High Frequency Laboratory
Experiment 4

Microwave FMCW Radar

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# 1 Theory

In this experiment, you will be familiarized with a 24 GHz FMCW Radar. During this experiment the performance of this type of radar, regarding its detection ability, range resolution, signal-to-noise ratio (SNR) and etc., will be investigated. For a detailed understanding of the FMCW Radar principle and to carry out the experiment successfully, it is essential to understand the underlying theory. Knowledge of the components such as voltage controlled oscillators (VCOs), power dividers, mixers and low noise amplifiers (LNAs) is also required. Therefore it is highly recommended that each student participating in the laboratory read the theoretical section of this text. Furthermore, as part of the overall course-score-evaluation, it is mandatory to solve the tasks at the end of this document before coming for the laboratory session.

## 1.1 Radar Introduction

The word RADAR was introduced in 1941 as an acronym for Radio Detection And Ranging. It was originally designed and established for military purposes in the first decades after its invention. Nevertheless, due to Radar’s high performance in object detection it gained more and more importance in civil applications such as in the automobile industry and geosciences in recent years.

Generally, a radar system consists of the following components: transmitter, receiver, data processing unit and displaying unit. The radar produces electromagnetic waves which are radiated by an antenna and propagate into free space. Targets in the path of the wave propagation scatter the electromagnetic energy in different directions. Some of this energy is then scattered back to the Radar, having experienced a slight change in wavelength if the target was moving (Doppler effect). The receiving antenna gathers the back-scattered radiation and feeds it to the receiver. The strength of the received signal can be increased through special antenna designs. The signals returned to the receiver are normally very weak and thus has to be amplified by electronic circuits. These signals are then processed and displayed.

Nowadays, different types of radar are developed to fulfill the requirements of various applications:

- Doppler radar
- Continuous-wave (CW) radar
- Frequency Modulated Continuous Wave (FMCW) radar
- Synthetic aperture radar (SAR).
In this experiment the FMCW Radar is discussed. Thanks to the frequency modulation scheme, the FMCW Radar achieves a number of significant advantages over traditional Radar systems. First, the basics of radar such as the Radar equation and the Doppler effect are introduced.

### 1.1.1 Radar Equation

Generally in radar applications the power of received signal can be calculated theoretically through the Radar equation. In most cases, this equation is derived for point targets. Point targets are objects whose dimensions $D$ are small compared to the illumination width $R \cdot \theta_{\text{BW}}$ (Range*Half-power beamwidth) of the radar at the target site as shown in Fig. 1.1.

![Fig. 1.1: Illustration of the principle of radar](image)

In a monostatic system for example, whose target is in the distance $R$. The power reflected back at the receiver is determined in Eq.(1.1) as,

$$P_R = \frac{P_T G_T G_R \lambda^2}{(4\pi)^3 R^4} \sigma$$

where $P_T$ is the transmitted power and the parameters $G_T$ and $G_R$ are the gain of transmitting and receiving antennas respectively. In a monostatic case the same antennas are used for transmitter and receiver and $G_T = G_R$. The measure of reflected power in direction of the radar is the Radar Cross-Section (RCS) $\sigma/m^2$. The RCS of a target is the fictional area and describes the amount and/or strength of this object scattering of the incident EM wave. The RCS is strongly frequency dependent and is influenced also by the size, shape, surface roughness and material of target [4].

The common way to define the Radar cross-section is with use of the Radar equation. For a homogenous, lossless, isotropic medium and a universally bistatic configuration according to Fig. 1.2, the Radar cross-section is defined by [4]:

$$\sigma = \frac{4\pi}{\lambda_0^2} \left( \frac{P_R L_R}{G_R} 4\pi R_T^2 \right) \left( \frac{L_T}{G_T P_T} 4\pi R_T^2 \right) L_P$$

(1.2)
1.1 Radar Introduction

Where $L$ is losses at the transmitter and receiver, $L_p$ is losses due to polarization effect, $E$ is radiated electric field strength, $L_T$ and $L_R$ are losses at the transmitter and receiver, $\theta$ and $\phi$ are azimuth and elevation angle of the incident wave front.

\[ \sigma = \lim_{R \to \infty} 4\pi R^2 \frac{\vec{E}_s \cdot \vec{E}_s^*}{\vec{E}_i \cdot \vec{E}_i^*} = \lim_{R \to \infty} 4\pi R^2 \frac{S_r}{S_i} \]  

(1.3)

Where $s=$scattered and $i=$incident.

Usually the cross-section is a hypothetical area with unit of m$^2$ and is stated in dB as

\[ \sigma / \text{dBm}^2 = 10 \log \left( \frac{\sigma}{1 \text{m}^2} \right) \]  

(1.4)

The given definition applies for the general bistatic case. In practice the positions of the receiving and transmitting antennas are often identical ($\theta^i = \theta^s$ and $\phi^i = \phi^s$). $\sigma$ can then be referred to as Radar Backscattering Cross-Section and the Radar geometry being monostatic. If the positions of the receiving and trasmitting antennas are not the same, but close to each other, one speaks of the quasi-monostatic Radar where the antennas are close to one another ($\theta^i \approx \theta^s$ and $\phi^i \approx \phi^s$).

In Table 1.1 the expressions for the RCS of typical radar reflectors are shown.
### Theory

#### Targets Dimensions

<table>
<thead>
<tr>
<th>Targets</th>
<th>Dimensions</th>
<th>RCS $\sigma$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cylinder</td>
<td>$\pi a^2$</td>
<td>$\frac{2\pi ab^2}{\lambda}$</td>
</tr>
<tr>
<td>Sphere</td>
<td>$\pi a^2$</td>
<td>$\frac{8\pi a^2b^2}{\lambda^2}$</td>
</tr>
<tr>
<td>Diplane</td>
<td>$\frac{4\pi a^4}{3\lambda^2}$</td>
<td></td>
</tr>
<tr>
<td>Triangular Trihedral</td>
<td>$\frac{4\pi a^4}{3\lambda^2}$</td>
<td></td>
</tr>
</tbody>
</table>

Table 1.1: RCS of different radar reflectors

#### 1.1.2 Doppler Effect

In the case of continuous observation of the target which moves relatively to the radar, the received signal contains a phase shift, known as the Doppler effect. This feature is utilized in some types of radar, to measure the velocity of detected objects, since the Doppler frequency is related to the relative velocity in the direction of radiation between the radar and the detected object, as shown in Fig. 1.3. When the radar transmitter and receiver are located very closely, the Doppler frequency $f_d$ can be written as[4]:

$$f_D = \frac{1}{2\pi} \cdot \frac{d\phi}{dt} = 2 \cdot \frac{1}{2\pi} \frac{2\pi}{\lambda} \cdot \frac{dR}{dt} = \frac{2v_t f_T}{c_0}$$  \hspace{1cm} (1.5)

where $v_t$ is the radial velocity of the target to the radar, $f_T$ is the transmit frequency and $c_0$ is propagation velocity of light.

![Fig. 1.3: Radial velocity of the target](image)
1.2 Basic theory of FMCW radar

Concerning the Doppler frequency, the relative velocity component in the direction of wave propagation between the target and the radar needs to be considered. The classic zero Doppler target is the one which moves tangentially to the direction of the antenna radiation. The target provides positive Doppler frequency, if it approaches the radar. Otherwise a negative Doppler frequency is determined for the situation where the target moves away from the radar [3]. Thereby, for instance, when the radial velocity is 30 m/s (approaching), the Doppler frequency at 24 GHz (K-band) is about +4.8 kHz.

1.2 Basic theory of FMCW radar

The idea of using FM signals for ranging to a reflecting object is very old. Such signals were used as long ago as the 1920s for ionospheric research. The practical application of FMCW radar started in 1928, when J.O. Bentley filed the American patent on an "airplane altitude indicating system". Industrial applications of this idea began only at the end of the 1930s, when the ultrahigh-frequency band was exploited. Nowadays the FMCW radar has been developed and applied on a large scale in military and civil industry [2].

1.2.1 Advantages

A major limitation of the CW radar is that it cannot determine target range because it lacks the timing mark necessary to allow the system to time accurately the transmit and receive cycle and to convert this into range. In pulse radar, this mark is provided by the pulse itself. Pulse radars measure object range by transmitting a short pulse of energy at the target and then waiting for the return signals of pulse echo. The time interval between pulse echoes then provides some information of the object range, since the echoes from close targets return sooner than that from more distant targets. However, short pulses require large peak power (typically ranging from kW to MW) to produce an average signal strength that is able to give a clear return over large distance. Additionally, the Doppler frequency is not detectable in this case. Thus FM modulation provides the necessary marks to allow both range and velocity information to be detected. Such radar is called the FMCW radar.

An FMCW radar sends out a continuous transmission wave with increasing frequency. With this technology, the emitted power of FMCW radar is possible to be largely reduced compared to that of the pulse radar.

