Research Article

Design of Special Impacting Filter for Multicarrier ABPSK System

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A rather intuitive technique known as pole-zero placement is introduced to illustrate the frequency response of the special impacting filters (SIFs) with a pair of conjugate zero-poles and deduce the equation of the pole radii. Based on that, the paper proposes an iterative scheme to derive the parameters of the cascade notch filter. The cost function is determined by the cascading notch filter's influence on impacting filters, converting the cost function's least square problem to a filter parameters' standard quadratic programming problem. Finally, a cascading notch SIF (CNSIF) designed to demodulate the ABPSK signals is realized.

1. Introduction

As we know, spectrum is an unrenewable resource, and the new wireless communication systems are occupying more and more transmission bandwidth. So many corporations and research institutions try to find the optimal method to utilize the spectrum. Against this background, the ultra-narrow band (UNB) technology was proposed by doctor Walker. Up until now, various modulation methods have been successively proposed with the progress of UNB development, for example, variable phase shifting keying (VPSK), enhanced VPSK, very minimum shifting keying (VMSK), pulse position phase reversal keying (3PRK), missing cycle modulation (MCM), suppressed cycle modulation (SCM), and minimum sideband modulation (MSB), and so forth. The key of the demodulation is the “zero group delay” filter. The “zero group delay” means the rise time of the filter is very short and the phase changes are preserved after the “zero group delay” filter. The core unit is a quartz crystal which is too unreliable, instable, and inflexible to be mass-produced [1–3]. Wu et al. analyzed the feasibility of the UNB and proposed a high-efficiency modulation called extended BPSK (EBPSK) [4, 5]. In 2009, they proposed continuous phase EBPSK (CP-EBPSK) based on the EBPSK [6], further tightening and reducing the EBPSK power spectrum

sideband. Subsequently, they imported the pseudorandom sequence and modified spectrum parameter to improve the CP-EBPSK, that is, modified CP-EBPSK (MCP-EBPSK). All of the above techniques have the same characteristic, that is “asymmetric.” They are uniformly referred to as asymmetric binary phase shift keying (ABPSK). ABPSK could increase the transport bitrate and the spectrum utilization efficiency with a more narrow band [7]. The modulation of the ABPSK signal relies on the special impacting filter (SIF), which could convert phase changes to amplitude impacts. Details of the SIF will be given in Section 3.

In recent years, multicarrier (MC) scheme has been recognized as a potential candidate for the physical layer of the wireless communication systems which could not only gain spectral efficiency but also lower out-of-band radiation. The first candidate of MC schemes is orthogonal frequency-division multiplexing (OFDM) [8]; however, a number of shortcomings of OFDM in the MC systems have been noted as well, such as large side lobes of the frequency response of the filters that characterize the subcarrier channels, high peak-to-average power ratio (PAPR), and sensitivity to carrier frequency offset. Compared with OFDM, the ABPSK allows a high level of subchannel spectral containment as Figure 1 shows. So we propose that the ABSPK technique can be used in MC schemes.

FIGURE 1: Power spectrum of 1/4MCP-EBPSK signal with three subcarriers.

2. Uniform Formula of ABPSK

Denote the following.

- (1) Data bit duration T , it lasts $N \geq 1$ carrier wave period, that is, $T = N/f_c$.
- (2) Phase modulation duration τ , it lasts K carrier wave period, that is, $\tau = K/f_c$ and $K \leq N$.
- (3) g_0 and g_1 are modulation waveforms corresponding to bit “0” and bit “1”, respectively.

Then the uniform formula of ABPSK is defined as

$$g_0(t) = A \sin 2\pi f_c t, \quad 0 \leq t < T,$$

$$g_1(t) = \begin{cases} B \sin(2\pi f_c t \pm \sigma), & 0 \leq t < \tau, \quad 0 \leq \Delta \leq 1, \quad 0 \leq \eta \leq 1, \\ A \sin 2\pi f_c t, & \tau \leq t < T. \end{cases} \quad (1)$$

When the phase changes abruptly, $\sigma \in [0, \pi]$; when phase changes continuously, $\sigma = \pm \Delta \sin(\eta * 2\pi f_c t)$. The modulated waveform has the following features.

