MILLIMETER WAVE MMIC AMPLIFIER LINEARIZATION BY PREDISTORTION

A THESIS SUBMITTED TO
THE GRADUATE SCHOOL OF NATURAL AND APPLIED SCIENCES
OF
MIDDLE EAST TECHNICAL UNIVERSITY

BY

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IN PARTIAL FULFILLMENT OF THE REQUIREMENTS
FOR
THE DEGREE OF MASTER OF SCIENCE
IN
ELECTRICAL AND ELECTRONICS ENGINEERING

DECEMBER 2006

Approval of the Graduate School of Natural and Applied Sciences

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ABSTRACT

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December 2006, 95 pages

For millimeter wave applications, MMIC is the best contemporary technology. Considering the requirements of the commercial and military applications on amplitude and phase linearity, it is necessary to reduce the nonlinearity of the amplifiers. There are several linearization techniques that are used to reduce the nonlinearity effects. In the context of the thesis, a special analog predistortion technique that is called “self cancellation scheme” is used to linearize a 35GHz MMIC amplifier. The amplifier to be linearized is used in the design of the predistorter, that is why it is called self cancellation.

This thesis contain the design of the amplifier, lumped element power divider and combiner circuits, and the complete analog predistortion linearizer. Layouts of linearizer system and its components are prepared and layout effects are taken into account.

Keywords: MMIC amplifier, predistortion, linearization

ÖZ

ÖN BOZUNUM İLE MİLİMETRİK DALGA MMIC YÜKSELTECİ DOĞRUSALLAŞTIRMASI

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Tez Yöneticisi: Doç. Dr. Şimşek DEMİR

Aralık 2006, 95 sayfa

Çağdaş milimetrik dalga uygulamalarında MMIC teknolojisi kullanılmaktadır. Endüstriyel ve askeri uygulamalardaki genlik ve faz doğrusallığı gereksinimleri değerledirildiğinde yükselteçlerin doğrusal performanslarının arttırılması ihtiyacı doğmaktadır. Yükselteç doğrusallaştırması için çeşitli yöntemler kullanılmaktadır. Bu tez kapsamında, analog ön bozunum tekniğinin özel bir kullanımı olan "kendi kendine sönümleme" yöntemi ile 35 GHz te çalışan bir MMIC yükselteç doğrusallaştırması yapılmıştır. Doğrusallaştırılması istenilen yükselteç önbozunum devresi tasarımında da kullanıldığından; yöntem, "kendi kendine sönümleme" olarak adlandırılmıştır.

Tezde, yükselteç, güç bölücü ve birleştirici devre tasarımları ve analog önbozunum doğrusallaştırıcı tasarımı içerilmektedir. Tasarlanan doğrusallaştırma devresi ve devre elemanlarının devre şemaları çizilerek devre şeması etkileri değerlendirilmiştir.

Anahtar Kelimeler: MMIC yükselteç, ön bozunum, doğrusallaştırma

To My Family

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CHAPTER 1

INTRODUCTION

Millimeter wave radars are employed in a wide range of commercial, military and scientific applications for remote sensing, safety, and measurements. Millimeter wave sensors are superior to microwave and infrared-based sensors in most applications. Millimeter wave radars offer better range resolution than lower frequency microwave radars, and can penetrate fog, smoke and other obscurants much better than infrared sensors.

Interest in millimeter-wave applications stems from the special properties exhibited by radar at these frequencies, as well as from the challenge of exploiting a region of the spectrum not widely used. The major attributes of the millimeter-wave region of interest to radar are the large bandwidth, small antenna size, and the characteristic wavelength. Large bandwidth means that high range-resolution can be achieved. It also reduces the likelihood of mutual interference between equipments, and makes more difficult the effective application of electronic countermeasures. The short wavelengths allow narrow beamwidths of high directivity with physically small antennas. Narrow beamwidths are important for high-resolution imaging radar and to avoid multipath effects when tracking low-altitude targets. The short wavelengths of millimeter waves are also useful when- exploring scattering objects whose dimensions are comparable to the millimeter wavelengths, such as insects and cloud particles. These are examples of scatterers whose radar cross sections are greater at millimeter wavelengths than at microwaves since they are generally in the resonance region at millimeter wavelengths, but in the Rayleigh region at microwaves (where the cross section varies as the fourth power of the frequency). Another advantage of the short wavelengths is that a doppler-frequency measurement of fixed accuracy gives a more accurate velocity measurement than at lower frequencies. The large attenuation of millimeter waves propagating in the clear atmosphere sometimes can be employed to advantage in those special cases where it is desired to reduce mutual interference or to minimize the probability of the radar being intercepted by a hostile intercept (elint) receiver at long range [1].

Millimeter wave radar operates at frequencies from 30 GHz up to 300 GHz, the term indicating the use of electromagnetic waves measuring between 10 and 1 mm in length. Millimeter waves are severely attenuated by absorption in the earth's atmosphere, and the frequencies near the attenuation minima, or RF "windows" at 35, 94, 140, 200 GHz are where most practical radar applications can be found [2].

Millimeter Wave Radars typically operate around 35 GHz and 94 GHz. They offer increased angular resolution as compared to microwave radars potentially wide transmission bandwidth. However, their operation is generally restricted to short range applications due to the considerable attenuation of millimeter waves, which occurs in the lower atmosphere. The millimeter wave radar could make use of propagation attenuation to achieve covert operation [3].

An ideal amplifier does not produce any distortion, that is no unwanted changes in the time domain waveform or frequency spectrum of the input signal; the output is related to the input in a purely linear fashion. In practice however, the components used in amplifier design, such as transistors, have nonlinearities resulting in distortion at the output. The degree to which a particular component is nonlinear depends to a large extent on the signal level and biasing arrangement and is often a complex relationship. The details in nonlinear behavior and bias techniques can be found in [4].

As a simple example that expresses the need for linearity can be the power amplifier stage of a loudspeaker in a music system. The goal of the music system is producing soft and loud music without distortion. The only way of achieving this is having a linear amplifier. Many power amplifiers have similar requirement, although the definition of acceptable distortion is more complex.

Since there is trade off between linearity and efficiency in amplifier design and if the aim is to achieve good linearity with reasonable efficiency, some type of linearization technique has to be employed. The main goal of linearization technique is to apply external linearization to a reasonably efficient but nonlinear amplifier so that the combination and the amplifier satisfy the linearity specification. In principle, this seem simple enough, bur several higher order effects seriously limit its effectiveness, in practice [5].

There are several linearization techniques exist, that are summarized in Chapter 2 and can be found in [5-8]. Stated briefly, linearization can be thought of as a cancellation of the distortion components, and especially as a cancellation of third order intermodulation distortion (IMD3), and where the achieved performance is proportional to the accuracy of the canceling signals. Unfortunately, the IMD3 components generated by the amplifier are not constant but vary as a function of many input conditions. Such as amplitude and signal bandwidth. Here, these bandwidth dependent phenomena are called memory effects. If the phase of an IMD3 component rotates 10° to 20°, or its amplitude changes 0.5 dB with increased tone spacing in a two tone test, it is usually does not have a dramatic effect on the adjacent channel power ratio (ACPR) performance of a standalone amplifier, nor is it especially of concern if the lower ACPR is slightly different from the upper one. However, the situation may be quite different if certain linearization techniques are used to cancel out the intermodulation sidebands; in fact the reported performance of some simple techniques may actually be limited not by the linearization technique itself, but by the properties if the amplifier, especially by memory effects.

Different linearization techniques have different sensitivities to memory effects. The details for the memory effects can be found in [5].

In the context of the thesis, the linearization techniques are examined and an application of analog predistortion is done. The thesis is given in five chapters.

In Chapter 1, an introduction to millimeter waves and power amplifier linearization is given.

In Chapter 2, the principals of linearization and linearity are described. The parameters that are used to specify the linearity of an amplifier are discussed. The basic linearization techniques that have been developed are told briefly. A comparison of the linearization techniques is given. Then the predistortion linearization technique is analyzed in detail and the studies on this technique are examined.

In Chapter 3, the selected predistortion method is analyzed in detail. The previous studies on this method are referenced and a simulation in ADS is performed using system components at a frequency of 1 GHz. The results of the analysis are given and the expected results for 35 GHz are discussed.

In Chapter 4, the implementation of the offered predistortion method is demonstrated in ADS using a MMIC amplifier at a frequency of 35 GHz. The design of the amplifier to be predistorted and the results of the amplifier analysis are given. The details of the predistorter circuit, design and analysis of the other components is given. The analysis of the overall system at 35 GHz is performed. The results obtained from ADS are given and discussed.

In Chapter 5, the conclusions obtained in the context of the thesis are given.

CHAPTER 2

LINEARITY AND LINEARIZATION

2.1 LINEARITY

An amplifier is said to be linear when the output voltage waveform is the replica of the input voltage waveform, i.e. no distortion is introduced into the output signal. But this is not the case in practice and many techniques are developed to have more linear amplifiers. The linearity of an amplifier is analyzed in [4-10] in detail.

Linearity should be considered only when the amplitude modulation is used. In modern communication methods AM is also used with frequency modulation (FM) and phase modulation (PM).

For an ideal amplifier that is perfectly linear, the output voltage is proportional to the input voltage. Assuming the two port system in Figure 2-1 as a linear amplifier with input voltage $V_{in}(t)$, output voltage $V_{out}(t)$ gives

$$V_{out}(t) = G\, V_{in}(t) \qquad (1)$$

where G is the gain of the amplifier and transfer function $T(\omega) = G$.

The amplifier being linear, the input voltage is amplified with the same gain regardless of the input level and the phase shift between the input and the output is fixed for a signal at a given frequency.

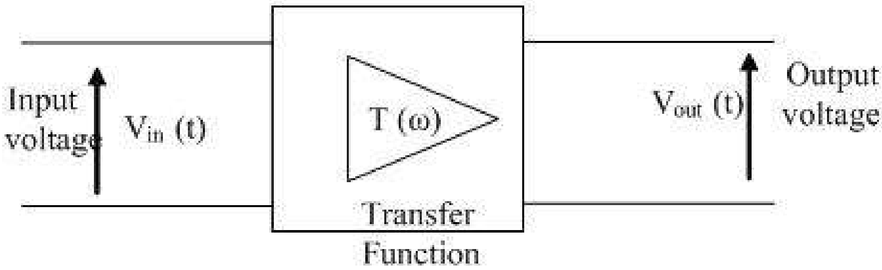

Figure 2-1 Block Diagram for a Power Amplifier

In frequency domain the transfer function T (ω) has constant characteristics over the bandwidth of the input signal, i.e. constant gain, linear phase and constant delay.

An ideal amplifier is memoryless which means that the output signal is only dependent to the input signal level at that time; it is independent from the previous input signal level. However, there can occur electrical memory effects due to frequency dependent envelope, fundamental and harmonic node impedances, i.e. bias impedance, and electrothermal memory effects due to electrothermal couplings. The details of memory effects can be found in [5].

In practice the amplifiers are not ideal; not linear. Due to nonlinearities of the components used in amplifier design, there occurs some distortion at the output of the amplifiers. The main nonlinearity effects are due to the semiconductor components in the amplifier; i.e. transistors. Without the semiconductor components, amplification is not possible and therefore nonlinearity is unavoidable.

The linearity requirement of the power amplifier is an important criterion in determining the design strategy. Linearity criterion can be achieved by using an amplifier with higher P_{1dB}. However, in this case, DC power consumption increases and efficiency decreases. Increasing linearity without decreasing efficiency is called linearization. In order to understand linearization better, in the following parts, nonlinearity effects are discussed.

2.1.1 AM to AM and AM to PM DISTORTION

For an amplifier the plot including output power level versus input power level is called the compression curve of the amplifier [10]. A typical compression curve is given in Figure 2-2. As it is seen from the curve the gain of an amplifier is usually constant at input power levels smaller than the maximum. So it can be said that in that region the input-output relation of the amplifier is linear. However increasing the input power level makes this relation nonlinear. In this region, that is also called the compression region, equal increments in the input power result in smaller increments in the output level. This means that the gain of the amplifier starts to decrease.

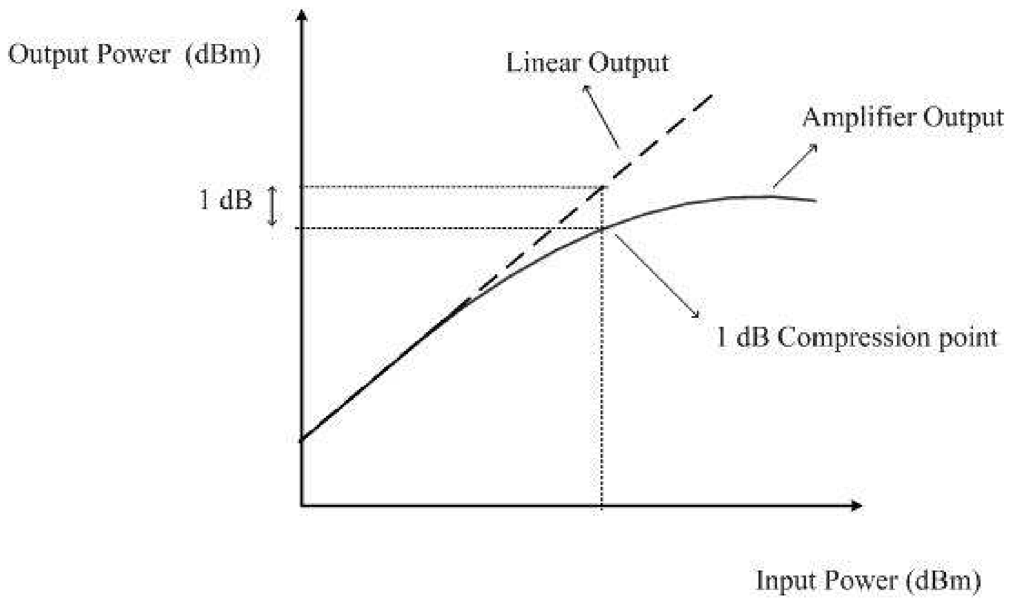

Figure 2-2 Compression Curve of an amplifier

Nonlinearity effects of the compression of the output power creates distortion as a function of amplitude; AM to AM distortion. Beside this distortion, the phase difference between the input and output signals which are also a function of input signal level introduces an extra distortion. This extra distortion due to phase difference is referred to as AM to PM distortion. AM to PM distortion may result from the storage delays in bipolar transistors or voltage dependent capacitance of device junctions.

The point at which the output power of the amplifier is dropped 1 dB below the output of a linear amplifier with same gain at the same input level is called the 1 dB compression point, P_{1dB}. In other words 1 dB compression point is the point at which the nonlinearity results in a 1 dB compression of the gain. P_{1dB} is used to assess the nonlinearity of the power amplifier.

In order to meet the linearity requirements, the maximum output power demand can be kept below P_{1dB} level. This is called as power back off; increased linearity in spite of decreased efficiency.

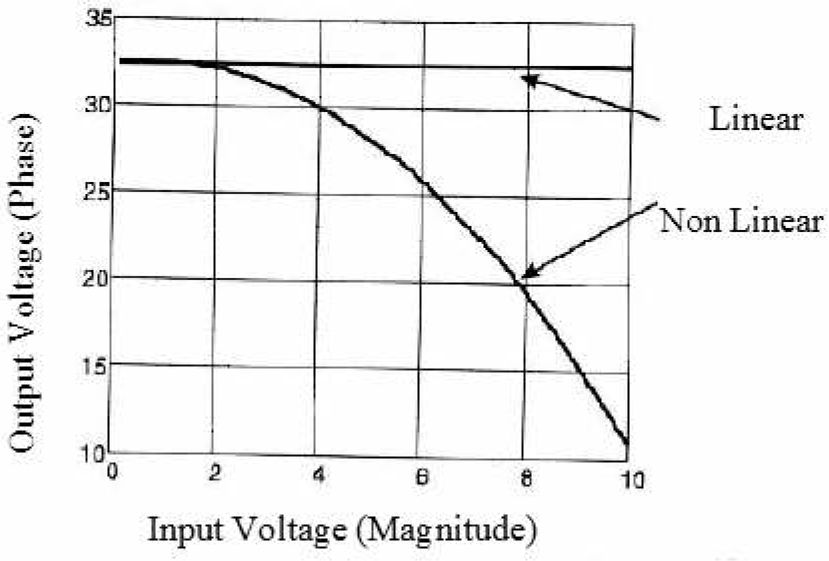

Figure 2-3 AM-PM Characteristic

2.1.2 ADJACENT CHANNEL POWER RATIO (ACPR)

Adjacent channel power ratio (ACPR) is a metric that can be used to measure the nonlinearity of a power amplifier. Nonlinearity of an amplifier may result in a spread of the output signal into adjacent frequency bands. The more the output signal is spread the more the amplifier is nonlinear.

ACPR is defined as the power ratio of the total power within a certain bandwidth, separate from the transmission channel but adjacent to the transmission channel, to the total power within the transmission bandwidth. A graphical representation of ACPR is given in Figure 2-4.

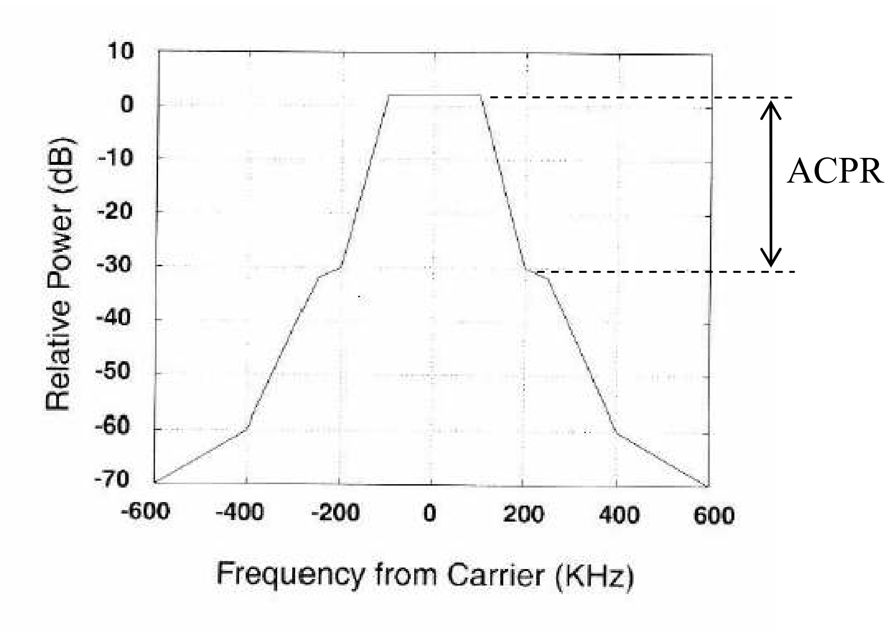

Figure 2-4 ACPR in Spectral Mask

2.1.3 THIRD ORDER INTERCEPT POINT (IP3)

The transfer function of an amplifier can be approximated with a polynomial. In the presence of two single tone sinusoidal input signals the polynomial approximation of the amplifier transfer function will result in extra signals at different frequencies, in addition to the two input frequencies. Input signal consisting of two intermodulation tones input can be shown as:

$$V_{in} = a_1 \cos(\omega_1 t) + a_2 \cos(\omega_2 t) \qquad (2)$$

The polynomial approximation of the amplifier transfer function will result an output in the form,

$$V_{out} = \sum a_{nm} \cos(n\omega_1 t \pm m\omega_2 t) \qquad (3)$$

where m and n are integers.

The third order nonlinearity of the transfer function will be at frequencies $(2\omega_1 - \omega_2)$ and $(2\omega_2 - \omega_1)$. Calling $(\omega_1 - \omega_2)$ as $\Delta\omega$ the third order nonlinearity terms become at frequencies $(\omega_2 - \Delta\omega)$ and $(\omega_1 + \Delta\omega)$. If the two input signals are not separated from each other compared to their own frequencies, that is, $\Delta\omega = \omega_1 - \omega_2 << \omega_1, \omega_2$ then the third order nonlinearities become so close to the desired signals making it impossible to filter them out. This situation results in third order spurious signals at the output with voltage level varying with the third order power of the input voltage.

Since the fundamental signal at the output linearly increases with the input while the third order terms increases with the third power of the input, there is a point at which the third order terms are equal to the fundamental signal. This point is called the third order intercept point (IP_3), see Figure 2-5.

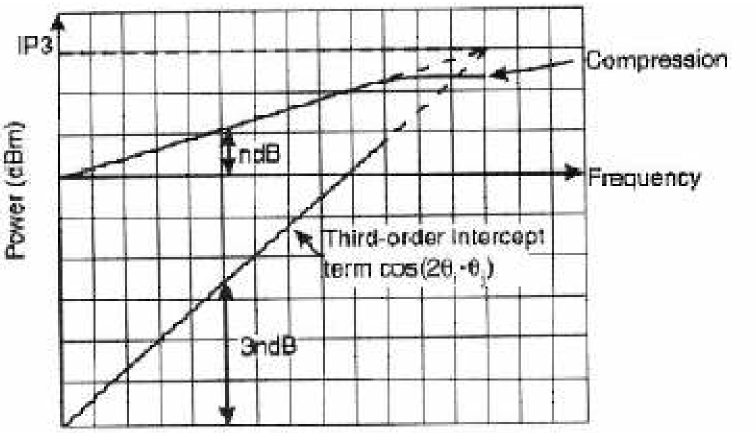

Figure 2-5 Definition of the third order intercept point as the intersection of the fundamental and the third order distortion component

In the final design, complete nonlinear transistor model is used. Thus, the simulation results are the expected measurement results at the intended drive level. IP3 as a criterion for nonlinearity refers to the small signal operation by definition. Thus, IP3 does not reflect the high power level operation condition. ACPR considers the total nonlinearity effects; however it is commonly used for communication systems. Therefore, neither IP3 nor ACPR by itself is the meaningful criterion; therefore we used a combination which is a sort of "effective IP3" definition.

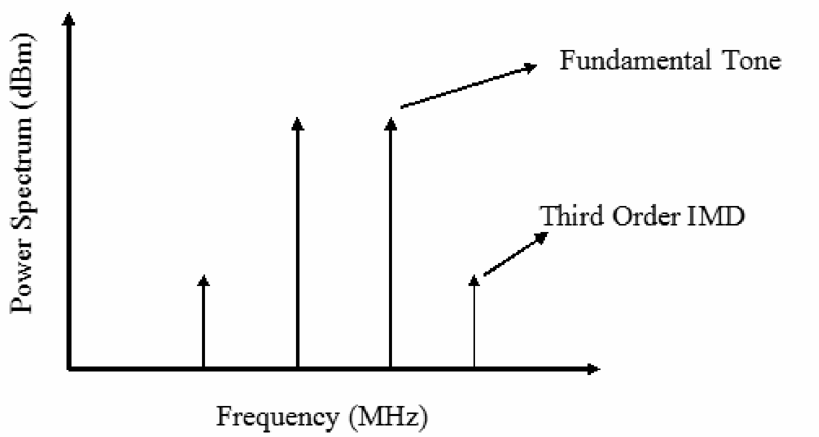

Figure 2-6 Typical Power Spectrum of fundamental and IMD components

In this thesis study, third order IMD components are identified as the components in the adjacent channel, see Figure 2-6. Thus ACPR criterion referred now on is the ratio of the fundamental component to the IMD component under two tone excitation.

2.2 LINEARIZATION TECHNIQUES

The main considerations of amplifier design are efficiency and linearity of the amplifier. Efficiency is important especially in mobile systems since it affects the battery life which is the main problem of a mobile system. But in some applications efficiency becomes the second consideration after linearity. Single and multi channel base station transmitter systems may be examples of such applications.

Multi channel power amplifier applications have heavier linearity requirements compared to that of single channel ones. Here the linearization techniques take place. The linearization techniques will be examined later in this chapter. But whatever method is used; the linearization will end up with an input-output relation something like in Figure 2-7

As impressed in Figure 2-7 the amplifier does not increase the intrinsic power capability, it just gives harder saturation characteristic to the amplifier. It is obvious that the linearization is ineffective at the saturation region and so the linearized amplifier should be backed off from its compression region.

Linearization techniques can be separated into two classes; closed loop linearization techniques and open loop linearization techniques. Closed loop linearization techniques such as feedback are better in precision but they are limited in modulation bandwidth. This limitation can make their usage inefficient especially in multi-channel applications.

On the other hand open loop linearization techniques such as predistortion are better in modulation bandwidth but they cannot reach the closed loop linearization technique precision.

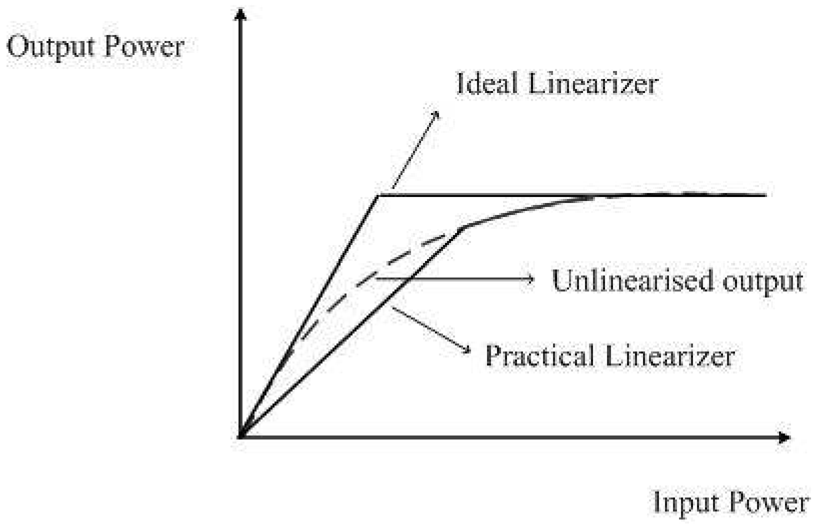

Figure 2-7 Output Power versus Input Power graphic of an amplifier

The feedforward technique is neither closed loop nor open loop technique. Feedforward offers the precision of closed loop linearization with stability and bandwidth performance of the open loop technique. But of course it is not the best one since its efficiency is worse due to the extra amplifier in the correction loop.

