

交调失真影响下星间微波光子链路接收灵敏度优化

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摘要 高输入信号功率时的交调失真是降低星间微波光子链路接收灵敏度的主要因素。考虑输入 N 路四相相移键控(QPSK)调制信号的情形,建立了强度调制/直接检测(IM/DD)星间微波光子链路模型。利用傅里叶级数展开、傅里叶变换/逆傅里叶变换和 Graf 加法定理,推导出了接收信号任意谐波和交调分量的解析表达式,在确定三阶交调失真个数的基础上,得到了接收光功率与信号噪声失真比(SNDR)之间的关系。着重分析了不同调制方式下链路接收灵敏度与信道数和调制系数之间的关系。当调制系数较小时,接收灵敏度对信道数变化不敏感。随着调制系数的增大,接收灵敏度先增大后减小,存在最佳的调制系数可以使链路接收灵敏度最高。

关键词 光通信;微波光子;星间链路;优化;接收灵敏度;三阶交调

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Receiver Sensitivity Optimization of Intersatellite Microwave Photonics Link Considering the Effect of Intermodulation Distortion

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Abstract The intermodulation distortion greatly deteriorates the receiver sensitivity of intersatellite microwave photonics link in the range of high input signal power. The situation of N quadrature phase shift keying (QPSK) modulation signal input is considered and the intensity modulation/direct detection (IM/DD) intersatellite microwave photonics link model is presented. An exact analytical solution for any harmonic and intermodulation components of receiver signal is deduced with the method of Fourier expansion, Fourier transform/inverse Fourier transform and Graf addition theory. Combined with the number of third-order intermodulation distortion, the relationship formula between receiver optical power and signal-to-noise and distortion ratio (SNDR) is derived. The analysis concentrates on the relationship between receiver sensitivity of intersatellite microwave photonic link and the number of channels, the modulation index under different modulation modes. When the modulation index is small, the receiver sensitivity is robust against the change of channel number. The receiver sensitivity increases and then decreases as the increase in modulation index. It shows that an optimum modulation index can maximize the receiver sensitivity of link.

Key words optical communications; microwave photonics; intersatellite link; optimization; receiver sensitivity; third-order inter-modulation

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1 引 言

将微波光子技术应用到卫星通信系统中,不仅

可以降低卫星载荷的体积、重量和功耗,提高射频隔离度,而且可以提供宽带、透明的数据传输和处理能

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力^[1~4]。星间微波光子链路中,卫星接收的来自地面站的多路微波信号调制到光载波上传输到另一颗卫星,信号经远距离传输损耗大,到达接收端时十分微弱,而且多路微波信号之间交调干扰严重,降低了系统信噪比。因此,需要优化星间微波光子链路的接收灵敏度。

目前微波光子链路性能建模分析主要集中在单路射频信号输入^[5~9],没有考虑交调失真的影响,而在多路射频信号输入条件下又采用小信号近似的方法,精度不高^[10~14]。本文建立了 N 路射频信号输入条件下星间微波光子链路模型,推导出了接收光功率与信号噪声失真比(SNDR)之间的关系式,在光双边带调制和光单边带调制下,分析了链路接收灵敏度与信道数和调制系数之间的关系,指出通过

调整调制系数在一定条件下可以使接收灵敏度达到最高。

2 星间微波光子链路建模

本文建立的强度调制/直接检测(IM/DD)星间微波光子链路模型如图1所示。静止轨道卫星1接收来自地面站的 N 路四相相移键控(QPSK)调制信号,通过马赫-曾德尔调制器(MZM)调制到激光二极管(LD)发出的光载波上,经掺铒光纤放大器(EDFA)放大后,由光学天线发送到空间信道。静止轨道卫星2的光学接收系统接收到光微波信号后,由光电探测器(PD)转换为电信号,经放大器放大后解调为基带数据信号。

