DEVELOPMENT OF FEEDFORWARD RF POWER AMPLIFIER

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Not for loan

Development of Feedforward RF Power Amplifier

by

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Declaration

I declare that this dissertation is my own work and is being submitted in fulfilment of the requirements for the degree of Magister Technologiae in Electrical Engineering at the Cape Peninsula University of Technology. It has not been submitted before for any degree at this or any other academic institution. All consulted literature is completely listed in the references.

P.A. Lötter

Candidate

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Synopsis

Electronic communication systems have become an integral part of our everyday lives. RF (Radio Frequency) power amplifiers form part of the fundamental building blocks of an electronic communication system. RF power amplifiers can also be one of the major causes of distortion in an electronic communication system.

This thesis describes the linearity requirement for a RF power amplifier that is used in a transmitter section of an electronic communication system. Furthermore, five different linearisation techniques are presented and their characteristics compared. Since a power amplifier employing the Feedforward linearisation technique was designed, built and tested, this thesis focusses on the Feedforward technique.

The design methods for the various Feedforward components are presented. The measured parameters of the Feedforward linearised amplifier are compared with the measured parameters of a non-linearised amplifier.

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Chapter 1

Introduction

1.1 Overview

In an electronic communication system, power amplifiers are commonly found in the transmitter. An information-bearing modulated signal is fed to the input port of a power amplifier. This modulated signal is then amplified and the output signal of the amplifier is then fed to an antenna, where it is transmitted. All wireless systems are required to cause minimal possible interference to other wireless users. These wireless systems must therefore maintain their transmissions within the allocated channel bandwidth and not radiate significant amounts of energy outside the bandwidth limitations. A large contributor to this interference are power amplifiers [1]. In some applications, a linearisation technique must be applied to the power amplifier to minimise this interference.

1.2 Linearity Requirement for Various Modulation Schemes

When a transmitter radiates a single carrier constant envelope modulated signal such as a FM (Frequency Modulation) signal, the output signal of the power amplifier will contain an amplified version of the input signal, as well as lower power versions of the output signal at multiples of the fundamental frequency [1]. This is shown in Figure 1.1.

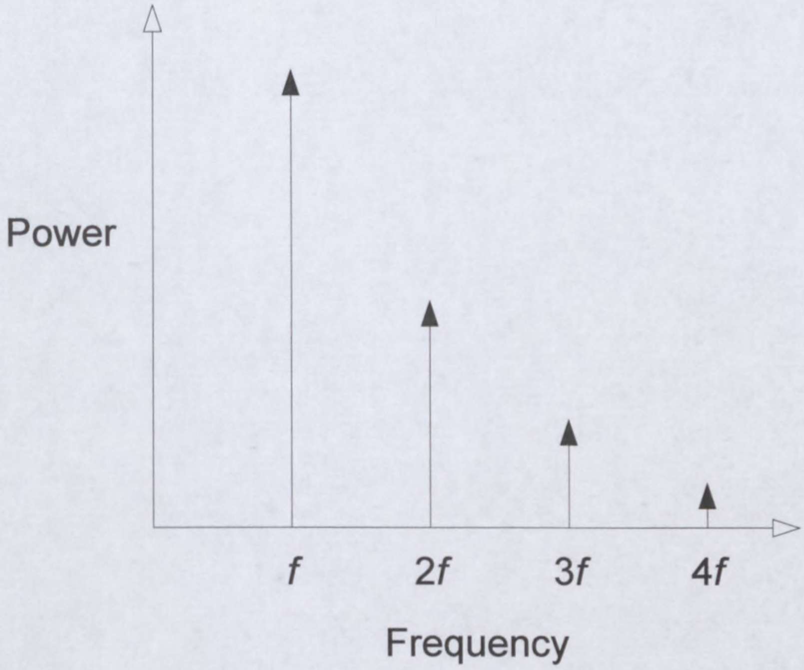


Figure 1.1: Output Spectrum of Power Amplifier with a Single Tone Input Signal

These unwanted signal components can easily be attenuated by means of a filter at the output of an amplifier. Therefore, it is possible to use highly non-linear amplifiers to transmit signals that use constant envelope modulation schemes [1].

The drawback to constant envelope modulation schemes is that they have a low data rate/bandwidth ratio. All wireless systems share a common transmission medium. Therefore, the available spectrum is limited and so the usage of the channel is directly associated with profit. Thus, in systems where higher data rates are required, more efficient modulation schemes are used, thereby minimizing channel bandwidth usage. These more efficient modulation schemes such as QAM (Quadrature Amplitude Modulation) contain information in the phase as well as the amplitude of the signal. Such modulation schemes are referred to as linear modulation schemes [2]. The drawback for this type of modulation scheme is that a more linear power amplifier is required, because if the modulated signal contains amplitude variations, any non-linearity in the power amplifier will cause a spreading of the transmitted signal spectrum, which is referred to as spectral re-growth. This spreading of the signal spectrum results in the information becoming distorted [2]. Figure 1.2 illustrates the frequency spectrum of a distorted 16QAM signal.

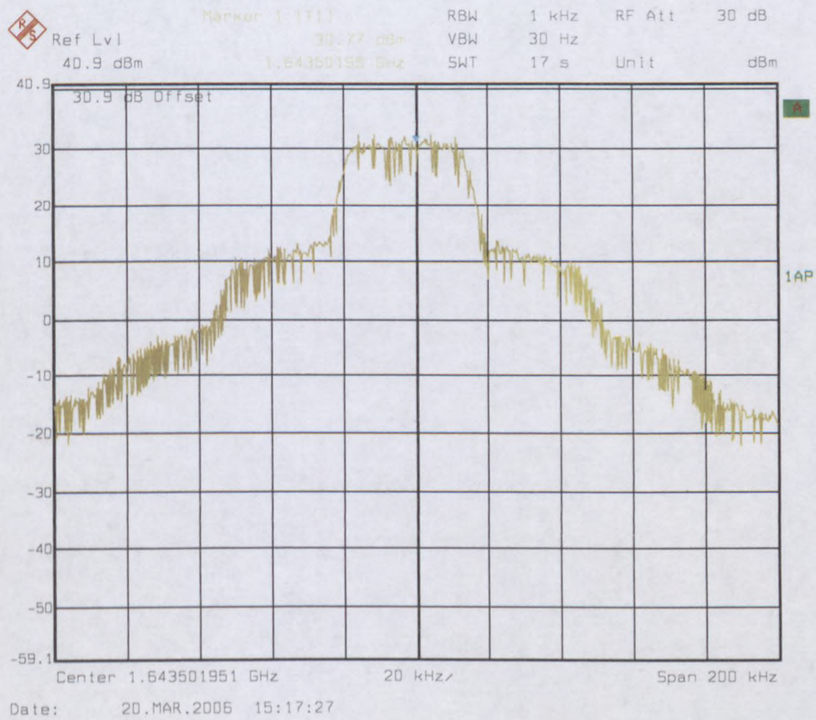


Figure 1.2: "Spreaded" Spectrum of Distorted 16QAM Signal

An essential requirement for some systems is to amplify multiple channels simultaneously. For the system to handle more than one user, the power amplifier must be able to handle more than one carrier at a time. Therefore, a multi-channel power amplifier is needed.

If more than one signal frequency is applied to a non-linear amplifier, the output spectrum contains additional signal components referred to as IMD (Intermodulation Distortion) products. These additional signal components are caused by inherent non-linearities of the amplifier. For an amplifier with input signals at frequencies f_1 and f_2, the output spectrum will contain signals at the following frequencies [3]:

$$nf_1 + mf_2,$$
$$\text{where: } n,m = 0, \pm1, \pm2, \pm3, \ldots\ldots \infty$$

The order of the IMD product is defined as $i = |n| + |m|$. Third order products (i=3) are a major concern because the third order intermodulation distortion products are very close to the fundamental components and cannot be filtered out. Figure 1.3 shows the amplified fundamental signal frequencies, the third order IMD products and the fifth order IMD products [3].

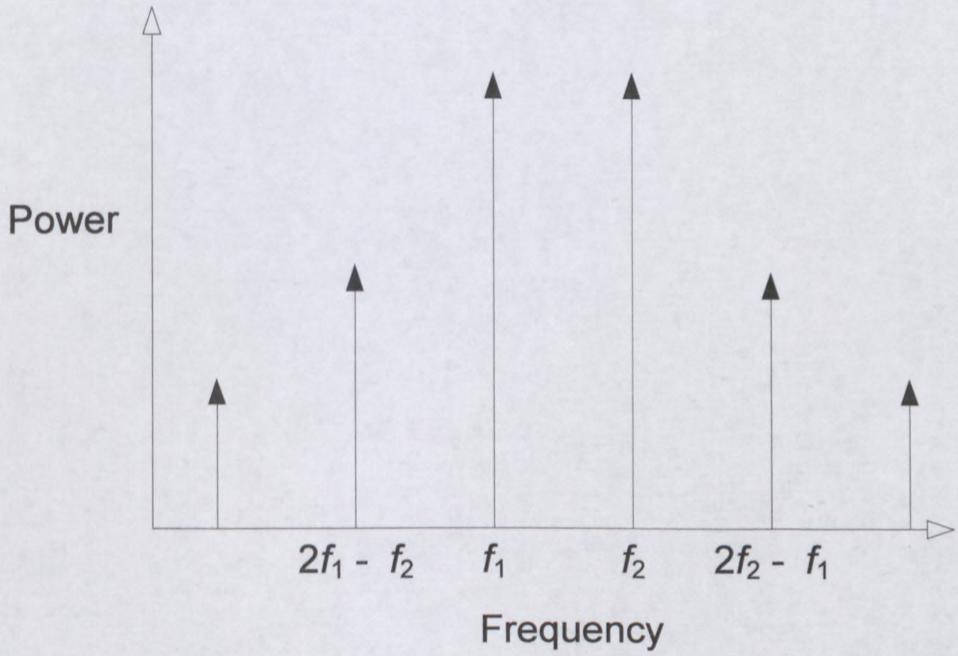


Figure 1.3: Output Spectrum of Amplifier with a Two-Tone Signal Applied

To suppress this intermodulation distortion, a linearisation scheme can be implemented to allow a single amplifier to amplify more than one carrier. The use of a linearisation technique has the advantage in it removes the need for a high power combiner with its associated loss, and a large number of individual single channel amplifiers. The alternative strategy would be to use a backed-off conventional Class-A or Class-AB amplifier. This means that the amplifier would be operating at a far lower power than it is capable of achieving, since the intermodulation performance will only then be satisfactory. However, this approach is not very power efficient. It is also not cost effective to use high power transistors when they are not being used to their full capabilities [1].

1.3 Thesis Structure

The aim of this research is to design, build and test a two-channel linearised power amplifier for use in the transmitter section of a satellite communication system. The satellite communication system makes use of the Inmarsat global satellite network and enables users to make telephone calls, send and receive faxes and have internet access. The systems have use in the maritime, land and aeronautical sectors.

The specifications of the two-channel amplifier are:

- Operating Frequency: 1626 MHz to 1660 MHz
- 1 dB compression point > 39.5 dBm
- Third order IMD product < -50 dBc when the Peak Envelope Power $= 39.5$ dBm

The structure of the thesis is as follows:

- **Chapter 2: Comparison of Linearisation Techniques:** Various linearisation techniques are presented, compared and discussed.
- **Chapter 3: Alternative Feedforward Method:** An alternative philosophy to the classical Feedforward method is presented.
- **Chapter 4: Practical Power and Gain Analysis:** The requirements for the various Feedforward components are determined and the Feedforward components are selected.
- **Chapter 5: Design of Feedforward Components:** The methods to design the different Feedforward components are presented. The Feedforward component measurements

and simulations are presented.

- **Chapter 6: Integration of the Feedforward System:** All the various Feedforward components are integrated. The Feedforward system measurements are presented.

Chapter 2

Comparison of Linearisation Techniques

There are a number of techniques that can be used to linearise power amplifiers. Some of the most common methods are: Cartesian Feedback, Analog Predistortion, Digital Predistortion and Feedforward [4,5]. Table 2.1 provides a comparison between the various methods of linearisation [2].

Linearisation Method	IMD Correction	Bandwidth	Efficiency
Cartesian Feedback	High	Narrow	High
Analog Pre-Distortion	Low	High	High
Digital Pre-Distortion	Moderate	Moderate	Moderate
Feedforward	High	High	Moderate

Table 2.1: Comparison of Different Methods of Linearisation

2.1 Cartesian Feedback Method

The Cartesian Feedback linearisation method is a form of Modulation Feedback and referred to in some texts as Baseband Feedback. In the Modulation Feedback technique, a sample of the output signal is demodulated and compared with the unmodulated input signal. At the output of the comparator there is a 'pre-distorted' version of the input signal. The pre-distorted signal is then modulated and the amplification of this pre-distorted signal will generate the intended signal at the output of the transmitter [1]. This method of linearisation is not viable for this application since access to the baseband signal is required. In this application, the baseband signal is not accessible.

2.2 Predistortion Method

The Predistortion Method involves the creation of a circuit with a distortion characteristic, which is precisely complementary to the distortion characteristic of the RF power amplifier. The RF amplifier and the distortion characteristic are then cascaded thereby eliminating distortion in the output signal. To implement this technique, the predistortion characteristic has to be determined and then a circuit has to be designed which closely resembles the required characteristic. Two types of Predistortion techniques are known as Analog (RF) Predistortion and Digital (Baseband) Predistortion [1].

In the RF Predistortion technique, the non-linear predistortion circuit operates at the final carrier frequency. The simplest form of a non-linear element, which acts as a predistorter, is a diode [1].

In the Digital Predistortion technique, the predistortion characteristic is typically stored as a table of gain and phase weighting values within a DSP (Digital Signal Processor) in order to pre-distort the baseband information before upconversion [1]. This linearisation method however is also not viable for this application because the baseband signal cannot be accessed.

2.3 Feedforward System

The Feedforward system presented in this thesis was chosen as the linearisation scheme for the two-channel amplifier because it has the best IMD performance and can operate over the widest bandwidth when compared to the other linearisation schemes [5]. Access to the baseband signal is also not required for this linearisation scheme.

The basic block diagram of a Feedforward system is shown in Figure 2.1 [1].

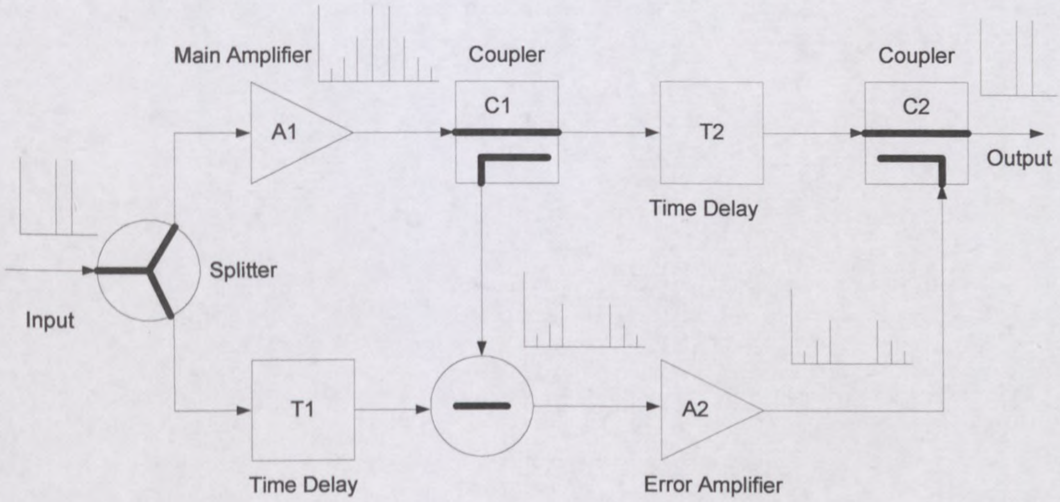


Figure 2.1: Feedforward Power Amplifier

The applied signal is split into two signals. The ratio in the split need not be the same. The signal in the upper channel is amplified by the main power amplifier. The non-linearities of this amplifier result in an output signal which contains intermodulation distortion products. The signal in the lower channel is time-delayed by the same amount as the delay introduced by the amplifier in the upper channel. The directional coupler at the output of the main amplifier provides a sample of the output signal which then is subtracted from the time-delayed signal in the lower channel. The output signal of the subtractor, referred to as the error signal, then contains only the intermodulation distortion products of the main amplifier output signal. Ideally, none of the original energy of the actual signal should remain. This error signal is then amplified by the error amplifier to the required level to eliminate the distortion in the signal in the upper channel. The upper channel signal is also time delayed by an amount equal to the delay caused by the error amplifier and is then applied to the output coupler in anti-phase to the amplified error signal. The final output signal will contain no distortion since the addition of the upper and lower channel signals will ensure that the distortion components add to zero [1].

In order for complete elimination of the intermodulation distortion products to be achieved and assuming the input splitter is an ideal 3 dB hybrid, the following conditions must occur [1]:

- The time delay of the main amplifier (A1) must equal the time delay (T1) in the lower channel.
- The coupling coefficient of the coupler (C1) must equal the gain of the main amplifier (A1).

If these conditions are realised, the signal at the input to the error amplifier will only contain the intermodulation distortion products present in the signal at the output of the main amplifier (A1).

- The time delay of the error amplifier (A2) must be equal to the time delay (T2) in the upper channel.
- The gain of the error amplifier (A2) must be equal to the coupling coefficient of the coupler (C1) plus the coupling coefficient of the coupler (C2).

Under these conditions, when the error signal of the lower channel and the distorted signal of the upper channel are added, the intermodulation distortion products at the output of the main amplifier will be eliminated [1].

Gain and phase compensation circuits, as is shown in Figure 2.2, are required in practice to enable the designer to optimise the circuit [4].

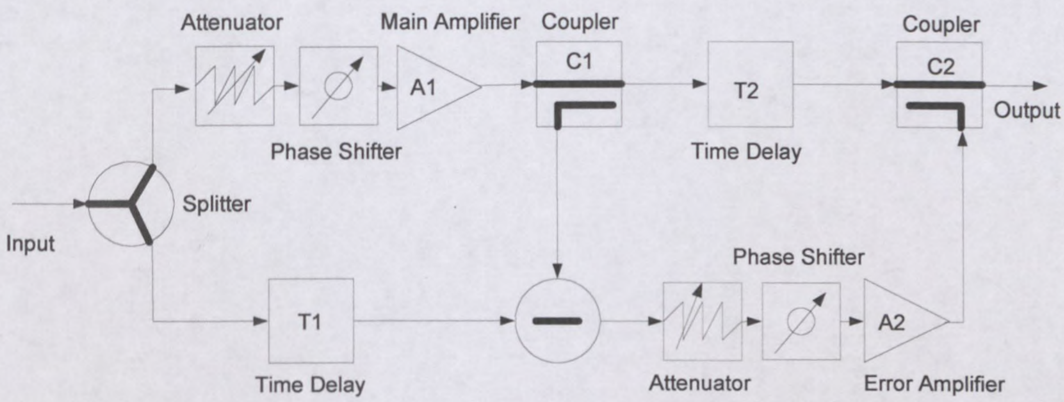


Figure 2.2: Feedforward Power Amplifier with Gain and Phase Compensation Circuits

The various components of the Feedforward system can be implemented in the following ways:

2.3.1 The Input Splitter

The input splitter is used to divide the input signal in an appropriate ratio to feed the main and error amplifiers. The original input signal need not necessarily be split into equal amplitudes. Different types of couplers that could be utilised are [1]:

- An equal ratio 3dB hybrid coupler, which could either be implemented using microstrip or a resistive splitter.
- Directional coupler.

2.3.2 The Main Amplifier

The main amplifier must be designed so that the gain versus frequency response in the frequency band of operation is optimally flat. The phase versus frequency response must be linear in the frequency band of operation. Any deviation from these characteristics will result in a reduction in the suppression of the main signal energy in the error signal path [1].

2.3.3 The Sampling Coupler

The sampling coupler (C1) can be implemented by means of a directional coupler. The requirement for this coupler is that it must have a high directivity to ensure that no reflected power is coupled back into the subtractor. It must also have a flat frequency response across the band of interest [1].

2.3.4 Subtractor

This circuit can be implemented by means of a hybrid coupler. An optimum phase and frequency response is required in order to obtain a high degree of suppression of the main signal energy.

2.3.5 Time Delay Elements

The time delay elements can be implemented by varying lengths of transmission line. An optimum amplitude and phase response is required [1].

2.3.6 Error Amplifier

The error amplifier uses a lower power active device than the main amplifier and the distortion levels must be lower than those of the main amplifier. The distortion level resulting from the error amplifier is critical in determining the overall system performance and is usually the limiting factor in determining the overall amplifier distortion performance [1]. Thus, the error amplifier should be as linear as possible.

2.3.7 Gain and Phase Compensation circuits

These circuits are incorporated to adjust the gain and phase of the signals in the upper and lower channels, to obtain optimum amplifier performance. As shown in Figure 2.2, there is a variable attenuator and variable phase shifter at the input to the main amplifier [1]. This allows the phase and amplitude of the output signal of the amplifier to be adjusted so that the coupled signal at the one input of the subtractor can be adjusted to the correct amplitude and phase. This adjustment ensures that the output signal of the subtractor contains minimal original signal energy. That is, so that only the intermodulation distortion products of the output signal of the main amplifier appear at the input to the error amplifier. The gain and phase compensation circuits can be implemented at alternate points in the circuit and not necessarily before the amplifiers. It is important that they are implemented in the low power sections of the circuit.

The circuits that are used to accomplish the gain and phase compensation necessary, involves the use of a 3dB quadrature hybrid coupler together with either PIN (Positive-Intrinsic-Negative) or varacter diodes respectively [1]. Since the PIN and varactor diodes are non-linear components, these compensation circuits could also cause intermodulation distortion, which would degrade the overall system performance.

