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Ka-Band Satcom Trends and Power

The ever-expanding appetite for data is driving strong growth in satellite communications (satcom), especially in higher-frequency bands. But, instead of concentrating on the lower frequency ranges, many companies are focusing on higher frequency Ka-band which provides more allocated spectrum for applications such as internet access and 5G.

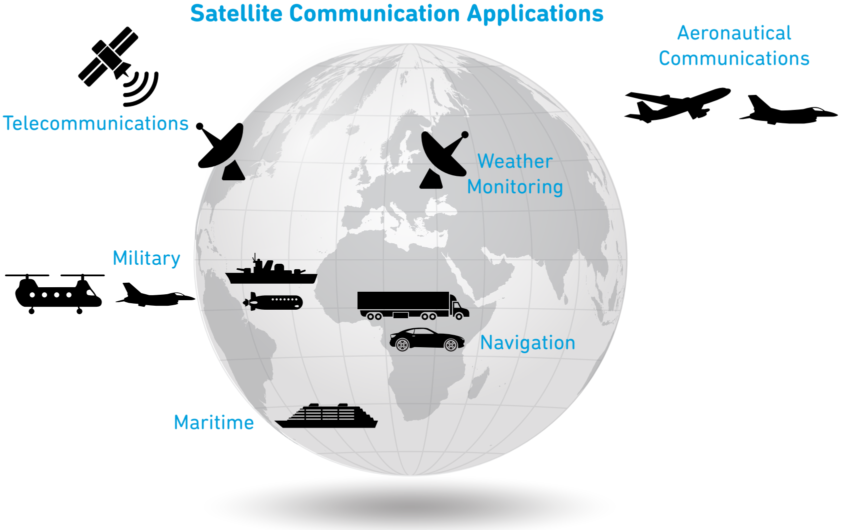
Operating in Ka-band creates new challenges in RF power amplification due to an increase in video bandwidth requirements and the need for high linear power to support multi-carrier systems. Thanks to advances in semiconductor technology, GaN PAs have increased performance and are now an attractive alternative to high-power vacuum tube amplifiers.

This article provides insights into current satcom trends and approaches to Ka-band power amplification.

Satcom's Changing Role in Global Communications

Satcom equipment plays a vital role in the daily lives of people across the world. It supports a broad variety of applications in telecommunications, weather monitoring, aeronautical communications, maritime and military uses, as well as navigation. Satcom in commercial aviation has satisfied the need to be always connected with in-flight connectivity services.

Satellite Communication Applications



The explosive growth in consumer data traffic, 5G, big data and artificial intelligence will further propel the global expansion of satellite systems and services – according to Strategy Analytics – which forecasts the market will grow at 5.4% a year to reach more than $93.6 billion in 2028.

This growth is driving change across the satcom landscape, with some companies planning huge constellations of high-throughput satellites to support the demand.

Ka-Band's Leading Role in Satcom

Responding to the increasing demand, the satcom industry has moved to higher-frequency bands, where more bandwidth is available. In the Ka-band, 3.5 GHz of bandwidth is available for satcom – over 4x more than in other commonly used bands. It's become widely used, especially for uplink (earth-to-satellite) connectivity.

The below table describes the bands used in the Satcom industry. In the Ka-band, many high-profile users include startups like Elon Musk's SpaceX and Amazon's Project Kuiper. These two organizations are planning to launch thousands of small satellites to provide high-speed internet access to consumers and businesses worldwide – including coverage for remote and underserved areas that are beyond the reach of other broadband services. Moreover, there are other organizations following a similar path to take advantage of the new revenue streams opened by the satcom market.

Satcom Bands

Satcom Frequency Band	Satcom Frequency Range	Satcom Bandwidth	Description
C-Band	5.85-6.425 GHz	575 MHz	Primarily used for satellite communications, full-time satellite TV networks or raw satellite feeds.
X-Band	7.9-8.4 GHz	500 MHz	Used for satcom uplink, military satcom. Also used in radar applications.
Ku-Band	13.75-14.5 GHz	750 MHz	Used for satcom uplink. Fixed satellite services & broadcast satellite services.
K-Band	17.3-21.2 GHz	3,900 MHz	Used for satcom downlink. Fixed satellite services & broadcast satellite services.
Ka-Band	27.5-31 GHz	3,500 MHz	Used for satcom uplink, military satcom, 5G telecommunications.

Because the Ka-band supports many revenue streams like stationary and mobile equipment, including satellite gateways, airborne and marine systems, and portable satcom man-packs, it is a key band to exploit.

Ka-Band Uplink Power Amplification Challenges

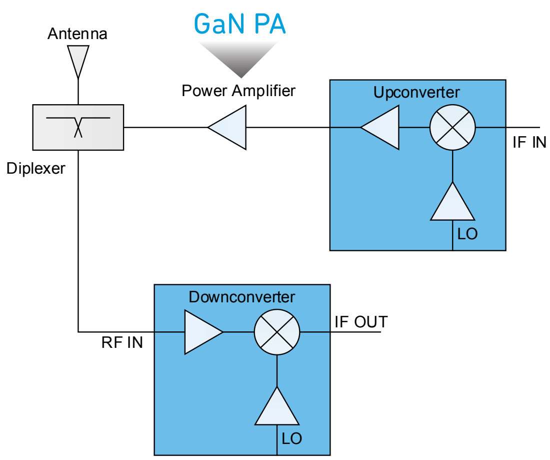
Ka-band transmission creates RF power amplification challenges: satcom equipment must be capable of transmitting at high power over wide bandwidth while maintaining high linearity. Also, the modulation schemes are increasing to enable more transmission data bits per second. Traditionally, QPSK modulation satisfied the tradeoff of data throughput versus signal noise. However, the recent push for higher modulation schemes in the 16 to 64 quadrature amplitude modulation (QAM) is driving the need for higher performance linear amplification.

Traditionally, traveling wave tube amplifiers (TWTAs), a type of vacuum tube, have been the mainstay for power amplification in satcom applications because of their ability to produce high power while maintaining high efficiency.

Advances in GaN semiconductors are changing that trend. This is in part due to RF performance improvements, though it is also attributed to robustness and reliability in comparison to tube amplifiers. Solid state power amplifier customers combine many of the latest GaN power amplifiers to achieve 100W+ of RF output power, making GaN an attractive alternative instead of TWTAs.

The figure to the right shows a typical satcom RF configuration using a GaN PA.

VSAT - Ka-Band

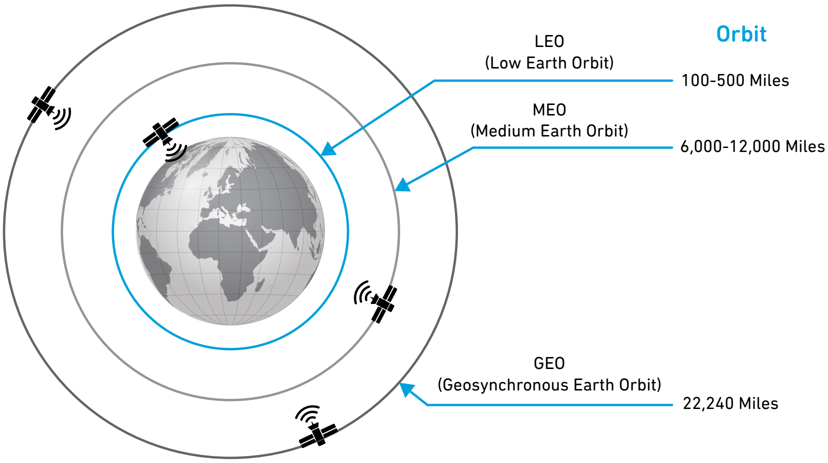


LEOs and Phased Arrays

GaN PAs are also well matched to a major change in satcom architecture: the shift from single antennas to phased arrays of multiple antennas. This change is taking place both on satellite and within ground terminals.

Some of the most ambitious new applications use thousands of satellites in low earth orbit (LEO) – 100-500 miles above earth's surface – much closer to earth than most traditional satellites. See figure on the following page. A big advantage of LEO is lower latency – roughly 20 ms round-trip – which is vital for applications like internet access.

LEO satellites orbit the earth much more rapidly than satellites at higher orbits – traveling from horizon to horizon in only 15 minutes. To maintain continuous communication links with these satellites, earth-based equipment must track them as they zoom across the sky.



This creates a new challenge. Traditionally, earth-based satcom equipment has used mechanically steered antennas to track satellites. But this approach isn't practical for LEO satellites due to the potential wear-and-tear on a system that has to sweep from horizon to horizon every 15 minutes; the maintenance and upkeep for the equipment needed to support very large LEO constellations would likely break the bank.

Electrically steered phased arrays avoid this problem, because they eliminate the need for mechanical movement to track satellites. Instead, these phased array systems have many small antennas that can continuously change the direction of the signal by adjusting the phase of the individual antennas within the array.

While designs based on a single TWTA were well suited to high-power single-antenna systems, designs based on GaN PAs are a natural match for multi-antenna phased arrays. GaN also makes it possible to build a system that's lower-cost, lighter and more compact.

Designing the Future: More Linear Power, Higher Frequencies and Even Greater Bandwidth

Looking to the future, satcom architectures are in a state of change. Higher linear power over wider bandwidths is the present trajectory in the Ku/K/Ka-bands. System operators are moving even higher in frequency to V-band (40-75 GHz) where bandwidth is even greater than Ka-band. As Ka-band was once a frontier in satcom, it is now an established and strong market segment with many operators and equipment providers. V-band is viewed as the new frontier in satellite communications. With the continued innovation we have seen across the satellite industry, the duration from adoption to maturity in V-band could be faster than anyone expects. It is expected that GaN will play in this higher frequency market as well.

Qorvo GaN technology is rapidly evolving to address the satcom market. Successive GaN generations offer increasing linear power output, with improvements in efficiency, enabling equipment manufacturers to use GaN PAs for more high-power applications that traditionally would have required TWTAs. The growing power capability per GaN PA also means equipment makers can build amplifiers with fewer GaN devices, resulting in a lower-cost solution to deliver the required power. More powerful devices help to make systems smaller, simpler to build and more reliable.

Qorvo offers a wide range of GaN devices for satcom ground-based applications and satellites. Some newly developed products recently added to our portfolio are shown below.

Part Number	Features	P_{SAT}	PAE	Small Signal Gain
QPA2210/QPA2210D	27-31 GHz	> 38.4 dBm	> 32%	25 dB
QPA2211/QPA2211D	27-31 GHz	> 41.5 dBm	> 34%	26 dB
QPA2212/QPA2212D	27-31 GHz	> 43.4 dBm	> 25%	22 dB

Foundry or Research Laboratory?

Where Innovation Comes to Life.

Qorvo's 30 Years of Foundry Service Produces Market-Changing Defense and Small Business Innovations

Maintaining a US based foundry is critical for the DoD, our country's defense primes and our key partnerships because it continues to fuel new technology, innovations and maturation of processes. Advancing technologies and manufacturing products in high volume not only helps drive costs down but brings new technologies into the commercial space – as we are seeing with 5G.

We sat down with Senior Director of Research Vijay Balakrishna to discuss Qorvo's foundry, its history and why it has a crucial role in the company's strategy.

Q1: Qorvo's Legacy in RF Innovations – Particularly its Foundry Services – has a Considerable History. Tell us about the Key Milestones.

Qorvo has been a main supplier of gallium arsenide (GaAs) for multiple markets since 1985. Our GaAs (III-V) semiconductor roots stem from several technology companies that came together to form Qorvo.

Altogether, it's been more than three decades of supporting a wide variety of customers with foundry services. We provided GaAs MESFET in the 1980s, pHEMT in the 1990s, gallium nitride (GaN) in the 2000s and we've continued advancing GaAs and GaN technology. Some may think of GaAs as antiquated but really, it's all around us, in airplanes and satellites for example, and just as necessary as ever – even for future 5G technology. We have also expanded foundry services to acoustic filters – surface and bulk acoustic wave (SAW, BAW) filter technologies – benefiting both the defense and commercial markets.

The Qorvo Manufacturing facility in Richardson, TX is fully accredited by the US DMEA as a Category 1A "Trusted Source".

Back in 2005, we worked with the defense advanced research projects agency (DARPA) on its wide bandgap semiconductors RF program (WBGS-RF) to advance critical GaN transistor technology for X-band radar and wideband EW applications. Since then, we've met challenging wideband monolithic microwave integrated circuit (MMIC)-level performance metrics in power, efficiency and bandwidth. Besides being the first to release the 0.25-micron GaN technology (GaN25) in 2008, and offering it as a foundry service in 2009, we were also the first to reach a 65-volt process for GaN. From UHF to Q-band, Qorvo's GaN product line has continued maturation, achieving manufacturing readiness level (MRL) 10 through the DoD's Title III GaN program.

