I. INTRODUCTION

Radar signals with large time-bandwidth product can be easily realized by repeating a basic waveform. The repetition may be continuous like CW signal, or have interval between pulses as coherent pulse train (CPT) [1]. The properties of phase-coded CW signals with perfect periodic autocorrelation functions (PACF) have been analyzed [2–4], and the receiver approaches towards these signals are also delivered [4–6]. Sequences with zero sidelobes in PACF are called perfect sequences, such as two-valued Golomb sequences [9], Ipatov ternary sequences [10], Frank sequences [11] and P codes [12, 13], etc. Perfect sequences are also employed in multi-carrier CW radar signals [7, 8] which have improved pulse compression ratio and high resolution ability. However, when long signal duration and good transmitter/receiver isolation are required simultaneously, interrupted CW (ICW) signal is preferred such as FMICW signal [15]. In this paper we mainly involve a phase-coded ICW signal and its digital receiver approach. Fig. 1 portrays the waveform of single-carrier ICW signal which is composed of a CW signal \( u(t) \) gated in each phase element by a CPT \( g(t) \), similar radar waveforms are applied in ionosonde StudioS [16]. The term “ICW signal” also means \( t_p / t_b \leq 0.5 \) and the exclusion of very narrow pulses, namely the cases when \( t_p \ll t_b \), where \( t_p \) and \( t_b \) denote the pulselength and pulse repetition time of \( g(t) \), respectively.

![Waveform of single-carrier ICW signal.](image)

Periodic ambiguity function (PAF) is a good tool to study the resolution ability of a radar signal. So we introduce the PAF of phase-coded ICW signals in Section II. Section III gives the analysis to the PAF of single-carrier ICW signals and the effects of \( g(t) \). Then we study the properties of multi-carrier complementary phase-coded (MCPC) ICW signals in Section IV and find out some interesting results. There are several designs of MCPC ICW signals, one of them bears resemblance to a single-carrier ICW signal and yields nearly perfect autocorrelation. In Section V, we suggest an efficient receiver approach to ICW signals with perfect sequences, which can reduce the computational burden of the processor. Meanwhile, the effect of bandwidth limitation on ICW signals is also investigated. Discussion and conclusion are made in Section VI.
II. PAF OF PHASE-CODED ICW SIGNALS

The PAF describes the response of a correlation receiver to a continuous signal modulated by a periodic waveform with period \( T \), thus the reference signal is of duration of \( MT \). The response is a function of both delay and Doppler shift, and the PAF is a two-dimension generalization of the PACF by including the effect of Doppler shift. To study the PAF of ICW signals, our discussion involves a coherent receiver to a continuous signal modulated by a periodic waveform with period \( T \). The illumination time is a few periods longer than \( MT \). Let the transmitted CW signal be with a periodic complex envelope \( u(t) \) with period \( T \),

\[
  u(t) = u(t - mT), \quad m = 0, \pm 1, \pm 2, \ldots \tag{1}
\]

Each period is divided into \( N \) slices by phase codes with bit duration \( t_b \), namely \( T = N t_b \). Then the expression of the ICW signal is given by

\[
  u_t(t) = u(t)g(t) \tag{2}
\]

where

\[
  g(t) = \begin{cases} 
    1, & m t_b \leq t < m t_b + t_p \\
    0, & m t_b + t_p \leq t < (m + 1)t_b 
  \end{cases} \tag{3}
\]

for all \( n \in \mathbb{Z} \). Obviously \( u_t(t) \) maintains the periodicity of \( u(t) \),

\[
  u_t(t) = u_t(t - mT). \tag{4}
\]

The PAF \( \chi_T \) and \( \chi_{MT} \) for CW signals were defined in [3]. Consequently for ICW signals, formula (5) is the single-period PAF and (6) is the PAF for \( M \) periods

\[
  \chi_T^1(\tau, f) = \frac{1}{T} \int_0^T u_t(t + \frac{T}{2}) u_t^*(t - \frac{T}{2}) e^{j2\pi f t} dt \tag{5}
\]

and

\[
  \chi_{MT}^1(\tau, f) = \frac{1}{MT} \int_0^{MT} u_t(t + \frac{T}{2}) u_t^*(t - \frac{T}{2}) e^{j2\pi f t} dt \tag{6}
\]

