The Impact of Laser Phase Noise on the Coherent Subcarrier Multiplexing System

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***Abstract*—In coherent optical subcarrier multiplexing systems, the laser phase noise may cause signal spectrum broadening, and hence, deteriorates the system performance seriously. In this paper, we analyze the influence of phase noise in terms of carrier to noise ratio, intermodulation distortion, and adjacent channel crosstalk. The optimal modulation index and carrier to noise ratio are also presented.**

I. Introduction

RECENTLY, multichannel transmission makes use of subcarrier multiplexing (SCM) techniques allowing multichannels with one optical carrier, utilizing the wide bandwidth of single-mode fiber and electrooptic components, and taking advantage of commercially mature microwave electronics. The coherent subcarrier multiplexing (CSCM) system combines the multichannel aspects of the SCM and high receiver sensitivity of coherent detection [1]. For the coherent optical fiber communication, a major problem is the phase noise of laser sources. The phase-noise penalty has been analyzed in [2] for a multichannel CPFSK coherent optical communication system in terms of sensitivity penalty by calculating the bit error rate. But such a system does not suffer from the intermodulation distortion (IMD) and the signal terms are not dependent on the phase modulation index β. In a CSCM system, on the other hand, the influence of phase noise is more complicated and serious. However, extreme narrow linewidth lasers and various phase-noise-cancellation schemes may alleviate the difficulties. For example, CSCM experiments [3]–[5] have been performed using narrow linewidth Nd YAG lasers ($\Delta v = 10$ KHz). With a proper phase-noise-cancellation circuit, a CSCM system has been demonstrated using commercially available DFB laser with large linewidth ($\Delta v = 50$ MHz) [6]. Nevertheless, it is important to evaluate the impact of the phase noise on a CSCM system. In this paper, we analyze the performance degradation of the CSCM system caused by phase noise in terms of carrier-to-noise ratio (CNR) penalty and present some numerical results.

II. Analysis

A. System Description

Consider a coherent optical SCM system consisting of N equispaced channels each with bandwidth B and spanning a total bandwidth of W(Hz) as shown in Fig. 1(a) and (b). Thus, the channel separation is $\Delta f = W/N$. In order to reduce the effects of the second-order intermodulation (IMD_2), we locate the frequency of the ith channel ($i = 1, \cdots, N$) at $f_i = (i - 1)\Delta f + F_{\min}\Delta f + \Delta f/2$, where $F_{\min}$ is an integer and $F_{\min}\Delta f + \Delta f/2$ denotes the center frequency of Channel 1. The offset frequency $\Delta f/2$ is employed to cause the spectrum of IMD_2 to locate at the center of two adjacent channels. In such an arrangement, the IMD_2 degrades channel signal least. As for the third-order intermodulation (IMD_3), its spectrum locates at the center of each channel as shown in Fig. 1(b). We can simplify the power spectra of the IMD_2 and IMD_3 as the convolution of the power spectrum of each channel and their magnitudes are determined by the signal level and the phase modulation index. The IMD_2's, which resulted from the convolution of Channels i and j, center at $(j - i)\Delta f$ and $(j + i)\Delta f (i, j = 1, \cdots, N$, and $i \neq j)$ with their bandwidths expanding to $2B$ as shown in Fig. 2(a). The IMD_3's with their bandwidth expanding to $3B$, as shown in Fig. 2(b), are obtained from the spectra of Channels i, j, and k, and center at $f_i \pm (k - j)\Delta f$, $f_j \pm (k - i)\Delta f$, and $f_k \pm (j - i)\Delta f (i, j, k = 1, \cdots, N$ and $i \neq j \neq k)$, respectively. Note that other IMD_3's, such as $i = j \neq k$, are relatively small and not considered here [7].

B. Number of Intermodulation Distortions

Considering the case $F_{\min} + 1 \leq N \leq 2F_{\min}$, the number of IMD_2 falling between channel $k - 1$ and k, $P_2(k)$ (shown in Fig. 3(a) and (b)) is given by [8]

$$P_2(k) = \begin{cases} N + 1 - F_{\min} - k, & \text{for } 1 \leq k \leq N - F_{\min} \\ 0 & \text{for } N - F_{\min} + 1 \leq k \leq F_{\min} + 1 \\ [(k - F_{\min} - 1)/2], & \text{for } F_{\min} + 2 \leq k \leq N \end{cases} \quad (1)$$

where $[x]$ in (1) denotes the largest integer not greater than x. From (1), we find that there is a central region in which IMD_2 does not exist. Out of the central region, each channel is contaminated by two groups of IMD_2 (indicated in Fig. 3(a)); but Channels $N - F_{\min}$ and $F_{\min} + 1$ are contaminated by only one group of IMD_2 (indicated in Fig. 3(b)). The number of IMD_3 falling in Channel k, $P_3(k)$, is given by [9] (shown in Fig. 3(c))

$$P_3(k) = k(N - k + 1)/2 + ((N - 3)^2 - 5)/4. \quad (2)$$

C. Carrier to Noise Ratio

The carrier to noise ratio (CNR) of a coherent optical SCM system with phase modulation (PM) is given by [3]–[5]

$$\mathrm{CNR} = \frac{2R^2 P_{LO} P_S J_1^2(\beta) J_0(\beta)^{2N-2}}{\sigma_{sh}^2 + \sigma_{th}^2 + \sigma_{2d}^2 + \sigma_{3d}^2 + k_{CT}\sigma_{CT}^2} \quad (3)$$

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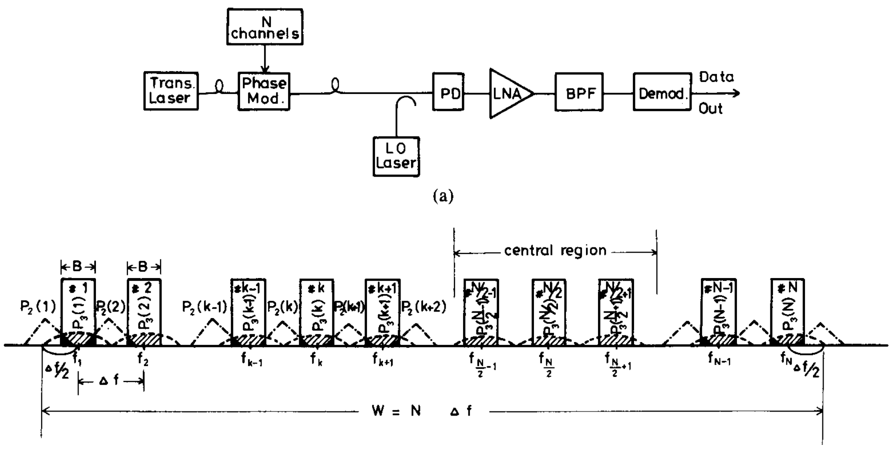

Fig. 1. (a) The system block diagram of CSCM. (b) The power spectrum at the output of low noise amplifier for the multichannel CSCM system.

