Microsatellite Radar Altimeter
Payload Design for Global Sea State Monitoring

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Summary

During the last two decades, microwave radar altimeters have proved an effective method for geographic, geodetic and oceanographic remote sensing. Compared with passive remote sensing methods, like optical CCD, altimeters have unique properties in geometric resolution independent of sensor altitude, and all-weather imaging. From 1973, nine radar altimeters have been flown on USA, French and Japanese satellites, but never yet on a micro/minisatellite. However, with the fast development of high performance solid-state devices which decreases the on-board power consumption dramatically and the maturing of small satellite platform & payload implementation technologies, the design of light-weight, low-power consumption, and highly accurate small satellite borne altimeters has now become feasible. This gives the Surrey Space Centre, which has pioneered the design of cost-effective micro/minisatellites, a unique opportunity to widen its research on remote sensing from passive to active techniques.

The overall aim of this project is to provide near real time altimetry measurements of significant wave height (SWH) and near sea surface wind speed to worldwide users to enhance safety at sea and optimise shipping routes. As the first phase of this project, the objective of this PhD study is to address the following aspects of work.

First is to undertake a detail feasibility study for a microsatellite altimeter. To analyse the performance and requirements of each sub-system respectively in order to identify the most significant problems in the whole platform and payload design. The feasibility study shows, that for a DC power-limited microsatellite environment, the DC-RF efficiency of the RF power amplifier is very critical as it is one of the most power hungry sub-systems in the whole payload. Alongside this, the antenna's off-pointing error compensation and correction is also very important in increasing received SNR.

Bearing in mind that the project application is for near-real time worldwide altimetry broadcasting, a 12 satellite constellation is proposed and simulated. This constellation is based on the consideration of altimetry measurement grid and minima waiting time for any possible user. The simulation shows the longest waiting time occurs for the users on equator. For this case, it is less than one satellite period which is around 100 minutes depending on the altitude.

A highly efficient power amplifier design has been demonstrated as the critical factor of the whole payload design, especially from a hardware point of view. Shown in the previous feasibility analysis, its DC-RF efficiency directly dictates the capability of the radar altimeter transmitter, received SNR, and consequently the satellite size, payload and platform cost. However, it is also well known that the higher the efficiency, the higher the phase distortion that may be expected. It is very common that this error may be larger than 40 degrees. However, a detailed understanding of the relationship between altimetry measurement, especially SWH measurement, and the phase distortion is still unclear. Therefore, the objective of this Ph.D. study is first to outline this relationship by a simulation using a model that considers the errors from both the signal source and the power amplifier. The simulation results show the power amplifier influence is more...
significant than that of signal source in SWH estimation, and that the phase errors influence is worse for lower SWH conditions. It is recommended from the simulation that the group delay error of the whole transmitter link, after the chirp generator, should be well controlled to be under 0.5ns.

In the payload design, Class-F is chosen as the amplifier operation mode due to its high efficiency and fewer harmonic frequency components. The difference between the operational principles of second and third harmonic peaking Class-F amplifiers have been illustrated by the simulation. Both of them can achieve high efficiency and high gain, however the third harmonic peaking Class-F is simpler to implement. Therefore it was chosen by the final design. In the simulation, a large signal STATZ model is set up, followed by the S-band Class-F amplifier design simulation and the implementation of third harmonic peaking Class-F amplifier. Based on this, an adaptive feedback group delay equalizer is proposed as a solution for the phase error compensation within the whole chirp signal swept bandwidth. A very simple but effective phase error detection and calculation circuit is designed, built and measured. The test branch results are very satisfying. Its small size and lower power consumption makes it very suitable for a compact microsatellite environment.

In summary, the possibility of a medium resolution microsatellite borne radar altimeter for optimising shipping routes is investigated in this study. A 12 satellites constellation is proposed for achieving near real time altimetry broadcasting. The key payload design problems are identified in a thorough feasibility study: the restriction corresponding to these main problems is quantified via the SWH estimation simulation. A feedback linearization method is proposed as a promising solution for the compact microsatellite design with high power efficiency requirements, demonstrated by both simulation and hardware implementation results.

Keywords: Altimeter, Microsatellite, SWH, Constellation, Phase distortion, Class-F amplifier, Linearization,

WWW: http://www.eim.surrey.ac.uk/
To My Parents, for Their Invaluable Support and Encouragement throughout My Life
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Chapter 1. Introduction

1.1 Background
On this planet, over 70% of the Earth surface is covered by ocean and so there is a great demand for monitoring global sea state: from locating ocean fronts, eddies and swells to provide real or near real time sea surface wave height and wind speed. In the past twenty years, several instruments have been developed to fit these requirements, among them radar altimetry has been demonstrated as one of the most effective and accurate methods for sea state monitoring.

Functionally, the applications of the radar altimeter can be grouped into two types: long-term and short-term measurements. A long-term history of altimetry data provides a profile of sea surface topography along the satellite track. These profiles can give fundamental information for oceanography as well as glaciology and land topography studies. On the other hand, the short term fast delivery altimetry significant wave height (SWH) and sea surface wind speed sensing data play important roles in commercial sea state alarm service such as shipping route management and safety at sea. Figure 1.1 shows the global sea surface significant wave height, while Figure 1.2 shows the global wind speed record. The measurements were carried out by the ERS-2 altimeter, the remote sensing satellite that was launched by the European Space Agency in 1992.

Figure 1.1 Global sea surface significant wave height (SWH), measured by ERS-2 (reference: http://www.satobsys.co.uk/)
Traditionally, radar altimeters are rarely launched as the main satellite payload. They were usually integrated and launched together with other radar remote sensing payloads such as SAR and scatterometer. These missions include SEASAT, ERS-1/2, and TOPEX, etc. The reasons are various. Among them, the cost of satellite platform and launch is one of the main considerations. As the radar altimeter is a reasonably small instrument, and the understanding of its measurement data will be restricted to certain level without the help from other sensing instruments. Fortunately, altimeter design techniques have developed very quickly. In the last two decades the payload transmitter power requirement has dropped from several kilowatts down to less than 10 Watts while the measurement accuracy has increased simultaneously. On the other side, the appearance of powerful small satellites reduces the platform cost dramatically. All these enable the design of a highly accurate and low cost radar altimeter.

The Surrey Space Centre (SSC), the world pioneer in low cost microsatellite design, has already demonstrated its great capability in the area of optical remote sensing. The latest development of microsatellite (TMSAT for ThaiPhutt) can provide 100 meters multi-spectral imaging. While the more powerful minisatellite (UoSAT-12) can now provide a 32 meters multi-spectral and 10 meters resolution black and white imaging. In the meaning while, the application of active remote sensing methods – radar remote sensing also has attracted lots of effort at Surrey. SSC has been
involved in several radar payload researches such as HF ionospheric top sounder, altimeter, synthetic aperture radar (SAR). Among these, the altimeter has been demonstrated as one of the most suitable radar payloads for the microsatellite, because it requires less transmitter power and on-board signal processing.

The major motivation behind sea state monitoring is the great concern for safety at sea and the insurance risk, followed by the consideration of ship routing economy. The project, jointly proposed by Satellite Observation System (SOS) and Surrey Satellite Technology Limited (SSTL), called GANDER (Global Altimeter Network Designed to Evaluate Risk), has attracted significant support and interest from a number of important customer segments, including the military. This has led to my Ph.D. project – Microsatellite Radar Altimetry Payload Design for Global Sea State Monitoring. The achievement of radar altimetry by microsatellite will no doubt, be a significant step towards SSC’s goal of designing final light SAR payload.

1.2 Ph.D. Study Road Map

This Ph.D. research program is a payload system study and has covered a wide range of areas – from the top level system feasibility analysis, & network constellation to the detailed specific payload subsystems design, which is believed are the most critical subsystem in the whole platform and payload design.

The structure of the Ph.D. research is shown by the following study road map.
Ph.D. Study Road Map

Objectives – World wide fast delivery sea state monitoring to measure sea surface significant wave height and wind speed by using microsatellite radar altimeter.

Radar altimeter + Microsatellite constellation

System

Constellation arrangement

Platform and payload feasibility analysis for limited platform and launch cost

identify main critical problems

Tight power budget
- Limited antenna size/power

Nadir pointing stability

Solution

Select Delay/Doppler altimeter & Identify key parameters in system design

1. Phase control of signal
2. DC-RF conversion efficiency

problems to be solved

Phase Errors
Phase error influence on SWH and wind speed on-board estimation
- Identify the different hardware subsystems influence on SWH estimation precision
- Quantify the link between hardware group delay error and SWH estimation under calm to rough sea conditions
- Give a hardware design guide for acceptable estimation

Amplifier Efficiency
Highly efficient Class-F power amplifier simulation and design
- Theoretical analysis of Class-F advantages on efficiency and output power capability
- Thorough study of Class-F group delay performance
  ◊ Impedance match network influence
  ◊ Major model parameters influence

Highly Efficient Class-F Power Amplifier
Feedback Group Delay Equalizer – Simulation design & bench implementation
- A compact & novel low power consumption phase detector design
A compact & novel low power consumption phase detector design

1.3 Outline of Thesis

Chapter 1 establishes the contents and requirements for the study. Chapter 2 is a detailed feasibility study for the microsatellite borne radar altimeter. The study focuses on analysing the radar payload requirements on satellite sub-systems and how the platform limitation influences the radar instrument performance. Trade-offs between platform capability and payload performance are unavoidable. During the feasibility study, several significant problems that exist in the hardware design have been identified and investigated.

To achieve real or near-real-time provision of data, a network of multiple satellites is required. Chapter 3 gives an illustration of a 12-satellite constellation arrangement. The most concern associated with the constellation characteristics is the user waiting time and altimetry map grid. They will be explored according to user requirement, and then be reviewed via the system simulation.

Chapter 4 outlines the relationship between system hardware performance, especially the payload transmitter/receiver phase and amplitude with regard to the altimeter’s fast on-board SWH estimation. The signal source and the power amplifier are the two subsystems that are most critical. The results are used as a design guide in quantifying the payload requirement for the subsystem hardware design. This leads to the selection of a linearised Class-F power amplifier.

In Chapter 5, one of the most critical sub-systems in the hardware design, a highly efficient Class F power amplifier, has been simulated and implemented. Theoretical analysis for two different types of harmonic peaking Class-F amplifier methods has been carried out. All the simulations are based on a self-developed MESFET large-signal model that has been acknowledged as one very promising design method for the highly nonlinear amplifier design. An amplifier has been built to verify the simulation, and the measurement results are presented.

Chapter 6 contains a step-by-step design breakdown for a simple, compact but very effective group delay feedback lineariser. Test bench measurement results are given in each step. The linearizer’s merits of low power consumption and compactness make it very suitable for microsatellite applications.

Chapter 7 summarises the research undertaken, the results and discusses the future work.
Chapter 2. Radar Altimeter Feasibility Study for Microsatellite

This chapter presents a system feasibility study for a microsatellite borne medium resolution radar altimeter. After the basic introduction of radar altimetry history and its operational principles, the feasibility analysis of payload and each platform sub-systems performance will be given. The analysis includes orbit determination, attitude determination and control system, station keeping, communication links, payload design and on-board DC power budget. At the same time, because the fast estimation and delivery of SWH and sea surface wind speed are the overall main concerns, especially during the bad weather condition, propagation effects have also been included in the study. The Delay/Doppler altimeter, as a newly proposed power efficient altimeter design idea, will be introduced in the final part of this chapter, as this concept is very promising for a platform with restricted capabilities. The question — “what are the most critical design problems for the microsatellite borne radar altimeter and what are the best trade-off solutions?” will be discussed.

2.1 Introduction

Up to now, satellite altimetry is the only known method by which oceanographers can precisely and accurately measure sea surface topography [Marth P. C. et al. 1993]. This technology also plays an important role in glaciology and land topography study, especially for remote areas that are difficult to survey.

The first space-borne microwave radar altimeter appeared in 1973 on SKYLAB [McGoogan J. T. et al. 1974], from then on more than ten space-borne altimeters have been launched and testified the important role of altimeters in climate related research. During the past two and a half decades, this technology has been developed from proof of concept, where oceanographic signals even as large as the Gulf Stream were ambiguous in the data, to a point where tracking of individual features such as eddies with only a few centimetre amplitude is possible. Table 2-1 is a summary of some of these altimeter missions and their main system parameters.

<table>
<thead>
<tr>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
<th></th>
</tr>
</thead>
<tbody>
<tr>
<td>Mean altitude (km)</td>
<td>435</td>
<td>800</td>
<td>800</td>
<td>800</td>
<td>800</td>
<td>1334</td>
</tr>
<tr>
<td>Orbit inclination (°)</td>
<td>13.9</td>
<td>108</td>
<td>108</td>
<td>108</td>
<td>108</td>
<td>108</td>
</tr>
<tr>
<td>Frequency (GHz)</td>
<td>13.5</td>
<td>13.5</td>
<td>13.5</td>
<td>13.5</td>
<td>13.5</td>
<td>13.6/5.3</td>
</tr>
<tr>
<td>PRF (Hz)</td>
<td>250</td>
<td>1020</td>
<td>1020</td>
<td>1020</td>
<td>1020</td>
<td>4000/1000</td>
</tr>
<tr>
<td>Range measurement precision (cm)</td>
<td>&lt;100</td>
<td>7</td>
<td>7</td>
<td>5</td>
<td>7</td>
<td>2</td>
</tr>
</tbody>
</table>

Table 2-1. Development of radar altimeters
Table 2-1 shows that, although the radar measurement accuracy has increased dramatically, the payload transmitted power also has decreased step by step. This trend is due to the adoption of pulse compression techniques, which relax the payload requirement for high output peak power by transmitting a longer pulse and compressing it in the subsequent signal processing.

Among these missions, POSEIDON is the first satellite-borne solid-state altimeter, and its appearance well represents the trend of altimeter development towards light weight and lower mass. The major characterises of POSEIDON are [Raizonville P. 1988]:

- Weight: 40 kg (without redundancy)
- Power consumption: 65 W DC power (average)
- Telemetry altimetry data rate: 1 kbps

As the first phase of this project to offer low cost, fast access altimetry measurement which can provide fast response towards the global sea state change, the objective of this study is, therefore, to carry out a detailed payload and platform design feasibility research.

2.2 Radar Altimeter Operation Principles

The principle of altimeter operation is conceptually simple: a nadir looking, high-resolution radar that measures the distance from the satellite to the ocean's surface with high accuracy.

![Altimeter range measurement diagram](image)
Chapter 2. Radar Altimeter Feasibility Study for Microsatellite

An accurate knowledge of the propagation delay will not be obtained without the help from other instruments, therefore it will influence the accuracy of the on board range measurement. Fortunately it does not influence the estimation of SWH which is based on the analysis of return waveform shape. This characteristic simplifies the payload design considerably which will be explained later in this chapter.

2.21 Beam-limited, Pulse-limited and Delay/Doppler Altimeter

Beam-limited and pulse-limited altimeters are the two well known basic altimeter concepts. Beam limited means the altimeters footprint is the region on the sea surface illuminated by the antenna beam angle defined by the half power point of the antenna gain pattern. On the contrary, pulse limited altimeter means the illumination area of sea surface is limited by the transmitted pulse width. Figure 2.2 shows their difference.

During the radar payload design, there is a trade-off in choosing the optimum pulse length and the antenna beam-width. For altimetric measurements of the mean sea surface elevation, the antenna beam width should be relatively wide so that the footprint is large enough to filter out the effects of waves on the sea surface so that to obtain a measure of mean sea level. At the same time, the footprint should be small enough to yield a meaningful measure of mean sea level (i.e., smaller than the order of 50 km) [Chelton D. B. et al. 1989]. Moreover, a broad beam width results in contamination of the measurement when land is present in the side lobes of the antenna pattern. A reasonable compromise is to design the altimeter so that the footprint diameter is of a few kilometres.

From a satellite altitude of 1000 km, an altimeter footprint with a 5 kilometres diameter could be achieved with an antenna beamwidth of about 0.3 degrees. Such a narrow beam-width requires a large antenna diameter of about 5 meters for a 13.6 GHz transmitted signal, which is impossible to be accommodated by a low cost microsatellite. Another even more stringent limitation of a beam-limited altimeter design is that the measurement accuracy is very sensitive to antenna pointing errors. While it is difficult for the microsatellite to measure and control the satellite pointing angle better than 0.5 degrees, or even to accurately acknowledge it.
Although up to now the entire launched altimeters are pulse-limited altimeters, beam limited altimeters are still very attractive to many designers due to their lower power requirements. However the requirements on precise satellite attitude control restricts the use of beam limited altimeter. The Delay/Doppler altimeter is a recently proposed altimeter concept by Raney [Raney R. K. 1997] which by nature, is a design concept between beam limited and pulse limited altimeters. Its merits are that it is less sensitive to terrain slope and requires less transmitted power. Thus this new concept is a very promising candidate for the next generation of altimeters. Currently the signal processing algorithms are under development in the Surrey Space Centre. We will explain its basic concept in more detail in section 2.5.

Due to the difference of the illumination area for pulse and beam limited altimeter, the radar equation for these two conditions are different, showed by equation (2-1) [Griffiths H. D. 1986]:

\[
\frac{P_r}{P_t} = \frac{\sigma^0 G^2 \lambda^2 c dL_p}{64\pi^2 h^3} \quad \text{(pulse-limited)}
\]

\[
\frac{P_r}{P_t} = \frac{\sigma^0 G^2 \lambda^2 L_p}{16\pi^2 h^2} \quad \text{(beam-limited)}
\]  

in here

- \(P_r\) is the received power, in Watts
- \(P_t\) is the transmitted power, in Watts
- \(G\) is the antenna gain
- \(h\) is the satellite-borne altimeter altitude above sea surface, in meters
- \(L_p\) is the two-way loss
- \(\sigma^0\) is the backscatter coefficient
- \(c\) is the velocity of light
- \(\lambda\) is the transmitted signal wavelength
- \(\tau\) is the radar pulse after compression

Equation (2-1) shows that pulse-limited and beam-limited altimeters have different dependence with satellite height and the beam limited altimeter has no relationship with the pulse length. The equation also shows that, if the received SNR is the same, the beam-limited condition may require less transmitted power. This is the advantage of beam-limited altimeter. However although the beam limited altimeter has the advantage of requiring less transmitted power, its strict requirement for larger antenna and highly accurate and stable nadir-pointing control sub-system make it unsuitable for a microsatellite-borne altimeter. Therefore, pulse limited altimeter becomes the common
operation choice. All the existing satellite borne altimeters missions are operating at this mode.

For comparison, it may be interest to have a quick look at the recently proposed Delay/Doppler altimeter concept. Its radar equation can be expressed as [Raney R. K. 1997]

\[ \frac{P_r}{P_t} = \frac{\sigma^0 \lambda^2 (TB) \sqrt{c\tau}}{(4\pi)^3 h^{3.5}} \]  

(2-2)

Equation (2-2) shows the radar received power has an \( h^{-3.5} (c\tau)^{3/2} \) dependence on the satellite orbit altitude \( h \) and the compressed pulse length \( \tau \). More efficient than the conventional pulse limited altimeter, which has an \( h^{-3} (c\tau) \) dependence. For example, the SNR improvement can be up to 10dB if we assume an altimeter has the same parameters as TOPEX while operating in the Delay/Doppler altimeter mode. This concept could be viewed as an operating between pulse limited and beam limited altimeters, combining their advantages and overcoming their disadvantages but at the price of slightly more complicated signal processing. We will address these in more detail in section 2.5.

### 2.22 Radar Altimeter Return Echoes and The Processing

In the pulse limited altimeter, the area of the circular footprint contributing to the signal received by the altimeter can be determined from the altitude \( h \) of the satellite above the nadir point and the slant range from the satellite to the outer perimeter of the circle. The two-way slant range for the signal received by the altimeter at time \( \Delta t \) is \( 2h + c\Delta t \) where \( \Delta t \) is measured relative to the time when the leading edge of the pulse reflected from satellite nadir is received by the altimeter.

[Chelton D. B. et al. 1989] shows the area \( A_{out} \) on the sea surface contributing to the radar return for time \( \Delta t < \tau_0 \) is

\[ A_{out}(\Delta t) = \frac{n \hbar c \Delta t}{1 + h / R_e} \quad \Delta t \leq \tau_0 \]  

(2-3)

In here,

- \( \Delta t \) is the two-way travel time difference measured relative to the time \( t_0 \) when the return signal from mean sea level at satellite nadir is received.
- \( t_0 \) is the two way travel time between the satellite and mean sea level at satellite nadir
- \( \tau_0 \) is the sweep period of the altimeter transmitted chirp
$R_e$ is the Earth radius.

After time $\tau_0$, the area defined by the outer boundary of the annulus contributing to the radar return continues to grow linearly with time according to (2-3). The area $A_{in}$ within the inner boundary of the annulus after time $\tau_0$ is given by

$$A_{in}(\Delta t) = \frac{\pi hc(\Delta t - \tau_0)}{1 + h / R_e}$$

which grows linearly from the time when the trailing edge of the pulse reflected from satellite nadir is received by the altimeter. The total area of the annulus is therefore given:

$$A_{ann} = A_{out} - A_{in} = \frac{\pi hc \tau_0}{1 + h / R_e}$$

It is evident from equation (2-5) that the footprint area contributing to the radar return remains constant after the trailing edge of the short pulse returns from the calm sea surface at satellite nadir.

The Significant Wave Height (SWH) corresponds approximately to the crest-to-trough height of one third of the largest waves in the altimeter footprint, generally denoted as $H_{1/3}$. For a wave height $H_{1/3}$, the travel time for the trailing edge of the pulse to propagate from wave crest to trough is $c^{-1} H_{1/3}$. Accounting for two-way travel time, the total rise time for the altimeter footprint area contributing to the radar return from a rough sea surface is therefore $\tau + 2c^{-1} H_{1/3}$. The maximum footprint area $A_{max}$ contributing to the radar return when the expanding circle becomes an annulus then becomes:

$$A_{max} = \frac{\pi h (c \tau + 2 H_{1/3})}{1 + h / R_e}$$

Thereafter, the footprint area is constant. The maximum footprint area contributing to the radar return increases linearly with SWH. The area on the sea surface contributing to the radar return can be shown by Figure 2.3.
Chapter 2. Radar Altimeter Feasibility Study for Microsatellite

Figure 2.3. Power received by a pulse limited satellite altimeter as a function of time

$t_c$ is the minimum time that radar can get the return signal

According to the nature of radar signals reflection, it is known that the two-way travel time, and hence the range to mean sea level, can be obtained by tracking the half-power point on the return waveform and that SWH can be determined from the slope of the leading edge of the waveform in the vicinity of the half-power point, as shown in Figure 2.3. The maximum footprint area contributing to the radar return increases linearly with SWH. For example, the TOPEX/POSEIDON of 1335 km altitude, the effective footprint size may increase from 2.0 km to 13.4 km for a SWH of 20 m.

Because the surface represents an extended, rather than point target, the received echo is not a replica of the transmitted pulse. Moore [Moore R. K. 1957] demonstrated that the average radar power return from a rough surface for near-normal incidence scattering could be expressed as a convolution of:

- the transmitted pulse shape
- a term which included effects of antenna pattern, off-nadir pointing angle, surface properties, and distance

Brown [Brown G. S. 1977] used this convolutional model with assumptions common to satellite radar altimeter systems to produce a simplified closed-form expression for the average rough surface impulse response function. The altimeter mean return waveform $W(t)$ is given by the convolution of three terms:

$$W(t) = P_{rs}(t) * q_s(t) * P_r(t)$$  \hspace{1cm} (2-6)
\[ P_{FS}(\tau) = \frac{G_0^2 \lambda c \sigma_0^2(0^\circ)}{4(4\pi)^2 L_p h^3} \cdot \exp \left[ -\frac{4}{\gamma} \sin^2 \xi - \frac{c\tau}{h} \left( \frac{4}{\gamma} \cos 2\xi + \alpha \right) \right] \cdot I_0 \left( \frac{4}{\tau} \sqrt{\frac{c\tau}{h}} \sin 2\xi \right) \] (2-7)

where

\( \gamma \) an antenna beamwidth parameter defined by assuming a Gaussian approximation to the antenna gain for an angle \( \theta \) off the antenna axis,
\[ G(\theta) = G_0 \exp \left[ -\frac{1}{2} \sin^2 \theta \right] \]

\( G_0 \) the radar antenna's bore-sight gain

\( L_p \) the two-way propagation path loss

\( \sigma_0(0^\circ) \) the ocean's radar back-scattering cross-section at normal incidence

\( g_s(t) \) : radar observed surface elevation probability density function. Usually is assumed to be the skewed Gaussian form, can be expressed as:
\[ \left( \frac{c}{2} \right) g \left( \frac{c\tau}{2} \right) = \frac{1}{\sqrt{2\pi} \sigma_s^2} \exp \left( -\frac{\tau^2}{2\sigma_s^2} \right) \] (2-8)

\( \sigma_s \) : the rms height of the specular points relative to the mean sea level;

\( p(t) \) : the effective transmitted pulse shape as observed by the receiver refer to point target response, and it has a theoretical shape of the \( \left( \sin^2 x \right) / x^2 \).

Figure 2.4 shows the processing of this convolution.
Chapter 2. Radar Altimeter Feasibility Study for Microsatellite

2.2 Convolutional Procedure

(a) Convolutional procedure

\[ S_a(t) \]

\[ * P_{RS}(t) \]

\[ * q(t) \]

\[ = P_r(t) \]

(b) Low SWH results (calm sea condition)

(c) High SWH results (serve sea condition)

Figure 2.4 The Brown mode convolution results over an ocean surface

See from Figure 2.4 that when the sea is calm, the slope of the return waveform leading edge will keep steep, while for the high SWH condition, it will take a longer time for the received waveform to reach its peak point, as we expect.

2.3 Microsatellite Platform for Altimeter Mission

In this section, the choice of a suitable low cost microsatellite platform is examined in the context of optimising the radar altimeter payload performance. Trades-offs have been investigated to identify the key issue during the platform and payload design.
Table 2-2 lists the characteristics of the current available microsatellites and minisatellite by Surrey Space Centre.

<table>
<thead>
<tr>
<th>Platform</th>
<th>Characteristics</th>
<th>Application to GANDER</th>
</tr>
</thead>
<tbody>
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<td><strong>MicroBus-70</strong></td>
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<td></td>
</tr>
<tr>
<td>Dimensions:</td>
<td>350X350X650mm</td>
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</tr>
<tr>
<td>Mass:</td>
<td>40-70kg</td>
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<tr>
<td>Raw bus power</td>
<td>21-43W</td>
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</tr>
<tr>
<td>Processor</td>
<td>80C186 / 80386EX</td>
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</tr>
<tr>
<td>Data storage</td>
<td>256Mbytes</td>
<td></td>
</tr>
<tr>
<td>TM/TC</td>
<td>9.6k-76.8kbps</td>
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</tr>
<tr>
<td>VHF/UHF</td>
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<tr>
<td>Power system</td>
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<td>Navigation</td>
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<td>Options</td>
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<tr>
<td>Station keeping</td>
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<tr>
<td><strong>MicroBus-130</strong></td>
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<td>Data storage</td>
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<tr>
<td>TM/TC</td>
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<tr>
<td>9.6kbps-1Mbps</td>
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<td>Power system</td>
<td>Centralised, 12V</td>
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<td>ACS</td>
<td>Spin, (3-axis), 0.5°</td>
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<td>Navigation</td>
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<tr>
<td>Options</td>
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<td>Station keeping</td>
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<td>Raw bus power</td>
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<td>Processor</td>
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<tr>
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<td>Distributed, 28V</td>
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</tr>
<tr>
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<td>3-axis (Spin), 0.1°</td>
<td></td>
</tr>
<tr>
<td>Navigation</td>
<td>GPS or NORAD</td>
<td></td>
</tr>
<tr>
<td>Options</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Station keeping</td>
<td>Cold gas</td>
<td></td>
</tr>
</tbody>
</table>

Table 2-2 Applicability of Surrey busses to the altimeter constellation mission

The total mission cost comprises the cost of platform, payload as well as launch. Estimation has been made based on these three different buses, shown in Table 2-3.
After this initial cost trade-off, the SSTL MicroBus-70 has been chosen as the target system platform, as even go for the MicroBus-130 will increase the cost by 80%. However this gives a big challenge to the platform & payload design, as the payload imposes a relatively high requirement on DC power supply, attitude control processing and station keeping. In the following sections, we will analyse the platform sub-system feasibility and impact on payload design one by one.

### 2.4 Subsystems Feasibility Analysis

A feasibility study for microsatellite radar altimeter is carried out in this section. The first half of this section is a payload requirement break down, which includes the requirements on orbit determination, attitude control, station keeping, communication link and the DC power supply sub-systems. It is then followed by a detailed discussion on rain effect on SWH measurement, as in bad weather the sea state condition is specially required by the customers.

#### 2.4.1 Precise Orbit Determination

The primary goal of most altimetry missions is to make precise measurements of mean sea levels for the study of global ocean circulation. In order to exploit fully the altimetric data, the satellite’s radial ephemeris must be known to sub-decimetre accuracy. From the early days of altimetry, enormous work has been put in this area and several methods have been developed to achieve that aim. For example, the TOPEX/POSEIDON altimeter, the most complex altimeter up to now, has four subsystems been implemented to give a precise orbit determination. They were:

- Satellite laser tracking (SLR) which is the nominal system for altimeter orbit determination computations;
- Doppler Orbitography and Radiopositioning Integrated by Satellite (DORIS) tracking;
- Global Positioning System (GPS) tracking;
- Tracking and Data Relay Satellite System (TDRSS) tracking;

On a 50-100 kg microsatellite, it is impossible to implement all the above subsystems on board. Fortunately the main task for this mission, as stated in the previous section, is different from the previous missions - we are only interested in the satellite relative height above sea surface and the related radar return waveform. Therefore the height of the satellite to the marine geoid is not critical for this mission.
However the altimeter payload does need the reasonably accurate orbit information to format output telemetry data for world-wide users to locate the source of the altimetry data. The on board GPS receiver that currently can provide an accuracy of up to 10ms will play the role. It is believed this resolution can easily satisfy the user requirements for commercial SWH applications.

2.42 Attitude determination and control system (ADCS) subsystem

The attitude determination and control system (ADCS) subsystem is one of the most critical elements that influences the altimeter measurement accuracy. The attitude errors can be divided into two parts:

- The vertical component of satellite velocity relative to the sea surface, which introduces a Doppler shift in the frequency of the return signal, received by the altimeter and therefore introduces an error in the altimeter range estimation.
- The pitch and roll error which will result in an on-board antenna pointing error away from nadir, and thus increase the altimeter measurement error.

2.42.1 Doppler shift errors

The vertical component of satellite velocity relative to the sea surface ranges from 0 - 7 m/s for a 800 km sun synchronise orbit, for the extreme case this value may increase up to 30 m/s. Velocities of this magnitude introduce a Doppler shift in the frequency of the return signal received by the altimeter. Since the frequency of this IF signal produced by pulse compression is directly related to the range to mean sea level, a Doppler shift in frequency introduces an error in the altimeter range estimate.

The two-way Doppler shift for a transmitted frequency $F$ is given by:

$$\Delta f_D = \frac{2v}{c} F \quad (2-9)$$

where $v$ is the vertical component of relative velocity. The height error corresponding to this Doppler shift is therefore:

$$\Delta h_D = \frac{c \Delta f_D}{2} = \frac{c \left( \frac{\Delta f_D}{Q} \right)}{2} = \frac{vF}{Q} \quad (2-10)$$

For the sweep rate of $Q = 3.125 \text{ kHz/ns}$, it corresponds to a Doppler shift range error of 13 cm for a vertical relative velocity of 30 m/s (maximum). This error is similar in nature to the EM bias in that the return waveform is shifted in frequency but otherwise unchanged in shape and therefore introduces an undetectable range error. As it does not influence the SWH estimation which relates to the returned waveform shape, and the
Doppler range error can only be corrected on ground post processing, we will omit it, as a 10 cm level range error is an acceptable value for the present service.

2.422 Mis-pointing influence

The most significant influence of the satellite mis-pointing angle is to decrease the received SNR, and therefore decrease the estimation of the backscatter cross section. There are two sources of mis-pointing errors of the satellite radar altimeter: a static component which must be traded off between the various distortion of the satellite, and a time varying component associated with the thermal distortion (due to the effect of temperature changing along the orbit) of the satellite. The latter has a period that will decide the time constant of the smoothing \( \alpha-\beta \) filter. The typical value of ERS-1 is 0.2° static mis-pointing error and a maximum harmonic error of 0.1°. It is only possible to keep the altimeter pointing normally to the ocean surface to within a certain accuracy; and the on-board estimates will be biased unless some correction is made for the "mis-pointing". For the well-known Brown model radar altimeter return waveform, the existence of mis-pointing error \( \xi \) will mainly influence the flat surface impulse response, which can be expressed as:

\[
P_{rs}(\tau) = \frac{G_0^2 \lambda c \sigma^0(0^\circ)}{4(4\pi)^2 L_p h^3} \cdot \exp \left[ -\frac{4}{\gamma} \sin^2 \xi - \frac{ct}{h} (\cos 2\xi + \alpha) \right] \cdot I_0 \left( \frac{4}{\gamma} \sqrt{\frac{ct}{h}} \sin 2\xi \right) \quad (2-11)
\]

in here,

\[
\frac{4}{\gamma} = (\ln 4) / \sin^2(\theta_a / 2)
\]

\[\theta_a\] usual antenna angular full width at half power

\[G_0\] radar antenna boresight gain

\[\lambda\] radar wavelength

\[\sigma_0(0)\] ocean's radar backscattering cross section at normal incidence

\[L_p\] two way propagation path loss

The mis-pointing error can give rise to three principal effects. The first effect, is a distortion of the leading edge of the average return. In general, a pointing error will give rise to a decrease in slope of the leading edge of the return and this could be misinterpreted as a manifestation of surface roughness effects. A second impact is to effectively reduce the level of backscattered power and, therefore, give rise to erroneously low values of \( \sigma^0 \). Also shows in Figure 2.6, see the mis-pointing error would result in nearly half energy loss in the processor, which means the \( \sigma^0(0) \) would be 3 dB lower than the real value.
The third effect is when the pointing angle is larger than the on-board antenna half-power angle, the trailing edge will have an increasing slope, which will lead to the tracking loss of the half power point. The situation is shown in Figure 2.5, where the mispointing angle is 0.5 degrees. In here the antenna half-power angle is set to 0.75 degrees.