\( \chi_T^1 \) and \( \chi_{MT}^1 \) maintain the symmetry and periodicity of \( \chi_T \) and \( \chi_{MT} \) in [3], especially we still have

\[
  \chi_{MT}^1(\tau, f) = \chi_T^1(\tau, f) \frac{\sin(\pi f MT)}{M \sin(\pi f T)} e^{j2\pi f (M-1)T}. \tag{7}
\]

Equation (7) is an important result indicating the relationship between \( \chi_T^1 \) and \( \chi_{MT}^1 \), and PAF of phase-coded ICW signals is given by (5) and (6) including the cases of single-carrier signals and multi-carrier signals.

III. PAF OF SINGLE-CARRIER ICW SIGNAL

For single-carrier CW signals, each period is constructed from a sequence of \( N \) bits of duration \( t_b \),

\[
  u(t) = \sum_{n=0}^{N-1} b_n(t - nt_b), \quad 0 \leq t \leq T \tag{8}
\]

where

\[
  b_n(t) = \begin{cases} 
    \exp(j\phi_n), & 0 \leq t < t_b \\
    0, & \text{elsewhere} \end{cases} \tag{9}
\]

The PACF values of \( u(t) \) at delays which are multiples of \( t_b \) are given by

\[
  C(k) = C(k t_b) = \frac{1}{N} \sum_{n=0}^{N-1} b_n b_n^*,
\]

\[
  k = 0, 1, 2, \ldots \tag{10}
\]

and the perfect PACF is described as follows,

\[
  C(k) = \begin{cases} 
    1, & k = 0(\text{mod} \ N) \\
    0, & k \neq 0(\text{mod} \ N) \end{cases}. \tag{11}
\]

Equation (3) in [11] gives continuous PACF of CW signal. From (2), (3), (5), and (10), we obtain the PACF of ICW signal

\[
  C_1(\tau) = \chi_T^1(\tau, 0) = \frac{1}{T} \int_0^T u_t(t + \frac{T}{2}) u_t^*(t - \frac{T}{2}) dt
\]

\[
  = \frac{1}{T} \sum_{n=0}^{N-1} \int_{nt_b}^{nt_b + t_p} u(t + \frac{T}{2}) u^*(t - \frac{T}{2}) dt,
\]

\[
  k t_b - t_p < \tau < k t_b + t_p
\]

\[
  = \frac{1}{T} \sum_{n=0}^{N-1} \int_{nt_b + t_p}^{nt_b + t_p + \tau} u(t + \frac{T}{2}) u^*(t - \frac{T}{2}) dt,
\]

\[
  k t_b - t_p < \tau < k t_b + t_p
\]

\[
  = \frac{t_p - |\tau - k t_b|}{t_b} C(k), \quad k t_b - t_p < \tau < k t_b + t_p
\]

\[
  = 0, \quad \text{elsewhere}
\]

which is the cut of PAF along the delay axis (zero Doppler). We use the P3 code to demonstrate the performances of single-carrier signals. We observe in Fig. 2 that \( g(t) \) confines the PACF mainlobe width of the single-carrier signal to as far as \( t_p \) by gating the pulsewidth of phase elements, and thus contributes to better range resolution. In other words, \( g(t) \) enlarges the bandwidth of the ICW signal to \( B = 1/t_p \). Fig. 2
also demonstrates that the ICW signal suffers from the same recurrent range ambiguities at \( \tau = mT \) as CW signal, where \( m = \pm 1, \pm 2, \ldots \).