2.2.1 FEEDBACK

Feedback linearization technique forces the output to follow the input. There are two ways of feedback. One is direct feedback applied directly to the RF amplifier which is also called RF feedback, and the other one is indirect feedback which is applied indirectly to the modulation (envelope, phase or I and Q components)

2.2.1.1 DIRECT FEEDBACK

In direct feedback strategy a portion of the RF output signal is taken and subtracted from the RF input signal.

The use of direct feedback is usually restricted to HF and lower VHF frequencies. The classical application of direct feedback is shown in the Figure 2-8 β being the feedback gain and A being intrinsic gain of the amplifier.

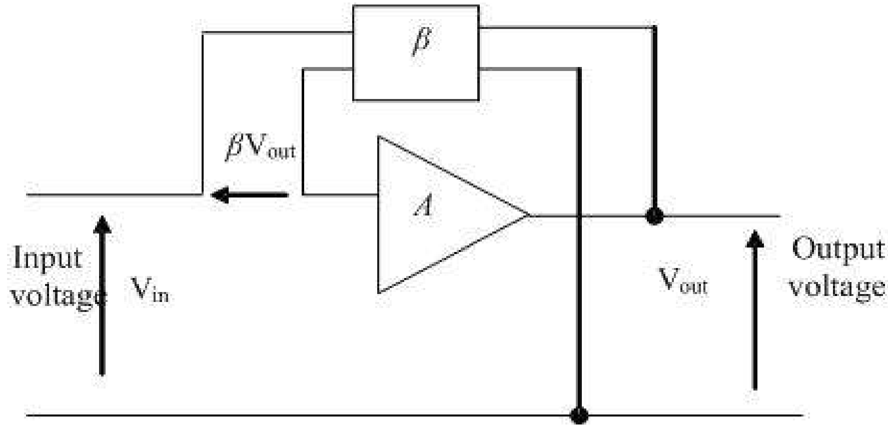

Figure 2-8 Classical application of direct feedback

Then the overall gain *G* can be evaluated as follows

$$\boldsymbol{G = A / (1+\beta A)} \qquad (4)$$

where *A* is assumed to incorporate a 180 degree phase shift for a single stage low frequency amplifier. Since *A* has a rapidly varying phase shift due to frequency at microwave frequencies, the phase of *β* should be adjusted so that at the point of feedback the feedback signal is subtracted from the main signal.

This analysis assumes that the process occurs instantaneously, that is there is no time delay on *β* and *A* components. In other words feedback technique ignores the time delay which results in difficulties in using feedback at higher frequencies

2.2.1.2 INDIRECT FEEDBACK

In envelope feedback strategy RF input signal is sampled by a coupler and the envelope of the input is detected by an envelope detector. The RF output signal is also sampled by a coupler and the envelope of the output is detected by an envelope detector. Then the input envelope and the output envelope are fed to a differential amplifier in order to amplify the difference between the input and output envelopes. The difference signal is used to drive modulator that modifies the envelope of the RF signal which drives the power amplifier.

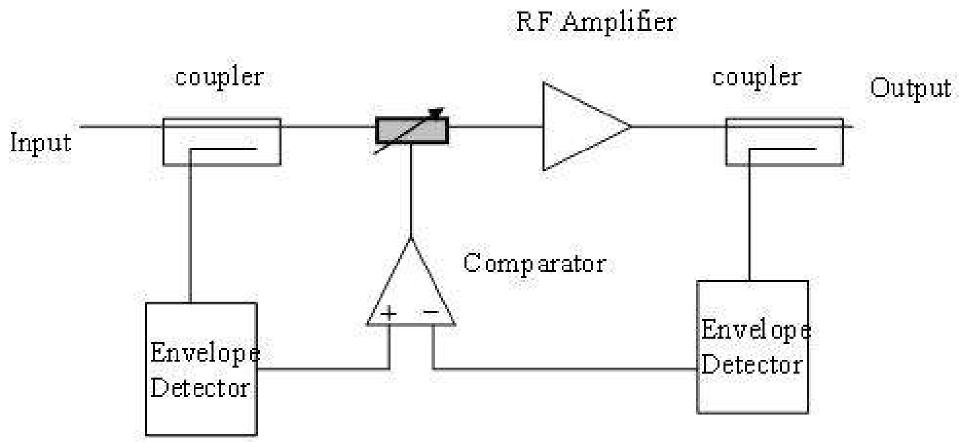

Figure 2-9 Indirect feedback

The working principle of indirect feedback is given in Figure 2-9. Provided that the amplifier is below the saturation this feedback loop will force the envelope of the output signal to replicate the input envelope.

We have said that the amplitude correction cannot increase the intrinsic power saturation. So as the envelope swings in to the compression region of the amplifier the effectiveness of the above procedure decreases.

Two alternative indirect feedback strategies are Cartesian Feedback and Polar Feedback, using the detected or down-converted input and output envelope amplitude and phase responses to generate error correcting loops.

2.2.1.3 POLAR FEEDBACK

Polar feedback is a logical extension of indirect feedback method. In polar feedback both phase and amplitude corrections are performed simultaneously. Polar feedback overcomes the fundamental inability of envelope feedback to correct for AM-PM distortion by adding a phase-locked loop to the envelope feedback system.

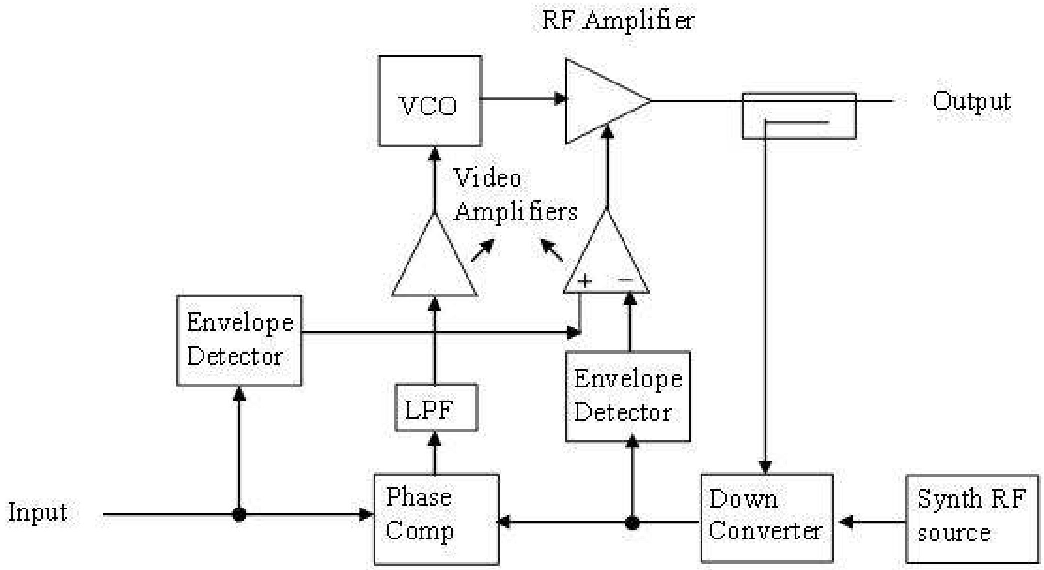

Figure 2-10 Polar feedback

The Figure 2-10 showing the implementation of polar loop seems to be complicated. Phase detection and correction is more complicated than that of amplitude. The phase locked loop in the figure is used to maintain a constant phase transfer characteristic.

The bandwidth requirement of the video amplifiers is a key issue in polar feedback systems. The phase amplifier will require higher bandwidth, but both of the amplifiers should have the same bandwidth and phase performance in order to avoid additional AM to PM distortion while correction process takes place.

2.2.1.4 CARTESIAN FEEDBACK

Cartesian Feedback has some benefits over the polar feedback. In modern digital systems it is more likely that the baseband signal will be available in *I* and *Q* format and so given that the modulation signal is in its baseband form it is possible to split the modulation signal into two quadrature channels; in order to use for amplitude and phase.

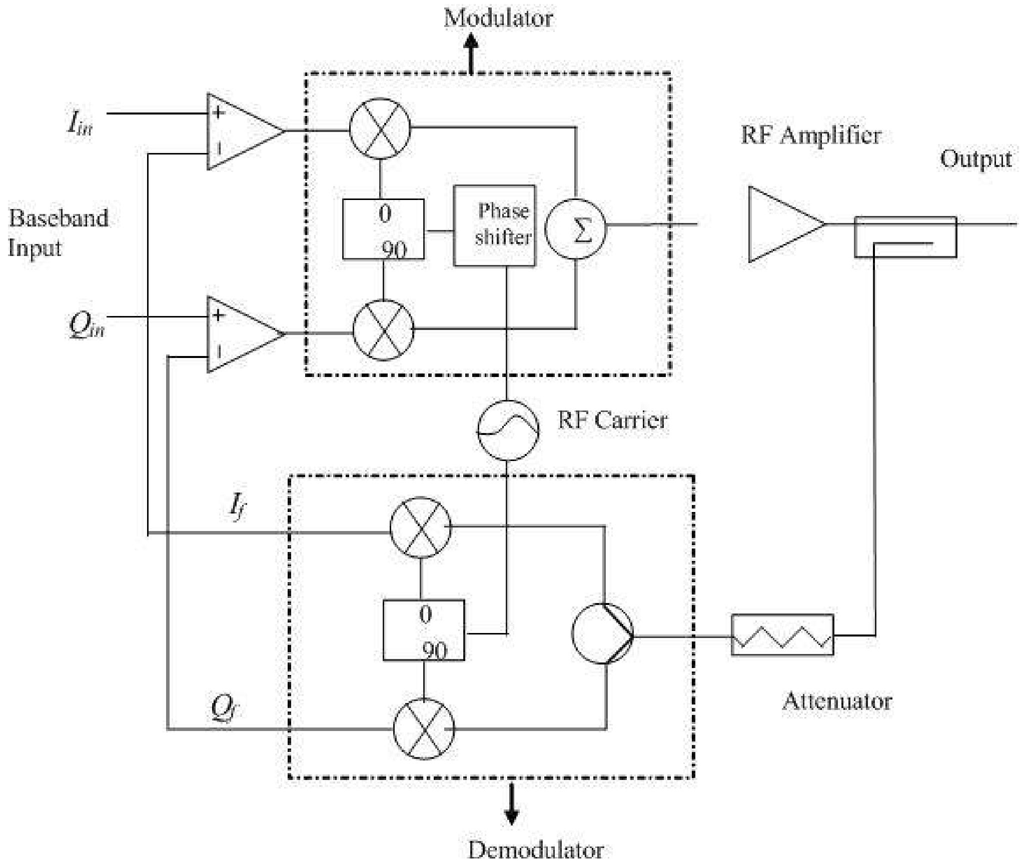

Figure 2-11 Cartesian feedback

In the Cartesian Feedback technique modulation feedback is applied on *I* and *Q* components so that it overcomes the problems associated with the wide bandwidth of the signal phase.

There are two identical feedback processes on the *I* and *Q* channels operating independently as shown in Figure 2-11.

The I and Q channel inputs are amplified and the amplifier output is sampled via a coupler. The sample is attenuated by a down-converter which has similar configuration with the up-converter used at the input. This process yields the distorted I and Q channels and these distorted signals are compared to the undistorted input signals in the input differential amplifiers. The output of the differential amplifiers will force the overall output I and Q channels to replicate the input I and Q channels.

The phase shifter shown in the up-converter local-oscillator path is used to align the phases of the up- and down-conversion processes.

2.2.2 PREDISTORTION

Predistortion is a commonly used linearization technique; that has a very simple idea. Since the amplifier has a distortion and the transfer function is known, an extra distortion that will eliminate the amplifier distortion at the output can be cascaded before the amplifier. The important point is that the distortion characteristic should be complementary to the amplifier distortion.

2.2.2.1 ANALOGUE PREDISTORTION

Analogue predistortion is ideal for use in linearization of wideband multi carrier systems due to its ability to linearize the entire bandwidth of an amplifier simultaneously [8]. There are many various applications of analog predistortion method [11-16].

The predistorter part of the linearization system is cascaded with the RF amplifier as shown in Figure 2-12.

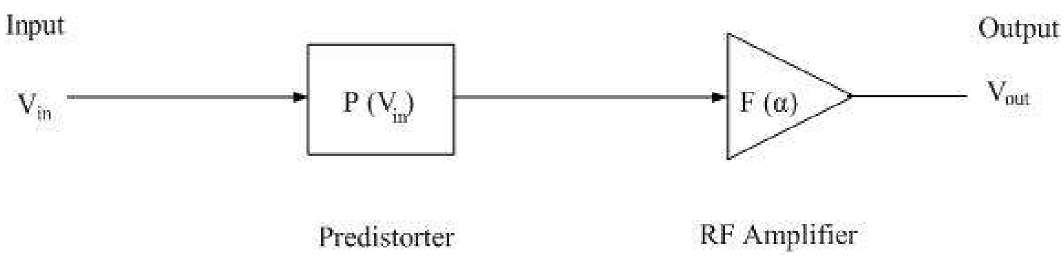

Figure 2-12 Block Diagram for Analogue Predistortion

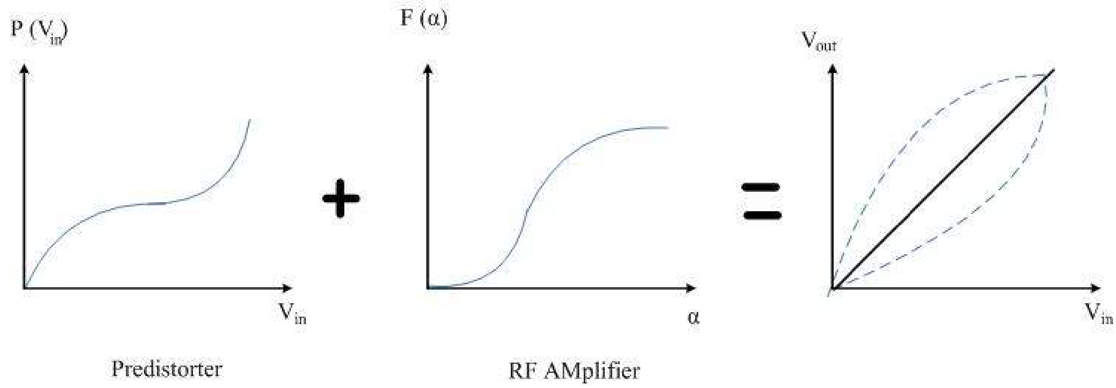

Figure 2-13 Operation of the Analogue Predistortion

Here, the aim is adjusting the predistortion block such that cascading the predistortion and the system will result in a linear transfer function.

The linearization operation of predistortion is shown in Figure 2-13. The first graphic is the transfer function of predistortion part, the second one is the transfer function of the system to be linearized and the last is the overall transfer function.

There are various examples for predistortion block some of which are given here.

2.2.2.2 CUBIC PREDISTORTER

In cubic predistorter case the aim is eliminating the third order distortion. The operation is basically adding the cubic component of the input signal with a correct phase difference in order to eliminate cubic distortion [13, 14].

As shown in Figure 2-14, the input signal is split into two in order to form two paths. The upper path is the main path and the lower one is secondary path.

The time delay element in the main path is due to compensate the time delay of components in the secondary path. The insertion loss of the time delay element may be neglected since it will operate at low power level.

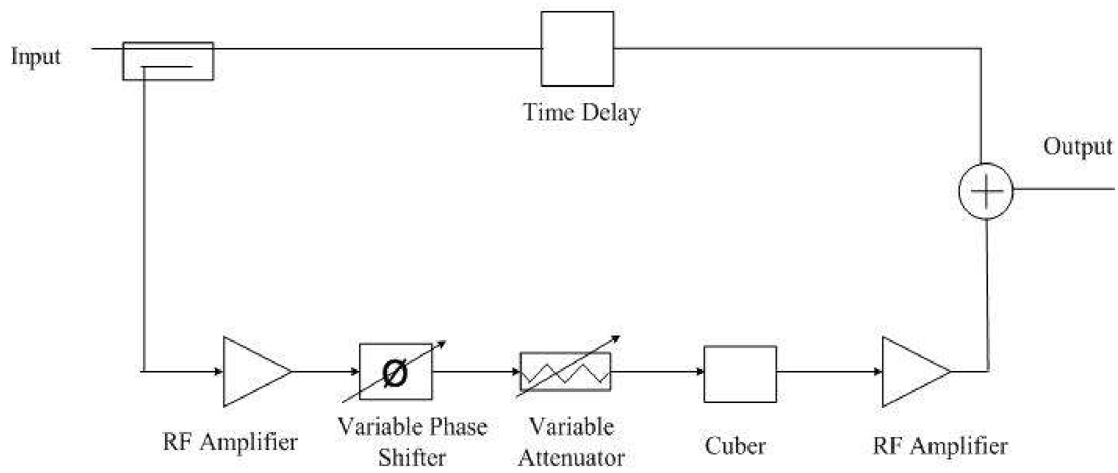

Figure 2-14 Block Diagram of Cubic predistorter

The phase difference between the main and the secondary paths is adjusted by the phase shifter in the secondary path so that the secondary path signal is subtracted from the main path signal.

The cubic circuit may be constructed in many ways most of which contains a diode or transistor having similar third order characteristics of an amplifier.

By the way the figure above is just an example; the order of the elements may be changed. The phase shifter can be used in the main path for example; of course satisfying a 180 degree phase difference between two paths.

2.2.2.3 DIGITAL PREDISTORTION

Digital predistortion operates by forming a complementary non-linearity to that of the PA, in the form of either a look-up table (LUT) or a polynomial approximation (or a combination of the two). This non-linearity is typically contained within a DSP, FPGA, or ASSP and is updated in response to a feedback signal from the output of the power amplifier. Updating can be based upon the minimization of adjacent channel energy, or upon coherent measurement of the error for each coefficient in the LUT or polynomial approximation [8].

A baseband voice or data input signal is converted to a suitable sampled I/Q format of the desired modulation scheme by the voice/data coder. The I/Q signals then undergo a complex multiplication with the relevant coefficients from the look-up table (or interpolated values derived from it) before DC elimination (if required) and adaptive error correction (based on the difference between the baseband I/Q and down-converted I/Q signals).

2.2.3 FEEDFORWARD

The feedforward technique details can be found in [4]. It has excellent linearity performance when controlled with an automatic control technique. This method can be applied more than one in order to have better linearity. The feedforward technique is relatively complex and so expensive. But it has been used in base station applications widely.

Basic configuration of the feedforward linearization technique is given in Figure 2-15.

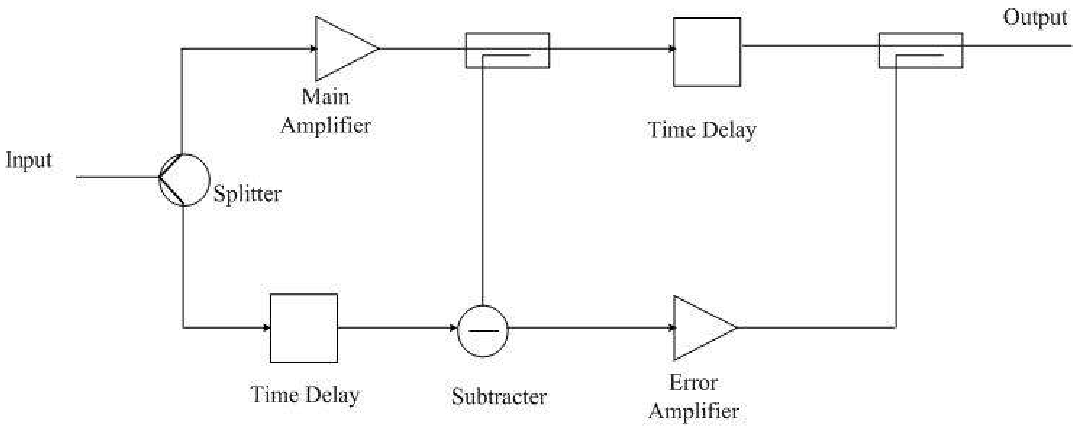

Figure 2-15 Feedforward

The input signal is split into two; main and secondary paths. The main amplifier is on the main path producing the amplified signal, of course with intermodulation and harmonic distortions added due to nonlinearity of the amplifier. A sample from the amplified signal is taken with the coupler and fed into subtractor. On the secondary path there is a time delay element to compensate the time delay of the main amplifier in the main path. The output of the subtractor contains only the nonlinear components which are the distortion components of the amplified signal; ideally none of the original signal energy would remain.

The signal in the secondary path amplified with an amount that will cancel the distortion elements of the main amplifier. The time delay element in the main path is due to error amplifier in the secondary path. Since the error amplifier output has 180 degree phase shift with main path signal and appropriate magnitude, the output signal will be the linearly amplified signal of the input. The distortion components are all cancelled out.

Note that the error amplifier is assumed to be distortion free and is accepted as linear. Since the power level of the error amplifier is much smaller than that of output signal, ignoring the error amplifier distortion is reasonable.

In feedforward applications high degree of matching capability is required on both amplitude and phase. Feedforward being an open loop system, i.e. lack of a feedback loop in the system, it can not control its own performance.

2.2.4 RF SYNTHESIS

Linear amplification with nonlinear components (LINC) is an example of RF Synthesis technique in narrowband linearization.

The LINC working principle is shown in the Figure 2-16. There are two phase modulated signals generated by using two voltage controlled oscillators (VCO). Both phase modulated signals are amplified with the nonlinear components.

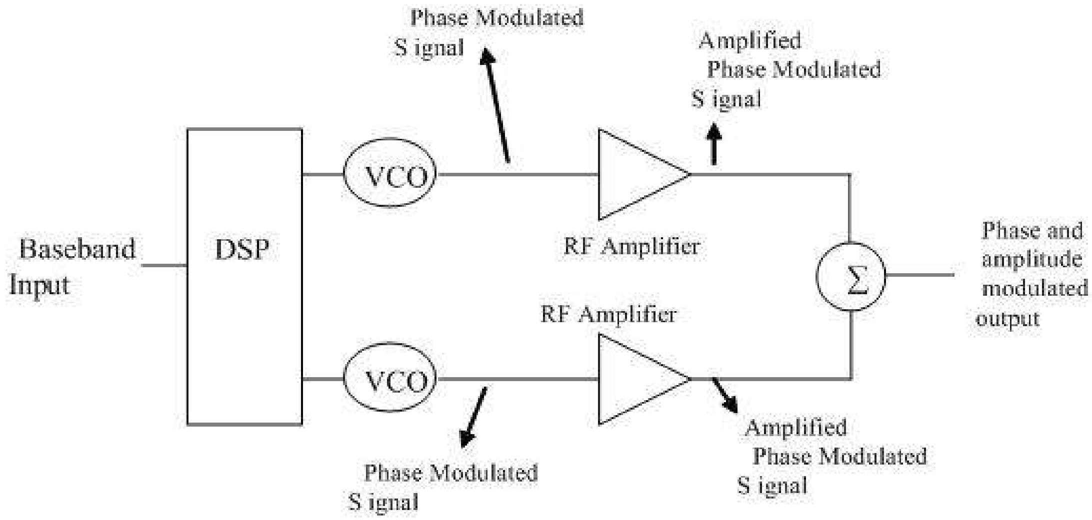

Figure 2-16 Linear Amplification with nonlinear components (LINC)

After combining the two amplified phase modulated signals the amplitude of the output is a function of the phase in the two signals.

There are some disadvantages of LINC method. It is being an open loop system, that is, there is no feedback; the two signal paths should be accurately matched. Also the vector combining may be a problem when the bandwidth of the phase modulated signal is large.

2.2.5 ENVELOPE ELIMINATION AND RESTORATION

Envelope elimination and restoration is a narrowband synthesis method which is also referred to as Kahn method. In this method the signal is split into two paths; one is going trough a limiter and the other going trough an envelope detector as shown in Figure 2-17. In the first path the output of the limiter becomes a constant envelope phase modulated signal since the limiter removes the amplitude modulation. In the second path, the output of the envelope detector is an amplitude modulated signal. Signals on the first and on the second paths are both amplified;

but with different amplifiers. The amplitude modulated signal at the second path is amplified with a low frequency amplifier while the phase modulated signal at the first path is amplified with an efficient but nonlinear RF amplifier. At the end the amplified phase modulated signal in the first path is modulated with the amplified amplitude modulated signal. The resulting output envelope wave shape will be the same as the input one.

In envelope elimination and restoration method, to have a linear output be sure that the time relationship between the amplitude modulated signal and the phase modulated signal is maintained, that is, the time delay of the first and the second paths are same.

