Chapter 2

Synthetic Aperture Radar

2.1 Introduction

Synthetic aperture radar (SAR) began with an observation by Carl Wiley (Hein, 2004) in 1951 that a radar beam oriented obliquely to the radar platform velocity vector will receive signals having frequencies offset from the radar carrier frequency due to the Doppler Effect. SAR is a type of radar in which a large, highly directional rotating antenna used by a conventional radar is replaced with many low-directivity small stationary antennas scattered over some area near or around the target area. The many echo waveforms received at the different antenna positions are post-processed to resolve the target. SAR can only be implemented by moving one or more antennas over relatively immobile targets, by placing multiple stationary antennas over a relatively large area, or combinations thereof.

In a typical SAR application, a single radar antenna is attached to the side of an aircraft for airborne operation or aboard a satellite for spaceborne operation. A single pulse from the antenna will be rather broad because diffraction requires a large antenna to produce a narrow beam. The pulse will also be broad in the vertical direction; often it will illuminate the terrain from the directly beneath the aircraft out to the horizon. If the terrain is approximately flat, the time at which echoes return allows points at different distance to be distinguished. Distinguishing points along the
track of the aircraft is difficult with a small antenna. However, if the amplitude (giving information about the surface) and the phase of the signal (giving the distance to object on the surface) returning from a given piece of the ground are recorded, and if the aircraft emits a series of pulses as it travels, then the results from these pulses can be combined, thus making the raw data complex (amplitude plus phase) signal. An appropriate coherent combination of the several pulses leads to the formation of a synthetically enlarged antenna - the so-called ‘Synthetic Aperture’. Maximum synthetic aperture size is the maximum distance traveled while target is illuminated.

The precision of a radar instrument is characterized by its spatial resolution (Tomiyasu, 1978). This resolution determines the minimum separation of two point targets being sensed that can be distinguished as separate by the system. To obtain fine resolution in the range direction, the classical technique of pulse compression (Cook, 1960) is typically used. In real aperture radar, the azimuth resolution is proportional to $\frac{\lambda_R}{A_L}$ where $\lambda_R$ is the wavelength of the imaging beam and $A_L$ is the antenna length. For imaging radar with a wavelength in the gigahertz (GHz) range, an
antenna length of 1000 times larger than those used for terahertz (THz) optical frequencies is necessary to achieve the same resolution.

2.2 Side-looking Radar

2.2.1 Scanning Configuration

To image a surface below, the side-looking real-aperture radar (SLRAR) is carried on a moving platform at speed $V_s$ relative to the surface in a straight line at a constant altitude $H$, as shown in fig. 2.2.

Figure 2.2: Scanning configuration of a space borne SLRAR
From this moving platform, the radar is directed with a squint angle $\theta_s$ relative to the flight path and downwards to the surface below with a look angle $\theta_L$, relative to vertical. The most common squint configuration is to position the radar beam to be directed perpendicular to the flight path, making $\theta_s=0$. Using this geometry, the look angle $\theta_L$ is the same as the incidence angle $\theta_i$, which is the angle between the radar beam and the normal to the Earth’s surface at the point of interest.

The side-looking radar images a surface in two dimensions. The dimension perpendicular to the flight path of the radar is called the **range dimension**, and the dimension parallel to the flight path is called the **cross-range or azimuth dimension**. The radar beam is wide in the vertical direction and so intersects the surface in an oval with the long axis extended in the range direction. The spread of the radar pulse in the azimuth direction, or the horizontal beam width $\theta_a$ of the radar, is determined by the antenna length $A_L$ and the radar wavelength $\lambda_R$ as

$$\theta_a = \frac{\lambda_R}{A_L} \quad (2.2.1)$$

The spread of the radar pulse in the range direction, or vertical beam width $\theta_r$ of the radar, is determined by the antenna width $A_H$ and the radar wavelength as

$$\theta_r = \frac{\lambda_R}{A_H} \quad (2.2.2)$$

The range extent of the radar beam footprint on the ground $W_g$, called the ground swath width, and azimuth extent of the beam footprint $W_a$ can be approximated as

