NTNU
Norwegian University of
Science and Technology

# Design of Wide-Band Supply Modulators for RF Power Amplifiers and Testing With LTE

Håkon Haugen

Master of Science in Electronics
Submission date: June 2018
Supervisor: Morten Olavsbråten, IES

Norwegian University of Science and Technology
Department of Electronic Systems

# Summary

In this thesis the effect of tracked RF amplifiers on modern LTE waveforms are studied. A tracked amplifier is an amplifier with a modulated supply voltage that track some wanted function of the base-band. LTE uses the particularly demanding OFDM modulation scheme on its down-link. Tracking amplifiers use a RFPA with modulated supply voltage to lower the voltage in parts of the transmission where the transmit power is low. Different tracking functions are compared both in simulation and with real results.

The design of high bandwidth trackers are also studied, with two trackers built and tested. The trackers use high frequency design techniques and deliver up to 100 MHz of usable tracker bandwidth. A design based on cascaded current-feedback operational amplifiers are used. Two THS3001 amplifiers deliver a combined gain of 86 V/V and a voltage buffer output stage with either 6x THS3001 or a single ADA4870 provide around 1 A of current.

The THS3001 tracker (the design that provided highest bandwidth) is used to test LTE performance with a RFPA based on a 10 W Gan HEMT. Tests are conducted with 10 and 20 MHz LTE channels and corresponding tracker waveforms with up to 40 MHz bandwidth.

The tracker functions tested are; max PAE, constant gain, power envelope and power envelope with a higher order polynomial. The PET and constant gain provide high linearity while the 2nd order PET and max PAE provide high efficiency. For LTE it is shown that power envelope tracking is the superior technique for modulating the supply voltage.

# Sammendrag

I denne oppgaven studeres effekten av modulert forsyning på radioforsterkere med moderne LTE signaler. En forsterker med modulert forsyning kan forbedre effektivitet og/eller linearitet ved at forsyningsspenningen følger en tilpasset funksjon av inngangssignalets basebånd. LTE setter ekstra store krav til effektforsterkere fordi ortogonalt frekvensdelt multipleksing benyttes. Det er en modulasjonsteknikk med stor forskjell mellom topp og gjennomsnittseffekt. Forsterkere med modulert forsyning vil da få redusert forsyningsspenning når effektbehovet er lavt. Overføringsfunksjonen som regner ut forsyningspenningen kan lages på utallige måter. I denne oppgaven blir et fåtall funksjoner simulert og sammenlignet med målinger på et ekte system.

Videre blir design av selve forsyningsmodulatoren studert. To slike forsyningsmodulatorer lages og testes. De er designet med omhu for bruk ved høye frekvenser og leverer opptil 100 MHz brukbar båndbredde. To THS3001 operasjonsforsterkere med strømsensitiv tilbakekobling er brukt for åoppnå 86 V/V forsterkning. Utgangssteget er laget enten med seks THS3001 eller en enkelt ADA4870 og kan levere rundt 1 A strøm ut til radioforsterkeren.

Forsyningsmodulatoren basert utelukkende på THS3001-forsterkere ble satt til å drive en GaN radioforsterker for å teste systemets ytelse på nevnte LTE signaler. Testene er utført med 10 og 20 MHz brede LTE signaler og medfølgende moduleringsfunksjoner som er begrenset oppad til 40 MHz.

De testede moduleringsfunksjonene er; maksimal tilført effektivitet, konstant forsterkning, effektommhylningskurven og effektommhylningskurven med høyere ordens polynom. Konstant forsterkning og effektomhylningskurven brukes for å oppnå høy linearitet. maksimal tilført effektivitet og effektommhylningskurven med høyere ordens polynom brukes for å oppnå høy effektivitet. Det viser seg at for bruk med LTE så er effektommhylningskurven overlegen de andre funksjonene som er testet.

# Table of Contents

# List of Tables

# List of Figures

# Abbreviations

| | | |
|---|---|---|
| ACPR | = | Adjacent Channel Power Ratio |
| ADC | = | Analog to Digital Converter |
| ADSL | = | Asymmetric Digital Subscriber Line |
| AWG | = | Arbitrary Waveform Generator |
| AWGN | = | Additive White Gaussian Noise |
| BB | = | Baseband |
| BPSK | = | Bipolar Phase Shift Keying |
| CDMA | = | Code Division Multiple Access |
| DAC | = | Digital to Analog Converter |
| DUT | = | Device Under Test |
| EVM | = | Error Vector Magnitude |
| FDD | = | Frequency Division Duplex |
| FFT | = | Fast Fourier Transform |
| FOM | = | Figure Of Merit |
| GaAs | = | Gallium Arsenide |
| GaN | = | Galium Nitride |
| GSM | = | Global System for Mobile |
| HEMT | = | High Electron Mobility Transistor |
| I | = | In-phase |
| IC | = | Integrated Circuit |
| IF | = | Intermediate Frequency |
| IFFT | = | Inverse Fast Fourier Transform |
| LTE | = | Long Term Evolution |
| MAE | = | Mean Absolute Error |
| MIMO | = | Multiple Input Multiple Output |
| NPR | = | Noise Power Ratio |
| OFDM | = | Orthogonal Frequency Division Multiplexing |
| OFDMA | = | Orthogonal Frequency Division Multiple Access |
| OOK | = | On Off Keying |
| PAE | = | Power Added Efficiency |
| PSD | = | Power Spectral Density |
| Q | = | Quadrature |
| (X)QAM | = | Quadrature Amplitude Modulation (X = constellation size) |
| QPSK | = | Quadrature Phase Shift Keying |
| RB | = | Resource Block |
| RF | = | Radio Frequency |
| RFPA | = | Radio Frequency Power Amplifier |
| R&S | = | Rohde & Schwarz |
| SC-FDMA | = | Single Carrier Frequency Division Multiple Access |
| SSB | = | Single Side Band |
| SSB-SC | = | Single Side Band Suppressed Carrier |
| TDD | = | Time Division Duplex |
| Telecom | = | Telecommunication |

Chapter 1

# Introduction

The modern society relies on the internet, and an increasing portion of internet traffic ends up in smart-phones. A key enabler of mass-data transfer to multiple mobile consumers is the communications scheme LTE, commonly advertised as 4G. LTE is the most modern standard supported by handsets and has proven to be so superior to 3G that Telenor (the largest telecom operator in Norway) decided to replace the entire 3G infrastructure (3). Such an upgrade is expensive and solely undertaken to make the most out of the limited spectrum won at government auctions.

The drawback with LTE is the high crest-factor that comes with it. The crest factor is usually well above 10 dB. This means that a powerful amplifier that can handle the peak powers is needed. If such an amplifier is to be made linear enough, it will waste most of the energy as heat. However, some techniques can lower the waste of energy and thereby help create cheaper base station amplifiers. Doherty and out-phasing amplifiers can do the job, but the technically elegant solution of modulating the supply is what this thesis will look closer into.

The concept of modulating the supply voltage of an RF amplifier along with the signal envelope has been around since the early days of AM modulation. In AM radio transmitters the amplifier can be smaller because less power is wasted as heat in the transmitter tube(s). When frequency modulation came around amplifiers could be highly non-linear, and could be operated at, or close to saturation, all the time with a constant transmit power. Constant envelope was also a decisive factor to why GMSK is used in the GSM mobile communication scheme (also known as 2G). In modern communication systems however, high spectral efficiency is required and obtained by modulating both phase and amplitude. The most modern system in use today is the LTE communications scheme. In LTE a multitude of carriers are stacked alongside each other (up to 1201 in downlink OFDMA) and modulated at up to 64 QAM. This results in the mentioned high crest factor and low-efficiency power amplifiers. LTE is explained further in the theory section.

A superb way of increasing efficiency would be to use envelope tracking. This, however, is not practical as the bandwidth needed to track the envelope would be way too high for any efficiency improvements (usually ca.5 times baseband bandwidth). M. Olavsbråten have pioneered a way of modulating the supply voltage that increases both linearity *and* efficiency. The primary method is to track the power envelope (PET), as the bandwidth is substantially reduced to around baseband levels. In addition, the baseband can be divided into multiple sections, and the sum of these envelopes can be used instead. A 2nd-degree function can also be used to track more closely to the envelope while doubling the theoretical tracker bandwidth.

The goal of this thesis is to insert LTE signals with different bandwidths and spacing to an RF amplifier driven with modulated supply voltage and evaluate RFPA performance. Tests are conducted with an amplifier and supply modulator explicitly developed for this purpose.

## 1.1 Supply modulator (tracker)

The supply modulator is, in fact, an amplifier that outputs a drive voltage to the RF amplifier. It is commonly referred to as the ”tracker” as it tracks the envelope or any other *tracker function*. When the baseband signal is produced, one has the in-phase and quadrature signals available to create an auxiliary output to the tracker. Using the absolute value would yield pure envelope tracking, but this is the best option only with an ideal RFPA and tracker. Instead, the I/Q signals are used as inputs to a transfer function that can be tailor-made to achieve better efficiency, better linearity and even both! For this thesis, a custom wide-band tracker is designed and built to test wide-band performance with different tracker functions. The designed tracker is linear and thus will not provide the best possible efficiency as the waste heat is simply transferred from the RFPA to the tracker.

## 1.2 RF power amplifier

The RF amplifier is based upon CGH40010, a 10 W GaN transistor from Wolfspeed. The critical aspect that sets this amplifier apart from an off-the-shelf amplifier is the Small 10 pF drain capacitor. If a large reservoir capacitor had been coupled directly on the drain, it would need continuous charging and discharge along with the tracker voltage. Aside from a small drain capacitor, it is a pretty standard wide-band RF amplifier with 540 MHz -1dB bandwidth centered at 2.6 GHz.

## 1.3 Tracker functions

The tracker functions that are used in this thesis is PET, 2nd order PET, maximal PAE and constant gain. These are compared to each other and constant voltage as a reference. How they are derived, and further details are located in the theory section.

# Chapter 2

# Theory

Within this chapter theory on envelope tracking, bandwidth reduction techniques and modern mobile communication systems are presented. This provides background for the thesis as envelope tracking with reduced bandwidth is applied to a PA fed with LTE signals. Also, some figures of merit used during the test are explained.

## 2.1 Envelope tracking

Envelope tracking is a way of improving efficiency in RF power amplifiers. By continuously adapting the supply voltage of the amplifier to the signal envelope, efficiency can increase. The level of efficiency increase is dependant on how far the given amplifier is into back-off. Consider now two ideal RF amplifiers, one with constant supply voltage and another with its supply modulated. If operating at max power output, there is no difference as both are operating at 0 dB back-off. When moving further and further into back-off, the efficiency stay constant and dissipated waste heat decrease in the tracked amplifier. On the constant supply amplifier however, the dissipated waste heat increase and efficiency plummet. When the amplitude change over time, the efficiency gain is dependant on signal characteristics. Higher crest-factor makes for a higher efficiency gain with tracking. Pure envelope tracking is illustrated in figure 2.1 and 2.2 below.

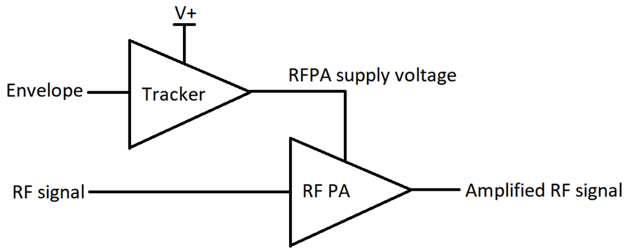



**Figure 2.1:** Envelope tracking illustrated

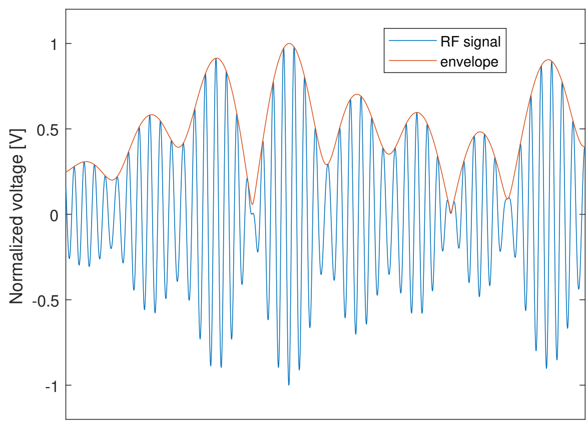



**Figure 2.2:** RF signal and its envelope

Although envelope tracking is an elegant solution, it is not practical with non-ideal amplifiers. First of all, the amplifier gain is dependant on compression level and supply voltage. Especially on GaN amplifiers gain is dramatically reduced. On the RFPA made for this thesis, gain goes from 15 to 5 dB when adjusting the voltage from 28 to 4 volts. The transfer function between gain and supply voltage is not linear either, further complicating the situation.

Additionally, the 180-degree phase shifts caused by zero-crossings makes sharp spikes in the time-domain envelope waveform. Sharp spikes in the time domain result in a wide spectrum in frequency domain. Numerically an envelope to a zero-crossing signal has infinite bandwidth(4). This property is a consequence of the square-root function in the calculation of absolute(envelope) value.

$$v_{signal}(t) = v_I(t) + j \cdot v_Q(t) \tag{2.1}$$

The base band signal is complex and composed of an in-phase and quadrature waveform (cartesian). The polar radius is the signal envelope.

$$v_{envelope}(t) = |v_{signal}(t)| = \sqrt{v_I(t)^2 + v_Q(t)^2} \tag{2.2}$$

The square-root function has no standard Fourier transform and provided the signal crosses zero, have an infinite bandwidth (4). To see the spectral properties of the envelope, it is better to use a reference signal and perform a DFT on the absolute value of the base band waveform. Below in fig.2.3 the power spectral density of the envelope of a 20 MHz wide LTE signal is presented. If one set the limit for what can be called the bandwidth at -50 dB from the DC component, then the bandwidth increases four times compared to the base band (shown for comparison).

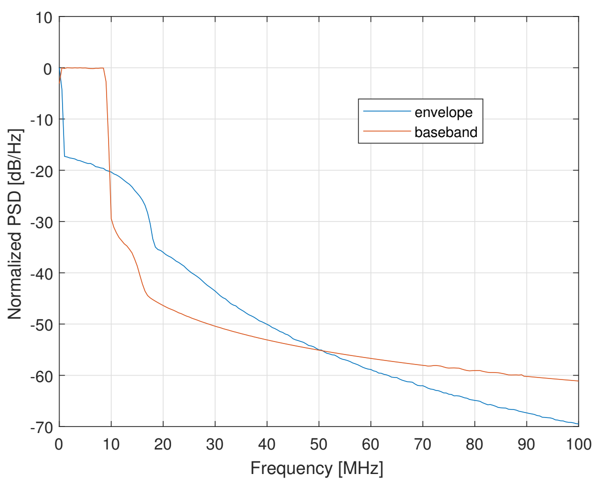



**Figure 2.3:** Power spectral density of 20 MHz LTE signal and its envelope, normalized with negative frequencies not shown

In practice, the tracker bandwidth needed is four to eight times higher than RF bandwidth(4)(5). With a 20 MHz LTE signal, this would result in a tracker bandwidth between 80 and 160 MHz. For comparison, conventional buck converters usually switch at a couple of 1 MHz or below. An efficient switching tracker with 100MHz bandwidth is not available on the market today, and would be a challenge to make with today's technology.

## 2.2 Complex modulation

The most basic modulation scheme aside from OOK is AM where the signal amplitude (envelope) is modulated in a non-optimal way for sending information. With pure AM, the occupied bandwidth is double the highest baseband frequency and most of the RF energy is spent in the carrier. With SSB or SSB-SC, this is reduced but at the cost of complexity. This reduction is accomplished in a variety of ways but at the core is the utilization of phase information.

In the digital domain, the same thing is accomplished in quadrature modulation. Instead of using a single varying voltage modulated with information that subsequently is up-converted and sent, two of these varying voltages is up-converted and summed, but in quadrature (90°out of phase). When up-converted these baseband signals are orthogonal, real and summed to a single RF waveform. To RF this would be like modulating the carrier with a complex vector changing in both phase and frequency. This way the double amount of information transmitted within the same bandwidth.

The two waveforms, Inphase(I) and quadrature(Q) can be coded with any number of voltage ranges where the ability to differ accurately enough (overcome random noise in the channel) is the decisive factor to how much information that can be carried by the channel. In other words, using complex modulation one can maximize the information transfer in any available AWGN channel with limited bandwidth. In modern communication schemes, the constellation sizes usually are (WI-FI and LTE) in the range from BPSK (two levels not complex) up to 64 QAM (8 levels on both I and Q). A set of I and Q values comprised is a symbol.