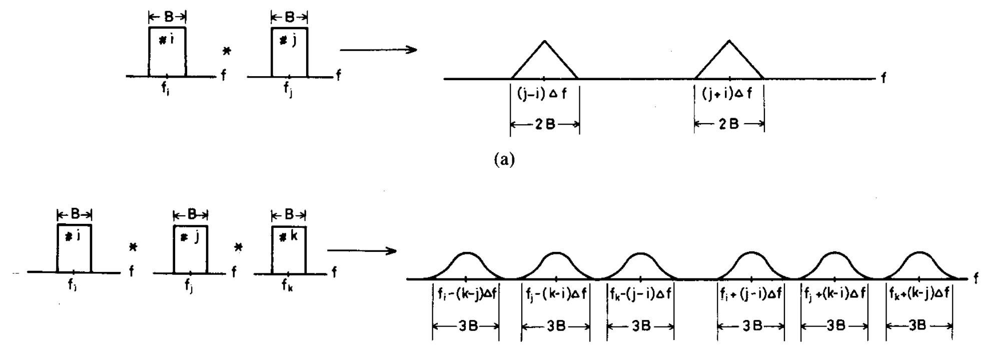

Fig. 2. The power spectrum of the second- and third-order IMD.

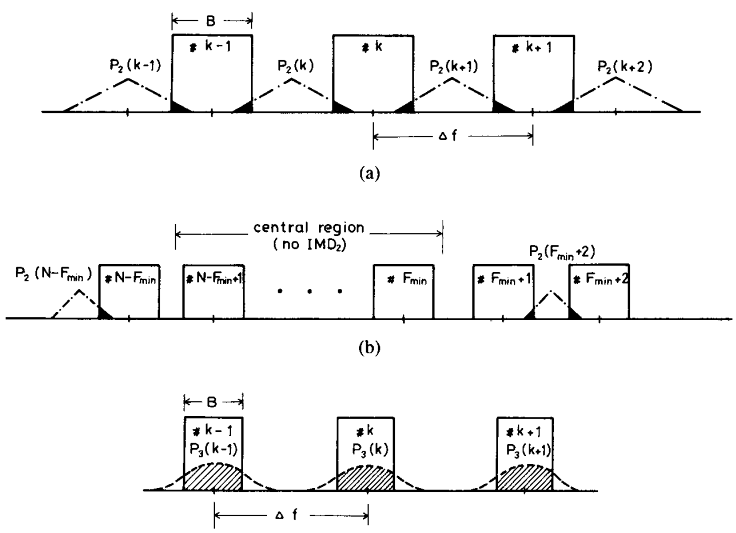

Fig. 3. (a) and (b) The channels contaminated by two and one groups of IMD_2, respectively. (c) The channel contaminated by one group of IMD_3.

where R is the photodiode responsivity, P_{LO} and P_S are the powers of the local oscillator and the received signal, and β is the phase modulation index. The terms $J_0(\beta)$ and $J_1(\beta)$ represent the zero- and first-order Bessel functions. σ_{sh}^2, σ_{th}^2, σ_{2d}^2, and σ_{3d}^2 stand for the variances of the shot noise, thermal noise, the second- and the third-order IMD. And we consider the crosstalk from adjacent channel as noise with variance $k_{CT}\sigma_{CT}^2$ and the crosstalk noise constant k_{CT} of Channels from 2 to $N-1$ is 2 (resulting from two neighborhood channels), but k_{CT} of Channels 1 and N is 1 (resulting from Channels 2 and $N-2$, respectively). It can be shown that the variances of IMD's (channel dependent) are given by (using approximation $J_1(\beta) \simeq \beta/2$ and $J_0(\beta) \simeq 1$ for small β)

$$\sigma_{2d}^2 = \begin{cases} A\Gamma_2\beta^4\{P_2(k)\}/4, & k = N - F_{\min} \text{ or } F_{\min} + 1 \\ 0, & k = N - F_{\min} + 1, \cdots, F_{\min} \\ A\Gamma_2\beta^4\{P_2(k) + P_2(k+1)\}/4, & \text{otherwise} \end{cases} \quad (4)$$

$$\sigma_{3d}^2 = A\Gamma_3 P_3(k)\beta^6/16 \quad (5)$$

where $A \equiv 0.5R^2P_{LO}P_S$.

And the sum of variances of the shot and thermal noises is

$$\sigma_n^2 = \sigma_{sh}^2 + \sigma_{th}^2 = (2qRP_{LO} + (\mathrm{NF})kT/r)B_{IF} \quad (6)$$

where q is the electronic charge, NF is the noise figure of the low noise amplifier (LNA), k is the Boltzmann's constant, T is the absolute temperature, and r is the input resistance of the amplifier and B_{IF} is the bandwidth of IF filter.

The crosstalk from an adjacent channel can be expressed as

$$\sigma_{CT}^2 = \Gamma_{CT}A\beta^2 \quad (7)$$

where the factor Γ_{CT} is the fraction of the signal power of a single adjacent channel within the passband of the BPF.

In (4) and (5), Γ_2 and Γ_3, related to the power spectra of the second- and third-order IMD in the neighborhood of the signal channel band, are the fractions of the powers within the passband of the bandpass filter [1], [2]. The values of Γ_2 and Γ_3 (see Appendix A) can be evaluated as

$$\Gamma_2 = \begin{cases} (3 - \Delta f/B)^2/8, & \text{for } \Delta f < 3B \\ 0 & \text{for } \Delta f \geq 3B \end{cases} \quad (8)$$

$$\Gamma_3 = 2/3. \quad (9)$$

In deriving (8) and (9), we have ignored the phase noise. For $\Delta f \geq 3B$, Γ_2 is equal to zero. It means that a CSCM system with the ratio of channel bandwidth over channel spacing less than 33.3% can render all the channels free from IMD_2 contamination. From (1), (2), and (4), we may obtain the number of the IMD for Channels 1 and $N/2$ as follows

CASE(A): $2F_{\min} > N > F_{\min}$

$$\Pi_2(1) = P_2(1) + P_2(2) = 2N - 2F_{\min} - 1 \quad (10)$$

$$\Pi_3(1) = P_3(1) = N/2 + ((N-3)^2 - 5)/4 \quad (11)$$

$$\Pi_2(N/2) = 0 \quad (12)$$

$$\Pi_3(N/2) = P_3(N/2) = N(N+2)/8 + ((N-3)^2 - 5)/4 \quad (13)$$

CASE(B): $N \leq F_{\min}$

$$\Pi_2(1) = \Pi_2(N/2) = 0 \quad (14)$$

$\Pi_3(1)$ and $\Pi_3(N/2)$ are the same as (11) and (13), respectively. $\Pi_2(1)$ and $\Pi_2(N/2)$ represent the number of IMD_2's contaminating Channels 1 and $N/2$, respectively; $\Pi_3(1)$ and $\Pi_3(N/2)$ denote the numbers of IMD_3's contaminating Channels 1 and $N/2$, respectively.

