APPLICATION OF SATELLITE-BASED SIDELOOKING RADAR IN MARITIME SURVEILLANCE

BY

EINAR-ARNE HERLAND

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The use of a satellite borne synthetic aperture radar (SAR) system for maritime surveillance is investigated. The system is analyzed, and algorithms for digital processing of SAR images from the raw data are developed. These algorithms are implemented on a general purpose computer and applied to data from SEASAT-1. An autofocusing algorithm for the azimuth compression is developed whereby SAR images can be processed without any additional information on satellite orbit and attitude. This makes operational use of satellite borne SAR feasible. A method for target detection from SAR images is also developed.
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APPLICATION OF SATELLITE-EASED SIDELOOKING RADAR IN MARITIME SURVEILLANCE
(SATELLITTBÆRET RADARSYSTEM MED SYNTETISK APERTURE FOR HAVOVERVÅKNING)

SUMMARY

The use of a satellite borne synthetic aperture radar (SAR) system for maritime surveillance is investigated. The main advantages of a synthetic aperture radar are high resolution combined with all-weather capabilities and independence of visual sight. The two last properties are obtained by using a frequency typically in the area 1 to 10 GHz. The high resolution is due to the application of the synthetic aperture principle. Resolution on the ground is typically 25mx25m, depending upon system parameters. By using a satellite platform large areas on the ground can be covered.

Synthetic aperture radar images can be obtained by rather extensive digital processing of the rawdata gathered with the satellite system. The necessary algorithms to process such images without use of externally supplied satellite orbit information have been developed and are applied to rawdata gathered by the radar system onboard the SEASAT-I satellite. The ability to distinguish ships in such images is demonstrated, and a target detection algorithm is developed.

1 INTRODUCTION

A radar operated in pulsed mode is actually a range measuring instrument. Combined with the angular resolving capability of the radar antenna, it is possible to generate an image of an area by means of radar. In one application the radar antenna rotates and the image is displayed on a plan position indicator (PPI) where the radar platform is located in the center of the image. An imaging radar thus resolves targets in two dimensions.
The resolving properties in range and angle are principally different, since the range resolution is given by the bandwidth of the transmitted pulse, and is thus range independent, while a constant angle resolution gives a resolution in meters that is proportional with range.

The resolving power of an antenna is typically limited by the physical size of the antenna itself. To overcome this difficulty, Carl Wiley of Goodyear Aircraft Corporation in 1951 proposed an improvement in angular resolution by frequency analysis of the target history obtained with a coherent, moving radar system. The first experimental demonstration was made by a group at the University of Illinois in 1953. They used an airborne radar system at X-band and achieved an angular resolution that was approximately ten times better than the resolution of the physical antenna. The first combat-surveillance system employing an airborne synthetic aperture radar was successfully demonstrated by a team from the University of Michigan in 1957. Since then synthetic aperture radar has been developed into a very useful instrument for imaging purposes.

The usual radar platform for synthetic aperture radar purposes has been an airborne one. In the summer 1978 a satellite called SEASAT-1 was launched which, together with 4 other instruments, contained a sideloooking radar designed as a synthetic aperture radar. The purpose of this satellite was to measure global ocean dynamics and physical characteristics and to provide monitoring of ice conditions. It was a proof of concept mission, and SEASAT-1 was planned to be the first in a series of similar satellites, and the purpose was to establish an operational system. Due to equipment malfunctioning the satellite failed in orbit after approximately 100
days of operation, and the launching of subsequent satellites was cancelled. During the time of operation a considerable amount of data was collected with the synthetic aperture radar.

This work is concerned with the use of satellite borne synthetic aperture radar in maritime surveillance. Although the SEASAT-1 SAR system was not designed for this purpose, it is shown to be possible to use such high resolution imagery as can be obtained with this kind of system for surveillance purposes.

To obtain images from the rawdata collected with a synthetic aperture radar requires rather extensive processing. This processing has in the past mainly been done by means of a coherent optical system. This is because the large amount of data requires a processing capacity offered only by optical processing, and also because film is a convenient storage medium for the rawdata, thus lending itself naturally to optical processing.

With the development of digital computers it is now possible to do real time digital processing for air-borne synthetic aperture radar, while for a satellite borne platform this has not yet been demonstrated.

In this work the necessary algorithms for processing, on a general purpose computer, synthetic aperture radar images from rawdata gathered by a satellite borne radar are developed and applied to rawdata from SEASAT-1. The numerical examples given also use parameters from the SEASAT-1 SAR system. Of new results, beside the SAR image processor, which has been developed independently of a few digital processors, are the autofocus method of target detection, where a tradeoff between resolution and look averaging
is made to maximize the signal to sea clutter ratio of a target. The autofocusing procedure is very important in operational use of a SAR system, since the time delay incurred when using externally supplied orbit information for the satellite may exclude operational use of the images.
2 BASIC THEORY

The theory for conventional antennas, both continuous and array antennas, can be found in (1). Here will only be given a short outline of a couple of things that are important to the understanding of synthetic aperture antennas.

2.1 Array antenna

Figure 2.1 shows the antenna diagram for an array antenna of seven equally spaced elements, element length \( \frac{D}{2} \) and element spacing \( D \).

![Antenna diagram for array antenna](image)

Figure 2.1 Antenna diagram for array antenna

The envelope of the diagram is the antenna diagram of a single element and the major peaks, or grating lobes, that repeat in angle, stem from the fact that it is an array antenna. Since a synthetic antenna is
an array antenna, it has a diagram of the same type as in figure 2.1. The important thing about this diagram is the grating lobes, which for a synthetic aperture radar would result in a periodic repetition in azimuth of every target in the image. When \( D > \lambda \), the transmitted wavelength, the grating lobes appearing in the main lobe of the envelope can be avoided by choosing the element length greater than the inter-element spacing. The corresponding requirement for a synthetic array antenna is somewhat different and is given in section 3.2.

An antenna can also be regarded as a spatial filter, a viewpoint which is useful when studying the analogy between synthetic aperture radar and holography.

2.2 Synthetic aperture

In a conventional array antenna all the antenna elements are present simultaneously. In a synthetic aperture antenna there is only one antenna element. This element is moved in space to different positions, and in each position it transmits and receives an electromagnetic signal. By combining the received signals from different element positions, a synthetic array antenna is generated. The geometry is shown in figure 2.2.

A series of different positions of the basic antenna is shown, together with the antenna beam. A target is shown, and it is seen that it will be inside the physical antenna beam for a certain time interval. This time interval then gives the longest possible synthetic aperture for the given target. Quantitative relationships will be given in section 3.2. For each antenna position the pulse echo is coherently recorded, i.e., the echo together with a phase reference is stored, and it
Figure 2.2  **Principle for synthetic array antenna**

is this phase reference that makes it possible to synthesize an antenna much longer than the physical one.

One difference between a physical and a synthetic array antenna is that in a physical array antenna all the elements receive the echo from the target of a wave transmitted by all the array elements or by a common transmitting antenna, while in the synthetic array each element receives only the echo of the wave transmitted by itself. As will be shown later, this gives the synthetic array two times the phase sensitivity of the corresponding physical array, and the angular resolution for a synthetic array of length D will be the same as the resolution for a physical array of length 2D.
For a target at distance $R$ from the antenna, the longest possible synthetic aperture is $R \cdot \theta$, where $\theta$ is the beamwidth of the physical antenna. The ground resolution for the synthetic antenna is then

$$\Delta_s = \frac{\lambda}{2} \frac{1}{R \cdot \theta} \cdot R = \frac{\lambda}{2 \theta}$$

If $D$ is the length of the physical antenna, we have

$$\theta = \frac{\lambda}{D} \Rightarrow \Delta_s = \frac{D}{2}$$

This shows that the ground resolution of a synthetic aperture antenna is independent of range and transmitted wavelength and is equal to half the length of the physical antenna. The wavelength independence is valid only as long as $D \gg \lambda$.

2.3 Focused and unfocused antenna

In the previous discussions it has been assumed that the target is in the far field of the antenna. The far field extends to infinity, and between the far field, or Fraunhofer region, and the antenna is the Fresnel region. The boundary between the Fraunhofer region and the Fresnel region is rather diffuse, but it can qualitatively be said that a point target is in the far field when a spherical wave expanding from the target can be regarded as plane across the antenna aperture. This is illustrated in figure 2.3.

Whether an incoming wavefront can be regarded as plane or not, depends upon the phase difference across the
Figure 2.3 Phase variation across antenna aperture

antenna aperture. This phase difference is given by the distance \( \delta \) in figure 2.3. We have

\[
R = R_0 + \delta
\]

and

\[
R = \sqrt{R_0^2 + \frac{D^2}{4}} \approx R_0 + \frac{D^2}{8R_0} \quad \text{when} \quad R_0 \gg D
\]

This gives \( \delta = \frac{D^2}{8R_0} \). The corresponding phase difference is

\[
\Delta \phi = \delta \cdot k = \delta \cdot \frac{2\pi}{\lambda} = \frac{\pi D^2}{4\lambda R_0}
\]

where \( k = \frac{2\pi}{\lambda} \) is the wave number of the incoming wavefront. If we require the phase difference across the antenna aperture to be less than \( \frac{\pi}{4} \) to regard the wavefront as plane, we have

\[
\frac{\pi D^2}{4\lambda R_0} < \frac{\pi}{4} \Rightarrow R_0 > \frac{D^2}{\lambda}
\]
This means that a target is said to be in the far field of the antenna when the distance between the antenna and the target is greater than \( \frac{D^2}{\lambda} \). When the target is closer to the antenna, it is said to be in the Fresnel region. When a target is moved into the Fresnel region of an antenna, there is seen to be destructive interference along the antenna because of the phase difference across the antenna aperture. To restore the constructive interference, the antenna will have to be formed to fit the wavefront, or in other words one will have to focus the antenna. For a target lying in the focus of the antenna, the beamwidth is given by \( \frac{\lambda}{D} \) as before. An unfocused antenna can be thought of as focused at infinity, and all the targets in the far field can thus be regarded as being at infinity. To find the actual ground resolution, however, one has to take into account the actual distance to the target, since the ground resolution for an antenna at a distance \( R \) is \( R \cdot \frac{\lambda}{D} \).

Focusing can be done in two ways. The antenna can be physically formed to fit the wavefront. One physical configuration will, however, place the focus in a given distance from the antenna, and this method is therefore impractical when one wants to focus at different distances simultaneously. Another method is to record separately the returns from the different positions of the antenna and subsequently focus by signal processing. In this way one can easily focus at all the distances one might want. Since this kind of data recording is inherent in the synthetic aperture principle, this method is used when generating a focused synthetic aperture. The processing can be done either optically or digitally, as will be explained later.

The high processing capacity needed for a synthetic aperture radar compared to a real aperture radar is due to the focusing, and not to the principle of
generating a synthetic aperture. Even if a physical antenna with the same length as the synthetically generated one should be available, it would require data recording and processing of the same complexity as with synthetic aperture because the physical antenna can only be focused on one distance at a time.
3 SYNTHETIC APERTURE RADAR PERFORMANCE

In this chapter the performance of a synthetic aperture radar system will be quantitatively evaluated and some particular phenomena encountered with SAR will be considered. General descriptions of synthetic aperture radar can be found in (2) and (3). A satellite borne system will be considered, a sketch of which is shown in figure 3.1.

![Figure 3.1 Geometry of satellite borne SAR](image-url)
Approximate values for SEASAT-1 are as follows:

\[ h = 800 \text{ km} \]
\[ R = 850 \text{ km} \]
\[ a = 240 \text{ km} \]
\[ \ell = 100 \text{ km} \]
\[ \theta = 20.5^\circ \]
\[ v = 7 \text{ km/sec} \]

Antenna beamwidth in azimuth, 1°
Antenna beamwidth in elevation, 6°

A general description of and detailed parameters for the SEASAT-1 SAR system can be found in (4).

In the following range resolution will be treated first and then azimuth resolution.

3.1 Range resolution

Resolution in range with a synthetic aperture radar is obtained in conventional manner. Although a focused system has an inherent range resolution given by the depth of focus, this resolution is highly inferior, as will be shown later. The range resolution is therefore obtained by pulsed operation of the radar. This is shown in figure 3.2 where the flight direction is into the paper.

If the transmitted pulse bandwidth is \( B \), then the resolution in radial direction from the antenna is given by

\[ \Delta R = \frac{C}{2B} \]
where \( c \) is the velocity of light. This formula is valid also when pulse compression techniques are used (5). The corresponding ground resolution is

\[
AR_g = \frac{AR}{\sin \theta}
\]

For SEASAT-1 a linear FM-pulse is transmitted, which gives true range 3 dB resolution \( AR = 7.1 \) m, where true range is the distance between antenna and target. Ground range is the distance from the nadir line to the target measured on the ground. For SEASAT-1 we have approximately \( 17^\circ < \theta < 24^\circ \), which gives the relationship shown in figure 3.3 between \( R_g \) and \( \theta \). The lower abscissa axis shows the corresponding ground range when the satellite's height \( h = 800 \) km.

