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Analysis and optimization of power spectrum on EBPSK modulation in throughput-efficient wireless system

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Abstract: In order to satisfy increasingly greater demand for the performance of communication systems, a throughput efficient wireless system based on the extended binary phase shift keying (EBPSK) modulation is presented. Simultaneously, corresponding analysis of power spectra is also given with a brief process. The optimal waveform is proposed without useful information loss, by removing linear spectra presenting periodic components. On this basis, the reasonable definition of bandwidth is discussed, which indicates that the EBPSK belongs to the category of the ultra narrow band (UNB) throughput-efficient communication. Meanwhile, the modulation parameters' effects on bandwidth, transmission rate and transmission performance are analyzed. Results illustrate the validity of theoretical analysis and spectrum optimization. Results also prove that this UNB system can obtain good bit error rate (BER) performance with high spectra efficiency.

Key words: extended binary phase shift keying; power spectrum; ultra narrow band; spectra efficiency

With the development of the information society, there is increasingly greater demand for the performance of communication systems. In order to increase the data rate of transmission, modern communication is developing in the ultra wide band (UWB) direction, which occupies more and more frequency resources, but the bandwidth efficiency is still not very high, lower than 4 bit/(s·Hz^{-1}) in mobile communication. Therefore, in the case of fixed bandwidths, how to transmit the information faster and how to improve the frequency bandwidth efficiency become very important. Recently, high efficiency transmission technologies, especially ultra narrow band (UNB) communication with outstanding transmission ability, are being paid more attention to by many researchers.

Walker started researches in the development and applications of high speed data transmission in the 1980s. Recently, he developed the modulation technology based mainly on very minimum-shift keying (VMSK) derived from early variable phase shifting keying (VPSK)[1] and improved VPSK. After that, several versions of VMSK[2] were gradually developed, such as PPM, PRK, MSB[3-4], etc. After cooperation with Photron Science Company, these kinds of patents were registered as ultra spectral modulation (USM), which has pretty high bandwidth efficiency. It is said that USM has been applied in many fields. In China, Wu et al. [5] started to focus on the research of UNB in 1999, and his research team proposed like-sine VMSK modulation in 2001 and very minimum waveform difference keying (VWDK)[6] modulation in 2003, respectively. Recently, they have just established another new high efficient modulation method, called extended binary phase shift keying (EBPSK)[7], which has good flexibility, universality and good possibilities for new high speed transmission and anti-jamming methods.

While, nowadays, schemes try to move as much power as possible into the sidebands (where the information resides) and away from the carrier signal, EBPSK and other UNB schemes do the opposite, placing most of the power in the carrier to keep sideband energy emissions negligible, which gives it the appearance of a UWB system. However, the power spectrum of a UNB system is so low in the adjacent bands that the legacy user of that spectrum would experience minimal or insignificant interference[8]. Actually, it is due to the UNB system's carrier-preserving and sideband-depressing characteristics that endows direct carrier-preserving modulation (DCPM) systems with the potential of frequency-hops to vacant spectrums, where is judged by cognitive technologies as more suitable for communications. These characteristics will allow for spectrum reuse of EBPSK systems in future generations of wireless communications[8].

1 System Description

In Ref. [7], Wu et al. introduced a unitive expression for a binary PM called an extended BPSK (EBPSK). It takes some current UNB modulations as its special examples, e. g., the missing cycle modulation (MCM)[9-10], the pulse position phase reversal keying (3PRK)[9], the suppressed cycle modulation (SCM)[11], and the traditional BPSK. Let the waveform modulated by bit "0" and bit "1" be $g_0(t)$ and $g_1(t)$, respectively; then a unitive expression of the arbitrary BPSK modulated signals is defined as below[7]:

$$g_0(t) = A\sin\omega_c t \qquad 0 \leqslant t < T$$

$$g_1(t) = \begin{cases} B\sin(\omega_c t + \theta) & 0 \leqslant t < \tau; 0 \leqslant \theta \leqslant \pi \\ A\sin\omega_c t & \tau \leqslant t < T \end{cases} \tag{1}$$

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where $T=2\pi N/\omega_c=NT_c$ is the symbol width of data, namely the temporal length of a code; θ is the jumping angle or modulating angle, $T_c=2\pi/\omega_c=1/f_c$ is the carrier cycle, namely T lasts $N\geqslant 1$ cycles of the carrier; and the temporal length $\tau=2\pi K/\omega_c=KT_c$ of a hopping waveform lasts $K\leqslant N$ cycles of the carrier.