2.3.8 Output Coupler

This coupler provides a very important function in the circuit, since it performs the final addition of the error signal and the distorted main amplifier signal. It must also be able to handle the high power signal. Besides having good directivity, the forward transmission loss must be minimal, since any loss may reduce the efficiency of the amplifier. The coupler must provide a constant gain and a linear phase response over the frequency range of operation. The coupler can be implemented by using microstrip technology [1].

Chapter 3

Alternative Feedforward Method

The main disadvantage of the classical Feedforward System, is that the power delivering capability of the error amplifier is wasted in terms of it ever contributing significant amounts of output power. The sole function of the error amplifier is to linearise the main power amplifier output signal. There is also a substantial reduction in the output power capability of the classical Feedforward System due to the loss in the output coupler and other necessary components [6].

There is an alternative philosophy which can be employed in the Feedforward System and which does not have the above mentioned disadvantages [6]. This Alternative Feedforward Method is shown in Figure 3.1.

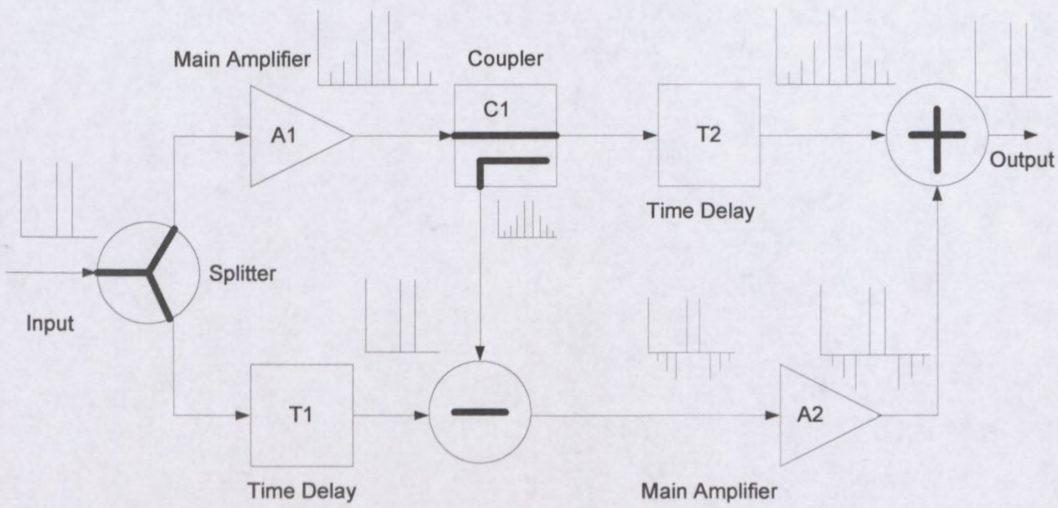


Figure 3.1: Alternative Feedforward Method

This method uses two main amplifiers. The input signal to the second main amplifier is pre-distorted in such a manner that the IMD products in the second amplifier output signal are in anti-phase to the IMD products produced by the first main amplifier. The output signals from the two amplifiers are then added to produce the final output signal which contains increased amplitude fundamental components and no IMD products.

For the subtraction to function optimally, the following must occur:
The two-tone input signal from the time delay (T1) must be greater in amplitude than the signal coupled from the output port of the first main amplifier. The subtraction occurs by ensuring that the coupled signal is 180 degrees out of phase with the two-tone signal from the time delay (T1). Since the two-tone signal from the time delay (T1) is greater in amplitude than that of the coupled signal, its phase will be assumed by the fundamental component at the subtractor output. The two-tone signal from the time delay (T1) contains no IMD products, therefore the IMD products at the output of the subtractor will be 180 degrees out of phase with the IMD products from the coupled signal.

This signal processing ensures that the input to the second main amplifier contains IMD products that are in anti-phase with the actual IMD products that the second main amplifier itself produces at its output port. This is a form of wanted predistortion that occurs at the second main amplifier. If the level of anti-phase IMD products at the input to the second main amplifier are large enough, the output signal of the second main amplifier will contain anti-phase IMD products. For the maximum linearisation to occur, the amplitude of the IMD products from the first amplifier must be exactly the same as the amplitude of the anti-phase IMD products at the output of the second main amplifier.

This alternative Feedforward method was simulated using Agilent Technologies' ADS (Advanced Design System). The non-linear Root model of the MRF282Z MOSFET was downloaded from the Freescale Semiconductor website and the model was imported into ADS. This model of the active device was used for the first and second main amplifiers. Most of the blocks used in the simulation were ideal, since the object of the simulation was to prove the concept. The ADS schematic of the alternative Feedforward method is shown in Figure 3.2.

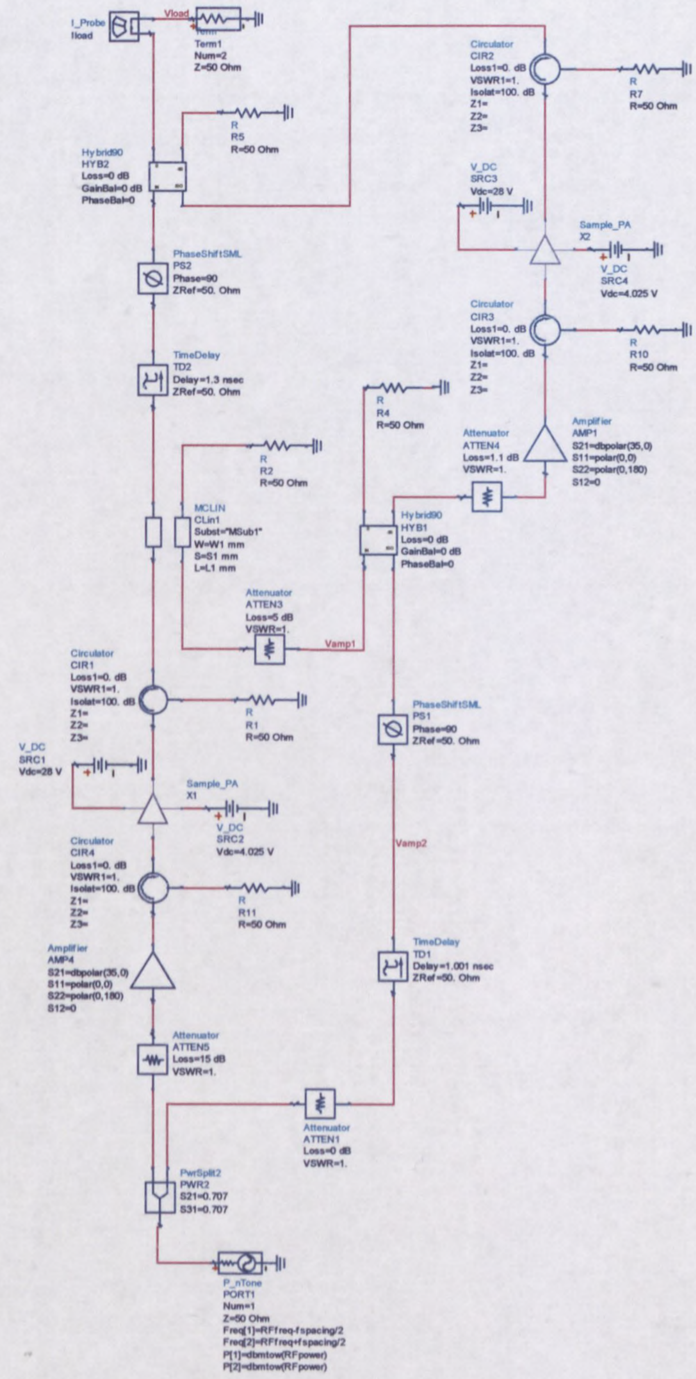


Figure 3.2: ADS Schematic of Alternative Feedforward Method

Simulations revealed a limitation in this alternative Feedforward method. The closer the output signal gets to the 1 dB compression point, the more the non-linearities of the second main amplifier start contributing to its output signal. A stage is reached where the IMD products at the output of the second main amplifier occur in phase with the IMD products produced from the first main amplifier. This causes the IMD products from the two amplifier stages to add. To counter this phenomenon, the input signal to the second main amplifier must be altered so that the level of its anti-phase IMD products are larger. This technique necessitates extreme complex tracking of the input signal to the second main amplifier.

For the above mentioned reason, as well as the fact that it is possible to obtain much better IMD suppression with the classical Feedforward method, the classical Feedforward was selected as the better option to implement.

Chapter 4

Practical Power and Gain Analysis of Feedforward Linearisation

This chapter presents the power and gain analysis of the Feedforward system. This analysis enabled the various Feedforward components to be selected. Before a gain analysis of the entire Feedforward system can be done, there are a number of apriori factors that must be known. It is necessary to first choose the main amplifier active device, the error amplifier output stage active device and then the output coupler. The choice of main and error amplifier active devices determines the final power delivering capability of the complete amplifier. Once these factors are known, the gain analysis can be performed and the remainder of the Feedforward components can be selected. A block diagram of the Feedforward system is shown in Figure 4.1.

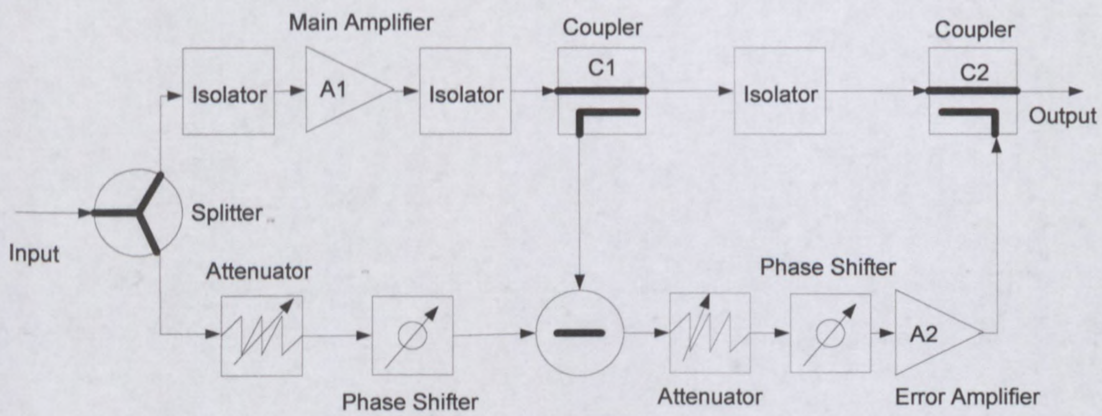


Figure 4.1: Block Diagram of Feedforward System

4.1 Choice of Active Device for the Main Amplifier

When selecting an active device, the peak power of the modulated input signal must be known. This is because the active device must be able to amplify the peak power, even though the amplitude of the peak power may be considerably higher than the average of the signal power.

A two-tone signal has a varying amplitude due to the phase of each tone adding or subtracting. Figure 4.2 shows the time domain representation of a two-tone signal.

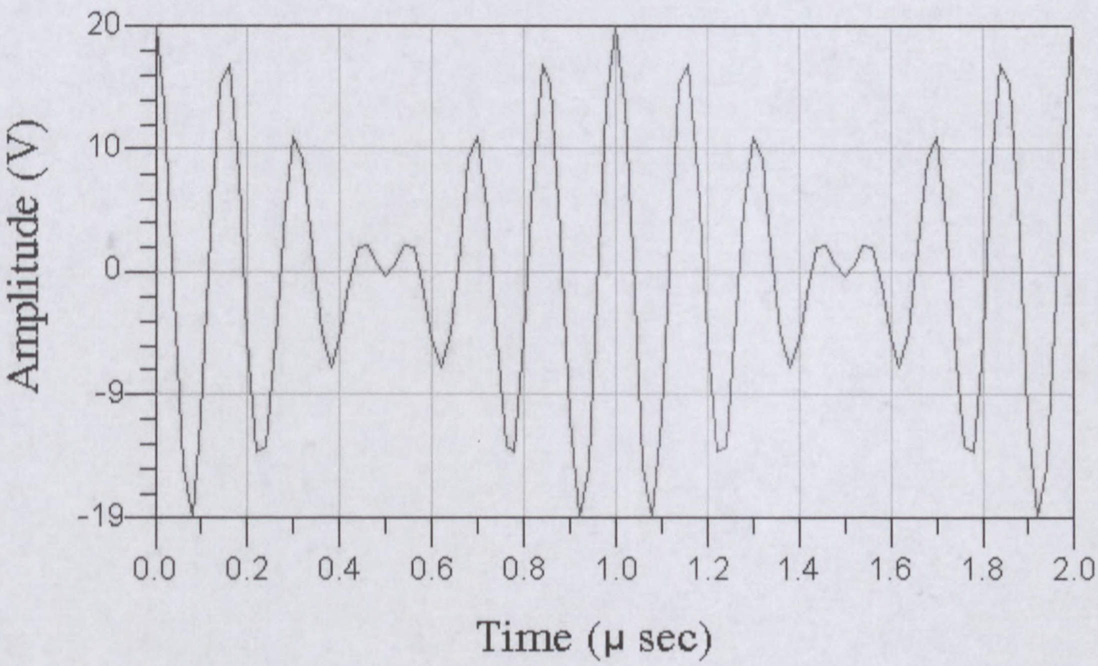


Figure 4.2: Time Domain Representation of a Two-Tone Signal

The average power of a two-tone signal is equal to the power of tone 1 plus the power of tone 2. If the two tones are equal in amplitude, the average power is therefore 3 dB higher than the power of tone 1. The peak power occurs when the two tones are exactly in phase and they add. The amplitude of the peak power is calculated using the following method:

The peak power of tone 1 is:

$$Power_{\text{Tone 1}} = \frac{(V_{\text{Tone 1}})^2}{R} \quad \text{Watt} \qquad 4.1$$

The total peak power is equal to:

$$Power_{\text{Total}} = Power_{\text{Tone 1}} + Power_{\text{Tone 2}} \quad \text{Watt} \qquad 4.2$$

When the tones are exactly in phase, $V_{\text{Tone 1}} = V_{\text{Tone 2}}$, then the peak power is:

$$Power_{\text{Peak}} = \frac{(2 \times V_{\text{Tone 1}})^2}{R} \quad \text{Watt} \qquad 4.3$$

$$= \frac{4 \times (V_{\text{Tone 1}})^2}{R} \quad \text{Watt}$$

Clearly, the total peak power is four times the peak power of tone 1. Therefore, the PEP (Peak Envelope Power) of a two-tone signal is 6 dB higher than the peak power of tone 1.

The active device chosen for the main amplifier stage (A1) was the MRF282Z MOSFET. This is a 10 W device manufactured by Freescale Semiconductor. This device was chosen because it was readily available and it can deliver the required amount of power to comply with the amplifier specification.

4.2 Choice of Active Devices for the Error Amplifier

4.2.1 Choice of Active Device for the Output Stage of the Error Amplifier

To select a suitable active device for the output stage of the error amplifier (A2), the amplitude of the worst case IMD products of the main amplifier, as well as the output coupling factor must be known. The main amplifier stage was simulated on ADS to determine the amplitude of the worst case IMD products. The Freescale Semiconductor non-linear Root model of the MRF282Z MOSFET, was used as the active device in this simulation.

When the output power of the main amplifier was 35.7 dBm per tone, the PEP (Peak Envelope Power) was 41.7 dBm (14.8W). At this stage the active device was driven slightly more than it's 1dB compression point. The 3^{rd} order IMD products were situated 24dB below the amplitude of the two-tone signal. The amplitude of the 3^{rd} order IMD products were calculated at 11.7dBm per tone. The PEP of the 3^{rd} order IMD products was 17.7dBm.

A coupling factor of 11 dB was chosen for the output coupler and the reasons therefore are explained in detail in the next section. Thus, the PEP that the error amplifier must produce to cancel the IMD caused by the main amplifier is:

$$11.7\,\text{dBm} + 11\,\text{dB} = 28.7\text{dBm}$$

In practice, some of the fundamental signal is presented at the input to the error amplifier, which necessitates that the error amplifier have more than sufficient power handling capabilities. The error amplifier should also operate well below it's maximum output power, so that it's IMD products do not have a significant effect on the Feedforward correction. The active device chosen was the Freescale MRF281Z MOSFET. This device was chosen because it has a 1dB compression point of 36 dBm.

4.2.2 Choice of Active Device for the Error Amplifier Driver

The required gain of the error amplifier is calculated in Section 4.4: Gain Analysis. The minimum gain required by the error amplifier was 36dB plus the minimum attenuation of the cascaded variable attenuator and phase shifter. That is, the minimum gain required is:

$$36\,\text{dB} + 4.5\,\text{dB} = 40.5\,\text{dB}$$

In order to attain this high gain, a three stage error amplifier was required. Since the error amplifier comprised more than one stage, the driver and pre-driver active devices were selected such that they provided the required amount of output power. The error amplifier output stage contributed 16dB of gain, thus the error amplifier driver must have a 1dB compression point of at least:

$$36\,\text{dBm} - 15\,\text{dB} = 21\,\text{dBm}$$

The performance parameters of the following active devices were scrutinised in order to select the appropriate devices:

- The NGA 386 MMIC. The 1dB compression point of this device was 15dBm, which was insufficient and therefore it was not selected.

- The NE5520279a MOSFET. This device had a 1dB compression point of 32dBm. This was a slight overkill in terms of the required power capability but was considered since there were few other active devices available which could meet the requirements.

- The MRF281Z MOSFET. As explained in the previous section, this device has a 1dB compression point of 36dBm. This device is also a slight overkill in terms if the required power capability but was considered because it is already used in the design.

The disadvantage of the NE5520279a MOSFET device was that it required a $5V_{dc}$ supply, while the other power devices used in the design required $28V_{dc}$. The DC power dissipation of the NE5220279a MOSFET was compared with that of the MRF281Z MOSFET. This comparison is shown in Table 4.1.

Device	DC Power Dissipated under Bias Conditions	DC Power Dissipated at 21 dBm Output Power
NE5520279a	2 Watt	3 Watt
MRF281	0.7 Watt	1.4 Watt

Table 4.1: Comparison of DC Power Dissipation Between Two Active Devices

Even though the NE5520279a had a lower output power capability, it's efficiency at the specific power of operation is lower than that of the MRF281Z.

The MRF281Z was thus chosen as the active device for the error amplifier driver stage. Even though this was an overkill, the MRF281Z has a better DC efficiency than the NE5520279a plus it operates off a $28V_{dc}$ supply.

4.2.3 Choice of active Device for the Error Amplifier Pre-driver

Since the error amplifier driver stage provided 16dB of gain, the error amplifier pre-driver must have a 1dB compression point of at least:

$$21\,\text{dBm} - 16\,\text{dB} = 5\,\text{dBm}$$

Agilent's High Linear MGA-53543 Amplifier, was selected. It has a 1dB compression point of 18 dBm and was readily available.

4.3 Choice of the Output Coupling Factor

The output coupling factor (C2) is one of the most critical elements in the Feedforward design. The choice thereof is a tradeoff between insertion loss, error amplifier power capability and efficiency. The weaker the coupling factor, the less the insertion loss, but the higher the power level requirement of the error amplifier.

The first step in determining the optimum output coupling factor, is to calculate the insertion loss of the coupler versus coupling factor. This is accomplished using equation 4.4 [6].

$$y = \sqrt{(1-B^2)} \tag{4.4}$$

where y is the transmission loss

B is the coupling factor

Figure 4.3 shows a graph of coupling factor in dB versus insertion loss in dB.

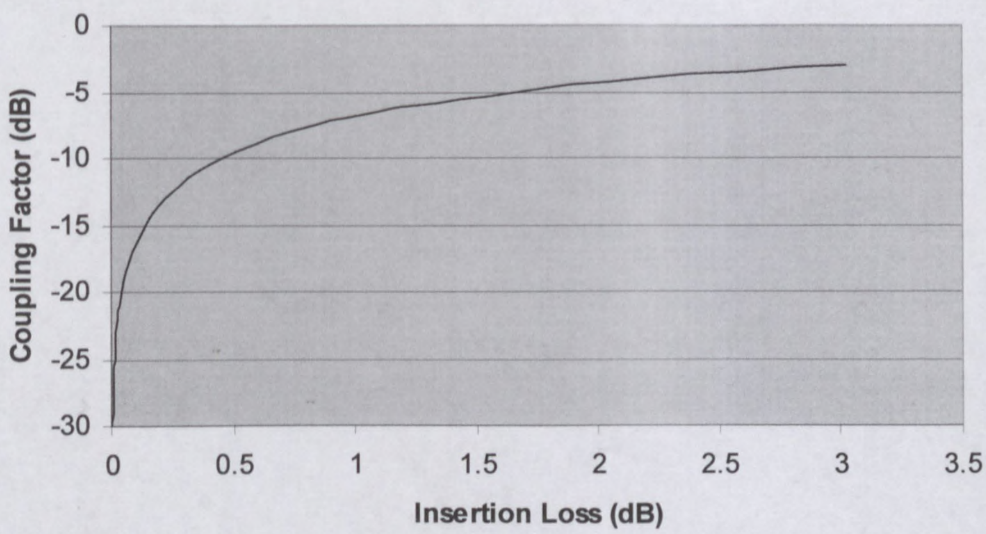


Figure 4.3: Coupling Factor versus Insertion Loss

The next step is to determine the efficiency of the Feedforward amplifier versus coupling factor. This was accomplished by using the following method:

A two-tone Harmonic-Balance simulation, including the main amplifier, error amplifier and output coupler was performed on ADS. The Freescale Root models for the MRF282Z and MRF281Z were used as the active devices for the main and error amplifiers respectively. The DC current drawn from the main amplifier and from the error amplifier was simulated for a range of output coupling factors and output power levels. The efficiency of the Feedforward system was then calculated for each coupling factor and power level pair using equation 4.5.

$$\eta = \frac{P_{\mathrm{RF}} \text{ at Coupler Output}}{I_{\mathrm{dc}} \text{ (Main Amp)} + I_{\mathrm{dc}} \text{ (Error Amp)}} \qquad 4.5$$

If the output coupling factor is too weak, a low insertion loss results, but the error amplifier will have to provide a high RF output power and therefore consume a large amount of DC power. If the coupling factor is too strong, a high insertion loss results and the error amplifier would be operating far below it's maximum power delivering capability. Therefore, the error amplifier will be operating far below it's optimum efficiency point.