- (1) Data bit “0” lasts N carrier cycles in T , the carrier cycle is $1/f_c$ and the amplitude is A .
- (2) Data bit “1” consists of K carrier cycles with an initial phase σ , amplitude B and $N-K$ carrier cycles with the same phase as bit “0”, and amplitude A .
- (3) If there are no phase changes or amplitude changes, the waveforms are normal sine wave. When $\sigma \in [0, \pi]$, it is the EBPSK; when $\sigma = \pm \Delta \sin(\eta * 2\pi f_c t)$, according to the sign of Δ and value of η , it is the CP-EBPSK and MCP-EBPSK, respectively.
- (4) When $B = 0$, $\sigma \in [0, \pi]$, (1) is MCM as has been mentioned in [9]; when $B = A$, (1) denote the following.
 - (a) When $\sigma = \pi$, $\tau = T$, it represents the classical BPSK; when $\tau = T/4$, it represents the 3PRK in [9].
 - (b) When $\sigma = \pi/2$, $\tau = T/4$, it represents 3PSK in [9].

FIGURE 2: Modulated and demodulated waveform of EBPSK.

- (5) The modulation degrees differ according to the phase change S and modulate duration. The phase changes could control the bandwidth: when the phase vanishingly changes, the spectrum is the narrowest, and when the changes are π , the spectrum is the widest; Denote $\tau/T = K/N$ as “modulated duty cycle.”

According to (1), choosing a proper value of Δ , η , σ , τ and using a better shaping-filter could tighten the bandwidth of the modulated signal. We call the signal defined by (1) the uniformly asymmetric binary phase shift keying (ABPSK).

3. Special Impacting Filter (SIF)

The demodulation of AEBPSK signal based on the special digital impacting filter is in [10, 11]. The SIF is a kind of digital IIR filters; when working on the proper frequency range, it can convert the phase changes to amplitude impacting. The places where phase is unchanged have much lower impacting than the changed places and most noise is removed at the same time (see Figure 2). Based on the output response, direct amplitude detection can be used for the symbol judgment in intermediate frequency instead of down-conversion to base-band. The filtering mechanism in that filter offers different band to signal and noise, which is discussed in [10, 11] in detail.

Different from normal multicarrier transmission system, the ABPSK subcarriers need not guard interval or satisfy the orthogonal condition. Every subchannel set differs from the others. A narrow pass band of the SIF is used to separate the subcarriers. In order to eliminate the adjacent channel interference, we need to set the notch frequency of the cascade notch filter the same as other carriers frequency to suppress other subchannels' interference and the output demodulation results [10]. The tiny phase change of the ABPSK makes the demodulation performance closely related to the sampling rate. Only by setting the proper parameter

FIGURE 3: Frequency response of the SIF.

of the filter could it reflect the signal change. Based on the above condition, the parameters of the SIF bank correspond with each subcarrier's sampling rate, respectively.

4. Design of Cascade Notch SIF (CNSIF)

The key to demodulate the multicarrier ABPSK signal is the design of the receive filter bank, that is, cascade notch filter based on the SIF. Each subchannel regards the other subchannels as a sinusoidal interference, the notch filter cancelling the interference and SIF demodulating the corresponding signal. Many methods have been mentioned in [10, 11] for the SIF design and performance analysis, but the SIF pole radius is specified by the designer arbitrarily. Here we present a new approach to the numerical solution of the SIF as well as use an iterative scheme to gain the pole angle of the notch filter that fulfills the cost function.

4.1. Calculation of the SIF Pole Radii. Consider an SIF with one pair of conjugate zero-poles whose transfer function is

$$H(z) = \frac{(1 - r_z e^{j\omega_c - \varepsilon} z^{-1})(1 - r_z e^{-j\omega_c - \varepsilon} z^{-1})}{(1 - r_p e^{j\omega_c + \mu} z^{-1})(1 - r_p e^{-j\omega_c + \mu} z^{-1})}. \quad (2)$$

Different from the conventional IIR filters, the zero angles are less than the corresponding carrier frequency and are less than the pole angles. Place the zeros on the unit circle with angle shift ν , that is, $r_z = 1$, $\omega_z = \omega_c - \nu$; the poles are close to the unit circle and are near to zeros with angle shift μ , that is, $0 \ll r_p < 1$, $\omega_p = \omega_c + \mu$. Generally the carrier frequency is situated in the maximum slope of the magnitude response. Figure 3 shows the amplitude-frequency response and phase-frequency response of the demodulation impacting filter with one pair of conjugate zero-poles, the normalized carrier frequency is 0.2π rad/sample; it has narrow notch-frequency-selecting performance near the carrier frequency.