Q2: Since the Merger Between TriQuint and RFMD, there have been some Misperceptions that Qorvo No Longer has an Open Foundry. Is the Foundry Open?

Yes, the foundry is open to strategic customers. We typically work with large defense companies and government research organizations, and we also work with small businesses and universities on research ideas. We engage with many labs, like DARPA, that pursue challenging initiatives and advanced functions that involve creating new designs, running tests and exploring "what if" scenarios on particular technology nodes.

When we have the availability, we're open to working through new designs and innovations and produce both custom and standard products. We also provide wafer and chip options for prototyping.

Q3: What are the Driving Factors of Qorvo's Foundry Services?

Working in lockstep with our customers affords us a hands-on perspective regarding new technology needs and use cases – from 3-5 years to 10 years and beyond. They're the canary in the coal mine, giving us the inside view into future research and shaping the products we develop.

For instance, DARPA's advanced research projects look at needs 10-20 years out. They're driving technology and programs with the future in mind. One example – what do we do with silicon (Si) after Moore's Law, how do we get more out of our chips?

The Army, Air Force, Navy and other research labs look 5-10 years out; the defense primes work approximately three years out. For small businesses and universities, it's hard to say. Their timelines really vary, depending on the customer and the needs of the industry.

Q4: You've Talked about the Technologies – GaN, GaAs, and the BAW/SAW Acoustic Filters – Produced in the Foundry. Why do your Foundry Customers come to Qorvo?

We see our customers evaluating four key services:

For our defense and DoD-based customers, US-based foundries are a critical resource. These organizations are looking for a secure, trusted environment in which to design, develop and test new ideas and technologies. We're proud to offer a DoD-accredited Trusted Source (Category 1A) facility, that has achieved manufacturing readiness level (MRL) 10.

Another service that we offer is a wide breadth of technology with the scalability to meet customer needs. We're one of the few companies that can scale rapidly to manufacture GaAs, GaN and BAW acoustic filters. The ability to scale also provides the benefit of driving production costs down.

In addition to providing quality and reliability, we do cost walkthroughs with our customers to make their solutions affordable. They give us a price point; we make recommendations on how to achieve their cost targets. We work with our customers to reduce test times, improve yields and deliver additional services or packaging. We also discuss specifications and volume needs to drive the costs to the next level of assembly.

Last but not least, we're the only vertically integrated foundry of its kind. This means we offer additional services like packaging, on-wafer tests, visual inspections, and extended foundry services all under one roof. Offering this variety of services is unique – including the ability to process wafers, die, components, packaging, provide consulting, run testing, grow the epitaxy.

Q5: Earlier you Mentioned that Qorvo Provides a Variety of Extra Services for Strategic Foundry Customers. What are some of these Services?

Our services are bucketed into a few different categories:

One is assembly and packaging technologies. Many customers prefer integrated, cost-effective packaging designs – so we offer various high-reliability RF packaging options including ceramic, plastic epoxy packaging, military-grade and high-grade custom metal packaging – all produced in house. Our automated microwave module assembly (AMMA), located in our Texas facility, is where we perform assembly functions. With the addition of die-attach equipment, we fabricate, test, package and ship die-level devices to customers from one secure location.

Another is design consulting. We have world-class researchers and designers available to engage with our customers. It's common that we may recommend small changes to design or custom testing to improve circuit performance or reduce variations and sensitivity. We offer consultations on electronic parameters and can adjust manufacturing characteristics, such as how we test and analyze a new product. We talk with our customers in great detail about our processing capabilities, to understand their needs and provide guidance regarding which processes will work best in their design.

Another service offered is technical assistance – various options in our processes and data acquisition analysis for customers to make better choices with their circuits. For example, the types of transistors, number of capacitors, allowable configurations for interconnect, different capabilities for on-wafer tests, inspection criteria, delivery characteristics are all areas we may offer insight. Our customer base produces highly differentiated products and we help them achieve that.

At times, our customers are looking for a secure foundry space in which they can either choose to develop new designs and technologies, while still maintaining their own IP – with the option to utilize Qorvo senior research fellows and designers for their expertise, or not.

Q6: What are some Significant Technologies that the Foundry has Produced?

One significant achievement is when we were working with a large defense prime contractor on an emergent need for the US Army. We were able to quickly ramp to scale, successfully respond to our customer's needs, and field the technology, thus, working to save lives in Afghanistan.

Another example is within space communications, most recently New Horizons. That mission was notable because it sent the first high-resolution pictures of Pluto and its moons back to Earth, nine years after launching. There's some remarkable history here – including that our space-qualified technology accompanied other explorations like the Curiosity Rover on Mars and the Cassini-Huygens spacecraft near Saturn as well.

We've had more than 250,000 Qorvo components launched in space, for more than three decades – aboard orbital payloads, such as communications and navigation satellites, including programs which support broadband data, telecommunications and global positioning services.

Qorvo has pioneered and perfected many of today's processes with GaN, GaAs and other materials that enable the most powerful RF modules on the market. Notably, GaN – matured through a defense funding model – is now extending data rates and operating frequencies for commercial 5G networks, more than ever envisioned for mobile wireless communications. These foundry-produced innovations are possible due to years of tests, proven reliability in harsh environments, and focus on quality. With a thriving foundry, and more than three decades of innovation under its belt, we're always looking to the future and new research developments.

Qorvo's headquarters in Greensboro, NC is also a design, sales, support center and GaAs/GaN manufacturing facility.

GaN-on-SiC is Driving Advances in Radar Applications

New and increasingly sophisticated threats are driving requirements for radar systems that deliver more instantaneous bandwidth, greater resolution, longer range and multi-beam function.

Traditionally, radar systems have had short pulse widths, narrow instantaneous bandwidths and relatively small duty cycles (e.g., 100 us pulse width and 10% transmit duty cycle). Today, there are requirements across all radar bands for 3x to 5x longer pulse widths and ≥ 50% duty cycle. In some cases, the requirements have been for near-continuous wave operation. Radar system requirements are pushing for more RF output power per element with minimal change in cooling requirements.

To support these system level requirements and reduce system operating costs, the RF hardware must have higher transmit output power and power-added efficiency (PAE), better thermal dissipation and lower receive path noise figure.

How These System Requirements Relate to the RF Module

A challenge for the component supplier is translating system-level needs to component-level capabilities. Generally, higher transmit power and lower receive noise figures equate to greater range and higher resolution – the radar can "see" a smaller target farther away and give the operator more time to react.

The downside of higher power along with near-continuous wave operation is more dissipated heat. In order to reduce the impact of increased heat on performance, the components must have better thermal dissipation and higher PAE. The desire for increased instantaneous bandwidth means more complex design, increased losses and sacrifices in performance that could be achieved in a narrower frequency band application.

Component Technologies

A mix of LDMOS (lower frequency operation), GaAs, SiGe and GaN products are being used in radar systems. LDMOS technology is mature, has high PAE and power density for the transmit path, and good thermal dissipation, but generally only supports relatively narrowband operation at S-band and below. LDMOS generally has lower recurring costs for the component level but requires board-level matching and additional surface mount components. Fully matched LDMOS components for radar applications are rare on the open market.

SiGe technologies allow for large-scale integration of RF and DC functions, low power operation, small component size and wide bandwidths, higher frequency bands and lower recurring costs. In contrast, SiGe has low power density and high non-recurring costs. In the receive path, SiGe components have higher noise figures than both GaN and GaAs technologies. The SiGe technology is well suited for low-power, shorter-range radar applications, signal control functions, large-scale phased arrays and/or high-volume applications.

Mature GaAs high electron mobility transistor (HEMT) technologies can support the bandwidth and the higher frequency bands, but have lower power density compared to GaN. GaAs HEMT remains a viable solution for both transmit and receive components where lower transmit power per element is viable and receive chain noise figure is key.

GaAs HEMT gate lengths continue to decrease, allowing for lower noise figures, which improves the radar resolution, range and sensitivity. Smaller gate lengths may improve RF performance, but the cost is ESD sensitivity and input power survivability. When higher transmit power is required, the GaAs HEMT device solution is to increase gate periphery by adding gates or by increasing the individual gate width. Increasing individual gate widths will limit the operating band of the device and stress the manufacturability of the product. Adding gates to the FET stack complicates the matching circuit, increases design and manufacturing risks and adds insertion loss.

GaN-on-SiC HEMT technologies have higher power density compared to GaAs HEMT, support wider bandwidths in all the operating bands of interest, use a thermally superior substrate (SiC) and are quickly being adopted by the radar markets as the PA solution of choice.

With the higher power density, matching circuit combining structures are simpler and lower loss vs. GaAs HEMTs. Like LDMOS, GaN operates at higher voltages compared to GaAs and SiGe. GaN-on-SiC HEMTs are viable solutions for high input power, robust LNAs. The noise figure of the GaN device is the similar to a GaAs LNA with an input protection limiter.

Qualitative Technology Comparison

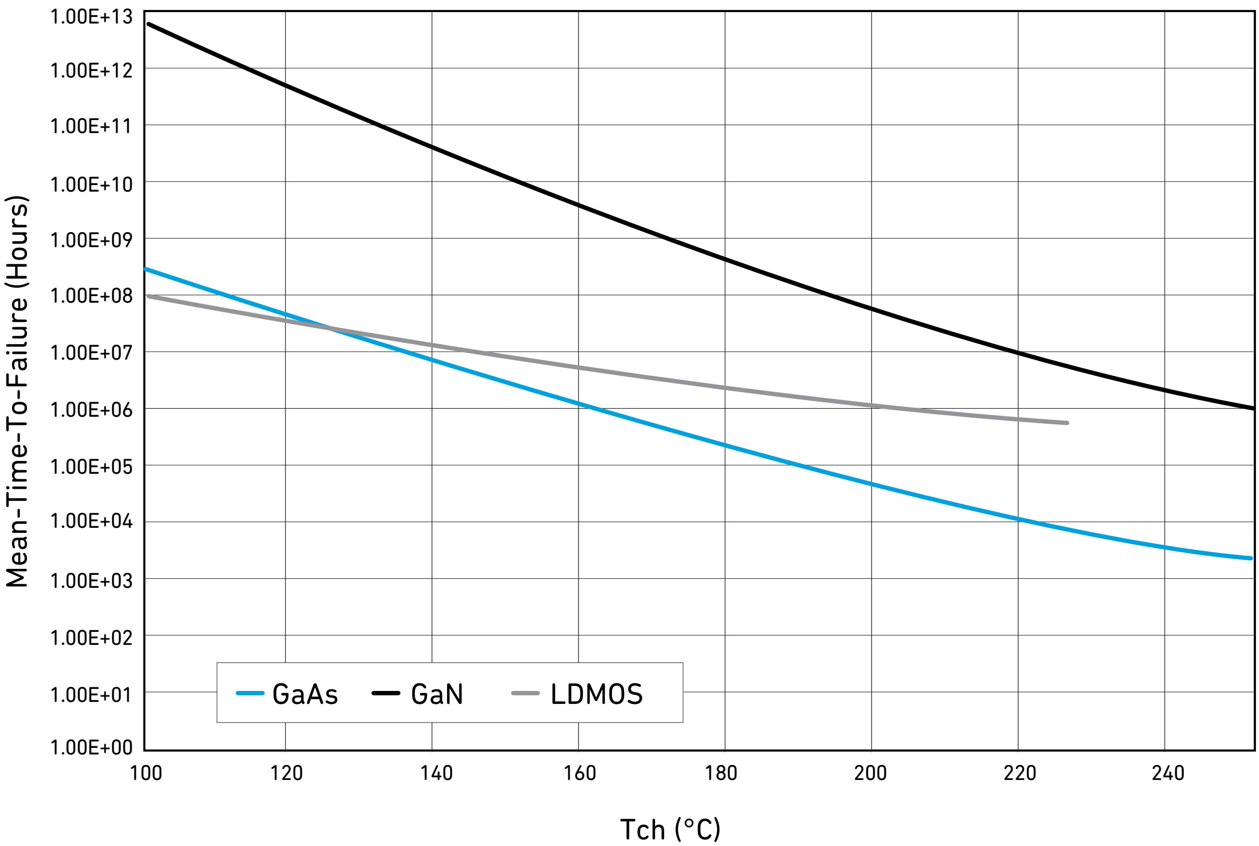
	GaN-on-SiC	GaAs	LDMOS	SiGe
Function	Tx PA & Rx LNA	Tx PA, Rx LNA, RF signal control	Tx PA	Rx LNA & RF signal control
Operating Voltage	High	Low	High	Low
Power Density	High	Low	High	Low
Operating Frequency	All bands	All bands	S-band and below	All bands
Bandwidth	Wide	Wide	Narrow	Wide

The Need for a New Approach

To meet new radar system demands, the products being designed today must meet tougher size, weight, power and cost (SWaP-C) requirements, and have higher PAE, lower channel temperatures and lower noise figures. SWaP-C is especially important in air and space-based systems where physical space and weight are at a premium.

