Linearisation Techniques for Microwave Direct-Carrier Transmitters

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UniS

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To my parents....... and my family
Abstract

A high bandwidth-efficiency modulation scheme is demanded for supporting a high-data rate communications, and so a highly linear transmitter is needed. Applying a linearisation technique to the transmitter can achieve this goal.

In this thesis, there are three main topics which are investigated. They are mixer linearisation, the vector modulator based on the reflection-type attenuator and transmitter linearisation.

For the first topic, there are two contributions in this thesis. The first technique is the application of the feedforward (FF) technique to linearise a downconversion mixer. It is shown for the first time that the FF technique for a mixer is simplified to be a single loop rather than the conventional double-loop structure, leading to a lower complexity and a high-linearity mixer. The second proposed technique applied to a mixer is a harmonic injection technique. The technique simply injects the difference-frequency tone to the input of a mixer. It is shown from the simulations and the experiments that the technique can improve the linearity significantly without trading off the power efficiency.

Apart from the mixer linearisation topic, there are three contributions concerning with the reflection-type attenuator (RTA). The first is the feedback reflection attenuator based on Field-Effect Transistors (FETs). It has been found that applying resistive feedback can improve the attenuation range of the RTA and also the phase-distortion by trading-off the input and output return losses. The RTA size is comparable to the size of the conventional RTA which suffers from the phase-distortion. For variable attenuator applications, this structure can improve the attenuation range over the conventional RTA. For bi-phase modulator application, the structure is 50% smaller than the balanced structure based on the conventional RTA, which is needed for correcting the phase-distortion. The second contribution for this topic is the demonstra-
tion of an improved structure for a vector modulator (VM) based on the RTA. To avoid the phase distortion problem, the full balanced structure is needed and hence a large chip area is consumed. A simple technique to compensate the phase distortion and balance the amplitude for the whole control voltage range is proposed by adding an extra source inductor and a shunt resistor at the MESFET's drain. The aforementioned problems are overcome and the circuit size is 50% of the balanced VM. In addition, the baseband signals for the proposed structure are reduced to 2, compared to 4 channels for the balanced VM. The third contribution is the study of the nonlinearity distortion in the RTA. The analysis technique is based on the power series model. The results provide the criteria for selecting the active devices to obtain the small nonlinearity distortion.

Linearisation techniques for the whole transmitter are also under the research in this thesis. There are three contributions to the topic: The first technique is applying the FF technique for the whole transmitter. The advantage of the technique is the capability of reducing the distortion not only from the main power amplifier but also for the modulator. The second technique is a proposed topology for a low-cost millimetre-wave transmitter. The structure has low complexity since it composes of only 3 main parts, i.e. a VM, a medium/high-power oscillator and a DSP processor. The DSP part provides multi-functionality to the structure. These functions include baseband predistortion, channel filtering, modulation technique, to name the few. The last contribution of this topic is the improved LINC structure, so called adaptive predistortion LINC, to correct the phase/gain imbalances. The proposed technique shows the capability to overcome this effect, which can degrade the distortion performance in LINC.
Key words: linearisation, microwave circuits, vector modulator, mixer, transmitter
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<td>ACI</td>
<td>Adjacent Channel Interference</td>
</tr>
<tr>
<td>AM-AM</td>
<td>Amplitude-to-Amplitude Modulation</td>
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<tr>
<td>AM-PM</td>
<td>Amplitude-to-Phase Modulation</td>
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<tr>
<td>AP</td>
<td>Adaptive Predistortion</td>
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<tr>
<td>ASIC</td>
<td>Application-Specific-Integrated Circuits</td>
</tr>
<tr>
<td>BER</td>
<td>Bit-Error-Rate</td>
</tr>
<tr>
<td>BPSK</td>
<td>Binary Phase Shift-Keying</td>
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<tr>
<td>CALLUM</td>
<td>Combined Analogue Locked-Loop Universal Modulator</td>
</tr>
<tr>
<td>CDMA</td>
<td>Code Division Multiple Access</td>
</tr>
<tr>
<td>dB</td>
<td>decibel</td>
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<tr>
<td>DBM</td>
<td>Doubled Balance Mixer</td>
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<tr>
<td>DC</td>
<td>Direct Current</td>
</tr>
<tr>
<td>DSP</td>
<td>Digital Signal Processing</td>
</tr>
<tr>
<td>EVM</td>
<td>Error Vector Magnitude</td>
</tr>
<tr>
<td>EER</td>
<td>Envelope Elimination and Restoration</td>
</tr>
<tr>
<td>FDI</td>
<td>Frequency-Different Injection Technique</td>
</tr>
<tr>
<td>FDMA</td>
<td>Frequency Division Multiple Access</td>
</tr>
<tr>
<td>FET</td>
<td>Field-Effect Transistor</td>
</tr>
<tr>
<td>FF</td>
<td>Feedforward</td>
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<tr>
<td>FPGAs</td>
<td>Field Programmable Gate-Array</td>
</tr>
<tr>
<td>GaAs</td>
<td>Gallium Arsenide</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Description</td>
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<tr>
<td>GPIB</td>
<td>General Purposed Interface Bus</td>
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<tr>
<td>HF</td>
<td>High Frequency</td>
</tr>
<tr>
<td>HPA</td>
<td>High Power Amplifier</td>
</tr>
<tr>
<td>IF</td>
<td>Intermediate Frequency</td>
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<tr>
<td>IMD</td>
<td>Intermodulation Distortion</td>
</tr>
<tr>
<td>IM&lt;sub&gt;3&lt;/sub&gt;</td>
<td>Third-order Intermodulation</td>
</tr>
<tr>
<td>ISI</td>
<td>Inter-Symbol Interference</td>
</tr>
<tr>
<td>kHz</td>
<td>kilo-Hertz</td>
</tr>
<tr>
<td>LINC</td>
<td>Linear Amplification using Nonlinear Components</td>
</tr>
<tr>
<td>LMS</td>
<td>Least-Mean Square</td>
</tr>
<tr>
<td>LO</td>
<td>Local Oscillator</td>
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<tr>
<td>MMICs</td>
<td>Monolithic Microwave Integrated Circuits</td>
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<tr>
<td>mm-wave</td>
<td>millimetre-wave</td>
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<tr>
<td>MSK</td>
<td>Minimum Shift-Keying</td>
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<tr>
<td>NF</td>
<td>Noise Figure</td>
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<tr>
<td>OFDM</td>
<td>Orthogonal Frequency-Division Multiplexing</td>
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<tr>
<td>PA</td>
<td>Power Amplifier</td>
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<tr>
<td>PCB</td>
<td>Printed Circuit Board</td>
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<tr>
<td>PD</td>
<td>Pre-Distortion</td>
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<tr>
<td>PN</td>
<td>Pseudonoise Sequence</td>
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<td>PSK</td>
<td>Phase Shift-Keying</td>
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<tr>
<td>PWM</td>
<td>Pulse-Width Modulation</td>
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<tr>
<td>QAM</td>
<td>Quadrature Amplitude Modulation</td>
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<tr>
<td>QPSK</td>
<td>Quatermary Phase Shift-Keying</td>
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<td>RF</td>
<td>Radio Frequency</td>
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<td>RFICs</td>
<td>Radio Frequency Integrated Circuits</td>
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<tr>
<td>RTA</td>
<td>Reflection-Type Attenuator</td>
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<tr>
<td>SCS</td>
<td>Signal Component Separator</td>
</tr>
<tr>
<td>Abbreviation</td>
<td>Full Form</td>
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<tr>
<td>SNR</td>
<td>Signal-to-Noise Ratio</td>
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<tr>
<td>SR</td>
<td>Software Radio</td>
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<tr>
<td>TOI</td>
<td>Third-order Intercept Point</td>
</tr>
<tr>
<td>WANs</td>
<td>Wireless-Local Area Networks</td>
</tr>
<tr>
<td>VM</td>
<td>Vector Modulator</td>
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List of Relevant Publications


6. M. Chongcheawchamnan, C. Y. Ng, N. Siripon and I. D. Robertson, "Reflection-type Bi-phase Amplitude Modulator with Improved Perfor-

Chapter 1

Introduction

1.1 The need for Linearisers

Over the past two decades, the rapid development of personal communications systems has led to a lot of research in both the academic and industrial sectors to develop new technologies. For digital communications, nonconstant envelope modulation schemes such as 16-QAM (16-Quadrature Amplitude Modulation) provide a better bit-error-rate for a given carrier-to-noise ratio compared to constant-envelope modulation schemes with the same bandwidth utilisation efficiency (such as 16-Phase Shift Keying). However, it is widely known that nonconstant envelope modulation schemes are sensitive to nonlinear components in the communication system. The nonlinear effects in the system create distortions which can be observed in the time, frequency or signal constellation domain.

Figure 1.1 shows a representation of the frequency plan for multichannel communications. Let the transmitted signal in CH$_1$ be comprised of two sinusoidal signals, which is a simple way to create a nonconstant envelope signal. The output signal spectrum from the nonlinear system is shown to have two
Chapter 1. Introduction

Figure 1.1: Nonlinearity effects in the frequency domain

Figure 1.2: Nonlinearity effects in the signal constellation diagram
1.1. The need for Linearisers

Extra intermodulation products, which appear in the adjacent channels, $CH_{i-1}$ and $CH_{i+1}$. These extra intermodulation distortion product components occurring in the adjacent channels interfere with the transmitted signal of those channels. Figure 1.2 shows the nonlinearity effects in the signal constellation diagram. At the transmitter, the nonlinearities in the modulator and the power amplifier cause amplitude and phase error which will be apparent after the received signal is demodulated to baseband (shown as the OUTPUT). From Figure 1.2, the magnitude difference between the ideal modulation vector (shown as the INPUT) and the actual modulator vector (shown as the OUTPUT), which has been changed by the nonidealities (nonlinearities, phase and gain imbalances in modulator and demodulator, phase noise, etc.), is defined as Error Vector Magnitude (EVM). The Bit-Error-Rate (BER) is increased since each detected bit is placed beyond the optimum detection region, and this is quantified by the EVM measurement. This degradation in BER is more critical with the higher level modulation schemes, such as 16-QAM or 64-QAM, since the detection region of each symbol is smaller.

The simplest technique to correct these problems is to employ a highly linear class-A HPA (High-Power Amplifier) in back-off mode. This technique, however, results in a very low power efficiency. The linearisation technique is a better approach to provide the required high linearity. Some linearisation techniques can achieve the high power efficiency demands by modern communications systems.

The aim of this thesis is to investigate the use of linearisation techniques in mixers and direct carrier modulation transmitters, with applications in communications systems operating from the 1 to 2 GHz range right up to the millimetre-wave band. In the course of the investigation, some of the individual components used in linearisers have been studied. These are an improved analogue variable attenuator and a vector modulator based on the reflection-
type topology.
1.2 Outline of the Thesis

The research presented in this thesis is concerned with the development of linearising techniques and the improved performance of the vector modulator and variable attenuator based on the reflection-type attenuator (RTA). The thesis is divided into five main chapters.

In Chapter 2, the fundamental concepts of existing linearisation techniques are described. These techniques are feedforward, feedback, predistortion, envelop elimination and restoration, linear amplification using nonlinear components, and harmonic injection. Their relative advantages and disadvantages are briefly discussed.

Chapter 3 presents the linearisation techniques for mixer applications. Two proposed techniques, the simplified feedforward technique and the Difference-Frequency Injection technique (DFI), are first applied to a double-balanced ring-diode mixer. A system-level analysis of these techniques which yields the initial design equations for implementing the linearisers is described in Section 3.2 and Section 3.3. A linearity comparison of the proposed techniques with the same ring-diode mixer proves the validity of these two techniques. The advantages and disadvantages of these two techniques are also given.

Three main sections describing the analogue RTA are presented in Chapter 4. Firstly, the feedback technique is applied to improve the attenuation range. The technique also provides an improvement in phase distortion, which is demonstrated experimentally. The next Section presents a simple technique to compensate for the parasitic effects in the Field-Effect Transistors (FETs) which produce a large phase distortion in the vector modulator. A detailed circuit-level analysis of the RTA basic cell, a cold FET variable-resistance termination, is given and the effect of external source inductance is discussed. The amplitude balance between the two extreme biasing points is improved
by shunting an external resistor across the transistor's drain-source terminal. Experimental results prove the validity of the technique. The advantage of this technique is its simplicity and the small circuit size. Many subsystems employ the RTA structure, including the variable attenuator, bi-phase modulator and vector modulator. The final section of Chapter 4 investigates the nonlinear distortion in the RTA. Closed-form expressions for the third-order nonlinear transconductance is derived, so that the IMD can be determined. The L-band experimental results and analysis provide a good agreement.

Three linearisation techniques for use in a direct-carrier transmitter are investigated in Chapter 5. They are the feedforward, baseband predistortion and adaptive predistortion LINC (Linear Amplification using Nonlinear Components) techniques. The feedforward technique is first applied to a direct-carrier transmitter. The concept is demonstrated at L-band using modular components. The aim of the work is to prove that the technique can reduce the distortion from the modulator, as well as that from the power amplifier, which normally receives all the attention. Hence the core modulator is operated in a strongly nonlinear region. A reference signal is generated from a second modulator operated in a weakly nonlinear region. The simulated and experimental results provide a promising result for both constant and non-constant envelope modulation signals. Section 5.3 proposes a low-complexity and hence a low-cost transmitter architecture, especially suited to millimetre-wave systems. It is composed of only a vector modulator and a medium-power oscillator. A multifunction transmitter is realised using baseband processing techniques. The technique is demonstrated at 60 GHz with a GaAs (Gallium Arsenide) MMIC (Monolithic Microwave Integrated Circuits) balanced vector modulator based on the RTA and a Gunn oscillator. The baseband processing is performed by a D/A card controlled by a personal computer. Many transmitter functions have been successfully implemented in the experiment.
1.2. Outline of the Thesis

Due to the problem of phase and amplitude imbalances in the LiNC transmitter, the technique of adaptive baseband predistortion is developed for LiNC in Section 5.4. An extra I-Q demodulator is added in the feedback path to demodulate the transmitted signal to baseband. The bandwidth trade-off from this technique is resulted from the time delay in the feedback path, especially in the DSP block. An adaptive algorithm is applied to automatically adjust the amplitude and phase of the baseband signal at the second channel. Simulation results in MATLAB™ after the algorithm converges to the optimum values show the successful operation of the technique.

The final chapter draws conclusions from the results and suggests some topics for future work which are relevant to this thesis.
Chapter 2

Linearisation Techniques

2.1 Introduction

With increasingly stringent performance requirements in modern communications systems, linearisation techniques are essential in transmitter and PA (Power Amplifier) design. The need to linearise transmitters and PAs applies from HF (High Frequency) right through to mm-wave (millimeter-wave) applications, from mobile communication [36] to high data-rate Wireless-local Area Network (WANs) [52]. Future mobile communication systems need a high-efficiency and high linearity transceiver.

Currently, there are only a few linearisation techniques that improve both linearity and efficiency. The linearity requirement arises as a result of spectral efficiency demands which lead to the use of non-constant envelope modulation [49]; for example, 4-QPSK (Quaternary Phase Shift Keying), 16-QAM (Quadrature Amplitude Modulation), $\frac{3\pi}{8}$-shifted 8-PSK, etc. However, the performance of any communication systems using these modulation techniques is considerably degraded by any nonlinear components in the system [3]. The result of nonlinearity affecting a modulated signal can be seen as distortion
in the time-domain, spectral regrowth in the frequency-domain \[42\] and error vector in the signal constellation diagram. In communication systems which utilise FDMA (Frequency Division Multiple Access), such effects appear as ACI (Adjacent Channel Interference), out-of-band and spectral distortion in-band, and ultimately increase the BER (Bit-Error-Rate) of the system. Although, theoretically, the constant-envelope modulation techniques, for instance MSK (Minimum Shift-Keying), 8-PSK (Phase Shift-Keying), etc., do not suffer from a linearity problem, spectral-shaping filter functions applied to limit the signal bandwidth whilst minimising ISI (Inter-Symbol Interference) cause the modulated signal to be a variable-envelope waveform. In addition, the transmitted signal from multicarrier systems, for example in base station, cable television, and OFDM (Orthogonal Frequency-Division Multiplexing) transmitters, is as a variable envelope signal, even though each individual channel might be a constant envelope signal.

It is well known that nonlinear effects not only cause an in-band spectral distortion but also out-of-band spectral regrowth. Practically, linearity measurement techniques of systems are various and can be classified as one of two types; either frequency domain or signal constellation measurements. Frequency domain measurement techniques \[42\], for instance, two-tone measurement, spectral regrowth measurement, etc., are simple to set up but do not take the effect of nonlinear distortion on the in-band spectrum into account. EVM (Error Vector Magnitude) measurement techniques which considers the nonlinear effect to be a cause of AM-AM (amplitude-to-amplitude modulation) and AM-PM (amplitude-to-phase modulation) on the signal, are now a new standard tool and grow ever more important. For example, in the enhanced data rate for GSM evolution (EDGE), in which the maximum-to-minimum of envelope values of the modulating signal is approximately 6.78 (17 dB), requires 7% of root-mean square EVM and 22% of peak EVM \[53\], which means
that a high-linearity transmitter is needed.

