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# Radar Performance of Ultra Wideband Waveforms

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## 1. Introduction

In the early days of radar, range resolution was made by transmitting a short burst of electromagnetic energy and receiving the reflected signals. This evolved into modulating a sinusoidal carrier into transmitting pulses at a given repetition interval. To get higher resolution in the radars the transmitted pulses got shorter and thereby the transmitted spectrum larger. As will be shown later the Signal-to-Noise Ratio (SNR) is related to the transmitted energy in the radar signal. The energy is given by the transmitted peak power in the pulse and the pulse length. Transmitting shorter pulses to get higher range resolution also means that less energy is being transmitted and reduced SNR for a given transmitter power. The radar engineers came up with radar waveforms that was longer in time and thereby had high energy and at the same time gave high range resolution. This is done by spreading the frequency bandwidth as a function of time in the pulse. This can be done either by changing the frequency or by changing the phase.

If the bandwidth is getting large compared to the center frequency of the radar, the signal is said to have an Ultra Wide Bandwidth (UWB), see (Astanin & Kostylev, 1997) and (Taylor, 2001). The definition made by FCC for an UWB signal is that the transmitted spectrum occupies a bandwidth of more than 500 MHz or greater than 25% of the transmitted signal center frequency. UWB signals have been used successfully in radar systems for many years. Ground Penetrating Radar (GPR) can penetrate the surface of the ground and image geological structures. Absorption of the radar waves in the ground is very frequency dependent and increases with increasing frequency. Lower frequencies penetrate the ground better than higher frequency. To transmit a low frequency signal and still get high enough range resolution calls for a UWB radar signal. The interest in using UWB signals in radar has increased considerably after FCC allocated part of the spectrum below 10 GHz for unlicensed use. Newer applications are through the wall radar for detecting people behind walls or buried in debris. Also use of UWB radar in medical sensing is seeing an increased interest the later years.

UWB radar signal may span a frequency bandwidth from several hundred of MHz to several GHz. This signal bandwidth must be captured by the radar receiver and digitized in some way. To capture and digitize a bandwidth that is several GHz wide and with sufficient resolution is possible but very costly energy and money wise. This has been solved in the impulse waveform only taking one receive sample for each transmitted pulse. In the Step-Frequency (SF) waveform the frequencies are transmitted one by one after each other. A

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general rule for UWB radars is that all of the different waveform techniques have different methods to reduce the sampling requirement. The optimal would be to collect the entire reflected signal in time and frequency at once and any technique that is only collecting part of the received signal is clearly not optimal.

This chapter will discuss how different UWB waveforms perform under a common constraint given that the transmitted signal has a maximum allowable Power Spectral Density (PSD). The spectral limitations for Ground Penetration Radars (GPR) is given in Section 2 together with a definition on System Dynamic Range (SDR). In Section 3 a short presentation on the mostly used UWB-radar waveforms are given together with an expression for the SDR. An example calculation for the different waveforms are done in Section 4 and a discussion on how radar performance can be measured in Section 5.

## 2. Radar performance

There are different radar performance measures for a given radar system. In this chapter only the SDR and related parameters will be discussed. Another important characteristic of a radar waveform is how the radar system behave if the radar target is moving relative to the radar. This can be studied by calculating the ambiguity function for the radar system. In a narrow band radar the velocity of the radar target gives a shift in frequency of the received waveform compared to the transmitted one. For a UWB-waveform the received waveform will be a scaled version of the transmitted signal. This is an important quality measure for a radar system but will not be discussed in this chapter.

### 2.1 Radiation limits

A comparison between an Impulse Radar (IR) and a Step Frequency (SF) radar was done by (Hamran et al., 1995). No constraint on the transmitted spectrum was done in that comparison. The new licensing rules for UWB signals put a maximum transmitted Power Spectral Density (PSD) on the Equivalent Isotropically Radiated Power (EIRP) peak emissions from the UWB device. The unit of the PSD is dBm/Hz and is measured as peak

Frequency Range (MHz)	Maximum Peak PSD
30 to 230	-44.5 dBm/120 kHz
230 to 1 000	-37.5 dBm/120 kHz
1 000 to 18 000	-30 dBm/MHz

Table 1. The maximum allowed measured radiated PSD for GPR/WPR imaging systems according to European rules

Frequency Range (MHz)	Maximum Mean PSD (dBm/MHz)
<230	-65
230 to 1000	-60
1000 to 1600	-65
1600-3400	-51.3
3400-5000	-41.3
5000-6000	-51.3
>6000	-65

Table 2. The maximum allowed mean PSD for GPR/WPR imaging systems according to European rules

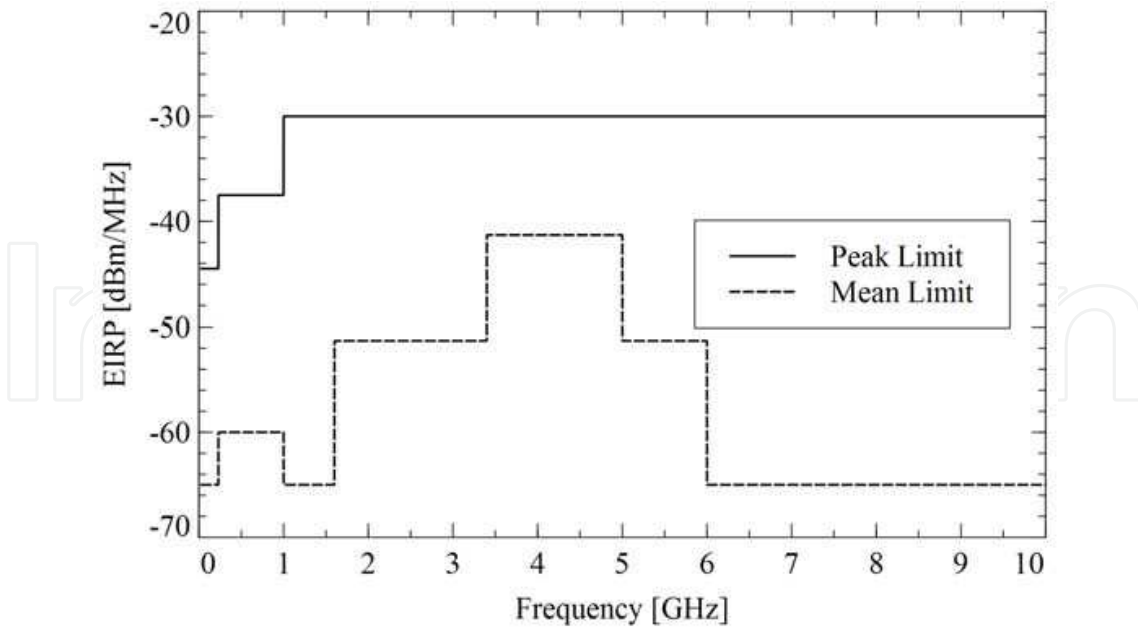


Fig. 1. The maximum allowed measured radiated PSD and mean PSD for GPR/WPR imaging systems

power in a 1 MHz measurement bandwidth (mB) above 1 GHz. There is no single set of emission limits for all UWB devices as both the level in dBm/Hz and the mB is changing with device and frequency. Table 1, Table 2 and Figure 1 give an example on how the PSD varies with frequency for GPR under European rules, (ETSI EN 302 066-1, 2007).

The limits in Table 1 are in principle the same for time domain systems like impulse radars and for frequency domain systems like SF-systems. The way the PSD is measured is however different see (ETSI EN 302 066-1, 2007) and (Chignell & Lightfoot, 2008).