D. The Impact of Phase Noise

The power spectrum of signal contaminated by the phase noise can be expressed as convolution of the power spectra of signal and phase noises [2]. Due to the phase noise, the signal spectrum is broadened and can be expressed as,

$$\begin{aligned} S(f) &= S_{NPN}(f) * G_{PN}(f) \\ &= \int_{f_k - B/2}^{f_k + B/2} \frac{2\,dx}{\pi\Delta v(1 + (2x/\Delta v)^2)} \\ &= \{\tan^{-1}((2f_k + B)/\Delta v) - \tan^{-1} \\ &\quad \cdot ((2f_k - B)/\Delta v)\}/\pi \end{aligned} \quad (15)$$

where

$$S_{NPN}(f) = \begin{cases} 1, & \text{for } f_k - B/2 \leq f \leq f_k + B/2 \\ 0, & \text{otherwise} \end{cases} \quad (16)$$

and

$$G_{PN}(f) = \frac{2}{\pi\Delta v(1 + (2f/\Delta v)^2)}, \quad 0 < f < \infty. \quad (17)$$

In (15), "*" denotes convolution, and $S(f)$, $S_{NPN}(f)$, and $G_{PN}(f)$ represent the power spectra of the signal contaminated by phase noise, the kth Channel signal (center frequency $= f_k$) without phase noise, and the phase noise (Lorentzian lineshape), respectively. The power spectrum density of signal with phase noise is shown in Fig. 4 in terms of normalized laser linewidth. Because of spectrum broadening, the bandwidth of the bandpass filter B_{IF} must be increased to avoid signal distortion caused by the filter (see Fig. 5). We can find the IF bandwidth required to accomodate 95% of signal power in the presence phase noise as $B_{IF(95\%)}$. The value of $B_{IF(95\%)}$ depends on Δv as shown in Fig. 6.

Γ_{2PN} and Γ_{3PN} (see Appendix B), related to the power spectra of the second- and third-order IMD's in the neighborhood of the signal channel band, are the fractions of their powers within the passband of the IF filter taking into consideration the phase noise. They can be expressed as

$$\Gamma_{2PN} = \begin{cases} (3 - \Delta f/B_{IF(95\%)})^2/8, & \text{for } \Delta f < 3B_{IF(95\%)} \\ 0 & \text{for } \Delta f \geq 3B_{IF(95\%)} \end{cases}$$

$$\begin{aligned} \Gamma_{3PN} &= 2/3 + \Delta_3 \\ &= 2/3 + (z^3 - 1)/48 + 3(1 - z^2)/16 + 9 \\ &\quad \cdot (z - 1)/16 \end{aligned} \quad (18)$$

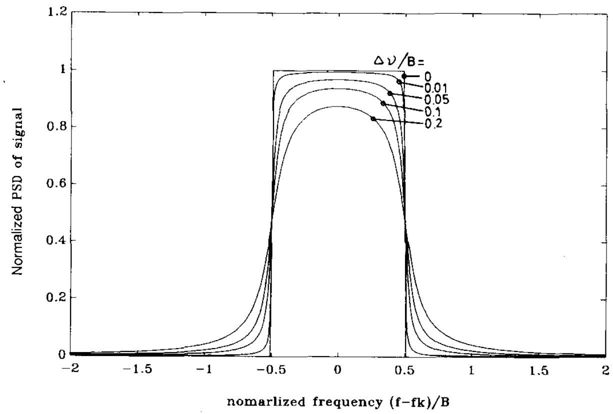

Fig. 4. Normalized power spectrum density (PSD) of signal contaminated by the phase noise for $\Delta v/B$ = 0, 0.01, 0.05, 0.1, and 0.2.

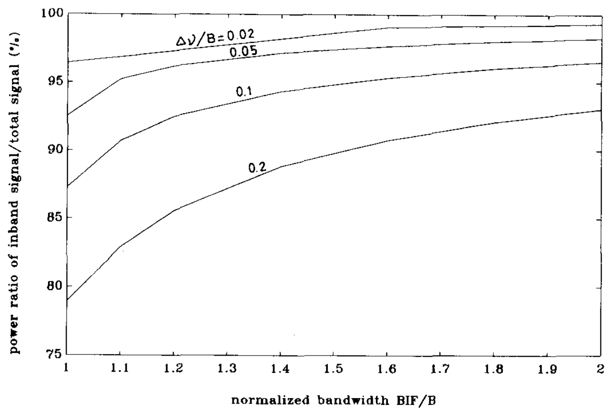

Fig. 5. Power ratio of signal within bandpass filter to total signal versus normalized bandwidth B_{IF}/B for $\Delta v/B$ = 0.02, 0.05, 0.1, and 0.2.

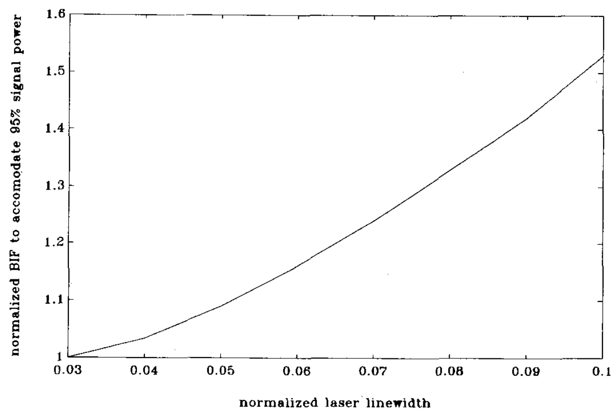

Fig. 6. The IF bandwidth accommodating 95% of signal power versus normalized laser linewidth ($\Delta v/B$) in the range of 0.03 to 0.1.

