Average Power Reduction Using Time-Varying Bias with Bandlimited Pulses in Multiple-Subcarrier Intensity-Modulated Optical Systems

Byron Yu, Joseph M. Kahn, and Roy You

Department of Electrical Engineering and Computer Sciences
514 Cory Hall
University of California
Berkeley, CA 94720

Telephone: (510) 643-8848
Facsimile: (510) 642-2739
Internet: jmk@eecs.berkeley.edu
URL: http://www.eecs.berkeley.edu/~jmk

July 7, 2001

Abstract:

We present a technique for average power reduction in intensity-modulated optical systems using multiple BPSK or QPSK subcarriers with bandlimited pulses, which are, necessarily, not time-limited. In order to insure the non-negativity of the transmitted intensity signal, we use a time-varying bias signal, which is a baseband pulse-amplitude modulation signal also using these pulses. We focus on the problem of designing this time-varying bias signal, which is complicated by the fact that the pulses are not time-limited. We consider pulses whose Fourier transform is a root-raised cosine with rolloff factor α, where $0 \leq \alpha \leq 1$. We use a heuristic approach in which, during a given symbol interval, the bias amplitude is computed considering the multiple-subcarrier amplitudes during that symbol interval, as well as several neighboring symbol intervals. We consider three cases in detail: $\alpha = 1$ with BPSK, $\alpha = 0.5$ with BPSK, and $\alpha = 1$ with QPSK. We show that, although using these pulses requires slightly more power than using rectangular pulses, the use of a time-varying bias yields a large gain in average power efficiency over a fixed bias, especially when the number of subcarriers is large.

Keywords: subcarrier multiplexing, intensity modulation, optical communication.

Subm. to IEEE Conf. on Global Commun., San Antonio, TX, November 25-29, 2001

CONTENTS

I. Introduction

In optical communication systems, multiple-subcarrier modulation (MSM) involves modulation of multiple digital and/or analog information sources onto different electrical subcarriers, which are then modulated onto a single optical carrier [1],[2]. The MSM electrical signal may be modulated onto the optical carrier using intensity, frequency or phase modulation. Most current practical MSM systems use intensity modulation (IM) with direct detection (DD), owing to its simple implementation. MSM permits asynchronous multiplexing of numerous, possibly heterogeneous information streams, and permits a receiver to demodulate only the streams of interest. For this reason, MSM with IM/DD is widely used in optical fiber distribution of video signals [1].

The main drawback of MSM with IM/DD is poor optical average power efficiency. This arises because the MSM electrical signal is a sum of modulated sinusoids, and thus takes on both negative and positive values. Optical intensity (instantaneous power) must be non-negative. Hence, a d.c. bias must be added to the MSM electrical signal in order to modulate it onto the intensity of an optical carrier.[1] As the number of subcarriers increases, the minimum value of the MSM electrical signal decreases (becomes more negative), and the required d.c. bias increases. Since the average optical power is proportional to this d.c. bias, the optical average-power efficiency worsens as the number of subcarriers increases [1],[2].

Optical wireless (OW) transmission with IM/DD is an attractive option for high-speed indoor and outdoor links [2],[5],[6]. MSM with has been considered for use in IM/DD OW systems, because the use of several narrow-band subcarriers promises to minimize ISI on multipath channels [2],[7], and because MSM can provide immunity to fluorescent-light noise near d.c. [8]. In most applications of OW, particularly those using portable transmitters, eye safety and power consumption limit the available average optical power. This dictates that in MSM systems, the number of subcarriers should be kept small, e.g., under ten.

In [9], we showed that the optical average power requirement in IM/DD MSM systems using BPSK or QPSK digital modulation with rectangular pulses can be reduced by replacing the fixed d.c. bias by a bias that varies on a symbol-by-symbol basis. During each symbol period, the minimum bias required to achieve non-negativity is utilized. The resulting time-varying bias signal is a baseband pulse-amplitude modulation (PAM) signal using rectangular pulses, and is piecewise constant. Although this technique provides good optical power efficiency, for some applications, it is necessary for the pulses to be bandlimited.

1. It is also possible to achieve non-negative intensity by clipping the negative part of the electrical signal. For example, optical fiber MSM IM/DD systems typically use tens to hundreds of subcarriers [1]. In an effort to reduce the optical average power requirement, the bias is often reduced to the point that some clipping occurs. Techniques intended to reduce the impact of clipping noise in optical fiber MSM systems have been investigated [3],[4]. However, clipping introduces clipping noise and interference, which are difficult to characterize analytically. Thus, we restrict our study to the non-clipping case.

In this paper, we describe a technique to use bandlimited transmit pulse shapes in IM/DD MSM systems that utilize BPSK or QPSK digital modulation. In particular, we consider pulses whose Fourier transform is the square root of a raised cosine, with rolloff factor α between 0 and 1. For brevity, in the remainder of this paper, we refer to these as "root-raised-cosine pulses". Biasing an MSM electrical signal with root-raised-cosine pulses is significantly more complicated than biasing with rectangular pulses, since the tails of the root-raised-cosine pulses extend infinitely in time in both directions. We describe in detail techniques for designing the bias signal in the following cases: $\alpha = 1$ with BPSK, $\alpha = 0.5$ with BPSK, and $\alpha = 1$ with QPSK. When implementing the methods described here, the calculations need only be performed once to generate a lookup table, and the methods can be implemented using table-lookup during transmission.

The remainder of this paper is organized as follows. The IM/DD channel model is described in Section II. Our multiple-subcarrier transmission scheme using bandlimited pulse shapes is presented in Section III. Design of the baseband bias signal is described in Section IV. The performance is evaluated and compared to using rectangular pulses in Section V. Concluding remarks are given in Section VI.

II. IM/DD Optical Channel Model

Indoor OW channels with IM/DD can be described by [2]:

$$y(t) = rx(t) \otimes h(t) + n(t), \tag{1}$$

where the symbol $\otimes$ represents convolution. The channel input $x(t)$ is the transmitted optical intensity, which must be non-negative:

$$x(t) \geq 0. \tag{2}$$

The average optical power P is given by the mean value of $x(t)$:

$$P = E[x(t)] \tag{3}$$

where, in this paper, all expectations are taken with respect to time and with respect to realizations of the signal $x(t)$, i.e. different transmitted messages. The channel output $y(t)$ is the received photocurrent. Here, r is the photodetector responsivity. The channel is a fixed, linear system having impulse response $h(t)$ and frequency response $H(j\omega)$. In this study, we will neglect multipath distortion, so that $h(t) = H(j0)\delta(t)$, where $\delta(t)$ is an impulse function. In OW systems with IM/DD, receiver thermal noise and intense ambient shot noise can be modeled as Gaussian and independent of the transmitted signal. We will model the noise $n(t)$ as white Gaussian and independent of $x(t)$, with two-sided power spectral density (PSD) N_0. The results obtained here are applicable also to fiber-optic channels provided that dispersion is negligible, and the noise is white Gaussian and independent of the transmitted optical signal.