The cut of PAF along the Doppler axis is obtained by setting \( \tau = 0 \) in (5) which yields

\[
|\chi_0^1(0,f)| = \frac{1}{T} \int_0^T |u(t)|^2 e^{2\pi ft} dt
\]

\[
= \frac{1}{T} \sum_{n=0}^{N-1} \int_{tb}^{nb+tb} |u(t)|^2 e^{2\pi ft} dt
\]

\[
= \frac{tp \sin(\pi ft_p)}{tb \pi ft_p} \sin(\pi fMtb) e^{\pi f(N-1)tb+tp}.
\]

(14)

Hence, from (7) and (14) we have that

\[
|\chi_{MT}^1(0,f)| = \frac{tp \sin(\pi ft_p) \sin(\pi fMT)}{tb \pi ft_p \sin(\pi ftb)}. \tag{15}
\]

Fig. 3 depicts the Doppler performance of the single-carrier signal. Compared with the CW signal, the ICW signal exhibits the same Doppler resolution that \( \Delta f = 1/(MT) \), and bears Doppler ambiguities at \( f = m/tb \) due to \( g(t) \). Fig. 4 explains these Doppler ambiguities clearly. The PAF of the CW signal (given in Fig. 4(a)) exhibits a ridge that crosses the Doppler axis at \( f = 0.5/tb \), and this ridge has a null at \( \tau = 0 \).

When \( tp = tb/2 \), \( g(t) \) shifts the nulls of the ridge to \( \tau = \pm tb/2 \) and thus causes the ambiguities of the ICW signal (depicted in Fig. 4(b)).
IV. PAF OF MULTI-CARRIER ICW SIGNAL

Now we come to the MCPC radar signal [7] which employs $N$ subcarriers simultaneously and the subcarriers are phase modulated by $N$ different sequences that together constitute an $N \times M$ complementary set. The subcarriers are frequency separated by the inverse of the duration of a phase element $t_p$, yielding orthogonal frequency division multiplexing (OFDM) which is well known in communications. An MCPC ICW signal can be produced by gating either on each carrier or on the final complex envelope. When an MCPC ICW signal is produced by gating on subcarriers, the bandwidth of each subcarrier will be enlarged to $B = 1/t_p$, so the mathematical expression of one period of an $N \times M$ MCPC ICW signal is given by

$$u_1(t) = \begin{cases} \sum_{n=1}^{N} W_n \exp \left[ j2\pi \left( \frac{N}{2} - n \right) t/t_p \right] & 0 \leq t \leq M t_b \\ \times \sum_{n=1}^{M} b_{n,m}[t - (m-1)t_b], & \text{otherwise} \end{cases}$$

(16)

where

$$b_{n,m}(t) = \begin{cases} \exp(j \phi_{n,m}), & 0 \leq t < t_b \\ 0, & \text{elsewhere} \end{cases}$$

(17)

$\phi_{n,m}$ is the $m$th phase element of the $n$th code sequence, and $W_n$ is the amplitude weight assigned to the $n$th subcarrier. There is a universal PACF of an MCPC ICW signal with perfect sequences which is a function only of $M$, $N$, the weights $W_n$ and the shifts $\theta_n$ (dimensionless integer), and the PACF is expressed as

$$R_l(k t_b + \eta) = \frac{t_p}{t_b} \bar{W}_0 \sum_{n=1}^{N} W_n W_n^* \bar{W}_0 \sum_{n=1}^{N} W_n W_n^*$$

$$\times \exp \left[ j2\pi \left( \frac{N}{2} - n \right) \frac{\eta}{t_p} \right]$$

$$\times [I_1(\mu_n - \mu_l + k + 1) + I_2(\mu_n - \mu_l - k)]$$

$$= \sum_{n=1}^{N} \sum_{l=1}^{M} R_{n,l}^{	ext{IS}}$$

(18)

where $\delta$ implies Kronecker delta, $[\theta]_M$ implies $\theta$ modulo $M$ and $R_{n,l}^{	ext{IS}}$ refers to the cross-correlation function (CCF) between the $n$th subcarrier and the $l$th subcarrier.