## 2.3 LTE

LTE is the most modern widely deployed mobile communication scheme. The focus has been increased spectral efficiency to provide large downlink data rates to as many connected devices as possible. A key technology in LTE is OFDM and the related OFDMA.

OFDM is the use of a range of sub-carriers stacked along each other in the spectrum, each modulated with complex modulation up to 64 QAM. Each carrier carries less information to have a bandwidth that doesn't distort the next carrier. This slicing and dicing of the spectrum has the added advantage of being more adaptable to fading dips because carriers with bad reception can be modulated with lower complexity. The receiver equalizer is also simplified as FFT is used to demodulate the baseband waveforms and each carrier can individually be multiplied by a complex vector to compensate for a varying fade(magnitude), Doppler spread(phase) and oscillator offset(phase).

While OFDM is implemented to battle fading dips, OFDMA is implemented alongside for flexibility, assigning different sub-carriers to different users. A user with low demands of bandwidth can have few assigned carriers while low demanding users can have more carriers assigned. In LTE carriers are divided into resource blocks(RBs) which are 12 sub-carriers 15 kHz wide and modulated with 7 symbols. Users are assigned RBs with regards to the data rate needed on channels up to 100 RBs wide in a 20 MHz spectrum.(6)

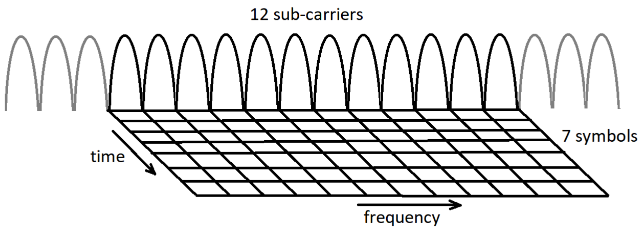



**Figure 2.4:** Illustration of one resource block

LTE can operate in both FDD and TDD modes of duplex (frequency and time respectively). When considering TDD over time, one can argue that with low down-link demand, the crest-factor will become even higher because of large amounts of time where the radio is inactive. This increases the motivation for the use of tracked amplifiers even further. The spectrum of a real 10 MHz wide LTE signal is shown in figure 2.5 recorded with a HackRF One(7) and SDR#(8).

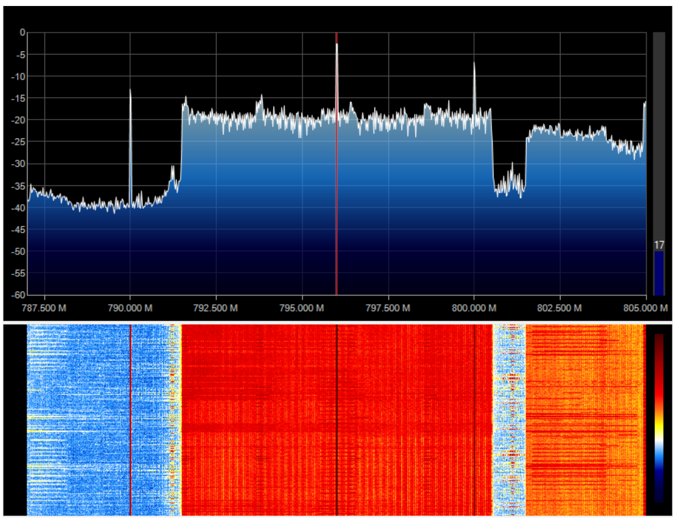


**Figure 2.5:** 10 MHz LTE at 796 MHz, spectrum snapshot and waterfall

LTE also has support for MIMO and it is key to achieve the highest data rates, with technologies like spatial multiplexing. This is of no concern to the design of the PA and will not be discussed further.

## 2.4 Efficiency

When measuring efficiency in an amplifier, it is common to use either *drain efficiency* or *power added efficiency*. Commonly denoted as $\eta_D$ and $PAE$ respectively. Drain efficiency is defined as output power divided by consumed supply power. Drain efficiency does not take into account the input power which PAE does, and it can therefore be considered a more describing FOM. They are expressed mathematically below in equations 2.3 and 2.4. Both figures are usually expressed in percentage.

$$\eta_D = \frac{P_{RFout}}{P_{DC}} \cdot 100\% \tag{2.3}$$

$$PAE = \frac{P_{RFout} - P_{RFin}}{P_{DC}} \cdot 100\% \tag{2.4}$$

## 2.5 Tracker functions

The large drawback with envelope tracking is "zero-crossings" as mentioned in section 2.1. In classical audio AM transmission, this is not a problem as the carrier is not modulated beyond 100% and the envelope will never cross through zero. In all modern communication systems, this is not the case, and zero-crossings are frequent.

To reduce the tracker bandwidth requirement, other transfer functions than the directly amplified signal envelope is needed. Even though a supply modulating function other than direct tracking of the envelope would not be defined as envelope tracking in a strict sense, it is considered descriptive and a "tracker function" may refer to a supply modulation function where the envelope or base-band functions are inputs. These tracker functions can be tailored to meet specific design criteria like maximal PAE, constant gain, simplicity or a given bandwidth. Four tracker functions are studied in this thesis and presented in subsections below.

### 2.5.1 Max PAE

PAE stands for Power Added Efficiency and describes how effectively the amplifier adds signal power compared to the supplied power. With a constant envelope input, max PAE is achieved at a specific level of compression (normally pretty deep into compression). By characterizing the RF amplifier, a set of max PAE points for each supply voltage is found. These points are then mapped to corresponding input levels, and a polynomial is made to fit as close as possible and generate a tracker function. In this thesis, an 8th-degree polynomial is used. A lower limit of 8 volts is set to prevent a total collapse of gain.

### 2.5.2 Constant gain

As with the maximal PAE case, one can seek to have the same level of gain for every input power. The same procedure is performed as with max PAE, mapping gain levels for given input powers and supply voltages to input levels. Again an 8th-degree polynomial is fitted to these data points to create a tracker function.

### 2.5.3 Power envelope tracking

It is favorable to have a properly defined tracker bandwidth and in a paper(4) by M.Olavsbåten and D.Gecan it is proposed to use the power envelope instead. The concept is that the RF power bandwidth is well defined and strictly limited by standards and laws. So why track at higher bandwidths? It turns out it is only feasible to track down to $V+/2$. If pushed further the power envelope will clip off some of the required voltage envelope. This results in a technique that is less efficient than pure envelope tracking but with the same required tracking bandwidth as the RF bandwidth. easily seen from the power envelope equation2.5 below.

$$P_{envelope}(t) = |v_{signal}(t)|^2 = \sqrt{v_I(t)^2 + v_Q(t)^2}^2 = v_I(t)^2 + v_Q(t)^2 \tag{2.5}$$

The actual waveform applied is limited to $V_{dmax}/2$ and the applied waveform is shown in equation2.6 below.

$$v_d(t) = V_{dmax}\frac{v^2_{signal}(t)+1}{2} \qquad 0 \leqslant |v_{signal}| \leqslant 1 \tag{2.6}$$

### 2.5.4 2nd order Power envelope tracking

Power envelope tracking can be extended to increase slew-rate (and the needed bandwidth) to achieve ”deeper” tracking and thereby better efficiency. The only modification is to add the square of the power envelope. This introduces another degree of freedom that can be used to track deeper and increase efficiency.

$$v_d(t) = V_{dmax}a_0 + a_2v^2_{signal}(t) + a_4v^4_{signal}(t) \qquad 0 \leqslant |v_{signal}| \leqslant 1 \tag{2.7}$$

In this thesis, 8 V is used as a lower bound for both the 2nd degree PET and max PAE. the ”a” factors used are derived below in equations 2.8. The factors can be different then these, and can be made to approach desired properties other then efficiency.

$$a_0 = V_{dmin} = 8 \qquad a_2 = 1.5(V_{dmax} - a_0) \qquad a_4 = V_{dmax} - a_2 - a_0 \tag{2.8}$$

## 2.6 Measurement figures of merit

Presented here is key figures of merit in regards to evaluating RFPA performance. Evaluation of PA performance in regards to linearity is not trivial, especially in LTE where hundreds of carriers are stacked alongside each other. The primary focus is, of course, to measure how well the amplifier will perform in a real environment. The transmitter is limited in regards to ACPR, which is a measure of how much power ends up in neighboring bands. Noise within band is measured with NPR. Another critical parameter is inter-symbol and inter-sub-carrier interference. These are harder to measure and cannot be evaluated merely by studying the emitted spectrum. This is why EVM and MAE are used.

### 2.6.1 EVM

EVM is the error vector divided by the symbol amplitude. Normally this is computed at every symbol. Computing EVM at every symbol is harder with OFDM signals as each carrier would need demodulation and an EVM for all those parallel symbols would be calculated individually. Alternately every base-band-IQ sample could be evaluated as a vector and a corresponding error vector to the received signal when aligned. This is what is done in measurements presented in section 4.4.4. Lower EVM makes for better performance and a larger eye opening at decision devices. It is normal to present EVM in percentage (%), and if there is more than one symbol evaluated, the RMS error vector and reference vector is used. EVM is expressed in equation 2.9 where the RMS error vector is numerator(ideal symbol - measured symbol), and the RMS reference vector is denominator(ideal symbol). EVM has one shortcoming though, it disproportionately overvalues single large deviations due to the squaring function and is why MAE is also computed.

$$EVM_{RMS} = \sqrt{\frac{\frac{1}{N}\sum_{n=1}^{N}|S_{ideal}[n]-S_{meas}[n]|^2}{\frac{1}{N}\sum_{n=1}^{N}|S_{ideal}[n]|^2}} \tag{2.9}$$

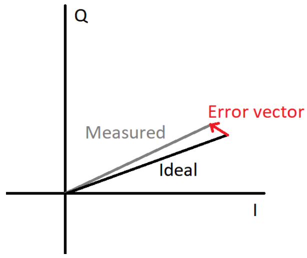



**Figure 2.6:** Illustration of measured EVM on a vector

### 2.6.2 MAE

As mentioned, EVM will weigh large error vectors more than small. That is not necessarily bad or not descriptive of performance, but with non-infinite measured waveforms, single samples can be too decisive when comparing EVM results. The closely related MAE(Mean Absolute Error) is therefore also used (9). Here the sent and received baseband waveforms are compared sample by sample and the mean error vector is calculated. To ensure the MAE results are correct, compared waveforms are normalized to have the same RMS amplitude before the MAE calculation to mimic a real receiver. The MAE calculation is shown below in equation 2.10

$$MAE = \frac{\sum_{n=1}^{N} |s_{ideal}[n] - s_{meas}[n]|}{N} \qquad RMS(S_{ideal}) = RMS(S_{meas}) = 1 \quad (2.10)$$

### 2.6.3 ACPR

The ACPR is used to measure out-of-band impairments. It is accomplished by simply integrating and comparing the spectrum in the adjacent channels and within the channel. The result is a lower and a higher adjacent channel power ratio.

### 2.6.4 NPR

Noise measured with ACPR is out of band. For measurements of in-band noise, NPR is used. A null-notch in the transmitted spectrum is introduced and the level within this notch is compared to the level outside the notch.

## 2.7 Two channel signal

One of the primary goals for this thesis is to measure the performance of different tracker functions with high baseband bandwidths. Like for example with two LTE channels with a spacing between them. If implemented in a real base station this would probably have been generated by having two base-bands up-converted with two carriers and summed before the RFPA. In the lab, an SMU200A signal generator with only a single up-converter is available. There are two primary solutions to this problem.(10)

The simplest way is to treat the generator base-band as an IF, up-convert a single base-band waveform to half of the wanted carrier spacing in MATLAB and then up-convert IF to RF. This is illustrated below in figures 2.7, 2.8 and 2.9.


**Figure 2.7:** Base-band in frequency domain


**Figure 2.8:** Baseband up-converted to IF in MATLAB


**Figure 2.9:** IF up-converted to RF in generator, negative frequencies not shown

Mathematically expressed below in equation 2.11 where $s_{BB}$ is the complex base-band in time domain and $S_{BB}$ is the same signal in frequency domain.

$$s_{BB} \cdot cos(2\pi f_{IF}t) \Leftrightarrow \frac{S_{BB}(f - f_{IF}) + S_{BB}(f + f_{IF})}{2} \qquad s_{BB}(t) \Leftrightarrow S_{BB}(f) \tag{2.11}$$

The other way is to use complex up-conversion in the IF and up-convert one base-band waveform to positive frequency and another base-band waveform to negative frequency.

That is, multiplying with a complex single sided frequency.

$$F(e^{(j2\pi f_{IF}t)}s_{BB}(t)) \Leftrightarrow S_{BB}(f - f_{IF}) \qquad s_{BB}(t) \Leftrightarrow S_{BB}(f) \tag{2.12}$$

When up-converted to RF this would yield much of the same spectral components if both base-band waveforms are modulated and filtered in the same way. The pro with this technique is that the two channels in RF would contain two different signals, like in the real world. The con is that these two would only in theory be orthogonal, and can intermodulate in IF up-conversion (and down-conversion at the signal analyzer) if these two functions are placed at an equal offset. If set on different offset frequencies the image frequencies would be in other regions of the spectrum but would require additional bandwidth in the signal generator. The simple IF up-conversion without suppression technique is therefore used when multiple channel LTE is generated.

Chapter 3

# Design of supply modulator (Tracker)

As mentioned in section 1.1 the supply modulator, or tracker, needs to provide varying voltage to the RF power amplifier and can be viewed as an amplifier itself. The important design parameters are the dynamic range, phase response, amplitude response and the S11 reflection viewed ”into” the tracker. The amplitude response is what determines the bandwidth of the amplifier where bandwidth is defined as the frequency with a 3 dB drop in amplitude. Phase response may also limit the bandwidth by introducing intermodulation and what level is acceptable may vary. However phase alone is easier to suppress with predistortion. The S11 parameter describes what the amplifier will see at the supply at RF and may decide if the RFPA is stable or not. In table3.1 below specification targets are presented.

**Table 3.1:** Initial Requirements

| Parameter | Target | Unit |
|---|---|---|
| Bandwidth (Ampl.) | $> 50$ | MHz |
| Bandwidth (Phase) | $> 50$ | MHz |
| Gain | $> 40$ | V/V |
| I out | $> 0.6$ | A |
| $\mid S_{11} \mid$ | <0 | dB |
| Dynamic range | 10 - 28 | V |
| Oscillate with amplifier as load | No | |
| Keep amplifier stable | Yes | |

## 3.1 Design

The modulator needs to behave like a high-frequency DC-coupled amplifier with relatively low output impedance, and when coupled to the RF amplifier - not oscillate. It is designed with a gain stage that can achieve high enough output voltage, and an output stage that buffers this voltage. When providing full RF output power, the RF amplifier will need 28 V and drain around 600 mA. This translates to ca. 50 Ω which is used as a test load. 50 Ω is also practical as a 50 Ω terminated coaxial cable (with 50 Ω characteristic impedance) will be ”seen” as purely resistive. Two output stage designs are evaluated.

## 3.2 Gain stage

The gain stage is needed to take the tracking waveform from around a volt to around 30 volts that drive the amplifier. For testing with envelope tracking, a signal generator with two IQ base-band generators and a single up-converter is used. Baseband A drives the up-converter with I and Q signals while base-band B’s I channel is fed to the tracker. That way, a base-band waveform and tracking waveform can be synthesized and synchronized. The output voltage is -1 to 1 V with 50 Ω output impedance. If the Q channel is also to be used, this is reduced to 0.7 V because the vector sum max is 1. With input termination of 50 Ω at the tracker, this voltage is further cut in half. The total gain therefore needs to be at least 80 V/V.

$$G = 28/0.7/2 \approx 80 \tag{3.1}$$

For simplicity and high-frequency performance, Texas instruments THS3001 series current-feedback op-amps are used. Current feedback operational amplifiers do not suffer from the gain-bandwidth-product limitation like conventional operational amplifiers. Instead, the bandwidth is primarily decided by the feedback resistance with a small dependency on the gain. Having operational amplifiers that can have G=10 and a 100 MHz bandwidth would require a regular voltage feedback op-amp with GBP of 1 GHz. The THS3001 also have an unmatched slew-rate of up to 6500 $V/\mu s$ s. The THS3001 family is therefore an ideal fit.

The gain-stage is comprised of two cascaded op-amps, each with a gain of +9.3, together providing a total of 86.5 V/V. With the classic inverting op-amp design the inverting input remains at virtual ground, and so practically all the error current goes directly into the inverting input. With the non-inverting design, the changing voltage at the inverting input makes some of the error current go through parasitic capacitance to ground. As a result, higher bandwidths can be achieved with inverting design. But the inverting topology has some drawbacks. Because of rail voltage limitations, it is impractical to have negative output voltages in the first gain stage. (The amps should not be used within 3.2 V of the rails). Additionally, more power is dissipated in the feedback resistor, as the entire output voltage is applied to the feedback resistor. Non-inverting architecture is therefore chosen. The 1 KΩ feedback resistor in the 2nd amplifier must be able to handle 28 V output continuously. That is $P = U^2/R = (28 - 28/9)^2/1000 = 0.62W$ therefore the feedback goes through two parallel coupled 2K resistors each rated for 0.5 W. below in figure 3.1 the schematics for the gain stage are shown.