As stated earlier, high-efficiency requirements can be very demanding, particularly in personal communication applications [36], [55], [45]. In a transceiver front-end, the majority of the DC power consumption is due to the PA. It is well known that to achieve a high-efficiency goal, the PA must be operated in the saturation region, but this results in a more distorted output signal. The trend to optimise this trade-off is to apply a powerful linearisation technique in a high-efficiency PA: for example, applying feedforward (FF) technique to a class AB amplifier. However, the complexity, cost, system requirements, sensitivity and reproducibility must be taken into account. The details of each technique will be discussed in the following sections.

Various linearised transmitter architectures have been proposed over many years [33]. In this section, the theory and practical implementation of these techniques as well as their advantages and disadvantages are discussed. These techniques are feedforward (FF), feedback, Pre-Distortion (PD), Envelope Elimination and Restoration (EER), Linear amplification using Nonlinear Components (LINC) and Harmonic Injection (HI) techniques.
2.2 Feedforward architecture

The FF system [54] has been proposed for application to a PA. Here assuming that the input signal of the system is two-tone, which is illustrated in frequency-domain as shown in Figure 2.1. The input signal is split into two parts by a directional coupler, one part is fed into a main HPA and the other provides as a reference signal to compare with the output signal from the HPA. The error signal is amplified to the same power level as the error signal from the HPA by the error amplifier. A signal without distortion is obtained by subtracting this amplified error with the output signal from the HPA. It should be noted that the distortion signal definition for the FF topology is the difference signal between the output signal from the HPA and the reference signal. Consequently, this topology is capable of reducing not only a nonlinear distortion but also a linear distortion, which is a term used to describe the nonideal gain and phase response across the operating bandwidth [34]. This technique is attractive due to the ease of practical implementation and excellent IMD reduction performance.

The FF structure has several shortcomings. First, the distortion reduction
performance considerably degrades with the gain and phase mismatch of both loops. It is reported [11] that to achieve a 20 dB intermodulation distortion reduction, if the first loop, the so-called signal cancellation loop, has perfect gain and phase match then the second loop (the error cancellation loop) requires a maximum 5% and 5° of gain and phase mismatch error. These requirements cause some difficulties in RFICs (Radio Frequency Integrated Circuits) and MMICs (Monolithic Microwave Integrated Circuits) unless an adaptive technique [11],[63] is employed in the configuration with the complexity and bandwidth trade-offs. Secondly, the power efficiency of the system drops significantly because of the DC power required for the error amplifier and because there is a power loss in the output coupler. For a 3-dB combiner, half of main signal power is internally dissipated in the output coupler. One technique to alleviate this is applying a more loose coupling factor together with a higher gain error amplifier [48]. However, this will in turn increase the DC power consumption of the error amplifier. Thirdly, the increased system complexity, for example, extra phase-shifter, delay, attenuator, etc., increases cost and reduces the overall power efficiency of the system [48].
Harold S. Black invented the two solutions, FF [6] and feedback [7], to overcome the distortion problem. Practical barriers in his day prevented him from realising the FF amplifier but the concept of the feedback seemed worthwhile and realisable in that time. Figure 2.2 shows the basic diagram of a feedback system. The overall gain, $A$, of this system is the reciprocal of the transfer function of the feedback network, i.e. $A = 1/f$, when the $af$ product is made to be much larger than unity. Even though the component of the forward path is very nonlinear, as long as the network in the feedback path is a linear component and the condition of the $af$ product being larger than unity is satisfied, the overall closed loop system behavior is still linear.

Based on the feedback concept mentioned above, there are various forms of feedback linearised transmitter, for example Cartesian loop, polar loop, and RF feedback loop [33],[25]. All the feedback systems use the same feedback condition with a linear device in the feedback path. Figure 2.3 demonstrates the Cartesian feedback topology. The forward path of the system is composed of an I-Q modulator and HPA. The output signal is sampled and fed back to an attenuator and I-Q demodulator. The linearity of the I-Q demodulator needs to be sufficiently good that the distortion generated from the I-Q demodulator is negligible. The output signal from this demodulator is compared with the
Chapter 2. Linearisation Techniques

Figure 2.3: Cartesian feedback topology

input signal in a differential amplifier producing the predistorted signal. Theoretically, the loop will be self-adaptive and a stable steady state is reached. The polar loop and the RF loop are similar except that the techniques used to produce a reference signal in the feedback path and predistorted signal are different. Generally, the polar loop is not used in practice since the technique needs a phase-locked loop for the phase feedback path. The loop can experience locking problems at low amplitude levels and a tracking problem at abrupt phase changes. Hence, the Cartesian and RF feedback loop are more practical than the polar loop.

The feedback techniques have many advantages. Sensitivity of the system is not as critical as the FF system. In addition, few additional components are needed which in turn means little increase in DC power. All these advantages make it possible to realise a system on a single chip. The control circuitry can be implemented using an ASIC (Application-Specific-Integrated Circuit) technology. However, the drawbacks of the system stem from a delay in the feedback path. Not only this effect can limit the system bandwidth but an
unstable condition may also occur.
2.4 Predistortion

\[
F(b(V_j)) = A
\]

Figure 2.4: Predistortion technique

Figure 2.4 shows the block diagram of the predistortion technique for HPAs. The fundamental concept of the technique is predistorting the input signal such that the distortion signal after the predistorter cancels the distortion parts in the HPA. There are two main types of predistortion technique; analogue and digital. The first technique is the simplest form of linearisation for an RF PA. The approach achieves linearity by creating a predistortion function block, which has characteristics complementary to those of the PA, so that cascading them results in a signal with little or no distortion. For a circuit implementation, this predistortion function block can be realised using diodes [60] or transistors [29], [35]. It has been reported [28] that the analogue predistortion technique can achieve 3 dB in adjacent-power improvement for the 1.25 MHz chip rate Code Division Multiple Access (CDMA) signal.

The digital predistorter is one of the most promising linearisation technique because of its high performance and adaptability [10], [56]. Feedback is used only for adaptation of the predistortion nonlinearity.

In the digital predistorter, several predistortion functions have been proposed. A predistorter based on a polynomial function [47] was developed to dominantly cancel the third-order intermodulation. Though the technique can
be extended to cancel the fifth- or seventh-order terms, the complexity increases rapidly. Moreover, the technique relies on good modelling of the PA with a polynomial function, which is not straightforward with some PA types, for example class AB ones. The complex gain predistorter [26] requires a dynamic phase-shifter which is not adaptive. The technique is implemented in digital baseband with RAM (random access memory) lookup tables, with an entry of for each predistorted point in the signal constellation. This is fast and requires very little memory, but is limited to a particular pulse-shaping filter type. The mapping predistorter [46] is the generalised lookup table approach, and is not restricted by the order and type of PA nonlinearity, modulation format, or pulse-shaping filter. The disadvantages of this technique are a large size of lookup table and a slow convergence of the system in the adaptation update. The gain-based predistorter [10] is similar to the mapping predistorter but requires less memory for the lookup table and converges quickly. It has been reported [61] that the digital predistortion technique can achieve nearly 10 dB improvement for the 1.25 MHz chip rate CDMA signal.

The predistortion technique is arguably the most efficient technique to linearise a PA or a whole transmitter for a narrowband application. The technique has advantages of power efficiency, effective distortion reduction, and ease of implementation. The disadvantage of the predistortion technique is a restriction to narrowband applications since it's very challenging to determine the nonlinear characteristic function of a nonlinear system over a large bandwidth. In addition, the distortion reduction capability of this technique is degraded when the linearised PA has a memory effect [59]. There are two important divisions in system theory: nonlinear systems and systems with memory. The fundamental difference between the two is that nonlinearities generate new spectral components, while memory only shapes the existing signal components, because the output signal is not only a function of the instantaneous
input signal, but also a function of previous input values. Consequently the PA's amplitude and phase distortions are amplitude and frequency dependent. Applying adaptive predistortion technique to correct the amplitude dependent characteristic creates a delay in the feedback path and hence the operation bandwidth is more reduced. Consequently, to achieve a large bandwidth operation and a whole dynamic range of RF power for the predistortion technique also needs some techniques to correct the memory effect in the PA.
2.5 Envelope Elimination and Restoration

The Envelope Elimination and Restoration (EER) [51] or Khan technique can be applied to a whole transmitter or only a PA. This technique combines a highly efficient but nonlinear PA with a highly efficient envelope amplifier to implement a high-efficiency linear PA or transmitter. Figure 2.5 shows the application of EER with a PA. A limiter is applied to eliminate the input signal envelope, yielding a phase-modulated signal which can be amplified efficiently by a high-efficiency PA such as class C, D, E or F. The signal envelope, which is a low-frequency component, is restored using an envelope detector and amplified with a very high-efficiency audio amplifier, i.e. a PWM (Pulse-Width Modulation) class-S. The output signal from the audio amplifier is effectively supplying the DC power to the RF amplifier, thereby applying high-level amplitude modulation. Theoretically, this technique can achieve 100% DC to RF power efficiency at all envelope levels of the modulating signals since both types of amplifiers are theoretically 100% efficient. The linearity of an EER transmitter does not depend on the linearity of the PA but upon the accuracy of reproduction of input signal's amplitude and phase information [50].

Figure 2.5: Envelope elimination and restoration topology
bandwidth of the audio amplifier and the differential delay between the envelope and phase modulation at the PA are additional linearity factors. With the existence of differential delay between the envelope and the phase path, an additional delay is needed which in turn limits the operation bandwidth of the technique. The low-frequency circuit part can be implemented in a single integrated circuit or using a digital signal processing technology.
2.6 Linear amplifier with nonlinear components (LINC)

The LINC [20],[38] transmitter and its derivative, the Combined Analogue Locked-Loop Universal Modulator (CALLUM) [5] method, are based on the immunity of a constant envelope signal to amplitude nonlinearities. The baseband signal or bandpass signal is separated into two constant envelope component signals. All of the amplitude and phase information of the original signal is contained in phase modulation of the component signals. Consequently both the PAs operated in LINC can be highly nonlinear yet high efficiency, for example class AB, C, D, E or F, which is a major attraction of this technique. The baseband processing can be realised by DSP, FPGAs (Field Programmable Gate-Arrays) or ASICs.

![Diagram of LINC transmitter]

Figure 2.6: LINC transmitter

There are some disadvantages of this technique. For example, the performance of LINC critically relies on the phase and gain match of the two PA paths. This problem can be corrected by applying adaptive feedback to the LINC transmitter [57],[58] which increases the complexity and limits the operational bandwidth of the transmitter due to the internal delay in the feedback.
path. More seriously, the architecture utilises a power combiner at the output which has an insertion loss: In the case of realising this component with a hybrid structure, the difference signal between the two PA paths appears at the difference port and is wasted in a 50 Ω termination, and this lowers the power efficiency dramatically[38]. This problem was recently alleviated by embedding RF-DC conversion circuitry at the difference port to reuse the power wasted in LINC [38].
2.7 Harmonic injection

This technique [1] is based on the assumption that the cubic term causes the main contribution to spectral regrowth, so that injected harmonic signals can be utilised to reduce the distortion. Figure 2.7 shows the diagram of this technique applied to a PA. The input signal is split into two paths, the main and the frequency multiplier paths, which generates the second harmonic product from the input signal. To reduce IMD in the HPA, the amplitude and phase of the second harmonic product must be adjusted to a particular value. This is achieved by using a variable attenuator and phase-shifter. Another form of this technique, the so called interstage second harmonic enhancement technique [32], is implemented using an additional amplifier to produce the second harmonic product and by applying a bandpass filter for the fundamental and the second harmonic product.

The advantage of the technique is its easy implementation and it has no stability problem. There is no circuitry required at the PA output, which minimises unwanted losses at this critical part of the transmitter. However the disadvantage of the system is that the IMD reduction performance of the technique is very sensitive to the amplitude and phase of the injected signal. The applications of this technique with the complex baseband signals is still under investigation.
2.8 Conclusion

Communication systems with spectrally-efficient modulation schemes and all multicarrier transmitters need a highly linear transmitter with good power efficiency. This can be achieved using a linearisation technique. Amongst the linearisation techniques described here, FF is by far the technique which yields the best IMD reduction performance. Predistortion yields an acceptable performance and a good power efficiency. It is widely used in transmitters for personal communications. The feedback technique provides good performance for narrow-band applications. The EER, LINC and harmonic injection techniques are emerging techniques. Some of these have been applied for certain applications; for example a Combined Analogue Locked-Loop Universal Modulator (CALLUM) base-station transmitter has been constructed for a Terrestrial Trunked Radio Access (TETRA) system. To summarise the PA linearisation techniques, their advantages and disadvantages are shown in Table 2.1.

Table 2.1: Comparison of the current linearisation techniques

<table>
<thead>
<tr>
<th>Technique</th>
<th>Linearity Improvement</th>
<th>Bandwidth</th>
<th>Complexity</th>
<th>Power Efficiency</th>
</tr>
</thead>
<tbody>
<tr>
<td>FF</td>
<td>high</td>
<td>wide</td>
<td>high</td>
<td>low</td>
</tr>
<tr>
<td>Feedback</td>
<td>moderate</td>
<td>narrow</td>
<td>medium</td>
<td>high</td>
</tr>
<tr>
<td>PD</td>
<td>moderate</td>
<td>moderate</td>
<td>low</td>
<td>high</td>
</tr>
<tr>
<td>EER</td>
<td>moderate</td>
<td>narrow</td>
<td>medium</td>
<td>high</td>
</tr>
<tr>
<td>LINC</td>
<td>high</td>
<td>wide</td>
<td>medium</td>
<td>medium</td>
</tr>
<tr>
<td>HI</td>
<td>moderate</td>
<td>narrow</td>
<td>low</td>
<td>high</td>
</tr>
</tbody>
</table>
Chapter 3

Linearisations Techniques for Microwave Mixers

3.1 Introduction

Nowadays modern communications transmitters, such as in mobile and satellite communication often employ a nonconstant envelope modulation technique to achieve high bit-rate data transmission when the spectrum resource is limited. Nonconstant envelope modulation techniques are sensitive to the nonlinearity of components, and in-band distortion and ACI are created [3]. This results in a communication performance degradation. Generally the nonlinearities are mainly produced by two nonlinear components in a transmitter; mixers and PAs [33]. In a receiver, a downconversion mixer is the key component which determines the IMD levels [41]. In addition, nonlinearity in downconversion mixer is one of the key factors to limit the sensitivity of the receiver. In this Chapter, two new techniques for linearising a mixer are proposed and experimentally investigated. The first technique is a simplified FF mixer. The second technique is the harmonic injection technique, adapted for a mixer.
For the simplified FF technique, the system topology is single-loop rather than the double-loop used in conventional FF system. Analysis of the technique will show how it operates. The sensitivity of the system is analysed and it is shown that this technique is less sensitive than the conventional technique.

The difference frequency injection (DFI) technique (the HI derivative) is proposed here for linearising a mixer. This technique has two main advantages over the related second-harmonic injection technique: (1) the injected signal is at a low frequency; (2) it has a good distortion reduction capability. Since the improvement produced by the technique depends on the amplitude and phase of the injected signal, the sensitivity of the system has also been analysed. The validity of the technique is experimentally demonstrated at L-band with a double-balanced mixer (DBM).
3.2 A Simplified Feedforward Mixer

The FF technique for linearising a mixer, the so-called modified FF mixer shown in Figure 3.1, was firstly proposed by Ellis [24]. The input signal to the system is assumed to be a two-tone signal, which is illustrated in frequency-domain. Similar to the FF PA, the system is composed of two loops; a signal cancellation loop and an error cancellation loop. The mixer that needs to be linearised is labelled as M_1. The major difference from the standard FF approach for amplifiers is that the modified FF system employs an auxiliary mixer (shown as M_2 in Figure 3.1) operated at a low-power level by the attenuator A to generate the error signal in the first loop. The error amplifier G amplifies the error signal such that it is equal to the same amplitude distortion level from the mixer M_1. The phase-shifters and attenuators are applied to match the phase and amplitude in both loops.

By using the same technique, a simplified FF approach for linearising a mixer, as shown in Figure 3.2 is obtained.