For impulse systems the following formula should be used:

$$PSD_{mean} = P_{peak} + conversion\_factor \tag{1}$$

with:

$$conversion\_factor = 10\log(PRF \times \tau)$$

where  $\tau$  is the pulse width of the GPR transmitter measured at 50 % amplitude points and  $PRF$  is the pulse repetition frequency of the pulse. The peak power and pulse length is measured using a wide bandwidth oscilloscope.

For systems using SF-waveforms the signal is formed by transmitting a sequence of discrete frequencies each having a Dwell Time ( $DT$ ) and a total sequence length called Scan Time ( $ST$ ). Calculating the mean PSD for a SF-system the following formula should be used:

$$PSD_{mean} = P_{peak} + conversion\_factor \tag{2}$$

with:

$$conversion\_factor = 10\log(DT/ST)$$

where  $DT$  is measured at 50 % amplitude points at a frequency near the maximum of the radiated spectrum, using a spectrum analyser in zero span mode and 1 MHz resolution bandwidth.

A simplified PSD spectrum will be used during the discussion in this chapter. The objectives are to see how the different type of UWB-waveforms performs under a given constraint on the radiated PSD. The simplified spectrum mask is just a constant maximum limit on the PSD over a bandwidth  $B$ . The PSD is measured for a given receiver measurement bandwidth,  $mB$ . For the mean PSD given in Table 2 the measurement bandwidth is  $mB = 1\text{MHz}$ .

## 2.2 System Dynamic Range (SDR)

The *SDR* is the ratio between the peak radiated power from the transmitting antenna and the minimum detectable peak signal power entering the receiver antenna. This number quantifies the maximum amount of loss the radar signal can have, and still be detectable in the receiver. Since the minimum detectable signal is strongly related to the integration time, the performance number should be given for a given integration time. In the literature the *SDR* measure is defined in many different ways and many names are used such as System Performance or System Q. Most of them excludes the antenna gain on the radar side and are different from what is used here. We will use the definition given in (Hamran et al., 1995) with the exception that we will call it the System Dynamic Range, *SDR*, of the radar system instead of only Dynamic Range. This number tells us that if a reflector has a maximum loss which is less than the radar system *SDR*, then the reflector can be detected by the radar system. This assumes that the reflected signal is within the Receiver Dynamic Range (*RDR*) of the radar system, thus some gain parameters in the radar system may have to be varied to fully make use of the *SDR*. To be able to calculate the *SDR* of a given radar system the matched filter radar equation must be used. Using the simple radar equation for a transmitting pulse having a time-bandwidth product equal unity makes one run into problems when trying to compare different system on a general basis. The following discussion is based on (Hamran et al., 1995) The Signal-to-Noise ratio (*SNR*) for a radar receiver that is matched to the transmitted signal is given by the following equation:

$$SNR = \frac{2E_R}{N_0} = \frac{2E_T G^2 \lambda^2 \sigma}{N_0 (4\pi)^3 R^4} \quad (3)$$

This is the matched receiver radar equation, where  $E_T$  = transmitted signal energy,  $E_R$  = received signal energy,  $N_0$  = noise spectral density,  $G$  = transmit/receive antenna gain,  $\lambda$  = wavelength,  $\sigma$  = target radar cross section and  $R$  = range to target.

For the matched filter in the white noise case, the *SNR* is dependent only on the signal energy and the noise spectral density, and is otherwise independent of the signal waveform, provided that the filter is matched to the signal, (DiFranco & Rubin, 1980). The power spectral density of the noise can be expressed by:

$$\frac{N_0}{2} = k_B T_0 F \quad (4)$$

where  $k_B$  is Boltzmann's constant ( $1.380650310^{-23} \text{m}^2 \text{kg} \text{s}^{-2} \text{K}^{-1}$ ),  $T_0$  is the room temperature in Kelvins (typically 290 K) and  $F$  is the dimensionless receiver noise figure. The signal-to-noise ratio is directly proportional to the transmitted energy. Thus, the longer the receiver integration time, the higher the signal-to-noise-ratio will be. The average transmitting power over a time period is given by:

$$\hat{P}_T = \frac{E_T}{\tau_i}. \quad (5)$$

Bringing all this into Equation 3 and splitting the radar equation in two separate parts where the first contains the radar system dependent parameters and the second the medium dependent parameters we have

$$\frac{\hat{P}_T \tau_i G^2}{k_B T_0 F(SNR)} = \left[ \frac{\lambda^2 \sigma}{(4\pi)^3 R^4} \right]^{-1}. \quad (6)$$

The left side contains the radar system dependent parts and the right the propagation and reflection dependent parts. The *SDR* of the radar system is defined as:

$$SDR = \frac{\hat{P}_T \tau_i G^2}{k_B T_0 F(SNR)}. \quad (7)$$

When comparing different radar systems, the same integration or observation time should be used. Comparing two radar systems having the same antennas and integration time, the following expression is obtained assuming the same *SNR* for both systems:

$$\frac{SDR_1}{SDR_2} = \frac{\hat{P}_{T1}}{F_1} / \frac{\hat{P}_{T2}}{F_2}. \quad (8)$$

We see that only the transmitted average power and receiver noise figure need to be considered. This is however only correct if the radar receiver is matched to the transmitted waveform. If the receiver is not matched, a mismatch loss must be introduced. Several factors are involved in this loss and could be taken into account by introducing a matching coefficient (=1 if matched) in the equations above.

In the following discussion we assume that all radars have the same antenna and the antenna gain has been set to 1. It is further assumed that all radars have the same noise figure that has been set equal to 5. The *SNR* is set to 1. The transmitted spectrum is the same for all waveforms and is given in Figure 2. The spectrum is flat with a power spectral density of *PSD* and a bandwidth of *B*.

### 3. UWB waveforms

Figure 3 show a pulse train that is made up of pulses with a 1 GHz center frequency and a Gaussian amplitude shape with a bandwidth of 80 % of the carrier frequency. The pulse repetition interval (*PRI*) is 10 ns. The frequency spectrum of the pulse train is shown in the lower plot in the figure. We see that the spectrum of the pulse train is a line spectrum centered at 1 GHz and a -4 dB bandwidth of 800 MHz. The distance between the lines is given by  $1/PRI$  and is 100 MHz. The pulse train and its spectrum represents the same signal. In stead of sending the pulse train one could send the line spectrum. In fact, if a line spectrum made by several fixed oscillator transmitting at the same time and amplitude weighted accordingly, the different sinusoidal signals would sum up and produce the pulse train. If the radar scattering scene does not move we could send out one line at a time and measure the reflected signal for each frequency component. A Fourier transform of the measured frequency response would produce the reflected time signal just like transmitting