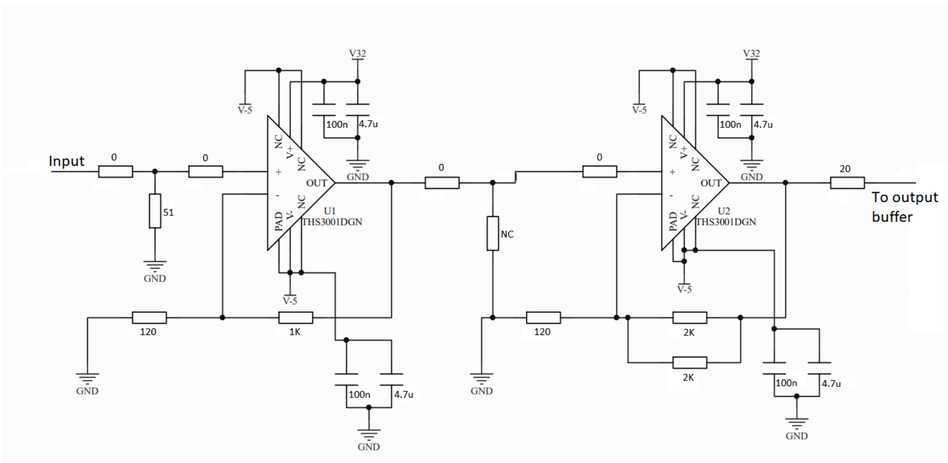



**Figure 3.1:** Tracker gain stage schematics

In layout, high-frequency design techniques are implemented. The input is 50 Ω, traces are kept short, via-stitching along sensitive signal paths, the thermal pad of the chips are at V- potential and ground plane removed from underneath the inputs. Passive components are in 0603' packages and the THS3001 chips in DGN-8 (PDSO-G8) packages. The complete tracker gain stage is presented below in figure 3.2.

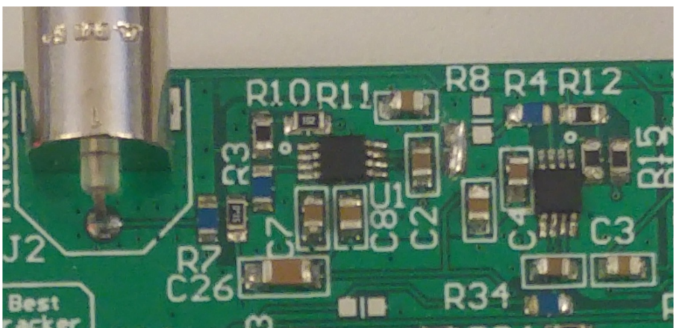



**Figure 3.2:** Tracker gain stage

## 3.3 Output stage

A tracker designed by M. Olavsbråten used LT1210 as the output stage, an operational power amplifier designed to drive video signals and ADSL lines. It is also a current feedback operational amplifier, and it is defiantly an elegant solution, but other output stages can provide higher bandwidths. To achieve good performance and to investigate different tracker designs, multiple output stage designs are used and described in section 3.3.1, 3.3.2 and 3.3.3.

### 3.3.1 LT1210

This amplifier chip only has a 35 MHz bandwidth(11) and does not suffice as it is no challenge to go further. Still, a DDPAK-7 version is implemented in a tracker and included for comparison, and It will not be discussed any further in length or detail.

### 3.3.2 Six THS3001 in parallel

With each of the THS3001 capable of delivering 120 mA, a total of six is sufficient to drive the RF amplifier. Each one of them has individual feedback through a 1 KΩ resistor and separated with 6.2 Ω output resistors. That way no significant current is lost to the feedback network (as would be the case with gain other than 1). Below is the output stage schematic in figure 3.3.

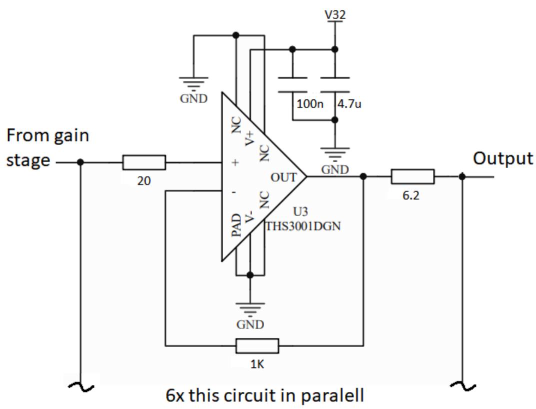



**Figure 3.3:** THS3001 output buffer

As with the gain stage, high-frequency design is also implemented here. Via-stitching is frequent, and the input signal is fed along the array of output amplifiers while the output is gathered from along the array from the opposite side. This is to make the electrical distance equal through all six paths and thereby make the amplifiers feed the same signal in phase. This is illustrated in figure 3.4 and the complete output stage is shown in figure 3.5.

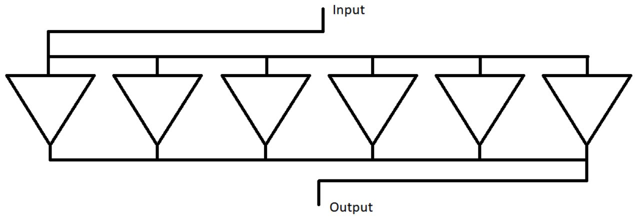



**Figure 3.4:** Tracker output stage feeding

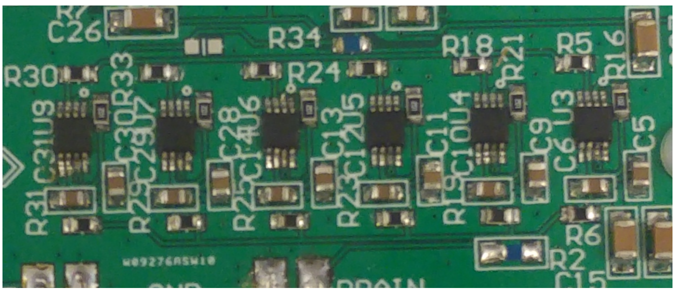


**Figure 3.5:** Tracker output stage with 6x THS3001

### 3.3.3 ADA4870

ADA4870 is a nimble circuit with an ability to drive over 1A and with smart packaging providing a nice feedback path and excellent thermal management. The feedback network is realized with only a single resistor from the dedicated feedback output and inverting input. Different power control and management functions are provided through jumper connections. Output stage schematics and layout are presented in figure 3.6 and 3.7 respectively.

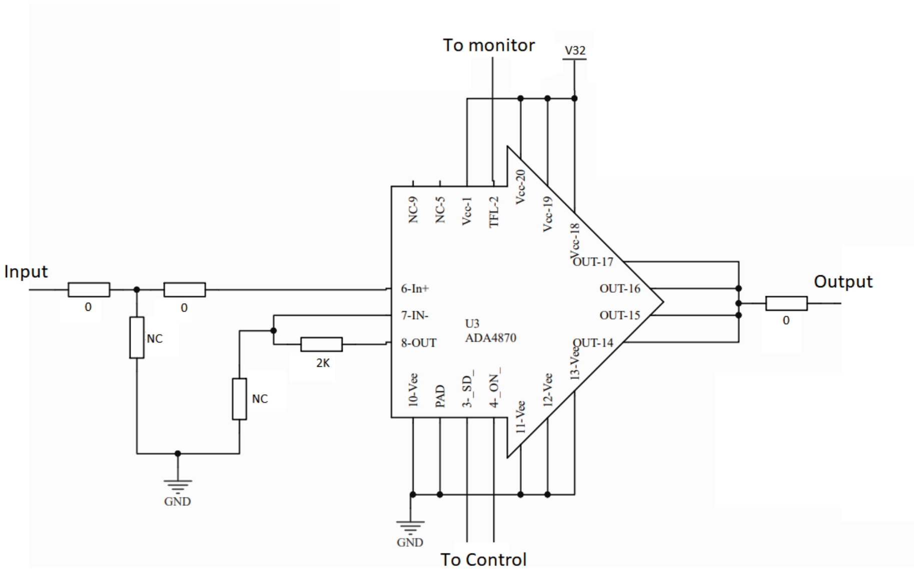



**Figure 3.6:** Tracker schematics with with ADA4870 output stage

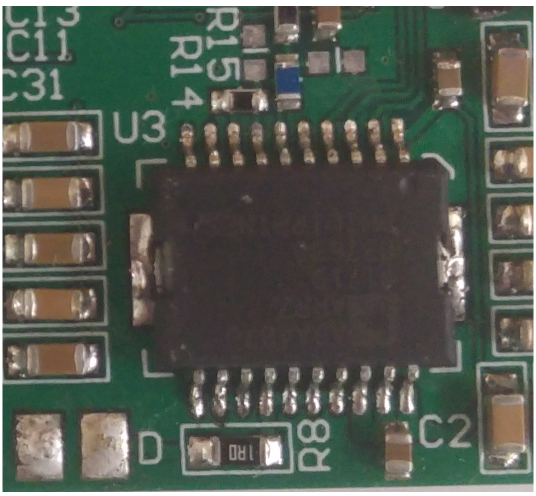



**Figure 3.7:** Tracker with with ADA4870 output stage

## 3.4 Layout

All designs were printed out on FR4 and laid out with high-frequency techniques incorporated. The output is made specifically for the RF amplifier, and the entire tracker mounts perfectly on the same heat-sink which is used to cool the ICs. A picture of the finished tracker and PA is shown in figure 4.21 on page 40.

## 3.5 Expected performance

The Tracker is configured as a tree stage amplifier with local feedback in the individual stages. Global feedback would have been too slow for stable operation, and would have required a considerable bandwidth reduction. The THS3001 data-sheet shows frequency response for different feedback resistors for 5 X gain in figure 3.8. In principle, a lower feedback resistor than 1 KΩ could provide higher gain especially since 9 X gain is used and not 5 but it proved to be unstable or conditionally unstable. Therefore 1 KΩ is chosen, as the AWG is bandwidth limited at 50 MHz and tracker functions are computed at 80 MS/s. The gain stage is then expected to have just above 100 MHz of bandwidth.

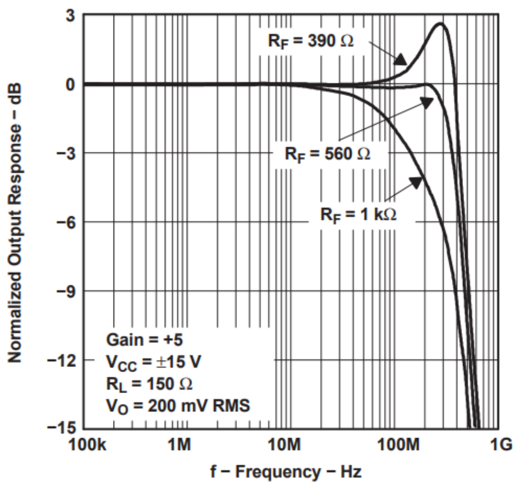



**Figure 3.8:** THS 3001 frequency response for 5X gain (datasheet(1))

Another factor that comes into play is that the op-amps do not have linear performance over input level. This behavior resides in the internal circuits dependence on drive voltage and slew-rate-booster circuits. When feedback is applied, it is not feasible to predict the outcome of this in the tracker but an idea of the time domain effect is achieved by studying large-signal effects from the datasheet in figure 3.9. These effects are suppressed when the entire bandwidth is not used, and so most of the open-loop gain suppress these large signal effects and linearize the response. In other words, this will probably not impact the tests as tracker functions applied are below 50 MHz.

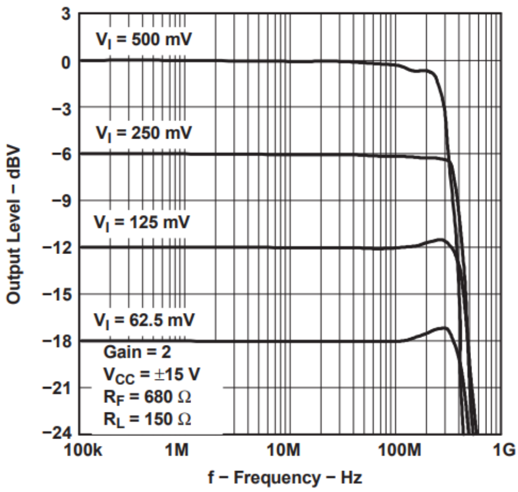



**Figure 3.9:** THS 3001 large signal response (datasheet(1))

For the output stage, the 6x THS3001 with unity gain has high bandwidth with 1 KΩ feedback resistance shown in figure 3.10. The output current is 120 mA (175 mA absolute max) with each, and so the total current capability is up to over 1050 mA and 720 mA recommended. This is more than enough for the amplifier that maximally draws ca. 600 mA. The output impedance shown in figure 3.11 is smooth over frequency and is never above 60 Ω up to 1 GHz where the drain capacitor will dominate the reflection seen by the RFPA. This is probably true for the tracker even though it is based on amplifiers with non-inverting topology as opposed to the test circuit in the datasheet.

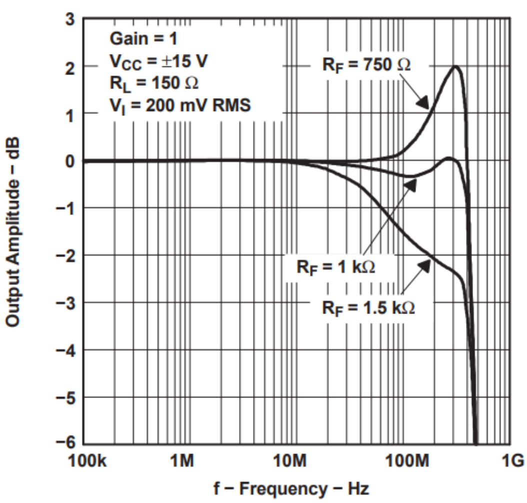



**Figure 3.10:** THS3001 frequency response for unity gain (datasheet(1))

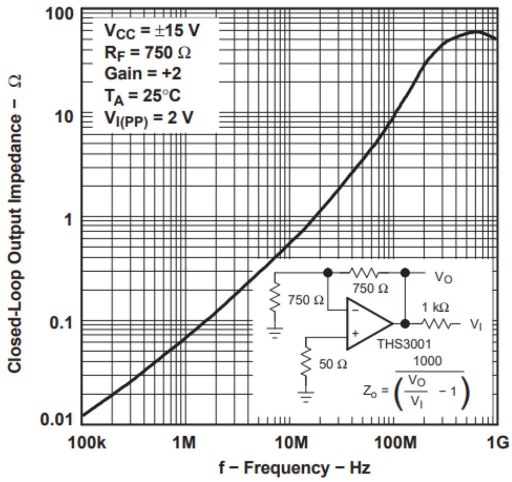



**Figure 3.11:** THS3001 output impedance (datasheet(1))

The tracker with ADA4870, on the other hand, will be bandwidth limited by the ADA4870 itself. The frequency response of unity gain is not shown in the datasheet but can be considered to be in the vicinity of the 2X gain. The 2X characteristics is shown in figure 3.12. Output impedance for unity gain is shown in figure 3.13. The output impedance phase is not detailed, and so the S11 phase is not estimated. The most important aspect is that it should not skyrocket and have a resonance after the open-loop gain go below 0dB.