A situation will be reached where the optimum coupling factor will result in the best overall efficiency. The efficiency versus coupling factor for various output power levels are shown in Figure 4.4, Figure 4.5 and Figure 4.6 respectively.

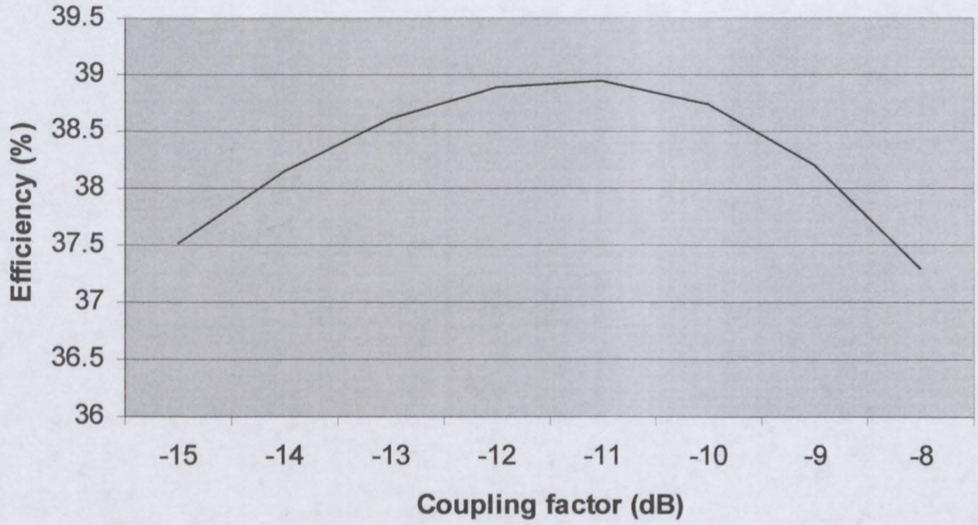


Figure 4.4: Efficiency vs Coupling Factor for Main Amplifier Peak Power of 42 dBm

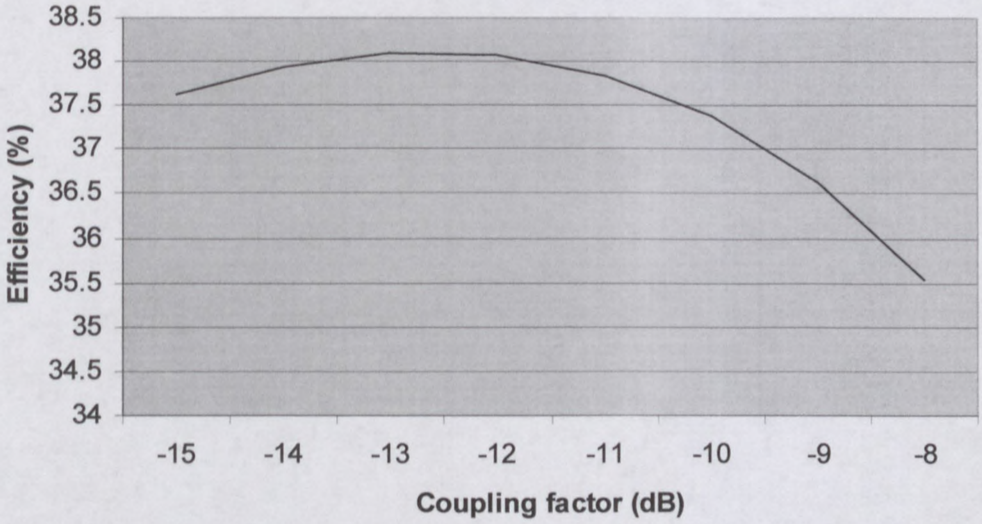


Figure 4.5: Efficiency vs Coupling Factor for Main Amplifier Peak Power of 41 dBm

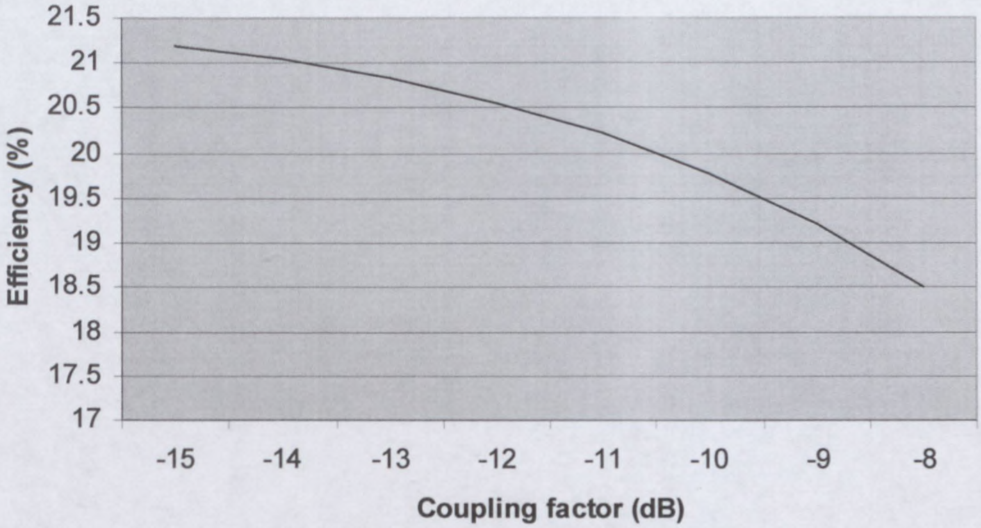


Figure 4.6: Efficiency vs Coupling Factor for Main Amplifier Peak Power of 38 dBm

At the lower main amplifier RF output power of 38 dBm, the optimum coupling factor is relatively weak. This is due to the output error amplifier operating at a low efficiency and increasing the error amplifier RF output power does not significantly increase efficiency.

At the higher main amplifier RF output power of 42 dBm, the optimum coupling factor is stronger. This is due to the error amplifier being driven more and operating at a more optimum efficiency.

A coupling factor of 11dB was chosen because it provides good efficiency at all RF power levels and the error amplifier still operates in it's linear region. This coupling factor of 11 dB also allows for the error amplifier to be driven somewhat harder and hence the capability of accommodating a higher level of fundamental signal power at it's input port. This allows for sub-optimum operation by the subtractor whereby 100% elimination of the fundamental signal components is not accomplished.

4.4 Gain Analysis

Figure 4.7 shows the block diagram of the Feedforward system with the first and second loops identified.

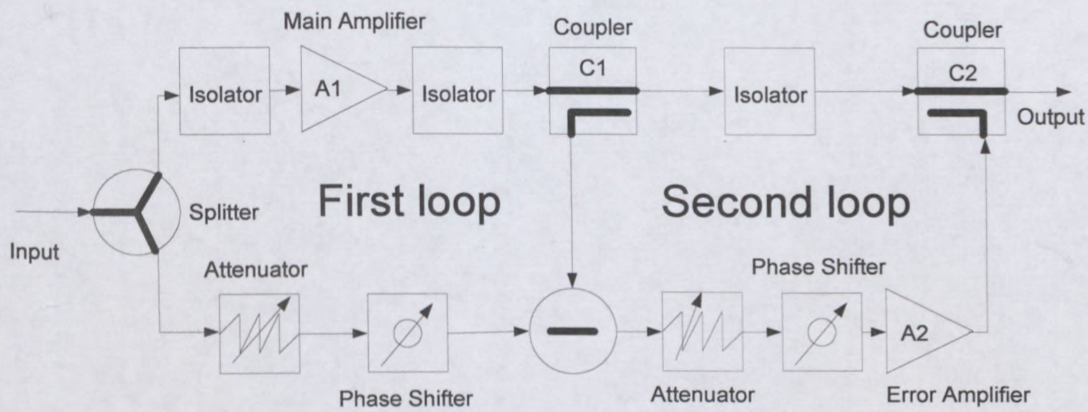


Figure 4.7: Block Diagram of Feedforward System with First and Second Loops Identified

An integrated circuit hybrid coupler was selected for the input splitter, as well as for the subtractor. The device selected was manufactured by Murata (Part number: LDC311G6003B-601) and it was chosen because of it's small size and it's availability.

It was decided that the variable attenuator and phase shifter in the first loop should be situated in the lower channel and not in the upper channel before the main amplifier. If the variable attenuator was situated at the input to the main amplifier and the variable attenuator was altered, the gain of the complete Feedforward system would change.

For optimum subtraction to occur, the gain of the main amplifier (A1) less the coupling factor (C1) must equate to the attenuation of the lower channel of the first loop. For the complete cancellation of the IMD products at the final output port to take place, the gain of the error amplifier (A2) must equal the coupling factor (C1) plus the coupling factor (C2). In both the first and second loop, the coupling factor (C1) influences the gain / attenuation required.

Since the minimum attenuation provided by the cascaded variable attenuator and variable phase shifter is 4.5dB, the coupling factor (C1) must be weak enough such that the attenuation required by the lower channel of the first loop is in excess of 4.5dB. The coupling factor (C1) should be even weaker than this so that the variable attenuator operates near the mid-point of it's range, enabling it to have a wide range of adjustment. The disadvantage of selecting a weak coupling factor (C1) is that the weaker the coupling factor selected, the higher the gain required by the error amplifier. A trade-off must be effected and an optimum coupling factor must be selected. Table 4.2 shows values of coupling factor versus attenuation required by the lower channel of the first loop and the gain required by the error amplifier.

Coupling Factor (C1)	Attenuation Required by Lower Channel of First Loop	Gain Required by Error Amplifier (A2)
18dB	5.2dB	32dB
19dB	6.2dB	33dB
20dB	7.2dB	34dB
21dB	8.2dB	35dB
22dB	9.2dB	36dB
23dB	10.2dB	37dB
24dB	11.2dB	38dB
25dB	12.2dB	39dB
26dB	13.2dB	40dB

Table 4.2: Coupling Factor (C1) versus Attenuation Required by Lower Channel of First Loop and Gain Required by Error Amplifier (A2)

The optimum coupling factor (C1) was chosen to be 22dB. With this value of the coupling factor, the forward transmission loss is low. The loss of the coupler did not significantly reduce the main amplifier output power.

An isolator was placed before and after the coupler (C1). These isolators are required to ensure a good 50Ω termination on the input and output ports of the coupler. The isolator on the output port of the coupler enhanced the directivity of coupler (C1). This isolator also prevents any reverse power, which is inserted into the output coupler (C2) by the error amplifier, from feeding back into coupler (C1) and thereby influencing the amplitude of the signal being fed to the subtractor.

Chapter 5

Design of Feedforward Amplifier Components

This chapter describes the methods used to design the following Feedforward components:

- Main Amplifier
- Error Amplifier
- Output Coupler and Main Amplifier Coupler
- Variable Attenuator
- Variable Phase Shifter

5.1 Main Amplifier

To obtain maximum power from an active device, its input and output ports are not matched to provide maximum gain. Rather, the output port of the active device is purposely mismatched to obtain maximum power to the load.

The reason for this is that an active device such as a FET, has a certain maximum current and voltage rating. If the active device is matched for maximum gain, it may reach its maximum voltage rating before it reaches its maximum current rating and thus will not deliver maximum power to the load. To deliver maximum power to the load, the output port of the active device must be presented with a specific load impedance such the maximum current and voltage ratings of the active device are simultaneously reached.

5.1.1 Efficiency versus Class of Amplifier

The efficiency of an RF amplifier is a very important parameter. Low efficiency results in excessive amounts of heat dissipated by the active device and excessive amounts of DC current drawn by the device. In high power applications, RF power amplifiers are never used in Class-A mode of operation. They are biassed for periods less than 360° of the input signal, therefore conserving DC power. This signal then gets re-generated by a tank circuit at the output of the active device. In practice, active devices are designed to operate in a specific class of operation. The datasheet of the device specifies the respective bias voltages and currents required to operate the device in that specific class of operation.

5.1.2 Design Method

The Freescale MRF282Z MOSFET was selected as the active device for the main amplifier as explained in the previous chapter. The basic steps designing a RF power amplifier are:

- Determine the specifications
- Decide on the biasing point
- Choose the load and source reflection coefficients for maximum power delivery to the load
- Check stability of device
- Design of matching networks
- Design of bias circuitry
- Simulate and Measure

5.1.3 Specifications of Main Amplifier

The specifications for the main amplifier are:

- Power Gain variation < 1.1dB over the frequency band of 1626 MHz to 1660 MHz
- 1 dB compression point > 40 dBm
- Gain > 13 dB
- Third order IMD product < -27 dBc when the PEP $= 40$ dBm

5.1.4 Bias Point Selection

This active device was designed to operate in Class-AB mode. The biasing points were specified on the datasheet as:

- $V_{DS} = 28\,V_{dc}$
- $I_{DS} = 75\,mA$

5.1.5 Load and Source Reflection Coefficient Selection

Three methods were investigated in determining the load and source reflection coefficients. These method were:

- Optimisation
- Load Pull Simulation
- Using Reflection Coefficients obtained from the manufacturer's datasheet

5.1.5.1 Optimisation

This method can be used if the non-linear model of the active device is accurate. The simulation can be setup to optimise the amplifier for a specific gain. The simulation is setup to optimise the source and load reflection coefficients so that a specific power gain is achieved, up to a specific input power level. The maximum input power level is adjusted to cause an output power that exceeds the rated output power. In this way the simulation optimises the 1dB compression point while keeping the gain constant. This method is illustrated in Figure 5.1. In this instance, the amplifier was optimised for 13dB gain.

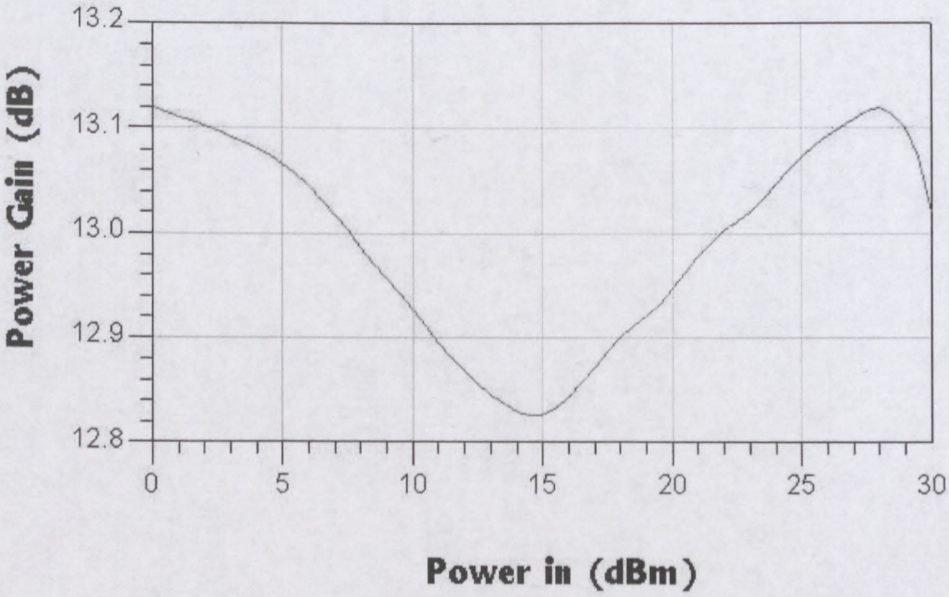


Figure 5.1: Power Gain vs Input Power

The optimisation procedure must be repeated for various appropriate levels of power gain. The following factors in the performance of the amplifier at various levels of power gain must then be compared:

- Output power at the 1dB compression point
- Gain flatness vs input power
- Efficiency
- Level of the IMD products when a two-tone signal is applied

When the optimum trade-off between these factors has been decided on, the source and load reflection coefficients can be selected and the matching networks are then designed.

5.1.5.2 Load Pull Simulation

The load pull simulation selects a number of load impedances and then for each load impedance, simulates the resulting output power. Load pull contours are then drawn on the Smith Chart. The load pull contours illustrate a locus of impedances that may be presented to the output port of the active device, to obtain a specific output power level. Figure 5.2 shows the different load pull contours on the Smith Chart.

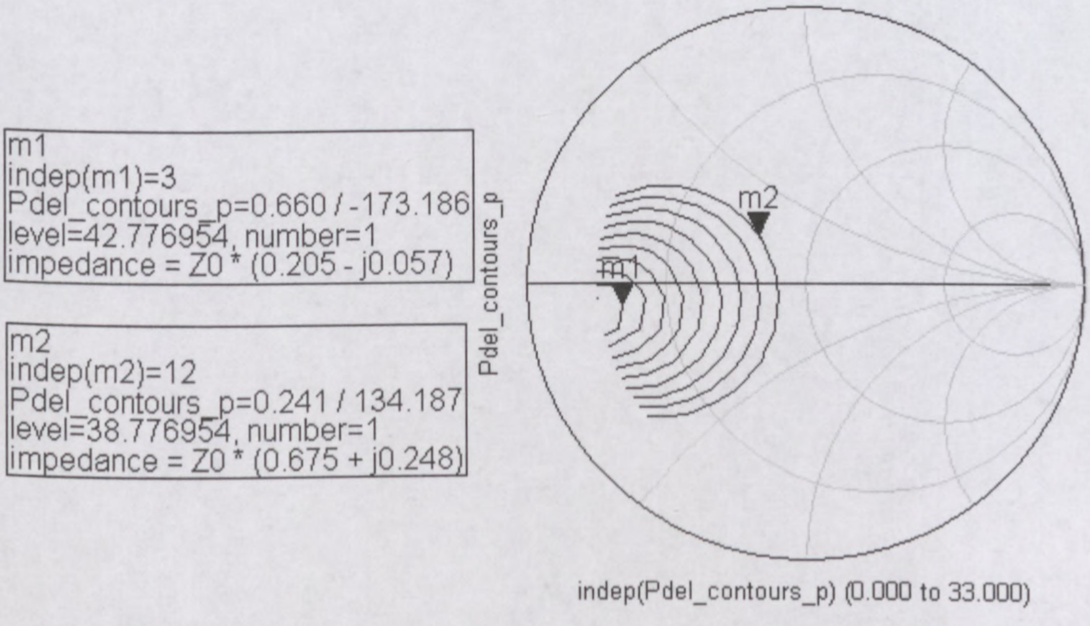


Figure 5.2: Load Pull Contours

Marker 1 is situated on the load pull contour that results in the maximum power delivered to the load. Marker 2 is situated on the load pull contour that results in much less power delivered to the load. The marker information blocks state the level of the output power for each marker as well as the specific impedance at that marker.

The load impedance that results in the maximum output power, is then selected. For this simulation, the input power has to be specified. If the input power is too low and the load impedance is chosen for a maximum output power, the amplifier will then be designed for maximum gain and not maximum power delivery to the load. In load pull simulations, it is important that the input power is set such that the amplifier is forced into compression. In this way it will be possible to see which load impedances provide the maximum output power.

Although the load impedance is the major contributing factor to the output power level, the source impedance also plays a role. This load pull simulation does not vary the source impedance at all. Therefore, for this simulation to be most useful, the optimum source impedance must be presented to the active device while varying the load impedance. The optimum source impedance could be obtained from the datasheet or by doing a source pull simulation.

5.1.5.3 Obtain Reflection Coefficients from Datasheet

The datasheet specifies the recommended source and load impedances for maximum power delivery to the load. It was decided that the matching networks would be designed according to the impedances recommended on the datasheet. According to the datasheet, at 1.64 GHz:

$$Z_L = 4.25 - j0.5\,\Omega$$

$$Z_S = 2.4 + j0.3\,\Omega$$

5.1.6 Stabilisation of Device

The device was found to be potentially unstable over the frequency band of operation, since the Rollet stability factor K was simulated and was found to be less than 1. Load and source stability circles were then drawn on the Smith Chart to confirm that the respective load and source impedances that were chosen, did not fall into areas of instability.

5.1.7 Design of Input and Output Matching Networks

In order to achieve optimum matching, it was decided to use a tunable microstrip matching network structure for both the input and output matching networks. A double open circuit stub matching network was selected, the topology of which is illustrated in Figure 5.3.

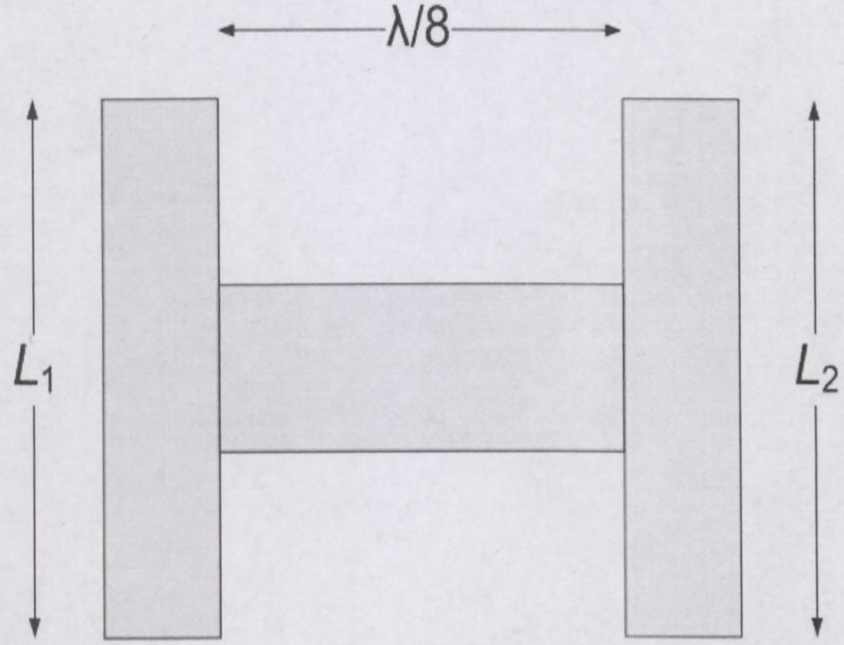


Figure 5.3: Structure of Matching Network

By adjusting the length of the two stubs (L_1 and L_2) it is possible to obtain a reasonably large range of input and output impedances. If the length of the first stub is varied, the impedance presented by the ports of the matching network structure is varied left and right on the Smith Chart (the resistive component is varied). If the length of the second stub is varied, the impedance presented by the ports of the matching network structure is varied up and down (the reactive component is varied). In this way, it is possible to obtain a large range of port impedances. The characteristic impedances of the lines were determined by a computer-aided optimiser. The range of obtainable impedances is illustrated in Figure 5.4.