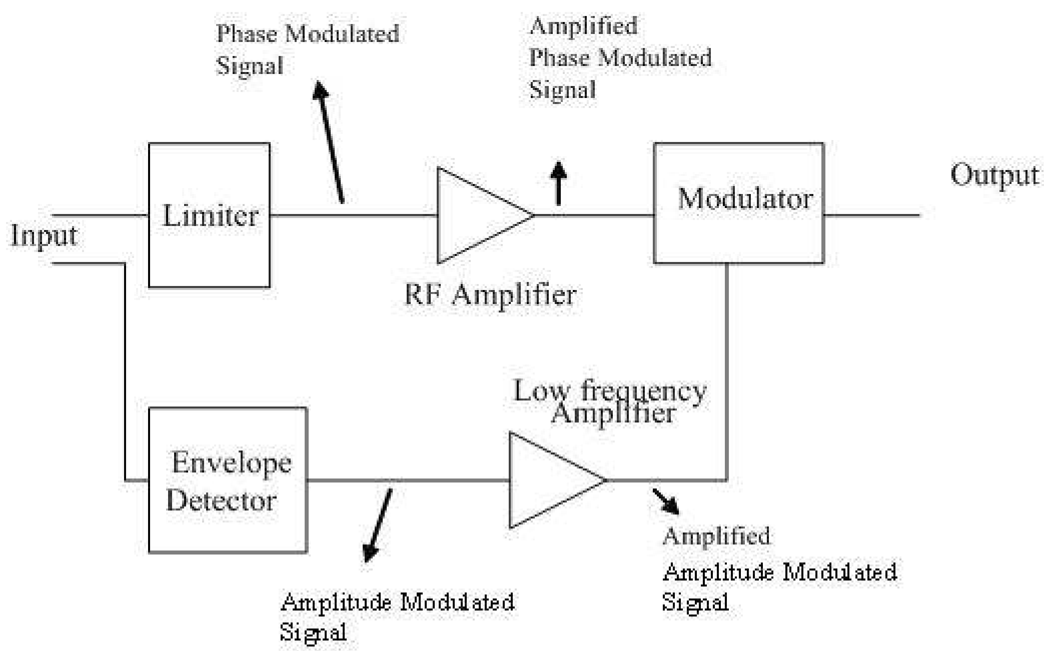

Figure 2-17 Envelope elimination and restoration

2.2.6 COMPARISON OF LINEARIZATION TECHNIQUES

Some of the advantages of feedforward as an amplifier linearization technique are detailed here [8]:

- Feedforward correction does not ideally reduce amplifier gain. This is in contrast to feedback systems in which linearity is achieved at the expense of gain.
- Gain-bandwidth is conserved within the band of interest. This is again in contrast to feedback systems which often require very wide feedback bandwidths in order to provide the required levels of correction.
- Correction is independent of the magnitude of the amplifier delays within the system. A high-gain RF amplifier will often have a significant group delay and this is potentially disastrous for any form of feedback system, due to the large potential for instability.
- Correction is not attempted based on past events, unlike feedback. The correction process is based on what is currently happening rather than what has happened in the recent past.
- The basic feedforward configuration is unconditionally stable. This is one of the most important advantages and follows from the points raised above.
- Cost is the main limiting factor to the number of stages (or loops) and hence the level of correction which may be achieved, although size and efficiency may also be important in some applications. In other words, an arbitrarily high level of correction is possible, as there is no theoretical limitation on the number of times which feedforward correction may be applied. In an ideal system, perfect correction could be achieved with just the basic system; however, in reality, the error amplifier itself will distort the error signal and this will appear directly at the output. Gain and phase matching throughout the system also affect the performance.

- The error amplifier ideally needs only to process the main amplifier distortion information and hence can be of a much lower power than the main amplifier. Thus, it is likely that a more linear and lower noise error amplifier can be constructed. This in turn will result in a lower overall system noise figure.
- Fault tolerance. In a single loop feedforward system, the failure of either amplifier will result in a degradation of performance and possibly a lowering of the final output power; however the system will not fail altogether. In the case of a feedback system there is only one forward-path amplifier, and if this fails, then the whole system has failed. If multiple feedforward loops are used, then the overall system will degrade gracefully if one or more amplifiers should fail.

Feedforward also suffers from some major disadvantages that have in the past led to its relative unpopularity when compared with feedback. These may be summarized as follows:

- Changes of device characteristics with time and temperature are not compensated. The open-loop nature of the feedforward system does not permit it to assess its own performance and correct for time variations in its system components. Thus, the performance of a basic uncompensated feedforward system can be expected to degrade with time.
- The matching between the circuit elements in both amplitude and phase must be maintained to a very high degree over the correction bandwidth of interest.
- Circuit complexity is generally greater than that of a feedback system particularly with the requirement for a second (error) amplifier. This usually results in greater size and cost.

Some of the advantages of predistortion as an amplifier linearization technique are detailed here:

- Good wideband performance can be achieved from very simple circuitry (with analogue RF predistortion).
- High levels of linearity improvement (greater than 25 dB) may be achieved over wide instantaneous bandwidths (greater than 20 MHz) with digital predistortion techniques.
- Very wide instantaneous bandwidths and operating bandwidths which are wider still (multi-octave) may be achieved with analogue RF predistortion.
- Gain bandwidth is conserved within the band of interest. This is in contrast to feedback systems that often require very wide feedback bandwidths in order to provide the required levels of correction.
- Correction is independent of the magnitude of the amplifier delays within the system. A high-gain RF amplifier will often have a significant group delay and this is potentially disastrous for any form of feedback system, due to the large potential for instability.
- Correction is not attempted based on past events, unlike feedback. The correction process is based on what is currently happening rather than what has happened in the recent past.
- An open-loop predistortion system is unconditionally stable, and even closed-loop systems are easily made stable. This is due to the effectively very narrow bandwidth of the feedback control system.

Predistortion also suffers from some disadvantages which can limit its applicability in some systems. These may be summarized as follows:

- Predistortion must, in general, take place at a low power level, as the devices and signal processing required are usually only available at such power levels. This is generally only a significant issue in booster type applications where a significant power may be available from the input signal. With most predistorters, this power must be attenuated to a low level before supplying the predistorter. This is clearly wasteful and hence potentially expensive.

- Changes of device characteristics with time and temperature are not compensated (other than in adaptive systems). The open-loop nature of the predistortion system does not permit it to assess its own performance and correct for time variations in its system components. Thus the performance of a basic (uncompensated) predistortion system can be expected to degrade with time (as with feedforward).
- The matching between the circuit elements in both amplitude and phase must be maintained to a very high degree over the correction bandwidth of interest. The levels of matching required are similar to those of feedforward, although in the case of baseband and IF predistortion systems, these must also be maintained in up-conversion and filtering stages, which can be difficult. Alternatively, digital filtering may be applied to counter the effects of these issues, in a digital baseband predistortion system.

CHAPTER 3

PREDISTORTION USING SELF CANCELATION

The predistortion type linearizer using self cancellation scheme given in [15] will be examined in more detail and the simulation results will be given in this chapter. A similar application is also given in [16].

3.1 THEORY

The main idea in this technique is determining the predistortion signal by using the same amplifier with the main amplifier. The basic scheme is presented in Figure 3-1. This is why the technique is called self cancellation. Using the same amplifier in the predistortion path enables the designer to make a very good prediction. The main advantage of the method is its performance stabilization over the environmental circumstances such as thermal effects. Since these effects can be predicted very well in the predistortion path, they can be avoided.

This method is a predistortion method, but it is similar to the feedback and feedforward techniques. The predistorter contains the main amplifier. The system is an open loop system but it has some of the advantage of closed loop systems.

There are two different approaches in RF power amplifier linearization. One is the “canceling the intermodulation distortion components” and the other is “predicting the intermodulation distortion components” [15]. First approach is the main idea of the feedback and feedforward methods, second approach is the main idea of the predistortion method. All of these methods are described in Chapter 2.

In feedback and feedforward, the intermodulation distortion signal generated from the output of the RF power amplifier is used to cancel its own intermodulation distortion components [15]. In the self cancellation scheme again the intermodulation distortion signal generated from the output of the RF power amplifier is used, but to generate the correct predistortion signal.

In predistortion, the RF power amplifier is driven with the input signal combined with the inverse of the predicted distortion. [15]. In the self cancellation scheme the same theory holds. The distortion prediction is done by the use of the RF power amplifier itself.

The offered self cancellation scheme combines the advantages of the two distinct approaches and it is expected that the simplicity of the predistortion and the linearity performance will be obtained in the design.

The working principle is shown in Figure 3-2.

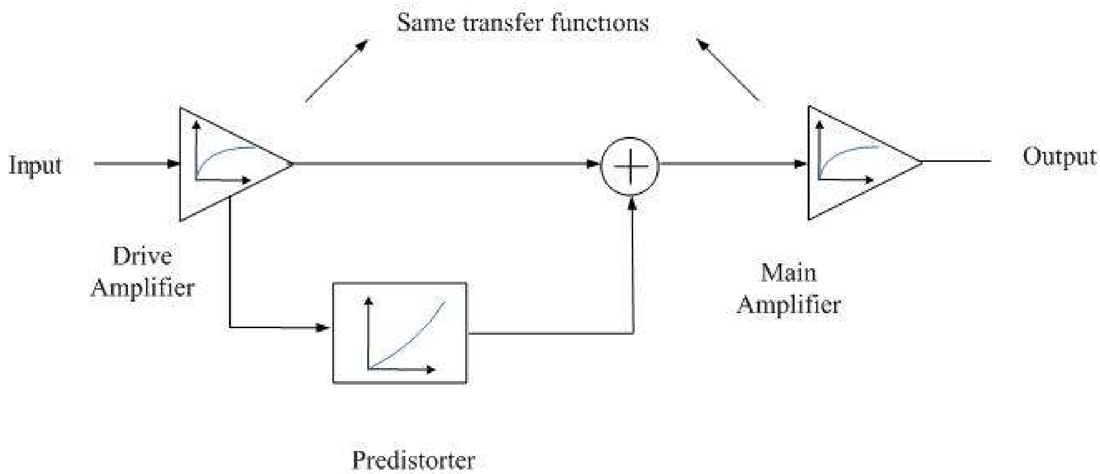

Figure 3-1 Basic schematic of self cancellation

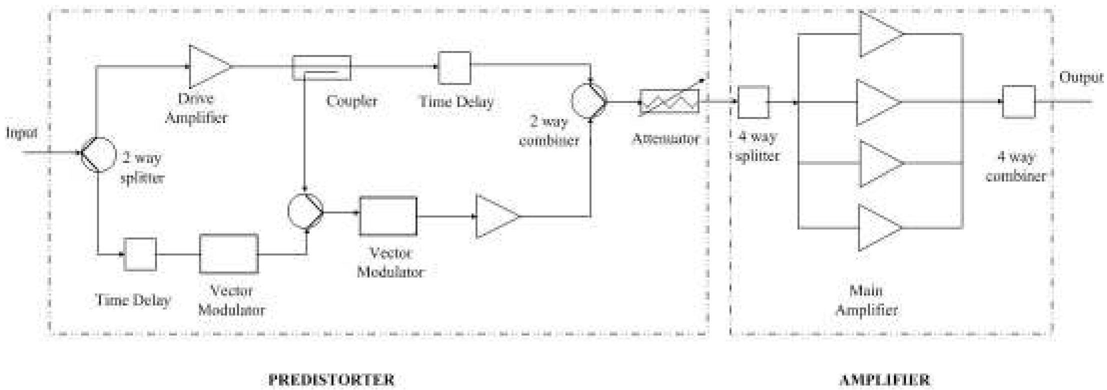

Figure 3-2 Detailed self cancellation scheme

As stated above the same amplifier is used for both predistortion and main amplifier design. The nonlinear characteristics of the predistortion part will cancel the distortion of the main amplifier. Since the amplifiers are same, they should be driven with the same input power level in order to have same nonlinear characteristics. So an optimization is needed for adjusting the main amplifier input; this is the reason of the attenuator prior to the main amplifier.

The input signal is divided into two; forming the main path and the prediction path. The signal on the main path is amplified with the drive amplifier and the intermodulation distortion components are generated. Then a sample of the amplified signal is taken and subtracted from the signal in the prediction path in order to have distortion components only. The amplitude and phase of the two signals are adjusted by the vector modulator in order to cancel the fundamental signal and have only the distortion components. The phase difference between the two signals should be 180 degrees to be able to use the power combiner as a subtractor. To do this, the phase delay of the amplifier should be considered.

The distortion of the amplifier is found; so the next step is producing the inverse characteristics of the distortion. By the use of vector modulator the phase and amplitude of the prediction path is adjusted to have the inverse of the distortion and it is added to the main signal. The phase difference between the distortion components of the two paths should be 180 degrees and the amplitude of the distortion component in the prediction path should be twice the amplitude of distortion component in the main path. The delay components are due to compensation of the drive amplifier and error amplifier delays.

The predistortion block is shown in Figure 3-3 in detail, including the fundamental and third order intermodulation distortion components.

Input power is adjusted to make the drive amplifier input power level as **α**, that is

$$Pin = \boldsymbol{\alpha}_{\ dBm} + 10\log(2) \qquad (5)$$

if we assume the power splitter divide the input power equally. Signal is amplified and the output power level of the drive amplifier is "**α**+**G**" at the fundamental frequency with intermodulation distortion components at sidebands. Note that the third order intermodulation distortion power level is evaluated as;

$$P_{IMD3} = 3\ Pout - 2\ IP3 \qquad (6)$$

Then, by neglecting the higher order terms, the third order intermodulation distortion component power level of an amplifier with a gain of **G** and a third order output intercept point of **A** is found as;

$$P_{\ IMD3} = 3\ (\boldsymbol{\alpha}+\mathbf{G}) - 2\mathbf{A} \qquad (7)$$

where P_{out} is the output power level of the amplifier and IP3 is the third order output intercept point of the amplifier.

Note that, at the P_{1dB} point of amplifier, gain is reduced by 1dB, $G_{1dB} = G - 1$.

The signal on the prediction path is subtracted from the sample of the amplified signal on the main path to have distortion only component of the amplified signal.

The main signal levels of both signals should be same, i.e. the loss due to coupling should compensate the gain of the amplifier and the loss of the time delay element.

The intermodulation distortion component of the signal is found and the predistortion that will be applied to the signal must be determined now. The idea is adding the inverse of the distortion component to the original signal. The predistortion signal will be added to the amplified signal. But, the predistortion signal should also compensate the distortion of the first drive amplifier so twice of the distortion should be added.

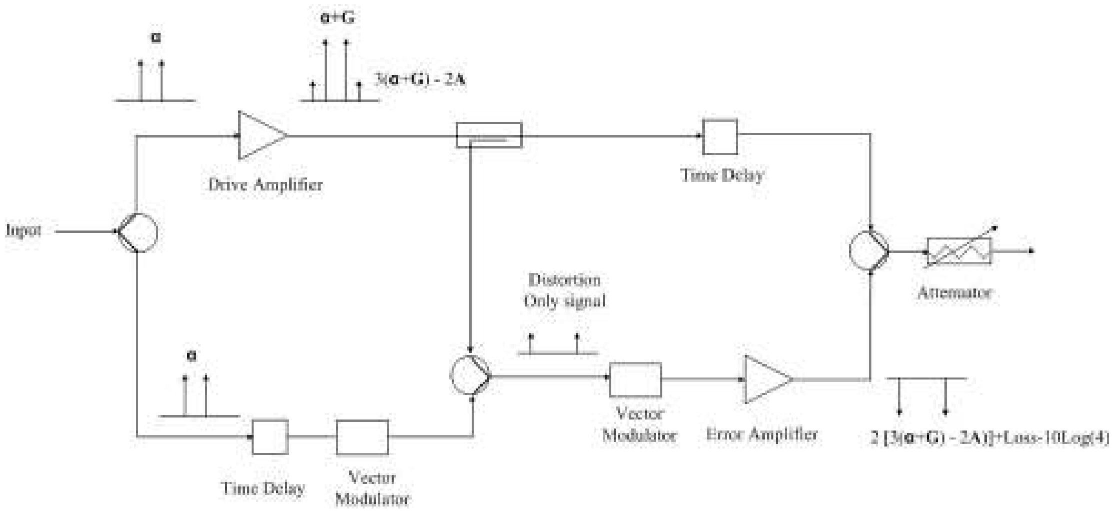

Figure 3-3 Predistorter block of self cancellation scheme

G: Gain of the drive amplifier and A is the third order output intercept point

Finally the predistorted signal at input of the attenuator in Figure 3-3 have distortion components with the same power ratio to the fundamental signal that have introduced at the output of the first drive amplifier, but with a 180 degree phase shift. To achieve this, the error amplifier is adjusted to have distortion components at power level;

$$2\,(3\,(\boldsymbol{\alpha}+\mathbf{G}) - 2\mathbf{A}) - L - 10\log(4) \qquad (8)$$

where L is due to loss of the coupler and time delay element.

The term (3 (**α**+**G**) – 2**A**) is the distortion introduced in the amplification process, and is considered twice, one for the drive amplifier in the predistorter block and the other for the main amplifier block. The 10log (4) component is due to four drive amplifiers in the main amplifier block.

Finally the attenuator should be adjusted to drive the main amplifier block such that the drive amplifiers in the main amplifier block are driven with the same input power level with the drive amplifier in the predistorter block in order to have the same nonlinear characteristics. The loss introduced in the attenuator should be adjusted to compensate the drive amplifier in the predistorter together with the loss of the coupler, time delay element, power splitter in the predistorter and the power splitter in the main amplifier. Then the four drive amplifiers are all driven with the same power level with the predistorter drive amplifier that is **α.** The attenuator loss should satisfy the following equation.

$$-L_{att} + 10\log(4) = G + L_C + L_{TD} + L_{PS\_pre} + L_{PS\_Amp} \qquad (9)$$

where L_{att} is the attenuator loss, L_C is the coupler loss, L_{TD} is the time delay element loss, L_{PS_pre} is the power splitter loss at in the predistorter circuit and L_{PS_Amp} is the power splitter loss at in the main amplifier circuit.

The self cancellation predistortion scheme offered here is said to be realized and the result are reported [15].

"Five identical class AB biased (I_{DQ} = 200 mA) MRF21030 LDMOS transistors, provided by the Freescale semiconductor, Inc., are used to design five drive amplifier blocks. The WCDMA test model I 64 DPCH signal with the test frequency of 2.15 GHz is used for the characterization of the fabricated LPA system. The targeted adjacent-channel-leakage-ratio (ACLR) specification is 50 dB, and the output power level is maintained in the range from 30 to 38 dBm."

3.2 SIMULATION WITH IDEAL COMPONENTS

The simulation of the self cancellation scheme is demonstrated in Advance Design System software (ADS) with ideal components. The two tone harmonic balance analysis is performed at a frequency of 1 GHz. The input power level of the drive amplifiers used in the design was 10 dBm. A system amplifier is defined to be used as the drive amplifier. The gain of the drive amplifier is defined as 20 dB and the third order intercept point as 50 dB.

The gain of the amplifier should be chosen greater than 12 dB to be able to drive the main amplifier with the same power level with the drive amplifier. There are four power dividers on the main path that the drive amplifier introduces, all of which have $S_{21}=S_{31}=0.707$ that is 3 dB in power level. In practice, these power dividers will be lossy and the design will only be available for the gains greater than that of 16 dB assuming the losses are about 1 dB. For the amplifiers with smaller gain a buffer amplifier can be used to drive the main amplifier as reported in [16]. Then the adjustment for driving the main amplifier with the same power level with the drive amplifier is done by the use of buffer amplifier.

The simulation configuration is set into ADS similar to the Figure 3-2. In ADS design, the defined system amplifier (gain of the amplifier, G=20 dB and third order output intercept point (IP3) A=50 dB) is used as the drive amplifier, the two way power splitter is chosen as a lossless equal power divider with;

$$S_{21}= S_{31}=0.707 \tag{10}$$

Instead of the coupler in Figure 3-2, a power divider and an attenuator is used. The attenuator is adjusted so that the signal level of the distorted signal and the undistorted signal becomes equal which will enable us to have distortion only signal. The time delay element and the vector modulator in Figure 3-2 is replaced by phase shifters. The phase shifters may be replaced by appropriate transmission lines or inductors in realization.

Let's examine Figure 3-4 to make the discussion more clear. The points A to H are determined for referencing the signals at those points. The frequency domain spectrums will be given at these points.

Basically at point A only the main signal is present. The amplified signal at point B contains both the main signal and the distortion components, which is the third order distortion components in our case. The phase of the signal at point C will be adjusted by the phase shifter and the amplitude signal at point D will be adjusted by the attenuator so that the main signal at point E will cancel out and only the distortion component will remain. Finally, the error amplifier will be adjusted to cancel the distortion at point H and the predistortion component is produced.

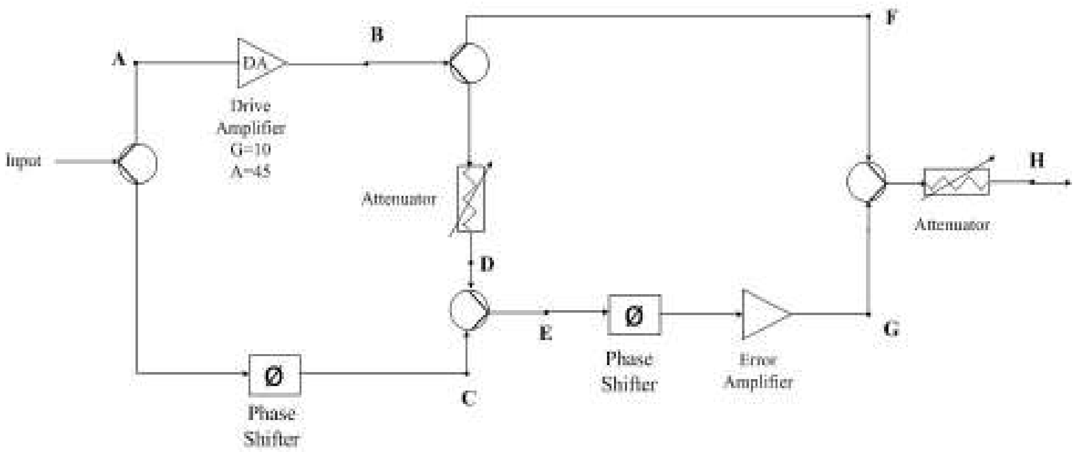

Figure 3-4 Self cancellation scheme with reference points

The signal spectrum at point A, the input of the drive amplifier, is given in Figure 3-5. The signal level is 10 dBm (α) with an ACPR of 10 – (–90) = 100 dBm.

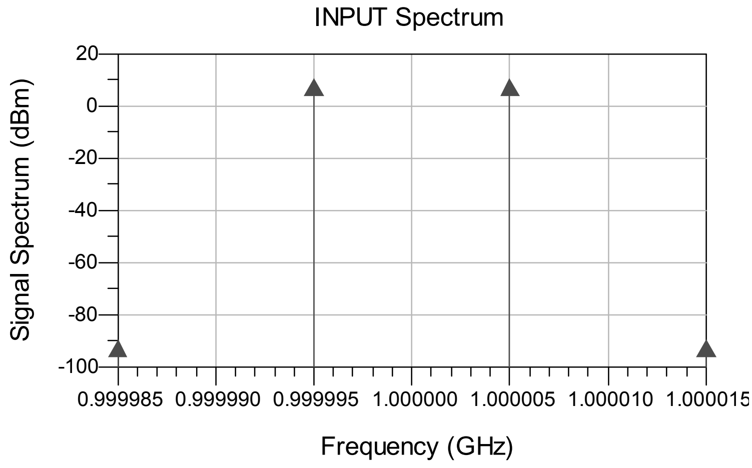

Figure 3-5 Input Spectrum at point A with ideal system components

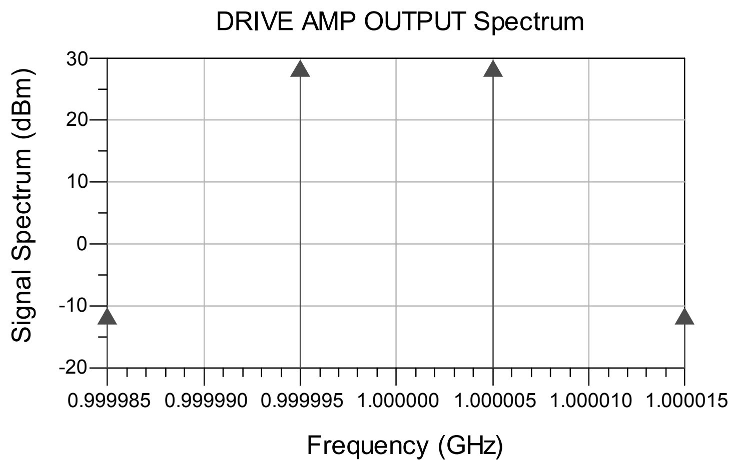

Figure 3-6 Drive Amplifier Output Spectrum at point B with ideal system components

The signal spectrum at point B, the output of the drive amplifier, is given in Figure 3-6. The signal level is 30 dBm (α+G) and the 3rd order intermodulation distortion (IMD3) component level is -10 dBm (3 ($\boldsymbol{\alpha}$+**G**) – 2**A**); where α=10, G=20 and A=50.

Then the ACPR is 30 – (–10) = 40 dBm. The signal spectrum at point C is same with point A but with a phase difference of 180 degrees. Since the drive amplifier is a system amplifier it does not have a delay so the input and output of the amplifier are with the same phase. In fact, the phase difference of the signals at points C and D should be 180 degrees.

The signal level at point D is same with point A, that is the amplification of the drive amplifier is canceled out by the power divider and the attenuator. At point D, while the main signal component is same as at point A, there is an IMD component. This IMD component is attenuated by the power divider and by the attenuator; total of which is equal to the gain of the drive amplifier.

The signal spectrum at point E, the spectrum of the IMD component is given in Figure 3-7. The main signal is canceled out and the IMD only components are left, of course with an attenuation of 3 dB due to the power combiner. The IMD component at point B was at -10 dBm. At point D the IMD component being attenuated by the power divider and the attenuator became at -30 dBm. Finally at point E, passing trough power combiner IMD component is at -33 dBm.

The signal spectrum at point F is 3 dB attenuated version of the signal spectrum at point B. That is the main signal of 17 dBm and IMD component of -13 dBm.

The signal spectrum at point G, the amplified version of the IMD component is given in Figure 3-8. The IMD component is 7 dBm, that is the error amplifier gain is 40 dB which amplifies the signal from -33 dBm to 7 dBm. The phase difference of the IMD components at point F and point G is adjusted to be 180 degree out of phase. Since the error amplifier is a system amplifier and does not have a delay, 180 degree phase shifter before or after the error amplifier will satisfy this phase difference. The IMD component at point G, being 180 degree phase shifted with point F, cancels out the distortion component at point F and it will add a predistortion component to the signal at point H.