2 Analysis and Optimization of Power Spectrum

2.1 Analysis of power spectrum

The theoretical power spectrum expression of EBPSK is given with a brief process description in the appendix. At $f=f_c$, $S_{xx}(f)$ can be approximated by

$$\frac{1}{4T}\frac{\tau^2}{4}(A^2+B^2-2AB\cos\theta)+\frac{1}{4T^2}\frac{1}{4}[\tau^2(A^2+B^2)+4T^2A^2-4T\tau A^2+4ABT\tau\cos\theta-2AB\tau^2\cos\theta] \tag{2}$$

On the other hand, at $f\neq f_c$, the power spectrum $S_{xx}(f)$ can be written as

$$\begin{aligned}S_{xx}(f)&=\frac{1}{4T}\left|G_1(f)-G_0(f)\right|^2+\frac{1}{4T^2}\sum_{m=-\infty}^{\infty}\delta\left(f-\frac{m}{T}\right)\left|G_1\left(\frac{m}{T}\right)+G_0\left(\frac{m}{T}\right)\right|^2=\frac{1}{16T\pi^2(f_c^2-f^2)^2}\cdot\\&[(2A^2f_c^2+2B^2f_c^2\cos^2\theta-4ABf_c^2\cos\theta+2B^2f^2\sin^2\theta)(1-\cos2\pi f\tau)]+\frac{1}{16T^2\pi^2}\sum_{\substack{m=-\infty\\ m\neq\pm N}}^{\infty}\delta\left(f-\frac{m}{T}\right)\frac{1}{\left(f_c^2-\left(\frac{m}{T}\right)^2\right)^2}\cdot\\&\left[\left(2A^2f_c^2+2B^2f_c^2\cos^2\theta-4ABf_c^2\cos\theta+2B^2\left(\frac{m}{T}\right)^2\sin^2\theta\right)\left(1-\cos2\pi\frac{m}{T}\tau\right)\right]\end{aligned} \tag{3}$$

Obviously, the power spectra are made up of a continuous spectrum and some linear spectra, where linear spectra present periodic components of the modulated EBPSK signal, and the continuous spectrum indicates random properties of modulated information. In order to reduce bandwidth, we can optimize this waveform, by removing all high order harmonics components except the main linear spectrum.

2.2 Optimization of power spectrum

In the optimization of the power spectrum, the removal of all high order harmonics components except the main linear spectrum is the key problem. Let $p_0(t)$ and $p_1(t)$ be new modulation waveforms corresponding to 0 and 1, respectively, then the power spectrum (3) can be revised as

$$S_{xx}(f)=\frac{1}{4T}\left|P_1(f)-P_0(f)\right|^2+\frac{1}{4T^2}\sum_{m=-\infty}^{\infty}\delta\left(f-\frac{m}{T}\right)\left|P_1\left(\frac{m}{T}\right)+P_0\left(\frac{m}{T}\right)\right|^2 \tag{4}$$

Let

$$p_n(t)=g_n(t)-x(t)\qquad n=0,1;0<t<T \tag{5}$$

First, we assume that $X(f)$ represents the Fourier transform of $x(t)$; that is to say, $X(f)=\int_{-\infty}^{\infty}x(t)\mathrm{e}^{-\mathrm{j}2\pi ft}\mathrm{d}t$, and then proceed to remove discrete harmonics components. While $m\neq\pm N$, $P_1\left(\frac{m}{T}\right)+P_0\left(\frac{m}{T}\right)=0$; i. e., $G_1\left(\frac{m}{T}\right)+G_0\left(\frac{m}{T}\right)-2X\left(\frac{m}{T}\right)=0$, so

$$X\left(\frac{m}{T}\right)=\frac{1}{2}\left[G_1\left(\frac{m}{T}\right)+G_0\left(\frac{m}{T}\right)\right]\qquad m\neq\pm N \tag{6}$$