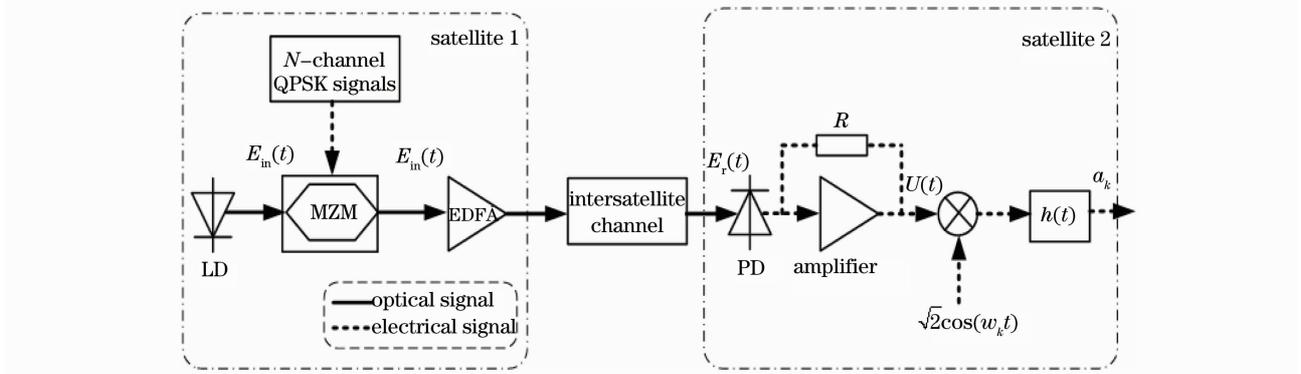


图1 IM/DD星间微波光子链路模型

Fig. 1 Model of IM/DD intersatellite microwave photonics link

由于调制器固有的非线性响应,输出光信号包络为

$$E_{out}(t) = \sqrt{\alpha} E_{in} \left\{ \gamma \exp \left[\frac{j\pi v(t)}{V_{\pi}} \right] + (1 - \gamma) \exp \left[\frac{j\pi v'(t)}{V_{\pi}} + \frac{j\pi V_{DC}}{V_{\pi}} \right] \right\} = \frac{\sqrt{\alpha}}{2} E_{in} \left\{ \prod_{k=1}^N \exp [jm \cos(\omega_k t + \Phi_k + \beta)] + \exp(j\theta) \prod_{k=1}^N \exp [jm \cos(\omega_k t + \Phi_k)] \right\}, \quad (1)$$

式中 E_{in} 为输入光信号幅度, α 为调制器插入损耗, $\gamma = 0.5$ 为分束比, $v(t) = \sum_{k=1}^N V \cos(\omega_k t + \Phi_k + \beta)$ 和 $v'(t) = \sum_{k=1}^N V \cos(\omega_k t + \Phi_k)$ 为相位差 β 的 N 路 QPSK 调制信号, V 为输入射频信号幅度, ω_k 为第 k 路信号角频率, Φ_k 为第 k 路信号包含的相位信息, $m = \pi V / V_{\pi}$ 为调制系数, V_{π} 为调制器半波电压, $\theta = \pi V_{DC} / V_{\pi}$ 为直流偏压引起的相移, V_{DC} 为直流偏置电压。

利用傅里叶级数展开可以得到光电探测器前的输入光信号为

$$E_r(t) = \frac{\sqrt{\alpha L G_{EDFA} G_t G_r}}{2} E_{in} \left\{ \prod_{k=1}^N \left\{ \sum_{p_k=-\infty}^{+\infty} a_{p_k} \exp [jp_k(\omega_k t + \Phi_k + \beta)] \right\} + \exp(j\theta) \prod_{k=1}^N \left\{ \sum_{p_k=-\infty}^{+\infty} a_{p_k} \exp [jp_k(\omega_k t + \Phi_k)] \right\} \right\}, \quad (2)$$

