SUCCESSIVE-MFCW MODULATION FOR ULTRA-FAST NARROWBAND RADAR

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ABSTRACT

Real time imaging radar requires short dwell time and low hardware complexity, limiting the application of many existing radar modulation methods. In this paper, we propose a novel Successive Multiple Frequency Continuous Wave (S-MFCW) modulation scheme to address these challenges. The S-MFCW signal model is formulated and the impact of noise is analyzed, leading to a very short dwell time and narrow signal bandwidth with excellent range and speed accuracy. From the numerical simulation of an automotive radar example, we show that S-MFCW achieves comparable range and speed performance, but orders of magnitude lower requirements on dwell time and bandwidth compared to the commonly used FMCW radar.

***Index Terms*—** Radar signal processing, millimeter wave radar, radar interferometry

1. INTRODUCTION

Real time scanning surveillance radar systems can be used in many commercial applications, such as automotive radar [1] and human feature extraction [2]. Typical scanning radars often operate under the constraint of only a limited number of pulses, thus real time estimation and detection of radar targets are quite difficult due to short dwell time (i.e., the time that an antenna beam spends on a target). Moreover, as modern radar systems are becoming increasingly sophisticated, systems with low hardware complexity and low cost while still achieving comparable high range and speed accuracy and high resolution are in strong demand.

For modern radars, frequency or phase modulation is normally needed to achieve desired radar performance. For example, the Frequency Modulated Continuous Wave (FMCW) technique is widely used due to its good range and velocity resolution [3, 4, 5]. However its resolution is inversely proportional to the signal bandwidth and it requires a long sampling window, hence a long dwell time, to complete a range measurement. Furthermore, FMCW radars usually use Fast Fourier Transform (FFT) to estimate Doppler frequencies, which results in high hardware complexity and high power consumption. The Multi Frequency Continuous Wave (MFCW) is another modulation scheme that continuously transmits multiple frequency tones in parallel with each other [2, 6, 7]. The phases of the returned signals are compared to obtain a range estimate. This technique allows a really short dwell time and a very small signal bandwidth. However, one transceiver is needed for each frequency tone, making the MFCW radar very costly in implementation and not suitable for large scale phased array integration.

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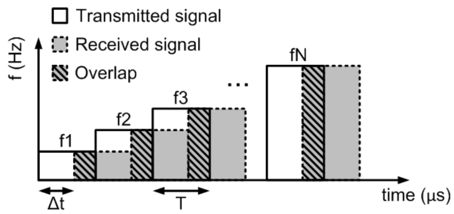

Fig. 1. S-MFCW transmitted and received signals with a sequence of frequency tones $f_1, \cdots, f_N$. T is the pulse width.

Prior work on radar modulation schemes such as the above mentioned FMCW and MFCW still face the challenges of balancing trade-offs of excellent radar performance against high hardware complexity and cost. To address this issue, in this paper we propose a new Successive-MFCW (S-MFCW) modulation scheme suitable for high speed radar estimation and detection and large scale integration for phased arrays. Unlike the typical MFCW modulation, the S-MFCW method sequentially transmits selected frequency tones from the same transceiver. Thus it results in comparable dwell time and signal bandwidth to the MFCW method, while drastically reducing the hardware complexity and enabling its use in large phased array systems. Although the successive transmission of the tones imposes constraints on the system performances such as range and velocity ambiguity, custom tone sequences can be devised for practical applications. Finally, the numerical simulation of a 77 GHz automotive radar is used as an example to verify the performance and noise impact of the proposed scheme.

2. SUCCESSIVE-MFCW SIGNAL MODEL

2.1. Signal Model

The transmitted S-MFCW signal consists of a sequence of N frequency tones $(f_1, \cdots, f_N)$ each with a pulse width of T (see Fig. 1 for illustration) and can be modeled as

$$s(t) = \sum_{i=1}^{N} \exp(j2\pi f_i t + j\phi_i(t)) \cdot \mathrm{rect}[(t-(2i-1)T/2)/T], \quad (1)$$

where the rectangular function is given by

$$\mathrm{rect}(t) = \begin{cases} 0, & |t| > 0.5 \\ 0.5, & |t| = 0.5 \\ 1, & |t| < 0.5 \end{cases} \quad (2)$$