Then

$$\begin{aligned}x(t)&=\frac{1}{T}\sum_{m\neq\pm N}X\left(\frac{m}{T}\right)\mathrm{e}^{\mathrm{j}2\pi mt/T}=\frac{1}{2T}\sum_{m\neq\pm N}\left[G_1\left(\frac{m}{T}\right)+G_0\left(\frac{m}{T}\right)\right]\mathrm{e}^{\mathrm{j}2\pi mt/T}=\frac{1}{2T}\sum_{\substack{m=-\infty\\ m\neq\pm N}}^{\infty}\left\{\frac{1}{2\pi}\frac{1}{f_c^2-(m/T)^2}\cdot\right.\\&\left.\left[(1-\mathrm{e}^{-\mathrm{j}2\pi\frac{m}{T}\tau})\left(Bf_c\cos\theta+\mathrm{j}B\frac{m}{T}\sin\theta-Af_c\right)\right]\right\}\mathrm{e}^{\mathrm{j}2\pi mt/T}\qquad 0<t<T\end{aligned} \tag{7}$$

Simultaneously, we can easily obtain $\left|P_1(f)-P_0(f)\right|^2=\left|G_1(f)-G_0(f)\right|^2$, which illustrates that this optimum process remains the power of a continuous spectrum. Based on this consideration, using $p_n(t)=g_n(t)-x(t)$, $n=0,1$, to modulate information bits "0" and "1", respectively, the high order harmonics components will not exist. In this case, a compact bandwidth is in accordance with Federal Communications Commission (FCC) standards, with no influence on demodulation performance, where the FCC standard bandwidth is defined as the 99% in-band power, i. e. the −20 dB crossing on a fractional out-of-band power chart, and it is even much stricter criterion of −60 dB regulated by the FCC.

3 Performance Simulation

In this section, computer simulation results are given, which illustrates the correctness of the power spectrum expression. Then the validity of optimization will be simulated to prove the UNB characteristic of EBPSK modulation. Finally, the system BER curves indicate that the optimization has no influence on the performance, i. e. , without useful information loss.

It should be noted that the infinite impulse function $\delta(f)$ is included in Eq. (3). In order to illustrate the power spectrum, we define narrow rectangular pulse to approximate it in simulations according to the definition of $\delta(f)$. At the same time, in the real spectra estimation, we will see the finite discrete impulses due to the effects of windows, the number of FFT and the number of samples. With the increase in these parameters, the height of the impulse at carrier frequency will increase when compared with the top of the continuous spectrum. All in all, it seems that the theoretical and real simulation results have a little difference, but the shapes of the spectra are identical.

First, we consider EBPSK power spectra with different modulation parameters, and provide results of theoretical calculations and Welch spectra estimations based on the Hamming window, respectively, as shown in Fig. 1 and Fig. 2. In our simulation, the number of samples is chosen as 10^4, carrier frequency $f_c = 40$ kHz, $N = 32$, $K = 2$, $A = B = 1$, θ is chosen as 0, $\pi/6$, $\pi/2$ and π, respectively.

