Portable Low-Power Fully Digital Radio-Frequency Direct-Transceiving See-Through-Wall Radar

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ABSTRACT A see-through-wall (STW) radar system based on a direct digital-logic radio-frequency (RF) pulse generator and an equivalent-time direct RF-sampling receiver chipset is proposed. Both the transmitter and the receiver of the radar system are both operated in the digital domain, considerably simplifying the design complexity and the difficulty of signal processing. In terms of data processing, we replace the conventional analog-domain mixing (self-mixing) and digital-domain mixing (self-mixing) modes by taking the absolute value to down convert the signal to the baseband. Thus, the mixer and the local oscillator are removed, which significantly reduces the power consumption while simplifying the signal processing. To verify the effectiveness of the proposed method, a relevant radar chipset was designed in Taiwan Semiconductor Manufacturing Company (TSMC) 65nm process. The prototype radar has a maximum transmission power of 17 dBm and a maximum equivalent sampling rate of 20 GS/s. The detection distance of the STW radar is 8 m with 30.0 cm thick wall, while the respiratory rate of the stationary life is measured through the wall.

INDEX TERMS See-through-wall radar, digital domain mixer, direct RF sampling, radar chipset.

I. INTRODUCTION
A STW radar system is intended for mapping building interiors and detecting static and moving objects hidden behind walls and other optically opaque obstacles. It can be used for a wide variety of purposes such as rescue missions, urban warfare, counter terrorism, non-contact medical monitoring, mine detection through ground penetration, and law enforcement [1]–[28]. In particular, during the current COVID-19 pandemic, a patient’s vital signs (such as the respiratory rate and the heart rate) are remotely monitored by a bioradar system and displayed on a monitor for the doctor to check the patient in a sterile environment, thereby avoiding the risk of contact with the patient and the consequent risk of infection.

These requirements and application scenarios require that STW and biodetection radar system have the characteristics of low-power, small-size, low-cost and long-endurance. However, existing STW and biodetection radars do not meet these requirements. At present, the radar systems considered for the abovementioned applications can be mainly divided into three categories. First, because of the difficulty and complexity associated with developing a custom STW system, existing commercial RF instruments such as oscilloscopes and network analyzers are used to build conceptual STW systems [20]. Although a conceptual system can provide the necessary experimental capability, it has the drawbacks of high-cost, low-portability, high-power, large-size, which cannot be used in real-time. Second, a low-cost STW platform composed of discrete components has become increasingly important, which may facilitate the development of various STW concepts or image-formation methods. The conventional implementation of a stand-alone, real-time ultra-wide-band STW platform is introduced in [16] [21]–[25]. Compared with the existing commercial RF instrument implementation, it has considerable advantages, but it still has low efficiency, complex implementation, high power consumption, lack of portability. Third, in order to further reduce power consumption, improve integration and support wearable applications, chip-level solutions have been proposed. Some new methods have been proposed in recent years [26]–[28], which sample the RF signals directly by using the front-end architecture of a comparator threshold scanning method. However, the accuracy degrades because
of the use of low sampling bits. As a result, these methods have a low scanning speed and thus their applications are limited. To solve the above-mentioned problems, in this article, we propose a novel wearable chip radar system with long-endurance.

In this study, the hardware system and the algorithm of the entire STW radar system are modified from the viewpoint of chip, laying a foundation for a wearable, long-endurance, array-based, and imaginable wall-penetrating radar system in the future. The rest of this article is organized as follows: The direct RF transceiver front-end architecture is analyzed and compared to the existing architecture in Section II. The STW system implementation, data acquisition, and data preprocessing are discussed in detail in Section III. Finally, the experimental results are presented in Section IV, followed by the conclusions in Section V.

II. DIRECT RF TRANSCEIVER FRONT-END ARCHITECTURAL

A. CONVENTIONAL IF SAMPLING SYSTEM

The principle conventional intermediate frequency (IF) sampling technique is shown in Fig. 1(a) [21]–[25]. This method is mainly implemented with multiple discrete devices. The analog front-end consists of two parts: the transmitter and the receivers. The transmitter is constructed with a pulse signal generator, a mixer, a filter, and a power amplifier (PA). The receiver mainly includes a low noise amplifier, a mixer, a low pass filter, an IF amplifier, and an Analog-Digital-Converter (ADC). The transceiver shares a local oscillator.