with

$$z = \frac{B_{IF(95\%)}}{B}. \tag{19}$$

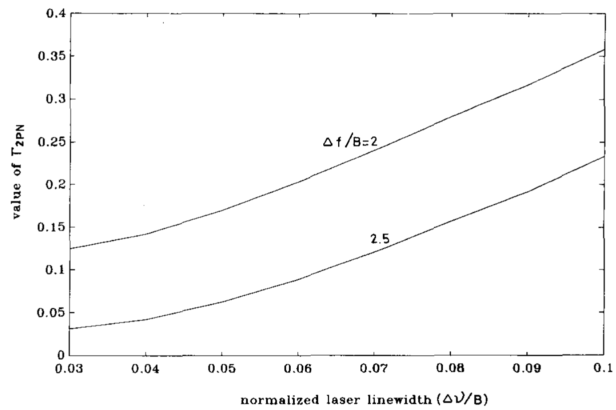

Fig. 7. The value of Γ_{2PN} versus normalized laser linewidth ($\Delta v/B$) in the range of 0.03 to 0.1 for $\Delta f/B$ = 2, 2.5.

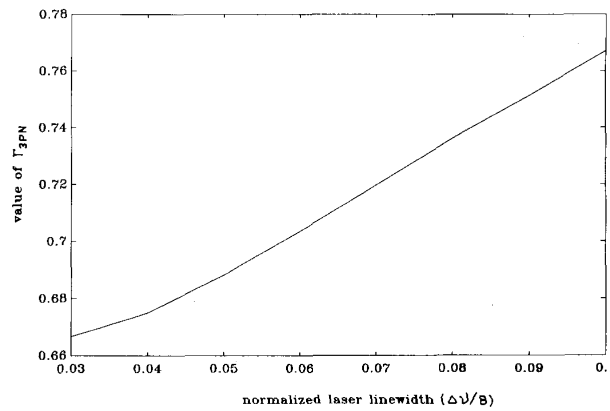

Fig. 8. The value of Γ_{3PN} versus normalized laser linewidth ($\Delta v/B$) in the range of 0.03 to 0.1.

In (18) and (19), Γ_{2PN} and Γ_{3PN} are seen to be functions of Δv as shown in Figs. 7 and 8, respectively.

The crosstalk from adjacent channels due to phase noise can be calculated as follows. First, we obtain the corresponding value of $B_{IF(95\%)}/B$ (for given $\Delta v/B$). Second, after choosing a $\Delta f/B$ to specify the location of adjacent channels, we can obtain the value of Γ_{CTPN} as a function of phase noise. Thereafter, the value of Γ_{CTPN}, related to the tail of the power spectrum of adjacent channel due to the phase noise, is the fraction of the power within the passband of the BPF which is shown in Fig. 9 as a function of Δv for a given $\Delta f/B$.

E. Optimum Phase Modulation Index

The CNR can be calculated from (3) by using the first-order approximations of $J_0(\beta)$ and $J_1(\beta)$. Assuming ail channels with the same modulation index, the CNR and the optimal modulation index that maximizes the CNR of a certain selected channel can be expressed as [1]

$$\text{CNR} = \frac{A\beta^2}{\sigma_n^2 + A\Pi_2\Gamma_{2PN}\beta^4/4 + A\Pi_3\Gamma_{3PN}\beta^6/16 + k_{CT}\Gamma_{CTPN}A\beta^2} \tag{20a}$$

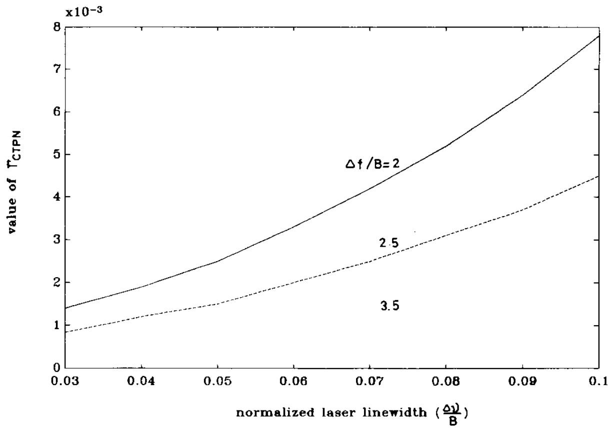

Fig. 9. The value of Γ_{CTPN} versus normalized laser linewidth ($\Delta v/B$) in the range of 0.03 to 0.1 for $\Delta f/B$ = 2, 2.5, and 3.5.

and

$$\Pi_3\Gamma_{3PN}\beta_{opt}^6 + 2\Pi_2\Gamma_{2PN}\beta_{opt}^4 - 16\sigma_n^2/(R^2P_{LO}P_S) = 0 \quad (20b)$$

where Π_2 and Π_3 are the number of IMD's contaminating the corresponding channel and we have used the definition of $A \equiv 0.5R^2P_{LO}P_S$ in (20b). From (20b), we may obtain the optimum phase modulation index as [10]

$$\beta_{opt} = (s1 + s2 - a2/3)^{1/2}$$
$$s1 = \left(r + (q^3 + r^2)^{1/2}\right)^{1/3}, \quad s2 = \left(r - (q^3 + r^2)^{1/2}\right)^{1/3},$$
$$a2 = 2\Pi_2\Gamma_{2PN}/(\Pi_3\Gamma_{3PN})$$
$$r = 8\sigma_n^2/(P_{LO}P_S\Pi_3\Gamma_{3PN}) - 8\left(\Pi_2\Gamma_{2PN}/(\Pi_3\Gamma_{3PN})\right)^3/27$$
$$q = -4\left(\Pi_2\Gamma_{2PN}/(\Pi_3\Gamma_{3PN})\right)^2/9. \quad (21a)$$

For case (B), there is no 2nd IMD, we can simplify (21a) as

$$\beta_{opt} = \left(16\sigma_n^2/(P_{LO}P_S\Pi_3\Gamma_{3PN})\right)^{1/6}. \quad (21b)$$

Using (21), we can calculate the optimal phase modulation index as a function of Δv for given σ_n^2, P_{LO}, and P_S. Eliminating σ_n^2/A in (20a) by using the relationship of (20b), we may obtain the optimal CNR in terms of β_{opt} as

$$(\text{CNR})_{opt} = \frac{1}{\Pi_2\Gamma_{2PN}\beta_{opt}^2/2 + 3\Pi_3\Gamma_{3PN}\beta_{opt}^4/16 + k_{CT}\Gamma_{CTPN}} \quad (22)$$

F. The CNR Penalty Due to Phase Noise

We define the penalty of CNR due to phase noise as

$$\textbf{Penalty} \equiv \frac{\text{CNR}_{(\text{without phase noise})}}{\text{CNR}_{(\text{with phase noise})}}. \quad (23)$$