From Fig. 1 and Fig. 2, with the decrease in modulation angle θ, the power spectrum and energy will be increasingly cen-

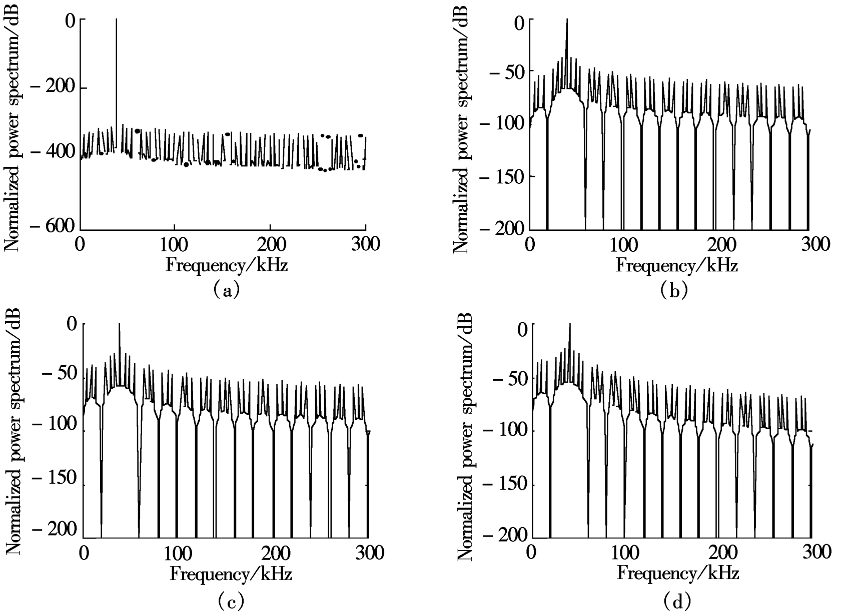

Fig. 1 Theoretical spectra according to Eq. (3). (a) $\theta = 0$; (b) $\theta = \pi/6$; (c) $\theta = \pi/2$; (d) $\theta = \pi$

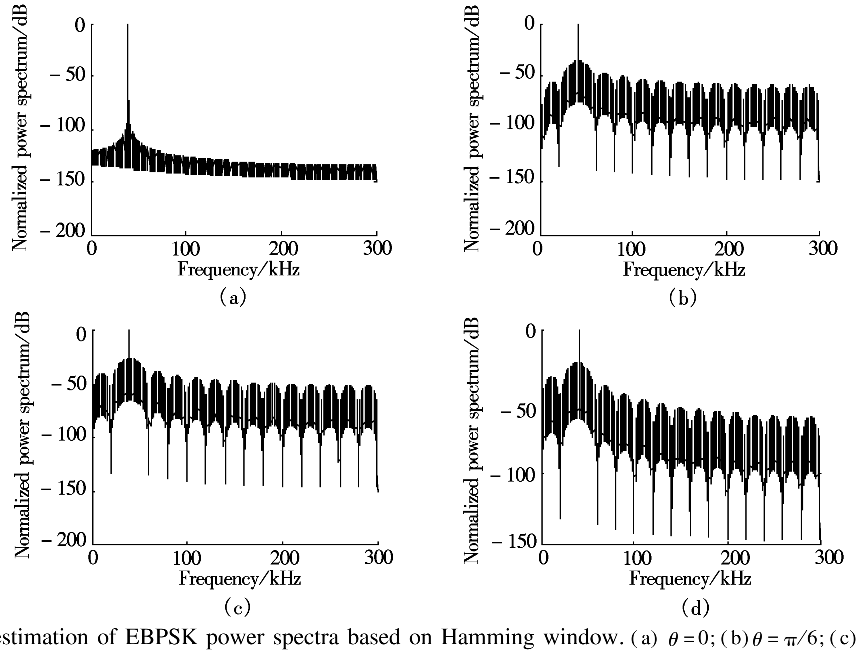

Fig. 2 Welch estimation of EBPSK power spectra based on Hamming window. (a) $\theta = 0$; (b) $\theta = \pi/6$; (c) $\theta = \pi/2$; (d) $\theta = \pi$

tralizedin EBPSK modulation; i. e., the bandwidth follows decreasing. When $\theta \to 0$, i. e., the modulation waveform approaches a pure sine-wave, the power spectrum is just an impulse at carrier frequency, and its bandwidth approaches 0, which agrees with the real case. On the other hand, due to different approximations of infinite impulse function $\delta(f)$, the height of the impulse at carrier frequency will be changed when compared with the top of continuous spectrum. But to be exact, the height of this impulse should be infinite in theory. So according to the FCC bandwidth definition below -60 dB, our system belongs to UNB category absolutely.

