Design of a wearable compact frequency-modulated continuous-wave radar

Olivier Caytan

Supervisors: Prof. dr. ir. Hendrik Rogier, Prof. dr. ir. Jan Vanfleteren

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Master's dissertation submitted in order to obtain the academic degree of Master of Science in Electrical Engineering

Department of Information Technology
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Olivier Caytan, August 11, 2014
Preface

The fact that this master’s dissertation is lying in front of you is not merely my own merit. I would like to take the opportunity in this preface to acknowledge all those who have made this work possible.

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Abstract

The design of a compact frequency-modulated continuous-wave (FMCW) radar system, capable of detecting and measuring the range and radial speed of human targets is discussed. The system operates in the 2.4GHz industrial, scientific and medical (ISM) radio band and relies on textile antennas. Both the compactness of the designed printed circuit board (PCB) and the application of textile materials, provide an outlook towards full textile integration of the radar system. This will lead to a wearable and very low-weight radar system, which will not hinder the operator. Such a radar will be practical in harsh conditions and demanding situations such as in burning buildings or on the battlefield. The possibility to detect humans hidden by smoke or fog, or even behind doors, offers interesting applications for numerous rescue, security and military purposes. Different measurement setups indicate that the designed radar system can successfully detect and measure human targets, as intended.

Keywords

Frequency-modulated continuous-wave radar (FMCW); Industrial, scientific and medical (ISM) radio band; compact; wearable
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Abstract—The design of a compact frequency-modulated continuous-wave (FMCW) radar system, capable of detecting and measuring the range and radial speed of human targets is discussed. The system operates in the 2.4GHz industrial, scientific and medical (ISM) radio band and relies on textile antennas. Both the compactness of the designed printed circuit board (PCB) and the application of textile materials for the antennas, provide an outlook towards full textile integration of the radar system. This will lead to a wearable and very low-weight radar system, which will not hinder the operator. Such a radar will be practical in harsh conditions and demanding situations such as in burning buildings or on the battlefield. The possibility to detect humans hidden by smoke or fog, or even behind doors, offers interesting applications for numerous rescue, security and military purposes. Different measurement setups indicate that the designed radar system can successfully detect and measure human targets, as intended.

Keywords—Frequency-modulated continuous-wave radar (FMCW); Industrial, scientific and medical (ISM) radio band; compact; wearable

I. INTRODUCTION

The purpose of a radar system (acronym for RAdio Detection And Ranging) is to detect the presence of reflecting targets in the environment and to measure their range and speed by means of electromagnetic radiation [1]. Range can be calculated by measuring the amount of time it takes for the radiation to propagate from the radar system antenna to the radar target and back. This round trip time (RTT) is related in a simple way to the range via the propagation speed. The radial speed can be measured using the Doppler effect. Electromagnetic radiation reflected by an object that moves in the direction of propagation of this electromagnetic radiation undergoes a frequency shift. This frequency shift is related in a simple way to the radial speed of the target.

The goal of this master’s dissertation, is to build a compact radar system operating in the 2.4GHz industrial, scientific and medical (ISM) radio band, capable of detecting and measuring the range and radial speed of human targets. The system has to operate in the frequency band from 2.4GHz to 2.5GHz, and the equivalent isotropically radiated power (EIRP) has to be limited to 20dBm. Moreover, as a step towards full textile integration of the radar system, the use of textile antennas is derived.

Such a radar system has numerous potential applications [2]. As the textile antennas can be integrated into the radar operator’s clothing, and the radar PCB is made very compact, the radar system will be wearable and very low-weight, and the operator will therefore not be hindered by the radar system. The wearer’s senses are enhanced by the radar system with minimal wearing discomfort. These properties make the use of this radar very practical in harsh conditions and demanding situations such as in burning buildings or on the battlefield. The possibility to detect humans hidden by smoke or fog, or even behind doors, are interesting applications for numerous rescue, security and military purposes.

II. RADAR SYSTEM ARCHITECTURE

First, an exploration of the different radar system architectures is conducted, in order to select the most suitable one to achieve the formulated goal. The adopted architecture is the linear frequency-modulated continuous-wave (linear FMCW) radar system [3]. Linear FMCW radar systems continuously transmit a linearly frequency-modulated (FM) signal. This type of radar system is ideal for low power and short range applications and leverages a very compact construction. Furthermore, FMCW systems are able to measure both target range and radial velocity. The signal processing is simplified because of the frequency modulation linearity.

A. Block diagram

The signal processing block diagram of the FMCW radar system is depicted in Fig. 1. An FM signal is generated and transmitted by an antenna (TX antenna). The expression for the instantaneous frequency $f_{TX}(t)$ of the transmitted signal is given by

$$f_{TX}(t) = f_c + \Delta f \cdot \gamma(t), \quad (1)$$

with $f_c$ the carrier frequency of 2.45GHz, $\Delta f$ the frequency swing of 100MHz and $\gamma(t)$ the linear symmetrical triangle frequency modulation function. This is a periodic function (modulation period $T_m$) of which a single period is shown in Fig. 2. This frequency modulation function will lead to an easy measurement of the target radial speed and range at the same time.

A radar target at range $R$ is illuminated by the TX antenna and returns a weak echo signal to the receive antenna (RX antenna). The received signal is an attenuated (can be calculated with the radar equation [1]) and delayed (propagation delay $\tau = 2 \cdot R/c$) version of the transmitted signal. As a consequence of the frequency modulation and propagation delay, the received signal has a frequency that is different from the transmitted signal. The dependence of the frequency difference on the target range and radial velocity constitutes the fundamental principle of FMCW radar systems.

The received signal and a sample of the transmitted signal are mixed, resulting in a signal that consists of a high-frequency term around 4.9GHz, which is useless and is filtered away by the subsequent band pass filter, and a baseband term called the converted signal. The power spectrum of the converted signal contains the range and radial velocity information of the radar target. This information is extracted by the signal processing circuit.
B. Converted signal processing

To measure the range and radial velocity \( v_{rad} \) of a target at time \( t_0 \), a spectral estimation of a signal fragment of the converted signal, centred at time \( t_0 \) is performed.

Considering the low speed and acceleration of human targets, the signal fragment length can be chosen adequately short so that both \( \tau \) and \( v_{rad} \) vary negligibly, while it is still adequately long to retain a good spectral estimation. Under these assumptions, the estimated power spectrum is discrete with frequency components at \( |k \cdot f_m + f_d| \) and \( |k \cdot f_m - f_d| \), with \( k \) an integer, \( f_m = 1/T_m \) the modulation frequency and \( f_d \) the Doppler shift. The Doppler shift \( f_d \) is given by

\[
f_d = f_c \cdot \frac{2 \cdot v_{rad}}{c}.
\]

(2)

Specifically in the case of a linear symmetrical triangle frequency modulation function, the spectral components at \( |k \cdot f_m \pm f_d| \), with \( k \) an integer, are equal in power.\footnote{Taking all this into consideration, the modulation frequency was limited on the modulation frequency is set by the assumption that \( \tau \ll T_m \), during the analysis of the converted signal spectrum.}

Assuming \( T_m \gg \tau \), it can be shown that the spectral envelope of the converted signal power spectrum exhibits a maximum around the frequency \( f_{max} \) given by

\[
f_{max} = 2 \cdot \Delta f \cdot \frac{\tau}{T_m}.
\]

(3)

A measurement of the target range can be acquired by determining the spectral component pair with maximal power, following (3). The frequency separation of this component pair is equal to twice the Doppler frequency. Equation (2) then yields the target’s radial speed. Note that it is impossible to determine whether the target is approaching or receding.

If there are two radar targets present, the converted signal spectrum will be equal to the sum of the converted signal spectra for each target alone. The range resolution is defined as the minimal range difference required between two targets such that they are still distinguishable by the radar system. The range resolution is inversely proportional to the used bandwidth \( \Delta f \).

C. Degradation effects

Two major degradation effects limit the performance of linear FMCW radar systems [4].

As indicated in Fig. 1, the receiver always directly receives a fraction of the transmitted power through TX/RX antenna leakage (and in a lesser extent through crosstalk). First, the strong directly leaked signal (generally the strongest signal at the receiver input) may overload the receiver. A second threat the directly leaked signal poses, is reduced sensitivity by masking of the radar targets by transmitter phase noise.

As the directly leaked signal has a very short delay, it appears as a series of strong low frequency spectral components in the converted signal power spectrum. The band pass filter of Fig. 1 can be used to suppress these spectral components.

Non-linearities in the frequency modulation cause a second performance degradation. Frequency modulation non-linearities degrade the range resolution of the radar system. Typically, the degradation increases with target range, and the range resolution is determined by the used bandwidth at low target ranges, and by the modulation linearity at high target ranges.

D. Modulation frequency

Several influences have to be taken into account when choosing the modulation frequency. A higher modulation frequency will increase the speed that can be reliably measured by the system and increase the time resolution of the system. An upper limit on the modulation frequency is set by the assumption that \( \tau \ll T_m \) during the analysis of the converted signal spectrum. Taking all this into consideration, the modulation frequency was chosen to be 10kHz.

III. DESIGN

Next, the FMCW radar system is realised with electronic hardware. All selected components are compact surface-mounted devices (SMDs).

A. Circuit design

A schematic block diagram of the designed electronic circuit is shown in Fig. 3.

The FM signal generator is implemented as a voltage controlled oscillator (VCO) with the tuning voltage controlled by a triangle wave oscillator. The VCO is selected on tuning characteristic linearity, output power, frequency band and phase noise.

![Fig. 1: Block diagram of the frequency-modulated continuous-wave radar](image-url)

![Fig. 2: One period of the symmetrical triangle frequency modulation function](image-url)
The linear triangle wave oscillator is realised with an operational amplifier (op-amp) circuit. This is one of the most simple and least linear implementation options, but it was expected that the acquired frequency modulation would be linear enough. This proved to be wrong as measurements on a prototype board indicated that the non-linearity might dominate the range resolution starting from a range of only 7.26m.

The signal generated by the VCO is applied to a power splitter. Half of the power generated by the VCO is amplified by the transmit amplifier (TX amplifier) and applied to the TX antenna. The TX amplifier is selected mainly on gain, frequency band and 1dB output compression point (well above 20dBm). The gain of the TX amplifier brings the power applied to the TX antenna close to 20dBm.

The other half of the power generated by the VCO is used to drive the local oscillator port (LO port) of the frequency mixer. We select a frequency mixer that can be driven with the power that is available. Other considered component properties are the frequency ranges, mixer conversion loss (as low as possible) and the radio frequency port (RF port) 1dB compression point (as high as possible).

The receive amplifier (RX amplifier) is chosen to be a cascade of a low noise amplifier (LNA) and a second power amplifier. The receive amplifiers will unavoidably amplify the directly leaked signal, and care has to be taken that the power of the amplified directly leaked signal stays well below the 1dB RF compression point of the frequency mixer. An isolation of 50dB between the TX and the RX antenna is assumed.

The chain formed by the two RX amplifiers and the frequency mixer has a low noise figure of 0.552dB, which will enhance the sensitivity of the system.

B. Printed circuit board design

A printed circuit board (PCB) is designed for the discussed circuitry. The PCB consists of two 18µm copper layers, of which one is a dedicated ground plane. The PCB substrate between the two copper layers is the low-loss I-Tera MT material of Isola Corporation (dielectric constant $\epsilon_r = 3.30$ and loss tangent $\tan\delta = 0.0036$) with a thickness of 0.5mm. This substrate is suited for high frequency applications up to 20GHz.

The PCB was designed to be as compact as possible (resulting PCB area 85mmx35mm). This compactness limits the losses, but might increase the crosstalk. Two design decisions were made to limit the amount of crosstalk. First, the on-board transmission lines are designed as grounded coplanar waveguides (GCPWs). The isolation between two GCPWs is usually higher than between two microstrips. Second, the transmitter transmission lines are routed perpendicularly to the receiver transmission lines.

EM circuit co-simulation of the PCB layout and component models performed with the Advanced Design System (ADS) software of Agilent shows that the losses are low and that the transmitter/receiver leakage due to PCB crosstalk is negligible compared to the transmitter/receiver leakage due to antenna leakage.

The board has one input and two outputs (SubMiniature version A connectors). The RX antenna is connected to the board RX input, and the TX antenna is connected to the board TX output. The other board output is connected to the intermediate frequency port (IF port) of the frequency mixer and is called the board IF output.

C. Antenna design

The TX and RX antennas are designed to exhibit a different circular polarisation. The orthogonal polarisation of the TX and RX antenna can lead to increased isolation between the two antennas and reduces multipath effects through rejection of second order reflections [2]. Circularly polarised rectangular ring topology patch antennas with a single coaxial feed are used [5]. The conducting patch and ground plane are fabricated with Flectron material, a copper-plated nylon fabric. The substrate between the patch and ground plane is black foam provided by Javaux ($\epsilon_r = 1.495$, $\tan\delta = 0.0168$) with a thickness of 4mm. The applied materials are flexible and light-weight and can be integrated into garments. The antenna pair is constructed and fixed on cardboard. The antennas are adequately separated (approximately 30cm separation between patch antenna centres) to achieve an isolation of approximately 50dB in the 2.4GHz ISM band.

IV. Validation

Tests with a fabricated prototype are performed in an anechoic chamber to validate the functionality of the radar system. The board IF output is connected to a spectrum analyser. First, the converted signal power spectrum is measured in the absence of targets (blank test). The result is shown in Fig. 4. Then, the converted signal power spectrum is measured when a human is walking at a normal speed in front of the radar system antennas at a distance of about 3m. The measured converted signal power spectrum is shown in Figs. 5 and 6.

The spectral envelope of the power spectrum depicted in Fig. 5 shows maxima around 20kHz and 50kHz. The maximum around 20kHz is already observed in the blank test and is caused by the directly leaked signal. The power of the spectral components around 50kHz is increased compared to the blank test, which leads to the detection of a radar target.

The maximum around 50kHz is depicted in more detail in Fig. 6. As the maximum of the spectral envelope is formed by a spectral component pair around 50kHz with a frequency separation of 42.1Hz, a moving radar target at a range corresponding to a spectral envelope maximum at 50kHz and giving rise to a Doppler shift of 21.05Hz is detected. Using (3) and (2), it can be calculated that this corresponds to a range of 3.75m and a
radial speed of 1.29 m/s. However, it has to be noted that the measured range is increased by 1.30 m by the antenna interconnection cables. The final conclusion is that a radar target moving at 1.29 m/s is detected at 2.45 m, which is a realistic measurement of the human target.

It can be seen that significant power is also present at exactly 50 kHz. As this power is only slightly higher compared to the blank test, no stationary target is detected at the range of 2.45 m.

V. Conclusion

The goal was to build a compact radar system operating in the 2.4 GHz ISM band, capable of detecting and measuring the range and radial speed of human targets. Several performed tests prove that the stated goal has been achieved. Moreover, the radar system has the potential to be fully integrated in garments.

VI. Outlook and Future Work

Both the compactness of the designed PCB and the application of textile materials for the antennas, provide an outlook towards full textile integration of the radar system.

Several aspects of the radar system itself can be improved.

Many improvement possibilities remain in the field of antenna design. The TX antenna could be realised as a phased antenna array as in [2]. The main antenna beam of a phased array is electronically steerable, enabling the system to sweep the environment looking for targets. The increased antenna gain of a phased array would reduce power consumption and consequently increase the compactness and textile integrability of the system. Higher gain of both the TX and RX antennas would lead to higher isolation between the antennas, and a higher angular resolution of the radar system.

Measurements indicated that the applied FM generator might fall short of linearity. Therefore, future implementations might use other topologies as described in [4] that are expected to be more linear.

More mature implementations should also keep the transmitted power and swept frequency band under better control.

REFERENCES

Ontwerp van een compacte frequentiegemoduleerde draaggolfradar voor integratie in kledij

Olivier Caytan

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I. INLEIDING

Het doel van een radarsysteem (acroniem voor RAdio Detection And Ranging) is de aanwezigheid van reflecterende doelen in de omgeving te detecteren en hun afstand en snelheid te meten aan de hand van electromagnetische straling [1]. De afstand kan worden berekend door te meten hoe lang de propageertijd is tussen de straling en de echo, en de radiale snelheid kan worden berekend door een eenvoudige manier van de afstand af te meten via de propageerzeer snelheid van het doel.

Het doel van deze masterthesis is om een compact radarsysteem te bouwen dat opereert in de 2.4GHz industriële, wetenschappelijke en medische (ISM) radioband en in staat is om mensen te detecteren en hun afstand en radiale snelheid te meten. Het systeem moet de frequentieband van 2.4GHz tot 2.5GHz gebruiken, en het equivalent isofoon uitgestraald vermogen (EIUV) dient beperkt te blijven tot 20dBm. Verder, als een stap in de richting van volledige textielintegratie van het radarsysteem in het vooruitzicht. Dit type radarsysteem is ideaal voor laagvermogens- en korrelafstandstoepassingen, en heeft het potentieel om erg compact te te worden gemaakt van de verschillende radiomodules, en in staat is om te worden gemeten met behulp van het Doppler-effect. Elektromagnetische straling die gereflecteerd wordt door een object dat beweegt in de propagatierichting van deze elektromagnetische straling, zal een frequentieschommeling ondergaan. Deze frequentieschommeling hangt op een eenvoudige manier af van de radiale snelheid van het doel.

Zijn licentie afgekeurd dat een radarsysteem in een omgeving met rook, mist of die zich achter een deur bevinden, biedt interessante toepassingen voor tal van reddings-, veiligheids-, en militaire toepassingen.

II. RADARSYSTEEMARCHITECTUUR

Vooropgesteld wordt er een studie gemaakt van de verschillende radarsysteemarchitecturen, om de meeste geschikte topologie voor het geformuleerde doel van te kiezen. Geselecteerde architectuur is de lineaire frequentiegemoduleerde draaggolfradar (lineaire FMCW) [3]. Lineaire FMCW radarsystemen zenden continu een lineair frequentie-gemoduleerd (FM) signaal uit. Dit type radarsysteem is ideaal voor laagvermogens- en korrelafstandstoepassingen, en heeft het potentieel om erg compact te worden geconstrueerd te worden. Verder kunnen FMCW radarsystemen zowel de afstand als de radiale snelheid van het doel meten. De signaalverwerking is vereenvoudigd door de lineairiteit van de frequentiemodulatie.

A. Blokdiagram

Het signaalverwerkingsblokdiagram van het FMCW radarsysteem wordt getoond in Fig. 1. Een FM signaal wordt opgewekt en uitgezonden met een antenne (TX antenne). Een uitdrukking voor de ogenblikkelijke frequentie \( f_{TX} \) van het verzonden signaal wordt gegeven door

\[ f_{TX}(t) = f_c + \Delta f \cdot \gamma(t), \]  

met \( f_c \) de draagolfrequentie van 2.45GHz, \( \Delta f \) de frequentiezwaaiv van 100MHz en \( \gamma(t) \) de lineair symmetrische driehoeksfrequentiemodulatiefunctie. Dit is een periodieke functie (modulatieperiode \( T_m \)) waarvan een enkele periode wordt getoond in Fig. 2. Deze frequentiemodulatiefunctie maakt het mogelijk op een eenvoudige manier tegelijkertijd afstand en radiale snelheid van het doel te meten.

Een dergelijk radarsysteem heeft tal van mogelijke toepassingen [2]. Aangezien de textielantennes kunnen gekanteld worden in de kledij van een operator, en het radarsysteem (PCB) zeer compact is ontworpen, zal het radarsysteem draagbaar en uiterst licht zijn, en bijgevolg de operator niet hinderen. De zintuigen van de drager worden verbeterd door het radarsysteem zonder verlies van comfort. Deze eigenschappen maken het radarsysteem erg praktisch in zware en veelvormige omstandigheden zoals in brandende gebouwen of op het slagveld. Mensen detecteren die onzichtbaar zijn door rook, mist of die zich achter een deur bevinden, biedt interessante toepassingen voor tal van reddings-, veiligheids-, en militaire toepassingen.

Trefwoorden—Frequentiegemoduleerde draaggolf (FMCW); industriële, wetenschappelijke en medische (ISM) radioband; compact; draagbaar
Het ontvangen signaal en een monster van het verzonden signaal worden gemengd, resulterend in een signaal dat bestaat uit een term met een frequentie rond 4.9GHz, die nutteloos en dus wordt weggefilterd door het daaropvolgende banddoorlaatfilter, en een basisbandterm die het geconverteerde signaal wordt genoemd. Het vermogensspectrum van het geconverteerde signaal bevat de informatie over de afstand en de radiale snelheid van het radardoel. Deze informatie wordt afgeleid door het signaalverwerkingscircuit.

B. Verwerking van het geconverteerde signaal

Om de afstand en radiale snelheid $v_{\text{rad}}$ van een doel te meten op tijdstip $t_0$, wordt een spectrale schatting van een signaalfragment gecentreerd op tijdstip $t_0$ van het geconverteerde signaal uitgevoerd.

De lage snelheid en de lage versnelling van mensen inacht genomen, kan de lengte van het signaalfragment voldoende kort gekozen worden zodat zowel $\tau$ als $v_{\text{rad}}$ verwaarloosbaar veranderen, terwijl het nog voldoende lang is om een goede spectrale schatting te behouden. Bij deze veronderstelling is het geschatte vermogensspectrum discreet met frequentiecomponenten $|k \cdot f_m \pm f_d|$, met $k$ een geheel getal, $f_m = 1/T_m$ de modulatiefrequentie en $f_d$ de Dopplerverschuiving. De Dopplerverschuiving $f_d$ wordt gegeven door

$$f_d = f_c \cdot \frac{2 \cdot v_{\text{rad}}}{c}.$$  \hspace{1cm} (2)

Specifiek in het geval van een lineaire symmetrische driehoeksfrequentiemodulatiefunctie, hebben de spectrale componenten op $|k \cdot f_m \pm f_d|$, met $k$ een geheel getal, eenzelfde vermogen.

Indien verondersteld wordt dat $T_m \gg \tau$, vinden we dat de spectrale enveloppe van het vermogensspectrum van het geconverteerde signaal een maximum vertoont rond de frequentie $f_{\text{max}}$, die gegeven wordt door

$$f_{\text{max}} = 2 \cdot f_c \cdot \frac{\tau}{T_m}.$$  \hspace{1cm} (3)

Een meting van de afstand van het radardoel kan worden bekomen door het spectrale componentpaar met maximaal vermogen te bepalen, en dan (3) te gebruiken. De frequentiescheiding van dit componentpaar is tevens gelijk aan tweemaal de Dopplerfrequentie. Met behulp van (2) wordt een meting van de radiale snelheid van het doel bekomen. Bemerk dat het onmogelijk is om te bepalen of het doel nadert of zich verwijdt.

Indien twee radardoelen aanwezig zijn, zal het spectrum van het geconverteerde signaal gelijk zijn aan de som van de spectra van de geconverteerde signalen voor elk doel afzonderlijk. De afstandsresolutie wordt gedefinieerd als het minimaal vereiste afstandverschil tussen twee doelen opdat ze nog steeds onderscheidbaar zouden zijn voor het radarsysteem. De afstandsresolutie is omgekeerd evenredig met de gebruikte bandbreedte $\Delta f$.

C. Prestatieverminderingen

Twee grote effecten begrenzen de prestaties van lineaire FMCW radarsystemen [4].

Zoals aangegeven op Fig. 1, ontvangt de ontvanger steeds direct een gedeelte van het uitgezonden vermogen ten gevolge van TX/RX antennelek (en in mindere mate ten gevolge van overspraak). Ten eerste kan het sterke direct gelekte signaal (in het algemeen het sterkste signaal aan de ontvangersingang) de ontvanger oversturen. Een tweede bedreiging, gevormd door het direct gelekte signaal, is verminderde gevoeligheid ten gevolge van maskering van de radardoelen door faseruis van de zender.

Daar het direct gelekte signaal een zeer korte vertraging heeft, manifesteert het zich als sterke laagfrequente spectrale componenten in het vermogenspectrum van het geconverteerde signaal. Het banddoorlaatfilter van Fig. 1 kan worden gebruikt om deze spectrale componenten te onderdrukken.

Niet-lineariteit van de frequentiemodulatie ligt aan de basis van een tweede prestatievermindering. Frequentiemodulatie niet-lineariteit verslechtert de afstandsresolutie van het radarsysteem. Typisch vergroot de verslechtering met de afstand van het doel, en is de afstandsresolutie bepaald door de gebruikte bandbreedte op korte afstand, en door de modulatielineariteit op grote afstand.

D. Modulatiefrequentie

Verscheidene invloeden moeten in rekening worden gebracht bij de keuze van de modulatiefrequentie. Een hogere modulatiefrequentie zal de snelheid betrouwbaar door het systeem kan worden gemeten verhogen, en zal de tijdsresolutie van het systeem verhogen. Een bovengrens op de modulatiefrequentie wordt opgelegd door de veronderstelling dat $\tau \ll T_m$ bij de analyse van het geconverteerde signaal. Rekening houdende met deze invloeden werd een modulatiefrequentie van 10kHz gekozen.

![FM generator](image)

Fig. 1: Blokdiagram frequentiegemoduleerde draaggolfradar

III. ONTWERP

Vervolgens wordt het FMCW radarsysteem gerealiseerd in elektronische hardware. Alle geselecteerde componenten zijn compacte surface-mounted componenten (SMD).

A. Circuitontwerp

Een schematisch blokdiagram van het ontworpen elektronisch circuit wordt getoond in Fig. 3.
De FM signaalgenerator wordt geïmplementeerd als een spanningsgestuurd oscillator (VCO) met de sturingspanning ingesteld door een driehoeksschering. De VCO is geselecteerd op basis van de lineairiteit van de sturingscharacteristiek, uitgangsvermogen, frequentieband en faserschuing. De driehoeksschering wordt gerealiseerd als een circuit met operationele versterkers (op-amps). Dit is een van de minst lineaire en meest simpele implementaties mogelijk, maar de verwachting was dat de frequentiemodulatie lineair genoeg zou zijn. Naderhand bleek dit verkeerd, aangezien metingen op een prototype aangaven dat de niet-lineairiteit de afstandsresolutie domineert vanaf een afstand van slechts 7.26m.

Het signaal opgewekt door de VCO wordt aangelegd aan een vermogenssplitter. De helft van het vermogen van de VCO wordt versterkt door de zendversterker (TX versterker) en aangelegd aan de TX antenne. De TX versterker wordt vooral geselecteerd op versterking, frequentieband en 1dB uitgangscompressiepunt (voldoende boven 20dBm). De versterking van de TX versterker wordt voor die doel van de TX uitgang van het bord. De overblijvende borduitgang is verbonden met de tussenfrequentiepoort (IF-poort) (zo hoog mogelijk).

De andere helft van het vermogen van de VCO wordt gebruikt om de lokale oscillatorpoort (LO-poort) van de mixer aan te sturen. Een mixer die kan worden aangestuurd met het beschikbare vermogen werd verwerkt. Andere componenten, die in overweging werden genomen zijn het frequentiebereik van de poorten, het omzettingsverlies (zo laag mogelijk) en het 1dB compressiepunt van de radiofrequentiespoort (RF-poort) (zo hoog mogelijk).

De ontvangstversterker (RX versterker) is een ketting van een lageruisversterker (LNA) en een tweede vermogensversterker. De ontvangstversterkers zullen onvermijdelijk het direct gelekte signaal versterken, en er moet worden opgelet dat het vermogen van het versterkte direct gelekte signaal voldoende onder het 1dB compressiepunt van de RF-poort van de mixer blijft. Een isolatie van 50dB tussen de TX en RX antenne wordt vooropgesteld.

De ketting gevormd door de twee RX versterkers en de mixer heeft een laag ruisgetal van 0.552dB dat de gevoeligheid van het systeem hoog maakt.

B. Printplaat ontwerp

Een printplaat (PCB) wordt ontworpen voor het besproken circuit. De PCB is opgebouwd uit twee 18μm koperlagen, waaronder een uitsluitend als grondvlak wordt gebruikt. Het PCB substraat tussen de twee koperlagen is het I-Tera MT materiaal van Isola Corporation (dielektrische constante $\varepsilon_r = 3.30$ en verliesstangens $tg\delta = 0.0036$) met een dikte van 0.5mm. Dit substraat vertoont lage verliezen en is geschikt voor hoogfrequente toepassingen tot 20GHz.

De PCB werd zo compact als mogelijk ontworpen (uiteindelijke PCB oppervlakte: 85mmx35mm). Deze compactheid houdt de verliezen laag, maar zal mogelijk de overspraak verhogen. Twee ontwerpbeslissingen werden gemaakt om de overspraak te beperken. Ten eerste werden de transmissielijnen op de PCB ontworpen als geaarde coplanaire golfgeleiders (GCPWs). De isolatie tussen twee GCPWs is doorgaans hoger dan tussen twee microstrips. Ten tweede zijn de transmissielijnen van de zender loodrecht op de transmissielijnen van de ontvanger geroutined.

