High frequency surface wave radar (HFSWR) simulator

Teo, Mia Koon

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High Frequency Surface Wave Radar (HFSWR) Simulator

TEO MIA KOON

School of Electrical and Electronic Engineering

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SUMMARY

High Frequency (HF) - Radar ,also known as Over-The-Horizon (OTH) Radar is able to detect targets that are over the horizon up to a few hundreds kilometers, beyond the range of a conventional radar. The intrinsic properties of a HF signal enable the radio frequency (RF) signal to propagate along the earth surface.

Surface wave radar requires extensive work which might incur unnecessary cost and expenses especially if the setup requires some modifications. Hence it will be an advantage if a simulation of the setup as well as the system can be done prior to the actual work being done to reduce this uncertainty aspect of the project.

This research focuses on the simulation of the HFSWR system from the transmitting waveform, power received from targets, Signal to Noise ratio, surface wave attenuation, signal processing and modeling of the waves clutters otherwise known as Bragg clutters. Targets of various Radar Cross Section (RCS), movement and speed will be simulated and displayed on the Range Doppler (RD) Map which is similar to what a real radar system would display. In summary, this simulator will allow the user to simulate the antenna placement, transmitting strength, various parameters of the radar system hardware as well as targets so as to determine the best operating conditions and parameters for the HFSWR system.

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CHAPTER 1 INTRODUCTION

1.1 Background

When the United Nations Convention on the Law of the Sea (UNCLOS) gives coastal nations sovereign rights over 200 nautical miles (nm) of sea, known as Exclusive Economic Zone, nations are required to establish and maintain administration, law enforcement and environmental protections over this new frontier, which is many times larger than their previous 12 nm territorial limits. Traditionally microwave radars are used to monitor the 12 nm territorial. However with this new requirement, the area that needs to be monitored are much larger than the range of the microwave radar where the coverage is limited by the line of sight, sometimes only as far as 100 km due to the curvature of the earth as well as mountainous terrain.

High Frequency Radar is used to overcome this shortcoming of microwave radar. By using high frequency band (3-30 MHz), it is shown that the high conductivity of sea water results in relatively low attenuation of the vertically polarized HF radio wave thus extending the detection limits of the ground based sea surveillance systems quite substantially to cover over the horizon or beyond line of sight (BLOS). High Frequency radar can cover an area of up to 260,000 square kilometers, and the potential number of detected targets can be extremely high. Moreover, rain or fog does not affect HF signals.

Figure 1.1 illustrates a simple High Frequency Radar.

Figure 1.1. High Frequency Radar.

HF Radar (HFSWR) offers “best value” and is ready for immediate deployment to protect and survey the littoral waters. Applications include:

- Counterdrug interdiction
- Homeland security
- Law enforcement
- Maritime domain awareness
- Illegal immigration
- Search and rescue

Despite its long-range sensing capability, there still exist quite a number of issues that will affect the integrity of the returned signal. Extensive testing of the HF Radar system, which has been conducted over the past years [1], [2] has shown that system performance during daylight hours is satisfactory, but the night performance can be degraded by high external interference, range wrapped ionospheric clutter and external environmental noise. Also due to its low grazing angle and ‘rough’ surface of the sea surface, sea clutter is inevitably a problem. The ocean is a rough surface, with water waves of many different periods. When

the radar signal hits ocean waves that are 3-50 m long, that signal scatters in many directions. In this way, the surface can act like a large diffraction grating.

But the radar signal will return directly to its source only when the radar signal scatters off a wave that is exactly half the transmitted signal wavelength, and that wave is traveling in a radial path either directly away from or towards the radar. The scattered radar electromagnetic waves add coherently resulting in a strong return of energy at a very precise wavelength. This is known as the Bragg principle, and the phenomenon 'Bragg scattering'. This scattering will result in the masking of targets which are moving at or very close to Doppler frequency of this Bragg scattering.

1.2 Literature Review

The setup of the HF radar station is usually quite elaborate. Figure 1.2 shows an example of HF radar receivers.

Figure 1.2 Receiving Antenna of High Frequency Radar System

As can be seen from Figure 1.2, typical HF radar antenna arrays can be more than a kilometer in length and massive work need to be done for setting up of the

HF Radar system. Therefore it is important to perform simulations with the operating parameters before the real system is deployed.

The data received by the receiving antennas, after the Analogue-Digital conversion, mixer, low pass filter etc and with beam forming, are processed to show the Range-Doppler map (RD map). RD map is often a good indication of the target position as well as velocity. With appropriate processing algorithm, angular tracking is also achievable. Figure 1.3 on the next page shows a typical RD map of a HF radar.

Several papers have been written on the simulation of HF radar. Levent Segvi [3] came up with a paper on the simulation of detection and tracking in HFSWR. In his model, terrain data are fed into the simulator by means of a specially designed graphical user interface. Then a scenario is prepared where the radar's location, coverage and operational parameters can be defined, together with different targets and their sailing routes. Propagation paths are extracted and the propagation losses are calculated according to the terrain data. Target detection is then performed in the frequency domain after calculation of the noise floor, signal to noise ratio and clutter to noise ratio.

Figure 1.3 RD Map of the Received Data

J C revell and D J Emery [4] have also came up with a HFSWR simulation. Their simulation is split into a four stage process:

1. Data gathering exercise to build the scenario database
2. Generating Scenarios.
3. Evaluating potential tracks using a radar performance model
4. Displaying of tracks on the real time radar display

J C revell and D J Emery's simulation, like Levent Segvi's, dealt with high level display, which do not take into account signal generation, signal waveform and signal processing. There also isn't any display of the RD map which is not of concern in their case.

1.3 Research Objectives

The objective of this research is to build a High Frequency Radar simulation program using MatLab [5] for study of radar system and radar signal processing purpose. This simulator will allow user to:

- Simulate HFSWR with own parameters for transmitting power, antenna gain, pulse width, pulse length, bandwidth and sampling frequency.
- Simulate HFSWR's targets with own parameters for target radar cross section (RCS), bearing of movement, angle it makes with the radar and velocity.
- Simulate Bragg clutter under controlled environment.
- Perform signal processing algorithm like Doppler processing and pulse compression on received data.
- Perform digital beam forming on received data to search for target outside broad sight beam width.
- Track Targets Using Mono-pulse Amplitude Comparison.

Unlike the simulation proposed by J C Revell/ D J Emery and Levent Sergvi, this mode of simulation will allow user to generate a RD map of the data as received by the receiving antennas. This simulation will be split into 2 sub-simulation program:

1. Simulation of transmitting/receiving waveform and target(s).
2. Simulation of Bragg 1st order and 2nd order clutter.

The data generated by this 2 sub-programs will be combined together to form the full RD map.

Figure 1.4 shows the flowchart for the simulation program.

Figure 1.4. Flow Chart of Simulation Program.

CHAPTER 2 HFSWR SYSTEM

2.1 HFSWR System Overview

Radar is a method of using electromagnetic waves to sense the position, velocity and identifying characteristics of targets. This is accomplished by illuminating a volume of space with electromagnetic energy and sensing the reflected energy by objects. This reflected energy is then processed to provide the necessary information, namely: distance, velocity and tracks.

A HFSWR system likewise operates on the same blockset. In the HFSWR system, the transmitting and receiving antennas are usually different. A transmitting antenna is usually a single monopole with a reflector to direct the energy whereas the receiving antennas are usually in the form of arrays so that beamforming can be done. Figure 2.1 shows a common set-up of the HFSWR system.

Figure 2.1 HFSWR System Block Diagram

HFSWR system is also commonly seen as Pulsed Doppler radar as the radar transmits in pulses rather than continuous wave. A Pulsed Doppler radar utilizes the Doppler effect to determine the radial component of the target velocity and it does this by pulsed transmission. When the signal hits the target, the signal will return to the receiving antennas as pulses as well.

These received data will pass through a Band-Pass filter to filter out the unwanted frequencies. After which the data will be sent to the Analogue-Digital Converter card where the analogue data are sampled to digital data, down-convert to the baseband. The data are then demodulated to I and Q, the in-phase and quadrature phase.

Appropriate signal processing will be performed on these data, which are in I and Q format, and the output will be the RD map. A more in-depth write-up of these processes can be found in the later chapters.

2.2 HFSWR System Interference

Besides receiving reflected energy from potential targets, a radar system will also receive unwanted reflected energy or external induced energy known commonly as interference. The interference will complicate the detection and target measurement process and if large enough, they can even mask the desired targets altogether. They can also cause the measured target parameters to be in error. There are five basic kinds of interferences: Noise, Clutter, Electronic Countermeasures, Electromagnetic interference and Spillover.

Other than deliberate jamming, HFSWR system is generally more affected by clutter interference, electromagnetic interference and noise interference.

2.2.1 Clutter Interference

Clutter is describes as unwanted signal echo from sea, land, weather or any unintended objects. As in the case of HFSWR system which is used to detect targets rather than observing the sea state, the backscattering from the sea surface usually posed a great challenge to target detection, its primary usage.

The most prevalent clutter interference encountered by HFSWR system will be the sea clutter also commonly known as Bragg Lines. As mentioned briefly in the Introduction section, the ocean is a rough surface, with water waves of many different periods. When the radar signal hits ocean waves that are 3-50 meters long, that signal scatters in many directions. In this way, the surface can act like a large diffraction grating.

The return echo will exhibit the highest signal only when the radar signal scatters off a wave that is exactly half the transmitted signal wavelength, and that wave is traveling in a radial path either directly away from or towards the radar. The scattered radar electromagnetic waves add coherently resulting in a strong return

of energy at a very precise wavelength. This is known as the Bragg principle, and the phenomenon 'Bragg scattering'. Bragg scattering can be decomposed into 2 components: 1st order Bragg scattering and higher order Bragg scattering.

1st order Bragg scattering - The ocean surface wave can be regarded as a fold of set of sinusoidal components with a wavelength L . When the EM signal of wavelength λ travel from one crest to another where the distance is $\lambda/2$, the propagating distance of the reflected echo wave is λ , and the total echo signal is the fold of all echo signals with same phases. For ocean waves with other wavelength, the total echo signal is the fold of all echo signals with different phases. Because of this, the wave intensity of signal with same phases will be much larger than the echo signals with different phases. This will results in 1st order Bragg scattering where the echo signals are of great intensity.