式中 L 为空间传输几何损耗, G_{EDFA} 为放大器增益, G_t 为发射天线增益, G_r 为接收天线增益, $a_{p_k} = j^{p_k} J_{p_k}(m)$, $J_{p_k}(\cdot)$ 表示 k 阶第一类贝塞尔函数。

将(2)式进行傅里叶变换得到

$$E_r(\omega) = \frac{\sqrt{\alpha L G_{\text{EDFA}} G_t G_r}}{2} E_{\text{in}} 2\pi \left\{ \prod_{k=1}^N \left\{ \sum_{p_k=-\infty}^{+\infty} a_{p_k} \exp[j p_k (\Phi_k + \beta)] \right\} \right\} \delta\left(\omega - \sum_{k=1}^N p_k \omega_k\right) + \exp(j\theta) \left\{ \prod_{k=1}^N \left[\sum_{p_k=-\infty}^{+\infty} a_{p_k} \exp(j p_k \Phi_k) \right] \right\} \delta\left(\omega - \sum_{k=1}^N p_k \omega_k\right), \quad (3)$$

考虑到光电探测器的平方律检测特性, 输出电流信号 $I(t) = \eta E_r(t) \overline{E_r(t)}$, $\overline{}$ 表示共轭, η 为光电转换效率, 则输出信号的频谱为

$$I(\omega) = \frac{1}{2\pi} \eta E_r(\omega) * \overline{E_r(\omega)} = \pi \eta P_r \left\{ \prod_{k=1}^N \left\{ \sum_{p_k=-\infty}^{+\infty} \sum_{p'_k=-\infty}^{+\infty} a_{p_k} \overline{a_{p'_k}} \exp[j(p_k - p'_k)(\Phi_k + \beta)] \right\} \right\} \delta\left[\omega - \sum_{k=1}^N (p_k - p'_k) \omega_k\right] + \exp(-j\theta) \left\{ \prod_{k=1}^N \left\{ \sum_{p_k=-\infty}^{+\infty} \sum_{p'_k=-\infty}^{+\infty} a_{p_k} \overline{a_{p'_k}} \exp[j p_k (\Phi_k + \beta) - j p'_k \Phi_k] \right\} \right\} \delta\left[\omega - \sum_{k=1}^N (p_k - p'_k) \omega_k\right] + \exp(j\theta) \left\{ \prod_{k=1}^N \left\{ \sum_{p_k=-\infty}^{+\infty} \sum_{p'_k=-\infty}^{+\infty} a_{p_k} \overline{a_{p'_k}} \exp[j p_k \Phi_k - j p'_k (\Phi_k + \beta)] \right\} \right\} \delta\left[\omega - \sum_{k=1}^N (p_k - p'_k) \omega_k\right] + \left\{ \prod_{k=1}^N \left\{ \sum_{p_k=-\infty}^{+\infty} \sum_{p'_k=-\infty}^{+\infty} a_{p_k} \overline{a_{p'_k}} \exp[j(p_k - p'_k) \Phi_k] \right\} \right\} \delta\left[\omega - \sum_{k=1}^N (p_k - p'_k) \omega_k\right], \quad (4)$$

式中 $*$ 表示卷积, $P_r = \frac{1}{2} \alpha L G_{\text{EDFA}} G_t G_r E_{\text{in}}^2$ 为平均接收光功率。

根据 Graf 加法定理^[15], (4) 式可以化简为

$$I(\omega) = \pi \eta P_r \left\{ \prod_{k=1}^N \{ j^{M_k} \exp[j M_k (\Phi_k + \beta)] J_{-M_k}(0) \} + \exp(-j\theta) \prod_{k=1}^N [\exp(j M_k \Phi_k) j^{M_k} \exp(j M_k \alpha) J_{-M_k}(R)] + \exp(j\theta) \prod_{k=1}^N \{ \exp[j M_k (\Phi_k + \beta)] j^{M_k} \exp(-j M_k \alpha) J_{-M_k}(R) \} + \prod_{k=1}^N [j^{M_k} \exp(j M_k \Phi_k) J_{-M_k}(0)] \right\} \delta\left(\omega - \sum_{k=1}^N M_k \omega_k\right), \quad (5)$$