$$W_g \approx \frac{\lambda_R H}{A_H \cos^2 \theta_L} \quad (2.2.3)$$

and

$$W_a \approx \frac{\lambda_R H}{A_L \cos \theta_L} \quad (2.2.4)$$

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The side-looking radar is an active system, and can determine the distance from the radar antenna to targets on the surface below by transmitting pulses. Because the radar is looking at the target downwards at an angle $\theta_L$, the distances between objects are measured in a direction which is at a slant to the ground and are said to be recorded in slant range. In imaging applications, the separation in slant range is not as desirable as the real separation on the ground, known as the ground range, and can be obtained by dividing the slant range by the sine of the look angle. In the same manner, the ground range swath width $W_g$ can be found from the slant range swath width $W_r$ by

$$W_g = \frac{W_r}{\sin \theta_L}$$

(2.2.5)

A diagram of the slant range geometry is shown in Fig. 2.3.

Figure 2.3: Slant range geometry of a side looking radar

The radar geometry results in a radar beam footprint of length $W_g$ in ground range. This geometry also defines the near range $R_n$ and far range $R_f$ which are the slant range to a target closest to the radar and farthest from the radar in the beam footprint.
Figure 2.4: Basic principle of strip-mapping SAR

When the radar is moving in space, it transmits a train of radar pulses at intervals determined by $f_{prf}$, the pulse repetition frequency (PRF). For each of these pulses radiated by the antenna, a pulse echo is received which provides information about the surface hit by the beam footprint. Between successive pulses, the radar platform moves a distance of $V_s / f_{prf}$, which is often much less than the azimuth extent of the beam footprint $W_a$. Because of this small platform displacement, the point targets in the radar beam footprint are hit by many different pulses from the radar. This type of radar imaging is known as strip-map imaging (Munson and Visentin, 1989) and is illustrated in Fig.2.4. Determining the appropriate PRF is a function of the geometry of the radar as after the radar pulse is radiated by the antenna, it goes into a listening mode to receive the echo. To avoid any echo overlap, the head of an echo from a
point target in the near range \( R_n \) must arrive after the tail of the previous echo from a point target in the far range \( R_f \) separated by an inter-pulse period \( t_{ipp} \). These times can be expressed in terms of distances as

\[
\frac{2R_f}{V_c} < \frac{2R_n}{V_c} + t_{ipp} + \tau_p \quad (2.2.6)
\]

SAR processing is then achieved by summing all the radar returns to form an azimuth beam which is governed by the length of the synthesized array and not the length of the physical antenna. Where \( V_c \) is the propagation speed of the radar pulse and \( \tau_p \) is the time duration of the radar pulse. Since the slant range swath width \( W_s \) is by definition the difference of the near range and the far range, the equation 2.2.6 can be written as

\[
W_s = R_f - R_n < \frac{V_c t_{prf}}{2} \quad (2.2.7)
\]

Where \( t_{prf} = t_{ipp} + \tau_p \). Equation 2.2.6 can then be combined with Eq. 2.2.3 to give \( t_{prf} \) in terms of the ground range swath width as

\[
\frac{1}{t_{prf}} \leq \frac{W_s V_c}{2 \sin \theta_L} \quad (2.2.8)
\]

Since the PRF \( f_{prf} \) is just the reciprocal of the \( t_{prf} \), Eq. 2.2.8 can then be combined with Eq. 2.2.2 to give \( t_{prf} \) as a function of the radar geometry

\[
f_{prf} \leq \frac{1}{f_{prf}} = \frac{A_H \cos \theta_L V_c}{2 \lambda g H \sin \theta_L} = \frac{A_H f_R}{2 H \tan \theta_L} \quad (2.2.9)
\]
2.2.2 Radar Range Resolution

To detect the distance to an object, a pulse is generated from the transmitter aboard the radar platform, and radiated by the antenna as shown in Fig. 2.5.