When $0 \leq \eta < t_p$,

$$I_1 = \begin{cases} \eta, & \text{when } n = l \\ \eta \sin(\beta) \exp(j\beta), & \text{when } n \neq l \end{cases}$$

(19)

$$I_2 = \begin{cases} t_b - \eta, & \text{when } n = l \\ -I_1, & \text{when } n \neq l \end{cases}$$

(20)

and when $t_p \leq \eta < t_b$,

$$I_1 = I_2 = 0.$$

(21)

$\theta_n$ tells us by how many elements is the sequence on the $n$th carrier shifted relative to the basic sequence. There are two designs of MCPC ICW signals based on $\theta_n$. The first one uses consecutive ordered cyclic shifts (COCS) along carriers, namely $\theta_n = n$, where $n = 1, 2, \ldots, N$. In the second design there are no shifts at all. The same code sequence modulates all the carriers, i.e., $\theta_n = 1$. The second design is also called identical sequence MCPC (IS MCPC) signal.

The performances of the MCPC ICW signal along the delay axis are demonstrated using 11-element Golomb code. We observe in Fig. 5 that the gate function $g(t)$ “compresses” the complex envelope of the MCPC pulse from $t_b$ to $t_p$, which is different from the single-carrier signal, and leads to the “compression” of $R_{n,l}^{	ext{IS}}$, the CCF of the MCPC CW signal (depicted in Fig. 6). So compared with the MCPC CW signal, the MCPC ICW signal obtains narrower PACF mainlobe width as far as $t_p/N$ and lower autocorrelation sidelobes within $|\tau| < t_b$ by this compression. Fig. 7 gives the PACF of the MCPC signal based on two designs.

However, the ICW signal maintains an unambiguous Doppler range of $-1/t_b < f < 1/t_b$ and obtains better range resolution than CW signal. Except for these, the performances of ICW signals in the delay-Doppler domain resemble the performances of CW signals in the single-carrier field.

We find out that decreasing $t_p/t_b$ can reduce PACF sidelobes of the MCPC ICW signal besides frequency weighting and increasing carrier number, but the reduction is limited since we exclude the situation of very narrow pulses. Meanwhile, only frequency weighting on subcarriers has no effect on sidelobes near $\tau = m t_p$ for the MCPC ICW signal based on COCS. To COCS MCPC ICW signals composed of Golomb sequence and Frank code, alternating the polarity of the weights between consecutive carriers can reduce their sidelobes further and make the PACF of the CW signal comparable with the PACF of the ICW signal. The PACF of the $33 \times 11$ MCPC ICW signal based on COCS and alternating polarity of Chebyshev weighting along carriers is plotted in Figs. 8 and 9.
Fig. 5. Partial complex envelope of MCPC signals based on Golomb code. (a) $11 \times 11$ MCPC signal based on COCS. (b) $11 \times 11$ IS MCPC signal with Chebyshev weighting.

Fig. 6. CCF of $11 \times 11$ MCPC signal based on COCS.

Fig. 7. Partial PACF of MCPC signals. (a) $11 \times 11$ MCPC signal based on COCS. (b) $11 \times 11$ IS MCPC signal.

It is interesting that the envelope of the IS MCPC CW signal bears a resemblance to the single-carrier ICW signal, and $g(t)$ compresses its envelope and PACF as well, depicted in Figs. 5(b) and 7(b). Fig. 10 presents the PACF of the $33 \times 11$ IS MCPC ICW signal, alternating polarity of Kaiser weighting along carriers and $t_p = t_b/2$. Comparison between Fig. 8 and Fig. 10 reveals that the IS MCPC ICW signal yields nearly perfect PACF.
We also examine the cut of PAF along the Doppler axis of the MCPC ICW signal and find out that it depends on the type of code and $\theta_n$. The Doppler performances of several MCPC ICW signals are plotted in Fig. 11. Similar to the single-carrier ICW signal, the Doppler resolution of the MCPC ICW signal remains $1/(MT)$ and the ambiguity at $f = 1/t_p$ is clear. Fig. 11(b) demonstrates the resemblance of the IS MCPC ICW signal to the single-carrier ICW signal in Doppler domain. Finally the Golomb code is the basis for the $11 \times 11$ MCPC ICW signals whose partial PAFs are plotted in Figs. 12(a) and 12(b), which present the signals based on two designs, respectively. The number of periods is eight and a smooth Hamming on-receive inter-period window is used. The frequency weighting, performed at both transmitter and receiver is the square root of Hamming.