Higher order Bragg scattering – Higher order Bragg scattering refers to the smaller peaks that exists around the 1st order Bragg at the Doppler spectrum plot. Usually the second order effects are more of the dominant contributors as compared to the 3rd order onwards and thus higher order scattering are sometimes also known as the 2nd order Bragg scattering. 2nd order scattering arises from the interaction of the ocean waves with different wavelengths thus producing irregular backscattering with no prominent addition of phase.

Figure 2.2 shows a Doppler spectrum of the Bragg scattering with 1st order Bragg and higher order (2nd order) Bragg [6].

Figure 2.2. Doppler Spectrum of Bragg Scattering [6].

Beside the Bragg scattering clutter, another clutter is known as the ionospheric clutter. Ionospheric clutter is the result of backscattering of the signal from the electrons that is present in the ionosphere layers.

The ionosphere is that region of the upper atmosphere of a planet where charged particles (electrons and ions) of thermal energy are present, which are the result of ionization of the neutral atmospheric constituents by electromagnetic and corpuscular radiation. The lower boundary of the ionosphere coincides with the region where the most penetrating radiation (generally, cosmic rays) produce free electron and ion pairs in numbers sufficient to affect the propagation of high frequency waves (D-region). The upper boundary of the ionosphere is directly or indirectly the result of the interaction of the solar wind with the planet. During a certain period of the year, there will be a high amount of solar flare activity with results in geomagnetic disturbance and thus there is a higher chance of the HF/SWR system being cluttered with ionospheric clutter.

Unlike the Bragg scattering where the 1st order Bragg clutter is spread in ranges, ionospheric clutter is usually characterized by the spread in Doppler. Figure 2.3 shows a typical ionospheric clutter in RD map.

Figure 2.3 RD Map of Ionospheric Clutter

As it can be seen from the figure, ionospheric clutter also posed a big challenge for HF radar designer as the clutter will also be able to mask potential targets in that particular range.

2.2.2 Electromagnetic Interference (EMI)

Electromagnetic Interference (EMI) is accidental or unintentional interference from friendly or neutral sources, such as radar, communications systems, and friendly jammers. These systems, during their operation, transmits signal with the same operating frequency or bandwidth that overlaps the carrier frequency of the HFSWR systems and thus resulting in EMI. This is shown on the RD map as a streak of line across the range cell at one particular frequency.

HFSWR system at lower frequency is more susceptible to EMI especially during at night because of the atmospheric conditions. During the day, HF users who utilize sky wave propagation will choose a higher frequency to avoid the large attenuation of signal by the D layer in the atmosphere. However during the night, this D layer disappears and thus user will prefer a lower frequency to increase

the skip zone, furthering the propagation distance. As a result, HFSWR system operating at lower frequency will encounter a number of EMI from other HF communications sources especially those who utilize the sky wave.

2.2.3 Noise Interference

The effect of noise on the HFSWR system cannot be ignored as it is generally external noise limited, that means the external noise are more dominant than the internal noise caused by the system. Noise is caused by the random motion of electrically charged particles which occurs within the system itself as well as external sources. At higher Doppler frequencies, the main constituent of the HF spectrum is the noise component. Depending on the factors listed above, the noise could be dominated by one or more of the following three sources: galactic, atmospheric and man-made, i.e. $\text{Total noise} = \text{Galactic Noise} + \text{Atmospheric Noise} + \text{Man Made Noise}$.

In HFSWR system, the receiving side can be designed to be external noise dominant. Radio noise external to the HSFWR system derives from the following causes [7]:

- radiation from lightning discharges (atmospheric noise due to lightning);
- unintended radiation from electrical machinery, electrical and electronic equipments, power transmission lines, or from internal combustion engine ignition (man-made noise);
- emissions from atmospheric gases and hydrometeors;
- the ground or other obstructions within the antenna beam;
- radiation from celestial radio sources.

The International Radio Consultative Committee (CCIR) [7] has compiled extensive data regarding these noise sources. Median values of man-made noise power for a number of environments are shown below in Fig 2.4.

Figure 2.4. Median values of man-made noise power [7].

In all cases, the results are consistent with a linear variation of the median value, F_{am} with frequency f of the form

$$F_{am} = c - d \log (f) \tag{2.1}$$

Where f expressed in MHz, c and d values taken from the table listed below.

Environmental category	c	d
Business (curve A)	76.8	27.7
Residential (curve B)	72.5	27.7
Rural (curve C)	67.2	27.7
Quiet rural (curve D)	53.6	28.6
Galactic noise (curve E)	52.0	23.0

With Singapore at a latitude of 1.3° North and 104.3° East, plots of the CCIR calculated noise level are shown from Figure 2.5 to Figure 2.10 for the HF band of 5, 10, 15, 20, 25 and 30 MHz. The four seasons in each figure refers to the timeline of the year. i.e. Winter (October – December), Spring (January – March), Summer (April – June) and Autumn (July – September).

Figure 2.5. External Noise Level for 5 MHz.

Figure 2.6. External Noise Level for 10 MHz.

Figure 2.7. External Noise Level for 15 MHz.

Figure 2.8. External Noise Level for 20 MHz.

Figure 2.9. External Noise Level for 25 MHz.

Figure 2.10. External Noise Level for 30 MHz.

It is important to note that the noise level do have a slight change over the different months of the year and thus it is important to take the month of the year into consideration when deciding the noise level. The noise for the simulation will be simulated using the Gaussian noise function with the level that is specific to the time and frequency. It can be seen that as the frequency of the carrier goes up, the noise level will also decrease. This phenomenon is generally due to the nature of the HF band where lower frequency tends to propagate further than higher frequency due to the reflection of the atmospheric condition. Unwanted carrier frequencies sent by other stations will become interference and noise.

2.3 High Frequency Surface Wave Propagation Loss

The propagation of HF consists of two waves. The sky-wave, that bounces from the ionosphere, and the surface-wave, that propagates along the ground. Because of the long wavelengths, HF communications are not effected by the

troposphere, and can usually bend around even large objects such as hills. In the case of sky-wave propagation, the ionosphere permits communication over great distances. The ionosphere is not dependable, however, and signal strengths can vary considerably -- effect is maximum in early evening, explaining the overcrowding of the medium wave-band in Britain with European transmissions in the evening.

The ground wave has both direct-wave and ground-reflected components, and under certain conditions a tropospheric ducting component. The direct-wave is limited only by the distance to the horizon from the transmitter plus a small distance added by atmospheric diffraction around the curvature of the earth. The ground-reflected portion of the radiated wave reaches the receiving antenna after being reflected from the earth's surface. There is also an induced ground-hugging surface-wave component known as the Norton surface wave. This wave is the result of electrical currents induced in the ground by refraction of a portion of the reflected-wave component at the earth-atmosphere interface. Upon reflection from the Earth's surface the reflected wave undergoes a 180deg phase reversal. When both transmitting and receiving antennas are on, or close to, the ground, and the distance between them becomes great, the direct and reflected components tend to cancel out, and the resulting field intensity is principally that of this surface wave. Because part of its energy is absorbed by the ground, the electrical intensity of the surface wave is attenuated at a much greater rate than inversely as the distance. It is the conductivity of the underlying terrain that determines the attenuation of the surface-wave field intensity as a function of distance.

As the surface wave passed over the surface of the earth, it is attenuated as a result of the energy absorbed by the earth due to the power loss resulting from the current flowing through the Earth's resistance. Energy is taken from the surface wave to supply the losses in the ground, and the attenuation of this wave is directly affected by the ground constants of the earth along which it travels.

The attenuation function or propagation loss factor is the ratio of the electric field from a short dipole over the lossy's earth's surface to that field from the same short dipole located on a flat perfectly conducting surface. At distances within the line of sight the surface field strength, E is given by

$$E = \frac{AE_o}{d} \quad (2.2)$$

where E_o is the electric field strength of the surface wave of a flat perfectly conducting surface, d is the distance between the transmitting and reception point and A is the attenuation function that take into account the ground losses.

There are a few ways to compute the attenuation factor of the surface wave propagation namely the residue series, small curvature expansion, power series and flat earth attenuation function.

David A. Hill [8] did a comparison of the different attenuation factor and found that for ranges greater than 10km, which is the norm for HF/SWR system, residue series with 50 terms provides the greatest accuracy of attenuation factor. The small curvature expansion is valid for d less than 50km and the power series is valid for d less than 1km. The flat earth approximation which is the first term for small curvature expansion, is valid for d less than 15km. Thus, residue series appears to be the most accurate formula in determining the attenuation factor of the surface propagation. Donald E Barrick[6] came up with a further extension of this attenuation factor based on the effective impedance of the sea water with relation to the surface roughness. Using the basis of Leontovich boundary condition, Barrick derive the effective surface impedance at grazing incidence of a slightly rough surface.

The mathematical formula for residue series as well as for the effective surface impedance will be discussed in the later chapter.

2.4 Signal Processing

Signal processing is usually performed on the data collected from the receiving antennas. Its objectives are:

- To improve both the signal-to-interference ratio and the detection of targets
- To make the radar less vulnerable to targets hiding in clutter
- To make the radar less vulnerable to ECM, and
- To extract information about the characteristics and behavior of the targets

There are 3 basic processes [9] in signal processing to separate the desired signal from the interference signal such that the echoes from the desired signal are enhanced and the echoes from the interference are suppressed. They are 1. Signal integration, 2. Correlation, and 3. Filtering and spectrum analysis.

- 1) Signal integration sums composite signals within the same range bin for several pulses. Desired targets, which are usually in-phase in the same range bin, will produce a larger sum than random data which often are noise interference.
- 2). Correlation is the process of measuring the similarity between two functions. Signal-plus-interference is compared to an ideal signal. The degree of match gives the likelihood that the composite signal contains a target echo. Pulse compression is the process of correlation in the time domain.
- 3). Filtering and spectrum analysis is done in the frequency domain where targets echo and interference are separated based on their Doppler

component. The components which concentrated into one or a few bins (in time and Doppler) are usually specific targets echo where as for noise or noise-like interference, they are spread evenly in the frequency domain. As the targets are usually moving away from or towards the radar, the non-zero Doppler will be separated from the clutter which are usually stationary at zero Doppler.

