On the Sensitivity Degradation Caused by Short-Range Leakage in FMCW Radar Systems

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Abstract

Frequency modulated continuous wave (FMCW) radar systems suffer from permanent leakage due to their continuous operation. Especially in integrated circuits this leads to the well known issue of on-chip leakage due to limited isolation between transmit and receive circuitry. In addition, we investigate *short-range* (SR) leakage resulting from signal reflections of an unwanted close object located a few centimeters distant from the antennas. We carry out an in-depth analysis of the SR leakage and show that its residual phase noise in the intermediate frequency signal exceeds the total noise floor of the system, hence degrading the target detection sensitivity. We prove our analytical derivations with a complete FMCW radar system simulation.

1 Introduction

Automotive distance measurement and safety systems are typically realized with frequency modulated continuous wave (FMCW) radars. In contrast to pulse based systems, the FMCW principle uses a linear chirp sequence as transmit signal. The distance information is extracted by downconverting the reflected waves with the instantaneous transmit signal. For a single static object in the channel, this results in a sinusoid with constant frequency, which is termed *beat frequency*. It is proportional to the round-trip delay time (RTDT) of the radio waves and therewith also to the target distance.

The main advantage of the FMCW radar principle is that the instantaneous transmit power can be significantly reduced compared to pulse based systems. However, it suffers from permanent leakage of the transmit into the receive path. Especially in semiconductors this is an issue since isolation between transmitter and receiver circuitry is limited. The resulting *on-chip leakage* generates a beat frequency close to zero, which is why it is often also termed *DC-offset issue*. There is a vast literature on the leakage cancelation of such [1,2,3,4]. Several contributions analyze the impact of other non-idealities in FMCW radar systems, such as phase noise (PN) or the non-linearity of the chirp [5,6,7]. It is shown that these artifacts have a severe impact on the overall system performance.

Differently, in this work we consider a setup with a fixed target in front of the radar antennas, e.g. a fixture or cover, whose intermediate frequency (IF) impact is unwanted and possibly interferes with the IF of other targets. We analyze the effects of the *short-range* (SR) leakage on the IF signal spectrum and show that the *decorrelated phase noise* (DPN) of the SR leakage exceeds the additive white Gaussian noise (AWGN) from the channel, hence the detection sensitivity is degraded significantly. Additionally, we point out the difficulties arising when seeking for an SR leakage cancelation concept. Finally, we carry out a full FMCW radar system simulation to evidence our analytical derivations.

The paper is structured as follows. In Section 2 the system model including the on-chip and SR leakage is introduced. Then an analytical analysis on the impact of the SR leakage is given in Section 3. Finally, in Section 4 the simulation results are presented.

2 System Model

The FMCW radar system model including the on-chip and SR leakage is depicted in Fig. 1. The PLL generates the chirp over a bandwidth B and duration T that is used as transmit signal defined as

$$s(t) = A\cos\left(2\pi f_0 t + \pi k t^2 + \varphi(t) + \Phi\right), \tag{1}$$

for $t \in [0, T]$. The peak amplitude is A and the chirp start frequency is f_0, $k = \frac{B}{T}$ is the sweep slope, $\varphi(t)$ is the instantaneous PN and Φ is a constant initial phase.

The channel comprises of the unwanted SR leakage as well as the targets that are to be detected. The SR leakage is modeled with an RTDT τ_S and a reflection factor A_S. Equivalently, targets are modeled with τ_{Tm} and A_{Tm} for $m = 1, ..., M$, where M is the number of targets in the channel. The channel noise $w(t)$ is modeled as white Gaussian noise and added to the receive signal prior to amplification by the low noise amplifier gain G_L. Lastly, the on-chip leakage is modeled with an isolation factor A_L and a delay τ_L. Note that due to the physical setup we have $\tau_L < \tau_S < \tau_{Tm}$.

