A Wall-Clutter Rejection Technique Using Two PLLs and a Phase Controller for Wall-Penetrating FMCW Radar

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***Abstract*— This letter proposes a novel frequency-modulated continuous-wave radar to achieve high wall-clutter rejection with a low-order high-pass filter (HPF). The proposed radar has two phase-locked loops (PLLs) and a phase controller. One transmitter PLL generates a chirp signal for transmitting (TX chirp) signal, the other a local-oscillator (LO) PLL generates a chirp signal for mixing (LO chirp), and the phase controller controls a phase of a reference clock entering the transmitter PLL. When the phase controller advances by a half-period of the reference clock, the PLL tracks and locks onto the advanced reference clock, and the TX chirp signal is advanced by a half-period of the reference clock. If longer advanced time is needed, the above-mentioned process (advance, track, and lock) is repeated after PLL locking. In this manner, long advanced time can be achieved with a fine time-resolution. The use of appropriate advanced time decreases the time gap between a wall-reflection signal and the LO chirp signal, and increases the ratio of target beat-frequency to wall beat-frequency. This enables a low-order HPF to fully attenuate wall-clutter. Moreover, this technique decouples the relationship between the wall's distance and the HPF's cut-off frequency. The radar was implemented and measured. The wall was located at 1.5 m and the target was located at 3 m. The measured results show that a second-order HPF attenuates by more than 20 dB for the wall beat-frequency signal, while it does not attenuate the target beat-frequency signal.**

***Index Terms*— Frequency-modulated continuous-wave (FMCW) radar, low-order high-pass filter (HPF), phase controller, two phase-locked loops (PLLs), wall-clutter rejection.**

I. Introduction

WITHIN recent years, wall-penetrating radar has gained attention from researchers [1]–[3]. Wideband frequency-modulated continuous-wave (FMCW) radars are good candidates for a high-resolution wall-penetrating detection system [2], [3]. The FMCW radar achieves high dynamic range by using a range-gating filter at intermediate frequency or baseband to fully eliminate wall-clutter. However, homodyne FMCW radars require a very high-order high-pass filter (HPF) to fully eliminate wall-clutter when a target

Manuscript received March 31, 2016; revised July 30, 2016 and November 26, 2016; accepted December 15, 2016. Date of publication February 17, 2017; date of current version March 3, 2017.

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Digital Object Identifier 10.1109/LGRS.2016.2643002

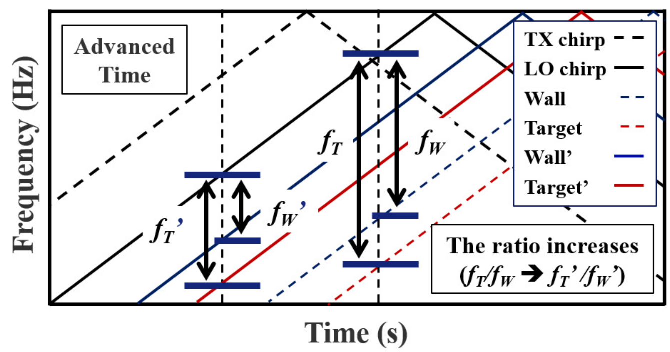

Fig. 1. TX chirp advanced effects. A TX chirp signal (black dotted line), an LO chirp (black solid line), a conventional wall signal (blue solid line), and a conventional target signal (red solid line), a wall signal with advanced TX chirp (blue dotted line), and a target signal with advanced TX chirp (red dotted line).

is located behind and in close proximity to the wall [4]. To overcome this problem, a delay-line is exploited in [4]; this delays a chirp signal, which enters a mixer local-oscillator (LO) port (LO chirp), and reduces the time gap between a received signal and the chirp signal. Consequently, the ratio of target beat-frequency to wall beat-frequency increases and a low-order HPF fully eliminates wall-clutter. It is equivalent to the transmitter chirp (TX chirp) being time-advanced, as shown in Fig. 1 (where f_T and f_T' are target beat-frequency, and f_W and f_W' are wall beat-frequency). Note that if the time gap is zero, then the wall beat-frequency is zero.