From figure 3.3 it is seen that a processed SAR-image that is linear in true range will be distorted in such a way that targets on the ground at near range will be compressed relative to ground targets at far range.
Figure 3.3  Ground resolution as a function of ground range

For SEASAT-1 the transmitted waveform is a linear FM-pulse, which can be written on complex form as

\[ s(t) = \exp\{j\pi kt^2\}, \quad -T < t < T \]

The carrier is removed from \( s(t) \) above since it does not affect the range resolution. The instantaneous frequency of \( s(t) \) is given by the time derivative of the phase \( \phi(t) \).

\[ f(t) = \frac{1}{2\pi} \frac{d\phi(t)}{dt} = \frac{1}{2\pi} \frac{d}{dt}(\pi kt^2) = kt \]

The bandwidth of \( s(t) \) is thus

\[ B = f(T) - f(-T) = 2kT \]

The pulse echo will not be an exact replica of the transmitted pulse because of the relative motion between the antenna and the targets which imposes a
dopplershift on the pulse echo. Figure 3.4 shows a so-called squinted system where the boresight direction makes an angle $\theta_s$ with the direction normal to the flight direction.

Figure 3.4  **Geometry of a squinted system**

It is seen that the greatest dopplershift occurs when the target is on the edge of the beam as shown in figure 3.4. At this point the relative velocity between antenna and target is

$$ v_r = -v \cdot \sin(\theta_s + \frac{\pi}{2}) $$

With $r(t) = R_0 + v_r \cdot t$, the distance between target and antenna for the case shown in figure 3.4, the pulse echo is

$$ u(t) = s[t - \frac{2r(t)}{c}] = s(t - \frac{2R_0}{c} - \frac{2v_r \cdot t}{c}) $$

The term $\frac{2R_0}{c}$ gives the phase modulation used when generating the synthetic aperture, but it is constant for a single pulse echo from one target. It is therefore removed here. This gives
\[ u(t) = s\left(1 - \frac{2v_r}{c}\right)t = \exp\{jmk(1 - \frac{2v_r}{c})^2t^2\} = \exp\{jmk't^2\} \]

where \( k' = k(1 - \frac{2v_r}{c})^2 \approx k(1 - \frac{4v_r}{c}) \) since \( v_r \ll c \). The FM-rate \( k' \) of the echo pulse is thus slightly different from the transmitted one, depending upon where in the antenna beam the target is, being equal for a target lying normal to the flight direction. The consequences of this depend upon the magnitude of the factor \( 1 - \frac{4v_r}{c} \). When compressing \( u(t) \) with a filter matched to the transmitted pulse \( s(t) \), i.e., with a filter with impulse response \( s^*(T-t) \), the normalized output as a function of time delay and the different chirp-rates will be

\[
I(k, k', \tau) = \frac{1}{2T} \int_{-T}^{T} \exp\{jmk't^2\} \cdot \exp\{-jmk'(t-\tau)^2\} dt
\]

\[
= \frac{1}{2T\sqrt{2(k-k')^2}} \exp\{-j\frac{kk'}{k-k'}\tau^2\} \int \exp\{j\frac{\tau^2}{2}\} dt \nonumber
\]

\[
= \frac{1}{2T\sqrt{2(k-k')^2}} \exp\{-j\frac{kk'}{k-k'}\tau^2\} \int \sqrt{\frac{2}{(k-k')^2}} \left(\frac{k}{k-k'}\right)^{\tau-T} \nonumber
\]

\[
= 0 < \tau < 2T, \ k > k'
\]

The value of \( I \) for \(-2T < \tau < 0 \) and \( k < k' \) is given from the formula above by symmetry. The numerical values of \( I \) are given in terms of the Fresnel integrals which are tabulated. By introducing \( k_r = \frac{k'}{k}, \tau_r = \frac{T}{T} \) and \( W = 2T \cdot B = 4kT^2 \) which is the time-bandwidth product of the linear FM-pulse, we can write
\[ I(k_r, \tau_r) = \frac{1}{2\sqrt{W \frac{2}{(1-k_r^2)}}} \exp\left(-j\frac{\pi}{4} \frac{k_r}{1-k_r} W \tau_r \right) \int \exp(j\frac{\pi}{2} t^2) dt \]

\[ \sqrt{W \frac{2}{(1-k_r^2)}} \frac{k_r}{1-k_r} (1-k_r) \tau_r + 1 \]

\[ 0 \leq k_r < 1, \ 0 \leq \tau_r \leq 2 \quad (3.1) \]

For \( k_r = 1 \) we have

\[ I(1, \tau_r) = \frac{\sin[\frac{\pi}{2} \tau_r (1-k_r^2)]}{\frac{\pi}{2} \tau_r} \quad , \quad 0 \leq \tau_r \leq 2 \quad (3.2) \]

For SEASAT-1 we have \( W = 634 \). \( |I(1, \tau_r)| \) is the magnitude of the response for a perfectly matched filter and it is shown in figure 3.5. Figure 3.6 shows \( |I(k_r, 0)| \) which gives the relationship between the maximum output value of the compression filter and the relative mismatch in FM-rate between the filter and the pulse echo.

The maximum mismatch with positive angles will be

\[ k_r = \left(1 - \frac{4v_r}{c}\right)^{-1} \approx 1 - \frac{4v_r}{c} \cdot \sin(\theta_s + \frac{\theta_a}{2}) \]

The maximum squint angle \( \theta_s \) for SEASAT-1 is approximately \( 1^\circ \), which together with \( \theta_a \approx 1^\circ \) and \( v \approx 7150 \) m/s gives

\[ k_r = 1 - 2.5 \cdot 10^{-6} \]
Figure 3.5 Time response for perfectly matched compression filter

Figure 3.6 Maximum output of compression filter as function of FM-rate mismatch
This is seen from figure 3.6 to have negligible effect on the range compression. If a 1 dB degradation is set as a limit, this corresponds to $k_r = 0.9968$, which means that the FM-rate has to be known to 3 decimals.

Another mismatch occurs because of the carrier wave used when transmitting the pulses. Because of the relative velocity $v_r$ between antenna and target, the carrier wave will be dopplershifted an amount $\frac{2v_r}{c} \cdot f$, where $f$ is the carrier frequency. The maximum doppler-shift for SEASAT-1 is approximately 1.5 kHz. This will shift the echo pulse slightly out of the matched filter bandwidth. Relative to the total bandwidth of 19 MHz this is $\frac{1.5 \text{ kHz}}{19 \text{ MHz}} = 7.9 \cdot 10^{-5}$, and since the output will be reduced with the same relative amount, it is totally negligible.

Since the effects of relative motion between antenna and targets are negligible, the range compression will be totally azimuth independent, and can be done independently of the azimuth compression.

3.2 Azimuth resolution

The basis for generating a synthetic array antenna is a measure of the distance between the physical antenna and the target as a function of time. Since this time history is different for targets at different azimuth positions, they may be resolved by subsequent processing of the recorded time histories. Figure 3.7 shows the physical configuration with a single target.

The position vectors as functions of time, measured from an arbitrary reference point, are $\dot{x}_a(t)$ for the antenna and $\dot{x}_t(t)$ for the target. If the antenna
Figure 3.7 Physical configuration with radar and a single target

transmits a continuous monochromatic wave \( s(t) \), we have on complex form

\[
s(t) = \exp(j2\pi ft)
\]

where \( f \) is the frequency of the transmitted wave. The wave scattered from the target will be

\[
u_1(t) = \exp\{j2\pi f[t - \frac{2}{c}|\vec{x}_t(t) - \vec{x}_a(t)|]\} = \exp\{j2\pi f[t - \frac{2}{c}|\Delta \vec{x}(t)|]\}
\]

If \( u_1(t) \) is coherently demodulated with the transmitted wave, we get

\[
u(t) = u_1(t) \cdot \exp(-j2\pi ft) = \exp\{-j\frac{4\pi f}{\lambda} |\Delta \vec{x}(t)|\}
\]

where the transmitted wavelength is \( \lambda = \frac{c}{f} \). The measuring of the distance between antenna and target is thus done as a phase measurement, or equally as a time delay measurement.
If the relative movement between antenna and target is linear, we have the case shown in figure 3.8, where the target is fixed in space, and the antenna moves with velocity $v$.

![Diagram](image)

Figure 3.8  **Physical antenna in linear motion**

In this case we have

$$|\Delta \hat{x}(t)| = r(t) = (R_0^2 + v^2 t^2)^{\frac{1}{2}}$$

The recorded phase history for the target is thus

$$u(t) = \exp\{-j\frac{4\pi}{\lambda}(R_0^2 + v^2 t^2)^{\frac{1}{2}}\} \quad , \quad T_1 < t < T_2$$

where $[T_1, T_2]$ is the time interval when the target is inside the antenna beam. $T_1$ and $T_2$ can be found from figure 3.8.
When $\theta_a$ and $\theta_s$ are small, and $R_0 >> vT_1, vT_2$, we can use the following approximations

$$\tan x \approx x \quad \text{and} \quad (1+x^2)^{\frac{1}{2}} \approx 1 + \frac{x^2}{2} \quad \text{when} \ x << 1$$

We then have

$$u(t) = \exp\{-j\frac{4\pi}{\lambda}(R_0 + \frac{v^2 t^2}{2R_0})\} = C \exp\{-j2\pi\frac{v^2 t^2}{\lambda R_0}\}, \quad C = \exp\{-j\frac{4\pi R_0}{\lambda}\}$$

$$T_1 < t < T_2, \quad T_1 = \frac{R_0}{v} \frac{\theta_a + \theta_s}{2}$$

$$T_2 = \frac{R_0}{v} \frac{\theta_s}{2}$$

The factor $C$ in $u(t)$ is a constant for a given target and is of no importance when generating the synthetic aperture.

Since a linear FM-pulse with chirp rate $k$ can be written $\exp\{-j\pi ft^2\}$, it is seen that the phase history $u(t)$ for a single target can be approximated with a linear FM-pulse with chirp rate $k_a = \frac{2v^2}{\lambda R_0}$. The frequency band actually recorded is given by $T_1^0$ and $T_2$ above.

The azimuth history can also be thought of as the doppler history of the target as it passes through the antenna beam, since the instantaneous frequency in the azimuth history is equal to the doppler shift caused by the relative motion between antenna and target.
Up to now it has not been considered that the radar operates in pulsed mode, rather than in continuous mode. It is the pulsed operation that makes the synthetic antenna an array antenna, and not a continuous antenna. The effect of using pulses is equivalent to sampling the azimuth history of a target with the pulse repetition frequency (PRF) of the radar.

In section 2.1 it was mentioned that there is a maximum spacing between adjacent elements in an array antenna if aliasing effects are to be avoided. In the case of a synthetic array antenna no aliasing will occur if the PRF is greater than or equal to the Nyquist frequency determined by the bandwidth of the azimuth history. This bandwidth is

\[ B_a = k_a \cdot (T_2 - T_1) = \frac{2v^2 \cdot R_o}{\lambda R_a} \cdot a = \frac{2v \cdot \lambda}{D} = \frac{2v}{D} \]

where \( \theta_a = \frac{\lambda}{D} \) is used. Since the sampling is complex, the Nyquist criterion for no aliasing is

\[ \text{PRF} \geq B_a = \frac{2v}{D} \]

The time \( T_s \) between to successive samples then obeys

\[ T_s = \frac{1}{\text{PRF}} \leq \frac{D}{2v} \]

and the corresponding spatial separation is

\[ \Delta_s = v \cdot T_s \leq \frac{D}{2} \]

which means there has to be taken at least one sample for an antenna displacement equal to half the length of the physical antenna. In section 3.2.2 this will be seen to be equivalent to at least one sample per resolution cell in the azimuth direction.
The effect of undersampling is that each point target will occur in the image for a series of different azimuth positions instead of one.

The synthetic aperture is generated by compression of the azimuth history, in a way similar to the compression in range. Depending upon the complexity of the compression filter, one can discern between unfocused and focused synthetic aperture radar.

3.2.1 Unfocused synthetic aperture

The azimuth history for a target was shown above to be given approximately by a linear FM-pulse

\[ u(t) = \exp\left(-j2\pi \frac{v^2}{\lambda R_0} t^2 \right), \quad T_1 < t < T_2 \]

An unfocused synthetic aperture is generated by compressing \( u(t) \) with a filter that has no phase corrections. If the length of the synthetic aperture measured in time units is \( t_u \), the azimuth point response is given by

\[ a(t) = \int_{t = \frac{t_u}{2}}^{t = \frac{t_u}{2}} u(\tau) d\tau \]

It can be shown that minimum angular 3 dB resolution occurs when \( t_u = 1.2 \cdot t_1 \), where \( t_1 \) is defined by

\[ \frac{v^2}{\lambda R_0} = \frac{\pi}{2} \Rightarrow t_1 = \frac{\sqrt{\lambda R_0}}{2v} \]
Figure 3.9 shows a target and a spherical wavefront emanating from it. Generation of an unfocused synthetic aperture corresponds to integrating the incoming wavefront over an area where the wavefront can be regarded as approximately plane.