and $\phi_i(t)$ represents the initial random phase noise of each tone. The i-th transmitted signal at frequency f_i is reflected from the target and detected by the receiver after a delay starting at $t = \Delta t_i = \Delta t((i-1)T)$. Let's consider a target located at the beginning of the measurement at a distance R, and moving in the direction of the radar beam at a constant speed v. The speed is considered positive if the target moves away from the radar. Thus, because the transmitted signal is an exponential function, we can find a delay that includes the doppler contribution in the form:

$$\Delta t(t) = 2 \cdot (R + v \cdot t)/c, \tag{3}$$

where c is the speed of light. Then the returned signal at the receiver antenna is:

$$\begin{aligned} r(t) &= A \cdot s(t - \Delta t(t)) + n(t) \\ &= \sum_{i=1}^{N} \{A \cdot \exp[j2\pi f_i(t - \Delta t(t)) + j\phi_i(t - \Delta t(t))] \\ &\quad + n_i(t)\} \cdot \mathrm{rect}[(t - \Delta t_i - (2i-1)T/2)/T], \end{aligned} \tag{4}$$

where A represents the propagation and reflection losses and $n(t)$ is the additive Gaussian noise. A has been assumed constant for each tone because we are going to consider a narrowband system ($f_1 \approx f_2 \approx \cdots \approx f_N$). We assume that $n(t)$ can be decomposed into N independent components, with $n_i(t)$ referring to a white bandpass Gaussian random process with zero mean and variance σ^2 due to signal return from the i-th tone. The received signal $r(t)$ will undergo processing to extract the phase information for range and speed estimation. To that end, the received signal is multiplied by the transmitted signal for demodulation, i.e., $r(t)^* \cdot s(t)$, where $(\cdot)^*$ is the phase conjugation, followed by a low-pass filter, yielding the post-processed signal

$$\begin{aligned} u(t) = \sum_{i=1}^{N} \Bigg\{ &[A \cdot \exp(j2\pi f_i \Delta t(t)) + n'_i(t)] \\ &\cdot \mathrm{rect}\left[\frac{t - (\Delta t_i + T)/2 - (i-1)T}{T - \Delta t_i}\right] \Bigg\}. \end{aligned} \tag{5}$$

$n'_i(t)$ follows a Gaussian distribution with zero mean and variance σ^2. We note that the $\phi_i(t)$ contribution disappears because the received phase noise is correlated to the transmitted phase noise [8]. From (5), we note that for each tone the received signal can be sampled only during the period of overlap between the transmitted and received signals (see the overlapped intervals in Fig. 1). To analyze the impact of the noise, we further model the noise for the i-th frequency tone as

$$n'_i(t) = \alpha_i(t)\exp[j\psi_i(t)], \tag{6}$$

where $\alpha_i(t)$ denotes the envelop of the noise and follows a Rayleigh distribution, while $\psi_i(t)$ represents the phase of the noise and has a uniform distribution within $[0, 2\pi]$. Next, we define the phase of each tone of the post-processed signal without noise as:

$$\Phi_i(t) = 2\pi f_i \Delta t(t). \tag{7}$$

Then, inserting (6) and (7) into (5), we obtain

$$\begin{aligned} u(t) = \sum_{i=1}^{N} \Bigg\{ &[A \, \exp(j\Phi_i(t)) + \alpha_i(t)\exp(j\psi_i(t))] \\ &\cdot \mathrm{rect}\left[\frac{t - (\Delta t_i + T)/2 - (i-1)T}{T - \Delta t_i}\right] \Bigg\}. \end{aligned} \tag{8}$$

Hence, the phase of the i-th frequency tone of the received signal can be derived as [6]:

$$\begin{aligned} \hat{\Phi}_i(t) &= \Phi_i(t) + \mathrm{phase}\{A + \alpha_i(t) \cdot \exp[j(\psi_i(t) - \Phi_i(t))]\} \\ &\approxeq \Phi_i(t) + \tan^{-1}\left\{\frac{\alpha_i(t) \cdot \sin[\psi_i(t) - \Phi_i(t)]}{A + \alpha_i(t) \cdot \cos[\psi_i(t) - \Phi_i(t)]}\right\} \\ &\approxeq \Phi_i(t) + \frac{w_i(t)}{A}, \end{aligned} \tag{9}$$

where

$$w_i(t) = \alpha_i(t) \cdot \sin[\psi'_i(t)], \tag{10}$$
$$\psi'_i(t) = \psi_i(t) - \Phi_i(t). \tag{11}$$