The basic features of an FMCW radar are [2]:

- Ability to measure small ranges to the target (minimal measured range being comparable to the transmitted wavelength);
- Ability to measure simultaneously the target range and its relative velocity;
- Very high accuracy of range measurement;
- Signal processing after mixing is performed at a low frequency range, considerably
simplifying the realization of the processing circuits;
• Safety from the absence of the pulse radiation with a high peak power.

1.2.2 Fundamentals of FMCW radar

In a CW radar, separation of transmitted and reflected signals in time is impossible, because continuous radiation is used. Hence, the reception of information about the range to a target is possible only when the transmission wave is modulated in amplitude or phase. Amplitude modulation is not used because it is practically impossible to select the reflected signal against the interfering background of the transmission. Thus, the reflected signal delay relative to the transmission is on the basis of the phase difference of these angularly modulated signals [2]. This operation can be easily carried out by multiplication of transmitted and reflected signals. After multiplication, two signals are formed: one with a phase equal to the difference of the phases of the multiplied signals with a frequency on the baseband, and the other with a phase equal to the sum of these phases. The latter signal is easily filtered out, as its frequency is twice of that of the radiated signal.

Several different modulations of FMCW signals are used (see Fig.1.4): Sawtooth, Triangle and Sinusoidal.

Shown in the first row are the transmitted signals (solid lines) and the received signal (dashed lines). Shown in the second row are the frequency differences between both the transmitted and received signals, called the intermediate frequency (IF). This IF signal depicts two important parameters: Beat frequency and Doppler frequency. Due to the fact that it is possible to measure both range and velocity of target using the triangle waveform, it has become more and more popular in applications [4]. Furthermore, the triangle waveform is easier to generate. Thus, we will use this waveform throughout the experiment.

The real form of a transmitted FM signal with triangle waveform can be described as

$$s_t(t) = \begin{cases} A_t \cos \left(2\pi f_{\text{min}} \hat{t} + \pi \gamma \hat{t}^2\right) & 2nT_{\text{sweep}} t < (2n + 1) \cdot T_{\text{sweep}} \\ A_t \cos \left(2\pi f_{\text{max}} \hat{t} - \pi \gamma \hat{t}^2\right) & (2n + 1) T_{\text{sweep}} t < 2(n + 1) T_{\text{sweep}} \end{cases}$$

(1.6)

where $A_t$ is the amplitude function, which is usually a constant, $\gamma$ is the chirp rate, which is equal to $\frac{B}{T_{\text{sweep}}}$, $T_{\text{sweep}}$ is the sweep time of up- or down-sweep, $f_{\text{min}}$ and $f_{\text{max}}$ are the minimum and maximum frequency of the transmitted FM signal respectively, and $\hat{t}$ is the remainder of $t$ divided by $T_{\text{sweep}}$, i.e.

$$\hat{t} = t \mod T_{\text{sweep}}$$

(1.7)

An example of the wave form of FMCW signal in time domain is shown in Fig. 1.5.

The signal received from a single point target at distance $r$ with radial velocity $v_r$ is

$$s_r(t) = \begin{cases} A_r \cos \left[2\pi f_{\text{min}} (\hat{t} - \tau) + \pi \gamma (\hat{t} - \tau)^2\right] & 2nT_{\text{sweep}} t < (2n + 1) T_{\text{sweep}} \\ A_r \cos \left[2\pi f_{\text{max}} (\hat{t} - \tau) - \pi \gamma (\hat{t} - \tau)^2\right] & (2n + 1) T_{\text{sweep}} t < 2(n + 1) T_{\text{sweep}} \end{cases}$$

(1.8)
1.2 Basic theory of FMCW radar

Since the received sweep is a delayed version of current transmitted sweep, it will stretch to the following transmitted sweep time, as the upper row of Fig. 1.4 indicates. However the expression of each received sweep remain constant during each transmitted sweep time \((2nT_{\text{sweep}} < (2n + 1) T_{\text{sweep}} \) or \((2n + 1) T_{\text{sweep}} < 2 (n + 1) T_{\text{sweep}}\)) as long as the delay \(\tau\) is much smaller than the modulation sweep time \(T_{\text{sweep}}\), which is also a consideration when designing radar parameters according to the application.

By mixing the received signal with a replica of transmitted signal and filtering out the higher frequency components, the IF signal is obtained, which expression can be written as

\[
 s_{\text{IF}} (t) = \begin{cases} 
 \frac{1}{2} A_t A_r \cos (2\pi \gamma \tau \hat{t} + 2\pi f_{\text{min}} \tau - \pi \gamma \tau^2) & 2nT_{\text{sweep}} < (2n + 1) T_{\text{sweep}} \\
 \frac{1}{2} A_t A_r \cos (2\pi \gamma \tau \hat{t} - 2\pi f_{\text{max}} \tau + \pi \gamma \tau^2) & (2n + 1) T_{\text{sweep}} < 2 (n + 1) T_{\text{sweep}} 
\end{cases}
\]

Since the delay \(\tau\) is small, thus \(2\pi f \tau \gg \pi \gamma \tau^2\). Then the IF signal becomes

\[
 s_{\text{IF}} (t) = \begin{cases} 
 \frac{1}{2} A_t A_r \cos (2\pi \gamma \tau \hat{t} + 2\pi f_{\text{min}} \tau) & 2nT_{\text{sweep}} < (2n + 1) T_{\text{sweep}} \\
 \frac{1}{2} A_t A_r \cos (2\pi \gamma \tau \hat{t} - 2\pi f_{\text{max}} \tau) & (2n + 1) T_{\text{sweep}} < 2 (n + 1) T_{\text{sweep}} 
\end{cases}
\]

To consider the Doppler effect, for a certain up- or down-sweep the delay \(\tau\) is expressed as a function of time \(t\) as

\[
 \tau = \frac{2 (r_0 - v_r t)}{c + v_r}
\]
where $r_0$ is the initial distance between the radar and the target. It changes between adjacent up- and down-sweeps, however it can be approximated to be the same value for a pair of adjacent up- and down-sweeps when the $v_r$ is much smaller than $c$. In this situation the expression of the delay $\tau$ can be written as

$$\tau = \frac{2 (r - v_r t)}{c}$$  \hspace{1cm} (1.12)

Note that to keep in consistence with most of the radar books, here $v_r$ is defined as being positive while approaching the radar. Therefore the instantaneous frequency of $s_{1F}(t)$ of a certain up- and down-sweep can be calculated as

\[
\begin{align*}
  f_1 &= (\gamma \tau t + f_{\text{min}} \tau)' = \gamma \cdot \frac{2v_r}{c} - \frac{2v_r}{c} \left(2\gamma t + f_{\text{min}}\right) & 0t < T_{\text{sweep}} \\
  f_2 &= (\gamma \tau \hat{t} - f_{\text{max}} \tau)' \hat{t} = \gamma \cdot \frac{2v_r}{c} - \frac{2v_r}{c} \left(2\gamma \hat{t} - f_{\text{max}}\right) & T_{\text{sweep}} t < 2T_{\text{sweep}}
\end{align*}
\]  \hspace{1cm} (1.13)

where $\hat{t}$ is defined the same as in (1.7); $\gamma \cdot \frac{2v_r}{c}$ is called beat frequency $f_B$, which includes the target’s distance information; $\frac{2v_r}{c} \left(2\gamma t + f_{\text{min}}\right)$ and $\frac{2v_r}{c} \left(2\gamma \hat{t} - f_{\text{max}}\right)$ are called Doppler frequency $f_D$, which include the velocity information. Although it is apparent that the Doppler frequency components are not constant during one sweep time, both $\frac{2v_r}{c} \left(2\gamma t + f_{\text{min}}\right)$ and $\frac{2v_r}{c} \left(2\gamma \hat{t} - f_{\text{max}}\right)$ can be approximated as

$$f_D = \frac{2v_r}{c} f_c$$  \hspace{1cm} (1.14)

for a narrow-band FMCW radar system with transmitted center frequency $f_c$. Then the IF frequency of a pair of adjacent up- and down-sweeps can be written as

\[
\begin{align*}
  f_1 &= \gamma \cdot \frac{2v_r}{c} - \frac{2v_r}{c} f_c & 0t < T_{\text{sweep}} \\
  f_2 &= \gamma \cdot \frac{2v_r}{c} + \frac{2v_r}{c} f_c & T_{\text{sweep}} t < 2T_{\text{sweep}}
\end{align*}
\]  \hspace{1cm} (1.15)

Note that $f_D$ is positive when the target approaches the radar and negative when the target moves away from the radar.
1.2 Basic theory of FMCW radar

The block diagram in Fig. 1.6 shows the structure of an FMCW radar, consisting of a transmitter, a receiver and a signal processing unit.

In the receiver, the received signal is mixed with the LO reference. The output IF signal is then sampled by an analog/digital converter and processed further to obtain the range and velocity information.

![Block diagram of the FMCW radar](image)

Fig. 1.6: Structure of the FMCW radar

1.2.3 Components of FMCW radar

In the following sections, the components of a FMCW radar and their corresponding important parameters are discussed.