From (23), we can obtain the penalty as

$$\textbf{Penalty} = \frac{\Pi_2\Gamma_{2PN}\beta_{opt}^2/2 + 3\Pi_3\Gamma_{3PN}\beta_{opt}^4/16 + k_{CT}\Gamma_{CTPN}}{\Pi_2\Gamma_2\beta_{opt}^2(\Delta v = 0)/2 + 3\Pi_3\Gamma_3\beta_{opt}^4(\Delta v = 0)/16} \quad (24)$$

where $\beta_{opt}(\Delta v = 0)$ represents for the optimal PM index without considering phase noise and it can be obtain from (21) by replacing Γ_{2PN} and Γ_{3PN} with Γ_2 and Γ_3. In deriving (24), we have assumed that the adjacent channel crosstalk is mainly due to phase noise; so Γ_{CTPN} is negligible for the system without phase noise [1].

III. Numerical Results and Discussion

A. System Parameters

Now we consider system with: 35-channel (N = 35), channel spacing 200 MHz (Δf = 200 MHz), total bandwidth 7 GHz (W = 7 GHz) and single-channel bandwidth 100 MHz (B = 100 MHz), hence the normalized channel spacing is 2 ($\Delta f/B$ = 2). In **CASE(A)** the frequency band is selected to be 4–11 GHz (F_{min} = 20, including C band and X [11]), which satisfies $2F_{min} > N > F_{min}$. In **CASE(B)** the frequency band is selected to be 7–14 GHz (F_{min} = 35) which satisfies $N \le F_{min}$. We now show the numerical results of these two systems.

B. CASE(A): $f_1 = 4.1, f_2 = 4.3, \cdots, f_N = 4.1 + 0.2(N - 1)$ GHz (N = 35 and F_{min} = 20)

From (1) and (2), we may obtain the number of IMD's for the kth channel as shown in Fig. 10(a) and (b). Channel 1 is contaminated by IMD_2 most and Channel $N/2$ suffers least from IMD_2 as shown in Fig. 10(a); but Channel $N/2$ is contaminated by IMD_3 most and Channels 1 and N suffer least from IMD_3 as shown in Fig. 10(b).

From (21), we can obtain β_{opt} for Channels 1 and $N/2$ versus laser linewidth with P_S = −35, −30, and −25 [dBm] as shown in Fig. 11(a). β_{opt} increases as the received signal power P_S decreases; β_{opt} of Channel 1, which increases as Δv decreases, is lower than that of Channel $N/2$, which increases as Δv increases. From Fig. 11(a) and (23), we can obtain the corresponding optimal CNR as shown in Fig. 11(b). $(\text{CNR})_{opt}$ increases as P_S increases and Δv decreases; the $(\text{CNR})_{opt}$ of Channel 1 is lower than that of Channel $N/2$. From (24), we can obtain the CNR penalties of Channels 1 and $N/2$ versus laser linewidth as shown in Fig. 11(c). The penalties decrease as P_S and Δv decrease. Channel $N/2$ has more penalty than Channel 1.

EXAMPLE—SYSTEM DESIGN FOR CASE(A): Assume that the laser linewidth is 5 MHz and the CNR of the worst channels (around Channel 1) must be larger than 17 dB. Now we may obtain the required power P_S such that the $(\text{CNR})_{opt}$ of Channel 1 is equal to 17 dB from Fig. 11(b) and substitute this power P_{SYSTEM} into (21) to get the β_{opt} of Channel 1, β_{SYSTEM}. Note that β_{SYSTEM} is only optimal for Channel 1 and is not the optimal PM index for other Channels. However, we guarantee that the CNR of each Channel is no less than 17 dB under this condition as shown in Fig. 12. The CNR of Channels 1 and N are better than their adjacent Channels 2 and $N - 1$ because k_{CT} of Channels 1 and N is 1 and k_{CT} of others is 2. The CNR of Channels at the central region (Ch. 16–20) are approximately equal since they have no IMD_2 and with roughly equal IMD_3's.

C. CASE(B): $f_1 = 7.1, f_2 = 7.3, \cdots, f_N = 7.1 + 0.2(N - 1)$ GHz (N = 35 and F_{min} = 35)

As discussed in Section II-C, there are no IMD_2's in **CASE(B)** and the number of IMD_3 is the same in Fig. 10(b).

From (21b), we can obtain the corresponding β_{opt} versus laser linewidth for P_S = −35, −30, and −25 [dBm] as shown in

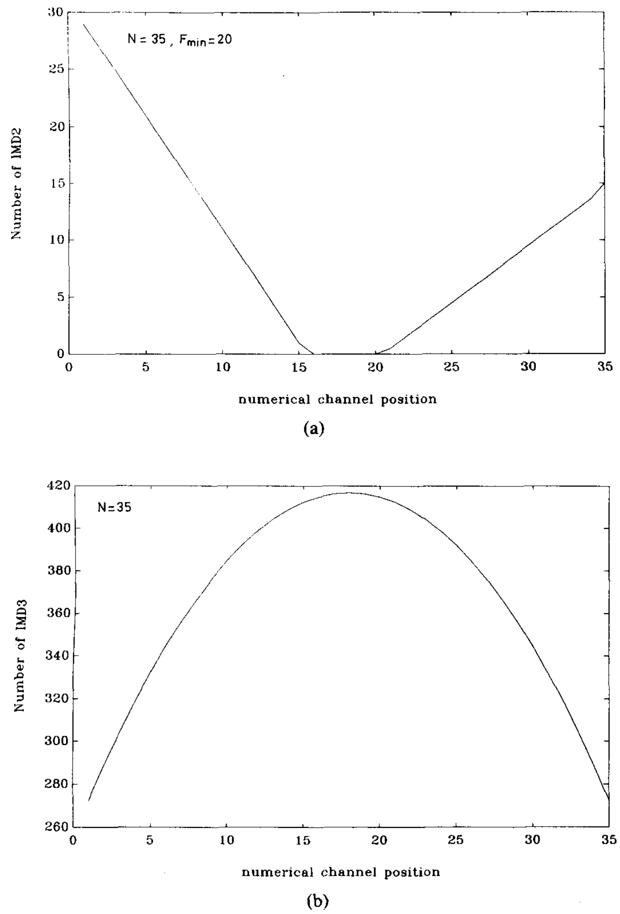

Fig. 10. (a) Number of interfering two-frequency second-order intermodulation (IMD_2) terms within a given channel, plotted against channel position for $N = 35$ with $F_{\min} = 20$. (b) Number of interfering three-frequency third-order intermodulation (IMD_3) terms within a given channel, plotted against channel position for $N = 35$.