![Figure 3.1: A modified feedforward mixer](image)

The proposed system is composed of two identical mixers, M_1 (the linearised mixer) and M_2 (the auxiliary mixer). In theory, these two identical mixers
provide the same intermediate frequency (IF) output power level and same IMD power level when operated at the same RF and LO (local oscillator) power level. The two-tone input signal is split in power by a power splitter. The input voltage signal to $M_2$ will be lower in power than the input signal at $M_1$ by $A$ due to the attenuator placed at the input of $M_2$. Since the IF and IM$_3$ powers vary with the RF input power, $M_2$ has an IF and IM$_3$ output power lower than $M_1$. The auxiliary amplifier $G$ amplifies the output signal from $M_2$ such that the distortion signals in $M_2$ are equal to $M_1$’s distortion signals (both are shown as $X$ dBm in Figure 3.2). These distortion signals can be made to cancel with suitable adjustment of the phase shifter but their IF signals are different in power level and so do not cancel. It is evident from Figure 3.1 and Figure 3.2 that the proposed system has a lower complexity than the modified FF mixer, but it still provides the same distortion reduction capability in principle.
3.2. A Simplified Feedforward Mixer

3.2.1 System Analysis

Considering Figure 3.2, assuming that the auxiliary amplifier in the system is linear. Let $v_i$ be the injected input signal power of the system, $A$ be the attenuation of the attenuator, $G$ be the auxiliary amplifier gain, $v_{i1}$ and $v_{i2}$ be the input signals in M1 and M2, respectively, and $v_{o1}$, $v_{o2}$ and $v_{o3}$ are as labelled in the diagram of Figure 3.2. For simplicity’s sake, the input signal of the system is assumed to be two-tone signal whose amplitudes are $A_1$, i.e. $v_{i1} = A_1 \sin(\omega_1 t) + \sin(\omega_2 t)$, so $v_{o2} = \frac{v_{o1}}{A}$. A mixer can be modelled as an RF switch which is controlled by the LO [42]. Consequently, a mixer output $v_{\text{mix}}$ is related to a mixer input signal, $v$, by the series approximation as,

$$v_{\text{mix}} = \left( a_1 v + a_3 v^3 + a_5 v^5 + \cdots \right) v_{LO} \quad (3.1)$$

where $a_1, a_3, \cdots$ is related to the bias condition, LO level and device characteristic in the mixer. It should be noted that the even-order terms are neglected since the IMD is mainly resulted from the odd-order terms. The higher-order terms will be neglected if the mixer is operated well below its intercept point.

Substitute $v_{i1}$ and $v_{i2}$ into (3.1) and $v_{LO} = 2\sin(\omega_{LO} t)$ then,

$$v_{o1} = \left( a_1 A_1 \left[ \sin(\omega_1 t) + \sin(\omega_2 t) \right] + a_3 A_1^3 \left[ \sin(\omega_1 t) + \sin(\omega_2 t) \right]^3 + \cdots \right) 2\sin(\omega_{LO} t) \quad (3.2)$$

$$v_{o3} = \left( a_1 A_1 \left[ \sin(\omega_1 t) + \sin(\omega_2 t) \right] + a_3 \left[ \frac{A_1}{A} \right]^3 \left[ \sin(\omega_1 t) + \sin(\omega_2 t) \right]^3 + \cdots \right) 2\sin(\omega_{LO} t) \quad (3.3)$$

This $v_{o3}$ is applied to the amplifier (gain = $G$) and phase-shifter, providing the signal $v_{o2}$. Hence,

$$v_{o2} = G \left( a_1 A_1 \left[ \sin(\omega_1 t) + \sin(\omega_2 t) \right] + a_3 \left[ \frac{A_1}{A} \right]^3 \left[ \sin(\omega_1 t) + \sin(\omega_2 t) \right]^3 + \cdots \right) 2\sin(\omega_{LO} t) \quad (3.4)$$

The amplifier and phase-shifter adjust the amplitude and phase of IM3s in $v_{o3}$ such that they are equal in amplitude but opposite in phase with those in $v_{o1}$. 
Chapter 3. Linearisations Techniques for Microwave Mixers

By taking only five nonlinearity order terms in the mixer model into account \((a_1 - a_5\), expanding (3.2) and (3.4), and collecting the IF and IM\(_3\) terms, one obtains:

\[
v_{b1} = \begin{cases} 
  a_1 A_1 + \frac{3}{4} a_3 A_1^3 + \frac{5}{8} a_5 A_1^5 & : \text{IF} \\
  -\frac{1}{4} a_3 A_1^3 - \frac{5}{16} a_5 A_1^5 & : \text{IM}_3 \\
  \frac{1}{16} a_5 A_1^5 & : \text{IM}_5 
\end{cases}
\]  

(3.5)

where IM\(_5\) is the fifth-order intermodulation distortion. With a condition of balancing only the IM\(_3\) terms, from (3.5) and (3.6) one obtains

\[
v_{b2} = \begin{cases} 
  G(a_1 \frac{A_1}{A} + \frac{3a_3}{4} \left[\frac{A_1}{A}\right]^3 + \frac{5a_5}{8} \left[\frac{A_1}{A}\right]^5) & : \text{IF} \\
  G(-\frac{a_3}{4} \left[\frac{A_1}{A}\right]^3 - \frac{5a_5}{16} \left[\frac{A_1}{A}\right]^5) & : \text{IM}_3 \\
  G(\frac{a_5}{16} \left[\frac{A_1}{A}\right]^5) & : \text{IM}_5 
\end{cases}
\]

(3.6)

(3.7)

The approximated solution for \(G\) \((G \approx A^3)\) is achieved by assuming that the input power to \(M_1\) is sufficiently small \((A_1 < 1)\). The IM\(_3\) reduction capability of the system is reduced when the RF becomes large, violating the truncated series approximation. Applying the approximated gain leads us to achieve a largely reduced power level of IM\(_3\) but an increased power level in the desired output IF. One major factor that can affect system operation is the nonlinearity of the amplifier which can cause an excess of IM\(_3\) power level. The other required amplifier property is a constant insertion gain and phase over the entire IF bandwidth, because the system is sensitive to phase and amplitude mismatch between the upper and the lower branch. This unbalanced condition of phase and amplitude can result from the amplifier, coupler and from having non-identical mixers. Such effects must be carefully estimated when realising this approach in system design.
3.2. A Simplified Feedforward Mixer

3.2.2 Gain and Phase Mismatch Effects

The sensitivity of the IM3 reduction performance of the proposed system is now analysed based on the block diagram of the generalised FF system shown in Figure 3.3. The functional blocks $M_1$ and $M_2$ represent the main and the auxiliary nonlinear components. The amplifier $\mu_1$ and $\mu_2$ represents the auxiliary amplifiers. The components $\alpha_1$-$\alpha_3$ are the gain blocks including the couplers, loss from the phase shifters and the attenuators in the FF system. The summing components in Figure 3.3 are assumed to be lossless.

Figure 3.3: Generalised FF mixer system for analysing gain and phase mismatch

Figure 3.4: The model of nonlinear components for studying the gain and phase mismatch effects
For studying gain and phase mismatch effects, the nonlinear components ($M_1$ and $M_2$) can be modelled as shown in Figure 3.4. Let $y$ be the output distorted signals and $x$ be the input signal, then

$$y = ax + \epsilon,$$

(3.8)

where $\epsilon$ is the error signal and $a$ is the gain of the nonlinear systems at the desired harmonic output. In Figure 3.3, let $\delta_i$ and $\Delta \phi_i$ represent gain and phase mismatch of the $i$th loop, respectively. It should be noted that this generalised system can be modified to represent the proposed system by setting $\mu_2$ to be one and $\alpha_3$ to be zero.

The effect of gain and phase mismatch is inherently significant due to the system being based on signal cancellation in the power combiners. Including the effect of these mismatches in the function of $\alpha_2$ and $\alpha_3$ regarding to the two power combiners in the system, one obtains:

$$\alpha_2 = |\alpha_2|(1 + \delta_1)e^{-j\Delta \phi_1}$$

(3.9)

$$\mu_2 = |\mu_2|(1 + \delta_2)e^{-j\Delta \phi_2}$$

(3.10)

Then the suppression of distortion of this generalised loop, $\varepsilon_G$, is

$$\varepsilon_G = \alpha_3 + \mu_2(1 + \delta_2)e^{-j\Delta \phi_2}[\alpha_2 - \mu_1\alpha_1^3(1 + \delta_1)e^{-j\Delta \phi_1}]$$

(3.11)

It should be noted that term $\alpha_1^3$ comes from the 3 to 1 slope of the IM$_3$ vs. RF input power characteristic.

The IM$_3$ sensitivity of the of the simplified FF mixer is simply obtained from (3.11) by setting $\alpha_3 = 0$ and $\mu_2 = 1$. Since the system is composed of one loop only then the gain and phase mismatches of the second loop must be all zero, i.e. $\delta_2 = \Delta \phi_2 = 0$. Hence, the magnitude of distortion suppression of the proposed system, $\varepsilon_a$ is

$$\varepsilon_a = |\sqrt{1 - 2\alpha_1^2\mu_1(1 + \delta_1)\cos \Delta \phi_1} + [\alpha_1^2\mu_1(1 + \delta_1)]|^2$$

(3.12)
3.2. A Simplified Feedforward Mixer

From (3.12), if the effect of phase and gain mismatches are eliminated, i.e. set $\delta_1 = 0$ and $\Delta \phi_1 = 0$, then one obtains the maximum distortion suppression of the proposed system when the auxiliary amplifier gain is equal to the inverse value of the third-power of the attenuation gain $(G = \frac{1}{\alpha_1^3})$. This is the same result obtained from (3.7), i.e. $G = A^3$.

Figure 3.5 illustrates the distortion suppression sensitivity of the simplified FF mixer to gain and phase mismatch obtained from (3.12). The results suggest that a 5% gain and 5° phase mismatch is needed for 20 dB IM3 improvement. Also, with the generalised FF distortion suppression, one can obtain the effect of gain and phase mismatch on the modified FF system by substituting $\alpha_1 = \alpha_2$ and $\alpha_3 = \mu_1 = 1$. Since the original FF mixer is composed of two loops, it suggests, by inspection, that the proposed system is less sensitive to gain and phase mismatches.
3.2.3 Measured Results and Discussion

To verify the system concept, two identical modular diode-ring mixer operating at 70 MHz IF are utilised. With the 6-dB attenuator at the front-end, 18 dB amplifier gain in the lower branch, estimated from the approximation formula in (3.7), is employed which is implemented by cascading a fixed-gain amplifier with a variable attenuator. In Figure 3.6, a standard reflection-type variable phase shifter [40] operating at 70 MHz is constructed. Both the phase shifter and a digital-step attenuator are employed to adjust the phase and amplitude. The narrowband property of this phase-shifter stems from the 90° hybrid coupler used in the phase-shifter. An experiment of this system shown in Figure 3.7 is set up with 1.93 GHz LO and two-tone RF signals, at 2.0 and 2.0001 GHz. The fixed-gain amplifier has 35 dB gain at 70 MHz and a digital-step attenuator is utilised to adjust the amplifier gain to achieve the maximum reduction of IM3. By measuring the insertion loss in the phase-shifter, digital-step attenuator and connectors, the overall gain applied at the M2 output is 17.8 dB, which is close to the estimated gain, 18 dB, obtained from (3.7). Figure 3.8 and 3.9 show the results which were achieved by applying a control voltage at $-5.15 \text{ V}$ in the phase-shifter giving the 180° phase difference. A closer look at Figure 3.8 shows that the output IF signal is increased by around 2 dB because of the amplifier in the lower branch. It is also
3.2. A Simplified Feedforward Mixer

Figure 3.7: The experimental setup of the proposed system

shown in Figure 3.8 that the 1-dB compression point of the proposed mixer is higher than that of a single mixer. The conversion loss of the single mixer is around 7.2 dB where that of the simplified FF mixer is 5.3 dB. The IM$_3$ of the single mixer and the simplified FF mixer are shown in Figure 3.9. It is evidently shown that there is a significant improvement of IM$_3$ from the proposed technique. The unequal IM$_3$ of the simplified FF mixer might come from the memory effect of the auxiliary amplifier and the narrow band property of the phase-shifter. The IM$_3$ improvement of the proposed technique over a single mixer is depicted in Figure 3.10, the maximum improvement of IM$_3$ is 32.8 dB at $-6.5$ dBm input RF level. At sufficiently high RF input levels, the system performance degrades due to amplifier nonlinearity and the 3-dB slope rule is violated. In Figure 3.11, a 900 MHz 96 Kbs/s bit-rate 16-QAM with a 0.35 roll-off factor raised cosine Nyquist filter is injected to the proposed system at $-8$ dBm power level. The vertical axis shows the normalised power spectral density since the proposed system has a gain causing an unequal power spectrum level between a single mixer and the proposed system. It is clearly shown that the adjacent channel interference is reduced in the proposed system.
Figure 3.8: Comparison of IF between a single mixer and the proposed system.

Figure 3.9: Comparison of IMD between a single mixer and the proposed system.
3.2. A Simplified Feedforward Mixer

Figure 3.10: IM₃ reduction performance compared to a single mixer

Figure 3.11: Comparison of 16-QAM power spectrum of the proposed system and a single mixer
3.3 Difference Frequency Injection Technique

The harmonic injection technique was first proposed and applied to a PA [1],[31]. The technique is applied in either of two forms, the second harmonic and the difference frequency technique. This technique is based on using the device nonlinearity itself to cancel the IM₃ by injecting a signal with appropriate amplitude and phase [1]. In this Section, the technique of difference frequency injection (DFI) is investigated for mixer linearisation for the first time. By assuming that the device is operated in a weakly nonlinear region, the injected amplitude and phase can be determined analytically. The analysis results are confirmed with the computer simulation results. The proposed technique is also experimentally demonstrated at 1.8 GHz.

![Diagram of Difference Frequency Injection System Block Diagram](image)

Figure 3.12: Difference frequency injection system block diagram

3.3.1 Theory of the DFI Technique

For simplicity sake, the linearised mixer is assumed to be an upconversion mixer. Figure 3.12 shows the block diagram of the difference frequency in-
Difference Frequency Injection Technique

3.3. Difference Frequency Injection Technique

jection technique applied to a microwave mixer used for upconversion, where the upper sideband output is desired. Let the input signal of the mixer be two-tones at \( \omega_1, \omega_2 \); then the difference frequency injection signal is \( \omega_2 - \omega_1 \). Hence,

\[
v_{in} = A \left[ \cos(\omega_1 t) + \cos(\omega_2 t) \right] + A_L \cos((\omega_2 - \omega_1) t + \phi_L) \tag{3.13}
\]

where \( A_L \) and \( \phi_L \) represent the amplitude and phase of the difference frequency injected signal. Representing the nonlinearity in the mixer with a power series, and neglecting the effect of AM-PM in the mixer, the output voltage signal from the mixer can be written as:

\[
v_{out} = \sum_{k=0}^{\infty} \cos(k\omega_{LO} t) \sum_{n=0}^{\infty} g_{mn} v_{in}^n \tag{3.14}
\]

where \( g_{mn} \) is the \( n^{th} \)-order voltage gain which depends on the nonlinearity of the device. We now assume that the mixer is weakly nonlinear and that the highest nonlinearity degree is 3. Applying (3.13) in (3.14) and considering the \( \text{IF} \) (3.15) and \( \text{IM}_3 \) (3.16) terms, one obtains:

\[
A \left[ g_{m1} + \frac{3}{2} g_{m3} \left( \frac{3A_L^2}{2} + A_L^2 \right) \right] \cos(\omega_{LO} t + \omega_2 t) + g_{m2} A_L A \cos(\omega_{LO} t + \omega_2 t - \phi_L) \tag{3.15}
\]

\[
g_{m2} A A_L \cos(\omega_{LO} t + (2\omega_2 - \omega_1) t - \phi_L) + \frac{3}{4} g_{m3} A^3 \cos(\omega_{LO} t + (2\omega_2 - \omega_1) t) + \frac{3}{4} g_{m3} A A_L^2 \cos(\omega_{LO} t + (2\omega_2 - \omega_1) t - 2\phi_L) \tag{3.16}
\]

From (3.16), it can be shown that to null the \( \text{IM}_3 \) harmonic contents, the amplitude and phase of the injected low-frequency signal need to be properly selected. With some mathematical manipulations, the conditions to cancel the \( \text{IM}_3 \) are obtained as follows:

\[
A_{LOpt} = \left[ \frac{2 g_{m2}}{3 g_{m3}} + \sqrt{\left[ \frac{2 g_{m2}}{3 g_{m3}} \right]^2 - A^2} \right] \tag{3.17}
\]
\[ \phi_{\text{opt}} = \begin{cases} \pi \cos^{-1} \left( \frac{2g_{m2}}{3g_{m3}} A \right) & \text{if } A \leq \left| \frac{2g_{m2}}{3g_{m3}} \right| \\ \pi & \text{else} \end{cases} \] (3.18)

From the result one can conclude that the required injected signal power and phase changes with the input power level. It is shown in (3.18) that at small input power, the phase of the injected signal is constant at 180°. Also, it is interesting to note that if the mixer even-order nonlinearity terms \( g_{m2}, g_{m4}, \cdots \) are zero, which is true for many balanced mixer structures, then the optimum amplitude and phase of the injected signal from (3.17) and (3.18) can be further simplified to be:

\[ A_{\text{opt}} = A \] (3.19)

\[ \phi_{\text{opt}} = \frac{\pi}{2} \] (3.20)

### 3.3.2 Sensitivity of the DFI technique

The sensitivity of this technique for reducing the IMD is now analysed for the case of a DBM. From (3.16), if the second-order nonlinearity term is small and the mixer is operating in a weakly nonlinear region and the injected signal amplitude and phase are \( A + \delta A \) and \( \phi + \delta \phi \), then the \( IM_3 \) term (in power) can be written as follows:

\[ IM_3 = 10 \log_{10} \left( \frac{\cos \phi - \left[ 1 + \frac{\delta A}{A} \right]^2 \cos(\phi + \delta \phi) \right)^2}{2 \text{Re}(Z_L)} \] (3.21)

where \( Z_L \) is the load impedance.