$\psi'_i(t)$ has a uniform distribution between $[0, 2\pi]$. So $w_i(t)$ follows a Gaussian distribution with zero mean and variance σ^2. We note that in the derivation of (9), we use the approximation $\tan^{-1}(x) \approx x$ under the assumption that $A \gg \alpha_i(t) \cdot \cos(\psi'_i(t))$ and $A \gg \alpha_i(t) \cdot \sin(\psi'_i(t))$. The approximations correspond to assuming either a medium or a high signal to noise ratio. Next, let f_s denote the sampling rate during the overlap interval $(i-1)T + \Delta t_i \le t \le iT$, the number of samples is given by:

$$N_s = (T - \Delta t_i) \cdot f_s. \tag{12}$$

Hence the average of the phase of the i-th tone during the overlap interval can be calculated by:

$$\begin{aligned} \bar{\Phi}_i &= \frac{1}{N_s} \cdot \sum_{k=1}^{N_s} \hat{\Phi}_i(\Delta t_i + (i-1)T + (k-1)/f_s) \\ &= 4\pi f_i \frac{R}{c} + 2\pi f_i \frac{v}{c}\left[\Delta t_i + (2i-1)T - \frac{1}{f_s}\right] + \frac{\bar{w}_i}{A}, \end{aligned} \tag{13}$$

where $\bar{w}_i$ is the average of the noise during the overlap interval for the i-th tone and follows a Gaussian distribution with zero mean and variance σ^2/N_s. Since Δt_i depends on the range and the speed of the target, it cannot be known in advance and it changes at every tone. To deal with this problem, we define a fixed start time for the sampling window of the return signal based on a pre-determined maximum range $R_{\max}$ so that:

$$\Delta t_{\max} = 2R_{\max}/c \ge \Delta t_i. \tag{14}$$

Using $\Delta t_{\max}$ instead of Δt_i, (13) becomes:

$$\begin{aligned} \bar{\Phi}_i(t) &= 4\pi f_i \frac{R}{c} + 2\pi f_i \frac{v}{c}\Bigg[\Delta t_{\max} + (2i-1)T \\ &\quad - \frac{1}{f_s}\Bigg] + \frac{\bar{w}_i}{A} = \tilde{\Phi}_i(t) + \frac{\bar{w}_i}{A}, \end{aligned} \tag{15}$$

where $\tilde{\Phi}_i(t)$ is the average of the phase without noise:

$$\tilde{\Phi}_i(t) = 4\pi f_i \frac{R}{c} + 2\pi f_i \frac{v}{c}\left[\Delta t_{\max} + (2i-1)T - \frac{1}{f_s}\right]. \tag{16}$$

2.2. Range and Speed Estimation

In this section, we present the algorithm for range and speed estimation. Let's first consider the case where noise is absent. We let $\bar{t}_i$ denote the average delay for the i-th tone,

$$\bar{t}_i = \frac{\tilde{\Phi}_i(t)}{2\pi f_i} = 2\frac{R}{c} + \frac{v}{c}\left[\Delta t_{\max} + (2i-1)T - \frac{1}{f_s}\right], \tag{17}$$

Using two different tones i and j we could calculate

$$\bar{t}_i - \bar{t}_j = 2v(i-j)T/c. \quad (18)$$

Then from (18) we could calculate the speed as:

$$v = \frac{c(\bar{t}_i - \bar{t}_j)}{2(i-j)T} = \frac{c}{2(i-j)T} \cdot \frac{f_j \cdot \tilde{\Phi}_i - f_i \cdot \tilde{\Phi}_j}{f_i \cdot f_j}. \quad (19)$$

However we should note that the phase measurement is wrapped. The phase constraint $\tilde{\Phi}_i \leq 2\pi$ must be imposed in order to avoid ambiguous estimation of speed and range. This requirement appears too stringent, because it would make the maximum unambiguous range be $\lambda/2$. Therefore, based on (16), we use the phase difference defined as follows:

$$\begin{aligned}\Delta\tilde{\Phi}_{ij} = \tilde{\Phi}_i - \tilde{\Phi}_j = 4\pi\frac{R}{c}(f_i - f_j) \\ + 2\pi\frac{v}{c}\left(\Delta t_{\max} - \frac{1}{f_s}\right)(f_i - f_j) \\ + 2\pi\frac{v}{c}T\left[(2i-1)f_i - (2j-1)f_j\right],\end{aligned} \quad (20)$$