EM circuit co-simulatie van het PCB ontwerp en de componentmodellen, uitgevoerd met de Advanced Design System (ADS) software van Agilent, toont aan dat de verliezen laag zijn en dat de zender/ontvanger-lek ten gevolge van PCB overspraak verwaarloosbaar is vergeleken bij de zender/ontvanger-lek ten gevolge van antennenelek.

Het bord heeft één ingang en twee uitgangen (SubMiniature version A connectoren). De RX antenne is verbonden met de RX ingang van het bord, en de TX antenne is verbonden met de TX uitgang van het bord. De overblijvende borduitgang is verbonden met de tussenfrequentiepoort (IF-poort) van de mixer is de IF-uitgang van het bord genoemd.

C. Antenneontwerp

De TX en RX antennes zijn zo ontworpen dat ze een verschillende circulaire polarisatie hebben. De orthogonal polarisatie van de TX en RX antenne leidt tot verhoogde isolatie tussen de twee antennes en vermindert multipadffecten door het verwerpen van tweede-orde reflecties [2]. Circulair gepolariseerde patchantennes met een rechthoekige ring topologie en enkele coaxiale voeding worden gebruikt [5]. De geleidende patch en het grondvlak worden vervaardigd van Flextron, een nylonweefsel waarop koper is afgezet. Het substraat tussen de patch en het grondvlak is zwart schuim geleverd door Javaux ($\varepsilon_r = 1.495$, $tg\delta = 0.0168$) met een dikte van 4mm. De gebruikte materialen zijn flexibil, licht en kunnen in kledingstukken worden geïntegreerd. Het antennepaar wordt geconstrueerd en op karton gefixeerd. De antennes worden verwerkt van elkaar geplaatst (ongeveer 30cm afstand tussen de centra) om een isolatie van ongeveer 50dB in de 2.4GHz ISM band te bereiken.

IV. Validatie

Tests met een gefabriceerd prototype werden uitgevoerd in een anechoïsche kamer om de functionaliteit van het radarsysteem te valideren. De IF-uitgang van het bord werd aangesloten op een spectrum analyser. Eerst werd het vermogensspectrum van het geconverteerde signaal gemeten in de afwezigheid van radardoeleinen in de kamer (blanco test). Het resultaat wordt voorgesteld in Fig. 4. Vervolgens werd een meting gedaan wanneer een mens op een afstand van ongeveer 3m vóór de radaran-
VCO
TX versterker TX antenne
mixer
RX versterker RX antenne
driehoeksc- oscillator

Fig. 3: Schematisch blokdiagram van het ontworpen elektronisch circuit

Fig. 4: Geen radardoelen aanwezig (blanco test). Vermogensspectrum IF-uitgang vanaf DC tot 75kHz.

driehoeksc-oscillator

Fig. 5: Wandeldende mens. Vermogensspectrum IF-uitgang vanaf DC tot 75kHz.

Fig. 6: Wandeldende mens. Vermogensspectrum IF-uitgang vanaf 49.5kHz tot 50.5kHz.

V. CONCLUSIE

Het doel was om een compact radarsysteem te bouwen dat opereert in de 2.4GHz ISM radioband, dat in staat is om mensen te detecteren en hun afstand en radiale snelheid te meten. Verscheidene tests tonen aan dat dit doel is bereikt. Verder heeft het radarsysteem het potentieel om volledig te worden geïntegreerd in kledingstukken.

VI. VOORUITZICHT EN TOEKOMSTIG WERK

Zowel de compactheid van de ontworpen printplaat als het gebruik van textielmaterialen voor de antennes, stelt een volledige textielintegratie van het radarsysteem in het vooruitzicht. Verscheidene aspecten van het radarsysteem zelf kunnen worden verbeterd.

Veel verbeteringsmogelijkheden bestaan op het vlak van antenneontwerp. De TX antenne zou kunnen worden gerealiseerd als een fasegestuurd antennerooster zoals in [2]. De antenne hoofdripol van een fasegestuurd antennerooster is elektronisch stuurbaar, wat het systeem in staat stelt de omgeving af te tasten voor radardoelen. De verhoogde antennewinst van een rooster zou het vermogenverbruik van het systeem verminderen, en bijgevolg de compactheid en textielintegratie van het systeem verhogen. hogere winsten van zowel de TX en RX antennes leiden ook tot hogere isolatie tussen deze antennes, en
een hogere angulaire resolutie van het radarsysteem. 

Meer volwassen implementaties zullen ook het vermogen en de bestreken frequentieband van het verzonden signaal beter onder controle kunnen houden.

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Glossary

ADS  Advanced Design System. 59
AGC  automatic gain control. 47
CW   continuous-wave. 6
DC   direct current. 9
DDS  direct digital synthesis. 25
EIRP equivalent isotropically radiated power. 2
EM   electromagnetic. 1
FFT  fast fourier transform. 31
FM   frequency modulation. 10
FMCW frequency-modulated continuous-wave. 10
GCPW grounded coplanar waveguide. 65
IF   intermediate frequency. 13
ISM  industrial, scientific and medical radio band. 2
LHCP left-hand circularly polarised. 77
LO   local oscillator. 13
op-amp operational amplifier. 25
PCB  printed circuit board. 2
PRF  pulse repetition frequency. 6
PRT pulse repetition time. 6
PSD power spectral density. 12
radar RAdio Detection and Ranging. 1
RBW Resolution Bandwidth. 52
RF radio frequency. 12
RHCP right-hand circularly polarised. 77
RMS root mean square. 17
RTT round trip time. 2
RX receiver. 11
SMA SubMiniature version A. 67
SNR signal-to-noise ratio. 13
SPICE Simulation Program with Integrated Circuit Emphasis. 37
TX transmitter. 1
VCO voltage controlled oscillator. 25
VGA variable gain amplifier. 43
Chapter 1

Introduction

1.1 What is radar?

The purpose of a RAdio Detection and Ranging (radar) system is to detect the presence of reflecting objects in the environment and to measure their range and velocity with electromagnetic (EM) radiation [1].

A very high level representation of a radar system is presented in Fig. 1.1. The transmitter of the radar system generates a signal with a certain type of waveform and emits it into the environment in the form of EM radiation using a transmitter (TX) antenna. Part of this EM energy hits a reflecting object (called the radar target) and is scattered in many directions by this object. A portion of this energy is scattered towards the radar receiver (RX) antenna. This energy is received by the RX antenna and is many magnitudes lower than the transmitted energy. In most cases this weak received echo signal is therefore amplified first, and then given to the radar receiver. This radar receiver decides on the presence of a radar target and extracts information about the radar target out of this signal. Typical information extracted about the target is its range to the radar system, its angular position and its radial velocity. Range can

![Fig. 1.1: High level representation of a general radar system](image-url)
be calculated by measuring the amount of time it takes for the radiation to travel from the TX antenna to the radar target, and then back to the RX antenna. This round trip time (RTT) is related in a simple manner to the range via the propagation speed. The radial velocity can be measured using the Doppler effect. A reflecting object that moves in the direction of propagation of EM radiation, introduces a frequency shift in the reflected radiation. This frequency shift is related in a simple way to the radial velocity of the target. The radial velocity can also be determined by repeated range measurements. The direction of arrival of the reflected signals reveals the angular position of the target.

1.2 Goal

The aim of this master’s dissertation is to design a compact radar system that operates in the 2450MHz industrial, scientific and medical radio band (ISM). The radar system has to be able to measure the range as well as the radial speed of the radar targets. Humans are the expected radar targets. Moreover, the antennas used by the system should be textile antennas.

The 2450MHz ISM radio band is subject to various regulations depending on the country. Typically, the frequency band ranging from 2.4GHz to 2.4835GHz is available for unlicensed use, if the radiated power is limited. In Belgium, this power is limited to 10dBm equivalent isotropically radiated power (EIRP) for non-specific short range devices (“niet-specifieke kortafstandsapparatuur”) and to 20dBm for wideband data transmission systems [2]. The goal, however, is not to build a radar system that complies to the regulations for the 2450MHz ISM band, but rather a radar system that may have potential to comply to those regulations. It was decided here to limit the used frequency band from 2.4GHz to 2.5GHz, and the EIRP to 20dBm.

The radar system should also be compact. Together with the demand for textile antennas, compactness paves the way for the full textile integration of this radar system as a future work.

1.3 Motivation

A radar system with the properties mentioned in the goal section has numerous potential applications. As the textile antennas can be integrated into the radar operator’s clothing, and the radar printed circuit board (PCB) is made very compact, the radar system will be wearable and very low-weight, and the operator will therefore not be hindered by the radar system. The wearer’s senses are enhanced by the radar system with minimal wearing effort. These properties make the use of this radar practical in harsh conditions and demanding situations such as in burning buildings or on the battlefield. The possibility to detect humans hidden by smoke or fog, or even behind doors offers interesting applications for numerous rescue, security and military purposes.
1.4 Content of this dissertation

This master’s dissertation consists of five chapters:

- Chapter 2 starts by introducing the radar equation and the Doppler effect. Next, it provides a brief overview of the different radar system architectures and selects the most suitable architecture for the stated goal. Subsequently, this particular radar system architecture is explained in more detail. A radar waveform is designed and the followed method to extract the range and radial velocity of the target from the received echo signal is presented. The performance limitations are discussed as well as implementation choices that can make a difference in performance. Finally, an upper limit is calculated on the radar operational range.

- In chapter 3 the radar system architecture that was decided on in the previous chapter is implemented in electronic circuitry. Components are selected and a PCB layout is designed. This PCB layout is partially simulated. The most important simulation is the isolation between the transmit subsystem and the receive subsystem of the radar board. Finally, the design of the transmit and receive antennas is discussed.

- Chapter 4 presents several measurements performed on prototype boards of the radar system. First, the correct operation of the different subsystems of the radar board is verified, insofar this is possible. This includes the testing of the TX and RX antennas. Next, the radar system operation is tested in several scenarios. Connecting the transmit and receive systems with a coax cable and attenuators, the radar system should be able to measure the length of the coax cable. Then, measurements using the designed antennas are performed in the anechoic chamber, testing the radar operation on stationary and moving targets.

- In chapter 5 the conclusion and outlook of this master’s dissertation is formulated.
Chapter 2

Design of the system architecture

2.1 Radar equation

A basic concept in radar is the radar equation \[ P_{TX} \cdot G_{TX} \cdot R^2 \]. The radar equation relates the power of the received echo signal to the power of the transmitted signal, the RX and TX antenna parameters, the propagation environment properties and the target properties. In this section, the simple form of the radar equation is discussed that is valid for the free space propagation environment.

When discussing the radar equation, Fig. 1.1 is instructive. Suppose \( P_{TX} \) is the power that is applied to the transmit antenna (without impedance mismatch between the antenna and the transmitter), and \( G_{TX} \) is the antenna gain of the transmit antenna in the direction of a reflecting object at range \( R \). We assume that the radar target is situated in the antenna far-field region. In this case the power density \( P_r \) (W/m\(^2\)) at the reflective object is given by:

\[
P_r = P_{TX} \cdot G_{TX} \cdot R^2.
\]

(2.1)

At this point the reflecting action of the target has to be described. This is described by the radar cross section with unit m\(^2\). The radar cross section is a measure for the amount of power that is intercepted by the target and reradiated towards the receive antenna. As seen from the receive antenna, the target behaves as an isotropic antenna that radiates a total power of \( \sigma \cdot P_r \). The estimated radar cross section of a human is 0.1m\(^2\) to 1m\(^2\) in the 2.4GHz ISM band, according to \[3\]. Consequently, the power density at the receive antenna is given by:

\[
P_r = P_{TX} \cdot G_{TX} \cdot \sigma \cdot R^2.
\]

(2.2)

The amount of power that the RX antenna receives and delivers to the receive circuitry is determined by the effective cross section (unit m\(^2\)) of the RX antenna \[4\]. The effective cross section of the RX antenna in the case of no impedance mismatch between the RX antenna and the receive circuitry, and in the absence of polarisation mismatch, is given by:

\[
A_e = \frac{\lambda^2 \cdot G_{RX}}{4\pi}.
\]

(2.3)
with $G_{RX}$ being the antenna gain of the RX antenna in the direction of the reflecting object at range $R$, and $\lambda$ the wavelength.

Finally, we can calculate the received echo signal power:

$$P_{RX} = P_{TX} \cdot \frac{G_{TX} \cdot \sigma}{(4\pi)^2 \cdot R^4 \cdot A_e} = P_{TX} \cdot \frac{G_{TX} \cdot \sigma \cdot G_{RX} \cdot \lambda^2}{(4\pi)^3 \cdot R^4}.$$  \hspace{1cm} (2.4)

This equation is the called the simple form of the radar equation.

If the minimum received signal power needed to detect a target is known, this equation can be used to determine the maximum operational range of the radar system. In our case, we mainly use it to determine the power that the received signal has after propagating from the TX antenna to the reflector, reflecting and propagating back to the RX antenna.

## 2.2 Doppler effect

In this section, the Doppler effect is discussed, an important effect that is exploited in many radar systems to measure the radial velocity of the radar targets. When the radar target has a radial speed relative to the radar system, the received echo signal frequency will be Doppler shifted with respect to the transmitted signal. The simplified expression for this frequency shift can easily be calculated when looking at Fig. 2.1. The radar system is stationary, while the target has a velocity $\vec{v}$. The radial velocity component $\vec{v}_{rad}$ is the projection of the vector $\vec{v}$ on the line between the radar system and the target. The radial velocity $v_{rad}$ is defined to be $+||\vec{v}_{rad}||$ when the target moves away from the radar system, and $-||\vec{v}_{rad}||$ otherwise. The range $R$ of the target to the radar system, as a function of time, is given by:

$$R(t) = R_0 + t \cdot v_{rad},$$  \hspace{1cm} (2.5)

with $R_0$ being the range of the target at time 0. The time $\tau$ it takes for a transmitted signal to propagate to the target, reflect, and propagate back, as a function of time is approximately given by:

$$\tau(t) = \frac{2 \cdot R(t)}{c} = \frac{2 \cdot (R_0 + t \cdot v_{rad})}{c},$$  \hspace{1cm} (2.6)

with $c$ being the propagation speed of EM waves in the considered medium. In vacuum, this is given by the speed of light. It is assumed the transmitted signal has a carrier frequency $f_c$, and is given by:

$$v_{TX}(t) = A \cdot \cos(2\pi f_c \cdot t).$$  \hspace{1cm} (2.7)
2.3 Overview radar types

The received signal is the transmitted signal delayed by a time $\tau$, so can be written as:

$$v_{RX}(t) = B \cdot \cos \left( 2\pi f_c \cdot (t - \tau(t)) \right) = B \cdot \cos \left( 2\pi f_c \cdot \left( t - \frac{2 \cdot (R_0 + t \cdot v_{rad})}{c} \right) \right). \quad (2.8)$$

This means the frequency of the received signal $f_{c,r}$ is given by:

$$f_{c,r} = \frac{d}{dt} \frac{1}{2\pi} \left( 2\pi f_c \cdot \left( t - \frac{2 \cdot (R_0 + t \cdot v_{rad})}{c} \right) \right) = f_c - f_c \cdot \frac{2 \cdot v_{rad}}{c}. \quad (2.9)$$

The expression obtained for the Doppler frequency shift is given by:

$$f_d = -f_c \cdot \frac{2 \cdot v_{rad}}{c}. \quad (2.10)$$

This means the reflection of a target moving away from the radar system is shifted down in frequency with respect to the transmitted signal, and a reflection of a target moving towards the radar system is shifted up in frequency.

2.3 Overview radar types

Radar systems can be distinguished based on several features \[1\]. They can be classified according to their intended application. Common radar applications include weather observation radars (used to observe rainfall, measure wind speed and other weather effects), surveillance radar systems (to detect the presence of a target, and measure its range and angular position), imaging radar systems (producing a two-dimensional image of a target, e.g. a satellite in orbit imaging the earth’s surface), guidance radar systems (to guide a device, e.g. a missile, towards another device, e.g. a boat), etc. Naturally, this is only a selection of the enormous application possibilities of radar systems.

Radar systems can also be classified according to the frequency band that they use.

A more important classification for this thesis is on basis of transmitter types. One can distinguish between pulse radar systems that periodically shut their transmitter down and continuous-wave (CW) radar systems that have their transmitter on continuously. This section will compare these two types of radar systems, and the type that best suits our application will be selected.

2.3.1 Pulse radar

Fig. 2.2 shows the general block diagram of a pulse radar system \[3\]. The generator, which is controlled by the pulse modulator, periodically generates a short burst of length $\tau$ of a high power single frequency carrier. These bursts are repeated at the so-called pulse repetition frequency (PRF). The inverse of the PRF is called the pulse repetition time (PRT). We now consider what happens during one PRT. Before the generator is switched on, the duplexer disconnects or shuts down the sensitive receive circuitry to protect it from damage caused by the high power signal of the generator and connects the generator to the antenna. Next, the generator is turned on
for a brief time $\tau$, and the generated signal is transmitted with the antenna. For the rest of
the PRT, the generator is disconnected from the antenna, and the receive circuitry is turned
on and connected to the antenna. During this time, the radar system "listens" for echo signals.
The PRT should be long enough to allow for the echoes with the longest expected delays to
arrive back at the radar system within the PRT. The receive circuitry has to detect and process
the received echo signals. Typically, the received signal is amplified with a low noise amplifier
(LNA). The signal processing circuit is responsible for deciding if an echo signal is present, and
if this is the case, extracting the range and radial velocity information. Range measurements
are performed by measuring the time difference between transmission of the pulse and reception
of the echo signal. Measurement of the radial velocity happens by exploiting the Doppler effect.
The echo burst received from a target that has a radial movement with respect to the radar,
will have a frequency that is different from the transmitted frequency as a consequence of the
Doppler effect.

A consequence of shutting down the receiver during the transmission of the high power pulse
is a so-called blind range. Only targets that are at a certain minimum range can be detected by
the system. The radar system is blind for targets that are closer. This constitutes a problem, as
we are building a radar that should work at a low range. It is easily understood that the blind
range $R_{\text{min}}$ is given by:

$$R_{\text{min}} = \frac{c}{2} \cdot (\tau + t_s),$$

with $t_s$ being the time the system needs to turn on the receiver circuitry, after the high power
burst of length $\tau$ is completely transmitted. Thus, pulse radar systems that can operate at a
close range require a very short burst length $\tau$. Moreover, when the burst is made shorter, its
instantaneous power has to become larger, if the same average power is to be held constant.
Therefore, pulse radar systems for close range will lead to complicated circuits that need to gen-
erate very short bursts and need to be able to handle very high instantaneous powers. Moreover,
in our case, the instantaneous transmitted power is limited to 20dBm EIRP. Therefore, we can
already conclude that a pulse radar system is not a suitable radar architecture to achieve our
goal.

### 2.3.2 Continuous-wave radar

The CW type radars have their transmitter turned on continuously. Their receive circuitry is
also always active, so it is expected that the problem of the blind range found in pulse radar
systems is avoided here.

The pulse radar systems separate the weak echo signal from the strong transmitted signal in
the time domain. As in the case of CW radars, the transmitter is continuously turned on, another
approach will have to be adopted here. This will become clear when discussing the subtypes
of the CW radar. CW radar systems can be subdivided into unmodulated and modulated CW
radar systems.
2.3 Overview radar types

Fig. 2.2: General pulse radar system

Unmodulated continuous-wave radar

Fig. 2.3: Block scheme of an unmodulated (a) and a frequency-modulated (b) CW radar

The block diagram for an unmodulated CW radar is given in Fig. 2.3a. The transmitted signal is an unmodulated CW carrier with fixed frequency $f_c$ and constant power. The frequency $f_{RX}$ of the received echo signal depends on the target radial velocity. If the target has a radial velocity different from zero, the received signal frequency exhibits a Doppler shift. In section 2.2 the received echo signal frequency was shown to be:

$$f_{RX} = f_c - f_c \cdot \frac{2 \cdot v_{rad}}{c},$$

(2.12)

with $v_{rad}$ being the radial velocity of the radar target with respect to the radar system. The received echo signal is mixed with a sample of the transmitted signal (homodyne system). The signal at the output of the mixer contains two frequency components: $|f_c - f_{RX}|$ and $f_c + f_{RX}$.
The interesting component of the two is the baseband component $|f_c - f_{RX}|$, also called the beat frequency $f_b$, and is equal to:

$$f_b = |f_c - f_{RX}| = f_c \cdot \frac{2 \cdot |v_{rad}|}{c}. \quad (2.13)$$

The presence of a target with a radial speed $|v_{rad}|$ will thus be evident by the presence of a frequency component $f_b$. We note that the system is not able to distinguish between approaching and receding targets, a consequence of being a homodyne system.

If there are multiple targets present, each target will independently produce a beat frequency at the output of the mixer (insofar the mixer is linear). Targets having an identical radial speed are indistinguishable.

Besides the echo signal originating from power of the TX antenna bouncing of the target towards the RX antenna, the RX antenna also receives a signal directly from the TX antenna. For the remainder of this text, this signal will be called the directly leaked signal. This power leakage is indicated in Fig. 2.3a by the dotted line pointing from the TX antenna to the RX antenna. The consequence of this directly leaked signal is that, at the output of the mixer, there are also frequency components at direct current (DC) and $2 \cdot f_c$, besides the intended frequency components at $|f_c - f_{RX}|$ and $f_c + f_{RX}$. These additional frequency components are always present, even if there are no reflections from targets at all. This DC signal will disable the system to detect stationary and very low speed targets.

The band pass filter that follows the frequency mixer filters away the ever-present frequency components at DC and $2 \cdot f_c$. The high-frequency component $f_c + f_{RX}$ also has to be suppressed. The high-frequency signals at $2 \cdot f_c$ and $f_c + f_{RX}$ usually fall outside the response of the following signal processing circuit, but low pass filtering is needed to reduce the noise power and to enhance the sensitivity of the system. The DC signal usually does fall in the frequency response and has to be filtered away, because its power may saturate the signal processing circuit. As a practical filter has a finite steepness, a minimum radial speed depending on the lower corner frequency of the band pass filter is needed for detection.

It is clear that this type of radar is completely unsuitable for the goal stated in the introduction, as only moving targets can be detected and the range of those targets cannot be measured. A solution for this is modulation of the transmitted signal.

**Frequency-modulated continuous-wave radar**

The reason why the unmodulated CW radar is unable to measure the range of the targets is that the transmitted signal is stationary. There is no way of measuring how long the signal needed to propagate from the TX antenna to the target, and back to the RX antenna. The solution to this is changing the transmitted signal properties as a function of time, or in other words, modulating the transmitted carrier. Modulating the carrier will broaden its frequency spectrum. Acquiring range measurements therefore requires the use of more bandwidth. Below, we will show that the accuracy of the range measurements directly depends on the used bandwidth.
In general, amplitude modulation or phase modulation (or equivalently, frequency modulation (FM)) can be performed. Using amplitude modulation is difficult because the received echo signal will always be superimposed on the relatively strong directly leaked signal. Usually, the frequency of the carrier is modulated. In this case, the received echo signal and the directly leaked signal will have different frequencies, and are distinguishable.

The block diagram for a frequency-modulated continuous-wave (FMCW) radar system is given in Fig. 2.3b. The transmitted signal is frequency-modulated with a certain modulation function. A few of the most used modulation functions will be discussed in the next section devoted entirely to the FMCW radar system.

2.3.3 Conclusion

Despite its ability to measure both target range and radial velocity, the pulse radar system architecture is not selected, because using such systems at a low range would require a very narrow pulse with a very high instantaneous power. This would require complicated circuits that would surely not be very compact. Moreover, in our case, the instantaneous transmitted power is limited to 20dBm EIRP.

The CW radar systems are more suitable for low range applications. Their receiver is continuously switched on and as a consequence their blind range is normally much lower than for pulse radar systems. In the case of CW radar systems, the instantaneous transmitter power is equal to the average transmitter power. A pulse radar system needs enormous instantaneous power to reach the same average power as a CW radar system because of the narrowness of the pulse. As average power determines the ability to detect targets [5], CW radars can reach the same ability to detect targets with lower power circuitry. Modulated subtypes of CW radar systems, like the FMCW radar systems, are also able to measure both target range and radial velocity. The main disadvantage of CW radar systems is the directly leaked signal.

It can be concluded that an FMCW radar system is the ideal architecture choice to achieve the goal. FMCW radar systems are discussed in the next section in more detail.

2.4 Frequency-modulated continuous-wave (FMCW) radar

Now that it has been established that the FMCW radar system is the best choice for our purpose, this radar system type will be studied in detail. An excellent resource on FMCW radar signal processing is [6].

2.4.1 Description of the block diagram

We will now explain in detail how the block diagram of Fig. 2.3b works.

The transmitted signal is frequency-modulated and has a instantaneous angular frequency
2.4 Frequency-modulated continuous-wave (FMCW) radar

given by:

\[ \omega_{TX}(t) = \omega_c + \Delta \omega \cdot \gamma(t). \]  

(2.14)

\( \omega_c \) is the angular carrier frequency equal to \( 2\pi \cdot f_c = 2\pi \cdot 2.45 \text{GHz} \). \( \gamma(t) \) is the FM function. This function is assumed to have a period \( T_m \) (modulation period) and \( |\gamma(t)| \leq 0.5 \). \( \Delta \omega = 2\pi \cdot \Delta f = 2\pi \cdot 100 \text{MHz} \) is the total angular frequency swing. It follows that the transmitted signal phase \( \phi_{TX}(t) \) is given by (phase assumed to be 0 rad at time 0):

\[ \phi_{TX}(t) = \int_0^t \omega_{TX}(u) \, du = \omega_c \cdot t + \Delta \omega \cdot \int_0^t \gamma(u) \, du. \]  

(2.15)

Consequently, the expression for the transmitted signal \( u_{TX}(t) \) is:

\[ u_{TX}(t) = A \cdot \cos \left( \phi_{TX}(t) \right) = A \cdot \cos \left( \omega_c \cdot t + \Delta \omega \cdot \int_0^t \gamma(u) \, du \right), \]  

(2.16)

with \( A \) the amplitude corresponding with the transmitted signal power.

Ideally, the received signal is a time delayed (delay \( \tau \)) and attenuated version of the transmitted signal, possibly with an extra phase jump \( \phi_0 \) originating of the reflecting target. The received signal phase \( \phi_{RX}(t) \) is given by:

\[ \phi_{RX}(t) = \phi_{TX}(t - \tau) + \phi_0 = \omega_c \cdot (t - \tau) + \Delta \omega \cdot \int_0^{t-\tau} \gamma(u) \, du + \phi_0. \]  

(2.17)

Consequently, the expression for the received signal \( u_{RX}(t) \) is:

\[ u_{RX}(t) = B \cdot \cos \left( \phi_{RX}(t) \right) = B \cdot \cos \left( \omega_c \cdot (t - \tau) + \Delta \omega \cdot \int_0^{t-\tau} \gamma(u) \, du + \phi_0 \right), \]  

(2.18)

with \( B \) the amplitude corresponding to the received signal power. In reality, also a directly leaked signal will be received, again indicated in Fig. 2.3b by the dotted arrow, but this will be ignored for now and discussed later in subsection 2.4.4.

A sample of the transmitted signal is mixed with the received signal resulting in the signal \( u_{IF}(t) \) given by:

\[ u_{IF}(t) = C \cdot \cos(\phi_{TX}(t) - \phi_{RX}(t)) + C \cdot \cos(\phi_{TX}(t) + \phi_{RX}(t)). \]  

(2.19)

The first term has a frequency at baseband, and the second term has a frequency around twice the carrier frequency \( \omega_c \). Both terms nominally have an equal amplitude \( C \) and power. This power is related to the power of the input signals of the mixer and the mixer itself. This is of no importance here. Of the two terms, only the baseband term is useful and the second term will be filtered out by the following band pass filter. The baseband term is called the converted signal \( u_c(t) \) and can be expanded as follows:

\[ u_c(t) = C \cdot \cos(\phi_{TX}(t) - \phi_{RX}(t)) \]

\[ = C \cdot \cos \left( \omega_c \cdot t + \Delta \omega \cdot \int_0^t \gamma(u) \, du - \left( \omega_c \cdot (t - \tau) + \Delta \omega \cdot \int_0^{t-\tau} \gamma(u) \, du + \phi_0 \right) \right) \]  

(2.20)

\[ = C \cdot \cos \left( \omega_c \cdot \tau + \Delta \omega \cdot \int_{t-\tau}^t \gamma(u) \, du - \phi_0 \right). \]
2.4 Frequency-modulated continuous-wave (FMCW) radar

The converted signal contains range and radial velocity information of the target. Subsection 2.4.3 will discuss the spectrum of the converted signal and how to extract the range and velocity information out of it. First, however, a few alternative FMCW radar systems will be discussed.

2.4.2 Alternative FMCW block diagrams

Two widely mentioned [1] [6] alternative FMCW systems are shown in Fig. 2.4.