Figure 3.13 shows the surface plot of \( IM_3 \) performance against amplitude deviation ratio \( \left( \frac{\delta A}{A} \right) \) and phase error \( \delta \phi \) which are deviated from the optimum values obtained from (3.19) and (3.20). It is very clear that the \( IM_3 \) reduction performance is very sensitive to amplitude and phase. However, this optimum
amplitude and phase can only be applied if the effect of AM-PM is negligible. This is true when a DBM is weakly nonlinear.

Figure 3.14 shows the simulated results on MATLAB™ of a mixer whose nonlinearity coefficients are \( g_{m1} = 0.45, g_{m2} = 0, g_{m3} = -1.92 \). The input is two equal-amplitude tones which mix with the LO. The input power is swept from \(-30\) to \(-5\) dBm. The low frequency signal was injected at the input port with the amplitude and phase obtained from (3.19) and (3.20). The simulation results show that significant IM3 reduction can be achieved but with an increase in IM5. The IM5 increase is produced from the effect of the odd-order nonlinearity to the difference-frequency tone. It is also interesting to see that at a certain high input power, the RF output power of the technique drops. This is a result of the effect of the DFI signal.
3.3.3 Measured Results

To study the technique, the ZEM-4300 DBM from Mini Circuits was used. The mixer is wideband, operating from DC-4300 MHz. The mixer is modelled using a PN-junction diode model and the obtained parameters are: \( R_s = 5 \Omega, I_s = 5 \ pA, \phi = 0.8, C_j = 0.1 \ pF \). The LO was set to 1.8 GHz with 7 dBm power. The RF signals were at 0.01 and 0.0105 GHz. The DFI signal phase was fixed at 90° while its power was equal to the IF input power. From Figure 3.15, it is shown that the IM3 is considerably reduced at low input power level. The IM3 reduction performance tends to be degraded when the input signal becomes large. It is possibly an AM-PM effect which causes the optimum injection phase to deviate from \( \frac{\pi}{2} \). However, this result shows that the equations for predicting the optimum amplitude are valid as long as the device is operated in weakly nonlinear region and the even-order nonlinearity terms are negligible. There is some slight degradation in IF output signal.
3.3. **Difference Frequency Injection Technique**

Figure 3.15: Comparison of simulated results DBM and DFI-DB mixer power when applying the DFI technique as predicted in (3.15).

![Graph showing comparison of simulated results](image)

Figure 3.16: Measured result before applying DFI technique

In the experimental setup, a 1.8 GHz 8 dBm LO signal was obtained from an Agilent ESG synthesised signal generator. The two RF input tones and the injected signal were obtained from an arbitrary waveform generator (TGA1224), since this gave a simple means of controlling the amplitude and phase of the
injected signal. The RF input frequencies were 10 and 10.5 MHz. This low frequency had to be used because after some experimentation it was revealed that the TGA1224 could only provide phase control at low sample rates. The output spectrum was obtained from an Agilent spectrum analyser. Figure 3.16 shows the output spectrum of the signal at 4.5 dBm input power before the proposed technique is applied. At this power level, the low frequency signal at 500 kHz with a power level at $-6.8$ dBm is applied. The low-frequency signal phase was set at $-10^\circ$. The large deviations from the optimum values estimated from (3.19) and (3.20) are resulted from the frequency-dependent characteristics of the DBM. The output spectrum of this mixer after applying this technique is shown in Figure 3.17. The $\text{IM}_3$ is reduced by more than 28 dB.

Figure 3.18 shows the intercept diagram of the mixer before and after applying the proposed technique. With a fixed phase and with the injected signal power proportional to the IF power, the input power was swept from $-5$ dBm to 4.5 dBm. The $\text{IM}_3$ reduction performance is degraded at the low power end, which is far away from the power level at which optimum $\text{IM}_3$ reduction was initially set.
3.3. Difference Frequency Injection Technique

Figure 3.18: The measured intercept diagram of the proposed technique and the DBM
3.4 Conclusions

A simplified FF system for mixer application which has a lower complexity but maintains the same IM$_3$ reduction performance with the previously reported technique has been presented. Two-tone experimental results show a greatly reduced level of intermodulation products, and the improvement in linearity has also been confirmed with an experiment on a digital communications signal. Compared with the original FF mixer, the analysis suggests that the simplified FF mixer is less sensitive to phase and gain mismatch for good IM$_3$ reduction performance. However, the nonlinearity of the auxiliary amplifier is of concern when realising this concept in practice. Fortunately, amplifiers have much higher intercept points than mixers so the technique is still viable.

The DFI technique has been applied to a double-balanced diode mixer at 1.8 GHz. The concept of this technique is to use the nonlinearity of the device itself to generate and internally cancel IMD. Based on a system analysis, the appropriate amplitude and phase of the injected signal have been determined. The validity of this technique has been proved through both simulation and experiment. It has been found that for a DBM, the phase of the injected signal is constant at 90° as long as the mixer is operated in a weakly nonlinear region. Also, the amplitude of the injected signal should be made proportional to the input signal. Although the experiment was performed for the upconversion mixer case, since it was easier to implement practically, the technique can also be applied to a downconversion mixer.

Both techniques presented in this chapter have been experimentally demonstrated at a system level using modular components. However, the techniques can also be implemented at a circuit level and this is expected to enable the realisation of high linearity mixers with a modest increase in complexity.
Chapter 4

Analogue Reflection-type Attenuator and Vector Modulator

4.1 Introduction

Analogue attenuators can be applied to a wide variety of microwave signal processing applications [52]. For example, they are often found in automatic gain control loops, vector modulators (VMs), adaptive beam-forming networks and FF amplifiers. The standard analogue reflection-type attenuator (RTA) [40] is well known and widely applied at microwave and mm-wave frequencies due to its simplicity. The circuit consists of a 90° hybrid coupler and two variable resistance terminations. The attenuation range is theoretically limited by the range of reflection coefficient provided by the devices used as variable resistors. In practice, parasitics in the devices cause amplitude and phase errors which are unacceptable in some applications, such as in modulators for digital communications. The phase distortion effect can be alleviated by applying a
balanced structure, but this increases circuit area and requires a complementary voltage control arrangement [4]. Consequently, the technique to improve the attenuator control range for the RTA is presented in Section 4.2. A resistor is applied to shunt to the input and isolation of the coupler which can be viewed as a feedback resistor for the RTA. It has also found that this feedback resistor improves the phase distortion in RTA, hence a high performance bi-phase modulator can be obtained from the modulator. The analysis and experimental results at 900 MHz are described.

"Mixer" and "modulator" are the important subsystems in a transceiver for translating frequency. Many textbooks and technical reports refer these two terms in many places without giving their definitions. For clarification sakes, the definition of these two nonlinear definitions from "Chapter 23: Design considerations for BJT active mixers " [43] are restated here without rephrasing:-

"MODULATORS : These devices can be viewed as "sign-changers". The two inputs generate an output which is simply one of these input, multiplied by just the sign of the other.

MIXERS : A mixer is a specialised modulator for frequency translation purposes, often placed near the antenna."

Nowadays, two orthogonal channel baseband input modulator, so called VM, is generally appeared in a transmitter for modern communication applications. A VM structure that is practically suitable for microwave and mm-wave applications is a two channels analogue attenuator rather than an analogue cascaded with a full 360° variable phase-shifter [52].

Due to the existence of parasitics in MESFETs, the VM employing the RTA technique needs a balanced structure to obtain a fully 360° signal constellation diagram [4]. The balanced technique, though providing a complete signal con-
4.1. Introduction

stellation, requires more than double the circuit area and increases the control complexity. The proposed technique provides a smaller circuit size and improve the amplitude balance and phase distortion. The analysis in Section 4.3 quantifies the phase distortion and gives an insight into the effects of these parasitics. A simple technique is proposed for improving the performance. Analytical expressions for the design are given. The experimental results at L-band are presented for the technique.

As mentioned earlier, the RTA has a wide range of applications. It is used as a basic cell for bi-phase and VMs in transmitters, amplitude and phase control devices in linearisers, and in adaptive phased array systems. Distortion in the modulator can degrade the performance of both transmitters and linearisers. Consequently, the linearity in the RTA is investigated in this Chapter. Section 4.4 presents the IMD analysis details for the RTA. It is shown that IMD is a function of attenuation level. The issue of optimum device selection to obtain good IMD performance is also addressed.
4.2 Improved Bi-phase Amplitude Modulator using Feedback

In this section, a resistive feedback technique is proposed which improves the attenuation range as well as a reduction phase error. The concept is proved in mathematical analysis and verified with simulation and measurement at 900 MHz with discrete component devices.

4.2.1 The Analogue RTA and its Limitations

Figure 4.1 shows the standard RTA. An input signal is split into two equal amplitude but orthogonal signals by the directional coupler. Two variable resistors ($R_r$) connected at the direct and coupled port are realised by using, for example, PIN diodes, MESFETs, or HEMTS. With MESFET devices which are biased in the "cold-FET" condition, the equivalent circuit consists mainly of the variable drain-source resistance [27] and some parasitic capacitance and inductance.

![Figure 4.1: RTA](image)

Figure 4.2 shows the small-signal equivalent circuit of a cold MESFET device. The devices are biased at zero drain-source voltage and a certain gate-source control voltage. In theory, this bias provides an ideal variable resistor
4.2. Improved Bi-phase Amplitude Modulator using Feedback

mainly resulting from the drain-source resistance ($R_{ds}$). Yet, practically, the parasitic components (except the gate resistance, $R_g$, and inductance, $L_g$) degrade the attenuator performance. It should be noted that both $R_g$ and $L_g$ do not affect to the modulator performance if $R_{ext}$ is sufficiently large.

The parasitic effect can be described in three different biasing operations. For a zero gate-source biasing voltage, the drain ($R_d$) and source ($R_s$) resistances dominate, causing an inherent loss of the modulator. The source ($L_s$) and drain ($L_d$) inductances cause the imaginary part of the input impedance of the FET at the drain ($Z_{in}$) to be positive. When the gate-source biasing voltage ($V_{gs}$) is more negative, the nonzero parasitic capacitances mainly dominate, and so the imaginary part of $Z_{in}$ is slightly negative. This causes $Z_{in}$ to deviate from the centre point of the Smith chart, which in turn limits the minimum attenuation. When the applied bias is near pinch-off, $Z_{in}$ is dominated by the junction capacitances ($C_{gs}, C_{gd}$). This parallel capacitance changes the insertion phase of the attenuator.
4.2.2 Mathematical Analysis

Here the analysis of the analogue RTA is presented. The number labels of the input and output port for the RTA shown in Figure 4.3 are port 1 and 2, respectively. The 90° hybrid coupler ports are labelled as follows: "a" for the input port, "b" for the direct port, "c" for the coupled port and "d" for the isolated port. For simplicity sake, the following assumptions are applied:

1) small signal condition
2) two identical reflection terminations, and
3) the hybrid coupler is perfectly symmetrical, \( s_{11c} = s_{22c} = s_{33c} = s_{44c}, s_{21c} = s_{12c} = s_{34c} = s_{43c}, s_{31c} = s_{13c} = s_{24c} = s_{42c} \) and \( s_{14c} = s_{41c} = s_{23c} = s_{32c} \).

After some mathematical manipulation, the insertion and reflection loss of the analogue RTA can be written as follows [40],

\[
\begin{align*}
    s_{21} &= s_{12} = s_{14c} + \frac{\Gamma_L}{(1 - s_{11c}\Gamma_L)^2 - (s_{14c}\Gamma_L)^2} \left[ 2s_{12c}s_{13c}(1 - s_{11c}\Gamma_L) + s_{14c}\Gamma_L(s_{12c}^2 + s_{13c}^2) \right] \\
    s_{11} &= s_{22} = s_{11c} + \frac{\Gamma_L}{(1 - s_{11c}\Gamma_L)^2 - (s_{14c}\Gamma_L)^2} \left[ (s_{12c}^2 + s_{13c}^2)(1 - s_{11c}\Gamma_L) + 2s_{12c}s_{13c}s_{14c}\Gamma_L \right]
\end{align*}
\]

(4.1)

(4.2)

where \( \Gamma_L \) is the reflection coefficient of the device and is voltage-dependent.

With an ideal 90° coupler, i.e. \( s_{11c} = s_{14c} = 0 \) and \( s_{12c} = js_{13c} = \frac{1}{\sqrt{2}} \), (4.2)
4.2. Improved Bi-phase Amplitude Modulator using Feedback

yields perfect matching \((s_{11} = s_{22} = 0)\), and the attenuation level from (4.1) only depends on the range of the reflection coefficient provided by the device which is given by:

\[ s_{21} = j\Gamma_L \]  

(4.3)

4.2.3 Use of Feedback to Improve the Attenuation Range

Figure 4.4: Feedback RTA

Figure 4.4 shows the proposed circuit with feedback, which improves the attenuation range and the modulator performance. The proposed technique trades the return loss with the insertion loss of the attenuator by shunting the standard attenuator with the admittance, \(Y_F\). If the reflection terminations are electrically identical and the 90° coupler is perfectly symmetrical, the insertion loss and reflection loss of the proposed circuit can be derived as follows:

\[
\overline{s_{11}} = \overline{s_{22}} = \frac{2s_{22} + Y_F[Z_o(s_{22}^2 + 2s_{21} - s_{21}^2 - 1)]}{2[Y_FZ_o(s_{22} - s_{21} + 1) + 1]} \]  

(4.4)

\[
\overline{s_{12}} = \overline{s_{21}} = \frac{2s_{21} + Y_F[Z_o(s_{22}^2 + 2s_{22} - s_{21}^2 + 1)]}{2[Y_FZ_o(s_{22} - s_{21} + 1) + 1]} \]  

(4.5)

where \(Z_o\) is the impedance of the source and load. Figure 4.5 shows the calculated performance of the feedback RTA based on the device NEC761084A with the various values of the resistive feedback admittance, \(Y_F\), for two different states, the miniumum \((@ V_{gs} = \text{pinched-off} = -V_p)\) and the maximum
attenuation state \( \Theta V_{gs} = -\frac{V_p}{2} \). It is noted that the maximum attenuation state is achieved when the device’s input impedance is close to 50 \( \Omega \).

From Figure 4.5, it is shown that the proposed technique can improve the maximum attenuation achievable at the expense of a small degradation in return loss. At the optimum point, a maximum attenuation state of \(-28\ dB\) is achieved (with the same minimum attenuation state insertion loss) when using a 225 \( \Omega \) shunting resistance value (or equal to 0.00444 \( \Omega^{-1} \) in Figure 4.5). The advantage of this technique over the balanced structure employed for attenuation range improvement is that a significant reduction in circuit complexity is obtained since the balanced structure needs four 90° hybrid couplers while the feedback technique needs only one coupler. The reduced number of couplers reduces the circuit area, meaning a lower manufacturing cost. The modulator performance degradation caused by non-perfect hybrid couplers is also reduced; for example phase and amplitude imbalance, inherent loss, etc.

Figure 4.5: The effect of shunting resistors on RTA performance.
4.2.4 Measured Results and Discussion

The standard analogue RTA and the improved circuit with feedback were designed to operate at 900 MHz and constructed on FR4 PCB (printed circuit board: substrate thickness $h=1.6$ mm, $\varepsilon_r = 4.55$). The Lange coupler was selected for the $90^\circ$ hybrid coupler due to its wide bandwidth. GaAs MESFET devices, the NE761084A, were selected for the reflection terminations.

![Image of the experimental bi-phase modulator](image)

Figure 4.6: The experimental bi-phase modulator

Figure 4.6 shows a photograph of the circuit with feedback. The gate bias resistor was chosen to be $600 \, \Omega$. This value is deviated from the calculated optimum value since there are some parasitic elements in the resistor package. The effect of the two $\lambda/8$ transmission lines in the feedback path is compensated for by applying a lumped surface mount capacitor. The gate-source control bias was varied from 0 to $-2$ V. All the measurements were performed on a network analyser test system, the HP8514A system. For the standard analogue RTA, the measurements show the minimum and maximum insertion loss being $-0.95$ dB and $-18$ dB at bias voltages around $-2$ V and $-1.1$ V, respectively. For the proposed circuit with feedback, minimum (ON state) and maximum (OFF state) insertion loss of $-2.14$ dB at $V_{gs} = -2.1$ and $-33.5$ dB at $V_{gs} = -1.1$, respectively, were obtained. Good input and output return loss for all the bias conditions was obtained from the proposed circuit from 600 MHz to 1 GHz.
The worst case matching is below $-10$ dB at the maximum attenuation, which is worse than the standard one by $8$ dB.