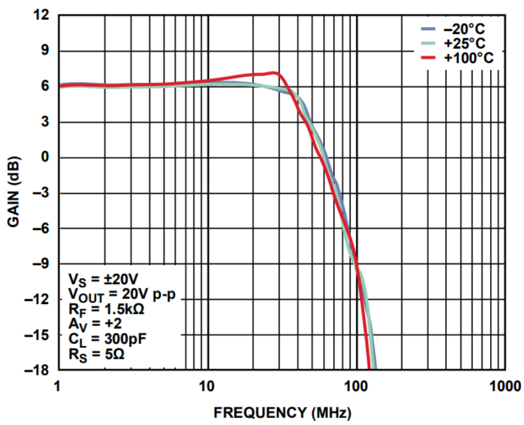



**Figure 3.12:** ADA4870 frequency response for 2X gain (datasheet(2))

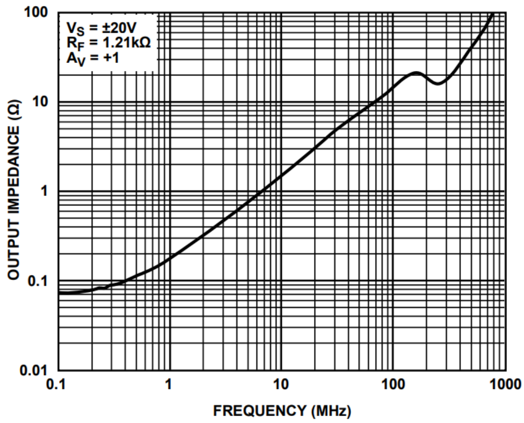



**Figure 3.13:** ADA4870 Output impedance closed loop (datasheet(2))

The expected tracker performance is summarized in table 3.2. The output impedance is defined at 500 MHz as the drain capacitor starts to dominate around this frequency (reactance is ca. 30 Ω at 500 MHz) and 10 MHz which is around base-band frequency. The output on the THS tracker is coupled through six 6.2 Ω resistors which are included in the output impedance calculation.

**Table 3.2:** Key expected tracker data

| Parameter | THS3001 | ADA4870 | Unit |
|---|---|---|---|
| Bandwidth | 100 | 50 | MHz |
| Gain | 86.5 | 86.5 | V/V |
| I out rec. | 0.72 | 1 | A |
| I out max. | 1 | 1.2 | A |
| Output impedance 500 MHz | 11 | 30 | Ω |
| Output impedance 10 MHz | 1.2 | 1.4 | Ω |
| Dynamic range | 3-29 | 3-29 | V |

Chapter 4

# Tracker functions and system measurements

To check the system performance, a set of measurements are performed. For these measurements, a set of base-band waveforms and corresponding tracker waveforms are needed. These waveforms are made in MATLAB and uploaded to the signal generators I/Q AWG. The signal generator used is an R&S SMU200A which have two sets of I/Q AWGs. One is used for the base-band signal which is up-converted to RF and drives the RFPA input. The other is uploaded with the corresponding tracker function and synchronized with a set time offset. Within this chapter further details on the tracker waveforms and tests of their performance are presented.

## 4.1 Tracker and baseband waveforms

As mentioned in section 2.5 the constant gain and max PAE tracker functions are 8th degree polynomial with the signal envelope as input. With PET and 2nd degree PET the power envelope and squared power envelope are multiplied by constants and summed, effectively a smaller polynomial. Polynomial factors are presented below in table4.1 and they result in transfer functions plotted in figure 4.1. The output polynomial is presented in equation 4.1 where y is output to tracker and x is the envelope.

$$y(x) = a_0 + a_1x^1 + a_2x^2 + ... + a_8x^8 \tag{4.1}$$

**Table 4.1:** Tracker transfer-function factors

| Function | $a_0$ | $a_1$ | $a_2$ | $a_3$ | $a_4$ | $a_5$ | $a_6$ | $a_7$ | $a_8$ |
|---|---|---|---|---|---|---|---|---|---|
| Const. Gain | 0.426 | -0.347 | 3.36 | -9.78 | 18.1 | -19.7 | 12.0 | -3.74 | 0.467 |
| Max PAE | 0.200 | -0.419 | 5.74 | -26.9 | 57.4 | -60.6 | 34.0 | -9.66 | 1.10 |
| PET 2nd deg. | 0.200 | - | 0.75 | - | -0.25 | - | - | - | - |
| PET | 0.400 | - | 0.307 | - | - | - | - | - | - |

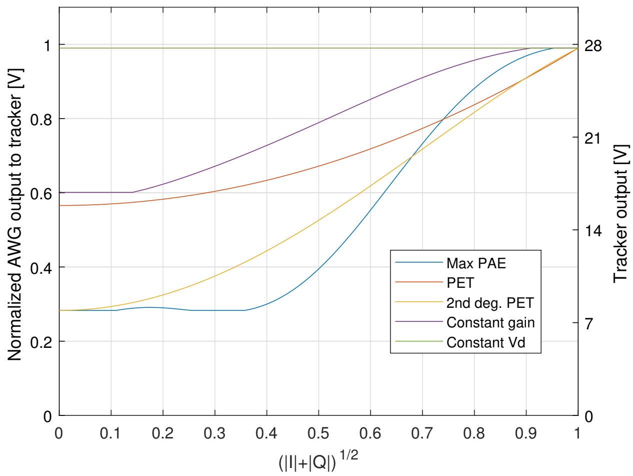



**Figure 4.1:** Tracker transfer functions

Note that the PETs have a slope with always increasing angle and have smoother transitions close to zero input. The LTE signal has a high crest factor and will spend most of the time in these low-level regions, and so characteristics in the lower power level regions are essential. Combined this results in less distortion with PETs as the RF gain won't get sudden changes. Regarding tracker "depth," the max PAE and 2nd order PET both track down to 8V output and will be more efficient while the constant gain and pure PET will have a higher gain and lower distortion. The smooth transitions in the PETs are what provides the lower tracker bandwidth required. In the subsequent tests, the tracker will however not be able to track above 40 MHz as the generator waveforms are put out at 80 MS/s. Having the not-PET waveforms bandwidth limited could result in more distortion but could also have an effect of forcing it to approach the PETs which in turn can have linearizing effects.

In figure 4.2 the max PAE waveform sent to the tracker is shown with generator voltage on the left and corresponding tracker output on the right (provided ideal amplification in tracker). For comparison, the base-band signal envelope is also shown. Note the sharp transitions and that output voltage is at 8 V most of the time. The abrupt transitions are what necessitate the high tracker bandwidths, and the low voltage is what provides the high efficiency as this tracking function makes the RFPA stay closest to or furthest into saturation. Using max PAE have a high impact on LTE as its crest factor is so high (ca. 12 dB in the 100 RB test waveform used in this example).

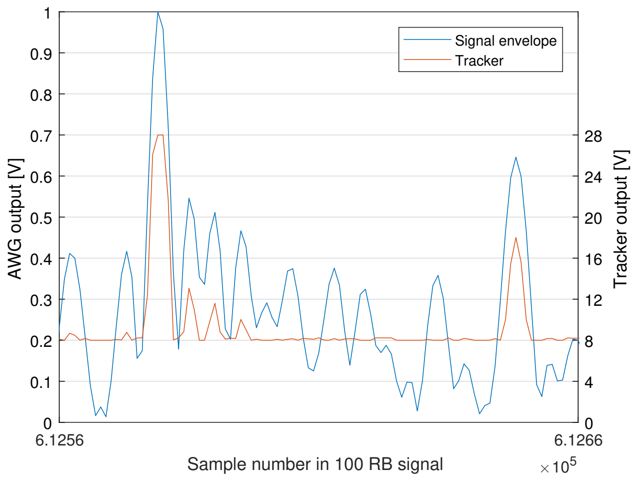



**Figure 4.2:** Max PAE tracker waveform example

Below in figure 4.3 is the 2nd degree PET. Compared to max PAE it does not spend as much time on 8 V, but on most of the spikes there is a ”natural detroughing”(12). Note however that there are also some sharp spikes like just after the peak voltage and that the slew-rate on these peaks are around the same. This is logical as the signal goes from a low to a high level one sample to the next and so there are no samples to distribute the smoothing features over. All facts considered the 2nd-order PETs expected to have lower distortion and somewhat lower efficiency than with max PAE.

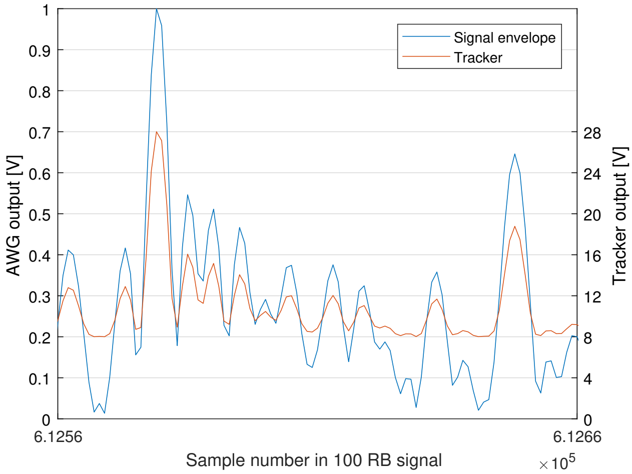



**Figure 4.3:** 2nd deg. PET tracker waveform example

In figure 4.4 is the constant gain waveform, tracking 12 dB gain. The most noticeable difference is the fact that it does not track as deep as max PAE or 2nd-order PET. This results in lower efficiency and lower distortion. The constant gain property results in ideally no AM-AM distortion, and hence only AM-PM distortion is left. To compensate for AM-PM alone is also much easier in digital predistortion.

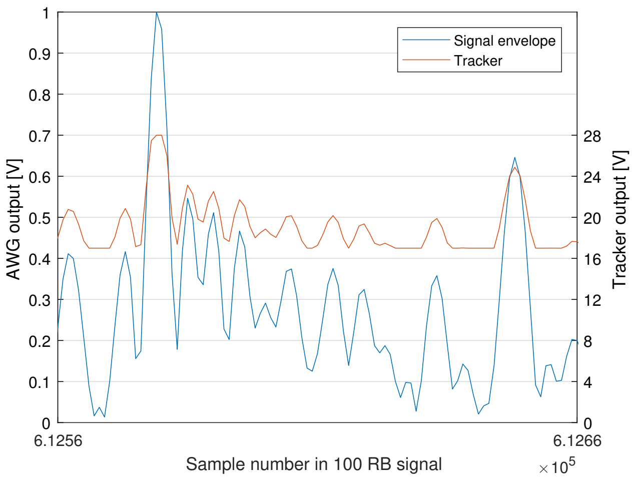



**Figure 4.4:** Constant Gain tracker waveform example

Finally, The PET waveform is presented in figure 4.5. It is pretty close to the constant gain waveform. The most significant difference in this plot is the slew-rate seen at the largest spike, and it is much lower with PET than any of the other waveforms. Additionally, PET is the most ”soft” waveform with the least sharp spikes and transients which results in reduced tracker bandwidth.

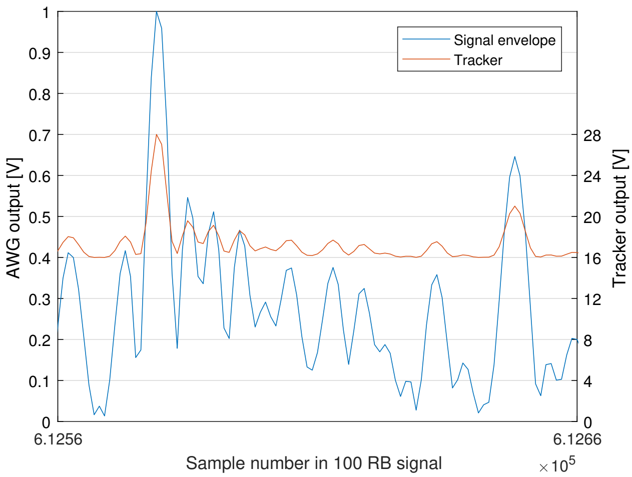



**Figure 4.5:** PET tracker waveform example

When calculating what is demanded by the tracker, a 2000 point Welch’s PSD estimate is used to analyze the spectral content of the tracker functions. The spectral density of the tracker waveforms and the corresponding base-band waveform are plotted below in figure 4.6. PET is the function that is lowest demanding followed by 2nd order PET, constant gain and max PAE. The base-band function is generated with the MATLAB LTE RMC waveform generation toolbox function and uses default filtering and sample-rate (30.72 MS/s)(6)(13). In the frequency domain, a nice ”brick wall” spectrum is produced down to -30 dB and then a small side-band before the residual spectrum is smeared out in the oversampling to 80 MS/s (in this plot all waveforms are further over-sampled to 1GS/s to see effect on the tracker spectrum). Had the base-band been ”brick wall” longer down there would probably have been a greater difference between the PETs and from the PETs to constant gain and max PAE. The fact that the 80 MS/s generator will limit the bandwidth of constant gain and max PAE further exemplifies the importance of having bandwidth limited tracker functions in real radio transmitters. This is because larger sample-rates will need to be processed in the transmitter system and faster DACs are needed to output the tracker waveforms. If the necessary bandwidth is not available, then accurate tracking will no be achievable. If the tracker waveforms are computed at higher samplerates they would follow the same characteristic further then 40 MHz seen in figure 4.6.

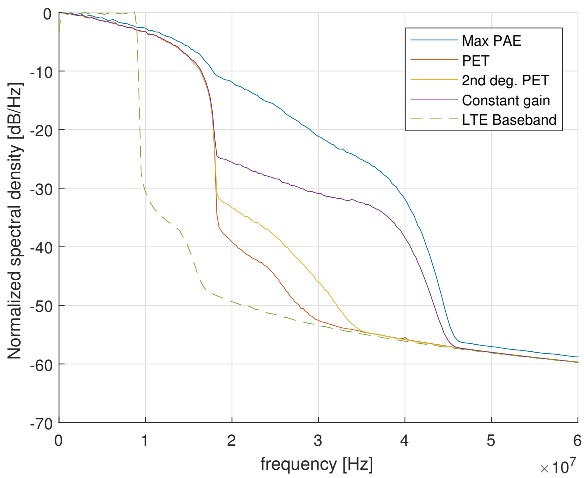



**Figure 4.6:** Comparison of tracker function spectral content

## 4.2 Simulated performance

When simulating expected performance a script from M.Olavsbråten was slightly modified and used to evaluate performance with LTE. The script uses the tracker functions and plots LTE waveform data points into interpolated RFPA data. The RFPA data used is from a December 2017 project and is simply logged power sweeps and measured input, output and DC power over different supply voltages. The mentioned project was carried out to ”break ground” for this thesis by designing a RFPA for use with envelope tracking.

Below in figure 4.7 gain vs. input power is shown, it is perhaps the most interesting plot from the simulation. PET and constant gain are driven with the same average input power (peak power - crest factor) and have the same peak point at 12 dB gain at ca. 28 dBm input power. As expected, the PET falls somewhat in gain to 1.5 dB as the voltage falls quicker and a bit lower than with constant gain. Both constant Vd, max PAE and 2nd order PET are driven to the same top point (max PAE at 28 V). Here the gain of max PAE goes up and down like a roller coaster. AM-AM distortion is significant as gain goes from 8 to 6 to 10 dBm in a ca. 10 dBm input span. It is clear that this will have a considerable effect on distortion performance. The 2nd order PET behaves much smoother and will not lose much efficiency as it pretty close to max PAE. The constant Vd also have a broad gain span, but it is a more seamless transition and only goes down which will make the amplifier behave as a peak compressor. A tempting thought now is to think that the best LTE performance regarding impairments would be the constant gain as AM-AM distortion is suppressed provided that the tracker can deliver voltage at the required bandwidth. There is, however, a significant discrepancy with the RFPA data: no phase measurements were available and so AM-PM distortion cannot be accurately simulated. S11 phase data is used instead and considered constant over power. This is NOT even remotely true for the amplifiers real performance.

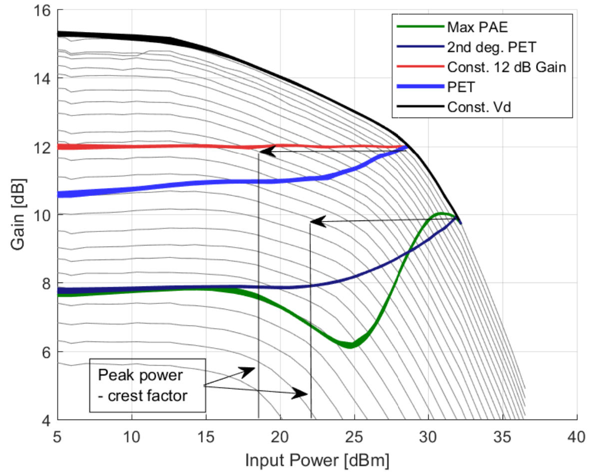



**Figure 4.7:** RFPA gain vs. input power

The gray lines in the plot are measured amplifier data swiping power over different voltages. Below in figure 4.8 is the same data but plotted for output power, an interesting note is how the top couple of dBm input power will be almost entirely clipped by the constant Vd waveform. It would probably be wise not to go that far into compression and maybe use the same input power as with the PET and constant gain.