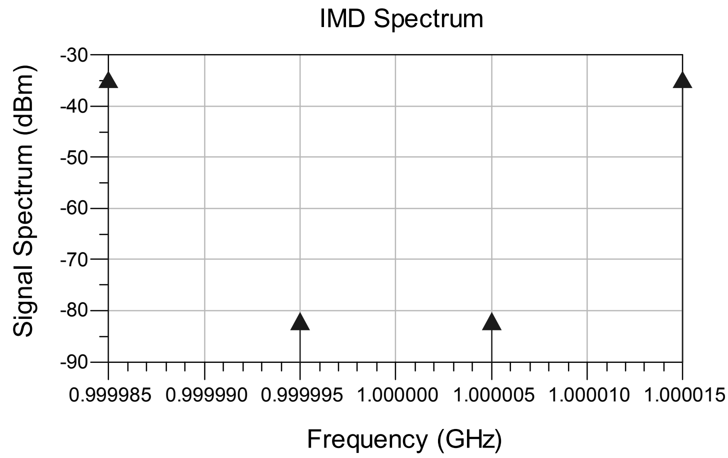

Figure 3-7 Intermodulation Distortion Signal Spectrum at point E with ideal system components

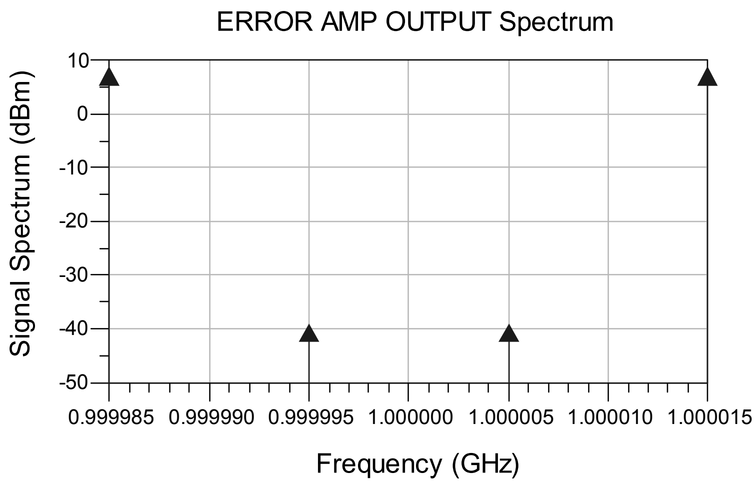

Figure 3-8 Error Amplifier Output Spectrum with ideal system components

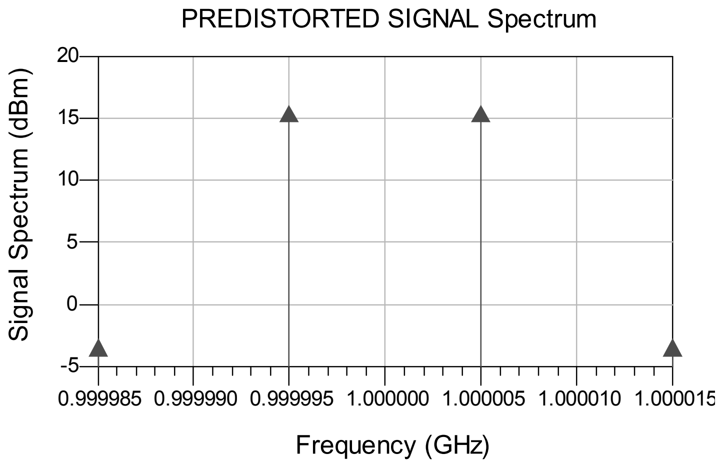

Figure 3-9 Predistorted Signal Spectrum at point H with ideal system components

The signal spectrum at point H, the predistorted signal spectrum is given in Figure 3-9. The main signal is at 16 dBm. It will be attenuated by 6 dB and result in a power level of 10 dBm at the input of the amplifiers.

The amplifiers in the main amplifier block will be driven with the same power level with the drive amplifier. The IMD component at point H is at -2.8 dBm. It will also be attenuated by 6 dB and then amplified by 20 dB, that is 14 dB amplification in total. And it is expected that the IMD components will cancel out at the output of the main amplifier.

The signal spectrum at the output of the main amplifier is given in Figure 3-10. The IMD components are at -58 dBm and -46 dBm after cancellation. The main signal s 34 dBm which gives an ACPR of 80 dB. The evaluation of ACPR is done by a simple subtraction of the main signal and distortion components in dBm. There are two different values for the ratio of the main signal to the distortion component on the high frequency and on the low frequency and the worse value is taken as the ACPR of the amplifier.

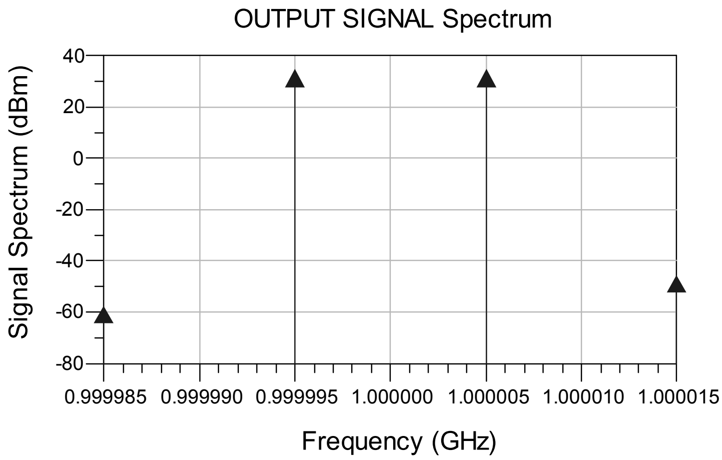

Figure 3-10 Output Signal Spectrum with ideal system components

The ACPR of the drive amplifier was 40 dB and the ACPR of the overall amplifier is 80 dB. A 40 dB improvement is achieved by the use of self cancellation scheme with ideal components.

The system amplifier and ideal system components use in the predistortion process resulted in an improvement of 40 dB in ACPR of the amplifier. However, the use of real components and a real amplifier will not give such a good improvement. The real components will have additional nonlinear characteristics and there will be unavoidable reflection that will decrease the overall performance. The amplifier will have an asymmetric behavior and since the offered predistortion method can only be optimized for one sideband, that is the worse one, the performance of the overall system will be restricted by the other sideband.

CHAPTER 4

MMIC IMPLEMENTATION OF SELF CANCELATION SCHEME

The method has introduced an improvement of 40 dB in ACPR of the amplifier designed with system components. But what is the case if real components are used. Can we achieve the same performance in real world? Probably no. For verification purposes MMIC technology is used for millimeter wave power amplifier design.

Using OMMIC foundry component library an MMIC amplifier is designed. The center frequency is 3 GHz. The aimed output power is 20 dBm. This level is set only for demonstration purposes. The same method can be used in higher output power.

The linearization performance is expected to be less than the performance of the amplifier with system components.

4.1 AMPLIFIER TO BE LINEARIZED

A Class-A amplifier is designed to be operated at the frequency of 35 GHz. The input and output matching are designed to be 50 ohm. Amplifier design and the demonstration is done with the use of design tool ADS.

In order to improve the gain of the amplifier two transistors used repeatedly. The transistors and the other components are taken from the OMMIC ED02AH TechInclude library.

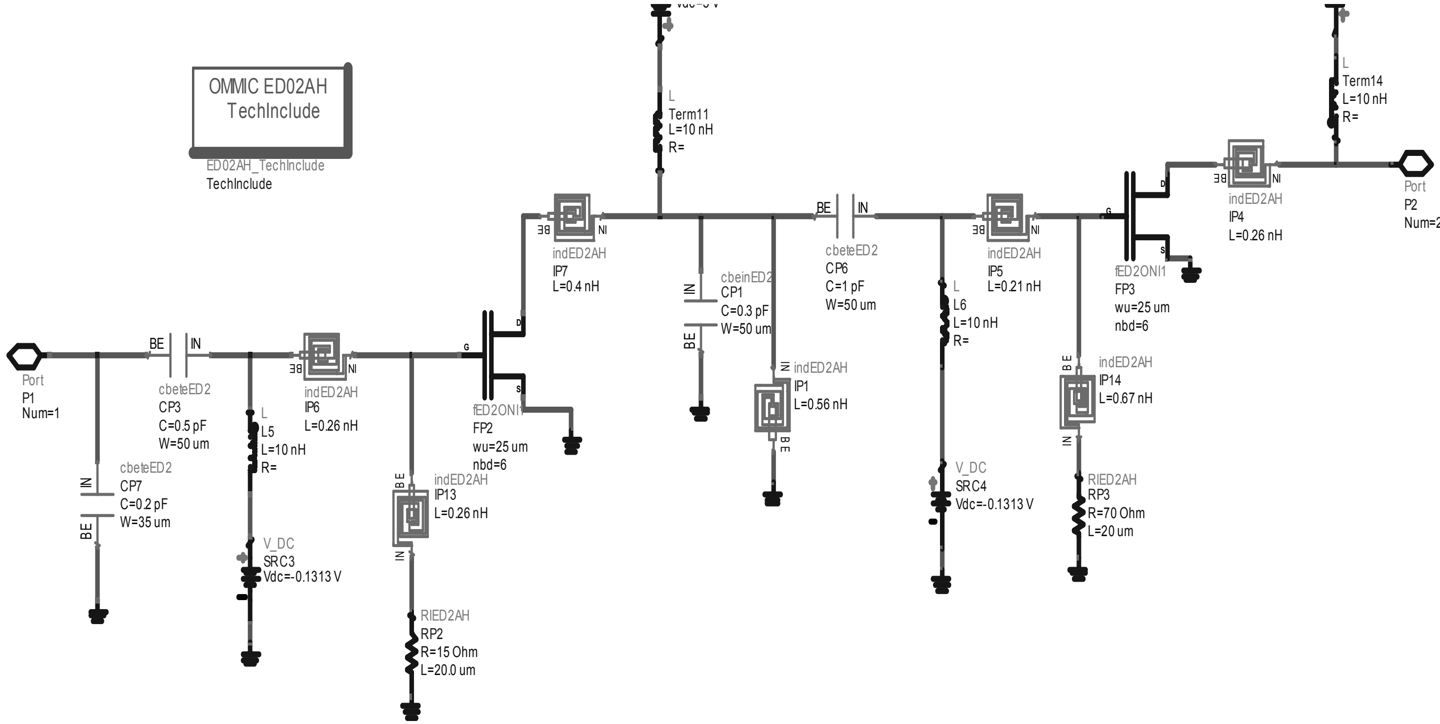

Figure 4-1 Drive Amplifier Design

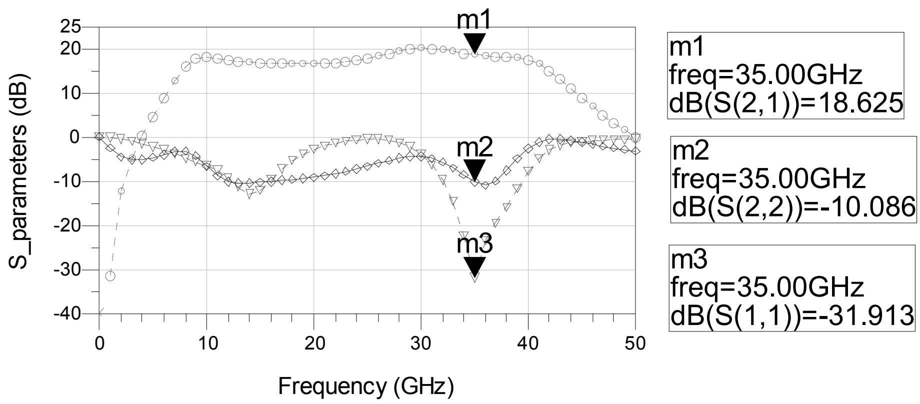

Figure 4-2 S Parameters of the Drive Amplifier 1 GHz – 50 GHz

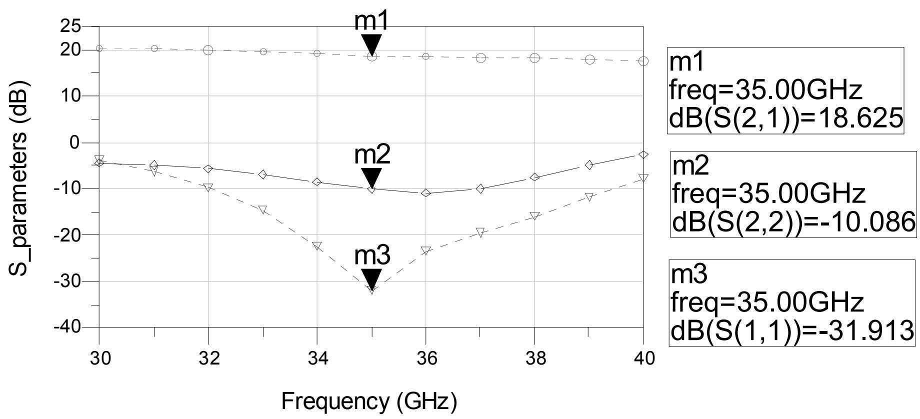

Figure 4-3 S Parameters of the Drive Amplifier 30 GHz – 40 GHz

The amplifier circuit that is designed is given in Figure 4-1 as the schematic view in ADS.

The S parameter simulation is done by use of ADS and the result is given in Figure 4-2. There is no unstability introduced in the frequency range up to 50 GHz. The S_{11} and S_{22} parameters are zero for all frequencies up to 50 GHz.

The gain of the amplifier is the main parameter that affects the linearization circuit design. The parameters, S_{11} and S_{22} will introduce reflections and the big values of these parameters will make the tuning process difficult. The input matching, S_{11}, is more important in the design, so the amplifier design parameters are adjusted to have a lower S_{11} value.

The S parameter simulation response in the range, 30GHZ - 40 GHz is given in Figure 4-3. The gain of the amplifier is 18.6 dB (the S_{21} parameter in the figure) at 35 GHz. The S_{11} and S_{22} parameters are -31.9 dB and -10 dB respectively.

The input matching of the amplifier is very good; resulting a -31 dB S_{11} and the output matching of the amplifier is acceptable.

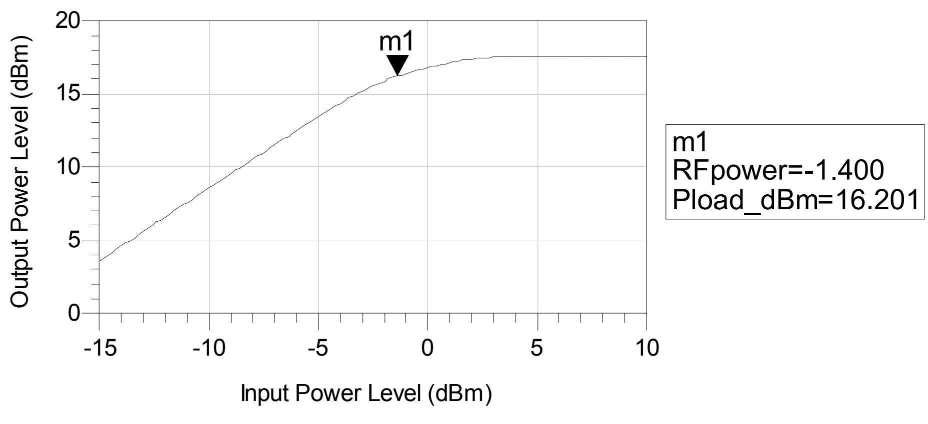

Figure 4-4 Amplifier input-output relationship

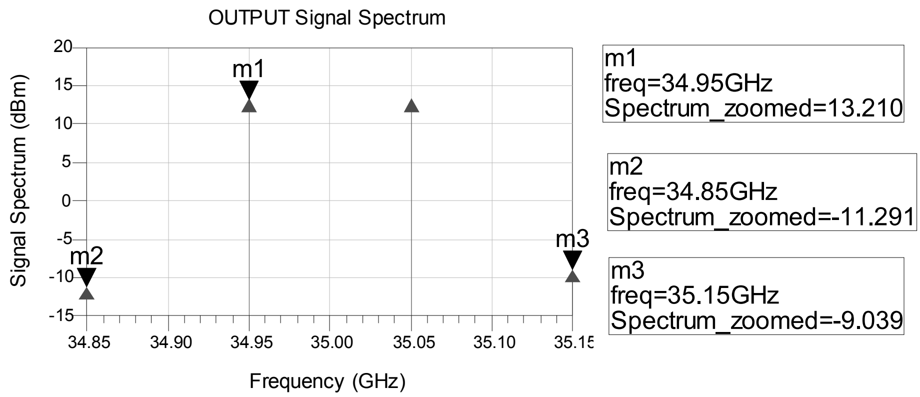

Figure 4-5 IMD components of the Drive Amplifier

After the S_Parameter analysis of the amplifier design, the two tone harmonic balance simulation is performed. The results of the two tone harmonic balance simulation are given below.

The input-output characteristic of the amplifier is given in Figure 4-4 which is obtained by the use of two tone harmonic balance simulation in ADS.

The 1 dB compression point of an amplifier is defined as the point at which the output power of the amplifier is dropped 1 dB below the output of a linear amplifier at the same input level. Since the gain of the amplifier is 18.6 dB, the 1 dB compression point here is the point at which the gain is 17.6 dB. Examining Figure 4-4 we see that the 1 dB compression point of the amplifier is 16.2 dBm corresponding to the input level of -1.4 dBm.

The amplifier is driven at the 1 dB compression point, i.e. the input power level is at -1.4 dBm and the two tone harmonic balance simulation is repeated. The result of the simulation is given in Figure 4-5. The main signal components and the IMD components of the amplifier can be seen in the figure.

The main signal level is 13.21 dBm and the IMD components are at -11.29 dBm and -9.04 dBm. The ratio of the main signal to the adjacent channel IMD component can be calculated as 13.21- (- 11.29) = 24.5 dB for lower frequency band and 13.21- (- 9.04) = 22.25 dB for higher frequency band. The evaluation of ACPR is done by a simple subtraction of the main signal and distortion components in dBm.

There are two different values for the ratio of the main signal to the distortion component on the high frequency and on the low frequency and the worse value is taken as the ACPR of the amplifier.

4.2 AMPLIFIER BLOCK WITH 4 PARALLEL DRIVE AMPLIFIER

The amplifier performance can be improved by the parallel combination of the same amplifier. Let us examine the performance of 4 parallel amplifiers. In parallel combination of the amplifiers; input signal is divided into four way and all of the four signals are amplified separately. Then the amplified signals are combined again. The characteristics of the power splitter and the power combiner used in the design are important that they affect the gain and linearity of the overall amplifier.

Firstly; Let us examine the performance with ideal power splitter with a perfect isolation. In ADS the system component two way power splitter with S_{21}=S_{31}=0.707 and 100 dB isolation is used. The schematic is given in Figure 4-6.

The combination of four amplifiers with ideal power splitters is expected to have the same gain, the same ratio of the main signal to the IMD components, but a 6 dB improvement of the 1 dB compression point.

Since there is no loss, the gain of the amplifier is same, 18.6 dB. The output power spectrum versus input power is given in Figure 4-7.

It is obvious from Figure 4-7 that the 1 dB compression point is 22.2 dB corresponding to a gain of 17.6, i.e. 1 dB less than the gain of the amplifier block. The 1 dB compression point of the 4 drive amplifier block is 6 dB more than the drive amplifier itself.

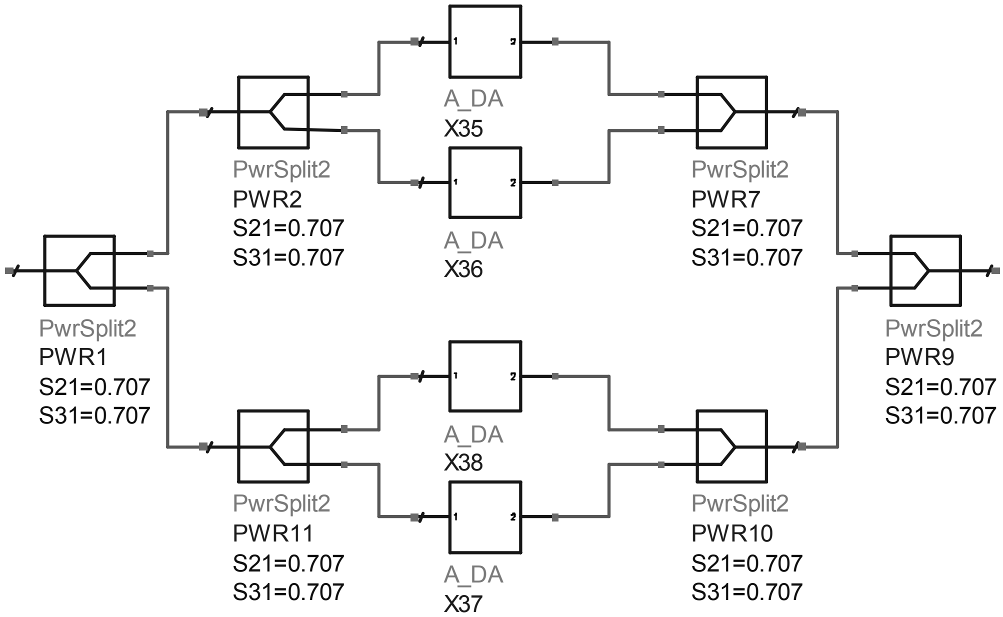

Figure 4-6 Parallel Combination of 4 Drive Amplifiers with ideal Power Splitters

To see the ratio of the main signal to the IMD components, the two tone harmonic balance simulation is repeated and the result is given in Figure 4-8.

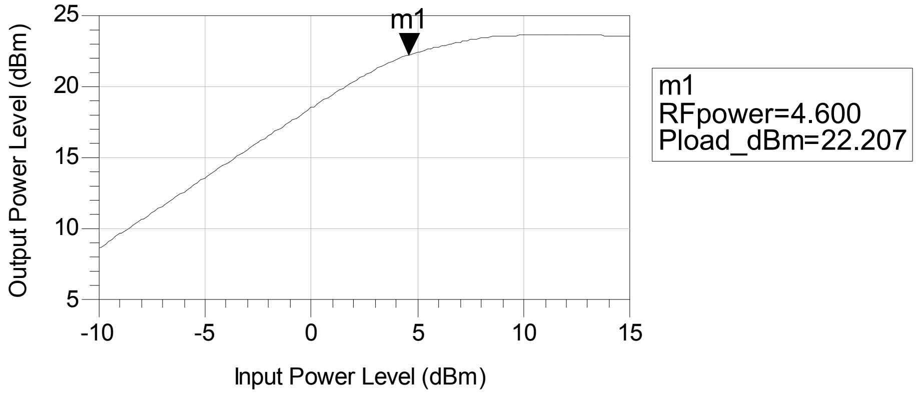

Figure 4-7 Input_Output Power Relation 4 Amplifier with ideal Power Splitters

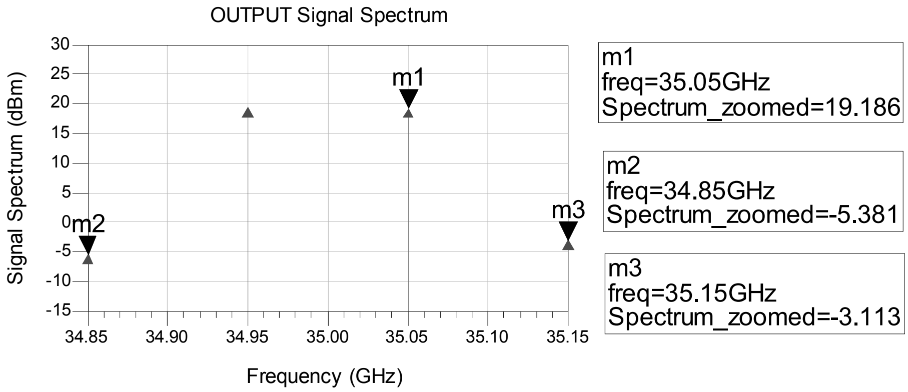

Figure 4-8 IMD components of the 4 Amplifier Block with ideal Power Splitters

The power ratio of the main signal to the adjacent channel IMD component is 19.18– (-5.38) = 24.56 dB for lower frequency band and 19.18– (-3.11) = 22.29 dB for higher frequency band. There is no change in the ratio of the main signal to the IMD components.

We see that the parallel combination of four amplifiers has the same gain and the same ratio of main signal to the IMD components with the amplifier itself. But an improvement of 6 dB in 1 dB compression point is obtained.

However, this is the case with ideal, lossless and perfectly isolated power splitter and power combiners. What if the power splitter is a lossy one?

4.3 POWER DIVIDER/COMBINER DESIGN

To see the performance of the 4 amplifier block with a real power divider/combiner a power combiner is designed. For the power divider design the power combiner design for 9-10 GHz in [16] is modified for 35 GHz.

The power combiner is designed to be symmetric and low loss. To have a better performance the power combiner is designed by a two stage structure. The design schematic is given in Figure 4-9.

The S_Parameter simulation of the power combiner is given in Figure 4-10. The isolation of the combiner is about 17 dB at 35 GHz. The S_Parameter values at 35 GHz are:

$$S_{21} = S_{31} = -3.155 \text{ dB} \quad (11)$$

$$S_{22} = S_{33} = -17 \text{ dB} \quad (12)$$

$$S_{11} = -30 \text{ dB} \quad (13)$$

At first the power combiner is used for both as the power combiner and as the power divider.