![Comparison of insertion loss between the feedback attenuator and the standard RTA](image)

Figure 4.7: Comparison of insertion loss between the feedback attenuator and the standard RTA

### Table 4.1: Summary of the feedback RTA performances @900 MHz

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$s_{11}$</td>
<td>$&lt;-10$ dB</td>
</tr>
<tr>
<td>$s_{22}$</td>
<td>$&lt;-10$ dB</td>
</tr>
<tr>
<td>attenuation range</td>
<td>$&gt; 31$ dB</td>
</tr>
<tr>
<td>loss@minimum attenuation state</td>
<td>$\approx 2.2$ dB</td>
</tr>
<tr>
<td>phase difference between ON-OFF</td>
<td>$179^\circ$</td>
</tr>
<tr>
<td>power consumption</td>
<td>$\approx 1$ mW</td>
</tr>
</tbody>
</table>

The comparison of the measured $s_{21}$ vs. control voltage characteristic of the two circuits is shown in Figure 4.7. The feedback technique does improve the amplitude symmetry and enlarge the $s_{21}$ dynamic range. It is also shown in Figure 4.8 that the phase error in the proposed system is reduced considerably.
Figure 4.8: Measured $s_{21}$ of the standard RTA and the proposed technique on a polar diagram.

These $s_{21}$ response prove that the proposed structure is more suitable for bi-phase modulator than the standard RTA structure. It should be noted that the power consumption of this technique is the same as the standard topology due to the negligible gate current. The proposed circuit consumes less than 0.5 mA DC current. The performance of the proposed attenuator is summarised in Table 4.1.
4.3 A Simple Technique for Compensation of FET Parasitics in VM based on the RTA Structure

Traditionally, a VM can be realised by two main approaches in microwave and millimeter wave applications. The first approach is based on two orthogonal bi-phase modulators combined with a 3-dB power combiner. Realizing a circuit with this technique has a 3-dB inherent loss penalty. The second approach is based on a variable attenuator and a 360° variable phase-shifter. Though the second approach is attractive in terms of having no inherent insertion loss, a high performance variable attenuator with a constant phase and a phase shifter with constant insertion loss are needed. These are notoriously difficult to design, and laborious amplitude and phase control schemes are often required for such a modulator. Hence, the first approach is more attractive in terms of realisation, especially in microwave and millimeter wave applications since the system is composed of a single circuit block, namely the bi-phase variable attenuator.

The analogue RTA using FETs was first proposed and applied to a VM by Devlin and Minnis [23]. A variable resistance reflection termination is used so that the attenuation level can be controlled. The MESFET device is preferred over a PIN diode because of its near-zero DC control power and widespread availability in foundry processes. However, it has a smaller dynamic range of resistance and larger parasitic elements. The balanced, or push-pull, configuration is one way to remove the amplitude and phase errors caused by the active device parasitics [4]. This structure employs a second attenuator operated in anti-phase, giving good constellation symmetry, but the chip area is more than doubled.
4.3. A Simple Technique for Compensation of FET Parasitics in VM based on the RTA Structure

In this Section, a simple technique for compensating for the FET parasitics is presented which achieves improved constellation accuracy without resorting to a balanced topology. First, the parasitic elements of the FET and their effects on the modulator are analysed. Then, the proposed simplified technique and analyses the improvement in constellation symmetry between zero bias and pinch-off. Based on the proposed technique, the experimental results on a VM demonstrated practically at 1.8 GHz will be described.

![Figure 4.9: Schematic diagram of analogue RTA](image)

4.3.1 Analysis of FET Parasitic Effects on the RTA

Figure 4.9 shows the configuration of the RTA. The cold-FETs are used as reflection terminations which theoretically provide an ideal variable resistance [2]. In practice, parasitic elements (junction capacitances, feed inductance, etc.) introduce phase and amplitude error, which restricts the application of the VM. Previously, the technique reported to overcome this problem is a balanced structure. Compared with the RTA in Figure 4.9, three additional hybrid couplers are required and an additional, complementary, baseband control, $I_t$, is required. Because circuit complexity increases a larger area for the circuit is needed. To implement a VM, the balanced topology requires nine hybrid couplers, one Wilkinson divider, eight active devices and four baseband signal channels. Thus, a large chip area is needed for MMIC realization which, in...
turn, results in higher manufacturing costs. Moreover, the larger number of signal controlling channels, which are $I$, $Q$, $\bar{I}$, and $\bar{Q}$, increases the complexity of the baseband circuitry. For example, a four-channel D-to-A converter is needed instead of the two-channel one needed for a conventional I-Q modulator.

Whilst the VM based on the balanced structure theoretically achieves only 3 dB insertion loss, in practice the insertion loss is increased considerably due to the combined losses of the nine hybrid couplers. The $s_{21}$ of this variable attenuator is directly related to $\Gamma_T$, while $s_{11}$ and $s_{22}$ are perfectly matched to $Z_0$. By assuming that the quadrature hybrid coupler is ideal, the effect of transistor parameters on the modulator performance is studied in this Section.

![Figure 4.10: A circuit diagram of an equivalent circuit of cold-FET](image)

The cold-FET and its equivalent small signal circuit is shown in Figure 4.2. The transistors are operated at $V_{ds} = 0$. The gate-drain and gate-source capacitances ($C_{gd}, C_{gs}$), which represent the variation in the depletion charge with respect to the applied voltage are equal because the depletion channel is symmetrical at this biasing point [27], i.e. $C_{gd} = C_{gs}$. Since the gate bias resistor $R_{ext}$ is quite large and $R_i$ is small, then the driving-point impedance at the drain of the model shown in Figure 4.2 is a three-element series network. There are two series R-L networks ($R_d \& L_d$ and $R_s \& L_s$) and one parallel
4.3. A Simple Technique for Compensation of FET Parasitics in VM based on the RTA Structure

RC network \((R_{ds} \& C_{ds}, C_{gs}, C_{gd})\). Because of this, the device model shown in Figure 4.2 can be reduced to the simplified equivalent circuit shown in Figure 4.10. Let

\[ C_z = C_{ds} + \frac{C_{gd}}{2} \]  

(4.6)

As shown in Figure 4.10, the input impedance \((Z_{in})\) of the cold-FET device, which can be visualised as the series resonant RLC circuit, is given by:-

\[ Z_{in} = R_{in} + jX_{in} \]  

(4.7)

\[ R_{in} = R_d + R_s + \frac{R_{ds}}{1 + (\omega R_{ds} \cdot C_s)^2} \]  

(4.8)

\[ X_{in} = L_{in} = L_d + L_s \]  

(4.9)

\[ C_{in} = \frac{1 + \omega^2 R_{ds}^2 C_s^2}{\omega^2 R_{ds}^2 C_s} \]  

(4.10)

where \(R_d\), \(R_s\) and \(R_{ds}\) are the drain, source and drain-source resistance, respectively. \(R_s\) and \(R_d\) represent the ohmic contacts and any bulk resistances leading up to the active channel. These two resistances are on the order of 1 \(\Omega\) [27]. The drain and source inductance \((L_d, L_s)\), primarily resulting from contact pads, are typically less than 10 \(\mu\)H[27]. The drain-source capacitance \((C_{ds})\) is related to the geometric capacitance effects between the source and drain electrodes.

From (4.8)-(4.10), the gate resistance \((R_g)\) and inductance \((L_g)\), result from the metallization of the gate Schottky contact and metal contact pad. These have no effect of \(Z_{in}\) since they are absorbed by a large gate biasing resistor which treats the transistor gate as an open circuit. Consequently, only the two model elements, \(C_{gd}\) and \(R_{ds}\), are voltage bias dependent cold-FET parameters.

From (4.8), the fixed parasitic resistances directly affect \(R_{in}\) and simply lead to an increase in insertion loss. The effect of parasitic inductance dominates as the applied bias is approaching zero since \(C_{in}\) becomes large. Since these parasitic inherent effects are small, these parameters can be assumed to be negligible. The equivalent circuit of the cold-FET will then be reduced to a
parallel circuit form \((R_{ds} \text{ and } C_x)\). The input impedance of this circuit (4.7) can then be written as,

\[
Z_{in} = \frac{1}{1 + \left(\omega R_{ds} C_x\right)^2} (R_{ds} - j\omega R_{ds}^2 C_x)
\]

Consequently \(\Gamma_T\) is given by,

\[
|\Gamma_T| = \sqrt\frac{(Z_o - R_{ds})^2 + (\omega R_{ds} C_x Z_o)^2}{(Z_o + R_{ds})^2 + (\omega R_{ds} C_x Z_o)^2}
\]

\[
\angle \Gamma_T = \tan^{-1} \frac{2\omega R_{ds}^2 C_x Z_o}{(Z_o^2 - R_{ds}^2) + (\omega R_{ds} C_x Z_o)^2}
\]

Figure 4.11 shows the effect of \(C_x\) on the characteristic of \(\Gamma_T\) for a FET with the parameters listed in Table 4.2. It is shown in the result that the existence of capacitance causes the phase trajectory of \(\Gamma_T\) at pinch-off voltage \(V_{TO}\) to deviate from the real-axis. Hence, 180° phase difference cannot be achieved. This \(\Gamma_T\) characteristic is not suitable for a bi-phase and VM implementation. The bias-dependent model parameter is now considered. Here, the Curtice
4.3. A Simple Technique for Compensation of FET Parasitics in VM based on the RTA Structure

FET model to analyse the cold-FET is used. It is shown that $R_{ds}$ increases when $V_{gs}$ is more negative since the channel is narrowed. Conversely, $C_{gd}$ is decreased when $V_{gs}$ is more negative since the charge in the depletion region is smaller.

Table 4.2: List of the Extracted Curtice Model Parameters of CFY30 FET

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\alpha$</td>
<td>0.285</td>
</tr>
<tr>
<td>$\beta$</td>
<td>0.1541</td>
</tr>
<tr>
<td>$V_{TO}$</td>
<td>$-1.9$</td>
</tr>
<tr>
<td>$C_{gds}$</td>
<td>1.891 pF</td>
</tr>
<tr>
<td>$C_{ds}$</td>
<td>0.1 pF</td>
</tr>
<tr>
<td>$V_{bi}$</td>
<td>0.7</td>
</tr>
</tbody>
</table>

4.3.2 A Proposed Technique to Improve the RTA

The technique presented here to compensate for the FET parasitics in the VM has two parts: Firstly, a series inductance is introduced which moves the left half of the $\Gamma_T$ trace to the upper half of the Smith chart, whilst only causing a small change in $\Gamma_T$ at the bias near to $V_{TO}$. By optimum choice of this added inductance, the phase difference can be set to 180°. Secondly, a shunt resistor is introduced to equalise the magnitudes of $\Gamma_T$ at $V_{gs} = 0$ and pinch-off.
The Effect of External Inductance on $|\Gamma_T|$ 

If an external series inductance, $L_w$, is deliberately applied, connected at either the drain or source, then $\Gamma_T$ is obtained as follows:

$$|\Gamma_T| = \sqrt{(Z_o - R_{ds})^2 + (\omega R_{ds} C_x Z_o)^2 + \Delta}$$

(4.14)

$$\angle \Gamma_T = \tan^{-1} \frac{2\omega Z_o [R_{ds}^2 C_x^2 - L_w \cdot (1 + \omega^2 R_{ds}^2 C_x^2)]}{(Z_o^2 - R_{ds}^2) + (\omega R_{ds} C_x Z_o)^2 - \Delta}$$

(4.15)

where $\Delta = \omega^2 [L_w^2 (1 + (\omega R_{ds} C_x))^2 - 2R_{ds}^2 C_x L_w]$.

The inductance $L_w$ is used to control $\Gamma_T$ at $V_{gs} = 0$, and causes only a small change of $\Gamma_T$ when the FET is biased near pinch-off. In this section, the sensitivity of $|\Gamma_T|$ with $L_w$ is investigated. If the sensitivity of $|\Gamma_T|$ is very small, this implies that the effect of $L_w$ on $|\Gamma_T|$ is also negligible. The sensitivity definition is applied to prove this for the bias voltage near pinch-off. Let $S_{\Gamma_T}^{[x]}$ be the sensitivity of a parameter $x$ to a function $f[x]$ which is defined as follows [22]:

$$S_{\Gamma_T}^{[x]} = \frac{\partial \ln f[x]}{\partial \ln x}$$

(4.16)

Hence, applying (4.16) to $|\Gamma_T|$ with respect to $L_w$ by using Maple™ software, one obtains:

$$S_{L_w}^{[\Gamma_T]} = \frac{4Z_o R_{ds} \omega^2 [L_w (1 + \omega^2 R_{ds}^2 C_x^2) - R_{ds}^2 C_x]}{((Z_o - R_{ds})^2 + (\omega R_{ds} C_x Z_o)^2 + \Delta) \cdot ((Z_o + R_{ds})^2 + (\omega R_{ds} C_x Z_o)^2 + \Delta)}$$

(4.17)

So for a bias point which is very close to pinch-off ($R_{ds} \to \infty$),

$$S_{L_w}^{[\Gamma_T]}(V_{gs} = V_{TQ}) \equiv \lim_{R_{ds} \to \infty} S_{L_w}^{[\Gamma_T]} = 0$$

(4.18)

Also at zero bias, $R_{ds}$ is very small ($R_{ds} \to 0$), hence

$$S_{L_w}^{[\Gamma_T]}(V_{gs} = 0) \equiv \lim_{R_{ds} \to 0} S_{L_w}^{[\Gamma_T]} = 0$$

(4.19)

This implies that adding $L_w$ to the termination circuit has no effect on $|\Gamma_T|$ at zero and pinch-off bias points.
4.3. A Simple Technique for Compensation of FET Parasitics in VM based on the RTA Structure

The Optimum $L_x$ for a Biphase Modulator

Applying $L_x$ causes a significant change in $\Gamma_T$ when $V_{gs}$ is near zero. The effect of $L_x$ can be predicted by considering (4.9) which makes the $|\Gamma_T|$ trace move upward to the upper plane of the Smith chart. We study this by applying different values of $L_x$ to the device (CFY30) (Table 4.2).

Figure 4.12: The effect of additional series inductance ($L_x$) on $\Gamma_T$

Figure 4.12 shows the $\Gamma_T$ obtained by applying values between $L_x = 0$ and $L_x = 3 \text{ nH}$. It is shown that the positions of $\Gamma_T$ at $V_{gs} \to 0$ depend on the value of $L_x$ and $|\Gamma_T|$ is also improved due to the reduction in $X_{in}$. The optimum value of $L_x$ is obtained by considering the phase difference of $\Gamma_T$ which should be $180^\circ$ between the two extreme biasing points, i.e. $V_{gs} = 0$ and near pinch-off voltage, that is:

$$\left(\angle \Gamma_T\right)_{V_{gs}=0} - \left(\angle \Gamma_T\right)_{V_{gs} \to V_T=0} = \pi$$  \hspace{1cm} (4.20)

At $V_{gs} = 0$, the value of parasitic resistance and drain-source resistance are very small compared with the reactance of the external inductance. Hence,
Chapter 4. Analogue Reflection-type Attenuator and Vector Modulator

one can simplify $\Gamma_T$ for this case as:

$$ (\Gamma_T)_{V_s=0} \approx \tan^{-1} \frac{-2\omega Z_o L_x}{Z_o^2 - (\omega L_x)^2} \quad (4.21) $$

As the bias is close to pinch-off, the reactance resulting from $L_x$ is very small when compared with $R_{ds}$ and $\omega C_x$, hence (4.14) becomes:

$$ (\Gamma_T)_{V_s\rightarrow V_{TO}} \approx \tan^{-1} \frac{2\omega Z_o C_x R_{ds}^2}{Z_o^2 - R_{ds}^2 + \omega R_{ds} C_x Z_o^2} \quad (4.22) $$

Substituting (4.22) and (4.21) into the condition defined in (4.20), the optimum value of $L_x$ which provides a $180^\circ$ phase difference of $\Gamma_T$ is obtained as follows:

$$ L_x = \frac{1}{2} \frac{Z_o^2 (1 + \omega R_{ds} C_x) - R_{ds}^2 + M}{(\omega R_{ds})^2 C_x} \quad (4.23) $$

where

$$ M = \sqrt{Z_o^2 (1 + \omega R_{ds} C_x)^2 + R_{ds}^4 (1 + 4\omega^2 Z_o^2 C_x^2) - 2\omega R_{ds}^3 C_x Z_o^2 - 2 R_{ds}^2 Z_o^2} \quad (4.24) $$

Since $R_{ds}$ at pinch-off is very large then one can approximate (4.23) as follows:

$$ L_{wa} \approx \lim_{R_{ds} \rightarrow \infty} L_x = \frac{1}{2} \frac{1 + 4\omega^2 Z_o^2 C_x (V_{TO})^2 - 1}{2C_x (V_{TO})^2 \omega^2} \quad (4.25) $$

Figure 4.12 shows that the best phase trajectory is obtained with $L = 1.4$ nH, which agrees with the optimum value of 1.33 nH obtained from (4.25). It is also evident from Figure 4.12 that the extra inductance provides a better dynamic range of attenuation, since $\Gamma_T$ now passes very close to the center point of the Smith chart.