To find the resolution obtainable with unfocused processing, we assume an allowable phase difference from center to edge of the antenna of $\frac{\pi}{4}$, as was used in section 2.3 to find the boundary between the Fresnel and Fraunhofer regions. From section 2.3 we then have

$$\frac{\pi D_u^2}{4\lambda R_o} = \frac{\pi}{4} = D_u = \sqrt{\lambda R_o}$$

The ground resolution obtained can be found by considering the bandwidth of the azimuth history along the synthetic aperture. This bandwidth is

$$B_u = |\mathbf{t}_u \cdot \mathbf{k}_a| = \frac{D_u}{v} \cdot \mathbf{k}_a = \frac{1}{v} \sqrt{\lambda R_o} \cdot \frac{2v^2}{\sqrt{\lambda R_o}} = \frac{2v}{\sqrt{\lambda R_o}}$$
Ground resolution is then

$$\Delta_u = \frac{v}{f_u} = \frac{1}{2} \sqrt{\lambda R_0}$$

The antenna diagram for unfocused processing is given by the point response

$$a(t) = \int_{t-u/2}^{t+u/2} u(t)dt = \int_{t-u/2}^{t+u/2} \exp\left(-j2\pi \frac{\nu t^2}{\lambda R_0}\right)dt$$

This can be evaluated in terms of Fresnel integrals. The antenna pattern in angle can be found by setting $\theta = \frac{\nu t}{R_0}$ for small values of $\theta$. Figure 3.10 shows $a(t)$ for different values of $t_u$, the length of the synthetic aperture in the time domain. $t_1 = \frac{\sqrt{\lambda R_0}}{2v}$ is used as a parameter.

**Figure 3.10** Antenna diagrams for different lengths of the unfocused synthetic aperture
3.2.2 Focused synthetic aperture

In section 2.3 it was shown that the need to focus an antenna only arises when the target is in the Fresnel region of the antenna. The target is said to be in the Fresnel region when the distance between antenna and target is less than \( \frac{D^2}{\lambda} \), where \( D \) is the length of the aperture. For SEASAT-I we have \( \lambda = 0.24 \text{ m} \) and typically \( D \approx 4 \text{ km} \), which gives \( \frac{D^2}{\lambda} \approx 65000 \text{ km} \). Since the actual distance between target and antenna is approximately 850 km, all targets are seen to be deep in the Fresnel region of the synthetic antenna.

With focused processing the received wavefront is phase compensated across the whole synthetic aperture. This is equivalent to matched filtering of the target's azimuth history. In this case the complete azimuth history is compressed, and the bandwidth was earlier found to be

\[
B = \frac{2v}{D}
\]

This gives ground resolution

\[
\Delta_f = \frac{v}{B} = \frac{D}{2}
\]

The reason for the range independence of the azimuth resolution is that while the azimuth bandwidth is range independent, the time it takes for a given target to pass through the beam of the physical antenna is proportional with range. The time-bandwidth product of the azimuth history is therefore proportional with range, and the azimuth resolution remains constant.

Since the azimuth history is a linear FM-pulse, the performance of the azimuth compression is expressed by formula (3.1). The time-bandwidth product of the
complete azimuth history for SEASAT-I is approximately 3200. The azimuth antenna diagram without weighting and with perfect focusing is given by (3.2) with $W = 3200$. An actual antenna diagram with weighting is shown in figure 4.15.

The effect of a focusing error is given by $|I(k_r,0)|$ and is shown in figure 3.11 for $W = 3200$. If an azimuth filter matched to targets at range $R_0$ is used also for targets at range $R_1$, $R_1 < R_0$, this corresponds to a focusing mismatch

$$k_r = \frac{2v^2}{\lambda R_0} \rightarrow R_1 = k_r R_0$$

From figure 3.11 3 dB degradation is found to give $k_r = 0.998$. With $R_0 = 850$ km we have $R_1 = 849.065$ km. This means that the true range interval over which the same azimuth filter can be used if 3 dB degradation is tolerated at the edges, is

$$\Delta_r = 2(R_0 - R_1) = 1870 \text{ m}$$

The corresponding interval for 1 dB degradation is $\Delta_r = 1105 \text{ m}$.

In section 3.4 it will be shown that the normal processing mode does not compress the whole azimuth history coherently. If shorter parts of the azimuth history are compressed coherently and afterwards incoherently summed, each part will be less sensitive to focusing errors because of the lower time-bandwidth products. The case evaluated numerically above can therefore be regarded as a worst case.
In section 3.1 it was mentioned that a focused antenna has an inherent range resolution given by the depth of focus. This is just the $\Delta_r$ calculated above, and it is seen to be inferior since it is of the order of kilometers.

The azimuth resolutions that can be obtained with the types of processing investigated are summarized below and numerical values for SEASAT-1 are shown.

**SEASAT-1**

- **Real aperture**
  \[ \Delta_a = \frac{R \lambda}{D} \]
  \[ \Delta_a = 18.2 \text{ km} \]

- **Unfocused synthetic aperture**
  \[ \Delta_a = \frac{1}{2} \sqrt{\lambda R} \]
  \[ \Delta_a = 223 \text{ m} \]

- **Focused synthetic aperture**
  \[ \Delta_a = \frac{D}{2} \]
  \[ \Delta_a = 5.5 \text{ m} \]
There are different kinds of distortions that appear in a synthetic aperture radar image of an area on the ground. These can be divided into deterministic and stochastic distortions, although the distinction between these two is partly a matter of definition.

The simplest type of distortion is the purely geometric distortion which gives a scale distortion, in general nonlinear, of the image. This kind of distortion was mentioned in section 3.1, where the difference between ground range and true range was treated. Similar distortions may also occur in the azimuth coordinate depending upon how the azimuth filters are generated. This is a correctable distortion, since it only causes a warping of the coordinate system in the processed image. It will not be treated here, since the absolute locations of targets in an image is of little importance for the purposes considered here.

First a type of distortion called range fold-over is treated. Then imaging of moving targets is studied. At last stochastic errors are briefly considered.

3.3.1 Range fold-over

Range fold-over is a distortion that stems from the difference between true range and ground range. Figure 3.12 shows a mountain and the way it will appear in the image.

All the three points A, B and C are at the same true range from the radar and will therefore be mapped to the same image point. This has the effect of folding the mountains towards the radar.
The displacement for an object at height \( h \) from its correct ground range position is approximately, from figure 3.12, 

\[
\delta_r = h \cdot \cotan \theta, \quad \text{for SEASAT-1} \quad \delta_r \approx h \cdot \cotan 20^\circ \approx 2.7 h
\]

This type of distortion is incorrectable and is easily seen in images of mountainous areas, as for example in figure 4.22.

3.3.2 Moving targets

As shown earlier, the principle of generating a synthetic aperture is based on relative movement between target and radar platform. When imaging the surface of the earth, a target that is in motion relative to the surface will be imaged differently from a target that is fixed on the surface. This will for example be the case for a moving ship, and even for the sea itself although the motion of the sea surface to a high degree may be considered as stochastic.
It was shown in section 3.1 that the dopplershift caused by a moving target is of no concern in range compression. The only effect of a moving target therefore appears in the azimuth compression.

It was shown in section 3.2 that the recorded phase history for a target is

\[ u(t) = \exp\left\{-\frac{j2\pi}{\lambda} |\vec{r}(t)| \right\}, \quad T_1 < t < T_2 \]

where \( \vec{r}(t) \) is the relative displacement vector between antenna and target. For quasi-linear relative motion we have

\[ |\vec{r}(t)| = (R_0^2 + v^2 t^2)^{\frac{1}{2}} \]

It is convenient to divide the target motion into components in range direction and azimuth direction. There will also be a component normal to both the range and azimuth directions, but this can be neglected for all targets of interest.

If terms of higher order than second are neglected, the motion of a target relative to the ground in the coordinates mentioned above can be described by four parameters, \( v_r, v_a, a_r \) and \( a_a \), where \( v \) is velocity and \( a \) acceleration, and the indices are \( r \) for range and \( a \) for azimuth. The length of the displacement vector for a moving target can then be written

\[ |\vec{r}(t)| = \left\{ (R_0 + v_r t + \frac{1}{2} a_r t^2)^2 + (v_a t + \frac{1}{2} a_a t^2)^2 \right\}^{\frac{1}{2}} \]

By retaining terms up to second order in time, we get

\[ |\vec{r}(t)| \approx R_0 + v_r t + \frac{1}{2} a_r t^2 \]
The corresponding phase history is

\[ \phi(t) = \frac{4\pi}{\lambda} \left| \mathbf{r}(t) \right| = \frac{4\pi}{\lambda} \left\{ R_0 + \mathbf{v} \cdot t + \frac{1}{2} \left( a_{x} \right) \frac{(v \cdot v_a)^2}{R_0} t^2 \right\} \]

We see that with the approximations made here, the acceleration \( a_a \) in the azimuth direction has no impact on the processing.

The difference in phase history between a moving target and a stationary target is

\[ \Delta \phi(t) = \frac{4\pi}{\lambda} \left\{ \mathbf{v} \cdot t + \left( \frac{1}{2} a_{x} + \frac{v \cdot v_a}{R_0} \right) t^2 \right\} \]

where the approximation \( 2v \cdot v_a + v_a^2 \approx 2v \cdot v_a \) has been used since \( v \gg v_a \).

The term \( \mathbf{v} \cdot t \) represents a frequency shift in the azimuth history of \( \frac{\mathbf{v} \cdot t}{\lambda} \) Hz. Since the azimuth history of a target is approximately a linear FM-pulse, this means that the effect of the velocity component \( v_x \) is a displacement of the target in the processed image in azimuth direction.

The effect of \( a_x \) and \( v_a \) is to change the chirp rate of the azimuth history. The chirp rate for a stationary target is \( k = \frac{2v^2}{\lambda R_0} \), and for a moving target we get chirp rate

\[ k' = \frac{\left( \frac{1}{2} a_x + \frac{v \cdot v_a}{R_0} \right)}{\frac{2v}{\lambda R_0}} k \]

Expressed in relative terms we have

\[ k_r = \frac{k}{k'} = \left( \frac{R_0 a_x}{2v^2} + \frac{2v}{v} + 1 \right)^{-1} \]

The effect of this mismatch is to cause defocusing of the target in the image.
The velocity component \( v_r \) may also have another impact on the processing since it changes the target trajectory in the raw data. Since the trajectory is oriented mostly in the azimuth direction, it is very little sensitive to \( v_a \). The change in trajectory will cause a smear of the target in the image.

The extra range migration for a moving target during the generation of the synthetic aperture is

\[
\Delta R = v_r \cdot \frac{\lambda}{D} \cdot R_0 \cdot \frac{1}{v} = \frac{v_r \cdot \lambda R_0}{vD}
\]

where \( D \) is the length of the real antenna. \( \frac{\lambda R_0}{D} \) is the length of the synthetic aperture.

The quantitative effects of these distortions are given below in terms of numerical values for SEASAT-1.

The dopplershift in the azimuth history due to the radial velocity \( v_r \) was above found to be \(-\frac{2v_r}{\lambda}\) Hz, where \( v_r \) is positive away from the radar. The dopplershift caused by a stationary target as it passes through the antenna beam is continuously decreasing. If a moving target causes an extra dopplershift, this corresponds to a target appearing earlier if the extra dopplershift is negative, and to a target appearing later if the dopplershift is positive. The displacement of the target in the image will therefore be as shown in figure 3.13.

An example of this phenomenon is shown in figure 3.14, where several ships moving in the radial direction are shown. The wake shows the direction of the ship's movement and where it actually is. The azimuthal displacement of the ships themselves is clearly seen from the figure. In figure 3.15 ships moving in the
Figure 3.13 Displacement in image of radially moving target

azimuth direction are shown, and no displacement from the wake is seen to occur.

A displacement in azimuth with one resolution cell, i.e. 25 m, gives

$$ v_r = \frac{2v}{\lambda k_a} = 25 \text{ m} $$

where $k_a$ is the azimuth chirp rate. Typical values for SEASAT-1 are $k_a \approx 500 \text{ Hz/sec}$ and $v \approx 7000 \text{ m/s}$, and we get

$$ v_r \approx 0.2 \text{ m/s} $$

With a depression angle of $70^\circ$, the corresponding ground range velocity is

$$ v_g = \frac{v_r}{\cos 70^\circ} \approx 0.65 \text{ m/s} = 1.3 \text{ knots} $$
Figure 3.14 Radially traveling ships

Figure 3.15 Azimuthally traveling ships
Typical displacements in figure 3.14 correspond to velocities of approximately 15 knots, which is seen to be quite reasonable.