The working principle of the transmitter is as follows: The transmitter first generates a baseband signal whose spectrum is close to zero. Upon mixing with the local oscillator, a double-sideband RF signal is generated. Then band-pass filters are used to filter the signal out of band. Finally, through power amplification, the signal is transmitted by an antenna. The receiver works as follows: echoes from a target are received by the receiving antenna and then, amplified via the LNA. Because of the bandwidth limit, ADC cannot quantify an RF signal directly; therefore, a mixer is used to transform the RF signal into an IF signal, the resulting signal consists of a baseband signal and an RF signal. Then, the RF signal is filtered out by the low-pass filters, and the baseband signal is amplified by the IF amplifier and then quantized by ADC. Finally, the resulting digital signal can be easily used for the subsequent digital signal processing. This method is responsible for the high power consumption and large size of the entire system, which are not conducive to portable and long battery life applications.

B. EQUIVALENT-TIME RF SAMPLING METHOD

To reduce the impact of discrete devices on power consumption, performance, and portability, we propose a new transceiver architecture, as shown in Fig. 1(b).

This architecture adopts direct RF transmission and sampling. Therefore, the mixers and filters for the transceivers part in the analog domain can be eliminated.

The RF signals generated directly by the transmitters usually consist of four different patterns: (1) Gaussian pulse, (2) Gaussian pulse derivative, (3) frequency shifted Gaussian pulse, and (4) rectangular modulation pulse. For the Gaussian pulse, the frequency band of the UWB signal is generally spread near Direct Current (DC). Accordingly, large antenna size is required, which is not suitable for miniaturized applications. For Gaussian pulse derivative, it is easy to meet the radiation limitation of Federal Communications Commission (FCC) for the system transmitting the signals. However, in this case, even though the complexity of the
the corresponding modulated signal is obtained as follows

certain repetition frequency period and duty cycle, and then, a sinusoidal signal is controlled by a square wave signal with a pulse. Further, $\varepsilon(t)$ represents a step signal.

As can be seen from Eq. (1), the rectangular envelope modulates the amplitude of the sinusoidal signal. Like the frequency-shifted Gaussian pulse modulation, such a signal is also a UWB signal based on the carrier modulation. The position of the rectangular wave spectrum can be adjusted by the carrier frequency for different applications. Moreover, adjusting the rectangular wave duty cycle can control parameters such as the spectrum bandwidth. In this study, we use a rectangular modulation structure for the transmitter design.

When a rectangular modulated pulse is generated using a combinatorial logic, the design of the entire transmitter is considerably simplified, but the carrier frequency of the output signal is uncertain. We can’t use LO (Local oscillator) to demodulated the RF signal. And so, the conventional IF transceiver architecture is not suitable for this application. However, in the nanoscale CMOS process, direct RF sampling is possible, which greatly simplifies the design of the entire receiver. This enables direct RF transmission and RF sampling reception. If RF under-sampling is used directly, the full waveform of the transmitted pulse cannot be acquired, and hence, an equivalent sampling method is required to retain the full waveform of the acquired signal, and then perform the subsequent digital signal processing.

When the half the sampling rate of the ADC is less than the direct transmitted signal frequency, the RF signal is folded back to the first Nyquist zone. The signals from other Nyquist zones are also folded back to the first Nyquist zone. Therefore, it is necessary to introduce an anti-folding filter before the input of ADC to filter out the out-of-band noise. An equivalent sampling method with a half-Nyquist bandwidth higher than the RF signal frequency can replace the analog filter with a digital band-pass filter. Fig. 3 shows that the RF signal directly corresponds to the high/low sampling rate ADC of the Nyquist zone. For the STW radar system developed in this study, the equivalent Nyquist sampling method is a cost-effective technique to sample the input at a high sampling rate so that the entire 1–6 GHz band fall in the first Nyquist zone. A digital band-pass filter is able to improve Signal to Noise Ratio (SNR) and the resolution while reducing the need for anti-aliasing analog filters.