![Basic radar block diagram](image)

The greatly attenuated radar echo reflected from the target is then sensed by the receiver. The distance to the target, or range, is calculated using the round-trip time of the radar pulse $t_R$, and the propagation speed $V_c$ of the electromagnetic wave, where $V_c$ is typically taken to be $c$, the speed of light in a vacuum. The slant range to the target, which is the distance along which the radar signals propagate, is then given by

$$R_s = \frac{V_c t_R}{2} \quad (2.2.10)$$

The slant range resolution $\delta_R$ of the radar defines the minimum slant range separation of two points that can be distinguished as separate by the system. Intuitively, we can see from Eq. 2.2.10, that if the head of an echo pulse arrives at the antenna at time $\tau > 0$ after the tail of the previous echo pulse, then they can be distinguished by the system as separate. As an example we consider the system shown in Fig. 2.6 and the resulting received signal shown in Fig. 2.7.
Figure 2.6: Propagation of a radar pulse and separation of the echoes (a) The radar radiates a pulse (b) The pulse hits then target A and an echo is reflected (c) An echo is reflected from the target B and the pulse interacts with the targets C and D (d) An echo from targets C and D is reflected.
The radar radiates a pulse of duration $\tau_p$ towards the surface at $t = t_0$. At $t = t_1$, the radar pulse hits the point target A and an echo is reflected back towards the radar. At $t = t_2$, an echo is reflected from point target B and the radar pulse interacts with point targets C and D. At $t = t_3$, echoes from A and B are in transit and separated by a non-zero time, but the echoes from C and D are not distinguishable in time as the point targets were too close on the ground. The resulting received signal shown in Fig. 2.7 shows that A and B are distinguishable, but there is no clear distinction between where C ends and D begins. The slant range resolution for the case of a single frequency pulse is therefore given by

$$\delta_R = \frac{V \tau_p}{2}$$  \hspace{1cm} (2.2.11)

Where $\tau_p$ is the time duration of the radar pulse.

Equation 2.2.11 suggests that reducing the time duration of the radar pulse will give improved range resolution indefinitely. However, as $\tau_p$ is reduced so is the energy contained within a pulse and this will eventually result in a pulse containing inadequate energy to produce a sufficient echo signal-to-noise ratio for reliable

\hspace{1cm}

Figure 2.7: Resulting received radar signal
detection (Curlander and McDonough, 1991). Increasing the pulse energy is not a complete solution either as the degree to which the radar pulse signal intensity can be increased as it is limited by the radar’s available power and heat dissipation capabilities. To improve the range resolution, a radar pulse compression technique (Cook, 1960) using a linear frequency modulated (FM) radar pulse is typically used where the frequency of the radar pulse changes linearly with time. A plot of a linear FM radar pulse is shown in Fig. 2.8.

![Chirp Signal](image)

The received echoes are passed through match filter which introduces a time lag that decreases linearly with frequency at exactly the same rate as the frequency of the echo increases. Being of progressively higher frequency, the trailing portions of an echo takes less time to pass through match filter than the leading portion. Successive portions then tend to bunch up. Consequently, when the pulse emerges from the filter its amplitude is much greater and its width much less than when it entered. Thus the pulse has compressed. One of the advantages of using a pulse compression technique is that even if the return from points at adjacent range intervals overlap in time, the shape of the pulse is distinctive enough for signal analysis to enable the components of the superimposed signals to be resolved using a matched filter (Curlander and McDonough, 1991; Olmsted, 1993). This technique also permits the use of an extended pulse at lower intensity, with lower power requirements, which will still emit enough energy to give a detectable return. The slant range resolution for radar using this pulse compression technique is given by
Where $B_R$ is the frequency bandwidth of the transmitted pulse. It can be further shown that this resolution can be made arbitrarily fine, within practical limits, by increasing the pulse bandwidth (Curlander and McDonough, 1991). The ground range resolution, which is the resolution along the surface, can be obtained by dividing this equation by the sine of the look angle to give

$$\delta_{GPC} = \frac{\delta_{RPC}}{\sin \theta_L} = \frac{V_c}{2B_R \sin \theta_L} \quad (2.2.13)$$