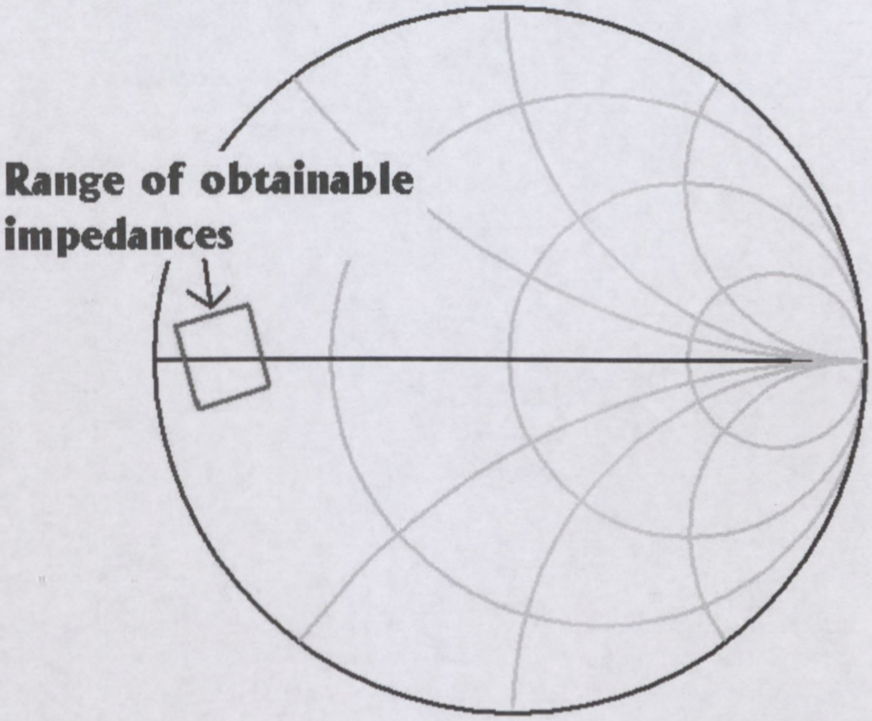


Figure 5.4: Range of Obtainable Impedances

5.1.8 Design of Bias Circuitry

Figure 5.5 illustrates the bias circuitry of the main amplifier.

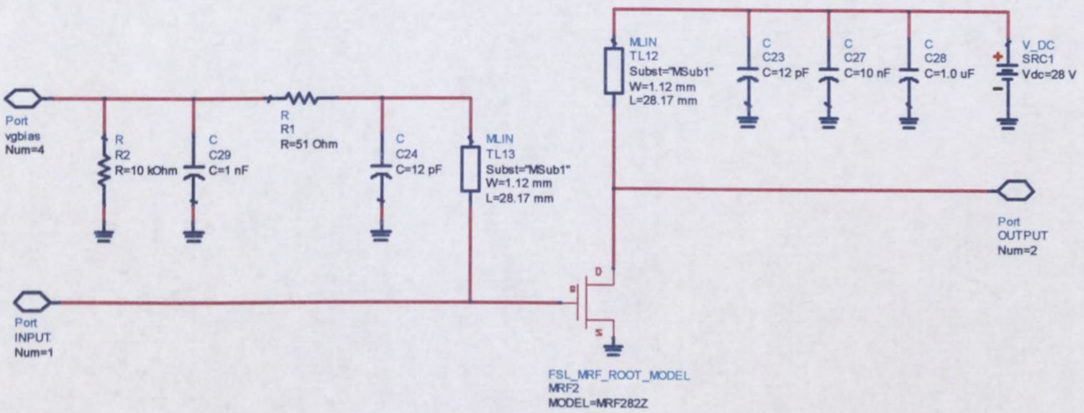


Figure 5.5: Bias Circuitry of Main Amplifier

A shorted quarter-wave length of transmission line is used to feed the DC to both the source and gate of the MOSFET respectively. This acts as a high impedance to RF, but a low impedance for DC. The 12pF capacitor acts as a short-circuit to the RF signal. On the drain bias circuitry of the device, there is a 1μF and a 10nF capacitor in parallel with the 12pF capacitor. These capacitors act as a low impedance source for the low frequency components contained in a modulated RF signal. On the gate bias circuitry, there a series 51Ω resistor with a shunt 1nF capacitor. This circuitry ensures that the gate of active device is presented with a resistive impedance at low frequencies. This ensures the stability of the active device at low frequencies. The 10k resistor in parallel with the shunt 1nF capacitor ensures that the gate of the MOSFET is never floating, since the gate is very susceptible to ESD (Electro-Static Discharge) damage. To bias the device, 28V DC is applied to the drain of the device. While measuring the drain current, the gate voltage is increased slowly until the required amount of gate current is measured.

5.1.9 Simulation and Measured Results

Agilent Technologies ADS (Advanced Design System) was used to do the non-linear simulations of the main amplifier. The non-linear Root model of the MRF282Z MOSFET was downloaded from the Freescale Semiconductor website and the model was imported into ADS. After the amplifier was simulated, it was then built and the matching networks were tuned for maximum output power.

To tune the amplifier, the following procedure was followed: The amplifier was biased and an input RF signal was applied to the input port. The amplitude of the input signal was sufficient such that the amplifier was driven slightly into compression. The output power was measured by a power meter. A power reading was taken, and then the amplifier was switched off. The output matching network was then tuned by either lengthening or shortening the stubs. After every iteration, the output power of the amplifier was then measured and the difference in output power noted. This process was repeated until the best output power performance of the amplifier was obtained. The input matching network was tuned to optimise the gain of the amplifier. A photo of the prototype main amplifier is shown in Figure 5.6.

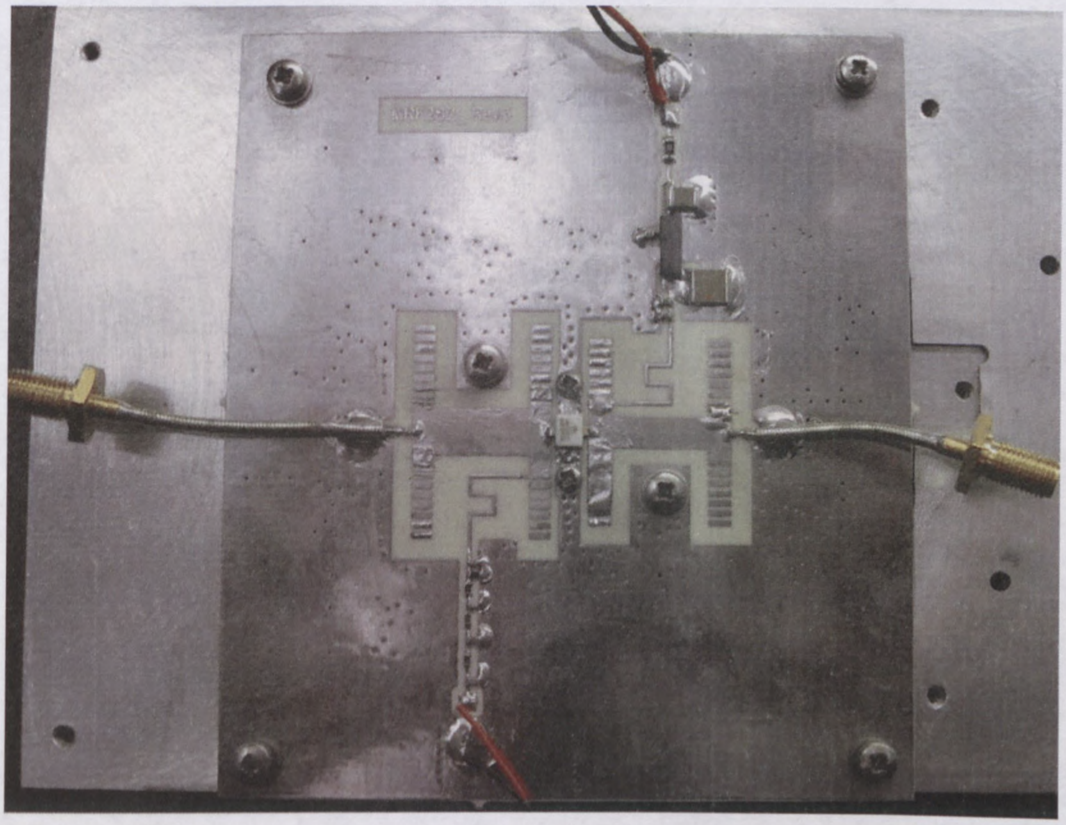


Figure 5.6: Prototype Main Amplifier

The simulated results of the main amplifier for a swept power single tone input signal at 1643 MHz are presented in Table 5.1. The measured results of the main amplifier for a swept power single tone input signal at 1643 MHz are presented in Table 5.2.

Output Power (dBm)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
36.99	13.99	0	37.6
37.94	13.94	-0.05	41.64
38.87	13.87	-0.12	45.84
39.7	13.7	-0.29	49.77
40.43	13.43	-0.56	52.94
40.7	13.2	-0.8	53.85
40.96	12.96	-1.03	54.76
41.17	12.67	-1.33	55.08
41.37	12.37	-1.62	55.4
41.65	11.65	-2.34	55.5

Table 5.1: Simulated Single Tone Results of Main Amplifier

Output Power (dBm)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
36.9	14.1	0	35.5
38.1	14	-0.1	40.26
39	13.9	-0.2	44.54
39.76	13.76	-0.34	48.17
40.35	13.55	-0.55	51.04
40.75	13.25	-0.85	52.86
41	12.9	-1.2	53.58
41.22	12.62	- 1.48	54.26
41.37	12.27	- 1.83	54.25
41.57	11.57	-2.53	54.14

Table 5.2: Measured Single Tone Results of Main Amplifier

The simulated 1 dB compression point is 40.96 dBm and the measured 1 dB compression point is approximately 40.9 dBm. The simulated output power correlates very well with the measured output power. The simulated efficiency also correlates very well with the measured efficiency. Both the 1 dB compression point and gain specification have been met.

The simulated and measured results of the main amplifier for a swept power two-tone input signal, are presented in Tables 5.3 and 5.4 respectively. The tones were set at 1643 MHz and 1644 MHz.

Output PEP (dBm)	IMD3 (dBc)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
38.2	-41	14.22	0	34
39.13	-38.4	14.13	-0.1	37.4
40	-32.3	13.97	-0.25	40.5
40.7	-27.3	13.7	-0.52	43.2
41.3	-23.5	13.31	-0.91	45.3
41.8	-20.6	12.8	-1.42	46.7

Table 5.3: Simulated Two-Tone Results of Main Amplifier

Output PEP (dBm)	IMD3 (dBc)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
38.3	-39	14.3	0	33.03
39.2	-36	14.2	-0.1	36.32
40.2	-32	14.2	-0.1	41.25
41.1	-28.1	14.1	-0.2	45.67
41.7	-24.2	13.7	-0.6	47.62
42.3	-21.2	13.3	-1	50.07

Table 5.4: Measured Two-Tone Results of Main Amplifier

The two-tone simulated results correlate very well with the measured results. When the PEP power is 40 dBm, the simulated third order IMD product is 32.3 dB below the carrier whereas the measured third order IMD product is about 32 dB below the carrier. These simulated and measured third order IMD products both exceed the required specification of less than 27 dB for a PEP of 40 dBm.

The simulated and measured gain vs input signal frequency responses are shown in Figures 5.7 and 5.8 respectively. The gain was measured using an input power level of -20 dBm. This power level was chosen so that the amplifier would be operating in its linear region.

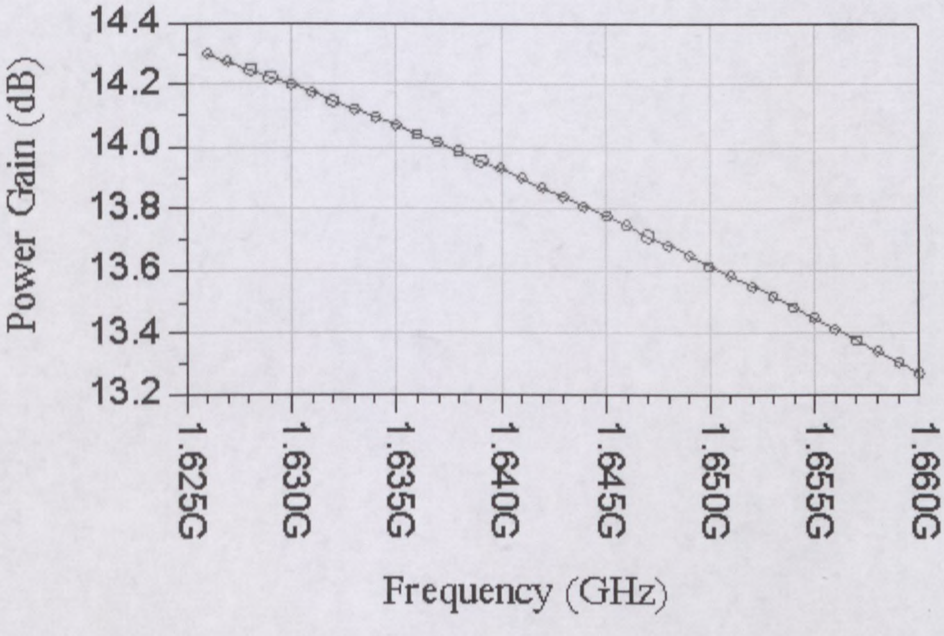


Figure 5.7: Simulated Power Gain vs Frequency of Main Amplifier

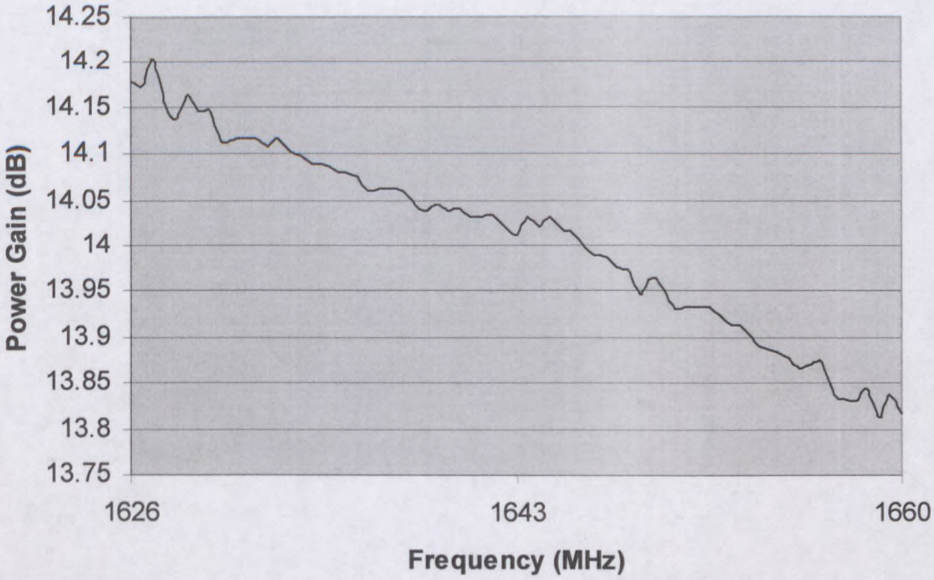


Figure 5.8: Measured Power Gain vs Frequency of Main Amplifier

The simulated power gain variation is 1 dB, whereas the measured power gain variation is 0.35 dB over the frequency range. Both the simulated and measured results exceed the required specification of less than 1.1 dB gain variation over the frequency range.

The measured load impedance presented to the main amplifier active device by the output matching network is shown on a Smith Chart in Figure 5.9. The measured source impedance presented to the main amplifier active device by the input matching network is shown on a Smith Chart in Figure 5.10. The measurement of each matching network was done using the following procedure: The active device was removed. A coaxial cable was soldered onto the active-device side of the matching network. The other port of the matching network was terminated with a 51Ω resistor. The network analyser was setup to display a Smith Chart. The centre conductor of the coaxial cable that was soldered onto the matching network, was short-circuited. The electrical length parameter of the network analyser was adjusted until the measured impedance was situated at the short-circuit point on the Smith Chart. The short-circuit on the coaxial cable was removed and the impedance of the matching network was measured.

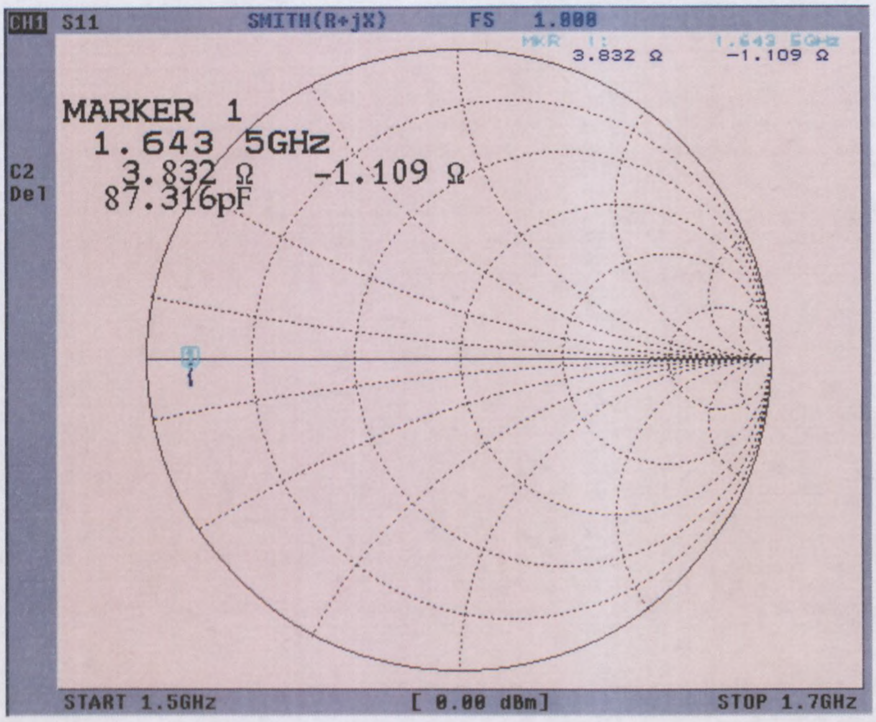


Figure 5.9: Load Impedance

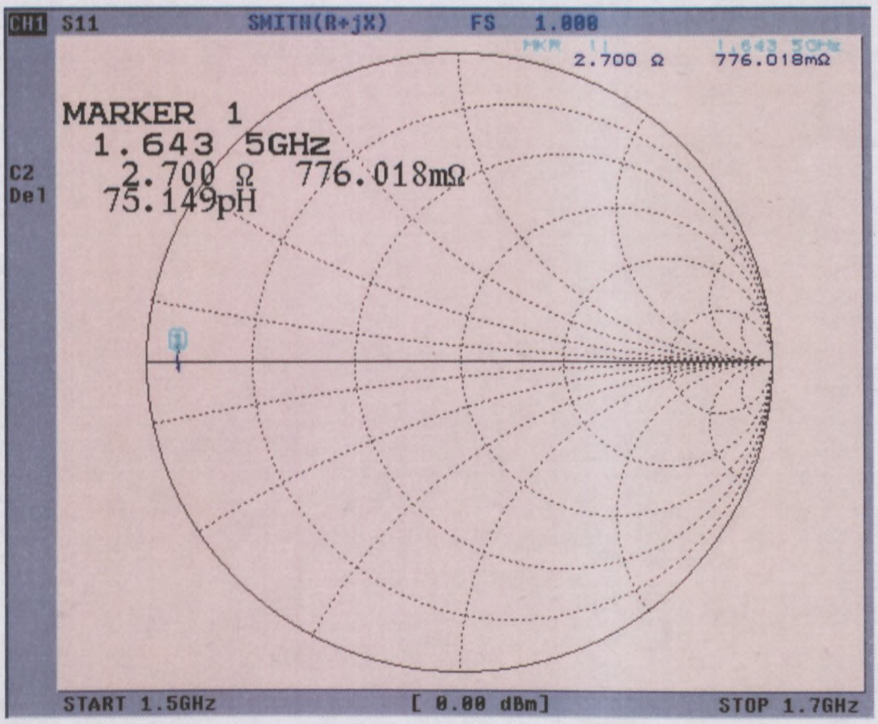


Figure 5.10: Source Impedance

The measured impedances at 1.644 GHz are:

$$Z_L = 3.8 - j\,1.1\,\Omega$$

$$Z_S = 2.7 + j\,0.78\,\Omega$$

These impedances were the optimised values obtained due to the tuning of the stubs of the matching networks for maximum output power. These impedances correlate quite well with the impedances that were recommended by the manufacturer's datasheet.

5.2 Error Amplifier

The power gain requirement of the error amplifier resulted in the use of three active devices. It was decided to use the MGA53543 MMIC as the pre-driver active device, as was explained in the previous chapter. The design of the MGA53543 MMIC pre-driver stage is not presented here since the design was obtained directly from the datasheet example. Previous experience with this device deemed it not necessary to evaluate it on its own, but it was evaluated in the final feedforward system. The MRF281Z MOSFET was selected as the active device for both the driver and final stages of the error amplifier. Therefore, one design will be done using the MRF281Z. This design will then be used for both the output and driver stages. This design is presented in the following section.