式中 R 和 α 可由(6)式得出

$$\begin{cases} m[1 - \cos \beta] = R \cos \alpha \\ -m \sin \beta = R \sin \alpha \\ R = \sqrt{2} m \sqrt{1 - \cos \beta} \end{cases}. \quad (6)$$

将(5)式进行逆傅里叶变换可以得出探测器输出电流

$$I(t) = \frac{1}{2} \eta P_r \left\{ \prod_{k=1}^N \{ j^{M_k} \exp[j M_k (\Phi_k + \beta)] J_{-M_k}(0) \} + \exp(-j\theta) \prod_{k=1}^N [\exp(j M_k \Phi_k) j^{M_k} \exp(j M_k \alpha) J_{-M_k}(R)] + \exp(j\theta) \prod_{k=1}^N \{ \exp[j M_k (\Phi_k + \beta)] j^{M_k} \exp(-j M_k \alpha) J_{-M_k}(R) \} + \prod_{k=1}^N [j^{M_k} \exp(j M_k \Phi_k) J_{-M_k}(0)] \right\} \exp\left(j \sum_{k=1}^N M_k \omega_k t\right), \quad (7)$$

则电放大器输出信号为

$$U(t) = I(t) R, \quad (8)$$

式中 R 为放大器偏置电阻。

为了解调出第 k 个信道的基带数据信号, 需将射频信号输入下变频器和低通滤波器。低通滤波器的脉冲响应为

$$h(t) = \begin{cases} 1/T_s, & 0 \leq t < T_s \\ 0, & \text{other} \end{cases} \quad (9)$$

式中 $T_s = 2\pi/\Delta\omega$ 为码元周期。则低通滤波器输出信号为

$$a_k = \frac{1}{T_s} \int_0^{T_s} U(t) \sqrt{2} \cos(\omega_k t) dt, \quad (10)$$

下变频器输出的噪声方差为

$$\sigma_{n,k}^2 = [(G_k + 1)k_0 T/R + 2q\eta P_r + (\eta P_r)^2 N_{RI}]R^2 B, \quad (11)$$

式中 G_k 为第 k 路射频信号增益, $k_0 = 1.38 \times 10^{-23}$ J/K 为波尔兹曼常量, $q = 1.6 \times 10^{-19}$ C 为电子电量, T 为绝对温度, N_{RI} 为激光器相对强度噪声, B 为下变频器单边噪声带宽, 可以表示为

$$B = \frac{1}{2} \int_{-\infty}^{\infty} [h(t) \sqrt{2} \cos(\omega_k t)]^2 dt = 1/2 T_s. \quad (12)$$

假设输入射频信号相位 $\Phi_k = 0$ 时低通滤波器输出信号的均值为

$$A_k = \langle a_k \rangle_{\Phi_k=0}. \quad (13)$$

在信道间隔相同的条件下, 落入信道带宽内的三阶交调分量是主要的干扰分量。低通滤波器输出的三阶交调信号方差为

$$\sigma_{iM,k}^2 = \langle (a_k - A_k)^2 \rangle_{\Phi_k=0}. \quad (14)$$

在 N 路射频信号输入的条件下, 落入中间信道 $k = \lceil N/2 \rceil$ 的三阶交调干扰数目最多^[16]。因此, 以中间信道为例讨论。则第 $k = \lceil N/2 \rceil$ 个信道的信号噪声失真比为