The receive signal is a superposition of the on-chip leakage, the SR leakage, the target reflections and the channel noise, that is

$$\begin{aligned} r(t) = &\underbrace{G_T A_L G_L\, s(t-\tau_L)}_{\text{On-chip leakage}} + \underbrace{G_T A_S G_L\, s(t-\tau_S)}_{\text{SR leakage}} \\ &+ \underbrace{\sum_{m=1}^{M} G_T A_{Tm} G_L\, s(t-\tau_{Tm})}_{\text{Target reflections}} + G_L\, w(t), \end{aligned} \tag{2}$$

where G_T is the transmission power amplifier gain. Since the reflected signal power decays steeply with the distance and the SR target is assumed to be only

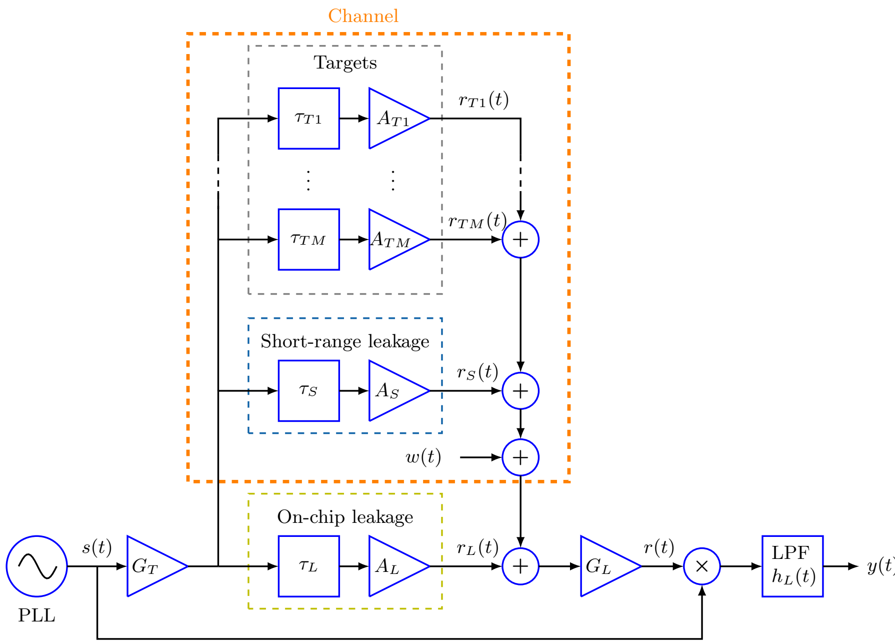

Fig. 1. System model with targets, on-chip leakage and SR leakage.

a few centimeters away from the radar antennas, A_S is in general significantly larger than A_L and A_{Tm}.

The receive signal is downconverted with the instantaneous transmit signal and lowpass filtered subsequently. Setting the initial phase $\Phi = 0$ for simplicity it is easy to show that the resulting IF signal is given as

$$\begin{aligned} y(t) &= [s(t)\, r(t)] * h_L(t) \\ &= \frac{A^2 G_T A_L G_L}{2} \cos\left(2\pi f_{BL} t + \Phi_L + \varphi(t) - \varphi(t-\tau_L)\right) \\ &+ \frac{A^2 G_T A_S G_L}{2} \cos\left(2\pi f_{BS} t + \Phi_S + \varphi(t) - \varphi(t-\tau_S)\right) \\ &+ \sum_{m=1}^{M} \frac{A^2 G_T A_{Tm} G_L}{2} \cos(2\pi f_{BTm} t + \Phi_{Tm} + \varphi(t) - \varphi(t-\tau_{Tm})) \\ &+ w_L(t), \end{aligned} \tag{3}$$

where $h_L(t)$ is the impulse response of a lowpass filter that eliminates the image originating from the mixing process, $f_{BL} = k\tau_L$, $f_{BS} = k\tau_S$, $f_{BTm} = k\tau_{Tm}$ are the beat frequencies and $\Phi_L = 2\pi f_0 \tau_L - k\pi\tau_L^2$, $\Phi_S = 2\pi f_0 \tau_S - k\pi\tau_S^2$, $\Phi_{Tm} = 2\pi f_0 \tau_{Tm} - k\pi\tau_{Tm}^2$ are constant phase terms. The respective channel noise in the IF domain is described as $w_L(t) = [s(t)\, G_L\, w(t)] * h_L(t)$.