Delay-lines have been exploited in FMCW radars, to delay a radio frequency (RF) signal or LO signal [4]–[7]. However, delay-lines in an RF signal path or LO signal path invoke several problems. First, a conventional long delay-line makes considerable signal loss at a high frequency, such as X-band or Ka-band [8]. It additionally requires high-frequency amplifiers. Second, the line-loss also introduces amplitude modulation due to the loss, depending on frequency. Therefore, amplitude modulation increases as radar bandwidth increases, representing a particular problem for high-resolution radar. This amplitude modulation can lead to large side-lobes near a target beat-frequency [9]. Third, the delay-line requires a large volume to produce a long delay. For example, to achieve a delay of several hundred nanoseconds, a microstrip line of several meters long is required on a conventional substrate [8]; alternatively, a multilayer substrate with additional process or expensive process, such as surface acoustic wave process, is required [5]. Finally, it is difficult to achieve a controllable

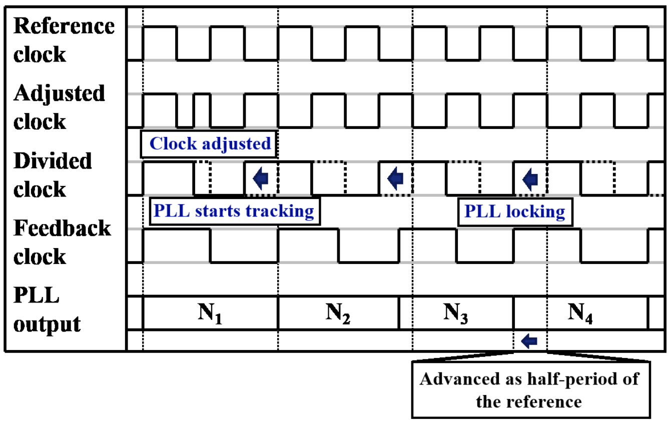

Fig. 2. Clocks and PLL response. The reference clock is advanced as half-period when a digital phase control function is activated ("clock adjusted"). Then, the PLL tracks and locks onto the time-advanced clock. It finally advances the time of the RF signals. In practice, the PLL requires some cycles for locking.

delay, from a short delay to a long delay, with a fine time-resolution. It requires abundant delay-lines, control circuits, and loss-compensate circuits, resulting in a bulky system.

In order to remove true-time delay-lines in RF or LO path, digital techniques can be used. These techniques create the original chirp and its replica entirely digitally. These radars have several advantages. The amplitude and the phase distortions occurring during the mixing process can be compensated by predistortion. And it can be easily controlled. However, to generate high frequency wideband chirp signals, it requires large power or additional blocks such as multipliers. It increases costs and complexity. Therefore, we propose a novel FMCW radar architecture that employs two phase-locked loops (PLLs) and a phase controller, as shown in Figs. 2 and 3. One PLL generates a chirp signal for transmitting (TX chirp), while the other PLL generates a chirp signal for mixing at the mixer (LO chirp). The PLLs share a reference clock, but the transmitter PLL input path includes a digital phase-controller. When a digital phase control function is activated, the controller advances the reference clock as a half-period by generating one more edge and inversing the following edges, as shown in Fig. 2. Each reference clock is divided by two and compared with the corresponding PLL's feedback clock. When the phase controller invokes a half-period advance, the transmitter PLL starts tracking. After some cycles, the PLL locks onto the advanced clock, producing a corresponding advanced time in the TX chirp. By repeating this process (advance reference, PLL tracking, and PLL locking), it is theoretically possible to produce an infinite time-difference. In practice, due to the finite period of the TX chirp and the LO chirp, the maximum time-difference is limited. This method solves all of the above problems: It does not result in any loss in RF or LO signals, nor produce any amplitude modulation including wideband FMCW radars; does not require greatly increased volume; and permits infinite time-delay with fine time-resolution. This method allows a low-order HPF to highly attenuate wall-clutter and also decouples the relationship between the wall's distance and the HPF's cut-off frequency. Sections II–IV provide a more detailed description and measurement results. Section II provides the design methodology, Section III provides measurement results, and Section IV presents the conclusions.

II. Design Methodology

In conventional homodyne FMCW radars that detect a target behind a wall, a very high-order HPF should be imposed to fully eliminate wall-clutter [4].