The direct equivalent RF sampling and quantization methods proposed in the application scenarios such as UWB short-range through-wall radar are shown in Fig. 1(b).

This chip design considerably simplifies the design of the STW radar, providing high flexibility, portability, and long battery life.

**C. DIGITAL-DOMAIN DOWN CONVERSION**

When the digital logic of the voltage-controlled delay method is used to generate the RF pulse waveform at the transmitter, instead of the traditional method which mixes the pulse with the local oscillator signal (shown in Fig. 4(a)), the carrier frequency of the transmitted signal cannot be estimated accurately. Consequently, the traditional method of mixing with the local oscillator in the analog domain is not appropriate for use in this study. For this problem, the self-mixing in the analog domain is proposed [29]. In this method, the local oscillator mixer is replaced with an input signal, as shown in Fig. 4(c). However, because the local oscillator is replaced by an input signal with a large dynamic range and a small swing, the input swing after mixing is significantly reduced, and the dynamic range of the entire receiving part is reduced.
as well. Figs. 4(b) and (d) show the basic methods of mixing for receiver in digital domain. Compared with the methods in analog domain as shown in Figs. 4(a) and (c), the corresponding local oscillator mixing and self-mixing can also be used in the digital domain, and thus the corresponding functions can be achieved. Although the above method is feasible, it is not optimal, and there is more flexibility for processing in digital domain. For example, down conversion can be directly realized by taking the absolute value in the digital domain shown in Fig. 4(e), but not that in the analog domain. Moreover, the band-pass filtering can be directly conducted in the digital domain without the use of a band-pass filter as in the analog domain, thereby decreasing the hardware cost considerably [16], [21]–[25].

III. STW RADAR SYSTEM IMPLEMENTATION, DATA ACQUISITION AND DATA PRE-PROCESSING

A. SYSTEM IMPLEMENTATION

The STW radar is usually realized by using discrete components and instruments [16], [21]–[25], as shown in Fig. 1(a). These devices are universal and applied widely, but they are primarily targeted toward universal applications with high power consumption and low efficiency. To overcome the above problems, discrete components are highly integrated into chipsets such as application specific integrated circuit (ASIC) in this study. The specific implementation is as follows. Concerning the transmitter, instead of using a pulse generator, a mixer, a local oscillator, a filter and a power amplifier (PA), we simply the architecture by using only two parts: the logic pulse generator and the PA, which are both integrated into one chip in TSMC 65nm process, as shown in Fig. 5(a). Further, the main parameters of the transmitter are shown in Table 1. The adjustable pulse width and the carrier frequency are directly generated by the logical gates, which led to the advantages of low power consumption, simple implementation, and less chip occupation. Moreover, the power-switchable and fast setting time mode is introduced in PA which only operated when a pulse is transmitted. It will greatly reduce power consumption, compared to transmitter in which PA always keeps operating. At a repetition period being 10MHz, the power consumption is only 22.5 mW, the output peak power is 17.6 dBm, and an area of 0.06 mm$^2$ is occupied.

Next the receiver is mainly consisted of a LNA, a variable-gain amplifier, ADC, a ring Phase Locked Loop (PLL), and an equivalent clock generator, as shown in Fig. 5. The parameters of the receiver are presented in Table 2, which is fabricated by the TSMC 65 nm process. The measured bandwidth of the receiver covers 0.5 GHz to 6 GHz, and the measured gains range from 20 dB to 40 dB while the 3-bit setting word changes from “000” to “111”. The measured minimum noise figure (NF) is about 6 dB at 6GHz and the maximum NF is approximately 8 dB at 6 GHz because of the small gain of RX when setting word is “000”. A 9-b direct RF sampling ADC is implemented by a time domain structure with a sampling bandwidth range from 500 MHz to 6 GHz and a power consumption of only 4 mW. The design flow of the entire receiver is simplified, and the entire signal conditioning process is transformed into the digital domain using the ADC, which considerably decreases the
power consumption. In addition, the equivalent sampling is realized by a PLL and an equivalent clock generator. In this way, an equivalent sampling rate of 20 GHz is achieved. The 3.0-6.0 GHz transmitted signal is contained within the first Nyquist zone, and the signal processing such as mixing and filtering can be implemented in the digital domain.