Fig. 13(a). β_{opt} increases as Δv increases and P_S decreases; β_{opt} of Channel 1 is always larger than that of Channel $N/2$. From Fig. 13(a) and (22), we can obtain the corresponding optimal CNR as shown in Fig. 13(b). $(CNR)_{opt}$ increases as P_S increases and Δv decreases; the $(CNR)_{opt}$ of Channel 1 is larger than that of Channel $N/2$. From (24), we can obtain the CNR penalties of Channels 1 and $N/2$ versus laser linewidth as shown in Fig. 13(c). The penalties decrease as P_S and Δv decrease; Channel $N/2$ always has more penalty than Channel 1.

EXAMPLE—SYSTEM DESIGN FOR CASE(B): Assume the given laser linewidth and CNR of the worst channel are the same as specified in the previous example ($\Delta v = 5$ MHz and CNR = 17 dB). Now we may obtain the required power P_S by setting $(CNR)_{opt}$ of Channel $N/2$ to be 17 dB from Fig. 13(b) and substitute this power P_{SYSTEM} into (21) to get the β_{opt} of Channel $N/2$. This optimal PM index, β_{SYSTEM}, is only optimal for Channel $N/2$ and is not for others. However, the CNR of all channels are larger than 17 dB as shown in Fig. 14. The CNR of Channels 1 and N are better than their adjacent Channels 2 and $N - 1$, because k_{CT} of Channels 1 and N is 1 and k_{CT} of other Channels is 2.

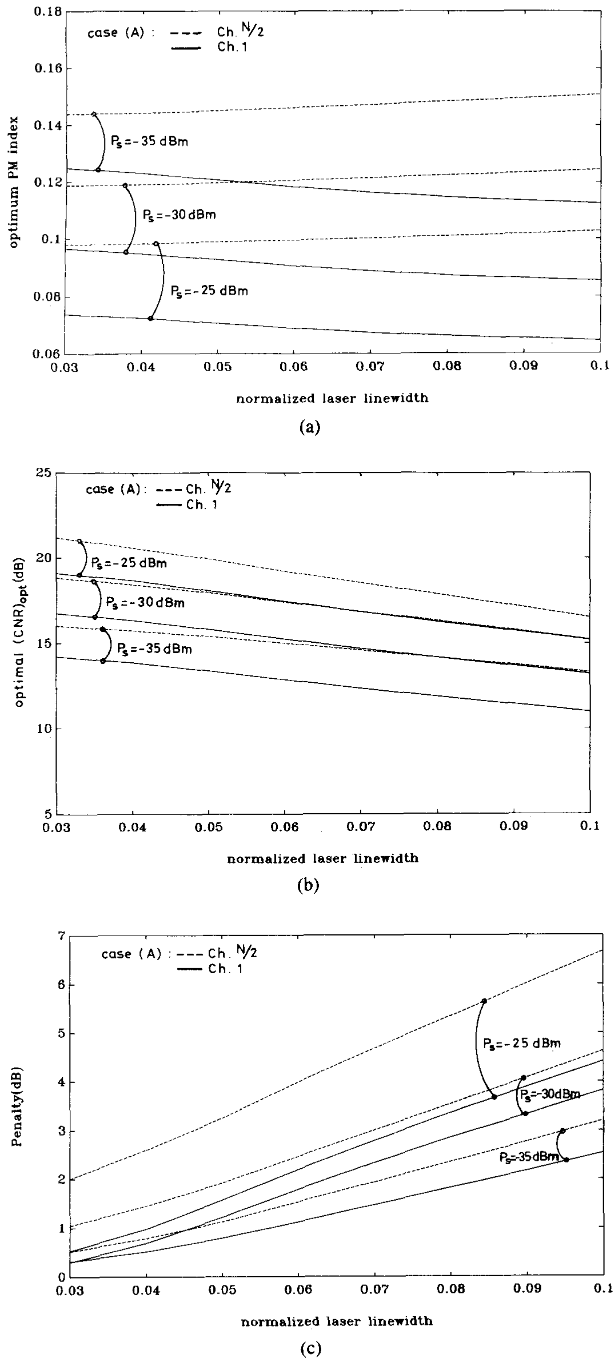

Fig. 11. (a)–(c) Optimal PM index, received power and CNR penalty, respectively, versus laser linewidth. The values of β_{opt}, P_S, and **Penalty** versus $\Delta v/B$ for the first channel ($k = 1$; solid line) and the center channel ($k = N/2$; dotted line). System parameters are given as in **CASE(A)**.

IV. Conclusion

We have analyzed the impact of phase noise on the CSCM system. Due to the phase noise, the signal spectrum is broadened so that we need to widen the bandwidth of the IF bandpass filter

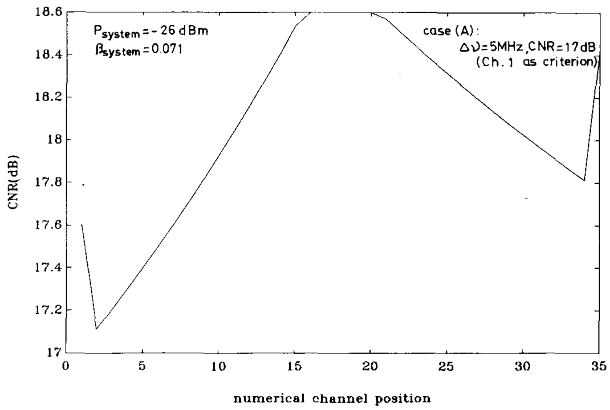

Fig. 12. The CNR versus numerical channel position for design example: $\Delta\upsilon$ = 5 MHz and CNR = 17 dB, other parameters are given as in **CASE(A)**. The P_{SYSTEM} and β_{SYSTEM} are -26 dBm and 0.071, respectively.