According to (2), we can evaluate the frequency response with graphical method (see Figure 4). In practice the poles and zeros are even much closer than Figure 4; the poles are

FIGURE 4: Schematic of zero-pole relative position.

even much closer to the unit circle. In the z -plane, the zero-pole vectors shift on the unit circle $e^{j\omega}$.

z_1, z_2 are zeros lying on the unit circle; p_1, p_2 are poles near the unit circle and zeros. $A_1, \psi_1, A_2, \psi_2, B_1, \theta_1, B_2, \theta_2$ represent the module and angle of z_1, z_2, p_1, p_2 , respectively. For the periodic frequency characteristic of the digital filters, we could just discuss the situation of $0 < \omega < \omega_s/2$, that is, $0 \sim \pi$.

From Figure 4, the frequency magnitude response of (2) can be rewritten as

$$|H(e^{j\omega})| = \frac{A_1 \cdot A_2}{B_1 \cdot B_2}. \quad (3)$$

Because the zeros and poles are close to the unit circle and the distances from the conjugate pole and zero to $e^{j\omega}$ are nearly equal when they shift in the region of $0 < \omega < \omega_z$ and $\omega_p < \omega < \omega_s/2$; that is, $A_1 \approx B_1, A_2 \approx B_2$. So

$$|H(e^{j\omega})| = 1, \quad 0 < \omega < \omega_z \cup \omega_p < \omega < \frac{\omega_s}{2}. \quad (4)$$

But when they shift in the region of $\omega_z < \omega < \omega_p$, the distance A_1, B_1 is much smaller than the distance A_2, B_2 ; we may state that $A_2 \approx B_2$. The frequency magnitude response can be written as

$$|H(e^{j\omega})| = \frac{A_1}{B_1}, \quad \omega_z < \omega < \omega_p. \quad (5)$$

Enlarge the region of $\omega_z < \omega < \omega_p$, the relative position between zero-pole vector and $e^{j\omega}$ depicted in Figure 5, here the influence of the complex conjugate zero-pole pair will be negligible.

FIGURE 5: Schematic of the enlarged relative position between zero-pole.

With the Pythagorean theorem, we obtain two equations:

$$A_1 = 2 \sin \left(\frac{\omega - \omega_z}{2} \right), \quad (6)$$

$$B_1 = \sqrt{h^2 + b^2} = \sqrt{r_p^2 - a^2 + b^2}, \quad (7)$$

where r_p is the module of pole p_1 .

Substituting $a = \sqrt{r_p^2 - r_p^2 \sin^2(\omega_p - \omega)}$, $a + b = 1$, into (7) leads to

$$B_1 = \sqrt{r_p^2 - 2r_p \cos(\omega_p - \omega) + 1}. \quad (8)$$

So the frequency magnitude response in the region of $\omega_z < \omega < \omega_p$ can be written as

$$|H(e^{j\omega})| = \frac{2 \sin[(\omega - \omega_z)/2]}{\sqrt{r_p^2 - 2r_p \cos(\omega_p - \omega) + 1}}, \quad \omega_z < \omega < \omega_p. \quad (9)$$

According to Moor's bandwidth definition [12], if the resonance gain over the DC gain exceeds 6 dB in absolute value, the frequencies at 3 dB below/above the peak/notch is the band edges, that is

$$|H(e^{j(\omega_c - B_w/2)})| = \frac{1}{\sqrt{2}} \cdot |H(e^{j\omega_c})|, \quad (10)$$

where B_w is the bandwidth of the SIF; (10) can be expressed as

$$\frac{\sin[(\omega_p - \omega_z)/2]}{\sin[(\omega_p - B_w/2 - \omega_z)/2]} \cdot \frac{\sqrt{r_p^2 - 2r_p \cos(B_w/2) + 1}}{1 - r_p} = \sqrt{2}. \quad (11)$$