Fig. 2.4a shows an FMCW radar system using only one antenna. By using a circulator component, one antenna can be used both for transmission and reception. The signal generated by the oscillator is passed on by the circulator to the transmit antenna. The signal received by the antenna will be passed on by the circulator to the frequency mixer. Its greatest quality is compactness, which is one of the main goals of this thesis. Its greatest disadvantage is limited isolation between transmitter and receiver (radio frequency (RF) port of the mixer). Initially, the isolation is limited by the match of the antenna to the system. Any reflection of power of the FM generator at the antenna will directly leak into the receiver. Typically an antenna is assumed to be impedance matched in a frequency band if it has a return loss of 10dB or more in that frequency band. This is a very poor transmitter/receiver isolation. If we match the antenna better, increasing the return loss, the isolation between the FM generator and the receiver will eventually be limited by the isolation of the circulator. A part of the power applied at port 1 of the circulator will appear at port 3. In this way, the isolation of this system is limited to 20dB-50dB.

We therefore decided to use a system that applies a separate antenna for transmission and reception. In this case, the isolation is generally limited by the isolation between the TX and RX antennas. Many opportunities exist to increase the TX/RX isolation. An evident way is by increasing the physical separation between the antennas. This, of course, makes the system less compact. Absorbing materials can be used to keep the antenna isolation high, while bringing them physically closer. Moreover, the antennas can be made more directional, decreasing the gain of the TX antenna in the direction of the RX antenna, and vice-versa. A more advanced way to increase the isolation even more, is by direct cancellation. It is attempted to cancel the directly leaked signal by adding an extra sample of the transmitted signal with correct phase and amplitude. If this is done with a control loop, an additional isolation of 30dB can be reached [1].

The system studied in subsection 2.4.1 is a homodyne system. The homodyne system mixes the received echo signal with a sample of the transmitted signal down to baseband. In electronic systems, besides white noise (thermal noise, shot noise, flat power spectral density (PSD)), there is also pink noise (flicker noise, PSD with 1/f profile) present. Below a certain corner frequency, the magnitude of the PSD of the pink noise is larger than that of white noise, and pink noise dominates. The PSD magnitude of the combined noise below the corner frequency is increased compared to the situation where only white noise is present. Above the corner frequency, pink noise power becomes negligible, and the magnitude of the PSD of the combined noise is
practically equal to the case where only white noise power is present. Thus, homodyne systems convert the received signal to a frequency region where noise power is enhanced. Fig. 2.4b shows a heterodyne FMCW radar system. A local oscillator (LO) $f_{IF}$ mixes with the generated FM signal (centre frequency $f_c$), and a band pass filter selects the down-mixed component with centre frequency $f_c - f_{IF}$. This signal is mixed with the received signal and the result is a signal with a centre frequency $f_{IF}$. If the intermediate frequency (IF) is chosen above the noise corner frequency, the 1/f noise no longer plays a role and the signal-to-noise ratio (SNR) will be better than is the case in homodyne systems. The signal will eventually have to be mixed to baseband, in order to process it, but first it is amplified with an IF amplifier, in order to keep the SNR high.

The disadvantage, however, is increased complexity and combined with this reduced compactness. Therefore, it was chosen not to implement this system as a heterodyne. As LNAs and power amplifiers are abundantly available for the frequency band of the received signal (2.4GHz ISM band), an amplifier will be directly placed behind the RX antenna, as in Fig. 2.4c (identical to Fig. 2.3b but with added amplifiers). If the amplification is high enough, a good SNR will be maintained, even in the presence of 1/f noise at baseband.

### 2.4.3 Spectrum of the converted signal

We will now investigate how to calculate the converted signal frequency spectrum for a general modulation function.

We assume that the target has a constant radial velocity $v_{rad}$. During a very short time interval, such as the time needed by the radar system to process and interpret this converted signal, order 10 milliseconds, this is always approximately the case. In this case the delay $\tau$ will vary linearly as a function of time, as given by the equation:

$$\tau(t) = \tau_0 + \frac{2 \cdot t \cdot v_{rad}}{c},$$

with $\tau_0$ being the delay at time 0. It follows that:

$$\omega_c \cdot \tau(t) = \omega_c \cdot \tau_0 + \frac{2 \cdot t \cdot v_{rad}}{c} = \omega_c \cdot \tau_0 + \omega_d \cdot t,$$

with $\omega_d$ being the angular Doppler frequency shift. Note that the Doppler frequency used here has an extra minus sign compared to the formula in section 2.2.

We can now rewrite the expression for the converted signal:

$$u_c(t) = C \cdot \cos \left( \omega_c \cdot \tau_0 + \omega_d \cdot t + \Delta \omega \cdot \int_{t-\tau}^{t} \gamma(u) du - \phi_0 \right)$$

$$= C \cdot \cos \left( \omega_d \cdot t + \omega_c \cdot \tau_0 - \phi_0 \right) \cdot \cos \left( \Delta \omega \cdot \int_{t-\tau}^{t} \gamma(u) du \right)$$

$$- C \cdot \sin \left( \omega_d \cdot t + \omega_c \cdot \tau_0 - \phi_0 \right) \cdot \sin \left( \Delta \omega \cdot \int_{t-\tau}^{t} \gamma(u) du \right).$$

(2.23)
2.4 Frequency-modulated continuous-wave (FMCW) radar

(a) Single antenna frequency-modulated continuous-wave radar

(b) Heterodyne frequency-modulated continuous-wave radar

(c) Frequency-modulated continuous-wave radar with extra amplifiers

Fig. 2.4: Alternative frequency-modulated continuous-wave radar block diagrams
The instantaneous frequency of the converted signal can be calculated by differentiating its phase:

$$\omega(t) = \frac{d}{dt} \left( \omega_c \cdot \tau_0 + \omega_d \cdot t + \Delta \omega \cdot \int_{t-\tau}^{t} \gamma(u) du - \phi_0 \right)$$

$$= \omega_d + \Delta \omega \cdot \frac{d}{dt} \left( \int_{t-\tau}^{t} \gamma(u) du \right)$$

(2.24)

With the converted signal written in the form of equation 2.23, only the frequency spectra of the factors

$$\cos \left( \Delta \omega \cdot \int_{t-\tau}^{t} \gamma(u) du \right) \quad (2.25)$$

and

$$\sin \left( \Delta \omega \cdot \int_{t-\tau}^{t} \gamma(u) du \right) \quad (2.26)$$

still need to be calculated in order to calculate the frequency spectrum of the converted signal.

In fact, we are not interested in the frequency spectrum of the entire converted signal. Usually, to measure the range and radial velocity of a target at time \(t_0\), a spectral estimation of a signal fragment of the converted signal, centred at time \(t_0\) is performed. In that fragment, \(\tau\) should vary negligibly so that the spectrum that we estimate is approximately equal to the spectrum that we would acquire when \(\tau\) is constant and equal to \(\tau(t_0)\). The length of the fragment should be long enough in order to have a good spectral estimation of the signal with adequate frequency resolution, and short enough to have a nearly constant \(\tau\). A shorter fragment length also allows to do more range and velocity measurements in given time, increasing the time resolution of the system.

If \(\tau(t)\) is assumed to be constant as a function of time in equations 2.25 and 2.26, these factors are periodic with a period \(T_m\), because \(\gamma(t)\) is periodic with that period. The frequency spectra are then discrete, consisting of components at the frequencies \(k \cdot (1/T_m)\). The frequency spectrum of the converted signal can be calculated by performing the Fourier series expansion of these factors and then using equation 2.23. The frequency spectrum of the converted signal is also discrete with frequency components at \(|k \cdot \omega_m + \omega_d|\) and \(|k \cdot \omega_m - \omega_d|\), with \(k\) an integer and \(\omega_m = 2\pi/T_m\) the angular modulation frequency. The magnitudes and phases of these components will depend on the chosen \(\gamma(t)\), \(\Delta \omega\) and most importantly \(\tau\).

Next, a few widely used FM functions are discussed, and we will select the most suitable one for our purpose. These are all linear FM functions because of their good properties. As the instantaneous frequency of the converted signal takes on constant values during parts of the modulation period, linear FM functions lead to converted signal spectra with minimum spectral width, and well-defined maxima in the spectrum.

**Sawtooth frequency modulation**

One period of this FM function is depicted in Fig. 2.5a.
2.4 Frequency-modulated continuous-wave (FMCW) radar

(a) One period of the sawtooth frequency modulation function

(b) One period of the general linear frequency modulation function

(c) One period of the symmetrical triangle frequency modulation function

Fig. 2.5: Linear frequency modulation functions
With equation 2.24, it can be calculated that the instantaneous frequency of the converted signal only takes two values [6, p. 35]:

\[
\omega_1 = \left| \omega_d + \Delta \omega \cdot \frac{\tau}{T_m} \right|, \quad t \in \left[ (n - 1/2)T_m + \tau, (n + 1/2)T_m \right]
\]

(2.27)

and

\[
\omega_2 = \left| \omega_d + \Delta \omega \cdot \left( \frac{\tau}{T_m} - 1 \right) \right|, \quad t \in \left[ (n + 1/2)T_m, (n + 1/2)T_m + \tau \right]
\]

(2.28)

with \( n \) being an integer. Both frequencies have a component caused by the target radial velocity, and a component caused by the target range. These expressions can also be derived by using diagrams as depicted in Fig. 2.6. The FM functions of both the transmitted and received signal are shown. The instantaneous frequency of the converted signal is given by the difference of these functions. The diagram of Fig. 2.6 assumes the Doppler shift is negative (approaching target), and that the frequency component caused by target range is in absolute value larger than the frequency component caused by the target radial velocity in absolute value. The two instantaneous frequencies \( f_1 \) and \( f_2 \) and the Doppler frequency \( f_d \) are depicted.

Usually, \( T_m \gg \tau \), so most of the time, the instantaneous frequency is equal to \( \omega_1 \). It was already proven that the frequency spectrum of the converted signal consists of components at the frequencies \( |k \cdot \omega_m + \omega_d| \) and \( |k \cdot \omega_m - \omega_d| \). Intuitively, we can expect the frequency components around \( \omega_1 \) to have maximal power. As \( \omega_1 \) linearly depends on the delay \( \tau \), we will be able to measure the range.
2.4 Frequency-modulated continuous-wave (FMCW) radar

If we assume $T_m \gg \tau$, the converted signal can be calculated to be [6, p. 36]:

$$u_c(t) = C \cdot \left( \begin{array}{c}
\sin (\pi \Delta f \tau) \\
\sum_{k=1}^{\infty} \frac{\sin (\pi (\Delta f \tau + k))}{\pi (\Delta f \tau - k)} \cdot \cos ((k\omega_m - \omega_d) \cdot t + \omega_c \cdot \tau_0 - \phi_0 - k\omega_m \tau / 2)
\end{array} \right)$$

$$= \sum_{k=-\infty}^{\infty} \frac{\sin (\pi (\Delta f \tau - k))}{\pi (\Delta f \tau - k)} \cdot \cos ((k\omega_m + \omega_d) \cdot t + \omega_c \cdot \tau_0 - \phi_0 - k\omega_m \tau / 2)$$

(2.29)

With $u_c(t)$ written in this way, the frequency spectrum is evident.

The angular frequency of the $k$-th term is $k\omega_m + \omega_d$, so if we replace $k \omega_m + \omega_d$ by $\omega$ in the coefficient of the $k$-th term, we acquire the envelope $w(\omega)$ of the spectral components:

$$w(\omega) = \text{sinc} \left( \pi \left( \frac{\Delta f \tau - \omega - \omega_d}{\omega_m} \right) \right)$$

(2.30)

The region of negative $\omega$ should be folded to the region of positive $\omega$. This envelope has a maximum at:

$$\omega = |\omega_d + \Delta f \tau \omega_m|$$

(2.31)

which is identical to equation 2.27. This means that the frequency components with the most power lie around this frequency. The total width of the main lobe of the envelope is $2 \cdot \omega_m$, so at most two frequency components fall in the main lobe. The first sidelobes are 13.5dB lower.

In the special case that $\Delta f \tau$ is an integer, only one frequency component $|\omega_d + \Delta f \tau \omega_m|$ is present in the spectrum.

It is now clear how to extract the range and radial velocity information of the target from the converted signal frequency spectrum. The magnitude of the Doppler frequency shift can be evaluated by measuring the separation of any two components $k \cdot \omega_m + \omega_d$ and $k \cdot \omega_m - \omega_d$. Hereby we assume that the frequency $k \cdot \omega_m - \omega_d$ is higher than $(k-1) \cdot \omega_m + \omega_d$, so no frequency components have switched places because of the Doppler shift. This limits the maximum Doppler frequency shift that can be reliably measured to $|\omega_d| < \omega_m / 2$. We can also derive the sign of the Doppler shift when looking at the frequency spectrum. If the target is receding (positive $v_{rad}$, positive $\omega_d$), then the component with the frequency higher than $k \cdot \omega_m$ will be the component with the largest power of the two. If the target is approaching, vice-versa. By measuring the frequency of the component with maximal power, we have a measurement of the target range, using equation 2.31.

We note that the range measurement is inherently granular. If we assume that the target is stationary, the most powerful frequency component is a multiple of $f_m$; suppose this is $k \cdot f_m$. Using equation 2.31, we obtain an expression for the range $R$:

$$\tau = \frac{2 \cdot R}{c} \Rightarrow R = \frac{c \cdot \tau}{2} = \frac{c \cdot k \cdot f_m}{2 \cdot \Delta f \cdot f_m} = \frac{c \cdot k}{2 \cdot \Delta f}.$$  

(2.32)
The measured range is granular because it is always a multiple of \( R_{\text{min}} = c/(2 \cdot \Delta f) \). We notice that, in the case of a sawtooth FM, the only way of increasing the accuracy of the range measurement, is by increasing the used bandwidth \( \Delta f \).

If two targets are present, the converted signal spectrum will be equal to the sum of the converted signal spectra for each target alone. The range resolution is defined as the minimal range difference there has to be between two targets (target one having delay \( \tau_1 \) and target two having delay \( \tau_2 \)) in order that they are still distinguishable by the radar system. Both targets will give rise to a maximum in the spectral envelope. The worst case situation is when the spectral envelopes of both targets have a maximum halfway between two spectral components. As the total width of the main lobe of the envelope is \( 2 \cdot \omega_m \), the envelope maxima have to be separated by at least 3 times the modulation frequency in order for these maxima to be reliably resolved. This leads to the expression for the range resolution for sawtooth FM:

\[
\Delta R = 3 \cdot c/(2 \cdot \Delta f) = 1.5 \cdot c/(\Delta f)
\]

We see that the sawtooth FM has good properties. However, it has one major disadvantage. Components \( k\omega_m + \omega_d \) and \( k\omega_m - \omega_d \) have a different power as seen in equation 2.29. It might be possible that one of the two components falls below the noise floor. In this case, we cannot measure the Doppler frequency as mentioned above. It is still possible to measure the Doppler frequency by measuring the difference between any frequency component and the closest harmonic of the modulation frequency. To this end, the modulation frequency needs to be known with great precision. In addition, a perfect sawtooth FM function cannot be realised because of the infinite steepness at times \( k \cdot T_m + T_m/2 \), with \( k \) an integer. We will take another linear FM function that does not have these disadvantages.

**General linear frequency modulation**

We now look at a more general linear FM function in order to find a waveform without the disadvantages mentioned above. The waveform we consider is depicted in Fig. 2.5b. The parameter \( \alpha \) can vary between 0.5 and 1 (the cases between 0 and 0.5 are equivalent). When \( \alpha = 1 \) the function degenerates to a sawtooth function.

If we still assume \( T_m \gg \tau \), the instantaneous frequency of the converted signal takes on two different constant values during the major part of each modulation period [6, p. 40]:

\[
\omega_1 = \left| \omega_d + \Delta \omega \cdot \frac{T_m}{T_m - \alpha} \right|
\]

and

\[
\omega_2 = \left| \omega_d - \Delta \omega \cdot \frac{T_m}{T_m \cdot (1 - \alpha)} \right|
\]

The converted signal frequency spectrum is given in [6, p.40]. The spectral envelope of the converted signal frequency spectrum now exhibits two maxima: a maximum around \( \omega_1 \) and a maximum around \( \omega_2 \). As the instantaneous frequency is part of the time \( \omega_1 \), and part of the
time $\omega_2$, with the fractions depending on $\alpha$, the main lobes of the maxima will in general have a different width and a different amplitude. The width of the main lobe around $\omega_1$, $\Delta \omega_1$ is given by:

$$
\Delta \omega_1 = \frac{4\pi}{\alpha \cdot T_m}
$$

(2.36)

and the width of the main lobe around $\omega_2$, $\Delta \omega_2$ is given by:

$$
\Delta \omega_2 = \frac{4\pi}{(1 - \alpha) \cdot T_m}
$$

(2.37)

The lobes are always broader than is the case with the sawtooth modulation function ($\alpha = 1$).

For $\alpha > 0.5$, the components with maximum power in the main lobe around $\omega_1$ of the component pairs $k \cdot \omega_m \pm \omega_d$ are the components $k \cdot \omega_m + \omega_d$. For the lobe around $\omega_2$, these are the components $k \cdot \omega_m - \omega_d$. In case $\alpha = 0.5$, the two main lobes coincide and have an equal amplitude. In this case the frequency components at $k \cdot \omega_m \pm \omega_d$ have an equal power. As this outcome is desirable, we choose $\alpha = 0.5$.

**Triangle frequency modulation**

If we choose $\alpha = 0.5$, we have a symmetrical triangle FM function. One period of this FM function is depicted in Fig. 2.5c.

The instantaneous frequency of the converted signal takes on two different constant values during the major part of the modulation period:

$$
\omega_1 = \left| \omega_d + 2 \cdot \Delta \omega \cdot \frac{\tau}{T_m} \right|
$$

(2.38)

and

$$
\omega_2 = \left| \omega_d - 2 \cdot \Delta \omega \cdot \frac{\tau}{T_m} \right|.
$$

(2.39)

Under the assumption $T_m \gg \tau$, the converted signal expression can be calculated to be [6, p.41]:

$$
u_c(t) = C \cdot \left\{ \begin{array}{l}
\sin \left( \frac{\pi f_1 \tau}{f_2} \right) \cdot \cos \left( \omega_d \cdot t + \omega_c \cdot \tau_0 - \phi_0 \right) \\
+ \sum_{k=1}^{\infty} \left( \frac{\sin \left( \frac{\pi f_2 \tau + k/2}{f_2} \right)}{2 \pi (f_2 \tau + k/2)} + (-1)^k \cdot \frac{\sin \left( \frac{\pi f_2 \tau - k/2}{f_2} \right)}{2 \pi (f_2 \tau - k/2)} \right) \\
\cdot \cos \left( (k\omega_m - \omega_d) \cdot t - \omega_c \cdot \tau_0 + \phi_0 - k\omega_m \tau/2 \right) \\
\end{array} \right. \\
+ \sum_{k=1}^{\infty} \left( \frac{\sin \left( \frac{\pi f_2 \tau - k/2}{f_2} \right)}{2 \pi (f_2 \tau - k/2)} + (-1)^k \cdot \frac{\sin \left( \frac{\pi f_2 \tau + k/2}{f_2} \right)}{2 \pi (f_2 \tau + k/2)} \right) \\
\cdot \cos \left( (k\omega_m + \omega_d) \cdot t + \omega_c \cdot \tau_0 - \phi_0 - k\omega_m \tau/2 \right) \right\}
$$

(2.40)

The frequency components at $k \cdot \omega_m \pm \omega_d$ have an equal power. The spectral envelope of the converted signal frequency spectrum exhibits two maxima: a maximum around $\omega_1$ and a
maximum around $\omega_2$. The widths of the main lobes around $\omega_1$ and $\omega_2$ are equal and given by:

$$
\Delta \omega_1 = \Delta \omega_2 = \frac{4 \cdot \pi}{T_m} = 4 \cdot \omega_m
$$

(2.41)

The lobes are twice as broad as in the case of sawtooth FM.

If we assume that the target is stationary, the most powerful frequency component is a multiple of $f_m$; suppose this is $k \cdot f_m$. Using equation 2.38, we acquire an expression for the range $R$:

$$
\tau = \frac{2 \cdot R}{c} \Rightarrow R = \frac{c \cdot \tau}{2} = \frac{c \cdot k \cdot f_m}{4 \cdot \Delta f \cdot f_m} = \frac{c \cdot k}{4 \cdot \Delta f}.
$$

(2.42)

It follows that:

$$
R_{\text{min}} = \frac{c}{(4 \cdot \Delta f)}.
$$

(2.43)

This means the range measurement when using a triangle FM is twice as good compared to sawtooth FM. This does not mean that the range resolution is also twice as good. As the main lobe when using triangle FM is twice as broad, the range resolution is practically equal compared to sawtooth FM.

Modulation function choice summary

We conclude by giving the advantages and disadvantages of the chosen symmetrical triangle modulation function over the sawtooth modulation function. The symmetrical triangle modulation function has components $k \cdot \omega_m \pm \omega_d$ with identical power, enabling an easy and accurate way of measuring the Doppler frequency shift. Moreover, a sawtooth modulation function would need twice as much bandwidth in order to have the same range measurement accuracy of a symmetrical triangle modulation function. As the bandwidth is limited in our case, this is a major advantage. Finally, there are no regions of infinite steepness, making the waveform much easier to approximate in a real electronic circuit. The disadvantage of the chosen modulation function is that the main lobe is twice as broad as in the case of the sawtooth modulation function. As a consequence, the maximum of the spectral envelope of the converted signal power spectrum is less distinct. Moreover, it will be impossible to tell whether the target is approaching or receding because both components $k \cdot \omega_m \pm \omega_d$ have equal power. This information, however, can also be extracted by repeated measurements of the target range.

The only thing left to choose is the modulation frequency. In order to choose a good modulation frequency, other considerations still need to be made. The modulation frequency will be finally decided on in subsection 2.4.6.

2.4.4 Performance limitations of linear FMCW radar systems

Direct transmitter/receiver leakage

The directly leaked signal has already been mentioned a few times. As the transmitter is always active in the case of CW radar, it is unavoidable that the receiver directly receives a fraction of
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the transmitted power. Generally, this is even the strongest signal at the receiver input. The RX antenna directly receiving a portion of the signal power that is transmitted by the TX antenna is normally the dominant leakage mechanism. Another mechanism is crosstalk on the PCB, but this is usually secondary to antenna leakage.

This directly leaked signal is the most important performance limitation for CW radar systems. It poses two threats.

Firstly, the relatively strong directly leaked signal may overload the receiver. The directly leaked signal, generally the strongest signal the system receives, is always present with approximately the same power. The echo signal of the radar target varies in power, depending on the target range. The directly leaked signal and the target echo signal will have a different frequency, because of the FM of the transmission, and the different propagation time of the directly leaked signal and the echo signal. Despite this frequency difference, it is impossible to separate these signals by filtering, because both the directly leaked signal and the echo signal are frequency-modulated, with a frequency swing that is much greater than the frequency difference. Hence, when amplifying the weak echo signal, the directly leaked signal will unavoidably be amplified as well. The power of the directly leaked signal limits the amount of amplification of the received signal that can be performed without overloading the input of the mixer. At the output of the frequency mixer, the directly leaked signal will be visible in the converted signal spectrum as a target at a very close range. Now it will be possible to suppress the low frequency components caused by the directly leaked signal, prior to extra amplification.

A second threat the directly leaked signal poses is reduced sensitivity due to masking of the radar targets by transmitter phase noise. To explain this, we look at the converted signal, restricted to a time interval where the FM function is rising, say \([-T_m/4, T_m/4]\). During this interval, the converted signal frequency spectrum contains two frequency components: \(\omega_{\text{direct}} = 2 \cdot \Delta \omega \cdot \tau_1 / T_m\) because of the directly leaked signal with a very small propagation delay \(\tau_1\), and \(\omega_{\text{target}} = |\omega_d + 2 \cdot \Delta \omega \cdot \tau_2 / T_m|\) because of a radar target at range \(R\) and propagation delay \(\tau_2 = 2 \cdot R/c > \tau_1\). As the directly leaked signal is usually much stronger than the echo signal, and the echo signal is at a low offset frequency with respect to the directly leaked signal, it is possible that the echo signal is masked by the phase noise accompanying the directly leaked signal. This is depicted in Fig. 2.7. The instantaneous frequency component caused by the target has a lower SNR due to the phase noise of the transmitter.

Because of transmitter phase noise correlation in time, the phase noise accompanying the directly leaked signal is reduced for short range targets, as described in [5, p. 45]. The expression for the transmitted signal including phase noise is given by:

\[
\begin{align*}
    u_{TX}(t) = A \cdot \cos (\phi_{TX}(t) + \phi(t)).
\end{align*}
\]  

(2.44)

The phase noise \(S_\phi(f)\) is defined as one half of the PSD of \(\phi(t)\). At first sight, it might be difficult to relate the definition of phase noise to an observable quantity. An alternative, more intuitive definition of phase noise \(S(f)\) of an oscillator is the ratio of the noise power in a 1Hz
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Fig. 2.7: Radar target masking by directly leaked signal phase noise

bandwidth at an offset frequency $f$ away from the carrier frequency, to the entire signal power. It can be shown [5] that $S_\phi(f) = S(f)$ for offset frequencies that are high enough. This transmitted signal is mixed with the received directly leaked signal, having a delay $\tau_1$, and whose expression including phase noise is given by:

$$u_{RX}(t) = B \cdot \cos(\phi_{RX}(t) + \phi(t - \tau_1)).$$ \hspace{1cm} (2.45)

The expression for the resulting converted signal is given by:

$$u_c(t) = C \cdot \cos(\phi_{TX}(t) - \phi_{RX}(t) + \phi(t) - \phi(t - \tau_1)).$$ \hspace{1cm} (2.46)

The phase noise $S_{\Delta\phi}(f)$ of the converted signal is consequently given by one half of the PSD of $\Delta\phi(t) = \phi(t) - \phi(t - \tau_1)$, and is related to the phase noise of the transmitter $S_\phi(f)$ in the following way [5] p. 46):

$$S_{\Delta\phi}(f) = 4 \cdot S_\phi(f) \cdot \sin^2(\pi \cdot l \cdot f/c),$$ \hspace{1cm} (2.47)

with $l = \tau_1 \cdot c$. The sine factor will significantly reduce the phase noise for low offset frequencies up to $f = \arcsin(0.5) \cdot c/(\pi \cdot l)$. This effect increases with decreasing $l$. For higher frequencies, the phase noise is sometimes reduced and sometimes enhanced depending on frequency. If the system is designed in such a manner that the delay from the FM generator to the mixer, and the delay of the directly leaked signal from the FM generator to the mixer are equal, the phase noise is theoretically perfectly cancelled.

Another remark we can make here, is that in the case of FMCW radars, increasing the transmit power by amplification will not necessarily extend the radar operational range. Simplifying things, we can say that for every decibel the transmit power is increased, the echo signal power of a certain target will also increase by one decibel. However, the phase noise power that leaks into the receiver will also increase by one decibel, because it is also amplified. If the phase noise at a certain frequency is already stronger than the other types of noise (thermal,
flicker, shot noise) that do not increase with the increased transmitter power, the SNR at that frequency cannot be increased by increasing the transmitter power. In this case, it can only be increased by reducing the phase noise of the oscillator at that frequency, or by increasing the transmitter/receiver isolation.

With regard to the masking of targets by transmitter phase noise, we note that the FMCW radar has the advantage over unmodulated CW radar. For FMCW radar systems in general, a more distant target, resulting in a weaker echo signal, will be detected as higher frequency components in the converted signal. Being at a higher offset frequency with respect to the directly leaked signal, they are less affected by the phase noise, because it is weaker there. In the case of unmodulated CW radar, this detection frequency is not related to the target range and echo signal power.

**Frequency modulation linearity**

In subsection 2.4.3 an analysis of the spectrum of the converted signal was performed. This analysis assumes perfectly linear FM functions. In reality, the FM can only be linear to a certain degree. The degree of linearity of the FM is a second major performance limitation for linear FMCW radar systems.

When the analysis of the converted signal is made taking the non-linearity of the frequency sweep into consideration, one will come to the conclusion that the range resolution of the system is not only limited by the total bandwidth $\Delta f$ as in formula 2.33 but also by the FM non-linearity. This can be intuitively understood by considering the instantaneous frequency of the converted signal. In the case of a perfect linear modulation, the instantaneous frequency will be constant during large intervals of each modulation period. If the modulation is not perfectly linear, this normally constant instantaneous frequency is spread in a way related to the shape of the non-linearity. As it is through the linear dependence of these instantaneous frequency on target range that we calculate the target range, this spread will result in a worse range resolution.

In [7, Appendix D], the special case of a quadratic chirp non-linearity was analysed. The expression for the instantaneous transmitted frequency affected by quadratic chirp non-linearity, within an interval in which the frequency is swept upwards, is given by:

$$\omega_{TX}(t) = \omega_c + A \cdot t + B \cdot t^2. \quad (2.48)$$

The conclusion of the analysis is that the range resolution, when only considering the modulation non-linearity, is given by:

$$\Delta R = R \cdot \text{Lin}, \quad (2.49)$$

with $R$ being the considered range, and $\text{Lin}$ the linearity of the sweep, defined as:

$$\text{Lin} = \frac{S_{\text{max}} - S_{\text{min}}}{S_{\text{min}}}, \quad (2.50)$$
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with $S_{\text{min}}$ being the minimal FM slope in Hz/s and $S_{\text{max}}$ the maximal FM slope. This means that the range resolution is lower for distant targets.