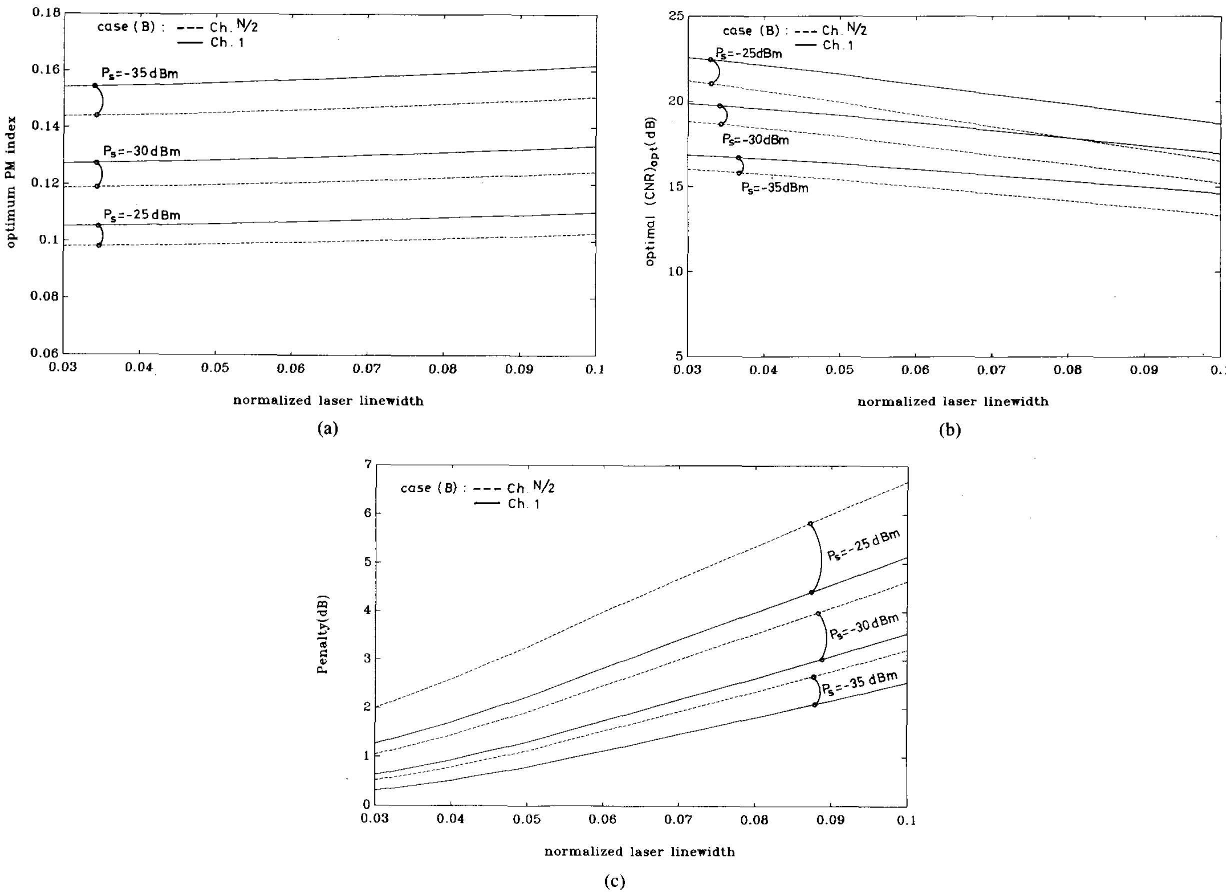

Fig. 13. (a)–(c) Optimal PM index, received power, and CNR penalty, respectively, versus laser linewidth. The values of β_{opt}, P_S, and Penalty versus $\Delta\upsilon/B$ for the first channel (k = 1; solid line) and the center channel ($k = N/2$; dotted line). System parameters are given as in **CASE(B)**.

to avoid signal distortion caused by the filter. In this paper, we take the bandwidth for 95% of signal power to be passed through the filter as a criterion for B_{IF}. Because of signal spectrum broadening and filter bandwidth widening, the system suffers from more shot noise, thermal noise, IMD, and adjacent crosstalk. In contrast, for the case of single subcarrier channel sys-

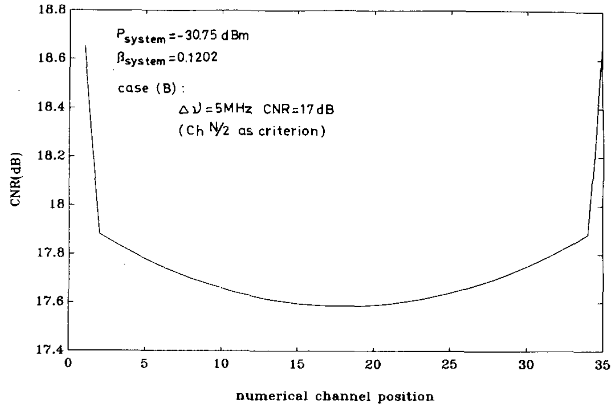

Fig. 14. CNR versus numerical channel position for design example: $\Delta v = 5$ MHz and CNR = 17 dB, other parameters are given as in **CASE(B)**. The P_{SYSTEM} and β_{SYSTEM} are -30.75 dBm and 0.1202, respectively.

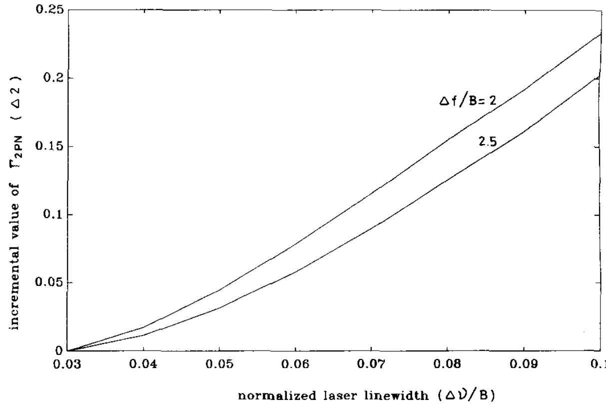

Fig. 15. The incremental value of $\Gamma_{2\text{PN}}(\Delta_2)$ due to phase noise.

tem (no crosstalk and IMD), the CNR penalty is mainly due to shot and thermal noises (the noise term in (20a) is only σ_n^2 and other noise terms are zero in such system) and the worst CNR penalty at $\Delta v/B = 0.1$ is only 1.85 dB for a normalized $B_{\text{IF}(95\%)}$ of 1.53 as shown in Fig. 6. Finally, the degradation of optimal CNR presented in this paper may be a useful guideline for CSCM system design.