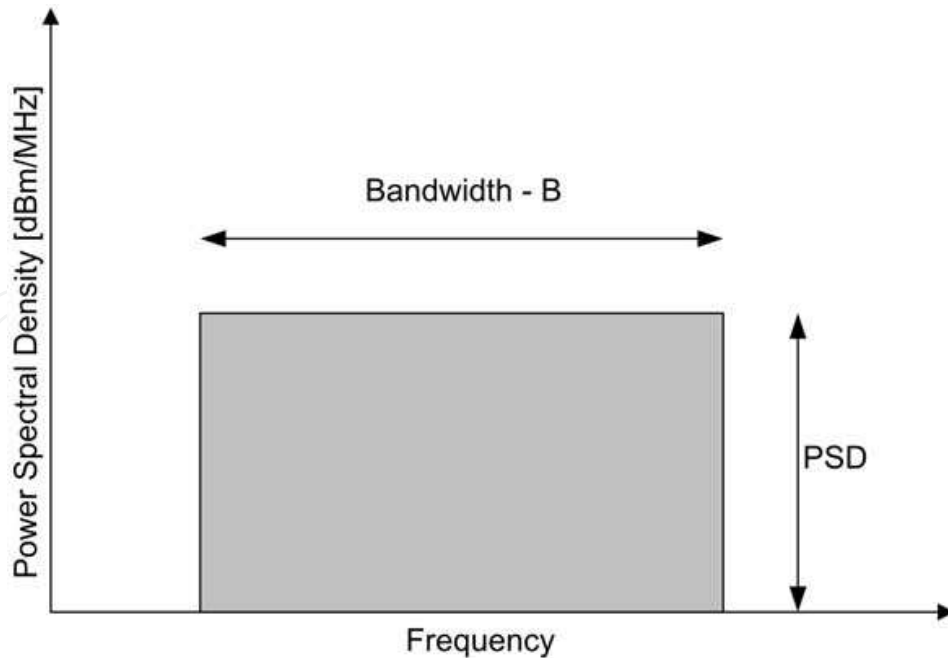


Fig. 2. For the comparison between the different radar waveforms the simple spectrum showed in the figure will be used. It has a flat spectrum with a power spectral density of  $PSD$  over the bandwidth  $B$ .

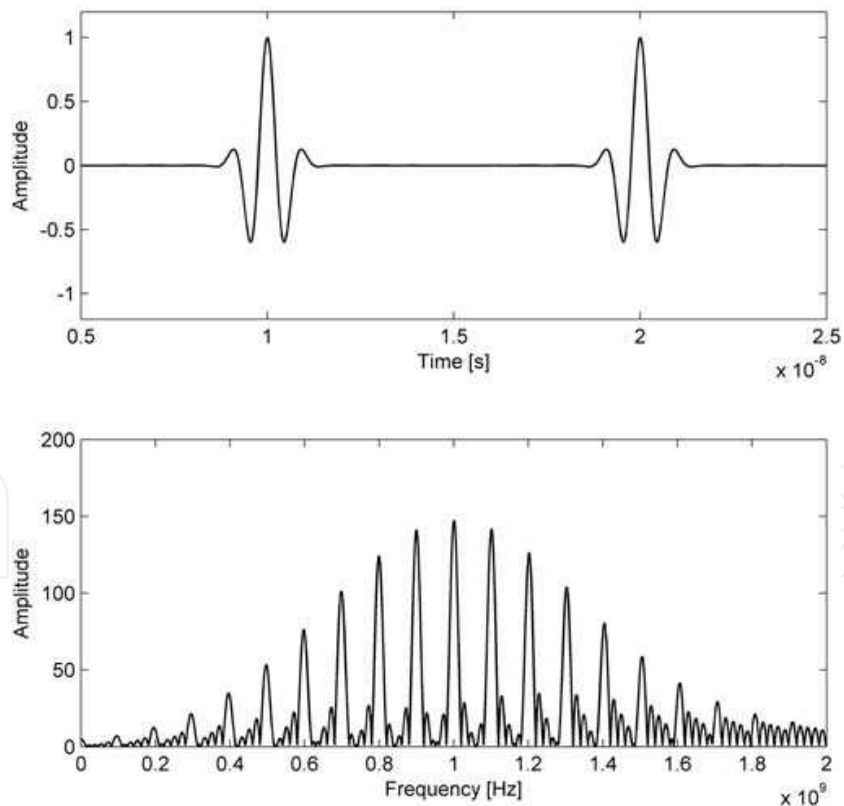


Fig. 3. The figure show a pulse train and its frequency spectrum. The pulse train has a Gaussian pulse with center frequency 1 GHz and a relative bandwidth of 80%.

a pulse and measuring the reflected signal. This method is used in radar and is called a Step Frequency (SF) radar.

All UWB radar waveforms have a way to reduce sampling speed in the receiver so that the full transmitted signal bandwidth does not need to be sampled directly. In impulse radars only one sample per transmitted pulse can generally be taken. The time between the transmitted pulse and the received sample is changed between each pulse and the range profile is then built up. Alternatively, several range cells are sampled with only one bit resolution and several transmitted pulses are needed to get more resolution. In Frequency Modulated Continuous Wave (FMCW) and Step-Frequency radars the frequencies are changed linearly with time or stepped with time respectively. Only the beat frequency between the transmitted signal and the received signal is sampled. In continuation of this section a short description of commonly used UWB radar waveforms will be given.

### 3.1 Impulse radar

In a normal pulsed radar system a local oscillator is switched on and off to generate the transmitted pulse. In impulse radar a short DC pulse is applied directly to the transmitter antenna. The transmitter antenna works as a filter and a band pass signal is radiated from the antenna. The shape and bandwidth of the radiated signal is determined by the antenna transfer function, in other word the parts of the spectrum that the antenna effectively can transmit. A filter may also be put in front of the antenna to shape the spectrum. The received signal needs to be sampled and stored for further treatment. Figure 4 illustrates the sampling process in impulse radar.

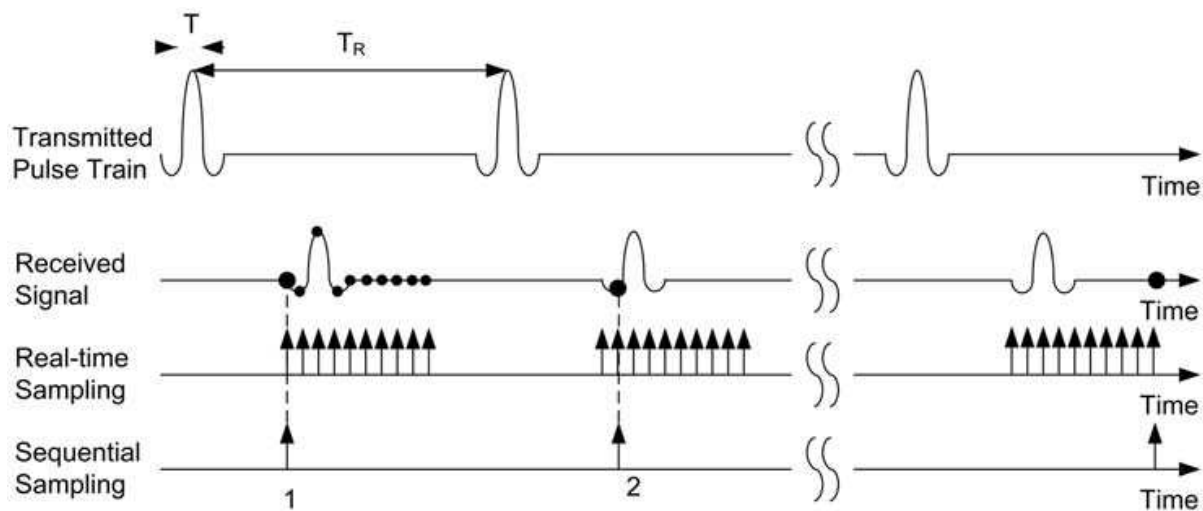


Fig. 4. Illustration of impulse radar with real time or sequential sampling, The transmitted pulse has a pulse width  $T$  and a pulse repetition interval  $T_R$ .