Higher PAE translates to less prime power, lower cooling requirements and lower operating costs. Specifically, in X-band, there are products now on the open market that are pushing >40% higher efficiency with 3 GHz of bandwidth. At L- and S-band radar frequencies, PAEs were in the 50% to 60% range for a discrete GaN HEMT – now products are pushing to 75% and higher. Multi-stage S-band MMICs are approaching 60% efficiency. In the past, these products were not available commercially due to technologies and design capabilities.

The SiC substrate used for GaN is the ideal enabler for the transmit portion of next-generation radars. SiC has higher thermal conductivity when compared with GaAs or Si. GaN-on-SiC transistors can operate at a higher channel temperature than GaAs or LDMOS transistors for the same mean-time-to-failure value.



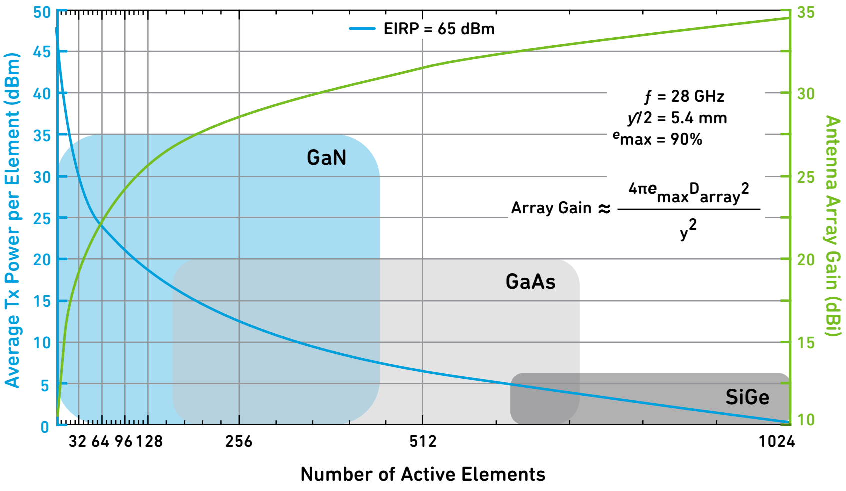
GaN transistors have operating voltage capability that is 2-5X higher than GaAs transistors. GaN operates between 20V and 50V. Higher operating voltages mean lower I^2*R losses are possible. Higher voltage operation also means fewer voltage step conversions between the power supply and the RF devices. These advantages translate to size reductions, less weight, fewer components, lower cost and increased system performance.

GaN offers a 3-5X increase in power density versus GaAs and even greater increase vs. SiGe. Increased power density means fewer components and size reductions.

From a radar antenna pattern perspective, when the equivalent isotropically radiated power (EIRP) remains constant, the PA power per channel increases as the number of elements in the array decrease. More transmit power can translate to fewer elements, smaller size and less complexity.

Use of GaN-on-SiC technology also opens the door to a range of manufacturing options with SWaP-C benefits:

- High thermal conductive materials for die attach and assembly, for increased thermal dissipation (weight and power)
- Direct Cu attach vs. CuMoly or CuW composites, for the highest thermal conductivity is possible



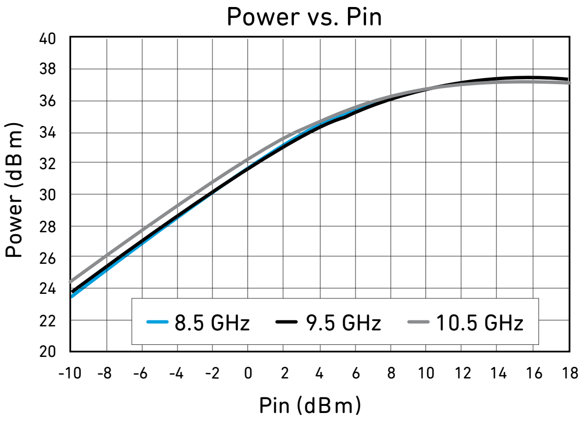
Integration of the PA and LNA into a single, front-end module (FEM) further reduces unit size and number of components. For example, designers working with traditional radar architectures often have five to seven components per channel plus all of the associated peripheral resistors and capacitors.

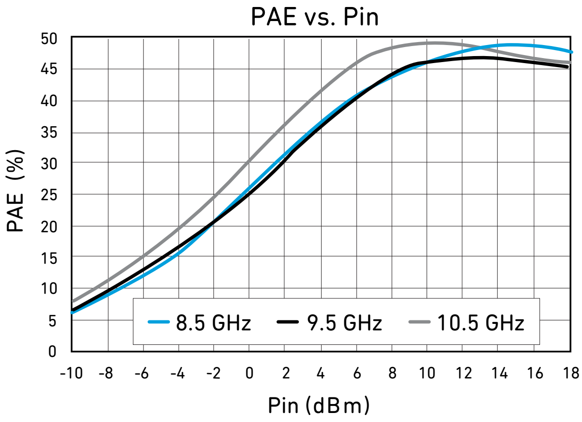
Replacing the traditional RF Tx/Rx module with an integrated GaN-only FEM or GaN/GaAs FEM reduces the number of components to one. This is a significant change in the BoM complexity of the board, simplifying designers' efforts to place the component closer to the antenna to reduce loss and provide higher dynamic range. Designers can create higher-density arrays and achieve greater range for the same power budget.

When the FEM is combined with a SiGe or GaAs "core" circuit, further component count reduction can occur. The core chip can replace the phase, attenuation and control circuits for one or more radar elements, depending on the radar architectures. It is feasible to see a greater than 50% component reduction for a 4-channel sub-array with the right combination of GaN, GaAs and SiGe.

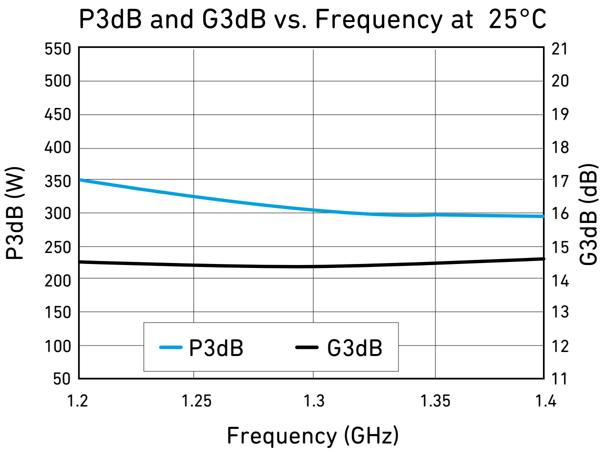
Below are a few examples from Qorvo of newer GaN-on-SiC products where high PAE, continuous wave operation and SWaP-C were primary design requirements in order to support next-generation military and civilian radar applications:

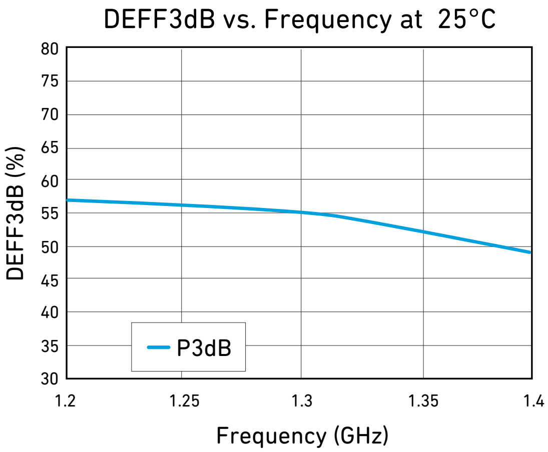
The Qorvo QPA1022 high-performance power amplifier supports X-band phased arrays. Built on the company's 0.15 um GaN-on-SiC process (QGaN15), this amplifier is an integrated, 4x4x0.85 mm package that can support tight lattice spacing requirements for phased array radar applications. In this PA, GaN technology enables best-in-class PAE of 45% at 4 watts RF power in the 8.5-11 GHz range. This is an increase in efficiency by 8% over previous products while providing 24 dB large signal gain.





The Qorvo QPD1006 supports 1.2-1.4 GHz frequency L-band applications and is capable of continuous wave operation at high voltages – 45V CW and 50V pulsed. Since it is fully matched to 50-ohms at the input and output, it supports a smaller module design and fewer components on the system board vs. the typical footprint for an unmatched FET design. The design has 55% drain efficiency for CW operation; 62.2% pulsed operation, with > 300 W CW power and > 450 W pulsed power.





Looking Ahead

To meet and defeat a new generation of threats, defense radars must leverage sophisticated RF technology to achieve higher efficiencies and greater bandwidth.

Designers are turning to GaN technology to deliver these operational enhancements, as well as SWaP-C improvements that are critical in the harsh, space-constrained environments where these systems operate. While GaN may be a relatively young technology compared to others, it continues to mature as a process and adopted for radar applications.

Further, marrying GaN PAs with GaAs LNAs into single front-end modules creates highly integrated, multifunction components that provide advanced capabilities. System operating and manufacturing costs reduce as component count reduces and PAE increases. These capabilities translate into maximum power with minimum heat, higher reliability and lower cost of operation.

Model-Based GaN PA Design Basics

GaN Transistor S-Parameters, Linear Stability Analysis & Resistive Stabilization

Introduction

S-parameter matching is used to maximize gain and gain flatness in simple linear RF/microwave amplifier designs. This same S-parameter data is used to develop matching networks that address amplifier stability. This article discusses the importance of using modeling for basic S-parameter and stability analyses in the GaN PA design process. It introduces the use of models and resistive stabilization techniques to help avoid device instabilities that can affect nonlinear and linear simulations.

In this article, we focus our attention on a simple two-port stability analysis derived from linear S-parameter calculations. We will use a nonlinear Qorvo GaN power transistor model from the Modelithics Qorvo GaN Library, in combination with simulation templates and keysight advanced design system (ADS) software.

Stability Explained

Stability refers to a PA's immunity from possible spurious oscillations. Oscillations can be full power, large-signal problems, or subtle spectral problems that might go unnoticed if not properly analyzed. Even unwanted signals outside your intended frequency range can cause system oscillations and gain performance degradation.

There are two types of stability and measures to analyze PA stability in your system.

- Conditional stability – a system design that is stable when the input and output see the intended characteristic impedance Z0 (50 ohms or 75 ohms) but may be subject to oscillations (exhibiting a negative resistance at the input or output port) for some other input or output impedance.

- Unconditional stability – a system that is stable in any possible positive real impedance inside of the Smith Chart. Note, that any system design can oscillate if it sees a real impedance that is negative (outside the Smith Chart). But generally speaking, if a system is defined as unconditionally stable, it is stable at all frequencies (where the device can have gain) and all positive real impedances.

Measures of stability

Let's begin with the well-known "k-factor" and stability measure "b" to determine frequency ranges that cause instability at a given bias. These are given by the following equations[1]:

$$k = \{1 - |S_{11}|^2 - |S_{22}|^2 + |S_{11}*S_{22} - S_{12}*S_{21}|^2\} / \{2*|S_{12}*S_{21}|\}$$

and

$$b = 1 + |S_{11}|^2 - |S_{22}|^2 - |S_{11}*S_{22} - S_{12}*S_{21}|^2$$

Unconditional stability is indicated by k > 1 and b > 0.

However, because this criterion requires two parameters to check for unconditional stability, a more compact formulation is given with the following "mu-prime" parameter[2]:

$$mu\_prime = \{1 - |S_{22}|^2\} / \{|S_{11} - conj(S_{22})*Delta| + |S_{21}*S_{12}|\}$$

If mu_prime > 1, it indicates unconditional (linear) stability.