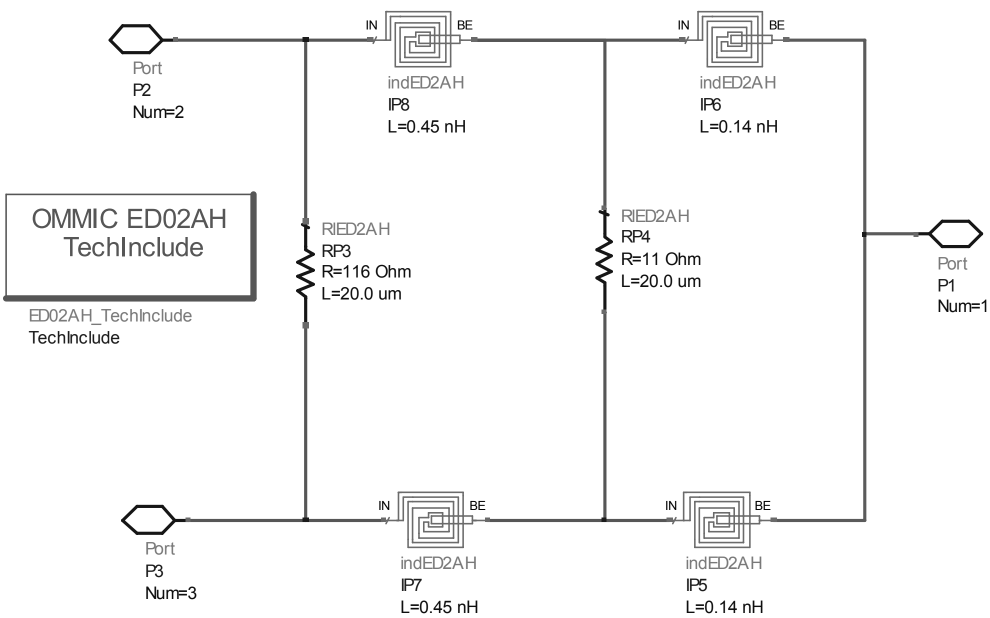

Figure 4-9 Power Combiner Design

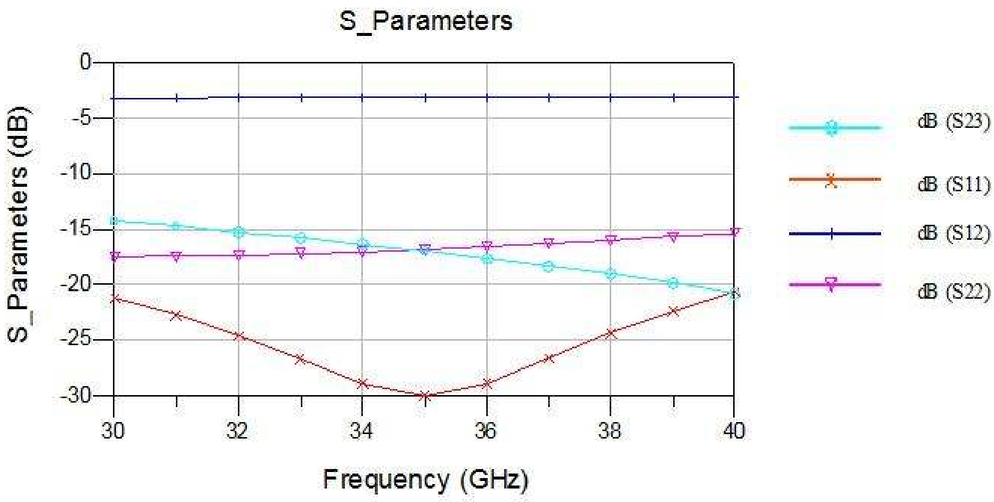

Figure 4-10 Power Combiner S_Parameter

The parallel combination of the four amplifiers with the designed power combiner as both power combiner and power divider is expected to have a less gain since there will be loss due to the power combiner.

The S_{12} parameter of the combiner is -3.155 dB, meaning loss of 0.155 dB. And since the signal passes trough the combiner four times, two in power divider case and two in power combiner case, the total loss 0.155 times 4 giving 0.62 dB. Then the gain of the amplifier block with designed power combiner is expected to be 0.62 dB less than the amplifier itself, i.e. about 18 dB.

The 1 dB compression point and the ratio of the main signal to the IMD components are expected to be same as the amplifier block with ideal power splitter.

The two tone harmonic balance simulation results are given in Figure 4-11 and Figure 4-12. From Figure 4-11 we see that the gain of the amplifier is decreased by 0.6 dB as expected. The amplifier block gain is 18 dB while the amplifier gain was 18.6 dB.

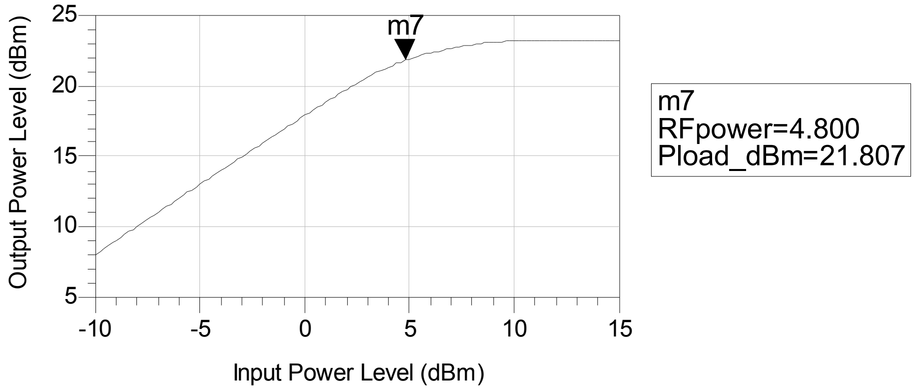

Figure 4-11 Input_Output Power Relation 4 Amplifier Block with Designed Power Combiner

Due to the decrease in the gain of the amplifier block the 1 dB compression point is also decreased to 21.8 dBm which was 22.2 dBm. In all cases the drive amplifiers are driven with the same power level, 1.4 dBm which is the 1 dB compression point of the amplifier. So due to the loss of the power combiner the input power level is increased in order to be able to drive the drive amplifier at its 1 dB compression point. In fact the drive amplifier 1 dB compression point does not change, but the 1 dB compression point decreases due to the loss introduced on the power combiners..

The decrease of 0.6 dB in gain causes a 0.4 dB decrease in 1 dB compression point with a 0.2 increase in the corresponding input power level.

The power ratio of the main signal to the adjacent channel IMD component is 18.89– (-5.04) = 23.93 dB for lower frequency band and 18.89– (-12.97) = 31.86 dB for higher frequency band. There is a slight decrease (0.5 dB) in the lower frequency band..

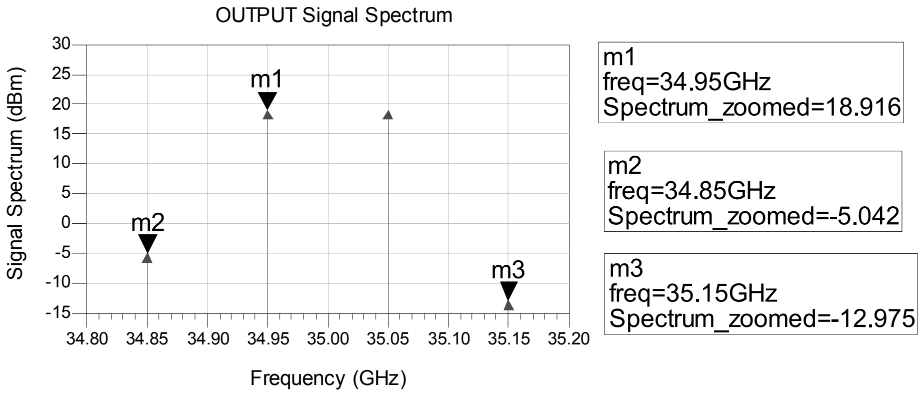

Figure 4-12 IMD components of the 4 Amplifier Block with Designed Power Combiner

However there is a big decrease in the higher frequency band. There are two different values for the ratio of the main signal to the distortion component on the high frequency and on the low frequency and the worse value is taken as the ACPR of the amplifier, i.e. 23.93 dB due to the characteristic of the power combiner, the higher frequency IMD component is decreased by 9.5 dB. The nonlinear characteristic of the power combiner and the reflections resulted in a cancellation of the higher frequency band IMD component.

As a result for the parallel combination of the four drive amplifiers with the use of power combiner designed, as both combiner and divider; there is a 0.6 dB decrease in gain, 0.4 dB decrease in 1 dB compression point, 0.5 dB decrease in lower frequency band IMD component and 9.5 dB decrease in higher frequency band IMD component.

4.4 PREDISTORTION APPLICATION AT 35 GHZ

The amplifier to be linearized is a Class-A amplifier with 18.6 dB gain. In the previous chapter it is said that the gain of the drive amplifier should be greater than 12 dB in this method. Otherwise, an additional buffer amplifier should be used.

Remembering the theory of self cancellation scheme; the same amplifier is used in both predistortion and main amplifier design. Hence, they should be driven with the same input power level in order to have the same nonlinear characteristics. So an optimization is needed for adjusting the main amplifier input. If the drive amplifier used has a gain less than 12 dB, then this optimization cannot be done unless an additional buffer amplifier is used. In our case we used an amplifier with 18.6 dB gain and we don't need a buffer amplifier.

The input power level of the system is chosen as 2 dBm. Hence there is a balanced power divider before the drive amplifier, introducing about 3 dB loss; the input power level of the drive amplifier is about -1 dBm. This power level is near to the 1 dB compression point of the drive amplifier.

The amplifier design contains non ideal components, and so there are reflections occurring due to amplifier and making the design difficult. So at first the amplifier is used with an isolator before it and the simulation is performed. The components are tuned for best performance.

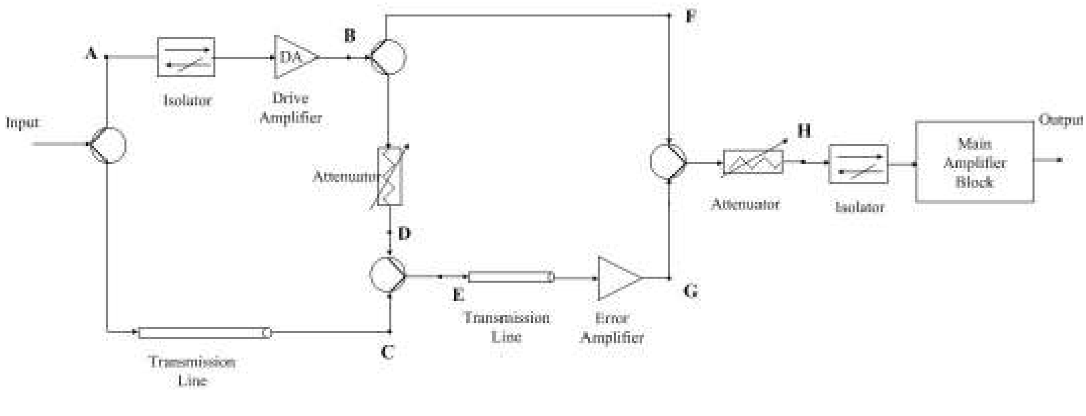

Figure 4-13 Preliminary MMIC implementation structure at 35 GHz

The configuration is very similar to the one in Figure 3-4. The drive amplifier is replaced with the amplifier designed and an isolator before it. The phase shifters are replaced with appropriate transmission lines. The preliminary configuration is given in Figure 4-13.

Finally the isolators are removed and the simulation is re-performed. Then the components are tuned without the isolators for better performance.

After these steps, the ideal components are replaced by the real ones one by one and the overall design is completed. All of the replacements of ideal components with real ones are done one by one and the remaining of the design components are tuned again.

The time delay elements are replaced by transmission lines with appropriate length. Then the length of the transmission lines, being so long, makes the use of transmission line difficult. And so, in the final design, they are replaced by inductors.

The attenuators can be replaced by TEE or PI structured resistances and in this case PI structure is preferred for both attenuators. The attenuators at points D and H are given in Figure 4-14 and Figure 4-15 respectively. After the replacement of the attenuator, the inductors need to be tuned again since the real attenuators change the phase of the signals.

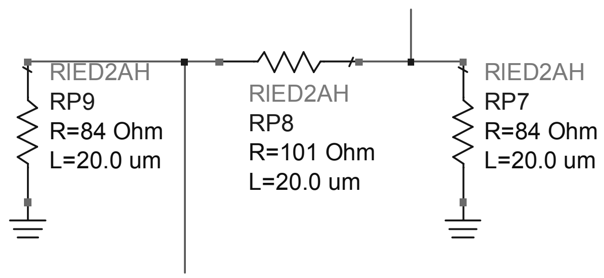

Figure 4-14 Attenuator at point D

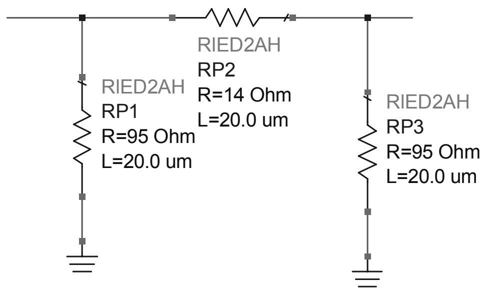

Figure 4-15 Attenuator before the Main Amplifier Block

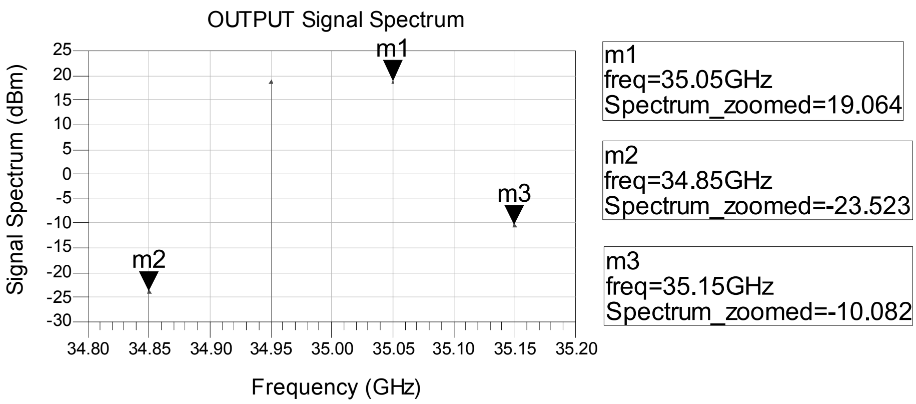

Figure 4-16 Predistortion Application

In the previous section, the combination of four parallel drive amplifiers is demonstrated and the simulation results are given in Figure 4-11 and Figure 4-12. The ratio of the main signal to the IMD component was 23.93 dB in the lower frequency band and 31.86 dB in the higher frequency band.

The 4 parallel drive amplifiers giving this response is used as the main amplifier block and the predistortion technique offered in Chapter 3 is applied to this main amplifier block. The output response is given in Figure 4-16.

The power ratio of the main signal to the adjacent channel IMD components is 19.06– (-23.52) = 42.5 dB in the lower frequency band and 19.06– (- 0.08) = 29.14 dB in the higher frequency band. The predistortion resulted in an improvement in the lower frequency band and a degradation in the higher frequency band.

There are two different values for the ratio of the main signal to the distortion component on the high frequency and on the low frequency and the worse value is taken as the ACPR of the amplifier, i.e. 29.14 dB.

The ACPR of the 4 amplifier block with designed power combiner was found as 23.93 dB. The improvement is 29.14 – 23.93 = 5.21 dB in ACPR. In fact this predistortion technique is used to improve the lower frequency IMD component and an improvement is obtained from 23.93 dB to 42.5 dB. But the overall performance is restricted by the higher frequency component. The predistortion technique can only be used to have an improvement of one IMD component. In the previous section we have concluded that the designed power combiner/divider effect the higher frequency band IMD component. Since the predistortion technique is successful in the lower frequency band and an improvement is achieved in the higher frequency band by the power combiner; the next step should be considering the power divider and power combiner design.

The same structure is being used as power divider and combiner. Designing the power divider again should be useful in order to have an improvement of the higher frequency band IMD component; the lower frequency band component will be eliminated by the predistortion path.

A new power divider is designed and used; but only in the main amplifier block so that the predistortion block does not need to be changed.

The power divider parameters are optimized for the best performance of the overall system. As it is seen from the schematic that the power divider is a one stage one unlike the power combiner. There is no need to have a better isolation like the power combiner in power divider case. For a power combiner isolation between the ports to be combined is more important because the main signals come from these ports and the transfer between the ports should be avoided. But for a power divider, the main signal is to be divided and the isolation between the ports is for the reflected signal only, and since the reflections are not so much the isolation becomes less important. New power divider has a different loss than that of the power combiner. Being a one stage divider the reflections from the other ports are greater compared to the two stage divider. So replacement of the power combiners with the new power divider in the main amplifier block requires tuning of the attenuator resistances prior to the main amplifier block. The schematic of the new power divider is given in Figure 4-17.

The final design of the predistorted amplifier is given in two figures; first one is the predistortion block and the second one is the main amplifier block, Figure 4-18 and Figure 4-19 respectively.

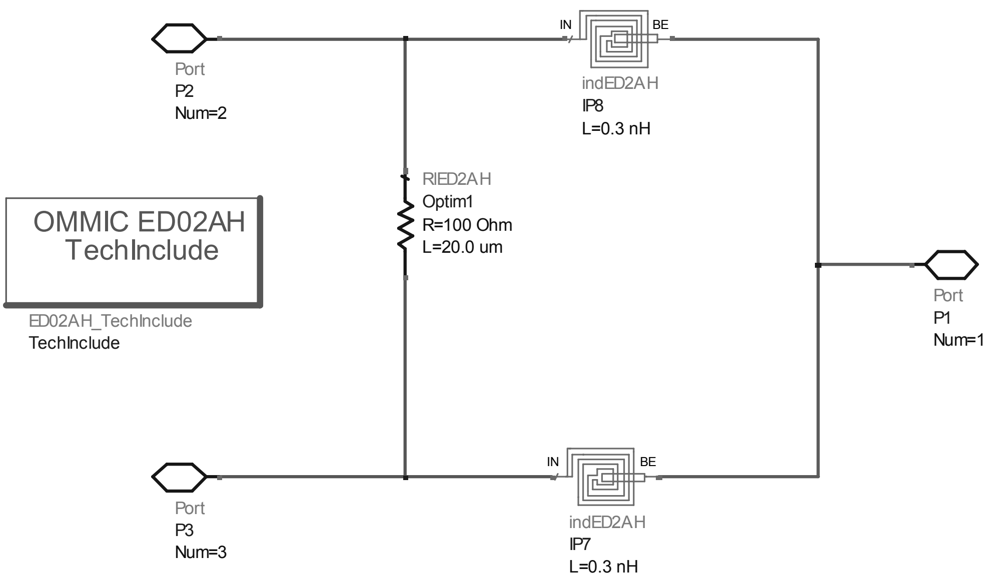

Figure 4-17 Power Divider Design

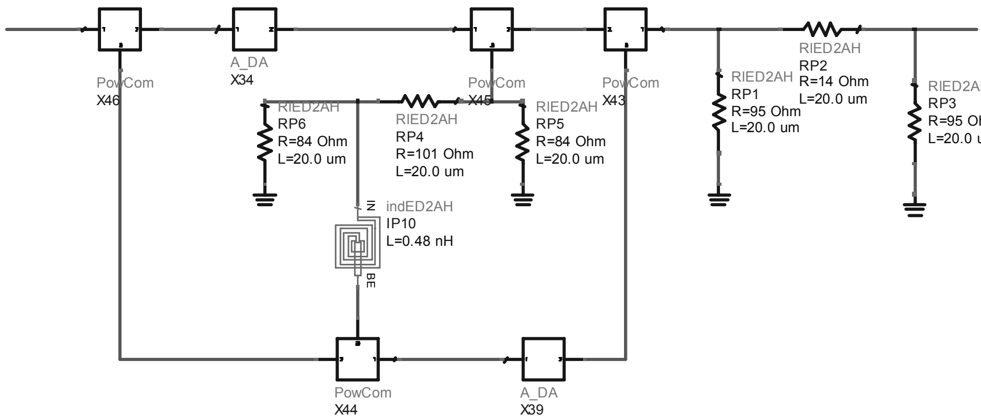

Figure 4-18 Predistortion Block

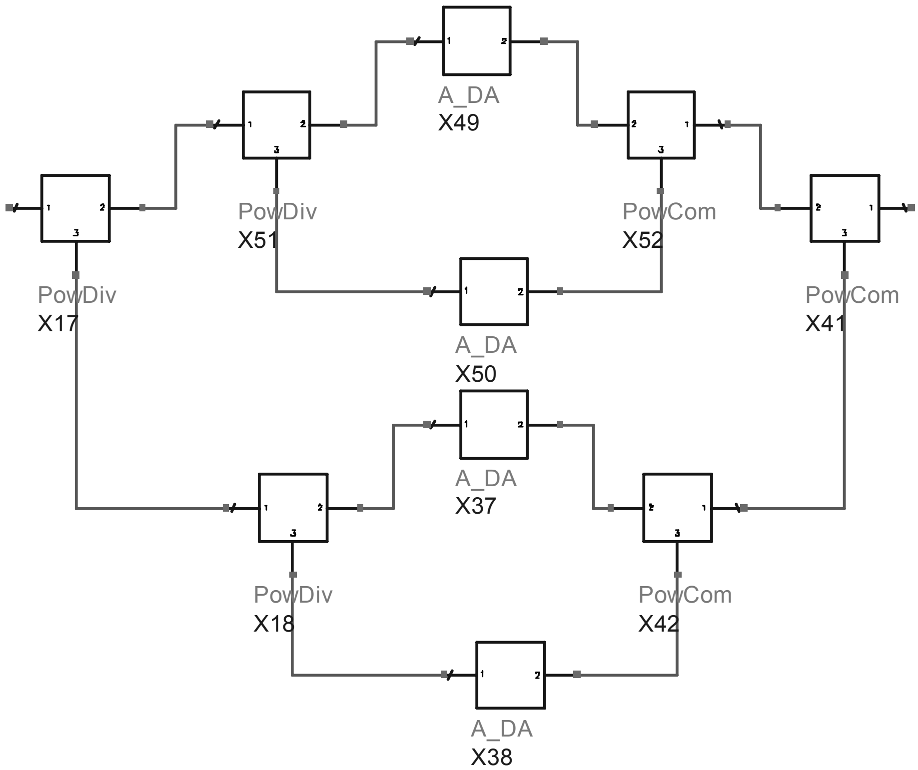

Figure 4-19 Main Amplifier Block

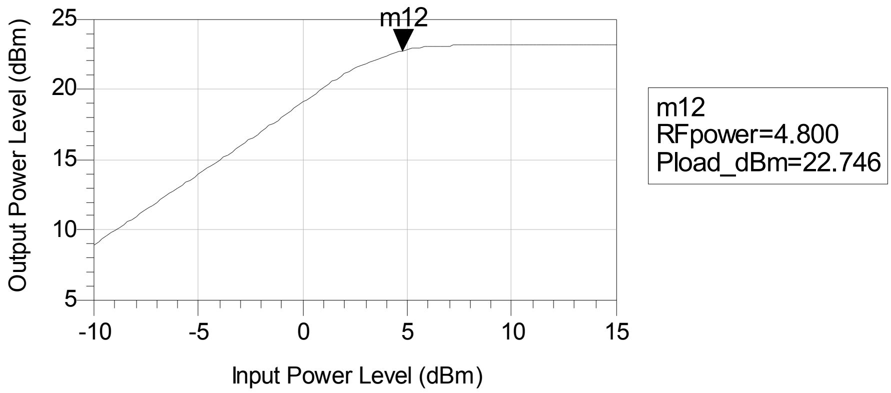

Figure 4-20 Input-Output Power Relation of the overall Amplifier

Drive amplifier output is sampled, attenuated and subtracted from the original signal in order to have the distortion only signal. The difference is adjusted to be 180 degrees by the use of inductor.

Distortion signal is amplified and added to the main path to perform the correct predistortion.

The gain of the amplifier is 18.9 dB. The 1 dB compression point of the amplifier is the point at which the amplification is 1 dB less than the gain which is 17.9 dB. From the graph showing output power versus input power of the amplifier in Figure 4-20, the 1 dB compression point is found as 22.75 dBm corresponding to the input power level of 4.8 dBm.

The output power of the overall amplifier versus input power is given in Figure 4-21. The power ratio of the main signal to the adjacent channel IMD component is 18.05– (-20.77) = 38.82 dB for lower frequency band and 18.05– (-17.89) = 35.94 dB for higher frequency band. In the previous case the predistortion was improved the lower frequency IMD component and the high frequency component was the dominant one. The higher frequency IMD component is decreased by the new power divider and the overall performance is improved.

There are two different values for the ratio of the main signal to the distortion component on the high frequency and on the low frequency and the worse value is taken as the ACPR of the amplifier, i.e. 35.94 dB. The ACPR of the 4 amplifier block with designed power combiner was found as 23.93 dB. The improvement is 35.94 – 23.93 = 12.01 dB in ACPR.

The power ratio of the main signal to the adjacent channel IMD component for the drive amplifier was 24.56 dB for lower frequency band and 22.29 dB for higher frequency band.

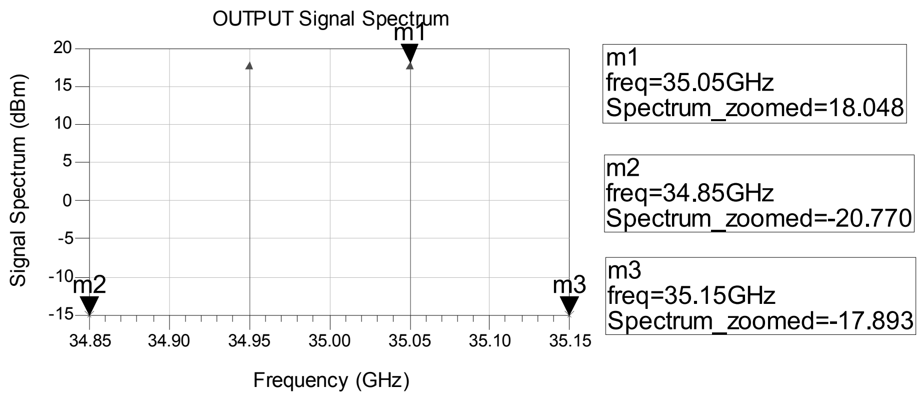

Figure 4-21 Predistorted Amplifier Output

There is an improvement in the ratio of the main signal to the IMD component compared to the drive amplifier alone: from 24.56 dB to 38.82 dB, 14.26 dB in lower frequency band and from 22.29 dB to 35.94 dB, 13.65 dB in higher frequency band.

The predistortion technique applied ends up with an ACPR of 35.94 dB. There is an improvement of 13.65 dB compared to the drive amplifier.

4.5 ANALYSIS OF THE AMPLIFIER

The performance of the amplifier is demonstrated at the specified conditions; at input power level of 1 dB compression point of the amplifier, at 35 GHz and at a frequency spacing of 100 MHz.

The results are summarized in the Table 4-1 for the drive amplifier, combination of four drive amplifiers in parallel and the overall amplifier.

ACPR is the Adjacent Channel Power Ratio of the total main signal on both frequency bands to the IMD components on both lower and higher IMD components.