The range resolution of an FMCW radar is therefore limited by two parameters. First, by the frequency bandwidth, which in our case is limited to 100MHz, and second, by the frequency chirp linearity. When assuming symmetrical triangle FM and a quadratic FM non-linearity, the combined effect can be summarised in the following formula:

$$\Delta R = \sqrt{\left(\frac{3 \cdot c}{4 \cdot \Delta f}\right)^2 + (R \cdot \text{Lin})^2}$$

(2.51)

At short range the frequency bandwidth limitation will dominate and at greater range, the chirp non-linearity will dominate. In our case the transmitter power is limited to 20dBm EIRP, limiting the range of the radar to a certain unknown range (unknown because several parameters like transmitter-receiver isolation, receiver chain noise figure, etc. are still unknown). The ideal case would be that the range resolution in the entire operational range of the radar is limited by the frequency bandwidth only and not the frequency chirp linearity, giving the radar system the maximum range resolution possible.

The degree of linearity of the FM depends on the implementation. There are several ways of implementing this frequency-modulated generator as described in [7, Appendix E.]. These different implementations mainly trade off frequency chirp linearity versus simplicity and cost.

The most simple design consists of a voltage controlled oscillator (VCO) with a tuning voltage provided by a linear triangular voltage oscillator. This is schematically presented in Fig. 2.8. The triangular voltage oscillator can be constructed with an operational amplifier (op-amp) circuit for instance. The average voltage and the peak-to-peak voltage of the generated triangular voltage are chosen so that the VCO output frequency sweeps the correct frequency interval. The generated frequency chirp will only be linear if the tuning voltage characteristic of the VCO is linear in the considered frequency interval. If the VCO tuning characteristic is not linear (enough), the tuning voltage can be predistorted so that the combination of the modulating oscillator and the VCO has a linear chirp characteristic. This complicates the implementation of the triangular voltage oscillator.

Another option is to use a direct digital synthesis (DDS) oscillator to generate the linear frequency chirp, and then up-convert this DDS output signal to the wanted centre frequency.

These mentioned implementations are all open loop configurations meaning that there is no feedback loop from the output to the input to improve the linearity. The advantage of an open loop configuration is its simplicity, its disadvantages are less linearity and temperature and ageing instability.

Closed loop configurations are also proposed in [7, Appendix E.]. They have a feedback loop that constantly monitors the linearity of the output signal and generates a correction signal that is added to the tuning voltage. This greatly improves the linearity of the frequency chirp, making it robust to temperature changes and ageing. This, however, makes the generator much more complicated.
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![Diagram of an open loop FM generator: VCO tuned by a triangle wave oscillator]

Finally, the most simple configuration, presented in Fig. 2.8 was chosen, where a VCO has its tuning voltage controlled by a triangular oscillator, because it was expected that the frequency chirp would be linear enough for the range resolution to be limited only by the frequency bandwidth. In this way, the design of the FM generator falls apart in two separate problems: firstly the choice of an appropriate VCO component, and secondly, the design of a linear triangular voltage oscillator.

2.4.5 Radar operational range

Upper limit

First, an upper limit on the radar operational range will be calculated. Fig. 2.4c is instructive. The assumptions that will lead to this upper limit are:

- Transmitted power $P_{TX} = 20\, dBm$,
- The directly leaked signal is ignored,
- Ideal receiver having a 0dB noise figure, and lossless components,
- TX and RX antennas have a gain $G_{TX} = G_{RX} = 0\, dB$ in the direction of the target,
- Only thermal noise is considered. The RX antenna acts as a 50$\Omega$ resistor with a noise temperature of 298K,
- Target radar cross section $\sigma = 0.1\, m^2$,
- Received power $P_{RX}$ can be calculated using the radar equation of equation 2.4,
- The range at which the SNR of the converted signal $SNR_r$ equals 0dB will be considered the maximum range $R_{max}$.

The received echo signal power available from the RX antenna of a radar target at the maximum range is given by the radar equation:

$$P_{RX} = P_{TX} \cdot G_{TX} \cdot \sigma \cdot G_{RX} \cdot \lambda^2 \cdot (4\pi^3 \cdot R_{max}^4)$$

(2.52)

This signal is mixed with a sample of the transmitted signal. If we assume that the mixer has no conversion loss, $P_{RX}$ also gives the power of the converted signal.
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The PSD of the thermal noise (dBm/Hz) available from the RX antenna is given by:

\[ \text{PSD}_{\text{th}} = k \cdot T = -173.85 \text{dBm/Hz}, \]  

(2.53)

The pass band of the filter after the frequency mixer will be chosen as narrow as possible to reduce the thermal noise power and maximise the SNR of the converted signal. As no directly leaked signal component has to be suppressed here, the band pass filter can be changed to a low pass filter with a bandwidth \( B \). The highest frequency components that can be expected in the converted signal spectrum correspond to the maximal target range. The maximum of the spectral envelope caused by a target at range \( R_{\text{max}} \) is at a frequency \( f_{\text{max}} \) given by equation 2.38. The cut off frequency of the low pass filter has to be higher than this value in order to be able to detect a maximum in the spectral envelope. For simplicity, however, we choose:

\[ B = 2 \cdot \Delta f \cdot \frac{2 \cdot R_{\text{max}}}{c} \cdot f_m. \]  

(2.54)

We have to take into account that the frequency mixer will convert both frequencies \( f_{LO} + f \) and \( f_{LO} - f \) to a frequency \( f \) at its output. Therefore, the noise PSD at its output is doubled (or in other words: the single sideband noise figure of a frequency mixer that does not add excess noise is equal to 3dB).

Using equations 2.52, 2.53 and 2.54, we express that the SNR \( SNR_r \) of the converted signal of a target at maximal range equals 0dB:

\[ SNR_r = 0dB = \frac{P_{RX}}{2 \cdot B \cdot \text{PSD}_{th}} = P_{TX} \cdot \frac{G_{TX} \cdot \sigma \cdot G_{RX} \cdot \lambda^2}{(4\pi)^3 \cdot R_{\text{max}}^4 \cdot 2 \cdot 2 \cdot \Delta f \cdot \frac{2 \cdot R_{\text{max}}}{c} \cdot f_m \cdot \text{PSD}_{th}}. \]  

(2.55)

In this last equation, another influence of the modulation frequency choice is visible. When the modulation frequency is increased, the frequency at which a fixed target is detected increases as well. The bandwidth of the receiver has to be made larger, increasing the noise power in the converted signal and reducing the SNR of the converted signal. Consequently, the maximum range decreases.

If the modulation frequency is chosen to be 1kHz, and the FM sweep has a bandwidth of \( \Delta \omega = 2\pi \cdot 100MHz \), the maximum range is given by:

\[ R_{\text{max}} = 92.78m. \]  

(2.56)

If the modulation frequency is chosen to be 10kHz:

\[ R_{\text{max}} = 58.54m. \]  

(2.57)
More accurate SNR

When the components are chosen, a more accurate estimate of the SNR will be possible. Here, the component properties will be treated as parameters. Again, Fig. 2.4c is instructive. The following assumptions are made:

- Transmitted power $P_{TX} = 20dBm$ EIRP,
- The directly leaked signal is taken into account: an isolation $I$ between the TX and RX antenna is assumed ($S_{12}$),
- The RX amplifier has a gain $G_{RXA}$ and a noise figure $NF_{RXA}$,
- The frequency mixer has a conversion loss $CL$ and a single sideband noise figure $NF_{mixer}$,
- The RX antenna has a gain $G_{RX}$ in the direction of the target,
- Only thermal noise is assumed. The RX antenna acts as a 50Ω resistor with a noise temperature of 298K,
- Target radar cross section $\sigma = 0.1m^2$,
- Received power $P_{RX}$ can be calculated using the radar equation of equation 2.4.

The receive chain formed by the RX amplifier and the frequency mixer has a total gain of $G_{RXA} - CL$. The converted signal has a component caused by the directly leaked signal, and a component caused by the received echo signal of the target. Both components are amplified by the RX amplifier and converted to baseband by the frequency mixer. Thus, the component caused by the directly leaked signal has a power equal to:

$$P_{leak} = 20dBm - I + G_{RXA} - CL.$$  \hfill (2.58)

The received echo signal power available from the RX antenna is given by equation 2.4, the radar equation. Thus, the component in the converted signal caused by the echo signal has a power of:

$$P_{echo} = P_{TX} + 10 \cdot \log \sigma + G_{RX} + 20 \cdot \log \lambda - 30 \log (4\pi) - 40 \log R + G_{RXA} - CL$$  \hfill (2.59)

The noise figure of a chain of devices can be calculated with the Friis formula:

$$F_{total} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1G_2} + \ldots + \frac{F_n - 1}{G_1G_2\ldots G_{n-1}}$$  \hfill (2.60)

with $F_n$ and $G_n$ respectively being the noise factor and the gain of the $n^{th}$ device in the chain. The noise figure is the noise factor expressed in decibel. We learn from the Friis formula that the noise factor of the chain of devices is at least larger than the noise factor of the first device, so the first device in the chain should have a low noise figure. We also notice that the SNR degradation
(the noise factor) of all devices following the first device is reduced by the first device’s gain. However, if the first device has negative gain, or attenuates, the first device increases the SNR degradation introduced by the following devices. This means that, in order for the receive chain noise figure to remain low, the received signal should not be applied directly to the RF port of the frequency mixer, because it introduces losses (conversion loss) and has a high noise figure. Therefore, a LNA is used as the first device in the receive chain. The noise figure of the receive chain for our situation is given by:

\[ NF_{\text{total}} = 10 \cdot \log \left( F_{RXA} + \frac{F_{\text{mixer}}}{G_{RXA}} - 1 \right), \] (2.61)

with \( F_{RXA} \) being the noise factor of the receive amplifier, and \( F_{\text{mixer}} \), the single sideband noise factor of the frequency mixer. The PSD (dBm/Hz) of the thermal noise available from the RX antenna is given by equation 2.53. The thermal noise PSD at the output of the frequency mixer is given by:

\[ PSD_{\text{th,conv}} = -173.85 \text{dBm/Hz} + G_{RXA} - CL + NF, \] (2.62)

The total noise power is given by:

\[ P_{\text{th,conv}} = PSD_{\text{th,conv}} \cdot B, \] (2.63)

with B being the bandwidth of the band pass filter. The upper cut off frequency of the band pass filter after the frequency mixer has to be chosen as low as possible to reduce the thermal noise power and maximise the SNR of the converted signal. The lower cut off frequency of the band pass filter can be chosen high enough in order to suppress the components caused by direct leakage, increasing the SNR.

Finally, the expression for the SNR is given by:

\[ \text{SNR}_{\text{conv}} = \frac{P_{\text{echo}}}{P_{\text{leak}} + P_{\text{th,conv}}}. \] (2.64)

### 2.4.6 Modulation frequency

The different influences of the modulation frequency \( f_m \) were discussed throughout this chapter.

First, the choice of modulation frequency will limit the maximum Doppler shift that can be reliably measured, and therefore also the maximum radial speed that can be measured. As the supposed radar targets are humans, radial speeds up to 10m/s should be measurable without any problems. This imposes a lower limit on the modulation frequency:

\[ |f_d| < f_m/2 \iff 2 \cdot 2.45 \text{GHz} \cdot \frac{2 \cdot 10 \text{m/s}}{c} = 326.66 \text{Hz} < f_m. \] (2.65)

Moreover, when choosing a higher modulation frequency, the system will need less time to process one measurement. One measurement of target range and velocity at time \( t_0 \) consists of the spectral estimation of a fragment centred on time \( t_0 \) with a certain length. For correct estimation of the power spectrum of a periodic signal, an adequate number of periods has to
be present in the signal fragment. The period here is the modulation period. In this signal fragment, the target range and radial velocity should vary negligibly, as supposed in subsection 2.4.3. The only way of achieving both is by choosing a sufficiently high modulation frequency.

At the same time, if the modulation frequency increases, the bandwidth of the signal processing circuit has to increase as well, increasing the total noise power. However, the noise PSD remains equal, thus increasing the modulation frequency is a good thing.

Finally, the most important limitation we have to consider is that the analysis of the converted signal spectrum in subsection 2.4.3 was performed assuming $\tau \ll T_m$. This puts an upper limit on the modulation frequency.

Taking all this into consideration, the modulation frequency was chosen to be 10kHz. A time fragment of 100 periods will then be 10ms. During this time no considerable change in target range or radial velocity for normal humans can occur. An upper limit on the radar operational range was estimated to be 58.54m in subsection 2.4.5. The corresponding delay is $3.9 \cdot 10^{-4} ms$, definitely ensuring the inequality $\tau \ll T_m$.

### 2.4.7 Band pass filter

In this subsection, an appropriate choice for the pass band of the band pass filter in the FMCW block diagram of Fig. 2.4c is made.

First, we consider the choice of the upper cut off frequency. The upper cut off frequency has to be chosen as low as possible to reduce the thermal noise power, without suppressing the useful converted signal. In the case of a modulation frequency of 10kHz, the upper limit on the radar operational range was calculated to be 58.54m. A radar target that is present at this maximum range will give rise to a maximum of the spectral envelope at a frequency $2 \cdot \Delta f \cdot \frac{2R_{\text{max}}}{c} \cdot f_m = 780.53 kHz$. Naturally, the upper cut off frequency has to be higher than this value, in order to be able to detect a maximum in the spectral envelope. Therefore, we round the value of 780.53kHz to 1MHz. For the rest of this text, an upper cut off frequency of 1MHz is assumed for the band pass filter.

The lower cut off frequency of the band pass filter can be chosen high enough in order to suppress the contribution of the directly leaked signal to the converted signal power spectrum. As the directly leaked signal has a very short delay, it is expected that the directly leaked signal gives rise to a maximum of the converted signal spectral envelope at DC. It is therefore decided that the DC component should be in the stop band, and the first harmonic of 10kHz in the pass band. This means the lower cut off frequency is chosen to be between DC and 10kHz. In this way, the band pass filter practically degrades to a low pass filter, and it is called a low pass filter for the rest of this dissertation.
2.4.8 Signal processing circuit

As the proposed method of acquiring the range and velocity information of the radar target is spectral, the signal processing circuit should be some form of a spectrum analyser. When integrating the spectral analysis of the converted signal in the radar system itself, the most simple way of doing this is by sampling the converted signal and using a microcontroller to analyse it with the fast fourier transform (FFT). This, however, is beyond the scope of this thesis. When doing measurements on the prototype board of the radar system, a spectrum analyser measuring the output of the mixer will be used to verify the correct operation of the system.
Chapter 3

Full-wave/circuit co-design of the radar

In the previous chapter we decided on the system architecture and its details. In this chapter electronic circuitry and antennas that implement this architecture will be designed. We will decide for every block of the system block diagram how it will be implemented and what components will be used for it in section 3.1. In this way we will come to a complete schematic. Subsequently, a PCB layout will be designed and partially simulated in section 3.2. In the last section 3.3 the antenna type and topology will be chosen, and the transmit and receive antennas will be designed.

3.1 Circuit design and component selection

We repeat the block diagram of the system architecture in Fig. 3.1. The different system blocks will now be further examined and translated to electronics. A 5V power supply is preferred for all components.

3.1.1 VCO and triangle wave oscillator

The intended functionality of these blocks is to generate a CW signal with a frequency that is modulated with the symmetrical triangle function described in subsection 2.4.3. The frequency has to be linearly swept between 2.4GHz and 2.5GHz at a modulation frequency of 10kHz. The signal of this generator will be used as an LO signal for the frequency mixer, and as the transmitted signal.

VCO component

To select the VCO component, we consider the supply voltage, the frequency band, tuning linearity, the phase noise, the harmonics levels and the output power. The chosen VCO component is the JTOS-2700V+ [8] component of Minicircuits.
The device has a supply voltage of 5V and draws a current of about 15.3mA when operating around 2.45GHz.

According to the datasheet the tuning characteristic is "linear" in the frequency interval ranging from 2050MHz to 2700MHz and the tuning sensitivity stays between 57MHz/V and 62MHz/V when operating between 2.4GHz and 2.5GHz.

As explained in subsection 2.4.4, the phase noise accompanying the directly leaked signal may render weak radar target reflections invisible. Therefore, we have to check whether the phase noise of the chosen VCO component is low enough to avoid this situation.

To examine this, the situation of Fig. 3.2 is considered. The FM signal with a power of 20dBm is transmitted with the 0dBi TX antenna. Part of the transmitted power unintentionally directly leaks to the 0dBi RX antenna. The isolation is assumed to be 50dB as indicated in the figure. As a result, the power of the directly leaked signal available from the RX antenna equals -30dBm. Another part of the transmitted power is sent back to the radar system by a stationary reflector at range R (radar cross section $\sigma = 0.1m^2$). The echo signal power $P_{RX}$ available from the RX antenna is given by equation 2.4:

$$P_{RX} = P_{TX} \cdot \frac{G_{TX} \cdot \sigma \cdot G_{RX} \cdot \lambda^2}{(4\pi)^3 \cdot R^4} = P_{TX} \cdot \frac{0.1m^2 \cdot \lambda^2}{(4\pi)^3 \cdot R^4},$$ (3.1)

with $P_{TX}$ being the transmitted signal power of 20dBm.

If the converted signal is observed restricted to a time interval where the FM function of Fig. 2.5c is rising, say $[-T_m/4, T_m/4]$, the converted signal frequency spectrum contains two frequency components. The first component $f_{direct}$ is caused by the directly leaked signal and
is given by:

\[ f_{\text{direct}} = 2 \cdot \frac{\Delta f}{T_m} \cdot \tau_1, \quad (3.2) \]

with \( \tau_1 \) being the very small propagation delay of the directly leaked signal. We are interested in the power spectral density of the phase noise accompanying this frequency component ("directly leaked phase noise"). The second component \( f_R \) is caused by the reflector at range \( R \) and is given by:

\[ f_R = 2 \cdot \frac{\Delta f}{T_m} \cdot \frac{2R}{c}. \quad (3.3) \]

As we have the intention to compare the power spectral density of the directly leaked phase noise to the power of the frequency component \( f_R \) and the conversion loss of the frequency mixer affects both these quantities equally, the conversion loss is arbitrarily assumed to be 0dB.

As \( f_{\text{direct}} \) usually is very small compared to \( f_R \), the frequency difference between the two components approximately equals \( f_R \). In other words, the frequency component caused by the reflector is approximately at an offset frequency \( f_R \) with respect to the frequency component caused by the directly leaked signal.

Combining equations 3.1 and 3.3 by eliminating the range \( R \), an expression of the echo signal power as a function of offset frequency \( f_R \) is obtained. A plot of this relation is presented in Fig. 3.3.

As discussed in subsection 2.4.4, the directly leaked phase noise PSD is given by equation 2.47 (at an adequate offset frequency with respect to the carrier). In this equation, \( S_\phi(f) \) is the phase noise of the signal generated by the VCO, which is provided in the datasheet. The length \( l \) is equal to \( \tau_1 \cdot c \) and is chosen to be 1m. This is reasonable as the antenna interconnection cables of a finished compact system are expected to be much shorter than 1m. A plot of the directly leaked phase noise PSD is also presented in Fig. 3.3.

Comparing the directly leaked phase noise PSD (dBm/Hz) to the power of the frequency component caused by the reflector (dBm) over the frequency range from 10kHz to 1MHz, it is clear that the targets are essentially not masked by the phase noise. The conclusion is that the chosen VCO component has an adequate phase noise performance.

The output power of the VCO in the frequency band 2.4GHz-2.5GHz is 7.06dBm-7.64dBm. Part of this power has to be used to drive the mixer LO pin and the other part has to be transmitted. A power splitter is used to split up the generated signal. The used power splitter is the BP2U1+ [9] component of Minicircuits. The splitter insertion loss is 3.54dB at 2.45GHz, so at both splitter outputs a signal of 3.52dBm-4.1dBm is available. The output of the VCO also contains harmonics of the configured carrier frequency. The power splitter S-parameters are only specified up to 4GHz, but because the input is no longer matched at 4GHz (\(|S_{11}| = -3.031dB, |S_{21}| = -6.90dB \) and \(|S_{31}| = -6.67dB\)), it is expected that the harmonics will be attenuated by the power splitter.

We notice that when the VCO is swept from 2.4GHz to 2.5GHz, the power will also be swept from 7.06dBm to 7.64dBm. In principle, this has an effect on the spectrum of the produced
3.1 Circuit design and component selection

Fig. 3.2: Situation to estimate impact phase noise accompanying the directly leaked signal

Fig. 3.3: Directly leaked phase noise PSD (dBm/Hz) and echo signal power (dBm) as a function of offset frequency
signal, but this will be ignored here.

The presence of harmonics poses two problems. First, the harmonics might be amplified and transmitted, meaning power is transmitted in a forbidden frequency band. In this way, the radar system would possibly interfere with other devices operating in the frequency band from 4.8GHz to 5GHz or even from 7.2GHz to 7.5GHz. This can be solved by filtering the signal before transmission, or, in a simpler manner, by selecting a transmitter power amplifier that attenuates these harmonic frequencies or design the transmit antenna not to radiate at these harmonic frequencies. More will be said about this in subsection 3.1.2 when the transmitter power amplifier is selected, and in section 3.3 when the antennas are designed.

Second, if the harmonics have enough power and are still in the mixer LO and RF frequency range, they are able to drive the mixer and produce extra unwanted mixing products. This will turn out to be not much of a problem. Since the mixer IF output frequencies of interest are very low, maximally 1MHz, the only unwanted mixing products that can disturb the wanted mixing products (not considering non-linearity) are the result of mixing a harmonic frequency at the LO port with the same harmonic frequency at the RF port. These harmonic frequencies at the RF port can originate from another transmitter or from the radar system itself. We have no control over signals from other transmitters, but we can attempt to make the radar system immune to these signals by designing the receive antenna to have poor performance at the higher harmonic frequencies. Possible sources originating from the radar system itself are reflection of the transmitted harmonic frequencies by a reflector back to the radar system, the directly leaked signal and crosstalk on the PCB. By selecting transmit and receiver amplifiers that amplify the fundamental frequency more than the harmonics, the power of the harmonics will fall relative to the fundamental frequency. As the power of the mixing product is largely determined by the power of the RF signal, the unwanted mixing products can be made much weaker than the wanted mixing product in this way. More will be said about this in subsection 3.1.2 about the transmit amplifier and subsection 3.1.4 about the receiver amplifier. Moreover, the power of the harmonic frequencies at the mixer LO port is at least 18.8dB weaker (VCO and power splitter datasheet) than the fundamental frequency, increasing the mixer conversion loss and further weakening the unwanted mixing products. Furthermore, we can expect that the TX/RX isolation is higher at the harmonic frequencies than at the fundamental frequency because the transmit and receive antennas will be designed for the 2.4GHz-2.5GHz frequency band.

**Linear triangular voltage oscillator**

Now we will construct the voltage oscillator that will drive the tuning pin of the VCO. As the tuning characteristic of the VCO is already linear, we will construct a linear triangular voltage oscillator. To sweep the VCO from 2.4GHz to 2.5GHz, a voltage sweep from about 9V to 11V is needed. It was chosen to implement this with an op-amp circuit. Sweeping from 9V to 11V with a 5V power supply is impossible, so the op-amp has to be fed by a higher voltage. The
op-amp chip that was selected is the general purpose TL-082 op-amp of Texas Instruments, which contains two op-amps with a +/-15V power supply. The total circuit is shown in Fig. 3.4. It consists of an integrator and a Schmitt trigger.

The Schmitt trigger is indicated in Fig. 3.4 as part a. Because of positive feedback, the only stable situations are when the op-amp is in positive ($V_{sat}$) or in negative saturation ($-V_{sat}$). It can easily be shown that the switching thresholds of this Schmitt trigger are given by:

$$V_{\text{switch},1} = V_{\text{bias}} \frac{R_1 + R_2}{R_2} + V_{sat} \frac{R_1}{R_2} \tag{3.4}$$

and

$$V_{\text{switch},2} = V_{\text{bias}} \frac{R_1 + R_2}{R_2} - V_{sat} \frac{R_1}{R_2}. \tag{3.5}$$

When the input voltage $V_{\text{Schmitt,in}}$ exceeds $V_{\text{switch},1}$, the Schmitt trigger saturates to a positive output $V_{sat}$. When the input voltage $V_{\text{Schmitt,in}}$ goes below $V_{\text{switch},2}$, the Schmitt trigger saturates to a negative output $-V_{sat}$.

The linear integrator is indicated in Fig. 3.4 as part b. As negative feedback is used, the nullator hypothesis can be used. In this way we can prove that the output of the integrator $V_{\text{int,out}}(t)$ is given by:

$$V_{\text{int,out}}(t) = V_{\text{int,out}}(0) - \int_0^t \frac{V_{\text{int,in}}(u)}{R_4C} du, \tag{3.6}$$

with $V_{\text{int,in}}(t)$ being the input of the integrator.

The output of the Schmitt trigger is not directly applied to the input of the integrator, but is clamped by a resistor and two Zener diodes. As the output voltage of the Schmitt trigger is always $+V_{sat}$ or $-V_{sat}$, there is always one Zener diode in forward conduction, and one in reverse conduction. In this way, the output voltage is clamped to $\pm V_{\text{clamp}} = \pm (V_{zener} + V_{\text{forward}})$, with $V_{zener}$ being the Zener diode breakdown voltage and $V_{\text{forward}}$ the diode voltage in forward conduction. The applied Zener diodes are Panasonic DZ2S051 types [12]. The nominal Zener voltage is 5.1V (good thermal stability) and the power rating is 125mW (so the maximum reverse current is about 125mW/5.1V=24.5mA). In order for the clamp circuit to work correctly, the current through the Zener diode has to lie in a region of low dynamic resistance. The current has to be high enough in order to allow for current to be delivered to the integrator while keeping the Zener diode current high enough to provide a stable regulated voltage, and the current has to be low enough in order to avoid needless power dissipation. The recommended reverse current in the datasheet is 5mA. This current will be set by the resistor $R_3$. The absolute value of the reverse current $I_{zener}$ through the Zener diode will be equal to:

$$|I_{zener}| = \frac{V_{\text{sat}} - (V_{zener} + V_{\text{forward}})}{R_3}. \tag{3.7}$$

The value of $V_{\text{sat}}$ also depends on the load resistance. A simulation with the Simulation Program with Integrated Circuit Emphasis (SPICE) models of the op-amp and the Zener diodes shows
3.1 Circuit design and component selection

Fig. 3.4: Linear triangular voltage oscillator
3.1 Circuit design and component selection

that a current of 5mA requires $R_3$ to be 1.5kΩ. The voltage is then clamped to a value of $\pm V_{\text{clamp}} = +/ - 5.92V$ and the op-amp saturation voltage $V_{\text{sat}}$ is equal to 13.39V.

Together, these two parts work as a triangular oscillator, with the oscillator output being the output of the integrator. The clamped output of the Schmitt trigger is the input of the integrator, and the output of the integrator is the input of the Schmitt trigger. If the Schmitt trigger is in its high output state, the integrator output voltage will fall linearly, eventually switching the Schmitt trigger to its low output state. The integrator output will then start to rise linearly, eventually switching the Schmitt trigger to its high output state, and so on. The voltage will sweep between the Schmitt trigger switching thresholds given in equation 3.4 and 3.5. In order to have switching threshold voltages of 9V and 11V we need an average voltage of 10V and a voltage swing of 2V:

$$V_{\text{switch},2} - V_{\text{switch},1} = 2V_{\text{sat}} \frac{R_1}{R_2} = 2V \iff 2 \cdot 13.39V \frac{R_1}{R_2} = 2V \iff \frac{R_2}{R_1} = 13.39$$  \hspace{1cm} (3.8)

$$\frac{V_{\text{switch},1} + V_{\text{switch},2}}{2} = V_{\text{bias}} \frac{R_1 + R_2}{R_2} = 10V \iff V_{\text{bias}} \left( \frac{R_1}{R_2} + 1 \right) = 10V \iff V_{\text{bias}} = \frac{10V}{1 + \frac{13.39}{1}} = 9.31V$$  \hspace{1cm} (3.9)

For $R_1$ and $R_2$ we choose 2kΩ and 26.78kΩ, respectively. For resistor $R_2$, a trimmer will be used so the difference between the switching thresholds can be accurately set. The bias voltage $V_{\text{bias}} = 9.31V$ is set with a voltage divider. For $R_5$ and $R_6$ we choose 1MΩ and 1.63MΩ respectively. The bias voltage influences the average of the switching thresholds. Resistor $R_6$ is also a trimmer resistor, in order to be able to accurately set the average of the two switching thresholds.