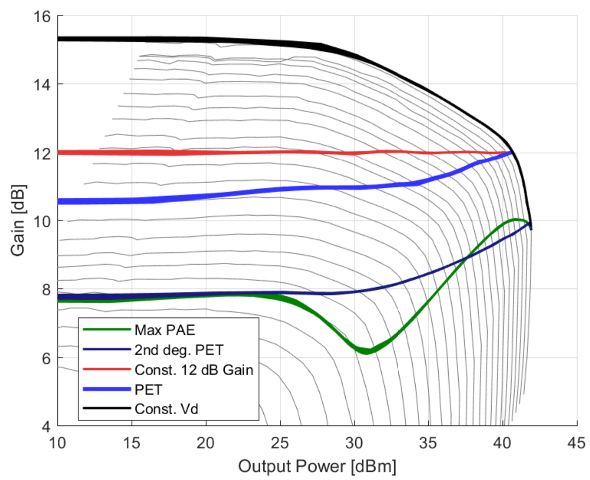



**Figure 4.8:** RFPA gain vs. output power

In figure 4.9 Below is the output power plotted for increasing input power. An ideal

PA would be a straight line which differentiated would be a flat gain. The constant gain is a straight line, and the PET is not deviating a lot. Note how the 2nd order PET is also pretty straight, especially compared to max PAE. As per definition, the constant Vd follow the constant 28V curve. Note also that the peak power output of constant Vd, max PAE and 2nd order PET is at the RFPA peak output power.

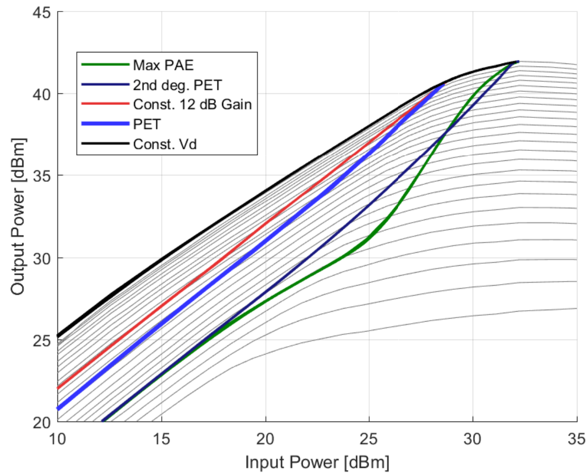



**Figure 4.9:** RFPA output vs. input power

In figure 4.10 we can see that the max PAE curve follow the peak PAE points, it is also interesting to note how close the 2nd order PET is. And so, for the small price paid in lost efficiency, one gets a lot back in linearity! Note also how PET is stable around 3 percentage points better than constant gain, not a lot but given the average PAE is 20% it would result in 15% better PAE. In figure 4.11 the drain efficiency is shown, and the same conclusions can be drawn.

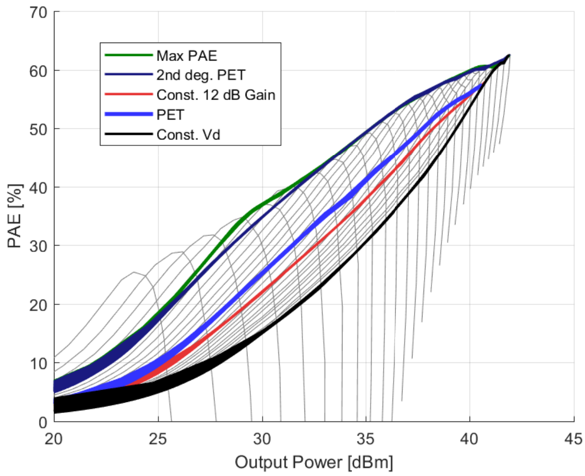



**Figure 4.10:** RFPA PAE vs. output power

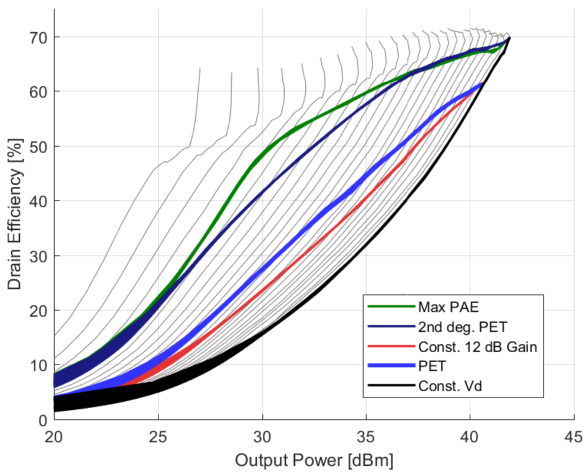



**Figure 4.11:** RFPA drain efficiency vs. output power

As mentioned, a significant discrepancy in the simulation data is the phase as it is not measured. S11 data is used instead and considered stable over power. This is not true, and as a result, distortion performance is expected to be worse than simulated. Below in figure 4.12 the phase versus the input power used in the simulations are presented. As an improvement, simulated phase with harmonic balance could have been used. This has not been done because of time restraints and the belief that overall the results would match pretty good anyways.

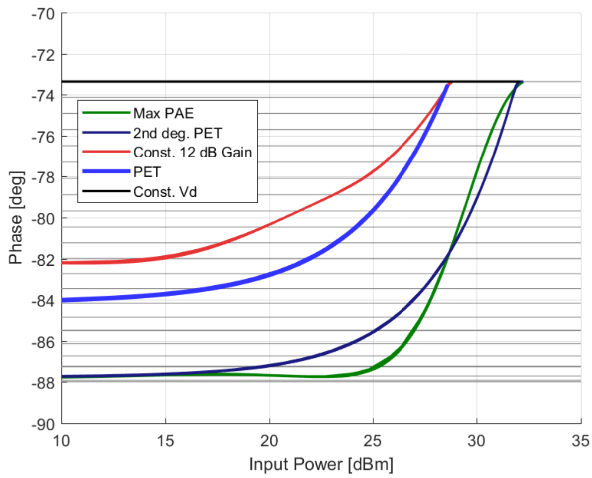



**Figure 4.12:** RFPA phase vs. input power

**Table 4.2:** Simulated performance for tracker functions

| FOM | Const.Vd | MAX PAE | PET2 | Const.gain | PET | Unit |
|---|---|---|---|---|---|---|
| ACPR | -30.3 | -28 | -36 | -41 | -40 | dB |
| EVM | 8.1 | 9.5 | 4.5 | 2.4 | 2.5 | % |
| PAE | 20 | 41 | 40 | 27 | 31 | % |
| NPR | -50 | -43 | -59 | -60 | -58 | dB |

## 4.3 Tracker measurements

The tracker based on 6x THS3001 and ADA4870 are studied closer and tested thoroughly. Frequency and phase response is examined and a simple time compensation added to the phase for simulation. Time compensation is added to emulate phase performance seen by an RFPA that has been adequately synced. S11 measurements have been conducted to analyze stability and to study what the amplifier supply terminal ”see” into the tracker. All tests are performed with the tracker seeing a 50 Ω load and outputting 20 V. The phase and amplitude of S11 could be in areas where the RFPA oscillate. The RFPA has shown to be prone to oscillations when only long wires are connected to a DC supply and stable when a 1 $\mu$F capacitor is added. Thus the amplifier is conditionally stable and depends on a supply that has close to capacitor like performance (look like short to RF). From here on, the tracker with LT1210, THS3001 and ADA4878 based output stage is abbreviated with LT, THS and ADA respectively.

Below in figure 4.13 the frequency response of the THS tracker is shown. When time compensated, it can work well up to ca. 100 MHz. 100 MHz is less than expected when taking into account its datasheet specifications, but the amplifiers are configured with a gain-feedback combination that is not specified, and the load is not connected between output and center of voltage supplies but tied to the negative supply at the output stage. Driving the sourcing network of the THS3001 chips harder than specified and driving the sinking network almost nothing results in an asymmetry. Time compensation is accomplished by mathematically removing the phase in an electrical length of 2 m (6.7 ns).

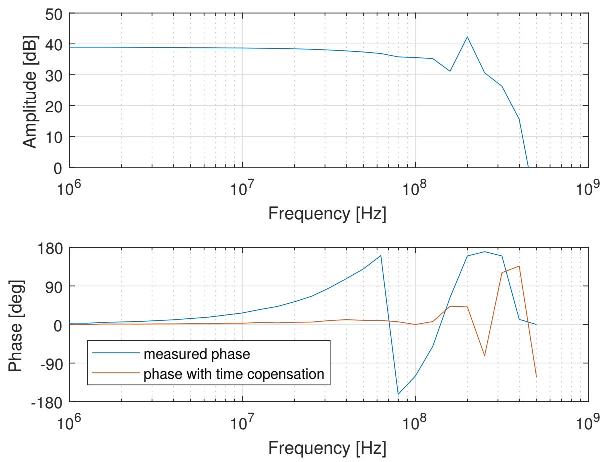



**Figure 4.13:** 6x THS3001 tracker, frequency response

The ADA tracker has a frequency response almost exactly as expected from the datasheet and work well up to ca. 40 - 50 MHz. Just like with the THS tracker, amplitude response is the limiting factor and not the phase when properly synchronized. Time compensation is accomplished by mathematically removing the phase in an electrical length of 3.5 m (11.7 ns).

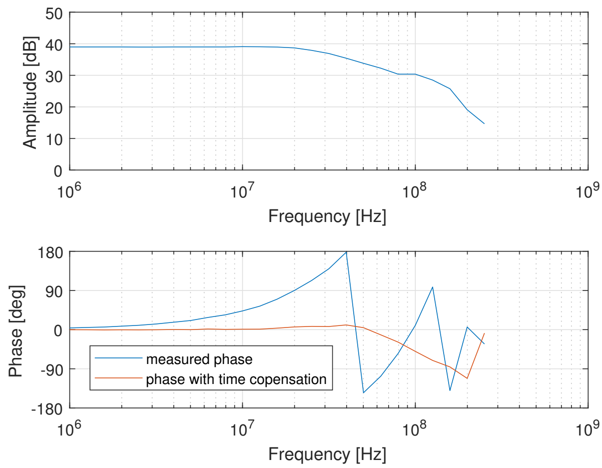



**Figure 4.14:** ADA4870 tracker, frequency response

The S11 parameters proved to be hard to get accurate. This is because the high voltage/power makes it necessary to protect the network analyzer with attenuators, and the reflected power will need to pass through this attenuation twice. Small changes in this total attenuation can thus result in large changes to reflected power. It was discovered when results changed real-time when a cheap off-brand cable was moved around. It turns out that a better cable, a custom-made dummy amplifier and consistent bend of the cable is what was needed to achieve good enough (ca 1 dB) accuracy. The cable is a Huber+Shuner SUCOFLEX 104 0.5 m. The dummy amplifier is a 50 Ω 10 dB attenuator that has the same physical dimensions as the RFPA. It can be cooled by the standard RFPA heat-sink and provide a reference plane for the RFPA at exactly the place of the RFPA power terminal. Consistent cable bend was accomplished by marking the spot of the heat-sink and keep it there during all measurements. The calibration was an Open Load Short type and accomplished by simply having nothing connected, shorting the terminals and by soldering two 100Ω resistors in parallel directly on the terminals. Below in figure 4.15 the setup is shown when calibrating 50Ω load. The upper frequency limit of this test setup was around 400 MHz, over this frequency the dummy amplifier proved not to be accurate enough, and noise levels became too high.

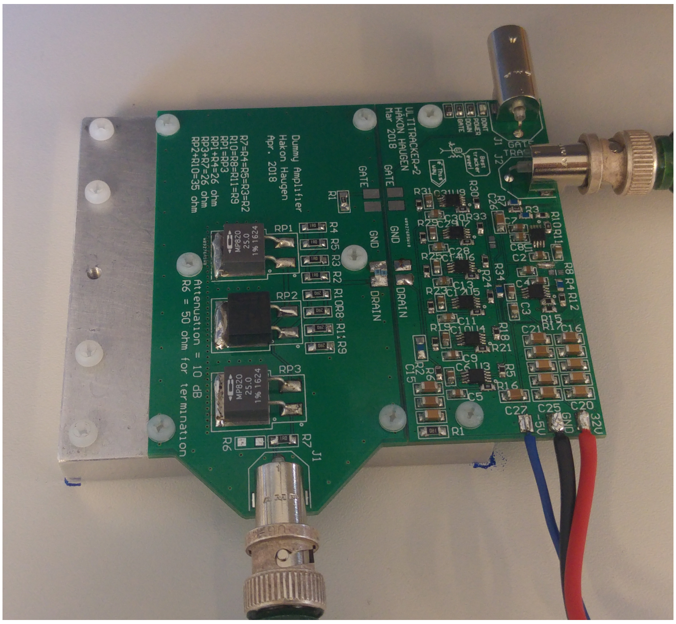



**Figure 4.15:** Tracker S11 test setup

Inf figures 4.16 and 4.17 THS tracker S11 performance is presented in a normal plot and Smith chart respectively. The results look good as S11 never is positive and is close to short circuit (0 dB ∠180 ). The spike at 80 MHz is believed to be an error in either the calibration or some switched function in the network analyzer as it is present in all the measurements. Tracker S11 performance is sustained even well over the frequency range it can track.

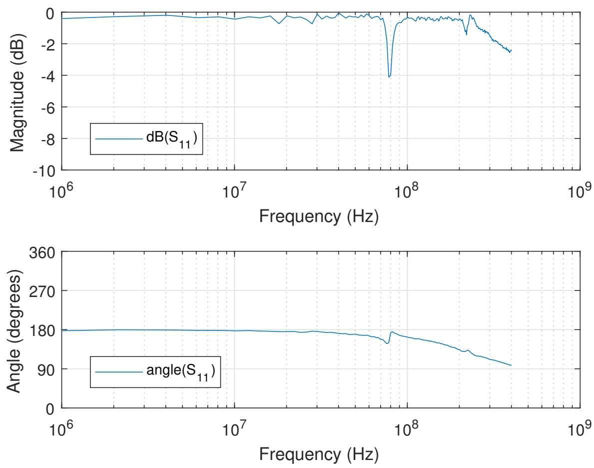



**Figure 4.16:** 6x THS3001 tracker, S11 parameters

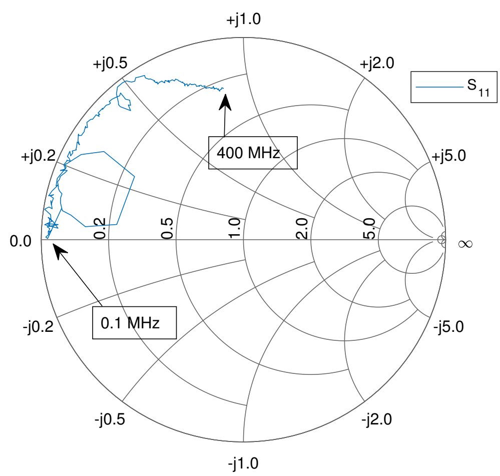



**Figure 4.17:** 6x THS3001 tracker, S11 smith chart

Inf figures 4.18 and 4.19 ADA tracker S11 performance is presented in a normal plot and Smith chart respectively. Just like with the THS tracker results look good as S11 never is positive and is close to short circuit. The spike at ca.80 MHz is also present and believed to be an error. The major difference with the ADA tracker is that it also shows dampening around 200 MHz and the response takes a loop in the smith-chart.

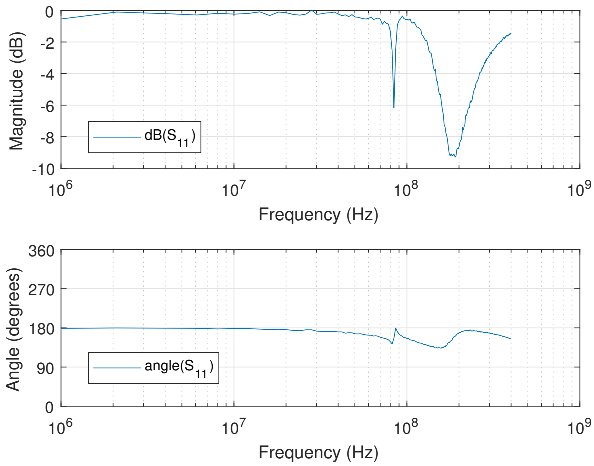



**Figure 4.18:** ADA4870 tracker, S11 parameters

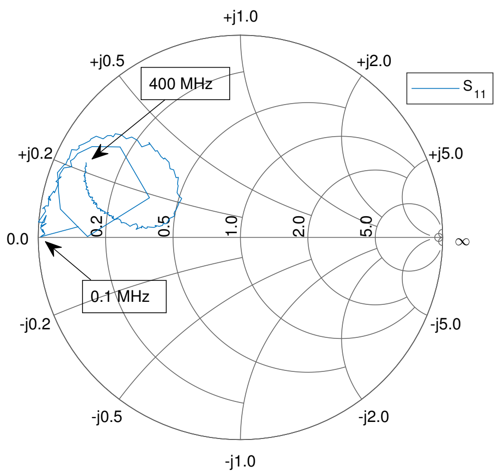



**Figure 4.19:** ADA4870 tracker, S11 smith chart

## 4.4 System performance with LTE

Testing the system performance is completed with the test setup illustrated in fig 4.20 and the tracker is pictured in figure 4.21 along with the RFPA. Average input power is adjusted to the target peak power minus the crest factor. This means one set of input powers for constant gain, PET and a through (through L) for reference and one set for constant Vd, max PAE, 2.nd order PET and a through (through H) for reference. Waveforms with a single 10 and 20 MHz LTE are tested for ACPR, NPR, EVM, MAE, and PAE. Two 10 MHz channels are tested at a different spacing with ACPR and NPR. A dual 20 MHz and quad 20 MHz signals are also tested with NPR and ACPR to push the limits and for pure interest. All tests conducted with all tracker waveforms and only with the THS tracker as there were no time to perform all these tests over again with the ADA tracker.