An MCPC ICW signal can also be produced by direct gating on the complex envelope of the MCPC CW signal. But this kind of the ICW signal doesn’t yield an ideal PACF because of the varying envelope of an MCPC signal and $g(t)$ doesn’t narrow the PACF mainlobe width significantly, depicted in Fig. 13. Moreover, Fig. 13 demonstrates the deterioration of PACF sidelobes due to $g(t)$. So this kind of the ICW signal is not an appropriate radar signal.
V. RECEIVER APPROACH TO THE ICW SIGNAL

So far we have discussed the theoretical PAF of phase-coded ICW signals. In this section we suggest an efficient receiver approach to these signals. Although there are differences between single-carrier signals and multi-carrier signals, the approaches towards them are similar.

The single-carrier signal is discussed first. Levanon and Getz [4, 5] have studied the matched filters for single-carrier CW signals and Popovic [6] employed the optimum algorithm to the aperiodic autocorrelation function (ACF). Here we emphasized our digital receiver approach upon PACF. Digital signal processing requires sampling of the received signal. In the phase-coded signal, sample rate should be no less than the signal bandwidth. To the CW signal, there is an obvious sampling choice: one sample per bit, i.e., sample rate \( f_s = 1/t_b \). But as to ICW signals we discussed in Section III, the sample rate should be increased to at least \( 1/t_p \) in order to achieve better range resolution. Inter-period weighting
is utilized in time domain to reduce the Doppler sidelobes. Let \( t_b = L t_p \), where \( L \) is an integer. After sampling at interval of \( t_p \) and weighting, we present \( M \) periods of digital received signals as follows:

\[
S = \begin{bmatrix}
    s_{11} & s_{12} & \cdots & s_{1N} \\
    s_{21} & s_{22} & \cdots & s_{2N} \\
    \vdots & \vdots & \ddots & \vdots \\
    s_{M1} & s_{M2} & \cdots & s_{MN}
\end{bmatrix}
\]  

(22)

where \( s_{mn} = [s_{mn}]^{d_t} \).

\( S \) is an \( M \times NL \) matrix. Similar to approaches to the CW signal, correlation is implemented for every Doppler cell following Doppler compensation. Coarse Doppler compensation is obtained by performing \( NL \) fast Fourier transforms (FFTs) along columns of \( S \) which compensate the Doppler shift between sequences, and then fine Doppler compensation is performed in the complex matrix (with elements \( d_e = e^{-j2\pi/MNL} \) which denotes the phase shift between \( t_p \)) that multiplies all the FFT outputs which compensates the Doppler shift between “consecutive” bits. Let \( S_1 \) and \( S_2 \) represent the coarse and fine Doppler compensation results. We have that

\[ S_2(m, p) = S_1(m, p) e^{j(m-1)(p-1)} \]

(23)

where \( m = 1, \ldots, M \) and \( p = 1, \ldots, NL \).

Next, we obtain a set of correlators.

Cross-correlation between the conjugate of the original signal and the received signal will result in the perfect PACF. Furthermore, \( P3 \) and \( P4 \) codes with original signal and the received signal will result in cross-correlation between the conjugate of \( S \) and \( S \) of \( NL \) periods of digital received signals as follows:

\[ \Delta f = \frac{2\pi T}{MNL} \]

(24)

where \( f_0 \) is the RF carrier and \( v \) denotes velocity of echoes. From (24) we can see that echoes with the same velocity will fall into a different Doppler cell if \( |f_0 - f_0^N| \geq \Delta f/2 = 1/(2MNt_b) \). To compensate this frequency shift offset, we apply chirp transformation algorithm (CTA, see [14, ch. 9.6]) instead of FFT along columns of \( S \). Let \( \omega_0 \) and \( \Delta \omega \) respectively represent frequency offset and frequency interval on unity. CTA can be expressed as

\[ S(\omega_0) = \sum_{m=0}^{M-1} s(m) e^{-j(\omega_0 + k\Delta \omega)m} \]