Appendix A

The value of Γ_2 which is defined as the ratio of IMD_2 power located within the signal band to its total power can be written as [1]

$$\Gamma_2 = \frac{\int_{-\infty}^{+\infty} S_i(f) * S_j(j)\,|H_{\text{BP}}(f)|^2\,df}{\int_{-\infty}^{+\infty} S_i(f) * S_j(f)\,df}. \quad \text{(A1)}$$

For the case of an ideal rectangular signal spectrum, we can obtain the Γ_2 as

$$\Gamma_2 = \begin{bmatrix} (3 - \Delta f/B)^2/8, & \text{for } \Delta f < 3B \\ 0, & \text{for } \Delta f \geq 3B \end{bmatrix}. \quad \text{(A2)}$$

The value of Γ_3 which is defined as the ratio of IMD_3 power within the signal band to its total power can be expressed as [1]

$$\Gamma_3 = \frac{\int_{-\infty}^{+\infty} S_i(f) * S_j(f) * S_k(f)\,|H_{\text{BP}}(f)|^2\,df}{\int_{-\infty}^{+\infty} S_i(f) * S_j(f) * S_k(f)\,df}. \quad \text{(A3)}$$

Similarly for the case of an ideal rectangular signal spectrum, we can obtain the value of Γ_3 as

$$\Gamma_3 = 2/3. \quad \text{(A4)}$$

In (A1) and (A3), $S_i(f)$, $S_j(f)$, and $S_k(f)$ represent the signal spectra of the ith, jth, and kth Channels, respectively; $H_{\text{BP}}(f)$ is the transfer function of the bandpass filter and "*" denotes convolution.

Appendix B

We assume that the power spectra of IMD_2, and IMD_3 are not influenced by the phase noise significantly. The system performance degradation due to the IMD_2 and IMD_3 caused by the phase noise comes mainly from the enlarging bandpass filter bandwidth so that more IMD_2 and IMD_3, shot and thermal noises will pass the IF filter. In such a case, the values of the power ratio Γ_2 and Γ_3 are determined by $B_{\text{IF}(95\%)}$. As shown in Fig. 6, the value of $B_{\text{IF}(95\%)}$ increases as the phase noise increases; hence Γ_2 and Γ_3 increase as the phase noise (or Δv) increases.

The value of $\Gamma_{2\text{PN}}$ (the ratio of IMD_2 power located within the signal band to its total power in the presence of the phase noise) can be expressed as

$$\Gamma_{2\text{PN}} = (3 - \Delta f/B_{\text{IF}(95\%)})^2/8 \quad \text{(B1)}$$

where the derivation of (B1) is the same as that of (A1) in Appendix A, but with B replaced by $B_{\text{IF}(95\%)}$. We express the incremental value of $\Gamma_{2\text{PN}}$, Δ_2, as

$$\Delta_2 = (3 - \Delta f/B_{\text{IF}(95\%)})^2/8 - (3 - \Delta f/B)^2/8. \quad \text{(B2)}$$

The value of $\Gamma_{3\text{PN}}$ (the ratio of IMD_3 power within the signal band to its total power taking into consideration the phase noise) can be derived as

$$\Gamma_{3\text{PN}} = 2/3 + \Delta_3 \quad \text{(B3)}$$

where Δ_3 is the incremental value of $\Gamma_{3\text{PN}}$ due to the broadening of B_{IF} and can be expressed as

$$\Delta_3 = \frac{\int_{-B_{\text{IF}(95\%)}/2}^{-B/2} 1/2(x + 3B/2)^2\,dx}{B^3}$$
$$= (z^3 - 1)/48 + 3(1 - z^2)/16 + 9(z - 1)/16$$

with

$$z = \frac{B_{\text{IF}(95\%)}}{B}. \quad \text{(B4)}$$

The values of Δ_2 and Δ_3 are shown in Figs. 15 and 16 and they increase as the phase noise increases.

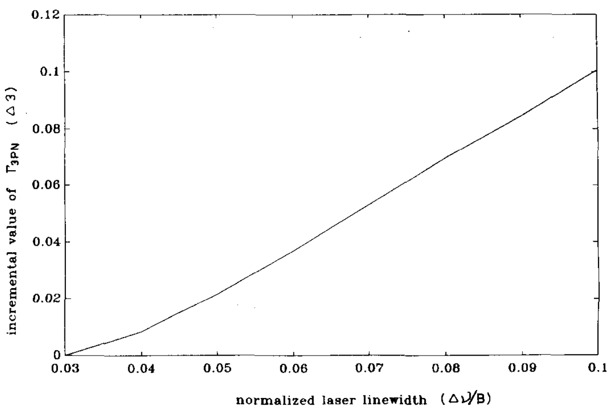

Fig. 16. The incremental value of $\Gamma_{3PN}(\Delta_3)$ due to phase noise.

REFERENCES

[1] R. Gross and R. Olshansky, "Multichannel coherent FSK experiments using subcarrier multiplexing techniques," *J. Lightwave Technol.*, vol. 8, no. 3, pp. 406–415, 1990.

[2] L. G. Kazovsky and G. Jocobsen, "Multichannel CPFSK coherent optical communication systems," *J. Lightwave Technol.*, vol. 7, no. 6, pp. 972–982, 1989.

[3] R. Gross, R. Olshansky, and P. Hill, "20 channel coherent FSK system using subcarrier multiplexing," *IEEE Photon. Technol. Lett.*, vol. 1, no. 8, pp. 224–226, 1989.

[4] R. Gross and R. Olshansky, "Third-order intermodulation distortion in coherent subcarrier-multiplexed system," *IEEE Photon. Technol. Lett.*, vol. 1, no. 4, pp. 91–93, 1989.

[5] R. Gross, R. Olshansky, and P. Hill, "Optical tunable 5 channel coherent subcarrier multiplex system," in *Tech. Dig. OFC'89* (Houston, TX), p. 43.

[6] R. Gross, R. Olshansky, and M. Schmidt, "Coherent FM-SCM system using DFB lasers and a phase noise cancellation circuit," *IEEE Photon. Technol. Lett.*, vol. 2, no. 1, pp. 66–68, 1990.

[7] B. Arnold, "Third order intermodulation products in a CATV systems," *IEEE Trans. Cable Television*, vol. CATV-2, no. 2, pp. 67–80, 1977.

[8] R. Olshansky, V. A. Lanzisera, and P. Hill, "Subcarrier multiplexed lightwave systems for broadband distribution," *J. Lightwave Technol.*, vol. 7, no. 9, pp. 1329–1341, 1989.

[9] M. T. Abuelma'atti, "Carrier-to-intermodulation performance of multiple FM/FDM carriers through a GaAlAs heterojunction laser diode," *IEEE Trans. Commun.*, vol. COM-33, no. 3, pp. 246–248, 1985.

[10] M. Abramowitz *et al.*, *Handbook of Mathematical Function.* New York: Dover, 1970, p.17.

[11] Samuel Y. Liao, *Microwave Devices and Circuits.* Englewood Cliffs, NJ: Prentice-Hall, 1985, p. 2.

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