**Voltage-Controlled Oscillator (VCO)**

A key component of the FMCW Radar is the VCO, acting as the source of signal. The VCO is an electronic oscillator designed to be controlled in oscillation frequency by a voltage input. Also modulating signals may be fed into the VCO to realize frequency modulation (FM) or phase modulation (PM). Theoretically the output frequency can be written as

\[
f_\circ(t) = K_0 v_{in}(t)
\]

where \(v_{in}(t)\) is the time signal of the input control voltage of the VCO. Therefore, the frequency modulation can be performed by using modulated \(v_{in}(t)\).

To achieve a FMCW radar with high performance, a very linear frequency sweep is needed. The main problem of VCO is that it uses an exponential converter, which is extremely temperature sensitive. It consequently exhibits a drifting oscillation frequency when the operation temperature changes. Since the triangle-waveform is selected in the experiment, the frequency-voltage curve of VCO should be characterized for the local operation conditions by the method called "predistortion".

Since the frequency signal generated by the VCO is too high to be sampled by a sampling device, hence to obtain the VCO characteristics, a frequency divider should be applied. By
applying a frequency divider before sampling device, the original high frequency component (GHz) is converted into a lower frequency component (GHz). In this way, the characteristic tuning curve of VCO can be configured ("predistortion") very precisely to assure the accuracy of the linearization of the output frequency of the VCO over the time.

In Fig. 1.7, the detailed process of a curve adjustment is illustrated. First, a series of output frequencies of the VCO corresponding to uniform distribution of input voltage is measured and recorded. The VCO characteristic is nonlinear (left figure). Using interpolation, the frequency-voltage characteristic curve can be generated, for which the frequency steps are equal (right figure).

**D/A converter**

The input voltage of VCO $v_{\text{in}}(t)$ is generated by a D/A converter. The primary parameters of a D/A converter are:

- output voltage range, which should correspond to the VCO’s input voltage range;
- resolution $b_{\text{DA}}$ (bits);
- update rate $f_{s,\text{DA}}$.

The impacts of $b_{\text{DA}}$ and $f_{s,\text{DA}}$ are similar to the impact of the nonlinear tuning curve of VCO, however it cannot be compensated by the "predistortion" method described before. Since it is impossible to utilize a D/A converter with infinite $b_{\text{DA}}$ and $f_{s,\text{DA}}$, in which case the impact of them is zero, a low-pass filter is needed after the output of D/A converter, which will filter out the high frequency components of the signal.

**Power Divider**

The frequency modulated signal output from the VCO is divided by the power divider into two channels: the transmitted signal and the LO reference. A low insertion loss of the power divider is needed. The insertion loss is the attenuation of signal power resulted from the insertion of a device in a transmission line. It is usually expressed in dB as a ratio of
output signal relative to the transmitted signal power. The ideal insertion loss of a two-way power divider is 3 dB. For higher frequency range a large insertion loss must be considered due to large loss of signal power in the transmission line and SMA connector. Furthermore, the bandwidth of the power divider should be as large as bandwidth of the FM signals.

Antennas

The antennas act as the transitional structure between a guiding device and free-space. In the bistatic radar two antennas are applied respectively for transmitting and receiving. When the radiated electromagnetic waves of transmitting antenna illuminate on targets, the waves can be reflected back and received by the receiving antenna.

The gain of antenna relates the intensity of an antenna in a given direction to the intensity that would be produced by a hypothetical ideal antenna (isotropic radiator) that radiates equally in all directions without losing energy. Although the gain of an antenna is directly related to its directivity, it is a measure that takes into account the efficiency of the antenna, which might be lowered through heat production or reflection of energy back towards the transmitter, as well as its directional capabilities influenced by the antenna pattern. The gain is unitless and usually expressed in logarithmic scale. Thus, antenna gain is denoted with dBi, where the index i refers to isotropic radiator [1].

For radar applications, antennas with very directive characteristics are needed. Increasing the antenna gain can be accomplished by increasing the electrical size of the antenna, since the antenna beamwidth is inverse proportional to the aperture of the antenna. To achieve this, some antennas can be aligned together as an antenna array. The radiation pattern of an antenna array is determined by the radiation pattern of single element and the array factor (for more details see the lecture of 'Antennen und Antennensysteme').

Low Noise Amplifier

Usually, the RF signals captured by the receiving antenna are very weak. Thus an amplifier is needed. The Low Noise Amplifier (LNA), placed at the front-end of the radio receiver circuit, is a special type of electronic amplifier to amplify very weak signals. Compared with common amplification devices, the LNA technology is based on devices with a sufficiently high gain at the targeted operating frequency with a low noise figure (NF). The details of the NF will be explained in Sec. 1.3.4.

Furthermore, the operating frequency and bandwidth of the LNA should be carefully considered for different applications. The gain at the desired frequency indicates the performance of this device. Usually, the gain of the LNA, written as $G_{LNA}$, is presented in terms of dB.
Mixer

The signal amplified by the LNA is mixed with the LO reference signal. The multiplication operation of transmitted and reflected signals can be achieved in this device. A modulated low frequency sinusoidal signal \( f_{IF} = |f_{RF} - f_{LO}| \), the main frequency of which is equal to the frequency difference between the two mixed signals, can be obtained from the output of the mixer. The high frequency component \( f_{IF} = |f_{RF} + f_{LO}| \) is filtered by the Low Pass (LP) Filter at the IF output of the mixer. Conversion loss is inevitable and should be taken into consideration because it indicates the SNR performance of the device.

Data Acquisition

The IF signal output from the mixer is sampled by A/D converter for processing later by a computer. The critical parameters of the A/D converter are its resolution \( b \) (bits) and sampling frequency \( f_s \) (Hz). During A/D conversion, the continuous analog signal is quantified and the number of quantified levels determines the resolution \( b \). For example, with the resolution of 3 bits, we can expect \( 2^3 = 8 \) quantization levels. On the other hand, the resolution \( b \) also determines the quantization noise which may influence the SNR of the whole system. According to the sampling theory, the sampling frequency \( f_s \) should be chosen to be at least twice higher than the bandwidth of signal \( f_{IF} \) to avoid frequency aliasing. Since \( f_{IF} \) is dependent on both the maximal range of radar detection and the maximal velocity of moving targets, \( f_s \) should be adjusted for specified applications of the FMCW radar (short range detection or long range detection).

1.3 Characters of FMCW Radar

1.3.1 Frequency Modulation

With the triangle waveform which is selected for this experiment, the signal frequency is ramped with a constant rate of change. As the signal propagates to the target and is reflected back to the radar, the waveform of received signal will be delayed by a certain time interval. Regarding the IF-signal, this time amount is proportional to a frequency variable \( f_B \), called the beat frequency. Thereby, a more distant target returns a larger beat frequency. In the case
of a moving target, the Doppler frequency may change the IF signal’s frequency additively or subtractively depending on the direction of target’s movement. In Fig. 1.9, an example of an approaching target with a positive Doppler frequency is shown.

![Triangle modulation with Doppler effect](image)

**Fig. 1.9:** Triangle modulation with Doppler effect

Sweep Time $T_{\text{sweep}}$ is defined as the period of frequency modulated signal. This value corresponds to the maximal detected distance, since the received signal may overlap the transmitted signal of the next period, which leads to confusion. Also, the $T_{\text{sweep}}$ determines the radar system’s velocity resolution $\Delta v_r$. Bandwidth $B$ is the frequency range of transmitted signal, which decides the radar system’s range resolution $\Delta r$.

### 1.3.2 IF Processing

Mathematically, both range and velocity information, which are represented by the beat frequency $f_B$ and Doppler frequency $f_D$ respectively, can simply be derived after the determination of $f_1$ and $f_2$ from IF signal in Fig.1.9 as

$$
 f_B = \frac{1}{2} \cdot (f_2 + f_1) \tag{1.17}
$$

$$
 f_D = \frac{1}{2} \cdot (f_2 - f_1) \tag{1.18}
$$

According to geometry, without considering velocity, the beat frequency $f_B$ is proportional to the delay $\tau$ between transmitted signal and received signal where

$$
 f_B = [f_{\text{min}} + \gamma \cdot (t - \tau)] - (f_{\text{min}} + \gamma \cdot t) = \gamma \cdot \tau \tag{1.19}
$$

Thus the target’s range $r$ can be derived as

$$
 r = \frac{c}{2\gamma} \cdot f_B \tag{1.20}
$$
Regarding the target’s velocity, according to (1.14), it only relies on $f_D$ as:

$$v_r = \frac{c}{2f_c} \cdot f_D$$  \hspace{1cm} (1.21)

### 1.3.3 Resolution

#### Range resolution

The range resolution $\Delta r$ indicates the ability of radar to distinguish two close targets. Since $r$ is deduced from beat frequency $f_B$, $\Delta r$ can be deduced from $\Delta f_B$ as well.

Since the length of each received IF signal is less than $T_{\text{sweep}}$, its frequency resolution is usually defined as the 4 dB bandwidth of a monochromatic sinusoid with a rectangle window of length $T_{\text{sweep}}$:

$$\Delta f_B = \frac{1}{T_{\text{sweep}}}$$  \hspace{1cm} (1.22)

Thus $\Delta r$ can be derived consequently as

$$\Delta r = \frac{c}{2B}$$  \hspace{1cm} (1.23)

#### Velocity resolution

Similarly, the radial velocity resolution $\Delta v_r$ can be derived from the Doppler resolution $\Delta f_D$.