Matching and tuning to attain stability

As noted above, S-parameter data is used to develop matching networks to attain amplifier stability. Figure 1 shows a single-stage amplifier configuration and the key parameters that affect gain and stability. In the unconditional stability region, maximum gain is achieved by setting Γ_s and Γ_L to conditions attaining a simultaneous conjugate match at both ports.[1]

Single-stage amplifier configuration showing key parameters affecting gain and stability calculations

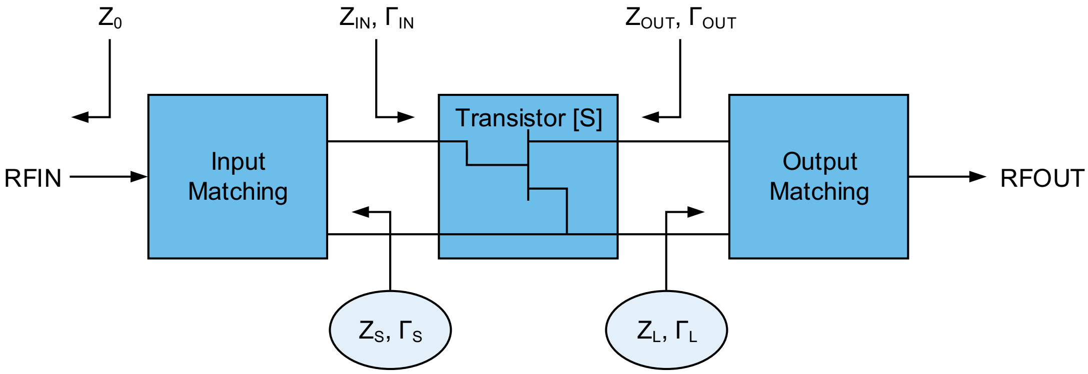


Figure 1.

Linear Stability Analysis

Stability measurements of untuned transistor

Let's consider an example. Figure 2 shows a simulation setup for linear S-parameter analysis of the nonlinear model for Qorvo's T2G6003028-FS GaN HEMT device, included in the Modelithics Qorvo GaN Model Library.

Setup with no stabilization added

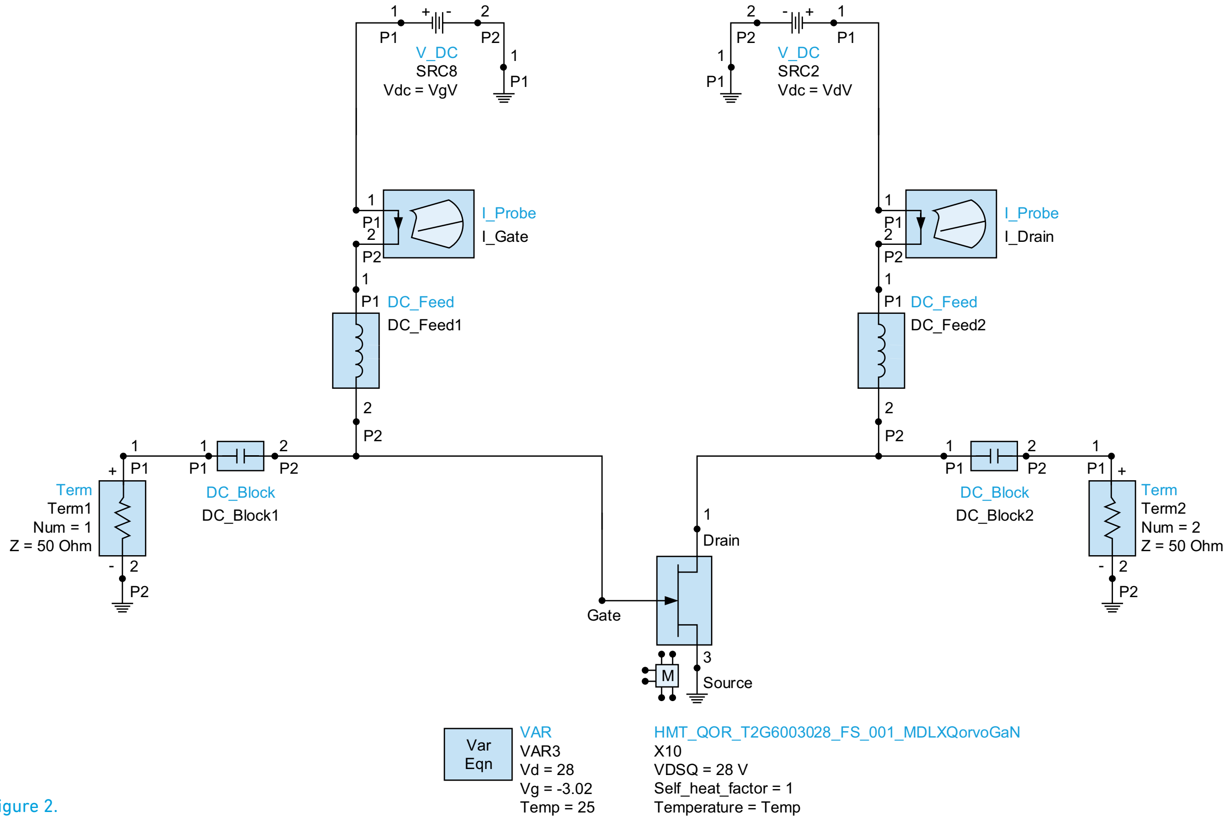


Figure 2.

Note: Bias condition for all simulations in this note is set to Vds = 28 V, Vgs = -3.02 V, which corresponds to a drain current of approximately 200 mA.

In the schematic above, icons represent parameters that can be calculated from device S-parameters, including stability k, b and mu_prime. The "MaxGain1" parameter is the maximum available gain. The "MaxGain1" parameter calculates the maximum available gain for frequency ranges where the device is unconditionally stable, and displays a value that is termed the maximum stable gain. This is calculated as simply |S21|/|S12| for regions of conditional stability.

Figure 3 shows the MaxGain1 parameter, the 50 ohm gain (S_{21} in dB) and stability factor k, measure b and mu_prime calculated from the schematic of Figure 2 (at m5). This plot shows that the stability measure b is > 0 and stability factor k > 1. The stability measurement parameters show a clear break point at about 1.85 GHz (m5). This is the transition frequency between conditional and unconditional stability regions. For 3.5 GHz the maximum gain indicated by this simulation parameter is approximately 18.4 dB (marker m3 in Figure 3). Note: The maximum available gain goes to 0 dB at about 10.4 GHz; this frequency is referred to as the maximum frequency or f_{max}. It is also a good practice to analyze stability from a very low frequency to at least f_{max}, which is why the frequency range for this example was set to sweep from 25 MHz through 12 GHz.

From this analysis, we can conclude the following:

- The device is unconditionally stable above 1.85 GHz.
- Frequencies below 1.85 GHz device are conditionally stable.

These S-parameters produced from the schematic simulation (Figure 2) are show in Figure 4. S_{11} and S_{22} are displayed on Smith Charts, while polar charts are used for S_{21} and S_{12}.

Notice the large difference between the gain for 50 ohm input and output ($|S_{21}|$ in dB) and the MaxGain1 value. This is due to the mismatch associated with S_{11} and S_{22} in a 50 ohm system.

Gain and maximum gain (left) and stability metrics (right)

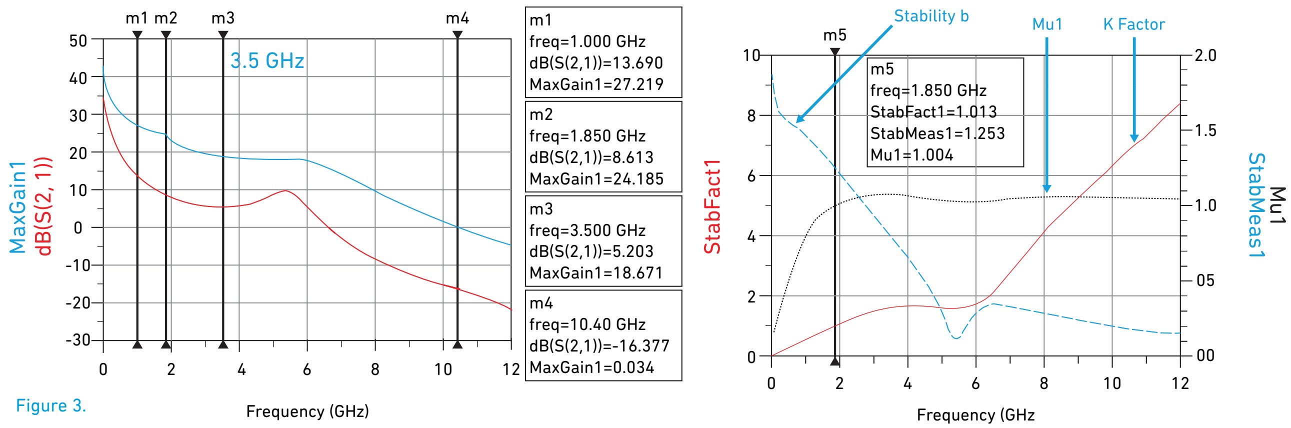


Figure 3.

Plotting the stability circles in the input and output planes provides additional insight. Also included in the schematic of Figure 2 are the icons for "S_StabCircle" and "L_StabCircle", which correspond to calculations of stability circles in the input and output planes.

The meanings of these circles can be described as follows. In the case of the input stability circle at 25 MHz, indicated by marker 14 in Figure 5, each point along that circle represents a Γ_s value that will result in a Γ_{out} value equal to 1 according the following relation.

$$\Gamma_{out} = S_{22} + S_{12}*S_{21}*\{ \Gamma_s / (1-S_{11}* \Gamma_s)\}$$

Eq. 1

S-parameters plotted from setup of Figure 2

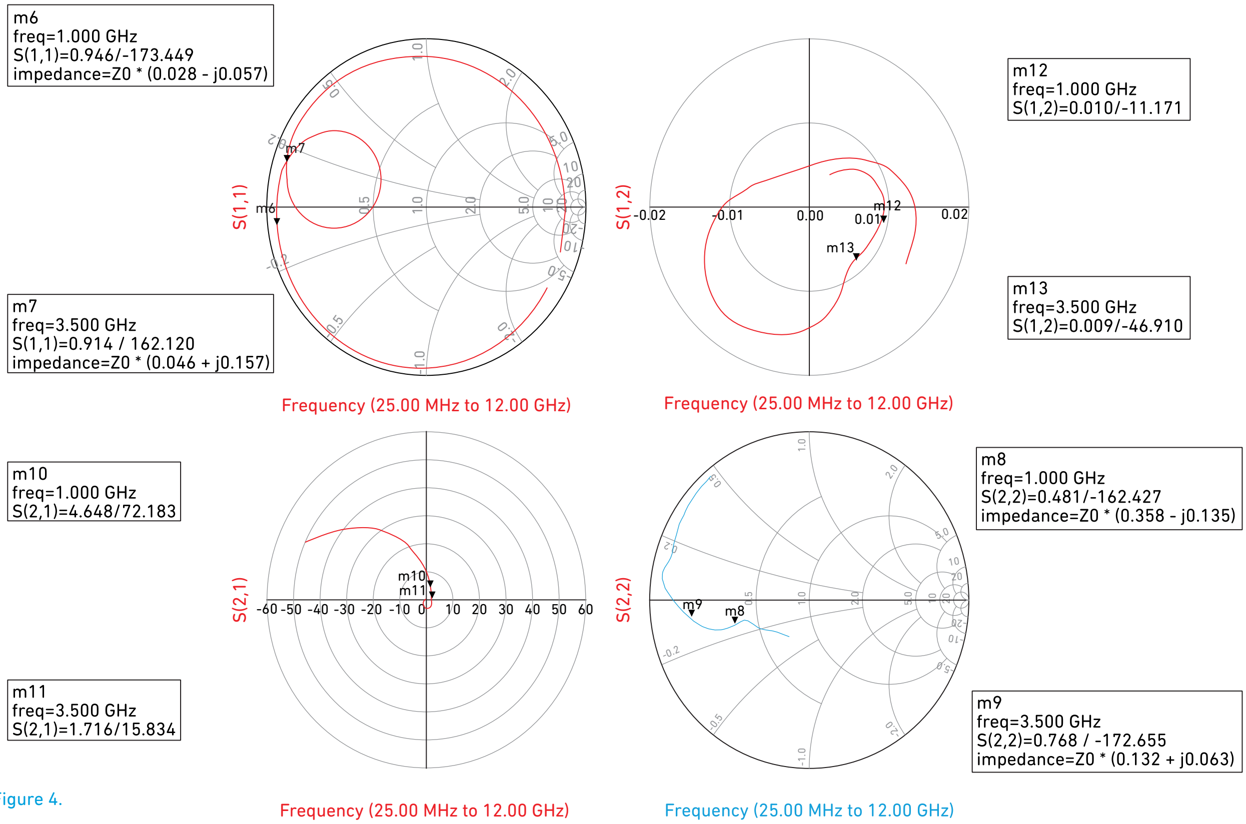


Figure 4.