Fig. 6(a) shows the simulated model of the gain-enhanced Vivaldi antenna. This antenna is consisted of two parts: (1) an additional strip at the top, which is printed on both sides of the substrate; (2) a conventional Vivaldi antenna at the bottom. There is a slot between ant1 and ant2, which benefits decoupling. The total dimension of radar system is 16.2 × 8.75 cm², which is comparable to the size of a mobile phone. Plainly, our radar system is very portable and hand-held.

The power consumption of the entire front-end system is only 141.7 mW, which can be battery-powered and has long battery life. Fig.6 shows the transceiver test board and measured S-parameter of the gain-enhanced antenna. As shown in Fig.6, the antenna is well adapted throughout a wide bandwidth from 3.5 GHz to 10 GHz with |S11| > 10 dB, which includes the frequency range of interest (from 4 GHz to 6 GHz). This is attributed to the slot at the ground such that |S21| is greater than 20 dB across the frequency range from 3.5 GHz to 7.5 GHz, which indicates that this antenna has low coupling.

**FIGURE 6.** Transceiver test board and (b) measured S parameters of the gain-enhanced antenna.

**FIGURE 7.** Equivalent time direct-RF sampling operational principle of STW radar.

B. DATA ACQUISITION, PULSE ACCUMULATION, AND DATA PRE-PROCESSING

Equivalent time data acquisition is a low-cost, high sampling rate solution for obtaining full waveforms of pulse radar. Fig.7 shows the equivalent time sampling timing diagram that we used in this study. To reconstruct a complete narrow pulse, the equivalent time sampling scheme utilizes a PLL-based time-delay generator on the chip and requires a precise programmable time delay to shift the ADC sampling clock. The echoes of the 32 repetition periods are continuously acquired by the 625 MHz ADC and then stored in the RAM of the Field Programmable Gate Array (FPGA). The sampling clock keeps shifting 32 times by 50ps intervals.

**FIGURE 8.** Data acquisition and pre-processing flowchart.

**FIGURE 9.** Echo pulses represented along the slow time and fast time axis. The sine signal represents the slow time variation in phase.

**FIGURE 10.** 48-order band-pass filter spectral characteristics.
after every 32 pulses repetition interval until it traverses through the entire equivalent sampling period T. After that the pulse can be recovered through these sampling points with different phases. Finally, the entire data of the 512 repetition periods as one-dimensional distance information is obtained; and an equivalent sampling rate of 20 GS/s is obtained. The raw data collected by the FPGA is transmitted to the computer through the USB interface for data processing and analysis.

The digital signal processing flow chart is given in Fig. 8. We will introduce and analyze the content of each part in detail. Different from communication applications, the radar system processes a number of sweeps, \( n_P \), and the SNR is increased by a factor of \( n_P \) after coherent integration:

\[
SNR_{NS} = n_P \cdot SNR = n_P \cdot \frac{\sigma_{P_{TX}G_{TX}G_{RX}L_w}}{\frac{\sigma}{2^2}R^4} \cdot \frac{1}{F_{RX}kT_0B_{RX}} \cdot RX\ Noise
\]  

(2)

We assume that \( kT_0 \) is approximately \(-174\) dBm. The distance between the radar system and the target is 0.5-10 m. The carrier frequency is 5 GHz. Because of the direct equivalent sampling, the signal bandwidth \( B \) is 2 GHz. The loss at the 30 cm thick wall is approximately 20 dB. The gains \( G_{TX} \) and \( G_{RX} \) of the antennas are nearly 8 dBi. The output power reaches 17.7 dBm. The noise of the receiver is 6 dB. In this study, the accumulation period \( n_P \) is set to 16.

Echoes for each of the 16 repetition periods are directly accumulated. The echo pulses are represented along the slow time and fast time axises shown in Fig. 9, and the blue signal represents the slow time variation in phase. As the analog front-end of the direct RF equivalent sampling has no band pass filter, various communication interferences are mixed into the analog front-end.