Smearing in the image caused by $v_r$ is said to occur when the radial displacement during the synthetic aperture is greater than one range resolution cell. This is approximately 8 m for SEASAT-1 and gives

$$\frac{v_r \cdot \frac{\lambda R_o}{V}}{D} > 8 \text{ m}$$

$D = 11 \text{ m}, V = 7000 \text{ m/s}$ and $R_o = 850 \text{ km}$ gives

$v_r > 3.1 \text{ m/s} \approx 6 \text{ knots}$

Corresponding ground velocity is

$v_g > 18 \text{ knots}$

In section 3.2 it was found that a mismatch in focus corresponding to $k_r = 0.9989$ gives a 3 dB degradation in peak response. It was found above that the components $a_r$ and $v_a$ give a mismatch in focus corresponding to

$$k_r = \left(\frac{R_o a_r}{v^2} + \frac{2 v_a}{v} + 1\right)^{-1}$$

With $R_o = 850 \text{ km}$ and $v = 7000 \text{ m/s}$ we have

$k_r = 0.9989$ and $a_r = 0 \Rightarrow v_a = 3.9 \text{ m/s} = 7.6 \text{ knots}$

$k_r = 0.9989$ and $v_a = 0 \Rightarrow a_r = 0.06 \text{ m/s}^2, a_g \approx 0.2 \text{ m/s}^2$
Since an acceleration of 0.2 m/s$^2$ is very high for a ship target, defocusing is likely to be due to the azimuthal velocity component $v_a$.

The numerical results derived above are summarized below. All velocities are ground velocities.

<table>
<thead>
<tr>
<th>Effect</th>
<th>Velocity (knots)</th>
</tr>
</thead>
<tbody>
<tr>
<td>Azimuth displacement</td>
<td>$v_r &gt; 1.3$</td>
</tr>
<tr>
<td>Smearing in range</td>
<td>$v_r &gt; 18$</td>
</tr>
<tr>
<td>Azimuth defocusing</td>
<td>$v_a &gt; 7.6$</td>
</tr>
</tbody>
</table>

If the azimuthal displacement from the wake is used to estimate the ship's radial velocity, $v_r = 1.3$ knots can be regarded as the velocity resolution.

### 3.3.3 Stochastic errors

Some of the errors can only be modelled as stochastic errors, and the effects expressed in terms of mean values, variances and correlation functions. The most important of this type of errors can be expressed as phase errors in the azimuth history of a target. If the azimuth history of a target is written

$$u(t) = a(t) \cdot \exp\{-j \cdot \phi(t)\}$$

where $a(t)$ is the amplitude and $\phi(t)$ is the phase history of a target, then a phase error $\delta\phi(t)$ can be included by writing

$$u(t) = a(t) \cdot \exp\{-j[\phi(t) + \delta\phi(t)]\}$$

Fluctuations in $a(t)$ have been neglected since they are in general of less importance than $\delta\phi(t)$. The phase error term $\delta\phi(t)$ includes random propagation delay...
through the atmosphere, phase jitter in the applied signals and random motions of targets and platform. The effect of $\delta \phi(t)$ on the compression operations will depend upon the statistical properties of $\delta \phi(t)$. This type of errors is not analyzed here, but references (6) and (7) contain results from investigations of such errors.

3.4 Speckle reduction

Because a synthetic aperture radar system is a highly coherent imaging system, the images generated are subject to speckle (8), (9). This phenomenon, which gives the images a granular appearance, can be explained by means of figure 3.16. This figure shows a single resolution cell on the ground, where its projection in true range is considered. Several specularly reflecting elements inside the resolution cell are shown.

Figure 3.16 Speckle in coherent imaging systems

If the signals received at the radar are designated

$$A_i \cdot \exp\{j\phi_i\}, \ i = 1, 2, 3$$
where $\phi_i$ includes both propagation delay and the internal phase of the scattering elements, the complex amplitude at the receiver will be

$$a(\psi) = \sum_{i=1}^{3} A_i \cdot \exp\{j\phi_i\}$$

where the dependency upon the viewing angle $\psi$ has been explicitly denoted. $\psi$ has a one-to-one correspondence with the azimuth time coordinate, and the received signal is therefore seen to vary in time in a manner determined by the spatial locations of the scattering elements inside the resolution cell.

In the general case with $N$ scattering elements inside the resolution cell the received signal will be

$$a(\psi) = \sum_{i=1}^{N} A_i \cdot \exp\{j\phi_i\}$$

and both $A_i$ and $\phi_i$, and consequently $a(\psi)$, can be regarded as random variables.

If all the $A_i$'s and $\phi_i$'s are statistically independent and the $\phi_i$'s are uniformly distributed on the interval $[-\pi, \pi]$, the addition of the complex amplitudes at the receiver can be regarded as a random walk in the complex plane (9). If the number $N$ of scattering elements is high, this process is similar to the narrowband noise process, and the probability distribution of $|a(\psi)|$ is equivalent to the distribution of the amplitude of narrowband noise, i.e., the Rayleigh distribution. We thus have

$$p(|a|) = \frac{2}{a_0} \exp\left(-\frac{|a|^2}{a_0^2}\right), \quad |a|>0$$
where \( a_0^2 = \langle |a|^2 \rangle = \int_0^\infty |a|^2 p(|a|) \, |a| \, da \)

The intensity of the received signal is \( I(\psi) = |a(\psi)|^2 \), the distribution of which is exponential

\[
p(I) = \frac{1}{I_o} \exp\left(-\frac{I}{I_o}\right)
\]

(3.3)

where \( I_o = \langle I \rangle \) is the expected value of the intensity.

The geometry shown in figure 3.16 is not directly applicable to the case of synthetic aperture radar, since each resolution cell is imaged by means of a synthetic aperture extending over an interval of the viewing angle \( \psi \). If, however, this imaging process is thought to occur separately for each scattering element inside the resolution cell, the resulting signal can be assigned for example to the middle value of the \( \psi \)-interval, and the geometry will be applicable also to the SAR case.

The assumptions about the statistical properties of the scattering elements, which are characteristic of what is called fully developed speckle, can not be expected to hold for very strong targets, as for example urban areas on land and ships on the sea. For most land areas and for the sea surface, however, the assumptions will in general be well satisfied. A resolution cell of 25mx25m on the sea surface will contain many independent scattering elements, and with a radar wavelength of 23.5 cm the phase excursions over the resolution cell will be large enough to give uniform phase distribution.
An example of an image of a sea area is shown in figure 3.17, where the speckly nature of the image clearly can be seen. Figure 3.18 shows the probability distribution function of the intensity calculated from the image and also the theoretical distribution according to formula (3.3) where the mean intensity calculated from the image has been used. They are seen to be in excellent agreement with each other.

![Image of sea area showing speckle](image)

Figure 3.17 Image of sea area showing speckle

As mentioned above the speckle in the images is caused by the coherence of the imaging process. The speckle effect can be reduced if several uncorrelated images of the same ground area are generated and afterwards incoherently added, that is added on an intensity basis.

The contrast of a speckle pattern is defined as

$$C = \frac{\sigma_I}{I_0}$$  \hspace{1cm} (3.4)
Figure 3.18  Intensity distribution for image in Figure 3.17

...the ratio between the standard deviation and the mean value of the intensity. With the probability distribution given in (3.3) we get $C = 1$. Addition of $N$ uncorrelated images gives a contrast $1/\sqrt{N}$.

The formulas above are for polarized speckle. By using two orthogonal polarizations two uncorrelated images are obtained. The usual method for obtaining uncorrelated images, and the one used here, is to generate images from different parts of the signal spectra. These images will be uncorrelated if they occupy nonoverlapping parts of the spectrum. This method represents a compromise between resolution and speckle reduction.
The two-dimensional signal spectrum is shown in figure 3.19, where it is divided into $N_r$ equal frequency bands in range and $N_a$ bands in azimuth.

![Diagram](image)

Figure 3.19 Division of signal spectrum in equal frequency bands

This will give $N_r \cdot N_a$ uncorrelated images. If the best resolutions that can be obtained by the system are $\Delta_r$ in range and $\Delta_a$ in azimuth, which corresponds to $N_r = N_a = 1$, this procedure gives resolutions $N_r \cdot \Delta_r$ and $N_a \cdot \Delta_a$, and a speckle contrast $1/\sqrt{N_r \cdot N_a}$.

In the case of SEASAT-1 we use $N_r = 1$ and $N_a = 4$. This gives a resulting ground resolution of approximately 25mx25m and a speckle contrast of 0.5. Since the azimuth history is a linear FM-pulse, the different frequency bands correspond to different time sections as shown in figure 3.20.

Each of the sections is called a "look". Since each look represents a particular viewing angle, uncorrelated images are seen to be obtained by angular diversity.
Partially overlapping looks will give a look correlation that depends upon the degree of overlap. The images processed here have the four looks placed in the time domain as shown in figure 3.21.

The window functions used for the looks are shown, and neighbouring looks have 1/3 overlap on the time axis. Because of the window functions, the effective overlap
will be less than $1/3$. The lowest curve in figure 4.25 shows that a contrast of very nearly 0.5 is obtained in this way for a sea area, which shows that the 4 looks are nearly uncorrelated.

Figure 3.22 shows the four different looks for an area in the Strait of Dover, and figure 3.23 shows the effect when adding the looks together. The reduction in speckle when averaging looks is easily seen.
Figure 3.22  Four different looks of the same area
Figure 3.23 Effect of look averaging
4 PROCESSING OF SYNTHETIC APERTURE RADAR IMAGES

Processing of SAR images from rawdata consists of matched filtering of the rawdata with a filter matched to the point target response. This is a twodimensional filtering operation in range and azimuth.

When generating a SAR image, the last operation is conversion from complex amplitude representation of the signals to intensity. Although this is a nonlinear operation, the quality of the image is well described by signal to noise ratios before intensity conversion because all the preceding operations are linear filtering operations.

The received power can be found from the radar equation. If the scattering cross section of a resolution cell on the ground is $A \cdot \sigma_o$, where $A$ is the area of the resolution cell and $\sigma_o$ is the radar reflectivity, the received power is

$$P_r = \frac{P_t \cdot G^2 \cdot \lambda^2 \cdot A \cdot \sigma_o}{(4\pi)^3 \cdot R^4}$$

where $P_t$ is transmitted power and $G$ is antenna gain. The system noise power is

$$N = kTB$$

where $k$ is Bolzmann's constant, $T$ is effective noise temperature and $B$ is receiver bandwidth. For SEASAT-1 we have

$P_t = 800$ W
$G = 35$ dB
$T = 650$ K
$B = 19 \cdot 10^6$ Hz
$R = 850$ km
For 1 look and full resolution we have $A \approx 150 \, \text{m}^2$ which gives signal to noise ratio

$$\frac{P_r}{N} = \frac{P_c \cdot G^2 \cdot \lambda^2 \cdot A \cdot \sigma_o}{(4\pi)^3 \cdot R^4 \cdot k \cdot T \cdot B} = 3.8 \cdot 10^{-4} \cdot \sigma_o = -34 \, \text{dB} + \sigma_o [\text{dB}]$$

For an average sea model we have in the middle of the swath (4), $\sigma_o = -14 \, \text{dB}$ which gives

$$\frac{P_r}{N} = -34 \, \text{dB} - 14 \, \text{dB} = -48 \, \text{dB}$$

The processing gain is given by the time-bandwidth products of the range and azimuth waveforms, 634 in range and approximately 3200 in azimuth. Processing gain is then

$$G_p = 634 \cdot 3200 = 2.03 \cdot 10^6 = 63 \, \text{dB}$$

This gives a resulting signal to noise ratio of $63 \, \text{dB} - 48 \, \text{dB} = 15 \, \text{dB}$.

Processing of SAR images has in the past mainly been done by optical processing, where a configuration of lenses and a coherent light source are employed. The rawdata signal is stored on film in a twodimensional format, the optical system scans this input film and the output is images which also are recorded on film.

The main reasons for doing the processing optically are processing speed, the suitability of film as a storage medium for large amounts of rawdata, and the fact that the azimuth signal for a target, used for generation of the synthetic aperture, exhibits self focusing properties in a coherent optical system.
Descriptions of signal formats and optical configurations used for optical processing of SAR images can be found in (10) and (11).

4.1 Digital processing

Although optical processing of SAR images still is used when really great throughput is needed, digital processing has some important advantages. The optical configuration needed is not very mobile and the optical recorder needed for recording the rawdata signal on film is a critical element. These are avoided with digital processing, and the ease with which processing parameters can be adjusted and optimized in digital processing tends to give digitally processed images higher quality than optically processed ones.

The SAR in SEASAT-1 was an L-band radar operating at 1.275 GHz. The pulse echoes were upconverted to S-band before transmission to ground. After reception on the ground the echoes were demodulated and converted to 5 bit samples and stored on magnetic tape.

The data recording is a twodimensional sampling operation, sampling in azimuth by transmission of pulses and in range by the A/D-conversion mentioned above. The sampling frequency depends upon whether the sampling is real or complex. If the azimuth sampling is complex, this gives a PRF that is only half the corresponding real sampling frequency, and it also makes it possible to use a recording window that has the double length of what it has in the case of real sampling. For SEASAT-1 the sampling is complex. An explanation of differences between these modes of operation can be found in (12).
The twodimensional filtering that has to be accomplished is performed digitally by doing two onedimensional filtering operations, one in range and one in azimuth.