To ensure the impacting effect, the pole and zero angle is specified by the designer. When the bandwidth B_w is known,

the results of the following two equations are constants, respectively, that is

$$\begin{aligned} \omega_p - \omega_z &= \phi, \\ \frac{2[1 - \cos(\phi - B_w/2)]}{1 - \cos \phi} &= M. \end{aligned} \quad (12)$$

Substituting (12) into (11), we can readily obtain

$$\begin{aligned} r_p &= \frac{M - \cos(B_w/2) \pm \sqrt{\cos^2(B_w/2) - 2M \cos(B_w/2) + 2M - 1}}{M - 1}. \end{aligned} \quad (13)$$

The poles and zeros are very close to each other, as a consequence, ϕ and B_w will be relatively small. To avoid $r_p > 1$, we abandon the adding situation. After some further simplification. The final result is

$$\begin{aligned} r_p &= \frac{M - \cos(B_w/2) - \sqrt{\cos^2(B_w/2) - 2M \cos(B_w/2) + 2M - 1}}{M - 1}. \end{aligned} \quad (14)$$

Equation (14) shows that the pole radii of SIF can be given in a quantitative way instead of arbitrary setting.

The procedure of the one-pair conjugate zero-poles SIF design could ensure the impacting effect, and we know that if the zeros and poles are far away from the unit circle, they will only affect the amplitude of the frequency response, but not the shape. From this standpoint, we can add more conjugate poles to tighten the bandwidth of the SIF and gain higher impacting effect. Generally we can add two conjugate poles whose angles are equal to the initial poles and are far away from the unit circle.

4.2. The Cascade Notch SIF (CNSIF). The multicarrier ABPSK scheme uses the CNSIF banks to separate and demodulate each subcarrier [13]. The key of the CNSIF is the narrow notch-frequency-selecting performance. When the overlapping signals pass the filter banks, the primary signal is enhanced and the noises or interferences are attenuated. Here the SIF enhances the primary signal and the notch filter decreases the disturbing signal. Let us start with the design of the notch filter in details.

Without loss of generality, we assume that the notch filter has k notch frequencies; the transfer function is given by

$$H_1(z) = \frac{B(z)}{A(z)} = \frac{\prod_{i=1}^k (1 - 2 \cos(\omega_{Ni}) z^{-1} + z^{-2})}{\prod_{i=1}^k (1 - 2r_i \cos(\omega_{pi}) z^{-1} + r_i^2 z^{-2})}, \quad (15)$$

where r_i is the pole radius. For stability and according to system requirements, the values of ω_{Ni} and r_i are given by the designer ($0 \leq r_i < 1$). Meanwhile the zeros are constrained to

locate on the unit circle at the notch frequencies ω_{N_i} , so the problem of designing a notch filter reduces to find the optimal value of ω_{p_i} .

Let $a_i = 2 \cos(\omega_{p_i})$, then $\omega_{p_i} = \arccos(a_i/2)$, $-2 \leq a_i \leq 2$, $-\pi \leq \omega_{p_i} \leq \pi$. The transfer function can be rewritten as

$$H_1(z) = \frac{\prod_{i=1}^k (1 - b_i z^{-1} + z^{-2})}{\prod_{i=1}^k (1 - r_i a_i z^{-1} + r_i^2 z^{-2})}, \quad (16)$$

where $b_i = 2 \cos(\omega_{N_i})$, $i = 1, 2, \dots, k$.