The frequency of the oscillation is given by:

$$f = \frac{V_{\text{clamp}}}{2(V_{\text{switch},2} - V_{\text{switch},1})R_4C}.$$  \hspace{1cm} (3.10)

The current that is drawn by the integrator should be low enough in order for the Zener diode to stay in reverse operation. $R_4 = 148k\Omega$ and $C = 1nF$ gives a frequency of 10kHz. The magnitude of the current drawn by the integrator $I_{\text{abs,integrator}}$ equals:

$$I_{\text{abs,integrator}} = \frac{5.92V}{148k\Omega} = 0.04mA.$$  \hspace{1cm} (3.11)

This is much lower than the 5mA that flows through the Zener diodes without the integrator, so the voltage stays regulated correctly.

When performing a SPICE simulation of the linear triangular oscillator with the mentioned component values and the SPICE models of the op-amp and the Zener diodes, a voltage sweep of 8.85V to 11.13V is observed, and a period of 0.115ms.

The +/-15V are generated with a boost converter that is fed with 5V. The used boost converter is the NTA0515MC [13] component of Murata Power Solutions. The decoupling capacitors
and choking coils recommended in the datasheet were added to the circuit. This boost converter can deliver an output current up to 33mA, and it has to be loaded at least 10% (3.3mA) in order to work correctly. This is the case as the current delivered by the output of the op-amp is 5mA. The switching frequency of the boost converter is 115kHz. This falls in the frequency range of interest of the mixer IF output, so we will have to make sure that the boost converter does not destroy the sensitivity of the system.

3.1.2 Transmitter power amplifier

We are allowed to transmit up to 20dBm EIRP in the ISM band. The power of the signal at the outputs of the power splitter is only approximately 3.52dBm-4.1dBm, so a power amplifier will be needed.

The most important properties of a power amplifier are gain, the frequency interval, 1dB output compression point, noise figure and output third order intercept point.

The gain has to bring the output power close to 20dBm (might be a few dB less, because the transmit antenna will also have gain). Preferably, the harmonic frequencies are attenuated, or at least significantly less amplified than the fundamental frequency. The 1dB output compression point has to lie beyond 20dBm to ensure that the amplifier stays in its linear region. The noise figure is of less importance here because the SNR is already high.

The chosen power amplifier is the MGA-30689 [14] component of Avago. It has a 5V power supply and draws a current of 105mA. It has a gain of 14.75dB, a 1dB output compression point of 22.5dBm and a noise figure of 3.5dB around 2.45GHz. Consequently, the output power lies in the 18.27dBm-18.85dBm range and the amplifier stays sufficiently far from its non-linear region.

A plot of the gain $|S_{21}|$ as a function of frequency is provided in Fig. 3.5. The nominal bandwidth is from 40MHz to 4GHz, but signals up to 5.14GHz are still amplified. The second and third harmonic of 2.45GHz are respectively amplified by 1.40dB and attenuated by 11.85dB. At the second and third harmonic, the in- and output are also no longer matched to 50Ω as shown in Fig. 3.6. This means that the harmonic overtones in the frequency spectrum of the signal applied to the transmit antenna are at least 18.8dB+14.75dB-1.40dB=32.15dB below the fundamental frequency carrier, even neglecting reflections at the power splitter input. The most powerful harmonic overtone in the frequency spectrum of the signal applied to the transmit antenna is the second harmonic with a power of -13.88dBm to -13.3dBm. Whether the power present at the harmonic overtone frequencies will actually be transmitted depends on the transmit antenna.

The device has 3 terminals: the RF input pin $RF_{in}$, the $RF_{out}$ pin and a mass pad. The $RF_{out}$ pin serves two purposes: it is both the power supply pin and the RF output pin. Both $RF_{in}$ and $RF_{out}$ are matched to 50Ω ($|S_{11}|,|S_{22}| < -10dB$) at 2.45GHz without the need of external matching components. The device is fed through a bias tee at the RF output pin. This is depicted in Fig. 3.7. The bias tee consists of the coupling capacitor $C_K$ and the inductor $L$ that are tied to the $RF_{out}$ terminal of the component. The RF signal can flow from $RF_{out}$ through the coupling capacitor $C_K$ to the output but is blocked by the choking coil L from
flowing to the DC feed network. The DC current can flow through the choking coil to feed the power amplifier, but it cannot flow to the output and disturb the next component.

$C_K$ is chosen to have a very low impedance and insertion loss at 2.45GHz. The chosen capacitor is a 8.2pF ceramic RF capacitor (251R14S8R2CV4S) from Johanson Technology with a self-resonance frequency of 2.98GHz. This capacitor has an impedance with magnitude 2.83Ω and an insertion loss of 0.026dB at 2.45GHz. The insertion loss is 0.25dB at 4.9GHz.

The inductor L is chosen to have a very high impedance at 2.45GHz. The chosen inductor is a 100nH RF inductor (LQW18ANR10G00) from Murata. The self-resonance frequency is around 2.7GHz and the insertion loss is about 37dB at 2.45GHz. The insertion loss is only 17.5dB at 4.9GHz, so the second harmonic is less blocked than the fundamental frequency.

The decoupling capacitors $C_1$, $C_2$ and $C_3$ are not essential, but they further decrease the amount of RF power that leaks into the DC feed network. Multiple capacitors are used in parallel in order to decouple over a wide frequency band: a ceramic RF capacitor of 2.7pF (251R14S2R7BV4S) to decouple around the second harmonic (self-resonance frequency 4.86GHz), a ceramic RF capacitor of 8.2pF (251R14S8R2CV4S) to decouple around the fundamental frequency and a ceramic capacitor of 2.2µF (GRM188R61A225KE34) to decouple for low frequencies (self-resonance frequency about 6 MHz).
3.1 Circuit design and component selection

Fig. 3.6: Transmitter power amplifier matching $|S_{11}|$ and $|S_{22}|$ as a function of frequency

Fig. 3.7: Transmitter power amplifier bias tee
3.1 Circuit design and component selection

3.1.3 Frequency mixer

The most important specifications of a frequency mixer are the LO drive power (“mixer level”, minimum needed LO signal power) that is needed, the conversion loss, noise figure (typically 0.5dB higher than conversion loss), the RF 1dB compression point, the third-order intermodulation distortion, the frequency range of the LO, RF and IF ports and the isolation between the mixer ports. These performance measures are defined in [15], [16] and [17].

The selected mixer is the MACA-63H+ [18] component from Minicircuits. This is a double balanced diode mixer. The required LO drive power is only 0dBm (a “level 0” mixer), but the LO power is allowed to be higher; the maximum rating is 10dBm. The LO signal in our case has a power ranging from 3.52dBm to 4.1dBm. This mixer is classified as an active mixer, as it internally amplifies the LO signal, and needs a power supply of 5V drawing 110mA. The frequency mixer can accept frequencies from 2GHz to 6GHz at the RF and LO ports, and the frequency range of the IF port extends from DC to 1GHz, which suits our application. The conversion loss for LO and RF frequencies around 2.45GHz, and an LO drive level of 3dBm is around 5.65dB. Knowing the conversion loss, we can estimate the noise figure to be 6.15dB.

For an LO drive power of 3dBm around 2.45GHz, the compression for an RF input signal of 10dBm around 2.45GHz is 0.56dB-0.77dB. The 1dB compression point usually increases with increasing LO power, and, as such, the 1dB compression point for our LO drive power of 3.52dBm-4.1dBm is certainly higher than 10dBm. When designing the RX amplifier chain, we will limit the gain so that the resulting power at the RF port of the frequency mixer stays well below this 1dB compression point. The other linearity measure, the input third order intercept point, is about 18-19dBm for an LO and RF signal around 2.45GHz and an LO power of 3dBm.

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For RF and LO frequencies around 2.45GHz and an LO signal of 3dBm, the LO-IF isolation is 33.92dB, the RF-IF isolation 50.99dB, and the LO-RF isolation 22.60dB. We have to remark that the LO-IF and RF-IF isolation figures are measured by Minicircuits with extra low pass filtering done at the IF output, so the isolation of the mixer alone will be much worse. The fact that the mixer LO-IF isolation is very low is not much of a problem. The three most important signals at the mixer IF output are the down-mixed signal extending from DC to mostly 1MHz, the up-mixed signal around 4.9GHz, and a leaked LO signal around 2.45GHz. Filtering these last two signals out will be possible.

3.1.4 Receiver low noise amplifier

The signal reflected off an object received by the RX antenna has a high dynamic range depending on the target range, as can be calculated with the radar equation. In a typical situation, a reflection of a target 1m away has a power of -41.22dBm at the output of a 0dBi reception antenna while a reflection of a target 10m away only -81.22dBm. The receive chain has to be able to deal with this. A variable gain amplifier (VGA) can be used to cope with this high dynamic range.
It is also important to keep the directly leaked signal in mind. When assuming a 50dB TX/RX isolation, the directly leaked signal has a power of -30dBm, meaning it will generally be the strongest received signal. The frequency of the directly leaked signal is almost the same as the frequency of the reflected signal, so it cannot be filtered out and it will be amplified together with the reflected signal. We have to beware that the amplifiers and frequency mixer do not saturate or operate in their non-linear region due to the strong directly leaked signal and no significant intermodulation products are created originating from the directly leaked signal and the reflected signal. We assume a 1dB compression point of the used frequency mixer of 10dBm. The power applied to the RF port of the frequency mixer has to stay sufficiently below this 10dBm. If an amplifier precedes the mixer RF port, the amplification has to stay well below 40dB.

Another important property of the receive chain is its sensitivity. In order to be sensitive, the receive chain should have a low noise figure, so the SNR is degraded as little as possible. A LNA should be the first device in the receive chain in order to achieve this.

The LNA that was chosen is the TQP3M9037 component of Triquint. This device has an operational bandwidth of 1.5GHz to 4GHz and is powered by 5V drawing 70mA of current.

For the frequencies from 2.3GHz-2.6GHz, the LNA has a very low noise figure of 0.43dB to 0.53dB, and is very linear having an OIP3 of 35.1dBm to 35dBm. At the same frequencies, the gain is 19.1dB to 18.1dB and the 1dB output compression point is 19.8dBm to 19.5dBm. When we consider the directly leaked signal of -30dBm to be the most powerful signal at the LNA input, the LNA amplifies this signal to about -11.4dBm. This is much lower than the 1dB compression point of both the LNA and the mixer RF port. A plot of the gain \( |S_{21}| \) versus frequency of the power amplifier is given in Fig. 3.8. We see that the LNA has a positive gain up to 5GHz, outside the operational bandwidth. The fundamental frequency gain and the gain of the second harmonic are annotated in the figure. The gain for the second harmonic is about 11.73dB. As the gain for the fundamental frequency is about 7.92dB higher, the power of the harmonic frequencies will fall by this amount relative to the power of the fundamental frequency. As explained in subsection 3.1.1, this helps to suppress the unwanted mixing products. Fig. 3.9 shows the matching of the in- and output \( (S_{11} \) and \( S_{22} \)). Both in- and output are no longer matched at the second harmonic frequency.

The input and output are internally matched to 50Ω so no external components are needed for matching. The LNA has four usable terminals: an RF input pin, a ground pad, a shutdown control pin and an RF output. The shutdown control pin is used for quickly shutting down and powering up the LNA in time division duplexing applications and will not be used in our design. This feature can be disabled by connecting this pin to ground. Just like with the transmit amplifier, the amplifier is also fed through a bias tee at the RF output pin. Exactly the same bias tee as in subsection 3.1.2, Fig. 3.7 is used, with the same components.

One can calculate with the Friis formula given in equation 2.61 the noise figure of the cascade of the LNA and the mixer. This results in a noise figure of 0.644dB. The total gain of
3.1 Circuit design and component selection

Low noise amplifier gain $|S_{21}|$ as a function of frequency

![Graph of Low noise amplifier gain $|S_{21}|$ as a function of frequency](image)

**Fig. 3.8:** LNA gain $|S_{21}|$ as a function of frequency

Low noise amplifier $|S_{11}|$ and $|S_{22}|$ as a function of frequency

![Graph of Low noise amplifier $|S_{11}|$ and $|S_{22}|$ as a function of frequency](image)

**Fig. 3.9:** LNA matching $|S_{11}|$ and $|S_{22}|$ as a function of frequency
the cascade is 12.95dB. As the amplification of the first amplifier is still far below the mentioned maximum amplification of 40dB, it was decided to put an extra amplifier between the LNA and the frequency mixer. Coincidently, the transmitter power amplifier has a gain that fits this application. The gain of the cascade of the LNA and the second amplifier is equal to 33.35dB around 2.45GHz. The directly leaked signal estimated to have a power of about -30dBm is amplified to a power of 3.35dBm, well below the input 1dB compression point (10dBm) of the frequency mixer. The gain of the cascade formed by the LNA, the second power amplifier and the frequency mixer is 27.7dB at 2.45GHz and its noise figure is 0.552dB. The advantages of the second power amplifier are that the noise figure is a bit smaller and the gain 14.75dB higher. An important disadvantage with this setup is that the second power amplifier can become unstable. Fig. 3.10 shows that there can be positive feedback around the second power amplifier via the DC feed network. Another disadvantage is that the RF input signal power comes close to the non-linear region of the frequency mixer. If the isolation between transmitter and receiver is less than 50dB, the frequency mixer will work in its non-linear region. A final disadvantage is a considerable increase in power consumption. To cover for these disadvantages, we will provide the possibility in the PCB layout to omit this second power amplifier.

3.1.5 Variable gain amplifier

The task of the low pass filter block is to filter out the unwanted signals at the frequency mixer IF output, among which the up-converted signal and the leaked LO signal have the highest power. This filtering will be discussed in subsection 3.1.6. The filtered signal is applied to a
VGA that compresses the large dynamic range of the signal to a nearly constant root mean square (RMS) output amplitude by automatic gain control (AGC).

The selected VGA is the AD8338 \cite{20} \cite{21} chip of Analog Devices. The nominal gain is adjustable from 0dB to 80dB within the nominal frequency band of DC to 18MHz. The device has an AGC circuit built in to control this variable gain.

The AD8338 with pinout and the configuring circuit components around it are depicted in Fig. 3.11. Step by step, the chosen components will be explained in the rest of this subsection.

The device is powered by 5V at pin VBAT drawing a current of 3.8-8.0mA depending on the gain. The power supply is bypassed with two capacitors (10\,\mu \text{F} and 0.1\,\mu \text{F}) to the ground pin COMM. The AD8338 uses an internal reference voltage of 1.5V. According to the datasheet the reference pin VREF has to be bypassed to ground with a capacitor (0.1\,\mu \text{F}).

The input signal always has to be provided differentially to this amplifier. There are two ways of applying the input signal to this amplifier: the INPR/INMR terminal pair can be used or the INPD/INMD terminal pair. The input signal has to be applied through coupling capacitors \( C_K \) in both cases, in order not to disturb the DC bias present on these pins. Disturbing the DC bias of these pins will degrade the noise performance of the amplifier. The chosen value of \( C_K = 100\,\mu \text{F} \) will be clarified later. The INPR/INMR terminal pair is a differential voltage input (input \( V_{\text{in},1} \)). The input impedance is 1k\,\Omega in parallel with a parasitic capacitance of 2pF. The input voltage range is 3V peak-to-peak. Another option is to use the INPD/INMD terminals. These are current input terminals and can be used in two ways: either the signal can be applied directly under the form of a current, or coupling resistors (\( R_P \) and \( R_N \)) can be used to convert a voltage signal into an input current (input \( V_{\text{in},2} \)). The input impedance of the amplifier is equal to the sum of the two coupling resistors, in parallel with a parasitic capacitance of 2pF. This means the input impedance is configurable. Essentially, the use of the INPR/INMR terminal pair is equivalent to using the INPD/INMD terminal pair with two 500\,\Omega coupling resistors. When using the INPD/INMD terminals, the INPR/INMR pair has to be shorted for stability reasons.

The gain has a span of 80dB and can be controlled by the voltage \( V_{\text{GAIN}} \) on pin GAIN. The gain control voltage is linear-in-dB over the voltage range of 0.1V to 1.1V with a slope of 1dB per 12.5mV. The voltage on pin MODE determines if the gain increases or decreases with increasing control voltage. When pin MODE is tied to 5V, the gain increases with increasing control voltage. The exact gain as a function of \( V_{\text{GAIN}} \) depends on which coupling resistors are used (when using the INPR/INMR terminals, the coupling resistors are assumed to be \( R_P = R_N = 500\,\Omega \)) and is given by the following formula (source: datasheet):

\[
G(\text{dB}) = 80 \cdot V_{\text{GAIN}} + 20 \cdot \log \left( \frac{2 \cdot R_{\text{FBK}}}{R_P + R_N} \right) - 34, \quad (3.12)
\]

where \( R_{\text{FBK}} \) is an internal feedback resistor of 9800\,\Omega. We see that the exact gain range can be shifted up or down by changing coupling resistors \( R_P \) and \( R_N \). When using the standard internal coupling resistors of 500\,\Omega, the minimum gain is 0dB and the maximum gain 80dB. When using
external coupling resistor, one also has to take into consideration that the bandwidth of the amplifier changes. The bandwidth of the amplifier when using external resistors \( R_P = R_N \) is given by the following formula (source: datasheet):

\[
 f_{BW} = 18 \text{MHz} \cdot \left( \frac{500\Omega \cdot R_P}{500\Omega + R_P} \right) \cdot \frac{1}{500\Omega} \quad (3.13)
\]

The change of the bandwidth will also change the noise profile of the amplifier, as explained in the datasheet.

The output of the amplifier has a peak-to-peak output swing of 2.8V and a common mode voltage of 1.5V. Care has to be taken that the amplifier is not overloaded resulting in a clipped output signal. The output has to be loaded with at least 1k\( \Omega \) in order to preserve the linearity. The output terminals of the amplifiers are left open circuited. Measurements will be performed on these pins with an oscilloscope. The output peak slew rate is 50V/\( \mu \)s for \( V\text{GAIN} = 0.6V \). The small signal bandwidth is nominally 18MHz but the peak slew rate limits the maximum frequency of a signal with full output swing that can be processed by the amplifier without distortion:

\[
 V\text{OUTP}(t) = 0.7V \cdot \sin(2\pi \cdot f \cdot t) \Rightarrow \frac{dV\text{OUTP}(t)}{dt} = 0.7V \cdot 2\pi \cdot f \cdot \cos(2\pi \cdot f \cdot t) \quad (3.14)
\]

\[
 \frac{dV\text{OUTP}(t)}{dt} = 50V/\mu s \Leftrightarrow f = 11.37\text{MHz} \quad (3.15)
\]

This is normally much higher than the frequencies that we have to amplify.

As shown by equation (3.12), the choice of input terminal pair and coupling resistors determines the gain range, which has to be chosen carefully in order to avoid output overloading and accompanying clipping. The IF output of the frequency mixer has to be terminated with 50\( \Omega \), so when a signal with frequency \( f \) and power \( P \) is available at this output, the amplitude of the output voltage is given by:

\[
 V_{IF} = \sqrt{2 \cdot 50\Omega \cdot P}. \quad (3.16)
\]

The output signal with the maximal power will have the maximal voltage. Generally, the directly leaked signal that is down-mixed to baseband will be the strongest signal that is present, estimated to have a power of -30dBm at the input of the LNA, so a power of -2.3dBm(-30dBm+27.7) at the mixer IF output. The corresponding voltage amplitude over the 50\( \Omega \) resistor is 0.243V. The maximum voltage gain without output clipping is:

\[
 G_{\text{max}} = 20 \cdot \log \left( \frac{1.4V}{0.243V} \right) = 15.21dB. \quad (3.17)
\]

In order for the minimum gain of the amplifier (for \( V\text{GAIN} = 0.1V \)) to be lower than this 15.21dB, the condition on the input coupling resistors is:

\[
 G_{dB,min} = 80 \cdot 0.1 + 20 \cdot \log \left( \frac{2 \cdot R_{\text{FBK}}}{R_P + R_N} \right) - 34 = 15.21dB \Leftrightarrow R_P + R_N > 170.51\Omega. \quad (3.18)
\]
In order to retain margin to overloading, we can limit the peak-to-peak output swing to 1.4V. We then obtain a maximum voltage gain:

\[ G_{\text{max}} = 20 \cdot \log \left( \frac{0.7V}{0.243V} \right) = 9.19\text{dB}. \]  \hfill (3.19)

The corresponding condition on the coupling resistors becomes:

\[ G_{dB,\text{min}} = 9.19\text{dB} \iff R_P + R_N > 341.01\Omega. \]  \hfill (3.20)

When feedback capacitors \( C_{FBK} \) between the pins OUTP/FBKP and OUTM/FBKM are applied, the bandwidth of the amplifier is reduced, reducing the output noise of the device.

\[ f_C = \frac{1}{2\pi \cdot R_{FBK} \cdot C_{FBK}} \]  \hfill (3.21)

with \( R_{FBK} \) internal resistors of 9800\( \Omega \). To acquire a cut off frequency of 1MHz, a capacitor \( C_{FBK} \) of 16.2pF is needed. The closest E12 value 15pF was taken.

The AD8338 also features an input offset correction circuit that cancels the DC offset that may be present in the amplifier. To enable this feature, a capacitor \( C_{OFSN} \) has to be connected between the OFSN and VREF pin. This circuit adds a high pass filtering characteristic to the transfer function of the amplifier with a 3dB cut off frequency \( f_{OFSN} \) that is given by:

\[ f_{OFSN} = \frac{1}{2\pi \cdot 400\Omega \cdot C_{OFSN}}. \]  \hfill (3.22)

As explained in chapter 2, the useful converted signal frequency spectrum consists of harmonics of the modulation frequency. The modulation frequency was chosen to be 10kHz. When we choose the capacitor \( C_{OFSN} \) to be 1\( \mu \)F, the 3dB cut off frequency \( f_{OFSN} \) is equal to 0.398kHz and the lowest frequency component of 10kHz will not be influenced by the input offset correction feature. This input offset correction circuit should not be used to suppress the directly leaked signal that is mixed to baseband. Instead, the coupling capacitors can be used for this. The proper choice of the coupling capacitors, however, will depend on the input impedance of the amplifier, which on its turn depends on whether we use the INPR/INMR terminal pair or the INPD/INMD terminal pair. We will choose one of the two when constructing the low pass filter.

The most important reason why the AD8338 chip was chosen is because it has an internal AGC circuit. The AGC circuit varies the gain of the amplifier to acquire a constant RMS output voltage, while the RMS input amplitude may fluctuate. By setting the voltage \( V_{AGC} \) at pin VAGC, the wanted output RMS amplitude is set to \( |V_{AGC} - V_{\text{REF}}| \). This voltage is set with a voltage divider, and is adjustable by the use of a trimmer. The actual AGC circuit is a current output RMS detector that outputs whether a current of 10\( \mu \)A or -10\( \mu \)A out of pin DETO depending on whether the output RMS amplitude is greater than the wanted output RMS amplitude or smaller. This current can be injected in a capacitor \( C_{AGC} \) to ground, to increase or decrease the voltage over it. This voltage can be used to control the GAIN pin. In order for this feedback loop to be negative, the gain slope of the amplifier has to be inverted.
by connecting the MODE pin to ground. In this way, the current output RMS detector will steer the gain pin so that the output RMS amplitude is driven towards the wanted value. As the current of pin DETO is fixed +/-10\(\mu\)A, the bandwidth of the AGC circuit is determined only by the value of the capacitor \(C_{AGC}\). The capacitor value has to be small enough, so that the gain can react quickly to a changing input voltage amplitude, but has to be large enough in order not to distort the input signal. The input signal of the amplifier is the filtered converted signal, containing power at harmonics of 10kHz. The smallest frequency component of 10kHz has a period of 0.1ms. The AGC circuit can change the gain of the amplifier by an amount \(\Delta G\) in this period:

\[
\Delta G = 80 \frac{dB}{V} \cdot \frac{0.1\text{ms} \cdot 10 \mu A}{C_{AGC}}.
\]  

(3.23)

If we want the gain to only change 0.01dB in this period, the capacitor \(C_{AGC}\) has to be equal to 8\(\mu\)F. We round this off to the nearest E12 value of 10\(\mu\)F. In this way, the AGC circuit will certainly be able to react quickly to changing input amplitudes. In our case the input signal amplitude depends on the range of the reflector. When the reflector moves radially away from the radar system, the received reflected signal power will decrease. If we consider a maximum target speed of 10m/s, moving from range 1m to range 11m, the decrease in received power is 41.66dB. The AGC circuit only needs 521ms to increase the gain by 41.66dB. A jumper to disconnect the AGC circuit from the GAIN pin was foreseen in case the AGC functionality were not to function.

3.1.6 Low pass filter

Frequency mixer terminations

The frequency mixer type that was chosen is a double balanced diode mixer. In the case of double balanced diode mixers, it is important that all three mixer ports are adequately terminated to 50\(\Omega\) in order to obtain a good performance, as discussed in [22]. A mismatch at the RF port of the mixer is least problematical; a relatively small increase of the conversion loss and a relatively small decrease of the RF input 1dB compression point may happen. A mismatch at the LO port of the mixer can degrade the linearity (third order intermodulation products and harmonic modulation products), but this does not influence the conversion loss or RF input 1dB compression point if the LO signal power remains adequately strong. The matching of the mixer RF and LO ports are taken care of in section 3.2 on the PCB layout. The most critical port to match is the IF port and this will be discussed here. When the mixer IF port is not properly matched to 50\(\Omega\) a significant increase in conversion loss and a significant decrease of linearity (RF input 1dB compression point, harmonic modulation products, third order intermodulation products) can be expected.
3.1 Circuit design and component selection

Fig. 3.11: AD8338
3.1 Circuit design and component selection

Low pass filter requirements

There are three requirements to the low pass filter block. Firstly, the stop band has to include the leaked LO signal around 2.45GHz and the mixing product around 4.9GHz. Moreover, in order to suppress the contribution of the directly leaked signal, the DC component should be blocked. The pass band has to include the useful converted signal frequencies from 10kHz to 1MHz. The insertion loss in the pass band has to be as low as possible. Choosing the pass band bandwidth minimal will reduce the noise power at the output.

Secondly, the mixer IF port has to be matched to 50Ω over a wide frequency band so no significant reflections occur back to the frequency mixer. This includes the stop band of the low pass filter. Therefore, a return loss of at least 10dB from DC up to at least 5GHz is required.

Finally, the low pass filter has to be designed so its output can be applied to the AD8338 chip.

We now consider several ways to accomplish this.

Standard lossless low pass filter

The low pass filtering cannot merely happen with the standard lossless LC filters like Butterworth or Chebyshev filters. An intrinsic property of lossless filters is that they can only provide attenuation by reflection ([23]). Consequently, the input of such filters is only matched to 50Ω in the pass band and not in the stop band.

The matching to 50Ω within the stop band of the low pass filter can be improved by inserting an attenuator as depicted in Fig. 3.12. Inserting an attenuator of 5dB, the return loss increases by twice this attenuation, 10dB. Of course, the 5dB attenuation is added to the insertion loss of the filter in the pass band. This increases the conversion loss of the mixer and the noise figure, which is unacceptable.

Low pass filtering with choke inductor

A simple solution without increased mixer conversion loss is presented in Fig. 3.13. The IF output of the mixer is loaded with an impedance $Z_{IF}$. This impedance $Z_{IF}$ is a parallel connection of a 50Ω resistor (with the IF voltage $V_{IF}$ across its terminals) and a branch, consisting of a series connection of a choke inductor $L_{choke}$, two coupling capacitors $C_K$ and the AD8338 chip.
3.1 Circuit design and component selection

with input impedance $R_{in}||C_{in}$. $C_{in}$ is given to be 2pF by the datasheet, and we can choose $R_{in}$ ourselves.

The parallel branch serves as a voltage divider, bringing a fraction $V_{AD8338}/V_{IF}$ of the IF voltage $V_{IF}$ over the input terminals of the amplifier. The intention is to give the transfer function $V_{AD8338}/V_{IF}$ a low pass characteristic with a pass band including 10kHz to 1MHz (useful converted signal frequencies) and a stop band including 2.4GHz to 5GHz (leaked LO signal and unwanted mixing product), while keeping the impedance $Z_{IF}$ matched to 50Ω from DC to over 5GHz.

The choke inductor has to have an impedance that is much higher than the input impedance of the amplifier at the frequencies in the stop band, so that almost nothing of the IF voltage is applied to the input of the amplifier. This impedance also has to be much higher than 50Ω, so that $Z_{IF}$ is matched to 50Ω. The input impedance of the amplifier has to be much higher than the impedance of the choke inductor at the frequencies in the pass band, so that almost all IF voltage is applied across the input terminals of the amplifier. The impedance of the parallel branch is then approximately equal to the input impedance of the amplifier. Consequently, the amplifier input resistor has to be chosen much higher than 50Ω, so that the parallel connection of the load resistor and parallel branch stays approximately 50Ω in the pass band.