III. Multiple-Subcarrier Transmission Scheme Using Bandlimited Pulses

A. MSM Signal and Bias Signal

Fig. 1(a) and (b) depict the transmitter and receiver design used in the proposed MSM transmission scheme with QPSK. The transmitter and receiver are identical for BPSK, except that all sine branches are omitted. Referring to Fig. 1(a), the transmitter uses a set of N subcarrier frequencies $\{\omega_n, n = 1, ... N\}$. During each symbol interval of duration T, it transmits a vector of K information bits $\boldsymbol{x}^{(m)} = (x_1^{(m)}, ..., x_K^{(m)})$, $m \in \{0, ..., M-1\}$, where $M = 2^K$. A bit-to-symbol mapper maps $\boldsymbol{x}^{(m)}$ to a corresponding vector of symbol amplitudes, $\boldsymbol{a}^{(m)}$. For QPSK, $K = 2N$, $M = 2^{2N}$, and each such vector is of the form:

$$\boldsymbol{a}^{(m)} = (a_{1c}^{(m)}, a_{1s}^{(m)}, ..., a_{Nc}^{(m)}, a_{Ns}^{(m)}), \tag{4}$$

where $a_{nc}^{(m)} \in \{-1, 1\}$, $a_{ns}^{(m)} \in \{-1, 1\}$, $n = 1, ..., N$. Similarly, for BPSK, $K = N$, $M = 2^N$, and:

$$\boldsymbol{a}^{(m)} = (a_{1c}^{(m)}, ..., a_{Nc}^{(m)}), \tag{5}$$

where $a_{nc}^{(m)} \in \{-1, 1\}$, $n = 1, ..., N$. Defining a pulse shape $g(t)$, the MSM electrical signal for QPSK is:

$$s(t) = \sum_{i=-\infty}^{\infty} \sum_{n=1}^{N} [a_{nc}^{(m_i)} \cos\omega_n t + a_{ns}^{(m_i)} \sin\omega_n t] g(t - iT). \tag{6}$$

The MSM electrical signal for BPSK is identical to (6), except that the sine terms are omitted.

Since $s(t)$ can be positive or negative, the transmitter adds to $s(t)$ a baseband bias signal $b(t)$, and scales the sum by a non-negative constant A to obtain the multiple-subcarrier optical signal $x(t)$. The bias signal $b(t)$ is chosen so that

$$x(t) = A[s(t) + b(t)] \geq 0. \tag{7}$$

The average optical power is:

$$P = AE[s(t)] + AE[b(t)]. \tag{8}$$

We choose the bias signal $b(t)$ to be of the form:

$$b(t) = b_{dc} + \sum_i b_i\left(\boldsymbol{a}^{(m_i)}; \left\{\boldsymbol{a}^{(m_j)}, j \neq i\right\}\right) g(t - iT), \quad (9)$$

which is the sum of a constant b_{dc} and a baseband PAM signal corrresponding to the second term in (9). With the proper choice of T, $\{\omega_n, n = 1, \ldots N\}$, and $g(t)$, each of the subcarriers is orthogonal to the others, and to the time-varying bias signal $b(t)$.

The coefficients $b_i(\ )$ for the baseband PAM signal depends, in general, on both $\boldsymbol{a}^{(m_i)}$, the vector of subcarrier amplitudes for the current symbol, and $\left\{\boldsymbol{a}^{(m_j)}, j \neq i\right\}$, the vectors of all past and future symbols. In the remainder of this paper, we will suppress the arguments of these coefficients and simply write these coefficients as b_i. Our goal is to determine how, given a sequence of symbols to be transmitted, $\left\{\boldsymbol{a}^{(m_i)}\right\}_{i=-\infty}^{\infty}$, we can find the constant b_{dc} and the sequence of coefficients $\{b_i\}_{i=-\infty}^{\infty}$ in (9) so as to minimize the average power given by (8) subject to the non-negativity constraint (7). When implementing the methods described here, calculation of the constant b_{dc} and the coefficients b_i need only be performed once to generate a lookup table and, during transmission, the coefficients employed can be determined using table-lookup.

Fig. 1(b) shows the receiver used under the proposed MSM techniques with QPSK; for BPSK, the receiver is identical, but omits the sine branches. Like a standard QPSK MSM receiver, the receiver of Fig. 1(b) uses a bank of hard decision devices to obtain a vector of detected symbol amplitudes, $\hat{\boldsymbol{a}} = (\hat{a}_{1c}, \hat{a}_{1s}, \ldots, \hat{a}_{Nc}, \hat{a}_{Ns})$, and then employs a symbol-to-bit mapper to map the vector $\hat{\boldsymbol{a}}$ to a vector of detected information bits, $\hat{\boldsymbol{x}} = (\hat{x}_1, \ldots, \hat{x}_K)$.

B. Average Power Requirement

1. Fixed Bias

When using a fixed bias signal, we set $b_i = 0$, $\forall i$, in (9) and use a fixed bias b_{dc}. With an arbitrary pulse shape $g(t)$, the smallest allowable b_{dc} is given by:

$$b_{dc}^{\min} = -\min_t s(t), \quad (10)$$

and the average optical power from (8) is:

$$P = AE[s(t)] + Ab_{dc}^{\min}. \quad (11)$$

2. Time-Varying Bias

With a time-varying bias, $b(t)$ is chosen to be of the form (9), i.e., the sum of a constant and a baseband PAM signal. For an arbitrary pulse shape $g(t)$, to minimize the average optical power *P*, it may be necessary to use both terms. In this case, the average transmitted optical power from (8) is:

$$P = AE[s(t)] + Ab_{dc} + AE[b_i]T^{-1}\int_{-\infty}^{\infty} g(t)dt. \quad \textbf{(12)}$$

C. Root-Raised-Cosine Pulse Shape

In the remainder of this paper, unless otherwise noted, we consider a transmit pulse shape whose Fourier transform is the square root of a raised cosine:

$$g(t) = \frac{4\alpha}{\pi}\left(\frac{\cos\left(\frac{(1+\alpha)\pi t}{T}\right) + T\frac{\sin\left(\frac{(1-\alpha)\pi t}{T}\right)}{4\alpha t}}{1-\left(\frac{4\alpha t}{T}\right)^2}\right), \quad \textbf{(13)}$$

where α is the roll-off factor taking on values in the range $0 \le \alpha \le 1$, and T is the symbol period. The Fourier transform of $g(t)$ is given by:

$$G(f) = \begin{cases} T & 0 \le |f| < \frac{(1-\alpha)}{2T} \\ T\sqrt{\cos\left(\pi T\frac{|f|-\frac{1}{2T}}{2\alpha}+\frac{\pi}{4}\right)} & \frac{(1-\alpha)}{2T} \le |f| < \frac{(1+\alpha)}{2T} \\ 0 & \text{otherwise} \end{cases}. \quad \textbf{(14)}$$

We choose the set of subcarrier frequencies $\{\omega_n\}$ to be:

$$\omega_n = n\frac{2\pi}{T}(\alpha+1), \quad \textbf{(15)}$$

where $n = 1, \ldots, N$. Assuming $E_m[a_{nc}^{(m)}] = 0$ for each *n* and using the fact that: $\int_{-\infty}^{\infty} g(t)\sin(\omega_n t)dt = 0$ for each *n*, we find that $E[s(t)] = 0$, which simplifies evaluation of (11) and (12).

1. Fixed Bias

For the special case of BPSK, for a given *N*, the smallest allowable fixed bias is given by:

$$b_{dc}^{\min} = -\sum_{i=-\infty}^{\infty}\sum_{n=1}^{N}[a_{nc}^{(m_i)}\cos\omega_n t]g(t-iT)\Big|_{t=0}, \tag{16}$$

where

$$a_{nc}^{(m_i)} = \begin{cases} -1 & \text{if } i = 0 \\ 1 & \text{if } i \neq 0 \end{cases} \tag{17}$$

for $n = 1, \ldots, N$. This is true because the most negative value of $s(t)$ among all possible choices of $a_{nc}^{(m_i)}$ (for a fixed N) occurs at $t = 0$ when the $i = 0$ symbol has all its coefficients equal to -1 and all other neighboring symbols have all their coefficients equal to 1.

Using the fact that $E[s(t)] = 0$, (11) becomes:

$$P = Ab_{dc}^{\min}, \tag{18}$$

where $b_{dc}^{\min}$ is given by (16).