Due to the special operation mode of HFSWR system, the signal processing techniques are different from that in the conventional microwave radar systems. First, in HF band, the cosmic noise, atmosphere noise and other artificial radio noise/interference are much stronger than the internal noise in radar receiver. Therefore the external noise is the main limitations for the maximum target detection range. In addition, the strong ground and sea clutter in the echo signal also restricts the target detection performance of HFSWR. For air target, whose Doppler frequency is high enough to be separated from the Bragg lines (sea clutter), the main detection limitation is the external noise level. For ship target, whose Doppler frequency is quite near to the Bragg lines, a very long coherent integration time (CIT) is necessary to improve the Doppler resolution and separate the ship target and sea clutter.

Figure 2.11 gives the block diagram of HFSWR signal processing scheme.

Figure 2.11. Block Diagram of HFSWR signal processing scheme.

A detailed description of the signal processing scheme used in the simulation can be found in the following chapters on the HSFWR simulation.

CHAPTER 3 HFSWR SIMULATION

3.1 Simulation of Pulse Doppler Radar with Targets

This is the main program used to simulate the HFSWR in the transmission as well as the receiving. As with the real radar, parameters used to describe the waveform as well the sampling frequency of the receiver card are input parameters in this simulation. They are namely: carrier frequency, transmission bandwidth, sampling frequency of the generator, sampling frequency of the receiver card, pulse width, pulse repetition interval, transmitter antenna gain and transmitter power. Receiver parameters will include: number of receiving elements, spacing of each elements, mixer loss, frequency of local oscillator, output power of local oscillator, cut off frequency of low pass filter and receiver antenna gain.

3.1.1 Signal and Pulse Generation

The signal waveform that is used in this simulation is the Linear Frequency Modulated waveform. (LFM). For LFM, the frequency is swept linearly across the pulse width in the upward direction. The advantage of using this modulation technique is to achieve large pulse compression ratio so that the radar is able to transmit a wide pulse (for energy and detection) and processing it to a narrow pulse (for range resolution).

A typical LFM waveform, $S(t)$, can be expressed as the following

$$S(t) = A(t) \exp(j2\pi(f_o t \pm \frac{\mu}{2} t^2) + \phi_o) \quad (3.1)$$

where $A(t)$ is the rectangular pulse envelope of pulse width τ , f_o is the radar carrier frequency, $\mu = (2\pi B)/\tau$ is the LFM coefficient, B is the transmission

bandwidth and ϕ_0 is the arbitrary phase angle. The $\pm$ indicates the up-chirp(+) or down-chirp(-).

In the simulation program, the transmitting waveform is simulated as:

$$S_{TX} = \cos(\pi(2f_0t + \frac{B}{\tau}t^2 - Bt)) \quad (3.2)$$

This signal is then multiply with the appropriate transmission power and passed through the pulse train modulator. In the pulse train modulator, pulses of chirp will be generate with the given pulse repetition interval. Figure 3.1 shows the plot of LFM wave form where the pulse width is 100 μ s. It can be seen that the frequency of the wave increase as the time increase.

Figure 3.1. LFM waveform of τ 100 μ s in time and frequency domain with $f_0 = 5$ MHz.

3.1.2 Target Generation

This portion will describe the simulation of the return echo of the transmitted signal after it has hit the target(s). Input parameters for the target include the number of targets, the distance of the target(s) to the radar, the angle of to the line of transmission of radar, the bearing of the target(s) movement (with respect to true north), and the speed of the target(s). A Matlab program used to calculate the updated bearing, distance and angle to the radar transmission. Power budget equation is used next to compute the power of the received signal to increase the fidelity of this simulation. The power budget equation used here is the conventional radar equation that is used across all frequencies of radar except for an additional attenuation factor which is specific to HFSWR.

3.1.2.1 Target Information

Target information comprises of the target(s) distance from radar, bearing of movement, angle to the radar transmission. Basing on the initial input of the velocity as well as the information listed above given by the user, this MatLab program will generate the updated information for each iteration of pulses. The newly generated pulse will take into account the previous time delayed, distance, updated angle as well as bearing of the target movement.

Doppler shift is created by the movement of the target, either away or towards the radar. When a target moves, the frequency of the return signal which hit the target, will be shifted slightly due to the movement of the target. Target radial velocity, v_r , is derived from the measured Doppler frequency from the following equation:

$$v_r = \frac{f_d \lambda}{2 \cos \theta} \quad (3.3)$$

where f_d is the Doppler frequency, λ is the wavelength of the carrier frequency and θ is the angle of the movement of the target that makes with radar's beam direction.

In the program, the Doppler frequency will be shown in the Range Doppler Map. There is no need of artificially adding this Doppler frequency into the pulse as the delay of the individual pulse has taken into account this phenomenon.

3.1.2.2 Target Radar Cross Section (RCS) Modeling

A primary characteristic of any radar target is its ability to return energy to the receiving antenna. The parameter used to describe this ability is known as the RCS of the target. Generally, a fix shape such as square, sphere will have a fix formula to calculate their RCS. However, for complex target with irregular shape such as aircraft, it is almost impossible to write a formula to calculate the RCS. Thus, to formulate the RCS of a complex target, Peter Swerling [10] in his classic paper on target modeling, established four statistical target RCS model:

Swerling case 1 - Targets exhibit large fluctuations at a slow rate. They are modeled as several scatterers of approximately equal radar cross-section. Case 1 is generally a good fit for complex targets such as aircraft to radars without pulse-to-pulse frequency agility, at either long ranges or flying toward or away the radar.

Swerling case 2 - Targets exhibit large fluctuations, occurring rapidly. This case is modeled the same as case 1. It fits complex targets where aspect is changing rapidly, as at close range. It also fits complex targets for radars with pulse-to-pulse frequency agility.

Case 1 and case 2 targets are described statistically as shown in figure 3.2, and their probability density is given in.

$$P(\sigma) = \left(\frac{1}{\mu} \right) e^{-\frac{\sigma}{\mu}} \quad (3.4)$$

where $P(\sigma)$ is the probability density of a certain RCS, σ is the RCS and μ is the median RCS. In the simulation, Swerling case 1 and 2 are modeled as chi-square distribution with 2 degree of freedom.

Swerling case 3 - Targets exhibit smaller fluctuations, typically an order of magnitude or less, and fluctuate slowly. They are modeled as one prominent scatter with several smaller scatters. This model fits simpler targets, such as some missiles, at ranges where aspect changes are small over a look. It also fit complex targets, which are augmented (have their CS made artificially larger) with a single augments.

Swerling case 4 - It is the same as case 3 except that the fluctuations occur rapidly, either from rapid aspect changes or from radar frequency agility.

Case 3 and case 4 targets are described statistically as shown in , and their probability density is given in..

$$P(\sigma) = \left(\frac{4\sigma}{\mu^2} \right) e^{-\frac{2\sigma}{\mu}} \quad (3.5)$$

where $P(\sigma)$ is the probability density of a certain RCS, σ is the RCS and μ is the median RCS. In this simulation, Swerling case 3 and 4 are modeled as chi-square distribution with 4 degree of freedom.

Figure 3.2 shows the probability density for the Swerling cases.

Figure 3.2. Probability Density for the Swerling Cases [10].

Figure 3.3 shows the returned echo, after the given time delay, with and without the Swerling model.

Figure 3.3. Returned Echo of Signal with and without Swerling Model.

It is known that in the arena of high frequency detection, the fluctuation of a target will not be as extensive as it is shown in the swerling case 1 and 2 as the

reflected beam width in the mono-static region are usually of a higher order than those in the L band and above. Swirling case 3 best describes the RCS fluctuation of a ship or any objects as variation in RCS occurs very infrequently. Below is a plot of RCS using Feko [11] of a simple pencil shaped object used to describe the aircraft for 5, 10, 15, 20, 25, and 30 MHz with the target facing 0 degree. These plots serve as an illustration that as the frequency goes lower, the beamwidth in the mono-static region is quite large and thus it is justified to use Swerling case 3.

Figure 3.4. Simple Model for the RCS computation.

Figure 3.5. 5 MHz RCS plot.

Figure 3.6. 10 MHz RCS plot.

Figure 3.7. 15 MHz RCS plot.

Figure 3.8. 20 MHz RCS plot.

Figure 3.9. 25 MHz RCS plot.

Figure 3.10. 30 MHz RCS plot.

As it can be seen from the RCS plot above, the beamwidth in the mono-static region is quite big where variation of the RCS occurs infrequently. Moreover, the above plots are for a pencil-shaped like structure like aircraft, which flies at much greater speed than ships thus it is suffice to use Swerling case 3 as an assumption to the RCS variation.

With regards to the absolute value of the RCS, the free space RCS of a vessel in m^2 can be approximated [12]:

$$\sigma = 52 f^{\frac{1}{2}} D^{\frac{3}{2}} \quad (3.6)$$

Where D is the full-load displacement of the vessel in kiloton and f is the radar frequency in MHz. The above empirical formula was derived from measurements made at X, S and L bands and extended as a rough approximation to the HF band for HFSWR. Generally the displacement of a small vessel can be of ~1000 ton whereas large vessels akin to cargo container ships can be of several tens of kilotons. Thus usually the RCS of a ship can range from ~1000 m^2 for a small vessel to 30,000 m^2 of a very large container ship.

3.1.2.3 HFSWR Power Budget

Power received for the signal level in this program is computed using the Radar power budget equation with an additional factor of attenuation loss. In a general Radar equation, it is given that:

$$P_r = \frac{P_t G_t G_r \sigma \lambda^2}{(4\pi)^3 R^4 L_s} \quad (3.7)$$

where

P_r = Received Peak Power

P_t = Transmitted Peak Power

- G_t = Transmitter Antenna Gain
- G_r = Received Antenna Gain
- σ = Target Radar Cross Section
- λ = Radar Wave Length
- R = Range of target
- L_s = System Loss

In the specification of the values of transmitter antenna gain as well as receiver antenna gain, 2 effects have to be taken into account to give achieve an accurate value. The first is the ground plan effect, which refers to the doubling of the field intensity or quadrupling of the power flux density when an antenna is radiating over a perfectly conducting plane. This will results in a 6 dB antenna gain. The second effect is the mutual coupling of the antenna (or target) and its image as its distance from the conducting surface decreases. A reduction of 3 dB in gain (or 6 dB in the target RCS) will result when the antenna (or target) is situated in the conducting surface. In the simulation, the gain of the transmitting and receiving antenna is assumed to be 10 dB and 9 dB respectively. These figures are taken from a trial done in Canada for a similar HFSWR system [13]. System loss is assumed to be 3 dB.