Building on the introduced system model, an analysis of the SR leakage in time and frequency domain is given in the next section.

3 Short-Range Leakage Analysis

3.1 Time-domain Analysis

The SR leakage's IF signal from (3) is given as

$$y_S(t) = \frac{A^2 G_T A_S G_L}{2} \cos\left(2\pi f_{BS} t + \Phi_S + \varphi(t) - \varphi(t - \tau_S)\right). \tag{4}$$

Typically the gain and the beat frequency f_{BS} are evaluated for target detection and distance estimation. In contrast, the DPN $\Delta\varphi_S(t) = \varphi(t) - \varphi(t - \tau_S)$ is a noise term. Its name originates from the fact that with increasing target distance, $\varphi(t)$ and $\varphi(t - \tau_S)$ become more and more uncorrelated and the DPN increases. On the other hand, that is why for the on-chip leakage the DPN can be neglected as τ_L can be considered to be negligibly small.

Applying the cosine sum identity to (4) we obtain

$$\begin{aligned} y_S(t) = &\frac{A^2 G_T A_S G_L}{2} \cos(2\pi f_{BS} t + \Phi_S) \cos(\Delta\varphi_S(t)) - \\ &\frac{A^2 G_T A_S G_L}{2} \sin(2\pi f_{BS} t + \Phi_S) \sin(\Delta\varphi_S(t)). \end{aligned} \tag{5}$$

Since the DPN can be considered sufficiently small, we can approximate $\cos(\Delta\varphi_S(t)) \approx 1$ and $\sin(\Delta\varphi_S(t)) \approx \Delta\varphi_S(t)$ such that

$$\begin{aligned} y_S(t) \approx &\underbrace{\frac{A^2 G_T A_S G_L}{2} \cos(2\pi f_{BS} t + \Phi_S)}_{y_{S1}(t)} - \\ &\underbrace{\frac{A^2 G_T A_S G_L}{2} \sin(2\pi f_{BS} t + \Phi_S)\, \Delta\varphi_S(t)}_{y_{S2}(t)}. \end{aligned} \tag{6}$$

The first summand $y_{S1}(t)$ in (6) is the actual beat frequency signal while the second summand $y_{S2}(t)$ is a noise term caused by the DPN $\Delta\varphi_S(t)$. These two summands are individually depicted in Fig. 2. Therein, the system parameters were chosen according to a typical automotive application scenario, that is a transmit power of 0 dBm, $G_T = 10\,\mathrm{dB}$, $A_S = -8\,\mathrm{dB}$, $\tau_S = 1\,\mathrm{ns}$ and $G_L = 20\,\mathrm{dB}$. The PN is generated based on a typical PN power spectrum of a 77 GHz PLL. Note that in Fig. 2 the second summand is scaled for a better visibility of the DPN's effect. It can be observed that since the $\sin(\cdot)$ term is 90° phase shifted to the actual beat frequency signal, the second summand is largest at the zero-crossings and smallest for the peak amplitude of $y_{S1}(t)$.

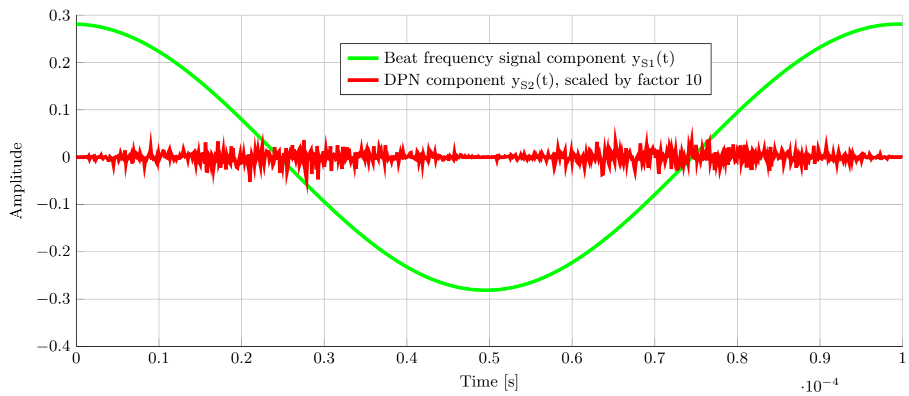

Fig. 2. Short-range leakage signal components in digital IF domain (approximation) over a single chirp for $\tau_S = 1\,\text{ns}$ ($d_S \approx 15\,\text{cm}$).