This circle sets a boundary between $\Gamma_{out} < 1$ and $\Gamma_{out} > 1$, the significance of which is that $\Gamma_{out} > 1$ corresponds to a negative resistance at the output port, which is a condition that can lead to an oscillation. The question then becomes whether the inside or outside of the circle is the unstable ($\Gamma_{out} > 1$) region. A quick check in the case of $\Gamma_s = 0$, which is the 50 ohm point. Note from Eq. 1, for this case $\Gamma_{out} = S_{22}$, which is less than 1 at all frequencies being analyzed here. From this, we can conclude the outside of the circle is the stable region and the inside is the unstable region.

The explanation of the output stability circles is basically the same, except here we are plotting circles of points for Γ_L for which $\Gamma_{in} = 1$, according to the Eq. 2. By a similar argument, we can conclude that it is the inside of the circles plotted on the right side of Figure 5 that correspond to the unstable regions. Note - the frequency plan of Figure 2 was reduced to show fewer circles in Figure 5 for clarity.

$$\Gamma_{in} = S_{11} + S_{12}*S_{21}*\{\Gamma_L / (1-S_{22}*\Gamma_L)\}$$

Eq. 2

Stability circles - the source (left plot) and load (right plot) reflection coefficient reference planes

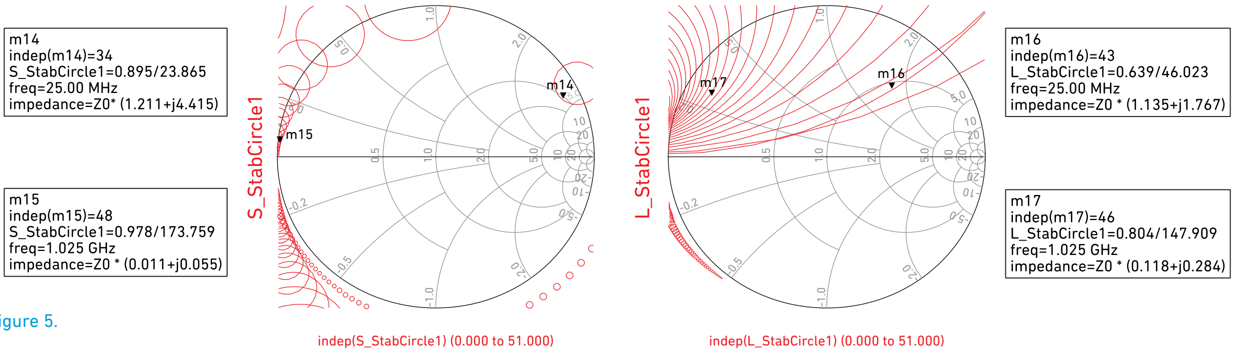


Figure 5.

Linear Stability Analysis

So, what if a device does not meet the requirements for unconditional stability, like in our example for frequencies below 1.85 GHz?

There are multiple matching methods to help stabilize your circuit. In this article we describe two methods. One is resistive and the second is frequency-dependent stabilization.

- Resistive: uses matching resistors to provide stabilization
- Frequency-dependent: uses resistors, inductors and capacitors to provide stabilization

Resistive stability for microwave PA design

Matching resistors can be employed in our example to help stabilize high-gain, low-frequency transistors in most microwave applications. These resistors can be series or shunt at the input or the output, can be in the parallel feedback loop, or included in the bias networks. For PAs, we want to maximize output power, so it's best to avoid resistors in the output network. Feedback amplifiers are outside the scope of this post, so we will concentrate on the series and shunt resistors in the input network.

Figure 6 shows where both series and shunt resistors have been added in the input network. The values are tuned to achieve unconditional stability over the entire 0.025 to 12 GHz frequency range. The resulting stability measurements are plotted in Figure 7. These show the transistor has unconditional stability over the entire frequency range. Note, however, fmax dropped from 10.3 GHz to about 8.75 GHz. Comparing the maximum gain estimation in Figure 7 (design frequency of 3.5 GHz [12.3 dB]) with Figure 3 achieved without this stabilization (18.4 dB), we can see we have incurred a 6 dB degradation in maximum available gain. This is caused by adding a purely resistive input stabilization network. The S-parameters of the resistively stabilized device are displayed in Figure 8, with the overlaid S-parameters of the non-stabilized device. We can see that S_{11} and S_{12} have been affected over the entire frequency range, and S_{21} is also reduced with only minimal change in S_{22}. It is gratifying to observe in Figure 9 that with the resistive stabilization network added, the stability circles are now all outside of the Smith Chart in both the source and load-planes.

Setup with resistive stabilization

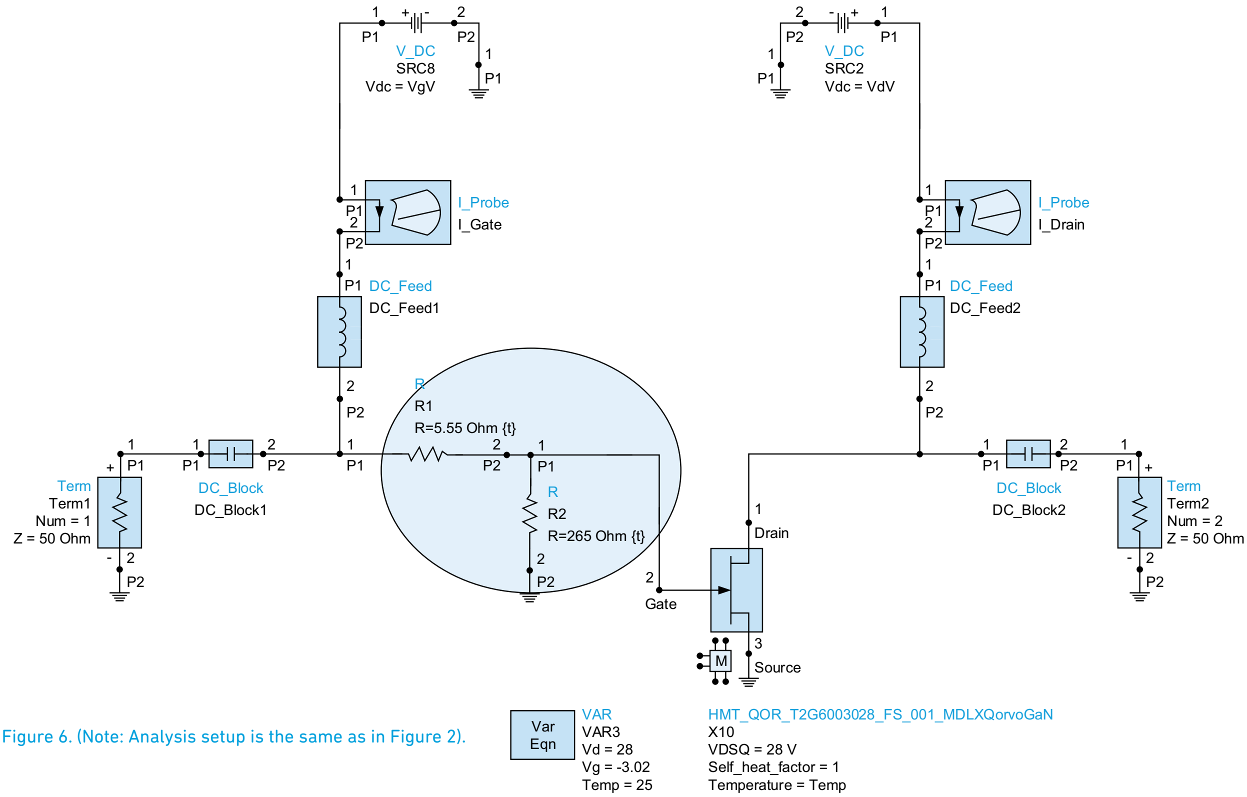


Figure 6. (Note: Analysis setup is the same as in Figure 2).

Resistively stabilized - gain and maximum gain calculated by ADS (left) and stability metrics (right)

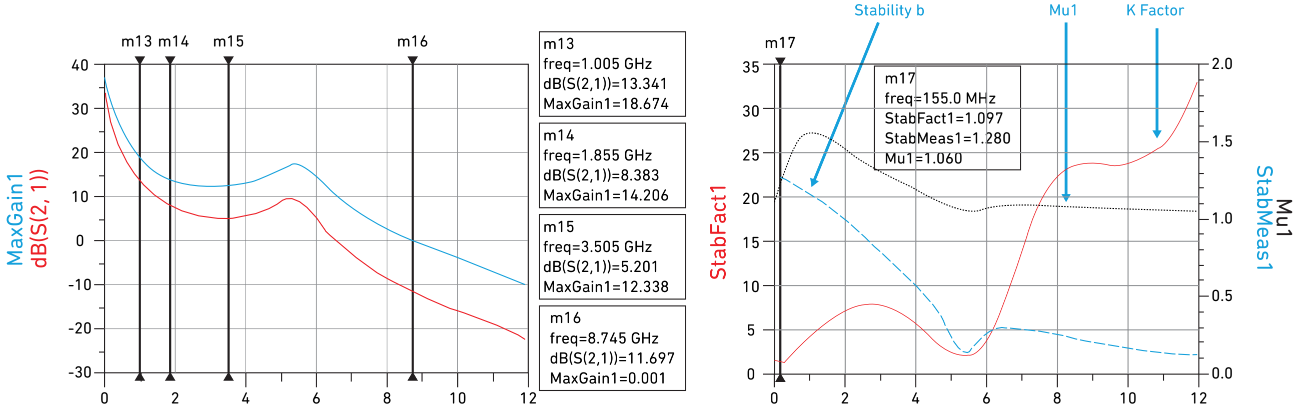


Figure 7.

S-parameters for resistively stabilized schematic versus the non-stabilized device

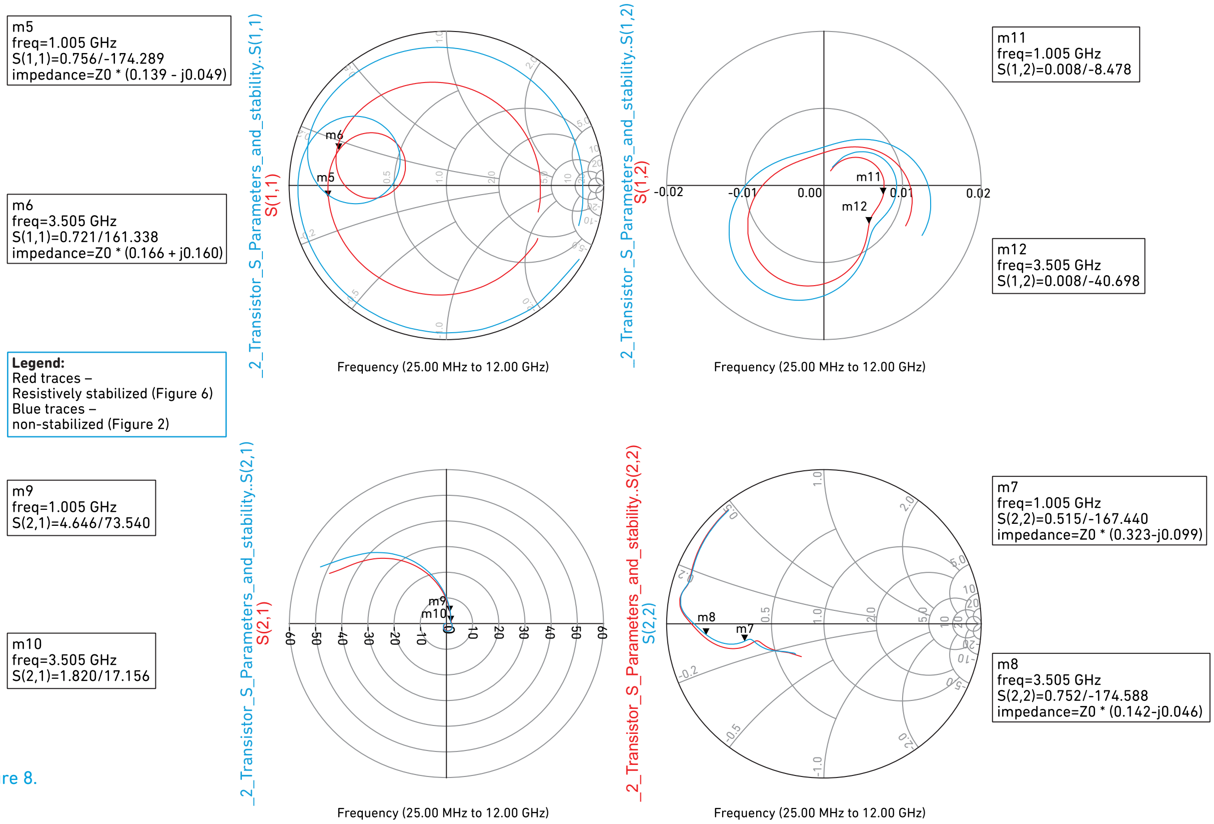


Figure 8.