Analogous to $\Delta f_B$, $\Delta f_D$ is restricted by the window length in time domain, i.e. $T_{\text{sweep}}$, so

$$\Delta f_D = \frac{1}{T_{\text{sweep}}}$$  \hspace{1cm} (1.24)

Thus, $\Delta v_r$ can be deduced as

$$\Delta v_r = \frac{c}{2 \cdot f_c} \cdot \Delta f_D = \frac{c}{2f_c \cdot T_{\text{sweep}}}$$  \hspace{1cm} (1.25)

### 1.3.4 Signal to noise ratio

#### Noise Figure

In radar system, noise figure is a very important parameter which measures the degradation of the signal to noise ratio ($SNR$), caused by components in the RF signal chain. It is a number by which the performance of a radio receiver can be specified. Mathematically, the noise figure of a system can be defined as
1.3 Characters of FMCW Radar

\[ F = \frac{SNR_{in}}{SNR_{out}} \]  

(1.26)

where SNR\(_{in}\) and SNR\(_{out}\) are the input and output SNR respectively. Alternatively, noise figure may be defined in terms of dB

\[ F_{dB} = 10 \cdot \log \left( \frac{SNR_{in}}{SNR_{out}} \right) = SNR_{in,dB} - SNR_{out,dB} \]  

(1.27)

where SNR\(_{in,dB}\) and SNR\(_{out,dB}\) are SNR in dB.

![System with components in series](image)

in \[ \begin{array}{cccc} \hline \circ & \circ & \circ & \circ \\ \hline \end{array} \]\( F_{sys} \)

out

Fig. 1.10: System with components in series

In the case of the system with components in series, as shown in Fig. 1.10, the system’s noise figure can be derived using Friis’ formula as

\[ F_{sys} = F_1 + \frac{F_2 - 1}{G_1} + \frac{F_3 - 1}{G_1G_2} + \cdots + \frac{F_n - 1}{G_1G_2 \cdots G_{n-1}} \]  

(1.28)

where \(G_n\) and \(F_n\) are the gain and noise figure of each component.

**SNR Calculation**

Here, we provide an example of SNR calculation for a 24 GHz FMCW radar. The system diagram is illustrated in Fig. 1.11 with the parameters of each hardware component depicted.

![Gain and noise figure of the radar components](image)

In this example, a triangular trihedral is placed in front of the FMCW radar. With the dimension \(a = 0.2 \text{ m}\), the RCS of the triangular trihedral is...
Table 1.2: Gain and noise figure of components

<table>
<thead>
<tr>
<th>Component</th>
<th>Gain or Loss</th>
<th>Noise Figure</th>
</tr>
</thead>
<tbody>
<tr>
<td>VCO</td>
<td>$P_{VCO,\text{dB}} = 12 \text{ dBm}$</td>
<td>-</td>
</tr>
<tr>
<td>Power Divider</td>
<td>$L_{PD,\text{dB}} = 3 \text{ dB}$</td>
<td>-</td>
</tr>
<tr>
<td>Antenna</td>
<td>$G_{\text{Tx,dB}} = G_{\text{Rx,dB}} = 17 \text{ dBi}$</td>
<td>-</td>
</tr>
<tr>
<td>LNA</td>
<td>$G_{\text{LNA,dB}} = 18 \text{ dB}$</td>
<td>$F_{\text{LNA,dB}} = 2.5 \text{ dB}$</td>
</tr>
<tr>
<td>Mixer</td>
<td>$L_{\text{mix,dB}} = 11 \text{ dB}$</td>
<td>$F_{\text{mix,dB}} = 11 \text{ dB}$</td>
</tr>
<tr>
<td>IF Amplifier</td>
<td>$G_{\text{IFA,dB}} = 35 \text{ dB}$</td>
<td>$F_{\text{IFA,dB}} = 15 \text{ dB}$</td>
</tr>
<tr>
<td>ADC</td>
<td>-</td>
<td>$F_{\text{AD,dB}} = 55 \text{ dB}$</td>
</tr>
</tbody>
</table>

\[
\sigma = \frac{4\pi a^4}{3\lambda_0^2} = 42.9 \text{ m}^2 \quad (1.29)
\]

The free space attenuation of the triangular trihedral, which is 3 meters away from the FMCW radar, can be calculated as

\[
D_{F,\text{dB}} = 10 \cdot \log \left( \frac{(4\pi)^3 \cdot R^4}{\sigma \cdot \lambda_0^2} \right) = 73.8 \text{ dB} \quad (1.30)
\]

The input power before the LNA at the receiver’s side is

\[
P_{\text{in,dB}} = P_{VCO,\text{dB}} - L_{PD,\text{dB}} + G_{\text{Tx,dB}} - D_{F,\text{dB}} + G_{\text{Rx,dB}} = -30.8 \text{ dBm} \quad (1.31)
\]

while the input thermal noise power can be calculated as

\[
N_{\text{T,dB}} = 10 \cdot \log (k_B T_0 B_N) = -126 \text{ dBm} \quad (1.32)
\]

where $k_B$ is Boltzmann constant, $T$ is the receiver’s temperature in Kelvin (here we assume $T = 290 \text{ K}$) and $B_N$ determined by the low pass filter before the A/D converter is the receiver’s noise bandwidth, which is 50 kHz in this example. So the input $SNR$ is

\[
SNR_{\text{in,dB}} = P_{\text{in,dB}} - N_{\text{T,dB}} = 95.2 \text{ dB} \quad (1.33)
\]

The noise figure of the whole receiver can be calculated according to (1.28) as

\[
F_{\text{sys}} = F_{\text{LNA}} + \frac{F_{\text{mix}} - 1}{G_{\text{LNA}}} + \frac{F_{\text{IFA}} - 1}{G_{\text{LNA}} \cdot G_{\text{mix}}} + \frac{F_{\text{AD}} - 1}{G_{\text{LNA}} \cdot G_{\text{mix}} \cdot G_{\text{IFA}}} \quad (1.34)
\]

Finally, the $SNR_{\text{out,dB}}$ can be calculated according to (1.27) as

\[
SNR_{\text{out,dB}} = SNR_{\text{in,dB}} - F_{\text{sys,dB}} = 80.7 \text{ dB} \quad (1.35)
\]

The total noise power at the A/D-converter input is determined by $B_{\text{Rx}}$ as shown in Fig. 1.12. It is influenced by the sweep time and the maximum range. Regarding the radar image SNR the noise power is determined by the bandwidth of the FFT (Fast Fourier Transformation) resolution cell $\Delta f$. 
Fig. 1.12: Signal spectrum
Bibliography


2 Experiment

To complement the theory that you have learned, this experiment will help you to get a better understanding of the FMCW technology. In this experiment, section 2.1 provides the descriptions of the hardware and their critical parameters. Questions are given in section 2.3. The answers must be handed in before the start of the experiment. With the preparation, you must finish part I and part II of the experiment tasks in the last section within two afternoons.

During the experiment you will

- review the principle and formulas of a FMCW radar system;
- understand the importance of linearity in FMCW signals;
- know how to configure the parameters of a FMCW radar to suit various applications;
- consider the shortages of a FMCW radar and the possible improvements.

2.1 Introduction to Hardware Components

All the components have been integrated as a radar demonstrator (see Fig. 2.1 and Fig. 2.2) and are not allowed to be disassembled during the experiment. The critical parameters of each component have been illustrated in Sec.1.2.2. Here we provide brief instructions with a picture or circuit diagram of them. For the purpose of further study, the data sheets of all the devices used in the experiments are attached in the appendix.

Fig. 2.1: Top view of the 24 GHz FMCW radar demonstrator (without USB-6251 and PC)
2.1.1 Low-pass Filter

The low-pass filter serves to clear the unwanted higher frequency components from the saw-tooth voltage generated by the DAC. It is implemented via a 2nd order active (Sallen Key) structure. The cut-off frequency can be calculated using the component values:

\[ f_c = \frac{1}{2\pi \sqrt{R_7 R_{10} C_{28} C_{38}}} = 18.936 \text{ kHz} \]

It is important to notice that for a faithful representation of the saw-tooth voltage, the cut-off frequency of the low-pass needs to be significantly higher than \(1/T\), with \(T\) being the duration of one period of the saw-tooth voltage, as the Fourier-series expansion of a saw-tooth is rich in harmonics.