3.2 Frequency-domain Analysis

With the approximation from (6) the DPN term $\Delta\varphi_S(t)$ was extracted from the $\cos(\cdot)$ term. From the auto-correlation of the DPN, that is

$$c_{\Delta\varphi_S\Delta\varphi_S}(u) = \mathrm{E}\{\Delta\varphi_S(t)\Delta\varphi_S(t+u)\}, \tag{7}$$

its power spectral density (PSD) can be computed with the *Wiener-Khintchine-Theorem* and is well known to be [8]

$$S_{\Delta\varphi_S\Delta\varphi_S}(f) = 2\,S_{\varphi\varphi}(f)\,(1-\cos(2\pi f\tau_S)). \tag{8}$$

To determine the PSD of the overall error signal $y_{S2}(t)$ containing the DPN we compute its auto-correlation

$$\begin{aligned} r_{y_{S2}y_{S2}}(t,u) &= \mathrm{E}\left\{y_{S2}(t)\,y_{S2}(t+u)\right\} \\ &= \frac{(A^2G_TA_SG_L)^2}{4}\mathrm{E}\left\{\Delta\varphi_S(t)\,\frac{1}{2j}\left(e^{j(2\pi f_{BS}t+\Phi_S)}-e^{-j(2\pi f_{BS}t+\Phi_S)}\right)\right. \\ &\quad\left.\Delta\varphi_S(t+u)\,\frac{1}{2j}\left(e^{j(2\pi f_{BS}(t+u)+\Phi_S)}-e^{-j(2\pi f_{BS}(t+u)+\Phi_S)}\right)\right\} \\ &= \frac{(A^2G_TA_SG_L)^2}{16j^2}\mathrm{E}\{\Delta\varphi_S(t)\,\Delta\varphi_S(t+u)\}\left[-\left(e^{j2\pi f_{BS}u}+e^{-j2\pi f_{BS}u}\right)\right. \\ &\quad\left.+\left(e^{j(2\pi f_{BS}(2t+u)+2\Phi_S)}+e^{-j(2\pi f_{BS}(2t+u)+2\Phi_S)}\right)\right] \\ &= \frac{(A^2G_TA_SG_L)^2}{16j^2}c_{\Delta\varphi_S\Delta\varphi_S}(u)\left[-2\cos(2\pi f_{BS}u)\right. \\ &\quad\left.+\,2\cos(2\pi f_{BS}(2t+u)+2\,\Phi_S)\right]. \end{aligned} \tag{9}$$

From Fig. 3 and (9) it can be observed that $y_{S2}(t)$ is not a stationary process. Due to the chosen parameters for a chirp duration of $T = 100\,\mu$s the beat frequency signal of the SR leakage is evaluated over a single period in our example. The signal $y_{S2}(t)$ can be considered as one period of a cyclostationary process with the average auto-correlation as

$$\bar{r}_{y_{S2}y_{S2}}(u) = \frac{(A^2 G_T A_S G_L)^2}{8} c_{\Delta\varphi_S\Delta\varphi_S}(u) \cos(2\pi f_{BS} u), \tag{10}$$

since the last term in (9) vanishes due to the averaging process. Further, with $S_{\Delta\varphi_S\Delta\varphi_S}(f)$ from (8) the average PSD evaluates to

$$\begin{aligned} \bar{S}_{y_{S2}y_{S2}}(f) = \frac{(A^2 G_T A_S G_L)^2}{8} & \left[S_{\varphi\varphi}(f - f_{BS}) \left(1 - \cos(2\pi(f - f_{BS})\tau_S)\right) \right. \\ & \left. + S_{\varphi\varphi}(f + f_{BS}) \left(1 - \cos(2\pi(f + f_{BS})\tau_S)\right) \right]. \end{aligned} \tag{11}$$