As a result, a digital band pass filter is required for one-dimensional distance data filtering. Fig. 10 shows the band-pass characteristics and the raw data of the filter. Fig. 11 shows the original data and their spectrum. Figs. 12(a) and (b) show the filtered data and the corresponding spectrum, respectively. As can be seen from the figures, the digital band-pass filter achieves good interference and clutter suppression.

IV. EXPERIMENT USING DEVELOPED UWB PD RADAR

The experimental setup is illustrated in Fig. 14. The FPGA acquisition and control platform are connected to the front-end board through a custom cable. In fact, the FPGA can also be integrated into the front-end Printed Circuit Board (PCB), which further reduce the size of the entire STW radar system. In the experiment, the STW radar is placed in the corridor, and its transceiver antenna is close to a 30 cm thick brick wall. The targets in an 8-m wide room are illuminated through the wall. The FPGA controls simultaneously timing of the transceiver front-end to produce the equivalent 20 GS/s sampling rate data. The measured echoes are stored in the FPGA memory, and then transferred to a Personal Computer (PC) for the application of the Matlab inversion.
algorithms codes. In the following parts, the results will be described with reference to the analytical, simulated, and measured data.

A. RADAR RANGING RESOLUTION

To measure the range resolution, the radar prototype is used to acquire the range of a stationary target placed at different distances without wall. The acquired target location is then compared with another accurate relative location. Fig. 13(a) shows the measured mean value and the standard deviation (sigma) of the range at different distances. It can be observed that the sigma is changed from 0.0114 m to 0.0145 m. Furthermore, Fig. 13(b) presents the range resolution as a function of the distance to the radar. A range resolution of approximately 1.1 cm is achieved at a distance from 1 m to 8 m. Thus, the radar system in this study has a sufficient range resolution to test human breath. Next, we find the live targets behind the brick wall.

B. MICRO-DOPPLER SIGNATURES OF A WALKING PERSON

Any target motion can be highlighted by removing the corresponding static reflection signals (e.g., signals due to background and stationary objects). At different pulse repetition durations, the echoes from the wall and the background surroundings are nearly constant in terms of both the amplitude and the time delay; therefore, their values in different columns of a matrix are very close. In contrast, as the echoes from a moving target keep changing, their statistical characteristics differ. Background subtraction is based on an average estimation of the background clutter over some periods of measurement time, which is then subtracted from the future measurements for another period of time before the estimation is updated. The process can be described as follows: Take the average of the pulse signals at each pulse repetition interval (PRI), and subtract the average from each element in every PRI; thus, the new signal can be expressed as follow:

$$\hat{x}(m, n) = x(m, n) - \frac{1}{N} \sum_{n=0}^{N-1} x(m, n),$$

$$m = 0, 1, \cdots, M - 1, \quad n = 0, 1, \cdots, N - 1 \quad (3)$$

Let the two-dimensional matrix be an M × N matrix. Further, x(m,n) denotes the original data of point (m,n), and
FIGURE 16. Transient wave for respiratory rate test. In first phase of measurement the target under test is breathing; In the second phase the person under the test is holding voluntarily his breath; In the third phase the normal respiration process is restored.

M and N denote the rows and columns of the image matrix, respectively.

The STW radar system and the test scenario are shown in Fig. 14. In the experiment, the STW radar system is placed on one side of a 30-cm thick brick wall, and one person walking toward the radar in an 8 m wide room is illuminated through the wall. Fig. 15(a) shows a sampled range profile with an accumulation of 16 cycles. Obviously, the clutter from the wall and the transmitter is dominating, but the reflected echo of the human target is much smaller than the clutter so as not to be easily observed. The original range profile indicates the location of the wall but presents quite an obscure trajectory of the person walking before the wall As shown in Fig. 15(b), the background reflections are eliminated, so the trajectory of the human target moving back and forth from 0.3 m to 7 m becomes much obvious (as shown in Fig. 15(c)).