Stability circles - the source (left plot) and load (right plot) reflection coefficient reference planes for resistively stabilized device (Figure 6)

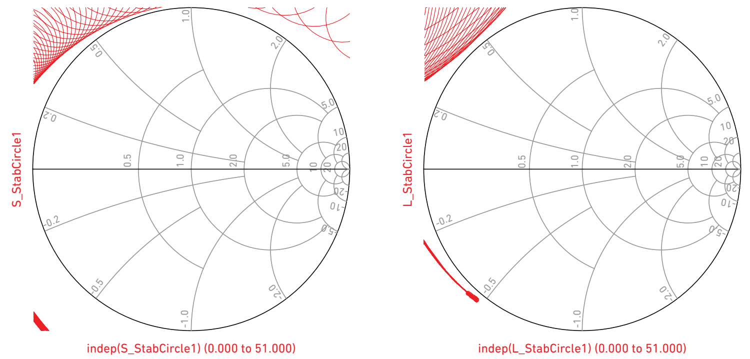


Figure 9.

Frequency-dependent resistive stability

If the design frequency is above 1.85 GHz (e.g., 3.5 GHz), we can implement a frequency-dependent resistive approach using the series-shunt stabilization network. Let's see if we can mitigate the above gain penalty using this approach.

In Figure 10, a resistor (R1) has been incorporated into a modified gate bias network. Additionally, a capacitor (C3) has been placed across the series stabilization resistor (R1). The value of this capacitance can be tuned to adjust what frequency the series resistor (R1) is - effectively shorting it out (making it not "seen"). This can help increase the available gain.

Setup with frequency-dependent stabilization

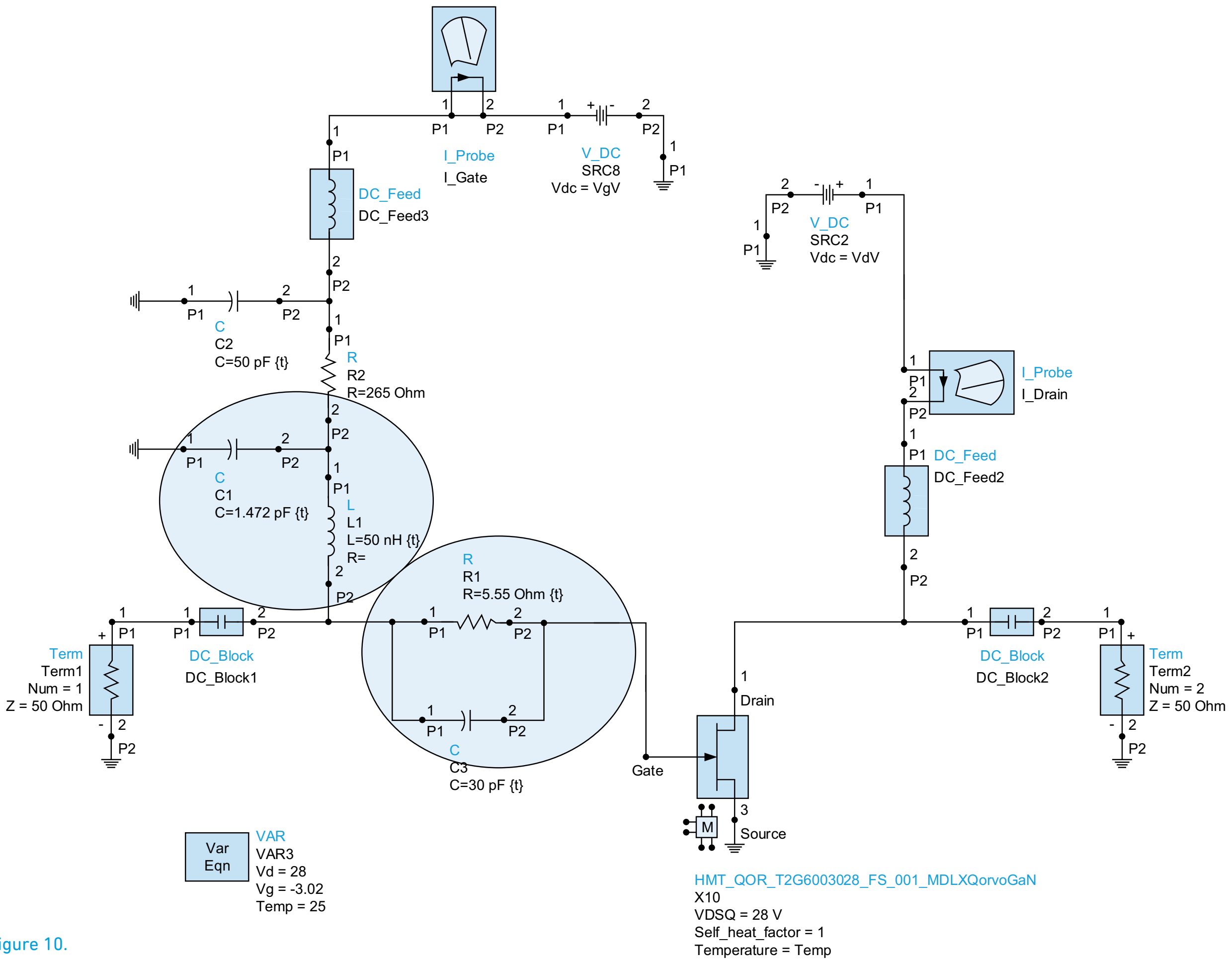


Figure 10.

The inductor (L1) and capacitor (C1) are used to create a low-pass filter. This prevents the resistor (R1) from being seen at higher RF frequencies, or lower frequencies for stabilization. The gain, stability and S-parameter analysis for this solution is shown in Figure 11, Figure 12 and Figure 13. As shown, the frequency-dependent stability network provides unconditional stability across the full frequency range, while reducing the impact on maximum available gain at 3.5 GHz. Note the gain at 3.5 GHz is now reduced by only about 1 dB compared to the non-stabilized device, and also the f_{max} is about the same as the non-stabilized device (~10.4 GHz). In examining the S-parameter comparison to the non-stabilized device as shown in Figure 12, we see that, in contrast to the resistively stabilized device, the S-parameters are not altered over the entire frequency range, but rather only at lower frequencies, as desired. Figure 13 just confirms that none of the stability circles overlap with the Smith Chart in either the source or load planes as expected for an unconditionally stable circuit.

Frequency-dependent gain and maximum gain (left) and stability metrics (right)

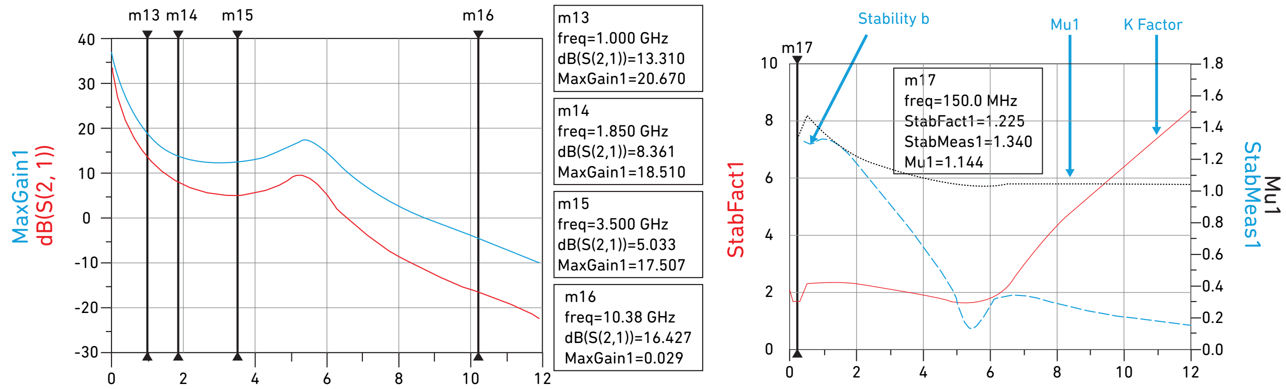


Figure 11.

S-parameters for frequency-dependent stabilized schematic versus the non-stabilized device

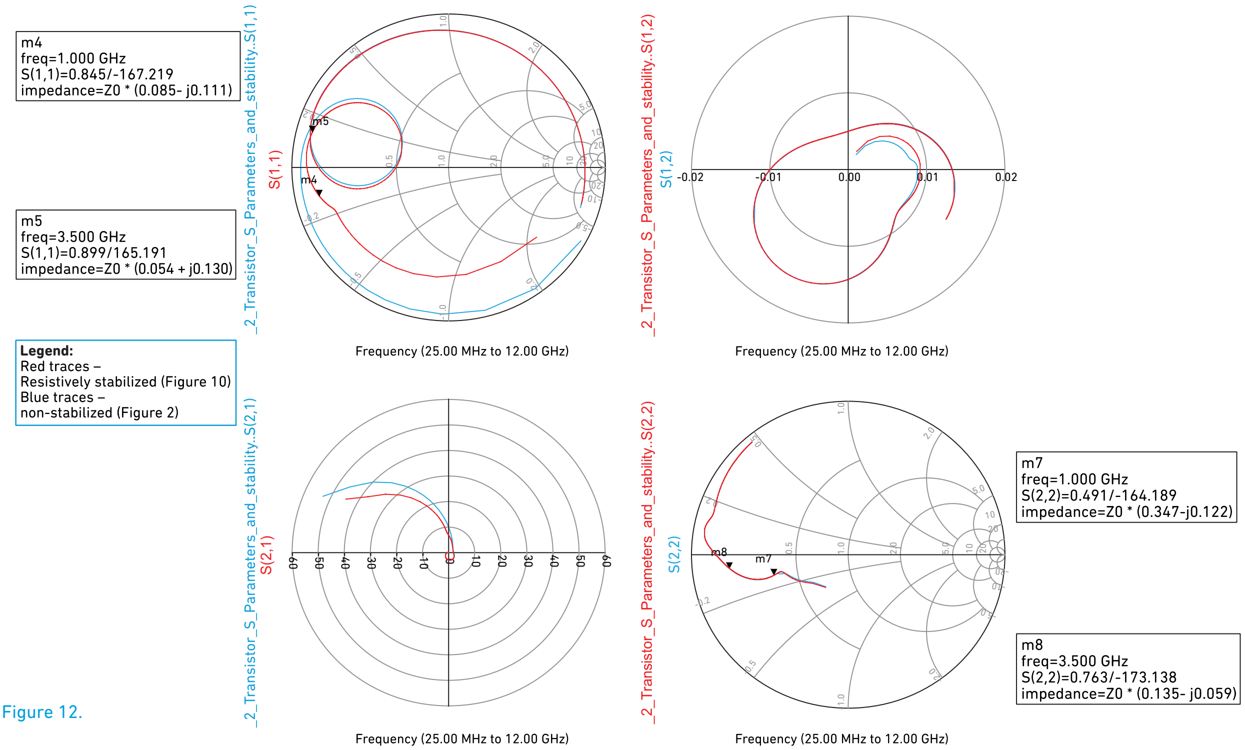


Figure 12.

Stability circles - the source (left plot) and load (right plot) reflection coefficient reference planes for frequency-dependent stabilized device (Figure 10)

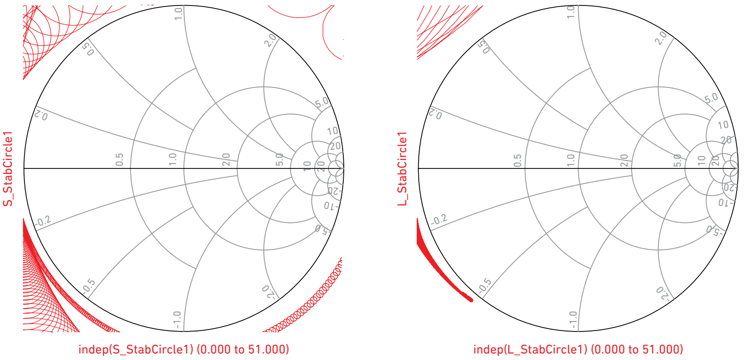


Figure 13.