Table 4-1 ACPR values of the main amplifier and predistorted amplifier

	IMD_low freq	IMD_high freq	ACPR
DA	24.5 dB	22.25 dB	22.25 dB
4 DA	23.93 dB	31.86 dB	23.93 dB
Predistorted Amplifier	42.5 dB	29.14 dB	29.14 dB
Power Divider Optimization	38.82 dB	35.94 dB	35.94 dB

"IMD_low freq" represents the ratio of the main signal to the lower frequency band IMD component and "IMD_high freq" represents the ratio of the main signal to the higher frequency band IMD component.

DA is the design amplifier that is used as the drive amplifier. 4 DA is the parallel combination of four drive amplifiers by the use of power combiner designed and Amplifier is the amplifier in which the predistortion technique is applied. Finally the last one is the final design with the power divider optimization.

How do these parameters affect the performance of the amplifier? Let's have a look at the performance of the amplifier when these parameter are changed.

In fact the amplifier is designed to operate at 35 GHz. So there should not be any expectation for a better performance at other frequencies, the predistortion block is tuned for best performance at 35 GHz and the components are dependent on frequency. Especially the phase shift of the components is being changed with changing frequency and the predistortion prediction becomes wrong and the predistortion fails to linearize the amplifier. The ACPR at 34 GHz and at 36 GH is nearly same with the parallel combination of four amplifiers.

The simulation results are sampled in table below. The ACPR is evaluated as the main signal to the adjacent channel IMD component. Since there are two tones, two different ACPR values for the low frequency band and high frequency band can be obtained, which are given in the following table as ACPR 1 and ACPR2 respectively. Then the overall ACPR is taken as the smaller value.

The ACPR of the predistorted amplifier does not change with bandwidth up to 230 MHz in Figure 4-22. The reason is that the dominant IMD component is the high frequency channel component up to a bandwidth of 230 MHz and is independent from the bandwidth.

The bandwidth is an important parameter for the amplifier. The design is optimized for a frequency spacing of 100 MHz and the performance will get worse with increasing value of the frequency spacing. The two tone harmonic balance analysis of the amplifier is re-perfumed with a sweep in variable BW which is the variable representing the frequency spacing. Frequency spacing is swept up to 1000 MHz with 10 MHz intervals and the ACPR versus BW is plotted in Figure 4-22.

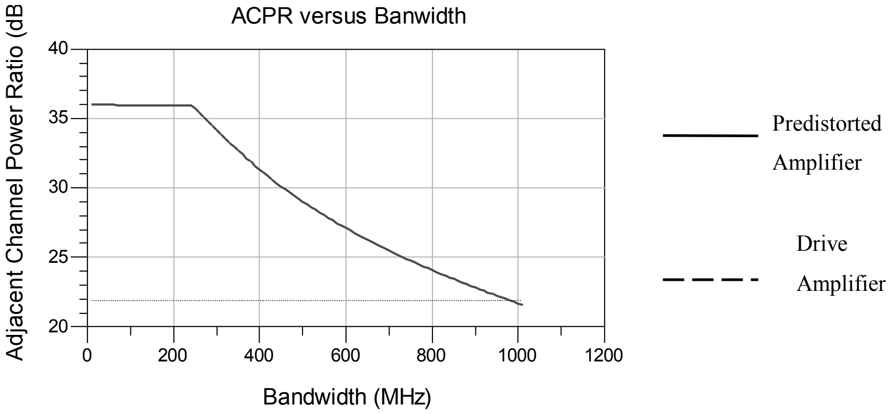

Figure 4-22 ACPR versus Bandwidth

As seen in Table 4-2, the ratio of the main signal to the IMD component at the high frequency band is independent from the bandwidth. We know that the offered predistortion technique is only for the low frequency component. The ACPR of the predistorted amplifier for the low frequency component decreases with increasing bandwidth due to the dependency of the circuit elements to the frequency. The characterization of the circuit changes with frequency and the circuit parameters are optimized for best performance with a bandwidth of 100 MHz.

The improvement in ACPR reduces with increasing BW. For BW value 1 GHz the improvement is zero and for the greater values of BW the ACPR becomes worse than the drive amplifier.

Table 4-2 ACPR variation with bandwidth

BW	ACPR1	ACPR2	ACPR
100.000	38.862	35.941	35.941
200.000	37.241	35.895	35.895
300.000	34.111	35.842	34.111
400.000	31.311	35.784	31.311
500.000	29.011	35.721	29.011
600.000	27.099	35.657	27.099
700.000	25.471	35.593	25.471
800.000	24.056	35.530	24.056
900.000	22.805	35.470	22.805
1000.000	21.684	35.413	21.684

To see the input power level effect on the ACPR the sweep parameter is changed to input power level and the simulation is re-performed. In Figure 4-23 the ACPR variation versus input power level is given.

The input power level corresponding to the 1 dB compression point of the drive amplifier is 2 dBm. There is an improvement in ACPR besides the input power level of 2 dBm. This predistortion method has a restricted performance on input power level.

The output power versus input power for a linear amplifier and a practical amplifier was given in Figure 2-1. The predistortion technique applied on the amplifier aims to have an output response closer to the output response of a linear amplifier.

The comparison of the output of the predistorted amplifier and the drive amplifier indicates the performance the predistortion application on linearization.

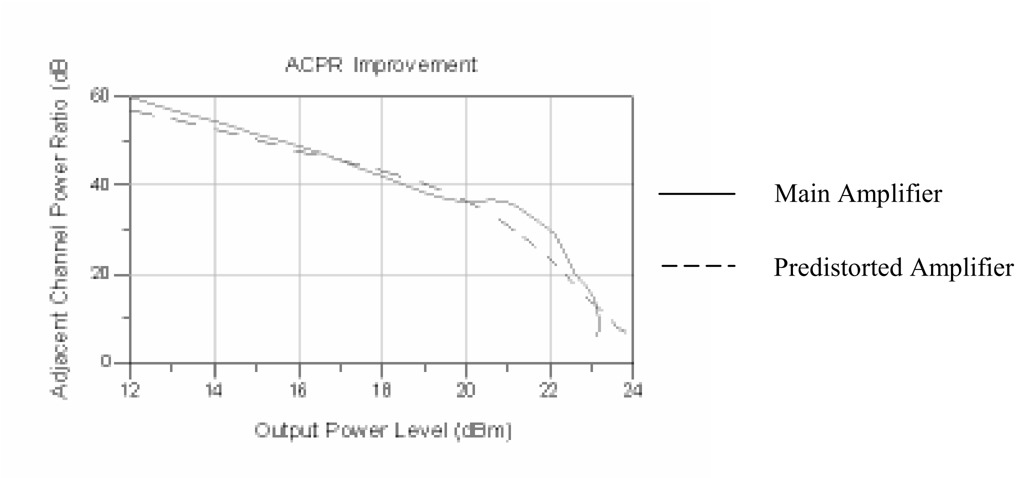

Figure 4-23 ACPR versus Input Power Level

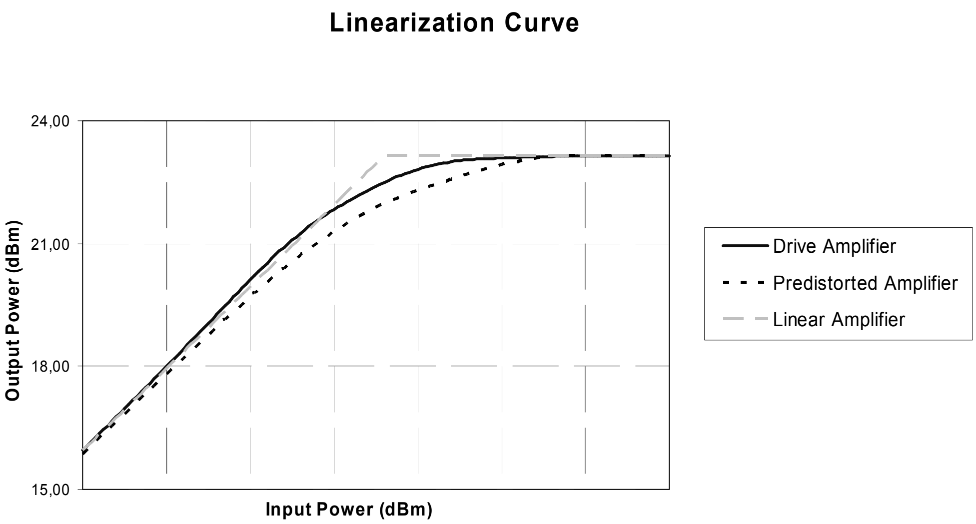

Figure 4.4-24 Output Power versus Input Power for drive amplifier and predistorted amplifier

In Figure 4.4-24 the output power level of the drive amplifier and the predistorted amplifier versus input power level is given. The predistorted amplifier has four drive amplifiers parallel to each other so the saturation of the predistorted amplifier corresponds to an output power level that is 6 dB more than the drive amplifier's. This 6 dB difference makes it difficult to make a fair comparison between the output power levels of the two amplifiers. The graphs in Figure 4.4-24 are normalized to make the comparison easier. So, the values of input power and output power levels on the x and y axis are out of concern.

4.6 LAYOUT OF THE AMPLIFIER

After the design and analysis of the amplifier, the layout of the system is prepared. The layouts for the drive amplifier, power divider and power combiner are given in Figure 4-25, Figure 4-26 and Figure **4-27** respectively.

In the design of main amplifier as a parallel combination of four drive amplifiers, the power divider and the power combiner could not be used because of the physical lengths. Therefore, a new pair of power combiner and power divider are designed, whose layouts are given in Figure 4-28 and Figure 4-29. The main amplifier layout is given in Figure 4-30 and the predistorter circuit layout is given in Figure 4-31.

The corresponding schematics of the layouts in figures from Figure 4-25 to Figure 4-31 are given in figures from Figure 4-32 to Figure 4-39.

In the schematic of the main amplifier, the power dividers, power combiners and the amplifiers are given as black boxes, since they are given separately.

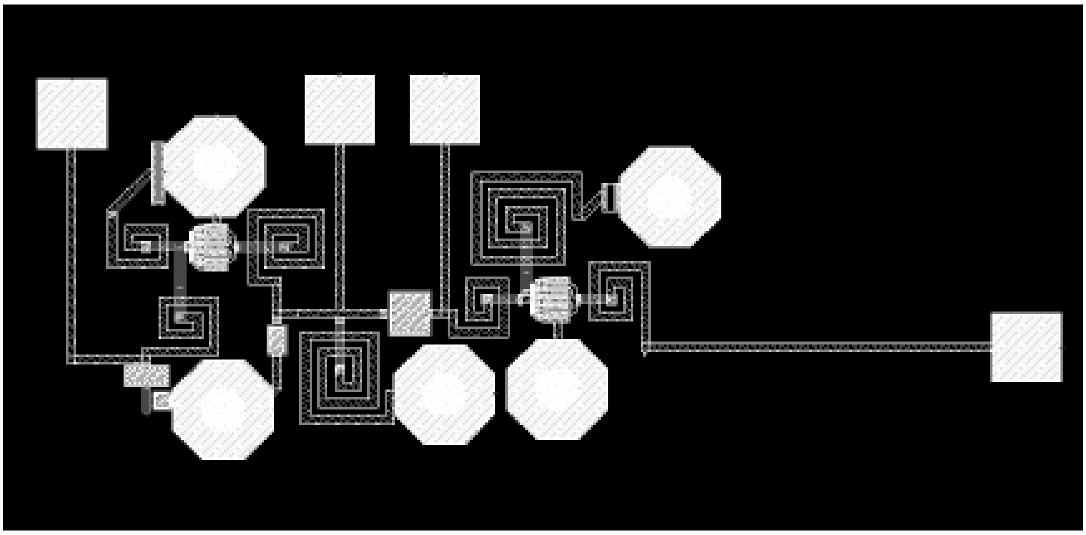

Figure 4-25 Layout for Drive Amplifier

Dimensions for the layout : 1.120 mm x 0.467 mm

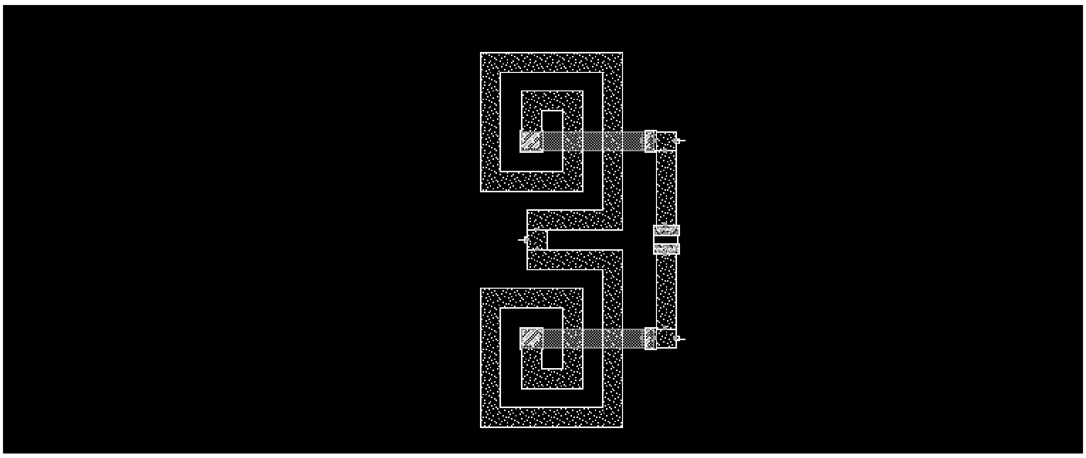

Figure 4-26 Layout for Power Divider

Dimensions for the layout : 0.100 mm x 0.195 mm

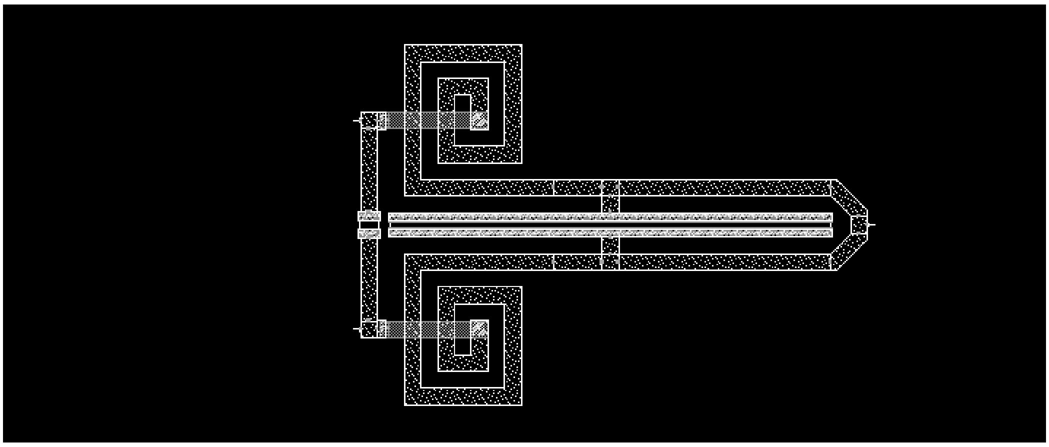

Figure 4-27 Layout for Power Combiner

Dimensions for the layout : 0.315 mm x 0.220 mm

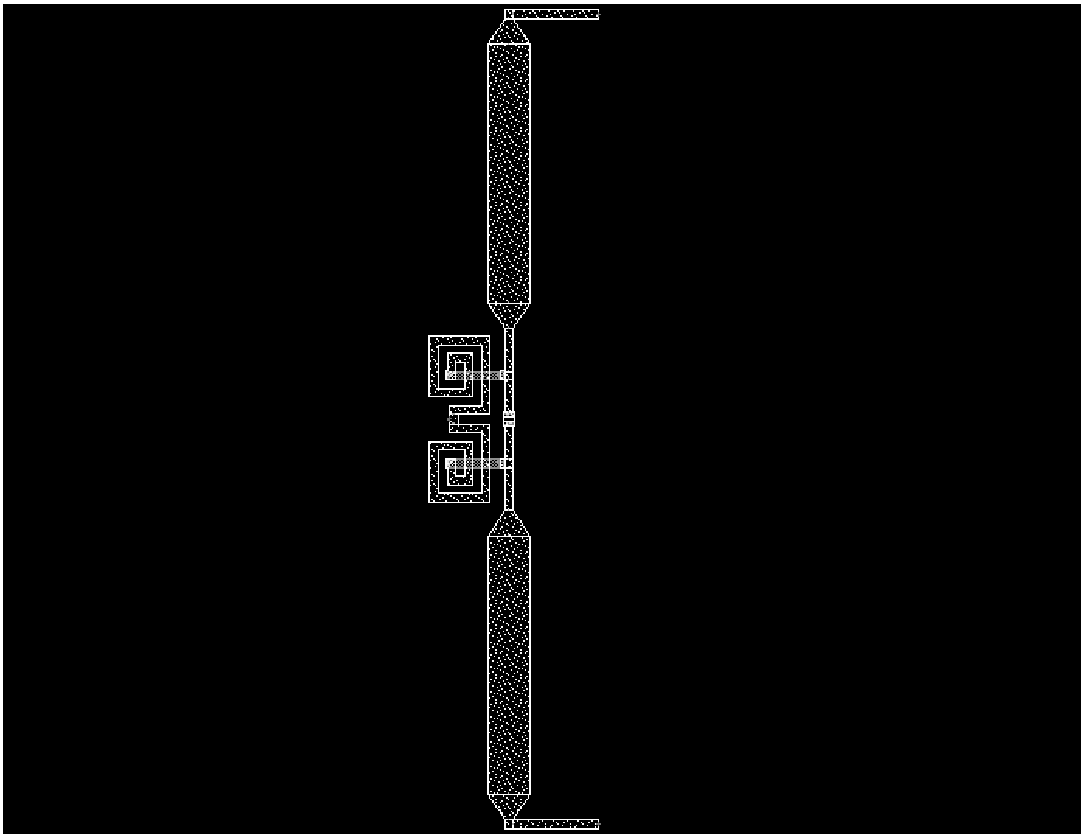

Figure 4-28 Layout for Power Divider 2

Dimensions for the layout : 0.205 mm x 0.975 mm

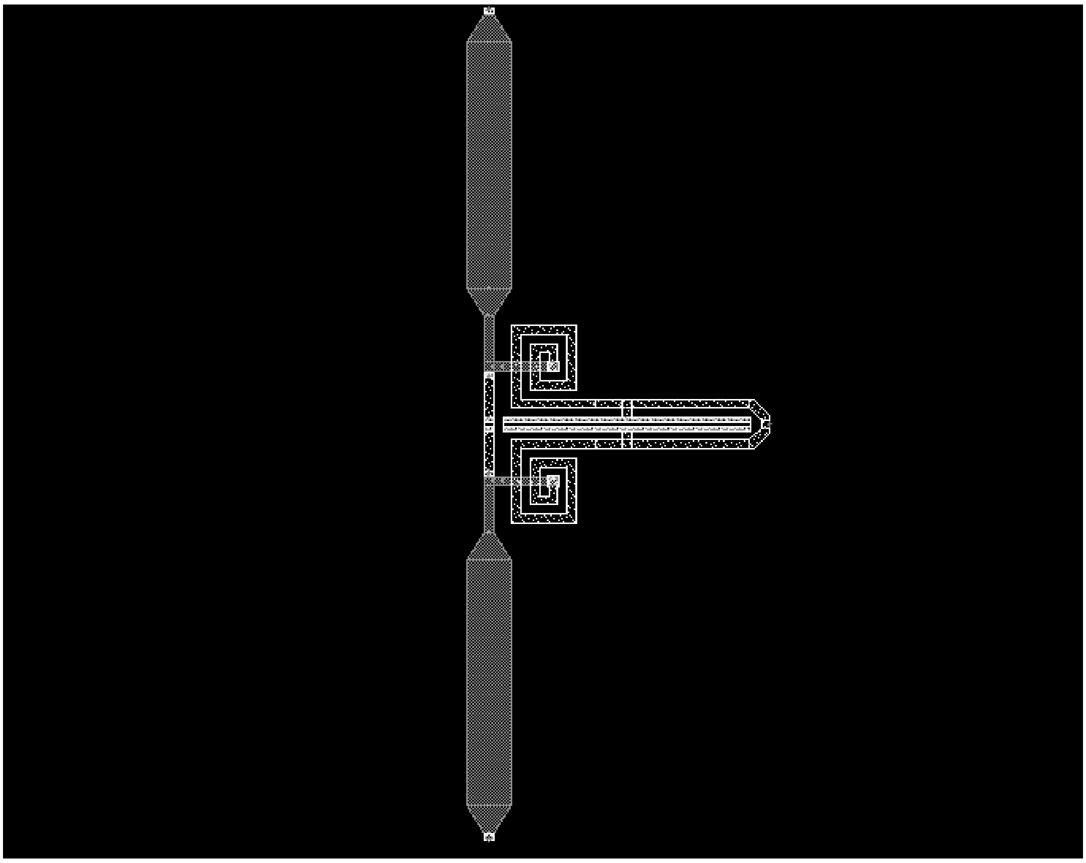

Figure 4-29 Layout for Power Combiner 2

Dimensions for the layout : 0.340 mm x 0.915 mm

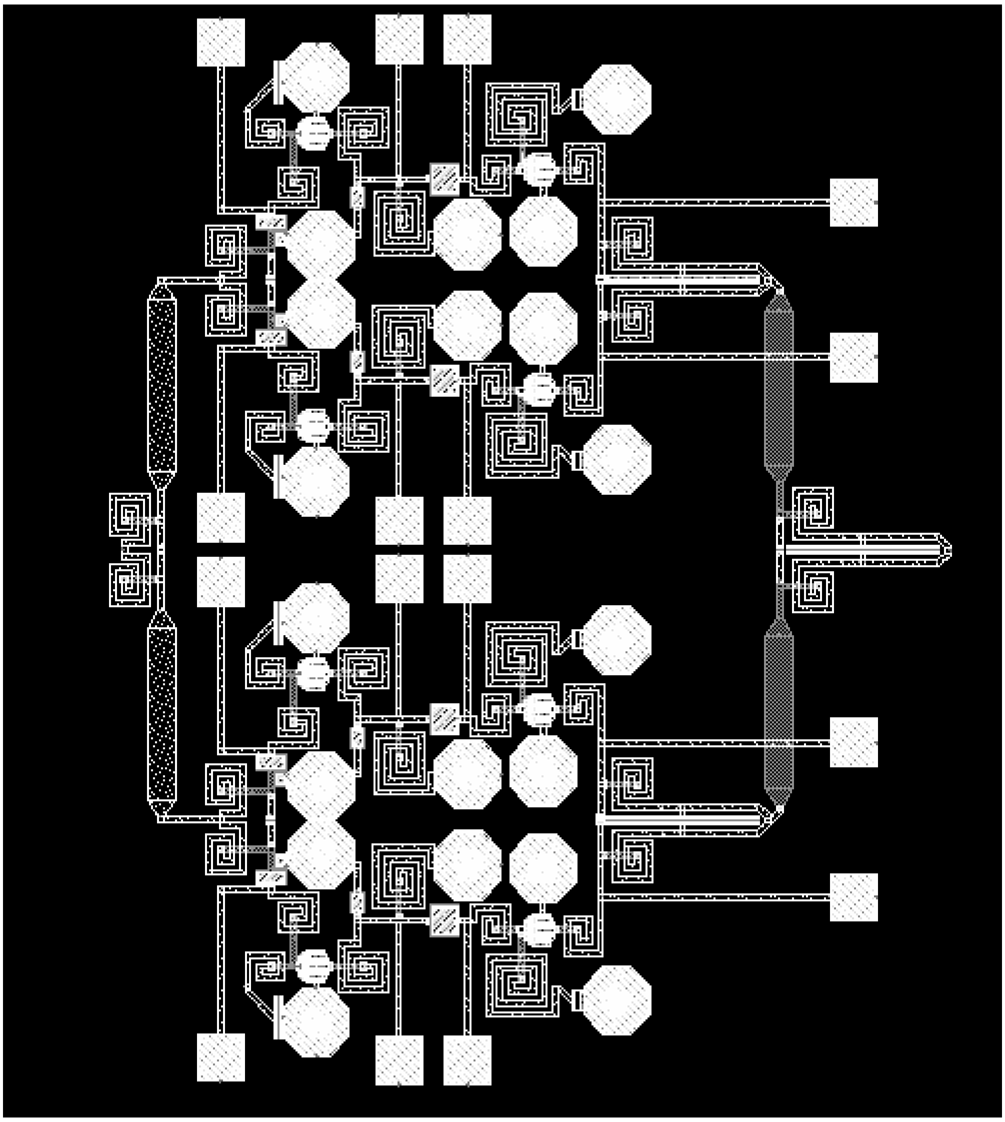

Figure 4-30 Layout for Main Amplifier

Dimensions for the layout : 1.520 mm x 1.880 mm

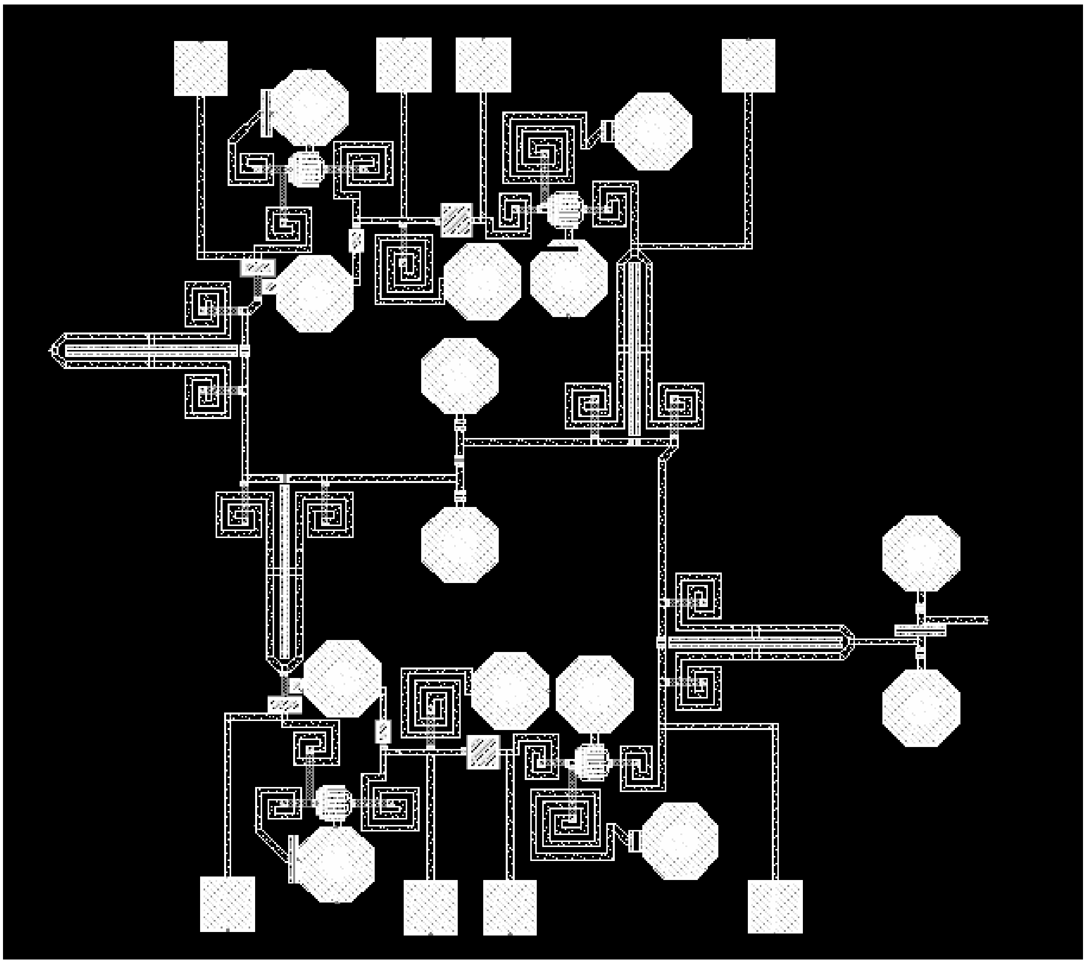

Figure 4-31 Layout for the Predistorter
Dimensions for the layout : 1.380 mm x 1.430 mm

The layout of te overall system has a dimension of 2.920mm x 1890 mm.