It is convenient to consider the recorded radar echoes as a twodimensional signal, and in the case of optical processing this is actually done when recording the signal on film. A similar format when using digital processing is shown in figure 4.1.

![Twodimensional digital data format](image)

**Figure 4.1** Twodimensional digital data format

The dots represent the recorded samples, i.e., one vertical line represents one sampled pulse echo. The area A shows how the echo from a single target is distributed in the recorded data when the radar antenna is pointing directly normal to the direction of relative motion between target and antenna. The distribution of the target history in range is caused by the length of the transmitted pulse, and the distribution in azimuth represents the synthetic aperture.
B is the target history for a squinted system and represents the typical target history with SEASAT-1.

The processing of a SAR-image from rawdata means to compress the target history in both range and azimuth to an extension equal to approximately one sample in both directions.

While the range compression is strictly onedimensional and, as shown in section 3.1, completely azimuth independent, the azimuth history is curved in the range direction, so called range migration, and it is also range dependent. In the next two sections range and azimuth compression will be treated separately.

A block diagram of the processor is shown in figure 4.2.

![Figure 4.2 Block diagram of digital processor](image)

After a suitable block of rawdata has been written to disc, the doppler centroid in the azimuth spectrum is estimated. The estimation algorithm is very simple and it is briefly explained in Appendix A. When the doppler centroid is estimated, a block of rawdata is compressed in range, and an interpolation is performed for each range compressed vector before it is written back to disc.
Since disc is a onedimensional storage medium, the range compressed datablock has to be transposed, or corner turned, before it can be accessed in the azimuth direction. This corner turning is done effectively by means of the so-called Eklundh's algorithm. It is explained in (13). After corner turning the azimuth filters are generated. To generate the correct matched filters for azimuth compression, three parameters related to the satellite orbit and attitude are needed. These are

1) distance between radar and target
2) effective squint angle of the system
3) relative velocity between antenna and target

The two last parameters will be treated in more detail later, but briefly, it can be said that 1) is given by the parameter called range pulse delay in figure 4.2, 2) is given by the doppler centroid in the azimuth spectrum, and 3) is estimated by the auto-focusing procedure described in section 4.4.

During azimuth compression the four looks are separately generated and afterwards incoherently added to give the final multi-look image.

To generate a complete image the sequence of processing steps is repeated a number of times except the doppler centroid estimation and the generation of the azimuth filters.

The theory of linear digital filtering is described in (14). Digital filtering can be done either in the time or frequency domain, where frequency domain filtering is more efficient, measured in number of operations, for filter lengths more than approximately 30 samples. Since the lengths of the matched filters
used here are typically several hundred samples, the pulse compressions are done in the frequency domain. Forward and inverse digital Fourier transforms (DFT) are calculated by means of the Fast Fourier Transform algorithm (FFT), described in (15).

The theory of matched filtering can be found in (5).

The details of the algorithms will depend on the radar system used and the actual system parameters. There will for example be operations that are necessary when processing data from a satellite borne synthetic aperture radar that would not be necessary if the radar was airborne, and vice versa. Although the algorithms developed here are for a satellite borne radar, they include all the principles of the synthetic aperture radar and could, with some modifications, be used also for airborne SAR.

A detailed description of the algorithms developed here together with computer programs written in FORTRAN that implement these algorithms on a general purpose computer are given in (16).

4.2 Range compression

During range compression each pulse echo is filtered with a filter matched to the transmitted linear FM-pulse. This corresponds to removing the range distribution of the target history as it is shown in figure 4.2.

The analog to digital conversion of each pulse echo is done to 5 bit real samples. During range compression these are converted into complex samples which reduces the sampling frequency in range with a factor 2. This
also reduces the range migration of the target histories with a factor 2 measured in number of samples. This makes the range migration correction during the azimuth compression easier, and it gives directly the phase information necessary for generating the synthetic aperture. The real to complex conversion is done as follows.

A DFT of a 2N-point real sequence gives a 2N-point complex sequence where the upper half of the sequence of frequency components is given as the complex conjugate of the lower half. By discarding the upper N complex frequency components and taking an N-point inverse DFT (IDFT) of the rest of the sequence, an N-point complex time sequence is obtained which represents the complex version of the 2N real samples.

Another method for obtaining the complex samples would be to take the Hilbert transform of the real samples, which would then give the imaginary parts of the complex samples.

The transmitted pulse is a linear FM-pulse with chirp-rate $5.63 \cdot 10^{11}$ Hz/sec. On complex form its duration is 768 samples, and the frequency spectrum is shown in figure 4.3.

Figure 4.4 shows the received frequency spectrum averaged over 100 pulse echoes.

When a suitable block of rawdata has been written to disc and the doppler centroid has been estimated, range compression is performed. This pulse compression is accomplished by filtering one range vector from the rawdata matrix at a time with a filter matched to the transmitted pulse, and storing the undistorted output values in matrix form on disc.
Figure 4.3 Frequency spectrum of transmitted pulse

Figure 4.4 Averaged received spectrum in range
The rawdata is packed with three 5 bit samples in a 16 bit word, and the unpacking necessary prior to range compression is done in the same routine that does the range compression.

Conversion from real to complex samples as mentioned above, is efficiently done by utilizing an N-point complex FFT routine for the transformation of 2N real samples, and afterwards descrambling the output values of the FFT routine to get the N complex frequency components wanted. The method is described in (15).

Figure 4.5 shows a block diagram for the range compression routine.

The input vector $x_1$ contains coded real 5 bit samples. $x_1$ is decoded to the complex vector $x_2(i)$, $i = 0, N-1$, where $N = 4096$ is a typical length chosen as processing vector length. The consecutive real samples are stored in $\text{Re}\{x_2(o)\}$, $\text{Im}\{x_2(o)\}$, $\text{Re}\{x_2(1)\}$, $\text{Im}\{x_2(1)\}$, etc. Fourier transformation gives

$$x_3(i) = \sum_{k=0}^{N-1} x_2(k) \cdot \exp\left(-j \frac{2\pi}{N} i \cdot k\right), \quad i = 0, N-1$$

Descrambling is done according to the following algorithm.
\[ \text{Re}(x_4(i)) = \left[ \frac{\text{Re}(i)}{2} + \frac{\text{Re}(N-1)}{2} \right] + \cos\frac{\pi i}{N} \left[ \frac{\text{Im}(i)}{2} + \frac{\text{Im}(N-1)}{2} \right] \\
- \sin\frac{\pi}{N} \left[ \frac{\text{Re}(i)}{2} - \frac{\text{Re}(N-1)}{2} \right] \]

\[ \text{Im}(x_4(i)) = \left[ \frac{\text{Im}(i)}{2} - \frac{\text{Im}(N-1)}{2} \right] - \sin\frac{\pi}{N} \left[ \frac{\text{Im}(i)}{2} + \frac{\text{Im}(N-1)}{2} \right] \\
- \cos\frac{\pi i}{N} \left[ \frac{\text{Re}(i)}{2} - \frac{\text{Re}(N-1)}{2} \right], \quad i = 0, N-1 \]

where \( \text{Re}(i) = \text{Re}(x_3(i)) \), \( \text{Im}(i) = \text{Im}(x_3(i)) \), and \( x_3(N) = x_3(0) \)

The matched filter's frequency components \( H(i) \), \( i = 0, N-1 \), are calculated and stored before range compression begins. Filter multiplication gives

\[ x_5(i) = x_4(i) \cdot H(i), \quad i = 0, N-1 \]

The received pulse echoes have a carrier frequency which is half the complex sampling frequency. Demodulation is given as

\[ x_6(i) = x_5[(i+\frac{N}{2}) \text{modulo } N], \quad i = 0, N-1 \]

Transformation back to the time domain

\[ x_7(i) = \frac{1}{N} \sum_{k=0}^{N-1} x_6(k) \cdot \exp\left( j \frac{2\pi}{N} i \cdot k \right), \quad i = 0, N-1 \]

Because of "wrap-around" in the filtering, the undistorted output values are

\[ x_7(i), \quad i = 0, M-1 \]

where \( M = N-767 \), since the length of the impulse response of the matched filter is 768 samples.
Interpolation in $x_7$ uses a 4-point interpolation filter which gives

$$x_8(i) = \sum_{k=0}^{3} x_7(i+k+m) \cdot \sin\left(\frac{(k-1-\tau)\pi}{(k-1-\tau)\pi}\right), \quad 0 \leq \tau < 1, \quad i = 0, M-4 \quad (4.1)$$

$m$ and $\tau$ vary from pulse to pulse and are determined as explained in section 4.3. The filtered vector $x_8$ is then written to disc.

No weighting is applied to the range pulse to reduce the sidelobe level, and the time response after compression is shown in figure 3.5. The highest sidelobes are at -13 dB.

To illustrate the effect of range compression, a portion of raw data is shown in figure 4.6 and it is seen to be completely noiselike. Noise or interference can be seen on some of the pulse echoes, but it is suppressed during the processing. Figure 4.7 shows the same data matrix after range compression, and it is seen that there is considerable structure in the range direction, while all the features in the azimuth direction are still smeared out.

4.3 Azimuth compression

After range compression of a suitable block has been completed, and corner turning is performed, the matrix has to be compressed in the azimuth direction, i.e., filtering rowwise in the data matrix in figure 4.1.

Of the two types of target histories shown in figure 4.1, type A can be regarded as a special case of type B. After range compression and before interpolation, a typical target history therefore appears as shown in figure 4.8.
Figure 4.6  Block of raw data

Figure 4.7  Datablock in figure 4.6 compressed in range
Figure 4.8 Target history after range compression

As described in section 3.2, the recorded target trajectory is given by

$$r(t) = (R_0^2 + v^2 t^2)^{1/2}, \quad T_1 < t < T_2$$

where $T_1$ and $T_2$ determine which part of the hyperbola by which $r(t)$ is given, is recorded by the system.

The time $(T_1 + T_2)/2$ corresponds approximately to the center of the antenna beam, and the doppler frequency of the recorded azimuth history corresponding to this value of $t$ is called the doppler centroid of the azimuth spectrum. Case A in figure 4.1 is thus seen to be the case of 0 doppler centroid.

A squinted system is a system where the doppler centroid is not equal to 0. Since the radar effectively samples the azimuth histories with the PRF, it sees all azimuth frequencies modulo PRF. This ambiguity can be resolved for example by calculating the variation in doppler centroid across the swath, since this variation will depend upon the squint angle of the system. More
effectively, however, the ambiguity is resolved by making a piece of an image for each possible value of the doppler centroid. Incorrect values of the doppler centroid will then give highly degraded images. It is here assumed that no such ambiguity exists since this turns out to be the typical case when using a satellite borne SAR.

A typical azimuth spectrum for a squinted system is shown schematically in figure 4.9 and figure 4.10 shows the azimuth spectrum for one data set from SEASAT-1 averaged over 100 azimuth vectors.

\[ W(f) \]

\begin{figure}[h]
\centering
\includegraphics[width=0.5\textwidth]{spectrum.png}
\caption{Typical azimuth spectrum for a squinted system}
\end{figure}

The range migration of a target is given by the target's range variation during the observation interval. Range migration complicates the azimuth compression, and it should be as small as possible. The range migration caused by the curvature of the target trajectory is unavoidable, but for a squinted system this is small compared to the range migration caused by the overall inclination of the target trajectory, as can be seen from figure 4.8. To minimize range migration, the matrix of range compressed
Figure 4.10: Averaged azimuth spectrum from SEASAT-1

data is therefore skewed to make the new azimuth vectors parallel to the straight line drawn as a tangent to the target trajectory in figure 4.8. This is the operation performed by the box termed interpolation in figure 4.5, where each range vector is shifted in range by means of a 4-point interpolation filter to fit the new azimuth vector direction.

To make the range migration as small as possible, the new azimuth lines should be tangential to the target trajectory at the point corresponding to the doppler centroid. The recorded phase history of a target is

\[ \phi(t) = \frac{4\pi}{\lambda} r(t), \quad T_1 < t < T_2 \]

The corresponding doppler frequency as a function of time is

\[ f(t) = \frac{1}{2\pi} \frac{d}{dt}[\phi(t)] = \frac{2}{\lambda} \frac{d}{dt}[r(t)] \]
This shows that the inclination of the target trajectory at the point corresponding to a given doppler frequency is directly proportional to this frequency, and does not depend on the actual distance and relative azimuthal velocity between antenna and target.

Since the inclination of the interpolation line is needed prior to range compression, only an estimate of the doppler centroid is needed before range compression begins. This is the reason why the doppler centroid estimation is done directly from the rawdata as shown in figure 4.2.

Typical values for the range migration measured in complex range samples is $100$ without interpolation and $8$ with interpolation, so the range migration is seen to be considerably reduced by the skewing procedure.