2. Time-Varying Bias

Using the facts that $\int_{-\infty}^{\infty} g(t)dt = T$ and $E[s(t)] = 0$, we can evaluate (12) to obtain the average transmitted power:

$$P = Ab_{dc} + AE[b_i]. \tag{19}$$

IV. Designing the Baseband Bias Signal

As stated previously, our goal is to determine how, given a sequence of symbols to be transmitted, $\left\{\boldsymbol{a}^{(m_i)}\right\}_{i=-\infty}^{\infty}$, we can find the constant b_{dc} and the sequence of coefficients $\{b_i\}_{i=-\infty}^{\infty}$ in (9) so as to minimize the average power given by (8) subject to the non-negativity constraint (7). In a typical implementation, computation of b_{dc} and the b_i will be performed once to generate a lookup table; during transmission, the coefficients employed can be determined using table-lookup. Instead of trying to find the optimal b_{dc} and b_i, we attempt to find "good" values by using heuristic techniques.

In the case of time-varying bias, to simplify the computation, we let $b_{dc} = 0$ in (9) and find the values b_i that satisfy the non-negativity criterion. Although we have not proven that the choice $b_{dc} = 0$ minimizes the average transmit power P, in our numerical studies, we have been unable to obtain a lower P by allowing b_{dc} to take on a nonzero value. When we set $b_{dc} = 0$, using (19), the average transmit power with time-vaying bias is simply:

$$P = AE[b_i]. \tag{20}$$

The following sections illustrate how to find the values b_i for each of the following three cases: $\alpha = 1$ with BPSK, $\alpha = 0.5$ with BPSK, and $\alpha = 1$ with QPSK.

A. $\alpha = 1$ with BPSK

Although b_i is a function of the current symbol m_i, and all past and future symbols $\{m_j, j \neq i\}$, the effects of symbols that are "far" from the current symbol are negligible and we need only consider those symbols in the immediate vicinity of the current symbol. In this paper, we consider only the current symbol m_i, four symbols to the immediate past $\{m_j, i–1 \leq j \leq i-4\}$, and four symbols to the immediate future $\{m_j, i+1 \leq j \leq i+4\}$. Let b_i be the sum of five components:

$$b_i = b^{(0)}(m_i) + \sum_{k=1}^{4} b^{(k)}(m_{i-k}, m_i, m_{i+k}), \tag{21}$$

where $m_i \in \{0, \ldots, M-1\}, -\infty < i < \infty$. The first term on the right in (21) is determined solely by the current symbol m_i. Each of the following terms inside the summation is determined jointly by the current symbol m_i and two neighboring symbols, m_{i-k} and m_{i+k}, where $k \in \{1, 2, 3, 4\}$. The superscripts in (21) denote the relative locations of the neighboring symbols that are being considered in each term.

1. Computing $b^{(0)}(m_i)$

To compute this term, we assume that there is no interference from the neighboring symbols and find the minimum bias necessary to achieve non-negativity within the current symbol period. Since we are using BPSK, we let:

$$s(t) = \sum_{n=1}^{N} [a_{nc}^{(m_i)} \cos\omega_n t] g(t) + b_0 g(t). \tag{22}$$

Then, $b^{(0)}(m_i)$ is the minimum value of b_0 satisfying $s(t) \geq 0$ for $-\frac{T}{2} \leq t \leq \frac{T}{2}$. There are M possible values of $b^{(0)}(m_i)$, one for each value of $m \in \{0, \ldots, M-1\}$.

2. Computing $b^{(k)}(m_{i-k}, m_i, m_{i+k})$

Each of these terms is a joint contribution from the $(i-k)$th and $(i+k)$th symbols, where $k \in \{1, 2, 3, 4\}$. It represents the additional bias that must be added due to the interference caused by the tails of the $(i-k)$th and $(i+k)$th symbols to preserve non-negativity in the current symbol period. Assuming we have already found the M possible $b^{(0)}(m_i)$, for a given $k \in \{1, 2, 3, 4\}$, let:

$$s(t) = \sum_{i=-k,0,k} \sum_{n=1}^{N} \left[a_{nc}^{(m_i)} \cos\omega_n t\right] g(t-iT) + \sum_{i=-k,k} b^{(0)}(m_i) g(t-iT) + \left(b^{(0)}(m_0) + b_{\text{resid}}\right) g(t). \quad \textbf{(23)}$$

Then, $b^{(k)}(m_{i-k}, m_i, m_{i+k})$ is the minimum value of b_{resid} satisfying $s(t) \geq 0$ for $-\left(\frac{T}{2} + \Delta\right) \leq t \leq \left(\frac{T}{2} + \Delta\right)$. Note that b_{resid} can be negative. Although it is intuitively appealing to set $\Delta = 0$ so that the width of the interval is one symbol period, in our numerical studies, we obtain better results if we let $\Delta = 0.05T$. For each $k \in \{1, 2, 3, 4\}$, there are M^3 possible values of $b^{(k)}(m_{i-k}, m_i, m_{i+k})$.

We can also improve the results by iterating one or more times when finding the $b^{(k)}(m_{i-k}, m_i, m_{i+k})$ terms. For example, during the second iteration, we should use $b^{(0)}(m_i) + b^{\text{avg}}(m_i)$ rather than $b^{(0)}(m_i)$ in (23), where $b^{\text{avg}}(m_i)$ is the mean of all possible values of $b^{(k)}(m_{i-k}, m_i, m_{i+k})$ given m_i:

$$b^{\text{avg}}(m_i) = \frac{1}{4M^2} \sum_{k=1}^{4} \sum_{m_{i-k}=0}^{M-1} \sum_{m_{i+k}=0}^{M-1} b^{(k)}(m_{i-k}, m_i, m_{i+k}). \quad \textbf{(24)}$$

3. Results

Fig. 2 shows an example with three subcarriers, $\alpha = 1$, and BPSK. The possible symbol vectors are $\boldsymbol{a}^{(m)}$, where $m \in \{0, \ldots, 7\}$. Let m be the base-10 value of the binary string $x_{3c}x_{2c}x_{1c}$, where $x_{nc} = 1$ if $a_{nc} = 1$ and $x_{nc} = 0$ if $a_{nc} = -1$, $n \in \{1, 2, 3\}$. We use a sample sequence of 13 symbols $\{m_i\}_{i=-6}^{6}$ = {5, 1, 3, 2, 7, 6, 3, 0, 5, 2, 4, 4, 7}. The corresponding b_i values are found to be $\{b_i\}_{i=-6}^{6}$ = {3.00, 1.59, 2.19, 1.66, 1.61, 2.49, 2.43, 3.66, 3.02, 1.83, 1.72, 1.69, 1.32}. Fig. 2(a) shows the MSM electrical signal $s(t)$, Fig. 2(b) shows the baseband bias $b(t)$, and Fig. 2(c) shows the transmittable signal $x(t) = s(t) + b(t)$. We use $T = 1$, $b_{dc} = 0$, and $A = 1$ for all three plots. In Fig. 2(c), note that there are edge effects because we do not have four neighboring symbols on each side for the edge symbols. The range of t within which the algorithm is in full force is [–2, 2] and, for all values of t in this range, the signal satisfies the non-negativity criterion.

B. $\alpha = 0.5$ with BPSK

Unlike the case of $\alpha = 1$ with BPSK, the symbol period T is not an integer multiple of the period of the sum of subcarriers in (6). This can be seen from (15), keeping in mind that all subcarriers originate from a single transmitter and are therefore synchronized. Letting $\alpha = 0.5$, the period of the $n = 1$ subcarrier is:

$$\tau_1 = \frac{2\pi}{\omega_1} = \frac{2T}{3}, \tag{25}$$

where T is the symbol period. From (25), it is apparent that the $n = 1$ subcarrier does not align in the same way with adjacent symbol centers. In fact, the period of the sum of subcarriers in (6) is $\frac{2T}{3}$ and the sum of subcarriers aligns in the same way with *every other* symbol center. To distinguish between the two types of symbol positions, let us call the symbol positions "even" for all even i in (6) and "odd" for all odd i.

Computing the b_i for the $\alpha = 0.5$ with BPSK case is nearly identical to computing the b_i for the $\alpha = 1$ with BPSK case. The only difference is that we must take into account the two types of symbol positions. We do this by converting our original symbol string using a process which is described below. Then, we can treat all symbols in the converted symbol string as symbols in even positions and apply the method presented in Part A.