Shunt Resistor to Correct Amplitude Asymmetry

The amplitude imbalance between $V_{gs} = 0$ and $V_{gs} = V_{TO}$ occurs due to the existence of small parasitic resistances at the drain and source which will dominate when the FET is zero biased. Normally, the $|\Gamma_T|$ at $V_{gs} = 0$ is
4.3. A Simple Technique for Compensation of FET Parasitics in VM based on the RTA Structure

smaller than $|\Gamma_T|$ near pinch-off. To correct this, a shunt resistor is applied at the drain to reduce the overall impedance when the FET is biased near to pinch-off. This decreases the $|\Gamma_T|$ at pinch-off to the $|\Gamma_T|$ value obtained when zero gate-source bias is applied. The full schematic diagram of the proposed improved technique for VM is shown in Figure 4.13. This modified RTA gives improved performance and can be used as a bi-phase amplitude modulator and in a VM, without resorting to the large balanced topology.

Figure 4.13: The proposed VM

Applying this structure to a complete VM, the full 360° phase rotation is achieved with very few additional components. It should be noted that the nonlinear phase and amplitude tuning characteristic can be corrected. One possible approach to correct this problem is by applying a predistortion function, which is straightforward to implement in software or by applying a predistortion circuitry.
4.3.3 Measured Performance

The implementation is performed with a packaged CFY30 device which has $f_{\text{max}} = 12$ GHz. All the circuits are realised on FR4 PCB using microstrip transmission lines. The substrate is epoxy glass ($h = 1.6\ mm, \epsilon_r = 4.55$, FR-4). All hybrid couplers are realised using two 8.4 dB couplers connected in a tandem fashion [44] to obtain a 3-dB coupler. A complete VM based on the proposed technique is constructed. Figure 4.14 shows the photograph of the VM. All the lumped elements on the PCB are surface mount (type 0805). The small external inductors are realised by a microstrip line and taking the via hole connecting line into account ($1\ mm \approx 1\ nH$). The parasitic inductance from via holes is minimized by using parallel via holes.

The measurement is performed with an HP8510C network analyser test system. The $s_{11}$ and $s_{22}$ of the VM are well below $-15$ dB at 1.8 GHz. Figure 4.15 shows the measurement results of the VM using the proposed technique with the bias voltages (I and Q) varying from 0 to $V_p$. The dynamic range of this VM is from $-5.2$ to $-60$ dB and a full 360° phase rotation is achieved for this VM. The amplitude and phase imbalance at 1.8 GHz are well below 0.4 dB and 3°. The amplitude and relative phase frequency response of the proposed VM from eight different bias points, listed in Table 4.3, are shown in Figure 4.16 and 4.17, respectively. The performance of the proposed VM is concluded in Table 4.4.
4.3. A Simple Technique for Compensation of FET Parasitics in VM based on the RTA Structure

Figure 4.15: Measured $s_{21}$ constellation of VM using the proposed technique

Figure 4.16: The amplitude response of the proposed VM
Figure 4.17: The relative phase response of the proposed VM

Table 4.3: Lists of 8 testing biasing points for measuring the frequency response of the proposed VM

<table>
<thead>
<tr>
<th>Bias point</th>
<th>I (Volt)</th>
<th>Q (Volt)</th>
</tr>
</thead>
<tbody>
<tr>
<td>A</td>
<td>-2</td>
<td>0</td>
</tr>
<tr>
<td>B</td>
<td>-1.8</td>
<td>-0.2</td>
</tr>
<tr>
<td>C</td>
<td>-1.6</td>
<td>-0.4</td>
</tr>
<tr>
<td>D</td>
<td>-1.5</td>
<td>-0.6</td>
</tr>
<tr>
<td>E</td>
<td>-1.7</td>
<td>-0.8</td>
</tr>
<tr>
<td>F</td>
<td>-1.4</td>
<td>-1</td>
</tr>
<tr>
<td>G</td>
<td>-1.2</td>
<td>-1.1</td>
</tr>
<tr>
<td>H</td>
<td>-1.1</td>
<td>-1.1</td>
</tr>
</tbody>
</table>
4.3. A Simple Technique for Compensation of FET Parasitics in VM based on the RTA Structure

Table 4.4: The performance of the proposed VM @ 1.8 GHz

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>$s_{11}$</td>
<td>$&lt;-15$ dB</td>
</tr>
<tr>
<td>$s_{22}$</td>
<td>$&lt;-15$dB</td>
</tr>
<tr>
<td>minimum $s_{21}$</td>
<td>-60 dB</td>
</tr>
<tr>
<td>maximum $s_{21}$</td>
<td>-5.4 dB</td>
</tr>
<tr>
<td>amplitude imbalance</td>
<td>0.4 dB</td>
</tr>
<tr>
<td>phase imbalance</td>
<td>$3^\circ$</td>
</tr>
<tr>
<td>power consumption</td>
<td>$&lt;1$ mW</td>
</tr>
<tr>
<td>size reduction compared to the balanced VM</td>
<td>$&gt;50%$</td>
</tr>
</tbody>
</table>
4.4 Analysis of IM₃ in RTAs

The distortion analysis of the FET-based RTA was presented in [8]-[9] using a power series approximation for single-tone distortion. In this section, a similar technique is applied to the RTA for the two-tone IMD case. The nonlinear mechanisms which contribute to nonlinear distortion in the RTA, implemented with GaAs FETs, are also investigated. The assumption that the nonlinear distortion is mainly due to the nonlinearity in the \( R_{ds} \) of the devices is exploited. The approach allows to estimate the degree of nonlinearity and provides criteria to help select the best devices for RTA design. The results are in good agreement with the experimental results at L-band.

4.4.1 Nonlinear Distortion Analysis of the RTA

Analysis model of Cold-FET Transistor in RTA

The RTA model for analysing the nonlinearity distortion is shown in Figure 4.18. The FET model with \( V_{ds} = 0 \) (the cold-FET) can be reduced to an RLC series circuit where the values of \( R \), \( L \), and \( C \) are obtained from (4.8)-(4.10). The inductances can be assumed to be negligible to simplify the
4.4. Analysis of IM<sub>3</sub> in RTAs

Analysis. From the model, only two bias-dependent components, \( R_{ds} \) and \( C_{gs} \), are taken into account. From the cubic Curtice model, the drain current source \( (I_{ds}) \) is modelled as:

\[
I_{ds} = \beta (V_{gs} - V_{TO})^2 (1 + \lambda V_{ds}) \tanh(\alpha V_{ds}) \tag{4.26}
\]

Since \( g_{ds} = \frac{dI_{ds}}{dV_{ds}} \) is the output conductance and \( C_{ds} \) is the drain-source capacitance, hence:

\[
g_{ds} = \frac{\alpha \beta}{\cosh(\alpha V_{ds})^2} (V_{gs} - V_{TO})^2 (1 + \lambda V_{ds}) + \beta \lambda (V_{gs} - V_{TO})^2 \tanh(\alpha V_{ds}) \tag{4.27}
\]

\[
C_{gs, gd} = C_{gs0, gd0} \left( 1 - \frac{V_{gs}}{V_{bi}} \right)^{0.5} \tag{4.28}
\]

In the \( g_{ds} \) function, which represents the DC-characteristic of device, \( V_{TO} \) is the threshold voltage, \( \alpha \) determines the voltage where the drain current saturates, \( \lambda \) is a parameter related to drain conductance, and \( \beta \) is a scaling parameter [42], [21]. The RF nonlinear parameter characteristic, \( C_{gs, gd} \), determined by \( C_{gs0, gd0} \) represents the depletion capacitance at zero gate bias. \( V_{gsi} \) is the gate-source voltage dropped at a junction, and \( V_{bi} \) is the built-in potential. It is shown in (4.27) that \( g_{ds} \) is both drain and gate bias dependent while \( C_{gs} \) and \( C_{gd} \) are only gate-source bias dependent since the variation of capacitances is small with the sufficiently small input signal at the drain-source. Considering the RTA structure in Figure 4.18, the RF signal is applied at the input coupler, and the drain voltage of the devices varies according to the input signal. Consequently, the nonlinear distortion of the RTA mainly results from the \( R_{ds} \) since it is the sole drain-bias dependent element. To simplify the problem, that the effect of AM-PM is small for weakly nonlinear operation is assumed, and then the device model of FET is composed of \( R_{ds} \) if \( \omega R_{ds} C_{gs} << 1 \). The analysis model shown in Figure 4.18 is composed of a linear circuit, which is a 4 port network representing the 90° hybrid coupler, two nonlinear resistors which are \( R_{ds} \) in series with a reactance term resulted from the inductance and capacitance in the transistors.
RTA Analysis Details

We start the analysis details with the ideal 90° hybrid coupler, for which the Y-matrix can be written as follows:

\[
Y = jY_o \begin{bmatrix}
0 & \sqrt{2} & 0 & 1 \\
\sqrt{2} & 0 & 1 & 0 \\
0 & 1 & 0 & \sqrt{2} \\
1 & 0 & \sqrt{2} & 0 \\
\end{bmatrix}
\]  

(4.29)

where \(Y_o\) is the coupler characteristic admittance. Applying Kirchoff Current’s Law at the coupled (node 3) and the direct ports (node 2), one obtains,

\[
i_2 = -Z_T (v_{ds}, v_{gs}) v_2
\]  

(4.30)

\[
i_3 = -Z_T (v_{ds}, v_{gs}) v_3
\]  

(4.31)

where \(Z_T\) is the reflection termination impedance. At the isolated port (node 4), a load impedance \((Z_L)\) is connected to the coupler, hence

\[
v_4 = -i_4 Z_L \equiv -i_4 Z_o
\]  

(4.32)

At node 1 one obtains the input current \(i_1\),

\[
i_1 = \frac{v_s - v_1}{Z_s} \equiv \frac{v_s - v_1}{Z_o}
\]  

(4.33)

For an ideal 90° hybrid coupler, one can write the relationship of \(v_2\) and \(v_3\) as follows:

\[
v_2 = jv_3
\]  

(4.34)

From these equations, one can write \(v_4\) in terms of \(v_s\) and \(v_2\) as follows:

\[
v_4 = -\frac{v_s}{2} - \sqrt{2}v_2
\]  

(4.35)

So if one can find \(v_2\), the output signal from the RTA will be obtained. It should be noted from (4.35) that the IM3 distortion at the output node is
3-dB more than that at the transistor drain output. From (4.29) to (4.34), one obtains:-
\[
\frac{v_s}{\sqrt{2}} = -[Z_T(v_3, v_{gs}) Z_0 + 1] v_3
\]
(4.36)
where \(Z_T\) in (4.36) is a function of \(v_3\), which is equal to \(V_{ds}\). Substituting the \(R_{ds}\) function and the transistor model components in (4.36) and solving (4.35) gives us the results of carrier output and the distortion for the RTA.

### 4.4.2 Analysis Details for a Practical Example

In this Section, the results obtained from (4.36) by applying the output conductance and gate source capacitance to \(Z_T\) are exploited. The transistor used in this study is the CFY30 GaAs FET from Siemens which was selected for the design of a 1.8 GHz variable attenuator for use in an adaptive FF amplifier. The Curtice model is extracted from the library model being available in Libra\(^{TM}\). The Curtice extracted model parameters for the CFY30 are listed in Table 4.2. The condition \(\omega R_{ds} C_0 \ll 1\) is satisfied except at the bias voltages near \(V_{TO}\). The value of parasitic elements are small, hence one can assume that the termination impedance, \(Z_T(v_{ds}, v_{gs})\), is approximately \(g_{ds}(v_{ds}, v_{gs})\).

We substitute (4.32) in (4.36), the results are obtained as follows:-
\[
v_s = -\sqrt{2} v_3 \left[ \frac{m Z_0}{\text{cosh}(\alpha v_3)} \right] \left[ \frac{1}{2} \sinh \alpha v_3 \cosh \alpha v_3 \right] + 1 \]
(4.37)

where \(m = \beta (v_{gs} - v_{TO})^2\). It is shown in (4.37) that there is no closed-form solution for \(v_3\). In such a case, one can approximate the function appearing in the right hand side of (4.37) by using a power series, that is \(v_s = a_1 v_3 + a_2 v_3^2 + a_3 v_3^3 + \cdots\). Simulation has shown that third-order approximation provides a good approximated result to the right term in (4.37). The approximated function can be written as:-
\[
v_s = -\sqrt{2} \left[ (m Z_0 \alpha + 1) v_3 + m Z_0 (\alpha v_3)^3 \right]
\]
(4.38)
Chapter 4. Analogue Reflection-type Attenuator and Vector Modulator

The second order term is neglected since the nonlinear distortion from the VM is produced from the odd order terms. Solving (4.38) for $v_3$ in a function of $v_s$, one obtains:

$$v_3 \approx a_1 v_s + a_3 v_s^3$$  \hspace{1cm} (4.39)

where $a_1$ and $a_3$ can be determined by:

$$a_1 = \frac{1}{\sqrt{2}(mZ_o\alpha + 1)}$$  \hspace{1cm} (4.40)

$$a_3 = \frac{3\sqrt{6} (mZ_o)^2 \alpha^2 \left(9mZ_o\alpha \lambda + 4\sqrt{\frac{2(1+mZ_o\alpha)^2}{mZ_o}}\right)}{64 (mZ_o \alpha + 1)^3 (\alpha m^2 Z_o^2) \sqrt{\frac{\alpha(mZ_o\alpha + 1)^3}{mZ_o}}^{1/2}}$$  \hspace{1cm} (4.41)

Substituting (4.39) into (4.36), the result is:

$$v_o = -j\frac{v_s}{2} - j\sqrt{2}(a_1 v_s + a_3 v_s^3)$$  \hspace{1cm} (4.42)

By applying (4.39) and (4.42), one can solve for the IM$_3$ and third-order intercept point (TOI) of the RTA. It should be noted that the detailed analysis can be extended to higher frequencies by modifying the relationships of drain voltage and current with a frequency-dependent function.

Considering the effect of DC parameters contributing to the nonlinear term (4.41), it is evident that $a_3$ is increased when $\lambda$ increases. This is because of an increased swing of output conductance which, in turn, produces more distortion. If $\alpha$ is increased, this means that the linear dynamic range of the FET transistor will be limited, hence the nonlinear distortion is increased.

### 4.4.3 A Comparison of Analysis and Experimental Results

To confirm the validity of the analysis presented here, an RTA based on CFY30 GaAs FET discrete devices was constructed and tested at L-band 1.8 GHz. The circuit was designed based on microstrip transmission lines. A tandem
coupler was used to obtain 3-dB quadrature planar hybrid coupler. To compare with the experimental results, the obtained extracted parameters from Table 4.2 were first inserted back to the Curtice FET model for comparison with the manufacturer-supplied library model in order to confirm the validity of the extracted parameters.

A two-tone measurement set up was used with signals at 1.8 and 1.801 GHz. The voltage control on the attenuator was swept from $-1.9$ to $-0.8$ V for minimum and maximum attenuation setting, respectively. The measurement setup was automatically controlled with HP-VEE$^{TM}$ software. The measurement data were retrieved over the GPIB bus (General Purposed Interface Bus) and the TOI was calculated by using MATLAB$^{TM}$.

![Graph](image)

Figure 4.19: Comparison of analysis and measured results at minimum attenuation level.

Figure 4.19 and 4.20 show the comparison of analysis and experimental results for the fundamental and IM$_3$ power at minimum and maximum attenuation level. It can be seen that the analysis results provide a good agreement with the experimental results. The IM$_3$ at a high attenuation setting is significantly worse than at the low attenuation levels.
Figure 4.20: Comparison of analysis and measured results at maximum attenuation level.

Figure 4.21 shows the TOI (referred to the input) obtained from the experimental results across the 5dB to 15.5 dB attenuation range. Both analysis and measurement results show that the TOI of the attenuator tends to be lower at a high attenuation level. This is because at a high attenuation level the device impedance is very close to 50 Ω, which maximises the power transfer to the devices, and because at this point the attenuation changes most rapidly with gate bias. In addition, since the TOI is calculated from the ratio of the output $IM_3$ and carrier power, the combination of lowest fundamental output signal and highest output $IM_3$ causes a significantly reduced TOI.
4.4. Analysis of IM$_3$ in RTAs

Figure 4.21: Comparison of analysis and measured results
Chapter 4. Analogue Reflection-type Attenuator and Vector Modulator

4.5 Conclusions

The analysis of the RTA modulator presented in this chapter confirms that the feedback technique can improve the attenuation range and phase error. The technique is simple to implement and design. The complexity increase is small and the circuit area is the same as the standard RTA. The measured result of a 900 MHz prototype circuit showed that a considerably improved performance was achieved. Though the experiment showed that the feedback RTA can be used as a bi-phase modulator, the technique involves a sacrifice of the input and output return loss. Consequently, another technique to eliminate the phase distortion for modulators was proposed.