The usual way of correcting the mispointing error is to apply a range window

$$W(\tau, \eta) = \exp\left(-\frac{4c\tau}{h\gamma}\cos 2\eta\right)I_0\left(\frac{4}{\gamma}\sqrt{c\tau/h\sin 2\eta}\right)$$

(2-12)

upon the received signal to correct the trailing edge of the real received signal before the real estimation for SWH and backscatter coefficient, in here $\eta$ is the estimated mispointing angle. Therefore the actual signal would become

$$R(\tau, \xi, \eta) = R(\tau, \xi) \cdot W(\tau, \eta)$$

(2-13)

Figure 2.6 shows the simulation results, the upper line is the received signal with mispointing error and after weighting ($\xi = \eta$), the middle line is the distorted received signal without weighting, and the lower one is the received signal without error.
There are two weighting effects we could read from Figure 2.6. First is the weighting window increases the received signal amplitude, which could compensate the backscatter coefficient estimation loss due to the mispointing influence. The second effect is the weighting window flattens the trailing edge of the return waveform and drives the slope of the trailing edge to zero. Further simulation shows when $\xi \neq \eta$, the slope of this trailing edge will change with the different condition of $\xi$ and $\eta$, shown in Figure 2.7.
Mathematical analysis shows the $R(\tau, \xi, \eta)$ could be approximated as:

$$R(\tau, \xi, \eta) = L\left[1 + \text{erf}\left(\frac{\tau}{\sigma_\tau \sqrt{2}}\right)\right]\left[1 + \left[\frac{4}{\gamma} \sqrt{c \tau / \bar{h}}\right]^2 \left(\xi^2 - \eta^2\right)\right]$$  \hspace{1cm} (2-14)

It can be seen from the equation that the slope of the trailing edge of the return is governed by the difference $\xi^2 - \eta^2$. If $\xi > \eta$, the slope is positive, if $\xi < \eta$ the slope is negative. Thus the trailing edge slope may be used to update the current estimate of $\eta$. The mispointing control loop attempts to drive the trailing edge slope of the range window return to zero, at which point the estimated and actual values of the mispointing angle will be the same. The above analysis shows the waveform after mispointing compensation can also serve as a reference for the satellite ADCS systems.

If one assumes the real time altimetry output data rate is 10 Hz which means the altimeter will give 10 samples of data per second, then 100 ms is the smallest time period on which the signal processor interprets the waveforms. For the typical SSTL microsatellite, the ADCS system will update its estimation around every 20 seconds. This is a very long time compares with the 100 ms. However, from Figure 2.8(a) the plot of the mispointing angle changes vs. time, we may notice the rate of mis-pointing changing is extremely slow. Therefore for this altimeter application, we could just use a constant value to weight the returned waveform during the 20 seconds period.

(a) Typical SSTL microsatellite mis-pointing plot 360second period
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Satellite off-pointing angle stabilise procedure within 12 hours

(b) Long term satellite ADCS operation procedure, starts from initial condition

Figure 2.8 Typical SSTL microsatellite off-pointing angle vs. time

Figure 2.8(b) is a simulation of mis-pointing error changing with time for a typical SSTL microsatellite, we can see once the satellite becomes stable, the mispointing error is within our requirements. Notice from the figure, there is a periodic high error period, that corresponds to the time when satellite pass the polar area, fortunately that is not the area that we are interested in. In fact, the altimeter spacecraft are likely to be launched in a 70° circular orbit, which should reduce this effect.

In summary, the attitude determination and control sub-system is very important in the radar altimeter operation. The mispointing error will directly influence of the altimeter return waveform, not only its leading edge but also the slope of the trailing edge. At the same time, if this error is larger than the on-board antenna half power angle, we will lost the track of the return waveform half power point. Therefore it is very important to have a highly accurate ADCS subsystem.

On the other hand, the radar altimeter waveform tracking implementation also can help the ADCS subsystem to determine the satellite attitude more accurately and frequently. We know the tracking function is usually achieved by a traditional second order $\alpha$-$\beta$ filter, shown in Figure 2.9. $T$ is the loop constant, depends on system requirements. The larger the $T$, the more accurate the estimation, however the longer time for the loop to settle down. Generally this filter is updated 50 times per second, which depends on the altimeter PRF. Such a high update rate enables the satellite to update its attitude knowledge much shorter than satellite ADCS’s 20 seconds updating time without increase the computation time much.
The $\alpha$-$\beta$ filter is given in the form

\[
\begin{align*}
\text{rate}(i) &= \text{rate}(i-1) + \beta e \\
\text{smth}(i) &= S(i) + \alpha e \\
S(i+1) &= \text{rate}(i) + \text{smth}(i)
\end{align*}
\]

(2-15)

in here $\text{rate}(i)$ is $i$-th signal rate, $\text{smth}(i)$ is $i$-th smooth signal, and $S(i+1)$ is $(i+1)$th estimated signal; $\alpha, \beta$ are two constants, generally in altimeter they are set as $\alpha = 1/4$, and $\beta = 1/64$ (Gao Z. 1993); $e$ is the error signal. The tracker only filters noise that is independent from one track interval to the next, but it only removes 75% of this noise power.

2.43 Satellite Operating Orbit and Station keeping

It is known that the higher the orbit, the lower atmosphere density and therefore the more stable the orbit. Research on TOPEX/POSEIDON satellite showed the atmosphere drag is not a significant problem for its 1334 km orbit height. Figure 2.10 is a comparison of off-nadir angle of two SSTL microsatellites both using gravity-gradient control. The orbit of S80/T in Figure 2.10(9a) is the same as T/P and the orbit of Cerise in Figure 2.10(b) is at a low orbit of 670 km. The figures show the off-nadir error of Cerise is around 3-5 degrees, while for higher orbit this error is generally around 1.5 degree. However the higher the satellite altitude, the higher the transmitted power is needed for the same received SNR. There is a trade-off in selecting the satellite altitude.
For any network constellation, station keeping is always a very critical question. However, due to the mass restriction of microsatellite, the fuel propulsion budget and therefore the tank size must be evaluated carefully to decrease the platform price.

Atmospheric drag, geomagnetic force and solar pressure are the main influence on the stability of station-keeping. Among these the requirements are dominated by orbital aerodynamic drag. The energy needed for differential in-track corrections between two adjacent satellites are negligible when compared to this. Cross track errors are small and can be tolerated in the mission, thus simplifying the thruster arrangement to a simple in track system. The acceleration of the spacecraft caused by its interaction with the Earth’s atmosphere can be described using the following equation:

\[ \vec{F}_D = -\frac{1}{2} C_D \frac{A}{m} \rho(h) \vec{V}_r \sqrt{\vec{V}_r} \]  

(2-16)

where

- \( C_D \) satellite drag coefficient, typical values ranges from 2 - 2.5
- \( \vec{V}_r \) satellite velocity relative to the atmosphere
- \( \rho(h) \) atmospheric density at the altitude \( h \)
- \( m \) satellite mass
- \( A \) satellite cross-sectional area projected normal to \( \vec{V}_r \)

To cope with this drag force, energy is required. A budget called \( \Delta V \) budget is traditionally used to account for this energy. Shown from the above equation, the drag force changes with different satellite altitude as the atmospheric density is different. Figure 2.11 shows this relationship.
It can be shown from Figure 2.11 that at 800 km altitude, assuming conditions near solar maximum, a delta-vee of approximately 0.75 m/s per year is required. Although the higher the altitude, the less the ΔV is required, due to the propagation attenuation of signal path, the higher the altitude, the larger the transmitted power is required.

The existing N₂ cold gas system design for the SSTL MicroBus is a good candidate for station keeping [Drum J. et al. 1997], it could employ up to 600 bar tanks in a 330x330x200 mm³ volume. The configuration of the system is illustrated in Figure 2.12. For a 100 kg spacecraft, the system offers a delta-vee of approximately 16 m/s, using a 0.01 N thruster. A delta-vee budget of 10 m/s would permit a constellation deployment into their respective slots along the orbit within 3 months, leaving 6 m/s for more than 7 years of station keeping fuel. For the 50 kg microsatellite, less fuel may be installed but this will reduce operational life to about 3 years.
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Figure 2.12 MicroBus cold gas station keeping system (after Deum J. et al 1997)

It must be pointed out, the above analysis is only a very brief discussion, the precise calculation will directly influenced by what kind of launch vehicle we will use and therefore what type of initial orbit the satellite is likely to be planned. Future detailed analysis will be required when the launcher is selected.

2.4.4 Communication Link

The consideration of communications links can be divided into two sections. One is low data rate real time broadcasting for worldwide users, the other is a high rate link for the scientific data transfer when satellite passes the main ground station.

For the real time data broadcast model, the communication capability requirement is very low, as stated in the previous section, a 10 Hz altimeter output data rate is acceptable for most of the applications considered. The output frame includes the basic information of satellite orbit position, SWH, satellite height and wind speed information, the total data need to be down loaded for broadcasting is around 200 bytes per second. Therefore a very simple VHF link could achieve this task.
At the same time, it is also very important to have a high capability down link so that when the satellite passes one main ground station, the data stored around the orbit on the data recorder will need a higher data rate communication link – depends on the operational models.

### 2.45 On-board Power Budget

The DC power supply is very critical to the satellite platform and payload design, especially for the power limited microsatellite environment. To understand how much average power could be available for a microsatellite, a simulation was carried out for a 1000 km orbit sun synchronise orbit small satellite condition. The assumptions are:

- typical SSTL microsatellite GaAs solar panel configuration
- 1 Sun synchronous orbit with similar Hour Angle as UoSAT-5 (around 10:30 o'clock) Vernal Equinox day is chosen
- Reasonable Nadir pointing with 10 minutes z-spin (typical SSTL attitude) is assumed.

Then the simulated orbit average DC power generation is:

<table>
<thead>
<tr>
<th>Satellite Orbit</th>
<th>Orbit average Power (W)</th>
</tr>
</thead>
<tbody>
<tr>
<td>h = 700 km ($\text{inclination}=98.2^\circ, \text{period}=98.77 \text{ minutes}$)</td>
<td>26.02</td>
</tr>
<tr>
<td>h = 1000 km ($\text{inclination}=90.5^\circ, \text{period}=105.12 \text{ minutes}$)</td>
<td>27.37</td>
</tr>
</tbody>
</table>

Table 2-4 Simulated average DC power generation for different orbit

Consider the efficiency of battery charge and discharge, power converter and so on, the possible orbit average available DC power will be around 20 W. Also consider the ocean area is approximately 70% of the total earth surface, we can then assume the orbit average available DC power is roughly 28.5 W.

### 2.45.1 Antenna

The antenna is another critical subsystem in the whole payload design, its size will not only influence the received signal amplitude, but also the satellite stability as well as the launch feasibility.

Generally speaking, dish a common choice for high gain antenna at Ku band. The sizes of the previous dish varies from 0.6 meter to 1.5 meters depends on the system requirements. The above diagrams show three possible combinations of two different satellite platforms, MicroBus-70 and MicroBus-130, and two size of dishes, 0.7 meter and 1 meter diameter dish respectively.
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To determine finally which combination is the best choice will mainly be decided by the link budget calculation of radar equation, as well as the satellite attitude control requirements.

2.452 Payload power consumption analysis

The single-pulse received SNR can be determined from the standard radar equation:

\[
\frac{S}{N} = \frac{P_G A \sigma \tau}{(4\pi)^2 k TN_f h^4} \quad (2-17)
\]
In here,

1. \( P \) is the peak transmitted power, assume 7 W for our application
2. \( G \) is the antenna gain, assumes a 0.7 m Ku band parabolic antenna will be used, therefore its gain is \( G = 17.8 + 20 \log D + 20 \log f = 37.5 \text{dB} \) (\( \eta = 0.55 \))
3. \( A \) is the antenna area, \( A = \pi R^2 = \pi (0.35)^2 = 0.38 m^2 \)
4. \( \sigma \) is the radar cross section of the illuminated surface, assumes the normalized cross section \( \sigma_0 = 6 \text{dB} \), then
   \[ \sigma = 2 \cdot \pi \cdot \sigma_0 \cdot r \cdot \sigma_0 = \pi c \cdot 3.125 \times 10^{-9} \cdot 800 \times 10^3 \cdot 4 = 0.94 \times 10^7 \]
5. \( \tau_0 \) is the transmitted pulse length, \( \tau_0 = 100 \mu s \)
6. \( k \) is the Boltzmann’s constant, \( k = 1.38 \times 10^{-23} \text{ J / K} \)
7. \( T \) is the antenna temperature, assumes \( T = 300 \text{ K} \)
8. \( N \) is the receiver noise figure, assumes \( N = 5 \text{ dB} \)
9. \( h \) is the satellite height, \( h = 800 \text{ km} \)

The calculated SNR for one transmitted pulse is therefore

\[
\frac{S}{N} = \frac{P \cdot G \cdot A \cdot \sigma \cdot T \cdot N \cdot \tau_0}{(4 \pi)^2 \cdot k \cdot T \cdot N \cdot f \cdot h^4} = 10 \log \left( \frac{7 \times 0.38 \times 0.94 \times 10^7 \times 100 \times 10^{-6}}{(4 \pi)^2 \times 1.38 \times 10^{-23} \times 300 \times (800 \times 10^3)^4} \right) + 37.5 - 5 = 12.2 \text{dB}
\]

(2-18)

12 dB SNR is not a very high value however it is still acceptable. It needs to be pointed out that the above equation does not consider the system loss, which in general is around 2-3 dB. However the received Rayleigh noise decreases as the square root of the number of waveforms in the average, and this will give a roughly 3 - 4 dB of gain. In this project we will average a group of 100 pulses before forming one output data format.

Using equation (2-18), we can also plot a diagram that shows the relationship between satellite altitude and transmitted power, shown in the following figure. In here, all other parameters are the same as in the above calculation.

![Figure 2.14 Required transmitted power vs. satellite orbit](image)
Although the lower the altitude, the less the transmit power required, considering the atmosphere drag on ADCS and station keeping requirement shown in Figure 2.12, there is still a trade-off.

Now let us start to analyse the power consumption of the payload first. The whole radar altimeter payload includes

1) transmitter link – digital chirp signal generator, up conversion, power amplifier;
2) receiver link – low noise amplifier, deramping mixer, down conversion;
3) signal processing unit -- A/D converter, power spectral analyser, tracking processor and $\alpha\beta$ filter.
4) antenna

Their power consumption could be summarised in the following, respectively:

- **Transmitted link:** its block diagram is shown by Figure 2.15.

![Figure 2.15 Altimeter transmitter link block diagram](image)

- Chirp generator – Qualcomm Q2368 130MHz Dual DDS (power = (15 mW/MHz) * (clock rate)=1.95 W) and Harris H15721 10 bit 125 MHz DAC (power = 650 mW) are chosen for the chirp generator. The total power consumption is estimated around 3W.

- Up-conversion – The output signal from chirp generator is around 0dBm, according to the above diagram it will be mixed twice and multiplied six times before it finally be fed to the Ku band power amplifier. The efficiency of mixer and multiplier, however, will be very low, roughly around 10%. During this conversion link, some drive amplifiers will possibly be adopted. Fortunately all these operations are in small signal regions, so an allocation of 1W will be enough for the whole up conversion link.

- Power amplifier – Power amplifier is the part that consumes most of the DC power when a chirp signal is transmitted. Generally speaking, the average efficiency of one stage amplifier plus power combiner is around 20%, if linear phase response is required. Also the higher the output power, the lower the efficiency we would expect. Therefore the power requirement for 7 W peak power output is 35 W. Consider the duty cycle influence (10 SWH data frame
per second, 100 signals average per data frame, 100 μs per pulse, then the total duty cycle is 10%), therefore the average power required is **3.5W**.

- **Receiver Link** – the receiver link’s LNA does not need much power, in here we allocate **0.5W** for the total requirement.

- **Signal processor** – After discussions with several people who have altimeter payload or simulator experience before, we believe **5W** DC power is purely enough for this mission by using the new generation powerful digital signal processor. The whole signal processing functional block diagram is shown in Figure 2.16:

![Functional block diagram of the altimeter signal processor](image)

- Power spectral analyser – for our application each 128 points FFT shall be finished within 900 μs. DSP is an ideal choice, ADSP21020 or TSC21020E are recommended due to their radiation tolerance and powerful functionality.

- Adaptive tracker unit – The ATU can be implemented by a general 80186 microprocessor, or a programmable DSP.

- Synchronizer/Acquisition/Calibrate unit (SACU) – The proposed altimeter payload is not a dual frequency operation, thus the synchronizer task will be much easier than that of TOPEX.

- Spacecraft interface – This subsystem receives the engineering data and commands, and format the telemetry data, 1W is enough for a general use.

### 2.4.53 Power budget

In summary, for the whole system – a 50 kg microsatellite platform plus a medium resolution radar altimeter payload, the average DC requirement is 28.5 Watts, the budget calculation is shown by Table 2-5.
Refer to the available average DC power analysis above, we could see it just can meet the system requirements without much margin that we would like to have. This means a very careful design is needed to achieve this tough task!

### 2.4.6 Propagation effect

According to the regulation, currently the frequency bands that are available to altimeter operation are C band (5.3 GHz) and Ku band (13.6 GHz). The propagation effects will be more serious in Ku band than that of C band. However, due to the antenna size/gain consideration, Ku band in the end is still the final choice. Therefore we must have a clear understanding of the propagation effects at this Ku frequency band.

The propagation effects are quite complex and in many conditions are very significant, especially at higher frequency such as Ku band. Figure 2.17 shows a measurement example when the ERS-2 altimeter passed the Tropical Cyclone. The map gives the location of the data - ‘+’ for good data values, and ‘.’ for those which failed the quality checks.

![Figure 2-17 Satellite altimetry results when passing heavy rain](image-url)
Basically the propagation effects can be divided into the following three classes:

- Amplitude attenuation - at Ku band, the rain and atmospheric molecules all may degrade the transmitted and received signal amplitude, especially in bad weather conditions. For some really bad case, this attenuation may as large as 20-30 dB.
- Change in signal phase-path distance due to the slight modification of the refractive index of the rain region.
- Direct backscatter from the raindrops themselves that fill regions of the altimeter range cells above the sea surface.

The first effect will reduce the absolute level of the entire altimeter echo strength, including the plateau. If ignored, a lower value of $\sigma_0$ will be deduced, and hence an erroneously high value of wind speeds. Examination of rain storm geometries and their statistics reveals, however, than rain often will not uniformly fill a horizontal cell equivalent to that seen by Seasat (at gate 60, the footprint diameter is 9 km, shown in Figure 2.18). The more intense the rainfall, the smaller the rain cell on the average. Hence people set up models to investigate the question of whether a typical size rain cell at an arbitrary location with respect to satellite nadir will produce distortion of the echo because of the increased attenuation within the rain region [Barrick D.E. et al. 1985]. The answers are that, even in moderate or light rain, the echo distortion will produce three errors or biases:

1) $\sigma_0$ on the plateau will be lower, and hence wind speed derived therefrom will be overestimated.
2) mean sea surface height will be in error by as much as tens of centimetres, depending on how the on board algorithm responds to the distorted waveforms.
3) significant wave height will not be appreciably biased by the presence of rain, however.

Conclusion 3 is good news for us, as SWH is the primary measurement of concern to this mission.
The next step is to try to figure out how much rain attenuation may appear for different weather, so that the figure can be severed as a reference to link margin design. Because the direct broadcast services (DBS) have operated at 12.7 - 13.3 GHz for several decades, a lot of research had already been done on the propagation influence for Ku and Ka band. Table 2-6, 7 are some information of the previous studies of the two-way propagation influence on the Ku band signal [Monaldo et. al. 1984].

<table>
<thead>
<tr>
<th>Frequency (GHz)</th>
<th>Normalised attenuation coefficient (dB/km)(gm/m$^2$)</th>
<th>Cloud type</th>
<th>Liquid water content (dB/km)(gm/m$^2$)</th>
<th>Total two-way attenuation through a 1 km thick cloud (dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>13.5</td>
<td>0.17</td>
<td>Stratiform</td>
<td>0.1 - 0.2</td>
<td>0.034 - 0.068</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Cumulus Humilis</td>
<td>1.0</td>
<td>0.34</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Cumulus Congestus</td>
<td>2.0</td>
<td>0.68</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Cumulonimbus</td>
<td>2.5</td>
<td>0.85</td>
</tr>
<tr>
<td>35</td>
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<td>Stratiform</td>
<td>0.1 - 0.2</td>
<td>0.22 - 0.44</td>
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<tr>
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<td></td>
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<td>2.2</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Cumulus Congestus</td>
<td>2.0</td>
<td>4.4</td>
</tr>
<tr>
<td></td>
<td></td>
<td>Cumulonimbus</td>
<td>2.5</td>
<td>5.5</td>
</tr>
</tbody>
</table>

Table 2-6. Two-way cloud attenuation levels
It can be seen from the above tables that for the worst case, the maximum attenuation may as large as 9 dB for a 5 km rain cell at 20 mm/hr rain rate. For that case, the altimeter definitely will lose track as the payload cannot provide such a large link margin due to the on board power limitation. Therefore for real time on board processing and tracking, some algorithms should be developed to indicate lose of track window in the heavy rain condition, or the window where the measured data is unbelievable. How large the window size might be will depend on the rain cloud size, and the on board radar signal tracking ability. Nevertheless, the link margin should take account of the customers’ requirements.

### 2.5 Delay/Doppler Altimeter

The original motivation for introducing Delay/Doppler altimeter is to overcome the difficulties that conventional pulse limited altimeters meet in continental ice measurement due to the influence of higher slope. However the great advantage of the Delay/Doppler altimeter in providing an extra 10dB SNR improvement also makes it very attractive to microsatellite.

The principal objectives of using the Delay / Doppler altimeter are to operate more efficiently and more effectively. The first objectives is met by compensating for systematic range delay errors; thus the entire (beam-limited) along track signal history contributes to height measurement rather than only the much smaller pulse-limited area. Stated another way, the Delay/Doppler altimeter uses much more of the instrument’s radiated energy than does a conventional beam-limited altimeter. The second objective is met by using Doppler selectivity to reduce the width of the post-processing along-track footprint; this minimises unwanted terrain dependency of the footprint size and position. Figure 2.19 shows the area that are making use of conventional and DD altimeter respectively.
The Delay/Doppler altimeter takes advantage of pulse-to-pulse correlation to realise better performance than a conventional incoherent radar altimeter. Pulse compression is used in the range dimension, just as is customary for incoherent radar altimeters. The range signal is a long linearly frequency modulated pulse, which is multiplied by a delayed replica FM pulse immediately upon reception and low-pass filtered. The deramp strategy transforms range into a continuous wave (cw) signal whose frequency is proportional to range, relative to track point. A conventional altimeter and the Delay/Doppler altimeter both complete range compression by application of an inverse Fourier transform to convert the cw signals into range. Each height estimate is the sum of a sequence of individual detected range measurements.

In this Delay/Doppler altimeter, zero Doppler corresponds to minimum range, which is the measurement objective of the altimeter. At all frequencies away from zero Doppler the range is slightly greater. Indeed, the range increases as a quadratic function of Doppler frequency (which is the underlying cause of the quadratic phase term in azimuth which if matched "focuses" the signal). In the altimeter signal processing, deramping, the "range" is represented as a CW signal in the range dimension. Usually, the frequency of the CW signal is relative to the track point, which by definition has zero frequency. Larger ranges have larger CW frequency. Thus, waveforms derived from off-nadir (non-Zero Doppler) positions have a known CW range frequency as a function of their Doppler location. The altimeter completes range measurement by an inverse range transform, converting CW frequency into range distance (height). The off-nadir range increase can be removed simply by shifting the range frequency of the non-zero Doppler by the right amount prior to the range inverse transform. The basic signal processing subsystem of a Delay/Doppler altimeters is shown in Figure 2.20.
The LO signal is:

\[ S_{LO} = \exp(-j\omega t + j2\pi f^2) \]  \hspace{1cm} (2-19)

The received signal is:

\[ S_{received}(t, \tau) = \exp\left[-j\omega(t - \frac{2r(\tau)}{c}) + j2\pi k\left(t - \frac{2r(\tau)}{c}\right)^2\right] \]  \hspace{1cm} (2-20)

in here:
- \( t \) refers to the range dimension fast time
- \( \tau \) refers to the along track dimension slow time
- \( r(\tau) = \sqrt{h^2 + v^2 \tau^2} \)
- \( v \) is the satellite along track velocity

The output signal from the deramping processing is therefore:

\[
S_{deramp}(t, \tau) = \exp\left[j2\pi k\left(t - \frac{2r(\tau)}{c}\right)^2 - t^2\right] \exp\left(j\omega \frac{2r(\tau)}{c}\right) \\
= \exp\left[j8\pi k \frac{r(\tau)}{c} t + j2\pi k \frac{4r^2(\tau)}{c^2}\right] \exp\left(j\frac{4\pi fr(\tau)}{c}\right) \\
= \exp(j8\pi kr(\tau)t/c) \exp\left(j8\pi k \frac{r^2(\tau)}{c^2}\right) \exp\left(4\pi fr(\tau)/\lambda\right) \hspace{1cm} (2-21)
\]
The aim of along-track FFT is to compensate the range, while in time domain this is a multi-valued question because echoes are received simultaneously from all individual scatters within the illuminated beam-width of the antenna. But we notice in along-track frequency domain, this problem could be reduced to a single-valued one, because the Doppler frequency is a measure of the slope of the signal history. So in Doppler frequency domain, \( r(\tau)h \) correspondingly be changed to \( R(f,h) \), their relationship can be deduced as in the following. We notice that only in \( r(\tau) \) has the along track slow time variety. By application of stationary phase [Raney R. K. 1994], the deramping signal after along track FFT is could be expressed as:

\[
S_{\text{deramp}}(t,f) = \int S_{\text{deramp}}(t,\tau) \exp(-j2\pi \tau) d\tau = \exp \left[ j8\pi k \left( \frac{R(f)}{c} t + \frac{R^2(f)}{c^2} \right) \right] \exp \left( \frac{4\pi R(f)}{\lambda} \right) 
\]

(2-22)

\( R(f) \) could be expressed as: \( R(f,h) = \frac{h}{\sqrt{1 - \left( \frac{Af}{2v} \right)^2}} \), if we apply the Taylor series

\[
(1 - x) = 1 + x + x^2 + ... + x^n \]

then we can get the \( R(f,h) \) is

\[
R(f,h) = \frac{h}{\sqrt{1 - \left( \frac{Af}{2v} \right)^2}} = \sqrt{1 - \left( \frac{Af}{2v} \right)^2} = h \left[ 1 + \frac{1}{2} h^2 \frac{A^2 f^2}{4v^2} \right] = h + \frac{h^2}{8v^2} f^2 = h + \Delta h(h,f) \]

(2-23)

if we ignore the 2nd phase factor and the last part of the above equation, the equation (2-23) can be simplified as:

\[
S_{\text{deramp}}(f,t) = \exp \left\{ j \frac{8\pi Kh}{c} t \right\} \exp \left\{ j \frac{8\pi K}{c} \Delta h(f,h)t \right\} \]

(2-24)

The next step is delay compensation, where the compensation phase multiplier coefficient can be expressed as:

\[
\Phi(f,t) = \exp \left\{ - j \frac{8\pi K}{c} \Delta h(f)t \right\} \]

(2-25)
the result after range compensation is:

\[ S_{\text{compensation}} = \exp \left\{ j \frac{8 \pi K h}{c} t \right\} \]  
(2-26)

where by the some later processing we can at last get the same final information of the range measurement information, even the signals come from a relatively longer strip along the satellite track. The Delay/Doppler altimeter final signal response is shown by Figure 2.21, in here TOPEX altimeter parameters are assumed.

![Figure 2.21 Delay/Doppler altimeter signal response with TOPEX parameters (after Raney R. K. 1997)](image)

Obviously the Delay/Doppler altimeter is very attractive to the power limited satellite platform, however the price is more complicated on-board signal processing. Up to now no altimeter was been designed and flown in orbit by using this design idea. Although APL and ESA have been involved in this topic research for a couple of years, the detailed processing algorithm is still under development.

Currently, the Surrey Space Centre is also working at the development of this algorithm through another Ph.D. project [Sun. Y 1999]. The initial research results, shown in Figure 2.22, indicate another advantage of the DD altimeter that it is not sensitive towards the off nadir pointing error. We can see the mis-pointing error mainly influences the trailing edge of the signal response, which is the not very important in the SWH estimation. That means the DD altimeter is not sensitive towards the satellite pointing error – and thus reduces the dependence on the satellite ADCS system.
Figure 2.22 DD altimeter return signals response with/without pointing errors (after Sun Y. 1999)

Figure 2.23 shows the SWH estimation results. First is the calculated SWH mean value, and then the difference between each estimation and the mean value are studied. The root mean square of the differences of all estimations will result the variance of the Y-axis in the diagram. We can see the difference between with and without pointing error is very small.

Figure 2.23 DD altimeter SWH estimation with/without pointing errors (after Sun Y. 1999)

Although DD altimeter is a very promising altimeter design concept, up to now there has been no altimeter mission adopt this idea, there are certainly some detail areas open to discuss and further research. It maybe better to use the conventional PL altimeter first, then upload the DD altimeter’s processing algorithm in orbit to investigate its capability.
2.6 Conclusion

The objectives of this microsatellite borne radar altimeter project are to measure the sea-state wave height and wind speed for optimizing shipping routes and supply increased safety at sea through providing world wide users a near real time altimetry data by a 12 satellites constellation.

To accommodate the altimeter payload on a 50 – 100 kg low mass & power limited microsatellite platform with very low cost, a preliminary feasibility study has been carried out. The target of the study is to identify critical subsystems areas that influence the mission design, such as link budget analysis, attitude control, power supply. The conclusion drawn from the above analysis is the SSTL Microbus-70 platform could just satisfy the payload requirements – DC power supply and attitude control, if the platform is designed with care.

The major characters of this proposed low cost microsatellite altimeter are summarised as follow:

- The SWH is expected to be measured with an accuracy better than 10% or ± 50 cm whichever is greater
- The backscatter coefficient is expected to be measured with a precision of ± 1 dB
- The antenna is a parabolic dish with 0.7 meters diameter with a transmitter peak transmit power of 7 W
- The Ku band is chosen due to the antenna size consideration. The signal pulse duration is 100 µs, and the bandwidth is 320 MHz
- The telemetry data rate is 10 Hz. Each output is an average of a group of 100 return waveforms, and the duty cycle of radar payload is 10%.

Two key questions, however, demand a more thorough research in the following areas in order to achieve an efficient radar altimeter compatible with a microsatellite platform:

a) A thorough understanding of hardware performance influence towards on board fast estimation for SWH and wind speed. This research should not only for the pulse-limited altimeter, but also for the Delay/Doppler altimeter. The analysis should not only consider the chirp signal generator but also the power amplifier and up-converter.

b) Highly efficient linear Ku band 7 W output solid state power amplifier design. According to the literature survey, for Ku band, the maximum output power of
available commercial solid state device is NEZ1414-8E, which could give an 8 W maximum power. However, the DC-RF efficiency of the usual Class-A amplifier design is very low, generally is around 30%. To improve the efficiency, high efficiency amplifier design method is highly desired. At the same time, the amplifier is also required to have high output power capability, as it will involve fewer pre-drive amplifier stages.

c) **More accurate attitude knowledge and more precise attitude control method.** Although the simulated satellite pointing performance is within the acceptable range, the real on board condition is still unclear yet as UoSAT microsatellite haven't fully explored the momentum wheel and reaction wheel performance more comprehensive algorithm is required for the real on-board processing.

Among them, the problem of (b), highly efficient linear power amplifier will form the major topic of this thesis.
Chapter 3. GANDER Network Constellation

Due to the project requirements for near real time altimetry data collection, a network or constellation is very necessary. However, due to the altimeter's natural restriction of very narrow swath width, the configuration of this constellation to maintain the altimetry map grid between different satellites as small as possible appears to be a critical issue in the whole network design. This chapter will try to address this question by illustrating a 12 satellite network constellation with simulation results run by STK simulation package will be given step by step.