1. Symbol String Conversion

For $\alpha = 0.5$, a symbol with $m \in \{0, \ldots, M-1\}$ in an odd position is equivalent to a symbol having a different $m \in \{0, \ldots, M-1\}$ in an even position. Let m_o be a symbol at an odd position and m_e be the equivalent symbol at an even position. The one-to-one mapping is described by:

$$\begin{cases} a_{nc}^{(m_o)} \rightarrow a_{nc}^{(m_e)} & n \text{ even} \\ -a_{nc}^{(m_o)} \rightarrow a_{nc}^{(m_e)} & n \text{ odd} \end{cases} \tag{26}$$

for $n = 1, \ldots, N$.

The following explains why this symbol string conversion works. Let the sum of subcarriers for BPSK be:

$$z(t) = \sum_{n=1}^{N} [a_{nc}^{(m_i)} \cos \omega_n t], \tag{27}$$

where ω_n is given by (15). Since (27) has period $\frac{2T}{3}$ for $\alpha = 0.5$, we can convert odd positions into even positions (and even positions into odd positions) by time-shifting the sum of subcarriers:

$$\begin{aligned} z\left(t + \frac{T}{3}\right) &= \sum_{n=1}^{N} \left[a_{nc}^{(m_i)} \cos \omega_n \left(t + \frac{T}{3}\right)\right] \\ &= \sum_{n=1}^{N} [a_{nc}^{(m_i)} \cos(\omega_n t + \pi n)] \end{aligned} \tag{28}$$

$$= \sum_{\substack{n=1 \\ \text{n even}}}^{N} [a_{nc}^{(m_i)} \cos(\omega_n t)] + \sum_{\substack{n=1 \\ \text{n odd}}}^{N} [-a_{nc}^{(m_i)} \cos(\omega_n t)]$$

Thus, when we go from an odd to even position (or, vice versa), the coefficients $a_{nc}^{(m_i)}$ with n even stay the same and the coefficients $a_{nc}^{(m_i)}$ with n odd change sign.

2. Results

Fig. 3 shows an example with three subcarriers, $\alpha = 0.5$, and BPSK. The possible symbol vectors are $\boldsymbol{a}^{(m)}$, where $m \in \{0, \ldots, 7\}$. As in Part A, let m be the base-10 value of the binary string $x_{3c}x_{2c}x_{1c}$, where $x_{nc} = 1$ if $a_{nc} = 1$ and $x_{nc} = 0$ if $a_{nc} = -1$, $n \in \{1, 2, 3\}$. We start with a sample sequence of 13 symbols $\{m_i\}_{i=-6}^{6}$ = {5, 1, 3, 2, 7, 6, 3, 0, 5, 2, 4, 4, 7}. We convert it using the symbol string conversion process described above and we get $\{m_i\}_{i=-6}^{6}$ = {5, 4, 3, 7, 7, 3, 3, 5, 5, 7, 4, 1, 7}. Now, we can assume that all of the symbols are in even positions and apply the same algorithm used in Part A. The corresponding b_i values are found to be $\{b_i\}_{i=-6}^{6}$ = {3.00, 1.87, 2.03, 1.41, 1.39, 2.38, 2.17, 3.74, 2.61, 1.73, 1.48, 1.78, 1.32}. Unlike in Part A where we used $\Delta = 0.05T$ to find the $b^{(k)}(m_{i-k}, m_i, m_{i+k})$, we use $\Delta = 0.15T$ here because it produces better results. Fig. 3(a) shows the MSM electrical signal $s(t)$, Fig. 3(b) shows the baseband bias $b(t)$, and Fig. 3(c) shows the transmittable signal $x(t) = s(t) + b(t)$. We use $T = 1$, $b_{dc} = 0$, and $A = 1$ for all three plots. As in Fig. 2(c), there are edge effects in Fig. 3(c). But, when t is in the range [–2,2] where the algorithm is in full force, the signal again satisfies the non-negativity criterion.

C. $\alpha = 1$ with QPSK

The process to find the b_i is nearly identical to the method described in Part A. The difference is that we now have an additional sine subcarrier term for each cosine subcarrier term in $s(t)$. To find the $b^{(0)}(m_i)$, we use the QPSK version of (22):

$$s(t) = \sum_{n=1}^{N} [a_{nc}^{(m_i)} \cos\omega_n t + a_{ns}^{(m_i)} \sin\omega_n t] g(t) + b_0 g(t). \quad \textbf{(29)}$$

To find the $b^{(k)}(m_{i-k}, m_i, m_{i+k})$, we use the QPSK version of (23):

$$s(t) = \sum_{i=-k,0,k} \sum_{n=1}^{N} \left[a_{nc}^{(m_i)} \cos\omega_n t + a_{ns}^{(m_i)} \sin\omega_n t\right] g(t-iT) + \sum_{i=-k,k} b^{(0)}(m_i) g(t-iT) + \left(b^{(0)}(m_0) + b_{\text{resid}}\right) g(t). \quad \textbf{(30)}$$

Even though there are strictly only *N* subcarriers, the complexity of the algorithm is the same as a BPSK case with *2N* subcarriers.

1. Results

Fig. 4 shows an example with two subcarriers, $\alpha = 1$, and QPSK. The possible symbol vectors are $\boldsymbol{a}^{(m)}$, where $m \in \{0, \ldots, 15\}$. Let m be the base-10 value of the binary string $x_{2s}x_{1s}x_{2c}x_{1c}$, where $x_{nc} = 1$ if $a_{nc} = 1$, $x_{nc} = 0$ if $a_{nc} = -1$, $x_{ns} = 1$ if $a_{ns} = 1$, and $x_{ns} = 0$ if $a_{ns} = -1$, $n \in \{1, 2\}$. We use a sample sequence of 13 symbols $\{m_i\}_{i=-6}^{6}$ = {3, 9, 7, 14, 12, 7, 0, 13, 7, 9, 12, 14, 11}. The corresponding b_i values are found to be $\{b_i\}_{i=-6}^{6}$ = {2.10, 2.88, 3.09, 3.14, 3.37, 3.17, 3.44, 2.18, 3.18, 2.94, 3.17, 3.04, 2.75}. Fig. 4(a) shows the MSM electrical signal $s(t)$, Fig. 4(b) shows the baseband bias $b(t)$, and Fig. 4(c) shows the transmittable signal $x(t) = s(t) + b(t)$. We use $T = 1$, $b_{dc} = 0$, and $A = 1$ for all three plots. Although there are edge effects in Fig. 4(c), when *t* is in the range [2,2] where the algorithm is in full force, the signal satisfies the non-negativity criterion.

V. Performance Evaluation

In this section, we evaluate the bandwidth and power requirements of the proposed MSM biasing scheme, comparing it to on-off keying (OOK). In both cases, R_b represents the information bit rate, *B* represents the total electrical bandwidth required at the receiver, and P_b represents the probability of information bit error.

We first consider a reference system using OOK with rectangular pulses of duration *T* (described by (13)) and symbol rate 1/*T*. Following [2], we have $R_b = 1/T$ and $B = 1/T$, where the bandwidth requirement *B* is taken to equal the first null in the PSD of the transmitted signal. For OOK, we have $P_b = Q\left(\sqrt{r^2P^2T/N_0}\right)$, where $Q(x) = (1/2)\mathrm{erfc}(x/\sqrt{2})$. Hence, the required average optical power to achieve a required bit-error probability $P_{b,\mathrm{req}}$ is given by:

$$P_{\mathrm{req}} = \sqrt{\frac{N_0}{r^2T}}Q^{-1}(P_{b,\mathrm{req}}) \text{ (OOK).} \tag{31}$$

For the proposed MSM scheme, the information bit rates are given by:

$$R_b = \frac{N-L}{T} \text{ (BPSK),} \tag{32}$$

and

$$R_b = \frac{2(N-L)}{T} \text{ (QPSK).} \tag{33}$$

With either BPSK or QPSK, the electrical bandwidth requirement is given by:

$$B = \frac{2N+1}{2T}(1+\alpha)\,. \tag{34}$$

These values of B correspond to the upper edge of the highest frequency component of the PSD of the MSM signal $s(t)$.