Prior to developing this technique to cancel the phase distortion, a deep insight of the RTA's basic cell (the cold FET device) was needed. The analysis of the effect of the cold-FET parasitic elements on the bi-phase modulator and VM has been described. It was shown that the junction capacitances cause the $\Gamma_T$ to deviate from the real axis of the Smith chart. A new simple technique to correct for phase distortion and extend the dynamic range of attenuation has been developed. The asymmetry of modulator insertion loss between the two bias extremes, $V_{gs} = 0$ and near pinch-off, is also corrected by simply adding a shunt resistor. Simulation and experimental results demonstrate the effectiveness of the proposed technique at L-band.

The analysis of nonlinear distortion in the RTA has been presented and experimentally verified at L-band. Closed-form expressions for nonlinear distortion have been obtained which can be used to predict the effect of FET model parameters on the severity of distortion. The technique can be extended to high frequency applications and to other circuits which use the RTA as a basic building block, such as the bi-phase modulator, phase-shifter and VM.
Chapter 5

Transmitter Linearisation Techniques

5.1 Introduction

Future digital radio communications systems require bandwidth-efficient modulation schemes. As stated in the previous chapters, these modulation schemes result in a non-constant envelope, and so whilst higher data-rates can be achieved, the performance will be considerably degraded by any nonlinear component in the system. Many linearisation techniques can be employed to address this problem. In this chapter, three linearisation techniques for a direct-carrier modulation transmitter have been proposed and developed. These are:

1) A FF technique applied to a direct-carrier transmitter. The concept was developed from the modified FF mixer in Chapter 3. The technique has extended to the whole transmitter rather than being used for only a HPA or modulator. In the FF direct-carrier transmitter, nonlinearity in both the modulator and
HPA are corrected by this technique. The linearity of this technique has been proven in simulation and experiment.

2) A low-cost direct-carrier transmitter using software radio techniques (SR). In this technique, the transmitter is composed of a VM, a medium-power oscillator and a baseband signal processing unit. Two circuit blocks which result in a main implementation cost are eliminated. They are the HPA and a channel-selecting filter. In addition, using the flexibility of the baseband processing unit, a number of different transmitter functions can be implemented. The technique has been implemented at V-band.

3) An adaptive baseband-predistorted LINC transmitter. This technique is developed for curing the amplitude and phase imbalances in LINC transmitter. An adaptive predistorter implemented in baseband has been proposed for this structure. An LMS (least-mean square) algorithm was adopted for automatically adjusting the amplitude and phase imbalances, due to its convergence robustness. Simulation in MATLAB™ shows an improvement of spectral regrowth in the output of the LINC transmitter.
5.2 FF Application to a Direct-Carrier Transmitter

The FF topology for linearising microwave amplifiers was first proposed by Seidel [54]. Fundamentally, the FF structure is composed of two loops, one is producing the estimated distortion and the other is canceling the distortion. Since the distortion signal is small when comparing with the main signal, then a second high-gain, low power, amplifier is needed.

In a direct carrier modulation transmitter the distortion from the modulator must be considered as well as the distortion from the power amplifier. A FF amplifier can correct for distortion from the main PA only. In order to cancel the distortion from both the modulator and the PA, a signal from the auxiliary modulator operating at a sufficiently low input power is used as a reference signal. Section 5.2.1 presents the details of the technique. A system analysis shows the required attenuation and gain from the first and the second loop, respectively. Section 5.2.2 discusses the MATLAB\textsuperscript{TM} simulation and experimental results from the proposed system.

5.2.1 FF Technique for Transmitter

Figure 5.1 shows the FF linearised direct-conversion transmitter. A baseband signal, which is band-limited by applying a pulse-shaping filter, is upconverted directly to RF and then amplified by a PA before transmission. Distortion is produced from both the modulator and power amplifier. As the modulator is made part of the linearisation system, its distortion can be reduced along with that of the PA. The distortion from the modulator is of concern, for example, in microwave systems, where a modulator can be operated as a switch which is controlled by a baseband signal or LO. A reference signal for the modified FF
transmitter is obtained by operating the auxiliary modulator at a low LO or baseband input level. In the case of baseband control of the second modulator for mm-wave applications, this can be implemented in software.

![Figure 5.1: FF technique applied to a direct carrier modulation transmitter](image)

The proposed FF transmitter, which takes account of distortion from the modulator, is shown in Figure 5.1. The system is composed of two loops: The first loop's function is to produce a distortion signal. This can be achieved by subtracting the output of the PA with a reference signal obtained from the auxiliary mixer output. The second loop's structure is similar to the second loop of the FF amplifier. Two variable phase shifters and two variable attenuators are included in the experimental FF transmitter in order to optimise the gain and phase balance adjustment so that all distorted signals from both modulator and PA will be considerably reduced.

Figure 5.2 shows a ring-diode mixer which is used as a balanced modulator. The four diodes operate as baseband switches which will be controlled by the pumped LO signal. This ring-diode mixer acts as a polarity-switching on the RF signal ($v_{RF}$) in response to the LO input ($v_{LO}$). Hence the output signal from this modulator can be written as an odd-order power series, i.e.

$$v_{m1} = \sum_{i=1,3,\ldots} a_i v_{RF}^i v_{LO} \quad (5.1)$$
5.2. FF Application to a Direct-Carrier Transmitter

Figure 5.2: Schematic diagram of a ring-diode mixer

where $a$ is related to the balun, load impedance, and the device. It is clearly shown that the output transfer voltage function has odd-symmetry; thus all the even harmonics of LO can be discarded.

From Figure 5.1, the RF input signal to the auxiliary modulator is made to be sufficiently small by applying the attenuator $\mu$, then the output signal from the modulator will be a very clean signal which can be used as a reference signal for the FF system. This can be shown by applying $v_{RF}$ in (5.1) with $\frac{v_{RF}}{\mu}$, hence:

$$v_{m2} = \left( a_1 \frac{v_{RF}}{\mu} + a_3 \left[ \frac{v_{RF}}{\mu} \right]^3 + \ldots \right) v_{LO} \approx a_1 \frac{v_{RF}}{\mu} v_{LO} \quad (5.2)$$

This output signal from the second modulator acts as a reference signal to create the error signal in the first loop. Similar to conventional FF PA, the distortion reduction degree of the FF transmitter depends on the quality of the reference signal.

The output signal from the first modulator is applied to the main PA. With the assumption of neglecting the memory effects (PA frequency-dependent characteristic), the input-output voltage-transfer function of the PA can be modelled as a power series which can be written as,

$$v_{out} = \sum_{j=1}^{\infty} a_j v^j \quad (5.3)$$
where \( v \) is the PA input signal, \( a_1 \) is PA voltage gain and \( a_j \) is the \( j^{th} \)-order nonlinearity. Applying the modulating signal from (5.1) into (5.3), collecting only the significant terms from the obtained amplifier output signal, the attenuation value, \( A \), to provide an estimated distortion from the first loop can be obtained as follows,

\[
A \approx \frac{a_1}{\mu}
\]  

(5.4)

Hence, a clean signal from the system can be obtained by selecting an error amplifier whose gain is a reciprocal of the attenuator, i.e. \( G = 1/A \). It should be noted that the conversion loss of the modulator has no effect on defining the gain in the error amplifier. This is also an additional advantage of the proposed system over the transmitter which employed a FF amplifier.

Figure 5.3: Two-tone simulation of the FF transmitter
5.2. FF Application to a Direct-Carrier Transmitter

5.2.2 Results and Discussions

Simulation Results

The voltage transfer function of the balanced modulator is obtained by employing a curve-fitting technique with a power series model. The obtained model provides a good agreement with the measurement data. We employed this modulator model to simulate the proposed system. Figure 5.3 shows the two-tone measurement results of the FF structure employing this modulator model. The effect of imbalance in the balun is also included to investigate the capability for LO feedthrough reduction. The amplifier power gain is selected at 13 dB, the attenuator at the baseband is selected at 20 dB, hence, from (5.4), the estimated attenuator level of the first loop and the error PA gain is 33 dB. The results obtained from simulation suggest an effective capability for reducing nonlinear distortion from both the modulator and PA. As shown in Figure 5.4, the quality of the output signal from the proposed system theoretically depends on the quality of the reference signal. It should be noted
that the effect of LO feedthrough is also reduced since this system considers the LO feedthrough effect to be a distortion signal.

![Figure 5.5: Measured BPSK PSD of FF transmitter](image)

### Experimental Results

An experimental 900 MHz FF direct-carrier modulator transmitter has been constructed using, primarily, modular components to verify the concept. Two analogue reflection-type phase-shifters [40] were designed and constructed. A continuous 160° phase variation from both phase-shifters was achieved. In this experiment, the input signal level of the second modulator is 6 dB lower than the first modulator, and due to the cubic-law, this causes approximately 18 dB reduction in intermodulation distortion. A 10-W 20-dB attenuator is used at the PA output. All spectra are measured on an HP8563E. The first modulator and the main PA are operated in the saturation region. A 10 dB reduction of LO feedthrough was achieved for an unmodulated case. Two modulation schemes, 20 kHz Binary Phase-Shift-Keying (BPSK) with Nyquist shaping filter(\(\alpha = 0.5\)) and 1 Msps 16-QAM with a square-root raised-cosine filter(\(\alpha = 0.5\)) are used in this experiment. Figure 5.5 shows the measured
output spectrum of unlinearised and linearised BPSK signals. 30-dB spectral distortion reduction is achieved. Figure 5.6 shows 15-dB spectral distortion reduction of the linearised 16-QAM system. The capability for reducing distortion of the system is related to the gain and phase adjustment of both loops, but it was not overly sensitive and the FF system could be made adaptive using a feedback control loop.

Figure 5.6: Comparison a measured output raised-cosine 16-QAM power spectrum from the proposed system and the main PA (20-dB attenuator is inserted at spectrum analyser input)
5.3 Direct-carrier Transmitter with Software Radio (SR) Technique

As communication technology continues its rapid transition from analogue to digital, more functions of contemporary radio systems are implemented in software. A software radio (SR) is a radio whose channel modulation waveforms are defined in software (www.ourworld.compuserve.com/homepages/jimitola). Employing SR techniques provides increased flexibility and multi-functionality in the transceiver. In this section, the combination of the proposed transmitter structure and the SR concept provides low complexity but multi-functionality mm-wave transmitter, which is suitable for short-range up to medium-range communication.

5.3.1 Fixed Frequency High-power Gunn Source Feeding a Vector Modulator

![Diagram](image)

Figure 5.7: A fixed-frequency high-power Gunn source feeding VM
5.3. Direct-carrier Transmitter with Software Radio(SR) Technique

Figure 5.7 presents a structure for a low-cost direct-carrier mm-wave transmitter proposed in this Section. The Gunn oscillator is connected to a multilevel balanced VM. Since a Gunn oscillator can generate a moderate power signal, the PA part which occupies a large chip size is not needed in many short range applications. Four baseband signals from an A/D card, produced by SR are fed into the modulator chip.

5.3.2 Spectral Shaping Filter and Digital PD

A balanced I-Q modulator employing analog RTAs operating at 60 GHz was used. However, such a modulator cannot be directly applied for a direct-carrier modulation transmitter due to the effect of LO feedthrough, and due to amplitude and phase distortion in the modulator. One technique to overcome these problems is to employ a balanced topology using another attenuator modulator which operates at a certain offset from the first modulator bias; called a complementary bias. This technique can alleviate the LO feedthrough and phase distortion but significant nonlinearity exists in the amplitude vs. control voltage characteristic. This effect leads to amplitude distortion, but this can be corrected at the baseband signal level. The digital PD technique implemented in signal processing is applied due to its simplicity. This approach provides a good performance for moderate power applications with the same power efficiency of the transmitter.

In the individual RTAs, each MESFET device is biased at $V_{ds} = 0$ while the input baseband is applied to the gate-source. This bias provides a large dynamic range input for the modulator but the transfer characteristic is not linear. The $V_{gs}$ signal will change the junction resistance and capacitance which in turn cause the driving-point impedance at the drain to vary from nearly zero to a certain large value of ohms. The real and imaginary part of this
impedance mainly results from the $R_{ds}$ and junction capacitances, respectively. For simplicity, the imaginary part of the impedance is neglected. An empirical equation of the drain-source resistance of the cold-FET is of the following form [4],

$$r_{ds} = \alpha_1 \tanh(\alpha_2 + \alpha_3 v_{gs})$$ \hfill (5.5)

where $\alpha_1$, $\alpha_2$ and $\alpha_3$ can be determined by curve fitting. Assuming that the hybrid coupler employed in the analog-reflection modulator type is ideal, then the insertion loss, $s_{21}$, of the modulator can be written as

$$s_{21} = s_{11x} = 1 + \frac{2\alpha_1}{(Z_o - \alpha_1)e^{2(\alpha_2 + \alpha_3 v_{gs})}}$$ \hfill (5.6)

$$s_{21} = -s_{11x} = 1 + \frac{2\alpha_1}{(Z_o - \alpha_1)e^{2(\alpha_2 + \alpha_3 (v_{gs} + v_{offset}))}}$$ \hfill (5.7)

where an overbar stands for the complementary modulator signal, $s_{11x}$ for a return loss of a cold-MESFET at the drain and $Z_o$ for a characteristic impedance. Since the modulator operates like a switch, the transfer function of the 1-cell modulator is an S-shaped function with the input baseband signal, $v_{gs}$. One may obtain a linear modulator by predistorting the baseband signal using the reciprocal function of the S-shaped transfer function with a targeted function, $v_{\text{linear}} = a + bv_{gs}$, hence,

$$s_{prD} = \frac{a + bv_{gs}}{s_{21}}$$ \hfill (5.8)

$$s_{prD} = \frac{a + b(v_{gs} + v_{offset})}{s_{21}}$$ \hfill (5.9)

where $v_{\text{offset}}$ is a DC offset bias for a complementary signal of balanced modulator. Cascading the PD block after spectral shaping prefiltering, one will obtain a signal with significant distortion reduction. It should be noted that the PD block here only corrects the problem of amplitude distortion since the balanced VM has an excellent phase distortion performance.
5.3.3 Channel Control Using a Direct Baseband Serrodyne Technique

Channel control can be achieved by tuning the LO signal or by controlling the baseband. Since the aim is to realise a low cost transmitter architecture, which requires a minimum of mm-wave components, the serrodyne modulator technique based on baseband implementation is selected. Originally, the serrodyne method uses a 360° phase-shifter injected with a sawtooth control signal, producing phase which increases linearly with time, which is equal to a frequency shift [37], [39]. The fundamental frequency of the sawtooth signal is related to the shift in frequency from the input carrier frequency, as follows,

\[ v_{serro} = V \sin(\omega_c t + \pi S[t]) \]  

where \( S[t] \) is the sawtooth function with unit amplitude. A greatly simplified system can be achieved by applying the serrodyne technique at baseband. The new approach for translating a linearised modulated signal using the serrodyne concept is proposed as follows. The I-Q modulation signal for driving a balanced I-Q modulator is given by the following form,

\[ v_o = (I - \bar{I}) \sin \omega_c t + (Q - \bar{Q}) \cos \omega_c t \]  

If the transmitter signal is shifted by \( \omega_s \) but the information baseband is unchanged, in this case the output can be written as

\[ v_{oshift} = (I - \bar{I}) \sin [\omega_c + \omega_s] t + (Q - \bar{Q}) \cos [\omega_c + \omega_s] t \]  

Using the trigonometric expansion of (5.12) and collecting the coefficient terms of \( \sin(\omega_c t) \) and \( \cos(\omega_c t) \) which represent the \( I_s \) and \( Q_s \) baseband signals for controlling the channel at spacing \( \omega_s \) from a carrier signal, respectively, then:

\[ I_s = I \cos \omega_s t + Q \sin \omega_s t \]  

(5.13)
Chapter 5. Transmitter Linearisation Techniques

\[ Q_s = Q \cos \omega_s t - I \sin \omega_s t \]  \hspace{1cm} (5.14)

\[ I_s = I \cos \omega_s t + Q \sin \omega_s t \]  \hspace{1cm} (5.15)

\[ Q_s = Q \cos \omega_s t - I \sin \omega_s t \]  \hspace{1cm} (5.16)

where subscript \( s \) stands for the serrodyne. It should be noted that the complementary baseband channels \( I_s \) and \( Q_s \) can also be obtained from (5.13)-(5.14) by adding the DC offset value such that the modulator is operated in a balanced mode. Applying (5.13)-(5.16), a new channel selection technique implemented at a baseband level is achieved.

\[ \text{Figure 5.8: Test bench setup for measuring the output spectrum} \]

5.3.4 Measured Results and Discussions

Figure 5.8 shows the test bench setup for the 60 GHz experiment on the proposed transmitter. A Gunn oscillator from Farran Technology was biased to provide a fixed 100 mW at 60 GHz. An MMIC VM chip, designed on the H40 process by colleague [4], was used in this experiment and setup on a Cascade Microtech Summit 9000 analytical probe-station. The output signal was connected to a HP8563E spectrum analyser with a harmonic mixer, and to ensure the harmonic mixer was linear a fixed 40-dB attenuator was used throughout.
the experiment. The four baseband signals were fed from an arbitrary waveform generator, with data files generated in MATLAB and downloaded from a PC. Figure 5.9 shows the output signal from the system with a 1.5 Mbps BPSK signal with no prefiltering. 20 mW output power is achieved with this transmitter. Figure 5.10 and 5.11 show the measured result for a QPSK signal from the modulator, with and without predistortion. The 20 kbps Pseudonoise sequence (PN_{23}) baseband signal is obtained from an A/D card and prefiltered by using a square-root raised-cosined filter with a shaping factor, $\alpha = 0.5$. It is clearly shown that the spectral regrowth obtained from the predistorted system is 12 dB less than unpredistorted signal.