Although the doppler centroid may vary with a couple of hundred Hz across the swath out of an azimuth bandwidth of approximately $1300$ Hz, the range migration is still maintained low, even if range migration is minimized only in the middle of the swath. If the whole swath is processed in several blocks, it is possible to use different inclination angles for the interpolation directions for the different blocks, but since the output images are correspondingly skewed, this has to accounted for if the images are mapped together. The images processed here have the same direction of the azimuth lines across the whole swath.

After range compression and skewing a target history looks like shown in figure 4.11.
It is seen that the azimuth history of a single target is distributed over several azimuth lines. One way to do the azimuth compression would be to divide the azimuth filter into several portions as shown by the seven rectangles in figure 4.11, filter each azimuth vector with all these filters, and afterwards combine the outputs in a proper way to get the total azimuth history compressed. The drawback with this method is, however, that a number of filters will have to be used for each azimuth vector, and this is very time consuming.

A method which requires only one azimuth filter is to do the straightening of the azimuth histories in the frequency domain. This is explained in figure 4.12, where F means Fourier transform.

To the left is shown the matrix form of range compressed data and the trajectories for two targets at the same range but at different azimuth locations. To the right is shown the trajectories after Fourier transformation of the azimuth vectors. The two trajectories are then seen to coincide. This is because the azimuth histories

Figure 4.11 **Target history after range compression and skewing**
are essentially linear FM-pulses. Because of this there is a one-to-one correspondence between position along the azimuth history and instantaneous frequency of the azimuth history. Said in another way, for a given azimuth vector the portions of the azimuth histories of the two targets shown in figure 4.12 that fall within this azimuth vector, contain the same frequency components. Therefore they will be mapped together when taking the Fourier transform of the azimuth vectors. The distinction between different targets in the frequency domain is represented by the complex phase factors of the frequency components.

Because of the different values that the doppler centroid $f_d$ can attain, it is seen from figure 4.12 that in general the azimuth histories will be wrapped around in the frequency domain.

In the frequency domain it is thus possible to straighten the trajectories, since this will affect all targets along an azimuth vector simultaneously,
while a similar straightening in the time domain would be valid only for a single target.

How well this straightening procedure works, depends on how accurate the relationship between the time and frequency domain representations is for a linear FM-pulse. This again depends on the time-bandwidth product of the chirp-pulse. It is seen from figure 4.11 that it is the frequency bands at the ends of the azimuth history that have the smallest time bandwidth products. In the case of SEASAT-1 these time bandwidth products are well above the value necessary to make the straightening procedure work well. If the skewing procedure for the range compressed data matrix were not to be done, this would have made the time-bandwidth products much smaller than when skewing is applied.

As shown in section 3.2.2 the same azimuth filter can be used for a range interval of 1105 m if 1 dB degradation is tolerated. This corresponds to 167 complex range samples. After skewing an azimuth vector does not any longer represent a single range cell. With a doppler centroid of 1000 Hz and a processing block of 4096 samples in azimuth, the additional range interval is 33 complex samples. To accommodate different doppler centroid values the same azimuth filter is therefore used for a block of 100 azimuth vectors for the processing done here. This gives a total of 61 azimuth filters across the whole swath. These are calculated and stored before the azimuth compression begins. Section 4.4 explains how they are generated.

To reduce the speckle in the images, the azimuth history is divided into four looks, the output images of which are added incoherently as explained in section 3.4. Each look has a duration of 1024 complex
samples and neighboring looks have 1/3 overlap. No weighting is applied to match the radar antenna diagram, but to reduce the sidelobe level each look is weighted with a window function as shown in figure 4.13. The frequency spectrum of a weighted look is shown in figure 4.14.

Figure 4.13 Window function applied to each look

Figure 4.14 Frequency spectrum for one look

The output for one look and a single target is given in figure 4.15. The sidelobe level is -20 dB, and
the 3 dB width of the peak corresponds typically to an azimuth resolution of 25 m.

Figure 4.15 Point response of azimuth filter

A block diagram describing the azimuth compression operation is given in figure 4.16.

Figure 4.16 Block diagram for azimuth compression
The range compressed data matrix on disc consists of a set of complex azimuth vectors, \( x_1, j(i), i = 0, N-1, j = 0, J-1 \), where \( j \) is a range index, increasing \( j \) meaning increasing range. \( N = 4096 \) has been used in this work.

Fourier transformation of an azimuth vector gives

\[
x_2, j(i) = \sum_{k=0}^{N-1} x_1, j(k) \cdot \exp(-j \frac{2\pi k}{N}), \quad i = 0, N-1, \ j = 0, J-1
\]

The range migration correction straightens the azimuth histories as shown in the following. \( M \) is chosen such that the number of available transformed azimuth vectors in memory, \( M+1 \), is great enough to include the whole trajectory plus a few extra azimuth vectors needed in the interpolation.

The doppler frequency history of a target is

\[
f(t) = \frac{1}{2\pi} \frac{d\phi(t)}{dt} = \frac{2 \frac{d}{dt}[\left( R_0^2 + v^2 t^2 \right)^{\frac{1}{2}}]}{\lambda} = \frac{2v^2}{\lambda} t \left( R_0^2 + v^2 t^2 \right)^{-\frac{1}{2}} \text{,} \quad T_1 < t < T_2
\]

If the doppler centroid is \( f_d \), the corresponding time \( t_d = \frac{1}{2}(T_1 + T_2) \) is given by \( f(t_d) = f_d \), or

\[
t_d = \frac{\lambda f_d R_0}{2v^2} \left( 1 - \frac{f_d^2}{4v^2} \right)^{-\frac{1}{2}}
\]

The inclination of the interpolation line used during range compression is

\[
a = \frac{d}{dt} x(t) \bigg|_{t=t_d} = \frac{\lambda}{2} f(t_d) = \frac{\lambda \cdot f_d}{2}
\]
α then determines the values of m and τ in (4.1). The remaining range migration after interpolation in range is

\[ \Delta r(t) = r(t) - \alpha \cdot t - [r(t_d) - \alpha \cdot t_d], \quad T_1 < t < T_2 \]

If the time-frequency correspondence for a linear FM-pulse is assumed preserved after Fourier transformation, the range migration as a function of frequency is

\[ \Delta r(f) = \Delta r(t(f)) = \Delta r(t = \frac{\lambda R_0 f}{2v^2} (1 - \frac{\lambda f_2^2}{4v^2})^{-\frac{1}{2}}) \]

The range migration is then corrected by means of a 4-point interpolation filter in range to fit the curved target trajectory. This is done according to the formula

\[ x_{3,j}(i) = \sum_{n=0}^{3} x_{2,j-M+n+K(i)}(i) \cdot \frac{\sin[(n-1-\tau)\pi]}{(n-1-\tau)\pi}, \quad 0 \leq \tau < 1 \]

\[ i = 0, N-1 \]

\[ K(i) = \lfloor \frac{\Delta r(i)}{\delta r} \rfloor_{\text{int}}, \quad \tau = \frac{\Delta r(i) - \delta r \cdot K(i)}{\delta r} \]

where \( \lfloor \cdot \rfloor_{\text{int}} \) means integer part. \( \delta r \) is the sample spacing in range measured in meters, and

\[ \Delta r(i) = \Delta r(f = \frac{PRF}{N}i) \]

If the actual azimuth frequencies are outside the band \([0, PRF]\), this must be accounted for when \( \Delta r(i) \) is calculated.
Referring to figure 4.11, \( \delta_r \) is the spacing between the azimuth lines, \( K(i) = 0 \) inside the mid rectangle, increasing with \( 1 \) when the trajectory enters a new set of rectangles, and \( \tau \) measures how close the trajectory is to the upper edge of the rectangle.

The frequency components in \( x_{3,j} \) outside the 4 looks are set to zero.

Multiplication with the matched filters for the four looks gives

\[
x_j^k(i) = x_{3,j}[i+\ell(k)] \cdot H_k(i), \quad i = 0, \frac{N}{4}-1, \quad k = 1, 4
\]

where \( \ell(k) \) is determined by where in the frequency region \([0, \text{PRF}]\) look no. \( k \) lies. The index \( i+\ell(k) \) is calculated modulo \( N \).

Transformation back to the time domain

\[
y_j^k(i) = \frac{1}{N} \sum_{n=0}^{N-1} x_j^k(n) \cdot \exp\left(j\frac{2\pi}{N}in\right), \quad i = 0, \frac{N}{4}-1, \quad k = 1, 4
\]

If the number of undistorted output values in \( y_j^k \) is \( L \), and neighboring looks overlap with \( 4L_1 \) samples (with sampling frequency equal to the PRF), \( L_1 < \frac{L}{3} \), look averaging gives

\[
\begin{align*}
|y_j^1(i)|^2, & \quad 0 \leq i \leq L_1 - 1 \\
|y_j^1(i)|^2 + |y_j^2(i)|^2, & \quad L_1 \leq i \leq 2L_1 - 1 \\
|y_j^1(i)|^2 + |y_j^2(i)|^2 + |y_j^3(i)|^2, & \quad 2L_1 \leq i \leq 3L_1 - 1
\end{align*}
\]
The corresponding azimuth vector from the previously processed block for the same range interval but a different azimuth interval is read from disc and \( z_j \) is added to this vector at the correct azimuth position and it is written back to disc. The image file on disc therefore always contains an image that is tapered in number of looks towards the end in the azimuth direction.

In section 4.2 an example of a rawdata image and the corresponding image after range compression was given. In figure 4.17 the range compressed image is repeated for comparison purposes, and the corresponding image after azimuth compression is shown in figure 4.18. The fully processed image has a resolution of 25mx25m and contains 4 looks. To make the fully processed image requires 7 times as much range compressed data in the azimuth direction as is shown in figure 4.1, but the relationship between the range compressed image and the mid section of the fully processed image can be seen.

Several examples of images processed to 25mx25m resolution and 4 looks are shown in figures 4.19-4.23.

It was earlier found that the signal to noise ratio for an average sea model after fully coherent processing is 15 dB. This is also the signal to noise ratio after
Figure 4.17 Range compressed data block

Figure 4.18 Datablock in figure 4.17 compressed in azimuth to 4 looks
Azimuth

Oslofjorden, scale 1:330000
25mx25m resolution, 4 looks

Figure 4.19 Example of digitally processed SAR image
Figure 4.20  Example of digitally processed SAR image
Arhus, Denmark, scale 1:330000
25mx25m resolution, 4 looks

Figure 1.21 Example of digitally processed SAR image
Figure 4.22 Example of digitally processed SAR image

Sunnfjord, scale 1:330000
25mx25m resolution, 4 looks

azimuth
Fredrikstad, scale 1:330000
25mx25m resolution, 4 looks

Figure 4.23 Example of digitally processed SAR image
Figure 4.24 Effect of accuracy in input data
the receiver for a pulse echo. Quantization on amplitude basis requires $\frac{15 \text{ dB}}{6 \text{ dB}}$ signbit $= 3.5$ bits. Figure 4.24 shows an image processed in 4 looks with different numbers of bits in the input signal. The same image with 5 bit input is the 4 look image in figure 3.23. Although the quantization with so few bits cannot be regarded as introducing additive noise, the high compression ratio tends to make the effect similar to the effect of additive noise, as can be seen from figure 4.24. It can also be seen that the estimate of 3.5 bits in the input to avoid quantization effects is quite good. Because of additional noise introduced, quantization to for example 1 bit to simplify processing might seriously degrade performance of the target detection procedure developed in chapter 5.

4.4 Autofocusing for generation of azimuth filters

Before the azimuth processing can be done, one has to have an estimate of the doppler centroid and of the relative velocity between the targets and the radar platform. The doppler centroid is estimated before the range processing, so only the relative velocity $v$, which determines the chirp rate of the azimuth history, remains to be estimated.

When $v$ is known, the azimuth history is known as a function of range. The matched filters for the 4 looks are then generated in the time domain by multiplying the azimuth history with the window function shown in figure 4.13, for 4 different, partially overlapping sections of the azimuth history, each 1024 complex samples long. The matched filters are then the complex conjugated Fourier transforms of the looks' time responses. These matched filters are all generated and stored before azimuth compression begins.
One way to estimate \( v \) is to calculate it from the orbit parameters of the satellite. The drawback with this procedure is, however, that this information has to be supplied separately from the synthetic aperture radar data collection itself, and this may take considerable time. Depending on the time delay, which for SEASAT-1 is on the order of days, this may exclude any operational use of the data. To avoid this difficulty an autofocus procedure is developed here, which means that only the data from the synthetic aperture radar is used to generate an estimate of the relative velocity \( v \).

The autofocus is based on maximization of the contrast in the image. The validity of this criterion can be seen as follows. If the azimuth compression is done with an incorrect value of the relative velocity \( v \), which means wrong chirp rate in the azimuth history, this will result in a smearing of point targets in the image. This smearing is equivalent to a reduction in contrast. It is therefore assumed that maximum contrast occurs for the correct value of \( v \).