For convenience, let $i = 1$, then the notch frequency is ω_{N_1} , pole radius is r_1 , and (16) can be expressed as

$$H_1(z) = \frac{B(z)}{A(z)} = \frac{1 - b_1 \cdot z^{-1} + z^{-2}}{1 - a_1 r_1 z^{-1} + r_1^2 z^{-2}}. \quad (17)$$

Assuming that $\omega_z < \omega_p < \omega_{N_1} < \omega_{N_2} < \dots < \omega_{N_k}$, $\omega_{N_1} \geq \omega_p + B_w/2$. With a given weighting function $W(\omega)$, we can define the cost function as

$$C(a) = \int_R W(\omega) |H(e^{j\omega}) - H_1(e^{j\omega})|^2 d\omega, \quad (18)$$

where the region $R = [\omega_p, \omega_{N_1} - \delta] \cup [\omega_{N_1} + \delta, \pi]$, δ is a prescribed small positive number, combining (4) and (18) yields

$$C(a) = \int_R W(\omega) |1 - H_1(e^{j\omega})|^2 d\omega. \quad (19)$$

To ensure the impacting characteristic of SIF and cancel the signal disturbing, the cascade notch filter should not affect the frequency response of the SIF; that is, the cost function should be minimized.

Substituting (17) into (19), we can obtain

$$\begin{aligned} C(a_1) &= \int_R W(\omega) \left| 1 - \frac{B(e^{j\omega})}{A(e^{j\omega})} \right|^2 d\omega \\ &= \int_R \frac{W(\omega)}{|A(e^{j\omega})|^2} |A(e^{j\omega}) - B(e^{j\omega})|^2 d\omega \\ &= \int_R \frac{W(\omega)}{|A(e^{j\omega})|^2} \left| [(r_1^2 - 1)e^{-2j\omega} + b_1 e^{-j\omega}] - r_1 a_1 e^{-j\omega} \right|^2 d\omega. \end{aligned} \quad (20)$$

Let $W(\omega)/|A(e^{j\omega})|^2$ be the new weighting function, since r_1 and b_1 are known numbers, we need to find a_1 to minimize the cost function. For $a_1 = 2 \cos(\omega_{p_1})$, $\omega_{N_1} - \delta \leq \omega_{p_1} \leq \omega_{N_1} + \delta$, we can use an iteration scheme to find the optimal value of a_1 ; each iteration refreshes the value of $A(e^{j\omega})$. We

FIGURE 6: The frequency magnitude response of the designed CNSIF.

denote the value of a_1 at the n th iteration as a_{1n} , $A(e^{j\omega})$ as $A_n(e^{j\omega})$, similarly. When $A(e^{j\omega})$ gets the new value, the weighting function is updated immediately.

For convenience, substitute

$$\begin{aligned} p(\omega) &\leftarrow (r_1^2 - 1)e^{-2j\omega} + b_1 e^{-j\omega}, \quad q(\omega) \leftarrow -r_1 e^{-j\omega} \\ \Rightarrow C(a_{1n}) &= \int_R \frac{W(\omega)}{|A_n(e^{j\omega})|^2} |p(\omega) + q(\omega) a_{1n}|^2 d\omega \\ &= \int_R \frac{W(\omega)}{|A_n(e^{j\omega})|^2} \cdot [a_{1n}^2 |q(\omega)|^2 + 2a_{1n} |p(\omega) q(\omega)| \\ &\quad + |p(\omega)|^2] d\omega \\ &= \int_R \frac{W(\omega)}{|A_n(e^{j\omega})|^2} [a_{1n}^2 |q(\omega)|^2 \\ &\quad + 2a_{1n} \cdot \text{Re}(p(\omega) q^*(\omega)) \\ &\quad + |p(\omega)|^2] d\omega. \end{aligned} \quad (21)$$

After some simplification, (21) is rewritten as

$$\begin{aligned} C(a_{1n}) &= \int_R W_n(\omega) [|q(\omega)|^2 \cdot a_{1n}^2 + 2 \cdot \text{Re}(p(\omega) q^*(\omega)) \\ &\quad \cdot a_{1n} + |p(\omega)|^2] d\omega, \end{aligned} \quad (22)$$

where

$$W_n(\omega) = \frac{W(\omega)}{|A_{n-1}(e^{j\omega})|^2}. \quad (23)$$

FIGURE 7: The waveform after the CNSIF bank filtering.