Figure 5.12 shows the output signal at two different channel settings, whilst simultaneously applying the spectral shaping filter and digital PD. The second channel setting is produced by introducing (5.13)-(5.16). 20 dB sideband rejection is achieved without using any external components. However the speed of the A/D part limits the communication bit-rate for this technique.
Chapter 5. Transmitter Linearisation Techniques

Figure 5.10: Measured output spectrum from modulator with prefiltering

Figure 5.11: Measured output spectrum from modulator with prefiltering and PD
Figure 5.12: Output spectrum obtained using baseband serrodyne technique
5.4 Adaptive Predistortion technique for a LINC Transmitter

The LINC technique, introduced in Chapter 2, creates a varying-envelope signal from the vector sum of two constant-envelope signals of varying phase. Each constant amplitude vector component can be amplified with high efficiency. However, the system is sensitive to gain and phase imbalances between the two signal paths. Many techniques [58], [57] have been proposed to mitigate the effect. Recently, a technique to correct phase/gain imbalance by employing calibration algorithm in baseband processing was proposed [62]. Since the phase and gain imbalances of the amplifier and/or power combiner are dependent on frequency, input power and temperature, the predistorted gain should be self-adjustable. In this Section, a novel baseband adaptive correction technique using predistortion is proposed. The least-mean-square (LMS) algorithm is applied due to its excellent convergence and low computational complexity. The sensitivity of LINC system to phase and gain imbalance, in terms of the baseband I-Q diagram, are discussed. The adaptive baseband PD technique to adjust the phase and gain imbalance is developed to cure this problem. Simulation results for the proposed system with a square root-raised cosine QPSK signals are presented. Significant improvement of spectral regrowth is achieved from the proposed system, compared to the standard LINC, when phase and gain imbalances are introduced to the system.

5.4.1 The LINC Lineariser with Adaptive Baseband Predistortion

The LINC transmitter relies on the fact that the odd-order nonlinearity in the system does not affect on spectral regrowth for constant envelope signals.
5.4. Adaptive Predistortion technique for a LINC Transmitter

The technique separates the nonconstant envelope signal into two constant envelope signals by introducing the auxiliary signals. After combining the two output signals from the two high-efficiency HPA, the auxiliary signals with opposite phase will be cancelled out and hence the perfect signal spectrum will be obtained.

Figure 5.13 shows the adaptive predistortion LINC system. Let \( s_k \) be the baseband representation of the bandpass RF signal, then \( s_k = r_k e^{j\phi_k} \). \( r_k \) and \( \phi_k \) are the amplitude and phase of the modulated signal, depending on the shaping filter type and modulation scheme. As shown in Figure 5.14, \( s_k \) will be split into two constant envelope signals \( (s_{1k}, s_{2k}) \) by introducing the auxiliary signal \( a_k \). Hence \( s_{1k} = r_{\text{max}} e^{j(\phi_k + \theta_k)} \) and \( s_{2k} = r_{\text{max}} e^{j(\phi_k - \theta_k)} \), where \( \theta_k = \cos^{-1}\frac{r_k}{r_{\text{max}}} \). These two signals will be upconverted at the carrier frequency \( \omega_c \) and fed into the HPAs. Let \( \Delta g_A \) and \( \Delta \theta_A \) be the gain and phase imbalances of HPA_2 relative to the HPA_1. If the combiner is lossless and its gain and phase imbalances are denoted by \( \Delta g_c, \Delta \theta_c \), then \( \Delta g \equiv \Delta g_A + \Delta g_c + \Delta g_A \Delta g_c \) and \( \Delta \theta \equiv \Delta \theta_A + \Delta \theta_c \), represent the gain and phase imbalances from both HPAs.
Chapter 5. Transmitter Linearisation Techniques

![Polar Diagram](image)

**Figure 5.14:** LINC on polar diagram

and combiner. The LINC transmitter output at the desired frequency band is:

\[ s_{out}(t) = G r_{max} \text{Re} [e^{j\omega_k t + \phi_k} (1 + \Delta g) e^{-j(\theta_k - \Delta \theta)}] \]  

(5.17)

where \( G \) is the voltage gain of the HPAs and \( \text{Re} [\cdot] \) means real part term. As shown in Figure 5.13, the output signal is coupled with the coefficient \( \frac{1}{2\pi} \) and is subsequently downconverted, hence the demodulated signal is

\[ s_{dh} = \frac{r_{max} e^{j\theta_k}}{2} (1 + \Delta g) e^{-j(\theta_k - \Delta \theta)} + n_k \]  

(5.18)

where \( n_k \) is the noise signal arising from the downconversion mixer. It is shown in (5.18) that if there is no gain and phase imbalances, then the amplitude and phase of the demodulated signal is \( r_k \) and \( \phi_k \). The error vector signal \( e_k \) is formed by subtracting the sampled demodulated signals \( s_{dh} \) with the input \( s_k \).

The magnitude and phase of this error signal can be written as follows:

\[ \Delta r_k = \sqrt{1 + (1 + \Delta g)^2 - 2(1 + \Delta g) \cos \Delta \theta + |n_k|} \]  

(5.19)

\[ \Delta \theta_k = \tan^{-1} \left( \frac{(1 + \Delta g) \sin(\theta_k + \Delta \theta - \phi_k) - \sin(\theta_k - \phi_k)}{(1 + \Delta g) \cos(\theta_k + \Delta \theta - \phi_k) - \cos(\theta_k - \phi_k)} + \theta_{nk} \right) \]  

(5.20)

where \( |n_k| \) and \( \theta_{nk} \) are the magnitude and phase of \( n_k \), respectively. It is shown in (5.19)-(5.20) that \( \Delta g \) and \( \Delta \theta \) are converted to the phase and gain errors in
the polar diagram of the demodulated signals, which is shown in Figure 5.14. Hence, one can introduce the predistortion complex gain $g_d$ to the second channel signal to counterbalance the errors. Note that $g_d$ should be self-adjustable since $\Delta g$ and $\Delta \theta$ are sensitive to frequency, input power, and temperature.

Figure 5.15: The mean-square-error of phase and amplitude imbalance in LINC transmitter

5.4.2 Adaptation Algorithm

Assuming that the process is stationary and the noise and signal are statistically independent. The mean square error (MSE) for each transmitted sample is obtained from the error between the original baseband signal ($s_k$) and (5.17), hence:

$$
\xi = E[s_k s_k^*] = 1 + (1 + \Delta g)^2 - 2(1 + \Delta g) \cos \Delta \theta + E[n_k^2 - E^2[n_k]] \quad (5.21)
$$

Figure 5.15 shows $\xi$ in dB units with the introduction of phase and gain imbalance under the assumption that the noise power is negligible. From Figure 5.13, the demodulated signals, $s_d$ are applied to the gain estimation and the adaptive parts. From the LMS algorithm [30], the $(k + 1)^{th}$ complex gain sample,
$g_d[k+1]$, is updated using,

$$g_d[k+1] = g_{dr}[k+1] + jg_{dq}[k+1] = g_d[k] + \mu s_k s_{dk}^*$$

(5.22)

where $\mu$ is the step-size parameter which determines the convergence speed and noise variance of the system and $[\cdot]^*$ is the complex conjugate operator. The estimated $g_d[k+1]$, multiplied with the modulated signal at the second channel to provide the predistorted signal, gradually converges to the optimum value (in a minimum mean-square error sense). The LMS algorithm used to predistort the signal for LINC requires, at most, 8 multiplications and 6 additions, which is relatively low complexity.

### 5.4.3 Simulated Performance

Simulations have been performed to demonstrate the adaptive predistortion LINC using MATLAB\textsuperscript{TM}. The input signal is a square-root raised cosined QPSK signal. The HPAs were mathematically modelled by introducing a cubic term and a saturation function. The coefficient for the cubic term is set to be 20% less than the linear term ($G = 10$). The driven input power was so high that the HPAs were operated in a strongly nonlinear region. A vector modulator was cascaded with the second channel HPA to introduce the effect of phase and gain imbalances.

Figure 5.16 and 5.17 show the real and imaginary part of the complex gain for case 1 ($\Delta \theta = 0, \Delta g = 0.65$) and case 2 ($\Delta \theta = 30^\circ, \Delta g = 0.2$). The adaptation algorithm converges to the optimum values for both cases. The signal spectrum from the LINC and from the proposed system after the complex gain are adapted to the optimum value are shown in Figure 5.18 and 5.19, respectively. It is clearly shown that the effects of phase and gain imbalance are reduced dramatically. However the delay occurred in the feedback path and the A/D speed decreases the communication bandwidth.
5.4. Adaptive Predistortion technique for a LINC Transmitter

![Graph 1](image1.png)
Figure 5.16: The convergence characteristics of $g_D$ where $\Delta g = 0.65$, $\Delta \theta = 0^\circ$ (case 1)

![Graph 2](image2.png)
Figure 5.17: The convergence characteristics of $g_D$ where $\Delta g = 0.2$, $\Delta \theta = 30^\circ$ (case 2)
Figure 5.18: The simulated PSD results of LINC and adaptive PD LINC transmitter (case 1: $\Delta g = 0.65, \Delta \theta = 0^\circ$)

Figure 5.19: The simulated PSD results of LINC and adaptive PD LINC transmitter (case 2: $\Delta g = 0.2, \Delta \theta = 30^\circ$)
5.5 Conclusions

In this Chapter, three techniques for linearisation of direct-carrier transmitters are presented. For the FF technique, by using the output signal from a second modulator, which is operating at a low input signal level, a reference signal can be generated. The results demonstrate the FF technique can reduce distortion from both the modulator and the PA. The overall gain requirement, which is independent of the conversion loss of the modulator, is also smaller than a conventional transmitter with a FF amplifier. The proposed system is suitable for transmitters in which the modulator operates at a high-level. The technique offers good linearity and low complexity.

A low-cost transmitter architecture using software radio techniques is proposed and successfully implemented at V-band. The circuit is composed of a medium-power stable oscillator and a VM chip. The baseband signals are generated from a DSP card. Multi-functionality in the transmitter is implemented entirely at baseband: the data-rate, modulation type, wave-shaping filter, baseband predistorter, and frequency shifting functions are all realised in software. Thus, the transmitter complexity is greatly reduced.

An adaptive PD LINC transmitter topology is presented. The LMS algorithm was applied due to its convergence robustness and low computation complexity. Simulation results of the adaptive system shows the capability for significantly reducing the spectral regrowth resulting from the phase and gain imbalances. The operation bandwidth of the proposed system mainly limited by the feedback path and the DSP speed.
Chapter 6

Conclusion

6.1 Contributions of this Thesis

Although several linearisation techniques have emerged in the last decade, only a few of them can be applied to very high frequency applications, for example beyond X-band. In this work, the baseband PD technique has been applied to a VM in a low-cost transmitter architecture at V-band [12]. The technique does not require any additional components and so high power efficiency (which is a difficult issue in the millimetre-wave range) is achieved. The only disadvantage of the technique is the need for a priori information about the vector modulator’s characteristics. The adaptive control loop has not been applied since the thermal-dependence of the VM is less than for the HPA. The measured results shows more than 15-dB improvement of spectral regrowth without power consumption increase.

The harmonic injection technique is one of the linearisation techniques which can achieve a high power efficiency. It has been shown that a high linearity mixer can be realised using this technique. As in the original approach applied to a HPA, the amplitude and phase of the injected signal are
critical, resulting in a time-consuming design and implementation stage. The initial estimated amplitude and phase, based on a system analysis, can accelerate the design stage of the system. The preliminary experiment for this technique shows a promising result of the validity of applying the technique to a mixer.

A linearisation concept in which the main complexity is shifted to baseband has been applied to LINC transmitters. The so-called adaptive baseband PD LINC transmitter has been proposed to cure the effect of phase and gain imbalances in the system. It has been analytically shown that the IMD reduction performance is very sensitive. With the proposed technique, the complex gain converges to the optimum value successfully. Significant reduction in spectral regrowth was observed from the simulations.

For MMIC implementation, where performance varies due to process variations, the FF technique for a direct carrier transmitter [14],[13] or the simplified FF mixer [18],[17] require more development due to their sensitivity to amplitude and phase imbalance. Adaptive techniques to control the amplitude and phase of the loop(s) are required. In addition, the power efficiency and power handling in the FF technique is significantly degraded in a MMIC implementation. However, the technique still can be used directly for HMICs, where tuning is possible in production, and in some applications at low frequency.

Having developed the low-cost short range transmitter architecture [12] which eliminates two major components - the HPA and tunable LO - it has been shown that a multifunctional and linearised transmitter at millimeter-wave frequency can be realised. The hardware complexity is largely reduced by handing the difficult tasks to the control software. These are the modulation scheme, data rate, linearisation, and channel selection. Not only are cost and complexity reduced, but this is also a versatile transmitter which has flexibility, easy maintenance, and future support for more advanced commu-
nication technology. Moreover, the total power efficiency is high due to the minimum number of components needed in the system. However, the transmitted signal quality mainly depends on the LO source which needs to be a stabilised medium power signal with low phase noise. The bandwidth of the system is determined by the speed of the D/A converters. The latter issue tends to be alleviated due to technology advances in the area of data conversion. Apart from these disadvantages, it is clear that this transmitter topology is very useful for some applications in the RF to millimeter-wave range.

Since the proposed transmitter topology employed the VM MMIC chip based on the RTA structure, improved VM circuits have been studied. Having thoroughly investigated and gained insights into the VM RTA circuit, a simple technique to reduce the circuit size, rather than employing the balanced structure, has been demonstrated [15]. The technique compensates the cold-FET parasitic elements using an inductor in series with the source terminal. The required inductance is small and not critical. The number of baseband control channels has also been reduced from four channels to two, because the complementary channels required in the balanced structure are eliminated. Although amplitude distortion still exists, the proposed technique combined with a baseband PD technique provides a good candidate for the VM in microwave and millimetre wave applications.

The feedback RTA [16] proposed in this Thesis is simple and easy to implement. No extra circuit area is required for the technique. The technique involves trading off the input and output return loss with the transmission performance. The linearity of the RTA was investigated [19] and the analysis provides the criteria required to design a circuit which provides a good linearity for the RTA. In addition, it has been analytically and experimentally observed that the distortion levels from the RTA at different attenuation values are not equal. It is concluded that at the maximum attenuation level, the main in-
put power is absorbed by the device while at the minimum attenuation level, the main input power is reflected to the output load. In brief, this nonlinearity investigation adds another consideration into the design of the variable attenuator or VM based on the structure.

In conclusion, the aims of the research as outlined in Chapter 1 have been achieved. It is shown that all the techniques that have been studied have application from RF to millimetre-wave frequencies.
6.2 Suggestions for Future Work

Various techniques have been described in this Thesis for improving the performance of the existing linearisers. Some linearising techniques have been applied to a mixer. Beside these, techniques for improving the performance of some necessary subsystems utilised in a lineariser, namely the VM and RTA, have been proposed. As a result of this research, a number of areas for further work have become apparent. A few of them have already been investigated by fellow researchers in the group.

The VM with the compensation technique, while requiring half the circuit area compared with the balanced structure, shares the same common problem; that of amplitude distortion. Amplitude distortion in the VM can be cured with a baseband PD technique, but a priori information is required. Based on the same fundamental concept of baseband PD, a circuit-level analogue PD technique can be applied to the VM. Using the I-V relationship in the cold-FET, the inverse function, which is an inverse hyperbolic tangent function, can be realised by analogue circuit design techniques.

Although the validity of the simplified FF mixer has been proved with a passive mixer at system level, the technique implemented at a circuit level based on an active mixer is attractive. It is clear that if an active mixer is used, the auxiliary amplifier in the loop will be replaced with an attenuator, which can be easily implemented and no extra DC power needed. Apart from this possibility, the noise figure of the simplified FF mixer should be investigated.

Also, the HI technique for mixer applications can be implemented at a circuit level. Both techniques in HI, second-harmonic and frequency difference, can be simply generated with a diode or a transistor cascaded with a band-pass or low-pass filter to select the desired harmonic components. The difference-frequency injection technique is more attractive due to its low frequency of
operation and better performance in reducing the inband nonlinear distortion.

The adaptive predistortion LINC concept proposed in this Thesis has not been yet implemented due to the complexity of the baseband systems and related hardware limitations. The signal processing part can be realised with DSP, FPGAs or ASICs. At the time of writing, this work is being undertaken by another researcher in the group.
Bibliography


Bibliography