For HSFWR, additional factor will include A' where A' is the attenuation factor for HF signal propagating over the sea water. The total propagation loss(one way) for the HFSWR, L_T is thus defined as:

$$L_T = \left(\frac{4\pi R}{\lambda A'} \right)^2 \quad (3.8)$$

where

- L_T = Total Propagation Loss (one way)
- A' = Modified Surface Wave Attenuation that includes sea state loss

The new HSFWR equation is hence rearrange as:

$$P_r = \frac{P_t G_t G_r \sigma 4\pi}{\lambda^2 L_s L_T^2} \quad (3.9)$$

The modified surface wave attenuation that includes sea state loss A' is a derivation of the more well-known Norton surface-wave field attenuation factor. This factor changes as the range changes. A' is primarily for surface waves propagating over a rough sea where the additional losses are due to the increased surface roughness. Barrick [14] uses a perturbation technique by which the effective surface impedance of the sea surface changes according to the sea surface roughness. Together with the basic propagation loss due to the spherical dispersion of the signal, the total propagation loss for one way is thus as shown as above. Further calculation of the attenuation factor can be found in the next chapter. This power budget will be further used to compute the Bragg Lines simulation with more parameters which will be discussed at a later chapter.

3.1.2.4 *Barrick's Attenuation Factor Based On Sea State*

As seen in (3.8), the HFSWR power budget equation comprises of an additional factor known as the Norton Attenuation Factor. This Norton attenuation factor can be obtained from the residue series as explained in the paper by J. R. Wait and K. P. Spies [15]. The calculation makes use of the surface impedance of the medium, which is the salt water in this case, and using the integral from:

$$W = \frac{1}{2} \left(\frac{ix}{\pi} \right)^{\frac{1}{2}} \int_{\infty \exp(-\frac{2\pi}{3})}^{\infty} \frac{\exp(ixt)w_1(t)}{w_1(t) - qw_1(t)} dt \quad (3.10)$$

to obtain the attenuation factor W .

This attenuation factor is basically for the case of a smooth conducting surface. But as we know, in the sea there are many different kind of states of sea out in the ocean. For surface wave propagation over a rough sea, there are bound to be additional losses due to the increase in surface roughness. Barrick [16] uses a perturbation method by which the additional losses are due mainly to the increase in sea-surface roughness and hence the increase in the effective surface impedance.

In the paper by Barrick [14], he started with the Rice technique of Fourier expansion for surface height. The basic expressions for a wave guided along a perfectly flat and highly conducting material of impedance Δ which are as follows:

$$\begin{aligned} E_x &\simeq E_0 \Delta \exp[ik_0(1 - \Delta^2)^{\frac{1}{2}}x - ik_0\Delta z] \\ E_y &= 0 \\ E_z &\simeq E_0 \exp[ik_0(1 - \Delta^2)^{\frac{1}{2}}x - ik_0\Delta z] \end{aligned} \quad (3.11)$$

where E_0 is the E field amplitude constant. From the above expressions, the total field of a rough surface above the interface can be express as a perturbation of the above equation as:

$$\begin{aligned} E_x &\simeq \Delta E(h, 0, z) + \sum_{m, n=-\infty}^{\infty} A_{mn} E(m + h, n, z) \\ E_y &= \sum_{m, n=-\infty}^{\infty} B_{mn} E(m + h, n, z) \\ E_z &\simeq E(h, 0, z) + \sum_{m, n=-\infty}^{\infty} C_{mn} E(m + h, n, z) \end{aligned} \quad (3.12)$$

where

$$E(m+h, n, z) \equiv E_0 \exp[ia(m+h)x + iany + ib(m+h, n,)z]$$

$$\text{and} \quad (3.13)$$

$$b(m+h, n) \equiv [k_0^2 - a^2(m+h)^2 - a^2n^2]^{\frac{1}{2}}$$

In equation 3.12, the presence of the roughness is expressed in the summation terms. Hence as the roughness disappeared, A_{mn} , B_{mn} , C_{mn} , will vanish leaving behind the terms in equation 3.11. By rearranging the terms and expressing them in terms of $E(h,0,z)$, we can get:

$$\begin{aligned} E_x^G &= (\Delta + A_{00})E(h,0,z) \\ E_y^G &= B_{00}E(h,0,z) \\ E_z^G &= E(h,0,z) \end{aligned} \quad (3.14)$$

By comparing the above equations with (3.12), we can define the effective impedance as

$$\bar{\Delta} = \Delta + A_{00} \quad (3.15)$$

As it can be seen in the above equation, the effective impedance is an addition of A_{00} with the impedance at a perfectly smooth surface, thus the roughness of the surface is define in A_{00} and hence the goal is to determine A_{00} .

Detailed explanation of the derivation can be found in Barrick's paper [14], [16] which shall not be explained here. Eventually, A_{00} can be expressed to be:

$$A_{00} = \frac{1}{4} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} \frac{p^2 W(p, q) dp dq}{b'}$$

and

$$b' = \frac{1}{k_0} \left[k_0^2 - (p + k_0)^2 - q^2 \right]^{\frac{1}{2}} \quad (3.16)$$

Where $W(p, q)$ is the wave spectrum which is defined as in Neumann-Pierson spectrum or Philips spectrum as follow:

$$W(p, q) = \frac{C(p \cos \alpha + q \sin \alpha)^2}{g^{\frac{5}{2}} (p^2 + q^2)^{\frac{13}{4}}} \cdot \exp \left\{ \frac{-2g}{U^2 (p^2 + q^2)^{\frac{1}{2}}} \right\} - \text{Neumann-Pierson spectrum} \quad (3.17)$$

where C is a constant empirically estimated to be $3.05 \text{ m}^2 / \text{sec}^5$,

U is the wind speed in m/s and,

g is the gravitational acceleration, 9.8 m/s^2

$$W(p, q) = \frac{4B}{\pi (p^2 + q^2)^2} - \text{Philips spectrum} \quad (3.18)$$

Where $B \cong 0.0005$.

Empirical evidence indicates that the wave height spectrum follows a law close to κ^{-4} in the saturated region ($\kappa^2 = p^2 + q^2$) and that it falls off rapidly to zero at the lower end when $\kappa < g/U^2$.

Using polar integral transformation and by letting $p = r \cos \theta$ and $q = r \sin \theta$, using Philip spectrum, it can be shown that A_{00} can be transformed into:

$$\begin{aligned}
 A_{00} &= \frac{1}{4} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} \frac{p^2 W(p, q) dp dq}{k_0 \left[k_0^2 - (p + k_0)^2 - q^2 \right]^{\frac{1}{2}}} \\
 &= \frac{1}{4} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} \frac{r^2 \cos^2 \theta \cdot 0.02 \cdot r dr d\theta}{k_0 \left[k_0^2 - r^2 \cos^2 \theta - k_0^2 - 2rk_0 \cos \theta - r^2 \sin^2 \theta \right]^{\frac{1}{2}} \pi (r^2 \cos^2 \theta + r^2 \sin^2 \theta)^2} \quad (3.19) \\
 &= \frac{1}{4} \int_{-\infty}^{\infty} \int_{-\infty}^{\infty} \frac{\cos^2 \theta \cdot 0.02 \cdot dr d\theta}{k_0 \left[-r^2 - 2rk_0 \cos \theta \right]^{\frac{1}{2}} \pi r}
 \end{aligned}$$

Since the spectrum is identically zero for $\kappa < g/U^2$, the integral for r can be from g/U^2 to a relatively large number like 100 and θ can be from 0 to π .

The figure below shows the plot of effective impedance against the wind speed and frequency in MHz. The graphs will be used to determine the correct effective surface impedance at the respective sea state and frequency.

Figure 3.11. Plot of effective impedance against the wind speed and frequency in MHz.

As it can be seen from the graph, the higher the sea state, i.e. the rougher the sea, the higher is the effective impedance. Before the signal is sent into the next stage, noise is added into this signal using the Matlab function. The noise power is taken from the data provided by the CCIR in dB.

3.1.3 Mixer / Low Pass Filter

Signal together with the noise are next passed into the mixer together with the local oscillator to translate the high frequency received signal to the intermediate frequency (IF). The IF is the difference between the signal frequency and that of the local oscillator. At this lower frequency, amplifiers and filters can be better implemented to match the requirement.

$$f_{IF} = f_{SIG} - f_{LO}$$

or

$$f_{IF} = f_{LO} - f_{SIG}$$
(3.20)

where f_{IF} is the intermediate frequency, f_{SIG} is the signal frequency and f_{LO} is the local oscillator frequency. Figure 3.12 shows a figure of the Mixer

Figure 3.12. Downconversion with a mixer.

After passing through the mixer, the signal is now at the baseband. The baseband signal is then passed through the low pass filter to remove any signal with frequency that is outside the stipulated bandwidth.

Figure 3.13 shows the plot of the signal after the mixer and after the low pass filter.

Figure 3.13. Signals v.s. Frequency Plots

(A: Local Oscillator 10 Hz, B: Received Signal; C: IF Signal; D: IF Signal after Low Pass Filter).

Notice that in Figure 3.13C, after the Mixer, the output IF signal has its frequency at 0 Hz and 20 MHz (recall equation (3.6)). The Low Pass Filter is to take out the upper bound frequency at 20 MHz as well as filtering out noise outside of its bandwidth.

3.1.4 Analogue-To-Digital Conversion Sampling and Data Storage

After the low pass filter, the signal is sampled at the ADC sampling rate as specified by the user. According to Nyquist theory, the returned signal should be sampled at the sampling frequency that is larger than 2 times the carrier frequency. In this simulation, as the signal has been down converted to baseband with only the bandwidth to take care of, it could be just simply larger than 2 times the bandwidth. However, as the signal is split into I and Q

format, it is suffice to sampled the signal at the bandwidth transmitted thus reducing the amount of data that need to be stored.