The matched filter for a pulse is a bandpass filter having the same bandwidth as the pulse and sampling the pulse at the maximum amplitude point. This can be done by sampling the received waveform with a sampling frequency of at least twice the highest frequency in the waveform. In impulse GPR they normally sample at least 10 times the center frequency of the transmitted waveform. In UWB impulse radars the highest frequency can be several GHz so that an Analog to Digital Converter (ADC) that can sample at this speed with several bit resolution is very difficult to manufacture and would be expensive and power consuming. This is called real time sampling and is illustrated in Figure 4. A solution to this problem is to take only one receive sample for each transmitted pulse. Then the ADC only needs to take one sample every pulse repetition interval (PRI) so the need for high speed



ADC is avoided. The lower part of Figure 4 illustrates the sequential sampling technique also called transient repetitive sampling or stroboscopic sampling. This technique is currently being used in oscilloscopes for capturing wideband repetitive signals.

Figure 5 show a block diagram of impulse radar with sequential sampling, see (Daniels, 2004). A noise generator is controlling the trigger of the transmitted pulse. This is to randomize the PRI of the pulse train and thereby reduce peaks and smooth the transmitted spectrum. The PRI-generator triggers the impulse transmitter. The trigger signal is delayed a certain amount before being used to trigger the receiver sampler to take one range sample. The delay is shifted before the next sample is taken. In this radar a multi bit high resolution ADC can be used.

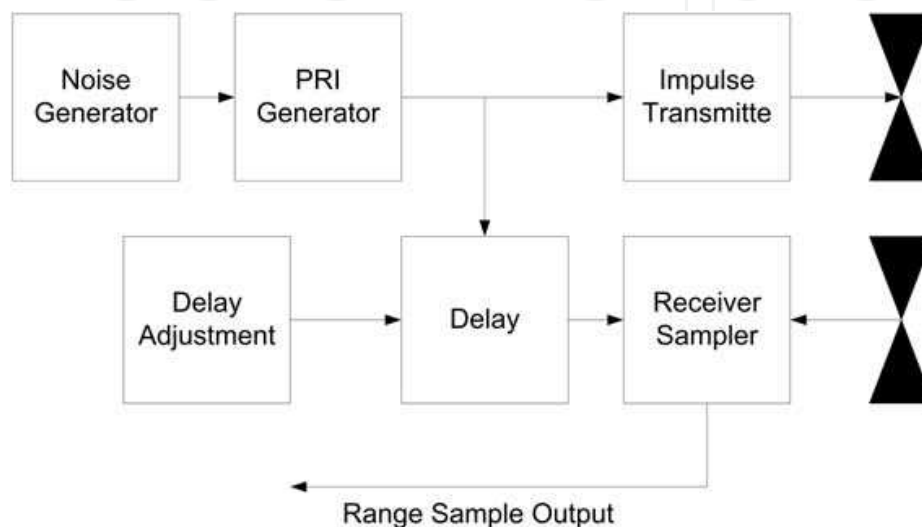


Fig. 5. Impulse radar with sequential sampling.

Another way to capture the reflected signal is to use a single bit ADC and collect several range cells for each transmitted impulse. Figure 6 show a block diagram of this type of impulse receiver. In a practical realization of this receiver one single bit ADC with dither is used and several delays are implemented with a counter on each tap, see (Hjortland et al., 2007). Since the delay is needed only on a single bit signal this is easily implemented using CMOS logic. However several transmitted impulse signals is needed to get sufficient resolution in the received radar profile. Each single bit signal from a range cell is fed to a counter effectively integrating up the single bit signal to yield a higher resolution and thereby higher dynamic range.

In both realizations of impulse radar described above several transmitted pulses are needed to build up one range profile. This means that the receiver is not optimally matched to the transmitted signal and that if the reflector is moving during the profile acquisition it may blur the resulting radar image.

The average transmitted power for an impulse radar with peak power  $P_T$ , pulse length  $T$  and pulse repetition interval  $T_R$  is:

$$\hat{P}_T = P_T \frac{T}{T_R} = PSD \times B. \quad (9)$$

The *SDR*, given in equation 7, for an impulse radar that has a real time sampling of all the receiver range gates and is therefore matched is given as:

$$SDR = \frac{PSB \times B\tau_i G^2}{k_B T_0 F(SNR)} \tag{10}$$

If the impulse radar has a sequential sampling the effective sampling is reduced by the number of range samples in the range profile,  $n_p$ . The radar transmits  $n_p$  pulses before it returns to the same range sample. The  $SDR$  for a sequential sampled impulse radar is given by:

$$SDR = \frac{PSB \times B\tau_i G^2}{n_p k_B T_0 F(SNR)} \tag{11}$$

If the radar has a single bit ADC, one transmitted pulse is needed for each signal level that needs to be discretized. If a 12-bits resolution is needed in the digitized range profile,  $n_b = 2^{12}$  transmitted pulses need to be transmitted before all the levels are measured. This gives the following expression for the  $SDR$ :

$$SDR = \frac{PSB \times B\tau_i G^2}{n_b k_B T_0 F(SNR)} \tag{12}$$

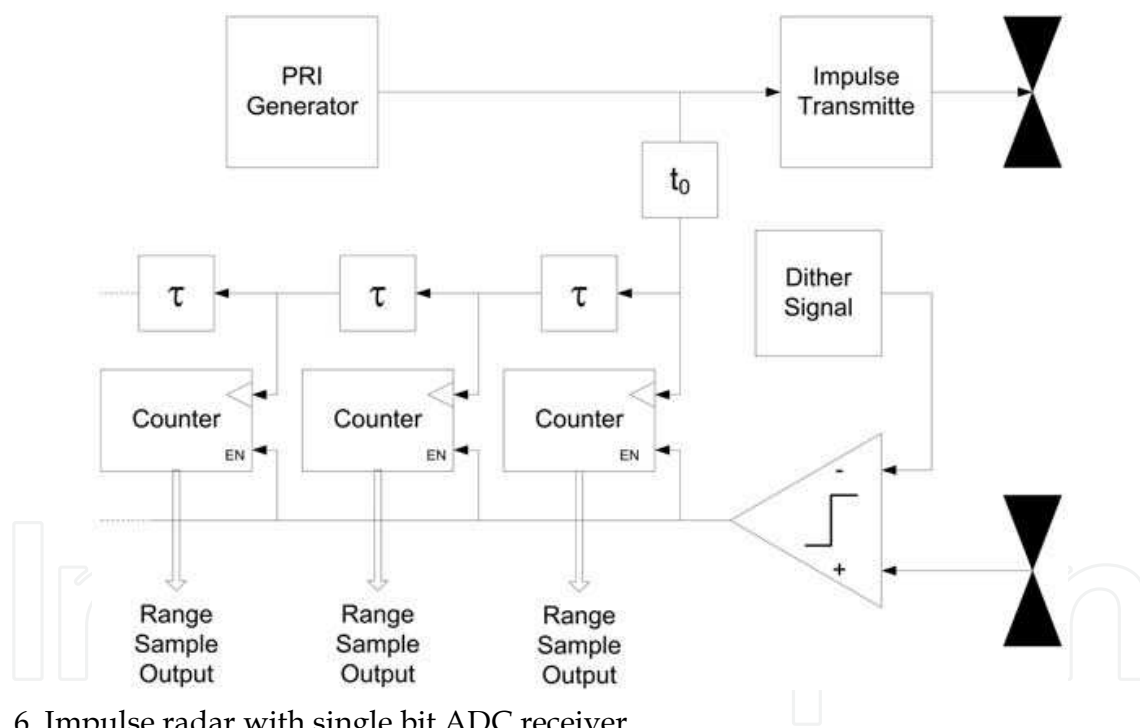


Fig. 6. Impulse radar with single bit ADC receiver

### 3.2 Step frequency radar (SFR)

A pulse train can be synthesized by sending single tone frequencies simultaneously as illustrated in Figure 3. The different sinusoidal frequencies will sum up, and the transmitted signal will be a pulse train. The width of each pulse is determined by the span between the highest and lowest frequency transmitted. The repetition interval is determined by the distance between the transmitted frequencies. A pulse train can also be synthesized by transmitting the different frequencies one at a time and summing them up afterwards. The step frequency radar does this and transmits single tone frequencies in sequence where only

one frequency is transmitted at a time. The phase and amplitude for each frequency is measured of the reflected signal and the radar reflectivity profile is obtained by taking and inverse Fourier transform of the received frequencies.