For MSM using BPSK or QPSK (assuming Gray coding with QPSK) and root-raised-cosine pulse shapes, it is easily shown that $P_b = Q\left(\sqrt{r^2A^2T/(2N_0)}\right)$. The amplitude A required to achieve a required bit-error probability $P_{b,\mathrm{req}}$ is $\sqrt{2N_0/(r^2T)}Q^{-1}(P_{b,\mathrm{req}})$. Using (20), we have:

$$P_{\mathrm{req}} = E[b_i]\sqrt{\frac{2N_0}{r^2T}}Q^{-1}(P_{b,\mathrm{req}})\,. \tag{35}$$

In comparing various modulation techniques, we consider the normalized bandwidth requirement B/R_b, and the normalized power requirement, P/P_{OOK}, which represents the average optical power required to achieve an arbitrary specified bit-error probability compared to that required by OOK to achieve the same bit-error probability, assuming a fixed bit rate R_b.[1]

Fig. 5 presents the normalized power requirement versus N, the total number of subcarrier frequencies. We consider root-raised-cosine pulses with fixed and time-varying bias. For comparison, we also consider rectangular pulses with fixed and time-varying bias. For each scheme, a label indicates the least-squares linear fit to the normalized power requirement (in dB) for $1 \le N \le 6$.

In Fig. 5(a), we consider the case $\alpha = 1$ with BPSK. Considering fixed bias, the normalized power requirement for root-raised-cosine pulses is approximately $2.9 + 5.00\ \log_{10}N$, while that for rectangular pulses is approximately $1.5 + 5.00\ \log_{10}N$. Both power requirements increase with N at about the same rate, but root-raised-cosine pulses require about 1.3 dB more power at each N. In the case of time-varying bias, the normalized power requirement for root-raised-cosine pulses is about $2.0 + 1.52\ \log_{10}N$, while that for rectangular pulses is about $1.5 + 1.52\ \log_{10}N$. Both power requirements increase with N at nearly the same rate, but root-raised-cosine pulses require approximately 0.5 dB more power at each N. We note that with rectangular pulses, for $N = 1$, the normalized power requirements for fixed and time-varying bias are identical. This occurs because the minimum value of $s(t)$ in a symbol interval is independent of whether a positive or negative amplitude is transmitted, and the time-varying bias reduces to a fixed bias. By contrast, with root-raised-cosine

[1]. We actually consider $10\ \log_{10}(P/P_{\mathrm{OOK}})$, which has units of optical dB.

pulses for $N = 1$, the fixed and time-varying bias signals are not equivalent, and time-varying bias is able to achieve a normalized power requirement that is about 0.8 dB lower.

In Fig. 5(b), we consider the case $\alpha = 0.5$ with BPSK, and obtain results that are very similar to the $\alpha = 1$ BPSK case.

In Fig. 6, we plot the normalized power requirement versus the normalized bandwidth requirement for our biasing scheme, comparing it to OOK. Fig. 6(a) shows the $\alpha = 1$ BPSK case and Fig. 6(b) shows the $\alpha = 0.5$ BPSK case.

VI. Concluding Remarks

We have described a technique to reduce the average transmitted power in MSM IM/DD systems using BPSK or QPSK modulation with root-raised-cosine pulses by using a time-varying bias signal. This time-varying bias signal is a baseband PAM signal using root-raised-cosine pulses. For each of the three cases we presented: $\alpha = 1$ with BPSK, $\alpha = 0.5$ with BPSK, and $\alpha = 1$ with QPSK, we can find the biasing coefficient b_i for each symbol by looking at the current symbol and four neighboring symbols on each side.

When implementing the algorithm described in this paper, it is better to perform the calculations off-line and then use table-lookup, especially if the number of subcarriers is large. This assumes that the storage requirement is not an issue. Typically, for BPSK, five tables need to be generated: one $M \times 1$ table containing the $b^{(0)}(m_i)$ values, and one $M \times M \times M$ table for each $k \in \{1, 2, 3, 4\}$ containing the $b^{(k)}(m_{i-k}, m_i, m_{i+k})$ values. Using table-lookup is preferred over calculating the on-the-fly because it can be performed in constant time. Since each table is indexed by the symbols $m \in \{0, \ldots, M-1\}$, no searching is necessary when a table value is needed.

Several problems remain for future study [9]. These include: (a) block coding techniques to further reduce the average power requirement for non-rectangular transmit pulse shapes, (b) obtaining analytical bounds on the transmit power requirements with block coding and/or time-varying bias, (c) design of block codes that minimize the transmit power required to achieve a specified information bit-error probability, (d) utilizing information contained in the time-varying bias signal to enhance detection efficiency [10], and (e) possible impairment caused by a time-varying bias signal on channels exhibiting memory or nonlinearity.

VII. Acknowledgments

This research has been supported by NSF Grant Number ECS-9710065.

VIII. References

[1] T.E. Darcie, "Subcarrier Multiplexing for Lightwave Networks and Video Distribution Systems", *IEEE J. on Sel. Areas in Commun*., vol. 8, (no.7), Sept. 1990, pp.1240-1248.

[2] J.R. Barry, *Wireless Infrared Communications*, Kluwer Academic Publishers, Boston, 1994.

[3] Qi Pan and R.J. Green, "Preclipping AM/QAM Hybrid Lightwave Systems with Bandstop Filtering", *IEEE Photon. Technol. Lett.*, vol. 8, (no.8), Aug. 1996, pp.1079-1081.

[4] Qun Shi, "Error Performance of OFDM-QAM in Subcarrier Multiplexed Fiber-Optic Transmission", *IEEE Photon. Technol. Lett.*, vol. 9 (no. 6), June 1997, pp. 845-847.

[5] F. R. Gfeller and U. H. Bapst, "Wireless In-House Data Communication via Diffuse Infrared Radiation," *Proceedings of the IEEE*, Vol. 67, No. 11, pp. 1474-1486, November 1979.

[6] D.J.T. Heatley, D.R. Wisely, I. Neild, and P. Cochrane, "Optical Wireless: The Story So Far", *IEEE Commun. Mag.*, vol. 36, (no.12), Dec. 1998. pp.72-74, 79-82.

[7] J.B. Carruthers and J.M. Kahn, "Multiple-Subcarrier Modulation for Non-Directed Wireless Infrared Communication", *IEEE J. Sel. Areas in Commun.*, vol. 14, pp. 538-546, April 1996.

[8] R. Narasimhan, M.D. Audeh and J.M. Kahn, "Effect of Electronic-Ballast Fluorescent Lighting on Wireless Infrared Links", *IEE Proceedings-Optoelectronics*, vol. 143, no. 6, pp. 347-354, December 1996.

[9] R. You and J.M. Kahn, "Average Power Reduction Techniques for Multiple-Subcarrier Intensity-Modulated Optical Signals", submitted to *IEEE Trans. on Commun.*, September 1999.

[10] S. Hranilovic and D.A. Johns, "A Multilevel Modulation Scheme for High-Speed Wireless Infrared Communications", *Proc. of IEEE Intl. Symp. on Circuits and Systems*, Orlando, FL, May 30-June 2, 1999.

[11] J.G. Proakis, *Digital Communications*, 3rd Edition, McGraw-Hill, 1995.

IX. Figures

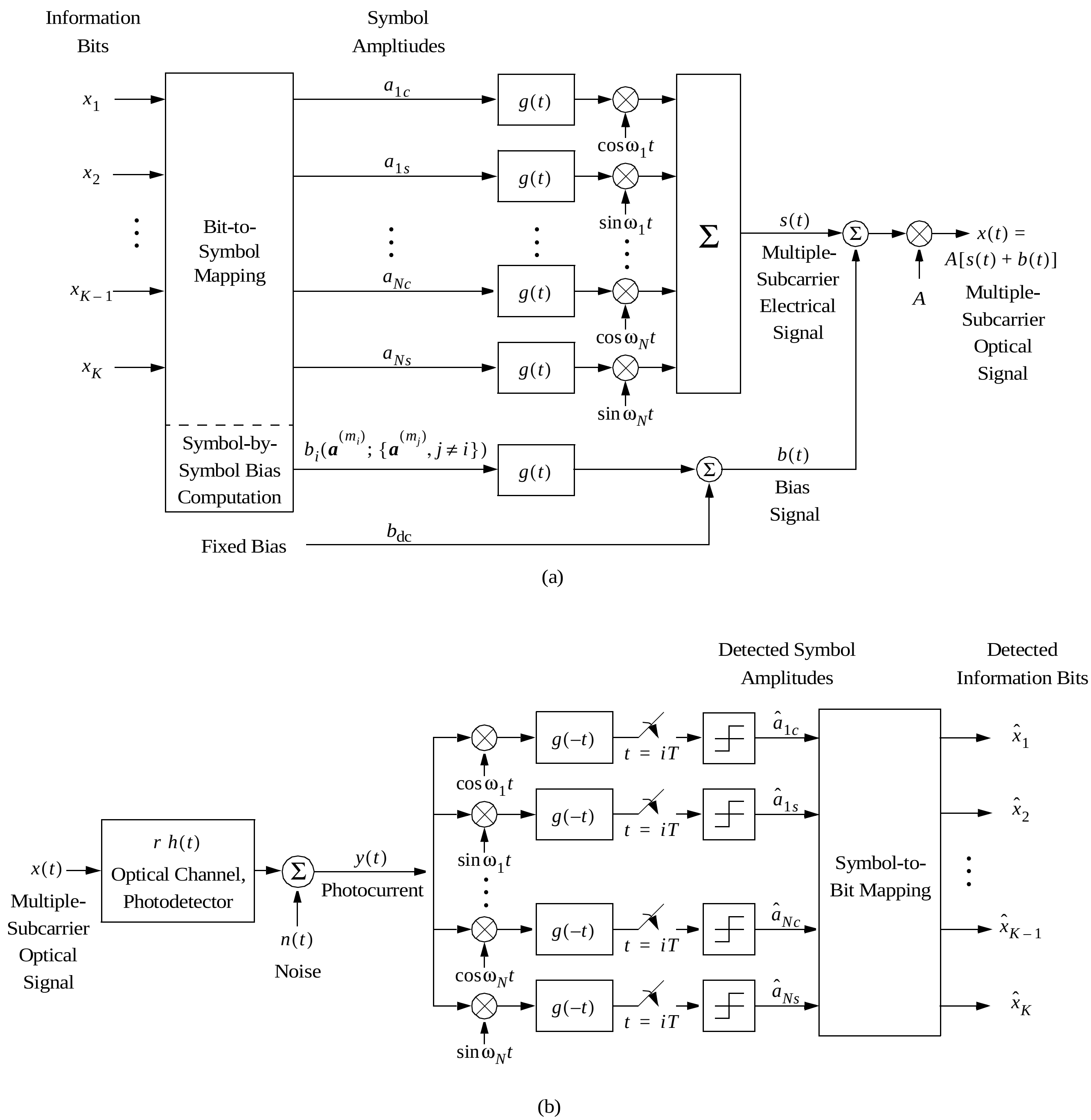

Fig. 1. Power-efficient multiple-subcarrier modulation scheme: (a) transmitter, and (b) channel and receiver. QPSK modulation is assumed in this figure.

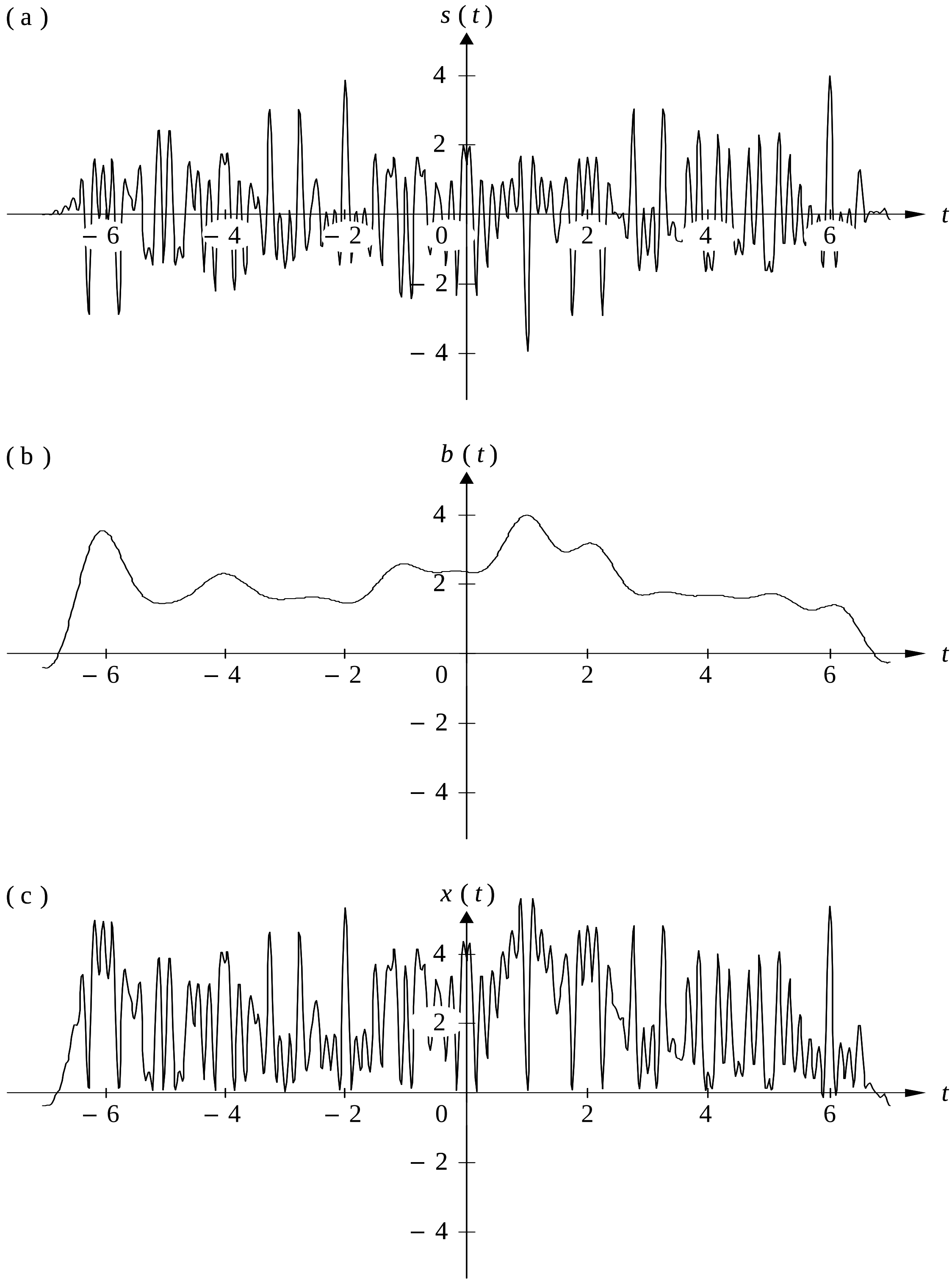

Fig. 2. Example of $\alpha = 1$, BPSK, 3 subcarriers. Symbol sequence: $\{m_i\}_{i=-6}^{6} = \{5, 4, 6, 2, 7, 3, 6, 0, 5, 2, 1, 1, 7\}$ (a) $s(t)$ (b) $b(t)$ (c) $x(t) = s(t) + b(t)$.