Finally, the beat frequency f_{BS} is comparably small, that is 10 kHz in our example. Thus, the average PSD from (11) can be approximated well as

$$\bar{S}_{y_{S2}y_{S2}}(f) \approx \frac{(A^2 G_T A_S G_L)^2}{4} S_{\varphi\varphi}(f) \left(1 - \cos(2\pi f \tau_S)\right). \tag{12}$$

We use (11) to investigate the sensitivity degradation caused by the SR leakage. For that, the same system parameters as in Section 3.1 are used. The resulting average PSD of $y_{S2}(t)$ is depicted in Fig. 3. It can be observed that the DPN's power contribution exceeds that of the AWGN from the channel for frequency offsets larger than 200 kHz, which is the actual IF frequency range of interest. Consequently, the overall noise floor of the system is increased and the target detection sensitivity degraded.

4 Simulation Results

In this section we carry out a full FMCW radar system simulation based on Fig. 1 to evidence our analytical derivations. We consider a typical automotive radar application scenario where the bumper is a few centimeters distant from the radar antennas. The PLL has an output power of 0 dBm and ramps from 6 to 7 GHz, thus $B = 1$ GHz. Note that state of the art automotive radars operate at 77 GHz, however the reduced frequency is used solely for computational reasons and does not affect the results as the SR leakage is analyzed purely in the IF domain.

The on-chip leakage is assumed with an isolation of $A_L = -40$ dB and a delay $\tau_L = 10$ ps, while the SR leakage has a reflection factor of $A_S = -8$ dB and a delay of $\tau_S = 1$ ns (distance $d_S \approx 15$ cm). Further, a single target is considered within the channel at approximately 50 m distance. The transmission power amplifier and the LNA have a gain of $G_T = 10$ dB and $G_L = 20$ dB, respectively.

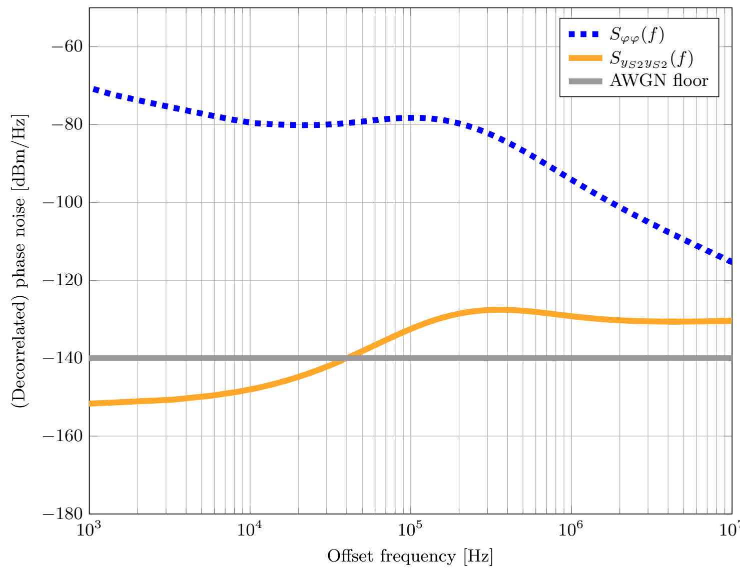

Fig. 3. Power spectral density of the PN and the DPN caused by the SR leakage for $d_S = 15\,\text{cm}$. It exceeds the AWGN noise floor from the channel for frequency offsets larger than 50 kHz, therewith degrading sensitivity of the FMCW radar.

As derived analytically in Section 3, the SR leakage's DPN exceeds the channel noise floor at $-140\,\text{dBm/Hz}$ in the IF domain. Consequently the target at 3.3 MHz cannot be resolved in the presence of the SR leakage. Also, cancelation of the beat frequency signal only, that is $y_{S1}(t)$ in (6), does not improve the sensitivity as the high-frequent noise remains in the IF signal.