This equation shows that the radar system ground range resolution is determined by the geometry of the radar and the bandwidth of the radar pulse. In fact, all conventional radar systems, be it real aperture or SAR, resolve targets in the range direction in the same way. It is the resolution of targets in the azimuth direction that distinguishes SAR from other radar systems (Curlander and McDonough, 1991)

### 2.2.3 SLRAR Azimuth Resolution

The azimuth resolution of a Side looking Real Aperture Radar (SLRAR) is defined as the minimum separation of two targets in azimuth that can be distinguished as separate by the system. Two such targets on the ground at the same slant range $R_s$ can only be distinguished if they are not both in the radar beam at the same time. The azimuth resolution $\delta_{AR}$ at a slant range $R_s$ is given by

$$\delta_{AR} = R_s \theta_a = \frac{R_s A_R}{A_L} \quad (2.2.14)$$

Equation 2.2.14 is very similar to the resolution in optics, and this equation demonstrates that the azimuth resolution of a SLRAR is proportional to the radar wavelength divided by the antenna length. This equation suggests that the azimuth resolution can be improved by simply increasing the antenna length. Unfortunately, this is not a possibility in all cases due to the mechanical problems involved in
constructing a precise antenna with an $A_L / \lambda_R$ ratio greater than a few hundred (Curlander and McDonough, 1991). For example, if we wanted to obtain the Radarsat-1 azimuth resolution of 25 m using a wavelength of $5.66 \times 10^{-3}$ m at a slant range of 850 km we would require an antenna length of about 2 km. Clearly, deploying an antenna this size in space is problematic at best, and an alternative is necessary.

### 2.3 Synthetic Aperture Radar Geometry

The goal of SAR is to achieve high resolution in both range and azimuth directions using a practically sized antenna. Its ground range resolution is determined by Eq. 2.2.13 where as the azimuth resolution is determined by the signal processing of the radar echoes. To better understand the SAR geometry; let us consider the radar echoes for an isolated point target as shown in Fig. 2.9.

![Figure 2.9: Radar target geometry](image)

Figure 2.9: Radar target geometry
When the radar, travelling parallel to the point target with velocity $V_s$, reaches $x_1$, the point target is illuminated by the radar beam. This point target continues to be illuminated until the radar reach $x_2$, travelling a distance $L$. The slant range $R_s$ to the point target is given by

$$R_s \approx \sqrt{R_0^2 + (x - x_0)^2} \quad (2.3.1)$$

Where $R_0$ is the nearest range to the target, $x$ is the position of the radar platform, and $x_0$ is the position of the radar platform when the range is $R_0$. The change in slant range $\Delta R_s$ of the target at any radar position $x$ is then approximately

$$\Delta R_s \approx \frac{(x - x_0)^2}{2R_0} \quad (2.3.2)$$

The phase change $\Phi(x)$ corresponding to a change in the slant range $\Delta R(x)$ as a function of radar position $x$ is given by the quadratic function

$$\Delta \Phi(x) = \frac{2\pi(x - x_0)^2}{\lambda R_0} \quad (2.3.3)$$

Assuming a constant velocity $V_s$ for the radar platform and substituting the displacement $(x-x_0)$ with $V_s(t-t_0)$, the phase change as a function of the time is

$$\Delta \Phi(t) = \frac{V_s^2(t - t_0)^2}{\lambda R_0} \quad (2.3.4)$$

Where the time variable $t$ corresponding to the time at radar position $x$. As a time rate of change of this equation 2.3.4 causes frequency shift, the Doppler frequency $f_D$, calculated by taking the first derivative of the equation 2.3.4 results,

$$f_D = \frac{2V_s^2(t - t_0)^2}{\lambda R_0} \quad (2.3.5)$$
Therefore, if the received signal is frequency analyzed at time $t_1$, any energy observed in the return at a time $t$ corresponding to a slant range $R_s$ and at Doppler frequency $f_{D1}$ will be associated with a target at coordinate

$$x_1 = \frac{\lambda_R R_s f_{D1}}{2V_s}$$  

(2.3.6)