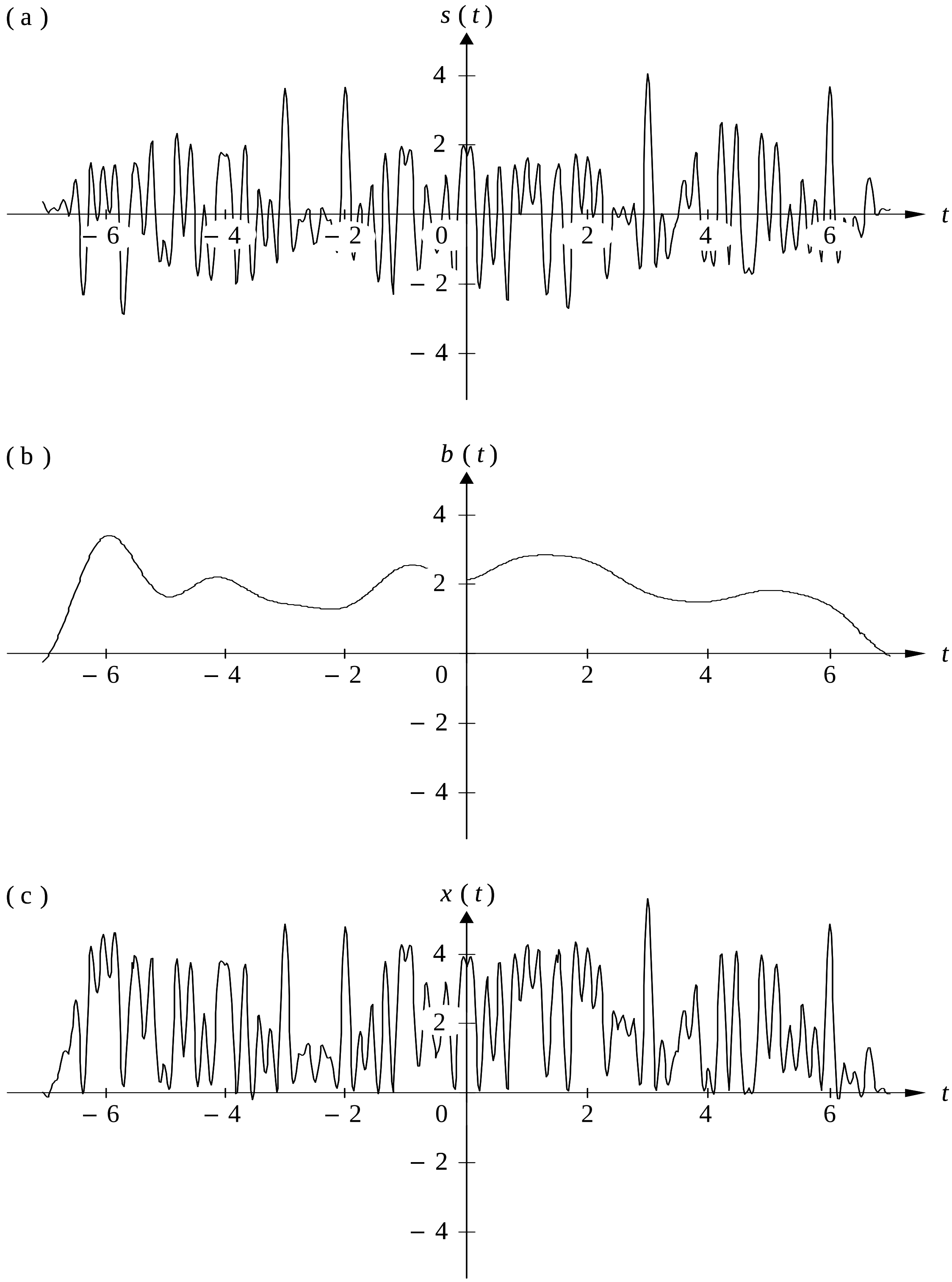

Fig. 3. Example of $\alpha = 0.5$, BPSK, 3 subcarriers. Symbol sequence: $\{m_i\}_{i=-6}^{6} = \{5, 4, 6, 2, 7, 3, 6, 0, 5, 2, 1, 1, 7\}$ (a) $s(t)$ (b) $b(t)$ (c) $x(t) = s(t) + b(t)$.

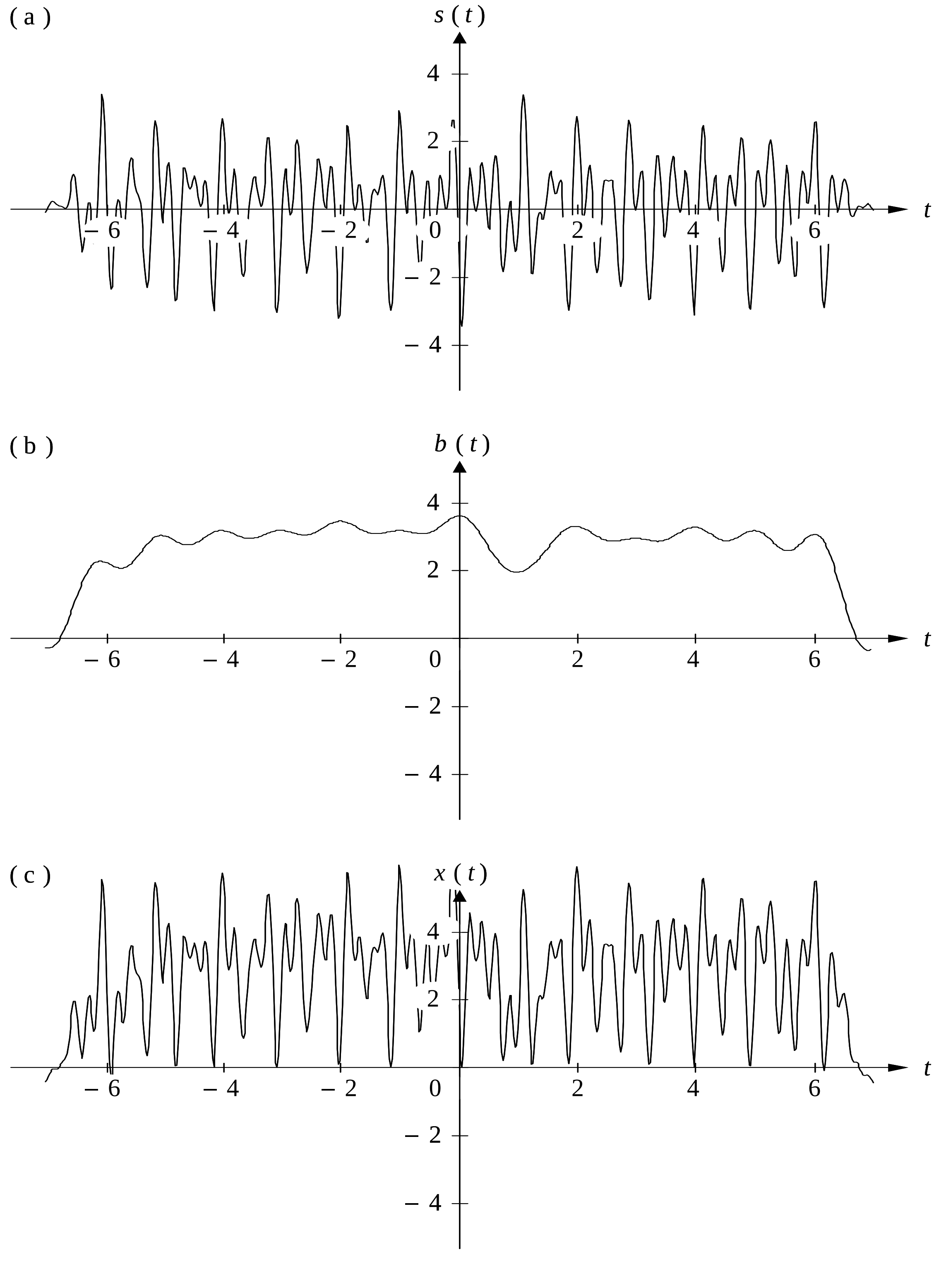

Fig. 4. Example of $\alpha = 1$, QPSK, 2 subcarriers. Symbol sequence: $\{m_i\}_{i=-6}^{6} = \{3, 9, 7, 14, 12, 7, 0, 13, 7, 9, 12, 14, 11\}$ (a) $s(t)$ (b) $b(t)$ (c) $x(t) = s(t) + b(t)$.