The coupling capacitors should be negligible at frequencies of 10kHz and up, and the transfer function is approximated by:

$$\frac{V_{AD8338}}{V_{IF}} = \frac{1}{s^2 \cdot L_{choke} \cdot C_{in} + s \cdot L_{choke}/R_{in} + 1}. \quad (3.24)$$

The resonance pulsation $\omega_{r,1}$ and the quality factor $Q_1$ are given by:

$$\omega_{r,1} = \frac{1}{\sqrt{L_{choke} \cdot C_{in}}}, \quad (3.25)$$

$$Q_1 = R_{in} \cdot \sqrt{C_{in} / L_{choke}}. \quad (3.26)$$

The IF impedance $Z_{IF}$ is approximated by:

$$Z_{IF} = \frac{R_{in} \cdot 50\Omega}{R_{in} + 50\Omega} \cdot \frac{s^2 \cdot L_{choke} \cdot C_{in} + s \cdot L_{choke}/R_{in} + 1}{s^2 \cdot L_{choke} \cdot C_{in} + s \cdot L_{choke}/R_{in} + 1 + s \cdot C_{in} \cdot R_{in} \cdot 50\Omega / L_{choke} + 1}. \quad (3.27)$$

The quality factor and resonance frequency for the numerator are also given by equation (3.25) and (3.26). The resonance pulsation $\omega_{r,2}$ and the quality factor $Q_2$ for the denominator are given by:

$$\omega_{r,2} = \sqrt{\frac{R_{in} + 50\Omega}{L_{choke} \cdot C_{in} \cdot R_{in}}}, \quad (3.28)$$

$$Q_2 = \frac{1}{\omega_r} \cdot \frac{R_{in} + 50\Omega}{L_{choke} + C_{in} \cdot R_{in} \cdot 50\Omega}. \quad (3.29)$$

The transfer function of equation (3.24) indeed has a low pass characteristic.
We first choose a value for $R_{in}$. As shown in equation 3.20, $R_{in}$ has to be at least 341Ω in order to keep a safe margin from amplifier overloading. We choose $R_{in} = 360\Omega$, being the series connection of two E12 values. This value in parallel with 50Ω yields a resistance of 43.9Ω. This corresponds to a return loss of 23.75dB.

Next, we choose the value of the inductor. The inductor value will determine the cut off frequency of the filter. If the value of the choke inductor is too low, equations 3.24 and 3.27 will be second order transfer functions and resonance of the choke inductor and the parasitic capacitor will occur. If this resonance has a high quality factor, the transfer functions exhibit resonance peaks, where the impedance of the parallel branch is significantly reduced and the match of $Z_{IF}$ to 50Ω is destroyed. To completely avoid resonances, the value of $L_{choke}$ should be at least 1.11$\mu$H, being an impractical value. Choosing a higher $R_{in}$ does not decrease the value of $L_{choke}$ needed to avoid resonance. For practical cases, the transfer functions will be second order.

A second order transfer function does not exhibit resonance peaks in its response if the quality factor is smaller than $1/\sqrt{2}$. For an E12 inductor value of about 470nH, the quality factors are: $Q_1 = 0.742$ and $Q_2 = 0.736$, and the resonance frequencies are $f_{r,1} = \frac{\omega_c}{\sqrt{2 \pi}} = 164.15$MHz and $f_{r,2} = 175.18$MHz. No significant resonance peaks will occur for this inductor. The cut off frequency is much higher than the mentioned 1MHz, but this is necessary in order to obtain practical values for the choke inductor. As no significant power is expected from 1MHz to about 170MHz in the IF output, this does not pose a serious problem. The only degradation this wider pass band has is on the area of noise. Low pass filtering with a cut off frequency of 1MHz at the VGA output will happen anyway.

The only thing left to dimension is the value of the coupling capacitors. At low frequencies, the choke inductor and the parasitic capacitance can be neglected. We obtain the following approximation for the transfer function:

$$\frac{V_{AD8338}}{V_{IF}} = \frac{s \cdot R_{in} \cdot C_K}{s \cdot R_{in} \cdot C_K + 2}.$$  
(3.30)

The break point is given by:

$$\omega_c = \frac{2}{R_{in} \cdot C_K}.$$  
(3.31)

This cut off frequency is the lower frequency of the pass band, and it therefore has to be much lower than 10kHz. An E12 value of 10$\mu$F yields a cut off frequency of 88.42Hz.

Figs. 3.14 and 3.15 depict the transfer function $\frac{V_{AD8338}}{V_{IF}}$ for the frequency interval 1Hz-10GHz. The pass band ranges from 88.42Hz to 173MHz as expected, with a negligible insertion loss. We also see that no phase distortion occurs in the frequency interval from 10kHz to 1MHz. The leaked LO signal at a frequency of 2.4-2.5GHz is attenuated at least 46.5dB. The unwanted mixing product even more so. As desired, DC is also blocked by the filter. Fig. 3.16 presents the matching of the IF load impedance $Z_{IF}$ to 50Ω. The return loss is the lowest in the filter pass band, where it is still 23.75dB. This means the match is good over the entire frequency band.
The calculations assume ideal components. The ideal 470nH choke inductor can be realised with the 470nH RF inductor (LQW18ANR47G00) of Murata up to frequencies of about 300-400MHz. The self-resonance frequency of this component is about 1.3GHz. Beyond this frequency the component does not behave as an inductor anymore, and shows regions where the insertion loss is low. The inductor impedance is not high enough for the higher frequencies of the stop band. The only way to achieve a high inductor impedance over the entire stop band (wide bandwidth of 173MHz to 5GHz) is by a series connection of multiple RF inductors with different values and resonance frequencies. A series connection of at least 4 inductors (LQW18ANR47J00, LQW18ANR39G00, LQW18ANR33J00 and LQW18ANR22J00) would be needed in order to obtain an insertion loss of at least 30dB over this frequency band.

Another more elegant way of filtering will be presented in the next subsection.
3.1 Circuit design and component selection

Fig. 3.14: Magnitude of transfer function $V_{AD8338}/V_{IF}$

Fig. 3.15: Phase of transfer function $V_{AD8338}/V_{IF}$
Diplexer

A diplexer is a passive three-port (S-port, L-port and H-port) device that performs frequency multiplexing. The input signal at the S-port, containing power in a certain frequency band, is split in two signals each containing power in disjoint frequency sub bands of the input signal. In other words, both the signals available at the L- and H-port are a filtered version of the input signal, with different pass bands. This separation of frequency bands should happen without too much loss of power. Typically, a high pass (hence port name H) filtering and a low pass (port name L) filtering with identical 3dB cut off frequencies is performed.

A diplexer can also be used in the opposite way, using the L-port and H-port as input ports and the S-port as output port. In this case, the input signals containing power in disjoint frequency bands, are combined into a single signal available at the S-port, containing power in both frequency bands.

Interesting for our application is that a diplexer with a high pass and low pass characteristic with identical 3dB cut off frequencies (also called crossover frequency) can be designed in such a manner that the S-port is matched for all input frequencies (at least for ideal lumped components). Such a diplexer can be used as a non-reflecting low pass or high pass filter by terminating one of the outputs with a dummy load. If, like in our case, a non-reflecting low pass filter is needed, the H-port should be terminated with a dummy load. In this way, a low pass filtered version of the input signal is available at the L-port, while the input is matched both in the pass band and stop band.

A naive way of designing such a diplexer would be designing a low pass and high pass filter...
with identical 3dB cut off frequencies separately, and then use these filters in parallel by using the same input port for both filters. This method will lead to poor performance with the input of the diplexer not being adequately matched around the 3dB cut off frequency.

The correct design procedure for diplexers with Butterworth and Chebyshev filter characteristics is described in [24]. This involves the use of singly terminated filters rather than the standard doubly terminated filters. A doubly terminated filter is a filter that is driven by a generator with a non-zero internal impedance at one side, and terminated with a resistor at the other side, while a singly terminated filter is driven by a zero internal impedance generator (or an ideal voltage source) at one side, and terminated with a resistor at the other side. Element values for prototype (cut off pulsation 1rad/s and load resistance 1Ω) Butterworth and Chebyshev singly terminated filters are tabulated in [24]. The choice of singly terminated Butterworth filters over singly terminated Chebyshev filters to design the diplexer, has the additional advantage that the input match is theoretically perfect for all frequencies. Therefore, it is chosen to design a Butterworth diplexer. The design procedure for a Butterworth diplexer can be summarised as follows:

1. The starting point is a prototype singly terminated Butterworth low pass filter. The filter order has to be chosen based on the desired properties. In [24], the filter topology is given and component values are tabulated.

2. Denormalise this prototype filter to a filter with the wanted 3dB cut off frequency \( f_c \) and generator impedance \( R_G \). Therefore, the following component replacement formulas can be used [23]:

   \[
   \text{Resistor } R \rightarrow \text{Resistor } R \times R_G, \tag{3.32}
   \]

   \[
   \text{Inductor } L \rightarrow \text{Inductor } \frac{L \times R_G}{2\pi f_c}, \tag{3.33}
   \]

   and

   \[
   \text{Capacitor } C \rightarrow \text{Capacitor } \frac{C}{R_G \times 2\pi f_c}. \tag{3.34}
   \]

3. Transform the prototype filter of the first step, to a high pass filter with a 3dB cut off frequency identical to the filter of step two. Therefore, the following replacement formulas can be used [23]:

   \[
   \text{Resistor } R \rightarrow \text{Resistor } R \times R_G, \tag{3.35}
   \]

   \[
   \text{Inductor } L \rightarrow \text{Capacitor } \frac{1}{2\pi f_c \times L \times R_G}, \tag{3.36}
   \]

   and

   \[
   \text{Capacitor } C \rightarrow \text{Inductor } \frac{R_G}{2\pi f_c \times C}. \tag{3.37}
   \]

4. Interconnect the filters of step two and three so that their input ports are common.
3.1 Circuit design and component selection

It was chosen to implement a 3\textsuperscript{rd} order diplexer to keep the number of components low. The schematic is shown in Fig. 3.17. Port 1 is regarded as an input port, and ports 2 and 3 as output ports. The upper half of the schematic, consisting of the components $C_2$, $C_3$ and $L_3$ is the high pass filter section of the diplexer, and the lower half, consisting of the components $L_1$, $L_2$ and $C_1$ is the low pass filter section. To use this diplexer as a non-reflective low pass filter, port three has to be terminated with a 50Ω resistor. In this case port one is the filter input, and port two the filter output. Different sets of component values for the capacitors and inductors are summarised in table 3.1. The second column gives the component values for a prototype Butterworth diplexer with a crossover pulsation of 1 rad/s and a generator impedance of 1Ω. As was stated in the low pass filter requirements, the pass band has to include frequencies at least up to 1MHz, and the input of the filter has to be matched to 50Ω up to frequencies of 5GHz. The component values for a Butterworth diplexer with a crossover frequency of 1MHz and a generator impedance of 50Ω are given in the third column of table 3.1. The component values are too large to be practical at frequencies of 5GHz. This diplexer is therefore not realisable. When the crossover frequency is increased, both the component values for the capacitors and inductors will decrease. However, the crossover frequency can not increase too much, because the frequencies of the leaked LO signal (2.4GHz-2.5GHz) and the up-mixed signal (4.8GHz-5GHz) still have to be adequately suppressed. Moreover, the noise power increases with increasing bandwidth. A good choice for the crossover frequency was found to be 1GHz, yielding practical values for the inductors and capacitors, while adequately suppressing the frequencies higher than 2.4GHz. The component values for a Butterworth diplexer with a crossover frequency of 1GHz and an input impedance of 50Ω are given in the fourth column of table 3.1.

A simulation of the S-parameters of the designed diplexer performed with the Advanced Design System (ADS) software of Agilent is given in Figs. 3.18 and 3.19. This simulation is performed with ideal lumped components. In Fig. 3.18 both $|S_{21}|$ and $|S_{31}|$ are plotted as

![Fig. 3.17: 3\textsuperscript{rd} order diplexer](image-url)
### 3.1 Circuit design and component selection

#### Table 3.1: Component values for the diplexer schematic

<table>
<thead>
<tr>
<th></th>
<th>prototype (F/H)</th>
<th>1MHz, 50Ω (nF/µH)</th>
<th>1GHz, 50Ω (pF/nH)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_1$</td>
<td>1.50</td>
<td>11.94</td>
<td>11.94</td>
</tr>
<tr>
<td>$C_1$</td>
<td>1.33</td>
<td>4.24</td>
<td>4.24</td>
</tr>
<tr>
<td>$L_2$</td>
<td>0.50</td>
<td>3.98</td>
<td>3.98</td>
</tr>
<tr>
<td>$C_2$</td>
<td>0.67</td>
<td>2.12</td>
<td>2.12</td>
</tr>
<tr>
<td>$L_3$</td>
<td>0.75</td>
<td>5.97</td>
<td>5.97</td>
</tr>
<tr>
<td>$C_3$</td>
<td>2.00</td>
<td>6.37</td>
<td>6.37</td>
</tr>
</tbody>
</table>

Fig. 3.18: Diplexer $|S_{21}|$ and $|S_{31}|$ as a function of frequency. Ideal lumped components used.

A function of frequency. The frequency multiplexing action is observed and has the correct crossover frequency of 1GHz. The leaked LO signal is suppressed by at least 22.83dB, and the up-mixed signal by at least 40.87dB. Fig. 3.19 shows the input matching of the diplexer, which is practically perfect over the entire plotted frequency range.

Next, real capacitor and inductor components are chosen. This will influence the performance in two ways. Firstly, the calculated values will have to rounded to the nearest available value. This will change the behaviour of the diplexer. Secondly, real capacitors and inductors have non-ideal frequency behaviour (component value changes with frequency, self-resonance, ...). The chosen components will need to be able to work correctly at frequencies at least up to 5GHz (up-mixed signal frequency). An important indicator for this is the self-resonance frequency of the capacitor or inductor, that is preferably well above 5GHz. Generally, smaller valued capacitors and inductors have a higher self-resonance frequency within a capacitor or inductor
3.1 Circuit design and component selection

Fig. 3.19: Input matching diplexer $|S_{11}|$ as a function of frequency. Ideal lumped components used.

Therefore, some values of the capacitors of the fourth column of table 3.1 are realised as parallel connections of smaller valued capacitors to make the behaviour at higher frequency better. The chosen capacitor and inductor components and their self-resonance frequencies are summarised in table 3.2. Table 3.3 summarises how the components of table 3.2 are used to achieve the needed capacitor and inductor values of table 3.1.

<table>
<thead>
<tr>
<th>Nominal value</th>
<th>Component</th>
<th>Self-resonance frequency (GHz)</th>
</tr>
</thead>
<tbody>
<tr>
<td>1pF</td>
<td>GQM1885c2a1r0cb01</td>
<td>7177MHz</td>
</tr>
<tr>
<td>1.1pF</td>
<td>GQM1885c2a1r1cb01</td>
<td>6841MHz</td>
</tr>
<tr>
<td>1.6pF</td>
<td>GQM1885c2a1r6cb01</td>
<td>5518MHz</td>
</tr>
<tr>
<td>3.9nH</td>
<td>LQW18an3n9d00</td>
<td>min. 6GHz</td>
</tr>
<tr>
<td>6.2nH</td>
<td>LQW18an6n2d00</td>
<td>min. 6GHz</td>
</tr>
<tr>
<td>12nH</td>
<td>LQW18an12nj00</td>
<td>min. 6GHz</td>
</tr>
</tbody>
</table>

Table 3.2: Used set of capacitor and inductor components for the diplexer

Another simulation of the S-parameters of the designed diplexer is performed using the component models of the chosen capacitor and inductor components. Fig. 3.20 depicts $|S_{21}|$ and $|S_{31}|$ as a function of frequency, both in the case when real components are used and when ideal components are used. The only noticeable difference in the figure is that $|S_{21}|$ is lower for the high frequencies of the stop band and this is advantageous. The leaked LO signal is now
3.1 Circuit design and component selection

<table>
<thead>
<tr>
<th></th>
<th>Ideal value</th>
<th>Combination</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_1$</td>
<td>11.94nH</td>
<td>12nH</td>
</tr>
<tr>
<td>$C_1$</td>
<td>4.24pF</td>
<td>1pF</td>
</tr>
<tr>
<td>$L_2$</td>
<td>3.98nH</td>
<td>3.9nH</td>
</tr>
<tr>
<td>$C_2$</td>
<td>2.12pF</td>
<td>1pF</td>
</tr>
<tr>
<td>$L_3$</td>
<td>5.97nH</td>
<td>6.2nH</td>
</tr>
<tr>
<td>$C_3$</td>
<td>6.37pF</td>
<td>1.6pF</td>
</tr>
</tbody>
</table>

Table 3.3: Component values for the diplexer schematic

suppressed by at least 23.63dB, and the up-mixed signal by at least 46.93dB. Around 7GHz, $|S_{21}|$ shows a downwards peak because of the resonance of one of the components. Another difference, but impossible to see in the figure is that $|S_{21}|$ is also slightly lower for the low frequencies of the pass band, meaning the insertion loss of the low pass filter will be slightly increased. At 1MHz, $|S_{21}|$ is now $-0.036$dB instead of the negligible $-4.41 \cdot 10^{-6}$dB. This still remains a negligible insertion loss. The same remark can be made for $|S_{31}|$, but this is of no interest because port three will be terminated with a resistor anyway. A more important deviation from the ideal case is that the low pass and high pass filter do not cross over perfectly at 1GHz anymore. The low pass filter has a 3dB cut off frequency of 999MHz while the high pass filter has a 3dB cut off frequency of 1025MHz. It is expected that the non-perfect crossover of the high pass and low pass section will result in a degraded input matching. The simulated input matching is depicted in Fig. 3.20. The input matching is indeed degraded, but stays excellent because $|S_{11}|$ is lower than $-20$dB up to 5.45GHz.

Another important factor that has a major influence on the input matching of the diplexer, is the matching of the terminations of port two and three. In order to have a good filter input match, port two (low pass output) has to be adequately terminated with 50Ω in the pass band and around the crossover frequency and port three (high pass output) has to be adequately terminated with 50Ω in the stop band and around the crossover frequency. This will now be ensured.

For our purpose, port three has to be resistively terminated with 50Ω up to well above 5GHz. A high frequency 50Ω thin film resistor (FC0603E50R0BST1) of Vishay was selected. This resistor has a value of 50Ω up to frequencies higher than 10GHz. Moreover, when designing the PCB layout of the diplexer, care will be taken to limit parasitics that would degrade the match of the termination.

The signal available at the low pass output of the diplexer will have to be applied to the VGA in some way. An interesting way would be by using the INPD/INMD input terminal pair, and select both coupling resistors to be $R_P = R_N = 25Ω$. In this way, the input impedance of the AD8338 chip is 50Ω in parallel with a parasitic capacitance of estimated to be 2pF in the
Fig. 3.20: Diplexer $|S_{21}|$ and $|S_{31}|$ as a function of frequency. Realistic component models used.

Fig. 3.21: Input matching diplexer $|S_{11}|$ as a function of frequency. Realistic component models used.
Fig. 3.22: Connection of the variable gain amplifier to the diplexer

datasheet. However, equation 3.19 shows that if the coupling resistors are chosen to be 25Ω, there is danger of overloading the amplifier. It was chosen to apply the signal available at the diplexer output as depicted in Fig. 3.22. This is identical to the situation of Fig. 3.13 when low pass filtering with a choking inductor was considered. Now, it is chosen to use the INPR/INMR input terminal. In this way, the input resistance of the VGA is 1kΩ which will improve the input matching of the diplexer at low frequencies.

The effect of the non-ideal terminations of ports two and three will not be simulated here, but in section 3.2 when simulations are performed that include the PCB layout.

The only thing left to dimension is the value of the coupling capacitors. A voltage divider is formed by the VGA input impedance and the two coupling capacitors $C_K$. As at low frequencies, the parasitic capacitance can be neglected, the transfer function $V_{AD}/V_{load}$ is approximated by:

$$\frac{V_{AD}}{V_{load}} = \frac{s \cdot 1k\Omega \cdot C_K}{s \cdot 1k\Omega \cdot C_K + 2}$$  

(3.38)

The break point is given by:

$$\omega_c = \frac{2}{1k\Omega \cdot C_K}$$  

(3.39)

This cut off frequency is the lower frequency of the pass band, so it has to be much lower than 10kHz. An E12 value of 2.2µF yields a cut off frequency of 144.69Hz.

3.2 PCB Layout

In this section, the design of the PCB that realises the electronic circuitry of section 3.1 is discussed. For modularity, it was chosen to design separate PCBs for the main radar system (containing the triangle voltage oscillator, VCO, amplifiers and frequency mixer) and what can
be called the IF system (containing the diplexer and VGA). Due to time shortage, only the diplexer part of the the IF system PCB was soldered and tested. Therefore only the PCB design of the diplexer part is discussed. First a few general remarks relevant to both PCBs are made.

The used PCB substrate is I-Tera MT material of Isola Corporation, which is suited for high frequency applications up to 20GHz [25]. The dielectric constant is specified to be $3.30 \pm 0.05$ and the loss tangent $0.0036 \pm 0.0005$ for frequencies ranging from 2GHz to 10GHz. Both PCBs have two copper layers of $18 \mu m$ and a substrate thickness of 0.5mm. One of these layers is reserved as a ground plane.

For on-board transmission lines, grounded coplanar waveguide (GCPW) are chosen instead of microstrip lines. A GCPW consists of four conductors. The bottom layer is a ground plane. On the top layer, a signal trace is defined separated by a narrow gap from two ground planes at each side. These two ground planes are connected with vias to the bottom ground plane layer. For our application, GCPW has several advantages over microstrip lines. Typically, GCPWs have less radiated emission than microstrip lines. Moreover, GCPW traces offer more freedom in choosing the signal trace width. In the case of microstrip lines, the characteristic impedance is entirely determined by the trace width (for a given frequency, conductor thickness and conductance, substrate height, dielectric constant and loss tangent). For GCPWs, the signal trace width together with the gap between the signal trace and the ground plane next to it determine the characteristic impedance, so several trace widths can give rise to the same characteristic impedance. The penalty for using more narrow signal traces is increased conductor loss. As two signal traces are always separated by a ground plane, the isolation between the waveguides is higher than for microstrip lines, making GCPW suited for compact circuits.

In our case, all transmission lines need to have a characteristic impedance of 50Ω. For the gap between the signal traces and the ground planes, 0.2mm is chosen. This is the minimal gap that can be defined in the copper for the used PCB manufacturing service. According to the Linecalc software of Agilent, for a gap of 0.2mm, a signal trace width of 0.86mm is needed to acquire a GCPW with a characteristic impedance of 50Ω. Typically for GCPW, the via spacing is kept lower than $\lambda/20$, with $\lambda$ the wavelength corresponding to the signal frequency. To determine the propagation speed in the GCPW, the effective dielectric constant $\epsilon_{eff}$ has to be known. This effective dielectric constant will be between the dielectric constant of air (essentially 1) and the dielectric constant of the Itera material. The propagation speed and wavelength are given by:

$$v = \frac{c}{\sqrt{\epsilon_{eff}}},$$

and

$$\lambda = \frac{c}{f \cdot \sqrt{\epsilon_{eff}}}.\quad (3.41)$$

It can be concluded that a higher effective dielectric constant will lead to a lower wavelength. Therefore, the minimum via spacing will be calculated with the largest dielectric constant of the two, being that of the Itera material (3.30). Consequently, for a frequency of 2.5GHz, the
minimum via spacing is given by:

$$\frac{\lambda}{20} = \frac{c}{f \cdot \sqrt{\varepsilon_{eff}}} = \frac{3 \cdot 10^8 m/s}{2.5 GHz \cdot \sqrt{3.3}} = 3.37 mm.$$  \hspace{1cm} (3.42)

The via spacing of the ground planes surrounding the signal trace of the GCPW is chosen to be 1mm.

Besides PCB traces that are used as signal waveguides, there are also PCB traces that are used to power the different active components. Some components draw a considerable amount of current, so care has been taken that these DC feed traces are wide enough to avoid substantial temperature rise. Moreover, the DC feed traces have to be dimensioned wide enough to keep the resistance adequately low, so the voltage drop over the DC feed traces remains negligible.

### 3.2.1 Layout of the radar system PCB

A picture of the fully soldered radar system PCB is given in Fig. 3.23. The designed radar system PCB is also depicted in Fig. 3.24. The top copper layer is depicted in red and the vias in white. These vias connect the bottom and top copper layers. All components that are soldered on the top copper layer are depicted as well. The bottom copper layer is a dedicated ground plane.

While designing, the high frequency components and transmission lines were kept separated from the low frequency components. This is done to avoid that the low frequency circuitry (which has a switching nature) disturbs the sensitive receive circuitry. This separation is indicated in the figure.

The bottom ground plane is unbroken except underneath the TL082 chip. There, a small slot is cut out of the ground plane, to allow for the placement of a decoupling capacitor of 0.22µF between the +Vcc and −Vcc pin of the op-amp. The interruption of the ground plane was chosen above the use of an extra copper layer because this interruption is harmless in the low frequency region where no transmission lines are present.

In Fig. 3.24 it is also indicated where the system is fed with 5V (lower right corner). Extra decoupling capacitors are provided at this location of 2.2µF (GRM188R61A225KE34), 8.2pF (251R14S8R2CV4S) and 2.7pF (251R14S2R7BV4S). These are identical decoupling capacitors as used for the bias tees of the power amplifiers. These decoupling capacitors are also applied at the power supply pin of the VCO and the frequency mixer. Decoupling capacitors are always placed as close as possible to the device they intend to decouple. The smallest capacitor values are placed the closest.

All the other components that are depicted in Fig. 3.24 have already been mentioned throughout subsections 3.1.1 to 3.1.4.

Jumpers are provided on the PCB, so some components can be powered down to facilitate debugging. The parts that can be powered down individually are the DC-DC converter, the op-amp, the receiver chain (consisting of the RX LNA, the RX amplifier and the mixer), the VCO and the TX amplifier.
SubMiniature version A (SMA) connectors are used for the board RX input, TX output and IF output.

As part of the goal of this thesis is to make a compact system, the main concern when designing the PCBs was compactness. The low interconnection lengths that accompanies this approach keeps the losses low. However, the danger of making the PCB too compact is increased crosstalk and consequently poor transmitter/receiver isolation. The approach was to design the PCB as compact as possible, and then check the level of isolation between transmitter and receiver by full EM simulation with the ADS Momentum software of Agilent. The power leakage from transmitter to receiver due to PCB crosstalk has to be negligible compared to the power leakage via the antennas.

A way to increase the transmitter receiver isolation, is by routing the transmitter transmission lines perpendicular to the receiver transmission lines on the board. In this way the crosstalk is kept to a minimum. It can be seen in Fig. 3.24 that the trace of the RX input is perpendicular to the trace of the TX output. The trace of the IF output is closer and parallel to the trace of the TX output, so the crosstalk between these two traces is expected to be higher. It is not necessary to make the IF output trace also perpendicular to the TX output trace, because the LO-IF isolation of the frequency mixer is already very poor, and the crosstalk will be negligible compared to the mixer LO-IF leakage.

The ADS Momentum software of Agilent allows to do full EM circuit co-simulation of the PCB layout and component models. S-parameter models are available only for the capacitors, the inductors, the two power amplifiers and for the power splitter, so only these can be included in the co-simulation. However, also the bias tee components were removed because they seemed to cause erroneous results in the simulation. Only the high frequency region of the PCB is interesting to simulate, depicted in detail in Fig. 3.25. Fig. 3.23 depicts the ports (with...
numbers 1 to 7) that are defined in the Momentum simulation, and for which the S-parameters will be calculated. All simulations are performed with an infinite ground plane with a finite thickness of 18µm and a finite conductivity of 5.8e7 Siemens/m (copper).
Fig. 3.24: Radar board PCB: top copper layer and components
Fig. 3.25: Top copper layer of radar board PCB design; high-frequency
3.2 PCB Layout

Transmission line matching

First, every transmission line segment of the PCB is simulated. The models of the coupling capacitors of 8.2pF are included in this simulation. The input and output matching to 50Ω and the transmission line loss is checked. This is necessary to ensure that the discontinuities that are present in the GCPWs have no major effect. The simulation results can be summarised as follows:

- All transmission line inputs and outputs have a return loss higher than 20dB, which is excellent.

- No transmission line has a loss higher than 0.15dB. The transmission that has a loss of 0.15dB is the transmission line from the board RX input to the input of the LNA, which is also the transmission line with the greatest length.

Power output of the transmitter

Next, we perform a simulation of the power of the transmitted signal. This power can be predicted using data of the VCO output power found in the datasheet, and the simulated gain from VCO output to board TX output. This gain is given by the S-parameter A plot of $S_{34}$. $S_{34}$ as a function of frequency is given in Fig. 3.26. According to the datasheet of the VCO, the output power varies from 7.06dBm to 7.64dBm if the output frequency varies from 2.4GHz to 2.5GHz. If we assume this variation is linear with the frequency, we can plot the power of the transmitted carrier as a function of frequency. This plot is given in Fig. 3.27. The output power varies from 18.51dBm to 18.81dBm, in accordance with the output power expected in subsection 3.1.2.
Fig. 3.26: S-parameter $|S_{34}|$ as a function of frequency

Fig. 3.27: Simulated transmitted carrier power as a function of frequency
3.2 PCB Layout

Receiver gain

The gain from the board RX input to the RF input is given by the S-parameter $|S_{71}|$. The simulated value of $|S_{71}|$ is plotted as a function of frequency in Fig. 3.28. $|S_{71}|$ ranges from 33.5dB to 32.95dB. In subsection 3.1.4 when selecting the power amplifier components, this gain was estimated to be around 33.35dB, which is in very good accordance to the gain acquired in this simulation.