Figure 7 show a block diagram of a step frequency radar. An oscillator generates the correct frequency that is transmitted by an antenna. The received signal is captured by the receiver antenna and amplified before it is multiplied with the same signal as transmitted. The signal after the mixer is low pass filtered and sampled in an I- and Q-detector. The I- and Q-detector can be implemented after the signal has been sampled. In a homo dyne receiver as shown in Figure 7 the signal oscillator signal need to have a low second and third harmonic content. If this is difficult to achieve a heterodyne receiver should be made.

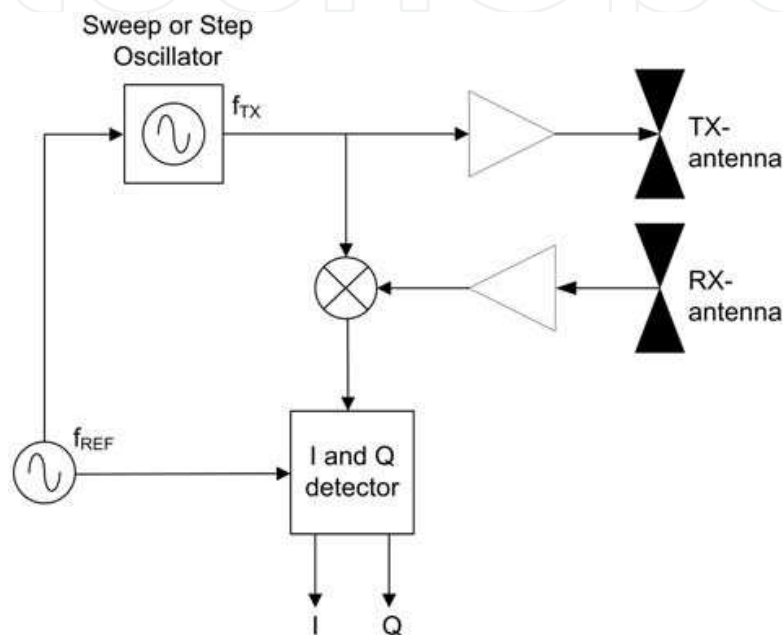


Fig. 7. Block diagram of a step frequency or FMCW radar with homo dyne receiver

When stepping to a new frequency by changing the frequency in the oscillator, it may take some time before the new frequency is stable and can be used. Figure 8 illustrates this where the oscillator needs  $T_{su}$  seconds before it is stable. If the effective integration time is  $T_i$  the reduction in integration time given by  $T_{su}/(T_{su} + T_i)$ . The  $T_{su}$  interval depends on the oscillator. If a Direct Digital Synthesizer - DDS is used the interval is very small and can be neglected. If however a PLL is used in the oscillator it will take some time before it is locked depending on the loop filter values.

The SF signal has a Dwell Time equal  $DT = T_{su} + T_i$  for each frequency and a Scan Time equal  $ST = DTn_f$  where  $n_f$  is the number of transmitted frequencies. If we assume that the dwell time is the inverse of the measurement bandwidth for the PSD, the average transmitted power for the step frequency signal is:

$$\hat{P}_T = P_T = PSD \times mB. \quad (13)$$

If this radar is used as a GPR the average power can be increased according to Equation 2 by a factor given by:

$$\frac{ST}{DT} = \frac{N(1/mB)}{1/mB} = \frac{B}{mB}. \quad (14)$$

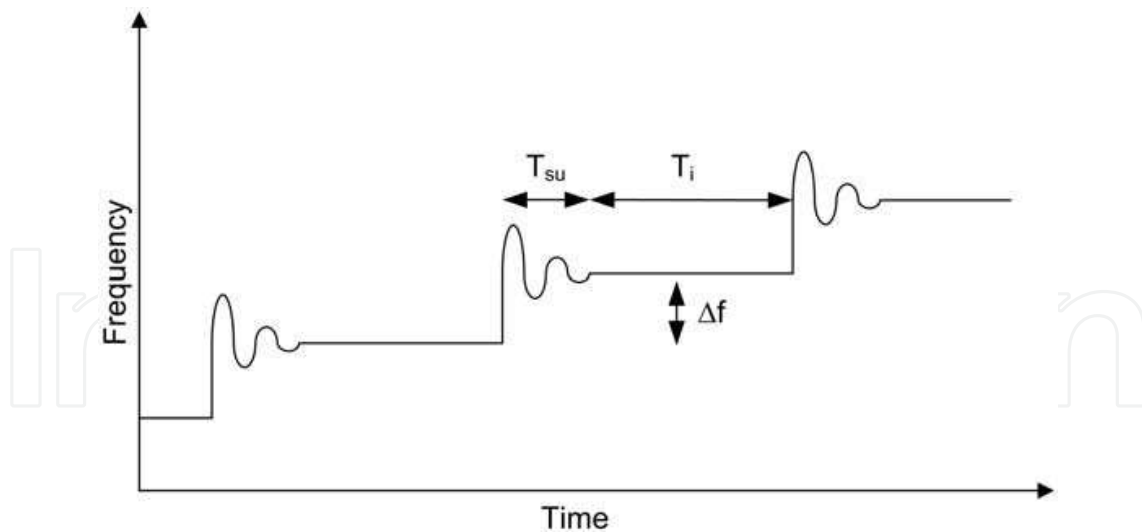


Fig. 8. Changing the frequency in a SF radar system may result in a need to wait until the frequency is stable

The new allowable average transmitted power is:

$$\hat{P}_T = PSD \times mB \frac{B}{mB} = PSD \times B \tag{15}$$

being the same as for the impulse radar in Equation 9. If the correction is not done the SF waveform would have a reduced average transmitted power of  $mB/B$  compared to impulse radar.

The SDR for a SF-radar is:

$$SDR = \frac{PSB \times B\tau_i T_i G^2}{k_B T_0 F (T_{su} + T_i) (SNR)} \tag{16}$$

### 3.3 Frequency modulated continuous wave (FMCW)

Another widely used radar technique in UWB systems is the Frequency-Modulated Continuous-Wave (FMCW) radar. This radar is also collecting the data in the frequency domain as with the SFR technique. In stead of changing the frequency in steps, as with SF-radars, the frequency is changed linearly as a function of time. The schematic block diagram is the same as shown in Figure 9 for the SFR except that the frequency oscillator is sweeping. The received signal, a delayed version of the transmitted signal is multiplied with the transmitted signal. The received signal will due to the propagation delay have a different frequency than the transmitted signal so the signal after the mixer will have a beat frequency proportional to the range of the reflector. The signal after digitization is the same as for the SF-radar and an inverse Fourier transform can be used to get the radar reflectivity profile.