$$R_{\text{SND}_k} = \frac{A_k^2}{\sigma_{n,k}^2 + \sigma_{iM,k}^2} = \begin{cases} \frac{P_{\lceil N/2 \rceil}}{\sigma_{n,\lceil N/2 \rceil}^2 + \left(\frac{N}{2} - 1\right) P_{2\omega_b - \omega_c} + \left(\frac{3}{8} N^2 - \frac{5}{4} N + 1\right) P_{\omega_b + \omega_c - \omega_d}}, & N = 4r \text{ or } 4r + 2, r = 0, 1, 2, \dots \\ \frac{P_{\lceil N/2 \rceil}}{\sigma_{n,\lceil N/2 \rceil}^2 + \left(\frac{N-1}{2}\right) P_{2\omega_b - \omega_c} + \left(\frac{3}{8} N^2 - \frac{5}{4} N + \frac{7}{8}\right) P_{\omega_b + \omega_c - \omega_d}}, & N = 4r + 1, r = 0, 1, 2, \dots \\ \frac{P_{\lceil N/2 \rceil}}{\sigma_{n,\lceil N/2 \rceil}^2 + \left(\frac{N-3}{2}\right) P_{2\omega_b - \omega_c} + \left(\frac{3}{8} N^2 - \frac{5}{4} N + \frac{11}{8}\right) P_{\omega_b + \omega_c - \omega_d}}, & N = 4r + 3, r = 0, 1, 2, \dots \end{cases} \quad (15)$$

式中 $P_{\lceil N/2 \rceil}$ 表示中间信道 $\lceil N/2 \rceil$ 的有用信号功率, $P_{2\omega_b - \omega_c}$ 、 $P_{\omega_b + \omega_c - \omega_d}$ 表示落入中间信道的频率为 $2\omega_b - \omega_c$ 、 $\omega_b + \omega_c - \omega_d$ 的交调信号功率。

2.1 光双边带调制信号噪声失真比

当调制器两臂输入射频信号相位差 $\beta = \pi$, 直流偏置相移 $\theta = \pi/2$ 时, 产生光双边带调制 (ODSB) 信号。根据 (7)、(8)、(10)、(13)、(14) 和 (15) 式, 可以得出光双边带调制中间信道 $\lceil N/2 \rceil$ 的信号噪声失真比为

$$R_{\lceil N/2 \rceil}^{\text{SND}} = \begin{cases} \frac{2\eta^2 P_r^2 R^2 J_1^2(2m) J_0^{2(N-1)}(2m)}{\sigma_{n,\lceil N/2 \rceil}^2 + \left(\frac{N}{2} - 1\right) \eta^2 P_r^2 R^2 J_0^{2(N-2)}(2m) J_1^2(2m) J_2^2(2m) + \left(\frac{3}{8} N^2 - \frac{5}{4} N + 1\right) \eta^2 P_r^2 R^2 J_0^{2(N-3)}(2m) J_1^6(2m)}, & N = 4r \text{ or } r + 2, r = 0, 1, 2, \dots \\ \frac{2\eta^2 P_r^2 R^2 J_1^2(2m) J_0^{2(N-1)}(2m)}{\sigma_{n,\lceil N/2 \rceil}^2 + \left(\frac{N-1}{2}\right) \eta^2 P_r^2 R^2 J_0^{2(N-2)}(2m) J_1^2(2m) J_2^2(2m) + \left(\frac{3}{8} N^2 - \frac{5}{4} N + \frac{7}{8}\right) \eta^2 P_r^2 R^2 J_0^{2(N-3)}(2m) J_1^6(2m)}, & N = 4r + 1, r = 0, 1, 2, \dots \\ \frac{2\eta^2 P_r^2 R^2 J_1^2(2m) J_0^{2(N-1)}(2m)}{\sigma_{n,\lceil N/2 \rceil}^2 + \left(\frac{N-3}{2}\right) \eta^2 P_r^2 R^2 J_0^{2(N-2)}(2m) J_1^2(2m) J_2^2(2m) + \left(\frac{3}{8} N^2 - \frac{5}{4} N + \frac{11}{8}\right) \eta^2 P_r^2 R^2 J_0^{2(N-3)}(2m) J_1^6(2m)}, & N = 4r + 3, r = 0, 1, 2, \dots \end{cases} \quad (16)$$