In Fig. 3, the power spectrum calculated with waveform optimization according to Eqs. (5), (6) and (7) is shown, where the summation from minus infinity to plus infinity in Eq. (7) can be approximated with a very large number, e. g., 10^5. Theoretically, this optimization can remove all the high order harmonics components clearly, but, in fact, for convenience, we just choose the real part of $x(t)$ and ignore its imaginary part. Therefore, in the simulation results, its high order harmonics components are not removed completely, but its amplitudes are decreased greatly, especially the harmonics around the main frequency are very small, which causes the bandwidth to decrease greatly. According to the FCC bandwidth definition below -60 dB, the even higher harmonics have no influence on the bandwidth, which we can ignore. When $\theta = \pi$, the imaginary part of $x(t)$ is just 0, so ignoring its imaginary part has no influence for itself. In this case, all high order harmonics are removed completely from the simulation results. The simulation results also validate the correctness of waveform optimization. Compared with the filter, this optimization compresses bandwidth without a useful information loss, and the phase leap is well remained.

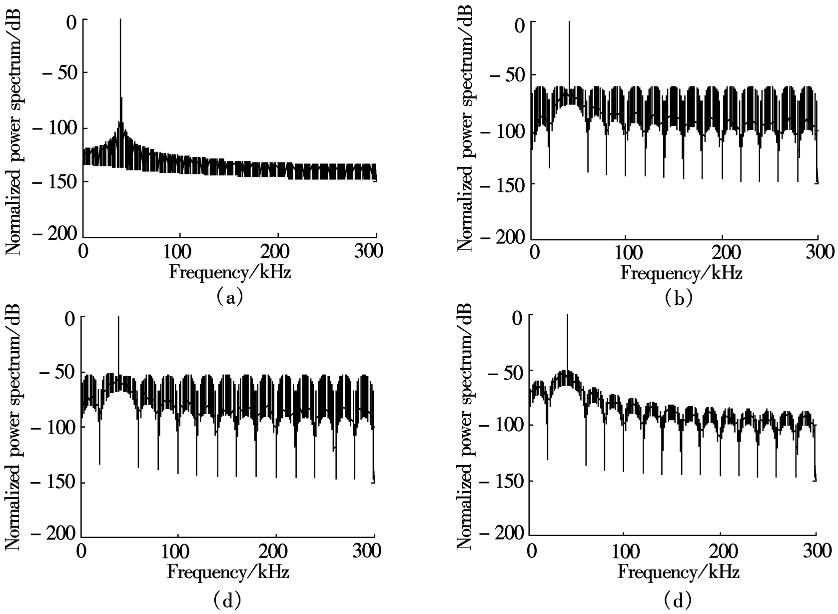

Fig. 3 EBPSK power spectra with waveform optimization. (a) $\theta = 0$; (b) $\theta = \pi/6$; (c) $\theta = \pi/2$; (d) $\theta = \pi$

Then the BER performance is simulated. For our EBPSK system, a phase lock loop (PLL) is used as its demodulator and a unique nonlinear BPF, called the geometric feature filter (GFF) and implemented with artificial neural networks (ANN), as its UNB filter, where a detailed discussion is presented in Ref. [12]. System parameters in the simulation are selected as $f_c = 100$ kHz; $N = 5$; $K = 1$; $A = B = 1$; and θ is chosen as $\pi/2$. The BER performances before and after optimization are given in Fig. 4. Obviously, both the two curves indicate good BER performance almost without difference. All in all, the optimization of power spectra is effective, which has no influence on BER performance. As a preliminary result, its spectra efficiency exceeds 15 bit/(s·Hz^{-1}) at BER $= 10^{-5}$ and $E_b/N_0 \approx 25$ dB[12].