- Cut-off frequency: 15.915 kHz

2.1.2 VCO

- Part-number: HMC533LP4
- Power output: 12 dBm
2.1 Introduction to Hardware Components

- Frequency range: 23.8 - 24.8 GHz
- Tune voltage: 2 - 13 V
- Factor of onchip frequency divider: 16

Fig. 2.4: VCO module

2.1.3 Power Divider

- Part-number: PS2-12
- Insertion loss: 4 dB at 24 GHz

Fig. 2.5: Power divider

2.1.4 Antennas

- Gain: 15 dBi
- Azimuth half-power beamwidth: 50.4°
- Elevation half-power beamwidth: 16.3°

2.1.5 LNA

- Part-number: HMC517LC4
- Frequency range: 17 - 26 GHz
- Gain: 18 dB
- Noise figure: 2.6 dB
Fig. 2.6: Transmitter (Receiver) antenna

(a) Elevation

(b) Azimuth

Fig. 2.7: Pattern of the $2 \times 4$ patch array in elevation and azimuth direction

Fig. 2.8: LNA module
2.1.6 Mixer

- Part-number: HMC292LC3B
- Frequency range: 16 - 30 GHz
- Conversion loss: 40 dB

![Mixer module](image)

Fig. 2.9: Mixer module

2.1.7 ADC

- Part-number: NI USB-6251
- Input channels: 16
- Sampling frequency: 1 MS/s
- Output channels: 2
- Update rate: 2.86 MS/s

![NI USB-6251](image)

Fig. 2.10: NI USB-6251

2.1.8 IF Amplifier

The IF amplifier is a rather straightforward Op-Amp circuit. The gain of the amplifier can be adjusted using $R_2$ and has been set to a voltage gain of about 50 i.e. $20 \times \log_{10} (50) = 34$ dB

- Gain: 33.26 dB
2.1.9 Frequency Divider

The frequency divider chain is made up of three stages. The first stage is integrated into the VCO chip and delivers a signal with frequency equivalent to that of the VCO output signal divided by 16, i.e. about 1.5 GHz. Next in line is a \( \mu\)PB1506GV prescaler IC configured to divide the approximately 1.5 GHz output by the VCO by a factor of 256. The output of the prescaler is then slightly shifted from its DC offset using the voltage divider formed by \( R_1 \) and \( R_5 \) so as to make the waveform pass the TTL-logic switching threshold. This becomes necessary because the last stage of the chain consists of a TTL-logic counter, providing a division by a factor of 16, thus leaving us with a frequency of about 366 kHz. As a side effect, the TTL-counter converts the rather ugly output of the prescaler IC into a nice square waveform, which can easily be processed by one of the digital inputs of the ADC/DAC card.

- Factor: 4096
2.2 Introduction to Software Tools

The software tools used in the experiment are LabView and MATLAB.

2.2.1 Tools programmed by LabView

LabView (short for Laboratory Virtual Instrumentation Engineering Workbench) is a platform and development environment for a visual programming language from National Instruments. The control and sampling functions of the radar system are programmed in LabView. However, programming with LabView is not required. Two executable programme are provided for controlling the radar during the experiment. The first one, T1_Control_Voltage.exe, is used for the first experiment to control the output voltage of the DA converter. As shown in Fig. 2.13, the GUI is quite intuitive, and the voltage can be changed in real time.

As shown in Fig. 2.14, the second programme provided is T2-4_FMCW_Radar.exe, which is used for the rest experiments. It is a compact programme for controlling the radar.

The radar’s parameters can be set before you push the Start button. As soon as the programme starts, user is asked to name a .dat for saving the recorded data. It is recommended to create a separate folder for saving all your recorded data. The Buffer progress bar right below Start button indicates the buffer of the USB-6251 onboard memory combined with the
assigned PC memory, it will grow when the processing speed of the PC is not fast enough, e.g. the sweep time is too small which means more FFT calculations are required or the A/D sampling frequency is too high. The program will report error and stop when the buffer is full. Moreover, the display of signal (in time domain or in frequency domain) will not be in real time when the Buffer progress bar is not zero which means current signal are still waiting in the buffer. So it is better to choose the radar parameters smartly when you perform the experiments (the experiments are designed properly that the buffer will remain zero if you choose the right parameters).

The Display Amplitude can be changed in real time to make the range of the signal (time domain) displaying properly. The Display Range is used for change the Range axis in the Up&Down-Sweep PSD Chart in real time.

If the radar works properly, then you can push the Record button, which will stop the display of the signal and start to record the data into the file designated by the user.

The Buffer progress bar indicates the buffer status in the same way as the above one. Finally, the Stop button simply stops the programme.

2.2.2 Tools programmed by MATLAB

The only MATLAB tool provided is a function that loads the recorded data into the workspace. The function is placed in the same folder with T2-4_FMCW_Radar.exe, its GUI is shown in Fig. 2.15. The usage of the function can be found by typing help FMCW_Radar_Loader.

2.3 Preparation for the experiment

In order to be successful in the following experiments, we have prepared four questions to help you review the principles of an FMCW radar.
2.3 Preparation for the experiment

Fig. 2.14: GUI of "T2-4_FMCW_Radar.exe"

Fig. 2.15: GUI of "FMCW_Radar_Loader.m"
2.3.1 Question 1: Doppler frequency

Assuming a car moving along x axis at the velocity $v = 75$ km/h, and the radar with a beamwidth of $\theta = 30^\circ$ captures the car when it moves from $x = 37$ m to $x = 10$ m, the geometry of the radar and car is illustrated in Fig. 2.3.1.

![Illustration of the Doppler frequency caused by the radial velocity](image)

Write down the expression of the Doppler frequency caused by the moving car, and draw the $f_D(t)$ curve using MATLAB ($t = 0$ when $x_{car} = 10$ m).

$$f_d = \text{expression}$$

print the $f_D(t)$ curve out!

2.3.2 Question 2: Signal to noise ratio ($SNR$)

Assuming a triangular trihedral with the length of side, $a = 0.25$ m put 3 m away in front of the radar, the A/D sampling rate is 100 kHz, calculate the $SNR$ of the output signal of AD converter, values of parameters shown in Fig. 2.3.2 are given in Tab. 2.1.

$$SNR = \text{calculation}$$

2.3.3 Question 3: Signal form

The FMCW mechanism can not only be applied on a radar, but also on an ultrasound sensor. Assuming that an FMCW ultrasound sensor with triangular modulation is used for distance measurement, whose parameters are listed in Table 2.2.
2.3 Preparation for the experiment

Fig. 2.17: Block diagram of a FMCW Radar

Table 2.1: Gain and noise figure of components

<table>
<thead>
<tr>
<th>Component</th>
<th>Gain or Loss</th>
<th>Noise Figure</th>
</tr>
</thead>
<tbody>
<tr>
<td>VCO</td>
<td>$P_{\text{VCO}} = 12 \text{ dBm}$</td>
<td>-</td>
</tr>
<tr>
<td>Power Divider</td>
<td>$L_{\text{PD}} = 4 \text{ dB}$</td>
<td>-</td>
</tr>
<tr>
<td>Antenna</td>
<td>$G_{\text{Tx}} = G_{\text{Rx}} = 15 \text{ dBi}$</td>
<td>-</td>
</tr>
<tr>
<td>LNA</td>
<td>$G_{\text{LNA}} = 18 \text{ dB}$</td>
<td>$F_{\text{LNA}} = 2.6 \text{ dB}$</td>
</tr>
<tr>
<td>Mixer</td>
<td>$L_{\text{mix}} = 40 \text{ dB}$</td>
<td>$F_{\text{mix}} = 40 \text{ dB}$</td>
</tr>
<tr>
<td>IF Amplifier</td>
<td>$G_{\text{IFA}} = 33.3 \text{ dB}$</td>
<td>$F_{\text{IFA}} = 19.1 \text{ dB}$</td>
</tr>
<tr>
<td>ADC</td>
<td></td>
<td>$F_{\text{AD}} = 59 \text{ dB}$</td>
</tr>
</tbody>
</table>

1. **Write down the expressions** of the transmitted and received RF signal in time domain: $s_{t,\text{up}}(t)$, $s_{t,\text{down}}(t)$ and $s_r(t)$ (assuming the target is located 1 m away).

   $$s_{t,\text{up}}(t) =$$
   $$s_{t,\text{down}}(t) =$$
   $$s_r(t) =$$

2. **Generate and plot** $s(t)$ for one period using MATLAB (attach the source code).

   **Print it out!**

3. **Generate the IF signal** $s_{\text{IF}}(t)$ according to the mechanism of an ideal mixer, and **plot** it for one period using MATLAB (attach the source code).

   **Print the $s_{\text{IF}}(t)$ out!**
Table 2.2: Parameters of a FMCW ultrasound sensor

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$c$</td>
<td>345 m/s</td>
</tr>
<tr>
<td>$f_0$</td>
<td>30 kHz</td>
</tr>
<tr>
<td>$B$</td>
<td>20 kHz</td>
</tr>
<tr>
<td>$T_{sweep}$</td>
<td>200 ms</td>
</tr>
</tbody>
</table>

2.3.4 Question 4: Range and velocity measurement

Assuming an FMCW radar with triangular modulation is used for baseball training, the parameters of the radar and the ball are given in Table 2.3.