2.2 光单边带调制信号噪声失真比

当调制器两臂输入射频信号相位差 $\beta = \pi/2$, 直流偏置相移 $\theta = \pi/2$ 时, 产生光单边带调制 (OSSB) 信号。根据 (7)、(8)、(10)、(13)、(14) 和 (15) 式, 可以得出光单边带调制中间信道 $\lceil N/2 \rceil$ 的信号噪声失真比为

$$R_{r \lceil N/2 \rceil}^{\text{SND}} = \left\{ \begin{array}{l} \frac{2\eta^2 P_r^2 R^2 J_1^2(\sqrt{2}m) J_0^{2(N-1)}(\sqrt{2}m)}{\sigma_{n, \lceil N/2 \rceil}^2 + \left(\frac{N}{2} - 1\right) \eta^2 P_r^2 R^2 J_0^{2(N-2)}(\sqrt{2}m) J_1^2(\sqrt{2}m) J_2^2(\sqrt{2}m) + \left(\frac{3}{8}N^2 - \frac{5}{4}N + 1\right) \eta^2 P_r^2 R^2 J_0^{2(N-3)}(\sqrt{2}m) J_1^6(\sqrt{2}m)}, \\ N = 4r \text{ or } r + 2, r = 0, 1, 2, \dots \\ \frac{2\eta^2 P_r^2 R^2 J_1^2(\sqrt{2}m) J_0^{2(N-1)}(\sqrt{2}m)}{\sigma_{n, \lceil N/2 \rceil}^2 + \left(\frac{N-1}{2}\right) \eta^2 P_r^2 R^2 J_0^{2(N-2)}(\sqrt{2}m) J_1^2(\sqrt{2}m) J_2^2(\sqrt{2}m) + \left(\frac{3}{8}N^2 - \frac{5}{4}N + \frac{7}{8}\right) \eta^2 P_r^2 R^2 J_0^{2(N-3)}(\sqrt{2}m) J_1^6(\sqrt{2}m)}, \\ N = 4r + 1, r = 0, 1, 2, \dots \\ \frac{2\eta^2 P_r^2 R^2 J_1^2(\sqrt{2}m) J_0^{2(N-1)}(\sqrt{2}m)}{\sigma_{n, \lceil N/2 \rceil}^2 + \left(\frac{N-3}{2}\right) \eta^2 P_r^2 R^2 J_0^{2(N-2)}(\sqrt{2}m) J_1^2(\sqrt{2}m) J_2^2(\sqrt{2}m) + \left(\frac{3}{8}N^2 - \frac{5}{4}N + \frac{11}{8}\right) \eta^2 P_r^2 R^2 J_0^{2(N-3)}(\sqrt{2}m) J_1^6(\sqrt{2}m)}. \\ N = 4r + 3, r = 0, 1, 2, \dots \end{array} \right. \quad (17)$$

3 数值仿真结果

光通信系统的接收灵敏度通常定义为使系统误码率达到某一量级时所需要的最小接收光功率,该最小接收光功率的值越小则接收灵敏度越高,通信链路的性能也就越好^[5]。QPSK 调制误比特率达到 10^{-7} 对应的信噪比要求为 17.3 dB,根据表 1 参数可以计算出 N 路射频信号输入条件下光双边带调制和光单边带调制接收灵敏度与信道数和调制系数之间的关系,如图 2 和 3 所示。