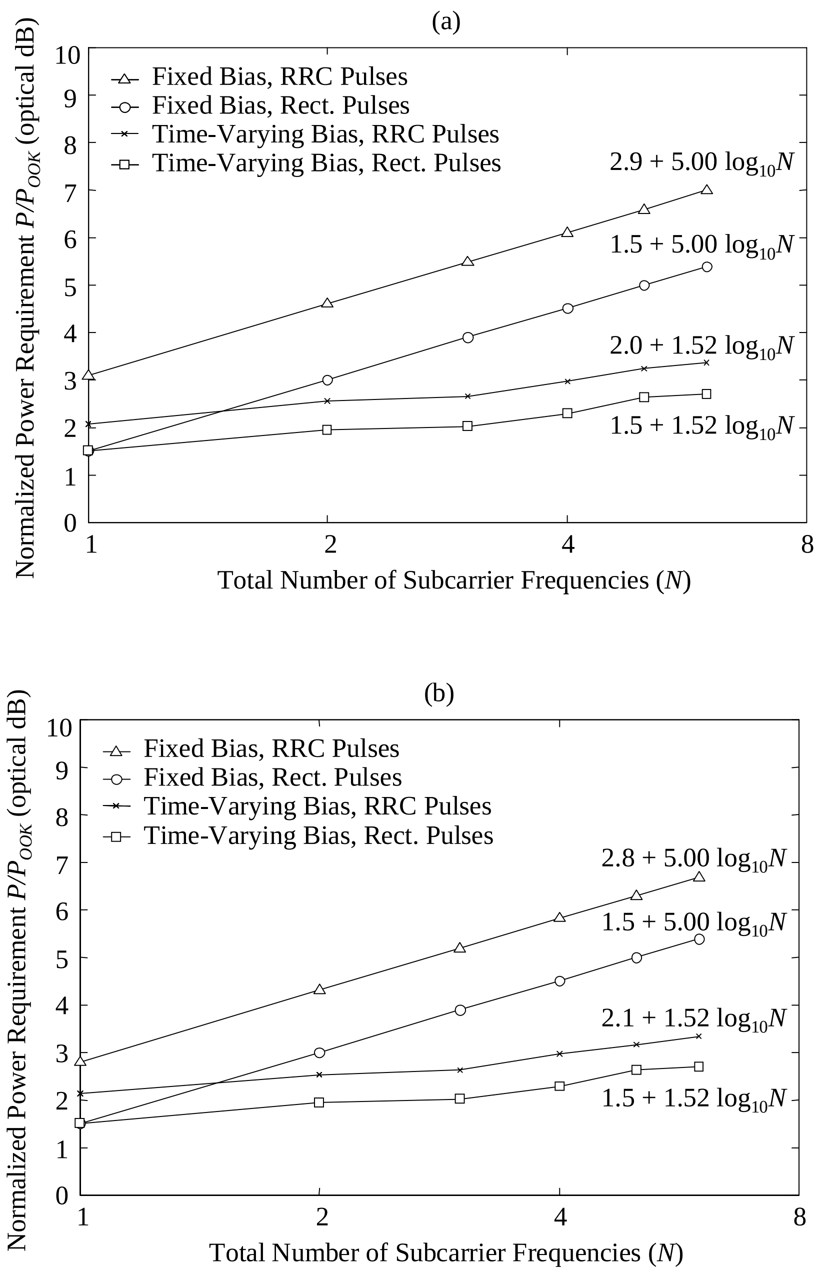

Fig. 5. Normalized optical power requirement versus number of subcarriers for BPSK (a) $\alpha = 1$ (b) $\alpha = 0.5$. For each scheme, a label indicates the least-squares linear fit to the normalized power requirement (in dB) for $1 \leq N \leq 6$.

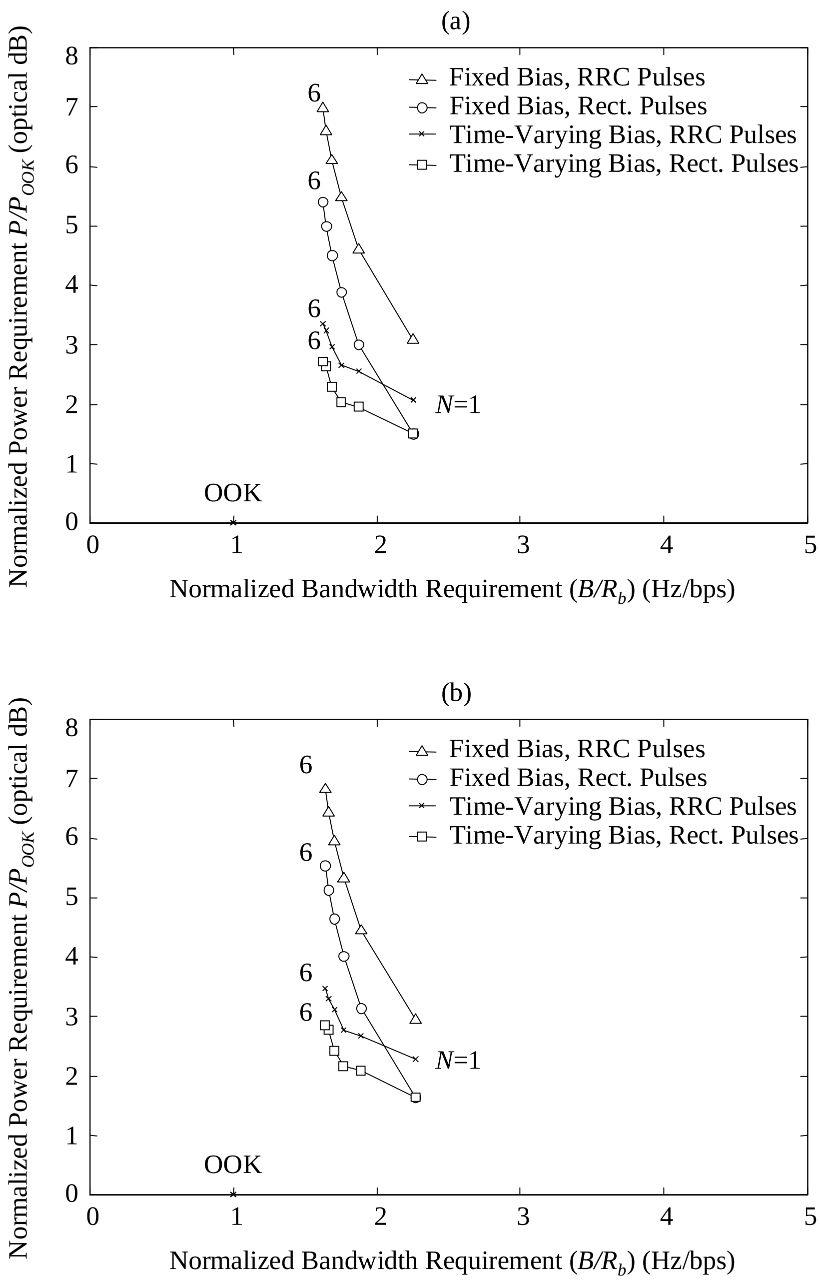

Fig. 6. Normalized optical power requirement versus normalized bandwidth requirement for BPSK (a) $\alpha = 1$ (b) $\alpha = 0.5$.