3.1.5 Signal Processing

Signal processing in this simulation is done in a separate program as compared to the simulation of the target and returned signal. As stated in section 2.4, there are 3 basic processes in signal processing to separate the desired signal from the interference signal. They are signal integration, correlation, and filtering and spectrum analysis.

In our simulation, the focus will be on correlation where the signal-plus-interference is compared to an ideal signal. This method is also sometimes known as match-filtering. The match filter function correlates the echo wave with a delayed copy of the transmitted signal. Pulse compression is also done at the same time.

To complete the matched filter processing for pulse compression, a baseband reference signal is generated by using the following equation:

$$s_{ref} = A(t) \cdot \exp(j\pi(\frac{B}{\tau}t^2 - Bt)) \quad (3.21)$$

where the centre frequency of the s_{ref} is 0hz and the frequency range is from $-B/2$ to $B/2$.

The pulse compression is equivalent to the matched filter processing. According to the matched filter theory, the output signal should be the convolution between the received signal and reference signal in time domain. To reduce the computational complexity of calculating convolution in time domain, the

equivalent processing technique in frequency domain is used. Fig. 3.14 shows the architecture of pulse compression processing in frequency domain.

Figure 3.14 Architecture of Pulse Compression Processing in Frequency Domain

The advantage of matched filtering and pulse compression is that when the returned echo of any 2 targets overlap each other, this process will be able to help to resolve the 2 targets' time delay.

Assuming a signal pulse width of $100\ \mu\text{s}$ and the time delay for the 1st target is $200\ \mu\text{s}$ and 2nd target is $230\ \mu\text{s}$, Figure 3.15a shows the returned echo of 1 pulse as received by the receiving antennas. It can be seen that the 2 returned echo overlap each other and thus it is difficult to distinguish the 2 targets range. However after matched filtering and pulse compression is done on the received signal, the plot easily distinguished the 2 targets with their time delay as shown in Figure 3.15b.

Figure 3.15. (a) Plot of Returned Signal before Signal Processing (top) and (b) after Signal Processing (bottom).

As it can be seen from the above plot, pulse compression usually is able to help to achieve a ~30 dB to 40 dB gain compared to the noise floor. Generally pulse compression gain can be defined as $B\tau$ (bandwidth x pulse width).

3.1.6 Digital Beam Forming

To enhance the SNR of potentials targets further, digital beam forming is performed on the received signal. Digital beam forming or commonly known as DBF allows the radar to digitally steer the beam to the angle in which the target might exist. By doing this, the SNR of the target will increase since it is lying in the main lobe of the steered beam rather than the side lobe. Digital beam forming performance depends heavily on the separation distance of the antenna as well as the total length of the antenna array.

Beam width (rad) = 0.88λ /total length of array

As it can be seen from above, if the length of array decreases, the beam width will increase thus affecting the azimuth detection. As the separation of antenna increase to λ , grating lobes will formed. Grating lobes refer to “side lobes” of the main beam with power that might be equivalent or comparable to the main beam power level. Grating lobes will also reduce the antenna gain of the array.

For a 10 MHz signal with λ at 30m, the following beam pattern plots are as shown for different spacing:

Figure 3.16. Array Pattern at 5m antenna spacing.

Figure 3.17. Array Pattern at 10m antenna spacing.

Figure 3.18. Array Pattern at 15m antenna spacing.

Figure 3.19. Array Pattern at 30m antenna spacing.

As you can see from the plots above, as the antenna spacing increase, grating lobes will appear at the side of the array or at the end fire. This might caused ambiguity in terms of the target azimuth detection especially if the detection occurred at end fire but it is mistaken at broad-sight. As the antenna spacing decrease, beam-width of the main beam increase causing difficulty in azimuth target detection. It is evident in the first plot for 5 m spacing, the beam-width is so broad that it does not allow any azimuth target detection at all. The most ideal pattern will be a sharp main beam with no grating lobe and low side lobes. However this is not possible as the spacing between the antenna as well as the length of the array affect the array pattern have an inverse relationship. Thus a spacing of $\lambda/2$ is always provided as guideline. In Figure 3.17, there is no grating lobe but the beam width is still reasonably big.

Figure 3.20 shows the array pattern for an ideal situation.

Figure 3.20. Array pattern at 10 MHz with digital beam forming at 5 angles.

Beam Forming is achievable at -25° , -12° , 0° , 12° degree and 25° . Assuming a target that is lying in the angle of -30° from the broad sight of the array, the SNR of this target will be relatively small if no digital beam forming is done. Looking at the plot in figure 3.20, the target signal is actually nulled! However with digital beam forming, the target will have a high SNR as compared to the rest due to the 1st beam that is digitally formed.

3.1.7 Target Range-Doppler Map Generation

Doppler processing is performed using the fast Fourier transform (FFT). As stated in chapter 3.1.5, it is another form of signal processing especially for moving target. As the returned echo of the target contains a specific Doppler shift, by doing Doppler processing, the target of very low SNR is able to be detected

depending on the length of integration time. Generally the gain from Doppler processing can be defined as $T_i f_p$ where T_i is the coherent integration time and f_p is the radar PRF. For the simulation where T_i can be 10second and f_p is 2 kHz, the Doppler processing gain be 86 dB, quite a high amount. A very important parameter for Doppler processing is the size of the FFT used. The appropriate size of the FFT is determined nominally by the length of the time series containing the target. Since a moving target will remain in a range resolution cell in a finite amount of time only, the size of the FFT is determined by the time required for a target with an expected maximum speed to transverse a range resolution cell.

From equation (3.3), as the Doppler frequency of a slow moving target (ship) traveling at 30 knots towards the radar of 30 MHz will be ~3 Hz. If the operating frequency is lower and the ship is traveling slower, the Doppler frequency of the target will be < 1 Hz. This might posed a problem of accuracy if the resolution of the Doppler frequency is not small enough. Thus the sample time should be as long as possible ~ 30,000 pulses (PRI = 500 μ s) for a Doppler resolution ~0.06 Hz.

$$\text{Doppler Resolution} = \frac{1}{CIT} \quad (3.22)$$

Doppler axis = -PRF/2 : Doppler Resolution : PRF/2

Usually the limit of the Doppler axis is not an issue as the range of interest in a HFSWR system is -30 Hz to 30 Hz. Fast moving craft can be observed within this range.

Below shows a figure of a Range Doppler Map simulated.

Figure 3.21. Range-Doppler Map showing the Targets at different radial velocity.

Figure 3.21 is simulated with the following parameters:

Carrier Frequency	10 MHz	Sampling Frequency (Generator)	99.2Mhz
Bandwidth	400 kHz	Transmitted Power	2KW
PRI	500 μs	Transmitting Antenna Gain (absolute)	10dB
Pulse Width	100 μs		

Table 3.1 Radar Transmitting Parameters

Sampling Frequency (ADC)	400 kHz	Transmitting Antenna Gain	7.9
No. of elements	8	Element Spacing	15m
Noise Level	-150 dbW	System Loss	3dB

Table 3.2 Radar Receiving Parameters

Velocity	15 m/s	Time Delay	250μs
Angle to Radar	10°	Target Movement Bearing	030
RCS	1000 m ²		

Table 3.3 Target 1 Parameters

Velocity	20 m/s	Time Delay	300 μ s
Angle to Radar	40°	Target Movement Bearing	180
RCS	1000m ²		

Table 3.4 Target 2 Parameters

As it can be seen, the Doppler frequency for the 1st target, using equation 3.22, is -0.984 Hz. Since the movement of the target is away from the Radar, the Doppler frequency will be negative in nature. As for the 2nd target, the Doppler frequency is 1.02 Hz. The target is moving towards the radar thus the frequency is positive in nature.

The noise level generate for this plot is -150 dBW. This is in accordance with the CCIR results given in chapter 2.2.3. We have taken -150dBW as the median value.

3.2 Simulation of Bragg Lines and Propagation Loss

In the pioneering work of Combie [17], through the experiments involving backscattering of HF radio wave with the frequency being 13.56 MHz, two Bragg peaks were observed. The most prevalent clutter that is encountered in the HFSWR system is the Bragg ocean clutter. Bragg clutter occurs when radio wave are reflected back from the ocean due to the unevenness of the ocean surface. It is akin to describe it as scattering of the electromagnetic waves from rough surfaces. The difference is that the roughness of the ocean surface is dynamic, i.e. they are moving constantly. The unevenness of the ocean surfaces refers to the raising and falling sea waves due to the gravitational pull of the earth as well as the surface wind. As the Radar signal reached these moving waves, the signal is then reflected back to the source. The following chapters will describe the simulation of the 1st order and 2nd order Bragg clutter which will be added to the Target Range Doppler Map generation.

3.2.1 Simulation of 1st order and 2nd order Bragg Lines

The ocean echo wave spectrum $P(\omega)$ consists of the first- and second-order spectrums. It is

$$P(\omega) = a(\phi^{(1)}(\omega) + \phi^{(2)}(\omega)) \quad (3.23)$$

where a is a constant, $\phi^{(1)}(\omega)$ is the cross section per unit area of the 1st order scattering and $\phi^{(2)}(\omega)$ is the cross section per unit area of the 2nd order scattering.