The greater the azimuth bandwidth used in the compression, the more sensitive the contrast is to changes in \( v \). Compression to 4 looks is therefore used also for the contrast calculation. This has the advantage that the routine used for azimuth compression of the whole image also can be used here.

The contrast \( C \) in an image is defined by (3.4), or equivalently

\[
C = \left[ \frac{\langle I^2 \rangle - \langle I \rangle^2}{\langle I \rangle} \right]^{\frac{1}{2}}
\]

where \( I \) is the intensity and \( \langle \rangle \) denotes expectation.
The estimate for $v$ is obtained by calculating the contrast curve $C(v)$, i.e., contrast as a function of $v$, for a suitable image area and choosing the value of $v$ that corresponds to the peak in the contrast curve. To get a reliable estimate for $v$, the area used for calculation of the contrast contains several thousand image points.

The contrast will be higher and the peak more pronounced for scenes containing strong specular points than for scenes with little or no strong point scatterers. Scenes characterized by fully developed speckle, where there are many weak specular scatterers inside a single resolution cell, should not be expected to show any pronounced peak in the contrast curve because the intensity distribution in this case, given in section 3.4, does not depend on focusing of the image.

Figure 4.25 shows four contrast curves obtained experimentally from one image. The two curves in the middle are from neighboring, nonoverlapping land areas. One of the areas is seen to have higher contrast than the other one, and the peak is more pronounced. The peaks occur, however, for the same value of $v$ for both areas, as was expected.

The lowest curve is for a homogeneous sea area. As shown in section 3.4, such an area is very well described by fully developed speckle. According to what was said above, this should not be an area suitable for focusing, and the contrast curve is indeed seen to be flat and close to the theoretical value of 0.5 for a four-look image.

Contrast curves for sea areas that have a pronounced structure have not been calculated and conclusions about focusing properties of such areas therefore cannot be drawn.
The uppermost curve in figure 4.25 is for a strong ship target. An autofocusing procedure based on such targets is not likely to be feasible, since it would require that the target was detected before focusing, and since it generally will be moving and therefore give a wrong estimate for \( v \) for generation of the whole image. It might, however, be possible to use the method to get an estimate of the targets azimuthal velocity. If the range component could be estimated for example from the ship's displacement from the wake, the total velocity vector could be estimated.
The ship is moving in the azimuth direction, but in the opposite direction of the satellite, and the peak in the contrast curve should therefore be expected to occur for a somewhat higher velocity than for the land areas. The curve is, however, seen not to have the same regular shape as for the land areas, and the peak corresponds to a target velocity between 35 and 40 knots, which is not likely to be true. Some kind of mean value obtained from the contrast curve might give a reasonable velocity, but no good justification for such a procedure is seen. What can be observed, however, is that the contrast curve is displaced in the correct direction relative to the curves for land areas. This curve is also necessarily calculated for considerably fewer image points than for the land areas, and the statistical properties therefore cannot be expected to be very good.

The autofocusing procedure has been tried for many cases and has been found to work well for land areas. When imaging coastal areas, the land areas can therefore be used for focusing which then also will be valid for the sea areas. By using a precise model for the satellite's orbital motion it might be possible to maintain correct focus setting during long passages over sea areas by updating the orbit parameters during passages over land.

A suitable search algorithm will make it necessary to calculate only relatively few points on the contrast curve before the estimation procedure converges, thereby minimizing the computational burden imposed by the autofocusing.
The ultimate goal when using synthetic aperture radar in maritime surveillance is not the images as such, but rather a detection of targets together with an indication of the targets' positions. When the targets are ships, they are approximately point targets. The detection procedure chosen is therefore maximization of the signal to noise ratio for each point. This leads to matched filtering where the filter is matched to the point target response. This is the processor developed earlier in this work, and the output can be interpreted as images. Target detection is then done in the images on an intensity basis.

The method of addition of different statistically independent looks was used to reduce the speckle in the images. This was done because it has a positive effect when the images are regarded by the human eye, and it also gives a better radiometric resolution, but it is not necessarily advantageous when applying an automatic detection procedure to the images. The effect of look addition is considered below.

The detection of ships in images can be done in several ways. One method is to let a human operator perform visual detection in processed images. Figures 3.14, 3.15 and 4.19-4.23 show several examples illustrating that this is fully possible.

This detection operation should not be done from photographs because of the poor dynamic range of intensity in photographic paper. Use of photographs would also incur an extra time delay which for operational purposes would be highly undesirable. In general it can be said that a photographic reproduction of a synthetic aperture radar image
that is scaled to give a pleasant view of land areas will, depending on sea state among other things, highly suppress strong point targets such as ships because of the poor dynamic range in photographs.

It is seen from the examples mentioned that an interesting way of detecting moving ships is by detection of the wake of the ship rather than of the ship itself, or both. With a human operator this is easily achieved.

While the detection of ships in images by means of the ship itself, or its wake, is easily combined in the case of a human operator, these two methods of target detection are completely different in the case of automatic target detection. A method based on detection of wakes in an image would be a pattern recognition technique, and this is not considered here. Besides, a wake is a distributed phenomenon, the detection of which might be done more optimally than by pattern recognition in an image. Another method might be less time consuming than full processing of images, and this would be very important in the case of operational use.

As mentioned above, the detection procedure developed here is based on the target's intensity in the image. In statistical terms this problem is called the binary hypothesis testing problem, (17). If $I$ is the intensity in a given image point, $s$ is the intensity caused by a target and $n$ is the intensity of the sea, we have to test $I$ against the following two hypotheses

\begin{align*}
H_1: &\quad I = s+n \\
H_0: &\quad I = n
\end{align*}

If $H_1$ is true, the point is a combination of a target and the sea, and if $H_0$ is true, there is no target in
the point. Depending on the size of the target and how much of the image point that is occupied by the target, an alternative set of hypotheses could be

\[
H_1': I = s \\
H_0': I = n
\]

In this case we have either a target or the sea in each image point and never a combination of both. Both sets of hypotheses will be treated.

The two most used test criterions are the ones due to Bayes and Neyman-Pearson, (17). The Bayesian approach requires a priori probabilities of there being a target or not and an assignment of costs to the four different combinations of two outcomes and true and false hypothesis. Because of the problems in assigning reasonable values to both, the Neyman-Pearson criterion is usually applied in radar detection problems.

When the Neyman-Pearson test is used, only two of the four possible combinations of outcomes and input hypotheses are used. These are the probability of detection, \( P_D \), which is the probability of saying that \( H_1 \) is true when it is true, and the probability of false alarm, \( P_F \), which is the probability of saying that \( H_1 \) is true when \( H_0 \) is true. The criterion used is to maximize \( P_D \) subject to the constraint \( P_F \leq \alpha \), i.e. the probability of detection is maximized while the probability of false alarm is kept below a certain threshold. This method thus emphasizes the importance of not giving false alarms.

It can be shown, (17) and (18), that both the Bayes criterion and the Neyman-Pearson criterion lead to a likelihood ratio test. The likelihood ratio is defined as
\[ \lambda(I) = \frac{p(I|H_1)}{p(I|H_0)} \]

that is the ratio between the conditional probabilities of the observed intensity given \( H_1 \) and \( H_0 \), respectively. The decision algorithm is then

If \( \lambda(I) \geq \eta \), say \( H_1 \) is true
If \( \lambda(I) < \eta \), say \( H_0 \) is true

The value of the threshold \( \eta \) depends on the values chosen for \( P_D \) and \( P_F \). The detector is thus a processor that generates the likelihood ratio \( \lambda(I) \) for each image point and decides which hypothesis is true by comparing \( \lambda(I) \) with the threshold \( \eta \).

To find the performance of the detector it is necessary to calculate the likelihood ratio \( \lambda(I) \) and the so-called "receiver operating characteristics" curves of \( P_D \) versus \( P_F \), for the different cases. We thus need the probability distributions for the intensity subject to the two hypotheses, and for different cases such as different numbers of looks.

The probability distribution for the intensity in a onelook image of a sun area is supposed to be given by the probability distribution for fully developed speckle. This distribution was given in Section 3.4, and if it is normalized by setting \( I_0 = 1 \), we get

\[ p_1(I|H_0) = \exp\{-I\}, \quad I \geq 0 \text{ and zero elsewhere.} \]

The subscript indicates the number of looks. The addition of several looks is equivalent to the addition of several independent, identically distributed stochastic variables. The resulting probability distribution is then the convolution of the individual distributions. This is most easily accomplished
mathematically by applying the moment generating function which is defined by

\[ M(s) = \int_{-\infty}^{\infty} p(x) \cdot \exp(-sx) \, dx \]

where \( p(x) \) is the probability distribution. The moment generating function for a sum of independent stochastic variables is the product of the individual moment generating functions. Since we here always have \( p(x) = 0, \, x < 0 \), the moment generating function is equal to the onesided Laplace transform of \( p(x) \).

We thus get

\[ M_1(s) = \int_0^\infty p_1(I|H_0) \cdot \exp(-sI) \, dI = \int_0^\infty \exp(-(s+1)I) \, dI = (s+1)^{-1} \]

The moment generating function for \( n \) looks is

\[ M_n(s) = (s+1)^{-n} \]

Inverse Laplace transform gives

\[ p_n(I|H_0) = \frac{1}{\Gamma(n)} I^{n-1} \cdot \exp(-I), \quad n \geq 1, \quad <I> = n \]

where \( \Gamma(n) \) is the gamma function.

Since we have \( H_0 = H_0' \), we also have \( p_n(I|H_0) = p_n(I|H_0') \)

To find \( p_n(I|H_l) \) and \( p_n(I|H_l') \) we must know the probability distribution of \( s \), i.e., the target's intensity distribution. Because of the inherent difficulties in defining this distribution, the most reasonable probability distribution is assumed to be a uniform distribution. We thus have
\[ p(s) = \frac{1}{S_0}, \quad 0 \leq s \leq S_0 \]

The significance of the choice of \( S_0 \) will be commented upon later.

When adding looks, the targets' intensities and the intensity of the sea background will be statistically independent, and the distribution \( p(I|H_1) = p(I|s+n) \) is the convolution of the probability distributions for target intensity and sea intensity. When processing images it has turned out, however, that there is strong interdependence between different looks for strong targets. This is not unexpected since a strong target may consist of perhaps only a single strong reflecting point. The intensity distribution for a target after look addition is therefore not obtained by convolution of the single look distribution with itself several times, but is better modelled by a uniform distribution even after look addition, properly scaled according to the new mean value.

By scaling all the speckle distribution functions to have mean values equal to 1, we get the following distributions.

\[ p_n(I|H_0) = p_n(I|H_0') = \frac{n}{I(n)} \cdot I^{n-1} \cdot \exp(-nI), \quad I \geq 0 \]

\[ p_n(I|H_1') = \frac{1}{S_0}, \quad 0 \leq s \leq S_0 \]

The distribution \( p_n(I|H_1) \) is given by the convolution of \( p_n(I|H_0) \) and \( p_n(I|H_1') \).

\[ p_n(I|H_1) = \frac{1}{S_0} \int_0^I p_n(x|H_0) \, dx = \frac{n}{S_0 I(n)} \int_0^I x^{n-1} \cdot \exp(-nx) \, dx \]

\[ = \frac{1}{S_0} \left[ 1 - \exp(-nI) \right] \sum_{m=0}^{n-1} \frac{(nI)^m}{m!}, \quad 0 \leq I \leq S_0 \]
The tail of the distribution $P_n(I|H_1)$ for $I > S_0$ is not calculated since it is not needed.

The likelihood ratios can now be calculated

$$\ell_n(I) = \frac{P_n(I|H_1)}{P_n(I|H_0)} = \frac{\Gamma(n)}{S_0 \cdot n^I} \cdot \frac{I^{-(n-1)} \cdot \exp\{nI\} - \sum_{m=0}^{n-1} \frac{(nI)^m}{m!}}{S_0 \cdot n^I}$$

$$0 \leq I \leq S_0$$

$$\ell_n'(I) = \frac{P_n(I|H_1')}{P_n(I|H_0')} = \frac{\Gamma(n)}{S_0 \cdot n^I} \cdot I^{-(n-1)} \cdot \exp\{nI\}, \quad 0 \leq I \leq S_0$$

Since

$$\exp\{nI\} = \sum_{m=0}^{\infty} \frac{(nI)^m}{m!}$$

we have

$$\ell_n(I) = \frac{\Gamma(n)}{S_0 \cdot n^I} \cdot I^{-(n-1)} \cdot \sum_{m=0}^{\infty} \frac{(nI)^m}{m!} = \frac{\Gamma(n)}{S_0 \cdot n^I} \cdot \sum_{m=0}^{\infty} \frac{(nI)^{m-n+1}}{m!}$$

We therefore have $\frac{d}{dI}[\ell_n(I)] > 0$ for $0 < I < S_0$ while for $\ell_n'(I)$ we have $\frac{d}{dI}[\ell_n'(I)] < 0$ for $0 < I < \frac{1}{n}$ and $\frac{d}{dI}[\ell_n'(I)] > 0$ for $\frac{1}{n} < I < S_0$. Also $\ell_n'(I) \to 0$ when $I \to 0$. This means that $\ell_n'(I)$ cannot be used for the interval $0 < I < \frac{1}{n}$.