Figure 9 show the frequency as a function of time. The parameters that characterize the sweep is the Sweep Time to sweep the whole bandwidth  $B$ ,  $ST = B/\alpha$  where  $\alpha$  is the sweep frequency rate of change. The Dwell Time is defined at the time the sweep signal stays within the measurement bandwidth  $mB$  and is given as  $DT = mB/\alpha$ .

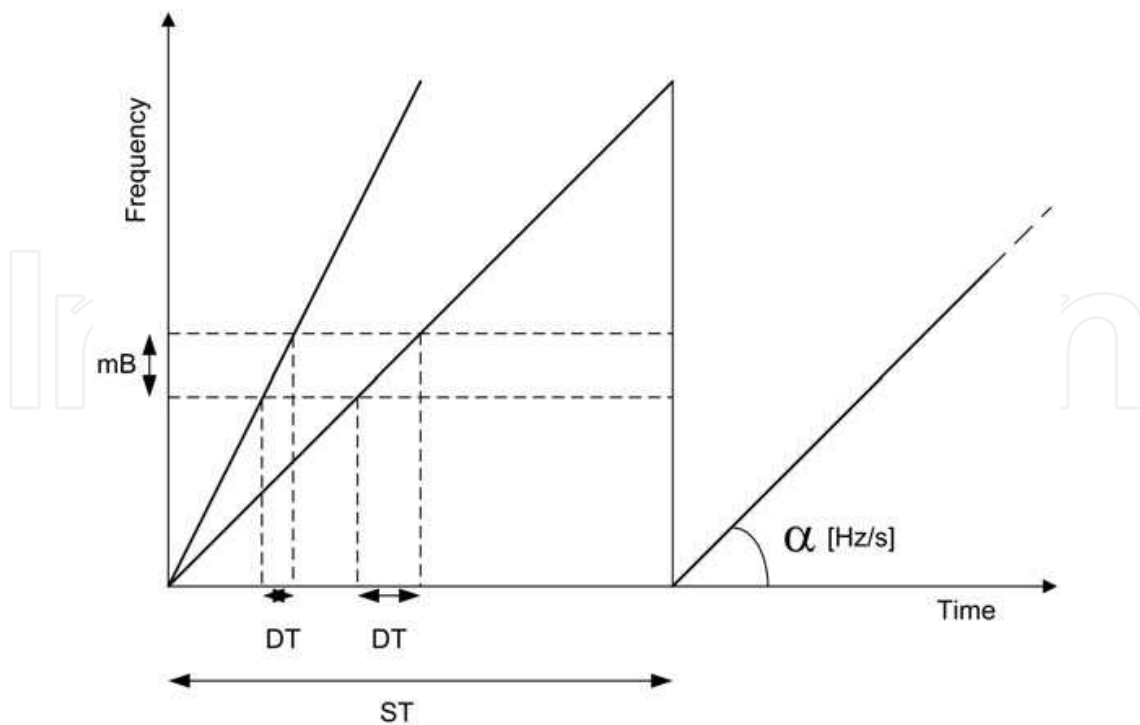


Fig. 9. The dwell time will depend on the sweep

The allowable average transmit power for the sweep is given as the PSD times the inverse of the time the sweep need to cross the  $mB$ :

$$\hat{P}_T = P_T = PSD \frac{1}{DT} = PSD \frac{\alpha}{mB}. \quad (17)$$

If the same correction that can be applied to the SF-radar also can be applied to the FMCW we get the following correction factor:

$$\frac{ST}{DT} = \frac{B/\alpha}{mB/\alpha} = \frac{B}{mB}. \quad (18)$$

The corrected average power then becomes:

$$\hat{P}_T = PSD \frac{\alpha}{mB} \frac{ST}{DT} = PSD \frac{\alpha}{mB} \frac{B}{mB} = PSD \times B \frac{\alpha}{mB^2}. \quad (19)$$

We see that the measurements bandwidth of the PSD and the sweep rate enters into the calculation. If  $mB = 1/DT$  we get  $\hat{P}_T = PSD \times B$ .

The SDR for a FMCW-radar is:

$$SDR = \frac{PSB \times B \alpha \tau_i G^2}{mB^2 k_B T_0 F(SNR)}. \quad (20)$$

### 3.4 Noise radar

A block diagram of a noise radar is shown in Figure 10. A random noise signal is transmitted by the antenna. The transmitted signal is reflected back and enters the receiver after a time given by the two way travel time delay. The received signal is amplified and

correlated with a reference signal that is a delayed version of the transmitted noise signal. When the reference signal is delayed the same amount as the received signal a strong correlation value would be expected.

The delay line in Figure 10 can be implemented in different ways. The most straight forward is a cable with a given length providing the correct delay. A mixer and a low pass filter can be used as a correlator. Another implementation is using a digital RF-memory where the transmitted signal is sampled and delayed digitally before converted to analog signal and correlated with the received signal. A fully digital correlation can be done where both the transmitted signal and the received signal is digitized and stored in a computer. Then the cross-correlation can be done by calculating the cross-spectrum by taking a Fourier transform of the transmitted and received signal. The correlation peak-to-noise ratio for a noise radar is close to the time bandwidth product. For a 40 dB peak-to-noise ratio 10 000 samples must be collected. For an analog correlator this number of new samples must be collected for each range cell. If a digital correlator is used, the same collected samples can be used for all the range cells. Thus the PRF is much higher for a digital correlator than for an analog correlator. The sampling in a noise radar can be done with a single bit ADC, see (Axelsson, 2003).

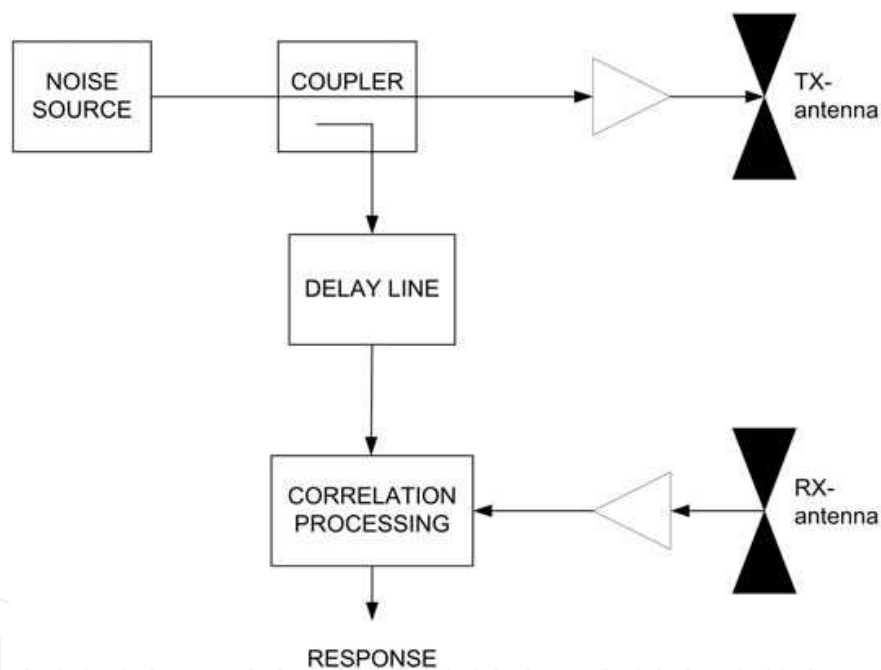


Fig. 10. Block diagram of a noise radar

If the transmitted noise signal is a pseudo random noise generated by a shift register a repetitive signal can be made. When the transmitted signal is repetitive a sequential sampling receiver like for the impulse radar can be made. This technique is described in (Sachs, 2004) and a schematic of the radar is given in Figure 11. The RF-clock generating the pseudo random noise signal is divided to make up the sampling clock. If the clock is divided by 4 it will take 4 pseudo random noise sequences to make up one measurements, see (Sachs, 2004) for more details.