3.2.1.1 1st Order Bragg Scattering

In a paper by Barrick [18], he presented the relationship between the first and second order scattering cross section and the heights of ocean wave at grazing

incidence by a beam of vertically polarized radiation. For the first-order scattering, the cross section per unit frequency per unit area of scattering region is given by:

$$\sigma^{(1)}(\omega) = 2^6 \pi k_0^4 \sum_{m=\pm 1} S(-2m\mathbf{k}_0) \delta(\omega - m\omega_B) \quad (3.24)$$

Where ω is the Doppler angular frequency, $\mathbf{k}_0$ is the wave vector of radar with its magnitude $k_0 = 2\pi/\lambda$. When considering the effect of first-order the resonance scattering, the wave vector of the resonance ocean wave is $\mathbf{k} = -2m\mathbf{k}_0 \mid m=\pm 1$, $S(\bullet)$ is the spectrum of the directional wave height of the ocean wave, ω_B is the angular frequency of the positive Bragg peak where,

$$\omega_B = 2\pi f_B^+ = \sqrt{2\pi g \cdot \frac{2}{\lambda} \cdot \cos \beta} \approx \sqrt{2\pi g \cdot \frac{2}{\lambda}} = \sqrt{2gk_0} \quad (3.25)$$

In equation 3.24, the δ function indicated the first ordered backscattered spectrum consists of two spikes at $\pm \omega_B$ under the ideal condition. In addition the direction wave spectrum in 3.24 which is decided by the sea wave frequency and the propagation direction can be expressed as follows:

$$S(\mathbf{k}) = S(k)d(\theta - \varphi_w) \quad (3.26)$$

Where $\mathbf{k}$ is the sea wave vector, $S(k)$ is the non directional wave spectrum. For a certain large sea state with wind speed, the non-directional wave spectrum can be expressed as follows:

$$S(\omega) = \frac{8.1}{10^3} \cdot \frac{g^2}{\omega^5} \exp \left[-0.74 \left(\frac{g}{\omega U} \right)^4 \right] \quad (3.27)$$

Where $\omega = \sqrt{gk}$, U is the wind speed at the height of 19.5 m above the sea surface. The directional factor is expressed as follows

$$d(\theta - \varphi_w) = A \cos^s \left(\frac{\theta - \varphi_w}{2} \right) \quad (3.28)$$

Where A is a constant, s is the extended factor with its typical value being 4, θ is the angle between the radar beam direction and the ocean wave direction, φ_w is the angle between the radar beam direction and the wind direction.

For the first-order spectrum, because of its large intensity, narrow width of spectrum, and the Doppler frequencies being $\pm 0.102 \sqrt{f_o}$ in Hz, where f_o is the operating frequency in MHz, it can be simulated to use two sinusoidal signals with the frequencies at $\pm 0.102 \sqrt{f_o}$. It is commonly accepted to simulate the 1st order scattering as Gaussian curve with their peak at the Bragg frequency. The equation is as follows:

$$\phi^{(1)}(\omega) = P_s \exp\left(\frac{-(\omega - \omega_B)^2}{2\sigma_\omega^2}\right) \quad (3.29)$$

where P_s is the 1st order scattering power, ω_B is the Bragg frequency and σ_ω is the spectral width of the 1st order scattering.

3.2.1.2 2nd Order Bragg Scattering

In 1974, Barrick [6] suggested that the complex band structures around the two Bragg peaks were explained by using the 1st and 2nd order scattering theory. This was backed up by many investigations which demonstrated that the main characteristics of the 2nd spectrum are agreement with those of Barrick's theory model [19]. Furthermore, through the experimental analysis, some investigators

suggested that there exist high-order scattering interactions [20]. Generally speaking, the 1st and 2nd order spectrums are dominant for total ocean wave backscattering. Donald Barrick [18] in his paper described the cross section per unit frequency interval per unit area of scattering for the 2nd order scattering as:

$$\sigma^{(2)}(\omega) = 2^6 \pi k_0^4 \sum_{m, m' = \pm 1} \int_{-\infty}^{\infty} \int |\Gamma|^2 S(mk) S(m'k') \delta(\omega - m\sqrt{gk} - m'\sqrt{gk'}) dp dq$$

and

$$\Gamma = \Gamma_H + \Gamma_{EM} , \quad (3.30)$$

$$\Gamma_H = -\frac{i}{2} \left[k + k' - \frac{(kk' - k \cdot k')(\omega^2 + \omega_B^2)}{mm' \sqrt{kk'}(\omega^2 - \omega_B^2)} \right]$$

$$\Gamma_{EM} = \frac{1}{2} \left(\frac{(k \cdot k_0)(k' \cdot k_0)}{-k_0^3 \Delta} \right)$$

Where Γ is the coupling or interaction coefficient between the two sets of water waves with Γ_H define as the hydrodynamic and Γ_{EM} define as the electromagnetic component, Δ is the normalized wave impedance on the sea surface by the vertically polarized incidence electromagnetic wave, k_0 is the radar wave vector with amplitude $2\pi/\lambda$. k and k' are the wave vectors of the 1st and 2nd column sinusoidal waves with the amplitudes being k and k' and the angular frequencies being $\sqrt{gk}$ and $\sqrt{gk'}$ respectively. p and q are the transverse axis and longitudinal axis respectively and are radar beam direction are in p axis. From Figure 2 in the paper by Barrick, the relationship the two wave vectors k and k' can be described as follows:

$$k + k' = -2k_0 \quad (3.31)$$

The value of m and m' takes the value of $+1$ and -1 thus giving rise to 4 different combinations. In the paper, the four combinations are given with their respective constraints.

With the different conditions for different regions of the frequency ω given in the paper ($\omega < -\omega_B$, $-\omega_B < \omega \leq 0$, $0 \leq \omega < \omega_B$, $\omega > \omega_B$), 4 different regions of the frequency ω can be simulated. It is convenient to convert the Cartesian coordinate system into the Polar coordinate system with the P axis being the referenced direction. Then equation (3.11) can be re-written as follows:

$$\phi^{(2)}(\omega) = 2^6 \pi k_0^4 \sum_{m, m' = \pm 1} \iint |\Gamma|^2 S(mk) S(m'k') \delta(\omega - m\sqrt{gk} - m'\sqrt{gk'}) k dk d\theta \quad (3.32)$$

Dispersing the variable k , namely, $k = k_n = n\Delta k$, - (3.13) can be simplified as

$$\begin{aligned} \phi^{(2)}(\omega) &= 2^6 \pi k_0^4 \sum_{m, m' = \pm 1} \sum_{n=1}^{\infty} n(\Delta k)^2 \int_{-\pi}^{\pi} |\Gamma|^2 S(mk_n) S(m'k') \delta(\omega - m\sqrt{gk} - m'\sqrt{gk'}) d\theta \\ &= \sum_{n=1}^{\infty} 2^6 \pi k_0^4 n(\Delta k)^2 \sum_{m, m' = \pm 1} \int_{-\pi}^{\pi} |\Gamma|^2 S(mk_n) S(m'k') \delta(\omega - m\sqrt{gk} - m'\sqrt{gk'}) d\theta \\ &= \sum_{n=1}^{\infty} \sigma_n^{(2)}(\omega) \end{aligned} \quad (3.33)$$

The computation on the second-order spectrum $\phi^{(2)}(\omega)$ can be converted into the computation on the second-order spectrum of sub-spectrum $\phi_n^{(2)}(\omega)$ at $k = k_n$. Firstly, for a certain value n , namely a certain value of k_n , a set of k' can be evaluated from the four intervals of ω . By using Cosine Theorem and recalling Figure 2 in Barrick's paper, the restricted condition in -(3.12) can be re-written as

$$\cos \theta = \frac{k'^2 - k_n^2 - 4k_0^2}{4k_n k_0} \quad (-1 \leq \cos \theta \leq 1) \quad (3.34)$$

Then, for the selected values of k' , the suitable values of k' can be obtained from (3.15), and the azimuths θ' are as follows:

$$\begin{aligned}\theta' &= \sin^{-1}\left(\frac{k_n \sin \theta}{k'}\right) - \pi & 0 \leq \theta \leq \pi \\ \theta' &= \sin^{-1}\left(\frac{k_n \sin \theta}{k'}\right) + \pi & 0 < \theta < \pi\end{aligned}\quad (3.35)$$

Substituting k_n , k' , θ , θ' into (3.14), it can be obtained the value of the second-order sub-spectrum $\phi_n^{(2)}(\omega)$ corresponding to n . The value of the second-order spectrum $\phi^{(2)}(\omega)$ can be obtained readily by pulsing all values of the second-order sub-spectrums. Finally, adding the second-order spectrums of the four sub-intervals and the two first-order spectrums, the full simulation clutter spectrums are obtained. Figure 3.22 shows the Bragg Scattering plot in frequency domain.

Figure 3.22. Bragg Scattering plot in Frequency Domain.

The above plot is simulated with a 10 MHz carrier frequency thus from equation 3.25, the Bragg lines can be seen at ± 0.3225 .

3.2.2 Simulation of HF Signal Propagation Loss

With the simulation of the 1st and 2nd order Bragg scattering, it is still insufficient to form the RD map as the simulation of the Bragg scattering can be only done in the frequency domain and not in a single pulse. To overcome this difficulty, the same Bragg scattering plot in frequency domain will be apply to all ranges and the propagation loss be multiply to obtain a true RD map.

In David A Hill's paper [8], he presented 4 possible ways of calculating the propagation loss function:

1. Residue series,
2. Small curvature expansion,
3. Flat earth (assuming no curvature)
4. Power series

In the paper, it is showed that only residue series will be able to provide the most accurate attenuation function in the longer range.

J. R. Wait and K. P. Spies [15] provided the following calculation on the ground wave attenuation factor. The ground wave attenuation function (residue series) may be written in the following integral:

$$W = \frac{1}{2} \left(\frac{ix}{\pi} \right)^{\frac{1}{2}} \int_{-\infty \exp(-\frac{2\pi}{3})}^{\infty} \frac{\exp(ixt)w_1(t)}{w_1'(t) - qw_1(t)} dt \quad (3.36)$$

where $x = (ka/2)^{1/3}(d/a)$, $k = 2\pi/\lambda$, a is the earth radius, $q = -i(ka/2)^{1/3}\Delta$, $\Delta =$ surface impedance, $w_1(t)$ is the Airy function and $w_1'(t)$ is $dw_1(t)/dt$.

The surface impedance Δ , which is a function of the ground constant of the earth's surface can be derived as followed [21]:

$$\Delta = \frac{\sqrt{\epsilon_{gc} - 1}}{\epsilon_{gc}}$$

and ϵ_{gc} is given by,

$$\epsilon_{gc} = \epsilon_g + \frac{\sigma_g}{j\omega\epsilon_0}$$
(3.37)

ϵ_g is the relative dielectric constant of ground, σ_g is the conductivity of the ground in simens per meter, ϵ_0 is the permittivity of free space 8.85×10^{-12} farads per meter and $\omega = 2\pi f$ where f is the operating frequency. Please note that the above surface impedance is only valid for vertical polarization.

By deforming the contour in (3.36) so that it encloses the poles of the integrand we obtain the classical residue series

$$W = \left(\frac{\pi x}{2} \right)^{1/2} \sum_{s=1}^{\infty} \frac{\exp(ixt_s)}{t_s - q^2}$$
(3.38)

Where the poles at $t = t_s$ are roots of

$$w_1'(t) - qw_1(t) = 0$$
(3.39)

$w_1(t)$ is defined by

$$w_1(t) = \pi^{1/2} (Bi(t) - iAi(t))$$
(3.40)

Where $Ai(t)$ and $Bi(t)$ are the Airy functions.