C. RESPIRATORY RATE MONITORING

The STW radar can help soldiers and rescuers see through a wall, judging a situation and determining the potential danger behind the wall. In disaster related applications, the STW radar system can help to find the signs of life under rubble and decide how to rescue the affected people after a disaster. Therefore, measuring the respiratory rate of a human target behind a wall is an important function of the STW radar system. In this study, the human target is set at one side of a 30-cm thick wall, and the radar system is set on the other side of the wall. Fig.16 shows that respiratory rate of this human target at two different statuses. In the first phase, the human target continues breathing and the respiratory rate is approximately 0.8 Hz. In the second phase, this person under test voluntarily holds his breath. In the third phase, the normal respiration process is restored. Note how the

STW radar system tracks the entire respiratory profile of the target. As shown in Fig.17, the respiratory rate of the human target seated normally behind the brick wall is measured. The respiratory rate of the human target is approximately 0.3 Hz, and the corresponding time domain waveform is shown in Fig. 17(a); Fig. 17(b) shows the frequency domain wave of the respiratory rate after Fast Fourier Transform (FFT).

Table 3 summarizes the performance of the radar system and compares it to previous works [21], [27], [30]. The proposed STW radar system integrated into a chipset in this study is superior to the compared discrete device architectures [21], [27] in terms of the power efficiency and the complexity. Compared with the system discussed in [30], the radar system proposed in this article has a higher output power level and bandwidth. The proposed system also has a better power consumption and chips area. Its penetration distance and scanning speed offer significant advantages over the other compared systems.
TABLE 3. Perfromance summary and comparison of STW radar.

| Parameter        | This work | [21] | [27] | [30] |
|------------------|-----------|------|------|------|
| Application      | STW       | STW  | Vital Signs +Ranging | UAV + SAR |
| Integration      | TX/RXF +ADC | Discrete devices | TX/RXF+ADC | TX/RX +ADC |
| RX Architecture  | Direct Sampling | Down sampling | Direct Sampling | Down Sampling |
| CMOS Technology  | 65nm | discrete devices | 55nm | 65nm |
| Center Frequency (GHz) | 3.2-5.8 | 0.4-0.72 | 7.3 & 8.8 | 15 |
| Pulse width (ns) | 0.75-1.2 | 3.125 | 1.8 | 1.48 |
| TX Power Level (dBm) | 17.46 | 30 | 8.6 | 13.3 |
| Power Consumption (mW) | 141.7 | >1000 | 118 | 260 |
| Die Area (mm²) | 0.97(TX) | 1.42(RX) | — | 8.6 | 4.06 |
| STW Range (m)@thick ness(m) | 8@0.3m | 5@0.19m | 9@ (no brick wall) | 110@ (no wall) |

V. CONCLUSION

An STW radar system based on direct digital-logic RF pulse generator and an equivalent-time direct RF-sampling receiver chipset is verified.

Section IV demonstrated that the proposed radar system in this article has a superior performance. A power consumption of 141.7 mW and an area of 2.39 (0.97 + 1.42) mm² are obtained. Therefore, the proposed radar system can realize the requirements of low-power, small-size, long-endurance. The measured results shown in Fig. 15 reveals that the STW radar can obtain the HRRP of the target behind the wall. Furthermore, Fig. 13 shows that a resolution of 1.1 cm is obtained. Further, the STW radar system can measure the live target’s respiratory rate. Finally, we test the human target’s respiratory rate and obtain the breath frequency to prove that the proposed STW radar system could be used in rescue missions, urban warfare, counter terrorism, non-contact medical monitoring, mine detection through ground penetration, and law enforcement scenarios.

The entire system test shows that the radar can not only achieve conventional through-wall radar applications, but also detect the vital signs of objects. Therefore, this miniaturized high-performance STW radar system provides effective support for military and civilian use.

REFERENCES

[1] L. Ren, N. Tran, F. Foroughian, and K. Naishadham, “Short-time state-space method for micro-Doppler identification of walking subject using UW impulse Doppler radar,” IEEE Trans. Microw. Theory Techn., vol. 66, no. 7, pp. 3521–3534, Jul. 2018.