Table 2.3: Parameters

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f_0$</td>
<td>24 GHz</td>
</tr>
<tr>
<td>$B$</td>
<td>1 GHz</td>
</tr>
<tr>
<td>$T_{sweep}$</td>
<td>2 ms</td>
</tr>
<tr>
<td>$r$</td>
<td>250 m</td>
</tr>
<tr>
<td>$v_r$</td>
<td>40 m/s</td>
</tr>
</tbody>
</table>

1. Assuming that the mixer is ideal, **write down** the expression of the dechirped IF signal $s_{IF}$, and **plot** it in time domain with MATLAB.

   $s_{IF} =$ 

   **Print it out!**

2. Use $s_{IF}$ to **measure** the target’s range and velocity with necessary procedure (including the power spectrum of the IF signal with notes, e.g. $f_1$, $f_2$, etc.).

   $R =$ 

   $v =$

3. **Calculate** the range resolution $\Delta r$ and the velocity resolution $\Delta v$ respectively.

   $\Delta r =$ 

   $\Delta v =$

4. Assuming $v_r = 0$, plot the spectrum of IF signal $s_{IF}$ and **measure** the $\Delta r$ on it. **Compare** the measured $\Delta r$ with the theoretical one.
2.3 Preparation for the experiment

Print it out!
2.4 Experiment tasks: Part I

2.4.1 VCO characteristic curve

1. Connect the "User 1" port, which output frequency is \( \frac{1}{16} \times 4096 \) (16*4096) of the output frequency of VCO \( f_{\text{VCO}} \), to an oscilloscope, change the control voltage \( V \) from 1.0 V to 7.0 V with a step of 0.4 V using "T1.Control_Voltage.exe" and record the frequency \( f_r \) for each step in a Table (see Table 2.4).

| \( V \)  | 1.0  | 1.4  | 1.8  | 2.2  | 2.6  | 3.0  | 3.4  | 3.8  | 4.2  | 4.6  | 5.0  | 5.4  | 5.8  | 6.2  | 6.6  | 7.0  |
|---------|------|------|------|------|------|------|------|------|------|------|------|------|------|------|------|
| \( f_r \) |      |      |      |      |      |      |      |      |      |      |      |      |      |      |      |      |
| \( f_{\text{VCO}} \) |      |      |      |      |      |      |      |      |      |      |      |      |      |      |      |      |

Fill in the results in the table!

2. Taking into account that a frequency divider circuit with a factor of 4096 is used here, calculate the output frequency \( f_{\text{VCO}} \) at VCO’s RFOUT port and record them in the third row of Tab. 2.4.

Write it down in the table!

3. Depict the VCO’s characteristic curve \( f_{\text{VCO}} (V) \) using MATLAB (use interp1() for interpolation and choose 'spline' for interpolation method, attach the source code to your final report).

Print it out!

4. Given \( f_0 = 24 \text{ GHz} \), \( B = 500 \text{ MHz} \), \( f_{\text{DA}} = 100 \text{ kHz} \) and \( T_{\text{sweep}} = 10 \text{ ms} \), specify the control voltage \( V_c \) for a triangular FMCW radar using the measured \( f_{\text{VCO}} (V) \) curve, and plot the \( V_c (t) \) curve using MATLAB.

Print it out!

5. Why is the VCO’s characteristic curve needed before the radar measurement?

2.4.2 Distance measurement

A corner reflector is placed in front of the radar \( (1.5 \text{ m} < r < 5 \text{ m}) \). The parameters of a triangular FMCW radar are given as follows: \( f_0 = 24 \text{ GHz} \), \( B = 1 \text{ GHz} \), \( T_{\text{sweep}} = 10 \text{ ms} \).

1. Estimate the minimal AD sampling frequency \( f_s \) for this experiment according to the Nyquist sampling criterion (the real sampling frequency should be 1.3 times of the
2.5 Experiment tasks: Part II

theoretical one).

\[ f_s = \]

2. Use "T2-4_FMCW_Radar.exe" to record the IF signal \( s_{IF} \) for one period, and load the data into MATLAB with function "FMCW_Radar_Loader.m". **Estimate** the target’s \( r \) from \( s_{IF} \).

**Write down the approach!** **Print the results out if necessary!** \( r = \)

3. By comparing the figure with the real scene setting, **indicate** the false target in the figure. What caused the resulting false target? How could this be removed?

4. **Measure** the noise floor of the radar by putting an absorber in front of the radar. By comparing the power spectrum of \( s_{IF} \) with the one with corner reflector, the \( SNR \) of the radar can be estimated. **Calculate** the theoretical \( SNR \) and **compare** it with the estimated one.

**Theoretical \( SNR = \)**

**Measured \( SNR = \)**

**2.5 Experiment tasks: Part II**

**2.5.1 Velocity measurement**

In this task, you will deal with a moving target and try to detect its maximal velocity during the measuring period. The target is supposed to be located 5 meters away from the radar, the velocity of the target is determined by you, the required \( \Delta r \) is 0.125 m and the required \( \Delta v_r \) is 1 m/s.

1. The proposed system parameters are given in Table 2.5 \( (f_0 = 24 \text{ GHz}) \). **select** the proper ones according to the knowledge that you have learned from the previous theoretical part and **write down** the reasons for these decisions in your final report.

<table>
<thead>
<tr>
<th>Parameter</th>
<th>value</th>
</tr>
</thead>
<tbody>
<tr>
<td>( B ) (MHz)</td>
<td>200</td>
</tr>
<tr>
<td>( T_{\text{sweep}} ) (ms)</td>
<td>2</td>
</tr>
<tr>
<td>( f_s ) (kHz)</td>
<td>100</td>
</tr>
</tbody>
</table>
2. You (or your partner) are required to move a corner reflector towards or away from the radar, while your partner (or you) records the IF signal $s_{IF}$ by "T2-4_FMCW_Radar.exe".

3. Load the recorded $s_{IF}$ into MATLAB by "FMCW_Radar_Loader.m", **estimate** the target’s $r$ and $v$ of each up- and down-sweep, and finally **plot** the $r(t)$ and $v(t)$ curves of the period during which the target is moving.

**Print it out with explanation including MATLAB code!**

### 2.5.2 Range Resolution

In this section, you are required to verify the parameters that affect $\Delta r$.

<table>
<thead>
<tr>
<th>$B$ (MHz)</th>
<th>$T_{\text{sweep}}$ (ms)</th>
<th>$\Delta r$ (m)</th>
</tr>
</thead>
<tbody>
<tr>
<td>200</td>
<td>10</td>
<td></td>
</tr>
<tr>
<td>500</td>
<td>20</td>
<td></td>
</tr>
<tr>
<td>1000</td>
<td>40</td>
<td></td>
</tr>
</tbody>
</table>

1. How is the $\Delta r$ dependent on the $B$ and $T_{\text{sweep}}$ theoretically?
2. **Fill out** Table 2.6 by calculating the theoretical $\Delta r$ of different $B$ and $T_{\text{sweep}}$.
3. Put a corner reflector in front of the radar at a certain distance as a point target, **measure** the IF signal for the variation of $B$ ($T_{\text{sweep}}=10$ ms) and the variation of $T_{\text{sweep}}$ ($B=500$ MHz). **Plot** the corresponding power spectrum of each setting and indicate the real $\Delta r$.

**Print it out with explanation!**

4. Conclude the relationship of $\Delta r$ with $B$ and $T_{\text{sweep}}$ based on the measured results. **Compare** it with the theoretical analysis.

**Conclusion:**
The TL08x JFET-input operational amplifier family is designed to offer a wider selection than any previously developed operational amplifier family. Each of these JFET-input operational amplifiers incorporates well-matched, high-voltage JFET and bipolar transistors in a monolithic integrated circuit. The devices feature high slew rates, low input bias and offset currents, and low offset-voltage temperature coefficient. Offset-matched, high-voltage JFET and bipolar transistors in a monolithic integrated circuit. The devices feature developed operational amplifier family. Each of these JFET-input operational amplifiers incorporates.

The M-suffix devices are characterized for operation over the full military temperature range of −55°C to 125°C. The devices are characterized for operation from −40°C to 85°C. The C-suffix devices are characterized for operation from −40°C to 85°C. The Q-suffix devices are characterized for operation from 0°C to 70°C. The devices are characterized for operation from −40°C to 85°C. The devices are characterized for operation from −40°C to 85°C. The devices are characterized for operation from −40°C to 85°C.

The C-suffix devices are characterized for operation from 0°C to 70°C. The Q-suffix devices are characterized for operation from −40°C to 85°C. The C-suffix devices are characterized for operation from 0°C to 70°C. The C-suffix devices are characterized for operation from 0°C to 70°C. The devices are characterized for operation from −40°C to 85°C. The devices are characterized for operation from −40°C to 85°C. The devices are characterized for operation from −40°C to 85°C. The devices are characterized for operation from −40°C to 85°C. The devices are characterized for operation from −40°C to 85°C. The devices are characterized for operation from −40°C to 85°C.
The HMC533LP4 & HMC533LP4E are MMIC VCOs with a Divide-by-16 feature. They integrate resonators, negative resistance devices, varactor diodes, and feature a divide-by-16 output. The VCOs maintain excellent phase noise performance over temperature, shock, and process due to their monolithic structure. Power output is +12 dBm typical from a +5V supply voltage. Prescaler function can be disabled to conserve current if not required. The voltage-controlled oscillator is packaged in a leadless QFN 4 x 4 mm surface mount package.