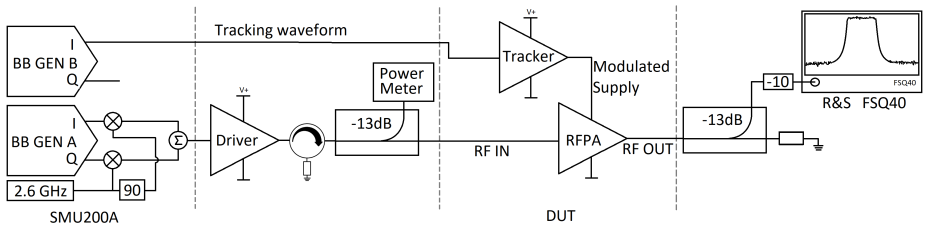



**Figure 4.20:** Test setup illustration

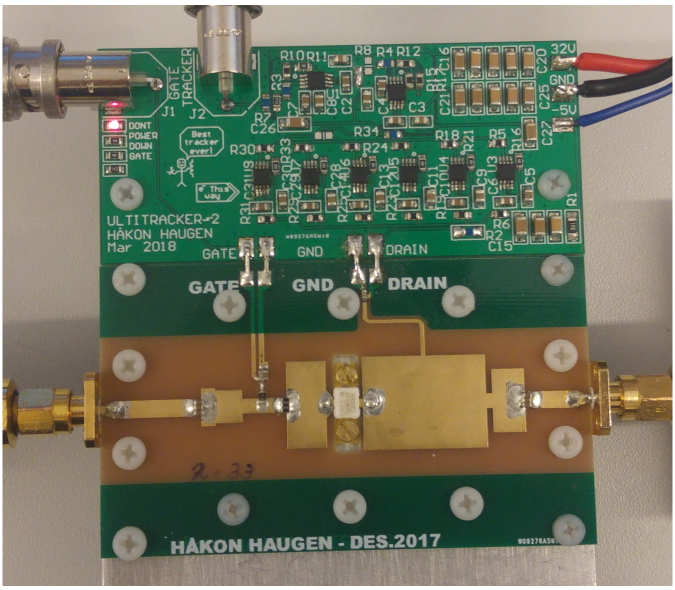



**Figure 4.21:** Tracker and RFPA in test setup

### 4.4.1 Synchronization

To time-align the tracker with the RFPA a special waveform is used illustrated below in figure 4.22. The waveform is fed to both the tracker and the up-converter path to the RFPA. The amplifier output is viewed on a high-frequency scope that can handle the RF frequency. If not aligned the waveform will be a flat bar with two spikes on it. The amplitude of the ”bar” corresponds to a specific RF input power and tracker voltage. One of the output spikes comes from an increase in gain because of the spike in supply voltage. The other output spike comes from the spike in input RF power. In the generator, an offset is inserted to align these two spikes. As a sanity-check, there will then be a spike also preceding the bar that comes from having a single sample with both RF power and tracker output. The same offset is kept throughout measurements.

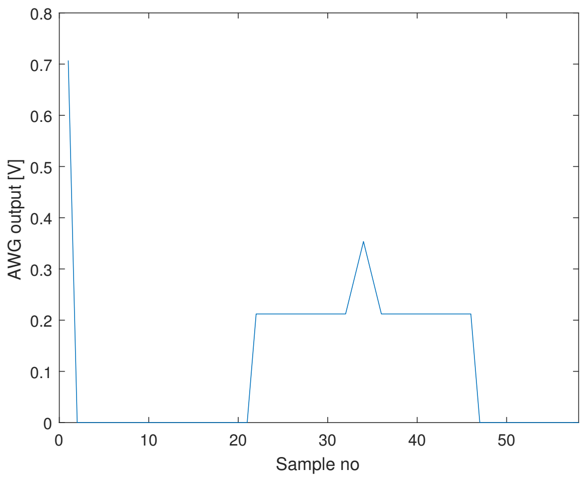



**Figure 4.22:** Synchronization waveform

### 4.4.2 ACPR

To measure ACPR, the signal analyzer is set to spectrum analyzer mode with 100 kHz resolution bandwidth. Power within the channel bandwidth is integrated and compared to the integrated power in neighboring channels. When measuring ACPR one thus get one lower and one upper ACPR value when comparing to the lower and upper-frequency bands respectively. To make the data more presentable and comfortable to compare, the average value in DB of the upper and lower is presented as a single ACPR value.

The ACPR measurements shown in figure 4.23 are somewhat as expected and comes with a few surprises. PET performs best among the different functions followed by constant gain even though single channel simulations predicted that constant gain would have a marginally better ACPR and their levels are ca. 5 dB worse than simulated. The constant Vd and max PAE performs pretty much as simulated. The 2nd order PET performs ca. 10 dB worse than simulated and is a significant deviation. The huge surprise in this measurement is how independent the ACPR level is of the channel spacing. One would expect that when the tracker waveform was limited in bandwidth that ACPR would increase along with channel spacing as the power envelope bandwidth increase. It is however believed that when the RFPA creates mixing products at base-band, the tracker has such a low output impedance over large enough bandwidth that these mixing products are effectively shunted. This, in turn, suppresses intermodulation that would have arisen if these products had been mixed up again to RF. In other words, these ACPR levels are more dependant on tracker output impedance than channel spacing. At least for the base-band frequencies that are available with 80 MS/s AWGs.

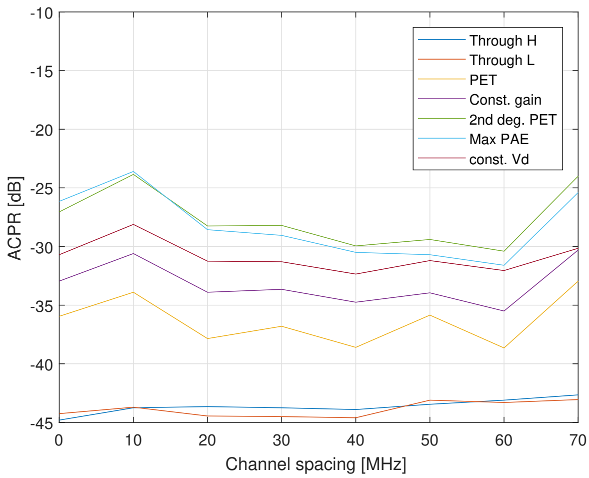



**Figure 4.23:** ACPR vs channel spacing on 10 MHz LTE

When measuring single and dual 20 MHz LTE, similar results were achieved, and they are presented below in figure 4.24. Again PET is the best performer, and 2. order PET is the worst. Mostly the dual channel spectrum has worse ACPR than a single channel; this is as expected. Somehow the max PAE performs better with the dual channel in this respect, maybe because the tracker function is limited in bandwidth and that way it doesn't track max PAE anymore. The ACPR on the Quad 20 MHz was not measurable, as the input signal had a way to high ACPR which would be amplified with an unknown gain by the RFPA.

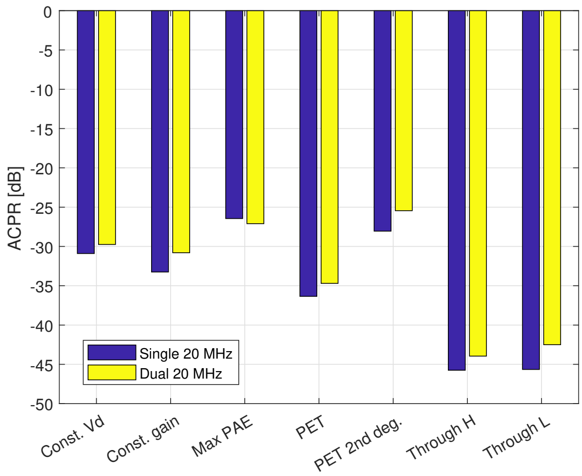



**Figure 4.24:** ACPR on 20 MHz LTE

### 4.4.3 NPR

When measuring NPR, the signal analyzer is set to spectrum analyzer mode just like with ACPR. The difference is that the signal fed to the analyzer has a null-notch ca 400 kHz wide. The power spectral density within this notch is then compared to the power spectral density 1 MHz outside the notch.

The NPR presented below in figure 4.25 reaffirm the ACPR results. PET is again showing best performance, and the level is as good as independent of the channel spacing. Again, this is not what was expected. The level also deviates from the simulations in all of the NPR measurements. For example, the 2nd order PET which is simulated to almost -60 dB and one dB better than PET whereas the real results are -20 dB and 10 dB worse than PET.

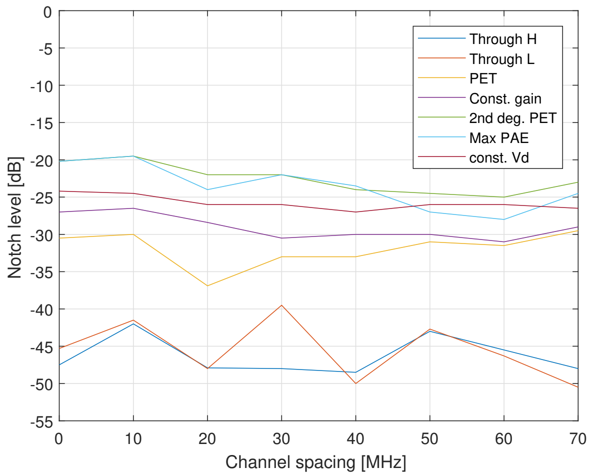



**Figure 4.25:** NPR vs channel spacing on 10 MHz LTE

In the NPR levels of single, dual and quad 20 MHz channel measurements in figure 4.26, we find a correlation to the corresponding ACPR in figure 4.24. The results on Quad 20 MHz (80 MHz of spectrum(!)) are quite surprising though as it on most tracker functions outperform one or both of the signals with lower bandwidths.

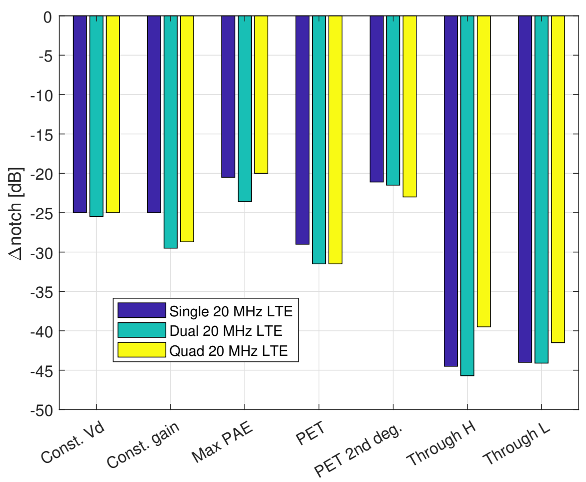



**Figure 4.26:** NPR on 20 MHz LTE

### 4.4.4 EVM

When EVM and MAE are measured the signal analyzer is used as a zero-IF receiver and collect 200 mega-samples of I/Q base-band data. The same LO does not feed the analyzers down-converter and the generators up-converter but the 10 MHz reference frequency is synchronized and that way it is assumed the frequencies are stable over the collected samples. The IF filter is now effectively the base-band filter and it is set to 50 MHz while data is sampled at 80 MS/s. The filter band-with is not that important in respect to blocking unwanted RF signals that would have created image frequencies as they are not present. However, it is found that internal tones are mixed in, and noise floor rises at higher sampling rates. The sampling rate and filter bandwidth are therefore chosen to get as best possible performance in the base-band frequency of the LTE signal. It can be favorable to over-sample the received data even more as the tracker functions are calculated at this rate as well, but for 20 or 10 MHz LTE in-band measurements, it is considered sufficient. In baseband, this means the instrument is collecting 10 or 5 MHz signals at the I/Q channels with a 40 MHz bandwidth. The sampled data is further processed digitally in MATLAB where the waveforms are filtered, aligned(frequency, phase, and RMS amplitude compensated) and compared to the waveforms uploaded to the generator.

The EVM measurements are also surprising. Again PET is proving to be the best performer and is even around the level of the through. This is where the problem arises with these measurements (also counts for MAE). It turns out the generated RF waveform at the RFPA input is already distorted. It is probably distorted because the generator up-converter, driver, or output amplifiers are driven too hard. If that is the problem, tests should be conducted with a driver amp with higher gain and with the instrument requirements relieved. Alternatively, the waveform generator and analyzer could be synchronized to the same portion of the signal in time domain. Then the through measurement could be used as a reference instead of the true base-band waveform. A single test with this technique showed good results. The PET then had a 4.8 % $\text{EVM}_{RMS}$ around the simulated value. This technique would probably be the correct way of doing these tests as both DUT input, and output paths are calibrated out. Either way, the EVM results gathered are still a good way of comparing the tracker waveforms against each other and time restraints made new measurements infeasible. Note also that on constant gain the distortion error levels are lower with 20 MHz than with 10 MHz, this could be due to the changed crest factor and correspondingly its further into back-off while somehow only affecting the constant Vd enough to go past the 10 MHz. It could also be a statistical lucky day, and most other sample areas were worse.

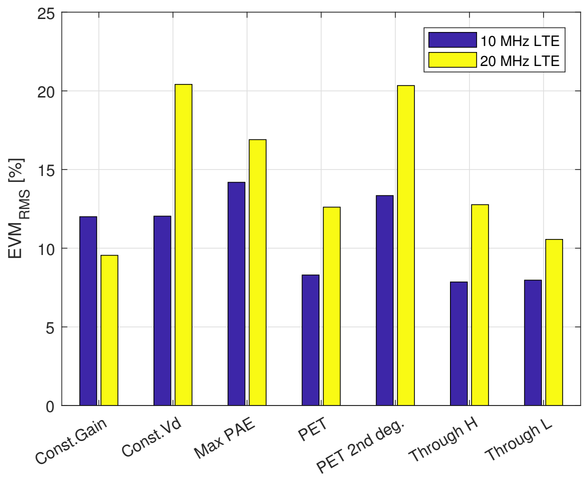



**Figure 4.27:** EVM performance

### 4.4.5 MAE

The MAE is just another way of weighing the error vectors that also produce EVM. Therefore these measurements also suffer from having an already distorted input signal. The most noticeable effect of the new weighing is that results are more even and that max PAE no longer performs the worst. This results probably because max PAE has the worst performance at the peaks and is therefore benefited by usage of a FOM without squaring of the error.

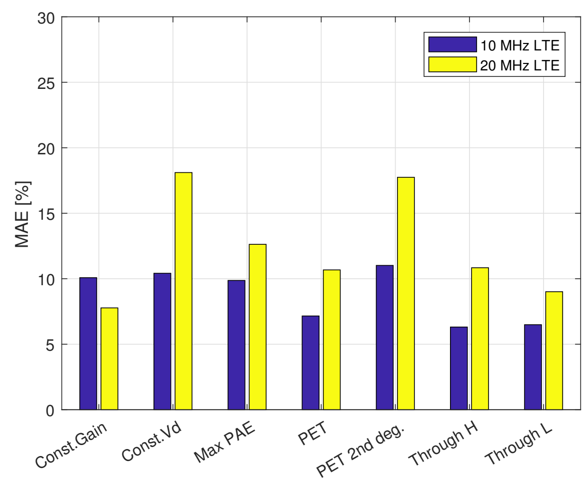



**Figure 4.28:** MAE performance

### 4.4.6 PAE

The PAE measurements are made by comparing the DC and input RF power to the output RF power. Output RF power is measured with attenuators and a power meter directly, and input RF power is measured with a directional coupler and power meter. The DC power is calculated by using a scope and measure voltage at 200 kS in 20 $\mu$s of the same peace of the waveform. The voltage is measured on each side of one of the output resistors on the THS3001 tracker with the same probe and scope channel. The probe is 10X attenuated, rated for 500 MHz and ground-coupled through a 1.5 cm ground spring. The RMS voltages are computed and translated to current and power using ohms law. This technique provides two inaccuracies. The current phase compared to voltage is not measured, and so the reactive part versus real part of the tracker load is unknown. Also, the entire repeating RF waveform is 10 ms long and so the 20 $\mu$s evaluated will only be an estimate. Because the same waveform peace is evaluated each time, static errors are canceled when comparing the results against each other. These factors should be kept in mind but not impact measurements to any noticeable degree, especially when comparing functions up against each other.