[2] F. Qi, F. Liang, M. Liu, H. Lv, P. Wang, H. Xue, and J. Wang, “Position-Information-Indexed classifier for improved through-wall detection and classification of human activities using UW bio-radar,” IEEE Antennas Wireless Propag. Lett., vol. 18, no. 5, pp. 437–441, Mar. 2019.

[3] Y. Ding, X. Yu, J. Zhang, and X. Xu, “Application of linear predictive coding and data fusion process for target tracking by Doppler through-wall radar,” IEEE Trans. Microw. Theory Techn., vol. 56, no. 7, pp. 3941–3952, Jul. 2019.

[4] W. Wang, D. Wang, B. Zhang, T. Li, and S. Jiang, “Through-wall multi-target target identification in smart and autonomous systems with UWB radar,” IEEE Internet Things J., vol. 5, no. 5, pp. 3278–3288, Oct. 2018.

[5] K. Wang, Z. Zeng, and J. Sun, “Through-wall detection of the moving paths and vital signs of human beings,” IEEE Geosci. Remote Sens. Lett., vol. 16, no. 5, pp. 717–721, May 2019.

[6] X. Wang, G. Li, Y. Liu, and M. G. Amin, “Two-level block matching pursuit for polarimetric through-wall radar imaging,” IEEE Trans. Geosci. Remote Sens., vol. 56, no. 3, pp. 1–13, Mar. 2018.

[7] Y. Kim and H. Ling, “Through-wall human tracking with multiple Doppler sensors using an artificial neural network,” IEEE Trans. Antennas Propag., vol. 57, no. 7, pp. 2116–2122, Jul. 2009.

[8] A. A. Mostafa, C. Debès, and A. M. Zoubir, “Segmentation by classification for through-the-wall radar imaging using polarization signatures,” IEEE Trans. Geosci. Remote Sens., vol. 50, no. 9, pp. 3425–3439, Sep. 2012.

[9] S. S. Ram, C. Christianson, Y. Kim, and H. Ling, “Simulation and analysis of multiple Doppler radars in through-wall environments,” IEEE Trans. Geosci. Remote Sens., vol. 48, no. 4, pp. 2015–2034, Apr. 2010.

[10] M. Ash, M. Ritchie, and K. Chetty, “On the application of digital moving target indication techniques to short-range FMCW radar data,” IEEE Sensors J., vol. 18, no. 10, pp. 4167–4175, May 2018.

[11] B. Vandersmissen, N. Knuiddé, A. Jalalvand, I. Couckuyt, A. Bourdoux, W. De Neve, and T. Dhaene, “Indoor person identification using a low-power FMCW radar,” IEEE Trans. Geosci. Remote Sens., vol. 56, no. 7, pp. 3941–3952, Jul. 2018.

[12] G. Wang, C. Gu, T. Inoue, and C. Li, “A hybrid FMCW-interferometry radar for indoor precise positioning and versatile life activity monitoring,” IEEE Trans. Microw. Theory Techn., vol. 62, no. 11, pp. 2812–2822, Nov. 2014.

[13] F. Fioranelli, S. Salous, I. Ndjip, and X. Raimundo, “Through-the-wall detection with gated FMCW signals using optimized patch-like and vivaldi antennas,” IEEE Trans. Antennas Propag., vol. 63, no. 3, pp. 1106–1117, Mar. 2015.

[14] F. Fioranelli, S. Salous, and X. Raimundo, “Frequency-modulated interrupted continuous wave as wall removal technique in Through-the-Wall imaging,” IEEE Trans. Geosci. Remote Sens., vol. 52, no. 10, pp. 6272–6283, Oct. 2014.

[15] N. Maarief, P. Millot, C. Pichot, and O. Picon, “Ultra-wideband frequency modulated continuous wave synthetic aperture radar for through-the-wall localization,” in Proc. Eur. Microw. Conf. (EuMC), Rome, Italy, Sep. 2009, pp. 609–612.

[16] Y. Yang and A. E. Fathy, “Development and implementation of a real-time See-Through-Wall radar system based on FPGA,” IEEE Trans. Geosci. Remote Sens., vol. 47, no. 5, pp. 1270–1280, May 2009.