3.1 Altimeter Swath Width

Due to the high requirements on range resolution combined with low transmitted power, the radar altimeter signal bandwidth is very wide, therefore the transmit pulse width after compression which directly links with the bandwidth as \( \tau = \frac{1}{B} = \frac{1}{320 \times 10^8 \text{ Hz}} = 3.125\, \text{ns} \) is very short. While the altimeter's swath width is decided by the compressed pulse length as well as the satellite altitude \( h \), expressed as:

\[
\text{Swath width} = 2c\, \text{th} = 2\sqrt{3\times10^8 \times 3.125 \times 10^{-9} \times 1000 \times 10^3} = 1.936\, \text{km}
\]  

(3-1)

in here we assume the satellite altitude is 1000 km, the swath width diagram is shown in Figure 3.1.

![Swath width diagram](image)

Figure 3.1 Radar altimeter swath width diagram

At the same time, we can calculate that for a sun synchronous orbit the distance between two continuous passes is:

\[
D = \frac{\theta}{360} \times 2\pi (R_s + h) = \frac{26.28}{360} \times 2\pi (6378.15 + 1000) = 3384.15\, \text{km}
\]  

(3-2)
where
\[ \theta = \frac{T}{24 \text{hours} \times 60 \text{min/ut}} \times 360^\circ = 26.28^\circ \]

\[ R_e: \text{earth radius} \]

\[ T = 105.2 \text{min} @ h = 1000 \text{km} \]

Figure 3.2 Satellite footprint for two passes

Obviously the altimeter swath width is extremely narrow when compared this distance with the distance between two continuous satellite passes for a 1000 km sun synchronise orbit. That is why a network constellation is extremely necessary for the real time global sea state monitoring. However, the arrangement of the network will depend on the altimetry map grid requirement resulting from the customers survey. The survey shows generally speaking, a meaningful global altimetry map should have a grid that is no larger than 600 km.

3.2 Network Constellation Simulation

If the satellite operates in a broadcast mode, a simple calculation would show the diameter of the satellite footprint circle will be as large as 6412 km for a 1000 km orbit satellite, shown by Figure 3.3. These characteristics enables a user to receive several altimetry broadcast results from several satellites simultaneously, if the network is arranged properly.

Figure 3.3. Broadcast antenna footprint
Notice that the satellite footprint diameter (6412 km) is larger than the distance between satellite two continuous passes (3384 km). It is, therefore, very natural to think of putting several satellites equally separated within the two continuous satellite tracks to decrease the altimetry map grid. That means for a user who is in the area within the two continues pass, he would receive the real time measurement from all these altimeters, even though some of the altimetry measurement place are taken quite far away from him.

The next question is how many satellites are required for this application? We know the worst case occurs when the ship is at the middle point between two satellites track at the equator, for this condition the distance is:

\[ d = \frac{D}{2(n-1)} \text{ km} \]  \hspace{1cm} (3-3)

where

- \( n \) the numbers of satellites.
- \( D \) the distance between two satellite continuous passes

If \( n = 6 \), then \( d \) is roughly 300 km, refer to the previous customer requirements survey, for most case it is an acceptable result. A simulation has been done, shown by Figure 3.4 the altimeter track for satellite 1 (RA1) and satellite 6 (RA2). We could see RA2 always repeat the track of RA1, this give the user who is near the track of RA1 some benefit of accessing the altimetry measurement result for a longer time (the round area is the satellite broadcast antenna footprint area).
However, to achieve real or near real time altimetry measurements for ship world wide, only six altimeters is not enough. Considering for the worst case that satellite 6 just passed a certain area (area A in Figure 3.5), the simulation shows the user would have to wait nearly half a day (12 hours) before the satellites came to close that area again.
Figure 3.5. Only six satellites condition, (a) the starting position, we could see the six satellite just leave area A (b) shows after nearly 10 hours satellite 1 return to the area A.

As the ships generally use six hours weather update, such a long waiting time is unacceptable. To decrease the users waiting time, another six satellites are then added in the constellation arrangement, in a second orbital plane, shown by Figure 3.6.

In here:

\[
\alpha/2 = 31.47 \, \text{deg} \quad \text{the half of broadcast antenna illumination angle (refer to Figure 3.3)}
\]

\[
\theta = 26.28 \, \text{deg} \quad \text{is the angle between satellite two continuous passes (refer to Figure 3.2)}
\]

Therefore we can get:
\[ \theta + \alpha / 2 + \alpha / 2 = 26.28^\circ + 54.6^\circ = 89.22^\circ \] (3-4)

The above equation presents a very interesting result. It tells us the angle between satellite No.1 and satellite No.7 is roughly 90 degree which all the satellites pass the same latitude simultaneously at different longitude, there will have no seam between the area illuminated by satellite No.6 and No.7, even near equator. Expressed another way, each group of six satellite footprints can cover nearly half of the earth, and the coverage is seamless between the two groups!

For this arrangement, when considering again the worst case – the ship at the equator, and satellites 6 just pass the ship, then the ship need only to wait another satellite period (100 minutes for 1000 km) before it appears in satellite 7's footprint area. Figure 3.7 shows within one satellite period, the footprint of the whole altimeters network could cover most area of the world. In here, in order to show clearly we omit RA2 ~ RA5 & RA8 ~ RA11, from the figure we could see in descending part, the satellite footprint covers from longitude 20 ~ -160 degree, while in the ascending part the satellite network footprint covers from longitude -150 ~ 30 degree.
The above results are very interesting, they clearly show this 12 microsatellite constellation arrangement provide a relatively narrow altimetry map grid as well as a reasonable low user waiting time.

3.3 Conclusion

In this chapter, to fulfil the customer requirements on real or nearly real-time broadcasting with narrow grid altimetry maps, a 12-satellite network constellation is proposed and simulated. The simulation demonstrates that for the proposed arrangement, the longest waiting time for any world wide user is no longer than one satellite period (around 90 minutes for 800 km sun synchronous orbit), and the largest mapping grid appears in equator, in where it is around 600 km.
Chapter 4. Signal Distortion Influence on Altimeter Significant Wave Height Estimation

The aim of this chapter is to discuss the hardware phase and amplitude distortion influence towards the on board radar altimeter significant wave height (SWH) estimation. The group delay is critical to understand and quantify the precise effects of signal amplitude and phase distortion on the resultant SWH so that enough emphases can be placed during the payload hardware design, especially the power amplifier.

Basically, the errors of transmitted and received signals introduced by the hardware design can be divided into two types - the uneven amplitude response and the non-linear phase response. Generally paired echoes are used to describe the amplitude and phase influence on radar signal analysis. However for radar altimeter application, the consideration will be different. The main concerns of this chapter’s analysis will concentrate on understanding the amplitude/phase distortion towards the return waveform’s noise floor, leading edge and the plateau edge, as they are the basic information from which to deduce SWH and wind speed.

Normally the hardware amplitude fluctuation characteristic is repeatable, thus they cannot be decreased by averaging a group of return waveforms as is usually adopted to decrease the Rayleigh fading noise. The same thing also applies to the phase distortion. Thus error models must be set up respectively to feed into transmitted signal and received signal for simulation analysis. In this chapter’s study, the error sources are considered to be amplifier and chirp signal generator which is believed are the two main possible error sources of the whole transmitter/receiver link. The chirp generator error will enter both the transmitted signal and the local deramping signal, whilst the amplifier error will only impact the transmitted signal.

The algorithm used for estimating SWH is SMLE (sub-optimal Maximum Likelihood Estimation) which was used in ERS altimeters. The simulation flow in this study is first to set up a phase and amplitude error model and generate the system impulse response based on this model, then convoluted it with the convolution product of height probability density and flat surface response to get the final return waveform. By analysing this waveform with SMLE algorithm, we can thus get the estimated SWH value. Compare it with the real SWH value, the changing percent will therefore indicate the different phase and amplitude errors influence on the on board fast estimation.

This chapter will first give an overview of the simulation processing flow, then it will be followed by setting up error models of different sources and analysing their impact on SWH estimation with SMLE algorithm. The objective is to obtain a good understanding for amplitude phase performance of the RF system, especially the power amplifier.
4.1 Simulation Model Overview

The widely used altimeter returned signal is the well established Brown model [Brown G. S. 1977], which can be represented as a triple-convolution of:

1) the point target response $P_{pr}$, due to the limited FFT length. After a receiver sidelobe weighting it will have a Gaussian shape;
2) flat surface response which includes effects of antenna pattern, off-nadir pointing angle $P_{fs}$;
3) the height probability density of the specular points $P_{pdf}$, a Gaussian shape is assumed in here;

Brown used this convolutional model with assumptions common to satellite radar altimeter systems to produce a simplified close-form expression for the radar return signal $P_r(\tau)$:

\[
P_r(\tau) = P_{pr}(\tau) * P_{fs}(\tau) * P_{pdf}(\tau) = \begin{cases} 
\eta P_{pr}(0) \sqrt{2\pi \sigma_p} \left[1 + \text{erf} \left( \frac{c\tau}{\sqrt{2\sigma_c}} \right) \right]/2, & \tau < 0 \\
\eta P_{pr}(\tau) \sqrt{2\pi \sigma_p} \left[1 + \text{erf} \left( \frac{c\tau}{\sqrt{2\sigma_c}} \right) \right]/2, & \tau \geq 0 
\end{cases} \quad (4-1)
\]

in here,

- $\eta$ is the pulse compression ratio,
- $P_T$ is the peak transmitted power,
- $\sigma_p$ is related to the point target 3 dB width ($T$) by $\sigma_p = 0.425T$, where $T$ is the radar signal transmit time
- $\sigma_c$ is determined from the $\sigma_c = \sqrt{\sigma_p^2 + (2\sigma_r/c)^2}$, where $\sigma_r$ is the rms height of the specular points relative to the mean sea level
- $\text{erf}(\cdot)$ denotes the error function,
- $\tau = t - 2H/c$ $H$ is satellite height, $c$ is the light velocity

4.1.1 Sea Surface Impulse Response

The payload hardware performance errors will influence the point target response, whilst for the altimeter average rough surface impulse response – the double convolution result of the flat surface response and probability density function, will still be the same. So the equation used in Brown model for this impulse response will still be adopted:
Chapter 4. Phase Error Influence to Significant Wave Height

\[ P_t(\tau) = P_{\phi}(\tau) * P_{\text{ref}}(\tau) = \begin{cases} P_{\phi}(0) \left[ 1 + \text{erf} \left( \frac{c\tau}{2\sqrt{2}\sigma_t} \right) \right]/2, \tau < 0 \\ P_{\phi}(\tau) \left[ 1 + \text{erf} \left( \frac{c\tau}{2\sqrt{2}\sigma_t} \right) \right]/2, \tau \geq 0 \end{cases} \quad (4-2) \]

In here the flat surface impulse response function \( P_{\phi}(t) \) is given by:

\[ P_{\phi}(t) = A_0 \exp(-\delta t) I_0(\beta^{1/2}) U(t) \quad (4-3) \]

in which

\[ \delta = \left( \frac{4}{\gamma} \right) \left( \frac{c}{H} \right) (1 + H/R_e)^{-1} \cos(2\xi) \quad (4-4) \]

and

\[ \beta = \left( \frac{4}{\gamma} \right) \left( \frac{c}{H} \right)^{1/2} (1 + H/R_e)^{1/2} \sin(2\xi) \quad (4-5) \]

where

- \( U(t) \) is the unit step function,
- \( I_0(\beta^{1/2}) \) is a modified Bessel function,
- \( R_e \) is the earth's radius,
- \( \xi \) is the absolute off-nadir pointing angle,
- \( \gamma \) is an antenna beam width parameter defined by assuming a Gaussian approximation to the antenna gain for an angle \( \theta \) off the antenna's axis

\[ G(\theta) = G_0 \exp\left[ -(2/\gamma) \sin^2 \theta \right] \quad (4-6) \]

if \( \alpha \) is the usual antenna angular full-width at half power, then

\[ 4/\gamma = (\ln 4)/\sin^2(\alpha/2) \quad (4-7) \]

the amplitude term \( A_0 \) in equation (4-3) includes several other constants:
\[ A_0 = \left[ G_0 \lambda_r^2 c \sigma^0(0) \right] / \left[ 4(4\pi)^2 L_p H^3 \left( 1 + H/R_e \right)^3 \right] \exp \left[ - \left( \frac{4\pi}{\gamma} \right) \sin^2 \xi \right] \]  \quad (4-8)

where

\( \lambda_r \) is the radar transmitted signal wavelength;
\( \sigma^0(0) \) is ocean’s radar backscattering cross-section at normal incidence;
\( G_0 \) is the radar antenna boresight gain,
\( L_p \) is the two way propagation path loss.

If we set the off-nadir pointing angle \( \xi = 0 \), the normalised average rough surface impulse response result will be shown by Figure 4.1 (This simulation assumes the satellite height is 1000 km, and the antenna is one meter diameter dish therefore the antenna half power beam width angle is 0.75 degree.)

![Figure 4.1 Noise free sea surface impulse response](image)

### 4.12 Point Target Response (PTR)

PTR is defined as the effective transmitted pulse shape observed by the receiver, so it is a FFT result of a time product of transmitted signal and the local dechirping signal. Consider the limited FFT length and after a receiver sidelobe weighting, it will have a quasi-Gaussian shape.

The transmitted chirp signal is expressed as:
Chapter 4. Phase Error Influence to Significant Wave Height

\[ s_r(t) = \exp\left(-j[2\pi ft + \mu t^2]\right) \quad (4-9) \]

where \( \mu = 2\pi B/T \)

\( B \) is the chirp signal bandwidth
\( T \) is the chirp signal transmit time

Let this transmitted signal be deramped by a local oscillator signal, misaligned by time \( \tau \) (\( \ll T \)) with respect to target echo, and with amplitude taper function \( A'(t) \):

\[ s_{local}(t) = A'(t)\exp\left(j[2\pi f(t - \tau) + \mu(t - \tau)^2]\right) \quad (4-10) \]

Therefore the deramped signal \( s_r(t) \) becomes:

\[ s_r(t) = s_{local}(t) \cdot s_r(t) = A'(t)\exp\left(j[-2\pi f\tau + \mu(\tau^2 - 2t\tau)]\right) \quad (4-11) \]

The amplitude taper function may have different choices where here we use Hamming weighting function:

\[ W_H(t) = 0.08 + 0.92 \cdot \cos^2 \frac{\pi(t - \tau/2)}{T} \quad (4-12) \]

Then the whole output signal becomes:

\[ s(t) = s_r(t) \cdot W_H(t) = A'(t) \cdot [0.08 + 0.92 \cdot \cos^2\left(\pi(t - 0.5T)/T\right)] \cdot \exp\left(j[-2\pi f\tau + \mu(\tau^2 - 2t\tau)]\right) \quad (4-13) \]

After FFT the signal of equation (4-13), the expected Gaussian shape point target response may be obtained. Figure 4.2(a) shows the graph of the point target response without weighting, Figure 4.2(b) is a comparison of point target response with and without weighting. We can see clearly that after the weighting, the first side-lobes pairs of PTR response are decreased to -40 dB lower than the main lobe.
Due to the chirp signal characterise there is a relationship between radar signal frequency and transmit time: $\Delta f = Q\Delta T$ ($Q = B/T = 320\text{MHz}/102.4\mu s$ in this altimeter's application), therefore correspondingly we can change the x-axis of Figure 4.2 (b) back to time domain to carry on the required convolution in the Brown model to produce the final altimeter return waveform expression.
there are two most important sub-systems that can seriously influence the system performance – the chirp generator and the power amplifier. And their influence on the radar signal PTR is different. The errors coming from the chirp signal generator may appear both in the received signal and the local deramping signal, while the errors coming from the power amplifier may only appear in the received signal. The general radar analysis uses paired echoes to describe the amplitude and phase influence, but in the radar altimeter, the concern will be different. The study in this chapter will concentrate on understanding the amplitude/phase distortion towards return waveform shape in terms of the influence on noise floor, leading edge and the plateau edges. We will give the detailed analysis in the later sections.

4.13 Rayleigh Noise

The above Brown model for altimeter return signal expressed by equation (4-1) only exists in the ideal condition or only applies to the average power of large numbers of return signals. In the real world, within the altimeter footprint on the sea surface there will always be many wave facets specularly reflecting the incident signal at a given range, therefore each return signal exhibits random fluctuation. The returned signal from a single wave facet can be expressed as an amplitude and a phase, or equivalently, in terms of real and imaginary components. Since the radar wavelength is short, the phase relationships between signals reflected from the various facets are random. The total signal received by the altimeter is the vector addition of the real and imaginary components from all individual specularly oriented wave facets in the altimeter footprint. By the Central Limit Theorem, the real and imaginary components of the total signal are both approximately Gaussian distributed random variables. The amplitude of the total returned signal is therefore Rayleigh distributed and the returned power at each frequency is an exponentially distributed random variable. Many individual waveforms must be averaged to obtain a mean waveform with the smooth shape, if each waveform is statistically independent (which depends on the pulse repetition rate and the satellite orbital velocity). The Rayleigh noise decreases as the square root of the number of waveforms in the average. The Seasat altimeter averaged a group of 50 waveforms, whilst on TOPEX the average number increased to 228 single returned waveform. Figure 4.4
shows the simulation procedure for generating the sea surface model when considering the Rayleigh noise.

\[ g(t) = \sqrt{u(t)} \]
\[ u(t): \text{altimeter average rough surface impulse response} \]

Figure 4.4 Simulation model for sea surface impulse response with Rayleigh noise

The Steps for setting up the simulation model with Rayleigh noise are:
1) Identify the convolution of the flat surface response and wave height distribution \( u(t) \), compute the square root of this, call \( g(t) \) in this model
2) Generate a file of \( N \) uncorrelated white noise samples of uniform variance, these noise samples are complex, should consist of two components real and imaginary which are each samples of a Gaussian random variable
3) Multiply each sample by gain \( g_k \) where these are samples of \( g(t) \) as specified above
4) \( x_k = g_k n_k \) and generate an input scaled white noise file of \( x_1 \) to \( x_N \),

Figure 4.5 shows the return waveform with the statistical noise and after a 50-group average.
4.14 Off-nadir Pointing Influence

Another important thing need to be considered in the return signal waveform simulation model is the satellite off-nadir pointing influence, which has been introduced briefly in section 2.422. Basically, due to the existence of off-nadir angle, the flat surface impulse in the Brown model has to be modified as [Brown G.S. 1977]:

$$ P_n (\tau) = \frac{G_0^2 \lambda_c \sigma_n^2 (0)}{4(4\pi)^2 L_n h^3} \cdot \exp \left[ -\frac{4}{\gamma} \sin^2 \xi - \frac{c \tau}{H} (\cos 2\xi + \alpha) \right] \cdot I_0 \left( \frac{4}{\gamma} \sqrt{\frac{c \tau}{h}} \sin 2\xi \right) $$ (4-14)
Chapter 4. Phase Error Influence to Significant Wave Height

The main effect of the off-nadir pointing in the final return signal waveform is to decrease the received signal amplitude as well as to change the slope of the signal plateau edge. To compensate this influence and get correct estimation, the usual way in signal processing is to apply a range window $W_R(\tau, \varepsilon)$ upon the received return waveform to correct the trailing edge of the real received signal before the real estimation for SWH and backscatter coefficient, shown by equation (4-15), in here $\varepsilon$ is the estimated mispointing angle:

$$W_R(\tau, \varepsilon) = \exp \left( -\frac{4c\varepsilon}{h\gamma} \cos 2\varepsilon \right) \frac{1}{I_0\left(\frac{4\gamma}{c\tau h} \sin 2\varepsilon\right)}$$

(4-15)

the actual signal therefore becomes:

$$R(\tau, \xi, \varepsilon) = P_\tau(\tau, \xi) \cdot W_R(\tau, \varepsilon)$$

(4-16)

Figure 4.6 shows a comparison of the return signal waveform with and without error, and the shape after weighting.

![Simulated altimeter received signal with and without nadir pointing weighting compensation](image)

Figure 4.6 Simulated altimeter received signal with and without nadir pointing weighting compensation

- upper line – distorted received signal after weighting ($\xi=\varepsilon$)
- middle line – distorted received signal without weighting
- lower line – received signal without error

Because the on board estimated off pointing angle may be different from the real off nadir angle, reflect in the waveform after weighting, the trailing edge will have different shape of slope, shown in Figure 4.7.
Chapter 4. Phase Error Influence to Significant Wave Height

Figure 4.7 Different trailing edge slope at different $\xi$, $\varepsilon$ condition (128 gates waveform is chosen)

---

Shown by Figure 4-7, the result of equation (4-16) will provide an error signal by judging the slope value of the trailing edge of the weighted return waveform. A typical second order $\alpha$-$\beta$ filter is generally adopted to drive this error to zero so that can get an updated estimation of the current satellite pointing condition. In the previous altimeter missions such as TOPEX, this compensation weighting was processed on ground. However for the fast delivery altimetry results, it is required that this processing be carried out on board, and that means this weighting must be applied before estimating the slope, AGC and height errors.

In the microsatellite GANDER project, a 0.7 meter parabolic antenna is proposed, so the antenna half power beam width will be slightly larger than the former missions which means the requirement on satellite attitude control will not be as restrictive as those previous altimeters. However, the accuracy required by this on board weighting to satisfy the customer's requirement is still an open question, which needs further careful evaluation.

For the simulation in this chapter which emphasises the influence of hardware performance, we will omit this processing by presuming the satellite has no off pointing error.

**4.15 SWH Estimation Algorithm**

There are several different methods in estimating SWH. The SMLE method we will adopt in here was developed by University College London (UCL) for the ERS-1, 2 altimeters. This algorithm derived depends on the following assumptions:

1. The form of the radar return is given by the Brown model
2. The SMLE tracker correctly positions the return so that gate 32 corresponds to the mean level.

3. The SMLE tracker applies a correction to the return waveform to remove the trailing edge decay due to the antenna beam pattern and the mis-pointing.

The most critical part in SWH estimation is, when given the altimeter return waveform, how to deduce $\sigma_e$ in equation (4-1), and then further use $\sigma_e$ to deduce the \textit{rms} height of the specular points relative to the mean sea level – SWH. Recall $P_r(\tau)$ in equation (4-1), it can be normalised as equation (4-17):

$$P_m(t) = 0.5 \left[ 1 + \text{erf} \left( \frac{t}{\sqrt{2}\sigma_e} \right) \right]$$  \hspace{1cm} (4-17)

$P_m(t)$ diagram can be shown by Figure 4.8.

![Diagram of area equalisation procedure](image)

Figure 4.8 Schematic illustration of the area equalisation procedure used by the SMLE to estimate the half-width of the altimeter return (after Wingham D. J., et al. 1986) (*original quality is poor)

It is known the higher the SWH, the longer time the return waveform requires to reach its peak power. Therefore the waveform slope is an important parameter in estimating SWH. Refer to Figure 4.8, it is reflected by the value of $\tau'$. The SMLE obtains a measure of the half width of the return $\tau'$ by equalising the shaded areas shown in the schematic representation of the return in Figure 4.8.

We may represent this procedure analytically by equating the area under $P_m(t)$ between A and B with the area of ABCO, thus,

$$\int_0^{\tau'} P_m(t) \, dt = 3\tau'/4$$ \hspace{1cm} (4-18)
where $T'$ measured in units of time, or in range gates. Alternatively, because of the asymmetry of $\text{erf}(\cdot)$, the area equalisation may be represented by

$$
\int_{-T'}^{T'} P_m(t) dt = \int_{-\frac{T'}{2}}^{\frac{T'}{2}} P_m(t) + \frac{T'}{2}
$$

(4-19)

combine the above equations, we may deduce that

$$
\int_{0}^{T'} \text{erf}\left(\frac{t}{\sqrt{2}\sigma_c}\right) dt = \frac{T'}{2}
$$

(4-20)

or

$$
\int_{0}^{\infty} \text{erf}(x) dx = \frac{y}{2}
$$

(4-21)

where

$$
y = \frac{T'}{\sqrt{2}\sigma_c}
$$

(4-22)

as the $\text{erf}(\cdot)$ is a standard function, equation (4-9) may be solved numerically in a straightforward manner to obtain the solution

$$
y = 1.0397
$$

(4-23)

combine the above result in equation (4-23) with $\sigma_c = \sqrt{\sigma_p^2 + (2\sigma_s/c)^2}$ and $\text{SWH} = 4\sigma_s$, we can then get

$$
\text{SWH} = 2c\sqrt{\left(0.6801T'\right)^2 - \sigma_p^2}
$$

(4-24)

Thus given the half width $T'$ from the SMLE tracker we may obtain the significant wave height $\text{SWH}$ from the above equation (4-24). In the equation $T'$ is given by the half width in range gates multiplied by the range bin width.
With the knowledge of the point target response model, the sea impulse response with / without Rayleigh noise, and the SWH estimation algorithm, the simulation flow chart for the study of evaluating hardware performance on SWH estimation can therefore be shown with the following structure:

![Simulation flow chart](image)

To simplify the simulation, we will manually choose the gate index for different SWH value instead of using an auto control loop.
4.2 Phase Distortion towards SWH Estimation

The hardware imperfect amplitude and phase response may distort all the wanted return signal waveform, not only the return waveform's leading edge but also its plateau edge. These lead to the SWH estimation error. In this section, the study mainly concentrates on phase distortion influence. The simulation analysis is based on setting the error models for power amplifier and chirp signal generator respectively. The recommended hardware performance requirement will be supplied at the end of analysis.

All the analysis in the below sections are using time domain phase / group delay distortion models, which are easier to understand for the hardware design people. This analysis method is based on stationary phase theorem, which is only valid when the chirp signal time-bandwidth product is very large.

To a chirp signal $u(t)$, after feed into the system, the output signal $v(t)$ can be expressed:

$$u(t) = e^{-j\mu^2} \quad \Rightarrow \quad v(t) = \sqrt{\frac{\pi}{j\mu}} e^{\frac{j\omega^2}{4\mu}} e^{j\phi(\omega)} e^{j\omega t} dt$$

According to the stationary phase principle, we can get that

$$\frac{2\omega}{4\mu} + \frac{d\phi(\omega)}{d\omega} + t = 0 \quad (4-25)$$

if we ignore $d\phi/d\omega$, then $\omega = -2\mu$. Therefore:

$$\frac{\omega^2}{4\mu} + \omega t = \frac{(-2\mu)^2}{4\mu} - 2\mu^2 = -\mu^2 \quad (4-26)$$

so

$$v(t) \propto e^{j\phi(-2\mu)} \cdot e^{-j\mu^2} \quad (4-27)$$

Equation (4-27) shows that any hardware phase distortion error can influence the signal by adding an extra time domain phase error part in the output signal.
4.21 Amplifier Phase Distortion Analysis

Ideally, the amplifier response in the time domain is to magnify the input signal amplitude and to delay the input signal by certain amount of time. This delay time is decided by the amplifier's group delay.

![Ideal amplifier response](image)

Figure 4.10 Ideal amplifier response

Group delay is defined as \( \tau = \Delta \phi / 2\pi f \), if the group delay of an amplifier is a constant value, \( \tau_0 \), then the amplifier output signal \( s(t) \) can be expressed as:

\[
S(t) = A \cdot h(t - \tau_0)
\]  

(4-28)

Due to the imperfect amplifier response, the amplifier group delay response usually is not a constant value within the whole bandwidth. For example, when measure an amplifier's group delay by a vector network analyser, it quite often displays a frequency dependent variation. This is because the amplifier's phase response is often frequency dependent, shown in Figure 4.11.

![Group delay variation](image)

Figure 4.11 Group delay variation under imperfect phase response of power amplifier

Bear in mind that the frequency of the transmitted altimeter transmits chirp signal \( f \) has a linear relationship with transmit time \( t \), expressed as:
Therefore, for the chirp signal, the amplifier group delay response \( \tau(f) \) can also be viewed as a time function \( \tau(t) \).

To study the influence of group delay distortion on altimeter’s on board SWH estimation, an accurate model must be set up. In this study, two simple but representative group delay distortion models have been used:

- assume \( \tau(t) \) is a sinewave function of time \( t \), \( \tau = \tau_0 + \tau_n \sum_n \cos(2\pi f_{GDn} t) \) (\( f_{GDn} \) is the Fourier frequencies of the hardware group delay distortion)
- assume \( \tau(t) \) is a linear function of time \( t \), \( \tau = \tau_0 + \tau_1 \cdot t \)

4.211 \( \tau(t) \) is sinewave function

It is known that for any time limited real function, it always can be expressed as the sum of a series of Fourier functions:

\[
\tau(t) = \tau_0 + \tau_1 \cos(2\pi f_{GD1} t) + \tau_2 \cos(2\pi f_{GD2} t) + \tau_3 \cos(2\pi f_{GD3} t) + \ldots -T/2 < t < T/2
\]

(4-30)

where \( \tau_0 \) is amplifier’s mean group delay; \( \tau_n \) is the peak value of the group delay variation at each \( f_{GDn} \) frequency.

If, for simplicity, one assumes there is only one dominant frequency component, then

\[
\tau(t) = \tau_0 + \tau_1 \cos(2\pi f_{GD1} t) \quad -T/2 < t < T/2
\]

(4-31)

Under this condition, the amplifier’s group delay will display a waveform shown in Figure 4.12.

![Figure 4.12 Group delay waveform for sinewave function model](image)
If the input altimeter chirp signal is \( h(t) = \exp[-j(2\pi ft + \mu t^2)] \), \( \mu = 2\pi Q \), then the amplifier output signal \( s_A(t) \) becomes
\[
s_A(t) = A \exp[-j(2\pi f(t - \tau_0 - t_1 \cos(2\pi f_{GD} t)) + \mu(t - \tau_0 - t_1 \cos(2\pi f_{GD} t))]  \]
\[-T/2 < t < T/2 \] (4-32)\]

The radar receiver signal is a delayed function of \( s_A(t) \), the delay time \( \tau_{\text{range}} \) is decided by the nadir range between satellite and sea mean surface:
\[
s_r(t) = A' (t - \tau_{\text{range}}) \cdot s_A(t - \tau_{\text{range}})
= A' (t - \tau_{\text{range}}) \cdot \exp[-j(2\pi f(t - \tau_0 - \tau_{\text{range}} - t_1 \cos(2\pi f_{GD} t)) + \mu(t - \tau_0 - \tau_{\text{range}} - t_1 \cos(2\pi f_{GD} t))]  \]
\[-T/2 < t < T/2 \] (4-33)\]

If one assumes the radar returned signal is tracked properly, and the group delay of the whole transmit link is well calibrated, the local oscillator-dechirping signal is:
\[
s_{\text{local}}(t) = B \exp\{j[2\pi f(t - \tau_0 - \tau_{\text{range}}) + \mu(t - \tau_0 - \tau_{\text{range}})]\}
\[-T/2 < t < T/2 \] (4-34)\]

The resulting point target response signal then becomes:
\[
S(t) = s_r(t) \cdot s_{\text{local}}(t)
= A' (t - \tau_{\text{range}}) \cdot B \cdot \exp\{j[2\pi f t_1 \cos(2\pi f_{GD} t) + \mu t_1 \cos(2\pi f_{GD} t) \cdot [2t - 2\tau_0 - 2\tau_{\text{range}} - t_1 \cos(2\pi f_{GD} t)]]\}
= A' (t - \tau_{\text{range}}) \cdot B \cdot \exp\{j[2\mu t_1 \cos(2\pi f_{GD} t) \cdot t + \cos(2\pi f_{GD} t) \cdot (2\pi f t_1 - 2\mu t_1 \tau_0 - 2\mu t_1 \tau_{\text{range}}) - \mu t_1^2 \cdot \cos^2(2\pi f_{GD} t)]\}
\[-T/2 < t < T/2 \] (4-35)\]

In the above equation, \( \exp[j \cdot 2\mu t_1 \cos(2\pi f_{GD} t) \cdot t] \) will decide the frequency of the point target response’s main lobe. If the value of \( 2\mu t_1 \cos(2\pi f_{GD} t) \) is nearly constant, the main lobe will be similar to the usual main lobe that without error. While if the variation of \( 2\mu t_1 \cos(2\pi f_{GD} t) \) is considerably large, this component will not only broaden the main lobe but also modulate the main lobe’s amplitude. The other two parts in equation (4-35), \( \exp[j \cdot \cos(2\pi f_{GD} t) \cdot (2\pi f t_1 - 2\mu t_1 \tau_0 - 2\mu t_1 \tau_{\text{range}})] \) and \( \exp[-j \cdot \mu t_1^2 \cos^2(2\pi f_{GD} t)] \), on the other hand, will have very little impact on the target response if the value of \( \cos(2\pi f_{GD} t) \) is nearly constant within the chirp signal time \( T \).
Taking the Fourier Transform of this function will yield the point target response spectrum in the frequency domain, expressed as:

\[ S(\omega) = F[A^r] = \int_{-T/2}^{T/2} S(t) \exp(-j\omega t) dt \]  

(4-36)

Due to the chirp signal character, \( f = Q^*t, AF \) of \( \omega \)-axis in \( S(\omega) \) plot always corresponds to a certain time amount \( \Delta t \). Thus, \( S(\omega) \) can also be viewed as a time function, \( PTR(t) \). Refer to equation (4-35), obviously the shape of this point target response will change with different \( f_{GD} \) values. The following three simulated diagrams in Figure 4.13 show the different point target responses at three different \( f_{GD} = 0, 10 \) kHz, \( 100 \) kHz conditions. The response amplitude is normalised to the maximum value, and plotted in \( dB \). All the parameters in the simulation, such as chirp signal time & bandwidth, are set according to the ERS altimeter.