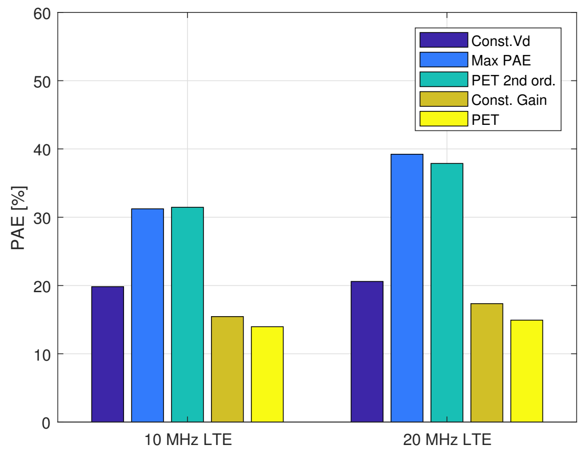



**Figure 4.29:** PAE performance

The PAE results are shown in figure 4.29. As expected, PAE is high with max PAE and 2nd order PET tracking and low with PET and constant gain tracking. Conformity with expectations ends here. Constant Vd is expected to be worst, and this is not the case. As mentioned the absolute level is not that accurate because of the small time sample evaluated, but results are consistently ca. 10 percentage points lower than expected. 10 percentage points is more than what could be expected of small differences in 20 us sample-spaces. The most surprising result is the fact that constant gain outperforms PET in terms of PAE. The difference is only by 2.5 percentage points but still surprising. The result is relatively consistent in both 10 and 20 MHz measurements. PET should be slightly more efficient as it always is tracked with a lower voltage and therefore less in back-off, this is true even though the tracker function only has a 40 MHz bandwidth that may limit true constant gain tracking. Compared to the expected results, just the constant Vd is performing on par (section 4.2). The problem in these measurements was realized to be an oversized DC-block capacitor at the RF output path, effectively putting a 50 Ω AC coupled load in parallel with the RFPA. All facts considered the PAE figures are not regarded as accurate and descriptive of tracker function performance.

## 4.5 Simulations comparison with measurements

In table 4.3 key simulated and measured performance data is presented. In many discrete cases, the simulation data does not co-inside with the measured data. But when taking into account the inaccuracies in simulation data and the measurements themselves, the figures make sense.

ACPR of the real system is consistently a couple of dB worse in the actual measurements. This can be a consequence of the intermodulation that is not taken into account in simulation to the correct degree. The 2nd order PET also performs allot worse than expected, and this could also be contributed to the phase, as it tracks just as ”deep” as the max PAE and the coefficients thus may not be optimal.

As mentioned, the EVM should have been measured with a through as reference, as this would cancel out errors in the test setup. Still, the results are useful for comparing tracker functions against each other, and there were no significant surprises. The PET performs best in class and in terms of linearity seems to outperform all other functions. Quite possibly because the constant gain tracker waveform(the closest rival) is bandwidth limited and that the AM-PM distortion might be higher.

PAE measurements are ruined because of the gigantic DC-blocking capacitor. It is worth noting that the simulated PAE of the constant Vd is almost a perfect match. In other words, it looks like the simulated PAE numbers are likely achievable with a smaller DC blocking capacitor.

NPR measure in-band noise spectral density and the system perform an entire 20 dB worse than expected values. Considering that the through performed so much better points to that these figures are real. It simply looks like the simulations show unrealistic performance, perhaps the way NPR is simulated is to blame.

**Table 4.3:** Simulated and measured performance for tracker functions

| FOM | Const.Vd | MAX PAE | PET2 | Const.gain | PET | Unit |
|---|---|---|---|---|---|---|
| ACPR(S) | -30.3 | -28 | -36 | -41 | -40 | dB |
| ACPR(M) | -31 | -26 | -27 | -33 | -36 | dB |
| EVM(S) | 8.1 | 9.5 | 4.5 | 2.4 | 2.5 | % |
| EVM(M) | 11 | 14 | 13 | 12 | 8 | % |
| PAE(S) | 20 | 41 | 40 | 27 | 31 | % |
| PAE(M) | 20 | 31 | 31 | 15 | 14 | % |
| NPR(S) | -50 | -43 | -59 | -60 | -58 | dB |
| NPR(M) | -31 | -26 | -28 | -33 | -36 | dB |

Chapter 5

# Discussion

## 5.1 Tracker polynomials

The created tracker functions are computed with polynomials, which is favorable if the system is to operate real-time as the function can be calculated in hardware with relatively simple multiply-and-accumulate circuits. In addition, it is relatively easy to fit a polynomial to selected points gathered from measured RFPA data. In other words, this technique of tracker function creation is implementable in real systems with today's technology.

LTE signal generation is specified and operate at a set FFT/IFFT size and sample-rate dependant on how many RBs wide the TX spectrum is. In the case of a 100 RB signal, the FFT size is 2048 and the I/Q sample-rate is 30.27 MS/s and when populated with 1201 sub-carriers is just below 20 MHz wide. Were this baseband data to be used directly as input to a tracker function one would be Nyquist limited to 15 MHz tracker bandwidth. That is even below the RF bandwidth! This means that the tracker function must be processed at a higher rate than what is possible with the standard waveform generator.

### 5.1.1 Constant supply voltage

The standard RFPA is very much outdated when used with modern communication systems. The main reason is a combination of high crest factor and a strict linearity requirement. If a base station is implemented with a constant Vd GaN amplifier, it would need to be driven a long way from compression. It is included in the measurements as a reference point, and at large it has proved to be among the worst performers all over.

### 5.1.2 Max PAE

Tracking the supply voltage for highest possible PAE is a challenge as the required tracking bandwidth is spread wide out and a comparatively small amount of power is left in the drain voltage DC component. The measurements have systematically shown max PAE to have the poorest results regarding linearity. This has corresponded well with simulations that hint towards huge AM-AM problems and large AM-PM distortion. It is clear that to use this as a tracking function is not wise at all with modern wide-band modulations. An interesting concept is that this could be used with great results on constant envelope modulations that slowly vary transmit-power. Say for example a GSM phone that is moving around and constantly need to adapt the transmit-power while the modulation itself is constant envelope GMSK. But again, with LTE, tracking for max PAE is plain out dumb.

### 5.1.3 2nd order PET

The 2nd order PET retain most of the favorable aspects associated with max PAE but cuts away the excessive tracker bandwidth need to a mere two times RF bandwidth. Simulations back this up as well and even estimated RFPA spectral properties are on par with PET. However, the phase results used in the simulation were way too optimistic, and thus the deeper tracking will probably result in more AM-PM distortion. If the 2nd order PET is to be used in a real LTE base station, it would need better linearity than what was both measured and simulated. Still, the level of efficiency reached makes it a good candidate for further linearization with digital predistortion. It is also worth mentioning that there is no single correct way of configuring factors, and a slight change in tracker function may have improved the measured performance.

### 5.1.4 Constant gain

The constant gain has a high level of linearity as AM-AM distortion is suppressed. In the simulation, it showed good results with an EVM almost down 4% (although phase is not accurately simulated). The fact that it has constant gain also have the added benefit of making digital predistortion easier because it would only need to compensate for phase. The huge drawback with constant gain however, is the large bandwidth needed. It is not as strict as max PAE and does not track as deep (lowering slew-rate), but this could be a good starting point for a relatively powerful amplifier that could be made pretty linear with some digital predistortion.

### 5.1.5 PET

The PET compared to constant gain is in many ways (at least with the coefficients used in this thesis) what 2nd order PET is to max PAE. Most of the wanted characteristics in constant gain remain while greatly reducing the tracker bandwidth. The PET come out good in both measurements and simulations. It has a strictly defined bandwidth and still linearizes the RFPA compared to constant Vd. In the simulation, its level of performance is around the same as constant gain, but with real phase distortion added, it could well be that the smother transitions provide better AM-PM performance. Anyways it is considered a superior tracking function as most of the characteristics are the same as with constant gain, while the tracker bandwidth is the lowest. The measured performance is also consistently the best in terms of linearity and is ”out of the box” the easiest solution to achieve performance that can be used in a real base station (ACPR = -44.2 dB) (14).

## 5.2 Multiple channel generation

It is not quite clear how the measurements turned out to be so stable with multiple basebands. One would expect impairments to increase when the channel spacing increased, but the results were largely the same. A theory is the trackers low output impedance shunts tones mixed to baseband. Another theory to why this happens is that the way multiple channels are created is to blame. As mentioned in section 2.7 the two channels are created by multiplying the base-band function with a virtual IF and then up-convert without image suppression. This creates two identical spectra, but in time the peaks would be located close to each other. So if complex up-conversion of two different base-band functions to a positive and one negative IF then maybe the result would be more distortion along with higher separation.

## 5.3 Tracker

The tracker proved not to be a trivial design exercise. High-performance current-feedback operational amplifiers like the THS3001 needed extra care, and high-frequency techniques were needed to achieve as high bandwidth and stability as wanted. All the tracker designs were also able to drive the potentially unstable RF amplifier with such an input impedance that it was stable. It was a fear that phase performance would limit the bandwidth and introduce distortion that way, but with proper time compensation, this proved not to be a problem as both trackers have excellent phase, and bandwidths that is limited by amplitude response. A test that maybe should have been conducted is frequency response over different output voltages. The performance could be different close to the power rails and can result in a potentially asymmetric rise and fall response.

### 5.3.1 THS3001

The TS3001 is a high-speed operational amplifier. The tracker based on 6x THS3001 was easily able to have better bandwidth than the I/Q base-band generator (>50 MHz). The tracker showed excellent wide-band performance and was used on all of the LTE performance tests. The tracker is not considered a limiting factor in the tests performed within this thesis.

### 5.3.2 ADA4870

ADA 4870 is a nice circuit with a power-pad package and perfectly placed pin-out. It provided performance in the range of what could be expected and also had good wideband performance. Even though it has lower bandwidth than THS3001, it can provide more power while its 50 MHz bandwidth is more than enough for most applications. The fact that it has a bandwidth on par with the signal generator points towards that tests would probably have had the same results with this tracker.

## 5.4 Simulation

The simulations have been accurate in the manner of pointing to good and bad LTE performance for different tracking waveforms. The large X-factor in the simulations is the phase response of the RFPA. As mentioned, harmonic balance simulation data should have been used instead of S11 data. Another is that tests are conducted with frequency swipes with constant voltage, and so the capacitance at drain is not taken into account. Combined this makes the simulated PAE close to what could be expected with the lower capacitance at the output DC-block. Additionally, the NPR results are considered to be a too optimistic estimate. All facts considered, the simulations at large are believed to be quiet accurate and could be used to estimate tracker function performance.

## 5.5 Tracker synchronization

The method of synchronization in time-domain by exploiting the voltage dependant gain in an amplifier and aligning a spike with the actual RF waveform proved to be highly efficient. It is easy to identify down to single sample precision level. The method depends on having a amplifier with varying gain dependant on supply voltage, which is perfect for GaN amplifiers, unlike for example GaAs based amplifiers. Still, the technique can be modified to have a larger supply voltage swing or even go all the way to power-gating.

## 5.6 ACPR & NPR

In the out-of-band ACPR measurements, levels are not sufficient for LTE. If implemented with digital predistortion however, reaching acceptable ACPR levels for LTE is within reach. The considerable surprise here is how bad the 2nd order PET perform. It was neither expected or simulated that it would perform on the same level as max PAE. A theory is that the waveforms are not over-sampled enough for the 2nd order PET to use its detroughing capability which in turn result in max PAE and 2nd order PET gets pushed towards each other in time and frequency domain.

## 5.7 EVM & MAE

Sadly the EVM measurements are heavily compromised by the test setup. Probably the mixers, driver amplifier or any internal amplifier is in or to close to compression. As mentioned, the through measured an EVM of almost 5%. With external syncing so the same peace of the waveform is compared each time and with the through signal used as a reference, much more accurate results could be achieved. That is because the system without the DUT could be calibrated out to a large degree. A single test with a 100 RB waveform and PET tracking show that such a setup could have around 5% EVM, which closely resemble the simulated levels.

## 5.8 PAE

The PAE results are deviating from the expected levels to a large extent. Especially emphasized by the fact that constant supply voltage showed better efficiency than PET and constant gain. Multiple factors make this measurement non-reliable, but there is one catch in particular.

What seems to be the cause is the DC-block capacitor which is too large. The drain capacitor on the RFPA is just 10 pF, not enough to make any significant difference (reactance ca.1600 Ω @ 10 MHz) this is by design as the RFPA was designed for envelope tracking. The transistor drain capacitance alike should be negligible. The big catch is the DC-block capacitor on the output. It is 300 nF (53 mΩ @ 10 MHz). This means that practically all AC voltage the tracker delivers is into the RFPA is also in parallel with 50 Ω and is a huge design flaw in the RFPA. The tracker has proven its capable of driving this load, and so the other measurements are not affected to any large degree, but all efficiency measurements except with constant voltage are effectively ruined. In a real system where the load is not a perfect 50 Ω from DC to daylight, it would also be preferable with a smaller capacitor that blocks the tracker function frequencies. Primarily because the antenna probably would be either a short or open circuit on these low frequencies and therefore have a high reflection that comes back to the tracker/RFPA at an unknown angle dependant on cable length.

The reason for why constant gain is performing better than PET is probably also a result of the ridiculously large DC-block. When the output power is measured with a power meter, the tracker function itself also deliver power to the load. This power, in turn, is dependant on amplitude and frequency in the tracker output. The constant gain has much larger high-frequency content and so will have a significant direct contribution to the delivered power. The fact that the tracker itself provides power to the load at tracker frequencies complicates the PAE measurements even further and underline the importance of a correct DC-block capacitor in the DC-block. In this case, ca. 15 pF could have been a good candidate with a reactance of 530 Ω at 20 MHz and 4 Ω at 2.6 GHz. An interesting note is that DC blocking at lower RF frequencies could be a challenge. There are for example LTE on 800 MHz, and sharper filters might be needed in those cases.

## 5.9 Application with other than GaN devices

It is not a given that these results are directly transferable to other technologies. The transconductance (gain) of GaN HEMTs have a significant dependence on the supply voltage. This is a characteristic that is not similar to different process technologies like for example GaAs. The study of effects on other devices is considered future work.

## 5.10 Implementation in mobile stations

In the up-link waveforms from mobile devices, SC-FDMA is incorporated instead of OFDMA. This is a multiplexing scheme that lowers the crest-factor to increase efficiency. The results from the tests presented here are therefore not applicable to up-link LTE transmissions, and the study of these effects are also considered future work.

# Chapter 6

# Conclusion

The purpose of this thesis is to investigate properties of different tracker functions with modern broadband LTE. In particular, power envelope functions compared to other functions that require more bandwidth. High bandwidth tracker design is also studied as its performance is paramount in a tracked power amplifier system.

## 6.1 Tracker design

The tracker design is discovered to be non-trivial when going up to tens of megahertz. High-frequency techniques like impedance matching, local supply decoupling, via-stitching and careful design of ground plane need to be addressed. Two designs are studied and both based on current feedback operational amplifiers. Operational amplifiers with current feedback can be challenging to keep stable, but their performance in high-frequency applications like this is unmatched. An especially interesting result is how well the tracker's phase response behave when properly time-aligned. Although not tested, it is believed that such tracker performance levels are not achievable with a voltage-feedback design. An interesting discovery in the system linearity test is how the increased bandwidth from two channels with increasing spacing had almost no effect. This could be contributed to good tracker performance because low output impedance suppress down-mixed intermodulation products.

## 6.2 Tracker functions

The tracker functions which has been studied are; max PAE, constant gain, the power envelope and the power envelope with an additional degree of freedom. It is shown that max PAE impact linearity the most, and is probably not useful with wide-band modulations unless extremely effective predistortion or error correction is implemented. Its efficiency is estimated and measured to be top of the class. It is measured to ca.30% PAE, but considering the DC-block design flaw, the simulated ca. 40% is a more likely number. The 2nd order PET shows many of the same characteristics being almost as efficient as max PAE and posses linearity in the same region. It is however believed that with another use of the 2nd degree of freedom could provide more linearity and around the same efficiency level.