Conclusion

In this work we investigated the sensitivity degradation caused by an SR leakage in an automotive FMCW radar transceiver. The decorrelated phase noise raises the total noise floor of the system and consequently limits the sensitivity of the radar. A full FMCW radar system simulation is employed to evidence the analytical derivations. For future work cancelation of the unwanted SR signal reflection is aspired, however, for that the instantaneous PN or DPN in the time-domain would need to be known.

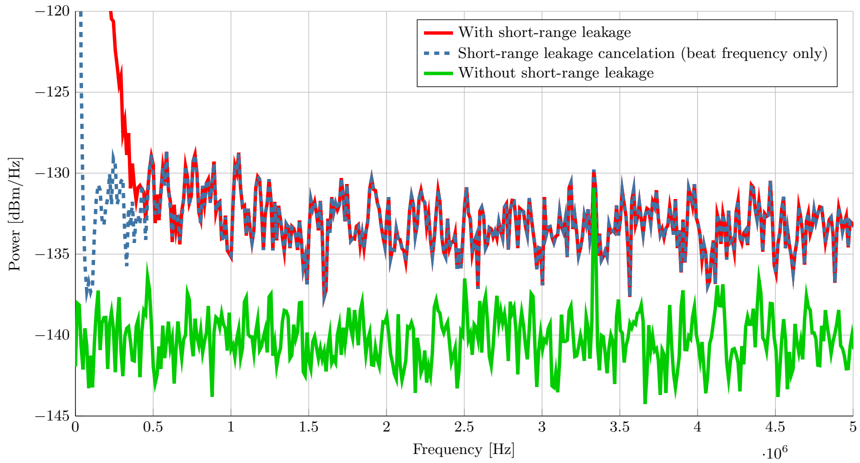

Fig. 4. Estimated power spectral density of the lowpass filtered IF signal. With the SR leakage the noise floor is increased and thus the target at 3.3 MHz is covered in noise.

References

1. Kaihui Lin, Yuanxun Ethan Wang, Cheng-Keng Pao, Yi-Chi Shih, “A Ka -Band FMCW Radar Front-End With Adaptive Leakage Cancellation,” In *IEEE Trans. on Microwave Theory and Techniques*, Vol. 54, No. 12, pp. 4041-4048, December 2006.
2. K. Lin and Y. Wang, “Transmitter Noise Cancellation in Monostatic FMCW Radar,” In *the IEEE/MTT-S International Microwave Symposium Digest, 2006*, pp.1406-1409, June 2006.
3. Jeiyoung Lee et al., “A UHF Mobile RFID Reader IC with Self-Leakage Canceller,” In *Proceedings of the Radio Frequency Integrated Circuits Symposium (RFIC’ 2007)*, pp. 273-276, Honolulu, June 2007.
4. A.G. Stove, “Linear FMCW radar techniques,” *In the IEE Proceedings F, Radar and Signal Processing, 1992*, Vol. 139, No. 5, pp. 343-350, October 1992.
5. K. Thurn and R. Ebelt and M. Vossiek, “Noise in Homodyne FMCW radar systems and its effects on ranging precision,” In *International Microwave Symposium Digest (IMS’ 2013)*, pp. 1-3, Seattle, USA, June 2013.
6. M. Pichler, A. Stelzer, P. Gulden, C. Seisenberger and M. Vossiek, “Phase-Error Measurement and Compensation in PLL Frequency Synthesizers for FMCW Sensors – I: Context and Application,” In *IEEE Trans. on Circuits and Systems I: Regular Papers*, Vol. 54, No. 5, pp. 1006-1017, May 2007.
7. C. Wagner, A. Stelzer and H. Jäger, “PLL Architecture for 77-GHz FMCW Radar Systems with Highly-Linear Ultra-Wideband Frequency Sweeps,” In *International Microwave Symposium Digest (IMS’ 2006)*, pp. 399-402, San Francisco, USA, June 2006.
8. M.C. Budge, Jr. and M.P. Burt, “Range correlation effects in radars,” In *Record of the 1993 IEEE National Radar Conference*, pp. 212-216, Lynnfield, USA, 1993.