5.2.1 Specifications of the MRF281Z stage

The specifications of the MRF281Z stage are:

- Power Gain variation $<$ 1.1dB over the frequency band of 1626 MHz to 1660 MHz
- 1 dB compression point $>$ 36 dBm
- Gain $>$ 15 dB
- Third order IMD product $<$ -27 dBc when the PEP $=$ 36 dBm

5.2.2 Bias Point Selection

This active device was designed to operate in Class-AB mode with the biasing points specified on the datasheet as:

- $V_{DS} = 28\,\mathrm{V}_{dc}$
- $I_{DS} = 25\,\mathrm{mA}$

5.2.3 Load and Source Reflection Coefficient Selection

It was decided to design the amplifier using the source and load impedances as specified on the datasheet. According to the datasheet, at 1.64 GHz:

$$Z_L = 14.7 + j\,12.5\,\Omega$$
$$Z_S = 3.1 + j\,3.8\,\Omega$$

5.2.4 Stabilisation of Device

The device was found to be stable over the frequency band of operation, since the Rollet stability factor K was simulated and was found to be greater than 1.

5.2.5 Design of Input and Output Matching Networks

In order to achieve optimum matching, it was decided to use an open circuit double stub tunable microstrip matching network structure for both the input and output matching networks.

5.2.6 Design of Bias Circuitry

The same bias circuitry that was designed for the main amplifier, was used for the MRF281Z amplifier.

5.2.7 Simulation and Measured Results

ADS was used to do the non-linear simulations of the MRF281Z amplifier. The non-linear Root model of the MRF281Z MOSFET was downloaded from the Freescale Semiconductor website and the model was imported into ADS. After the amplifier was simulated, it was then built and the matching networks were tuned for maximum output power. The same procedure that was used to tune the main amplifier, was also used here. A photo of the prototype MRF281Z amplifier is shown in Figure 5.11.

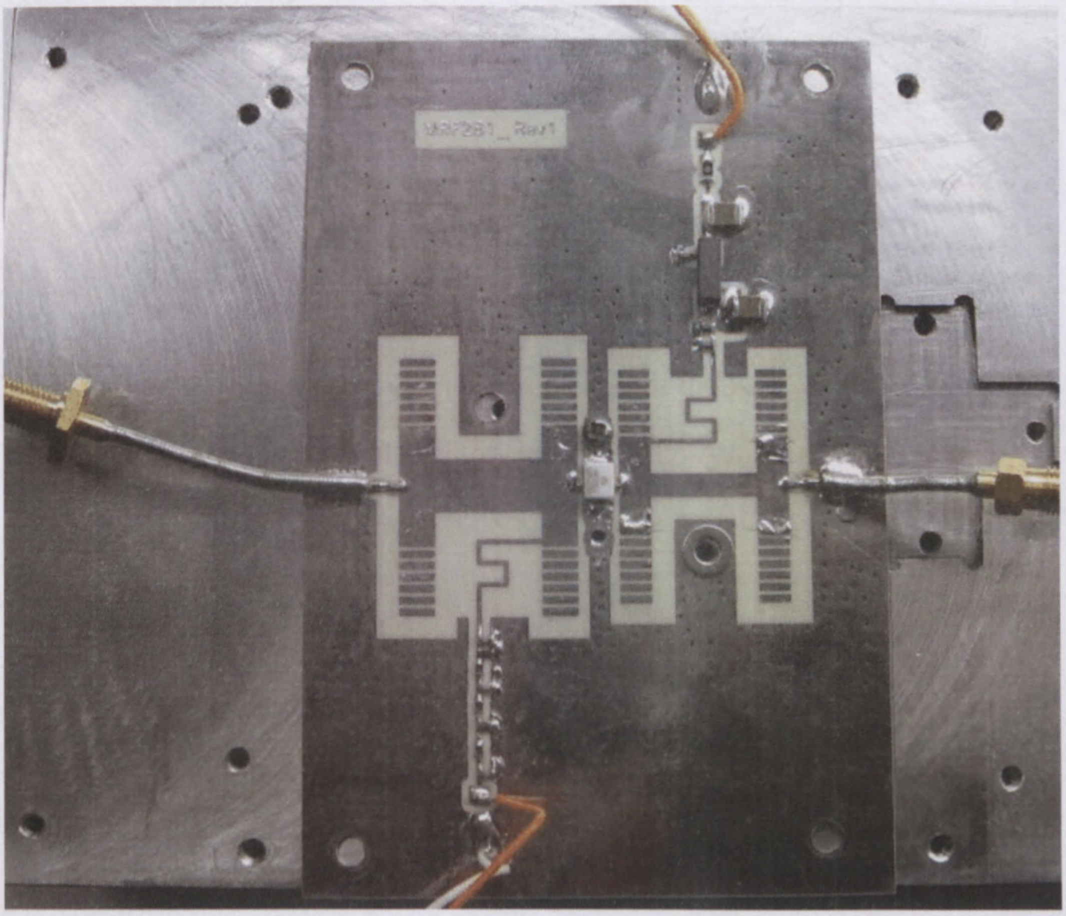


Figure 5.11: Prototype MRF281Z Amplifier

The simulated results of the MRF281Z amplifier for a swept power single tone input signal at 1643 MHz are presented in Table 5.5. The measured results of the MRF281Z amplifier for a swept power single tone input signal at 1643 MHz are presented in Table 5.6.

Output Power (dBm)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
33.18	17.19	0	21.14
34.13	17.18	-0.01	21.77
35.04	17.13	-0.06	22.46
35.89	17.04	-0.15	23.2
36.63	16.9	-0.29	24
37.23	16.6	-0.59	24.87
37.71	16.2	-0.99	25.66
38.16	15.7	-1.49	26.4
38.5	15.16	-2.03	27.2

Table 5.5: Simulated Single Tone Results for MRF281Z Amplifier

Output Power (dBm)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
33.2	15.94	0	32.9
34.13	15.86	-0.08	36.6
34.99	15.71	-0.23	40.17
35.77	15.48	-0.46	43.68
36.41	15.11	-0.83	46.6
36.9	14.6	-1.34	48.4
37.33	13.76	-2.18	49.6
37.57	13	-2.94	49.7
37.76	12.22	-3.72	49.5

Table 5.6: Measured Single Tone Results for MRF281Z Amplifier

The simulated results did not correlate that well with the measured results. Investigations revealed an error in the DC characteristics of the non-linear model of the MRF281Z MOSFET. This conclusion was drawn by the fact that the simulation provided unrealistic results for the DC current, hence this conclusion. In spite thereof, the simulated 1 dB compression point of 37.7 dBm and the measured 1 dB compression point of approximately 36.6 dBm showed a reasonable correlation. Both the simulated and measured 1 dB compression point and gain exceed the specification requirements of 36 dBm and 15 dB respectively.

The simulated and measured results of the main amplifier for a swept power two-tone input signal, are presented in Tables 5.7 and 5.8 respectively. The tones were set at 1643 MHz and 1644 MHz.

Output PEP (dBm)	**IMD3 (dBc)**	**Gain (dB)**	**Gain Compression (dB)**	**Efficiency (%)**
33.19	-35.8	17.19	0	5.3
34.17	-33.9	17.17	-0.02	6.5
35.12	-31.8	17.12	-0.07	7.9
36.03	-29.6	17.03	-0.16	9.5
36.87	-27	16.87	-0.32	11.2
37.63	-24.4	16.63	-0.56	12.9

Table 5.7: Simulated Two-Tone Results of the MRF281Z Amplifier

Output PEP (dBm)	IMD3 (dBc)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
33.1	−32.8	16.1	0	25.58
34.1	−31.5	16.1	0	28.94
35.2	−29.5	16.2	0.1	32.55
36	−28	16	−0.1	35.94
36.8	−24.7	15.8	−0.3	38.86
37.5	−22.4	15.5	−0.6	41.47

Table 5.8: Measured Two-Tone Results of the MRF281Z Amplifier

The two-tone simulated results correlate reasonable well with the measured results. When the PEP power of the output signal is 36 dBm, the simulated third order IMD product is 29.6 dB below the carrier whereas the measured third order IMD product is about 28 dB below the carrier. These simulated and measured third order IMD products both exceed the required specification of less than 27 dB for a PEP of 36 dBm.

The simulated and measured gain vs input signal frequency responses are shown in Figures 5.12 and 5.13 respectively. The gain was measured using an input power level of -20 dBm. This power level was chosen so that the amplifier would be operating in its linear region.

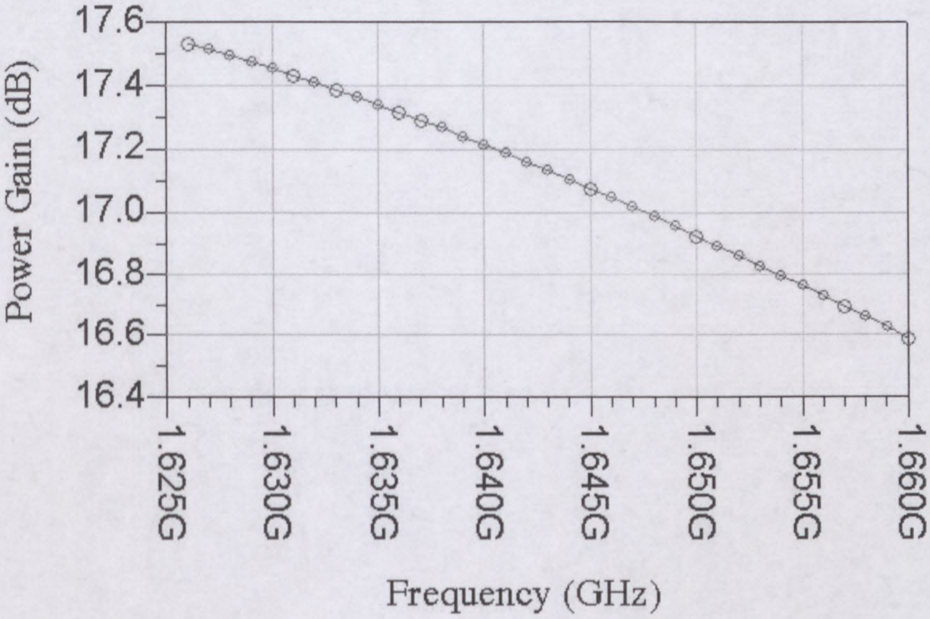


Figure 5.12: Simulated Power Gain vs Frequency of MRF281Z Amplifier

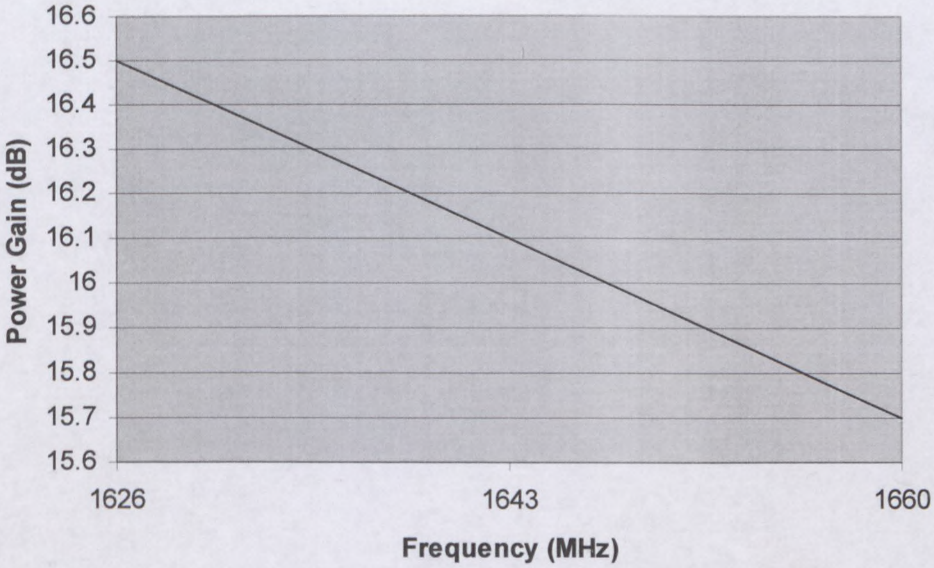


Figure 5.13: Measured Power Gain vs Frequency of MRF281Z Amplifier

The simulated power gain variation is 0.9 dB, whereas the measured power gain variation is 0.8 dB over the frequency range. Both the simulated and measured results exceed the required specification of less than 1.1 dB gain variation over the frequency range.

The measured load impedance presented to the MRF281Z by the output matching network is shown on a Smith Chart in Figure 5.14. The measured source impedance presented to the MRF281Z by the input matching network is shown on a Smith Chart in Figure 5.15. The same procedure that was used to measure the matching networks of the main amplifier, was also used here.

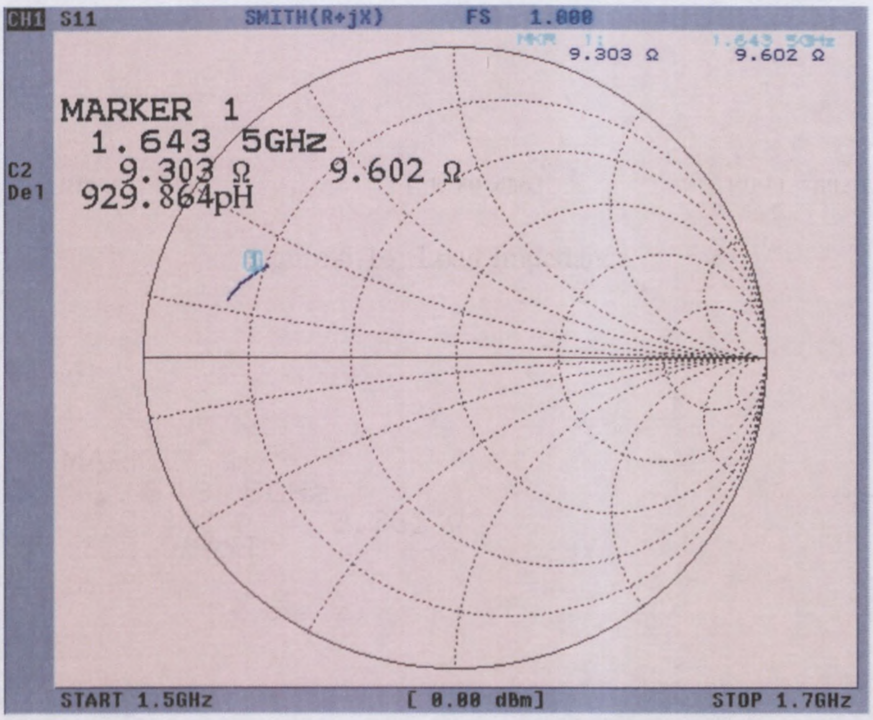


Figure 5.14: Load Impedance

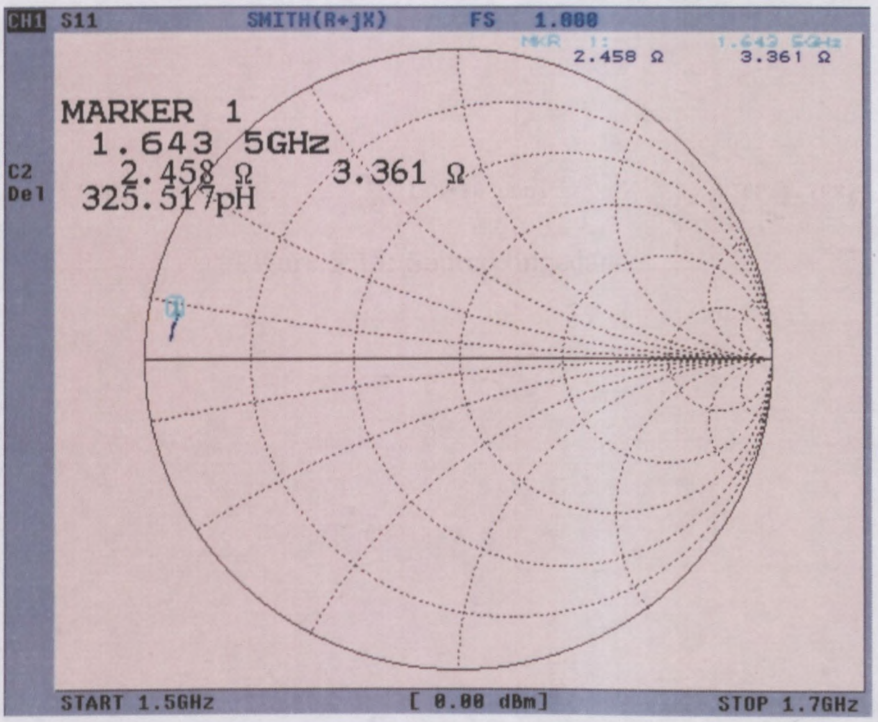


Figure 5.15: Source Impedance

The measured impedances at 1.644 GHz are:

$$Z_{L} = 9.3 + j9.6\,\Omega$$
$$Z_{S} = 2.46 + j3.36\,\Omega$$

These impedances were the optimised values obtained due to the tuning of the stubs of the matching networks for maximum output power. These impedances correlate reasonably well with the impedances that were recommended by the manufacturer's datasheet.

5.3 Output Coupler and Main Amplifier Coupler

The basic operation of a directional coupler is presented with the aid of Figure 5.16 [7].

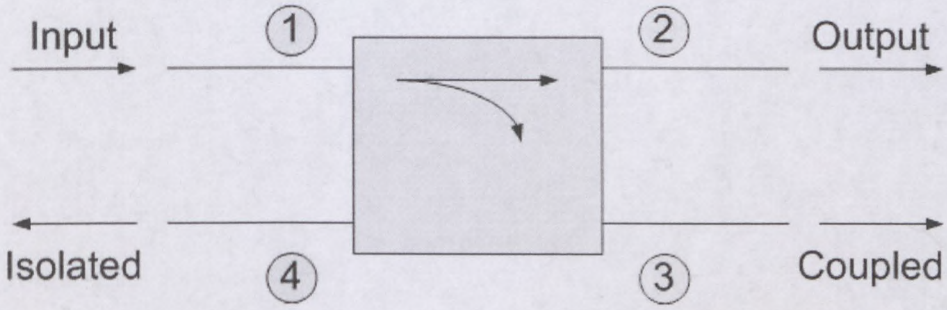


Figure 5.16: Commonly used symbol for Directional Coupler and Power Flow Convention

Power supplied to port 1 is coupled to port 3 (coupled port), while the remainder of the input power is delivered to port 2 (output port). In an ideal directional coupler, no power is delivered to port 4 (isolated port). The following three quantities are generally used to characterise a directional coupler [7]:

$$\textit{Coupling Factor} = 10 \log\left[\frac{P_1}{P_3}\right] \text{dB} \qquad 5.2$$

$$\textit{Directivity} = 10 \log\left[\frac{P_3}{P_4}\right] \text{dB} \qquad 5.3$$

$$\textit{Isolation} = 10 \log\left[\frac{P_1}{P_4}\right] \text{dB} \qquad 5.4$$

The coupling factor indicates the fraction of input power which is coupled to the coupled port. The directivity and isolation parameters are a measure of the coupler's ability to separate foward and reverse power. The ideal directional coupler would have an infinitely high directivity and isolation [7].

When two unshielded transmission lines are in close proximity, power can be coupled between the lines due to the interaction of the electromagnetic fields surrounding each line. Such lines are referred to as coupled transmission lines [7]. For the Feedforward application, it was decided to use a microstrip coupled line structure as the output and main amplifier couplers. Figure 5.17 illustrates the microstrip coupled line structure.

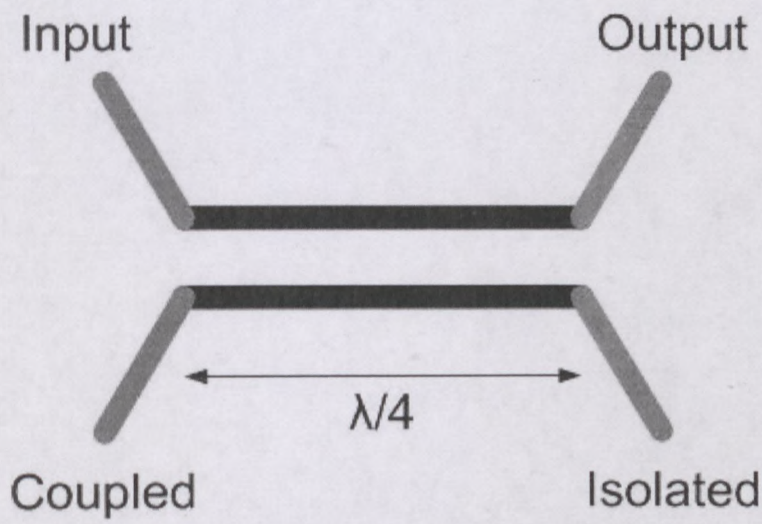


Figure 5.17: Microstrip Coupled Line Structure

The microstrip coupled line coupler consists of two quarter-wave length transmission lines in close proximity of one another. The distance between the lines and the characteristic impedance of the lines determine the coupling factor and directivity of the coupler.

5.3.1 Output Coupler

The output coupler was designed to have a coupling factor of 11 dB. Table 5.9 shows the measured results of the output coupler.

Coupling Factor	Isolation	Insertion loss	Input Return Loss	Output Return Loss
11.1 dB	28.5 dB	0.55 dB	27 dB	29.3 dB

Table 5.9: Measured Results of Output Coupler

The measured coupling factor is very close to the design goal of 11 dB. Since the coupling factor is relatively strong, the insertion loss of the coupler is relatively high. This coupler has a reasonably good directivity of approximately 17.5 dB and the input and output return losses are also very good. This coupler is therefore suitable for operation in the Feedforward system.

5.3.2 Main Amplifier Coupler

The main amplifier coupler was designed to have a coupling factor of 22 dB. Table 5.10 shows the measured results of the main amplifier coupler.