which yields to a more relaxed constraint $\Delta\tilde{\Phi}_{ij} \leq 2\pi$. This makes the maximum range in case of $v = 0$ to be:

$$R_{\max} = c/[2(f_i - f_j)]. \quad (21)$$

Since (20) has two unknowns R and v, we need two different equations in order to solve the system. This implies that we need two sets of $\Delta\tilde{\Phi}_{ij}$ with two different $f_i - f_j$. It suffices to use three frequencies f_i, f_j and f_k to meet this requirement. Furthermore, to simplify the solution of the system, we choose two of the tones having equal frequencies, i.e., $f_i = f_j$. In this case (20) becomes:

$$\Delta\tilde{\Phi}_{ij} = \tilde{\Phi}_i - \tilde{\Phi}_j = 4\pi vTf_i(i-j)/c. \quad (22)$$

Thus the speed can be calculated as:

$$v = \frac{c(\tilde{\Phi}_i - \tilde{\Phi}_j)}{4\pi(i-j)Tf_i}. \quad (23)$$

In order to determine if the target is moving with a positive or a negative speed relative to the radar beam direction, we need to consider $\Delta\tilde{\Phi}_{ij}$ in the interval $[-\pi, \pi]$. Using the value of π to replace $\Delta\tilde{\Phi}_{ij}$ in (23) as the maximum difference in phase, we can calculate the maximum detectable speed:

$$v_{\max} = c/[4(i-j)Tf_i]. \quad (24)$$

Since the speed has been calculated using the tones i and j ($f_i = f_j$), the range can be derived from (20) using the tones j and k ($f_j \neq f_k$). In this case the phase has to be considered in the interval $[0, 2\pi]$. Thus we obtain

$$\begin{aligned}R = \frac{c(\tilde{\Phi}_j - \tilde{\Phi}_k)}{4\pi(f_j - f_k)} - \frac{v}{2}\left(\Delta t_{\max} - \frac{1}{f_s}\right) \\ \frac{v}{2}T\left[\frac{(2j-1)f_j - (2k-1)f_k}{f_j - f_k}\right].\end{aligned} \quad (25)$$

In reality, the noise on the measurement must be considered. It is straightforward to calculate the speed in the presence of noise. Replacing $\Delta\tilde{\Phi}_{ij}$ in (23) by $\Delta\bar{\Phi}_{ij}$, we obtain

$$\begin{aligned}\hat{v} = \frac{c(\bar{\Phi}_i - \bar{\Phi}_j)}{4\pi(i-j)Tf_i} = v + \frac{c}{4\pi(i-j)Tf_i} \\ \cdot\left(\frac{\bar{w}_i - \bar{w}_j}{A}\right) = v + \frac{c}{4\pi(i-j)Tf_i} \cdot \frac{\bar{w}_v}{A},\end{aligned} \quad (26)$$

where $\bar{w}_v$ follows a Gaussian distribution with zero mean and variance $2\sigma^2/N_s$, and it is defined as:

$$\bar{w}_v = \bar{w}_i - \bar{w}_j. \quad (27)$$

Similarly, by (25) we obtain the range estimation in the presence of noise as follows:

$$\begin{aligned}\hat{R} = \frac{c(\bar{\Phi}_j - \bar{\Phi}_k)}{4\pi(f_j - f_k)} - \frac{\hat{v}}{2}\left(\Delta t_{\max} - \frac{1}{f_s}\right) \\ - \frac{\hat{v}}{2}T\left[\frac{(2j-1)f_j - (2k-1)f_k}{f_j - f_k}\right].\end{aligned} \quad (28)$$

Since f_i is the carrier frequency, we can assume $f_i \approx f_j \approx f_k$ and (28) can be simplified as

$$\begin{aligned}\hat{R} &= R + \frac{c}{4\pi(f_j - f_k)} \cdot \left\{\frac{\bar{w}_j - \bar{w}_k}{A} - \frac{\bar{w}_v}{A}\right. \\ &\quad \left.\cdot\left[\frac{1}{2}\left(\Delta t_{\max} - \frac{1}{f_s}\right) \cdot \frac{f_j - f_k}{(i-j)Tf_i} + \frac{j-k}{i-j}\right]\right\} \\ &= R + \frac{c}{4\pi(f_j - f_k)} \cdot \frac{\bar{w}_R}{A},\end{aligned} \quad (29)$$