To moderate the HPF specification, a novel FMCW radar architecture is proposed, as shown in Fig. 3. One PLL (TX PLL) produces a chirp signal for transmitting (TX chirp) and a second PLL (LO PLL) produces a chirp signal for mixing (LO chirp). If there is a proper constant time-advance of the TX chirp compared to the LO chirp, then the time gap between the received signal and the LO chirp decreases. Consequently, the beat frequencies of the wall and of the target decrease, which are calculated as follows:

$$f_T' = \left(\frac{2R_T}{c} - \tau\right) C_R \quad f_W' = \left(\frac{2R_W}{c} - \tau\right) C_R \tag{1}$$

where f_T' is the target's beat-frequency, f_W' is the wall's beat-frequency, R_T is the target's distance, R_W is the wall's distance, c is the speed of light, τ is the time-advance, and C_R is the chirp rate. Note that when the time-advance equals the flight-of-time of the wall, $2R_W/c$, then the wall beat-frequency goes to zero and a simple first-order HPF or a dc block capacitor fully rejects the wall-clutter.

As explained above, the two PLLs and the phase controller allow for a specific time-shift. Note that time-shifting is quantized with a specific time-resolution. A higher reference frequency provides a higher time-resolution.

From the HPF standpoint, a high frequency is preferred to achieve a very fine time-resolution, in order to reduce the wall beat-frequency to zero or a very low frequency. For example, radar with a 25 000-GHz/s chirp rate detects a target located at 3 m, and a wall located at 1.5 m. If the reference frequency is 10 GHz, then the half-period is 50 ps. By repeating the half-period time-advance function, the TX chirp has a 10-ns time-advance. Then, the beat-frequency of the wall decreases to zero, and that of the target decreases to 250 kHz. Then, a first-order HPF fully attenuates the wall-clutter whereas it attenuates the target by less than 3 dB [10]. However, it is important to select an appropriate input reference frequency. Such a high frequency (10 GHz) increases the complexity of the design and implementation of components for phase-shifting, increases the costs and power requirement of components, and makes it difficult to achieve a stable, high-quality, low-cost reference source.

If the proposed technique is applied to a radar system positioned at a specific location and with a fixed radar-wall distance, the reference clock frequency is easily determined. Because the distance is fixed, the flight-of-time of the wall and the required time-advance can be precisely calculated. The time-resolution is chosen either by the equivalent of the calculated wall flight-of-time or the calculated value divided by an integer. The period of the reference clock is set to double the time-resolution. For example, if the radar-wall distance is 4.5 m, then the flight-of-time is 30 ns. The time-resolution can be selected as 30, 15, 7.5, and so on. For time-resolution of 15 ns, the period of the reference clock is 30 ns and the corresponding frequency is 33.333 MHz.

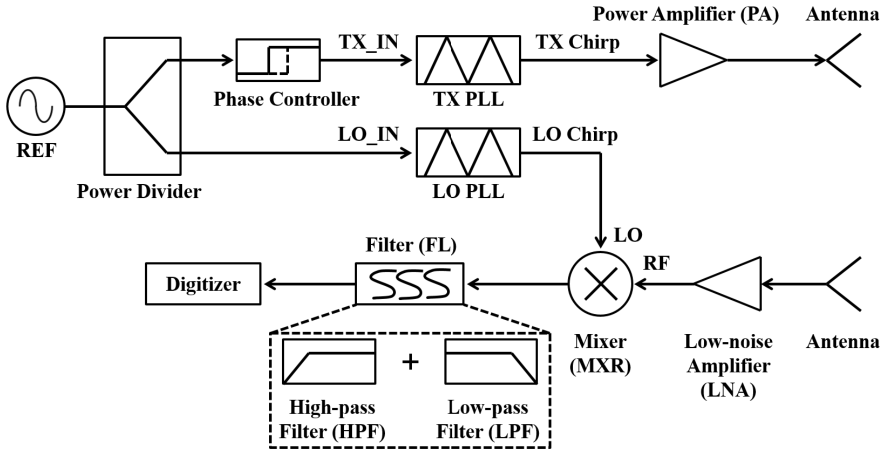

Fig. 3. Proposed FMCW radar architecture.