Coupling Factor	Isolation	Insertion loss	Input Return Loss	Output Return Loss
21.2 dB	28.3 dB	0.2 dB	23 dB	24 dB

Table 5.10: Measured Results of Main Amplifier Coupler

The measured coupling factor is very close to the design goal of 22 dB. Since the coupling factor of this coupler is quite weak, the insertion loss of this coupler is much lower than the insertion loss of the output coupler. Unfortunately, the weaker coupling factor also means that it is more difficult to obtain a good directivity. The directivity of this coupler is 7.1 dB.

The effects of the relatively low directivity in the Feedforward system would be that some of the power injected into the output coupler from the error amplifier, would then be coupled back to the error amplifier's input via this main amplifier coupler. To combat this, an isolator was placed at the output port of the main amplifier coupler. This method unfortunately has the disadvantage of the added insertion loss of the isolator.

The input and output return loss of this coupler is reasonably good. This coupler with the isolator on its output is therefore suitable for operation in the Feedforward system.

5.4 Variable Attenuator

The variable attenuator topology that was selected, consisted of a 3 dB hybrid coupler and two PIN diodes. The schematic of this topology is shown in Figure 5.18.

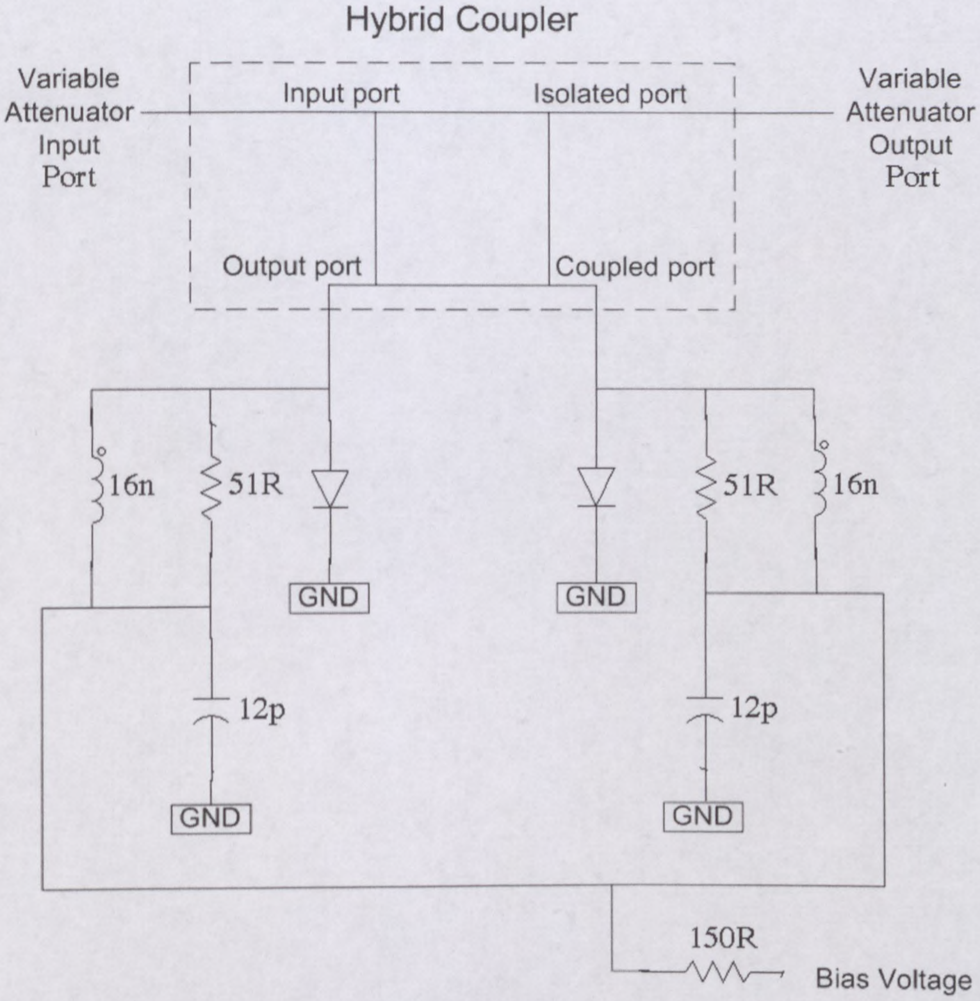


Figure 5.18: Variable Attenuator Topology

Agilent Technologies HSMP4810 PIN diodes were selected for the variable attenuator. The PIN diode is a current dependant device and acts as a variable resistor. As the current through the PIN diode increases, the resistance of the PIN diode decreases.

The input signal to the variable attenuator is connected to the input port of the hybrid coupler and the output signal of the variable attenuator is available at the isolated port of the hybrid coupler. If the output and coupled ports of the hybrid coupler are terminated in 50 Ω, the input signal of the variable attenuator will be isolated from the output port of the variable attenuator, and maximum attenuation of the input signal will occur under these conditions. This situation occurs when no current flows through the PIN diodes, which is due to the PIN diodes presenting a maximum resistance when there is no current flowing through them. Since the 12pF capacitor presents a virtual short circuit at the operating frequency, the 51 Ω resistor in parallel with the PIN diodes cause the output and coupled ports of the hybrid coupler to be presented with an impedance of approximately 50 Ω. As the current increases through the PIN diodes, the impedance that the output and coupled ports are presented with starts decreasing. This results in power being reflected back to the isolated port of the hybrid coupler, which results in the attenuation of the variable attenuator becoming less. The greater the mismatch on the coupler's output and coupled ports, the more the power will be that gets reflected back to the isolated port, the less attenuation the variable attenuator will have. Thus, as the current through the PIN diodes is increased, the attenuation of the variable attenuator is decreased.

The properties that are desirable for the variable attenuator in the Feedforward application are as follows:

- High attenuation range
- Minimum variation in attenuation over the frequency band of operation
- Constant phase over the attenuation range
- Good input and output match over the attenuation range

A variable attenuator was designed, built and measured. The measured results of the variable attenuator are shown in Table 5.11.

Current (mA)	Attenuation (dB)		Phase Shift (Degrees)		S_{11} (dB)	S_{22} (dB)
	1626 MHz	1660 MHz	1626 MHz	1660 MHz		
0	20	20.7	-18	-36	-15.2	-21.6
0.1	20	20.7	-5	-22	-15.3	-21.6
0.2	20	20.6	7.2	-6	-15.3	-21.7
0.5	18.5	18.9	40	25	-15.2	-21.7
1	14.9	14.9	43	48	-15.3	-22
1.5	12.5	12.4	69.8	54	-15.37	-22.4
2	10.7	10.7	72.8	56.6	-15.4	-22.8
2.5	9.5	9.5	74.2	57.7	-15.5	-23.3
3	8.6	8.5	74.8	58	-15.5	-23.8
3.5	7.9	7.8	75.1	58.3	-15.6	-24.4
4	7.3	7.3	75.2	58.3	-15.7	-24.8
4.5	6.8	6.8	75.2	58.3	-15.8	-25.4
5	6.4	6.4	75.2	58	-15.8	-25.8
10	4.3	4.3	74	56.4	-16.4	-29.8
20	3	3	72.6	54.5	-16.8	-31.8
30	2.6	2.5	71.9	53.5	-17	-31
40	2.36	2.3	71.5	53.1	-17	-31
50	2.2	2.2	71.3	52.7	-17	-31

Table 5.11: Measured Parameters of Variable Attenuator

The variable attenuator has a 17.8 dB range of attenuation. The variation of attenuation over frequency is minimal. It should be observed that above 1.5 mA of current flowing through the PIN diodes, the phase is reasonably constant over the attenuation range. The input and output return loss values obtained are acceptable.

Since the variable attenuator is situated at one of the output ports of the input splitter, it must be able to tolerate a reasonably large power level at its input port. It must be able to handle a worst case average power of 31 dBm. At this power level, the variable attenuator must not overheat and it must not add any intermodulation distortion to a two-tone signal. The worst case situation occurs when the variable attenuator is set to maximum attenuation. In this situation, the full 31 dBm of power will be dissipated as heat.

To verify this requirement, the variable attenuator was set to maximum attenuation and subjected to a single tone signal of 31 dBm for 20 minutes. Even though the temperature of the attenuator rose significantly, no damage occurred.

A two-tone signal, with a power level 28 dBm per tone, was applied to the variable attenuator. The carrier to third order IMD ratio was measured at the input and output of the variable attenuator and was the same in both cases. Clearly, the variable attenuator did not add any intermodulation distortion to the two-tone signal.

5.5 Variable Phase Shifter

The variable phase shifter topology selected, consisted of a 3 dB hybrid coupler and two varactor diodes. The schematic of the topology this shown in Figure 5.19.

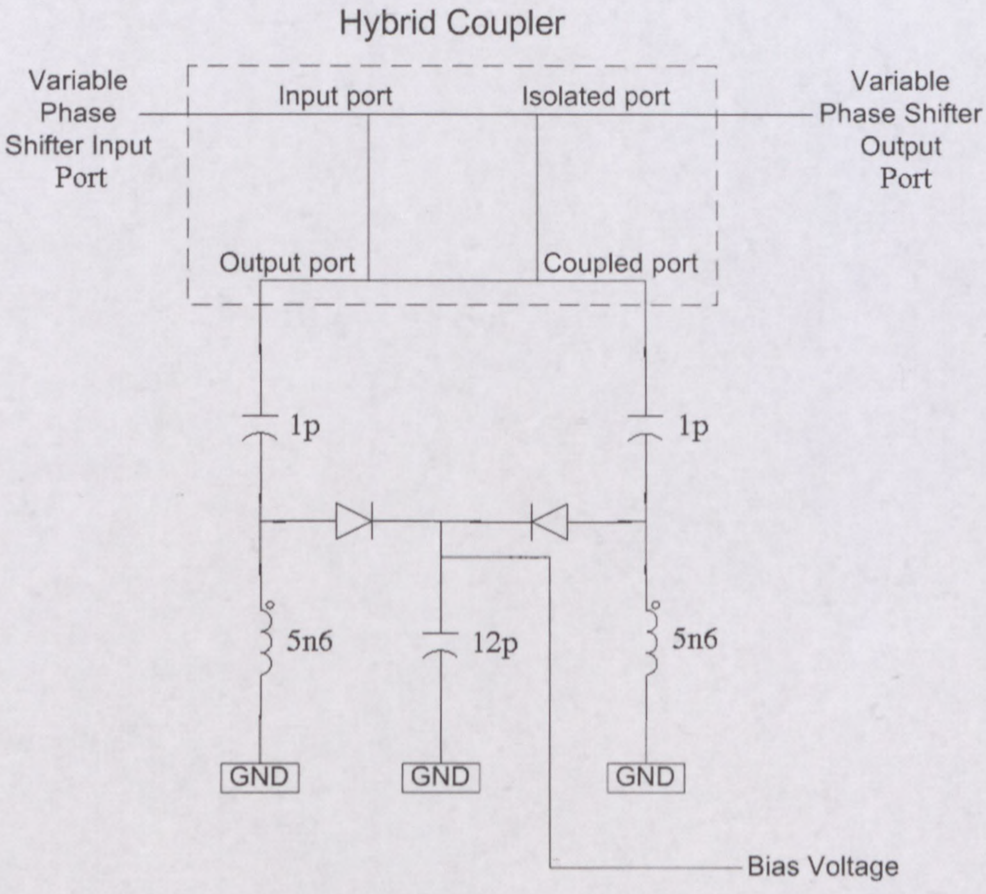


Figure 5.19: Variable Phase Shifter Topology

The Skyworks SMV1247 varactor diode was selected for use in the variable phase shifter. The varactor diode is a voltage dependant device and acts as a variable capacitor. As the voltage across the varactor diode increases, the capacitance of the varactor diode decreases. Since the 12pF capacitor presents a virtual 0Ω impedance at the frequency of operation, the cathodes of the varactor diodes are effectively connected to ground.

The input signal to the variable phase shifter is connected to the input port of the hybrid coupler and the output signal of the variable phase shifter is available at the isolated port of the hybrid coupler. Since the output and coupled ports of the hybrid coupler are not terminated in 50Ω, most of the power at these ports will be reflected back to the isolated port of the hybrid coupler and there will be minimum attenuation between the input and the output ports of the variable phase shifter. The phase shift between the input and output signals of the variable phase shifter is determined by the phase of the reflected power from the output and coupled ports of the hybrid coupler. The phase of the reflected power is determined by the phase of the impedance at the output and coupled ports of the hybrid coupler. If the voltage across the varactor diodes change, the capacitance of the varactor diodes change, causing the phase of the impedance at the output and coupled ports to change, resulting in the phase change of the variable phase shifter.

The properties that are desirable for the variable phase shifter in the Feedforward application are as follows:

- High phase shift range
- Low attenuation
- Constant attenuation over the phase shift range
- Good input and output match over the phase shift range

A variable phase shifter was designed, built and measured. The measured results of the variable phase shifter are shown in Table 5.12.

Bias Voltage (V)	Phase Shift (Degrees)		Attenuation (dB)		S11 (dB)	S22 (dB)
	1626 MHz	1660 MHz	1626 MHz	1660 MHz		
0	178.4	146	2	1.75	-14	-16.3
1	-175.7	152	2.09	1.78	-13.7	-16
2	-160	167	2.2	1.85	-13.3	-15
3	-123	-157	2.3	1.8	-13.3	-13.8
4	-99	-135	2.3	1.8	-14.3	-14
5	-85.6	-122.6	2.27	1.8	-15.3	-14.6
6	-78	-115	2.23	1.8	-16.2	-14.9
7	-75	-111	2.22	1.8	-16.5	-15.3
8	-71	-109	2.23	1.8	-16.9	-15.2

Table 5.12: Variable Phase Shifter Measurements

The variable phase shifter has approximately a 110° range in phase shift. The attenuation of the phase shifter is less than 2.3 dB and the attenuation variation of the phase shifter over the frequency band of operation is about 0.5 dB. The input and output return loss of 15 dB $\pm$ 2 dB over the phase shift range range is acceptable.

Chapter 6

Integration of the Feedforward System

The final circuits comprising the Feedforward amplifier were built and measured. A schematic of the entire Feedforward system is presented in Appendix A. A photo of the final Feedforward amplifier is shown in Figure 6.1.

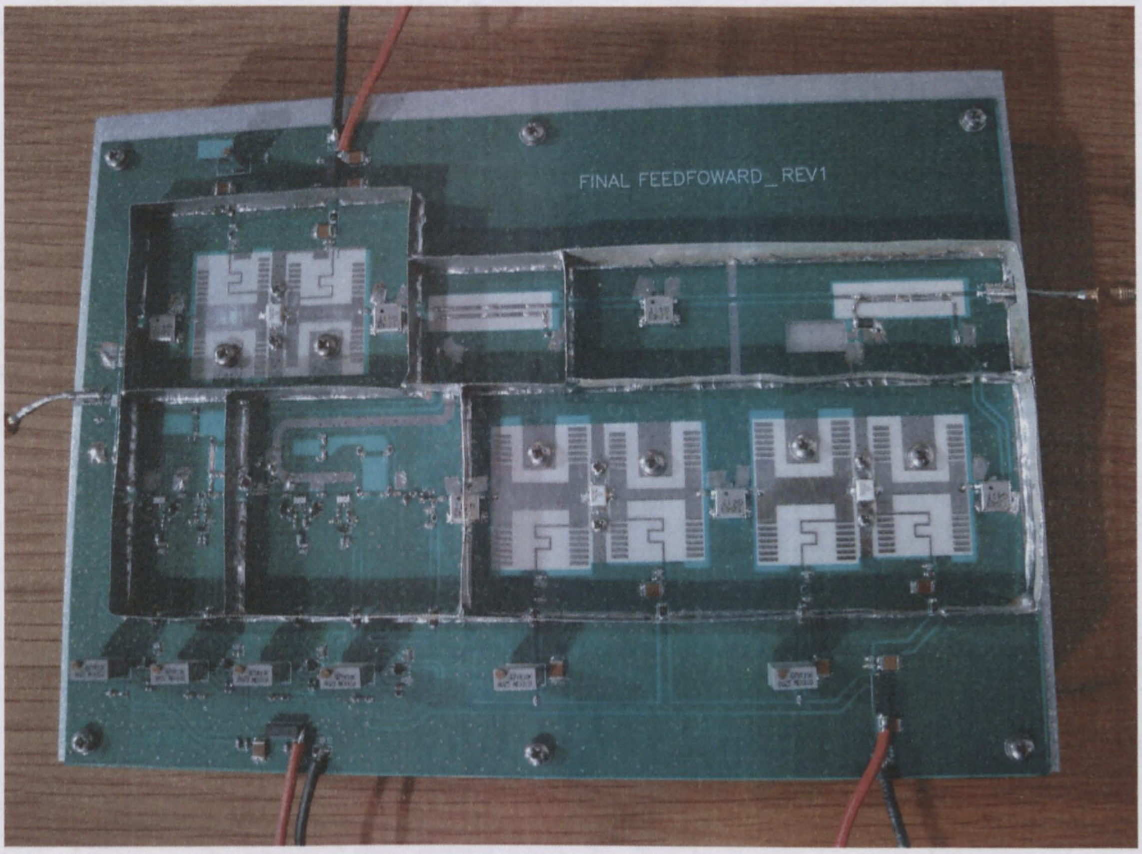


Figure 6.1: Complete Feedforward Amplifier System

The procedure used to tune and optimise the Feedforward amplifier was as follows:

- Measure the output signal of the subtractor with a spectrum analyser.
- Apply a two-tone signal to the input port of the Feedforward amplifier. Adjust the amplitude of the input signal such that the main amplifier peak envelope output power is equal to the 1dB compression point of the main amplifier, that is 40dBm.
- Tune the first loop variable phase shifter and attenuator until the amplitude of the two fundamental tones are significantly lower than the amplitude of the IMD products.
- Connect the output signal of the subtractor to the input port of the error amplifier.
- Measure the output signal of the Feedforward system using a spectrum analyser.
- Apply a two-tone signal to the input port of the Feedforward amplifier. Adjust the amplitude of the input signal such that the output peak envelope power is equal to the 1dB compression point of the main amplifier, that is 40dBm.
- Tune the second loop variable phase shifter and attenuator until the amplitude of the IMD products are significantly lower, that is −50dBc, than the amplitude of the two fundamental tones.

Two signals at 1643MHz and 1644MHz were applied to the input port of the Feedforward amplifier. During measurements of the output signal of the subtractor, an unwanted change in the amplitude of the IMD products occurred when the first loop variable attenuator and phase shifter were tuned. Only the amplitude of the fundamental two tones should change. The output signal of the first loop variable phase shifter was then measured, while the two tones were applied to the input port. Figures 6.2 and 6.3 show the output signal of the first loop variable phase shifter for two different variable attenuator settings. The variable phase shifter output power levels of the two measurements were 14dBm per tone and 3.2dBm per tone respectively.

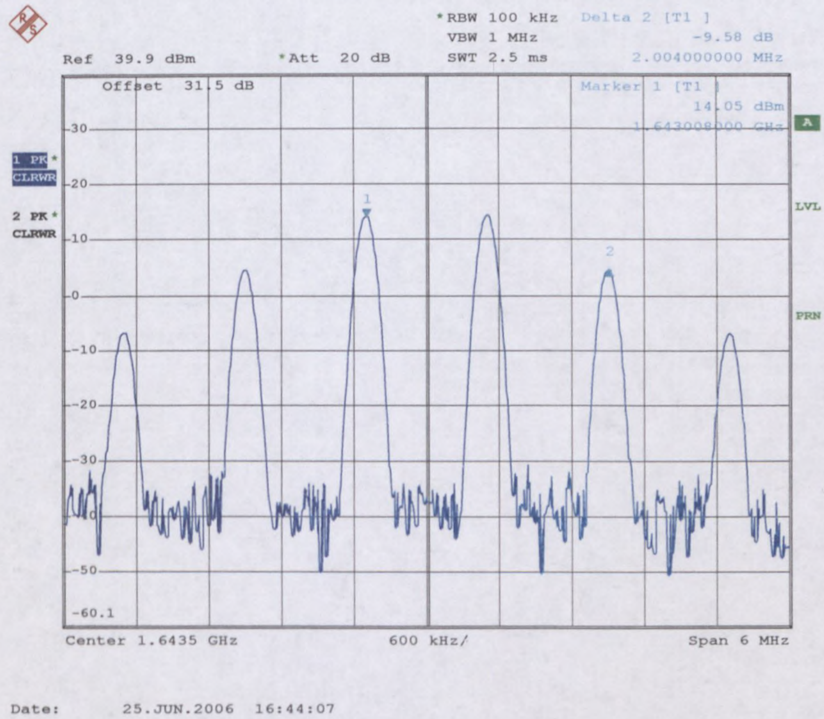


Figure 6.2: Output of Variable Phase Shifter

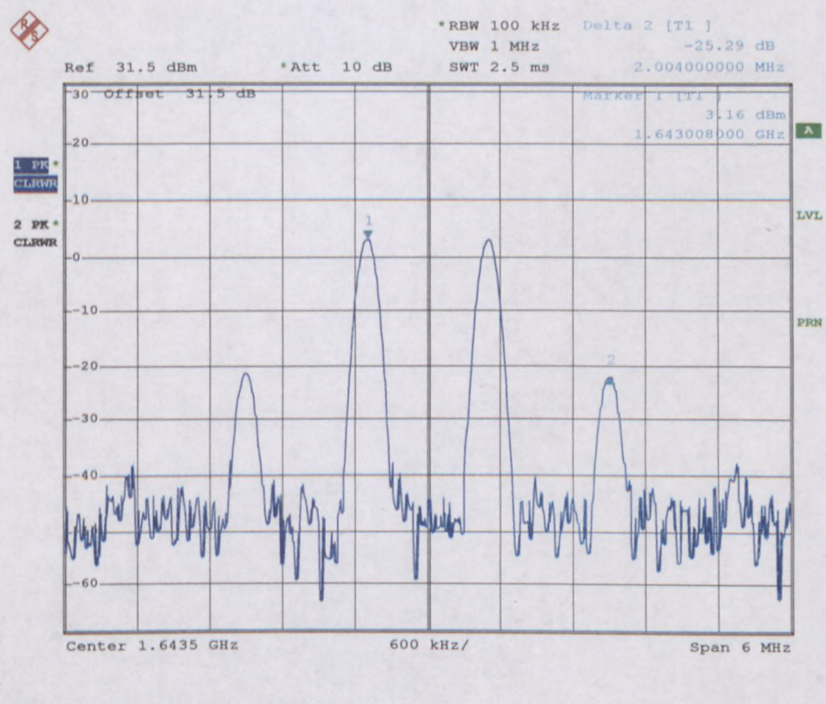


Figure 6.3: Output of Variable Phase Shifter

The results clearly indicate a high level of intermodulation distortion caused by the variable phase shifter. The level of intermodulation distortion at this point in the circuit should be non-existent, as the Feedforward system cannot tolerate this unwanted distortion. This evidence proved that the variable phase shifter was extremely non-linear at the operational power level and was apparently due to the variable phase shifter using a varactor diode, which is a non-linear device.

To overcome this problem, the first loop variable phase shifter was bypassed and the lengths of transmission line were manually tuned until the correct phase shift was achieved. Figure 6.4 depicts the manually tuned line.

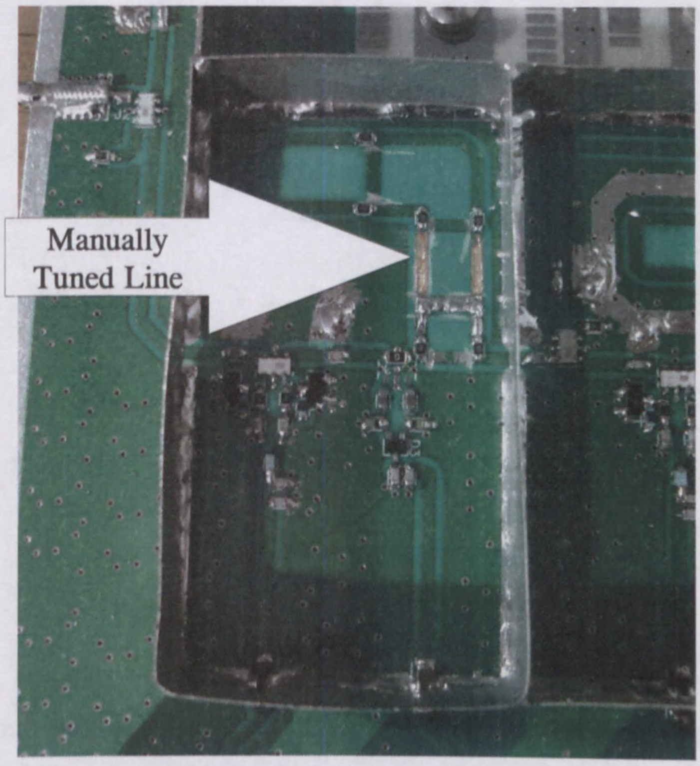


Figure 6.4: Manually Tuned Line Length

This method of manually tuning the lengths of line was not optimum, resulting in the suppression of the fundamental two tones at the output port of the subtractor being sub-optimum. Fortunately, the error amplifier had an excess in power handling capability, thus the error amplifier could tolerate a significant amount of fundamental signal input power. Figure 6.5 shows the spectrum of the signal at the output of the subtractor.

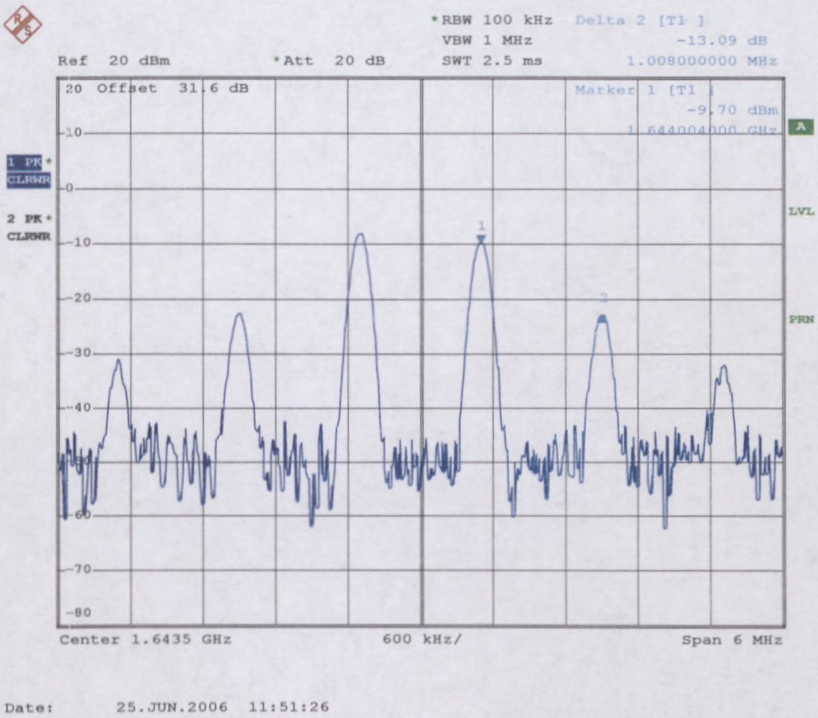


Figure 6.5: Output of Subtractor

Results show that the amplitude of the two fundamental tones were 13dB above the amplitude of the third order IMD products. This means that the subtractor suppressed the fundamental two tones by:

$$34\,\text{dBm} - 22\,\text{dB} - (-9.7\,\text{dBm}) = 21.7\,\text{dB}$$

Ideally, the amplitude of the two fundamental tones should be lower than the amplitude of the IMD products. However, the error amplifier can tolerate this amount of fundamental input power.

The output signal of the subtractor was then connected to the input port of the error amplifier. The output signal of the Feedforward amplifier system was then measured with a spectrum analyser. The second loop variable attenuator and phase shifter were then tuned until maximum suppression of the IMD products occurred. The non-linearities of the second loop variable phase shifter did not cause any unwanted distortions, since the input signal power to this phase shifter was significantly lower. Figure 6.6 shows the output spectrum of the Feedforward system when the peak envelope power was set to 40 dBm.

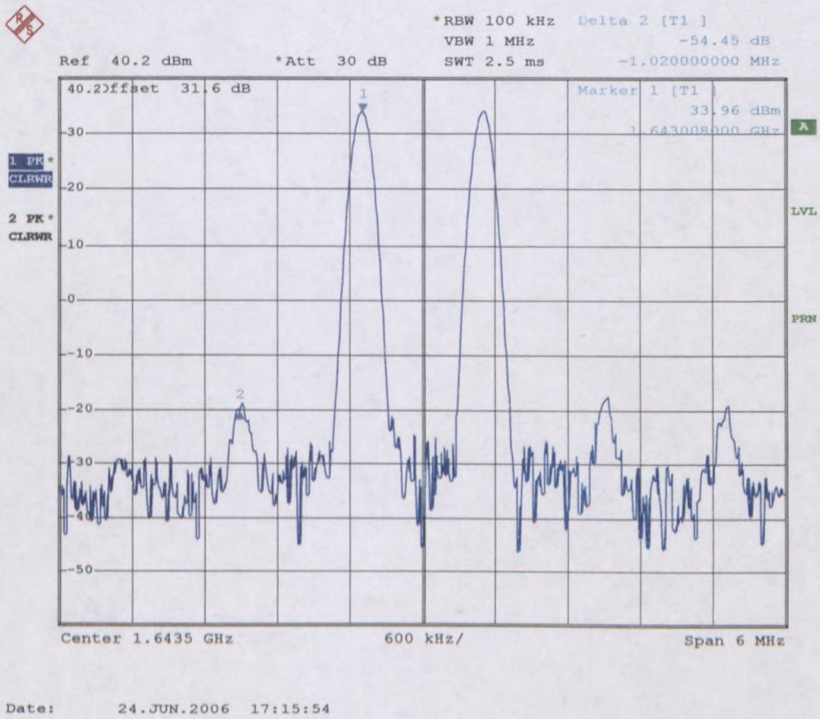


Figure 6.6: Output Spectrum of the Feedforward Amplifier System

The third order IMD products were situated 54 dB below the amplitude of the two-tone signal. This is a 24 dB improvement of intermodulation distortion performance, compared with the non-linearised amplifier. At this output power level, the main amplifier is operating just above its 1 dB compression point. Table 6.1 shows the measured parameters for a swept power two-tone input signal to the Feedforward amplifier with no linearisation applied. To perform these measurements, the input signal to the error amplifier was disconnected and the bias voltages to the error amplifier were disabled. Table 6.2 shows the measured results for a swept power two-tone input signal to the Feedforward amplifier with linearisation applied. The input signal to the error amplifier was re-connected and the bias voltages to the error amplifier were enabled for these measurements. Table 6.3 presents the improvement in the intermodulation distortion parameters that the linearisation process introduced. To calculate the IMD performance improvement, the amplitude of the third order IMD products with no linearisation applied, was subtracted from the amplitude of the third order IMD products with linearisation applied.

Output PEP (dBm)	IMD3 (dBc)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
33	-42	8.9	0	12.7
34	-41.5	8.9	0	14.3
35	-41	8.9	0	16.2
36	-41.3	8.9	0	18.4
37	-40	8.9	0	20.9
38	-37.7	8.9	0	23.6
39	-34.7	8.8	-0.1	26
39.5	-30.7	8.6	-0.3	28
40	-27.5	8.5	-0.4	30.4
41	-22.6	8	-0.9	31
42	-16.6	6	-2.9	28.7
43	-13.3	3.8	-5.1	23.9

Table 6.1: Measured Results for a Two-Tone Input Signal with No Linearisation Applied

Output PEP (dBm)	IMD3 (dBc)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
33	−55	8.7	0	8.4
34	−55	8.7	0	9.6
35	−55	8.63	−0.07	11.1
36	−55	8.73	0.03	12.8
37	−58	8.65	−0.05	14.8
38	−59	8.6	−0.1	17
39	−56	8.6	−0.1	19.9
39.5	−54	8.65	−0.05	21.4
40	−53.4	8.63	−0.07	23.4
41	−48	8.7	0	26.3
42	−36.7	8.6	−0.1	28.5
43	−23.5	8	−0.7	29.5

Table 6.2: Measured Results for a Two-Tone Input Signal with Linearisation Applied

Output PEP (dBm)	IMD improvement (dB)
33	13
34	13.5
35	14
36	13.7
37	18
38	21.3
39	21.3
39.5	23.3
40	25.9
41	25.4
42	20.1
43	10.2

Table 6.3: Third Order Intermodulation Distortion Performance Improvement

The results clearly show that the Feedforward system greatly improves the intermodulation distortion in the output signal. However, the efficiency versus output power of the system decreases when the linearisation is applied. When comparing efficiencies, one cannot however only look at the output power.

The efficiency at a specific IMD level should be compared. The efficiency of the linearised amplifier at a certain carrier to IMD ratio, should be compared with the non-linearised amplifier with the same carrier to IMD ratio. If this is done, it will be found that the linearised amplifier is much more efficient than the non-linearised amplifier.

When the linearisation is applied and the main amplifier is operating at its 1 dB compression point, the third order IMD is 53.4 dB below the two-tone power level and the efficiency is

23.4%. With no linearisation applied, the power level delivered by the main amplifier must be reduced until its third order IMD is 53.4 dB below the two-tone power level. The efficiency at this point must be measured and compared with the efficiency of the linearised amplifier. The main amplifier was simulated on its own and it was found that when the third order IMD products are 53 dB below the two-tone power level, the output PEP was only 24 dBm. The efficiency at this power level is 5.2%. Thus, the efficiency versus IMD level of the linearised amplifier is much better. This is the main reason why linearisation schemes are used as opposed to backed-off non-linearised amplifiers.

The single tone performance of the Feedforward amplifier at 1643 MHz was measured. The performance of the amplifier with no linearisation applied, was again compared to the performance with the linearisation applied. Figure 6.7 shows the output power versus gain with linearisation applied and with no linearisation applied, plotted on the same set of axes. The measured results of the Feedforward amplifier for a swept power single tone input, with no linearisation applied and with linearisation applied are presented in Tables 6.4 and 6.5 respectively.

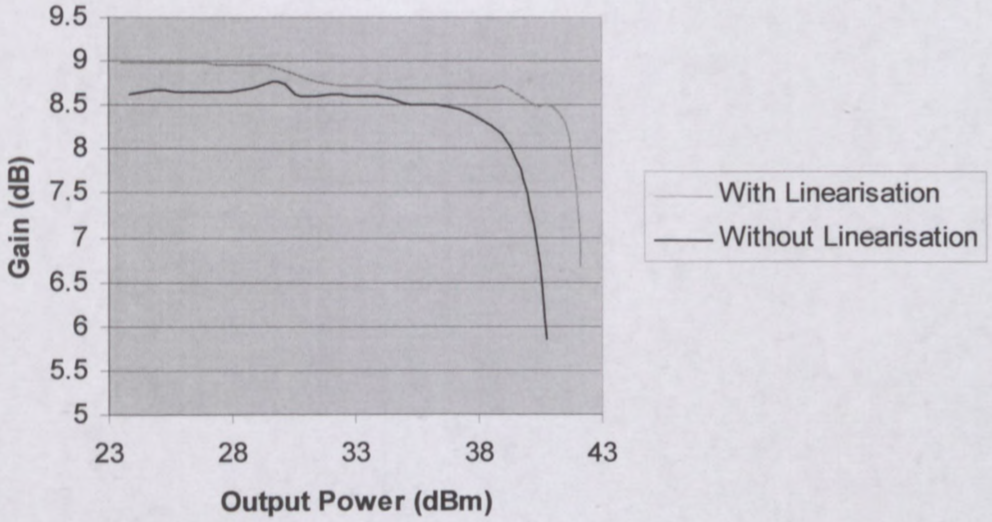


Figure 6.7: Output Power versus Gain with Linearisation and without Linearisation

Output Power (dBm)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
32.94	8.61	0	17.16
33.94	8.61	0	19.31
34.92	8.53	−0.09	21.66
35.85	8.5	−0.12	23.97
36.76	8.48	−0.14	26.52
37.67	8.4	−0.22	29.3
38.55	8.26	−0.36	32.04
39.28	8.04	−0.58	34.19
39.85	7.65	−0.97	35.15
40.36	6.99	−1.63	35.23
40.57	6.4	−2.22	34.05
40.72	5.86	−2.76	32.76

Table 6.4: Measured Single Tone Results with No Linearisation Applied

Output Power (dBm)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
32.46	8.73	0	11.05
33.42	8.71	-0.02	12.67
34.4	8.7	-0.03	14.45
35.34	8.69	-0.04	16.27
36.33	8.7	-0.03	18.67
37.3	8.7	-0.03	21.4
38.24	8.69	-0.04	24.4
39.19	8.7	-0.03	28.43
40.28	8.5	-0.23	32.17
41.05	8.49	-0.24	33.77
41.52	8.3	-0.43	33.5
41.79	7.92	-0.81	32.7
41.95	7.59	-1.14	31.65
42.04	7.24	-1.49	30.65
42.11	6.95	-1.78	29.93

Table 6.5: Measured Single Tone Results with Linearisation Applied

The output power capability of the amplifier is improved when the linearisation is applied. The 1dB compression point improved from 39.85 dBm to 41.85 dBm. The following two figures are used to aid the explanation for the 1 dB compression point improvement.

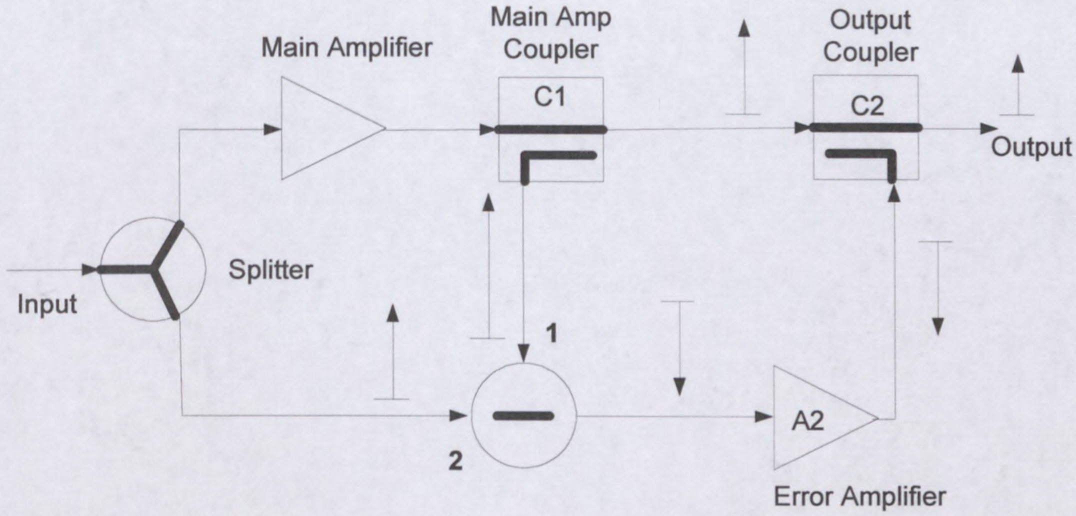


Figure 6.8: Decrease of Ouput Power Due to Subtraction of Error Amplifier Output

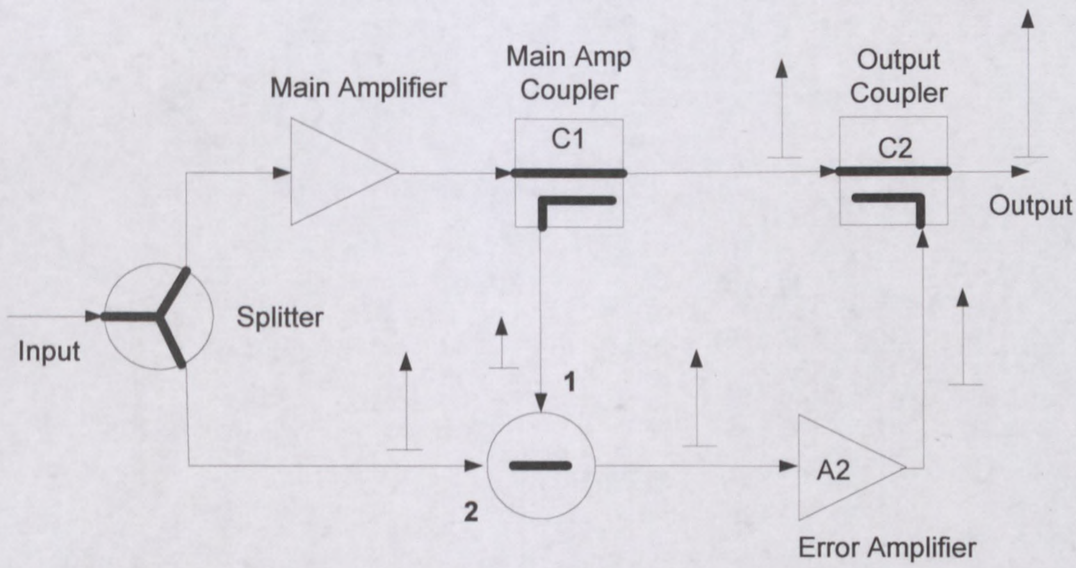


Figure 6.9: Increase of Output Power Due to Addition of Error Amplifier Output

If the amplitude of the signal at port 1 of the subtractor is larger than the amplitude of the signal at port 2 of the subtractor, the phase of the output signal of the subtractor will be such that the signal injected into the output coupler will subtract from the main amplifier signal. Ideally, the amplitude of the single tone inputs to the subtractor will be equal in amplitude.

As the main amplifier is driven into compression, the amplitude of the signal at port 1 of the subtractor will start decreasing and the amplitude of the signal at port 2 of the subtractor will be larger. Thus, the phase of the signal at the output of the subtractor will be such that the signal injected into the output coupler will add to the main amplifier signal. Therefore, the reason why the linearisation technique improves the 1 dB compression point of the amplifier, is because the error amplifier adds a signal to the output of the main amplifier that is in phase with the main amplifier output signal. Figure 6.10 shows the time domain signal at the output of the subtractor as the input power level is increased and the main amplifier starts compressing.

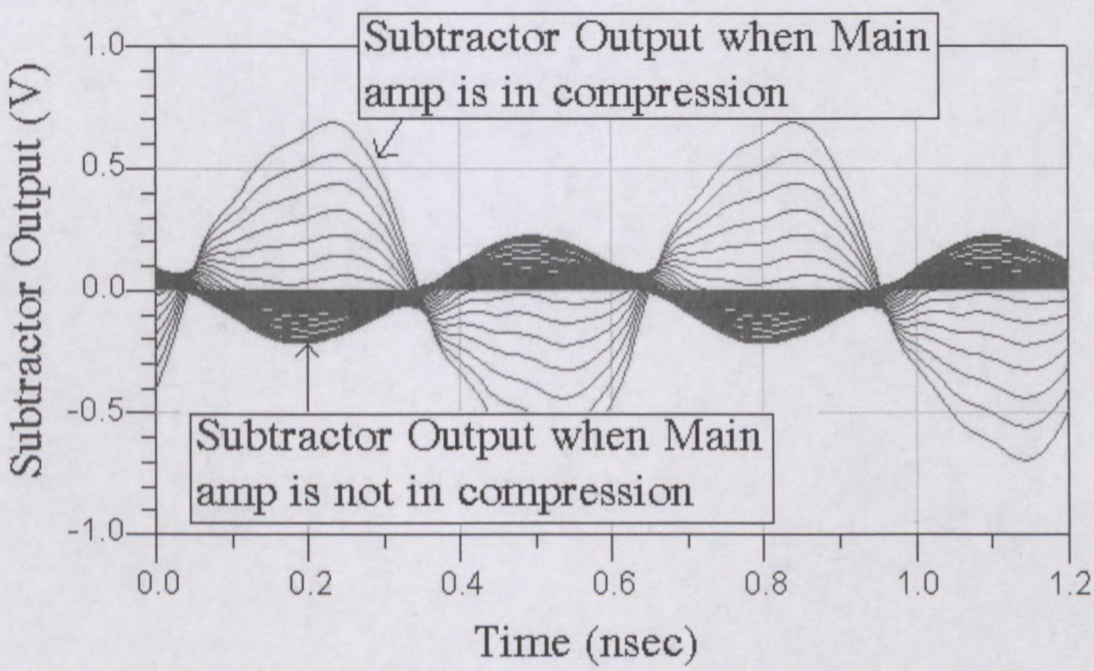


Figure 6.10: Time Domain Representation of Subtractor Output Signal as Main Amplifier Starts Compressing

Initially, the phase of the signal in Figure 6.10 is determined by port 1 of the subtractor, since it is larger in amplitude. As the main amplifier starts compressing, the amplitude of the signal at port 2 of the subtractor becomes the larger of the two inputs and it determines the phase at the output of the subtractor. Thus, at this stage, the output of the subtractor changes in phase by 180°. It is at this stage that the signal injected into the output coupler adds to the signal from the main amplifier.

The system specification of the 1 dB compression point being greater than 39.5 dBm and the third order IMD product being smaller than 50 dB below the two-tone signal when the PEP is 39.5 dBm, has been achieved at 1643 MHz.

These measurements were repeated at 1626 MHz and at 1660 MHz. The results of these measurements are presented in Appendix B. The performance of the Feedforward amplifier at these two frequencies is not as good as the performance at 1643 MHz. This is because the Feedforward amplifier had to be retuned at these frequencies. Since the phase at the input to the subtractor had to be manually tuned by changing the lengths of line, the suppression of the fundamental at the output of the subtractor, was not good as it was at 1643 MHz. Since the amplitude of the fundamental in the error loop was larger than it was at 1643 MHz, the performance of the Feedforward amplifier was degraded.

Chapter 7

Conclusions and Recommendations

7.1 Conclusions

A Feedforward power amplifier system, including all the Feedforward components, has been designed, built and tested. The Feedforward system greatly improves the intermodulation distortion performance of the power amplifier. The specifications have been achieved at the operating centre frequency of 1643 MHz. Since the first loop variable phase shifter caused large amounts of intermodulation distortion at the higher input power levels, this phase shifter had to be replaced with lengths of tunable transmission line. The phase was then manually tuned by changing the length of the line, which was not optimum. For this reason, it was not possible to obtain optimum suppression of the two-tone components at the output of the subtractor. Consequently, the performance of the Feedforward system at the two frequency extremes (1626 MHz and 1660 MHz) were not as good as the performance at the centre frequency.

The drawback of the Feedforward system is that it is an open loop system. Therefore, changes in device characteristics with respect to time, temperature, voltage and signal level are not compensated for. For the Feedforward system to operate in harsh environmental conditions, some kind of automatic control scheme must be implemented whereby gain and phase are continuously adjusted to obtain the best signal cancellation and output linearity.

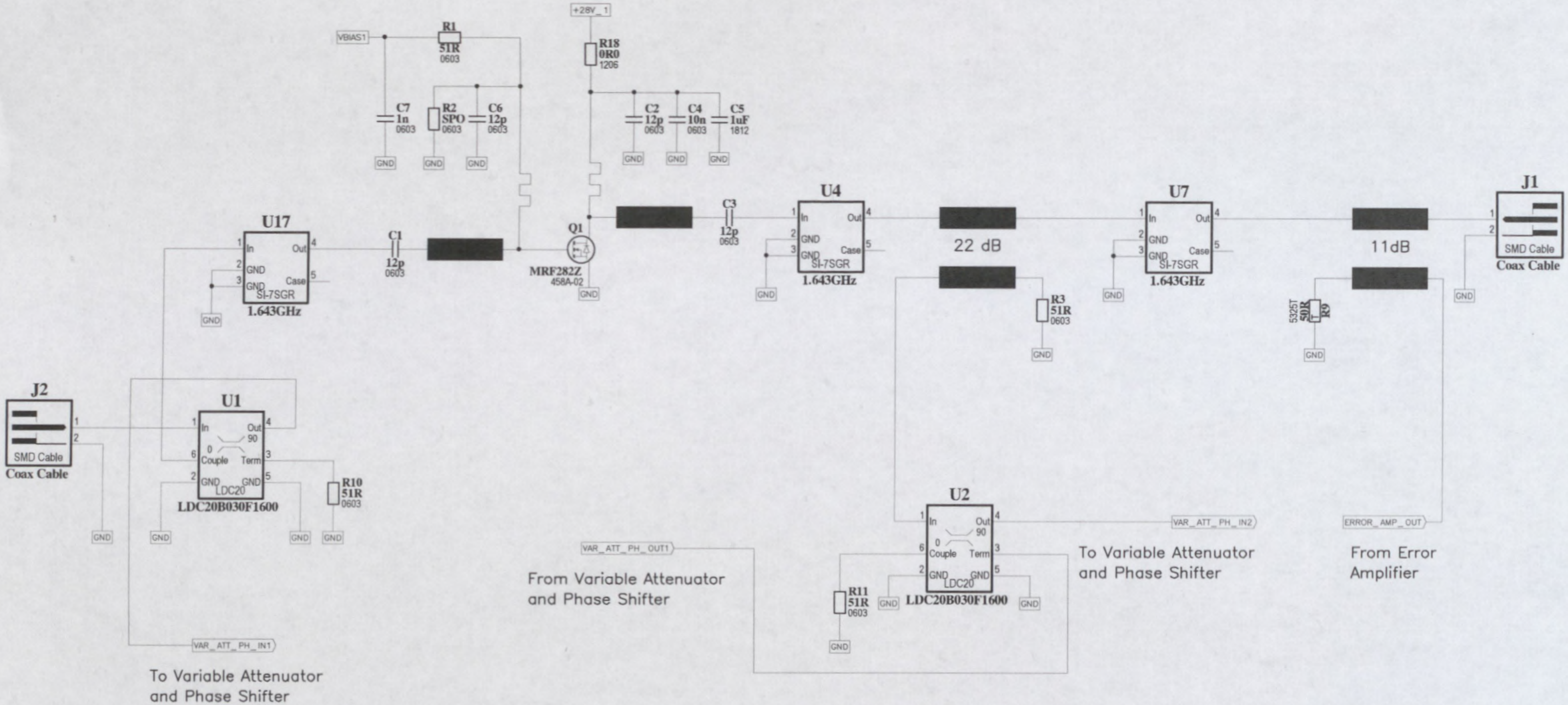
7.2 Recommendations

Further work could be done by implementing a variable phase shifter that does not contribute intermodulation distortion to the input signal at the higher input power levels. This would significantly improve the performance of the Feedforward system.

Further research could also be done to develop an adaptive Feedforward system, whereby the gain and phase are continuously adjusted so that the Feedforward system can operate optimally over largely changing environmental conditions.

Appendix A Feedforward Schematic

The following four pages present the complete schematic of the entire Feedforward system.



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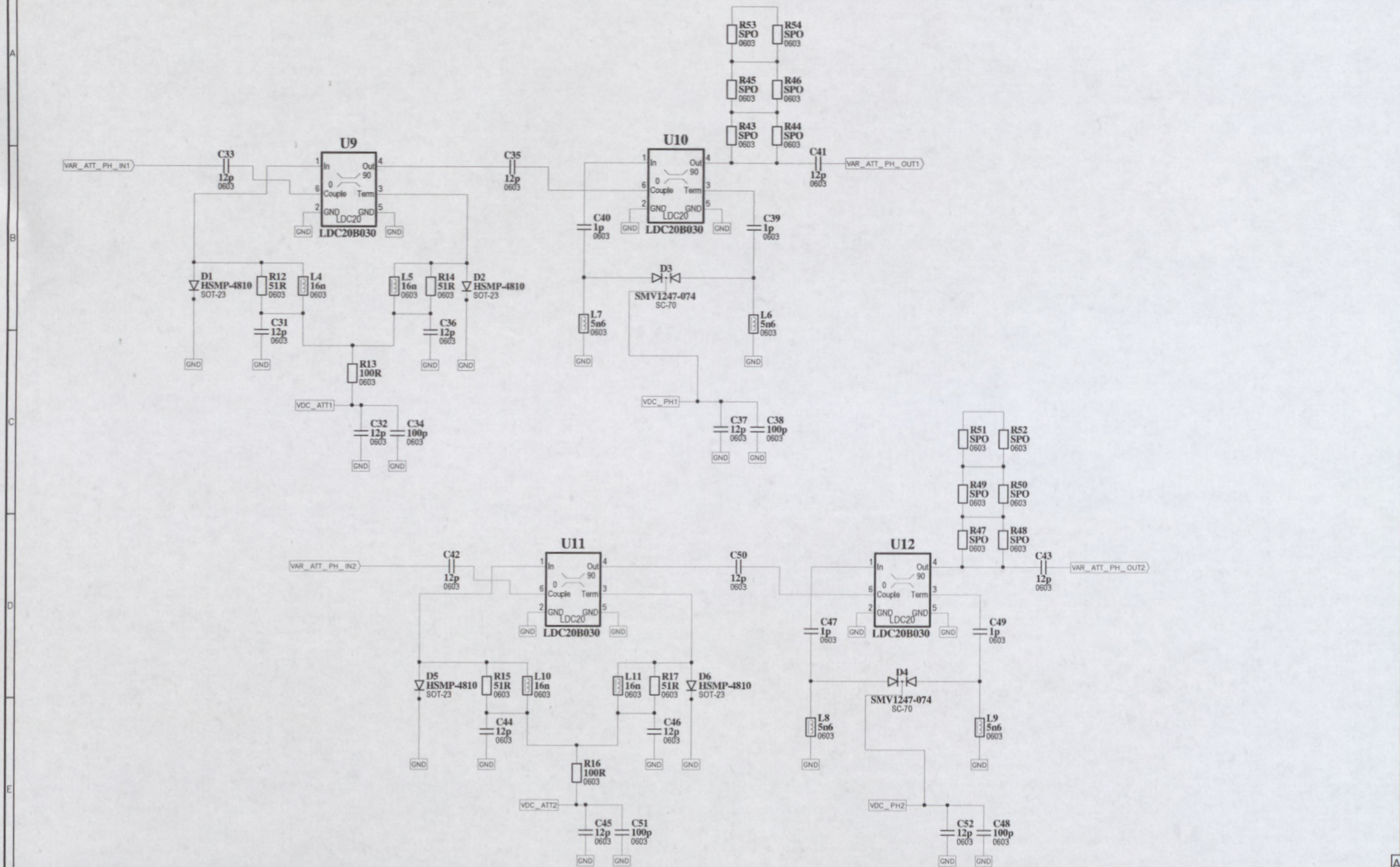

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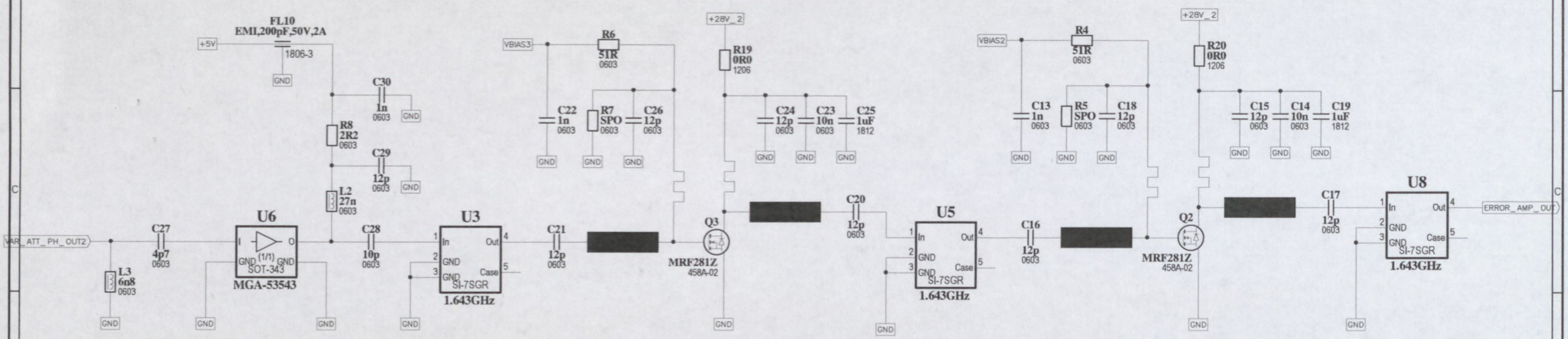

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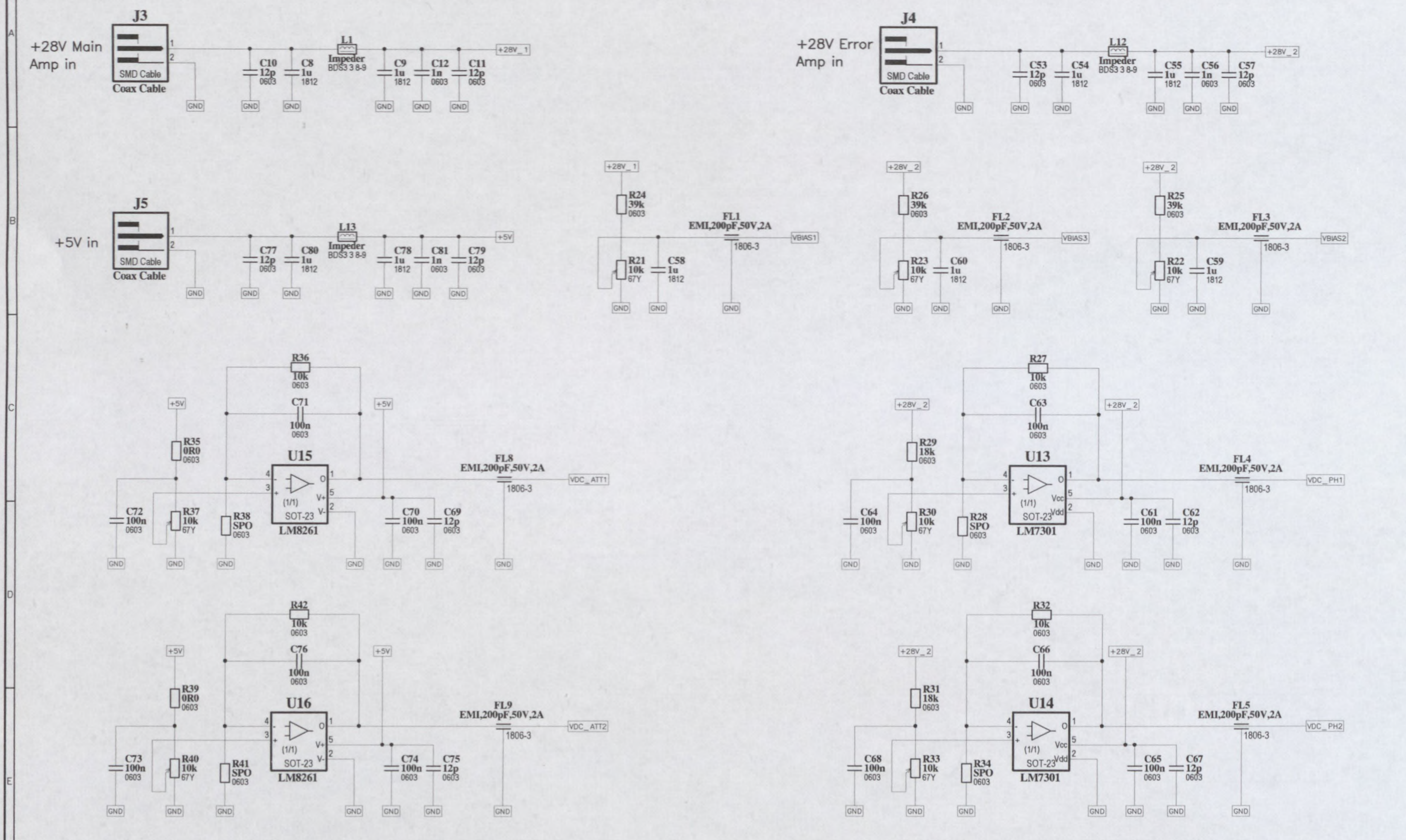

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Appendix B Measured Feedforward Results

The following tables present the measured results of the Feedforward amplifier with linearisation applied and without linearisation applied, at 1626 MHz and at 1660 MHz.

Output PEP (dBm)	IMD3 (dBc)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
33	-42	8.9	0	12.2
34	-41.5	8.9	0	13.8
35	-41.5	8.9	0	15.6
36	-42	8.9	0	17.6
37	-41	8.9	0	19.8
38	-38	8.9	0	22
39	-35	8.8	-0.1	24.8
39.5	-31	8.6	-0.3	27.4
40	-28	8.5	-0.4	29.2
41	-23.3	8	-0.9	30.15
42	-16.2	6	-2.9	26.83
43	-14.3	3.8	-5.1	23.7

Table B.1: Measured Two-Tone Results with No Linearisation Applied
(F1 = 1626 MHz; F2 = 1627 MHz)

Output PEP (dBm)	IMD3 (dBc)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
33	-50	7.7	0	7.2
34	-50	7.7	0	8.22
35	-50	7.72	0.02	9.36
36	-51	7.71	0.01	10.62
37	-51	7.71	0.01	11.75
38	-50	7.65	-0.05	13.26
39	-46	7.7	0	14.99
39.5	-40.9	7.6	-0.1	15.88
40	-35.5	7.55	-0.15	16.7
41	-27.6	7.5	-0.2	18.7
42	-21.5	7.1	-0.6	20.22
43	-17.3	6.25	-1.45	20.97

Table B.2: Measured Two-Tone Results with Linearisation Applied
(F1 = 1626 MHz; F2 = 1627 MHz)

Output PEP (dBm)	IMD3 (dBc)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
33	-43	8.9	0	12.93
34	-42.4	8.9	0	14.6
35	-43	8.9	0	16.4
36	-42.5	8.9	0	18.5
37	-40.5	8.9	0	21.2
38	-37	8.9	0	23.23
39	-34	8.8	-0.1	26
39.5	-29.9	8.6	-0.3	28.6
40	-26.3	8.5	-0.4	30.32
41	-22	8	-0.9	31.54
42	-16.7	6	-2.9	29.5
43	-13.5	3.8	-5.1	24.2

Table B.3 Measured Two-Tone Results with No Linearisation Applied
(F1 = 1659 MHz; F2 = 1660 MHz)

Output PEP (dBm)	IMD3 (dBc)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
33	-61	7.8	0	7.86
34	-63	7.8	0	8.87
35	-62	7.8	0	10.07
36	-59	7.9	0.1	11.57
37	-58	7.8	0.	12.89
38	-57	7.8	0	14.56
39	-52	7.66	-0.14	16.14
39.5	-46	7.6	-0.2	16.92
40	-43	7.7	-0.1	18.14
41	-29.5	7.4	-0.4	19.97
42	-22.6	7.1	-0.7	21.98
43	-17.8	6.41	-1.39	22.94

Table B.4: Measured Two-Tone Results with Linearisation Applied
(F1 = 1659 MHz; F2 = 1660 MHz)

Output Power (dBm)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
32.35	8.95	0	15.43
33.34	8.88	-0.07	17.34
34.31	8.83	-0.12	19.34
35.27	8.79	-0.16	21.64
36.38	8.82	-0.13	24.65
37.3	8.73	-0.22	27.14
38.16	8.58	-0.37	29.57
38.96	8.31	-0.64	31.69
39.65	7.97	-0.98	33.48
40.14	7.44	-1.51	33.78
40.47	6.89	-2.06	33.49
40.69	6.32	-2.63	32.35

Table B.5: Measured Single Tone Results with No Linearisation Applied (F = 1626 MHz)

Output Power (dBm)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
30.92	7.65	0	7.86
31.9	7.65	0	8.81
32.86	7.76	0.11	9.9
34	7.75	0.1	11.31
34.97	7.7	0.05	12.55
35.93	7.68	0.03	13.96
36.88	7.62	-0.03	15.4
37.84	7.64	-0.01	17.17
38.83	7.66	0.01	19.06
39.9	7.74	0.09	24.91
40.38	7.43	-0.22	26.14
40.65	7.05	-0.6	25.79
40.8	6.6	-1.05	24.68
40.93	6.19	-1.46	23.4
41.08	5.94	-1.71	22.4

Table B.6: Measured Single Tone Results with Linearisation Applied (F = 1626 MHz)

Output Power (dBm)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
32.83	8.53	0	17.53
33.9	8.51	-0.02	19.76
34.86	8.53	0	22.17
35.84	8.59	0.06	24.86
36.78	8.51	-0.02	27.58
37.7	8.48	-0.05	30.53
38.49	8.33	-0.2	33.06
39.24	8.09	-0.44	35.27
39.81	7.68	-0.85	36.3
40.17	6.99	-1.54	35.63
40.47	6.31	-2.22	34.69
40.72	5.52	-3.01	32.82

Table B.7: Measured Single Tone Results with No Linearisation Applied (F = 1660 MHz)

Output Power (dBm)	Gain (dB)	Gain Compression (dB)	Efficiency (%)
31.51	7.69	0	9.22
32.5	7.68	-0.01	10.41
33.48	7.69	0	11.75
34.49	7.72	0.03	13.29
35.48	7.73	0.04	14.98
36.39	7.76	0.07	16.67
37.37	7.9	0.21	18.79
38.35	7.94	0.25	20.98
39.35	7.9	0.21	23.57
40.22	7.87	0.18	25.94
41.08	7.67	-0.02	28.48
41.46	7.32	-0.37	28.86
41.74	6.99	-0.7	28.86
41.91	6.64	-1.05	28.41
42.05	6.36	-1.33	27.99

Table B.8: Measured Single Tone Results with Linearisation Applied (F = 1660 MHz)

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