Similarly, energy at a different Doppler frequency $f_{D2}$ will be assigned to a corresponding coordinate $x_2$. This allows targets to be discriminated even though they are in the radar beam at the same time. With the use of Doppler analysis of radar returns, the azimuth resolution $\delta_A$ is now related to the resolution of the measurement of the Doppler frequency $\delta_{f_D}$. From the equation 2.3.5 the azimuth resolution of the SAR is given as

$$\delta_A = \frac{\lambda_R R_s \delta_{f_D}}{2V_s}$$  

(2.3.7)

Which results

$$\delta_A = \frac{A_L}{2}$$  

(2.3.8)

### 2.3.1 Nature of SAR data

The backscatter returns from a synthetic aperture radar instrument can be modeled as the superposition of many small scatters within the antenna beam footprint (Kwok and Johnson, 1989). The coherent radar returns $A(x, y, z)$ at observation point $(x, y, z)$, which is a complex-valued function of space, is then given by
A(x, y, z) = \sum_{k=1}^{N} |a_k|e^{i\Phi_k} \tag{2.3.9}

Where \(a_k\) is the reflectance amplitude, \(\Phi_k\) is the phase delay, and the sum is over all elementary phasor contributions \(N\) (Goodman, 1975). To illustrate this, Fig. 2.10 shows a time scale waveform of 1000 samples from the European Space Agency Environmental Imaging Satellite – Advance Synthetic Aperture Radar (ESA-ENVISAT-ASAR) test set.

Since the calculation of the signal statistics at point \((x, y, z)\) is the statistical problem of a random walk in the complex plane, the randomly phased phasors can be assumed to have the following assumptions (Goodman, 1975):

1. The amplitude \(a_k\) and phase \(\Phi_k\) of the \(k\)th elementary phasor are statistically independent of each other and of the amplitudes and phases of all other elementary phasors; and
2. The phases \(\Phi_k\) are uniformly distributed on the interval \([-\pi, \pi]\).
Assumption (2) is a result of the scatters having an unknown range and the range resolution of the SAR being much greater than the wavelength of the transmitted SAR signal. This then produces the result that phase excursions of many times $2\pi$ radians produce a uniform distribution on the interval $[-\pi, \pi]$ (Kwok and Johnson, 1989). These two properties allow the determination of the statistical characteristics of the real and imaginary parts, along with the magnitude and phase, of the received SAR signal.

The real and imaginary parts of the received SAR signal are given by

$$I = \text{Real}[A] = \sum_{k=1}^{N} |a_k| \cos \Phi_k$$  

(2.3.10)

$$Q = \text{Imag}[A] = \sum_{k=1}^{N} |a_k| \sin \Phi_k$$  

(2.3.11)

The expected values of the real and imaginary parts of the received SAR signal is given by

$$E[|I|] = \sum_{k=1}^{N} E[|a_k| \cos \Phi_k] = \sum_{k=1}^{N} E[|a_k|] E[\cos \Phi_k] = 0$$  

(2.3.12)

$$E[|Q|] = \sum_{k=1}^{N} E[|a_k| \sin \Phi_k] = \sum_{k=1}^{N} E[|a_k|] E[\sin \Phi_k] = 0$$  

(2.3.13)

Where Assumption (1) allows the calculation of the expected value of $|a_k|$ and $\Phi_k$ separately, and Assumption (2) assures a value of zero for the expected values of both $\cos \Phi_k$ and $\sin \Phi_k$. 