Key Results

So what are the key findings? As shown in the below data, stability and gain are optimized when using frequency-dependent stability.

No stabilization: 18.373 dB maximum available gain at 3.5 GHz – Figure 3
- Unconditionally stable above 1.85 GHz
- Conditionally stable below 1.85 GHz

Resistive stability applied: 12.334 dB maximum available gain is at 3.5 GHz – Figure 7
- Unconditionally stable over entire 0.025 to 12 GHz frequency range
- 6 dB maximum available gain degradation

BEST RESULT: Frequency-dependent stability applied – 17.5 dB maximum available gain at 3.5 GHz – Figure 11
- Unconditionally stable over entire 0.025 to 12 GHz frequency range
- Increase in maximum available gain of 5.166 dB above resistive stability

Summary

Modeling helps address common design problems such as stability prior to testing your application on the bench. By accurately modeling and implementing stability techniques, we can match and tune for optimal S-parameter performance while maintaining unconditional stability.

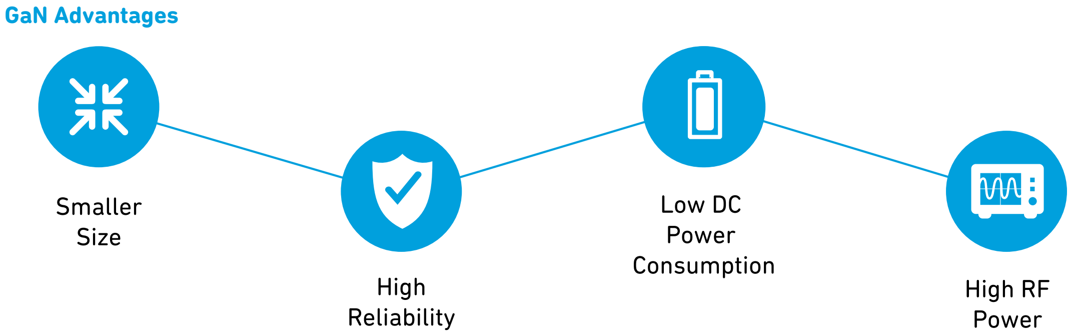
As a final note, the stabilization networks explored here used ideal lumped elements. In an actual microwave design, you will need to include microstrip interconnects and accurate parasitic models for all RLC components, whether you are doing a MMIC design or a board-based hybrid design with lumped elements.

How GaN is Changing the Satcom RF Front-End

Introduction

Solid-state technologies such as GaN are transforming satcom. GaN's advantages of high RF power, low DC power consumption, high reliability and smaller size (which reduces system weight) are opening new markets and revolutionizing the RF front-end (RFFE) in existing satcom applications.

For many years the TWTA and GaAs were the go-to RFFE technologies for power amplification in satcom – with TWTAs used for high-power applications and GaAs employed in lower-power applications and as a pre-driver. But the situation has changed rapidly in recent years, due to advancements in GaN. Now, GaN and GaAs semiconductor technologies are becoming the go-to solutions; GaN has been replacing TWTAs due to its high-power performance and reliability combined with a small form factor. GaN and GaAs are enabling a wide variety of commercial and military satcom applications, such as 5G backhaul, ultra-HD TV transmission, satcom-on-the-move, internet access for aircraft passengers, and manpack (portable) terminals.



Satcom Trends

Satcom equipment plays vital roles in the global communications ecosystem and the daily lives of people across the world. It supports a broad and expanding variety of applications in telecommunications, weather monitoring, aeronautical communications, maritime applications, military uses and navigation (see Figure 1). According to MarketsandMarkets research, the satcom equipment market is projected to grow at about 8.5% a year to reach $30B by 2022. Strategy Analytics forecasts spending on global military communications systems and services will grow to over $36.7 billion in 2026, representing a CAGR of 3.5%.

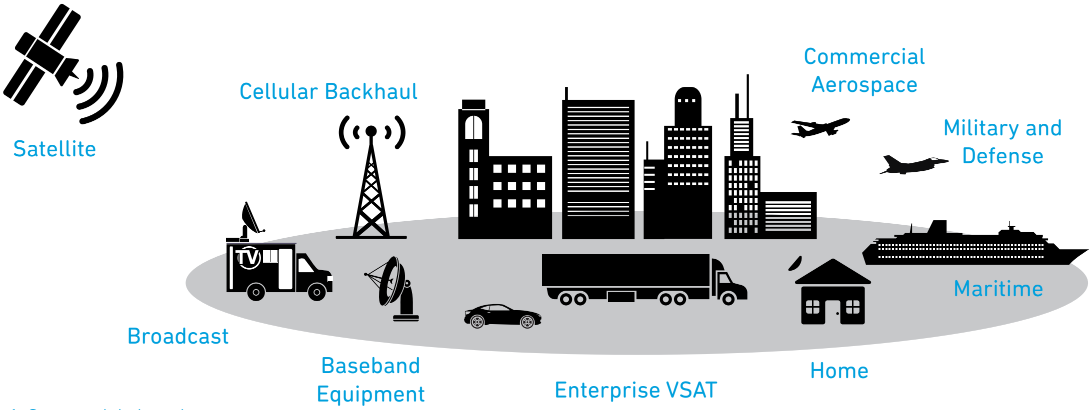


Figure 1. Satcom global markets.

Several broad trends are creating new and more challenging requirements for the RFFE in satcom equipment. The increasing use of smaller satellites and portable, mobile satcom devices is driving a need for more compact, lightweight components with lower power consumption. Additionally, components need to handle much greater bandwidth and data throughput to support advances such as 5G, ultra-HD TV, un-interrupted and secure communications. There's also pressure, to reduce development costs and increase reliability.

These trends are propelling the transition from TWTAs to solid-state devices that support higher data throughput and smaller form factors. Though GaAs and Si have been used in some systems, GaN offers significant advantages for high-power amplification in satcom applications. Its high saturation velocity, high breakdown voltage, and thermal conductivity result in an order of magnitude improvement in power density and high reliability under thermal stress. As a result, GaN is uniquely suited to the high-power requirements of satcom, very small aperture terminal (VSAT), point-to-point (PtP) and base station applications, as shown in Figure 2.

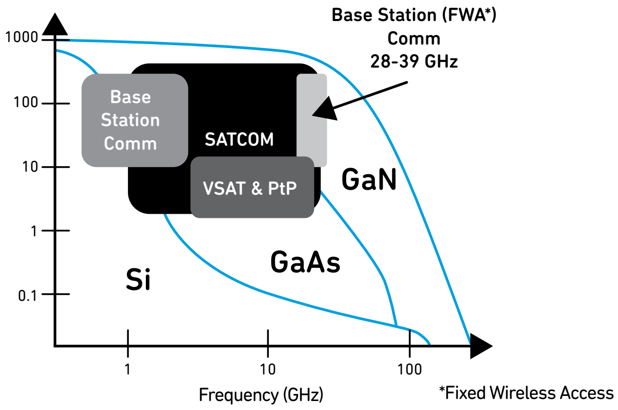


Figure 2. Suitability of semiconductor technologies to different applications.

GaN's potential for the space and satellite communications sectors is only beginning to be realized. The high RF power, low DC power consumption, lightweight, small form factor and high reliability will enable manufacturers to downsize the RFFE. For example, GaN is expected to facilitate weight reduction in satellites and aeronautic applications.

Frequency Bands

The satcom industry has progressively moved to higher-frequency bands to support growing demand for bandwidth, including the X, Ku, K, and Ka-bands as shown in Figure 3. GaN easily supports high throughput and wide bandwidth across these higher frequencies. Today, many of the same satcom components are used across multiple military, space and commercial applications in these bands.

IEEE Microwave Band	Frequency Range	Description
L-Band	1-2 GHz	Global positioning system (GPS) carriers and also satellite mobile phones, some communications at sea, land and air
S-Band	2-4 GHz	Weather radar, surface ship radar, and some communications satellites
C-Band	4-8 GHz	Primarily used for satellite communications, for full-time satellite TV networks or raw satellite feeds
X-Band	8-12 GHz	Primarily used by the military, also used in radar applications
Ku-Band	12-18 GHz	Used for satellite communications, fixed satellite services and broadcast satellite services
K-Band	18-26 GHz	Used for fixed satellite services and broadcast satellite services
Ka-Band	26-40 GHz	Communications satellites, uplink in either the 27.5 GHz and 31 GHz bands, and high-resolution, ctlose-range targeting radars on military aircraft

Figure 3. IEEE microwave bands.

Replacing TWTAs

Until recently, TWTAs were the mainstay in many satcom applications because solid-state devices weren't capable of producing similar power levels. However, power combining techniques now make it possible to generate much higher power using GaN, enabling the replacement of TWTAs with more-reliable solid-state devices.

The GaN power-combining approach aggregates the output power from several single power amplifier MMICs using fully isolated coupling networks. An example is Qorvo's Spatium®, which is a spatial combining product that uses a patented spatial combining technique to offer high RF power, high efficiency and broadband operation. Spatium uses broadband antipodal fin-line antennas as the launch to and from the coaxial mode, splitting into multiple microstrip circuits (see Figure 4). It then combines the power from these circuits after amplification with a power MMIC. A typical Spatium design combines 16 devices, with a combined loss of 0.5 dB. Spatium is used in Ka-band satellite earth stations that operate at 100 Watts and 27-31 GHz, covering both military and commercial bands. Within these stations, it is used in the transmitter side at the antenna hub in block up-converters (BUCs).

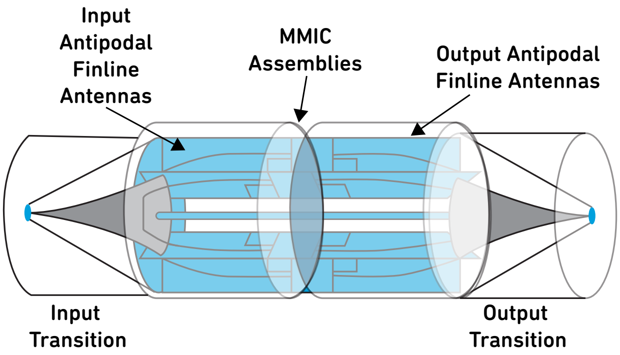


Figure 4. Spatium spatial combining patented design.

This solid-state power-combining approach offers several performance advantages over TWTAs.

- TWTAs need to warm up before they can attain stable RF performance. Warming the tube can take a few minutes. Mitigating the problem requires backup systems running in hot standby. This results in large energy costs. These back-up systems are not required when using Spatium, as no warm up is required.
- TWTAs require high voltage power supplies, typically in the multi-kV range, thus increasing system power consumption. GaN devices do not require high-voltage power supplies.
- Solid-state GaN generates lower noise and has better linearity than TWTAs. Noise figures for medium power TWTAs can be around 30 dB, versus about 10 dB for a solid-state GaN MMIC PA. Another operational benefit of the GaN transmitter is the reduced harmonic content in the output signal.

GaN Advantages for Satcom

GaN offers a range of other advantages over both TWTAs and other solid-state technologies for satcom applications.

Reliability and ruggedness. Reliability is extremely important in satcom applications. GaN offers much higher reliability than TWTAs for several reasons. With TWTAs, a failure in the tube causes a total performance breakdown. In contrast, a spatial combining technique like Spatium increases robustness and reliability. The failure of one transistor does not mean the entire unit shuts down; instead it continues to function using the remaining GaN amplifier MMICs. Each solid-state device is also highly reliable: although the lifetime of a transistor is limited due to electromigration, time-to-failure is typically over 100 years.

The higher power-efficiency of GaN also reduces heat output, which further contributes to higher reliability. Furthermore, wide-bandgap GaN tolerates much higher operating temperatures, so the cooling requirements in compact areas may be relaxed without compromising performance and reliability. This reduces the need for cooling fans and heatsinks, which reduces the weight and size of satellites and therefore the cost of launching them into orbit.

Small, lightweight devices. Weight and size are becoming critical factors in satcom applications, with the trend to smaller satellites and the growth of other on-the-move satcom applications. GaN's high-power output and on-resistance and breakdown voltage allows satellites and other applications to reach target power output levels with smaller devices. Higher power density results in less weight and size per given unit of power output. The high breakdown field allows higher voltage operation and increased efficiency and helps to ease impedance matching requirements, reducing the need for tuning components and helping to decrease board size.