(25)

As uniform sampling is implemented in Doppler domain, we obtain that

\[ \Delta \omega = 2\pi T \Delta f = 2\pi Nt_b \frac{1}{MNL} = \frac{2\pi}{M} \]

(26)

where \( T = Nt_b \). Assume that the Doppler frequency shift caused by \( f_0 \) is without frequency offset, i.e.,

\[ \frac{2v}{c} f_0 = k \Delta f = \frac{k}{MNt_b}, \quad k = 0, 1, \ldots, M - 1 \]

(27)

From (24)–(27), the frequency offset \( \omega_0 \) of the \( n \)th subcarrier is given by

\[ \omega_0 = 2\pi T \left( f_d - \frac{2v}{c} f_0 \right) \]

\[ = 2\pi Nt_b \times \frac{2v}{c} \left( \frac{N + 1}{2} - n \right) \frac{1}{t_p} \]

\[ = \frac{2\pi k}{Mt_pb} \left( \frac{N + 1}{2} - n \right) \]

(28)

Thus the Doppler frequency shift of each subcarrier equals that of \( f_0 \) by using CTA, and the offsets within subcarriers are removed. CTA of the \( n \)th
Fig. 15. Bandwidth limitation of single-carrier ICW signal based on 64-element P3 code and \( t_p = t_b/2 \). Sample rate \( f_s = 8/t_b \).

Fig. 16. CCF of bandwidth limited single-carrier ICW signal based on 64-element P3 code and sample rate \( f_s = 8/t_b \).

subcarrier is implemented with coefficients determined by (26) and (28). Let \( S^c \) and \( S^f \) represent the coarse and fine Doppler compensation results. We have that

\[
S^f(m,p) = S^c(m,p) \exp[-j(2\pi/MNNL)(m-1)(p-1)]
\]

(29)

where \( m = 1, \ldots, M \) and \( p = 1, \ldots, NNL \).

Of course when \( |f_r - f_m| < 1/(2MNt_b) \), the frequency shift offset might be ignored and direct FFT can be implemented in Doppler compensation.

After Doppler compensation, a bank of PSCs will be employed to every Doppler cell. Equation (18) demonstrates that computing the PACF of MCPC signals includes cross-correlation performed between each one of the \( N \) received sequences and the corresponding one of \( N \) reference sequences. It means each Doppler cell needs \( N^2 \) PSCs. The cross-correlation complex outputs are then combined to yield the postcompression output. The major modification in PSC for the MCPC signal is that it will reduce the sample rate of input data by a factor of \( NL \) and thus it has \( NL \) correlator units. After correlation, it has to recover the sample rate by the same factor. The numbers of complex multiplications required by direct correlation are given by \( ML^2N^6 \) while PSC needs \( MLN^3 \) complex multiplications. For the case \( N = 11, L = 2 \) the PSC needs less than 5% of the number of multipliers required by direct correlation. In case of MCPC signal based on P3 or P4 codes, the number of multipliers will decrease to \( MLN^3 \log_2 N + 1 \), which will reduce the computational burden of the processor a great deal.

Considering the fact that oversampling is usually applied in real receiver, our approach is in fact much more efficient than we have analyzed above. On the other hand, precompression bandwidth limitation will be found in any well-designed radar receiver to avoid out-of-band noise foldover and signal aliasing. The effect of bandwidth limitation on ICW signals under consideration is investigated by bandlimiting echo signal to be compressed by a digital half-band finite-duration impulse response (FIR). Fig. 15 exhibits the process of bandwidth limitation to a single-carrier ICW signal, and Fig. 16 gives the CCF after bandlimiting. Fig. 17 presents the CCF of MCPC ICW signals after bandlimiting. Simulation results reveal that when the bandwidth of the receiver equals the bandwidth of the ICW signal, such as \( 1/t_p \) for single-carrier signal and \( N/t_p \) for multi-carrier signal, adequate peak signal to peak sidelobe ratio (PSR) of ICW signals will maintain if appropriate code type and code length are chosen.