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Proceeding in a similar fashion, the second order statistics of the real and imaginary parts of the received SAR signal are given by

\[
E[|Re\{A\}|^2] = \frac{1}{N} \sum_{k=1}^{N} \sum_{m=1}^{N} E[|a_k||a_m||E|\cos \Phi_k \cos \Phi_m|] \\
= \frac{1}{N} \sum_{k=1}^{N} E[|a_k|^2] = \sigma^2
\]  

(2.3.14)

\[
E[|Im\{A\}|^2] = \frac{1}{N} \sum_{k=1}^{N} \sum_{m=1}^{N} E[|a_k||a_m||E|\sin \Phi_k \sin \Phi_m|] \\
= \frac{1}{N} \sum_{k=1}^{N} E[|a_k|^2] = \sigma^2
\]  

(2.3.15)

\[
E[Re\{A\}Im\{A\}] = \frac{1}{N} \sum_{k,m=1}^{N} E[|a_k||a_m||E|\cos \Phi_k \sin \Phi_m|] \\
= 0
\]  

(2.3.16)

Here, we have used the fact that independent and uniformly distributed phases exhibit the expectations

\[
E[\cos \Phi_k \cos \Phi_m] = E[\sin \Phi_k \sin \Phi_m] = \begin{cases} 
\frac{1}{2} & k = m \\
0 & k \neq m 
\end{cases} 
\]  

(2.3.17)
If we now suppose that the number $N$ of the elementary phasor contributions is very large, the central limit theorem dictates that the real and imaginary parts of the received SAR signal are Gaussian. Accordingly, the probability density function of the real and imaginary parts of the received signal can be modeled as

\[
f_X(x) = \frac{1}{\sigma \sqrt{2\pi}} e^{-\frac{x^2}{2\sigma^2}}
\]

(2.3.18)

\[
f_Y(y) = \frac{1}{\sigma \sqrt{2\pi}} e^{-\frac{y^2}{2\sigma^2}}
\]

(2.3.19)

Thus, the real and imaginary parts of the received SAR signal can be modeled as uncorrelated Gaussian distributions with zero means and identical variances.

The magnitude $M$ and phase $\theta$ of the received SAR signal can be related to the real and imaginary parts of $A$ using the transformation

\[
M = \text{Re} \{A\}^2 + \text{Im} \{A\}^2
\]

(2.3.20)

\[
\theta = \arctan \frac{\text{Im}\{A\}}{\text{Re}\{A\}}
\]

(2.3.21)

This gives the probability density function of the magnitude $M$ as

\[
f_M(M) = \frac{1}{2\sigma^2} e^{-\frac{M}{2\sigma^2}}
\]

(2.3.22)

And the probability density function of the phase $\theta$ as

\[
f_\theta(\theta) = \frac{1}{2\pi}
\]

(2.3.23)
Thus the probability density function of the magnitude $M$ of the received SAR signal can be modeled as Rayleigh distribution and the phase as uniformly distributed on the interval $[-\pi, \pi]$. Also,

$$ f_{M,\theta}(M, \theta) = f_M(M)f_\theta(\theta) \quad (2.3.24) $$

This equation suggests that the magnitude and the phase are statistically independent at any point of time. The magnitude $|A|$ and the phase $\phi$ of the received SAR signal are related to the real and imaginary parts of $A$ by

$$ |A| = \sqrt{I^2 + Q^2} \quad (2.3.25) $$

And

$$ \phi = \arctan \frac{Q}{I} \quad (2.3.26) $$

Thus, $|A|$ is Rayleigh distributed, $\phi$ is uniformly distributed on $[-\pi, \pi]$, and they are independent.

### 2.4 Image Formation and Characteristics

Once the raw SAR data have been received at the ground station, it must be processed to form the SAR image. The processing consists of correlating all Doppler components for ground targets. Originally, this processing was done using optical lenses (Cutrona et al., 1960); however, all current processing is done using numerical computers. Various processing steps designed in MATLAB 7.0 as a part of the SAR processor in this thesis to process the raw data is shown in the flowchart figure 2.11.
The raw signal space SAR input is the two-dimensional signal which is first analyzed as a series range time signals for each azimuth bin. Each range time signal undergoes matched filtering (Pulse Compression) in the range frequency/azimuth time domain through range Fast Fourier Transform (FFTs) applied to the range time signals. After
each signal is transformed back into the range time/azimuth time domain, the result is
the range compressed signal as the matched filtering was performed in the range
frequency domain. In order to obtain azimuth compression, azimuth matched filtering
must be performed. The range compressed signal is then composed into a series of
signals with respect to azimuth time at different range bins. Each azimuth signal is
Fourier transformed via an azimuth FFT. After azimuth matched filtering of each
signal and azimuth inverse fast Fourier transforms (IFFTs), the final target image is
obtained.