It is mentioned here that the gain from the board RX input to the RF input of the frequency mixer is equal to 32.95dB at 2.5GHz, as this is a value that will be used in the next chapter.

![Fig. 3.28: S-parameter $|S_{71}|$ as a function of frequency](image)

Transmitter receiver isolation

Now a simulation of the transmitter/receiver isolation will be performed. The goal is to compare the transmitter/receiver leakage due to PCB crosstalk to the transmitter/receiver leakage due to antenna leakage. The transmitter/receiver isolation is defined here as the division of the power delivered to the RF input of the frequency mixer by the power generated by the VCO. Then, the transmitter/receiver isolation when only considering PCB crosstalk is equal to the S-parameter $|S_{74}|$. When an antenna isolation of 50dB is assumed, the transmitter/receiver isolation when only considering antenna leakage is given by $|S_{34}|$-50dB+$|S_{71}|$. Both isolations are plotted as a function of frequency in Fig. 3.29.

The isolation when only considering crosstalk is approximately 24dB higher than the isolation when only considering antenna leakage. Therefore can be concluded that the transmitter/receiver leakage due to PCB crosstalk is negligible to the transmitter/receiver leakage due...
It has to be remarked that the simulated isolation is optimistic due to the fact that the bias tees are omitted from the simulation. Some power will unavoidably leak through the bias tee inductors of the transmitter amplifier and the receiver amplifier, lowering the transmitter/receiver isolation.

![Transmitter/receiver isolation as a function of frequency](image_url)

**Fig. 3.29: Transmitter/receiver isolation as a function of frequency**

### 3.2.2 Layout of the IF system PCB

The IF system PCB contains the diplexer discussed in subsection 3.1.6 and the VGA circuit discussed in subsection 3.1.5. Here, only the layout and simulation of the diplexer part is discussed.

A picture of the finished diplexer PCB is given in Fig. 3.30. The diplexer PCB layout is depicted schematically in Fig. 3.31. Again, the top copper layer is in red, the vias are in white and all components that are soldered on the top copper layer are shown. The diplexer is configured as a low pass filter by the 50Ω resistor.

Several layouts of the diplexer PCB were simulated with ADS Momentum to acquire a layout with a good performance. The EM circuit co-simulation uses the component models of the capacitors, inductors and 50Ω resistor. Fig. 3.31 depicts the ports (one and two) that are defined in the Momentum simulation for which the S-parameters are calculated. All simulations are performed with an infinite ground plane with a finite thickness of 18µm and a finite conductivity of 5.8e7 Siemens/m (copper).

The first layout version was made as compact as possible, to approximate the ideal lumped component situation as good as possible. This design had a poor performance due to the parasitic
3.2 PCB Layout

capacitance and inductance as a consequence of the compactness. The parasitic capacitance at several places was lowered by making the design less compact, leaving a larger gap between the signal traces and the nearby ground planes. Of course, the gap of the GCPW was not increased, to maintain the 50\(\Omega\) characteristic impedance. Parasitic inductance in series with the shunt capacitors was reduced by taking a larger number of vias to ground.

The simulation results of the final layout are depicted in Figs. 3.32 and 3.33. The plots also compare the simulation results to the case when the simulation did not include the layout (Figs. 3.20 and 3.21). Fig. 3.32 shows \(|S_{21}|\) as a function of frequency. Again, the most noticeable difference is observed in the stop band. Initially, the attenuation in the stop band is increased compared to the previous simulation. The leaked LO signal is now suppressed by at least 25.32\(\text{dB}\), and the up-mixed signal by at least 51.43\(\text{dB}\). Starting from 5.5\(\text{GHz}\), the low pass filter roll-off decreases. This is due to parasitics of the PCB layout. This parasitic behaviour only starts to dominate at frequencies beyond interest. Furthermore, the attenuation of the filter remains higher than 50\(\text{dB}\) for frequencies at least up to 7\(\text{GHz}\). The insertion loss of the low pass filter has only risen negligibly and is now equal to 0.04\(\text{dB}\) at 1\(\text{MHz}\). A difference that is not noticeable in Fig. 3.32 is that the 3\(\text{dB}\) cut off frequency has decreased and is now equal to 951\(\text{MHz}\). As both cut off frequencies of the low pass filter and high pass filter will have deviated, it is expected that input matching compared is degraded to the previous simulation. The simulated input matching \(S_{11}\) is depicted in Fig. 3.33. The input matching is indeed degraded, but remains acceptable. \(|S_{11}|\) is lower than -14.45\(\text{dB}\) at least up to 7\(\text{GHz}\).

Fig. 3.30: Picture of the finished radar system PCB
Fig. 3.31: Top copper layer of the diplexer PCB design

Fig. 3.32: Low pass filter $|S_{21}|$ as a function of frequency. EM circuit co-simulation of the PCB layout and component models.
3.3 Antenna design

This section describes the design of the transmit and receive antennas that will be used for the radar system. The transmit and receive antennas are designed to exhibit a different circular polarisation (e.g. transmit antenna right-hand circularly polarised (RHCP) and receive antenna left-hand circularly polarised (LHCP)). This has several advantages as described in [26]. A major advantage for CW radar types is that the orthogonal polarisation of the transmit and receive antennas can lead to increased isolation between the two antennas. Moreover, multipath effects are reduced by the rejection of second order reflections.

Patch antennas with a rectangular ring topology and a coaxial feed are used, first proposed in [27]. This antenna type can be circularly polarised with the use of a single coaxial feed point. The patch antenna consists of a conducting top patch and a conducting ground plane with a dielectric in between. The top patch (red) geometry along with the design parameters L, W, l, w, \( x_f \) and \( y_f \) are schematically depicted in Fig. 3.34. A conducting patch with dimensions \( L \times W \) has a slot cut out of dimensions \( l \times w \). The top patch is coaxially fed at a position \( (x_f,y_f) \) with respect to the centre of the patch. The introduction states that the used antennas have to be textile antennas. This restricts the materials that can be used to construct the antennas. The conducting patch and bottom ground plane are fabricated with Flectron material, a copper-plated nylon fabric. The substrate between the patch and ground plane is black foam provided by Javaux with a thickness of 4mm. This material has a dielectric constant of 1.495 and a loss tangent of 0.0168. The applied materials are flexible and light-weight and can be integrated into
3.3 Antenna design

![Rectangular ring topology with coaxial feed](image)

Fig. 3.34: Rectangular ring topology with coaxial feed

The designed antennas have to meet the following specifications:

- Both antennas have to be matched to 50Ω in the frequency band from 2.4GHz to 2.5GHz. Quantitatively, this means that the input reflection coefficient has to be lower than -10dB in that frequency band.

- The transmit antenna has to be RHCP around 2.45GHz. Preferably, the axial ratio has to be lower than 3dB from 2.40GHz to 2.50GHz.

- The receive antenna has to be LHCP around 2.45GHz. Again, preferably, the axial ratio has to be lower than 3dB from 2.40GHz to 2.50GHz.

- The antennas should be directional, meaning they should have a main lobe with a gain larger than 0dBi.

The antenna design parameters are chosen and optimised through simulation with the ADS Momentum software of Agilent. The resulting antenna design parameters are summarised in table 3.4. Remark that the LHCP receive antenna is the RHCP transmit antenna mirrored with respect to the y-axis. The simulation results for the RHCP antenna are given in Figs. 3.35 and
3.3 Antenna design

Table 3.4: Design parameter choice for the transmit and receive antenna

<table>
<thead>
<tr>
<th>Design parameter</th>
<th>RHCP transmit antenna (mm)</th>
<th>LHCP receive antenna (mm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>W</td>
<td>43.5</td>
<td>43.5</td>
</tr>
<tr>
<td>L</td>
<td>47.5</td>
<td>47.5</td>
</tr>
<tr>
<td>w</td>
<td>7</td>
<td>7</td>
</tr>
<tr>
<td>l</td>
<td>4</td>
<td>4</td>
</tr>
<tr>
<td>xf</td>
<td>6.75</td>
<td>-6.75</td>
</tr>
<tr>
<td>yf</td>
<td>11</td>
<td>11</td>
</tr>
</tbody>
</table>

Fig. 3.35a shows that the antenna is matched from 2327MHz to 2593MHz, containing the wanted frequency band. Table 3.5 shows that the radiation efficiency is about 76% in this frequency band. Best circular polarisation is expected at 2448MHz (see annotation Fig. 3.35a). Table 3.5 shows that the simulated axial ratio at 2.45GHz in the +z direction is indeed very close to 0dB (0.564dB), indicating circular polarisation. Acquiring axial ratios below 3dB at 2.4GHz and 2.5GHz was not achieved, as the circular polarisation is in a too small band. The polarisation is nevertheless elliptical at 2.4GHz and 2.5GHz, still partially providing the advantages of circular polarisation.

The antenna gain at 2.45GHz is plotted as a function of elevation angle $\theta$ in Fig. 3.36. The maximum gain is in the +z direction and equal to 7.316dB. The 3dB beamwidth of the main lobe of the antenna is also annotated and is equal to 76°. Table 3.5 shows that the gain in the +z direction of the antenna is practically constant in the frequency band from 2.4GHz to 2.5GHz.

In subsection 3.1.2 it was predicted that the most powerful harmonic in the frequency spectrum of the signal applied to the transmit antenna is the second harmonic around 4.9GHz with a power of about -13.6dBm. Fig. 3.35b shows the input reflection coefficient of the antenna from 3GHz to 6GHz, and the return loss at 4.90GHz is annotated to be 4.182dB. Taking this return loss and the radiation efficiency of 53.3% at 4.9GHz into account, the radiated power at the second harmonic frequency $P_{rad,2}$ can be calculated to be:

$$P_{rad,2} = -13.6dBm + 10 \cdot \log \left(1 - 10^{-4.182/10}\right) + 10 \cdot \log (0.533) = -18.42dBm. \quad (3.43)$$

The power radiated at the second harmonic is still significant. Modifying the radar system to comply to frequency spectrum regulations will presumably require additional filtering of the signal before transmission.
### 3.3 Antenna design

![Graph 1](image1.png)

(a) Simulated input matching of RHCP antenna from 2GHz to 3GHz

![Graph 2](image2.png)

(b) Simulated input matching of RHCP antenna from 3GHz to 6GHz

---

*Fig. 3.35: Simulated input matching of RHCP antenna*
Elevation gain pattern ($\phi=0^\circ$) transmit antenna at 2.45GHz

Fig. 3.36: Elevation gain pattern ($\phi = 0^\circ$) transmit antenna at 2.45GHz

<table>
<thead>
<tr>
<th>frequency (GHz)</th>
<th>Axial ratio (dB)</th>
<th>Efficiency (%)</th>
<th>Gain (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>2.40</td>
<td>5.093</td>
<td>75.3</td>
<td>7.254</td>
</tr>
<tr>
<td>2.45</td>
<td>0.564</td>
<td>75.7</td>
<td>7.316</td>
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<td>2.50</td>
<td>4.786</td>
<td>76.6</td>
<td>7.409</td>
</tr>
<tr>
<td>4.90</td>
<td>-</td>
<td>53.3</td>
<td>-</td>
</tr>
</tbody>
</table>

Table 3.5: Simulation results RHCP antenna. Axial ratio and gain are for the +z direction ($\theta = 0^\circ$ and $\phi = 0^\circ$)
Chapter 4

Validation of the radar prototype

4.1 Verification the subsystems

Prior to validating the full radar system, the correct operation of the different parts of the system will be verified. Jumper are provided on the PCB to be able to power only specific parts of the system in order to test them separately. The parts that can be powered separately are the DC-DC converter, the op-amp, the receiver chain (consisting of the RX LNA, the RX amplifier and the mixer), the VCO and the TX amplifier.

4.1.1 VCO and TX amplifier

First, the cascade of the VCO, power splitter and TX amplifier is tested. Therefore, the VCO, TX amplifier and frequency mixer all have to be powered. The frequency mixer (so also the entire receive chain) has to be powered so it provides a 50Ω load to the power splitter (the frequency mixer has an internal amplifier built in its LO path). The board RX input and IF output are terminated with a 50Ω dummy load and the board TX output is connected to a spectrum analyser. The spectrum analyser used for all measurements performed in this chapter is the Rohde & Schwarz FSV Signal Analyzer 10Hz-40GHz. The tuning voltage $V_{tuning}$ of the VCO is set with an external power supply. With the tuning voltage of the VCO set to a fixed value, the output signal is expected to be a fixed frequency carrier with harmonics. The power $P_c$ and frequency $f_c$ of the carrier are measured as a function of $V_{tuning}$. The power $P_{2c}$ of the second harmonic is also measured. The resulting measurements are tabulated in table 4.1. The power measurements are performed with the spectrum analyser Resolution Bandwidth (RBW) set to 10MHz, and the frequency measurements are performed with the RBW set to 100kHz. Details about measurement equipment settings are from now on annotated in the tables or plots containing the corresponding data.

The carrier power $P_c$ is plotted as a function of the tuning voltage $V_{tuning}$ in Fig. 4.1. The plot also shows the predicted carrier power as simulated in subsection 3.2.1 but now as a function of the tuning voltage. The measured output power is a bit lower than simulated,
but never deviates for more than about 0.7dB. This small difference has various causes. The component models and information in the datasheet are only valid for a certain temperature (25°C), which is different from the temperature of the measured hardware. The PCB substrate parameters will also be slightly different than foreseen. Moreover, in reality, additional losses compared to the simulation are unavoidable: soldering of connectors and components is not ideal.

It is observed that the carrier power is not constant as a function of the tuning voltage. The carrier power varies between 17.77dBm and 18.65dBm. This means that, when sweeping the VCO tuning voltage, the transmitted signal will also have an amplitude modulation besides the intended FM. This amplitude modulation is not modelled in the analysis of subsection 2.4.3, and it is assumed to have a negligible influence on the resulting converted signal spectrum.

In subsection 3.1.2, it is predicted that, if the VCO is tuned to 2.45GHz, the 2nd harmonic is at least 32.15dB below the carrier. Indeed, it is measured here that the 2nd harmonic is 33.9dB below the carrier if the VCO is tuned to about 2.45GHz.

The carrier frequency $f_c$ is plotted as a function of the tuning voltage $V_{\text{tuning}}$ in Fig. 4.2. The goal of measuring carrier frequency as a function of tuning voltage is measuring the linearity of the FM. Table 4.1 includes the calculation of the tuning sensitivity (MHz/V). The linearity as defined in equation 2.50, calculated with the data of table 4.1, is equal to 0.31. This linearity is very poor according to [7]. This non-linearity results in a reduced range resolution, because the instantaneous frequency of the converted signal is spread. The non-linearity is not quadratic, and because of this equation 2.51 cannot be used. It is therefore very difficult to estimate the effect on the converted signal spectrum. If it is assumed that equation 2.51 is nevertheless valid, the range resolution is dominated by the bandwidth at ranges below 7.26m, and by the FM linearity above 7.26m.

It is expected that the linearity of the generated frequency sweep will be even worse, because the tuning voltage, generated by the triangle voltage generator, will also be slightly non-linear. These non-linearities have to be combined to determine the total non-linearity of the frequency sweep.
4.1 Verification the subsystems

Fig. 4.1: Carrier frequency $f_c$ as a function of tuning voltage $V_{tuning}$

Fig. 4.2: Carrier power $P_c$ as a function of tuning voltage $V_{tuning}$
4.1 Verification the subsystems

Table 4.1: Carrier power and frequency as a function of VCO tuning voltage

<table>
<thead>
<tr>
<th>( V_{\text{tuning}} ) (V)</th>
<th>( P_c ) (dBm)</th>
<th>( P_{2c} ) (dBm)</th>
<th>( f_c ) (MHz)</th>
<th>Tuning sensitivity (MHz/V)</th>
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</thead>
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<td>RBW 10MHz</td>
<td>RBW 100kHz</td>
<td>calculated</td>
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4.1.2 Triangle voltage oscillator

Table 4.1 shows that the tuning voltage should sweep between approximately 9.4V and 10.93V in order to make the carrier frequency sweep between 2.4GHz and 2.5GHz. This is an average voltage of 10.165V and a peak-to-peak amplitude of 1.53V. The trimmer resistors can be adjusted to give the output triangle voltage the correct average and peak-to-peak amplitude.

The tuning characteristic varies with temperature, meaning the trimmers need to be adjusted frequently to correct this. A control loop would be needed in a real system in order to keep the transmitted signal within the correct frequency band.
The triangle voltage generator output, set to sweep from approximately 9.5V to 11V at approximately 10kHz was measured with an oscilloscope. The measurement is depicted in Fig. 4.3. It is easily observed that the acquired voltage triangle is not very clean. As these non-linearities might severely influence the range resolution, and the converted signal spectrum in general, it was decided to use a function generator to generate the tuning voltage of the VCO in the following measurements.

4.1.3 TX and RX Antennas

In this subsection measurements are performed on the constructed RHCP and LHCP rectangular ring antennas that were designed in section 3.3. The used network analyser is the Agilent N5242A PNA-X Network Analyser. As the circular polarisation is preferred to be optimal at 2.45GHz, a large number of antennas were constructed, and the best ones were selected. First, the input matching to 50Ω is measured, and next the isolation between the two antennas.

The measured input reflection coefficient of the RHCP antenna, together with the simulated input reflection coefficient is depicted in Figs. 4.4a and 4.4b. Fig. 4.4a shows the input reflection coefficient for frequencies around the carrier frequency from 2.4GHz to 2.5GHz. The measured input reflection coefficient is in good accordance with the simulated values. The measured antenna has a slightly larger bandwidth (2.32GHz to 2.61GHz) than simulated. The measured matching is worse than the simulation, but still excellent (<18dB) in the entire frequency band of interest (2.4GHz to 2.5GHz). The frequency at which the antenna will be optimally circularly polarised is about 2.435GHz, very close to the wanted value of 2.45GHz. Fig. 4.4b shows the input matching of the antenna around frequencies of the second harmonic. The measured
matching for the second harmonic (4.8GHz-5GHz) is worse than simulated. If it is assumed that
the antenna efficiency is equal to the simulated antenna efficiency, the radiated power at the
second harmonic frequency is a bit less than calculated in section 3.3.

The measured input reflection coefficient of the LHCP antenna, together with the simulated
input reflection coefficient is depicted in Figs. 4.5a and 4.5b. Fig. 4.5a shows the input reflection
coefficient for frequencies around the carrier frequency from 2.4GHz to 2.5GHz. Almost identical
remarks can be given about this antenna. The measured bandwidth (2.315GHz to 2.605GHz)
is slightly larger than simulated. The measured matching is worse and more asymmetrical
than simulated, but still good (< 14.82dB) in the entire frequency band of interest (2.4GHz to
2.5GHz). This difference is caused by variation in the antenna construction. The frequency at
which the antenna will be optimally circularly polarised is about 2.455GHz, again very close
to the wanted value of 2.45GHz. Fig. 4.5b shows the input matching of the antenna around
frequencies of the second harmonic. Exactly the same remark can be made as for the previous
antenna.

Next the antenna isolation is considered. The antenna isolation will determine the power of
the directly leaked signal, and it was assumed through chapter 3 that the antenna isolation is
50dB. As the used patch antennas are circularly polarised, all four edges are radiating edges.
Both antennas have to be placed close enough together so they can be fitted on one piece of
garment, while keeping the isolation between the two antennas high enough. Therefore, the
patch antennas are arranged in such a manner that the two patches share the same diagonal. In
this way, no pair of radiating edges face each other directly, and the isolation is as good as it can
be for a given separation. Fig. 4.6 is a picture of the two antennas mounted on cardboard in this
way. When the two patch antennas centres are separated by approximately 30cm, an isolation of
about 50dB is obtained in the frequency band from 2.4GHz to 2.5GHz. The measured isolation
in the frequency band from 2GHz to 3GHz is depicted in Fig. 4.7.

4.1.4 RX amplifiers stability

In this test, only the receive chain (RX LNA, the RX amplifier and the mixer) is powered.
The VCO and the TX amplifier are turned off. As already explained in subsection 3.1.4 it is
possible that the cascade of the LNA and RX amplifier becomes unstable. In this case, the RX
amplifier will oscillate at high power (and be in saturation) due to positive feedback via the DC
feed network. The output power of the RX amplifier would certainly be greater than the 1dB
output compression point (about 22.5dBm) in this case. This oscillation would be observed at
the IF output of the frequency due to finite RF-IF isolation (order 50dB for the used frequency
mixer) with a power of at least 22.5dB-50dB=-27.5dBm. This is a power level that can be easily
observed with a spectrum analyser.

Therefore, the board RX input and TX output are terminated with a dummy load, and the
IF output of the frequency mixer is observed with a spectrum analyser from DC to 10GHz. The
RBW is set to 1kHz and the amplitude reference to -40dBm in order to make the noise floor
very low. The noise floor in this case is about -110dBm/Hz. No oscillations originating from
the radar board are observed.

This unstable oscillation would also be observed at the TX output of the board (because
of propagation over the DC feed network) and also at the RX input of the board (because of
propagation over the DC feed network and the RX LNA not being a perfect unilateral element).
4.1 Verification the subsystems

Moreover, no oscillations originating of the radar board are observed at these ports.

4.1.5 RX amplifiers and frequency mixer

Next, the correct operation of the frequency mixer is investigated. The gain provided by the RX LNA and RX amplifier is also checked. The receive chain, the TX amplifier and the VCO
4.1 Verification the subsystems

Fig. 4.6: Both antennas mounted on cardboard

![Antennas mounted on cardboard]

TX/RX antenna isolation as a function of frequency

![Graph showing antenna isolation]

Fig. 4.7: Isolation of the RHCP and LHCP antenna from 2GHz to 3GHz

are all powered. The tuning voltage of the VCO is set with a power supply, so that the signal frequency at the board TX output is 2.5GHz. To this end, a voltage of 10.963V was needed. As the signal generated by the VCO is only needed as the mixer LO signal, the board TX output is terminated with a 50Ω dummy load. A signal generator generates a signal of -50dBm at 2.4GHz at the board RX input. The actual generated power was -51.57dBm (measured with the spectrum analyser). This signal is amplified by the RX LNA and the RX amplifier, and is then applied to the RF port of the frequency mixer. The desired mixing product will have
4.1 Verification the subsystems

A frequency of 100MHz with a power determined by the gain of the amplifier cascade, and the mixer conversion loss.

As simulated in section 3.2, the gain from the board RX input to the RF input of the frequency mixer is equal to 32.95dB at 2.5GHz. The power of the signal at the RF input of the frequency mixer is then predicted to be -18.62dBm (-51.57dBm+32.95dB). According to the frequency mixer datasheets, the conversion loss around the frequency of 2.45GHz is equal to about 5.65dBm. Therefore, a 100MHz mixing product with a power of -24.27dBm (-18.62dBm-5.65dB) is expected.

The IF output of the frequency mixer is observed with the spectrum analyser. The measured power spectrum from DC to 10GHz is depicted in Fig. 4.8. It is observed that signals with significant power are present at baseband (the mixing product), around the carrier frequency of 2.45GHz and around harmonics of this 2.45GHz.

Fig. 4.9 shows the baseband region from DC to 200kHz in more detail. There is one clear signal at 102.3MHz of -25.17dBm. This signal is the wanted mixing product. Comparing its power with the predicted value of -24.27dBm, there are about 0.9dB more losses than expected. These extra losses will likely be caused by a combination of slightly more substrate losses than simulated, slightly less amplifier gain, slightly higher mixer conversion loss, losses due to non-ideality of soldered connections and in particular the soldering of the SMA connector to the PCB. Moreover, the information found in the datasheets is tied to a temperature of 25°C, which did not apply to the tested hardware. Higher order mixing products are not observed in Fig. 4.9, so these are well below the first order mixing product.

Fig. 4.10 shows the frequency band from 2GHz to 3GHz in more detail. It is remarkable that the leaked LO signal that is observed at the mixer IF output is approximately equally strong as the LO signal at the mixer LO port itself (about 4.30dBm). This is caused by the amplifier in the LO path embedded in the frequency mixer. It was thoroughly checked that the poor LO-IF isolation is only caused by the mixer component itself, and not by propagation of the LO signal over the DC feed network or by crosstalk. This leaked LO signal has a frequency that is much higher than the frequencies of the wanted mixing products around baseband so it can be filtered away without further problems. In addition, two spurious products are observed in Fig. 4.10.

Fig. 4.8 also shows groups of signals around 5GHz and 7.5GHz. Both groups consist of a harmonic of the VCO leaked from mixer LO port to IF port, and two spurious products. All of these unwanted signals can be filtered away without further problems.
4.1 Verification the subsystems

Fig. 4.8: Power spectrum IF output (dBm) from DC to 10GHz (RBW 100kHz, reference level 10dBm)

Fig. 4.9: Power spectrum IF output (dBm) from DC to 200MHz (RBW 100kHz, reference level 0dBm)
4.1 Verification the subsystems

Fig. 4.10: Power spectrum IF output (dBm) from 2GHz to 3GHz (RBW 100kHz, reference level 15dBm)

4.1.6 Diplexer

This subsection describes the testing of a soldered prototype of the diplexer part of the IF system PCB described in subsection 3.2.2. When testing the diplexer with the components mentioned in tables 3.2 and 3.3, the input matching of the diplexer was not meeting the minimum return loss of 10dB from DC to 5GHz as stated in subsection 3.1.6. Input matching was not matched to 50Ω anymore for frequencies higher than 2.83GHz. Slightly different choices for the components of the high pass filter section were made in an attempt to solve this. The components mentioned in tables 4.2 and 4.3 result in a diplexer meeting all specifications. The most important change was the replacement of the $C_3$ 4 component parallel combination (4x1.6pF) by a 3 component parallel combination (2x2.7pF+1pF).

Fig. 4.11 depicts the measured filter characteristic $|S_{21}(f)|$. The low pass filter has a 3dB cut off frequency of 929MHz. The insertion loss for the useful converted signal frequencies from 10kHz to 1MHz is maximally 0.08dB. The leaked LO signal around 2.45GHz is suppressed by at least 24.85dB and the frequency component around 4.9GHz by at least 44.19dB. The measured input reflection coefficient of the soldered diplexer is shown in Fig. 4.12. The measured return loss is higher than 10dB for frequencies up to 7GHz. At 2.40GHz and 4.80GHz, frequencies where considerable power at the frequency mixer IF output is expected, the return loss is equal to 17.30dB and 12.52dB respectively. It can be concluded that all low pass filter requirements stated in subsection 3.2.2 are met.
### 4.1 Verification the subsystems

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Table 4.2: Soldered set of capacitor and inductor components for the diplexer

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Table 4.3: Component values for the diplexer schematic

![Low pass filter $|S_{21}|$ as a function of frequency](image.png)

Fig. 4.11: Measured filter characteristic soldered diplexer $|S_{21}|$ as a function of frequency.
4.2 Validation of the radar prototype

In this section, the full radar system is tested and validated. After the initial tests performed in the previous section, it is expected that every component functions normally. To generate the VCO tuning voltage, however, an external function generator is used instead of the on-board triangle voltage oscillator, for reasons mentioned above. The function generator used is the Agilent 33220A 20MHz Function / Arbitrary Waveform Generator. For the rest of this section, it is assumed that the function generator is set to a symmetrical triangle function with an average voltage and peak-to-peak amplitude appropriate to sweep the VCO between 2.4GHz and 2.5GHz. The sweep repetition frequency is set to 10kHz.

4.2.1 Direct coaxial connection

The first test involves connecting a coax cable with extra attenuators between the board TX output and the board RX input. The board IF output is observed with the spectrum analyser. The radar system should be capable of measuring the length of the coax cable, by taking into account the correct propagation speed. As no antennas are used, the directly leaked signal is limited to PCB crosstalk. In this way, this test is a validation of the correct operation of the radar system in case no significant directly leaked signal is present.

Coaxial cable of 16m

The coax cable used is a 50Ω coax cable, type 104, of Spectrum Elektrotechnik GmbH. The length is 16m, the velocity factor is specified to be 84% ± 2% and the attenuation at 2.45GHz
is specified to be 0.29dB/m. The cable attenuation is thus expected to be 4.64dB. A number of extra attenuators is also added (VAT-20+ [29], two VAT-10+ [30], VAT-9+ [31], VAT-8+ [32], VAT-7+ [33], VAT-5+ [34] fixed SMA attenuators from Minicircuits) to avoid receiver overloading, making the total expected attenuation to be 73.64dB.