VI. DISCUSSION AND CONCLUSION

We discuss a phase-coded ICW radar signal in this paper. Phase codes used here are with perfect PACF in order to study the performances of ICW signals.
Fig. 17. CCF of bandwidth limited MCPC ICW signal based on 11-element Golomb code and \( t_p = t_b/2 \). Sample rate \( f_s = 132/t_b \).
(a) 33 \( \times \) 11 MCPC ICW signal based on COCS with Chebyshev weighting, \( t_p = t_b/2 \). (b) 33 \( \times \) 11 IS MCPC ICW signal with Kaiser weighting, \( t_p = t_b/2 \).

in the delay-Doppler domain. The introduction of \( g(t) \) to single-carrier signals achieves better range resolution and brings Doppler ambiguity at the same time. Actually, \( g(t) \) enlarges the bandwidth of single-carrier ICW signal to \( B = 1/t_p \). Although direct gating on the complex envelope of a multi-carrier CW signal doesn’t yield a satisfactory radar signal, there is a convenient way to produce multi-carrier ICW signals just by modulating \( N \) single-carrier ICW signals simultaneously which are frequency separated by the inverse of the pulsewidth of a phase element \( t_p \), and an MCPC ICW signal is a good example. Different from single-carrier signals, the gate function \( g(t) \) compresses the complex envelope and PACF mainlobe width of the MCPC signal, and suppresses its autocorrelation sidelobes as well, especially within \( |\tau| < t_b \). Although the cut of PAF along Doppler axis of the MCPC ICW signal depends on the type of code and \( \theta_p \), its Doppler resolution remains \( 1/(MT) \) and the ambiguity at \( f = 1/t_b \) due to \( g(t) \) is clear like a single-carrier ICW signal. Fortunately, the Doppler ambiguity of the ICW signals won’t affect their Doppler resolution ability within a Doppler range of \(-1/t_b < f < 1/t_b\).

There are two designs of MCPC ICW signals based on \( \theta_p \). Decreasing \( t_p/t_b \) can reduce PACF sidelobes of MCPC ICW signals besides conventional methods, and the polarity of the weights should alternate between consecutive carriers for both designs. Simulation results reveal that the PACF sidelobe level of the MCPC ICW signal based on COCS is comparable with that of the CW signal while the IS MCPC ICW signal yields nearly perfect PACF. The resemblance of the IS MCPC signal to the single-carrier ICW signal is the inherent reason and \( g(t) \) helps to reduce the autocorrelation mainlobe width and integrated sidelobe level. Fig. 5 exhibits that MCPC ICW signals have the same requirement of linear power amplifier (LPA) in the transmitter as MCPC CW signals. But generally speaking, the peak-to-average power ratio (PAPR) of the ICW signal is larger than that of the CW signal because the transmitter will be shut down when receiving radar echoes. So we exclude the scheme of very narrow pulses. Low PAPR MCPC signal design was also covered in [7, 8]. On the other hand, if the MCPC ICW signal is produced by generating each carrier separately and power combining takes place in the air, the PAPR will be \( t_b/t_p \).

High range resolution means enlarged signal bandwidth which leads to increased sample rate and computational load at the receiver end. So we suggest an efficient digital receiver approach to phase-coded ICW signals. The MCPC ICW signal can be treated as a single-carrier ICW signal after being divided into \( N \) subcarriers. But CTA may be implemented
instead of FFT in Doppler compensation due to the Doppler shift offsets within subcarriers. After Doppler compensation, PSC is employed in correlation to reduce the computational burden of the processor and utilize the good properties of P codes. Analysis results show that the number of complex multipliers will be decreased spectacularly when PSC is used to pulse compression of MCPC ICW signals. At the same time, simulation results also demonstrate the robustness of the ICW signals to the bandwidth limitation of receivers.

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