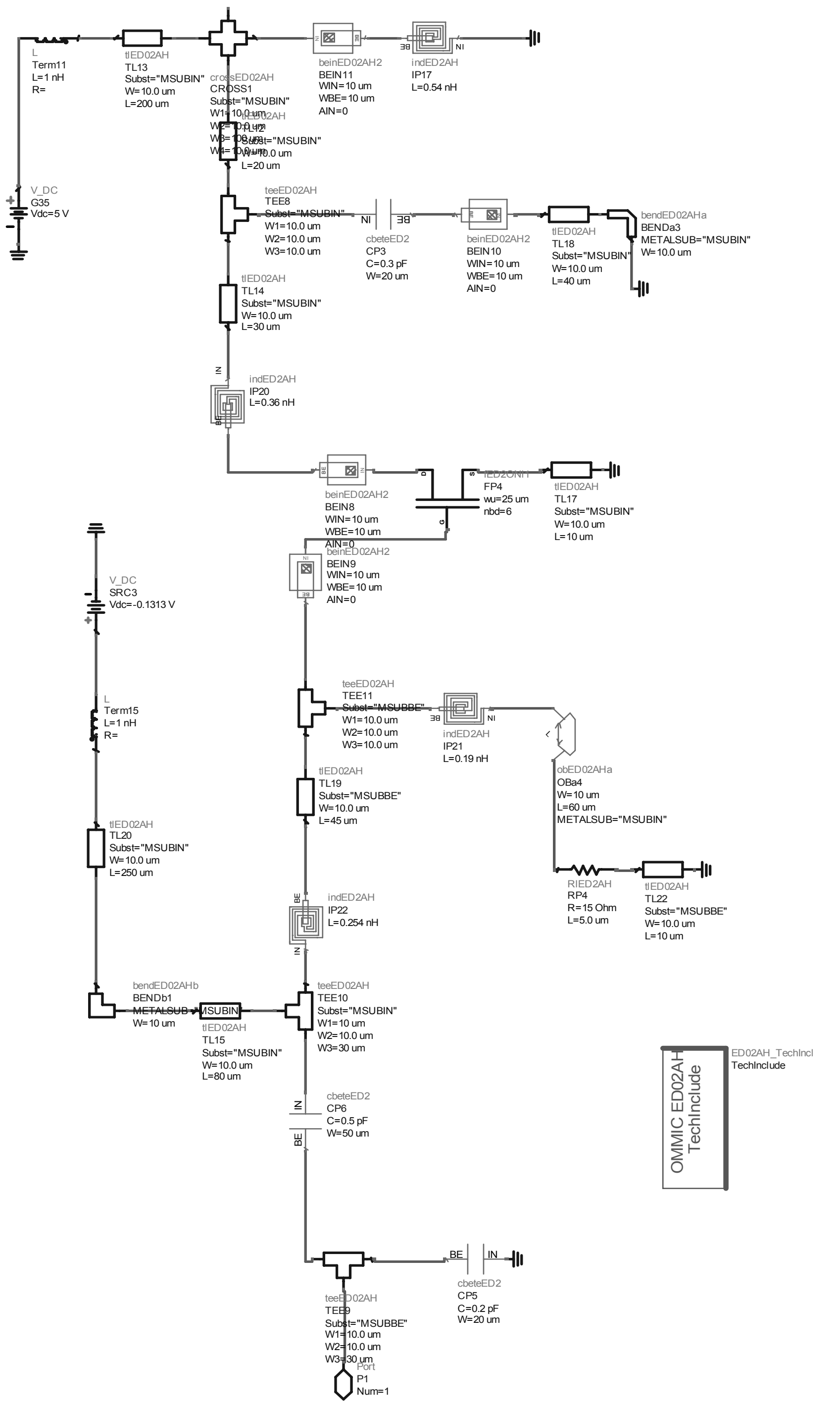

Figure 4-32 Schematic for the Drive Amplifier Layout (Stage 1)

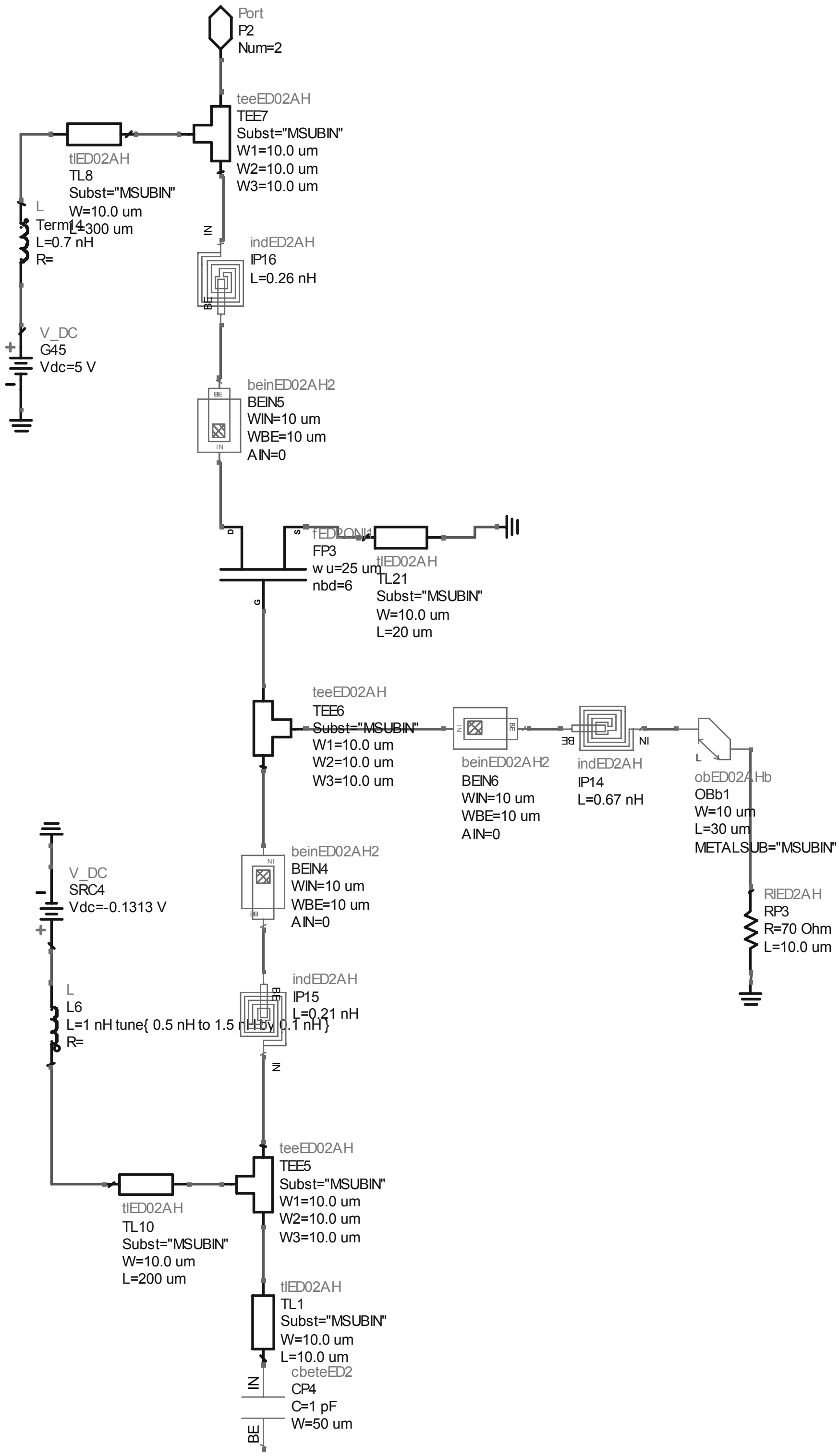

Figure 4-33 Schematic for the Drive Amplifier Layout (Stage 2)

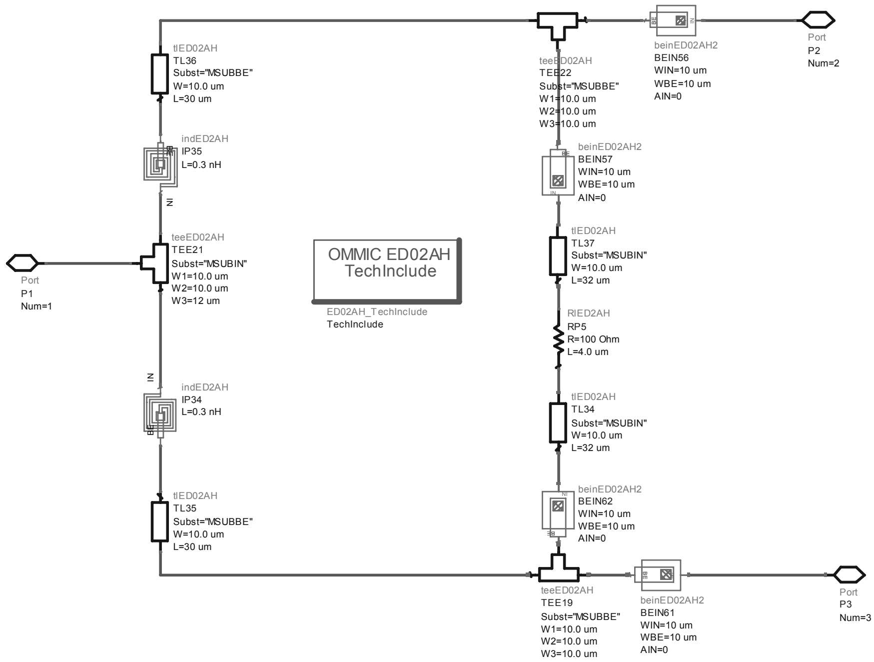

Figure 4-34 Schematic for the Power Divider Layout

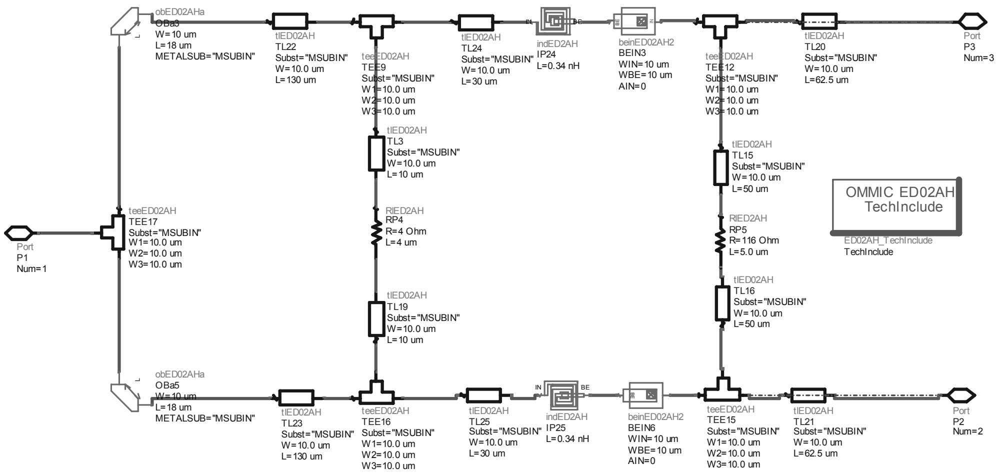

Figure 4-35 Schematic for the Power Combiner Layout

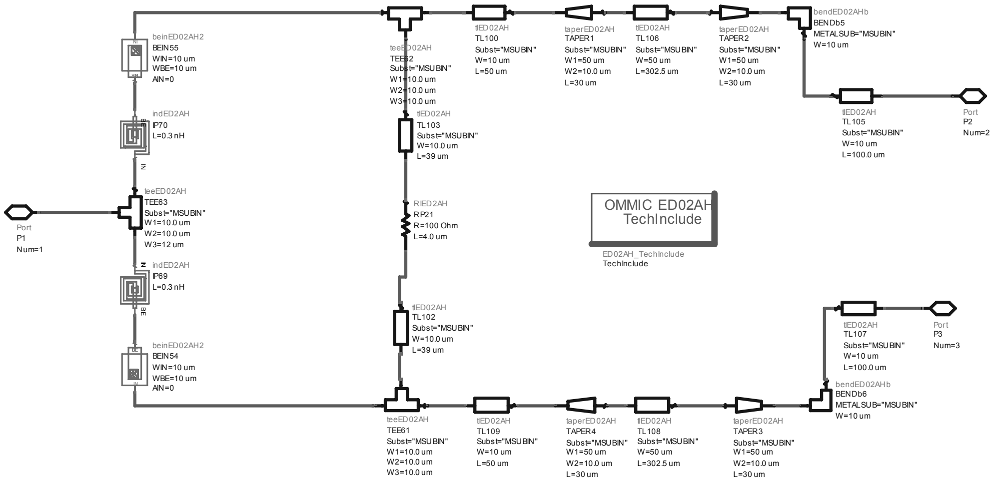

Figure 4-36 Schematic for the Power Divider 2 Layout

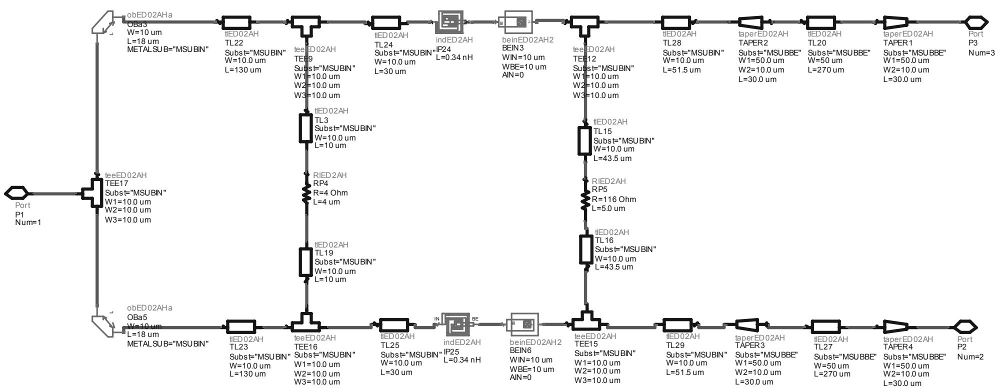

Figure 4-37 Schematic for the Power Combiner 2 Layout

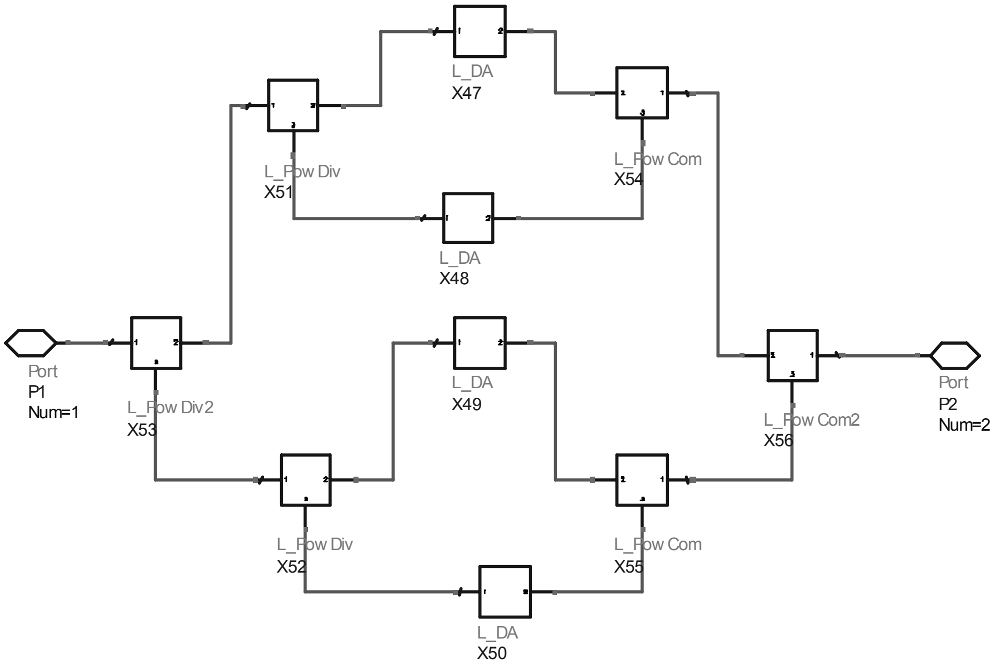

Figure 4-38 Schematic for the Main Amplifier Layout

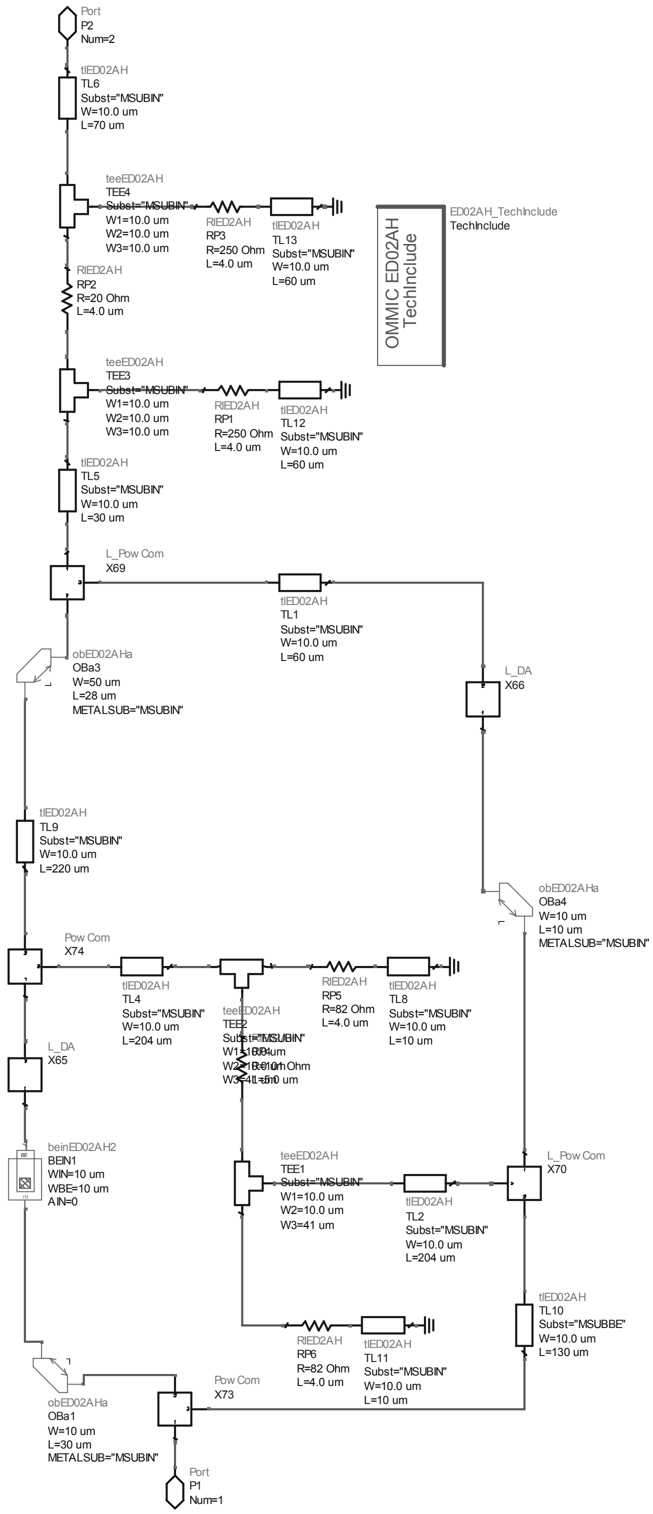

Figure 4-39 Predistorter Circuit Schematic

During the preparation of the layouts, some of the circuit element values had to be changed. The layout schematic of the drive amplifier given in Figure 4-32 and Figure 4-33 is not exactly the same with the drive amplifier given in Figure 4-1. The S parameter analysis of the drive amplifier layout schematic in the whole frequency spectrum from 1 GHz to 50 GHz is given in Figure 4-40 and in frequency band from 30 GHz to 40 GHz is given in Figure 4-41. Additional components only degraded the gain by 0.1dB, however; input reflection is changed considerably. Special design effort is not consumed to improve the input reflection at this stage, but instead, later, overall performance of the system is optimized.

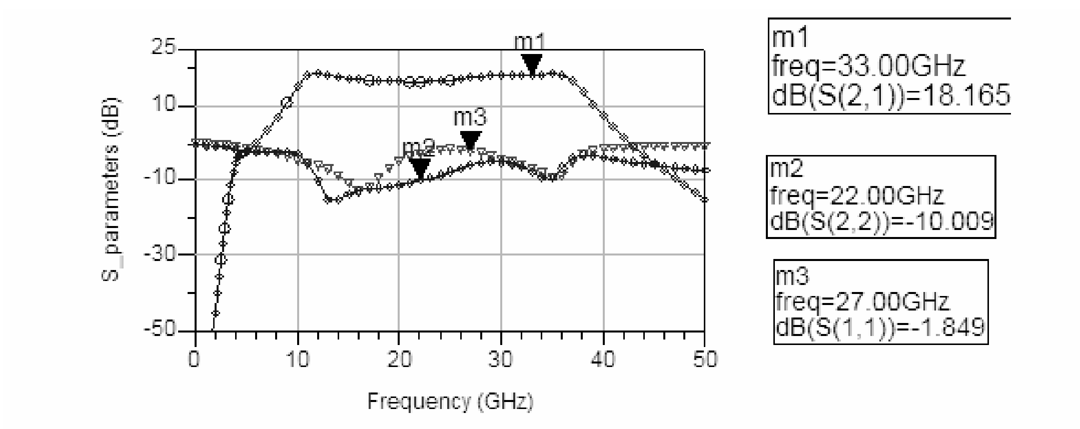

Figure 4-40 S Parameters of the Drive Amplifier Layout 1 GHz – 50 GHz

The two tone harmonic balance simulation results of the drive amplifier layout schematic (all parts of the layout are included in the simulation), the input output relation and the fundamental and IMD components are given in Figure 4-42 and Figure 4-43 respectively. P_{1dBm} point of the drive amplifier is 16.3 dBm and ACPR of the drive amplifier is 13.3 – (-9.2), i.e. 22.5 dB. These values are very similar to the initial design.