[17] Y. Wang, Q. Liu, and A. E. Fathy, “CW and pulse–Doppler radar processing based on FGPA for human sensing applications,” IEEE Trans. Geosci. Remote Sens., vol. 51, no. 5, pp. 3097–3107, May 2013.

[18] X. Zhang, X. Xi, M. Li, and D. Wu, “Comparison of impulse radar and spread-spectrum radar in through-wall imaging,” IEEE Trans. Microw. Theory Techn., vol. 64, no. 3, pp. 699–706, Mar. 2016.

[19] E. Schires, P. Georgiou, and T. S. Lande, “Vital sign monitoring through the back using an UWB impulse radar with body coupled antennas,” IEEE Trans. Biomed. Circuits Syst., vol. 12, no. 2, pp. 292–302, Apr. 2018.

[20] Y. Yang and A. E. Fathy, “Development and implementation of ultra-wideband See-Through-Wall imaging system based on sampling oscilloscope,” IEEE Antennas Wireless Propag. Lett., vol. 7, pp. 465–468, 2008.

[21] Y. Wang, Y. Yang, and A. E. Fathy, “A reconfigurable UWB system for real-time through-wall imaging applications,” in Proc. IEEE Radio Wireless Symp. (RWS), Santa Clara, CA, USA, Jan. 2010, pp. 633–636.

[22] P. H. Chen, M. C. Shastry, C. P. Lai, and R. M. Narayanan, “A portable real-time digital noise radar system for through-the-wall imaging,” IEEE Trans. Geosci. Remote Sens., vol. 50, no. 10, pp. 4123–4134, Oct. 2012.

[23] Q. Liu, Y. Yang, and A. E. Fathy, “A compact integrated 100 GS/s sampling module for UWB see through wall radar with fast refresh rate for dynamic real-time imaging,” in Proc. IEEE Radio Wireless Symp., Santa Clara, CA, USA, Jan. 2012, pp. 59–62.
Y. Wang, Q. Liu, and A. E. Fathy, “Simultaneous localization and respiration detection of multiple people using low cost UWB biometric pulse Doppler radar sensor,” in IEEE MTT-S Int. Microw. Symp. Dig., Montreal, QC, Canada, Jun. 2012, pp. 59–62.

Y. Wang and A. E. Fathy, “Advanced system level simulation platform for three-dimensional UWB through-wall imaging SAR using time-domain approach,” IEEE Trans. Geosci. Remote Sens., vol. 50, no. 5, pp. 1986–2000, May 2012.

H. A. Hjortland and T. S. B. Lande, “CTBV integrated impulse radio design for biomedical applications,” IEEE Trans. Biomed. Circuits Syst., vol. 3, no. 2, pp. 79–88, Apr. 2009.

N. Andersen, K. Granhaug, J. A. Michaelsen, S. Bagga, H. A. Hjortland, M. R. Knutsen, T. S. Lande, and D. T. Wisland, “A 118-mW pulse-based radar SoC in 55-nm CMOS for non-contact human vital signs detection,” IEEE J. Solid-State Circuits, vol. 52, no. 12, pp. 3421–3433, Dec. 2017.

D. T. Wisland, K. Granhaug, J. R. Pleym, N. Andersen, S. Stoa, and H. A. Hjortland, “Remote monitoring of vital signs using a CMOS UWB radar transceiver,” in Proc. 14th IEEE Int. New Circuits Syst. Conf. (NEWCAS), Vancouver, BC, Canada, Jun. 2016, pp. 4008–4011.

J.-Y. Kim and W.-Y. Choi, “30 GHz CMOS self-oscillating mixer for self-heterodyne receiver application,” IEEE Microw. Wireless Compon. Lett., vol. 20, no. 6, pp. 334–336, Jun. 2010.

Y. Wang, L. Lou, B. Chen, Y. Zhang, K. Tang, L. Qiu, S. Liu, and Y. Zheng, “A 260-mW Ku-band FMCW transceiver for synthetic aperture radar sensor with 1.48-GHz bandwidth in 65-nm CMOS technology,” IEEE Trans. Microw. Theory Techn., vol. 65, no. 11, pp. 4385–4399, Nov. 2017.

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