Given the roots t_s satisfy the differential equation,

$$\frac{dt_s}{dq} = \frac{1}{t_s - q^2} \quad (3.41)$$

And the initial condition

$$t_s|_{q=0} t_s(0) = \alpha_s e^{-i\pi/3} \quad (s = 1, 2, 3, \dots) \quad (3.42)$$

Where the α_s ' are the zeros of $Ai'(-\alpha)$

Given q with $|q| \leq 1$, a sequence of $R + 1$ points is formed in the complex q -plane equally space on the straight line from the origin to the given value of q .

$$0 = q_0, q_1, q_2, q_3, q_4, \dots, q_R = q \quad (3.43)$$

With the known solution at $q_0 = 0$ from equation, the subsequent q_1 to q_R can be calculated in succession using a fourth order Runge-Kutta formula:

$$\begin{aligned} t_s(q_i + 1) &= t_s(q_i) + \frac{\Delta q}{6} (p_1 + 2p_2 + 2p_3 + p_4) \quad ; \\ p_1 &= f(q_i, t_s(q_i)), \\ p_2 &= f(q_i + \frac{1}{2}\Delta q, t_s(q_i) + \frac{1}{2}p_1), \\ p_3 &= f(q_i + \frac{1}{2}\Delta q, t_s(q_i) + \frac{1}{2}p_2), \\ p_4 &= f(q_i + \Delta q, t_s(q_i) + p_3), \\ f(q, t) &= \frac{1}{t - q^2} \end{aligned} \quad (3.44)$$

The integral R was chosen so that $|\Delta q| \sim 0.01$.

For q with $|q| > 1$, let $Q = 1/q$ and the differential equation will be

$$\frac{dt_s}{dq} = \frac{1}{1 - Q^2 t_s} \quad (3.45)$$

With the initial condition

$$t_s|_{Q=0} t_s(\infty) = \alpha_s e^{-i\pi/3} \quad (s = 1, 2, 3, \dots) \quad (3.46)$$

Where the α_s are the zeros of $\text{Ai}(-\alpha)$

Likewise previously in equation 3.43, the $R + 1$ points are also formed for Q :

$$0 = Q_0, Q_1, Q_2, Q_3, Q_4, \dots, Q_R = Q = 1/q \quad (3.47)$$

And applying the same equation (3.25) except that $f(q, t)$ will be replaced by $f(Q, t)$

Figure 3.23 shows the attenuation function of the Surface wave propagation plotted from Matlab.

Figure 3.23. Attenuation Function of the Surface Wave Propagation.

3.2.3 Bragg lines Range-Doppler Map Generation

As the Bragg clutter is extremely difficult to be simulated in the time domain into the individual pulses, it is therefore simulated in the frequency domain. A propagation factor generated in section 3.2.2 will then be multiply throughout the respective range cell to get the true power spectrum of the Bragg scattering as seen from the transmission point.

After obtaining the attenuation function across the range axis and the RCS of the 1st order and 2nd order Bragg lines, it is possible to simulate a high fidelity fading of the Bragg lines across the ranges. Recall from section 3.1.2.3, the power budget of the HFSWR for any target can be described as:

$$P_r = \frac{P_t G_t G_r \sigma 4\pi}{\lambda^2 L_s L_T^2} \quad (3.48)$$

and

$$L_T = \left(\frac{4\pi R}{\lambda A'} \right)^2 \quad (3.49)$$

where

P_r = Received Peak Power

P_t = Transmitted Peak Power

G_t = Transmitter Antenna Gain

G_r = Received Antenna Gain

σ = Target Radar Cross Section

λ = Radar Wave Length

R = Range of target

L_s = System Loss

L_T = total propagation loss(one way)

A' = Modified Surface Wave Attenuation that includes sea state loss

As the above power budget equation did not take into account the signal processing gain, which are namely the pulse compression gain as well as the Doppler processing gain, it would have to be manually added in to reflect the true value of the SNR of the Bragg lines.

Pulse compression, as describe in earlier chapter, refers to the correlation of a reference pulse chirp with the reflected echo. In this way, the reflected echo, when correlate with the reference signal, will produce a peak at the delay (equivalent to distance) of the target. Pulse compression gain can be normally be equate to the product of bandwidth and pulse width.

Another signal processing gain is the Doppler processing gain. Since coherent integration is always employed in HFSWR signal processing, it is thus necessary to include the Doppler processing gain when calculating the power budget equation. In target simulation, however, the power budget of the target does not include Doppler processing gain because the data are the simulated reflected echo and not the processed signal. In Bragg simulation, the data simulated are supposed to be the processed signal due to the difficulties of simulating a return echo for Bragg lines as explained earlier. For a coherent integration time (CIT), there are $(CIT \times f_{prf}) = N$ pulses where f_{prf} refers to the pulse repetition frequency of the HFSWR. If the target echo is present in the returns of all pulses, the coherent integration over the N pulses will enhance the SNR by a factor of N. Hence the resulting SNR over N pulses will be given by:

$$SNR = \frac{P_t(\tau f_{prf})CIT_{seconds}G_tG_r\sigma 4\pi}{\lambda^2 L_s L_T^2 N_o} \quad (3.50)$$

By removing the noise component from the SNR, we can obtained the estimated power received of the Bragg lines which will be used to determined the fading characteristic over the range axis.

Figure 3.24 shows the Range-Doppler map of the Bragg scattering with the bragg lines visible at the appropriate frequency for 10 Mhz.

Figure 3.24. Range-Doppler map of Bragg scattering.

The above plot is simulated with the following conditions:

Carrier Frequency	10 MHz	Direction of Wind to Radar Beam	60°
Wind Speed	10 m/s	CIT	40s

3.3 Combination of Bragg Scattering & Target Range-Doppler Map

When the simulation of the Bragg scattering and target Range-Doppler map is complete, the task now is to combine the 2 Range-Doppler map together to form the complete HFSWR Range- Doppler map. Since the power budget for both the target and the Bragg clutter has been simulated on the same scale, both the

simulated data can be added up easily. Below show the results of the combined Range-Doppler map. Figure 3.25 shows the Combined Range-Doppler map.

Figure 3.25. Combined Range-Doppler map.

It can be seen that the side-lobe level of the target will be amplified if the range of the dB level increases. More signal processing scheme will have to be done to cancel out the side-lobe or bring the level down.

CHAPTER 4 SIMULATIONS DATA

This chapter will showcase some of the simulation done based on different parameters. These parameters will be stated in the table.

Also discussed in this chapter will be a simple form of tracking of target using mono-pulse amplitude comparison.

4.1 Simulation 1: 2 Fast Targets Moving Away From Radar Over 100s

2 targets traveling at 100m/s and 200m/s has been simulated over 100s (200,000 pulses with PRI at 500μs). Below are the parameters set for these 2 targets

	Target 1	Target 2
Distance from Radar	30 km	37.5 km
Speed Travelling	200 m/s	100 m/s
Initial Angle to Radar	-30°	10°
Bearing of Movement with respect to true north	000	030
RCS of target	1000 m ²	

Table 4.1 Parameters for simulation 1

A pictorial illustration of the above simulation setup is shown in Figure 4.1.

Figure 4.1. Illustration of Simulation 1 Setup.

Figure 4.2 shows the Range Doppler Maps of the simulated data over a CIT of 4 sec. Using the data given above and the Doppler equation, it can be computed that target 1 has a Doppler frequency of -11.54 Hz. As it starts to move in the north 000N bearing, the angle of it make with the radar become smaller resulting in a higher negative Doppler shift as it moves away from the radar(increasing Doppler shift). As for the target 2, the initial Doppler is -6.26 Hz. As it starts to move in the direction of bearing 030N, the angle it makes with the radar become bigger thus resulting in the negative Doppler shift becoming smaller(decreasing).

(a) RD Map at 4 sec

(b) RD Map at 12 sec

(c) RD Map at 20 sec

(d) RD Map at 28 sec

(e) RD Map at 36 sec

(f) RD Map at 44 sec

Figure 4.2 (a) – (f). RD Map generated from Simulation 4.1

As the range of the Doppler is expanded to take into account the high Doppler frequency of the 2 targets, the Bragg lines are not really as visible as before. It could be seen that target 1 with a faster velocity, overtake target 2.

4.2 Simulation 2: 1 Fast Targets Moving Across Radar Broad-Sight from Negative Doppler to Positive Doppler

This simulation is to show the movement of a target as it moving parallel to the radar array crossing from the positive Doppler region, passing the zero Doppler region, to a negative Doppler region. In the Pulse Doppler radar, target usually tries to maneuver perpendicular to the Pulse Doppler radar. Because the Pulse Doppler radar depends on the target movement for detection, it is thus critical to get a non-zero Doppler from the target to distinguish it from the non-moving or stationary clutter. Below are the values for the simulation.

	Target 1
Distance from Radar	30km
Speed Traveling	200m/s
Initial Angle to Radar	10°
Bearing of Movement with respect to true north	270
RCS of target	100m ²

Table 4.2 Parameters for Simulation 2

The following figure shows a pictorial illustration of the simulation.

Figure 4.3 Illustration of simulation 2 Setup

Figure 4.4 shows the Range Doppler Maps generated over a CIT of 4s. Using the data given above and the Doppler equation, it can be computed that the target 1 has an initial Doppler frequency of +2.15Hz. As it starts to move in the north 270N bearing (westward), the velocity angle of the target make with the radar become smaller resulting in a smaller positive Doppler shift. This continues on until its velocity makes a perpendicular angle with the radar broad-sight resulting in a zero Doppler shift as cosine 90 is 0.

The target continues to move westward at the constant velocity thereby resulting in an increasing Doppler shift in the negative region.