The transmitted average power for a noise radar is:

$$\hat{P}_T = PSD \times B. \tag{21}$$

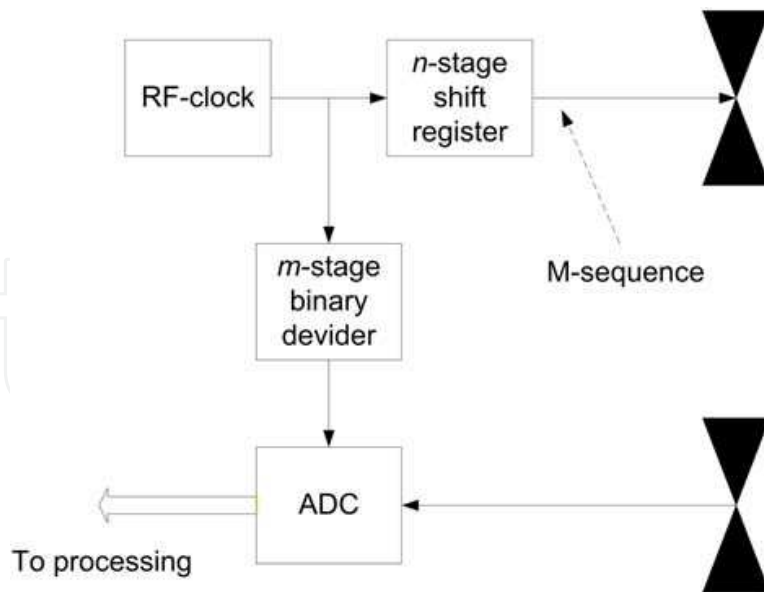


Fig. 11. Block diagram of a noise radar using pseudo random noise and sequential sampling in the receiver

If both the transmitted noise signal and the reflected signal is digitized and sampled the SDR is given by:

$$SDR = \frac{PSD \times B\tau_i G^2}{k_B T_0 F(SNR)}. \quad (22)$$

If pseudo random noise and sequential sampling is used the above equation must be multiplied with a factor  $1/n_m$  where  $n_m$  is the number of pseudo random sequences needed to collect all the data. The SDR becomes:

$$SDR = \frac{PSD \times B\tau_i G^2}{n_m k_B T_0 F(SNR)}. \quad (23)$$

#### 4. Examples on radar performance

A general comparison between the different waveforms is not possible as the SDR depends on the number of range samples needed and the bandwidth of the radar signal. To compare the waveforms a radar with the parameters given in Table 3 will be used. The radar has a bandwidth of 1 GHz and a center frequency of 1 GHz. The range window that should be imaged is 200 ns long. The integration time for collecting one trace is 1 ms. Table 4 gives the calculated SDR for the different waveforms. The impulse real time sampling waveforms has an ADC that can collect all samples in the radar trace for each transmitted impulse. The receiver is matched and the radar system gets the maximum possible SDR = 131.5 dB. The impulse sequential sampling is having an equivalent sampling at 10 times the radar center frequency. This means that it needs  $n_p = 2000$  transmitted pulses to build one radar trace. This gives a reduction of 33 dB compared to the real time sampling. The single bit receiver changes the threshold levels to get 12-bit resolution in the receiver. The radar needs to transmit  $n_b = 2^{12}$  pulses to get one radar trace at 12-bit resolution. This gives a reduction of 36.1 dB compared to the real time sampling, only 3.1 dB lower than the sequential sampling receiver.

Radar parameters	Value
$f_c$	1 GHz
$B$	1 GHz
$\tau_i$	1 ms
$G$	1
$F$	1
$SNR$	1
$T_0$	300 K
Range window	200 ns

Table 3. General radar parameters used for all the radar systems under study

Radar waveform	SDR
Impulse real time sampling	131.5 dB
Impulse sequential sampling	98.5 dB
Impulse single bit sampling	95.4 dB
Step frequency	91.5 dB
Step frequency corrected	121.5 dB
FMCW	101.5 dB
FMCW corrected	131.5 dB
Noise	131.5 dB
Noise sequential sampling	115.5 dB

Table 4. SDR for the different waveforms for the radar parameters given in Table 3

In the step frequency waveform the oscillator set up time is  $T_{su} = 100$  ns and the integration time per frequency is  $T_i = 900$  ns. The dwell time is  $DT = 1/mB$ . The SDR is calculated to be 91.5 dB that is 40.4 dB lower than impulse real time performance. This is due to the much lower allowable transmitted average power given by Equation 13. If the correction factor in Equation 2 is applied the SDR is 121.5 dB, still 10 dB lower than impulse real time sampling. If a DDS is used as a oscillator with practically no set up time between the frequency steps the, SDR would increase and be the same as for the impulse real time sampling.

The sweep in the FMCW is adjusted so that  $DT = 1/mB = 1\mu s$ . This gives  $\alpha = 1\text{MHz}/\mu s$  and one sweep takes 1 ms. The SDR for the FMCW is 101.5 dB and if corrected by Equation 13 becomes 131.5 dB.

A noise radar that collects all the bandwidth with a real sampling techniques gets a SDR of 131.5 dB If a sequential sampling where 40 sequences are needed to get one radar trace the performance is reduced by 16 dB.

The SDR numbers given in Table 4 are depending on several parameters and will change depending on radar center frequency, signal bandwidth and radar range window. Table 4 show that the SDR is varying more than 40 dB for the different waveforms.

## 5. Measuring radar performance

Figure 12 illustrates the different signal levels present in the radar receiver. The signal levels will be referenced to the transmitter signal level  $P_T$  either in dBm or in dB relative to the transmitter power level. The transmitted power level is indicated by a horizontal dashed line. The next level in the receiver is the level where the ADC has reached full scale, ADC-



FS. The ADC-FS should be given by what signal level in dBm at the receiver antenna input is driving the ADC full scale. This level is also indicated by a horizontal line in Figure 12. The next level is the thermal noise,  $N$ , in the receiver. This is the thermal noise at the receiver input given by Equation 4. There is also an internal coupling in the radar system that is illustrated on the left side as a sloping dashed line. The internal coupling level will decrease with time in the receiver time window. The internal coupling may be characterized by the maximum signal level in the receiver and the slope given in  $[dB/ns]$ . The coupling ring down depends among other things on the oscillator phase noise and the receiver IF-mixer. The different levels indicated in Figure 12 can be measured by some simple measurements.