Electrical Specifications, $T_a = +25^\circ$C, $Vcc$1, $Vcc$2, $Vcc$3 = +5V

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Min.</th>
<th>Typ.</th>
<th>Max.</th>
<th>Units</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency Range</td>
<td>23.8 - 24.8 GHz</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Power Output RFOUT, RFOUT/16</td>
<td>$+9$ dBm</td>
<td>$-7$ dBm</td>
<td>$+15$ dBm</td>
<td></td>
</tr>
<tr>
<td>SSB Phase Noise @ 100 kHz Offset, $V_{tune}$ = +5V @ RFOUT</td>
<td>$-95$ dBc/Hz</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Tune Voltage $V_{tune}$</td>
<td>$2.1$ V</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Supply Current $I_{cc}$</td>
<td>$180$ mA</td>
<td>$220$ mA</td>
<td>$260$ mA</td>
<td></td>
</tr>
<tr>
<td>Tune Port Leakage Current ($V_{tune}$ = 13V)</td>
<td>$10$ μA</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Output Return Loss</td>
<td>$3$ dB</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Harmonics/Subharmonics 1/2</td>
<td>$26$ dBc</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Harmonics/Subharmonics 3/2</td>
<td>$37$ dBc</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Pulling (into a 2.0:1 VSWR)</td>
<td>$13$ MHz</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Pushing @ $V_{tune}$ = 5V</td>
<td>$80$ MHz/V</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Frequency Drift Rate</td>
<td>$2.3$ MHz/°C</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

For further details and to place orders, please visit our website or contact Hittite Microwave Corporation.

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20 Alpha Road, Chelmsford, MA 01824 Phone: 978-250-3343 Fax: 978-250-3373
Order On-line at www.hittite.com
The µPB1506GV and µPB1507GV are 3.0 GHz input, high division silicon prescaler ICs for analog DBS tuner applications. These ICs divide-by-256, 128 and 64 and contribute to produce analog DBS tuners with the use of 17 K series DTS controller or standard CMOS PLL synthesizer IC. The µPB1506GV/µPB1507GV are shrink package versions of the PB586G/PB588G or PB1505GR so that these smaller packages contribute to reduce the mounting space replacing from conventional ICs.

The µPB1506GV and µPB1507GV are manufactured using NEC’s high fT NESAT™IV silicon bipolar process. This process uses silicon nitride passivation film and gold electrodes. These materials can protect chip surface from external pollution and prevent corrosion/migration. Thus, these ICs have excellent performance, uniformity and reliability.

FEATURES
- High toggle frequency: fIN = 0.5 GHz to 3.0 GHz
- High-density surface mounting: 8-pin plastic SSOP (175 mil)
- Low current consumption: 5 V, 19 mA
- Selectable high division: 256, 128, 64
- Pin connection variation: µPB1506GV and µPB1507GV

APPLICATION
- These ICs can used as a prescaler between local oscillator and PLL frequency synthesizer included modulus prescaler. For example, following application can be chosen:
  - Analog DBS tuner's synthesizer
  - Analog CATV converter synthesizer

ORDERING INFORMATION

<table>
<thead>
<tr>
<th>PART NUMBER</th>
<th>PACKAGE</th>
<th>MARKING</th>
<th>SUPPLYING FORM</th>
</tr>
</thead>
<tbody>
<tr>
<td>µPB1506GV-E1</td>
<td>8-pin plastic</td>
<td>1028</td>
<td>Embossed tape/8 mm wide. Pin 1 is in tape pull-out direction. 1 000/pack.</td>
</tr>
<tr>
<td>µPB1507GV-E1</td>
<td>SSOP (175 mil)</td>
<td>1907</td>
<td></td>
</tr>
</tbody>
</table>

Remarks: To order evaluation samples, please contact your local NEC sales office.

(Part number for sample order: µPB1506GV, µPB1507GV)

PRODUCT LINE-UP

<table>
<thead>
<tr>
<th>Features (division, Freq.)</th>
<th>Part No.</th>
<th>Icc (mA)</th>
<th>Vcc (V)</th>
<th>Package</th>
<th>Pin connection</th>
</tr>
</thead>
<tbody>
<tr>
<td>256, 128, 64, 2.5 GHz</td>
<td>µPB586G</td>
<td>28</td>
<td>0.55 to 2.5</td>
<td>4.5 to 5.5</td>
<td>8-pin SOP 225 mil NEC original</td>
</tr>
<tr>
<td>256, 128, 64, 2.5 GHz</td>
<td>µPB588G</td>
<td>26</td>
<td>0.55 to 2.5</td>
<td>4.5 to 5.5</td>
<td>Standard</td>
</tr>
<tr>
<td>256, 128, 64, 3.0 GHz</td>
<td>µPB1505GR</td>
<td>14</td>
<td>0.55 to 3.0</td>
<td>4.5 to 5.5</td>
<td>Standard</td>
</tr>
<tr>
<td>256, 128, 64, 3.0 GHz</td>
<td>µPB1506GV</td>
<td>19</td>
<td>0.55 to 3.0</td>
<td>4.5 to 5.5</td>
<td>8-pin SOP 175 mil NEC original</td>
</tr>
<tr>
<td>256, 128, 64, 3.0 GHz</td>
<td>µPB1507GV</td>
<td>19</td>
<td>0.55 to 3.0</td>
<td>4.5 to 5.5</td>
<td>Standard</td>
</tr>
</tbody>
</table>

Remarks: This table shows the TYP values of main parameters. Please refer to ELECTRICAL CHARACTERISTICS.

- µPB586G and µPB588G are discontinued.

INTERNAL BLOCK DIAGRAM
Dual 4-bit synchronous binary counter  74HC/HCT4520

FEATURES
• Output capability: standard
• IC category: MSI

GENERAL DESCRIPTION
The 74HC/HCT4520 are high-speed Si-gate CMOS devices and are pin compatible with the '4520' of the '4000B' series. They are specified in compliance with JEDEC standard no. 7A.

The 74HC/HCT4520 are dual 4-bit internally synchronous binary counters with an active HIGH clock input (nCP0) and an active LOW clock input (nCP1), buffered outputs from all four bit positions (nQ0 to nQ3) and an active HIGH overriding asynchronous master reset input (nMR).

The counter advances on either the LOW-to-HIGH transition of nCP0 (nCP1) if nCP1 is HIGH or the HIGH-to-LOW transition of nCP1 (nCP0) is LOW. Either nCP0 or nCP1 may be used as a clock input to the counter and the other clock input may be used as a clock enable input. A HIGH on nMR resets the counter (nQ0 to nQ3 = LOW) independent of nCP0 and nCP1.

APPLICATIONS
• Multistage synchronous counting
• Multistage asynchronous counting
• Frequency dividers

QUICK REFERENCE DATA
GND = 0 V; T amb = 25 °C; t r = t f = 6 ns

<table>
<thead>
<tr>
<th>SYMBOL</th>
<th>PARAMETER</th>
<th>CONDITIONS</th>
<th>TYPICAL</th>
<th>UNIT</th>
</tr>
</thead>
<tbody>
<tr>
<td>tPD</td>
<td>propagation delay</td>
<td>nCP0, nCP1 to nQ0</td>
<td>nCP1 = LOW, nCP0 = HIGH</td>
<td>24</td>
</tr>
<tr>
<td>tPD</td>
<td>propagation delay</td>
<td>nMR to nQ1</td>
<td>nMR = LOW</td>
<td>13</td>
</tr>
<tr>
<td>fmax</td>
<td>maximum clock frequency</td>
<td></td>
<td></td>
<td>68 MHz</td>
</tr>
<tr>
<td>C I</td>
<td>input capacitance</td>
<td></td>
<td></td>
<td>3.5 pF</td>
</tr>
<tr>
<td>C PD</td>
<td>power dissipation capacitance per counter</td>
<td>notes 1 and 2</td>
<td></td>
<td>29 pF</td>
</tr>
</tbody>
</table>

Notes
1. C PD is used to determine the dynamic power dissipation (P D) in µW:
   \[ P D = C PD \times V CC^2 \times f i + \sum (C L \times V CC^2 \times f o) \]
   where:
   - f i = input frequency in MHz
   - f o = output frequency in MHz
   - \( \sum (C L \times V CC^2 \times f o) \) = sum of outputs
   - C L = output load capacitance in pF
   - V CC = supply voltage in V

2. For HC, the condition is V CC = 0 V to V CC
   For HCT the condition is V CC = 1.5 V

ORDERING INFORMATION
See "74HC/HCT/HCU/HCMOS Logic Package Information".
HMC292LC3B
GaAs MMIC FUNDAMENTAL MIXER, 16 - 30 GHz

Features
- Passive: No DC Bias Required
- Input IP3: +20 dBm
- LO/RF Isolation: 40 dB
- Wide IF Bandwidth: DC - 8 GHz
- RoHS Compliant 3x3 mm SMT Package

General Description
The HMC292LC3B is a general purpose passive double balanced mixer in a leadless RoHS-Compliant SMT package that can be used as an upconverter or downconverter between 16 and 30 GHz. This mixer requires no external components or matching circuitry. The HMC292LC3B provides excellent LO to RF and LO to IF suppression due to optimized balun structures. The mixer operates with LO drive levels above +9 dBm. The HMC292LC3B eliminates the need for wire bonding, allowing use of surface mount manufacturing techniques.