The coax cable will cause a delay given by:

$$\tau = \frac{16m}{0.84 \cdot c} = 63.49\,\text{ns}. \quad (4.1)$$

The delay caused by the attenuators is not specified by the datasheets, and it is assumed that this delay is negligible. Hence, the total delay is given by $\tau$. It follows from subsection 2.4.3 that the envelope of the spectral components of the converted signal will be maximal around the frequency $f_{max}$, given by equation 2.38, with the Doppler shift $\omega_d$ set to zero:

$$f_{max} = 2 \cdot \Delta f \cdot \frac{\tau}{T_m} = 2 \cdot 100\,\text{MHz} \cdot \frac{63.49\,\text{ns}}{0.1\,\text{ms}} = 126.98\,\text{kHz}, \quad (4.2)$$

with $\Delta f$ the total frequency swing of 100MHz and $T_m$ the modulation period of 0.1ms. It is therefore expected that the $13^{th}$ harmonic of 10kHz will be the spectral component with the maximum power.

The measured power spectrum of the IF output signal is depicted in Fig. 4.13. The frequency interval is restricted to the region where significant power is present (DC-200kHz). As expected, it is observed that the spectral envelope is maximal at the $13^{th}$ harmonic. The spectral component power is annotated in Fig. 4.13.

The spectral envelope also shows a maximum around 10kHz. As a coax cable with attenuators is used to connect the transmitter with the receiver, the directly leaked signal is limited to PCB crosstalk. Nevertheless, it still has an observable effect on the power spectrum of the IF output signal. Usually, as the directly leaked signal has a very short delay, a maximum of the converted signal spectral envelope at DC is expected. As the IF output is connected to the spectrum analyser with a DC blocker (to protect the spectrum analyser) this is not observable, and the maximum occurs at 10kHz instead.
4.2 Validation of the radar prototype

This test was repeated with a different coax cable. This time, the used coax cable is a 50Ω coax cable type 100 of Spectrum Elektrotechnik GmbH with a length of 8m, a velocity factor of 75% ± 2%, and an attenuation of 0.39dB/m at 2.45GHz. The cable attenuation is thus expected to be 3.16dB. A number of extra attenuators is added (VAT-20+, VAT-10+, VAT-9+, VAT-7+, VAT-5+ fixed SMA attenuators from Minicircuits) to avoid receiver overloading, making the total expected attenuation to be 54.16dB.

The spectral envelope of the frequency components of the converted signal should be maximal around the frequency $f_{\text{max}}$, again given by equation 2.38 with $\omega_d = 0$:

$$f_{\text{max}} = 2 \cdot \Delta f \cdot \frac{\tau}{T_m} = 2 \cdot 100MHz \cdot \frac{0.75 \cdot c}{0.1ms} = 71.11kHz.$$  \hspace{1cm} (4.3)

The measured IF output signal power spectrum is shown in Figs. 4.14 and 4.15. Fig. 4.14 shows that the power of the spectral components is only significant in the region from DC to about 200kHz. Fig. 4.15 is restricted to this region.

As the attenuation is now lower, the power of the spectral components in Fig. 4.15 is greater compared to Fig. 4.13. The component with maximum power has a power of -42.17dBm at 70kHz. The component just below 70kHz has a power of -49.93dBm, and the component just above 70kHz has a power of -45.12dBm, indicating that the maximum of the spectral envelope lies just above 70kHz. This is in correspondence with the calculated value of equation 4.3.
4.2 Validation of the radar prototype

Fig. 4.14: Power spectrum IF output for type 100 8m coax cable from DC to 1MHz (RBW 1kHz, reference level 0dBm)

Fig. 4.15: Power spectrum IF output for type 100 8m coax cable from DC to 200kHz (RBW 1kHz, reference level 0dBm)

4.2.2 Anechoic chamber

The tests in this subsection are performed in the anechoic chamber using the constructed TX/RX antenna pair from subsection 4.1.3. Both the RX and TX antennas are connected to the radar
system with 1m HUBER+SUHNER SUCOFLEX 104 coax cables with a velocity factor of 77% \[36\]. Every range measurement will have to take this extra length into account. The board IF output is observed with the spectrum analyser. Now, due to leakage between the two antennas, a directly leaked signal with significant power is present at the board RX input. Care has been taken when fixing the TX/RX antenna pair on cardboard, so as to ensure that the isolation between the antennas is adequately high, in order for the directly leaked signal to be sufficiently weak. Several scenarios involving stationary, moving, human and non-human targets are tested.

**No targets (blank test)**

First, a measurement is performed of the converted signal power spectrum when no radar targets are present in the anechoic room. In this case the only contribution to the converted signal power spectrum originates from the directly leaked signal, and the effect of the directly leaked signal on the converted signal power spectrum is measured.

The measured converted signal power spectrum from DC to 500kHz is shown in Fig. 4.16. As expected, the directly leaked signal mainly affects the lower frequency spectral components. Significant power in the spectral components from DC to about 100kHz is observed. This frequency region is shown in more detail in Fig. 4.17. The component with maximal power at 20kHz has a power of \(-48.4\) dBm. This is the only spectral component that exceeds -60dBm. The significant power that is always present in these spectral components caused by the directly leaked signal, may mask out targets with a low radar cross section that give rise to these spectral components.

In subsection 3.1.6, attention was given to suppress the contribution of the directly leaked signal to the converted signal. This only involved blocking the DC component, while here the most powerful spectral component caused by the directly leaked signal is at 20kHz. This is a consequence of connecting the radar system to the transmit and receive antennas with coax cables of non-negligible length (compared to the radar resolution of 75cm). If the radar system is to be made compact, the connection cables will be much shorter, and the most powerful spectral component in the converted signal spectrum caused by the directly leaked signal will be at DC.
4.2 Validation of the radar prototype

The next test is performed with a single stationary radar target present in the anechoic room. An almost flat patch of copper tape, perpendicular to the line of sight of the radar system,
4.2 Validation of the radar prototype

with an area of 14cm x 14cm is positioned at a range of 4.92m from the radar system antennas. According to [1], the approximate radar cross section $\sigma$ for a flat plate with area $A$ perpendicular to the line of sight of the radar system is given by:

$$\sigma = \frac{4\pi \cdot A^2}{\lambda^2} = \frac{4\pi \cdot (0.14 \cdot 0.14m^2)^2}{\lambda^2} = 0.322m^2,$$  \hspace{1cm} (4.4)

with $\lambda$ being the wavelength of the radar signal, equal to about $c/2.45GHz = 12.2cm$. The estimated radar cross section of a human is $0.1m^2$ to $1m^2$ around the used frequencies, according to [3], so the copper tape patch has a radar cross section that is comparable to that of a human.

The measured IF output power spectrum is shown in Fig. 4.18. Only the frequency region where spectral components with significant power are present is shown. When comparing this power spectrum to the situation when no radar targets are present, as in Fig. 4.17, it is observed that the spectral component power at 90kHz is now much stronger and exceeds -60dBm. It can be concluded that a radar target is detected. The spectral envelope shows a maximum at a frequency slightly higher than 90kHz. The target range that corresponds to the 9th harmonic is given by equation 2.42:

$$R = 9 \cdot \frac{c}{4 \cdot 100MHz} = 6.75m.$$  \hspace{1cm} (4.5)

As already mentioned, this result includes the 1m connection cables of the radar system to its antennas. Correcting for two cables of 1m with a velocity factor of 77%, a measured range $R_{eff}$ is acquired, given by:

$$R_{eff} = 6.75m - 1m/0.77 = 5.45m.$$  \hspace{1cm} (4.6)

This result is wrong, as the real range of the copper patch is 4.92m. This, however, can be explained as follows. In the used anechoic chamber, a fixed antenna positioning system is present. When unused, as is the case in our setup, a large metallic part of this antenna positioning system is exposed in the anechoic chamber. As this exposed metallic part is collinear with the copper patch and the radar system antennas, and as it is a radar target with a very large radar cross section, the experimenter has tried to cover these metallic parts with absorbers to the best of his ability. Nevertheless, the reflections of the metallic parts of the antenna positioning system still manage to mask out the signal reflected by the copper patch. In this way, the 5.45m is actually the measurement of the distance from the radar system antennas to the antenna positioning system. This is in good accordance with the actual distance of 5.5m from the radar antennas to the antenna positioning system.
4.2 Validation of the radar prototype

Stationary target with large radar cross section

As in the previous test, the target with a human-like radar cross section was masked out by a nearby target with a radar cross section that is much larger, this test will be repeated with a stationary radar target that deliberately has an even larger radar cross section. To this end, a square copper plate of 50cm x 50cm is used. The radar system antennas point at the copper plate from a distance of 4.75m. The approximate radar cross section $\sigma$ can be calculated with the already mentioned formula:

$$\sigma = \frac{4\pi \cdot A^2}{\lambda^2} = \frac{4\pi \cdot (0.25m^2)^2}{\lambda^2} = 52.38m^2. \quad (4.7)$$

Compared to the estimated radar cross section of a human, this is a radar target with a very large radar cross section.

The measured IF output power spectrum is shown in Figs. 4.19 and 4.20. The first figure shows that the power of the spectral components is only significant in the region from DC to about 200kHz. Fig. 4.20 only shows this region.

It has to be remarked that this measurement is part of a different measurement series. This is important, as in the second series of measurements, it was noticed that the effect of the directly leaked signal on the converted signal power spectrum had drastically changed. This is the consequence of dismounting and remounting the radar antennas. Therefore, power spectra acquired during this test cannot be compared with the blank power spectra of Figs. 4.16 and 4.17. Instead it is mentioned here that the blank test for this measurement series showed that
the directly leaked signal gives rise to strong components at the frequencies of 20kHz (-45dBm) and 30kHz (-45dBm).

The converted signal spectral envelope shows maxima around 20kHz and around 80kHz. In the blank test, it was observed that the directly leaked signal gives rise to a maximum in the spectral envelope of the converted signal around 20kHz, with a spectral component at 10kHz of approximately -60dBm, at 20kHz of -45dBm and at 30kHz of -45dBm. Approximately the same spectral component powers are observed in Fig. 4.20 (exact values: 10kHz -60.23dBm, 20kHz -44.71dBm, 30kHz -45.08dBm). Taking this into consideration, the spectral envelope maximum around 20kHz, is caused by the directly leaked signal only, and does not lead to a detected radar target.

The powers, of the components around the maximum of the spectral envelope at about 80kHz, are annotated in Fig. 4.20. When the 8th harmonic of 10kHz is the most powerful component in the converted signal power spectrum, while the components at 70kHz and 90kHz have approximately the same power, it is expected that a reflector is present at a range $R$ given by equation 2.42:

$$R = \frac{8 \cdot c}{4 \cdot 100 MHz} = 6m.$$  \hspace{1cm} (4.8)

Correcting for the 1m antenna connection cables that appear as a 1.30m range increase to the radar system, a range of 4.70m is measured. This measured target range is in very good accordance to the actual target range of 4.75m. Compared to the previous test setup, the situation is reversed. The antenna positioning system is now masked out by the copper plate.

Fig. 4.19: Radar target with $\sigma = 52.38m^2$. Power spectrum IF output from DC to 1MHz (RBW 500Hz, reference level 0dBm)
4.2 Validation of the radar prototype

Fig. 4.20: Radar target with $\sigma = 52.38m^2$. Power spectrum IF output from DC to 200kHz (RBW 500Hz, reference level 0dBm)

Moving human target

The previous attempt to detect a target with a radar cross section that is human-like failed, because this target was masked out by the nearby antenna positioning system that has a substantially larger radar cross section. This masking out can be avoided by a greater separation between the interfering antenna positioning system and the human target. The converted signal power spectrum will be measured at the exact moment that a human subject is walking at a normal speed halfway between the antenna positioning system and the radar system antenna. The theory of subsection 2.4.3 is applicable to this situation. This showed that a target moving with a constant radial velocity gives rise to spectral components at the frequencies $|k \cdot f_m \pm f_d|$, with $k$ an integer, $f_m$ the modulation frequency of 10kHz and $f_d$ the Doppler frequency shift caused by the target’s radial velocity.

The measured converted signal power spectrum is shown in Fig. 4.21. When comparing this power spectrum to the blank test of Fig. 4.17, it is noticed that the power of the spectral components ranging from 30kHz to 100kHz has increased. The power of these spectral components has increased by one or more radar targets. When looking at the data in more detail, it can be seen that there are two kinds of spectral components present. Firstly, there are spectral components at frequencies $|k \cdot f_m \pm f_d|$, corresponding to a target moving with a certain radial speed (related to the Doppler shift $f_d$). It is also possible that there are multiple moving targets with an identical radial speed. Secondly, there are spectral components at frequencies $k \cdot f_m$ corresponding to stationary targets. As the two different kinds of spectral components are certainly caused by different radar targets, separate spectral envelopes have to be considered for
these two kinds of spectral components, in order to discern the moving and stationary targets.

The spectral envelope of the components at frequencies $|k \cdot f_m \pm f_d|$ has a maximum around 50kHz. This maximum is depicted in more detail in Fig. 4.22. Consequently, a moving target is detected at a range corresponding to the 5th harmonic, with a radial speed corresponding to the Doppler frequency shift $f_d$ that is given by half the frequency separation of the two peaks in Fig. 4.22. The exact frequencies of the two peaks are annotated in Fig. 4.22. With this Doppler frequency shift $f_d$, the moving target’s radial speed $|v_{rad}|$ can be calculated in the following way:

$$
\frac{f_d}{2} \cdot \frac{\left|v_{rad}\right|}{c} \cdot 2.45GHz \Rightarrow \left|v_{rad}\right| = \frac{f_d \cdot c}{2 \cdot 2.45GHz} = \frac{50.0207kHz - 49.9786kHz}{2 \cdot 2.45GHz} \cdot c = 1.29m/s.
$$

It should be noted that there is no way to determine whether the target is approaching or receding in any other way than by repeated range measurements. The target range corresponding to the 5th harmonic is given by equation 2.42:

$$
R = 5 \cdot \frac{c}{4 \cdot 100MHz} = 3.75m.
$$

Correcting for the 1m antenna connection cables that appear as a 1.30m range increase to the radar system, a range of 2.45m is measured. This is approximately the range of the human target when the power spectrum was measured. It can be seen that also power is present at exactly 50kHz. This power, however, is only slightly higher compared to the blank test of Fig. 4.17 when only the directly leaked signal is present. Therefore, no stationary target is detected at the range of 2.45m.

The spectral envelope of the components at frequencies $k \cdot f_m$ has a maximum around 90kHz. This maximum is depicted in more detail in Fig. 4.23. This maximum corresponds to the antenna positioning system at a range of 5.5m again, which was already seen in the test "Stationary target with human radar cross section". Moreover, weak spectral components are present at $|90kHz \pm f_d|$, which are the sidelobes of the spectral envelope corresponding to the moving target present at 2.45m.
4.2 Validation of the radar prototype

Fig. 4.21: Walking human. Power spectrum IF output from DC to 110kHz (RBW 5Hz, reference level 0dBm)

Fig. 4.22: Walking human. Power spectrum IF output from 49.5kHz to 50.5kHz (RBW 5Hz, reference level 0dBm)
4.2 Validation of the radar prototype

![Power spectrum IF output](image)

Fig. 4.23: Walking human. Power spectrum IF output from 89.5kHz to 90.5kHz (RBW 5Hz, reference level 0dBm)

**Loudspeaker target**

This time, as a moving radar target, a large loudspeaker, with copper tape stuck to the speaker cone is used. The copper area is about $14\text{cm} \times 14\text{cm}=196\text{cm}^2$. The same formula can be used to estimate the approximate radar cross section $\sigma$ of the loudspeaker cone with area $A$:

$$\sigma = \frac{4\pi \cdot A^2}{\lambda^2} = \frac{4\pi \cdot (196\text{cm}^2)^2}{\lambda^2} = 0.322\text{m}^2. \quad (4.11)$$

This is a radar target with a cross section that is comparable to that of a man. When a signal of 10-20Hz is applied to the speaker, the cone makes a large movement that should be detectable by the radar system. The loudspeaker is set to produce a sine wave tone of 10Hz. The radar antennas are pointed at the loudspeaker cone from a distance of 4.87m. In addition, a stationary target is present at 5.5m, again the fixed antenna positioning system in the used anechoic chamber. This stationary target is collinear with the loudspeaker and the radar antennas.

The advantage of using a loudspeaker as a moving radar target is its reproducibility and stability. The main disadvantage is that the loudspeaker cone is not a radar target with a fixed radial velocity. It will give rise to a converted signal frequency spectrum that is different from the one that was calculated in subsection 2.4.3 for a moving target with constant radial velocity. The measured converted signal spectrum, however, will be interpreted. Another disadvantage is that the loudspeaker is a very small radar target that will produce only a weak echo signal that may be masked out by the larger stationary target near it. This, however, will not pose a major problem as will be seen further on.
The measured IF output power spectrum for the frequency band from DC to 200kHz is shown in Fig. 4.24. Again, a maximum of the spectral envelope can be observed around 20kHz (10kHz -59.4dBm, 20kHz -45.27dBm, 30kHz -47.15dBm), originating from the directly leaked signal. Another spectral envelope maximum can be recognised around 90kHz. More detailed measurements around the frequencies 70kHz, 80kHz and 90kHz are respectively shown in Figs. 4.25, 4.26 and 4.27. In subsection 2.4.3 it was calculated that a target moving with a constant radial velocity gives rise to spectral components at the frequencies \(|k \cdot f_m \pm f_d|\), with \(k\) an integer, \(f_m\) the modulation frequency of 10kHz and \(f_d\) the Doppler frequency shift caused by the target’s radial velocity. In Figs. 4.25, 4.26 and 4.27 spectral components at multiples of 10Hz are recognised. It is no surprise that the measured power spectrum differs from the case of a target with fixed radial velocity, as the measurement of the power spectrum happens on a fragment longer than 0.05s, and in this case, a loudspeaker cone vibrating at 10Hz is not a target with a fixed radial velocity but a target that is constantly accelerating. To explain the acquired power spectrum, instead of doing the entire analysis conducted in subsection 2.4.3 again for this specific case, it will be reasoned how the power spectrum for this case should look like. The calculation of the frequency spectrum of the converted signal starts with equation 2.20. As the loudspeaker cone moves towards the radar system and away from it at a frequency of 10Hz, the delay \(\tau\) is periodic with a frequency 10Hz in this equation. As a consequence, the entire argument of the cosine is periodic with a frequency 10Hz, and the frequency spectrum will consist of harmonics of 10Hz. The measured power spectrum indeed consists of frequency components spaced by 10Hz, as can be observed in Figs. 4.25, 4.26 and 4.27 by counting the numbers of peaks. If it is assumed that during each half period of 10Hz, the radial velocity \(v_{rad}\) of the cone is almost constant, the loudspeaker cone constantly switches between being a target with radial velocity \(v_{rad}\) and being a target with radial velocity \(-v_{rad}\). At the same time, the loudspeaker cone stays at approximately the same range from the radar system. Consequently, just like in the case of a target with fixed radial velocity, the converted signal instantaneous frequency takes on constant values during the major parts of the FM periods given by the equations 2.38 and 2.39. It is therefore expected that the spectral components (now harmonics of 10Hz) around these two frequencies will have maximal power.

Now it is clear that the vibrating speaker cone at 4.87m will likely not be masked out by the stationary target at 5.5m. The stationary target will give rise to frequency components in the converted signal power spectrum at multiples of the modulation frequency of 10kHz. The power of the spectral components of the converted signal caused by the loudspeaker cone at multiples of the modulation frequency, on the contrary, is expected to be low, as the power is contained in the frequency components around \(|k \cdot f_m \pm f_d|\) instead. In this way, both radar targets do not interfere. To detect both targets at the same time, a separate spectral envelope for the harmonics of 10kHz and the other 10Hz (excluding harmonics of 10kHz) should be considered. The spectral envelope of the frequency components at the exact multiples of 10kHz will detect the stationary target, while the other spectral envelope will detect the moving target. As can
be verified with the annotations of Figs. 4.25, 4.26 and 4.27 the spectral envelope of the fixed target has a maximum between 90kHz and 100kHz. The component of 90kHz corresponds to a range \( R_{90kHz} \) given by:

\[
R_{90kHz} = 9 \cdot \frac{c}{4 \cdot 100 MHz} - 1.30 m = 5.45m.
\] (4.12)

As the spectral envelope’s maximum is a bit higher than 90kHz, the stationary target range is also a bit greater than 5.45m. In the same figures, it can be seen that the spectral envelope of the moving target has a maximum that lies just above 80kHz. This frequency component corresponds to a range \( R_{80kHz} \) given by:

\[
R_{80kHz} = 8 \cdot \frac{c}{4 \cdot 100 MHz} - 1.30 m = 4.70m.
\] (4.13)

The range of the moving target is therefore a bit greater than 4.70m. The measurements of the ranges of both the fixed and loudspeaker target are in good correspondence with the actual ranges.

Despite the fact that the loudspeakers do not appear to the radar system as targets with a fixed velocity, it has been verified that the radar system can detect moving targets with a radar cross section comparable to humans in the presence of nearby stationary targets.

Again, it has to be remarked that this measurement is part of a different measurement series, and Figs. 4.16 and 4.17 are of no relevance here.

![Power spectrum IF output from DC to 200kHz (RBW 1Hz, reference level 0dBm)](image.png)

Fig. 4.24: Power spectrum IF output from DC to 200kHz (RBW 1Hz, reference level 0dBm)
4.2 Validation of the radar prototype

Fig. 4.25: Power spectrum IF output from 69.5kHz to 70.5kHz (RBW 1Hz, reference level 0dBm)

Fig. 4.26: Power spectrum IF output from 79.5kHz to 80.5kHz (RBW 1Hz, reference level 0dBm)
4.2 Validation of the radar prototype

4.2.3 Radar system measurements conclusion

As stated in this master’s thesis goal, humans are the expected radar targets, so the measurements involving radar targets with a human-like radar cross section are the most important. It was shown that the radar system can successfully detect a moving human target at a close range, and measure its range and radial speed. Due to their relatively small radar cross section, however, humans are easily masked out by metallic objects that have the same radial speed. In the measurements it was shown that a stationary human-like target is masked out by a stationary metallic object. When the human-like target was moving, however, and the metallic object was stationary (loudspeaker measurement), the two targets could be resolved. This was shown in the measurement setup with the loudspeaker.

Fig. 4.27: Power spectrum IF output from 89.5kHz to 90.5kHz (RBW 1Hz, reference level 0dBm)
Chapter 5

Conclusion

In this final chapter, suggestions for future work around the subject, and a global conclusion are formulated.

5.1 Future work

In this section, a few suggestions for an improved design are made.

In the field of antenna design, there is certainly room for improvement. The transmit antenna could be realised as a phased antenna array. This would offer several benefits. An antenna array would have increased gain compared to the single element antenna used in this thesis. In this way, the power generated by the radar system can be lower, while the EIRP remains the same. This will reduce the power consumption and heat dissipation of the system. With a portable application and a certain battery lifetime in mind, the reduced power consumption will make a physically smaller battery possible, increasing the compactness and textile integrability. Another advantage the increased transmit antenna gain will offer, is increased isolation between the transmit and receive antenna, making the radar system more sensitive. The same is true for receive antennas with a higher gain. Higher gain antennas in general will also enhance the angular resolution of the radar system. A phased antenna array also has a main antenna beam that is electronically steerable. In this way, the radar system itself would be able to sweep the environment looking for targets, and the wearer of the radar system would not have to do this.

In our setup, it was carefully checked that the isolation between the transmit and receive antenna is adequate. However, if these two antennas are to be integrated into garments, the antenna substrates will be bent. It has not been verified that the isolation between the transmit and receive antenna, or the antenna characteristics itself are adequately robust against this situation.

In retrospect, it would have been better to omit the second power amplifier in the RX chain. This amplifier consumes a lot of current (105mA at 5V), which has an adverse effect on the compactness. The only advantage this extra component offers is a slightly reduced
receiver chain noise figure (0.552dB instead of 0.644dB), and a higher gain (14.75dB extra). However, this extra component poses the risk of instability, as explained in subsection 3.1.4 and overloading of the frequency mixer by the directly leaked signal. It is more advantageous to do the amplification of the received signal after the frequency mixer, making it easier to suppress the directly leaked signal component. In this way, the risk of overloading the frequency mixer would be eliminated.

Another opportunity to drastically decrease the consumed power and dissipated heat is by choosing a passive frequency mixer instead of the active mixer.

In this prototype board, a fixed gain transmit amplifier was used. Therefore, the actual transmitted power was not exactly 20dBm EIRP. Newer versions of the radar board might use a VGA, enabling the system to correct the transmitted power to 20dBm EIRP. Using a transmitter amplifier with variable gain would also decouple the radar system board from the used antennas. Different antennas with different gains could be used while keeping the transmitted power equal to 20dBm EIRP.

It was also noticed that the swept frequency band easily drifts with temperature influences. For a real product this would of course be unacceptable. To solve this, a control loop could be implemented to keep the frequency sweep within the correct band.

A few notes can also be made on the used FM generator. In this thesis, it was chosen to use the simplest topology, consisting of a VCO controlled in open loop by a triangle voltage oscillator realised with an op-amp circuit. This topology was chosen for its low complexity. It was expected that the linearity would have been good enough so that the range resolution would be limited by the available bandwidth of 100MHz rather than by the frequency sweep linearity. Measurements indicated that the non-linearity might dominate the range resolution starting from a range of only 7.26m. There are other open loop topologies that are expected to be more linear, when, for example, using a DDS generator. Many DDS chips can be programmed to perform a certain frequency sweep. The frequency sweep generated by the DDS will need to be up-converted to the correct frequency band, however, requiring additional hardware, and making the system less compact. Additional hardware is also needed to control the DDS.

5.2 Conclusion

The goal of this master’s thesis formulated in the introduction, was to build a compact radar system operating in the 2.4GHz ISM band, capable of detecting the range and radial speed of human targets. Moreover, as a step towards full textile integration, the use of textile antennas was required. It can be concluded that the goal stated in the introduction has been achieved.

First, an exploration of the different types of radar system was conducted, in order to find the most suitable radar system to reach the formulated goal. The choice fell on the FMCW type of radar systems. This type of radar system is ideal for low power and short range applications and leverages a very compact construction. Furthermore, both target range and radial velocity can
be measured. A detailed analysis of FMCW radar systems followed. An analysis of the power spectrum of the converted signal for different FM waveforms was performed. This analysis resulted in a method to detect radar targets and to extract their range and radial velocity information from this power spectrum. The most appropriate FM waveform was selected, the symmetrical linear triangle waveform, which leads to attractive properties of the converted signal spectrum such as a low spectral width, well-defined maxima and easy measurement of the target radial speed and range at the same time. A well thought-through choice of the FM function was made. The most important performance limitations of linear FMCW radar systems, the directly leaked signal and the modulation linearity are also discussed. As the FM linearity is very dependent on the used hardware, different hardware implementations of the FM generator and their linearity were considered. The most simple FM generator architecture was selected, consisting of a VCO controlled in open loop by a triangle voltage oscillator, because it was expected that the chirp non-linearity would not limit the range resolution.

Subsequently, the attention was moved towards realising this FMCW radar system in electronic hardware. Component specifications were formulated and components were selected. All chosen components were compact surface-mounted devices. Next, a compact PCB layout was designed. EM circuit co-simulation of the PCB layout and component models are performed with the Agilent ADS software to check if the compactness does not lead to excessive amounts of crosstalk on the PCB. The last piece of hardware that needed to be designed were the antennas. Circularly polarised patch antennas with a rectangular ring topology were selected. Textile materials were chosen to fabricate these patch antennas: black foam provided by Javaux as a substrate and Electron as patch and ground plane material.

Finally, a prototype board of the radar system was fabricated. First, simple tests were performed on the prototype board, to verify the correct operation of every subsystem of the board. Subsequently, the functionality of the radar board in its entirety was tested. First the radar board transmitter was directly connected to the radar board receiver with a coax cable with extra attenuators. The radar system was capable of measuring the length of the coax cable. Next tests with the designed transmit and receive antennas connected to the radar system were conducted in the anechoic chamber. The first attempt to measure the range of a stationary radar target with a radar cross section comparable to that of a human failed because the human-like target was masked out by nearby metallic parts of an antenna positioning system that was not covered well with absorbers. In a second attempt, a real human walked towards the radar system, and a measurement was made when the human was halfway between the interfering antenna positioning system and the radar system antennas. This attempt was successful. The acquired converted signal power spectrum was correctly interpreted, and a measurement of the human range and speed, as well as the range of the stationary antenna positioning system was made at the same time. In a final measurement, we returned to the situation of the first attempt that failed. The only difference is that now the human-like target that is nearby the interfering antenna positioning is moving. As a moving target, a loudspeaker with copper
stuck to the speaker cone was used. The area of the copper was chosen so that its radar cross section is comparable to that of a man. The moving speaker cone and the stationary target were successfully detected at the same time, and both of their ranges to the radar system were successfully measured.
Bibliography


[34] Coaxial SMA Fixed Attenuator VAT-5+, Minicircuits, 2013.
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