$$\begin{aligned}\bar{w}_R &= \bar{w}_j - \bar{w}_k - \bar{w}_v \cdot \left[\frac{1}{2}\left(\Delta t_{\max} - \frac{1}{f_s}\right)\right. \\ &\quad \left.\cdot \frac{f_j - f_k}{(i-j)Tf_i} + \frac{j-k}{i-j}\right].\end{aligned} \quad (30)$$

where $\bar{w}_R$ follows a Gaussian distribution with zero mean and variance:

$$\begin{aligned}\sigma^2_{\bar{w}_R} = \frac{2\sigma^2}{Ns}\left\{1 + \left[\frac{1}{2}\left(\Delta t_{\max} - \frac{1}{f_s}\right)\right.\right. \\ \left.\left.\cdot \frac{f_j - f_k}{(i-j)Tf_i} + \frac{j-k}{i-j}\right]^2\right\}.\end{aligned} \quad (31)$$

From (26) and (29), the error on the speed and range measurement due to noise can be calculated as:

$$v_{\text{err}} = \hat{v} - v = \frac{c}{4\pi(i-j)Tf_i} \cdot \frac{\bar{w}_v}{A}, \quad (32)$$

$$R_{\text{err}} = \hat{R} - R = \frac{c}{4\pi(f_j - f_k)} \cdot \frac{\bar{w}_R}{A}. \quad (33)$$

Thus, we obtain the expression for the error variance of the speed and range estimation, respectively as follows:

$$\sigma^2_{v_{\text{err}}} = \left(\frac{c}{4\pi(i-j)Tf_i}\right)^2 \cdot \frac{1}{Ns} \cdot \frac{1}{\text{SNR}} \quad (34)$$

$$\begin{aligned}\sigma^2_{R_{\text{err}}} &= \left(\frac{c}{4\pi(f_j - f_k)}\right)^2\left\{1 + \left[\frac{1}{2}\left(\Delta t_{\max} - \frac{1}{f_s}\right)\right.\right. \\ &\quad \left.\left.\cdot \frac{f_j - f_k}{(i-j)Tf_i} + \frac{j-k}{i-j}\right]^2\right\} \cdot \frac{1}{N_s} \cdot \frac{1}{\text{SNR}},\end{aligned} \quad (35)$$

where $\text{SNR} = A^2/(2\sigma^2)$. Equations (34) and (35) reveal that a large number of samples and a high SNR will reduce the error both on the speed and range estimation. A high carrier frequency and a long pulse width are necessary to reduce the noise variance on the speed. However, a reduction of noise variance comes with a trade off with the maximum detectable speed based on (24). Furthermore, (26) and (29) show that both the speed and the range are deterministic unknowns with additive Guassian noise of zero mean. Hence, the maximum likelihood estimator (MLE) is the optimum unbiased estimator to obtain the estimates for speed and range. So the error variances that are given in (34) and (35) achieve the classic Cramer-Rao lower bound [9].

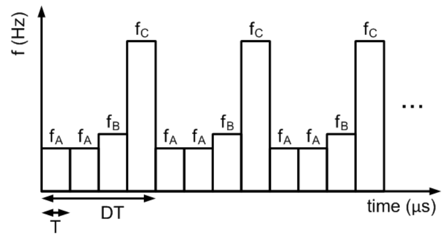

Fig. 2. Proposed sequence of frequency tones for the S-MFCW radar.

Table 1. Simulation Parameters, Results and Comparisons.

	S-MFCW	FMCW
Target Maximum Distance	150 m	112 Km
Target Maximum Speed	46 m/s	606 m/s
Target Initial Distance	120 m	120 m
Target Speed	-30 m/s	-30 m/s
Target Cross Section	3 m^2	3 m^2
Carrier Frequency	77 GHz	77 GHz
Δf_{AB}	1 MHz	N/A
Δf_{BC}	21 MHz	N/A
Bandwidth	**22 MHz**	**700 MHz**
Tone Period (T)	24 μs	N/A
Dwell Time (DT)	**96** μ**s**	**1.5 ms**
Samples (N_s)	23	N/A
FFT Points	N/A	2048
Sampling Rate (f_s)	1 MHz	3 MHz
SNR	15 dB	15 dB
$\sigma^2_{v_{\mathrm{err}}}$	$0.232\ (\mathrm{m/s})^2$	$0.23\ (\mathrm{m/s})^2$
$\sigma^2_{R_{\mathrm{err}}}$	$35.5\ \mathrm{cm}^2$	$35\ \mathrm{cm}^2$