So we can determine the value of a_{1n} by solving the constrained optimization problem

$$\begin{aligned}
 \min \quad & \int_R W_n(\omega) d\omega \cdot [|q(\omega)|^2 \cdot a_{1n}^2 + 2 \cdot \text{Re}(p(\omega) q^*(\omega)) \\
 & \cdot a_{1n} + |p(\omega)|^2] \\
 \text{s.t.} \quad & 0 \leq r_1 < 1 \\
 & -2 \leq a_{1n} \leq 2 \\
 & \omega_{N1} - \delta \leq \omega_{p1} \leq \omega_{N1} + \delta.
 \end{aligned} \tag{24}$$

We can see that the cost function is a parabola with its mouth opened up, so it could be transformed to a closed form and have an optimal solution. At the n th iteration, minimizing (22) subjected to constraints of (24) is a typical quadratic programming problem. We can compute the solution efficiently.

5. Design Example

5.1. Design Parameters. Set the normalized three notch frequencies as 0.21π , 0.25π , and 0.3π . Easily derive the initial values that $a_1 = 0.521292$, $a_2 = 0.555360$, $a_3 = 0.587785$. Let $\delta = 0.001\pi$; the convergent values are 0.52206411, 0.55613211, and 0.58855711. Given the following specifications: $r_1 = r_2 = r_3 = 0.95$; weighting function $W(\omega) = 1$; $\omega_z = 0.6282$ rad, $\omega_p = 0.62855$ rad, $B_w = \omega_p - \omega_z = 0.00035$ rad ($1.114e^{-4}\pi$ rad). The pole radii r_p defined in (14) is 0.999999969375. Adding another three pairs of conjugate poles with angle 0.62855rad , radii are 0.9, 0.92, and 0.95, respectively. The frequency magnitude response of the final designed CNSIF is shown in Figure 6.

5.2. Performance. We choose the MCP-EBPSK signal as a source, simulating the performance of two subcarriers. The parameters are as follows: $f_{c1} = 10$ MHz, $f_{c2} = 10.04$ MHz, $f_{s1} = 160$ MHz, $f_{s2} = 100.4$ MHz, $A = 1$; $B = 1$; $\Delta = 0.1$; $\eta = 0.5$; $K = 2$; $N_1 = 100$; $N_2 = 160$; the entire test symbols are

FIGURE 8: BER performances of two subcarriers.

3 million. Set the notch frequencies as 0.1255π and 0.1992π , we can obtain $a_1 = 1.84656$, $a_2 = 1.62098$, after six iterations converge to 1.84855455 and 1.6219734, respectively. Let $r_1 = r_2 = 0.95$, $\delta = 0.001\pi$, weighting function $W(\omega) = 1$.

We can obtain the parameters of the CNSIF bank as

$$\begin{aligned}
 a_1 = & [1, -6.09611187753762, 16.418167515832877, \\
 & -25.326641789930786, 24.298560015038895, \\
 & -14.721898165092053, 5.447828484314318, \\
 & -1.116449367584231, 0.098333159057181];
 \end{aligned}$$

$$\begin{aligned}
 b_1 = & [0, 0, 0, 0, 1, -3.694313616799019, 5.411987912102056, \\
 & -3.694313616799019, 1];
 \end{aligned}$$

$$\begin{aligned}
 a_2 = & [1, -5.406525327518239, 13.652579283244823, \\
 & -20.517799618098030, 19.861352592576445, \\
 & -12.536073668382514, 4.996805854728122, \\
 & -1.139889481405723, 0.120047843082279];
 \end{aligned}$$

$$\begin{aligned}
 b_2 = & [0, 0, 0, 0, 1, -3.239006288658281, 4.622788277906668, \\
 & -3.239006288658281, 1].
 \end{aligned}$$

Figure 7 shows the waveform of the component signal after the CNSIF bank filtering. We can see that the designed CNSIF could separate the component MCP-EBPSK signals. The BER performance is shown in Figure 8.

6. Conclusions

With the intuitive method known as zero-pole placement, we analyzed the frequency response of the SIF in a quantitative way. The equation about the pole radii under specified bandwidth is derived. The designed CNSIF could not only maintain the impacting characteristic but also attenuate the interchannel interference (ICI). From Figure 6, we can also see that adding notch zeros could speed up the decaying of the response curve and tighten the bandwidth; this will help to filter the noise. Compared with OFDM, the multicarrier scheme of ABPSK could have quite low PAPR ratio. It can also be combined with MIMO, cognitive radio, and so forth to get better performance, which is our next work to emphasize.

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