表 1 接收灵敏度计算参数

Table 1 Calculation parameters of receiver sensitivity

Parameter	Symbol	Value	Unit
Photodiode responsivity	η	0.6	A/W
Channel spacing	$\Delta\omega/2\pi$	40	MHz
Noise bandwidth	B	20	MHz
Temperature	T	500	K
Relative intensity noise	N_{RI}	-150	dB/Hz
Transimpedance	R	50	Ω
Modulator switching voltage	V_π	3.35	V
Signal-to-noise ratio	R_{SN}	17.3	dB

从图 2 可以看出,光双边带调制接收灵敏度随着调制系数的增大先增大后减小,当调制系数 $m < 0.05$ 时,接收灵敏度对信道数的变化不敏感,因为此时三阶交调分量很小,随着调制系数的增大,三阶交调分量增大,接收灵敏度随信道数的增加下降很快,当 $m = 0.25$ 时, $N = 16$ 比 $N = 8$ 时接收灵敏度下降约 2.5 dB, $N = 32$ 比 $N = 16$ 时接收灵敏度下降约 7.3 dB。从图 3 可以看出,光单边带调制接收灵敏度随着调制系数的增大先增大后减小,当调制系

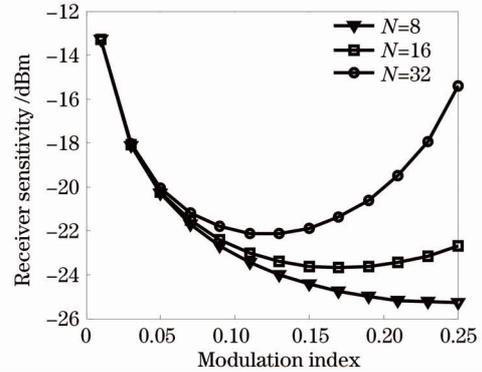


图 2 光双边带调制接收灵敏度与调制系数和信道数的关系

Fig. 2 Receiver sensitivity as a function of the modulation index under ODSB modulation according to three sets of user channels

数 $m < 0.07$ 时,接收灵敏度对信道数的变化不敏感,随着调制系数的增大,三阶交调分量增大,接收灵敏度随信道数的增加下降很快,当 $m = 0.25$ 时, $N = 16$ 比 $N = 8$ 时接收灵敏度下降约 1.25 dB, $N = 32$ 比 $N = 16$ 时接收灵敏度下降约 3 dB。

图 4 表示最高接收灵敏度与信道数之间的关系。从图 4 中可以看出,信道数由 1 增加到 10,光双边带调制最高接收灵敏度由 -31.08 dBm 减小到 -24.75 dBm,光单边带调制最高接收灵敏度由 -29.57 dBm 减小到 -23.17 dBm,这主要是由落入中间信道的三阶交调分量个数增加引起的。相同信道数条件下,光双边带调制最高接收灵敏度优于光单边带调制,这与光单边带调制造成 QPSK 调制信号功率衰减一半有关。图 5 表示最佳调制系数与信

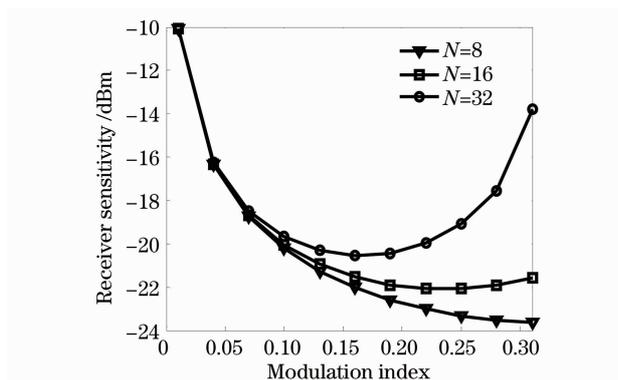


图 3 光单边带调制接收灵敏度与调制系数和信道数的关系

Fig. 3 Receiver sensitivity as a function of the modulation index under OSSB modulation according to three sets of user channels