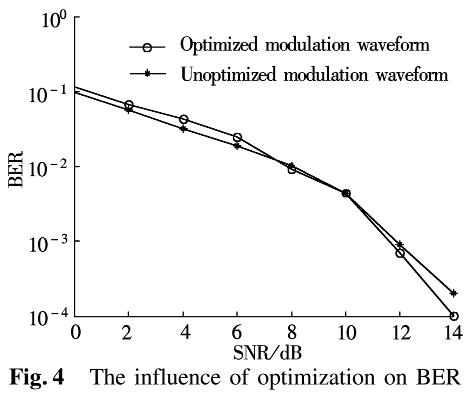

Fig. 4 The influence of optimization on BER

4 Conclusion

In this paper, aiming at the UNB characteristics of EBPSK, we deduce the power spectrum expression and present the optimal waveform without useful information loss. Then some influences of modulation parameters on bandwidth, transmission rate and performance are discussed. Finally, the simulation results prove the correctness and validity of the theoretical analysis.

In the EBPSK modulation, with the decrease in the modulation angle θ, the power spectra and energy are increasingly centralized, that is to say, with much narrower bandwidths and higher efficiency. But in this case, the corresponding demodulation performance decreases, so there should be a good tradeoff between parameters. When K/N becomes smaller and smaller, i. e., fewer periods of phase changing in modulation waveform of logic "1", its spectra with higher main frequency component are much closer to that of the sine-wave. Consequently, it is easier to meet the bandwidth standard of UNB.

According to the traditional FCC standards, the bandwidth is defined as a bandwidth cut off below -60 dB. Therefore, we choose proper parameters and consider the modulation waveform with optimization. Undoubtedly, it belongs to the UNB. Because it has a compact bandwidth defined by FCC standards, having no effects on other communication services, a reliable and high speed communication is achieved without very high power.

Appendix The deduction of the theoretical power spectrum expression of EBPSK.

The transmitted symbols are assumed to be independent and are taken from a finite alphabet set $A = \{0, 1\}$. Hence, the EBPSK signals can be defined as

$$x(t) = \sum_{n=-\infty}^{+\infty} (s_n g_1(t - nT) + (1 - s_n) g_0(t - nT)) \tag{A1}$$

It is easy to see that $x(t)$ is a wide-sense cyclostationary signal with a period T, so the time average of the auto-correlation function can be described as

$$\begin{aligned}\bar{R}_{xx}(\tau) \triangleq \lim_{c\to\infty}\frac{1}{2c}\int_{-c}^{c} R_{xx}(t+\tau, t)\,\mathrm{d}t = \frac{1}{T}\int_{-T/2}^{T/2} R_{xx}(t+\tau, t)\,\mathrm{d}t = \frac{1}{T}\sum_m R_{ss}(m)\int_{-\infty}^{\infty} g_1(t+\tau-mT)g_1(t)\,\mathrm{d}t + \\ \frac{1}{T}\sum_m [m_s - R_{ss}(m)]\int_{-\infty}^{\infty} g_1(t+\tau-mT)g_0(t)\,\mathrm{d}t + \frac{1}{T}\sum_m [m_s - R_{ss}(m)]\int_{-\infty}^{\infty} g_0(t+\tau-mT)g_1(t)\,\mathrm{d}t + \\ \frac{1}{T}\sum_m [1 - 2m_s + R_{ss}(m)]\int_{-\infty}^{\infty} g_0(t+\tau-mT)g_0(t)\,\mathrm{d}t\end{aligned} \tag{A2}$$

Let $G_1(f) = \int_{-\infty}^{\infty} g_1(t)\mathrm{e}^{-\mathrm{j}2\pi ft}\mathrm{d}t$, $G_0(f) = \int_{-\infty}^{\infty} g_0(t)\mathrm{e}^{-\mathrm{j}2\pi ft}\mathrm{d}t$, so

$$\int_{-\infty}^{\infty}\mathrm{e}^{-\mathrm{j}2\pi f\tau}\mathrm{d}\tau\int_{-\infty}^{\infty} g_1(t+\tau-mT)g_0(t)\,\mathrm{d}t = \mathrm{e}^{-\mathrm{j}2\pi fmT}G_1(f)G_0^*(f) \tag{A3}$$