3. EXAMPLE AND SIMULATION RESULTS

To demonstrate the performances of the proposed S-MFCW modulation, we use the design configuration and requirements of a 77 GHz automotive radar given in [3]. As derived in (21), a small frequency difference between the transmitted tones is required to obtain a long maximum unambiguous range. Meanwhile a large frequency difference is required to obtain a small variance of the range estimation error, as shown in (35). Hence the requirements for a maximum range and for a small variance of the range error are contradicting each other. This can be mitigated if a three-different-frequency (f_A, f_B and f_C) tone sequence is utilized as follows. f_B can be chosen to be very close to f_A, while f_C has to have a much larger separation from f_A and f_B. Then the range estimations derived from the two pairs (f_A, f_B) and (f_B, f_C) can be combined to obtain the superior performance: the first pair determines the maximum range and the second pair determines the variance of the range error. To estimate the speed of the target, two tones having equal frequencies are needed (see Section 2), so the proposed S-MFCW tone sequence is depicted in Fig. 2 and is repeated every time a measurement is completed. The resulting dwell time for one range measurement is $\mathrm{DT} = 4T$.

A Matlab simulation model is used to validate the proposed S-MFCW radar modulation and compare it to a typical FMCW radar as in [3]. For both radars the time domain simulation uses dense sampling to generate the high frequency waveforms, while uses the respective sample frequency (Table 1) for the baseband signals. The Phased Array Toolbox has been used to generate and receive the waveform signals as well as to emulate the propagation and reflection on the target. The scenario is taken from the long range automotive radar. First the FMCW radar has been simulated using the same system parameters as in [3] (see Table 1). The variance on the speed and range error over 500 measurements has been simulated to be respectively $0.23\ (\mathrm{m/s})^2$ and $35\ \mathrm{cm}^2$. Then we used the equations developed in Section 2 to obtain the S-MFCW parameters that would give the same variance on the speed and range errors (see Table 1). To obtain the desired result, first we estimate the speed using (26) on the pair of tones using the same frequency f_A. Then, two range estimations are calculated using (28). A coarse range (R_{coarse}) is determined by the close spaced frequency pair (f_A, f_B) of tones 2 and 3. While a fine range (R_{fine}) is determined by the far spaced frequency pair (f_B, f_C) of tones 3 and 4. Next, the integer n that solves the equation

$$n = \arg\min_n (R_{\mathrm{coarse}} - R_{\mathrm{fine}} - n \cdot R_{\mathrm{max},34}) \tag{36}$$

must be found, where $R_{\mathrm{max},34}$ represents the maximum range calculated using the 3-rd and 4-th tones. Finally the accurate target range can be found as:

$$R_{\mathrm{final}} = n \cdot R_{\mathrm{max},34} + R_{\mathrm{fine}}. \tag{37}$$

Using the presented sequence of tones and reconstruction method, the target can be detected at the maximum range set by the difference between f_A and f_B, with the error set by the difference between f_B and f_C. From the simulation results, the variance on the speed and range errors over 500 measurements are respectively $0.232\ (\mathrm{m/s})^2$ and $35.5\ \mathrm{cm}^2$. They are in good agreement with the expected analytical results of $0.23\ (\mathrm{m/s})^2$ and $35\ \mathrm{cm}^2$.

From simulation, we demonstrated that both S-MFCW and FMCW can detect targets that meet the long range automotive radar requirements. FMCW could detect targets even at a much longer distance (maximum distance in Table 1). But the S-MFCW modulation requires a much smaller bandwidth, a much shorter dwell time and does not need FFT. FFT is very area and power consuming. So not requiring FFT is a hardware advantage especially for large scale integration, like phased arrays. As shown in Table 1, S-MFCW uses a bandwidth 32 times smaller and it is 15.6 times faster than the FMCW radar. These advantages enable S-MFCW to be a viable modulation scheme for real time imaging radar applications that often require short dwell time and low hardware complexity.

4. CONCLUSIONS

In this paper we have proposed a novel modulation technique that allows fast radar detections with narrow bandwidth requirements. The new modulation method has been validated through Matlab simulations and compared to a FMCW radar. Simulation results demonstrate that the proposed S-MFCW modulation achieves significantly lower system requirements while providing comparable performance for range and speed estimation to a FMCW radar.

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