The detector will therefore operate as follows for the different intensity intervals.

Hypotheses $H_0$ and $H_1$:

$0 \leq I \leq S_0$: $H_0$ true if $\ell_n(I) < n$

$H_1$ true if $\ell_n(I) \geq n$

$I > S_0$: $H_1$ true
Hypotheses $H_0'$ and $H_1'$:

$0 \leq I < \frac{1}{n} : H_0'$ true

$I - \frac{1}{n} \leq I < S_0 : H_0'$ true if $\ln' (I) < n$

$H_1'$ true if $\ln' (I) \geq n$

$I > S_0 : H_1'$ true

$n$ is the decision threshold.

To find the receiver operating characteristics it is necessary to calculate the probability of detection $P_D$ and the probability of false alarm $P_F$ as a function of the intensity threshold $\gamma$ and the number of looks $n$.

With $\ln (\gamma) = n$ or $\ln' (\gamma) = n$ we have

$$P_D = \int_{\gamma}^{\infty} p_n (I | H_1') dI$$

$$P_F = \int_{\gamma}^{\infty} p_n (I | H_0') dI$$

and correspondingly for the primed case.

$$P_D = \int_{\gamma}^{\infty} p_n (I | H_1') dI = 1 - \int_{0}^{\gamma} p_n (I | H_1') dI$$

$$= 1 - \int_{0}^{\gamma} \frac{1}{S_0} \left[ 1 - \exp \{-nI\} \sum_{m=0}^{n-1} \frac{(nI)^m}{m!} \right] dI$$

$$= 1 - \frac{\gamma}{S_0} + \frac{1}{S_0} \sum_{m=0}^{n-1} \frac{n^m \gamma}{m!} \int_{0}^{\infty} \exp \{-nI\} dI$$

$$= 1 - \frac{\gamma - 1}{S_0} + \frac{1}{S_0} \exp \{-n\gamma\} \sum_{m=0}^{n-1} \frac{m (n\gamma)^k}{k!}, \quad 0 \leq \gamma \leq S_0$$

$$P_D' = \int_{\gamma}^{\infty} p_n (I | H_1 '') dI = 1 - \frac{\gamma}{S_0}, \quad 1 - \frac{1}{n} \leq \gamma \leq S_0$$
The probability of detection $P_D$ is seen to depend strongly upon $S_0$, which means that a somewhat arbitrary choice of $S_0$ will have considerable effect upon the resulting probability of detection. Since one can write

$$P_D = 1 - \frac{1}{S_0} \cdot f(\gamma, n)$$

where $f(\gamma, n)$ is a function of $\gamma$ and $n$, the dependence of $P_D$ upon $n$, the number of looks, can be found by studying $f(\gamma, n)$ which does not depend on $S_0$.

The reason why the number of looks $n$ is an important parameter with regard to target detection is the relationship between the coherence time of a target and the coherent integration time in the processor. By coherence time of a target is meant the longest observation time during which the target response can be regarded as coherent. The peak response for a target will increase more strongly for an increase in coherent integration time when this integration time is shorter than the target's coherence time than when it is longer. The coherence time of the sea background, regarded as speckle noise, is of no importance in this case since the speckle distribution functions are independent of coherence time. Since the length of the synthetic aperture is fixed, there is a compromise between coherent integration time (integration time per look) and number of looks.
When comparing the receiver operating characteristics for different numbers of looks, this must be done with the same amount of input information in each case to get relevant results. This means that the total integration time used for a given number of looks must be the same for all cases compared. This total integration time, given by the length of the whole synthetic aperture, is assumed constant for all cases considered.

The formulas developed above are all derived by means of speckle distributions which have an expected value of 1. This represents an intensity scaling. \( S_0 \) is also the same in all the formulas.

As a parameter in the comparison of different cases a coherence factor \( k_c \) will be introduced. This factor is defined below.

If the case of a single look, i.e. \( n = 1 \), is described by an expected speckle intensity of 1 and a target intensity distribution that is uniform on the interval \( 0 \leq I \leq S_0 \), the case of \( n \) looks can be described by an expected speckle intensity of 1 and a target intensity distribution that is uniform on the interval \( 0 \leq I \leq k_c S_0 \). We will always have \( \frac{1}{n} < k_c < 1 \), where a high value of target coherence time relative to the coherent integration time in the processor corresponds to a low value of \( k_c \). This can be seen as follows.

Given an image which contains \( n \) looks. If the coherent integration time is increased to the case of one look, the peak intensity of the target relative to the sea in the onelook image will change from the case of \( n \) looks depending on the coherence properties of the target. As mentioned above the speckle distribution is independent of sea coherence time, and the mean value of the speckle intensity is proportional to the bandwidth of the chirp-filter when the input speckle noise is considered white.
in the frequency band of interest. Mean speckle intensity is therefore proportional to coherent integration time. If the target is coherent over the whole observation time, an increase in coherent integration time with a factor n will result in a peak intensity increase for the target with a factor \(n^2\). Since the mean sea intensity increases with a factor n, this gives \(k_c = \frac{1}{n}\). The other extreme case is when the target coherence time is shorter than the coherent integration time corresponding to n looks. In this case target peak intensity increases incoherently, i.e. proportionally with coherent integration time, and we have \(k_c = 1\).

Figures 5.1 and 5.2 show the receiver operating characteristics for \(k_c = 1\), \(k_c = \frac{1}{n}\) and \(n = 1,2,3,4\). \(S_0\) is chosen 100. Only the primed case is shown since the unprimed case exhibits the same principal behaviour.

The figures show that in the case of \(k_c = 1\), the maximum number of looks should be chosen, and when \(k_c = \frac{1}{n}\), one look should be chosen. This is intuitively logical since in the first case there is nothing to be gained by increasing the coherent integration time, while averaging of several looks will suppress the upper tail of the sea intensity distribution, thereby increasing the detection probability for a given false alarm probability. Likewise for \(k_c = \frac{1}{n}\), the target detection probability is highest when using a single look.

Since the order of the receiver operating characteristics in figures 5.1 and 5.2 for different values of n is seen to be reversed, there is obviously a value of \(k_c\) for which the receiver operating characteristics for two given numbers of looks are equal for any chosen value of false alarm probability \(P_F\).
To be able to choose the optimum number of looks, the coherence properties of typical targets must be known. These properties can be expressed by a function that shows how the expected peak intensity of a target changes with coherent integration time, that is integration time per look.

If $T$ is the total integration time, i.e., the length of the azimuth history, then $g(\tau) = \frac{1}{T}$ gives the normalized expected intensity of the sea, where $\tau$ is the coherent integration time, $0 \leq \tau \leq T$. $f(\tau)$ is the corresponding curve for a target, and it is closely related to the correlation function of the target's azimuth history. $f(\tau)$ can best be obtained empirically. Figure 5.3 shows $g(\tau)$ and a possible normalized $f(\tau)$.
Figure 5.2 Receiver operating characteristics with $k_c = \frac{1}{n}$

Figure 5.3 Principal behaviour of expected peak target intensity and expected mean sea intensity as functions of coherent integration time
The parabolic portion of \( f(\tau) \) gives the integration times for which the target response is highly coherent. The linear portion of \( f(\tau) \) represents an incoherent increase in peak intensity as the coherent integration time is increased.

To find the optimum number of looks the receiver operating characteristics must be plotted. The curves \( g(\tau) \) and \( f(\tau) \) will then determine the coherence factor \( k_c \), which is given as a function of \( n \) by the expression

\[
k_c(n) = \frac{f\left(\frac{T}{n}\right)}{g\left(\frac{T}{n}\right)} = n \cdot f\left(\frac{T}{n}\right)
\]

If the functional form of \( f(\tau) \) were known, the optimum number of looks could be calculated. Instead a single case with a ship target has been tried experimentally by processing to 1 and 3 looks, respectively. By comparing the peak intensity of the ship with the mean intensity of the surrounding sea, a coherence factor of 0.59 was found. Figure 5.4 shows the receiver operating characteristics for 1 and 3 looks with \( k_c = 0.59 \). It is seen from the figure that in this case it is advantageous to use 3 looks if a false alarm rate less than approximately \( 5 \cdot 10^{-3} \) is desired.

A generalization not considered here is to use overlapping looks. This would give partially correlated looks and the resulting receiver operating characteristics would be somewhat different from the ones derived above.

Since the total integration time is constant, the azimuth resolution will be degraded as the number of looks is increased. Depending on the target size,
this may lead to a suppression of the target if it becomes much smaller than the resolution cell. For a SAR system like the one onboard SEASAT-1 it seems unlikely that it should be advantageous to use so many looks that this might occur.
6 CONCLUSION

As has been shown in the previous chapters, a satellite borne synthetic aperture radar can very well be used for maritime surveillance purposes. It is seen from the examples of processed imagery that in many cases ships can be seen clearly against the sea background. An especially interesting phenomenon is the clear visibility of the ship’s wake, which probably appears more clearly and lasts longer than it would in images obtained in the optical frequency band.

The autofocusing procedure developed has been tried for many different cases and works well for land areas. Such a method is necessary when operational use of the SAR imagery is intended, and may, depending upon the quality of the estimation algorithms used for the satellite orbit parameters, be adequate for maintaining correct focus setting also for sea areas by updating the estimates during passes over land.

When using the SEASAT-1 SAR system to evaluate the usefulness of a satellite borne SAR system for maritime surveillance, it should be remembered that this system was not designed for such a purpose. Some of the design criteria were in fact the opposite of what would be chosen for surveillance purposes. The depression angle of \(70^\circ\) was for example, among other reasons, chosen because this gives a rather high backscatter coefficient for the sea. When looking for ships this results in a high sea clutter level which makes the target detection more difficult. A somewhat smaller depression angle would result in a lower sea clutter level.

The main problem when using a satellite borne SAR system for maritime surveillance, and digital processing of images, is the processing time. To be of any
operational value, the images would have to be processed within a few hours after the satellite had passed the area. The average processing rate would also have to be equal to the data acquisition rate. If the data rate is measured in km\(^2\)/sec, the acquisition rate for SEASAT-I was approximately 700 km\(^2\)/sec. The images shown here are processed with a general purpose computer and with a rate of approximately 60 km\(^2\)/hour = 0.017 km\(^2\)/sec. If the average acquisition rate for an operational system should be of the order of 10 km\(^2\)/sec, this would require an increase in processing speed with a factor 600.

The target detection algorithm developed in chapter 5 will require processing to obtain empirical values for coherence times for typical targets. This has not yet been done, but a single example is given which illustrates the effect of changing the coherent integration time.
DOPPLER CENTROID ESTIMATION

The doppler centroid of the azimuth frequency spectrum is estimated directly from a block of rawdata. The real samples are converted to complex samples by demodulating the range vectors so that the lowest sideband is centered around 0 frequency, and then lowpass filtering each range vector to reduce the sampling frequency with a factor 2. Each azimuth vector is then Fourier transformed, the spectrum is averaged over 100 azimuth vectors, each of length 512 complex samples, and the doppler centroid is estimated from the phase of the first harmonic of the frequency spectrum.

Rawdata block with real samples

$$a(i,j), \quad i = 0,208, \quad j = 0,511, \quad <a(i,j)> = 0$$

i is range index and j is azimuth index. The lowest sideband in range is centered at 11.38 MHz, which is the real sampling frequency divided by four. Demodulation gives

$$b(i,j) = a(i,j) \cdot \exp{-\frac{j\pi}{2} \cdot i}, \quad i = 0,208, \quad j = 0,511$$

Lowpass filtering with a truncated interpolation filter gives the rawdata matrix on complex form

$$c(i,j) = \sum_{k=0}^{10} b(2i+k,j) \cdot \frac{\sin[(k-5)\frac{\pi}{2}]}{(k-5)\frac{\pi}{2}}, \quad i = 0,99, \quad j = 0,511$$

Fourier transformation in azimuth

$$H_1(k) = \sum_{j=0}^{511} c(i,j) \cdot \exp{-\frac{j\pi}{256} k \cdot j}, \quad k = 0,511, \quad i = 0,99$$
Averaged frequency spectrum

\[ W(k) = \frac{1}{10} \sum_{i=0}^{99} |H_1(k)|, \quad k = 0,511 \]

First harmonic of \( W(k) \) is

\[ d = \sum_{k=0}^{511} W(k) \cdot \exp\{-j \frac{\pi}{256} k\} = r_d \cdot \exp\{-j\phi_d\}, \quad 0 \leq \phi < 2\pi \]

The doppler centroid estimate \( \hat{f}_d \) is then

\[ \hat{f}_d = \frac{\phi_d}{2\pi} \cdot \text{PRF} \]

PRF is the pulse repetition frequency of the radar. Through use of this estimation algorithm it seems that an accuracy of better than 100 Hz is obtained with PRF = 1647 Hz.
References