According to the Wiener-Khinchin theorem[13-14], the power spectrum density of the wide-sense cyclostationary stochastic process is the Fourier transform of the corresponding auto-correlation function.

$$\begin{aligned}S_{xx}(f) = \int_{-\infty}^{\infty}\bar{R}_{xx}(\tau)\mathrm{e}^{-\mathrm{j}2\pi f\tau}\mathrm{d}\tau = \frac{1}{T}S_{ss}(f)[\,|G_1(f)|^2 + |G_0(f)|^2 - G_1(f)G_0^*(f) - G_1^*(f)G_0(f)] + \\ \frac{1}{T}\sum_m \mathrm{e}^{-\mathrm{j}2\pi fmT}[m_s G_1(f)G_0^*(f) + m_s G_1^*(f)G_0(f) + (1-2m_s)\,|G_0(f)|^2]\end{aligned} \tag{A4}$$

Assume that the binary information sequence, with equal probability, is independent identical distribution (i. i. d.), that is to say, $m_s = 1/2$, $R_{ss}(m) = \begin{cases}1/2, & m=0\\ 1/4, & m\neq 0\end{cases}$. So

$$S_{ss}(f) = \sum_{m=-\infty}^{\infty} R_{ss}(m)\mathrm{e}^{-\mathrm{j}2\pi fmT} = \frac{1}{4} + \frac{1}{4}\sum_{m=-\infty}^{\infty}\mathrm{e}^{-\mathrm{j}2\pi fmT} = \frac{1}{4} + \frac{1}{4T}\sum_{m=-\infty}^{\infty}\delta\left(f - \frac{m}{T}\right) \tag{A5}$$

where the Poisson formula is used, $\sum_{m=-\infty}^{\infty}\mathrm{e}^{-\mathrm{j}2\pi fmT} = \frac{1}{T}\sum_{n=-\infty}^{\infty}\delta\left(f - \frac{n}{T}\right)$.

Then its power spectrum density can be simplified as

$$S_{xx}(f) = \frac{1}{4T}|G_1(f) - G_0(f)|^2 + \frac{1}{4T^2}\sum_{m=-\infty}^{\infty}\delta\left(f - \frac{m}{T}\right)\left|G_1\left(\frac{m}{T}\right) + G_0\left(\frac{m}{T}\right)\right|^2 \tag{A6}$$

The Fourier transforms of $g_0(t)$ and $g_1(t)$ can be calculated as

$$G_0(f) = \int_0^T A\sin(2\pi f_c t)\mathrm{e}^{-\mathrm{j}2\pi ft}\mathrm{d}t = \frac{A}{2\pi}\frac{f_c}{f_c^2 - f^2}(1 - \mathrm{e}^{-\mathrm{j}2\pi fT}) \tag{A7}$$

$$G_1(f) = \frac{1}{2\pi}\frac{1}{f_c^2 - f^2}[B(1 - \mathrm{e}^{-\mathrm{j}2\pi f\tau})(f_c\cos\theta + \mathrm{j}f\sin\theta) + A(\mathrm{e}^{-\mathrm{j}2\pi f\tau} - \mathrm{e}^{-\mathrm{j}2\pi fT})f_c] \tag{A8}$$

So

$$\begin{aligned}|G_1(f) - G_0(f)|^2 = [\mathrm{Re}(G_1(f) - G_0(f))]^2 + [\mathrm{Im}(G_1(f) - G_0(f))]^2 = \frac{1}{4\pi^2(f_c^2 - f^2)^2}\cdot \\ [(2A^2f_c^2 + 2B^2f_c^2\cos^2\theta - 4ABf_c^2\cos\theta + 2B^2f^2\sin^2\theta)(1 - \cos2\pi f\tau)]\end{aligned} \tag{A9}$$

$$G_1(f) + G_0(f) = \frac{1}{2\pi}\frac{1}{f_c^2 - f^2}[B(1 - \mathrm{e}^{-\mathrm{j}2\pi f\tau})(f_c\cos\theta + \mathrm{j}f\sin\theta) + A(1 + \mathrm{e}^{-\mathrm{j}2\pi f\tau} - 2\mathrm{e}^{-\mathrm{j}2\pi fT})f_c] \tag{A10}$$