Matched filtering is the correlation of a template signal with an unknown signal,
which is the equivalent of convolution of an unknown signal with a time reversed
template, to detect the presence of the template signal in the unknown signal. This
detection is effective even in low signal to noise ratio (SNR) cases. In the Figure 2.12
below, the transmitted radar signal is denoted as $s(t)$ and the received radar signal is
modeled as a time delayed version of $s(t)$. The matched filter template, $h(t)$, is the
time reversed version of $s(t)$ and the convolution of the two produces a compressed
pulse of energy centered around the time delay of radar reflection.

\[
s(t) \quad s(t-t_0) \quad y(t)
\]

\[
t \quad t_0
\]

\[
h(t) = s(-t)
\]

\[
t
\]

\[
y(t)
\]

Figure 2.12: Radar Range finding Match Filter

Instead of correlation in the time domain, multiplication by the complex conjugate in
the Fourier domain, which is an equivalent operation, is performed for speed as it is
equivalent and less processing intensive. Matched filtering in the simulation is termed pulse compression as the energy of the received SAR signal converges to or is compressed to the regions of template signal detection. This process is enhanced by the chirp signal used in the transmitted radar signal construction as there is more information embedded for detection. To improve computational efficiency the FFT is used, which is a radix-2 algorithm for efficient computation of the discrete Fourier transform, (DFT) and its inverse. The radix-2 feature of the FFT constrains the number of processed time samples to be an integer multiple of two. FFTs use block processing through simultaneous computations of different inputs which makes them highly efficient.

2.4.1 SAR Image Characteristics

Since the raw SAR signal is a complex number, one can form either a complex image, image of the real part, the imaginary part image, the phase image, the intensity image or the log intensity image. The intensity image $A = \sqrt{I^2 + Q^2}$ is most commonly used. $|A|$ is Rayleigh distributed and it follows that the intensity is exponentially distributed. Mathematically

$$P_X(x) = \frac{1}{\sigma} e^{-\frac{x}{\sigma}} \text{ for } x \geq 0, \quad 0 \text{ otherwise} \quad (2.4.1)$$

2.5 Data Presentation

The data sets used in this thesis are from European Environmental Remote Sensing Satellite ERS-2, Canadian Radar Satellite RADARSAT-2 and German Space Agency DLR’s TerraSAR-X satellites. Figures 2.13 to 2.15 show histograms of the block of data from one of these test sets showing normal probability distribution.
Figure 2.13: Histogram of the real data from the Terra-SAR data set

Figure 2.14: Histogram of imaginary data from Terra-SAR data set

Figure 2.15: Histogram of intensity image from Terra-SAR
The large probability at the tails of the histogram of real data is attributed to the limited dynamic range of the ADC used to quantize the data. The normal probability plot shows that the block of raw SAR data follow a normal distribution, except at the tails having little bit more deviation.

2.6 Summary

The general SAR systems principle has been presented in this chapter. The range resolution of side-looking radar is shown to be independent of the radar-target distance, and by means of the pulse compression technique, high range resolution can be achieved. High azimuth resolution can also be achieved by SAR processing. However, because of geometric and radiometric distortions, the azimuth processing, unlike the pulse compression, can be very complex. Under Goodman conditions, the statistics of the I and Q components of the raw SAR data are totally uncorrelated to zero mean Gaussian, with identical variances. This model was confirmed on real SAR data sets and any compression technique should be adapted to it as will be seen in the subsequent chapters.