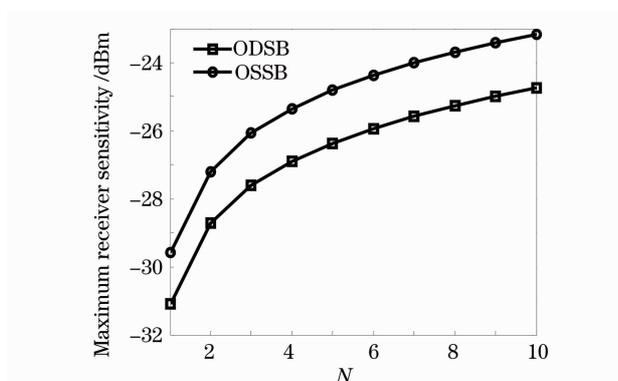


图 4 最高接收灵敏度与信道数的关系

Fig. 4 Optimum receiver sensitivity versus number of channels N under ODSB and OSSB modulation

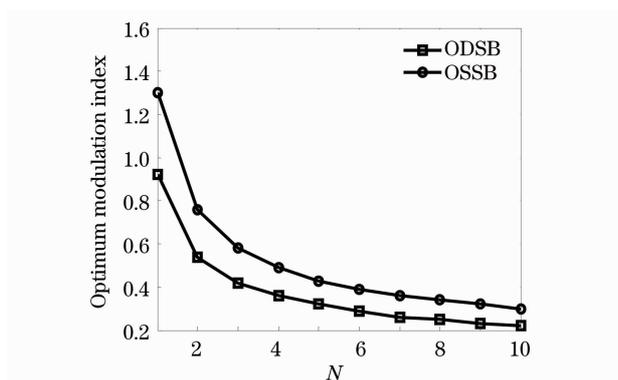


图 5 最佳调制系数与信道数的关系

Fig. 5 Optimum modulation index versus number of channels N under ODSB and OSSB modulation

道数的关系。从图中可以看出,信道数由 1 增加到 10,光双边带调制最佳调制系数由 0.92 减小到 0.22,光单边带调制最佳调制系数由 1.3 减小到 0.3,高于光双边带调制。

4 结 论

本文建立了交调失真影响下星间微波光子链路模型,推导出了接收光功率与信号噪声失真比之间的关系式,以三阶交调影响最严重的中间信道 $R_{\text{SND}} = 17.3 \text{ dB}$ ($f_{\text{BER}} = 10^{-7}$) 为例,分析了光双边带和光单边带调制接收灵敏度与信道数和调制系数之间的关系,结果表明,光双边带调制接收灵敏度随着调制系数的增大先增大后减小,当调制系数 $m < 0.05$ 时,接收灵敏度对信道数的变化不敏感,因为此时三阶交调分量很小,随着调制系数的增大,三阶交调分量增大,接收灵敏度随信道数的增加下降很快,当 $m = 0.25$ 时, $N = 16$ 比 $N = 8$ 时接收灵敏度下降约 2.5 dB, $N = 32$ 比 $N = 16$ 时接收灵敏度下降约 7.3 dB. 光单边带调制接收灵敏度随着调制系数的增大先增大后减小,当调制系数 $m < 0.07$ 时,接收灵敏度对信道数的变化不敏感,随着调制系数的增大,三阶交调分量增大,接收灵敏度随信道数的增加下降很快,当 $m = 0.25$ 时, $N = 16$ 比 $N = 8$ 时接收灵敏度下降约 1.25 dB, $N = 32$ 比 $N = 16$ 时接收灵敏度下降约 3 dB. 相同信道数条件下,光双边带调制最高接收灵敏度优于光单边带调制,这与光单边带调制造成 QPSK 调制信号功率衰减一半有关。通过调整调制系数可以优化星间微波光子链路接收灵敏度,然而直接探测的方式限制了接收灵敏度的进一步提高,下一步可以考虑采用相干接收体制进行研究。

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