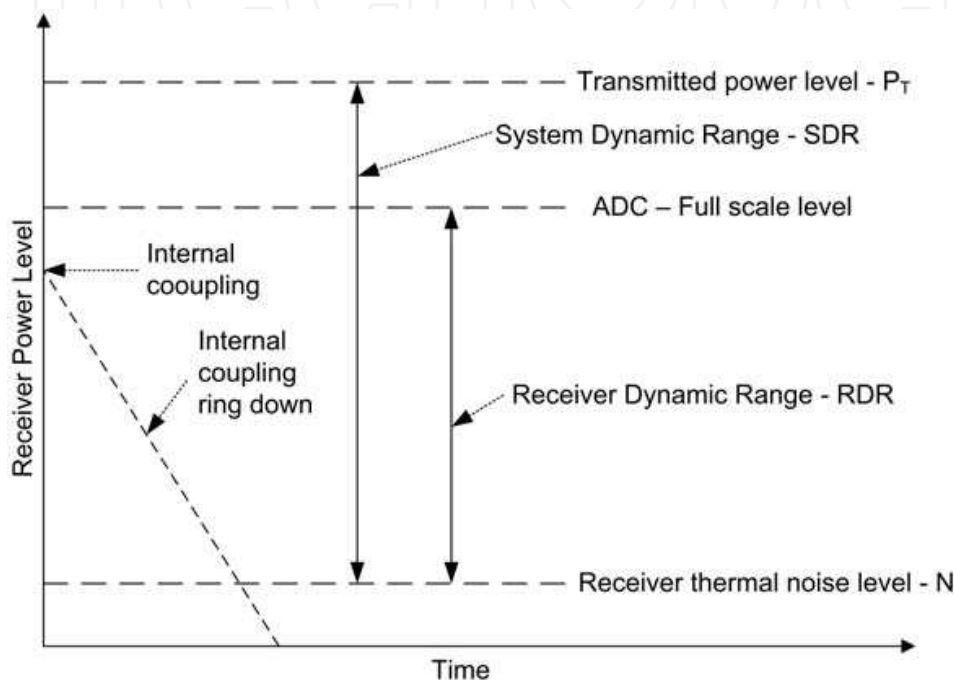


Fig. 12. Illustration of the different signal levels in the radar receiver

### 5.1 Transmitter power

UWB radars use wide band microwave amplifiers that do not necessarily have the same gain for all the frequencies. Normally the gain decrease as a function of frequency. If the gain change by 2 dB over the used bandwidth for each amplifier and the radar have 5 such gain stages in the transmitter chain the transmitted spectrum will change by 10 dB over the band. The transmitted spectrum should be measured using a calibrated spectrum analyzer.

### 5.2 Receiver gain

The receiver level should be calibrated relative to the transmitted power measured in 5.1. If the transmitter power is known in dBm the receiver can be calibrated in dBm. This can be done by connecting a cable with a known attenuator as shown in Figure 13. The cable will give a signal in the receiver with a level given as  $P_T - Att$ . The cable signal level should be much higher than the thermal noise level to give a high signal-to-noise ratio. If the cable and attenuator are frequency independent the receiver gain can be measured. Measuring the received signal from the cable and subtracting the input signal to the receiver given by  $P_T - Att$  the receiver gain can be calculated as a function of frequency. The gain is the total receiver gain from the receiver antenna input to the ADC.

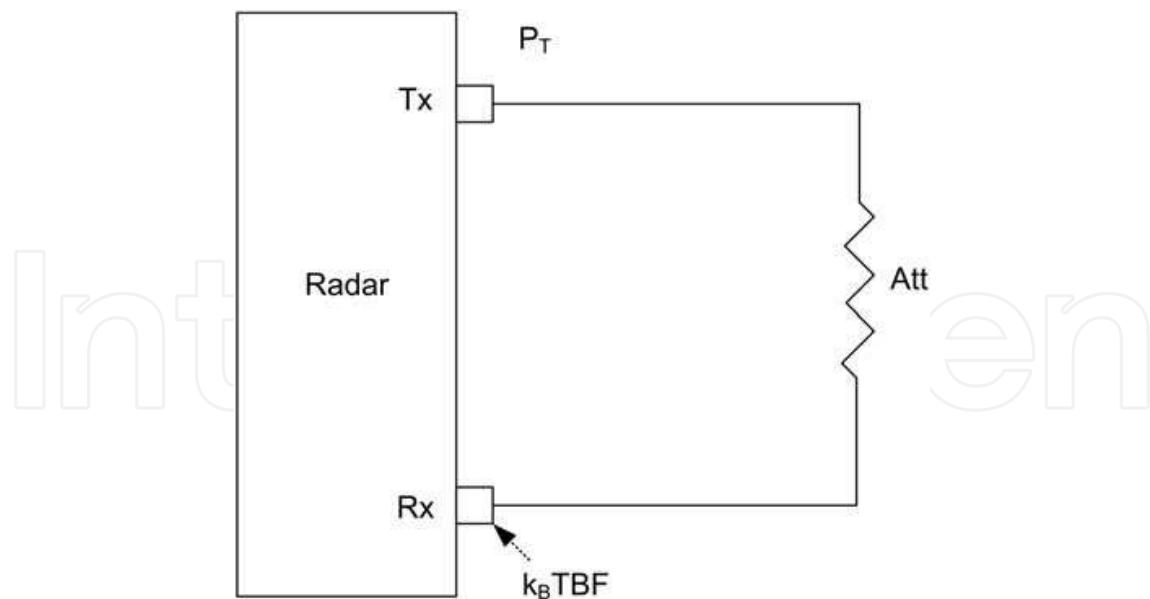


Fig. 13. Measuring radar system performance. The transmitter output is connected to the receiver input via a cable and an attenuator

### 5.3 Internal coupling

The internal coupling should be measured by terminating both the transmitter output and the receiver input to their characteristic impedances. Then a normal radar measurement should be done. The coupling should be characterized by its maximum signal level given in dBm and its slope given in  $[dB/ns]$ . The internal coupling should be as low as possible and the slope as steep as possible.

### 5.4 Receiver noise

The measurement above can also be used to measure the receiver noise level. The internal coupling will slope downward as a function of travel time in the radar receiver window. At a certain time the signal will level out and become horizontal. The level of this horizontal part in the receiver is the receiver noise level. Since the receiver is calibrated in dBm the noise level can be measured in dBm and should be compared with the theoretical value given by equation 4. The receiver thermal noise level will depend on the integration time of the radar system.

### 5.5 System Dynamic Range - SDR

The difference between the measured transmitted power and the measured receiver noise is the System Dynamic Range. The SDR gives a measure for how much loss the radar wave can have and still be detectable in the radar receiver. This is a key measure for the performance of the radar system. The SDR is dependent on the radar integration time. If the integration time is made 10 times longer then the SDR will increase by 10 dB.

### 5.6 Receiver Dynamic Range - RDR

The difference between the ADC full scale value and the noise level is called the Receiver Dynamic Range. This value tells us what signal difference there can be between two reflected signals and that both can be imaged in the receiver window. Increasing the integration increases the RDR.

### 5.7 Spurious free receiver dynamic range

Increasing the integration will increase the RDR and eventually lead to spurious signals to show up in the receiver window. The difference between the ADC-FS and the peak of the spurious signals will give the spurious free receiver dynamic range. This tells us how large the RDR can be made without having unwanted signal in the receiver window. If the spurious signals are stable and does not change from measurement to measurement they can in principle be removed by subtracting an average of the measurements.

## 6. Conclusion

UWB-waveforms are very often used in radar system that are looking through structures like the ground or walls. The propagation of the waves through these structures may attenuate the waves considerably. For the radar system to be successful in these applications it must have a high SDR. This chapter have developed equations for some popular UWB waveforms so that the SDR can be estimated based on parameters for the different waveforms. The SDR varies more than 40 dB between the different UWB-waveforms under study here. The achieved radar performance for a given system can be measured and a simple way to do this has been explained.

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