Obviously, Eq. (A10) has no definition at $f = f_c$, so only while $m \neq N$, i. e. $m/T \neq f_c$, the following equation holds.

$$G_1\left(\frac{m}{T}\right)+G_0\left(\frac{m}{T}\right)=\frac{1}{2\pi}\frac{1}{f_c^2-(m/T)^2}\left[B(1-e^{-j2\pi\frac{m}{T}\tau})\left(Bf_c\cos\theta+jB\frac{m}{T}\sin\theta-Af_c\right)\right]=G_1(f)-G_0(f)\mid_{f=\frac{m}{T}} \tag{A11}$$

According to Eq. (A6), the power spectrum is made up of two parts, i. e., a continuous spectrum and some linear spectra. While $f\to f_c$, both the first and the second parts of the spectrum are types of 0/0. Separately, we can obtain its limit at f_c by the L'Hospital rule.

$$\lim_{f\to f_c}\mid G_1(f)-G_0(f)\mid^2=\frac{\tau^2}{4}(A^2+B^2-2AB\cos\theta) \tag{A12}$$

$$\lim_{f\to f_c}\mid G_1(f)+G_0(f)\mid^2=\frac{1}{4}[\tau^2(A^2+B^2)+4T^2A^2-4T\tau A^2+4ABT\tau\cos\theta-2AB\tau^2\cos\theta] \tag{A13}$$

So, at $f=f_c$, $S_{xx}(f)$ can be approximated by

$$\frac{1}{4T}\frac{\tau^2}{4}(A^2+B^2-2AB\cos\theta)+\frac{1}{4T^2}\frac{1}{4}[\tau^2(A^2+B^2)+4T^2A^2-4T\tau A^2+4ABT\tau\cos\theta-2AB\tau^2\cos\theta]$$

Substituting the above into Eq. (A6), when $f\neq f_c$, the power spectrum $S_{xx}(f)$ can be written as

$$S_{xx}(f)=\frac{1}{16T\pi^2(f_c^2-f^2)^2}[(2A^2f_c^2+2B^2f_c^2\cos^2\theta-4ABf_c^2\cos\theta+2B^2f^2\sin^2\theta)(1-\cos2\pi f\tau)]+\frac{1}{16T^2\pi^2}\cdot$$
$$\sum_{\substack{m=-\infty\\ m\neq\pm N}}^{\infty}\delta\left(f-\frac{m}{T}\right)\frac{1}{\left(f_c^2-\left(\frac{m}{T}\right)^2\right)^2}\left[\left(2A^2f_c^2+2B^2f_c^2\cos^2\theta-4ABf_c^2\cos\theta+2B^2\left(\frac{m}{T}\right)^2\sin^2\theta\right)\left(1-\cos2\pi\frac{m}{T}\tau\right)\right] \tag{A14}$$

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高效无线系统中 EBPSK 调制的功率谱分析及优化

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摘要:为了满足人们对通信系统传输率等性能越来越高的需求,提出了基于扩展的二元相移键控(EBPSK)调制的高效无线系统,详细推导了 EBPSK 功率谱解析表达式. 在不损失有用信息的前提下,通过直接在表达式中去除表示周期分量的线谱,导出新的频谱优化方法. 在此基础上讨论了合理的带宽定义方式,表明了 EBPSK 属于超窄带高效通信的范畴. 分析了 EBPSK 调制参数对带宽、传输速率及传输性能的影响. 实验结果表明了理论谱表达式及优化的正确性,也证明这种超窄带系统在得到较高的频谱效率的同时可以达到很好的误码率性能.

关键词:扩展的二元相移键控;功率谱;超窄带;频谱效率

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