(a) Point target response at \( f_{GD} = 1 \) kHz & \( f_{GD} = 0 \)
In all the above three diagrams, the x-axis represents FFT points, and the $\Delta f$ value between these FFT points are the same. It can be clearly seen that when $f_{GD} = 1$ kHz the main lobe width is around $40*\Delta f$, while if $f_{GD} = 1$kHz the main lobe width will increase to roughly $200*\Delta f$. This width will become enormously high when $f_{GD}$ is even higher, around $3000*\Delta f$ when $f_{GD} = 100$ kHz.
Chapter 4. Phase Error Influence to Significant Wave Height

The simulation results indicate that the existence of sine wave function type of group delay distortion will broaden the main lobe of the point target response. Once the distortion period is of the same order or smaller than the chirp signal length of 100\mu s, the main lobe of the point target response will become extremely bad with high frequency spurious components appearing.

It must be pointed out here, that the asymmetric PTR response in Figure 4.13 (b) and (c) is due to whether we choose sine or cosine wave model in expressing group delay distortion. Figure 4.14 show the two different model response when \( f_{gd} = 10 \) kHz.

![Figure 4.14 PRT response with different group delay models (in here two curves represent \( \tau_0 = 10\text{ns} \), \( \tau_1 = 2\text{ns} \), \( 4\text{ns} \) respectively)](image)

(a) \( \tau = \tau_0 + \tau_1 \cos(2\pi f_{gd} t) \)  
(b) \( \tau = \tau_0 + \tau_1 \sin(2\pi f_{gd} t) \)

At the same time, if \( f_{gd} \) is fixed to a certain value while changing \( \tau_1 \) in equation (4-31), \( \tau(t) \equiv \tau_0 + \tau_1 \cdot \cos(2\pi f_{gd} t) \), the simulation results show that the higher the \( \tau_1 \), the wider the main lobe of point target response becomes, and consequently the worse the impact on SWH estimation. Figure 4.15 shows a group of simulation results with \( f_{gd} = 1 \) kHz, 5 kHz and \( \tau_1 \) varies from 1 ns to 4 ns.
Chapter 4. Phase Error Influence to Significant Wave Height

When the point target response convolutes with sea surface impulse response, the final radar return waveform will be obtained. This waveform, when fed into the altimeter SWH estimation algorithm SMLE, the SWH value can be calculated.

Table 4-1 is a group of simulation results of SWH estimation changing percent under different $\tau_1$ and SWH conditions when $f_{GD} = 1$ kHz. It clearly shows that the group delay distortion influence is more serious in low SWH condition than that in higher SWH condition. For example, when $\tau = 9$ ns, the SWH estimation changing percent is nearly 30% at SWH = 1 m condition, while it drops down to only 4% when SWH = 5 m. Figure 4.16 is a 3-D view of this SWH estimation changing percent.
Table 4-1 SWH estimation error at different $\tau_1$ and SWH conditions ($f_{GD} = 1 \text{ kHz}$)

<table>
<thead>
<tr>
<th>$\tau_1$ (ns)</th>
<th>SWH = 1 m</th>
<th>SWH = 1.5 m</th>
<th>SWH = 2 m</th>
<th>SWH = 2.5 m</th>
<th>SWH = 3 m</th>
<th>SWH = 3.5 m</th>
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</tr>
</tbody>
</table>

Figure 4.16 3-D view of SWH estimation changing percent vs. SWH and $\tau_1$ ($f_{GD} = 1 \text{ kHz}$)

$S1$ to $S9$: SWH = 1 m to SWH = 5 m step 0.5 m

It must be pointed out that when $f_{GD}$ is set to 1 kHz, during the whole chirp signal sweep time window (100 µs) the group delay value $\tau$ only changes from $\tau = \frac{4}{f_{GD}}$ to $\tau = \frac{4}{f_{GD} + 0.1}$, which is not a big change. While in the real application, the distortion period may be much smaller than the sweeping time window of chirp signal, in other words, $f_{GD}$ is larger than $f = \frac{1}{100 \mu s} = 10 \text{ kHz}$. Figure 4.17 shows an example, a measured S band amplifier group delay performance.
Thus another group of simulation therefore must be run with higher $f_{GD}$ value ($f_{GD} = 10$ kHz). The results are shown in Table 4-2. The 3-D view results are shown in Figure 4.18.

<table>
<thead>
<tr>
<th>$\tau_1$ (ns)</th>
<th>SWH = 1 m</th>
<th>SWH = 1.5 m</th>
<th>SWH = 2 m</th>
<th>SWH = 2.5 m</th>
<th>SWH = 3 m</th>
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<td>58%</td>
<td>53%</td>
<td>48%</td>
<td>33%</td>
<td>31%</td>
</tr>
<tr>
<td>2.50</td>
<td>396%</td>
<td>159%</td>
<td>142%</td>
<td>123%</td>
<td>76%</td>
<td>72%</td>
<td>64%</td>
<td>46%</td>
<td>43%</td>
</tr>
<tr>
<td>3.00</td>
<td>495%</td>
<td>206%</td>
<td>180%</td>
<td>155%</td>
<td>94%</td>
<td>89%</td>
<td>80%</td>
<td>58%</td>
<td>55%</td>
</tr>
<tr>
<td>3.50</td>
<td>690%</td>
<td>246%</td>
<td>214%</td>
<td>191%</td>
<td>114%</td>
<td>104%</td>
<td>96%</td>
<td>70%</td>
<td>65%</td>
</tr>
<tr>
<td>4.00</td>
<td>861%</td>
<td>291%</td>
<td>264%</td>
<td>222%</td>
<td>131%</td>
<td>122%</td>
<td>113%</td>
<td>81%</td>
<td>76%</td>
</tr>
<tr>
<td>4.50</td>
<td>1081%</td>
<td>360%</td>
<td>305%</td>
<td>254%</td>
<td>148%</td>
<td>138%</td>
<td>127%</td>
<td>91%</td>
<td>88%</td>
</tr>
</tbody>
</table>

Table 4-2. SWH estimation error at different $\tau_1$ and SWH conditions ($f_{GD} = 10$ kHz)
Chapter 4. Phase Error Influence to Significant Wave Height

Figure 4.18 3-D view of SWH estimation changing percent vs. SWH and \( \tau_1 \)

\[ S1 \text{ to } S9: \text{SWH = 1 m to SWH=5 m step 0.5 m} \]

The results in Table 4-2 show, under very low SWH condition (SWH = 1 m), when \( \tau_1 = 4.5 \) ns, the estimated SWH value will change for over 1000% which is an extremely large error. Compared the results in Table 4-1 and 4-2, it is obvious to believe that the SWH estimation distortion will become worse when \( f_{GD} \) increases to higher value.

The usual acceptable results of radar altimeter SWH measurement are ±0.5 m or ±10% whichever is higher. Therefore for SWH lower than 5 m condition, an error no larger than 0.5 m is an acceptable value. Thus, for SWH = 1 m condition, the SWH changing percent should be controlled under 50%. Refer to Table 4-2, that means \( \tau_1 \) should be well controlled under 0.5 ns.

Another questions need to be answered is whether the absolute value of \( \tau_0 \) will have any serious impact on SWH estimation. Table 4-3 is a group of simulation results to address this question. In the simulation, all the parameters are the same except \( \tau_0 \) is set to 10 ns and 60 ns respectively.

<table>
<thead>
<tr>
<th>( \tau_1 ) (ns)</th>
<th>SWH = 1 m</th>
<th>SWH = 4 m</th>
<th>SWH = 8 m</th>
<th>SWH = 12 m</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>144%</td>
<td>15%</td>
<td>3.2%</td>
<td>1.3%</td>
</tr>
<tr>
<td>2</td>
<td>194%</td>
<td>27%</td>
<td>6.4%</td>
<td>2.6%</td>
</tr>
<tr>
<td>3</td>
<td>207%</td>
<td>29%</td>
<td>7.2%</td>
<td>2.9%</td>
</tr>
<tr>
<td>4</td>
<td>207%</td>
<td>30%</td>
<td>7.2%</td>
<td>3.1%</td>
</tr>
<tr>
<td>5</td>
<td>208%</td>
<td>30%</td>
<td>7.3%</td>
<td>2.9%</td>
</tr>
<tr>
<td>6</td>
<td>219%</td>
<td>32%</td>
<td>7.7%</td>
<td>3.1%</td>
</tr>
<tr>
<td>7</td>
<td>233%</td>
<td>34%</td>
<td>8.2%</td>
<td>3.3%</td>
</tr>
<tr>
<td>8</td>
<td>243%</td>
<td>36%</td>
<td>8.8%</td>
<td>3.7%</td>
</tr>
<tr>
<td>9</td>
<td>253%</td>
<td>37%</td>
<td>9.3%</td>
<td>3.9%</td>
</tr>
</tbody>
</table>

(a) \( f_{GD} = 10 \text{ kHz} \) \( \tau_0 = 60 \) ns
Chapter 4. Phase Error Influence to Significant Wave Height

<table>
<thead>
<tr>
<th>$\tau_1$ (ns)</th>
<th>SWH Estimation Changing Percent</th>
<th>SWH = 1 m</th>
<th>SWH = 4 m</th>
<th>SWH = 8 m</th>
<th>SWH = 12 m</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td></td>
<td>144%</td>
<td>15%</td>
<td>3.2%</td>
<td>1.3%</td>
</tr>
<tr>
<td>2</td>
<td></td>
<td>195%</td>
<td>27%</td>
<td>6.4%</td>
<td>2.6%</td>
</tr>
<tr>
<td>3</td>
<td></td>
<td>207%</td>
<td>29%</td>
<td>7.0%</td>
<td>3.0%</td>
</tr>
<tr>
<td>4</td>
<td></td>
<td>207%</td>
<td>30%</td>
<td>7.2%</td>
<td>2.8%</td>
</tr>
<tr>
<td>5</td>
<td></td>
<td>209%</td>
<td>30%</td>
<td>7.4%</td>
<td>3.0%</td>
</tr>
<tr>
<td>6</td>
<td></td>
<td>220%</td>
<td>32%</td>
<td>7.7%</td>
<td>3.1%</td>
</tr>
<tr>
<td>7</td>
<td></td>
<td>228%</td>
<td>34%</td>
<td>8.3%</td>
<td>3.3%</td>
</tr>
<tr>
<td>8</td>
<td></td>
<td>243%</td>
<td>36%</td>
<td>8.8%</td>
<td>3.8%</td>
</tr>
<tr>
<td>9</td>
<td></td>
<td>253%</td>
<td>37%</td>
<td>9.3%</td>
<td>3.7%</td>
</tr>
</tbody>
</table>

(b) $f_{GD} = 10$ kHz $\tau_0 = 10$ ns

Table 4-3 SWH estimation changing percent at different $\tau_0$ values.

The simulation results in Table 4-3 (a) and (b) show that the percentage change of SWH estimation is roughly the same at different $\tau_0$ values. The reason lies in the fact that it is the first term in equation (4-32), $\exp\left[\frac{j \cdot 2 \mu_{p} \tau_{1} \cos(2\pi f_{GD} t) \cdot t}{T_{o}}\right]$, that decides the main lobe of the point target response, while that term does not contain $\tau_0$, presuming the altimeter is well tracked.

Finally, recalling the simulation results in Figure 4.14, the last question that needs to be studied is whether choosing a sine or cosine function in establishing the group delay distortion model will have any serious effect in SWH estimation results. Table 4-4 shows the simulation results. We can see the changing percentage of the final SWH estimation result does not have significant difference.

<table>
<thead>
<tr>
<th>Changing percent (%)</th>
<th>$\tau_1 = 1$ ns</th>
<th>$\tau_1 = 4$ ns</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\tau = \tau_0 + \tau_1 \cos(2\pi f_{GD} t)$</td>
<td>1.1%</td>
<td>15.7%</td>
</tr>
<tr>
<td>$\tau = \tau_0 + \tau_1 \sin(2\pi f_{GD} t)$</td>
<td>1.2%</td>
<td>15.9%</td>
</tr>
</tbody>
</table>

Table 4-4 Simulation results for different group delay model ($f_{GD} = 5$ kHz, $\tau_0 = 10$ ns)

4.212 $\tau(t)$ is linear function of time $t$

If the amplifier group delay has a linear distortion response as: $\tau = \tau_0 + \tau_1 \cdot t$, 
Then the amplifier output signal in equation (4-25) becomes:

\[ s_r(t) = A \cdot \exp \left[ -j \left( 2\pi f \left( t - \tau_o - \tau_1 \cdot t \right) + \mu \left( t - \tau_o - \tau_1 \cdot t \right)^2 \right) \right] \quad (4-37) \]

the received radar signal is thus:

\[ s_r(t) = A' \left( t - \tau_{\text{range}} \right) \cdot \exp \left[ -j \left( 2\pi f \left( t - \tau_o - \tau_{\text{range}} - \tau_1 \cdot t \right) + \mu \left( t - \tau_o - \tau_{\text{range}} - \tau_1 \cdot t \right)^2 \right) \right] \quad (4-38) \]

If it is still assumed that the radar signal is tracked properly, the local deramping signal is therefore:

\[ s_{\text{local}}(t) = B \exp \left[ j \left( 2\pi f \left( t - \tau_o - \tau_{\text{range}} \right) + \mu \left( t - \tau_o - \tau_{\text{range}} \right)^2 \right) \right] \quad (4-39) \]

The deramping signal is:

\[ S(t) = A' \left( t - \tau_{\text{range}} \right) \cdot B \cdot \exp \left[ j \left( 2\pi f \tau_1 \cdot t + \mu \tau_1 \cdot t \cdot \left( 2t - 2\tau_o - 2\tau_{\text{range}} - \tau_1 \cdot t \right) \right) \right] \]

\[ = A' \left( t - \tau_{\text{range}} \right) \cdot B \cdot \exp \left[ j \left( 2\pi f \tau_1 \cdot t + \left( -2\mu \tau_1 \tau_o - 2\mu \tau_1 \tau_{\text{range}} \right) \cdot t + \left( 2\mu \tau_1 - \mu \tau_1^2 \right) t^2 \right) \right] \quad (4-40) \]

Generally speaking, \( \mu \tau_1 \left( 2 - \tau_1 \right) \) is a very small value, therefore the point target response is still approximately a sinc function with its centre frequency determined by the function...
of \( \exp\left[j(2\pi \tau_1 \cdot r)\right] \) in equation (4-40). Figure 4.20 shows the simulated point target response at different \( \tau_1 \) values.

![Figure 4.20 Point target response under linear function group delay influence](image)

Note in here, \( \tau_1 \) represents the changing slope of amplifier group delay \( \tau \) which has a different meaning than that is used in the triangle function model in last section.

Table 4-5 and Figure 4.21 show the simulated SWH estimation error vs. SWH and \( \tau_1 \). It can be seen that to control the SWH estimation changing percent lower than 50% at SWH = 1 m condition, \( \tau_1 \) must be smaller than 30\( \mu \).

<table>
<thead>
<tr>
<th>( \tau_1 (\mu) )</th>
<th>SWH = 1 m</th>
<th>SWH = 1.5 m</th>
<th>SWH = 2 m</th>
<th>SWH = 2.5 m</th>
<th>SWH = 3 m</th>
<th>SWH = 3.5 m</th>
<th>SWH = 4 m</th>
<th>SWH = 4.5 m</th>
<th>SWH = 5 m</th>
</tr>
</thead>
<tbody>
<tr>
<td>10</td>
<td>5.2%</td>
<td>1.3%</td>
<td>1.2%</td>
<td>0.5%</td>
<td>0.2%</td>
<td>0.5%</td>
<td>0.4%</td>
<td>0.02%</td>
<td>0.3%</td>
</tr>
<tr>
<td>20</td>
<td>17%</td>
<td>4.1%</td>
<td>3.5%</td>
<td>2.8%</td>
<td>1.3%</td>
<td>1%</td>
<td>1%</td>
<td>0.7%</td>
<td>0.6%</td>
</tr>
<tr>
<td>30</td>
<td>38%</td>
<td>9.5%</td>
<td>8.0%</td>
<td>6.0%</td>
<td>2.8%</td>
<td>2.8%</td>
<td>2.5%</td>
<td>1.3%</td>
<td>1.5%</td>
</tr>
<tr>
<td>40</td>
<td>64%</td>
<td>16%</td>
<td>14%</td>
<td>11%</td>
<td>5.3%</td>
<td>4.8%</td>
<td>4.2%</td>
<td>2.6%</td>
<td>2.4%</td>
</tr>
<tr>
<td>50</td>
<td>94%</td>
<td>26%</td>
<td>21%</td>
<td>17%</td>
<td>8%</td>
<td>7.6%</td>
<td>6.4%</td>
<td>3.9%</td>
<td>3.9%</td>
</tr>
<tr>
<td>60</td>
<td>125%</td>
<td>36%</td>
<td>30%</td>
<td>24%</td>
<td>12%</td>
<td>11%</td>
<td>9.4%</td>
<td>5.8%</td>
<td>5.4%</td>
</tr>
<tr>
<td>70</td>
<td>157%</td>
<td>47%</td>
<td>39%</td>
<td>31%</td>
<td>16%</td>
<td>15%</td>
<td>12.6%</td>
<td>7.8%</td>
<td>7.5%</td>
</tr>
<tr>
<td>80</td>
<td>190%</td>
<td>61%</td>
<td>50%</td>
<td>40%</td>
<td>20%</td>
<td>19%</td>
<td>16%</td>
<td>10%</td>
<td>9.6%</td>
</tr>
<tr>
<td>90</td>
<td>224%</td>
<td>74%</td>
<td>61%</td>
<td>49%</td>
<td>26%</td>
<td>23%</td>
<td>20%</td>
<td>13%</td>
<td>12%</td>
</tr>
</tbody>
</table>

Table 4-5. Simulated SWH estimation changing percent at different \( \tau_1 \) values
Still the same, we can answer that the SWH estimation error will mainly be determined by the absolute value of $\tau_1$ while has little relationship with $\tau_0$.

### 4.2.3 Changing System Parameters

In all the above simulations, the altimeter signal parameters, such as bandwidth and pulse length are set according to the ERS altimeter. It will be interesting to have a close look at the phase error influence when changing the signal parameters to different values.

Table 4-6, 7 are the simulation results for changing signal bandwidth and sweep time. In here, the group delay model still adopts the model in equation (4-31). The simulation results show that when signal bandwidth is decreased, the corresponding estimation error is also decreased. When signal bandwidth is only quarter of the original bandwidth, the error influence is very tiny. This is because that the decrease of signal bandwidth will reduce the resolution, therefore the estimation will be less sensitive to the influence of error.

<table>
<thead>
<tr>
<th>SWH estimation error (%)</th>
<th>$\tau_1 = 1$ ns</th>
<th>$\tau_1 = 2$ ns</th>
</tr>
</thead>
<tbody>
<tr>
<td>B = 320 MHz</td>
<td>1.1%</td>
<td>4.3%</td>
</tr>
<tr>
<td>B = 160 MHz</td>
<td>0.47%</td>
<td>1.2%</td>
</tr>
<tr>
<td>B = 80 MHz</td>
<td>0.08%</td>
<td>0.27%</td>
</tr>
</tbody>
</table>

(a) SWH = 5 m
Chapter 4. Phase Error Influence to Significant Wave Height

<table>
<thead>
<tr>
<th>SWH estimation error (%)</th>
<th>$\tau_l = 1\text{ns}$</th>
<th>$\tau_l = 2\text{ns}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>B = 320 MHz</td>
<td>6.1%</td>
<td>22.9%</td>
</tr>
<tr>
<td>B = 160 MHz</td>
<td>1.4%</td>
<td>5.78%</td>
</tr>
<tr>
<td>B = 80 MHz</td>
<td>0.68%</td>
<td>1.5%</td>
</tr>
</tbody>
</table>

(b) SWH = 2 m

Table 4-6 Phase error influence at different transmit bandwidth

Table 4-7 shows the simulation result when changing the altimeter signal sweep time. We can see when estimation error decreases when decrease the sweep time. While if we double the sweep time, the estimation error will become enormously high.

<table>
<thead>
<tr>
<th>SWH estimation error (%)</th>
<th>$\tau_l = 1\text{ns}$</th>
<th>$\tau_l = 2\text{ns}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>T = 50 $\mu$s</td>
<td>2.2%</td>
<td>7.6%</td>
</tr>
<tr>
<td>T = 100 $\mu$s</td>
<td>6.1%</td>
<td>22.9%</td>
</tr>
<tr>
<td>T = 200 $\mu$s</td>
<td>44.6%</td>
<td>1012%</td>
</tr>
</tbody>
</table>

Table 4-7 Phase error influence at different signal sweep time

The above simulations show that the SWH estimation error can be reduced by decrease the signal bandwidth or sweep time. However to complete the same bandwidth at shorter sweep time will be very difficult to achieve in hardware. Therefore decrease the signal bandwidth will be the only realistic choice.

4.214 Conclusion

In this section, the relationship between amplifier’s group delay distortions and on board SWH estimation changing errors are studied in detail by using two simple but representative group delay distortion models. Several conclusions can be drawn from the above simulation analysis:

- The existence of group delay distortion will broaden the main lobe of the point target response, as well as introduce in modulated spurious distortion on the main lobe amplitude.
- Group delay distortion has more serious impact on low SWH estimation than that for the high SWH condition.
- In choosing group delay distortion model, whether to adopt sine waveform or cosine waveform function will have very little effect on simulation results.
- The transmitter link’s absolute group delay value has very little effect on SWH estimation.
- When the group delay distortion model is a sine waveform function, the higher the distortion frequency, the more distortion the response’s main lobe will have. It is recommended that the peak value of the group delay distortion should be lower than 0.5 ns.
- If the group delay distortion is a linear function, it is recommended that the distortion slope should be smaller than 30 μ.
- To decrease the SWH estimation error, decrease the signal bandwidth while still keep the same sweep time will be a good choice. When bandwidth is decreased to quarter the original bandwidth, the estimation error can be ignored.

4.22 Chirp Signal Source Phase Distortion Analysis

Regarding the traditional digital chirp signal generator, the phase discontinuity is the main concern of signal distortion. However with the appearing of new powerful DDS chip which can operate in the specific chirp mode, this problem has been decreased to a less significant level. Nevertheless, there still possibly exist different phase variation. Figure 4.22 is a diagram that shows the possible chirp signal phase error condition [Sheehan D.V. et al. 1992].

![Figure 4.22 Example of skew symmetric phase error in chirp signal (after Sheehan D.V. et al. 1992)](image)

If we assume that the phase error generated by the chirp signal source is a function of frequency $E_d(f)$. Using the same principle as applied in power amplifier phase distortion analysis, this distortion can also be expressed as $E_d(t)$, expressed as:

$$ E_d(t) = A_c \cos(2\pi f_c \cdot t) $$

(4-39)

in here: $A_c$ is the peak value of the phase distortion from chirp signal generator

$f_c$ is the dominating frequency of the distortion

If only consider the influence of chirp signal generator, the transmitter signal then becomes:
Chapter 4. Phase Error Influence to Significant Wave Height

\[ s_c(t) = \exp[-j(2\pi ft + \mu t^2 + E_c(t))] = \exp[-j(2\pi ft + \mu t^2 + A_c \cos(2\pi f_c t))] \] (4-40)

the local deramping signal is:

\[ s_{local}(t) = A(t) \exp\left[j(2\pi f(t - \tau) + \mu(t - \tau)^2 + A_c \cos(2\pi f_c (t - \tau))\right)] \] (4-41)

the deramped signal is then:

\[ s(t) = s_c(t) \cdot s_{local}(t) = A(t) \exp\left[-j(2\pi f_c \tau + \mu(2t \tau - \tau^2) + A_c \left[\cos(2\pi f_c t) - \cos(2\pi f_c (t - \tau))\right]\right)] \] (4-42)

when using the above equation to simulate the point target response and final return signal and feed it to the SMLE algorithm, the SWH estimation changing percent can be obtained. The results are listed in Table 4-8.

<table>
<thead>
<tr>
<th>(f_2)</th>
<th>1kHz</th>
<th>10kHz</th>
<th>100kHz</th>
<th>Changing Percent (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>(A_2 = 0.2) (rad)</td>
<td>0.9278</td>
<td>0.9278</td>
<td>0.9278</td>
<td>0%</td>
</tr>
<tr>
<td>(A_2 = 0.4) (rad)</td>
<td>0.9278</td>
<td>0.9278</td>
<td>0.9188</td>
<td>1%</td>
</tr>
<tr>
<td>(A_2 = 0.6) (rad)</td>
<td>0.9278</td>
<td>0.9278</td>
<td>0.9278</td>
<td>0%</td>
</tr>
<tr>
<td>(A_2 = 0.8) (rad)</td>
<td>0.9278</td>
<td>0.9278</td>
<td>0.9278</td>
<td>0%</td>
</tr>
</tbody>
</table>

(a) SWH = 0.9278 (calm sea state)

<table>
<thead>
<tr>
<th>(f_2)</th>
<th>1kHz</th>
<th>10kHz</th>
<th>100kHz</th>
<th>Changing Percent (%)</th>
</tr>
</thead>
<tbody>
<tr>
<td>(A_2 = 0.2) (rad)</td>
<td>9.3025</td>
<td>9.3025</td>
<td>9.3025</td>
<td>0%</td>
</tr>
<tr>
<td>(A_2 = 0.4) (rad)</td>
<td>9.3025</td>
<td>9.3025</td>
<td>9.3016</td>
<td>0.01%</td>
</tr>
<tr>
<td>(A_2 = 0.6) (rad)</td>
<td>9.3025</td>
<td>9.3025</td>
<td>9.3025</td>
<td>0%</td>
</tr>
<tr>
<td>(A_2 = 0.8) (rad)</td>
<td>9.3025</td>
<td>9.3025</td>
<td>9.3025</td>
<td>0%</td>
</tr>
</tbody>
</table>

(b) SWH = 9.3025 m (serve sea state)

Table 4-8 Phase error in chirp signal generator influence towards SWH estimation

Comparing with the results in section 4.21, the results in Table 4-8 clearly show that the phase error generated by the chirp signal source has much less impact on SWH estimation. That is because when compared with the transmitted pulse length \(\tau\), the delay
time $\Delta t_d$ in Figure 4.23 (b) is very short. The appearance of this $\Delta t_d$ may introduce in an error in time delay, therefore influencing the height estimation. It does not, however, change the waveform shape itself much, and thus does not change SWH estimation, presuming the half power point is well tracked. Therefore, most of the phase error in the chirp signal would cancel with each other during the deramping processing. Whilst for the phase error in the power amplifier, as it only appears in the transmitted signal so does not have a counterpart to cancel with, therefore its influence is more serious.

(a) Schematic representation of a chirp transmitted by a satellite altimeter at time zero. A deramping chirp is generated internally by the altimeter at time $t_d$, which is intended to match the arrival time of the reflected chirp from nadir mean sea level.

(b) $\Delta t_d$ is timing error, generally is very small

Figure 4.23 Deramping principles, the time relationship of transmitted signal, desired and actual deramping signals (after Chelton D. B. et al. 1989)

4.23 Phase Error Influence in Half Power Point Tracking

The proper SWH estimation is based on the proper tracking of return waveform’s half power point. Only a single gate shift may possibly result in a large SWH error. Shows in Table 4-9, for lower SWH condition, the influence is very significant as the estimation result may be doubled.
**Table 4-9. Uncorrected half power point tracking influence in SWH estimation**

<table>
<thead>
<tr>
<th>Changing percent</th>
<th>One gate advance</th>
<th>One gate behind</th>
</tr>
</thead>
<tbody>
<tr>
<td>SWH = 0.82 m</td>
<td>2.28 / 178%</td>
<td>2.37 / 189%</td>
</tr>
<tr>
<td>SWH = 3.75 m</td>
<td>4.03 / 7.45%</td>
<td>4.02 / 7.3%</td>
</tr>
</tbody>
</table>

Table 4-10 shows the simulated result of the possible shifted gate under different distortion conditions. In here the simulation still uses the equation (4-31) group delay model. We can see that when \( \tau_i \) is a small value, the distortion will not result in more than one gate shift in tracking.

<table>
<thead>
<tr>
<th>Shifted gate</th>
<th>( \tau_i = 1\text{ns} )</th>
<th>( \tau_i = 2\text{ns} )</th>
<th>( \tau_i = 4\text{ns} )</th>
</tr>
</thead>
<tbody>
<tr>
<td>SWH = 1 m</td>
<td>0.1</td>
<td>0.3</td>
<td>1</td>
</tr>
<tr>
<td>SWH = 3 m</td>
<td>0.1</td>
<td>0.3</td>
<td>1</td>
</tr>
</tbody>
</table>

**Table 4-10 Shift in half power tracking**

### 4.24 Rayleigh noise influence

All the simulations in section 4.21 & 4.22 are for ideal conditions, which do not consider the influence of Rayleigh noise in a real application. Due to the existence of Rayleigh noise, it is difficult to judge the individual SWH estimation under different phase errors. So a 500 sample histdiagram is drawn by the simulation, shown in Figure 4.24. Figure 4.24(a) shows the SWH estimation with the consideration of Rayleigh noise but does not include any phase error in the model. While in Figure 4.24(b), the simulation considers the Rayleigh noise influence, as well as the phase distortion by setting amplifier group delay distortion. Compared with 4.24 (a) and (b), we can see very clearly that the existence of noise increases \( \sigma \) in the Gaussian shape histdiagram. This means that the \( \alpha-\beta \) filter will require more time to settle down to the proper value.
Chapter 4. Phase Error Influence to Significant Wave Height

(a) 500 times distribution results without error

(b) 500 times average distribution with consideration of error ($\tau_1 = 2\text{ns}$)
The simulations show, if no error is considered, the 500 samples average value of the SWH estimation is 1.99 m. Whilst if the group delay is considered \((\tau_1 = 2\,\text{ns})\), the 500 times average value will increase to 2.89 m. The changing percent for this case is 45%. Figure 4.24(c) shows when the error is controlled within the recommend value \((\tau_1 = 0.5\,\text{ns})\), the estimation result after 50 times average is 2.24 m, which can still satisfy the design target.

One thing we should bear in mind is that experience from the previous altimeter missions shows that by using on board tracking filter, nearly 75% of the noise energy can be removed. Whilst as the hardware imperfect performance is repeatable instead of randomly distributed, the error introduced by the hardware will remain the most the significant impact on the SWH on board estimation.

### 4.3 Amplitude Distortion Analysis

The expression for amplitude error is quite similar to what has been used for phase error distortion. Still we can divide the errors into two types, one comes from chirp generator and another comes from power amplifier stage. Compares with chirp generator, power amplifier's amplitude influence is much larger. So if we omit the amplitude variation that comes from chirp generator and only consider the amplifier's influence, the signal amplitude variation can be expressed as:

\[
A = a_0 + a_1 \sin(2\pi f_o t) \quad (4-43)
\]
Chapter 4. Phase Error Influence to Significant Wave Height

in here,

\[ a_0 \]  the ideal response of the amplifier
\[ a_1 \]  the maximum fluctuation of the error response
\[ f_a \]  the dominating frequency of amplitude variation

The diagram is shown by Figure 4.25. Although the real hardware response does not behaviour exactly like the diagram in Figure 4.25, via Fourier analysis, we still can simplify the complex shape into a combination of several simple expressions.

If the power amplifier’s amplitude distortion is repeatable, the error will not be decreased by averaging a group of received waveforms (what we apply for decreasing the Rayleigh noise). Using the same method in the previous sections, we can also simulate and analyse the amplitude influence towards SWH estimation. The transmitted and received signals can be expressed as:

\[ s_r(t) = |a_0 + a_1 \sin(2\pi f_a (t - \tau))| \exp[j(2\pi f(t - \tau) + u(t - \tau)^2)] \]  \hspace{1cm} (4.44)

\[ s_r(t) = \exp[j(2\pi f t + u t^2)] \]  \hspace{1cm} (4.45)

\[ S(t) = s_r(t) \cdot s_s(t - \tau) \]  \hspace{1cm} (4.46)

by changing to different \( f_a \) and \( a_1 / a_0 \) values, we can then get a group of results which represents the error influence toward SWH estimation, shown in the following table and figures for SWH=1.3 m and SWH = 4 m.
Table 4-11 Amplitude fluctuation influence at SWH=1.3 m (calm sea state)

<table>
<thead>
<tr>
<th>Changing Percent (%)</th>
<th>$a_1/a_0 = 0.1$ (0.9 dB)</th>
<th>$a_1/a_0 = 0.2$ (1.7 dB)</th>
</tr>
</thead>
<tbody>
<tr>
<td>100 Hz</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>1 k Hz</td>
<td>-0.032</td>
<td>-0.06</td>
</tr>
<tr>
<td>10 kHz</td>
<td>-3.14</td>
<td>-6.97</td>
</tr>
<tr>
<td>50 kHz</td>
<td>-5.32</td>
<td>-3.02</td>
</tr>
<tr>
<td>100 kHz</td>
<td>1.08</td>
<td>9.31</td>
</tr>
</tbody>
</table>

Table 4-12. Simulation results for SWH=4. m (medium sea state)

<table>
<thead>
<tr>
<th>Changing Percent (%)</th>
<th>$a_1/a_0 = 0.1$</th>
<th>$a_1/a_0 = 0.2$</th>
</tr>
</thead>
<tbody>
<tr>
<td>100 Hz</td>
<td>0</td>
<td>0</td>
</tr>
<tr>
<td>1 kHz</td>
<td>-0.002</td>
<td>-0.004</td>
</tr>
<tr>
<td>10 kHz</td>
<td>-0.299</td>
<td>-0.651</td>
</tr>
<tr>
<td>50 kHz</td>
<td>-0.347</td>
<td>0.440</td>
</tr>
<tr>
<td>100 kHz</td>
<td>0.383</td>
<td>2.341</td>
</tr>
</tbody>
</table>
The above results show that the amplitude error contribution towards SWH estimation is quite trivial and:

- The worst case occurs at calm sea state condition. When SWH is higher than 4 m, the SWH changing percent is nearly always less than 1%
- When the amplitude variation is $a_1/a_0 = 0.2$ (1.7 dB variation), the maximum SWH estimation changing percent is only about 7%. However, in the real design, the amplifier response is always controlled within 1 dB variation. Observe from Table 4-11 and 4-12, when $a_1/a_0 = 0.1$ (0.9 dB variation), the worst case of estimation occurs when $f_a = 50kHz \Rightarrow T = 20\mu s$, whilst the radar pulse duration is 100 $\mu s$ for the simulation. That means the error will vary 5 cycles within the amplifier sweep bandwidth, which is very unlikely to happen in the real world. Consequently we can believe even in very low SWH condition, the amplitude wavering will not result in great impact to the whole system performance.
- Although the results give us confidence in predicting amplitude impact, we still must calibrate the amplifier very carefully when considering ageing and possible temperature effects.

### 4.4 Hardware Error Influence in Delay/Doppler Altimeter

The Delay/Doppler altimeter takes advantage of pulse-to-pulse correlation to realise improved performance in requiring less transmitted power than the conventional pulse limited altimeter. This lead to a consideration of whether this special characteristic would impose more strict requirements on the payload hardware design.

Fortunately referring to Raney’s paper, it turns out that this Delay/Doppler altimeter processing is an unfocused synthetic aperture processing method. That means the...
altimeter resolution cell is smaller than the first Fresnel zone, which has a diameter of
\[ L = \sqrt{2 \lambda, H}, \]
shown in Figure 4.28. For the Ku band 800 km altitude radar, the diameter is only
\[ L = \sqrt{2 \lambda, H} = \sqrt{2 \times 24, 800} = 189\text{m}. \]
So in the real processing, all the echoes are
added without extra phase shift corrections.

![Figure 4.28 Unfocused method processing zone](image)

Referring to the previous experience in unfocused SAR processing [Tomiyasu K. 1978],
the requirement on hardware phase performance is not high, generally lower than 30
degrees distortion may satisfy the requirements.

### 4.5 Hardware Distortion towards Wind Speed Estimation

All the above discussions are focused on the SWH estimation. In this section, we will try
to address some analysis on wind speed estimation and the possible error influence.

Up to now there is no very accurate algorithm in estimating wind speed, especially at
high wind speed condition. For fast delivery estimation, within the range of 4 - 24 ms\(^{-1}\),
the accuracy is restricted to \( \pm 2\text{ms}^{-1} \). This result will become more accurate if can get the
help from other instruments such as Scanning Radiometer in the post ground processing.

Wind speed is derived from backscatter coefficient \( (\sigma^0) \), through a look up table, whilst
\( \sigma^0 \) is determined from the in-flight AGC data. The algorithm for calculating \( \sigma^0 \) that was
used in Seasat-1 is shown as [Townsend 1980]:

\[
\sigma^0 = C - Cal_a - \Delta AGC_a + L_{att} + 30 \log_{10} \frac{H}{796.44} + L_{aom} + B \quad (4-45)
\]
where

\[ C \]

Constant, determined on the basis of pre-launch thermal-vacuum data. In Seasat it is set as 39.93 dB

\[ C_{\text{CAL}} \]  
the measured value of the calibrate mode attenuator for step coarse attenuation steps

\[ \Delta AGC_a = AGC_a - AGC; \]  
i.e., the difference between the calibrate mode AGC for a step and the measured AFC of the ocean surface; \( a \) is chosen to minimise \( |\Delta AGC_a| \)

\[ L_{\text{att}} \]  
loss in antenna gain at the nadir due to the off nadir pointing

\[ L_{\text{atm}} \]  
the atmospheric loss in dB

\[ B \]  
bias determined from evaluating on orbit data in dB

From backscatter coefficient \( (\sigma^0) \) to wind speed, several look-up table based on empirical experience is implemented. Table 4-13 shows two possible altimeter wind algorithms to construct this look-up table. One is proposed by Brown [Brown et al. 1981], the other is proposed by Chelton [Chelton D.B. et al. 1985], the table shows the largest differences occurs at speed greater than 10 m/s\(^2\) whereas at moderate winds (7 – 8 m/s\(^2\)) there is very good agreement.

<table>
<thead>
<tr>
<th>Brown et al Algorithm</th>
<th>Wind speed (m/s)</th>
<th>Chelton &amp; McCabe algorithm</th>
<th>Wind speed (m/s)</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \sigma^0 ) (dB)</td>
<td></td>
<td>( \sigma^0 ) (dB)</td>
<td></td>
</tr>
<tr>
<td>6.7</td>
<td>24.6</td>
<td>8.4</td>
<td>24.2</td>
</tr>
<tr>
<td>6.8</td>
<td>23.7</td>
<td>8.5</td>
<td>23.0</td>
</tr>
<tr>
<td>6.9</td>
<td>22.8</td>
<td>8.6</td>
<td>21.9</td>
</tr>
<tr>
<td>7.0</td>
<td>22.0</td>
<td>8.7</td>
<td>20.8</td>
</tr>
<tr>
<td>7.1</td>
<td>21.3</td>
<td>8.8</td>
<td>19.8</td>
</tr>
<tr>
<td>7.2</td>
<td>20.6</td>
<td>8.9</td>
<td>18.9</td>
</tr>
<tr>
<td>7.3</td>
<td>19.9</td>
<td>9.0</td>
<td>18.0</td>
</tr>
<tr>
<td>7.4</td>
<td>19.3</td>
<td>9.1</td>
<td>17.1</td>
</tr>
<tr>
<td>7.5</td>
<td>18.7</td>
<td>9.2</td>
<td>16.3</td>
</tr>
<tr>
<td>7.6</td>
<td>18.1</td>
<td>9.3</td>
<td>15.5</td>
</tr>
<tr>
<td>7.7</td>
<td>17.5</td>
<td>9.4</td>
<td>14.8</td>
</tr>
<tr>
<td>7.9</td>
<td>16.5</td>
<td>9.5</td>
<td>14.1</td>
</tr>
<tr>
<td>8.1</td>
<td>15.6</td>
<td>9.6</td>
<td>13.4</td>
</tr>
<tr>
<td>8.3</td>
<td>14.8</td>
<td>9.7</td>
<td>12.7</td>
</tr>
<tr>
<td>8.5</td>
<td>14.1</td>
<td>9.8</td>
<td>12.1</td>
</tr>
<tr>
<td>8.8</td>
<td>13.1</td>
<td>9.9</td>
<td>11.5</td>
</tr>
</tbody>
</table>
Chapter 4. Phase Error Influence to Significant Wave Height

From the simulation experience on SWH estimation in this chapter we already know the hardware influence will be more significant at low SWH condition than in bad weather. Similarly, for the wind speed estimation that means the serious impact will occur at low wind speed condition. Refer to the above Table 4-12, we can see for that condition even quite large step change of backscatter coefficient $\sigma^0$ will only result in the small change of wind speed estimation. For example, when $\sigma^0$ increases from 12 dB to 12.5 dB, the wind speed estimation only changes 0.1 ms$^{-1}$.

Due to the high non-linearity of wind speed estimation, it is quite difficult to have a very clear understanding of the hardware impact on that before we have large amount of real data. However, we still can conclude that the hardware distortion will not introduce much impact on wind speed, if the correction based on climatological data were incorporated into look up table well ahead of actual processing. Fortunately, this project is intended in measuring wave height and wind speed at severe weather condition as calm sea has little threat towards shipping industry. Thus the errors at small SWH and wind speed are not of great importance to this applications.

4.6 Conclusion

In this chapter, the influence of hardware distortion on SWH and wind speed estimation has been evaluated in detail. The analysis mainly concentrates on the study of power amplifier and chirp signal source performance in terms of amplitude and phase distortion. There are several conclusions that can be drawn from the simulation:

- Hardware distortion influence is more serious in low SWH condition than that in higher SWH condition.
• According to the simulation, the amplitude fluctuation impact on SWH estimation is trivial. Even at the worse case, a 1.7 dB amplitude variation may only introduces in 7% estimation error.

• For the phase error influence, the error due to the power amplifier is more serious than the errors from the chirp signal generator.

  > The existence of group delay distortion in power amplifier will broaden the main lobe of the point target response, as well as introduce a modulated spurious distortion on the main lobe amplitude.

  > When the group delay distortion is a sine wave function, the higher the distortion frequency, the more serious the impact on response’s main lobe.

  > It is recommended that the peak value of this distortion should be lower than 0.5 ns.

• The existence of Rayleigh noise will increase the group delay error influence, especially when the estimation average time is not large enough.

• The hardware distortion does not have a great impact on wind speed estimation.

These results then impose the target specification for the design of a high efficient and highly linear power amplifier in the next chapter.
Chapter 5. Highly Efficient Class-F Power Amplifier – Design and Implementation

This chapter gives the simulation and implementation of a highly efficient S-band Class-F power amplifier. The objective of this study is first to demonstrate the feasibility of using a large-signal model to design a nonlinear power amplifier; second is to understand the design and operational principle of a highly efficient Class-F amplifier; and third to build the amplifier for the novel linearizer described in chapter 6.

The large-signal modelling of the MESFET was first studied. A comparison of different large-signal models, their DC and AC performance and ability in simulating various amplifier distortions were examined. The study concentrated on the models that are available in the HP-EEsof simulation software — COMMS, which is the package that was used for this microwave simulation design.

Based on the large-signal study, a highly efficient S-band Class-F power amplifier has been simulated in COMMS’s harmonic test bench, by using the STATZ large-signal model. The amplifier’s characteristics, such as large-signal S-parameters, power added efficiency (PAE), amplifier’s gain, drain harmonic-to-fundamental frequency voltage ratio, output signal phase distortion, and the output power spectrum will be given. The advantages/disadvantages of using this large signal modelling method, for amplifier design, and its accuracy will be discussed. Finally, the simulated Class-F amplifier was implemented. The test bench results of PAE, DC-RF efficiency, output power and gain are measured.

5.1 MESFET Large-signal Modelling

With the maturing of comprehensive microwave simulation packages and a deeper understanding of the device operational principles at higher frequencies, the computer aided design of power amplifiers has been more and more acknowledged as a very useful and accurate tool for predicting practical amplifier’s performance.

Large-signal MESFET models are best introduced by starting from the MESFET’s small-signal model, shown in Figure 5.1, as it has long served as an accurate model for virtually all GaAs MESFETs and HEMTs. The model has the capability to accurately reproduce measured S-parameters to frequencies of at least 60 GHz. However, this small-signal model lacks the ability to accurately represent the nonlinearity when the device is operating in large-signal condition. Therefore, a new model must be set-up for large-signal design. In the last two decades, enormous efforts have been put in for setting-up a useful MESFET large-signal model, and this leads to more and more confidence in designing various highly nonlinear circuits.
Chapter 5. Highly Efficient Class-F Power Amplifier - Design and Implementation

Currently, the large-signal models popularly used for circuit design and analysis are empirical models that incorporate analytical functions to represent the MESFET's non-linearity dependence on $V_{gs}$ and $V_{ds}$. They can be considered to consist of two parts. The first part is a static DC I-V portion that relates to the device's resistive elements and the transconductance. The second part is the dynamic portion that accounts for the reactive capacitive elements. If extrinsic interconnections or a package were to be included in the device model, the extracted associated inductances, capacitances, or distributed transmission elements would form additional elements in the dynamic portion.

The device's DC and C-V characteristics are required to obtain the input and output characteristics, such as breakdown voltages, contact junction ideality factors, and static capacitance behaviour. The parasitic elements may be extracted by using a series of "cold" and "hot" measurements. "Cold" measurements are performed with the transistor biased at $V_{ds} = 0$ and pinched-off (no current flow) and S-parameters are measured over a range of frequencies. A number of low frequency S-parameters measurements are usually needed to minimise inductive effects when extracting the pad capacitances. "Hot" measurements are performed with the gate in forward-bias. The frequency independent values of the parasitic elements (capacitances, resistances, and inductances) associated with packaging, mounting, and probing are then extracted by converting the measured data to Z-and/or Y-parameters. Following the extraction of the parasitic elements, an extensive set of S-parameter measurements is required over a wide range of DC bias conditions. In the case of power transistors this will include the breakdown, pinch-off, and gate forward bias conditions. It is also necessary to carry out these measurements over a frequency range extending from well below 1 GHz (to account for dispersion) to above the highest harmonic frequency of interest. The bias dependence of the intrinsic...
static transconductance and output conductance obtained from the DC results are then compared with the bias dependent S-parameter results. Differences between the two sets of results are used to determine the dispersion (often resulting in a fourfold increase in output conductance over the frequency range 1 Hz to 1 MHz). Dispersion is usually modelled with the addition of an RC network in the output circuit of the FET.

Although the above steps are very useful in extracting accurate device’s large-signal parameters of the device, these measurements are extremely difficult to fulfil for most of the microwave laboratories. The main reason is that the measurement requires a very accurate network analyser, which can operate up to a very high frequency plus an accurate probing system. The general simplified method people use, therefore, is only to measure the device’s small-signal S-parameters and its I/V curves, and then extract the model parameters by fitting these curves.

There are several existing large-signal models available to the designer, such as the Curtice model, Materka model, Statz model and the TOM model. The following section is a study summary for the characteristic of each model.

5.1. The Curtice Quadratic Function MESFET Model

The Curtice quadratic function large-signal MESFET model was introduced in 1980 [Curtice W. R. et al. 1980]. The model assumes a square law dependence of drain current $I_{ds}$ on gate bias $V_{gs}$ by using the following analytical function:

$$I_{ds} = \beta \left( V_{gs} + |V_{p}| \right)^2 (1 + \lambda V_{ds}) \tanh(\alpha V_{ds}) \quad (5-1)$$

where:

- $\beta$ is a transconductance scaling parameter
- $\lambda$ defines the DC value of $R_{ds}$
- $V_{p}$ is the device’s pinch off voltage
- $\alpha$ adjusts the knee of $I_{ds}$ versus $V_{ds}$

The hyperbolic tangent is a continuous function that provides versatility in fitting the knee of the $I_{ds}$ versus $V_{ds}$ characteristics. The $\tanh$ function has been adopted for all of the popular, empirical, large-signal MESFET models. For computational simplicity the $\tanh$ function can be approximated with good accuracy by:

$$\tanh(\alpha V_{ds}) \approx \begin{cases} 1 - \left(1 - \alpha V_{ds} / 3 \right)^3 & 0 < V_{ds} < 3 / \alpha \\ 0 & V_{ds} \geq 3 / \alpha \end{cases} \quad (5-2)$$
At $V_{ds} = 0$, both $tanh$ function and its approximation have a slope of $\alpha$. The $(1 + \lambda V_{ds})$ in equation (5-1) models the DC output conductance slope of the $I_{ds}$ versus $V_{ds}$ curves in the saturated region. The measured $I_{ds} - V_{gs}$ characteristic of a typical transistor is plotted together with the Curtice quadratic fitted model in Figure 5.2.

![Figure 5.2 Curtice quadratic model fitting to the $I_{ds}$-$V_{gs}$ characteristic of a typical MESFET (after Walker J. L. 1993)](image)

In COMMS, this model is represented by the Curtice2 model. Its simple schematic diagram is shown in Figure 5.3.

![Figure 5.3 Curtice2 model schematic in COMMS](image)

The merit of the Curtice quadratic model is its simplicity, which leads to quick computations and good convergence qualities in simulations. However, the square law model for the $gm$ nonlinearity means only second harmonics can be generated by $gm$. Thus $gm$ does not contribute to higher than third order harmonics and inter-modulation distortion. That is the major disadvantage of Curtice quadratic model.
5.12. Materka Model

The MESFET DC I-V characteristic indicates that the value of $V_{ds}$ at which the drain current is pinched off is seen to vary with $V_{gs}$. This characteristic is referred to as a soft pinch-off. The Crutice quadratic model is unable to fit this soft pinch-off behaviour. The Materka model [Kacprzak T. et al. 1983] has generated interest largely because it can represent this soft pinch-off characteristic. The analytical function is given by the followings:

$$I_{ds} = \left(1 + \frac{V_{gs}}{V_p}\right)^2 \tanh\left[\alpha V_{ds} \left(V_{gs} + V_p\right)\right]$$  \hspace{1cm} (5-3)

where $V_p = \left|V_p\right| + \gamma V_{ds}$. The coefficient $\gamma$ provides the adjustment of pinch-off voltage $V_p$ as a function of $V_{ds}$ from a nominal $\left|V_p\right|$.

5.13 The Statz Model

The STATZ model [Statz H.P. et al. 1987] introduces two important advances over the previous MESFET models. First, the model incorporates an improved analytic DC I-V formulation. Second, the model offers an improved charge model representation of $C_{gs}$ and $C_{gd}$, as functions of both $V_{gs}$ and $V_{ds}$. The simple Statz DC I-V modelling function, which has proven to be highly versatile in fitting a broad variety of MESFETs, contains only two coefficients. It is given by

$$I_{ds} = \left[\frac{\beta(V_{gs} + \left|V_p\right|)^2}{1 + \beta(V_{gs} + \left|V_p\right|)\tanh(\alpha V_{gs})}\right](1 + \lambda V_{ds}) \tanh(\alpha V_{gs})$$  \hspace{1cm} (5-4)

This equation provides usually good control over the contour in the transition of $I_{ds} = f(V_{gs})$ as it goes from square law to linear behaviour. The broad latitude in $I_{ds}$ - $V_{gs}$ fitting using the Statz function is illustrated in the family of curves in Figure 5.4. A hypothetical device with $I_{ds} = 1A$ and $\left|V_p\right| = 3V$ was assumed throughout. By adjusting $\beta$ and $b$ in equation (5-4) over a broad range, $I_{ds}$ - $V_{gs}$ behaviour will vary from a square law throughout ($b=0$) to a linear throughout (very large $b$).
Despite the improved advancement provided by the Statz formulation, it lacks the means of representing the soft pinch-off characteristic common to many MESFETs. However, the STATZ DC I-V function can be expected to produce a very good fitting to the \( g_m \) non-linearities near pinch-off, making the model a good choice for simulating MESFET mixers and class AB and Class B amplifiers. COMMS also provides the STATZ model, shown in Figure 5.5.

![Figure 5.5 STATZ model schematic diagram in COMMS](image)

It is interesting to note that schematic diagrams are nearly the same as that for the Curtice2 model. The difference lies in that the two models have two different methods in presenting gate-source and gate-drain junctions.
5.14. The TOM Model
The TOM model [McCamant A. J. et al. 1990] incorporates the soft pinch-off feature of the Materka model into a significantly modified Statz model. Some of the TOM features demonstrated in these characteristics are:

1) A simple and widely controllable means for fitting gm as a function of Vgs as the transistor changes from square law dependence on Vgs towards a linear dependence.

2) A soft or gradual pinch-off characteristic, that is, a pinch-off voltage $V_p$ whose magnitude increases with increasing $Vds$;

3) A simple means for modelling the dependence of $Rds$ on $Vgs$, $Vds$ and channel temperature.

Figure 5.6 is the equivalent circuit for the TOM model in COMMS.

![Figure 5.6 Equivalent circuit for TOM model in COMMS](image)

5.15. The Curtice - Ettenberg Model
The Curtice - Ettenberg model is based on the work of Curtice [Curtice W. R. et al. 1985], it uses a third order polynomial to fit the I versus V characteristic and is often referred to as the cubic model. In COMMS, the drain current of this model (Curtice3), is computed with the following expression:

$$I_{ds} = I_{ds0} \cdot \tanh(\gamma \cdot V_{ds})$$  \hspace{1cm} (5-5)

where
\[ I_{d0} = \left[ A0 + A1 \cdot V_1 + A2 \cdot V_1^2 + A3 \cdot V_1^3 \right] \]
\[ V_1 = V_{gs} \left[ 1 + \beta (V_{dso} - V_{ds}) \right] \]

\( \beta \) controls the change in the pinch-off voltage with \( V_{ds} \),

\( V_{dso} \) is the drain source voltage at which the \( Ai \) coefficients are evaluated.

To some extend, this analytic model function can fit I-V characteristics with a soft pinch-off. However, this model has a poor fit to the region of curvature close to pinch-off. Except for the Class-A amplifier, the accurate fitting of the pinch-off region is important for producing realistic simulations of the output power as a function of the input drive signal.

### 5.16. Discussion

For quasi-linear class-A large-signal operation, where harmonic and inter-modulation distortion (IMD) simulations are not of principal interests, the accurate fitting of the \( I_{ds} - V_{gs} \) characteristics around the quiescent bias region of the load-line should be emphasised. The Curtice quadratic function model is a good choice as it requires less simulation time and provide reasonable accuracy. The Curtice-Ettenberg model should be capable of achieving the most accuracy in representing the third-order nonlinearities. Because of all the models considered, only the third-order polynomial function offers direct control of third-order nonlinearity through the coefficient \( A3 \) of the third order term. However, for simulating the electrical performance characteristics of Class-B & C amplifiers, the focuses should be on fitting of the \( I_{ds} - V_{gs} \) characteristics near pinch-off. For this case, STATZ model is a good choice.

It is important to note that in large signal modelling, for calculating the power efficiency for Class-B & C amplifiers, the fixed value of \( R_{ds} \) in the SPICE model (such as \( R_c \) in Figure 5.2) should be set to a large value when simulating the operating efficiency of Class-AB to C stages. The reason for this is that under Class-B or C operating conditions, where the device should be non-conducting for at least half the cycle, \( R_{ds} \) should be very large. A fixed low value that is extracted from small-signal Class-A bias S-parameters could not represent it very well. It will result in power dissipation, which causes simulations of DC-RF efficiency is lower than what can be achieved in practice. So it will be more accurate if \( R_{ds} \) is eliminated by setting it to be approximately an open circuit.
5.2 MESFET Large Signal Model Simulation in COMMS

After studying the popular large-signal MESFET models, STATZ model was chosen as the most suitable one for evaluating the AM-AM and AM-PM distortions generated by Class-AB to Class-C biasing amplifier.

5.21 Determining Large-Signal Model Parameters

The quadratic dependence of the drain current with respect to the gate voltage is computed with the following expression (in the region $V_{ds} \geq 0$ V).

$$I_{ds} = \beta \cdot (V_{gd} - V_{TO})^2 \cdot (1 - \lambda \cdot V_{ds}) \cdot \tanh(\alpha \cdot V_{ds}) \quad (5-6)$$

- $V_{TO}$ is threshold voltage
- $V_{ds}$ is the gate-to-source voltage
- $\lambda$ is the parameter related to drain conductance
- $\beta$ is a parameter
- $\alpha$ determines the voltage at which drain current characteristics saturate

The STATZ model in COMMS also includes a number of features that (although not described in Statz's paper) are generally accepted to be important features of a GaAs MESFET model. These include a gate delay factor ($\tau$), and input charging resistance ($RI$), gate junction forward conduction and gate-drain breakdown.

The modified drain-source current is, therefore, given by the following expressions:

For $0 < V_{ds} < 3/\alpha$

$$I_{ds} = \frac{\beta (V_{gd} - V_{TO})^2}{1 + \Theta (V_{gd} - V_{TO})} \left[ 1 - \left( 1 - \frac{\alpha V_{ds}}{3} \right)^3 \right] (1 - \lambda V_{ds}) \quad (5-7)$$

For $V_{ds} \geq 3/\alpha$

$$I_{ds} = \frac{\beta (V_{gd} - V_{TO})^2}{1 + \Theta (V_{gd} - V_{TO})} (1 + \lambda V_{ds}) \quad (5-8)$$

The current is set to zero for $V_{gd} < V_{TO}$.
Here $\Theta$ is a parameter with a unit of $V^{-1}$. The more gradual the doping the more the profiles appear to give a lower value of $\Theta$. Neglecting $\Theta$ cannot be tolerated in most circuit simulations. Table 5-1 gives the meaning of each different parameters.

<table>
<thead>
<tr>
<th>Name</th>
<th>Meaning</th>
<th>Unit</th>
<th>Default value</th>
</tr>
</thead>
<tbody>
<tr>
<td>BETA</td>
<td>Transconductance parameter</td>
<td>A/V^2</td>
<td>0.01</td>
</tr>
<tr>
<td>VTO</td>
<td>Threshold voltage</td>
<td>V</td>
<td>-2.5</td>
</tr>
<tr>
<td>ALPHA</td>
<td>Current saturation parameter</td>
<td>I/V</td>
<td>2.0</td>
</tr>
<tr>
<td>LAMBD A</td>
<td>Output conductance parameter</td>
<td>I/V</td>
<td>0.0</td>
</tr>
<tr>
<td>THETA</td>
<td>Parameter that controls $I_d-V_{gs}$ characteristic transition from quadratic to linear behaviour (called &quot;b&quot; in Statz's paper)</td>
<td>I/V</td>
<td>0.0</td>
</tr>
<tr>
<td>TAU</td>
<td>Transit time under gate</td>
<td>S</td>
<td>0.0</td>
</tr>
<tr>
<td>VBR</td>
<td>Gate-drain junction reverse bias breakdown voltage (gate-source junction reverse bias breakdown voltage with $V_{ds}&lt;0$)</td>
<td>V</td>
<td>infinity</td>
</tr>
<tr>
<td>IS</td>
<td>Gate junction reverse saturation current (diode model)</td>
<td>A</td>
<td>$10^{-14}$</td>
</tr>
<tr>
<td>N</td>
<td>Gate junction ideality factor (diode model)</td>
<td>-</td>
<td>1.0</td>
</tr>
<tr>
<td>VBI</td>
<td>Build-in gate potential</td>
<td>V</td>
<td>1.0</td>
</tr>
<tr>
<td>FC</td>
<td>Coefficient for forward bias depletion capacitance (diode model)</td>
<td>-</td>
<td>0.5</td>
</tr>
<tr>
<td>RC</td>
<td>Used with CRF to model frequency dependent output conductance</td>
<td>ohm</td>
<td>infinity</td>
</tr>
<tr>
<td>CRF</td>
<td>Used with RC to model frequency dependent output conductance</td>
<td>F</td>
<td>0.0</td>
</tr>
<tr>
<td>RD</td>
<td>Drain ohmic resistance</td>
<td>ohm</td>
<td>0.0</td>
</tr>
<tr>
<td>RG</td>
<td>Gate resistance</td>
<td>ohm</td>
<td>0.0</td>
</tr>
<tr>
<td>RS</td>
<td>Source ohmic resistance</td>
<td>ohm</td>
<td>0.0</td>
</tr>
<tr>
<td>RIN</td>
<td>Channel resistance</td>
<td>ohm</td>
<td>0.0</td>
</tr>
<tr>
<td>CGSO</td>
<td>Zero bias gate-source junction capacitance (diode model)</td>
<td>F</td>
<td>0.0</td>
</tr>
<tr>
<td>CGDO</td>
<td>Zero bias gate-drain junction capacitance (diode model)</td>
<td>F</td>
<td>0.0</td>
</tr>
<tr>
<td>DELTA1</td>
<td>Capacitance saturation transition voltage parameter</td>
<td>-</td>
<td>0.3</td>
</tr>
<tr>
<td>DELTA2</td>
<td>Capacitance threshold transition voltage parameter</td>
<td>-</td>
<td>0.2</td>
</tr>
<tr>
<td>CDS</td>
<td>Drain-source capacitance</td>
<td>F</td>
<td>0.0</td>
</tr>
<tr>
<td>CGS</td>
<td>Gate-source capacitance</td>
<td>F</td>
<td>0.0</td>
</tr>
<tr>
<td>CGD</td>
<td>Gate-drain capacitance</td>
<td>F</td>
<td>0.0</td>
</tr>
<tr>
<td>KF</td>
<td>Flicker noise coefficient</td>
<td>-</td>
<td>0.0</td>
</tr>
<tr>
<td>AF</td>
<td>Flicker noise exponent</td>
<td>-</td>
<td>1.0</td>
</tr>
<tr>
<td>TNOM</td>
<td>Nominal ambient temperature at which these model parameters were derived</td>
<td>°C</td>
<td>27.0</td>
</tr>
<tr>
<td>XTI</td>
<td>Temperature exponent for saturation current</td>
<td></td>
<td>3.0</td>
</tr>
<tr>
<td>EG</td>
<td>Energy gap for temperature effect on IS</td>
<td>eV</td>
<td>1.11</td>
</tr>
<tr>
<td>VTOTC</td>
<td>$V_{TO}$ temperature coefficient</td>
<td>V/°C</td>
<td>0.0</td>
</tr>
<tr>
<td>BETATC E</td>
<td>Drain current exponential temperature coefficient</td>
<td>%/°C</td>
<td>0.0</td>
</tr>
<tr>
<td>FPE</td>
<td>Flicker noise frequency exponent</td>
<td></td>
<td>1.0</td>
</tr>
</tbody>
</table>

Table 5-1. Statz model parameters (COMMS menus)
The output from the radar altimeter is a Ku-band 7W chirp signal. However, due to the unavailability of Ku band power amplifier devices at the time of simulation, an S-band devices has been chosen as a proof-of-concept. It is a Mitsubishi Semiconductor power GaAs FET device – MGF0906B. The maximum output power is 37 dBm at 1 dB point, and its typical gain is 11 dB at 2.3 GHz.

Due to the lack of accurate probe for testing the device’s S-parameters, it was not possible to carry out the S-parameters measurement at different biasing points. Therefore, only the S-parameters provided by the manufacturer, which are tested under the Class-A biasing condition, can be used. This feature will obviously impose some limitations on the model accuracy. However, after referring to some previous works [Walker J. L. B. 1993], it is believed that by carefully adjusting the parameters to fit the DC curves and small-signal S-parameters, the model still can be reasonably accurate in simulating power amplifier.

To determine large-signal model parameters for MGF0906B, a circuit schematic bench is set up in COMMS, shown in Figure 5.7. FET3 in Figure 5.7 is the STATZ model, that has 32 parameters needed to be determined. The aim of the inductors L3, L4, L5 in the figure are to connect the drain, source and gate terminal of the device are to enable the phase of the simulated S parameters to fit with the available small-signal S-parameters as much as possible. All of them are very small values.

In the COMMS’s STATZ model, the user is provided with two options for modeling the junction capacitance of a device. The first is to model the junction as a linear element, i.e. a constant capacitance. The second is to model the junction using a diode depletion capacitance model. If a non-zero value of $C_{gs}$ is specified in the model, then the gate-source junction will be treated as a linear element. A zero value for either $C_{gs}$ or $C_{gd}$ will force the use of the diode depletion capacitance model for that particular junction. The two junctions are independent of each other. So it is possible to model one junction as a linear element while the other is modelled nonlinearly.
5.22 Simulation Results

First the model's DC performance is examined in COMMS's DC bias test bench. It has been found that the first five parameters and $R_p$, $R_g$ and $R_s$ in STATZ's model, play the most important roles in deciding the model's DC performance.
Figure 5.8 (a) (b) and (c) are the simulated $I_{\text{drain}}$ curves vs. different $V_{gs}$ and $V_{ds}$ biasing. The STATZ model parameters are set-up according to the values in Figure 5.8. Figure 5.8(c) shows that the device breakdown voltage is actually changing with $V_{gs}$; the more negative the device gate is biased, the earlier the drain current enters into the breakdown region. This presents a very clear soft pinch-off phenomena.
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Figure 5.8 Simulated DC I-V curves

(c) Drain current when $V_{ds}$ approaches the breakdown value (VBR in STATZ model is set to 30V)

$C_{GS0}$, $C_{GDO}$ and $C_{DS}$ are the three most important parameters in deciding model's AC performance and S-parameters. To determine suitable values requires considerably patience in tuning. The final values of these parameters combine with the parameters that are decided by the DC bench will generate the device’s time domain voltage and current waveforms as well as the small/large-signal S-parameters.

The time domain $V_{\text{drain}}(t)$ performance is also studied. Figure 5.9 is the simulated $V_{\text{drain}}(t)$ curves vs. different input power conditions. The results are simulated in the linear test bench. It can be seen that when the gate is biased to Class-B ($V_{gs} = -3$ V), $V_{\text{drain}}(t)$ curves display the expected half-sinusoid and square-wave shapes. The curves also show that the higher the input power, the more obvious half-sinusoid and squaring of the signal can be observed.
Figure 5.9. Simulated $V_{\text{drain}}(t)$ vs. input power for Class-B ($V_{gs} = -3$ V, $V_{ds} = 8$ V) biasing condition.

Figure 5.10 shows a group of curves that illustrate the differences between the extracted S-parameters decided in this work and the measured S-parameters provided by the manufacturer. The simulation frequencies sweep from 0.5 to 3 GHz.

(a) $S_{11}$, $S_{22}$ results
The above diagrams show that the extracted and measured results are reasonably close with each other for all the S-parameters. For the interested frequency band, the difference between simulated S21 and measured S21 is less than 0.5 dB. However, for the S22 and S11 results, there still exist some phase difference between the real and simulated results, especially at the upper edge of the sweeping frequency. Fortunately, for an operation at 2.2 GHz, the extracted small-signal S-parameter results are very close to the values provided by the manufacturer.

5.23 Temperature Influence

It is well known that the space environment changes dynamically with temperature. It generally requires that all electronic components must work well at least within the -10°C to 50°C temperature range. Therefore, before carry out a detailed simulation analysis for Class-F power amplifier design by using the above large signal model, it is very helpful to have a clear understanding of the amplifier performance under a dynamic temperature-changing environment.

In the STATZ model, it specifies a parameter, TNMO, which represents the nominal temperature at which the model parameters are calculated or extracted. In order to simulate the device's performance at temperatures other than TNOM, several model parameters must be scaled with temperature. The temperature at which the device is simulated is specified by the TEMP parameter on the device item. For example, the transconductance $\beta$ varies as:
\[ \beta^{\text{NEW}} = \beta \cdot 1.01^{\text{BETATCE}(\text{TEMP}-\text{TNOM})} \quad (5-9) \]

and saturation current \( IS \) scales as:

\[ IS^{\text{NEW}} = IS \cdot \exp \left[ \left( \frac{\text{TEMP}}{\text{TNOM}} - 1 \right) \cdot \frac{q \cdot \text{EG}}{k \cdot N \cdot \text{TEMP}} + \frac{\text{XTI}}{N} \cdot \ln \left( \frac{\text{TEMP}}{\text{TNOM}} \right) \right] \quad (5-10) \]

Figure 5.11 shows a plot of amplifier gain vs. temperature performance. The amplifier here is a Class-A amplifier and designed by using the above extracted MGF0906A model. The parameter \( \text{BETATCE} \) in the device model is set to 0.1%/°C. The diagram clearly shows that the higher the temperature, the lower the amplifier power gain will be.

![Figure 5.11 Amplifier's power gain vs. environment temperature](image-url)
5.3 Highly Efficient Class-F Power Amplifier Design

This section will give a detailed theoretical analysis for two types of highly efficient Class-F amplifiers: second harmonic peaking Class-F amplifier and the third harmonic peaking Class-F amplifier. The analysis on each type of amplifier will be followed by the simulation results based on the large-signal model developed in the previous section. Finally, the influence of the device large-signal model parameters on amplifier group delay performance will be discussed in detail.

Class-F has long been known as one kind of highly efficient power amplifier [Tyler V. J. 1958] [Snider D. M. 1967] [Goto N. et al. 1995] [Raab F. H. 1997]. The main idea of this concept is to make use of the unwanted harmonic frequencies to shape the drain current and drain voltage so that when one reaches its peak the other is always around zero. Thus, to reduce the DC power dissipation and to achieve a higher efficiency at the amplifier's output. Although the design concept will allow the existence of harmonics at the drain, by properly choosing the output-matching network, the harmonics will be prevented from being transmitted to the load resistance.

If the device is assumed to be an ideal voltage controlled current generator, the drain voltage will be the multiply product of drain current and load impedance. Its waveform can then be controlled by a proper combination of fundamental and higher order harmonic load terminations. The drain voltage can be shaped either by even harmonic or odd harmonic components. This leads to two Class-F amplifier design methods: second harmonic peaking and third harmonic peaking, which makes use of second harmonic drain component and third harmonic drain component respectively. The detailed theoretical analysis for these two amplifier design concepts will be given in the following sections.

5.3.1 Second Harmonic Peaking

Second harmonic peaking makes use of the drain voltage component of second harmonic to shape the drain voltage. The advantage of this design concept is that the amplifier requires only a very low drain DC supply voltage to enable the drain voltage to swing within the full range. This advantage makes the amplifier very attractive for low DC supply applications, such as mobile phone handsets.

5.3.1.1 Theoretical Analysis

The design concept requires that the fundamental and drain voltage components of second harmonic are out-of-phase. If one omits the higher drain harmonic components, the drain voltage can be expressed as:

\[ v_{\text{drain}}(\theta) = V_{\text{drain,dc}} - V_{\text{drain,1}} \cos(\theta) + V_{\text{drain,2}} \cos(2\theta) \quad (5-11) \]

Here \( \theta = \omega t \), and \( \omega \) is the fundamental frequency.
If one omits the knee voltage influence and assumes that the swing range of $v_{\text{drain}}(\theta)$ is [0,1], and at the same time normalizes $v_{\text{drain}}$ to $V_{\text{drain},1}$ by using $\beta$ as $\beta = \frac{V_{\text{drain},2}}{V_{\text{drain},1}}$, then $v_{\text{drain}}(\theta)$ can be expressed as:

$$v_{\text{drain}}(\theta) = V_{\text{drain,dc}} - V_{\text{drain},1} \cos(\theta) + \beta \cos(2\theta) \quad (5-12)$$

If one increases $\beta$ in equation (5-12) while still keeping the $v_{\text{drain}}(\theta)$ swing between zero and its maximum value, the values of $V_{\text{drain,dc}}$ and $V_{\text{drain},1}$ need to be adjusted correspondingly to satisfy equation (5-12). Obviously, the overall drain voltage shape will be different for different $\beta$. Its waveforms for different $\beta$ are shown in Figure 5.12.

![Figure 5.12 Drain voltage waveforms at different $\beta$](image)

It can be seen that with the increase of $\beta$, the drain voltage waveform changes from a pure sinusoid waveform to a waveform that has narrower wave crest and wider wave trough. Table 5-2 gives values of $V_{\text{drain,dc}}$, $V_{\text{drain},1}$ and $V_{\text{drain},2}$ for different $\beta$. The table shows clearly that when $\beta$ increases from $\beta = 0$ to $\beta = 0.7$, both the $V_{\text{drain,dc}}$ and $V_{\text{drain},1}$ decrease. However, the $V_{\text{drain,dc}}$ decreases faster than that of $V_{\text{drain},1}$. By doing this way, it will effectively drop the amplifier’s DC power consumption.

<table>
<thead>
<tr>
<th>$\beta$</th>
<th>$V_{\text{drain,dc}}$</th>
<th>$V_{\text{drain},1}$</th>
<th>$V_{\text{drain},2}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.0</td>
<td>0.5</td>
<td>0.5</td>
<td>0</td>
</tr>
<tr>
<td>0.1</td>
<td>0.45</td>
<td>0.5</td>
<td>0.05</td>
</tr>
<tr>
<td>0.2</td>
<td>0.4</td>
<td>0.5</td>
<td>0.1</td>
</tr>
<tr>
<td>0.3</td>
<td>0.3554</td>
<td>0.4959</td>
<td>0.1488</td>
</tr>
<tr>
<td>0.4</td>
<td>0.3373</td>
<td>0.4734</td>
<td>0.1893</td>
</tr>
<tr>
<td>0.5</td>
<td>0.3333</td>
<td>0.4444</td>
<td>0.2222</td>
</tr>
<tr>
<td>0.6</td>
<td>0.3356</td>
<td>0.4152</td>
<td>0.2491</td>
</tr>
<tr>
<td>0.7</td>
<td>0.3407</td>
<td>0.3878</td>
<td>0.2715</td>
</tr>
</tbody>
</table>

Table 5-2. $V_{\text{drain,dc}}$, $V_{\text{drain},1}$ and $V_{\text{drain},2}$ values for full swing condition at different $\beta$. 

5-19
If one assumes that the drain current waveform is the same for different \( \beta \), which means the gate biasing point does not change with all \( \beta \) values, then the amplifier DC-RF efficiency will only be decided by the ratio value of \( V_{\text{drain},1}/V_{\text{drain},\text{dc}} \). The simulation results are plotted and shown in Figure 5.13. Here it is assumed that the amplifier gate supply is biased to Class-A which is the mid-point between zero and pinch off.

![Graph showing V_{drain}, V_{\text{drain},1}, V_{\text{drain},2} and amplifier efficiency at different \( \beta \)](image)

Figure 5.13 shows that when \( \beta \) varies between \([0, 0.375]\), the requirement for drain DC supply drops down while the DC-RF efficiency increases simultaneously. For this simulation condition, the maximum DC-RF efficiency, \( \eta \), is \( \eta = 70.7\% \), which occurs at \( \beta = |V_{\text{drain},2}/V_{\text{drain},1}| = 0.375 \), where \( V_{\text{drain},\text{dc}} = 0.3429 \), \( V_{\text{drain},1} = 0.4849 \) and \( V_{\text{drain},2} = 0.1722 \). It can be calculated that by properly tuning the fundamental and harmonic terminations, at the maximum efficiency point, compared with the normal Class-A amplifier, the drain DC supply requirement of the Class-F second harmonic peaking amplifier drops down by 31.4\%, while the maximum amplifier output power capability only drops by 3\%.

The drain voltage waveform for maximum efficiency is shown in Figure 5.14.
An immediate question that may arise after studying the basic second harmonic peaking Class-F principle is, how does one get the proper load termination value so that one can obtain the proper ratio of \( \frac{V_{\text{drain,2}}}{V_{\text{drain,1}}} \) that is required for the amplifier design concept? Moreover, how does one determine the drain biasing voltages at different gate biasing points, and how can one achieve the required 180° out-of-phase relationship between the fundamental and second harmonic drain voltage?

It is clear that if one assumes that the FET transistor is an ideal voltage-controlled current generator and its gate is biased at the middle point between zero and pinch-off point, at a certain fixed drain biasing point, the drain current conduction angle is also fixed. This can be expressed as:

\[
\alpha = 2 \cos^{-1} \left( \frac{I_{\text{drain,DC}}}{I_{\text{drain,DC}} - I_{\text{max}}} \right) \tag{5-13}
\]
The drain current can, therefore, be written as a function of conduction angle, $\alpha$, and the maximum drain current, $I_{\text{max}}$, as:

$$I_{\text{drain}}(\theta) = \frac{I_{\text{max}}}{(1 - \cos(\alpha/2))} \left[ \cos \theta - \cos \left( \frac{\alpha}{2} \right) \right]$$

$$= 0$$

$$-\pi < \theta < -\alpha/2, \quad \alpha/2 < \theta < \pi$$

(5-14)

The drain current waveform, as a function of conduction angles, is shown in Figure 5.16. The three waveforms in the plot represent the drain current for Class-A, Class-AB and Class-B respectively. Their conduction angles are $2\pi$, $\pi$, and $\pi/2$ respectively.

![Figure 5.16 Drain current at different conduction angle](image)

Each individual harmonic drain current component values, as the function of conduction angle, can be calculated by a simple Fourier series as:

$$I_{\text{drain},n}(\alpha) = \frac{1}{2\pi} \int_{-\alpha/2}^{\alpha/2} I_{\text{drain}}(\theta, \alpha) d\theta$$

$$I_{\text{drain},1}(\alpha) = \frac{1}{\pi} \int_{-\alpha/2}^{\alpha/2} I_{\text{drain}}(\theta, \alpha) \cos \theta d\theta$$

$$I_{\text{drain},2}(\alpha) = \frac{1}{\pi} \int_{-\alpha/2}^{\alpha/2} I_{\text{drain}}(\theta, \alpha) \cos 2\theta d\theta$$
Figure 5.17 shows a diagram that illustrates the relationship between conduction angle $\alpha$ and drain current’s DC, fundamental, second and third harmonic components. In the simulation, the conduction angle changes from $\alpha=2\pi$ (Class-A) to $\alpha=0$ (Class-C). All the current components are normalized to $I_{\text{max}}$.

The above figure shows that within the area of $0 \leq \alpha \leq 2\pi$, the second and fundamental current components are always in phase. Referring to the drain voltage requirement for the second harmonic peaking Class-F design, it seems impractical to find a fundamental frequency load which has a negative real impedance and zero image impedance to satisfy the voltage shaping requirement in equation (5-11).

The possible solution is to introduce a second harmonic gate voltage to combine with the fundamental gate voltage to control the drain current, to obtain a correct phase relationship between drain fundamental and second harmonic current components. Thus, when they multiply with corresponding loads, the requirement for drain voltage can be satisfied.

For example, if one assumes that the fundamental and second harmonic drain currents are $I_{\text{drain,1}} \cdot e^{j(2\pi f_1 t + \theta_1)}$, and $I_{\text{drain,2}} \cdot e^{j(2\pi f_2 t + \theta_2)}$, where $f_1$ and $f_2$ represent fundamental and second harmonic frequencies respectively, and fundamental load is a pure resistor while the second harmonic load is a pure reactance. Then the requirements for $\theta_1$ and $\theta_2$ are calculated as follow:

$$V_{\text{drain,1}} = \text{Re} \{ I_{\text{drain,1}} \cdot e^{j(2\pi f_1 t + \theta_1)} \cdot R_1 \} = I_{\text{drain,1}} R_1 \cos[2\pi f_1 t + \theta_1] \Rightarrow \theta_1 = \pi \quad (5-16)$$

The above expression shows that the phase of the second harmonic gate voltage should be $\pi$ to ensure that the second harmonic component is in phase with the fundamental component.
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and

\[ V_{\text{drain,2}} = R_c a d \left[ I_{\text{drain,2}} e^{i(2\pi f_2 + \theta)} \cdot R_2 e^{i\pi/2} \right] = R_c a d \left[ I_{\text{drain,2}} \cdot R_2 e^{i(2\pi f_2 + \theta + \pi/2)} \right] \]

\[ = I_{\text{drain,2}} \cdot R_2 \cos(2\pi f_2 t + \theta + \pi/2) \Rightarrow \theta_2 = -\pi/2 \] (5-17)

The calculation requires that there must be a \( \pi/2 \) phase difference between the fundamental and second harmonic drain current. This obviously cannot be achieved by only using the fundamental gate voltage to control the drain current. As we mentioned before, second harmonic gate voltage is introduced to combine with the fundamental gate voltage to control the drain current.

The next task is to find out the required optimal fundamental and harmonic loads values that are functions of conduction angle \( \alpha \) and \( V_{\text{drain,2}}/V_{\text{drain,1}} \) ratio \( \beta \). Here, for simplicity, it is assumed that the drain current values at different harmonic frequencies are generated only by the fundamental gate voltage, while the second harmonic gate voltage is used to control the correct phase relationship between drain components. Therefore, the drain current values are the same as those shown in Figure 5-17.

According to the previous analysis, for any fixed \( \beta \), to make the drain voltage swing within its full range, the \( V_{\text{drain,DC}}, V_{\text{drain,1}} \) and \( V_{\text{drain,2}} \) values are fixed. At the same time, at any fixed conduction angle \( \alpha \), the values for \( I_{\text{drain,dc}}, I_{\text{drain,1}}, I_{\text{drain,2}} \) are also fixed. Therefore, the requirements for fundamental load and second harmonic termination \( R_1 \) and \( R_2 \) can be easily calculated as:

\[ R_1 = V_{\text{drain,1}}/I_{\text{drain,1}} \text{ and } Z_2 = V_{\text{drain,2}}/I_{\text{drain,2}} \] (5-18)

Thus, \( R_1 \) as the function of \( \alpha, \beta, R_1(\alpha, \beta) \) can be simulated according to equation (5-12). Figure 5.18 shows the simulation results. Here, \( \alpha \) ranges from \( 2\pi \) to 0 which means the device’s gate is biased from normal Class-A condition to Class-C, while \( \beta \) varies from 0 to 1. Here, when \( R_1 = 1 \) represents the load requirement for the Class-A biasing condition. All the calculations in the simulations are normalised to this value.
It can be seen from Figure 5.18 that in the region of $\alpha \in [0, 2\pi]$ and $\beta \in [0, 0.375]$, the drop of $\alpha$ and the increase of $\beta$ will both result in an increase of $R_f$. This means that when the amplifier's design is changed from Class-A to second peaking Class-F, a higher fundamental load is required by the design.

In the same way, the required load impedance for the second harmonic $R_2(\alpha, \beta)$ can be calculated and the results are shown in Figure 5.19 and Table 5-3.
Table 5-3 $R_2$ as the function of $\alpha$ and $\beta$

<table>
<thead>
<tr>
<th>$\alpha/\beta$</th>
<th>0</th>
<th>0.1</th>
<th>0.2</th>
<th>0.3</th>
<th>0.4</th>
<th>0.5</th>
<th>0.6</th>
<th>0.7</th>
<th>0.8</th>
<th>0.9</th>
<th>1</th>
</tr>
</thead>
<tbody>
<tr>
<td>1.8$\pi$</td>
<td>0</td>
<td>15.58</td>
<td>31.16</td>
<td>46.35</td>
<td>58.99</td>
<td>69.24</td>
<td>77.62</td>
<td>84.58</td>
<td>90.43</td>
<td>95.42</td>
<td>99.70</td>
</tr>
<tr>
<td>1.6$\pi$</td>
<td>0</td>
<td>2.09</td>
<td>4.20</td>
<td>6.24</td>
<td>7.94</td>
<td>9.32</td>
<td>10.46</td>
<td>11.39</td>
<td>12.18</td>
<td>12.85</td>
<td>13.43</td>
</tr>
<tr>
<td>1.4$\pi$</td>
<td>0</td>
<td>0.70</td>
<td>1.41</td>
<td>2.10</td>
<td>2.67</td>
<td>3.14</td>
<td>3.52</td>
<td>3.83</td>
<td>4.10</td>
<td>4.32</td>
<td>4.52</td>
</tr>
<tr>
<td>1.2$\pi$</td>
<td>0</td>
<td>0.35</td>
<td>0.71</td>
<td>1.06</td>
<td>1.35</td>
<td>1.59</td>
<td>1.78</td>
<td>1.94</td>
<td>2.08</td>
<td>2.19</td>
<td>2.29</td>
</tr>
<tr>
<td>$\pi$</td>
<td>0</td>
<td>0.23</td>
<td>0.47</td>
<td>0.70</td>
<td>0.89</td>
<td>1.04</td>
<td>1.17</td>
<td>1.27</td>
<td>1.36</td>
<td>1.44</td>
<td>1.50</td>
</tr>
<tr>
<td>0.8$\pi$</td>
<td>0</td>
<td>0.191</td>
<td>0.38</td>
<td>0.56</td>
<td>0.71</td>
<td>0.84</td>
<td>0.94</td>
<td>1.02</td>
<td>1.09</td>
<td>1.15</td>
<td>1.21</td>
</tr>
<tr>
<td>0.6$\pi$</td>
<td>0</td>
<td>0.18</td>
<td>0.36</td>
<td>0.54</td>
<td>0.69</td>
<td>0.81</td>
<td>0.91</td>
<td>0.99</td>
<td>1.06</td>
<td>1.12</td>
<td>1.17</td>
</tr>
<tr>
<td>0.4$\pi$</td>
<td>0</td>
<td>0.22</td>
<td>0.44</td>
<td>0.65</td>
<td>0.83</td>
<td>0.98</td>
<td>1.10</td>
<td>1.20</td>
<td>1.28</td>
<td>1.36</td>
<td>1.41</td>
</tr>
<tr>
<td>0.2$\pi$</td>
<td>0</td>
<td>0.39</td>
<td>0.78</td>
<td>1.16</td>
<td>1.48</td>
<td>1.73</td>
<td>1.95</td>
<td>2.12</td>
<td>2.27</td>
<td>2.39</td>
<td>2.50</td>
</tr>
</tbody>
</table>

Simulation results show that when the device’s gate is biased around the highest efficiency point ($\beta = 0.375$) of Class-F second harmonic peaking condition, the requirement on $R_2$ changes dynamically – from 50 times of $R_1$ to only half of $R_1$.

We must bear in mind, however, that all the above calculations are based on a basic theoretical understanding. Due to the introduction of the second harmonic gate voltage, to get the approximate proper drain current phase relationship between the fundamental and second harmonic frequency, the value ratio between current components are no longer the values shown in Figure 5.17. Therefore, it needs to be carefully tuned in the real design to achieve the maximum performance point.

### 5.3.12 Simulation Results

After finished all the theoretical analysis, the next step is to carry out a Class-F second harmonic peaking amplifier design simulation in COMMS. The simulation procedures are:

1) Generate device’s large-signal S-parameters under specific input power and bias conditions.

2) Set-up and optimize the input/output match network, according to the large-signal S-parameters.

3) Control the gate voltage to manipulate the proper phase relationship between drain harmonic current components that is required by the design concept.

4) Set proper harmonic terminations at the output port of the output match network, to satisfy the shaping requirement.

5) Evaluate the amplifier performance in the harmonic test bench, and re-check the input/output match condition by testing the input/output return loss.

The model in this simulation is the MGF0906B STATZ large-signal model, that was set-up in the previous section. The complete amplifier schematic diagram decided by the simulation is shown in Figure 5.20. In the design, device’s drain is biased to a low value $V_{ds} = 5V$, and TL5 in the input match network is used to control the gate voltage so that can get the proper drain current that is required for the amplifier’s operation.
Figure 5.20 Schematic diagram for second harmonic peaking simulation

The simulation results for the amplifier's DC-RF efficiency, power added efficiency and gain are shown in Figure 5.21. It can be seen that within the 50 MHz sweep bandwidth, the maximum DC-RF efficiency is over 80%, and the gain is around 9 dB with only 0.2 dB variation, while the PAE is about 13% lower than the DC-RF efficiency.
Figure 5.21 Simulation results for DCRF, PAE and gain

Figure 5.22 shows the simulated drain voltage $v_{\text{drain}}(t)$ for the second harmonic peaking.

Figure 5.22 Drain voltage waveform
Figure 5.22 shows that although the device’s drain is biased at $V_{ds} = 5V$, the maximum drain voltage swings all the way up to 20V, which clearly demonstrates the characteristic of the Class-F second harmonic peaking amplifier.

The amplifier’s performance of DC-RF efficiency, PAE and output power vs. input power are shown in Figure 5.23. The --V-- curve represents output power. The 1dB compression occurs at $P_{in}=25$dBm. The maximum DC-RF efficiency is over 80%, and maximum PAE is about 68%. They both also occur at the 1dB compression point.

The amplifier’s group delay performance, one of the result that we are most concerned with, is plotted in Figure 5.24. It can be seen that the group delay varies from 0.94ps to 1.15ps, the maximum variation is about 10%. This is compared with less than 0.5 ns distortion that is required by the analysis in Chapter 5, and is quite a small value. However, it is generally accepted that the current simulation package does not have the ability to accurately predict the absolute value of the group delay performance but rather study the large-signal model’s individual parameter on group delay performance. We will discuss this in detail in Section 5.33.
In summary, the second harmonic peaking Class-F amplifier requires that the fundamental drain voltage and second harmonic drain voltage are 180° out-of-phase. To achieve this target, a proper input match network must be adopted to control the gate’s fundamental and second harmonic voltages so that to insure a correct phase relationship between drain current components. Simulation show that the amplifier can provide up to 80% DC-RF efficiency and relatively high gain. However, the most significant advantage of this amplifier is that it only requires a very low DC drain biasing voltage, which makes it very suitable for mobile handset applications.

5.32 Third harmonic peaking

The third harmonic peaking Class-F amplifier is somehow the most popular and widely accepted Class-F design method. Indicated by its name, the design concept is to make use of the fundamental and third harmonic drain voltage components to shape the drain voltage to the wanted waveform. The key design of the amplifier is the output impedance matching network which aims to achieve proper fundamental and harmonics voltage and current waveform shaping. The objective of this matching network is to prevent the higher harmonics from reaching the load while allowing their presence at the input to achieve the desired voltage and current shapes. The basic schematic diagram of third harmonic peaking Class-F amplifier is shown in Figure 5.25.
5.32.1 Theoretical Analysis

Ideally, voltage and current shaping can be accomplished with any combination of harmonics. However, when considering both the power efficiency and power output capability, it is better to use odd harmonics to force one waveform to approximate a square-wave and even harmonics to force the other to approximate a half-sinusoid, or vice versa. Quite often, only the second and third harmonic frequencies are considered in a real design. In Raab’s paper, the drain voltage and current are expressed as:

\[ V_{\text{drain}}(\theta) = V_{\text{drain,dc}} + V_{\text{drain,1}} \sin \theta + V_{\text{drain,3}} \sin(3\theta) + V_{\text{drain,5}} \sin(5\theta) + \ldots \] (5-19)

and

\[ I_{\text{drain}}(\theta) = I_{\text{drain,dc}} - I_{\text{drain,1}} \sin \theta - I_{\text{drain,3}} \cos(2\theta) - I_{\text{drain,4}} \cos(4\theta) + \ldots \] (5-20)

where the subscript denotes the order of harmonics and \( \theta = \omega t \). In Raab’s paper, he defines the basic drain waveform parameters \( \gamma_v, \gamma_i, \delta_v, \delta_i \) relating the DC component with the fundamental component and the peak of harmonic waveform, expressed as:

\[ V_{\text{drain,1}} = \gamma_v V_{\text{drain,dc}} \]
\[ V_{\text{Dmax}} = \delta_v V_{\text{drain,dc}} \]
\[ I_{\text{drain,1}} = \gamma_i I_{\text{drain,dc}} \]
\[ i_{\text{Dmax}} = \delta_i I_{\text{drain,dc}} \] (5-21)

the basic DC to RF efficiency can then be calculated by:
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\[ \eta = \frac{P_1}{P_{dc}} = \frac{\left( V_{drain,1}^2 / 2R_L \right)}{V_{drain,dc}^2 \gamma_1} = \frac{\gamma_1^2 V_{drain,dc}^2 / 2R_L}{\gamma_1 V_{drain,dc}^2 / \gamma_1 R_L} = \frac{\gamma_1}{2} \]  

(5-22)

Furthermore, Raab demonstrated that for the maximum flat drain voltage waveform condition, if one only considers the second and third harmonic frequencies, the maximum DC to RF efficiency occurs when \( \gamma_1 = 9/8, \gamma_2 = 4/3 \). For this case,

\[ \eta = 75\% \]

\[ V_{drain,3} = (1/8)V_{drain,dc} \]

\[ I_{drain,3} = (1/4)I_{drain,dc} \]

The drain voltage and current waveform \( V_{ds}, I_{ds} \) are shown in Figure 5.26(a) (b).

![Vds and Ids waveforms](image)

Figure 5.26 Desired drain voltage and current waveforms for third harmonic peaking Class-F amplifier design when only considering second and third order harmonic frequencies. (assumes operating frequency = 100Hz)

Obviously, the more harmonic frequencies that be considered, the higher the efficiency that can be expected. In the ideal case, if all the harmonic frequencies are considered, the efficiency can be increased to 100%.

The key issue behind this concept is that if the drain voltage has a maximum permissible amplitude swing of \( V_{max\_swing} \) (usually, this is decided by the transistor's breakdown voltage and knee voltage), the presence of a third harmonic component will allow the fundamental component amplitude to be increased from \( V_{drain,1} = V_{max\_swing} \) to a higher value \( V_{drain,1} = V_{max} / \gamma_1 \). This corresponds directly to an increase in fundamental power, given by the factor \( \gamma_1^2 \), if one assumes the fundamental power and the DC supply are unaffected by the generation of a small amount of third harmonic voltage.
Following this idea, it is very easy to demonstrate that $\gamma_v = 9/8$ is actually not the point that can give the maximum fundamental power increase. It is only the point that will give the maximum flatness in the drain voltage waveform. For $\gamma_v > 9/8$, the waveform starts to overshoot and will have a double peak. In equation (5-9), if only the third harmonic voltage component is considered, the drain voltage can then be expressed as:

$$v_{\text{drain}}(\theta) = v_{\text{drain,dc}} + v_{\text{drain,1}} \sin(\theta) + v_{\text{drain,3}} \sin(3\theta)$$  \hspace{1cm} (5-23)

Calculations show the maximum $\gamma_v$ that can keep the peak drain voltage value lower than $V_{\text{max,swing}}$ is $\gamma_v = 2/\sqrt{3}$ and $V_{\text{drain,3}} = 0.1844V_{\text{drain,1}}$, correspondingly. Figure 5.28 shows the drain voltage waveform under this condition. If only the second harmonic drain current component is considered, and the drain current coefficient of fundamental to second harmonic is still $\gamma_i = 4/3$, the drain DC-RF efficiency becomes $\eta = \frac{1}{2} \cdot \frac{2}{\sqrt{3}} \cdot \frac{4}{3} = 77\%$. If an ideal half sinusoid wave current is considered, which means all the even harmonics are shorted to ground properly, then the DC-RF efficiency will increase to 88.4%.

![Figure 5.27 Drain voltage waveform for $\gamma_v = 2/\sqrt{3}$ and $V_3 = 0.1844V_1$.](image)

Although the double peak condition can provide a slightly higher efficiency than the maximum flat waveform case, it is believed [Cripps S. C. 1999] that the maximum flat case is more practically realizable.

The simulation shows that if third harmonic peaking Class-F and Class-A amplifier both operate at full swing conditions, the maximum output power of third harmonic peaking Class-F amplifier is about 1.2 dB higher than what we can get from Class-A amplifier. This directly indicates a good improvement in the amplifier's gain. Recall that it is well known that for Class-B amplifier, to get the same output power, the amplifier need to be driven 6 dB higher. That is one of the most significant advantages of the third harmonic
peaking Class-F amplifier in that it can provide the high DC-RF efficiency as well as the high output gain.

During the design process it is critical to keep the correct phase relationship between the fundamental and third harmonic drain voltages. Figure 5.28 show these two drain voltages in the out-of-phase condition. It can be seen that the incorrect combination of \( V_{\text{drain,1}} \) and \( V_{\text{drain,3}} \) will result in a drain voltage that exceeds the \( V_{\text{max,swing}} \) and, therefore, this will lead to voltage clipping.

![Figure 5.28 Out-of-phase condition (voltage normalized to \( V_{\text{max}} \) )](image)

In Raab's paper, to achieve the design target, he suggested that the third harmonic load should be set to infinite, while the second harmonic load should be kept to zero. However, if we still assume that the device is a voltage-control current generator, immediately we can see that these requirements for harmonic loads are not correct. The following is the detail analysis.

Recall in Figure 5.17, when the conduction angle \( \alpha \) is larger than \( \pi \), the second and fundamental current component are in-phase, while the third and fundamental current components are out-of-phase. Therefore, to satisfy the Class-F design requirements on drain voltage in equation (5-19), the amplifier can not be biased from Class-B to Class-C region. In other word, the conduction angle must be set between \([\pi, 2\pi]\). At the same time, the third harmonic impedance appears at the device’s drain terminal will be a pure resistor. In the following discussion, we will concentrate on studying the case when the conduction angle is within the \([\pi, 2\pi]\) region.

To maintain the drain voltage to operate within the maximum swing, \( V_{\text{max,swing}} = (V_{\text{max}} - V_t)/2 \), while still considering the maximum flattened waveform case, the fundamental drain voltage component becomes:

\[
V_{\text{drain,1}} = \frac{9}{8} V_{\text{max,swing}} = \frac{9}{8} \frac{V_{\text{max}} - V_t}{2} \tag{5-24}
\]
Therefore, the required fundamental load is:

\[
R_{L,1}(\alpha) = \frac{V_{\text{drain},1}}{I_{\text{drain},1}(\alpha)} \tag{5-25}
\]

The simulation result is shown in Figure 5.29. Here, the resistor is normalized to,

\[
R_{L,A} = \frac{(V_{\text{max}} - V_{\text{th}})/2}{I_{\text{max}}/2},
\]

which is the optimum load for ClassA biasing.

With a third harmonic peaking Class-F amplifier, the maximum swing of the fundamental drain voltage is larger than \((V_{\text{max}} - V_{\text{th}})/2\), i.e., the maximum swing voltage for the tuned load condition. While the drain currents are the same for these two cases, therefore, the fundamental load required by third harmonic peaking Class-F is larger than the tuned load by a ratio of 9/8.

The ideal third harmonic load becomes:

\[
R_{L,3} = \frac{V_{\text{drain},3}}{I_{\text{drain},3}(\alpha)} = \frac{1}{9} \cdot \frac{V_{\text{drain},1}}{I_{\text{drain},1}(\alpha)} \tag{5-26}
\]

Figure 5.30 shows the impedance ratio between \(R_{L,1}\) and \(R_{L,3}\) as a function of conduction angle \(\alpha\) within the range \(\alpha \in (2\pi, \pi)\).
The results in Figure 5.30 shows clearly that Raab’s assumption is only valid at the Class-B biasing point where the $R_{load,3}$ is required to be infinite. The reason is $I_{\text{drain,3}}$ is only zero at the $\alpha = \pi$ point.

The DC power consumption, RF output power at the fundamental and third harmonic frequencies, as a functions of conduction angle $\alpha$, are given by:

$$P_{\text{dc}}(\alpha) = V_{\text{drain,dc}} \cdot I_{\text{dc}}(\alpha)$$  \hspace{1cm} (5-27)

$$P_1(\alpha) = \frac{V_{\text{drain,1}}(\alpha)}{\sqrt{2}} \cdot \frac{I_{\text{drain,1}}(\alpha)}{\sqrt{2}} = V_{\text{drain,1}}(\alpha) \cdot I_{\text{drain,1}}(\alpha)/2$$  \hspace{1cm} (5-28)

$$P_3(\alpha) = \frac{V_{\text{drain,3}}(\alpha) \cdot I_{\text{drain,3}}(\alpha)}{2}$$  \hspace{1cm} (5-29)

The amplifier’s DC-RF efficiency is also a function of conduction angle and is given by:

$$\text{Efficiency}(\alpha) = \frac{P_1(\alpha)}{P_{\text{dc}}(\alpha)}$$  \hspace{1cm} (5-30)

Figure 5.31 shows the efficiency when the amplifier operates within the area where conduction angle changes from $[2\pi, \pi)$. 

---

Figure 5.30 Impedance ration between fundamental and third harmonic
Figure 5.31 efficiency vs. conduction angle for third harmonic peaking Class-F and tuned load amplifier

Figure 5.31 shows the comparison between the conventional tuned load Class-AB amplifier and the third harmonic peaking Class-F design. The Class-F design method can provide an ~ 7% increase in DC-RF efficiency.

Because there is always a third harmonic power in the amplifier output, it is important to understand the power ratio between fundamental and third harmonic. The simulation results for \( \log_{10}\left(\frac{P_1}{P_3}\right) \) as the function of conduction angle \( \alpha \), are plotted in Figure 5.32.

Figure 5.32 shows that when the amplifier is biased from Class-AB to deep Class-AB region, the output power of third harmonic frequency is about 25dB lower than the
fundamental output power. For most of the applications, this is low enough to cause any serious trouble.

**Overdriven Case**

It is well known that deliberately overdriving the Class-A amplifier may increase the amplifier power efficiency. So what will happen if we overdrive a Class-F amplifier?

For overdriven condition, the drain current can be expressed as:

\[
I_{\text{drain}}(\theta) = \begin{cases} 
\frac{I_{\text{max}}}{(1 - \cos(\alpha/2))} \left[ \cos \theta - \cos\left(\frac{\alpha}{2}\right) \right] & -\alpha/2 < \theta < -\gamma/2, \quad \gamma/2 < \theta < \alpha/2 \\
0 & -\pi < \theta < -\alpha/2, \quad \alpha/2 < \theta < \pi \\
I_{\text{max}} & -\gamma/2 < \theta < \gamma/2
\end{cases}
\] (5-31)

Here, \(\gamma\) represents the degree of overdrive, \(\gamma = \cos^{-1}\left(\frac{I_{\text{max}}}{I_{\text{rms}}}\right)\), \(I_{\text{max}}\) is the peak amplitude of the output current. Figure 5.33 shows the drain current when the amplifier is overdriven by 0.5 dB, 1 dB and 2 dB.

![Figure 5.33 Time domain overdrive plot](image)

Using Fourier analysis on the drain current, Figure 5.34 shows the drain current’s DC, fundamental and higher order harmonic components, when the amplifier is overdriven by 0.5 dB, 1 dB and 2 dB. We can see that the results of the overdrive case are to extend the third and fundamental out-of-phase region be larger than \(\pi\). The larger the overdrive, the wider this region.
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The amplifier's DC-RF efficiency under different overdrive conditions is shown in Figure 5-35. The simulation results show when the amplifier is biased between the possible operating region. The higher the overdrive, the lower the efficiency. However, the maximum DC-RF efficiency of 1dB overdrive is 90.1%, and the possible operating region is $[2\pi, 0.92\pi]$, while for 2dB overdrive the efficiency is 90.8% and the operating region is $[2\pi, 0.88\pi]$.

5.322 Third Harmonic Peaking Class -F Amplifier Simulation Results

This is the same as the procedures for the second harmonic peaking case. The third harmonic peaking Class-F amplifier can also be simulated as shown in Figure 5.36.
The simulated amplifier performance of output power & gain vs. input power, drain voltage waveform and DC-RF efficiency, PAE & gain vs. sweep bandwidth are plotted in Figures 5.37 to 5.39.
The time domain drain voltage waveform in Figure 5.38 shows that the ratio between third harmonic drain voltage and fundamental drain voltage is higher than the recommended value of maximum flattened case. The reason is because there is an interaction between device's drain and gate terminals, so in the simulation the device is
not an ideal voltage-controlled current generator. Therefore, it is very difficult to control the phase of the harmonic drain voltage components, as required by the design concept. However, the simulation results of the amplifier’s performance of DC-RF efficiency, PAE and gain are still quite satisfying, as shown by Figure 5.39.

Finally, the amplifier group delay performance is also simulated, as shown in Figure 5.40. It can be seen that it varies from 5.2 ns down to 0 ns, which is a very poor performance.
In summary, the third harmonic peaking amplifier makes use of the third harmonic drain voltage component to flatten the drain voltage waveform. To achieve the target, it requires that the drain current's fundamental and third harmonic components are 180° out-of-phase, when the conduction angle $\alpha$ is in the range of $(2\pi, \pi)$. When the device is viewed as an ideal voltage-control current generator, this requirement can easily be satisfied. Within the operating bandwidth, the simulated maximum DC-RF efficiency is around 80% with a gain of 8.8 dB. The most significant advantage of the third harmonic peaking Class-F amplifier is that it can provide high efficiency as well as the high output gain. This makes it very attractive for microsatellite payload design.

5.33 Large-Signal Parameter Influence on Group Delay Performance

The final question that the simulation will try to answer is what kind of influence the individual parameter in the large-signal model will have on amplifier group delay performance? There are 32 parameters in the STATZ large signal model, and the functions of these parameters can basically be divided into three types:

- DC parameters, such as $\alpha$, $\beta$, $\theta$, VTO, $\lambda$
- AC parameters, such as $C_{gse}$, $C_{gdo}$ and $C_{ds}$
- Temperature characteristics, such as TNOM, VTOTC, BETATCE

Among these $C_{gse}$, $C_{gdo}$ and $C_{ds}$ are the three most important parameters in deciding the device's large-signal S-parameters. Therefore, the simulation will concentrate on these three parameters to analysis the group delay performance by comparing parameter influence on the final result one by one. Here, the analysis will only consider the group delay performance of third harmonic peaking Class-F case, as its group delay has more distortion than that for second harmonic peaking.

5.33.1 Simulation Results

The simulation procedure will change the original $C_{gse}$, $C_{gdo}$ and $C_{ds}$ values by ±5% and ±10%, individually, and observe the group delay performance according to the variation. Figure 5.41 shows the simulation results for $C_{gse}$. Here, $C_{gse} = 30 \text{ pF}$ is the original value that is decided by the large-signal model. The simulation varies $C_{gse}$ value from 27 pF to 33 pF, with step of 1.5 pF.

The curve shows that the change of $C_{gse}$ introduces an obvious group delay variation effect within the amplifier's operating bandwidth. It also results in a change of the peak group delay value. When $C_{gse}$ decreases by 10%, the maximum group delay value increases by about 0.7 ns, while when $C_{gse}$ increases by 10%, its maximum value drops by approximately 1 ns. The change is relative high, when considering that the amplifier mean group delay distortion is around 2.5 ns.
Note, in Figure 5.41, that when $C_{gs0}$ decreases from 30 pF, the group delay falls down to below zero at a small region. This is due to the imperfect response of harmonic balance software that cannot be obtained in the real test-bench measurement. Nevertheless, the simulation results still give a guide of the $C_{gs0}$ influence on amplifier group delay.

In the same way, the group delay performance as a function of $C_{gd0}$ are also plotted, as shown in Figure 5.42. $C_{gd0}$ changes from 0.63 pF to 0.77 pF with 0.035 pF steps. The maximum group delay change is around 0.7 ns, which occurs when $C_{gd0}$ drops by 10%. Compared with the influence of $C_{gs0}$, the $C_{gd0}$ effect also introduces less variation within the operating bandwidth.
Finally, the influence of $C_{ds}$ is plotted, as shown in Figure 5.43. Here, $C_{ds}$ changes from $C_{ds} = 1.35\, pF$ to $C_{ds} = 1.65\, pF$, with $0.075\, pF$ step. Compared with the results from $C_{gdo}$ and $C_{gso}$, the change of $C_{ds}$ will have much less effect. The maximum group delay change is less than $0.5\, ns$, and the change in the value of $C_{ds}$ also has much less variation effects within the bandwidth.
5.3.3.2 Result Analysis

A detailed understanding of the simulation results are out of the scope of this study. However, by looking at on the device's large-signal model, it can be seen that the above simulation results are not surprising. First, in the large signal model, \( C_{ds} \) is a constant, while the junction capacitance \( C_{go}, C_{gd} \) are dictated by \( C_{gso}, C_{gdo} \) if the diode depletion capacitance model is used. In Statz's paper \( C_{go} \) and \( C_{gd} \) are expressed as:

\[
C_{gs} = \frac{C_{gso}}{\sqrt{1 - \frac{V_{new}}{V_B}}} \left[ 1 + \frac{V_{eff1} - V_T}{\sqrt{(V_{eff1} - V_T)^2 + \delta^2}} \right] \times \frac{1}{2} \left[ 1 + \frac{V_{gs} - V_T}{\sqrt{(V_{gs} - V_T)^2 + \left(\frac{1}{\alpha}\right)^2}} \right]
\] (5-32)

\[
+ C_{gdo} \left[ 1 - \frac{V_{gs} - V_{gd}}{\sqrt{(V_{gs} - V_{gd})^2 + \left(\frac{1}{\alpha}\right)^2}} \right]
\]

\[
C_{gd} = \frac{C_{gdo}}{\sqrt{1 - \frac{V_{new}}{V_B}}} \left[ 1 + \frac{V_{eff1} - V_T}{\sqrt{(V_{eff1} - V_T)^2 + \delta^2}} \right] \times \frac{1}{2} \left[ 1 - \frac{V_{gs} - V_{gd}}{\sqrt{(V_{gs} - V_{gd})^2 + \left(\frac{1}{\alpha}\right)^2}} \right]
\] (5-33)

\[
+ C_{gdo} \left[ 1 + \frac{V_{gs} - V_{gd}}{\sqrt{(V_{gs} - V_{gd})^2 + \left(\frac{1}{\alpha}\right)^2}} \right]
\]

where \( V_{new} = \frac{1}{2} \left[ V_{eff1} + V_T + \sqrt{(V_{eff1} - V_T)^2 + \delta^2} \right] \)

and \( V_{eff1} = \frac{1}{2} \left[ V_{gs} + V_{gd} + \sqrt{(V_{gs} - V_{gd})^2 + \delta^2} \right] \)

Research [Walker L. B. 1993] shows the variation of \( C_{gs} \) and \( C_{gd} \) as the function of \( V_{gs} \) and \( V_{ds} \) are different, as shown in Figure 5.44.
Chapter 5. Highly Efficient Class-F Power Amplifier - Design and Implementation

It is known that in the variations of $V_{gs}$ and $V_{ds}$, $C_{gs}$ accounts for the next major nonlinearity after $g_m$. Also, the variation of $C_{gs}$ with $V_{gs}$ causes the reactance variation which causes AM-PM conversion in amplifiers, and FM noise in oscillators. Figure 5.44 also shows that the $V_{gs}$ and $V_{ds}$ dependence of $C_{gs}$ is slight, except when $V_{ds}$ is lower than 1V, which means there is only a small bias variation of $C_{gs}$ in the active region of transistor operation. Therefore, as the simulation results predict, $C_{gs}$ is the main reason for the group delay performance variation, which reflects the device's reactance phase change.

5.34 Conclusion

In this section, the theory for Class-F second and third harmonic peaking amplifiers design have been analysed. The amplifiers have been simulated by using the large signal MGF0906B model that has been set up in the previous section. The simulation results are good. Both amplifiers can achieve up to 80% DC-RF efficiency, with around 9dB gain over the whole swept bandwidth. The most significant advantage of second harmonic peaking Class-F amplifier is that it only requires very low DC supply and the merit of the third harmonic peaking Class-F amplifier is that it can provide high DC-RF efficiency as well as high output gain.

The amplifier's group delay performance has also been simulated. It can be seen that due to the nonlinear characters of the device, the group delay is not a constant, as expected by the ideal case. Instead, it has a large dynamic variation. According to the analysis in Chapter 4, this variation will introduce an estimation error in the SWH predicting. Therefore, some methods must be adopted to correct this group delay variation. A novel method to correct for group delay variation will be given in Chapter 6.
5.4 Practical Circuit Implementation

The simulated power amplifier from section 5.32 has been given. Figure 5.45 shows the complete circuit diagram, when all the transmission line discontinuities have been considered.

Figure 5.45 Complete circuit schematic diagram

Figure 5.46 shows the PCB layout. The substrate chose for the amplifier is TLX-8-0300 whose ε = 2.55 and having a thickness of 0.76mm. The advantage of this substrate is that it has a thick metal carrier directly underneath. This enables the heat generated by amplifier be transmitted very effectively.
The test bench set-up is shown in Figure 5.47. A wide-band drive amplifier is connected with the main amplifier. The amplifier performance is tested using the HP network analyser HP 8753D.
Figure 5.48 shows the measured $S_{21}$ performance of the drive amplifier connected with the main amplifier. The results show that the amplifier's optimum range is a bit lower than the design frequency. However, within the 50 MHz bandwidth, the amplifier output is very flat.

Figure 5.48 Measured amplifier $S[21]$ performance

Figure 5.49, 50 are the measured amplifier Efficiency, Pout & Gain vs. Input power. The diagrams show that the maximum PAE is 50%. The results show that the amplifier's 1 dB compression point is around 27 dBm, which is the same as its Class-A operation that
suggested by the device handbook. At 1 dB compression point, amplifier gain of the power amplifier is around 9 dB. This demonstrates again the advantage of Class-F amplifier over Class-B, C in it need not sacrifice gain to get high efficiency.

**DCRF & PAE vs. Input power**

![Graph showing DCRF and PAE vs. Input power](image)

*Figure 5.49 Measured efficiency vs. Input power (f = 2GHz)*

**Pout & Gain vs. Pin**

![Graph showing Pout and Gain vs. Pin](image)

*Figure 5.50 Measured output power & gain vs. Input power*

It should be pointed out that although the measured PAE result is lower than the simulation results, it still demonstrates that the device's large-signal works well in the nonlinear power amplifier design. The possible reason for the decrease of efficiency is the large signal model still could not completely correctly describe the device's large-
signal S-parameters, and therefore generated the mismatch between the drive amplifier and main amplifier.

The measured amplifier's group delay performance is shown in Figure 5.51. The diagram shows that the group delay has a sine wave function shape distortion as expected by the simulations.

![Figure 5.51 Measured amplifier group delay performance](image)

### 5.5 Conclusions

In this chapter, a highly efficient S-band Class-F amplifier has been simulated and implemented. The study starts with the demonstration of the feasibility of using the large-signal model to design nonlinear power amplifier. It is followed by providing a detailed theoretical analysis for two very different types of Class-F cases: second harmonic and third harmonic peaking modes. The characteristics of theoretical results of these two types of amplifier have been demonstrated by the nonlinear amplifier simulation in the COMMS's harmonic test bench. Finally, the amplifier has been implemented and characterised in hardware on a real test bench.

It is believed that this study is the most detailed and comprehensive design guide for the two different types Class-F amplifier yet undertaken. It is also believed that it is the first time that a Class-F power amplifier has been comprehensively studied, compared and simulated by using a real self-developed device large signal model.

The conclusions that can be drawn from the above work are:

- A device large signal model can be deduced with the knowledge for device's small signal S parameters and DC measurements.
Chapter 5. Highly Efficient Class-F Power Amplifier - Design and Implementation

- The most significant advantage of the Class-F amplifier is that it can provide high efficiency without having to reduce the gain.

- Second harmonic peaking Class-F amplifiers require that the fundamental drain voltage and second harmonic drain voltage are out-of-phase. Therefore a proper input matching network must be adopted to control the gate’s fundamental and second harmonic voltages so that to insure a correct phase relationship between drain current components. The significant advantage of this amplifier is that it only needs very low drain DC voltage biasing, which makes it very suitable for mobile phone handset application.

- Third harmonic peaking amplifier makes use the third harmonic drain voltage component to flatten the drain voltage waveform. It is required that the drain current fundamental and third harmonic components are out-of-phase, for an ideal voltage control current generator, it is easier to achieve than the second harmonic peaking case.

- Among $C_{gs}$, $C_{gd}$ and $C_{ds}$, $C_{gs}$ is the main reason that accounts for the amplifier's group delay performance variation, which by nature reflects the device’s reactance phase change.

- The test bench measurements show that the large signal-model can achieve relatively accurate results in non-linear power amplifier design and simulation.

- Highly efficient amplifiers cannot achieve good group delay performance simultaneously with high DC-RF transfer efficiency. Therefore, an extra circuit is required to correct for the phase distortion introduced.
Chapter 6. Feedback Linearization of Highly Efficient Power Amplifiers

The purpose of this chapter is to design a linearization circuit to correct the group delay error introduced by the highly efficient power amplifier that has been designed in the previous chapter. The design goal, however, is different from the usual meaning of linearization technique. Here the goal is to concentrate on trying to flatten the group delay within the whole sweep bandwidth of the chirp signal. This means not caring too much about the absolute phase error but rather the changing rate of this phase error. As determined from the simulation analysis in Chapter 4, the group delay variation within the chirp signal’s swept bandwidth will introduce in the estimation error for the altimeter on-board fast SWH calculation.

This chapter starts with a detailed literature survey of linearization techniques, which includes the introduction of the main design ideas of the most popular linearization techniques – predistortion, feedback, feedforward, LINC, etc. After analysis of the specific requirements for this project, adaptive feedback linearization was finally chosen. This adaptive feedback circuit is a very simple but effective linearization technique. Its compact size and low power consumption make it very suitable for the proposed microsatellite radar payload. The simulation results, as well as the implementation results of each subsystem, are given step-by-step in this chapter.

6.1. Literature Survey of Linearization Techniques

Linearization technology has recently become a very popular technique at microwave frequencies, due to the high demand from the mobile industry for high efficiency, and high linearity power amplifier. Where phase sensitive modulation methods, such as π/4-DQPSK and QAM are used. The techniques are employed either at baseband or in the microwave band to correct for added AM/PM distortion, or introduce the out-of-band suppression, as well as to maintain high power added efficiency.

Linearization techniques have long been used in satellite communications to improve TWTA's linearity. A diode or FET linearizer is a common choice for the designer. Recently, due to the requirements of multi-carrier communications on low leakage adjacent channels, low DC power consumption, high efficiency and high linearity, techniques have been developed for a much wider application region, with more design choice. There are several well-known existing linearization methods, such as feedforward, feedback, predistortion, and LINC.

6.11. Predistortion (PD)

The commonly used baseband predistortion method is to add a predistorting signal to the baseband modulating signal so as to cancel the distortion error. Figure 6.1 shows a basic configuration for baseband predistortion with Cartesian feedback [Cavers J. K. 1990] linearization. With the predistortion technique, as its name indicates, the amplitude and phase non-linearity of the baseband signal are pre-distorted so as to impart an inverse
characteristic of the transmitter. These methods can be achieved either at RF, IF or baseband. Baseband predistortion can be accomplished through using a look-up table. Recently, DSP is often adopted to achieve the adaptive compensation aim. With IF and RF predistortion, diode or non-linear FETs are still the common choice, although DSP sometime also plays a role in the design.

**Baseband Predistortion**

Baseband predistortion has become popular due to the fast development of digital techniques. The commonly used baseband predistortion method is to add a predistorting signal to the baseband modulating signal so as to cancel the distortion error. Figure 6.1 shows a basic configuration for baseband predistortion with Cartesian feedback [Cavers J. K. 1990]. The linearizer creates a predistorted version, $V_d(t)$, of the desired modulation $V_m(t)$, making use of measurement $V_r(t)$ of the actual amplifier output $V_a(t)$ to correct for the amplitude and phase error. Note that the same oscillator is used in the up and down conversion, for coherence, and in some methods, a phase shifter is required for stability. The virtue of this method is that of simple, however, its linearity and its bandwidth are critically dependent on the loop delay.

![Figure 6.1 Generic configurations for adaptive linearization (after Cavers J. K. 1990)](image)

A modified version is to implement in digital baseband, a random access memory (RAM) look-up table (LUT) [Nagata Y. 1989], with an entry for each pre-distorted point in the signal constellation to achieve fast correction with very little memory. Cavers called it mapping predistortion. Figure 6.2 (a) & (b) show the configurations.

![Figure 6.2 (a) Basic configuration](image)
The mapping PD generalises the look up table to provide the predistorted equivalent $V_d(t)$ of any input value $V_m(t)$, therefore mapping the complex plane into itself. It is unrestricted by the order and type of PA nonlinearity (provided it is memoryless). Since this permits pulse shaping prior to predistortion, it is not restricted by modulation format either. However, powerful though it is, the mapping PD has several drawbacks. First it requires a large size look up table. Second it needs a phase shifter in the feedback path for stability in the adaptation update, and convergence is very slow.

Based on this, Cavers [Cavers J. K. 1990] [Wright A. S. et al 1992] proposed a gain base predistorter, shown in Figure 6.3. The authors defined the relationship of input $V_d(t)$ and the output $V_a(t)$ as:

$$\nu_a(t) = \nu_d(t) G\left(\left|\nu_d\right|^2\right)$$

Here, $G(x)$ represents the power amplifier’s AM/AM and AM/PM characteristics.

Note that $G(x_d)$ depends only on the instantaneous power of the input and, therefore, this predistortion method is called the gain based predistorter.
Chapter 6. Feedback Linearization for Highly Efficient Amplifier Payload Design

Compared with mapping PD, the advantages of gain PD are:
- It has a reduced memory requirement
- It reduces convergence time
- It removes the need for a phase shifter or PLL in the feedback path

However, the improvement is at the expense of increased computation, both in real time multiplication of each sample by the complex gain, and in the adaptation update step. This restricts the use in wide-band applications.

**Discussion:** The baseband predistortion technique is a very interesting and flexible idea. However, analysis from Chapter 4 shows that the design efforts should concentrate on the errors that only appears once in the deramping process, while the baseband signal always appears twice in the deramping. Therefore, baseband predistortion is not a candidate for the altimeter application.

**Predistortion with source (emitter) inductance or diodes**

This method is mainly proposed by the power amplifier designers of Mitsubishi Electric Cooperation [Mori K. et al. 1997] [Nakayama M. et al. 1995] [Yamauchi K. et al. 1997] for their L & S-Band linearizer. The basic ideas is based on the observation that the power amplifier displays negative gain and positive phase slope with an increase in input power, while the inductance or diode at the source (emitter) of a FET (BJT) could achieve the opposite performance. Figure 6.4 is the system diagram for this method.
Chapter 6. Feedback Linearization for Highly Efficient Amplifier Payload Design

(b) from [Yamauchi K. et al 1997]

Figure 6.4 Predistortion with source (emitter) inductance or diodes

The advantages of this design are they are simple and exhibit with good performance, as shown in Figure 6.5. The phase deviation can be successfully corrected from 40 degree to less than 10 degree. Accurate large-signal modelling will play an important role in this approach, which is a difficult task.

Figure 6.5 Measured gain compression and phase of the power amplifier with and without the use of the predistortion linearizer, from [Nakayama M. et al. 1995]

**RF & IF predistortion**

Diodes [Cahana D. et al. 1985] [Satoh G. et al. 1983] or non-linear FETs [Drury D. M. et al. 1985] [Kumar M. et al. 1985] are traditional choices for RF & IF predistorter. The amplifiers can be linearized either directly at the transmitted RF frequency or alternatively at their respective intermediate frequencies. Usually, the Schottky diodes are chosen as inter-modulation generators. Varactor diodes are chosen as phase shifters, for setting the appropriate conditions for inter-modulation suppression. The experimental results in Cahana's paper showed an IF predistorted, a Ku-band TWTA linearized at C-band. This is essentially the same as that of the amplifier linearized directly at Ku band.
Chapter 6. Feedback Linearization for Highly Efficient Amplifier Payload Design

The advantage of IF linearization is that one linearizer could be used to predistort the input signals of two or more power amplifiers.

The principle of the RF diode linearizer is to offset the nonlinear gain and phase lag characteristics of an amplifier by producing drive-dependent gain expansion and phase advance in cascade with the amplifier. Figure 6.6 is a schematic diagram of transmission-type predistortion. The output $V_T$ is the vector sum of $V_D$ and $V_P$. The dynamic impedance of the diode is different for low and high drive levels. The desired shape of gain and phase of predistorter can be obtained by carefully choosing $L$ (inductor), $\phi$ and the diode.

![Figure 6.6 Schematic diagram of RF diode predistorter, from [Cahana D. et al. 1985]](image)

For QPSK/TDMA operation, at or beyond saturation, the desirable TWTA characteristics are those of a soft-limiter. That is, linear gain below saturation and constant gain beyond saturation, with constant phase throughout these ranges. Theoretical and experimental results have shown that, beyond saturation, the effect of conventional linearization is usually lost and the transfer characteristics of the linearized amplifier resemble those of the TWTA. However, by adding a soft-limiter before the linearizer, the power output and phase shifter of the linearized TWTA can be held nearly constant for a significant overdrive range. Figure 6.7 shows the basic configuration of the soft-limited type linearizer [Satoh G. et al. 1983]

![Figure 6.7 Basic configuration of the soft-limited type linearizer, from [Satoh G. et al. 1983]](image)
Discussion: The advantages of IF & RF predistortion methods are that they are simple. There is less dc power consumption and they are compact, especially when MMIC techniques are adopted. However, as mentioned in many papers, the predistortion method can not achieve a very good linearity performance, especially over a wide bandwidth. It is recommended, therefore, that predistortion combined with other predistortion techniques that would achieve much better results.

Other methods

Another interesting method is called the Quasi-linear amplifier, which uses a self-phase distortion compensation technique, reported by Hayashi [Hayashi H. et al. 1995]. Their work uses the common-source FET (CSF), which delivers the output power out-of-phase while the common-gate FET (CGF) does so in-phase. Their AM-PM conversion near the saturation region is also expected to have opposite behaviours. Therefore, they combine the CSF and CGF to create a cascade connection FET (CCF).

(a) Phase deviation versus the output power of the CSF and CGF

(b) Circuit configuration

Figure 6.8 Block diagram of self phase distortion method, (after Hayashi H. et al. 1995)
The results are very good. The experimental CCF power amplifier has more than 20dB gain and 50% PAE at an operating voltage of only 3V. The authors claimed, among the PA's designed for the mobile phone handsets, that this amplifier achieves the highest efficiency.

6.12. Feedback

Feedback linearization techniques are another classic method. The greatest advantage of feedback techniques, over other methods, is its complete independence of the minimum nonlinear performance of the amplifier produced by the effects of drift. However, feedback has a limited effectiveness because the linear output power is not the maximum the amplifier is capable of providing. In this case, the reduction in gain at the output of the feedback amplifier is proportional to the return difference.

Cartesian feedback is the most commonly used feedback methods [Briffa M. A. et al. 1994], as illustrated in Figure 6.9. The technique is to demodulate the actual transmitted signal and to use the baseband in-phase (I) and quadrature phase (Q) values as feedback signals. The demodulated I and Q signals are fed back to the modulator for adaptive predistortion of the signal constellation. They always combine with the predistortion to achieve the desired linearization. However, its instability restricts its wider use.

![Figure 6.9 Dynamically biased Cartesian feedback transmitter (after Briffa M. A. et al. 1994)](image)

Envelope feedback is a widely used form of feedback linearization, as illustrated in Figure 6.10. Gain and phase correction can be achieved by two envelope-feedback loops respective [Cardinal J. S. et. al. 1995]. This idea combines dynamic biasing, phase predistortion, and envelope feedback. However, when the envelope feedback control method does not operate ideally, amplitude distortion occurs. The two main possible causes are: (a) nonlinear characteristics of a diode used for the envelope detector, and (b)
a change of loop gain by the envelope variation. Therefore, in most feedback designs, the feedback loop is the most important part of the whole system.

![Diagram of adaptive double envelope feedback amplifier block diagram](after Cardinal J. S. et al. 1995)

**Discussion:** The disadvantage of the feedback linearizer is its instability, due to the loop delay. This also restricts wide bandwidth applications. For the altimeter application, the chirp signal's bandwidth is 2.3%. This is considered as a narrow band case, which means feedback can be used. Also, its adaptiveness makes it very suitable to work in a dynamic temperature changing environment.

### 6.13. Feedforward

Compared with other linearization, the merit of feedforward is its wide bandwidth. However, due to the need for an auxiliary amplifier, its power efficiency is usually not very satisfying.

**General construction**

Feedforward is another widely used linearization method. The principle of operation is based on two cancellation loops, as illustrated by the block diagram shown in Figure 6.11. In the first loop, the input carriers are suppressed at arm B of the second coupler, producing an error signal proportional to the distortion of the main amplifier. In the second loop, this error signal is amplified to the appropriate level by the linear auxiliary amplifier and is subtracted from the main amplified distorted signal at the output coupler, resulting in an error free signal at the linearizer output.
The main advantage of feedforward is it can operate over wide bandwidths, compared with other methods. However, its disadvantages are also obvious. Due to the adoption of an extra amplifier, the overall efficiency of the whole system is correspondingly decreased. In the classical feedforward linearizer circuit, both the main and auxiliary amplifiers have the same power rating. However, the auxiliary amplifier only amplifies the error signal, which contains the inter-modulation products and the remnants of the carrier signals and could, therefore, have a lower power rating without degrading the performance of the circuit, resulting in improved system efficiency.

The analysis of Parsons [Parsons K. J. et al. 1994] pointed out that the maximum feedforward efficiency is markedly reduced with even a small amount of delay insertion loss. With zero delay loss the maximum feedforward efficiency that can be achieved is 42%. This is reduced to 35% with 1 dB of delay insertion loss, and to only 28% with 2 dB of delay insertion loss. This corresponds to 7%/dB drops in efficiency. Therefore, great care should be put into the design of the delay line.

Feedforward linearization is very sensitive to component changes, due to environmental effects and ageing. Therefore, adaptive interference cancellation is proposed [Grant S. J., et al. 1996] [Talwar A. K. 1994]. Grant proposed a DSP controlled adaptive feedforward amplifier linearizer, shown in Figure 6.12. Here, the adaptive complex coefficients $\alpha$ and $\beta$ represent the attenuation and phase shift introduced in both the signal and error cancellation circuits, respectively. Control of $\alpha$, based on the linear estimation of power amplifier output $Va(t)$, $Pe$ with basis $Vm(t)$, is designed to minimise the average power of the error signal. The control of $\beta$ proceeds in a similar manner except that the basis of the estimate of $Va(t)$ is $Ve(t)$ and the estimation error is $Vo(t)$. The experimental result shows that with the use of DSP it can overcome the problems of mixer DC offsets and masking of strong signals by weaker ones, that would otherwise compromise other previously proposed analogue implementations.
Combine with predistortion

Due to the poor power efficiency characteristics of feedforward linearization technique, it is not attractive for handset applications. To overcome this, various techniques have been proposed [Parsons K.J. et al. 1995]. For example, to remove the time delay to increase the power efficiency as well as reduce the size in the narrow bandwidth condition. However, this will reduce the linearity improvement. One solution is to combine the predistortion and feedforward methods to achieve a better performance. Results show the overall feedforward efficiency is dependent upon the level of predistortion and the error amplifier efficiency, as shown in Figure 6.13. By improving the predistortion factor, the overall efficiency could be increased. At the same time reducing the power rating of the error amplifier will reduce the size and cost of the whole system.

Figure 6.13. Predistortion and feedforward amplifier efficiency (after Parsons K.J. et al. 1995)
Combine feedforward and feedback

Faulkner [Faulkner M. et al. 1995] proposed a feedforward system using an RF feedback loop as the first signal cancellation loop, as shown in Figure 6.14. The coefficient \( g_2 \) is adjusted to give the correct amplitude and phase for the feedforward cancellation loop.

The feedforward-feedback system has a number of advantages. The additional linearisation provided by the feedforward system allows the loop gain of the RF feedback system to be reduced and this can be traded for other features, such as a greater tolerance in the adjustment of the loop phase compensation circuit or increased bandwidth. Also the loop gain of the feedback system reduces the need for critical adjustment in the two feedforward loops.

![Analytical model of feedback-feedforward proposal](image)

**Figure 6.14.** Analytical model of feedback-feedforward proposal. The coefficient \( g_1 \) controls the signal cancellation loop and \( g_2 \) the error cancellation loop (after Faulkner M. et al. 1995)

**Discussion:** The large size and low efficiency of feedforward linearization technique limit its use in our application. However, for wide bandwidth applications, this idea will be very suitable, especially when combined with other methods to correct the environmental influence.

### 6.14. Linear amplification with Nonlinear Components (LINC)

The basic principle of the LINC transmitter is to represent any arbitrary band pass signal, which may have both amplitude and phase variations, by means of two signals which are of constant amplitude and only have phase variations. These two angle modulated signals can be amplified separately using efficient high power, non-linear devices. Finally, the amplified signals are passively combined to produce an amplitude modulated signal. The potential DC-RF conversion efficiency of this type of amplifier can approach 100%. This technique was first introduced in 1930. It is only now, with the advent of DSP to generate two angle modulated signals that this technique has attracted more attention. The basic system diagram and the popular design ideas are shown in Figure 6.15.
Chapter 6. Feedback Linearization for Highly Efficient Amplifier Payload Design

(a) Basic diagram of LINC;

(b) Analogue component separator from (after Cox D. C. 1974) (*original quality is poor)

(c) Example of LINC transmitter implementation with a digital component separator, (after Sundsrom L. 1996)

Figure 6.15 LINC linearizer diagrams

The implementation of the signal component separator is the most difficult part of this method. Cox [Cox D. C. 1974] proposed a analogue component separator which is very complex, as shown in Figure 6.15(b). The complexity of these systems has prevented the technique from being widely accepted. Today, it is already possible to implement the component separator in software, by using a standard DSP technique, as shown, in Figure 6.15(c).
Another difficulty is the suppression of broadband phase modulation components at the summation port, this relies on a very tight tolerance in the gain and the phase match in both paths. (0.1 dB gain error and 0.1° phase error will give a component suppression of only 54 dB. Such a tolerance is impractical to achieve with an open loop system, where significant gain and phase variations occur as a result of changes in operating frequency, output power, temperature and RF power devices. Bateman [Bateman A. 1993] proposed a possible solution by using a combined analogue locked loop universal modulator (CALLUM). The basic diagram is shown in Figure 6.16.

![Figure 6.16 Block diagram of Callum LINC (after Bateman A. 1993)](image)

The Cartesian (I and Q) signals demodulated from the CALLUM output signal are fed back and compare with the input I and Q signals. The resulting I and Q error signals are used to control two VCOs respectively. The outputs of the two VCOs are then combined, producing an amplified replica of input signals. With appropriate design, the CALLUM modulator can maintain a very small error signal. Therefore compared with the conventional LINC transmitter, a major advantage of the CALLUM modulator is its ability to correct the mismatch between two signal processing arms. With the feedback control mechanism, any mismatch (gain and phase) existing in the two processing arms will result in an error signal, which is then used to control the VCO’s phase, so as to compensate for the mismatch.

Discussion: LINC is designed to be used in the non-constant envelope condition. Therefore, it is not appropriate for the FMCW signal case.

6.15. Dynamic Biasing

The operational principle of dynamic biasing is it uses an external circuit to control dynamically, the gate’s ‘DC’ bias voltage of the FET with the envelope of the input RF signal such that the drain DC bias current is proportional to the envelope [Saleh A. A. et. al. 1985]. Thus, the DC bias power would vary up and down with the signal envelope so
as to reduce the DC power consumption and increasing the efficiency. The block diagram is shown in Figure 6.17.

![Block Diagram](image)

Figure 6.17 Dynamic biasing block diagram (from Saleh A. A. et. al. 1985)

Dynamic biasing can also be combined with other linearization methods to achieve a better performance. Ghannouchi [Ghannouchi F. M. et al. 1995] reported a novel adaptive envelope feedback lineariser using a dynamic gate bias for gain stabilisation. This linearity enhancement is accompanied by an 10% increase (in average) in the power-added efficiency from 35% to 45%.

### 6.16 Discussion

The above study is a summary of existing linearization techniques. Although recently many new linearization ideas have been proposed, most of them are still based on the above basic concept of predistortion, feedforward, feedback and LINC. The development trend is to combine several methods together. For example, feedforward plus feedback, predistortion plus LINC, etc. To make maximum use of each method's advantage, as well as to compensate each one's disadvantages.

For the radar altimeter payload project, which requires high DC-RF power efficiency, compact size, dynamic operating temperature, and high linearity, adaptive feedback is a good choice. Here extra care must be taken during the design phase to monitor the loop stability.
6.2 Feedback Linearizer – Implementation and Measured Results

It is known that for a general amplifier the higher the input dynamic range, the higher the phase distortion we would expect, as shown in Figure 6.18. The Figure shows that the phase distortion between mark 1 and mark 4 is larger than 40 degrees. Therefore, it is important to correct the phase distortion when the amplifier is driven to the high power condition.

![Figure 6.18. Phase distortion as a function of input power](image)

However, for the radar altimeter application, the condition is a bit different. The input signal is an FMCW signal, sweeps from 13.6 to 13.9 GHz, within 100 μs, whilst the input signal power does not change very dramatically. The phase error we expect in here is the phase deviation error within the whole sweep bandwidth. In other words, because the phase error at each individual frequency point is different within the whole bandwidth, therefore the measured group delay will not be constant within the swept frequency bandwidth. Because the delay relates with the phase information directly, therefore, the aim of this feedback loop can be deemed as to correct the delay time difference based on a reference delay. From this point of view, the proposed linearization circuit can be viewed as a group delay equalizer.

Compared with other linearization methods, such as feedforward and predistortion, the feedback linearization method has the advantages in that it is simple, accurate, adaptive, and needs no extra amplifier. As a result, it has been chosen as the linearization method for proposed radar altimeter payload project.

The general role of a linearizer is to correct for the AM/PM distortion as well as the AM/AM distortion. However, for the altimeter radar payload, the previous analysis for
the altimeter's significant wave height estimation shows that the phase error has a more serious influence upon SWH than that of the amplitude fluctuation error. This means the lineariser design can be simplified by concentrating on compensating for the power amplifier's phase error, instead of considering both the phase and amplitude errors. The basic implementation principle of this linearization method is, shown in Figure 6.19.

The output of the phase error detector, at point A in Figure 6.19, is a DC voltage that changes quasi-linearly with the changing of the phase distortion, when this phase error is a reasonable small value. This DC signal is used to control the voltage-controlled analogue reflection-type phase shifter. The phase shifter is added before the power amplifier, instead of after the amplifier, due to the diode's limited power handling capability. With reference to Figure 6.19, the higher the phase difference between the amplifier's input and output signals, at points B and C, the larger the output voltage from the phase error detector at point A. Therefore, the higher the related phase shift will be introduced to the transmitted signal before it is fed to the power amplifier. Operating in this way, the phase error introduced by the power amplifier can be compensated for.

Feedback linearization has its disadvantage of conditional stability when it is operated in a wide bandwidth condition. However, for the altimeter application, the bandwidth for this altimeter chirp signal is very narrow -- only 320 MHz/13.6 GHz = 2.35%, therefore it is very easy to stabilise the loop within such a narrow bandwidth.

The following sections will explain, in detail, the operational principles of this feedback linearizer loop, especially the phase error detection and calculation sub-system. This is the most important part of the whole linearization design. It will then be followed by the discussion of hardware design simulation and implementation results. The task has been broken down into three main steps, and in each step the simulation and implementation results will be given in detail. Finally, the whole feedback linearization loop test results will be given.

### 6.21 Phase Difference Detector

The phase detector is the core part in the whole feedback lineariser loop. The general approach in the design is to feed the amplifier's input and output signals to power dividers and then feed their output signals to a phase difference calculator and amplitude calculator, in order to get the final control voltage for the phase shifter. The phase detector makes use of a 3dB branch line hybrid coupler to detect the relative phase.
imbalance between the amplifier’s input and output. The circuit diagrams are shown in Figure 6.20.

![Circuit Diagrams](image)

Figure 6. 20. Conventional design for phase imbalance detection

The disadvantage of the above design is that it is quite large, as the size of the power divider is inversely proportional to the operating frequency. Also, the two output ports of the Wilkinson power divider shown in Figure 6.20(b) are not yet matched to 50Ω. Therefore, there is a poor impedance match at the divider’s output ports.

The novel idea introduced in this chapter’s design, is to remove the in-phase power divider in both feedback loops and combine two 3dB quadrature couplers together to achieve the same goal. First let’s look at the operation principle of the 3 dB branch line coupler.

![Branch Line Coupler Diagram](image)

Let \( \tilde{a} \) & \( \tilde{b} \) represent the two input signals of the branch line coupler. \( \tilde{c} \) & \( \tilde{d} \) are the two output signals. Ideally, it can be shown that:
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\[ c = \frac{\sqrt{2}}{2}(\bar{a} - j\bar{b}) \]

and

\[ d = \frac{\sqrt{2}}{2}(-j\bar{a}) \quad \text{(6-1)} \]

If the two inputs are connected with two power detector diodes, and an Op-amp is used to calculate the difference between the diode outputs, a control voltage, \( V \) can be obtained. Now, let \((A, \theta_1)\) and \((B, \theta_2)\) represent the amplitude and phase of the two input signals respectively, \( c \) & \( d \) can be calculated as:

\[ \bar{c} = \frac{\sqrt{2}}{2}(Ae^{j\theta_1} - jBe^{j\theta_2}) = \frac{\sqrt{2}}{2}[(A \cos \theta_1 + B \sin \theta_2) + j(A \sin \theta_1 - B \cos \theta_2)] \]

\[ \bar{d} = \frac{\sqrt{2}}{2}(Be^{j\theta_2} - jAe^{j\theta_1}) = \frac{\sqrt{2}}{2}[(A \sin \theta_1 + B \cos \theta_2) + j(B \sin \theta_2 - A \cos \theta_1)] \]

\[ V = G_{op}(|\bar{c}|^2 - |\bar{d}|^2) = \frac{G_{op}}{2} (4AB \sin(\theta_2 - \theta_1)) = 2G_{op} AB \sin(\theta_2 - \theta_1) \quad \text{(6-2)} \]

Combined with the knowledge of the two input signal amplitudes, an output signal can be determined, which only relates with phase difference of the two input signals. The normal method is to use a power divider to divide the amplifier input & output signals simultaneously and then use another set of Op-amps to calculate the signal amplitude. However, it will make the design more complex and increases the substrate size. The diagram for the new design method is shown in Figure 6.22.

![Figure 6.22. Design of the novel branch line coupler](image)

According to the operation of the branch line coupler, a simple re-analysing of equation (6-2) gives:
Equation (6-3) shows that the combination of the two branch line couplers can separate the two input signals and output them to two different ports. If these signals are fed to the power diode detectors, we can then get amplitude information for these two input signals. The basic configuration of this power detection diagram is shown in Figure 6.23.

\[
\tilde{c} = \frac{1}{2} \frac{\sqrt{2}}{2} (\tilde{a} - j \tilde{b}), \quad \tilde{d} = \frac{1}{2} \frac{\sqrt{2}}{2} (\tilde{b} - j \tilde{a})
\]

\[
\tilde{a} = \frac{\sqrt{2}}{2} (\tilde{c} - j \tilde{d}) = \frac{1}{4} [\tilde{a} - j \tilde{b} - j(\tilde{b} - j \tilde{a})] = -j \frac{1}{2} \tilde{b}
\]

\[
\tilde{f} = \frac{\sqrt{2}}{2} (\tilde{d} - j \tilde{c}) = \frac{1}{4} ([\tilde{b} - j \tilde{a}] - j(\tilde{a} - j \tilde{b})] = -j \frac{1}{2} \tilde{a}
\]  

Figure 6.23. Novel design of the phase difference detection and calculation sub-circuit

The final output voltage can be expressed as:

\[
V_{out} = \frac{V_1}{V_2} = \frac{V_c - V_d}{V_c \cdot V_f} = \frac{G_{op1}}{G_{op2}} \sin[\theta_1 - \theta_2]
\]

By carefully adjusting the gain of the two operational amplifiers, the slope of the output voltage can be the same as the voltage-tuning characteristic of the voltage controlled phase shifter. Operating in this way, the linearization aim can be achieved.

6.22. Hardware Implementation Results

In this section, the implementation of the phase detector and calculator has been broken down into several major steps:
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a) The branch line coupler simulation and implementation test results
b) The branch line coupler plus RF power detector performance
c) Op-amp performance
d) The phase shifter design and implementation
e) The whole RF and DC analogue design measurement

Branch line coupler implementation

For the hybrid coupler design in Figure 6.20, it is very important to keep the amplitude imbalance between port 3 and port 4 as small as possible. At the same time the phase difference between \( S_{31} \) & \( S_{32} \) and \( S_{41} \) & \( S_{42} \) should also be kept as close to 90 degree as possible. Table 6-1 shows the measured S-parameters results for port 3 and port 4. Here, the substrate chosen for the design was FR4 \((h = 1.6 \, mm, \varepsilon = 4.45)\)

| Frequency (GHz) | \( |S_{31}| \) (dB) | \( \angle S_{31} \) (deg) | \( |S_{32}| \) (dB) | \( \angle S_{32} \) (deg) | \( |S_{41}| \) (dB) | \( \angle S_{41} \) (deg) | \( |S_{42}| \) (dB) | \( \angle S_{42} \) (deg) |
|----------------|------------------|------------------|------------------|------------------|------------------|------------------|------------------|------------------|
| 2.2            | -6.64            | -142.7           | -7.4             | -142.7           | -7.4             | -142.7           | -6.29            | 121.7            |
| 2.205          | -6.75            | -143.8           | -7.47            | -143.8           | -7.48            | -143.8           | -6.40            | 120.7            |
| 2.21           | -6.85            | -144.2           | -7.55            | -144.2           | -7.52            | -144.2           | -6.49            | 119.9            |
| 2.215          | -6.97            | -144.8           | -7.63            | -144.8           | -7.38            | -144.8           | -6.6             | 119.1            |
| 2.22           | -7.12            | -145.7           | -7.73            | -145.7           | -7.69            | -145.7           | -6.75            | 118.3            |
| 2.225          | -7.24            | -145.9           | -7.80            | -145.9           | -7.75            | -145.9           | -6.83            | 117.7            |
| 2.23           | -7.33            | -145.8           | -7.86            | -145.8           | -7.81            | -145.8           | -6.93            | 117.4            |
| 2.235          | -7.41            | -146.2           | -7.91            | -146.2           | -7.87            | -146.2           | -7.00            | 117.0            |
| 2.24           | -7.45            | -146.1           | -7.94            | -146.1           | -7.88            | -146.1           | -7.09            | 116.8            |
| 2.245          | -7.54            | -143.1           | -7.96            | -143.1           | -7.88            | -143.1           | -7.14            | 116.6            |

(a) \( S_{31}, S_{32}, S_{41}, S_{42} \) hybrid measurement results
Table 6-1(b) shows that the amplitude imbalance between the signals of port 1 and port 2 is nearly always smaller than 1 dB, which could generally satisfy the design requirements. Also, the phase difference is very close to 90 degrees. The maximum error is lower than 12.52 degrees for the worst case, within the error range.

The measured S-parameters of ports 5 and 6 are listed in Table 6-2. The measurement results show that both the $|S51|$ and $|S62|$ are very small, lower than $-17$ dB. Thus, it can be assumed that the output of port 5 only presents the input signal from port 2, and the output of port 6 only presents the input signal from port 1. Similarly, the amplitude imbalance between $S52$ and $S61$ is also very small — generally this is around 0.1 dB within the swept bandwidth. This provides a good premise for the later design for power detection.
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| Frequency (GHz) | $|S51|$ (dB) | $\angle S51$ (deg) | $|S52|$ (dB) | $\angle S52$ (deg) | $|S61|$ (dB) | $\angle S61$ (deg) | $|S62|$ (dB) | $\angle S62$ (deg) |
|----------------|------------|-------------------|------------|-------------------|------------|-------------------|------------|-------------------|
| 2.2            | -17.37     | 42.5              | -6.05      | 32.6              | -5.94      | 29.2              | -17.83     | 0.8               |
| 2.205          | -17.49     | 42.1              | -6.12      | 31.6              | -6.03      | 28.1              | -18.02     | 0.7               |
| 2.21           | -17.64     | 41.9              | -6.2       | 30.7              | -6.08      | 27.3              | -18.23     | 0.7               |
| 2.215          | -17.8      | 41.8              | -6.28      | 30.1              | -6.17      | 26.5              | -18.45     | 0.7               |
| 2.22           | -18.03     | 42.1              | -6.4       | 29.3              | -6.29      | 25.8              | -18.67     | 0.6               |
| 2.225          | -18.16     | 42.4              | -6.44      | 28.8              | -6.32      | 25.1              | -18.85     | 0.6               |
| 2.23           | -18.28     | 42.5              | -6.5       | 28.3              | -6.38      | 24.8              | -19.03     | 0.5               |
| 2.235          | -18.36     | 42.8              | -6.55      | 27.9              | -6.43      | 24.5              | -19.23     | 0.5               |
| 2.24           | -18.56     | 43.2              | -6.59      | 27.5              | -6.48      | 23.8              | -19.38     | 0.5               |
| 2.245          | -18.56     | 46.8              | -6.61      | 27.1              | -6.51      | 23.4              | -19.45     | 0.4               |
| 2.25           | -18.61     | 44.1              | -6.62      | 26.6              | -6.5       | 22.9              | -19.53     | 0.4               |

Table 6-2. Port 5 and port 6's S-parameters measurement results

In summary, the design of this novel branch line coupler is very successful. Both the amplitude and phase imbalance is very small, which provide a good premise for the next step of the design.

**Hybrid plus power detector implementation results**

The next step is to measure the performance of a hybrid coupler with the power detector. Recall in equation (6-2), the output voltage from the detector $V$ in the equation should be $V = (V_c - V_d) / (V_e \cdot V_f) = G \sin \theta$. However in this step, the op-amplifier used to calculate the final output voltage $V$ hasn't been adopted yet. But $V$ can still be obtained by measuring the voltages of $V_c, V_d, V_e, V_f$ using meters. The schematic diagram and measurement set up are shown in Figure 6.24.

---

Figure 6.24 System configuration diagram and test set up
One of the most critical design points in this step is to choose four identical power detection diodes, because any difference in response among those diodes will introduce an extra phase error. During the design, the power detector diode chosen was a high performance surface mounted Schottky diode, HSMS-2860. This is a high detection sensitivity diode, and it can work up to 12 GHz. At S band, its detection sensitivity is 35 mV/µW. Figure 6.25 shows the detector’s circuit diagram.

![Figure 6.25 RF power detector circuit diagram](image)

In the design of detector, the proper choice of $R_1$ in Figure 6.23 is very important. As when the detector is connected with different pin of the op-amplifier, it equals to connect with different value resistors. According to the test experience, it is suggested that $R_1$ for $V_c$ and $V_d$ should be set around 50 Ω, while for $V_e$ and $V_f$ the value of $R_1$ values should be very large, say 100 kΩ.

Figure 6.26 shows the theoretical (diamond line) and measured (solid line) output voltages when changing the phase difference between the two hybrid inputs. It can be seen that there is a constant voltage difference between the measured and theoretical values. This voltage can be removed by setting an offset bias voltage in the op-amp.

![Figure 6.26 Measured and theoretical performance of the detectors](image)
In the implementation, it is very important to choose a proper diode forward bias voltage, as it will influence the accuracy of the output voltage, especially when the amplitude of the two input signals are not in balance.

**Op-amp performance**

The last step of the power detector design is to add the Op-amps, and put both the operational amplifier PCB and RF signal detectors PCB together on one board, to test their performance. The analogue devices are used to achieve the functions of minus, multiply and divide. All of the Op-amps are Burr-Brown devices – MPY634 for multiplier & divider, and INA114 for the minus function. Figure 6.27 shows their schematic.

![Minus circuit configuration diagram](image)

(a) Minus circuit configuration diagram

![Multiplier circuit connection diagram](image)

(b) Multiplier circuit connection diagram
Chapter 6. Feedback Linearization for Highly Efficient Amplifier Payload Design

The INA114 is a precision low off-set instrumentation amplifier. It has a wide DC supply voltage range, from ±2.25V to ±18V.

MPY634 is a wide bandwidth precision analog multiplier. Its differential X, Y, and Z inputs allow its configuration as a multiplier, squarer, divider, square-rooter, and other functions while maintaining high accuracy. The error of the multiplier and divider can be compensated for by the external off-set voltage in Y1 and Y2. Figure 6.27(b) is an operational amplifier non-converter connection. By changing the resistor values, the output multiplication gain can be changed.

There are several things that must be considered carefully during the design:

- For the INA114, although the input impedance is extremely high, a path must be provided for the input bias current of both input diodes. This will be achieved by connecting a resistor R1 to ground after the power detector diodes.

- For the MPY634 divider, because the output amplitude of the multiplier of the two input signals amplitude is very small, generally is around 0.1 V, it is quite easy that the divide product is higher than 15V, the DC power supply voltage. For that case, the divider will not work properly. Thus, it is important to provide some higher gain in the multiplier design by changing the ratio of the two resistors between output and Z1 & Z2 in the divider stage.

- It is better to supply a trimming voltage (device allows it varies from 100mV to 10V) to the MPY634 divider to correct the error in the final output voltage.

The whole DC and RF component side's board is shown in Figure 6.28 (a) (having dimensions of 78×37mm², for an FR4 substrate). As the RF and IF sections are very
close to each other, during the implementation, it was very important to prevent RF signal leakage into the DC supply resulting in unwelcome noise.

Figure 6.28 Detector board

To test the detector's performance, a test-bench was set up, as shown in Figure 6.29. Here, the branch line coupler split the input signal to two equal amplitude and equal phase paths. One is directly fed into the detector, while the other is fed to P2 through a phase shifter, which is used to control the phase difference between the two input paths, P1 and P2.
Chapter 6. Feedback Linearization for Highly Efficient Amplifier Payload Design

The test results for the functions of three different Op-ampifiers are good, as shown in Figure 6.30 (a), (b) & (c). Each plot gives the comparison of the measured and theoretical results for the device of achieving minus, multiply and divide functions. Here, the theoretical measurement results are obtained by calculating the output from the power detectors. Their values are different from the Op-ampifier’s output values by a factor of $G$, which depends on different Op-ampifiers, and can be changed by setting different $SF$ values.

(a) Measured and theoretical results for minus function results
Chapter 6. Feedback Linearization for Highly Efficient Amplifier Payload Design

The final detector's output voltage vs. the phase difference between the two input ports is shown in Figure 6.31. The circuit is tested at three different operating frequencies: 2.15 GHz, 2.2 GHz and 2.25 GHz. In each plot of Figure 6.30 the theoretical output is a sine function of phase difference between the two inputs, and has been plotted to compare with the measured output.
Detector output vs. phase difference

(a) $f = 2.15$ GHz

Detector output vs. phase difference

(b) $f = 2.20$ GHz
Detector output vs. phase difference

Figure 6.31 Final Output vs. phase difference at three test frequencies

Figure 6.31 shows that the detector circuit operates best around 2.15 GHz. In all the figures, the measured and theoretical values are plotted on two Y-axes. In the real operation, the value of $G$ should be chosen by equation (6-4), $\Delta V = \frac{V_c - V_d}{V_c - V_f} = G \sin \theta$, will be decided by the phase shifter performance. Because it is only if these two have the same voltage/phase slope could the whole design achieve the required feedback linearization performance.

Phase shifter design and implementation

The design of the analogue phase shifter is also very critical for a good performance lineariser. The performance requirements include proper tuning characteristics, high linearity, low insertion loss, etc. The first design step is to run a phase shifter simulation in COMMS by setting a capacitor and changing its value to observe its $\Delta V$. The prototype of the phase shifter is shown in Figure 6.32 (a). The simulation is based on the FR4 substrate. Figure 6.32 (b) shows the substrate layout of this S-band phase shifter. The thin line, near the output port, is a quarter wavelength high impedance line, and it should have no influence on the RF signal. It is connected to the ground at its end to provide the control voltage ground reference.
A Siemens BB 825 diode was chosen as the tuning varactor. It can work up to 2.8GHz and has a high tuning capacitance ratio of $C_{V_{1}}/C_{V_{28}} = 14$. Figure 6.31 shows the measured $\text{Ang}(S21)$ changing vs. control voltage. Figure 6.33(a) shows that this phase shifter can achieve more than 120 degrees phase changing within a 10 V control changing range. However, its linearity performance is not very good, which is mainly due to the inherent nature of the device physics. Nevertheless, up to a 50 degree phase change is already sufficient enough to satisfy the design requirement. Thus, the control voltage will be restricted within a range of -4 to -8 V, to control the output phase changing slope linearity, as shown in Figure 6.33(b). For this condition, the phase change could still achieve up to 80 degrees, but the linearity has been improved greatly.
Figure 6.34 shows the measured phase shifter’s $Mag[S21]$ and $Ang[S21]$ performance. The result shows that the phase shifter will introduce around 2 dB of insertion loss. This is partly because of the substrate that is used in the design. To improve the performance, it is suggested a high quality microwave substrate should be adopted, such as Duriod.
**Final linearization test**

With all the necessary sub-circuits tested and ready, the whole feedback linearization loop can finally be evaluated. The test-bench is set-up according to the block diagram in Figure 6.35.

![Figure 6.35 Lineariser test bench diagram](image)

A photograph of the whole test bench is shown in Figure 6.36. An S-band phase shifter is used as an adjustable delay to regulate the delay time of the delay line in Figure 6.35. This delay is set to the maximum value of the power amplifier’s delay time. An attenuator is used between the drive amplifier and the main amplifier, to control the input power of the main amplifier to the right level as well as to serve as an isolator to improve the main amplifier’s performance. It is also important to make sure that the input power of the phase detector at port A and port B are the same, otherwise the extra power imbalance will introduce in an error signal the output signal at port C.

The test frequency for the lineariser is chosen between 2.1 GHz to 2.15 GHz. This is because, although the amplifier is designed for 2.2 GHz, test results in Chapter 5 show that the amplifier actually works best at 1.9 GHz. When the amplifier operates near 2.2 GHz, its output amplitude is not very flat. Therefore, after a trade-off, 2.1-2.15 GHz was chosen as the lineariser’s operating frequency range. The bandwidth is 2.3% (50 MHz /2.15 GHz), the same as required by the Ku-band altimeter radar payload.
In the photograph:

(a) Branch line coupler to split the input power to two equal amplitude and equal phase paths
(b) Voltage control phase shifter, the control voltage comes from the detector’s output
(c) Drive amplifier
(d) Adjustable attenuator, used as isolator between the drive amplifier and the main amplifier
(e) Main amplifier
(f) Coupler
(g) Phase shifter used as adjustable delay line
(h) Detector

Figure 6.36 Photograph of lineariser test bench

Figure 6.37 shows the final measurement result before the lineariser is applied. Before the feedback loop is applied, the measured amplifier group delay varies around 1.5ns within the bandwidth (mark 1 to mark 4 in Figure 6.37).
Figure 6.37. Group delay over the whole frequency band before feedback linearization is applied.

Figure 6.38 shows the measured group delay performance result after the lineariser is applied. The plot shows that between mark 2 to mark 4, the amplifier’s group delay is very flat – the group delay variation is less than 0.2 ns. However, the plot also shows that the lineariser only works well between 2.11 – 2.15 GHz frequency band, which is a bit lower than what was expected. The reason is due to the amplifier’s uneven gain response at f = 2.1 GHz. The power difference between the detector’s two input ports is largest, which is about 1.2 dB. This power imbalance then introduces an error control voltage in the detector’s output.

The result in Figure 6.38 shows that although the analysis in Chapter 4 indicated that the amplifier’s amplitude variation does not play an important role in SWH estimation error, this amplitude variation will influence the amplifier’s group delay by manipulating the lineariser detector’s output. According to test experience, it is recommended that the amplitude difference should be controlled under 0.5 dB.
The power consumption of the lineariser is very low, the circuit’s current is only 7mA at ±15 V dc supply. The gain loss of the total link under test is around 2.5 dB, which is mainly due to the phase shifter’s performance. So it is important that a low insertion loss phase shifter be adopted.

The comparison of Figure 6.37 and Figure 6.38 shows that this feedback linearizer successfully flattens the amplifier group delay, as expected. At the same time, the power consumption of the extra phase detector is very small – only around 200mW. The overall amplifier gain only drops by around 2 dB due to the voltage-controlled phase shifter’s insertion loss.

6.3 Conclusion

In this chapter, a detailed design and implementation of a novel feedback linearization loop, for chirp signal power amplifier group delay equalization, is given. The work in this chapter includes the simulation & design for a novel 3 dB phase difference detector & calculator, and the design of a linear voltage-controlled phase shifter. The novel part of this work is the phase difference detector, which aims to get rid of the conventional, more lossy power splitter thus decrease the circuit board size to one third of the original size.
and minimise the losses. This makes the lineariser very suitable for the mass and power limited microsatellite application.

The final complete lineariser performance measurement results are good – the peak-to-peak group delay variation dropping from 1.5 ns, before the linearization loop is applied, to lower than 0.3 ns, after the lineariser is applied.

The advantages of this feedback linearization are its relative simplicity, compactness, very low extra DC power consumption, as well as its adaptive nature. These advantages make the design well suited for the power and mass limited microsatellite payload applications.
Chapter 7 Conclusion and Future work

7.1 Work Summary

Although there is growing commercial interest in launching a low-cost microsatellite radar altimeter constellation, to achieve real or near real-time global sea state monitoring, up to now there has been no specific mission that is designed to achieve this aim. However, the analysis in Chapter 2, for both the design of a microsatellite platform and the radar altimeter payload, has shown that current technology for both are mature enough to fulfill this target. At the same time, the detailed feasibility study in this thesis has also identified several critical problems that may influence the platform and payload design. The first and the most important is the onboard DC power supply. A trade-off must be made between the limited power resource and the payload’s acceptable measurement resolution. Another important issue is the satellite attitude determination and control. The altimeter payload requires that the satellite’s off-nadir pointing angle must be smaller than the half power angle of the onboard antenna, and this mis-pointing angle knowledge must be estimated very accurately. The final question is the station keeping, as the fuel mass will directly decide the satellite’s platform size.

To achieve the goal of real or near real-time global sea state monitoring, a 12 satellites constellation network has been studied and simulated. This constellation arrangement is mainly decided by the user’s requirements, such as the maximum user waiting time and the altimetry map grid.

The phase and amplitude distortion influence on the onboard fast SWH estimation has also been investigated. The simulation results show that for the same hardware distortion, the worse case occurs at low SWH condition. The simulation also shows that in SWH estimation, the power amplifier influence has a more significant impact than that of the signal generator. At the same time, the influence of the hardware’s phase error, in other words the amplifier’s group delay influence is much more critical than that of amplitude influence. It is recommended, from simulations that within the whole chirp signal swept bandwidth, the peak group delay error value should be well controlled under 0.5ns. For the amplifier’s influence, in the worst case, a 2 dB amplifier’s amplitude variation will only introduce an 7% estimation error, which is within the design tolerate range.

The highly efficient power amplifier has been identified as one of the most critical problems in the payload design. Therefore, a high efficiency and high amplifier has been introduced in this research. The study gives a detail analysis of the Class-F amplifier’s design principles of second and third harmonic peaking concept, and also demonstrates them in the simulation. The measured third harmonic peaking Class-F amplifier’s DC-RF efficiency is 60%, compared with the normal Class-A amplifier, which only has around 30% DC-RF efficiency, the amplifier efficiency has been doubled.

It is well known that the amplifier’s power efficiency and phase distortion contradict with each other. To solve this problem, a feedback linearization technique has been proposed, as a good choice for satisfying both requirements. A novel detector circuit has been
Ch. 7. Conclusion and Future Work

proposed to make the circuit suitable for using in the compact microsatellite application. The measurement results show that the linearizer can successfully control the peak group delay distortion value under the required 0.5 ns target.

Although the designed amplifier is at S-band with a 3 W output, while the radar payload requires a Ku-band 7 W amplifier. It is believed that the design concept in this study can be applied to the real application. Therefore if one assumes the same amplifier’s DC-RF efficiency, and the radar duty cycle is 50%, only 11 W of DC power is required by the altimeter payload. According to the system feasibility study, this requirement can be satisfied by a microsatellite power budget.

In summary, the contributions to the state-of-art from this Ph.D. study are as following:

- A detailed and comprehensive system feasibility study for the microsatellite radar altimeter payload design and the corresponding requirements on each platform subsystem has been provided.

- A unique network constellation concept for near real time global monitoring has been presented.

- A clear and detailed understanding for the relationship between system amplitude/phase response and the on-board SWH and wind speed estimation has been outlined. The design target for the hardware performance has been specified.

- The possibility and accuracy of using large signal MESFET model for power amplifier design have been demonstrated. A detailed analysis of two types of Class-F operation including principles and design requirements has been given.

- A novel group delay feedback linearizer design method has been demonstrated.

7.2 Future Work

The Delay/Doppler altimeter, a method that can improve the received SNR by up to 10dB, has been proposed as a very promising solution for power limited microsatellites. It has fewer requirements on the satellite attitude control and determination subsystem. However there is no available algorithm for the SWH and wind speed estimation for this new concept. At the same time, how much on-board processing should be required is also not well understood. This leads to another Ph.D. study in SSC.

Although the current microsatellite attitude control for antenna mis-pointing could satisfy the basic requirement of that altimeter, a 0.5° mis-pointing does introduce a loss in the received SNR by 3 dB. At the same time, the trailing-edge of the altimeter return signal
waveform could help in the mis-pointing estimation. So it would be interesting to develop an algorithm that can rapidly determine the satellite pointing condition.

It is quite clear in this project that the fast delivery significant wave height estimation and sea surface wind speed are the two issues that are of most interest. However, the altimeter’s application in sea topography is also attractive to us. The main problem is how to quickly and accurately determine the satellite orbit without much help from other instruments. Detailed work must be done to explore the possibility.

In Chapter 5, the accuracy and capability of using MESFET large-signal modelling at higher frequencies using commercially available software, has been researched in detail. The study has clearly shown that the nonlinear circuit design will remain a very difficult problem for the RF engineer for a long time. There is still lots of work needed to do for the device modelling, however, this requires highly accurate measurements that might be a good MSc project.

Finally, with new and powerful DDS chips, the design of chirp signals will be much easier and the circuit will be more compact with less power consumption. So it will be quite interesting to use this chip to design the new generation of chirp signal generators.


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