(a) RD Map at 4 sec

(b) RD Map at 8 sec

(c) RD Map at 12 sec

(d) RD Map at 16 sec

(e) RD Map at 20 sec

(f) RD Map at 24 sec

(g) RD Map at 28 sec

(h) RD Map at 32 sec

(i) RD Map at 36 sec

(j) RD Map at 40 sec

(k) RD Map at 44 sec

(L) RD Map at 48 sec

(m) RD Map at 52 sec

(n) RD Map at 56 sec

(o) RD Map at 60 sec

Figure 4.4(a) – (o). RD Map generated from Simulation 4.2

The breakdown of the different stages of the signal are as shown below for the pulse before adding noise, after adding noise and after the various signal processing algorithm gain.

Figure 4.5 Returned echo before the addition of noise

Figure 4.6 Returned echo are the addition of noise of -150 dB/Hz

It can be noticed that the returned echo of the pulse is completely covered by the noise with a SNR of -57.5 dB. In such cases, it will definitely not be detected by any radar depending on a single pulse. Therefore in this situation, it will require signal processing to identify the results.

Figure 4.7 Different Stages of Signal Processing

The blue colour dash line indicates a single pulse before any signal processing is done. The red colour line with '+' as marker represents the level of a single pulse after pulse compression. As it can be seen, the SNR of the target is so low that even with a pulse compression gain of 32 dB ($400\text{ kHz} \times 100\text{ }\mu\text{s}$) is not able to increase the signal to a visible level. However with Doppler processing, the signal is able to be resolved from the noise at its Doppler frequency. For a 4 sec CIT, the processing gain is around 78 dB which is enough to increase the SNR to +52 dB.

4.3 Tracking of Target Using Mono-pulse Amplitude Comparison

Mono-pulse amplitude comparison is the simplest way to determine the azimuth angle of a target. Using the ideal array antenna pattern, the amplitude of the return echo in the different beams formed in the Digital Beam Forming is able to be used to determine exactly what angle the signal is coming into the array. From Figure 4.8, assuming 2 adjacent beams, $G_1(\theta)$, $G_2(\theta)$ are formed. When a target is detected and found to have the maximum amplitude at $G_1(\theta)$, its adjacent beams $G_0(\theta)$ and $G_2(\theta)$ will be used to determine the next higher return echo amplitude. In the case below, $G_2(\theta)$ will be found to have the next highest echo amplitude. With both the echo amplitude at beam $G_1(\theta)$ and $G_2(\theta)$, the ratio will be taken $G_1(\theta)/G_2(\theta)$ and will be compared to a pre-calculated table of the different ratios of $G_1(\theta)/G_2(\theta)$. In this way, the ratio will be able to point to the respective angle.

Figure 4.8 Monopulse Amplitude Comparison

Below figures shows the tracking results done on the simulated data in section 4.2.

(a)

(b)

(c)

(d)

Figure 4.9 (a) - (d) Tracking of target in RD map (enclosed by the square)

The resultant angles that are computed based on 4s CIT from this mono-pulse comparison tracking are as follows:

Figure 4.10 Comparison of Range with Ground Truth and Processed Data

Figure 4.11 Comparison of Angle with Ground Truth and Processed Data

It can be seen from the range comparison graph, as there is a range discrepancy of 300 m – 400 m, which is around the range resolution of the data. Recalled previously range resolution is $c/2B$ where c is the speed of light and B is the bandwidth, thus with $B = 400$ KHz, the range resolution will be 375 m. The angular tracking can be considered quite accurate. The downside to this simple algorithm is that the accuracy is very susceptible to noise.

From the above example, at 51-52 sec time frame, the target supposed to the highest amplitude at the 2nd beam. The next highest amplitude is found to be at beam 1 though both beam 1 and beam 3 has amplitude that differs only by 0.001. Due to beam 1 having higher amplitude, the ratio of the power difference will be compared between beam 1 and beam 2 instead thus creating the error of around 8°. However if the ratio has been compared between beam 2 and beam 3, the angle will be at the correct location of around -10° instead of -18°. This phenomenon can be attributed to the noise which causes the discrepancy in amplitude of the intended track in beam 1 and beam 3. Despite this drawback, as the amplitude of the signal increase in the different beams, the azimuth angle will be re-adjust to the correct angle as shown in the around 62 sec. This discrepancy last around 10 sec.

Beside this drawback, tracking using mono-pulse amplitude comparison depends heavily on the receiving array pattern of the receiving elements. In this simulation, the receiving pattern are simulated to be ideal without any distortion, thus when the mono-pulse amplitude comparison is used for tracking of the target, it is able to return an error of 1° at around 35 km. However, if the receiving array pattern has distortion such that the amplitude ratio between the beams are not ideal, it will result in a azimuth angle error which sometimes could be 5° – 10° out depends on how bad the distortion is.

4.4 Simulation 3: 5 Moving Targets At Varying Speed In A Single Data File

This simulation is to show the movement of 5 targets simultaneously in a data file. Below are the values for the simulation.

	Target 1	Target 2	Target 3	Target 4	Target 5
Distance from Radar km	37.5	45	34.5	52.5	25.5
Velocity in m/s	20	10	100	15	200
Initial Angle to Radar	-10°	0°	8°	-15°	20°
Bearing of Movement with respect to true north	150	20	60	200	340
RCS of target in m ²	40	100	60	80	20

Table 4.3 Parameters for Simulation 3

The following figure shows a pictorial illustration of the simulation.

Figure 4.12 Illustration of simulation 3 setup

In this simulation, the decimation and CIT for the processing of the data are alternate to show the slowing moving target (with low Doppler frequency) and

fast moving target (with high Doppler frequency). To see slow moving target, usually the CIT are smaller with higher decimation. This is to improve the resolution especially for very slow moving target with Doppler frequency $< 1\text{Hz}$. Below is the Range Doppler Map of CIT – 40 sec and Decimation of 200 pulses.

(a) Range Doppler Map at 40 sec

(b) Range Doppler Map at 80 sec

Figure 4.13 (a) – (b). RD Map for CIT 40 sec and Decimation 200 pulses

By expanding the Doppler frequency range, we will be able to see additional Target 3 and 5 with their Doppler frequency lies beyond ± 3 Hz.

Figure 4.14a. Range Doppler Map at 4 sec

As the target moves away from the Radar, the SNR of the target, because of its RCS and losses, will become “dimmer”. Especially for Target 5 with a RCS of only 20 m², it disappear from the Range Doppler map at 10th image which is around 36 sec to 40 sec.

(b). Range Doppler Map at 32 sec

(c). Range Doppler Map at 36 sec

(d). Range Doppler Map at 40 sec

Figure 4.14 (a) – (d). RD Map for Simulation 3 for Fast Moving Targets

Notice that Target 5 disappear from the Range Doppler Map as it moves further away from the Radar due to the low SNR.

CHAPTER 5 CONCLUSIONS AND FUTURE WORKS

5.1 Conclusion

Recalling the research objectives in Section 1.3, the objective of this research is to build a High Frequency Radar simulator program for study of radar system and radar signal processing purpose so that the user can use it to:

- Simulate HFSWR with own parameters for transmitting power, antenna gain, pulse width, pulse length, bandwidth, sampling frequency.
- Simulate HFSWR's targets with own parameters for target RCS, bearing of movement, angle it makes with the radar, velocity.
- Simulate Bragg clutter under controlled environment.
- Perform Signal processing algorithm like Doppler processing and pulse compression on received data.
- Perform digital beam forming on received data to search for target outside broad sight beam width.
- Track Targets Using Mono-pulse Amplitude Comparison

In chapter 2, we have discussed about the various interference that the HFSWR will encounter. This is especially true for the sea clutter which plays a major role in the contribution to the clutter for the radar system. Noise interference is also discussed and calculated to show the level of noise in Singapore. This will affect

the SNR of the targets that we are simulating. Chapter 2 also discussed about the various signal processing schemes that will be used in this HFSWR Simulator.

Chapter 3 shows the simulation of the targets as well as the bragg clutter. In chapter 3, parameters of the HFSWR as well as the targets are decided and used for the simulation. This simulation allows user to key in the pulse width, pulse repetition interval, bandwidth, sampling frequency etc. The signal is then generated using chirp waveform for transmission. With the target generation program, the user is able to key in the target information such as distance from Radar, RCS, angle it makes with the Radar, movement bearing. The echoes from these targets are then simulated into the receiving elements as stated by the user with parameters such as number of receiving elements, elements spacing etc. The power received by the receiving elements are simulated using the HFSWR power budget which will take into account the transmitting power, the transmitting antenna gain, receiving antenna gain, the Norton factor loss and System Loss. This gives a high fidelity of the power received into the receiving elements. After this, the echo is simulated to pass into a mixer and low pass filter to bring the signal to baseband and the Analogue to Digital Conversion is simulated just like in the real system where the data is converted to digital format to be stored. The data are then displayed in a Range Doppler map after passing through Doppler processing and pulse compression. It is also shown that before all these processing scheme, the target is not visible at all, well covered by the noisy environment. However, when the processing scheme is done, the SNR of the target raises to allow detection. Beam forming are performed also to enable the HSFWR to look beyond broad sight angle especially at those angle where nulling occurs for broad sight beam.

The Bragg clutters are also simulated in the frequency domain using Barrick's method so as to show the Bragg clutter in the Range Doppler Map. The 1st order and the 2nd order Bragg are simulated and together with the power budget equation, with the pulse compression gain and Doppler processing gain added

into the power budget, the Bragg lines are able to show on the Range Doppler Map. However the drawback of this method is that the phase values of the Bragg lines are ignored totally as Barrick's equation only computes the amplitude and not the phase.

In chapter 4, 3 simulations are run for various scenarios and mono-pulse tracking is performed on one of the generated data. The mono-pulse tracking shows that that despite its simplicity, it is able to perform tracking though at certain SNR, monopulse amplitude comparison is unable to give a clear indication of the angle due to the power received from the target at different beam.

5.2 Future Works

Future works on this HFSWR Simulator could comprise of the simulation of Bragg lines in its EM environment where phase and amplitude could be included in individual pulse. This will negate the necessity of performing an additional simulation for the Bragg clutter as well as increase the fidelity of the simulator especially when signal processing gain of the Bragg clutter is done together with the target(s).

Beside Bragg clutter, other interferences like jamming, lightning, atmospheric interferences could be added to enable users to come up with clutter, interference cancellation so that the target is able to retrieve in a highly cluttered, highly noisy environment. This is especially important if tracking is to be done as this clutters or inference will hide the intended target(s) of concern.

6. REFERENCES

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