If the technique is applied to nonfixed radars, the reference clock frequency should be determined more carefully because the radar-wall distance is not fixed. For example, radar that uses a 50-MHz reference frequency can dramatically reduce the beat-frequency of a wall at 1.5 m but not at 1 m.

In order to achieve the lowest HPF order, we must maximize the ratio of target beat-frequency to wall beat-frequency, r, as calculated as

$$r = \frac{f_T'}{f_W'} = \frac{2R_T - \tau c}{2R_W - \tau c} \quad \text{for } 0 \le \tau \le \frac{2R_W}{c} \tag{2}$$

$$r = \frac{f_T'}{f_W'} = \frac{2R_T - \tau c}{\tau c - 2R_W} \quad \text{for } \frac{2R_W}{c} < \tau \tag{3}$$

where

$$\tau = T_{\text{RES}} N \tag{4}$$

where T_{RES} is the time-resolution and N is an integer. The time-advance is quantized because N is an integer.

Because (2) is a monotonically increasing function, the ratio increases as the time-advance increases unless the time-advance does not exceed the flight-of-time of the wall. Otherwise, the ratio decreases as the time-advance increases when the time-advance exceeds the flight-of-time of the wall. Therefore, N_O or $N_O + 1$ is the optimum value and they meet the following conditions:

$$r = \frac{T_{\text{FIX}} + \alpha}{\alpha}, \quad \text{when } N = N_O \tag{5}$$

$$r = \frac{T_{\text{FIX}} + \alpha - T_{\text{RES}}}{T_{\text{RES}} - \alpha}, \quad \text{when } N = N_O + 1 \tag{6}$$

where

$$0 \le \alpha < T_{\text{RES}}. \tag{7}$$

Therefore, the maximum achievable ratio, r_{MAX}, is calculated as follows:

$$r_{\text{MAX}} = \max\left(\frac{T_{\text{FIX}} + \alpha}{\alpha}, \frac{T_{\text{FIX}} + \alpha - T_{\text{RES}}}{T_{\text{RES}} - \alpha}\right) \tag{8}$$

where T_{FIX} is the wall-target distance. Equation (8) determines the time-resolution or the reference frequency when the HPF order, desired wall-attenuation, wall-target distance, and the bounds of the radar-wall distance have been set. For example, if the demanded ratio is above 4, the wall-target distance is 1.5 m, and the radar-wall distance is within the range of 3–4 m, then the time-resolution is determined to be 5 ns and this is because, this time-resolution always makes a ratio value greater than 4.2 in this case.

Although (8) determines the time-resolution or the reference frequency, it requires some calculation to find the optimum condition and reconfigurable HPF. The modified (9) and (10) allow us know the lower boundary of the achievable maximum ratio, $r_{\text{MAX\_LOW\_BOUND}}$, and determine a cut-off frequency, $f_{\text{CUT\_OFF}}$, of the HPF independently, wherever the wall is located

$$r_{\text{MAX\_LOW\_BOUND}} = \frac{T_{\text{FIX}}}{T_{\text{RES}}} \tag{9}$$

$$f_{\text{CUT\_OFF}} = T_{\text{FIX}} C_R. \tag{10}$$

Note that in order to design an HPF with conventional FMCW radars, the upper and lower bounds of radar-wall distance and target-wall distance must be predefined. However, to design an HPF with the proposed radar system, we only predefine a wall-target distance. For example, when detecting a target located 1.5 m behind a wall, in order to achieve the minimum ratio of 8, the time-resolution is set to 1.25 ns. The HPF cut-off frequency is set to 250 kHz when the chirp rate is 25 000 GHz/s. In the above-mentioned condition, a second-order HPF with 250-kHz cut-off frequency can attenuate wall-clutter by at least 30 dB and attenuate target signal by less than 3 dB in all cases where the wall-target distance exceeds 1.5 m.

Note that 400 MHz is a good candidate reference frequency: as it is not a particularly high frequency, many low-cost commercial components available, including phase controller, while the required filter order is only two, to attenuate wall-clutter more than 30 dB where a target located 1.5 m from a wall.

III. Measurement Results

A radar is implemented to verify the proposed technique. A reference clock is generated by an external signal generator.