Electrical Specifications, $T_e = +25^\circ$ C, IF = 1 GHz, $LO = +13$ dBm*

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency Range, RF &amp; LO</td>
<td>16 - 26</td>
<td></td>
<td></td>
<td>26 - 30</td>
<td></td>
<td></td>
<td>GHz</td>
</tr>
<tr>
<td>Conversion Loss</td>
<td>0.0</td>
<td>11.0</td>
<td></td>
<td>0.5</td>
<td>12.5</td>
<td></td>
<td>dB</td>
</tr>
<tr>
<td>Noise Figure (dBm)</td>
<td>8.0</td>
<td>11.0</td>
<td></td>
<td>6.5</td>
<td>12.5</td>
<td></td>
<td>dB</td>
</tr>
<tr>
<td>LO to IF Isolation</td>
<td>24</td>
<td>32</td>
<td></td>
<td>28</td>
<td>34</td>
<td></td>
<td>dB</td>
</tr>
<tr>
<td>IP2 (dBm)</td>
<td>14</td>
<td>23</td>
<td></td>
<td>24</td>
<td>30</td>
<td></td>
<td>dB</td>
</tr>
<tr>
<td>IP3 (dBm)</td>
<td>18</td>
<td>27</td>
<td></td>
<td>20</td>
<td>30</td>
<td></td>
<td>dB</td>
</tr>
<tr>
<td>1 dB Gain Compression (dB)</td>
<td>12</td>
<td>19</td>
<td></td>
<td>15</td>
<td>26</td>
<td></td>
<td>dB</td>
</tr>
</tbody>
</table>

*Unless otherwise noted, all measurements performed as downconverter, IF = 1 GHz.
HMC517LC4

SMT PHEMT LOW NOISE AMPLIFIER, 17 - 26 GHz

Typical Applications
The HMC517LC4 is ideal for use as a LNA or driver amplifier for:
- Point-to-Point Radios
- Point-to-Port Radios & VSAT
- Test Equipment and Sensors
- Military

Features
- Noise Figure: 2.5 dB
- Gain: 19 dB
- IP3: +23 dBm
- Single Supply: +3V @ 67 mA
- 50 Ohm Matched Input/Output
- RoHS Compliant 4 x 4 mm Package

General Description
The HMC517LC4 chip is a high dynamic range GaAs PHEMT MMIC Low Noise Amplifier (LNA) housed in a leadless “Pb free” RoHS compliant SMT package. The HMC517LC4 provides 19 dB of small signal gain, 2.5 dB of noise figure and has an output IP3 of +23 dBm. The P1dB output power of +13 dBm enables the LNA to also function as a LO driver for balanced, IQ or image reject mixers. The HMC517LC4 allows the use of surface mount manufacturing techniques.

Electrical Specifications, \( T_{\text{io}} = 25^\circ \text{C}, V_{\text{dd}} 1, 2, 3 = +3V \)

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Min</th>
<th>Typ</th>
<th>Max</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency Range</td>
<td>17 - 22 GHz</td>
<td>22 - 26 GHz</td>
<td></td>
</tr>
<tr>
<td>Gain</td>
<td>16</td>
<td>19</td>
<td>15</td>
</tr>
<tr>
<td>Gain Variation Over Temperature</td>
<td>0.02</td>
<td>0.03</td>
<td>0.02</td>
</tr>
<tr>
<td>Noise Figure</td>
<td>2.5</td>
<td>2.2</td>
<td>2.8</td>
</tr>
<tr>
<td>Input Return Loss</td>
<td>15</td>
<td>17</td>
<td>15</td>
</tr>
<tr>
<td>Output Return Loss</td>
<td>15</td>
<td>17</td>
<td>15</td>
</tr>
<tr>
<td>Output Power at 1 dB Compact (P1dB)</td>
<td>15</td>
<td>17</td>
<td>15</td>
</tr>
<tr>
<td>Saturated Output Power (Psat)</td>
<td>15</td>
<td>17</td>
<td>15</td>
</tr>
<tr>
<td>Output Third Order Intercept (IP3)</td>
<td>24</td>
<td>24</td>
<td>24</td>
</tr>
<tr>
<td>Supply Current (Idd)(Vdd = +3V)</td>
<td>67</td>
<td>67</td>
<td>67</td>
</tr>
</tbody>
</table>

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20 Alpha Road, Chelmsford, MA 01824  Phone: 978-250-3343  Fax: 978-250-3373
Order On-line at www.hittite.com
The NE/SA/SE5534/5534A are single high-performance low noise operational amplifiers. Compared to other operational amplifiers, such as TL083, they show better noise performance, improved output drive capability, and considerably higher small-signal and power bandwidths.

This makes the devices especially suitable for application in high quality and professional audio equipment, in instrumentation and control circuits and telephone channel amplifiers. The op amps are internally compensated for gain equal to, or higher than, three. The frequency response can be optimized with an external compensation capacitor for various applications (unity gain amplifier, capacitive load, slew rate, low overshoot, etc.).

Features
- Small-Signal Bandwidth: 10 MHz
- Output Drive Capability: 600 \( \frac{\text{V}}{\text{A}} \)
- Input Noise Voltage: 4 nV
- DC Voltage Gain: 100000
- AC Voltage Gain: 6000 at 10 kHz
- Power Bandwidth: 200 kHz
- Slew Rate: 13 V/\( \mu \text{s} \)
- Large Supply Voltage Range: 3.0 to 20 V
- Pb-Free Packages are Available

Applications
- Audio Equipment
- Instrumentation and Control Circuits
- Telephone Channel Amplifiers
- Medical Equipment

See detailed ordering and shipping information in the package dimensions section on page 8 of this data sheet.

ORDERING INFORMATION

PIN CONNECTIONS

DEVICE MARKING INFORMATION

See general marking information in the device marking section on page 8 of this data sheet.

ORDERING INFORMATION

See detailed ordering and shipping information in the package dimensions section on page 8 of this data sheet.

MAXIMUM RATINGS

<table>
<thead>
<tr>
<th>Rating</th>
<th>Symbol</th>
<th>Value</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Supply Voltage</td>
<td>( V_S )</td>
<td>±22 V</td>
<td></td>
</tr>
<tr>
<td>Input Voltage</td>
<td>( V_{\text{IN}} )</td>
<td>±0.5 V</td>
<td></td>
</tr>
<tr>
<td>Differential Input Voltage (Note 1)</td>
<td>( V_{\text{DIFF}} )</td>
<td>±0.5 V</td>
<td></td>
</tr>
<tr>
<td>Operating Temperature Range</td>
<td>N</td>
<td>0 to 70°C</td>
<td></td>
</tr>
<tr>
<td>S</td>
<td>-40 to 85°C</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Se</td>
<td>-55 to 125°C</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Junction Temperature Range</td>
<td>( T_J )</td>
<td>-5 to 125°C</td>
<td></td>
</tr>
<tr>
<td>Power Dissipation at 25°C</td>
<td>P</td>
<td>150 W</td>
<td></td>
</tr>
<tr>
<td></td>
<td>D Package</td>
<td>750 mW</td>
<td></td>
</tr>
<tr>
<td></td>
<td>N Package</td>
<td>750 mW</td>
<td></td>
</tr>
<tr>
<td></td>
<td>D Package</td>
<td>130 mW</td>
<td></td>
</tr>
<tr>
<td></td>
<td>N Package</td>
<td>130 mW</td>
<td></td>
</tr>
<tr>
<td></td>
<td>N Package</td>
<td>158 mW</td>
<td></td>
</tr>
<tr>
<td>Thermal Resistance, Junction-to-Ambient</td>
<td>R</td>
<td>150 °C/W</td>
<td></td>
</tr>
<tr>
<td>Output Short-Circuit Duration (Note 2)</td>
<td>-</td>
<td>Infinite</td>
<td></td>
</tr>
</tbody>
</table>

Because exceeding Maximum Ratings may damage the device, Maximum Ratings are stress ratings only. Functional operation above the Recommended Operating Conditions is not implied. Extended exposure to stresses above the Recommended Operating Conditions may affect device reliability.

1. Diodes protect the inputs against overvoltage. Therefore, unless current-limiting resistors are used, large currents will flow if the differential input voltage exceeds 0.5 V. Maximum current should be limited to 10 mA.

2. Output may be shorted to ground at \( V_S = 15 \) V, \( T_J = 25 \) °C. Temperature and/or supply voltages must be limited to ensure dissipation rating is not exceeded.