With constant gain and pure PET, a lot of the same linearity is achieved. Considering the smaller bandwidth requirement of PET, it is the superior tracker function when implemented for use with LTE. Regarding efficiency, the measurements are ruined by the RFPA DC-block design flaw but simulations predict a PAE of around 30 %. For reference, the constant supply is measured and simulated to have a PAE of 20% and linearity in the middle of the tracker functions.

## 6.3 Discrepancies in measurements

There are some key discrepancies in these measurements. The largest is, as mentioned, the RFPA design flaw, where the DC blocking capacitor is too large and ruin efficiency measurements. When measuring in-band linearity in terms of EVM and MAE, the reference waveform should not have been the ideal master waveform created in MATLAB, but the received waveform in a thorough test. The simulated levels of around 4% with PET is assumed a more accurate estimate.

## 6.4 Further work

Tracking amplifiers and their tracker functions is a large field of research and an important one considering today's requirements of high efficiency, high bandwidth, and high linearity. Further work in this field can go along multiple paths. The possible benefit of digital predistortion and its implementability can be studied. This would further outline the usability in base stations, and one could go further in depth of what this would relate to regarding real throughput. The gain in GaN transistors are highly dependant on the supply voltage, and the effect on other technologies could be studied. The impact tracked amplifiers have on LTE up-link waveforms with SC-FDMA is an exciting study. The incentives for having efficient amplifiers in cell phones are substantially larger and studying the impact of efficiently tracked amplifiers in cell phones could have a significant impact on future phone battery life. There could also be a study of the economic aspect, where a lifetime analysis of tracked vs. non-tracked amplifiers is compared. It is not given that efficient amplifiers are the best when telecom operators are selecting base station equipment.

## 6.5 Final conclusion

Its clear that for wide-band LTE modulation power envelope tracking is the superior tracker function. It can serve to both linearize a RFPA and make it more efficient at the same time(!). Much of the added efficiency benefit depends on an efficient tracker(non-linear), but it is still a point to relieve RFPA heat dissipation. With digital predistortion, power envelope tracking is a prime candidate for base station RFPA topology. Besides, a wide-band tracked system can even be used on multiple LTE channels, a useful feature as modern mobile devices can aggregate multiple channels. The downlink is the primary focus of this thesis, but implementation into mobile devices can be beneficial as power consumption is directly impacting battery life. The use of SC-FDMA and low voltage power amplifiers though makes it unclear to what degree these tracker functions impact efficiency. On a final note, all the LTE base stations that are going to be deployed in the near future are probably already designed. This thesis is however still relevant as future mobile communication schemes probably won't lower the amplifier's requirements.

# Bibliography


[1] *THS3001 Data Sheet*, Texas Instruments.Inc, 2009. [Online]. Available: http://www.ti.com/lit/ds/symlink/ths3001.pdf

[2] *ADA4870 Data Sheet*, Analog Devices.Inc, 2018. [Online]. Available: http://www.analog.com/media/en/technical-documentation/data-sheets/ADA4870.pdf

[3] M. Lorentzen, *Telenor legger ned 3G:- Dette er helt nødvendig*, 2018, 13.06.2018. [Online]. Available: https://e24.no/digital/telenor/telenor-legger-ned-3g-dette-er-helt-noedvendig/24367285

[4] M. Olavsbråten and D. Gecan, "Bandwidth Reduction for Supply Modulated RF PAs using Power Envelope Tracking," *IEEE Microwave and Wireless Components Letters*, vol. 27, pp. 85–87, 2017.

[5] S. Cripps, *RF Power Amplifiers for Wireless communications*. ARTECH HOUSE, INC., 2006.

[6] *LTE Physical Layer Overview*, Keysight technologies, Inc, 2018. [Online]. Available: http://rfmw.em.keysight.com/wireless/helpfiles/89600b/webhelp/subsystems/lte/content/lte_overview.htm

[7] *HackRF One*, Great Scott Gadgets, 2018. [Online]. Available: https://greatscottgadgets.com/hackrf/

[8] *SDR#*, Airspy, 2018. [Online]. Available: https://airspy.com/download/

[9] C. J. Willmott and K. Matsuura, "Advantages of the mean absolute error (MAE) over the root mean square error (RMSE) in assessing average model performance," *CLIMATE RESEARCH*, vol. 30, pp. 79–82, 2005.

[10] S. Haykin and M. Moher, *COMMUNICATION SYSTEMS*, 5th ed. Jhon Wiley & Sons, INC., 2010.

[11] *LT1210 Data Sheet*, Analog Devices.Inc, 2018. [Online]. Available: http://www.analog.com/media/en/technical-documentation/data-sheets/lt1210.pdf

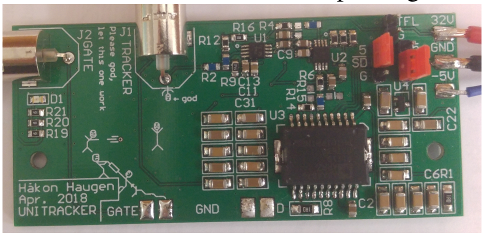
[12] M. Olavsbråten and D. Gecan, "Efficiency Enhancment and Linearization of GaN PAs using Reduced-Bandwidth Supply Modulation," *Proceedings of the 47th European Microwave conference*, 2017.

[13] 3GPP, 2017, lTE, Evolved Universal Terrestrial Radio Access (E-UTRA) Requirements for support of radio resource management (3GPP TS 36.133 version 14.3.0 Release 14).

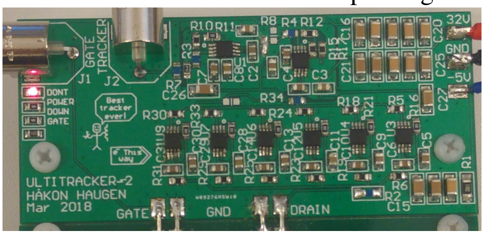
[14] *Measuring ACLR Performance in LTE Transmitters*, Agilent Technologies, Inc, 2010. [Online]. Available: http://ccontrols.ch/cms/upload/applikationen/LTE/5990-5089EN.pdf

# Appendix

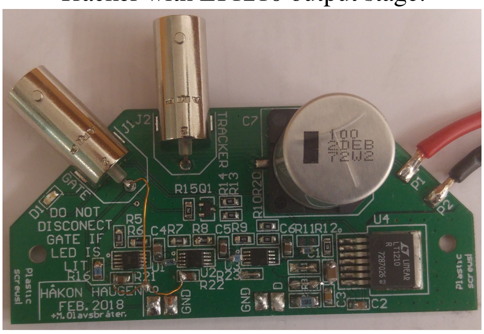
Tracker with ADA4870 output stage:



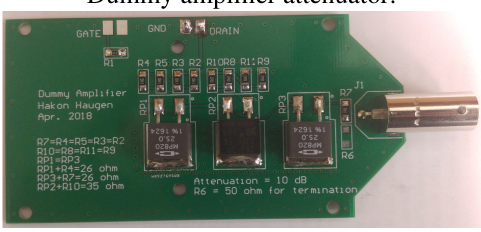

Tracker with 6x THS3001 output stage:



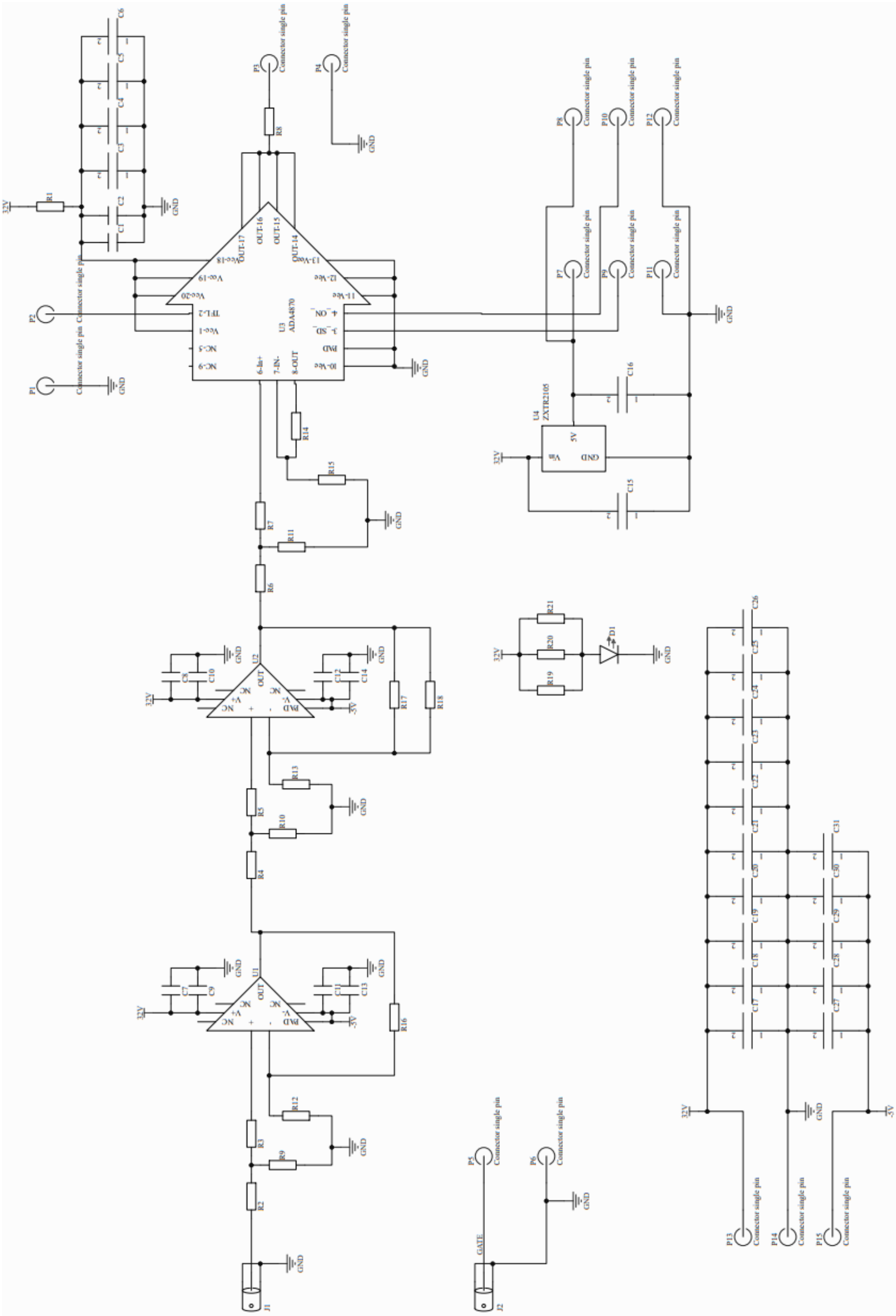

Tracker with LT1210 output stage:



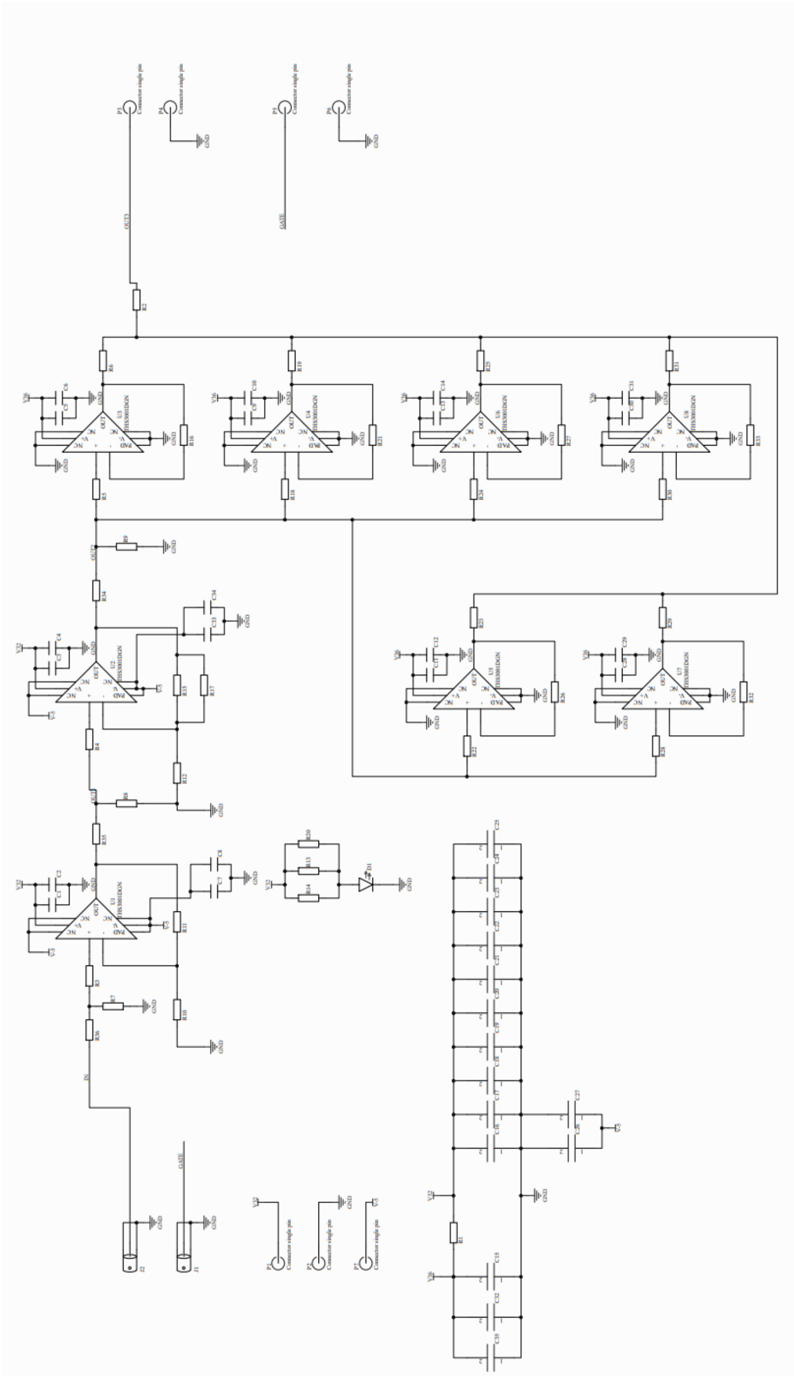

Dummy amplifier attenuator:



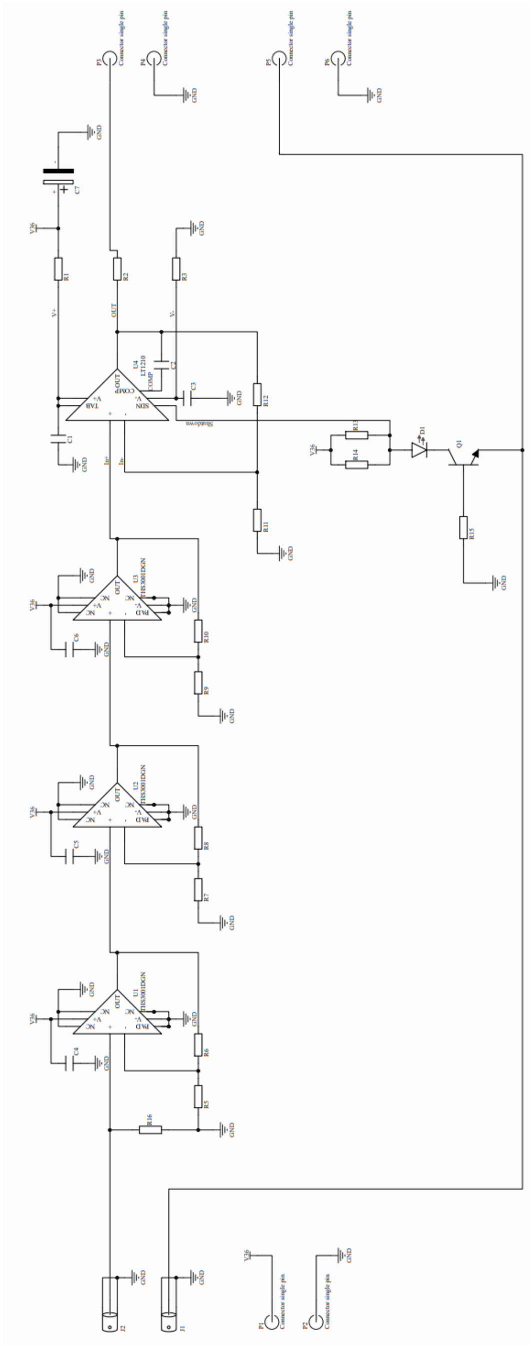
Tracker with ADA4870 output stage schematic:



Tracker with 6x THS3001 output stage schematic:



Tracker with LT1210 output stage schematic:



Dummy amplifier attenuator schematic: