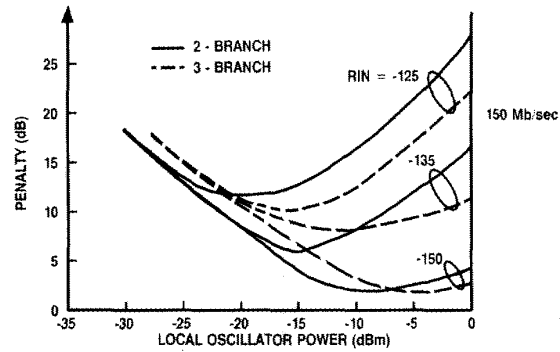


Fig. 8. Receiver sensitivity penalty versus local oscillator power for two types of phase-diversity receivers [14].

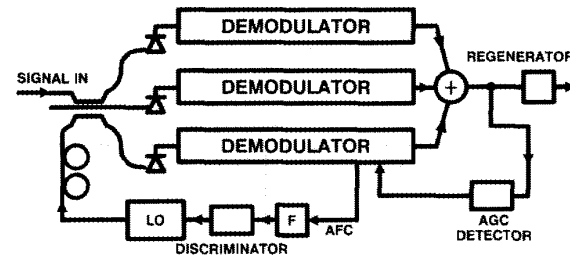


Fig. 9. Experimental setup used in three-branch phase-diversity experiments, after [24]–[26]. HeNe/DFB lasers; ASK, DPSK, single-filter FSK; 140, 320, 600 Mbit/s.

2. Three-Branch Receivers: Fig. 9 shows an experimental setup used in a wide variety of experimental studies of three-branch phase-diversity receivers [19], [24]–[26]. These are the most fully engineered experiments to date as they include both automatic gain control (AGC) and AFC loops. Using this setup, experiments were conducted with both HeNe and DFB lasers at bit rates of 140, 320, and 680 Mbit/s employing ASK, DPSK, and single-filter FSK modulation formats. In the latest wide-linewidth experiments [25], [26], the authors demonstrated multichannel operation and confirmed experimentally BER floor elimination using a wide-band filter in the IF demodulation path. Fig. 10 shows BER versus received optical power for a single-filter FSK system with DFB lasers; the bit rate is 140 Mbit/s and the linewidth-to-bit-rate ratio is 0.36. The data for Fig. 10(a) was taken with a reduced receiver bandwidth, while the data for Fig. 10(b) was taken with full receiver bandwidth. Comparison of Fig. 10(a) with Fig. 10(b) reveals that a wide IF bandwidth does eliminate the BER floor as predicted by theory [4].

A compact 565-Mbit/s DPSK three-branch phase-diversity receiver was reported in [27]; the authors used HeNe lasers as light sources in both transmitter and receiver.

C. Summary: Phase-Diversity Studies

Table II summarizes phase-diversity research results. Inspection of Table II reveals impressive progress in

TABLE II
PHASE-DIVERSITY STUDIES

Ref	[15]	[19]	[24]	[4]	[25]	[26]	[16]	[21]	[18]	[14]	[17]
Journal or conference	Electronics Letters	Electronics Letters	J. L. T.	J. L. T.	Electronics Letters	ECOC-87	Electronics Letters	Electronics Letters	ECOC-87	J. L. T.	Electronics Letters
Authors	Hodgkinson et al.	Davis et al.	Davis et al.	Kazovsky et al.	Pettitt et al.	Davis et al.	Kazovsky et al.	Welter Kazovsky	Schneider Pietzsch	Kazovsky et al.	Kazovsky
Firm	BTRL	STC	STC	Bellcore	STC	STC	Bellcore	Bellcore	Siemens	Bellcore	Bellcore
Number of branches	2	3	3	Arbitrary	3	3	2	2	3	2	3
Wavelength	1.5 μm	1.5 μm	1.5 μm		1.5 μm	1.5 μm	1.5 μm	1.5 μm	1.5 μm	1.5 μm	1.5 μm
Bit Rate	140 Mb/sec	680 Mb/sec	320 Mb/sec	Theory	140 Mb/sec	140 Mb/sec	140 Mb/sec	150 Mb/sec	565 Mb/sec	150 Mb/sec	Theory
IF	0	6.8 MHz	3.2 MHz		70 MHz	70 MHz	0	0	5 MHz	< 75 MHz	
Modulation	DPSK	ASK	DPSK Single-filter FSK	ASK FSK	Single-filter FSK	Single-filter FSK	ASK	ASK	DPSK	ASK	ASK FSK
Sensitivity	-44 dBm	-42 dBm	-47.5 dBm	-	-51.8 dBm	-51.8 dBm	-	-51 dBm	-40 dBm	-56.5 dBm	-
Linewidth/Bit Rate	0	~ 0	~ 0	~ 1	0.36	0.36	0.57	0.53	~ 0	0.61	Up to 1
Lasers	One: HeNe	Two: HeNe/ Injection Locked	Two: HeNe	Wide linewidth	Two: DFB	Two: DFB	One: DFB	One: DFB	Two HeNe	Two: DFB	Wide Linewidth
Technical Significance	First 2-branch experiment	First 3-branch experiment; first experiment with 2 lasers	First 3-branch DPSK experiment	First prediction of Wide linewidth tolerance	First 3-branch experiment with semiconductor lasers	First multichannel experiment with diversity receivers	First 2-branch experiment with semiconductor lasers	First 2-branch BER measurements with semiconductor lasers	Compact receiver	First 2-branch experiments with 2 semiconductor lasers	First study of the impact of reflections

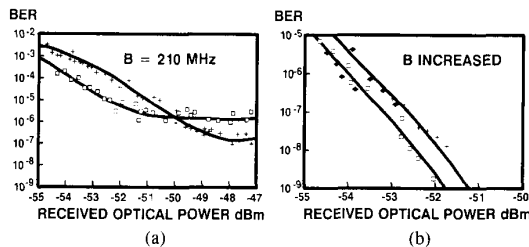


Fig. 10. Elimination of BER floor using wide-band IF filtering after [25]. (a) Insufficient receiver bandwidth. (b) Sufficient receiver bandwidth.

phase-diversity technologies from the early zero-line-width demonstrations to recent experiments with the linewidth in excess of 50 percent of the bit rate [14], [16], [21]; this is not the limit, however, as the theory predicts that the linewidth can be even larger than the bit rate with only a small penalty [4]. The highest bit rate to date was 680 Mbit/s [19], and higher bit rates can be expected as the advantages of baseband operation are explored.

D. Impact of Imperfections on Phase-Diversity Receivers

Phase-diversity receivers can suffer from several imperfections [4], [14]–[17], [19], [21], [24]–[26]. In the rest of this subsection, a brief discussion of important imperfections is provided.

1. *Receiver Thermal Noise:* A phase-diversity receiver employs N detector/preamplifier modules ($N \geq 2$) and is affected by thermal noises of all of them. To suppress the impact of these noise, the LO power needs to be N times larger than that needed for a single detector receiver employing the same detector/preamplifier module. On the other hand, the detector/preamplifier modules needed for phase-diversity receivers have a smaller bandwidth than those needed for single detector heterodyne receivers, and smaller bandwidth modules tend to have less noise. In practice, the net result of the two foregoing factors have

been in favor of phase-diversity receivers: for a given bit rate, better sensitivities have been achieved with phase-diversity homodyne receivers than with single detector heterodyne receivers.

2. *Reflections:* [17], [23], [28], [30] Reflections at discontinuities in connectors, splices, and other components are known to cause impairments in fiber transmission systems. The following three mechanisms are important:

- Reflection of the incoming power backwards into the laser [30] can cause linewidth broadening, frequency hopping, reflection-induced noise, and other degradations; this effect can be minimized by incorporating an optical isolator in the laser package.
- Refraction index discontinuities in the signal transmission path can form Fabry-Perot cavities which convert the signal phase noise into high levels of intensity noise downstream of the isolator [23]; this effect leads to roughly the same penalty in both phase-diversity homodyne and in heterodyne systems; to ensure reliable system performance, it is desirable to suppress all reflections in the signal path to well below -20 dB for both types of systems [17], [28].
- Refraction index discontinuities in the LO path can also form Fabry-Perot cavities which convert the LO phase noise into high levels of intensity noise, similar to item b) above; however, the impact of this phenomenon on homodyne phase-diversity receivers is vastly different from its impact on heterodyne receivers. The reason is that the spectrum of the dominant component of the resulting intensity noise has a low-pass shape, i.e., is concentrated around zero frequency. Thus, it is substantially attenuated in heterodyne systems if the IF is sufficiently large. Homodyne systems, however, operate in the baseband, and have no mechanism for protection against this noise. As a result, a well-designed heterodyne system can tolerate reflections of up to -20 dB [28]

whereas a phase-diversity homodyne system requires suppression of reflections to well below -35 dB [17].

3. Polarization Fluctuations: Two-branch phase-diversity receivers are less tolerant to polarization fluctuations than conventional heterodyne receivers [15]: they incur a 1-dB penalty when the polarization deviates by $\sim 10^\circ$ from the optimum value, while conventional heterodyne receivers can tolerate much larger polarization fluctuations. The reason for this phenomenon is that practically used 90° hybrids rely on the correct SOP for proper operation, so that once the SOP deviates from the design value, the phase relationship between the branches is disturbed, too. Three-branch phase-diversity receivers use 3×3 couplers as 120° hybrids; they do not rely on polarization for proper phase relationship between the branches. As a result, three-branch receivers are less sensitive to polarization fluctuations than two-branch receivers if the 3×3 couplers used do not exhibit polarization dependence; their tolerance to polarization fluctuations is identical to that of heterodyne receivers.

4. Laser Phase Noise: With proper design [4], [5], [9], [10], [14], [25], both phase-diversity and heterodyne receivers can be essentially immune to phase noise.

5. Local Oscillator Intensity Noise: Both heterodyne and phase-diversity receivers [16] are sensitive to the LO intensity noise and require RIN of less than -150 dB/Hz for proper operation unless a balanced configuration is employed.

6. Gain Mismatch: Branch-to-branch gain and/or quantum efficiency mismatch as small as 1 dB leads to a noticeable sensitivity penalty in phase-diversity receivers [14].

7. Time Delay Mismatch: Branch-to-branch time delay mismatch as small as 1 ns leads to a noticeable sensitivity penalty in 150-Mbit/s phase-diversity receivers employing DFB lasers [14]: the wide linewidth of DFB lasers broadens the signal spectrum, leading to a higher penalty for a given time delay mismatch. Higher bit rate receivers are expected to be even more sensitive to the time delay mismatch.

8. Phase Mismatch: The phase response of all the branches must be matched over the entire signal bandwidth; however, no quantitative estimates of the tolerances needed are available at this time.

9. Nonideal Rectifiers: The rectifiers in Fig. 4 should have an ideal square-law characteristic $y = x^2$. If their characteristic is different from the ideal, the sensitivity of the receiver deteriorates. For example, the characteristic $y = |x|$ leads to an estimated 1.5-dB sensitivity penalty [14].

10. Loss in the Hybrid: Two-branch receivers require 90° hybrids. The intrinsic loss of symmetric 90° hybrids is at least 2.3 dB [14], and all known implementations have a 3-dB intrinsic loss (see Section II-B-1). Three-branch receivers require 120° hybrids. Such hybrids can be implemented with 3×3 couplers with no intrinsic loss.

11. Summary: All the foregoing imperfections have to and can be controlled; phase-diversity receivers have been reported to operate within a few decibels of their theoretical performance limits.

IV. POLARIZATION-DIVERSITY RECEIVERS: RECENT RESEARCH RESULTS

A. Theory

A polarization-independent diversity receiver with combining of demodulated signals is proposed in [29]. The author analyzes the DPSK format and shows that the sensitivity penalty is, in this case, only 0.4 dB as compared to the ideal case of a perfect polarization controller. Similar results are obtained for FSK polarization-diversity receivers in [31]; the authors propose a simple combining circuit that turns off the noisy branch and shows that the resulting performance is close to the ideal case. Enning *et al.* [32] deal with ASK polarization-diversity receivers. The theory of [32] shows that polarization diversity leads to a penalty of up to 3 dB if synchronous demodulation is used, and only 0.4 dB if square-law nonsynchronous demodulation is used. The authors also study envelope (linear) nonsynchronous demodulation and find that in this case the penalty is at most 0.6 dB if AGC is employed on both branches and at most 1.4 dB with no AGC.

B. Experimental Research

The first proposal for polarization diversity [33] suggested that the IF currents obtained from the two orthogonal polarizations should be combined directly, as shown in Fig. 11. One problem with this approach is that the relative phase of the two IF currents can vary with time, leading to signal fades or cancellation. To avoid this problem, the authors suggested and experimented with an automatic phase control loop whose purpose is to adjust the relative phase, keeping the output signal at maximum [33]. Due to the complexity of this approach, further experiments used the baseband combining of Fig. 3 rather than the IF combining of Fig. 11.

In a recent 100-kbit/s polarization-diversity experiment [34], the signal and the LO field were derived from the same HeNe laser leading to zero linewidth; a Bragg cell was used to impose FSK modulation, and a PZT modulator was used to introduce polarization fluctuations artificially. The IF currents were demodulated separately and then combined as per Fig. 3. No BER measurements were reported, but suppression of signal fluctuations due to polarization fluctuations was demonstrated.

The advantages of phase and polarization diversity can be combined in a single receiver shown in Fig. 12 (after [35]). In principle, such a receiver can tolerate both polarization fluctuations and large linewidth if ASK or large-deviation FSK are used. However, in this particular experiment the DPSK modulation format was used, limiting the linewidth to a small fraction of the bit rate. As a result, the authors had to use external cavity lasers as light sources. The receiver consists of two phase-diversity re-

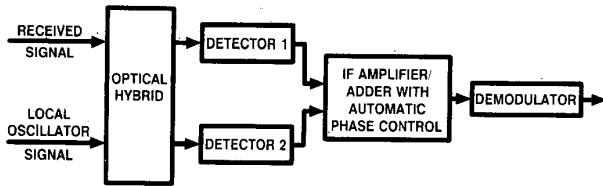


Fig. 11. Polarization-diversity receiver with IF combining, after [33].

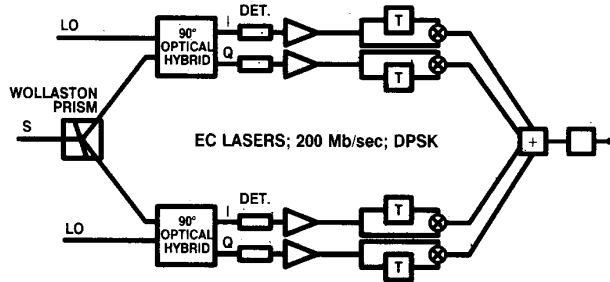


Fig. 12. A coherent receiver comprising phase and polarization diversity, after [35].

ceivers, each recovering an orthogonal polarization of the signal via post-detection combining. At 200 Mbit/s, the authors measured a sensitivity of -49 dBm at $\lambda = 1.3$ μm .

A block diagram of a higher bit-rate (560 Mbit/s) experiment is shown in Fig. 13 (after [32]). A unique feature of this experiment is the use of variable gain (AGC) blocks in each demodulation branch. Using the ASK modulation format, the authors measured the BER under a wide variety of conditions. They found a 2-dB power penalty due to dividing the limited LO power between two photodetectors. Receiver sensitivity varied by up to 1 dB as a function of the received SOP when the AGC blocks were switched off; when the AGC blocks were switched on, sensitivity variations were suppressed to a negligible level.

In all the foregoing experiments, only one output of the combining coupler is used. The field at the other output is discarded; the resulting loss of the signal and local oscillator power leads to an additional sensitivity penalty. An elegant way to avoid this penalty is shown in Fig. 14 (after [20]). The polarization controller PC1 ensures that the polarization of the signal reaching a given detector from fiber *B* is orthogonal with that reaching the same detector from fiber *A*. The LO power is divided equally between both detectors, and the currents generated by the two polarizations are in phase if

$$\Delta L = \pi V / (\omega_{LO} - \omega_S) \quad (15)$$

where ΔL is the length of the fiber coil in Fig. 14, V is the speed of light in the fiber, ω_{LO} is the LO frequency, and ω_S is the signal frequency. With this arrangement, all of the signal and LO power are used, but only one photodiode is required per branch. An experimental setup implementing the foregoing scheme [20] was demonstrated

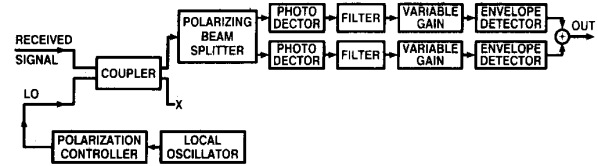


Fig. 13. A 560-Mbit/s polarization-diversity experiment, after [32].

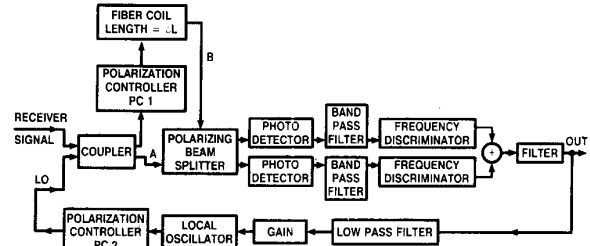


Fig. 14. An FSK polarization-diversity receiver employing both outputs of the combining coupler, after [20].

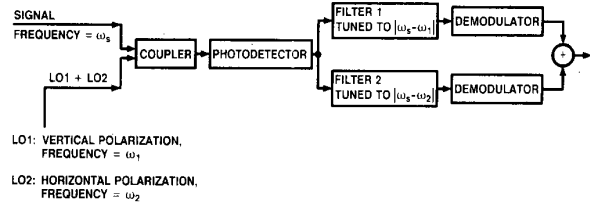


Fig. 15. A polarization-insensitive dual IF heterodyne receiver, after [36]. This receiver employs only one photodetector.

using external cavity lasers, with the signal laser employing a two-section chip that enabled FSK modulation at bit rates of up to 100 Mbit/s. The bit rate was 50 Mbit/s with a sensitivity of -55 dBm and a polarization dependence of ± 1 dB.

Polarization diversity can be achieved with one detector if the two orthogonal polarizations of the LO have different frequencies as shown in Fig. 15 (after [36]). With this arrangement, the mixing between the signal and LO light produces two IF components in the detected output. These two components are centered around different IF frequencies and fade in antiphase with SOP variations in the signal. Consequently, information encoded on the signal light is always recoverable by combining the output from the two IF channels. In the experimental setup [36], the two orthogonally polarized LO fields were derived from the same laser, and a 750-MHz frequency offset was created with a Bragg cell modulator. The authors FSK modulated the transmitter at 100 Mbit/s; no BER measurements were reported.

Dynamic operation of a polarization-diversity receiver was demonstrated in [37]. The authors introduced a periodic 30-Hz polarization fluctuation by applying a time-varying mechanical pressure to the signal fiber. Stable receiver operation was demonstrated with no additional performance deterioration as compared with the static case.

TABLE III
POLARIZATION-DIVERSITY STUDIES

Ref.	[33]	[39]	[34]	[29]	[35]	[32]	[20]	[36]	[40]	[37]
Journal or conference	IOOC'83	ECOC'86	Electronics Letters	J. L. T.	Electronics Letters	OFC-88	Electronics Letters	Electronics Letters	Electronics Letters	Electronics Letters
Authors	Okoshi, Ryu, Kikuchi	Kawahara et al.	Kreit and Youngquist	Glance	Okoshi Cheng	Enning et al.	Darcie et al.	Kersey et al.	Tzeng et al.	Ryu Yamamoto Mochizuki
Firm	Tokyo University	Fujitsu	University College, London	AT&T Bell Labs	Tokyo University	Heinrich Hertz Institute	AT&T Bell Labs	Naval Research Laboratory	AT&T Bell Labs	KDD
Wavelength	Unknown		1.5 μm		1.3 μm	1.3 μm	1.5 μm	0.83 μm	1.5 μm	1.5 μm
Bit rate	-		100 kbit/sec		200 Mbit/sec	560 Mbit/sec	50 Mbit/sec	100 Mbit/sec	40 Mbit/sec	560 Mbit/sec
IF	40 MHz	Proposal	40 MHz	Theoretical analysis	30 MHz	1.5 GHz	800 MHz	Two IF's 750 MHz apart	80 MHz	980 MHz
Modulation	-		FSK		DPSK	ASK	FSK	FSK	DPSK	FSK
Sensitivity	-		-		-49 dBm	-35 dBm	-55.5 dBm	Has not been measured	Unknown	-42 dBm
Linewidth/Bit Rate	-		~0		0.007	0.009	<<1	0.3	~0	0.0009
Lasers	Unknown		One HeNe		Two EC DFB	Three cavity + EC	Two external cavity	Two semiconductor	HeNe	Two DFB EC
Technical significance	First proposal of polarization diversity	First proposal of dual balanced polarization diversity	First FSK experiment	First proposal of polarization-independent receiver with combining of demodulated signals.	First receiver comprising phase and polarization diversity	Bit rate of 560 Mb/s achieved for the first time with polarization diversity	Signal and LO powers are fully used	Polarization diversity with one photodetector	First demonstration of balanced polarization diversity receiver	First demonstration of dynamic operation

Receiver sensitivity in both static and dynamic case was 2 dB worse than that of a conventional single-detector receiver. The authors estimate that this penalty is due to mismatch of the two branches, excess noise stemming from the use of two optical receivers, and post-detection combining. Other details of this experiment are provided in Table III.

C. *Balanced Diversity Receivers*

Balanced receivers use the signal and the local oscillator powers efficiently and substantially reduce the impact of the LO intensity noise [38]; in a multichannel environment, they also reduce the impact of channel-cross-channel interference [6]. These advantages can be combined with that of the polarization-diversity approach if balanced polarization-diversity receivers are constructed [39]. A prototype of such a receiver [40] is shown in Fig. 16. The authors used a 40-Mbit/s DPSK self-heterodyne setup to study receiver properties. The maximum variation in the amplitude of the recovered data was only about 0.8 dB under random variation of the signal polarization.

D. *Summary: Polarization-Diversity Studies*

Table III summarizes the polarization-diversity studies. Inspection of Table III reveals that a series of recent experimental demonstrations has brought the bit rates to 560 Mbit/s [32]; demonstrated efficient use of the signal and local oscillator powers with just two photodetectors [20]; proposed polarization diversity with a single photodetector [36]; and showed that the advantages of polarization-diversity receivers can be combined with those of balanced receivers [39], [40].

E. *Impact of Imperfections on Polarization-Diversity Receivers*

Polarization-diversity receivers can suffer from several imperfections [20], [29], [32]-[36], [39], [40]. In the rest of this subsection, a brief discussion of important imperfections is provided.

1. *Receiver Thermal Noise:* A polarization-diversity

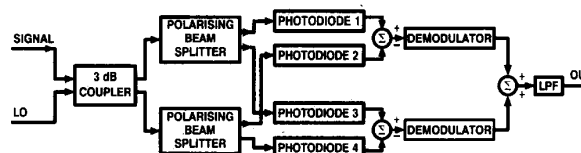


Fig. 16. A balanced polarization-diversity receiver, after [40].

receiver employs N detector/preamplifier modules ($N \geq 2$), and is affected by thermal noises of all of them. To suppress the impact of these noises, the LO power needs to be N times larger than that needed for a single-detector receiver employing the same detector/preamplifier module. If the LO power is limited, the receiver can incur a sensitivity penalty. The magnitude of the penalty in one reported experiment [32] was 2 dB.

2. *Reflections:* [17], [23], [28], [30] Reflections have the same impact on polarization-diversity receivers as on other heterodyne receivers; for comparison, we recall (Section III-D) that phase-diversity homodyne receivers are more susceptible to reflections than heterodyne receivers.

3. *Polarization Fluctuations:* Polarization diversity receivers have been theoretically predicted [29], [32] and experimentally confirmed [20], [32], [40] to be virtually independent from polarization fluctuations, with the performance variations suppressed to less than 1 dB.

4. *Laser Phase and Intensity Noise:* The impact of the laser phase and intensity noise on polarization-diversity receivers is the same as on single-detector heterodyne receivers.

5. *Gain, Time Delay, and Phase Mismatch Between the Branches:* The impact of these imperfections on polarization-diversity receivers is the same as on phase-diversity receivers (see Section III-D).

6. *Nonideal Demodulators:* The knowledge in this area is sparse. Some insight into this problem is provided in [32] where ASK polarization diversity receivers are analyzed. This analysis shows that ideal square-law nonsynchronous demodulators lead to the 0.4-dB penalty only

while envelope nonsynchronous demodulators lead to a penalty of up to 1.4 dB.

7. *Loss in the Hybrid:* A hybrid for polarization-diversity receivers does not have to have any intrinsic loss whatsoever. Indeed, some implementations [20] have a 0-dB intrinsic loss (their actual implementation loss is, of course, larger than 0 dB) while others have an intrinsic 3-dB loss on both the LO and signal part; still others have different losses on the signal and on the LO parts (see Section II-B-2).

8. *Summary:* All the foregoing imperfections have to and can be controlled; polarization-diversity receivers have been reported to operate within a few decibels of their theoretical performance limits.

V. CONCLUSIONS

Phase- and polarization-diversity receivers are more complex than conventional coherent receivers, and need larger LO powers and more precise signal processing. However, diversity receivers have important advantages:

- they are able to tolerate large polarization fluctuations if polarization diversity is employed, large LO intensity noise if a balanced receiver is employed, and wide laser linewidth;
- they require a smaller receiver bandwidth and slower photodetectors than heterodyne receivers if phase diversity is employed.

In the long run, the advantages of diversity receivers may outweigh the disadvantages; so, intensive research in this area is likely to continue. Balanced polarization-diversity receivers seem to be particularly attractive, and are a good candidate for further efforts.

APPENDIX A

In this appendix, the principle of operation of the all-single-mode-fiber 90° hybrid [12] shown in Fig. 2(b) is explained. We assume, for simplicity, that the input fields at the "signal" and "LO" ports of the device are identically linearly polarized. Then, the complex vector electric fields at the two input ports can be written as

$$\mathbf{E}_1 = E_1 [1 \ 0]^T \quad (\text{A1})$$

$$\mathbf{E}_2 = E_2 \exp(j\Phi) [1 \ 0]^T \quad (\text{A2})$$

where E_1 and E_2 are the scalar amplitudes of the signals at the two input ports, Φ is the relative phase between the two inputs (input 1 is selected as the phase reference); and $[\]^T$ denotes transposition. Equations (A1) and (A2) employ the Jones polarization matrix calculus [41] describing the electric field with an arbitrary state of polarization (SOP) by a (generally complex) vector consisting of two spatially orthogonal components. Polarization controller 1 rotates the SOP by 45°, while polarization controller 2 converts the linear SOP into a circular SOP. Hence, the complex vector electric fields at the outputs of the two polarization controllers (i.e., at the inputs of the 3-dB

coupler) are

$$\mathbf{E}_3 = \frac{1}{\sqrt{2}} E_1 [1 \ 1]^T \quad (\text{A3})$$

$$\mathbf{E}_4 = \frac{1}{\sqrt{2}} E_2 \exp(j\phi) [1 \ \exp(j\theta)]^T \quad (\text{A4})$$

where $\theta = 90^\circ$. Assume the coupling coefficient of the 3-dB coupler to be polarization independent. Then the output signals of the 3-dB coupler are, respectively:

$$\begin{aligned} \mathbf{E}_5 &= \frac{1}{\sqrt{2}} \mathbf{E}_3 + \frac{1}{\sqrt{2}} \mathbf{E}_4 \exp(j90^\circ) \\ &= 0.5 \begin{bmatrix} E_1 + E_2 \exp[j(\Phi + 90^\circ)] \\ E_1 + E_2 \exp[j(\Phi + \theta + 90^\circ)] \end{bmatrix} \end{aligned} \quad (\text{A5})$$

$$\begin{aligned} \mathbf{E}_6 &= \frac{1}{\sqrt{2}} \mathbf{E}_3 \exp(j90^\circ) + \frac{1}{\sqrt{2}} \mathbf{E}_4 \\ &= 0.5 \begin{bmatrix} E_1 \exp(j90^\circ) + E_2 \exp(j\Phi) \\ E_1 \exp(j90^\circ) + E_2 \exp[j(\Phi + \theta)] \end{bmatrix}. \end{aligned} \quad (\text{A6})$$

The components of the hybrid are interconnected with a conventional (*not* polarization-preserving) single-mode fiber. The beat length for a conventional unstrained single-mode fiber is around 50 m, while the length of interconnecting pigtailed of the hybrid is several centimeters. Therefore, the polarization states at the inputs of the polarization analyzers are essentially the same as the polarization states at the outputs of the 3-dB coupler. Since the polarization analyzers are aligned orthogonally with respect to each other, their outputs are

$$\mathbf{E}_7 = 0.5 \begin{bmatrix} E_1 + E_2 \exp[j(\Phi + 90^\circ)] \\ 0 \end{bmatrix} \quad (\text{A7})$$

$$\begin{aligned} \mathbf{E}_8 &= 0.5 \begin{bmatrix} 0 \\ E_1 \exp(j90^\circ) + E_2 \exp[j(\Phi + \theta)] \end{bmatrix} \\ &= 0.5 \exp(j90^\circ) \begin{bmatrix} 0 \\ E_1 + E_2 \exp[j(\Phi + \theta - 90^\circ)] \end{bmatrix}. \end{aligned} \quad (\text{A8})$$

For the 90° hybrid, $\theta = 90^\circ$, and (A8) yields

$$\mathbf{E}_8 = 0.5 \exp(j90^\circ) \begin{bmatrix} 0 \\ E_1 + E_2 \exp(j\Phi) \end{bmatrix}. \quad (\text{A9})$$

Comparison of (A7) with (A9) reveals that the device of Fig. 2(b) should operate as a 90° hybrid with a 3-dB intrinsic loss. Experimental confirmation of the foregoing theory is contained in [12]. Note that if the pigtailed connecting the 3-dB coupler with the polarization analyzers (Fig. 2(b)) are too long or are under stress, they can

change the SOP's of the optical fields E_5 and/or E_6 . To ensure proper operation of the hybrid in this case, an additional polarization controller has to be inserted between one of the outputs of the 3-dB coupler and the input of the corresponding polarization analyzer. Then, this controller has to be adjusted to compensate for the change of the SOP in the fiber pigtail.

APPENDIX B

In this appendix, expression (12) of Section II-B is derived.

Squaring (11a) and (11b), we obtain

$$i_1^2 = L^2 R^2 d P_S P_{LO} \cos^2 \theta [1 + \cos(2\omega t + 2\phi)] \quad (\text{B1})$$

$$i_2^2 = L^2 R^2 d P_S P_{LO} \sin^2 \theta [1 + \cos(2\omega t + 2\phi + 2\Psi)] \quad (\text{B2})$$

where we have used the fact that $d^2 = d$ since d is either zero or one. Adding (B1) to (B2), we obtain

$$i_1^2 + i_2^2 = L^2 R^2 P_S P_{LO} (d + xd + yd) \quad (\text{B3})$$

where

$$x = \cos^2 \theta \cos(2\omega t + 2\phi) \quad (\text{B4})$$

and

$$y = \sin^2 \theta \cos(2\omega t + 2\phi + 2\Psi). \quad (\text{B5})$$

Both x and y are bandpass random processes. Their spectra are concentrated around twice the IF frequency 2ω . The angles θ and Ψ stem from the random polarization, while the angle ϕ stems from the random phase. If semiconductor lasers are employed, then ϕ varies much faster than θ and Ψ , and the bandwidth of both x and y is primarily determined by the factor 2ϕ . Hence, the FWHM of both x and y is approximately $4\Delta\nu$ where $\Delta\nu$ is the FWHM linewidth at the "IF" given by (4).

Now let us discuss the spectra of xd and yd . Both spectra are concentrated around 2ω since d is a low-pass process, and x and y are bandpass processes. The bandwidth of xd and yd is approximately [9], [10]

$$B_{x,y} = B_x = B_y \cong 2R_b + 4k\Delta\nu \quad (\text{B6})$$

where R_b is the system bit rate and k is between 5 and 10.

Assume now that the quantity $i_1^2 + i_2^2$ is processed by a low-pass filter with the bandwidth B , and the following condition is satisfied:

$$2f_{IF} - 0.5B_{x,y} \gg B \geq 0.7R_b \quad (\text{B7})$$

where $f_{IF} = \omega/2\pi$ is the IF frequency in hertz. Then, the component proportional to d (see (B3)) will pass through the filter while the components proportional to xd and yd will be rejected. Hence, the output signal of the low-pass filter is

$$i_{TOT} = \text{lowpass} \{i_1^2 + i_2^2\} = L^2 R^2 d P_S P_{LO} \quad (\text{B8})$$

where $\text{lowpass} \{ \cdot \}$ denotes the lowpass part of $\{ \cdot \}$. Note that the result is independent both of polarization (of θ and Ψ) and of the phase noise (of ϕ). It can be shown

that this result holds true even if ϕ does not vary much faster than θ and Ψ , as assumed in the foregoing discussion. Combining (B7) with (B6), we obtain

$$f_{IF} \gg 0.85R_b + k\Delta\nu. \quad (\text{B9})$$

Thus, the physical meaning of the condition (B7) for proper receiver operation (as indicated by (B8)) is fairly simple: the intermediate frequency must be much larger than the signal bandwidth stemming from the modulation and phase noise; in addition, the IF bandwidth must be wide enough to pass the signal bandwidth (B6) without noticeable distortion. If the rate of change of θ and Ψ is nonnegligible, then the foregoing conditions have to be modified to read: the intermediate frequency must be much larger than the signal bandwidth stemming from the modulation, phase noise, and polarization fluctuations; in addition, the IF bandwidth must be wide enough to pass the signal bandwidth without noticeable distortion.

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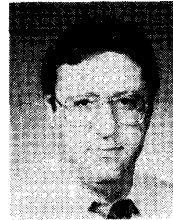
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