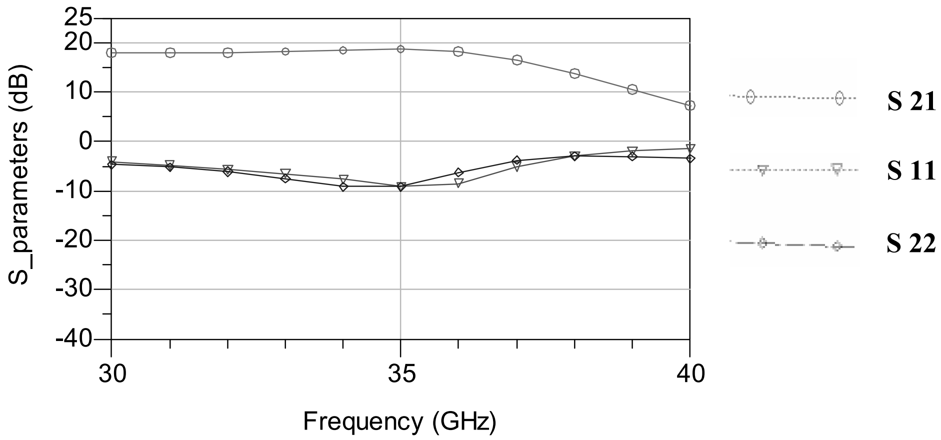

Figure 4-41 S Parameters of the Drive Amplifier Layout 30 GHz – 40 GHz

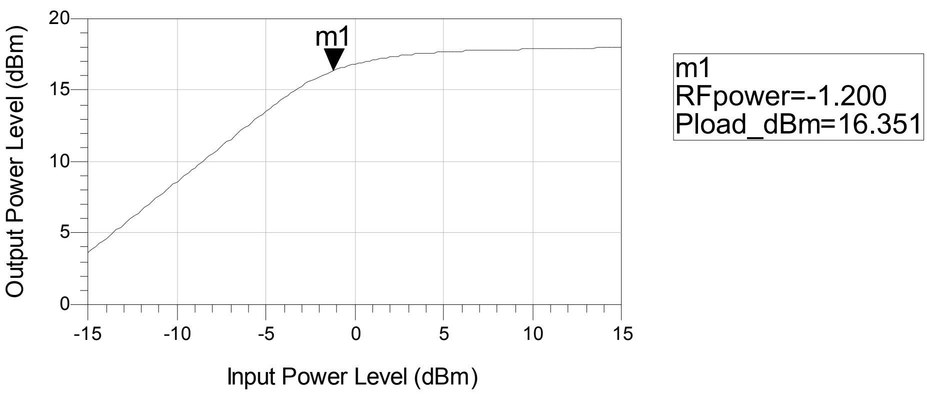

Figure 4-42 Input_Output Power Relation of Drive Amplifier Layout Schematic

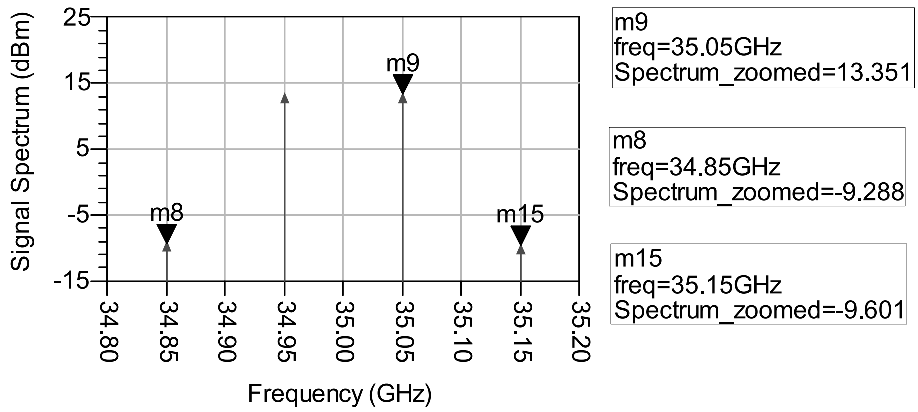

Figure 4-43 Fundamental and IMD components of the Drive Amplifier Layout Schematic

After the drive amplifier layout is completed, the next step is preparation of the layout for the main amplifier block. For the main amplifier block, the layouts of the power divider and power combiner are prepared first. The layouts and schematics of the power divider, power combiner and main amplifier block are given above together with the others. The two tone harmonic balance analysis of the designed main amplifier is given below.

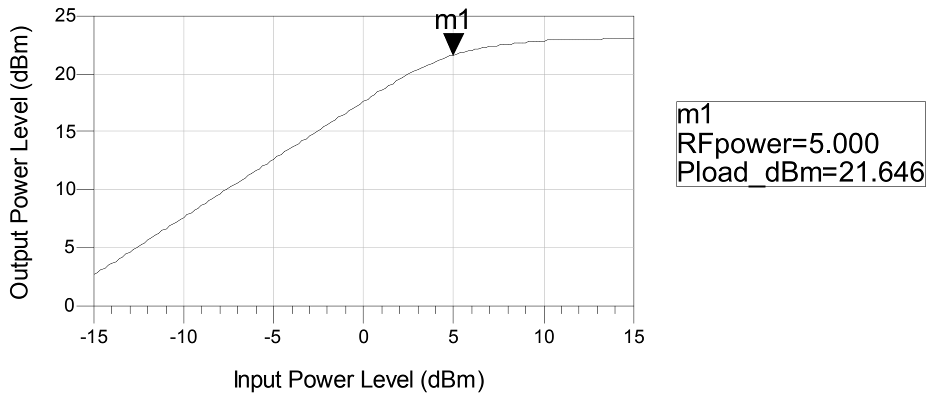

Figure 4-44 Input_Output Power Relation of Main Amplifier Layout Schematic

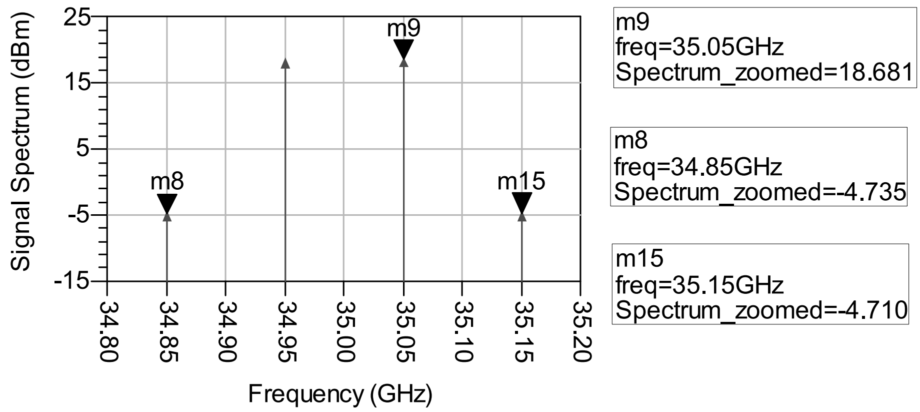

Figure 4-45 Fundamental and IMD components of the Main Amplifier Layout Schematic

The input-output relationship is given in Figure 4-44 and the fundamental and IMD components is given in Figure 4-45 for the main amplifier block.

The P_{1dB} point of the main amplifier block is 21.6 dBm, i.e. 5.3 dB more than the drive amplifier itself. It is expected to be 6 dB more than that of the drive amplifier, but the power combiner loss introduces a 0.7 dB loss in the output of the main amplifier. The ACPR for the main amplifier is 18.6 – (-4.7), i.e. 23.3 dB. The values are similar to those of the initial design.

The last step of the layout study is the predistorter design, and combining of the predistorter with the main amplifier. The predistorter layout is given in Figure 4-31 and the analysis of the predistorted amplifier is given below. The two tone harmonic balance simulation results, input-output relation of the overall system is given in Figure 4-46 and the fundamental and IMD components are given in Figure 4-47.

The P_{1dB} point of the overall amplifier is 22 dBm, 0.5 dB more than the main amplifier. This improvement is mainly because of the linearization. The IMD component in the lower frequency band is dropped from –4.7 dBm to -13 dBm. The improvement in the lower frequency band IMD component is more than 8 dB but the ACPR improvement is limited by the higher frequency IMD component. This is the expected result because cancellation depends highly on the phase distribution of the spectrum. The two IMD components are different phases thus the cancellation circuit can only deal with one of them.

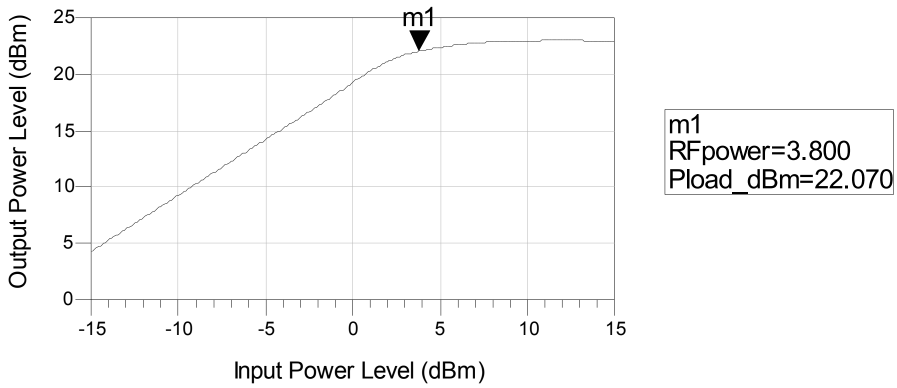

Figure 4-46 Input_Output Power Relation of the Predistorted Amplifier Layout Schematic

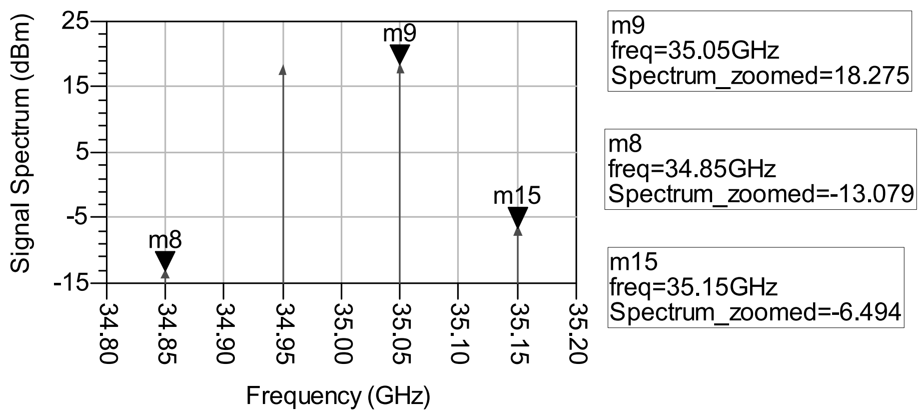

Figure 4-47 Fundamental and IMD components of the Predistorted Amplifier Layout Schematic

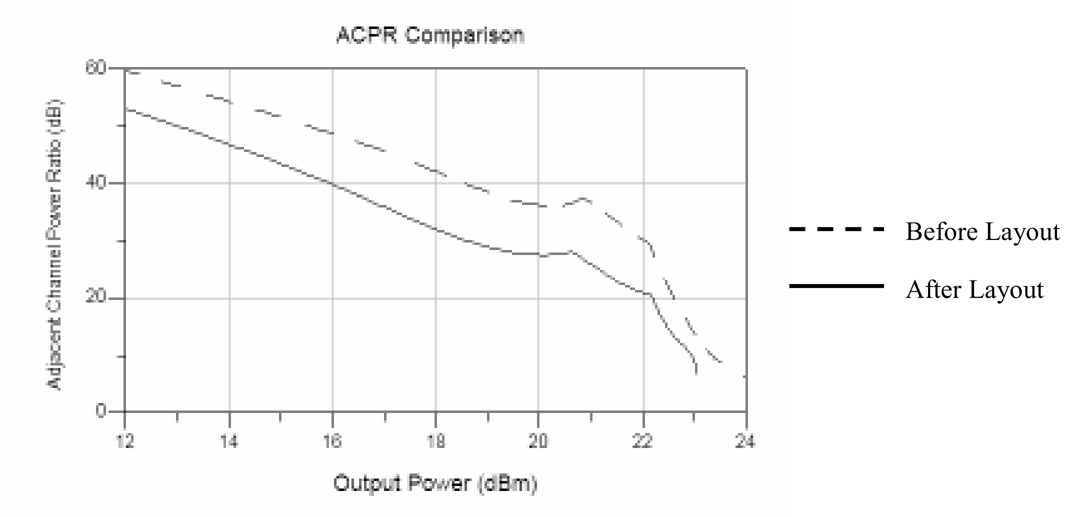

Figure 4-48 ACPR Comparison of the Predistorted Amplifier Layout Schematic with the Predistorted Amplifier before the layout

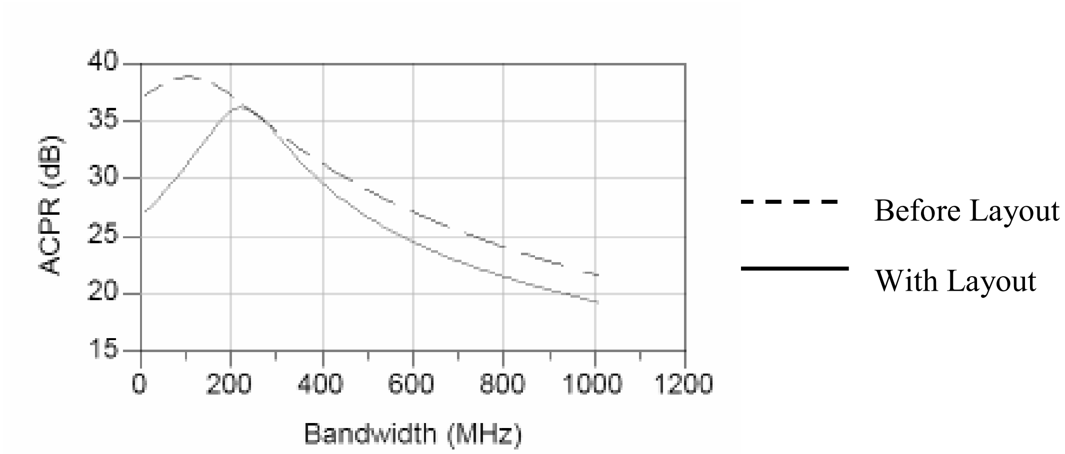

Figure 4-49 Fundamental signal ratio to the IMD component in low frequency band for the Predistorted Amplifier before the layout and with the layout

The ACPR of the overall system, that is the predistorted amplifier is given in Figure 4-48 with the ACPR of the system before the layout. The characteristic of the ACPR versus Output power is very similar, but the layout effects of the components make the performance worse. The reasons might be the increased levels of internal reflections and the amplitude unbalance due to the additional lossy components. A further optimization of the full layout may improve.

The fundamental signal ratio to the IMD component at the lower frequency band versus bandwidth is given in Figure 4-49. The system before the layout and the system with the layout effects can be compared by use of Figure 4-49. The characteristic of the variation is similar but the optimum bandwidth for the systems are different. This shows that the system can be optimized to a desired bandwidth in an application.

CHAPTER 5

CONCLUSION

Linearity is an important measure for systems using modulation schemes having phase and amplitude modulation at the same time. The phase and amplitude relation of the input and output signals are expected to be independent of the input power level. However, due to the nature of the semiconductor components, nonlinearity is inherently present. In order to decrease the negative effects, a certain back-off can be considered in spite of increased DC power consumption which decreases the efficiency. On the other hand, in some cases attainable maximum power level is limited due to the technology. In such cases, maximum linearity at that power level is aimed whatever the expense is. Linearization circuits are used to increase the linearity at a certain power level. Cost in this case is the increased circuit complexity; both in design and implementation.

In this thesis an analog predistortion scheme, "self cancellation", is investigated. The simulation of the self cancellation scheme is demonstrated in Advance Design System software (ADS) with ideal components and real components separately.

The two tone harmonic balance analysis is performed at a frequency of 1 GHz with ideal components. The input power level of the drive amplifiers used in the design was 10 dBm. A system amplifier is defined to be used as the drive amplifier. The gain of the drive amplifier is defined as 20 dB and the third order intercept point as 50 dB.

The linearity of the drive amplifier can be evaluated as the ratio of the main signal to the third order distortion component, which is also called ACPR, the difference of the signal level in dB. The linearity of the final amplifier is improved significantly compared to the drive amplifier itself.

The simulation with ideal components resulted in an improvement of 40 dB in ACPR. The drive amplifier itself had a 40 dB and the over all design had a 80 dB ACPR.

The two tone harmonic balance analysis is repeated with a Class-A amplifier and real components at a frequency of 35 GHz with a frequency spacing of. A Class-A amplifier with a gain of 18.6 dB is designed to be used as the drive amplifier.

Firstly, the designed Class-A amplifier is analyzed alone and the ACPR of the drive amplifier is evaluated as 22.25 dB.

Secondly, four identical drive amplifiers are combined in parallel to form an amplifier block. Input signal is divided into 4 signals, amplified and combined again. The analysis of the amplifier block is performed two times; one is with the perfect power splitters used as power divider and combiner, the other is with the designed power splitter used as the power divider and combiner. The ideal power splitter case resulted in an ACPR of 22.29 dB which is slightly different from the drive amplifier ACPR. In the designed power splitter case the third order IMD components were asymmetric. The higher frequency component is decreased due to the power splitter design. The reflections of the power splitter caused the higher frequency component to be decreased. But the ACPR is evaluated as 23.93 dB and there is no improvement is recorded. The high frequency IMD component is decreased but the low frequency component remains so the ACPR is not changed.

It is seen that the IMD components of the amplifier block are asymmetric. Since the offered predistortion method is used to eliminate the one of the IMD component which should be the dominant one, the other IMD component will affect the linearization performance.

The predistortion application on the amplifier block resulted in an improvement of 20.25 dB in the lower frequency band compared to the drive amplifier, from 22.25 to 42.5. But the ACPR is evaluated as 29.14 dB which is 9 dB greater than the drive amplifier. The ACPR improvement is restricted by the higher frequency band IMD component. The higher frequency band IMD component is decreased by the use of non ideal power combiner. This implies that; the power divider and power combiner design can decrease the higher frequency band IMD component and a better performance may be obtained by a new power divider instead of using the designed power combiner as a power divider. So a new power divider is designed and tuned to have a better overall performance.

The simulation result with the new power divider, the ACPR is evaluated as 35.94 dB. There is an improvement of 12 dB compared to the combination of four amplifier block, from 23.93 dB to 35.94 dB and an improvement of 13.69 dB compared to the drive amplifier, 22.25 dB to 35.94 dB.

The improvement of the method at 35 GHz on the MMIC amplifier is about 14 dB in ACPR of the designed amplifier.

The linearization performance decreases with increasing bandwidth. Up to 1 GHz there is a decreasing improvement in IMD; but greater values than 1 GHz the amplifier linearity decreases.

Finally the layout for the overall system is prepared, the production of the amplifier planned. In the layout study, frequency dependence and AM-PM effects degraded the overall performance of the system. These effects can be investigated as a future work.

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APPENDIX

The formulas in order to calculate the desired power spectrums in ADS are given.

Firstly the formulas to find the input spectrum are given. Secondly the drive amplifier output spectrum formulas are presented. The third group is for the error signal spectrum. Then the predistorted signal spectrum is evaluated and lastly the formulas for evaluation of the output amplifier spectum is given.

The formulas for input power spectrum.

```
Eqn Spectrum_W_in=0.5*real(Vin*conj(Iin.i))
Eqn Spectrum_in=10*log(Spectrum_W_in+1e-12) +30
Eqn TOIoutput_low_in=1.5*mix(Spectrum_in,{1,0},Mix)-0.5*mix(Spectrum_in,{2,-1},Mix)
Eqn TOIoutput_high_in=1.5*mix(Spectrum_in,{0,1},Mix)-0.5*mix(Spectrum_in,{-1,2},Mix)
Eqn TOIinput_low_in=TOIoutput_low_in-P_gain_transducer_in
Eqn TOIinput_high_in=TOIoutput_high_in-P_gain_transducer_in
Eqn FifthOIoutput_low_in=1.25*mix(Spectrum_in,{1,0},Mix)-0.25*mix(Spectrum_in,{3,-2},Mix)
Eqn FifthOIoutput_high_in=1.25*mix(Spectrum_in,{0,1},Mix)-0.25*mix(Spectrum_in,{-2,3},Mix)
Eqn FifthOIinput_low_in=FifthOIoutput_low_in-P_gain_transducer_in
Eqn FifthOIinput_high_in=FifthOIoutput_high_in-P_gain_transducer_in
Eqn P_gain_transducer_in=Pin_dBm-RFpower

Eqn Spectrum_zoomed_in=mix(Spectrum_in,tones,Mix)
Eqn Pin_W1=mix(Spectrum_W_in,{1,0},Mix)
Eqn Pin_W2=mix(Spectrum_W_in,{0,1},Mix)
Eqn Pin_dBm=10*log(Pin_W1 +Pin_W2)+30
```

The formulas for drive amplifier output spectrum

```
Eqn Spectrum_W_da=0.5*real(Vda*conj(Ida.i))
Eqn Spectrum_da=10*log(Spectrum_W_da) +30
Eqn TOIoutput_low_da=1.5*mix(Spectrum_da,{1,0},Mix)-0.5*mix(Spectrum_da,{2,-1},Mix)
Eqn TOIoutput_high_da=1.5*mix(Spectrum_da,{0,1},Mix)-0.5*mix(Spectrum_da,{-1,2},Mix)
Eqn TOIinput_low_da=TOIoutput_low_da-P_gain_transducer_da
Eqn TOIinput_high_da=TOIoutput_high_da-P_gain_transducer_da
Eqn FifthOIoutput_low_da=1.25*mix(Spectrum_da,{1,0},Mix)-0.25*mix(Spectrum_da,{3,-2},Mix)
Eqn FifthOIoutput_high_da=1.25*mix(Spectrum_da,{0,1},Mix)-0.25*mix(Spectrum_da,{-2,3},Mix)
Eqn FifthOIinput_low_da=FifthOIoutput_low_da-P_gain_transducer_da
Eqn FifthOIinput_high_da=FifthOIoutput_high_da-P_gain_transducer_da
Eqn P_gain_transducer_da=Pda_dBm-RFpower

Eqn Spectrum_zoomed_da=mix(Spectrum_da,tones,Mix)
Eqn Pda_W1=mix(Spectrum_W_da,{1,0},Mix)
Eqn Pda_W2=mix(Spectrum_W_da,{0,1},Mix)
Eqn Pda_dBm=10*log(Pda_W1 +Pda_W2)+30
```

The formulas for error signal spectrum.

Eqn Spectrum_W_error=0.5*real(Verror*conj(Ierror.i))

Eqn Spectrum_error=10*log(Spectrum_W_error+1e-12) +30

Eqn TOloutput_low_error=1.5*mix(Spectrum_error,{1,0},Mix)-0.5*mix(Spectrum_error,{2,-1},Mix)

Eqn TOloutput_high_error=1.5*mix(Spectrum_error,{0,1},Mix)-0.5*mix(Spectrum_error,{-1,2},Mix)

Eqn TOlinput_low_error=TOloutput_low_error-P_gain_transducer_error

Eqn TOlinput_high_error=TOloutput_high_error-P_gain_transducer_error

Eqn FifthOloutput_low_error=1.25*mix(Spectrum_error,{1,0},Mix)-0.25*mix(Spectrum_error,{3,-2},Mix)

Eqn FifthOloutput_high_error=1.25*mix(Spectrum_error,{0,1},Mix)-0.25*mix(Spectrum_error,{-2,3},Mix)

Eqn FifthOlinput_low_error=FifthOloutput_low_error-P_gain_transducer_error

Eqn FifthOlinput_high_error=FifthOloutput_high_error-P_gain_transducer_error

Eqn P_gain_transducer_error=Perror_dBm-RFpower

Eqn Spectrum_zoomed_error=mix(Spectrum_error,tones,Mix)

Eqn Perror_W1=mix(Spectrum_W_error,{1,0},Mix)

Eqn Perror_W2=mix(Spectrum_W_error,{0,1},Mix)

Eqn Perror_dBm=10*log(Perror_W1 +Perror_W2)+30

The formulas for predistorted signal spectrum.

```
Eqn Spectrum_W_pre=0.5*real(Vpre*conj(Ipre.i))
Eqn Spectrum_pre=10*log(Spectrum_W_pre) +30
Eqn TOIoutput_low_pre=1.5*mix(Spectrum_pre,{1,0},Mix)-0.5*mix(Spectrum_pre,{2,-1},Mix)
Eqn TOIoutput_high_pre=1.5*mix(Spectrum_pre,{0,1},Mix)-0.5*mix(Spectrum_pre,{-1,2},Mix)
Eqn TOIinput_low_pre=TOIoutput_low_pre-P_gain_transducer_pre
Eqn TOIinput_high_pre=TOIoutput_high_pre-P_gain_transducer_pre
Eqn FifthOIoutput_low_pre=1.25*mix(Spectrum_pre,{1,0},Mix)-0.25*mix(Spectrum_pre,{3,-2},Mix)
Eqn FifthOIoutput_high_pre=1.25*mix(Spectrum_pre,{0,1},Mix)-0.25*mix(Spectrum_pre,{-2,3},Mix)
Eqn FifthOIinput_low_pre=FifthOIoutput_low_pre-P_gain_transducer_pre
Eqn FifthOIinput_high_pre=FifthOIoutput_high_pre-P_gain_transducer_pre
Eqn P_gain_transducer_pre=Ppre_dBm-RFpower

Eqn Spectrum_zoomed_pre=mix(Spectrum_pre,tones,Mix)
Eqn Ppre_W1=mix(Spectrum_W_pre,{1,0},Mix)
Eqn Ppre_W2=mix(Spectrum_W_pre,{0,1},Mix)
Eqn Ppre_dBm=10*log(Ppre_W1 +Ppre_W2)+30
```

The formulas for output spectrum.

```
Eqn Spectrum_W=0.5*real(Vload*conj(Iload.i))
Eqn Spectrum=10*log(Spectrum_W) +30
Eqn TOIoutput_low=1.5*mix(Spectrum,{1,0},Mix)-0.5*mix(Spectrum,{2,-1},Mix)
Eqn TOIoutput_high=1.5*mix(Spectrum,{0,1},Mix)-0.5*mix(Spectrum,{-1,2},Mix)
Eqn TOIinput_low=TOIoutput_low-P_gain_transducer
Eqn TOIinput_high=TOIoutput_high-P_gain_transducer
Eqn FifthOIoutput_low=1.25*mix(Spectrum,{1,0},Mix)-0.25*mix(Spectrum,{3,-2},Mix)
Eqn FifthOIoutput_high=1.25*mix(Spectrum,{0,1},Mix)-0.25*mix(Spectrum,{-2,3},Mix)
Eqn FifthOIinput_low=FifthOIoutput_low-P_gain_transducer
Eqn FifthOIinput_high=FifthOIoutput_high-P_gain_transducer
Eqn P_gain_transducer=Pload_dBm-RFpower

Eqn Spectrum_zoomed=mix(Spectrum,tones,Mix)
Eqn Pload_W1=mix(Spectrum_W,{1,0},Mix)
Eqn Pload_W2=mix(Spectrum_W,{0,1},Mix)
Eqn Pload_dBm=10*log(Pload_W1 +Pload_W2)+30

Eqn mix_1=[{1,0},{0,1}]
Eqn mix_3=[{1,0},{0,1},{2,-1},{-1,2}]
Eqn mix_5=[{1,0},{0,1},{2,-1},{-1,2},{3,-2},{2,-3}]
Eqn mix_7=[{1,0},{0,1},{2,-1},{-1,2},{3,-2},{2,-3},{4,-3},{3,-4}]
Eqn tones=if (Max_IMD_order[0]<2) then mix_1 elseif (Max_IMD_order[0]<4) then mix_3 elseif (Max_IMD_order[0]<6) then mix_5 else mix_7
```