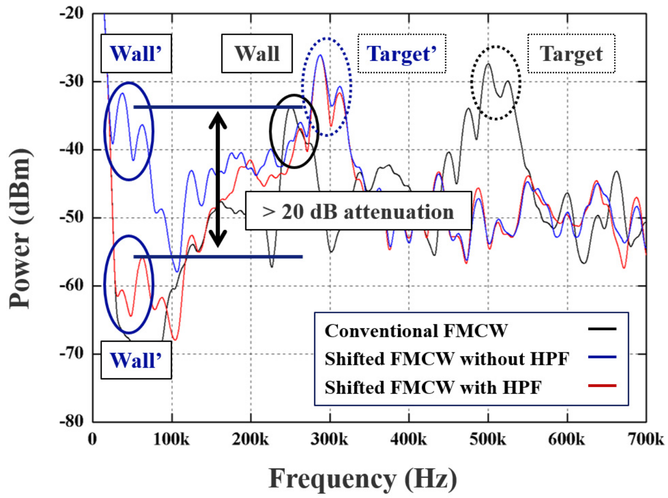

Fig. 4. Measured results. The wall is located at 1.5 m and the target is located at 3 m. Conventional FMCW radar beat frequency results (black curve), TX chirp time advanced FMCW radar without HPF beat frequency results (blue curve), and TX chirp time advanced FMCW radar with HPF beat frequency results (red curve).

The reference frequency is chosen as 400 MHz to achieve a 1.25-ns time-resolution. An HMC988LP3E chip [11] is exploited to advance the reference clock entering into a TX PLL. Two LMX2492 chips [12] are used to generate a 9–10-GHz chirp signal with 40 μs as rising and falling time. The generated TX chirp signal are amplified and emitted by an antenna. The emitted chirp signals propagate in free-space and are reflected by a wall and a target. The reflected chirp signals are amplified and entered into the mixer RF port.

When the LO chirp enters the mixer LO port, the mixer produces a beat-frequency corresponding to the time gap between the LO chirp and the received signal. The produced beat-frequency signal passes a second-order HPF with a 220-kHz cut-off frequency, and is measured by a spectrum analyzer (MS2830A [13]). Note that any low-pass filter (LPF) was not implemented in the final system for the test. However, the mixer and the spectrum analyzer act like the LPF. The mixer attenuates high-frequency signals such as the sum signal generated during the mixing process. And the spectrum analyzer eliminates out-range frequency signals.

The test was performed in a room. A wall and a target were located at the middle of the room. Before the measurements, the radar was calibrated. After calibration, the beat-frequencies were measured by the spectrum analyzer. After that, time-shifting was performed and the beat-frequencies were measured by the spectrum without HPF and with HPF. The measured results are shown in Fig. 4.

The wall was located at 1.5 m and the target was located at 3 m. The corresponding beat-frequencies are 250 and 500 kHz with the chirp rate 25 000 GHz/s. After radar calibration process, the measured wall beat-frequency was 250.8 kHz and the target beat-frequency was 500.2 kHz. The measured powers were −33.575 and −27.346 dBm, respectively. After the TX chirp time-shift, the beat-frequencies were measured without HPF and with HPF. In the case without HPF, the measured wall beat-frequencies were 37.4 and 62.6 kHz. The measured powers were −31.679 and −36.31 dBm, respectively. The target beat-frequency was 287.6 kHz and the power was −26.073 dBm. In the case with HPF, the measured wall beat-frequencies were 36.8 and 62.8 kHz. The measured powers were −60.623 and −55.842 dBm, respectively. The target beat-frequency was 287.6 kHz and the power was −26.075 dBm. Note that the wall beat-frequency signal is attenuated by more than 20 dB while the target beat-frequency signal is not attenuated by the second-order HPF.

IV. Conclusion

This letter proposes a novel FMCW radar that strongly attenuates wall-clutter with a low-order HPF by utilizing two PLLs and a phase controller. Additionally, this technique decouples the relationship between the radar-wall distance and the HPF's cut-off frequency. To design an HPF with the proposed system, we only predefine the wall-target distance when a chirp rate and desired wall attenuation are set. The proposed system allows for the use of radars in diverse environments. The measurement results show that a second-order HPF attenuates by more than 20 dB for a wall located at 1.5 m and does not attenuate for a target located at 3 m. The proposed radar is highly appropriate for wall-penetrating detection applications.

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