Low current consumption means lower operating costs and less heat to dissipate. Lower currents also helps to reduce system power consumption and demand on power supplies. The result is reduced expense for manufacturers and operators.

Reducing the thermal rise in a system makes it easier to increase performance and cuts cost for the application. Because GaN technology is highly power-efficient and tolerates higher operating temperatures, GaN technology can help system designers work within tighter thermal related margins, allowing extra performance to be delivered from the RFFE.

Frequency bandwith. Increased bandwidth is being used across the entire communications industry to provide greater capacity to support the ever-growing number of users and insatiable demand for data. The high-power density of GaN and its lower gate capacitance enables greater operational bandwidth and higher speeds. Today's GaN modules and power amplifiers deliver broadband operation to support the unprecedented bandwidth requirements of 5G and other emerging applications.

Integration is now appearing in satcom RFFEs. Demand for smaller solutions for aeronautic applications and satellites is prompting suppliers to replace large multi-technology discrete RF front-ends with monolithic fully integrated solutions. GaN manufacturers are catching this wave, and are developing fully integrated solutions that combine the transmit and receive chain in a single package. This will further reduce system size, weight and time to market for manufacturers.

Key GaN Satcom Applications

GaN is making its way into many commercial and military satcom applications, including satellites, manpack, satcom-on-the-move, commercial aircraft, and very small aperture terminal (VSAT) terminals. In the space industry, GaN is replacing Si and GaAs due to advantages such as size, weight and efficiency. The smaller die size of a GaN device, compared to Si, enables performance improvements in power-switching applications. Parasitics such as output capacitance and layout inductance are reduced, resulting in lower switching insertion loss and higher-frequency operation.

Moreover, new all-electric satellites are currently under study. GaN will be a key enabler in these developments as size reduction, weight and low power consumption are important for success. Some GaN suppliers like Qorvo have space-qualified their technology, underlining the clear opportunities for GaN in this sector.

GaN is also poised to transform the lower-power VSAT satcom sector. The use of VSAT systems is expanding: they are employed for a wide variety of applications including fixed and portable wideband systems for consumer, commercial, defense and maritime communications, as well as transaction processing, data acquisition and remote monitoring. GaN is replacing and teaming up with traditionally all GaAs systems used in VSAT due to its ability to provide higher output power, which supports higher speeds and increased bandwidth allowing greater data throughput. This accelerates demanding applications, such as commercial two-way data transfers of video and other large files. GaN also outperforms Si in PA-related applications.

GaN's reliability under harsh environmental conditions is also important in this sector. VSAT devices are typically used in environments where they are subjected to harsh conditions. Advances in thermal management using unique packaging are further enhancing GaN's high reliability under these conditions. A two high-level VSAT applications using GaN PA are shown in Figure 5.

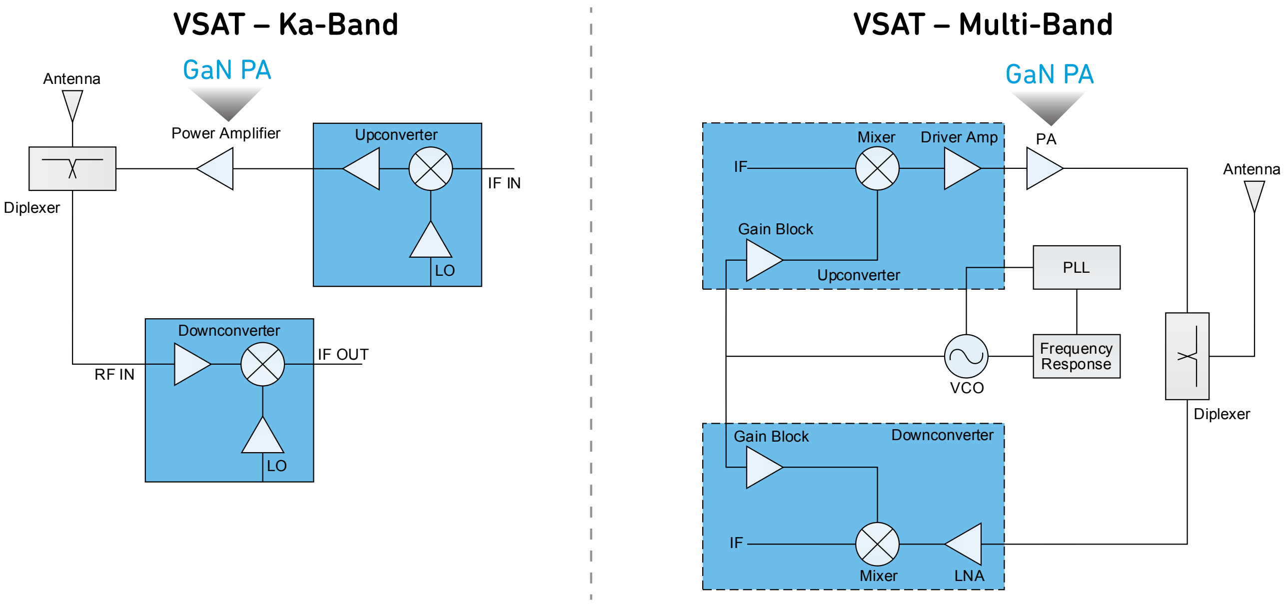


Figure 5. Satellite terminal system, VSAT block diagrams.

Conclusion

GaN is transforming the RF front end across multiple satcom application sectors. GaN is replacing incumbent technologies such as TWTAs, GaAs, and Si, because it is more reliable, more efficient, smaller, and offers higher power density and lower power consumption. Satcom manufacturers are using GaN to improve current satcom products – and to explore new developments and potential new applications.

Reaching New Levels of Linearity in Passive Mixers with GaN Technology

When microwave engineers hear the word GaN, one thought usually comes to their minds: power amplifiers. Due to its physical properties of high-breakdown voltage and high-thermal conductivity, GaN is ideally suited over its GaAs counterpart for the development of highly efficient, highly effective power amplifiers. Until now, the benefits of GaN technology beyond amplification have not been pursued.

Over the past year, Qorvo has been investigating the use of GaN for passive mixers. As a result of this work, we have discovered the high-linearity properties of GaN amplifiers do indeed translate to mixers. This realization has let us develop a new family of ultra-linear mixers operating from 1 to 20 GHz. With these mixers, microwave engineers can now approach levels of linearity that used to seem impossible.

Current State of Affairs

The passive, high-frequency mixer market of today is dominated by circuits constructed from various GaAs processes including MESFET, pHEMT and HBT. A key linearity-performance criterion used to judge different mixers is the third-order intercept point (IP3), as the higher the IP3, the better the linearity. One key factor that can influence the IP3 is the LO drive level. Many GaAs mixers achieve an IP3 that is only slightly higher than the LO drive level. To examine this phenomenon, we define the new metric, Linear Efficiency, which is the difference between IP3 and LO drive level. In Table 1 below, we summarize the linear efficiencies of some commercially available mixers.

In this table, we note that many of the mixers have a linear efficiency of 3 to 8 dB, with no clear dependence on frequency or process. Such linearity performance has been relatively unchanged over the past twenty years, when GaAs-based MMIC mixers were first introduced into the marketplace. Since then, system linearity requirements have only increased. Today, microwave engineers are routinely asking for mixers with IP3 levels above 30 dBm, if not 40 dBm. Based on the information in Table 1, GaAs-based mixers would require extraordinary amounts of LO power, greater than 30 dBm, to achieve these high-linearity levels. Such a cost is usually prohibitive in most systems, and is one reason why highly linear GaAs mixers have not been introduced to the market.

Technology	Topology	LO, RF Freq. (GHz)	IF Freq. (GHz)	Conv. Gain (dB)	LO Drive (dBm)	Input IP3 (dBm)	Linear Efficiency (dB)
pHEMT	Double Balanced	2.5-7	DC-3	-9	13	18	5
pHEMT	Double Balanced	3-10	DC-4	-9	17	24	7
pHEMT	Double Balanced	3.5-8	DC-1.6	-7	13	17	4
MESFET	Double Balanced	5.5-14	DC-6	-7.5	15	21	6
HBT	Double Balanced	6-14	DC-4.5	-6.5	13	16	3
HBT	Double Balanced	16-26	DC-9	-6.5	13	17	4
MESFET	Double Balanced	18-32	DC-8	-9	13	19	6
HBT	Double Balanced	20-32	DC-10	-7	13	18	5
pHEMT	Double Balanced	24-40	DC-17	-8	13	21	8

Table 1. Summary of some commercially available, passive, double-balanced GaAs MMIC mixers, including their linear efficiencies.

GaN to the Rescue

As developers of passive GaAs mixers, we at Qorvo have studied this problem in detail and asked what GaN can do to help solve this problem. One compelling feature of GaN processes is that they can generally support two types of mixing elements: a 2-terminal Schottky diode, and a 3-terminal FET. Each of these mixing devices offers its own unique benefits for increased linearity through physical properties and topology choice. Consider the 2-terminal GaN Schottky diode. In Figure 1, we present a typical DC I-V trace for a GaN diode as compared to a typical GaAs pHEMT diode.

In this figure, we note the pHEMT diode has a turn-on voltage of approximately 0.8 V, whereas the GaN diode has a turn-on voltage closer to 1.1 V. A higher turn-on voltage translates directly into a higher IP3 level, but at a cost of more LO drive. However, as we will see, the GaN diodes deliver a much higher LO efficiency than their GaAs pHEMT counterparts. Although not shown, a similar trend can be seen in the three terminal GaN FET devices, which have a more negative pinch-off voltage and higher breakdown voltage than their GaAs counterparts. Such performance allows for higher levels of linearity in passive mixers, again at the cost of higher LO drive.

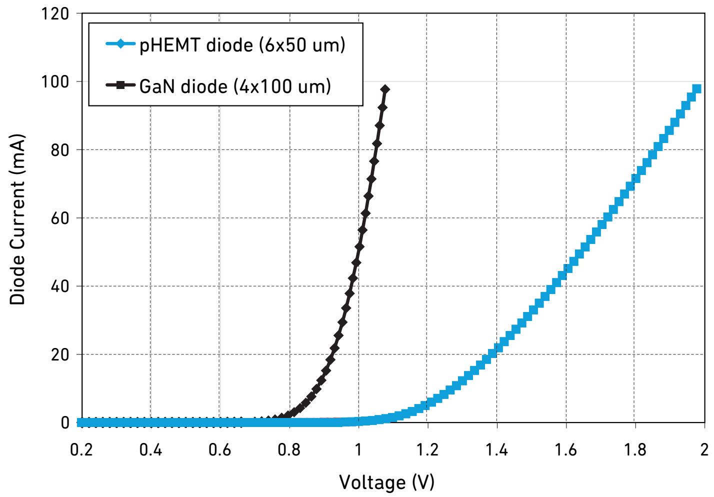


Figure 1. DCI-V comparison of a typical GaAs pHEMT diode vs. a GaN diode.

GaN Mixer Higher Linearity and Superior Efficiency

As the result of design and manufacturing process improvements and enhancements, Qorvo has been successful in developing a family of highly linear mixers with input IP3 levels well above 30 dBm and linear efficiencies of greater than 10 dB. A summary of these designs is shown below in Table 2.

In the sections below, we take a closer look at three of the most recent mixers, which are bolded in the table below.

Technology	Topology	LO, RF Freq. (GHz)	IF Freq. (GHz)	Conv. Gain (dB)	LO Drive (dBm)	Input IP3 (dBm)	Linear Efficiency (dB)
GaN	Diplexed FET	1.4-2.6	DC-0.8	-8	21	35	14
GaN	**Diplexed FET**	**2.6-3.8**	**DC-1.1**	**-8**	**22**	**37**	**15**
GaN	Double Balanced	6-11	DC-5	-6	19	30	11
GaN	**Double Balanced**	**7-13**	**DC-1.5**	**-8**	**22**	**35**	**13**
GaN	**Single Balanced**	**13-19**	**DC-2**	**-9**	**24**	**34**	**10**

Table 2. Summary of GaN mixers developed by Qorvo, showing very high linear efficiencies.

S-band GaN FET Mixer (3-Terminal Device)

The first design is a 3-terminal FET mixer as shown in Figure 2. The mixer has an operational bandwidth of 2.6 to 3.8 GHz and an IF response from DC to 1.1 GHz.

In this figure, we note the LO is applied to the gate through a filter network, whereas the RF/IF is applied to the drain via a diplexer constructed from a spiral transformer. The transformer does not introduce any balance into the circuit, but does allow for an enhanced isolation of the signals. Additional filtering elements in the signal path provide further port-to-port isolations.

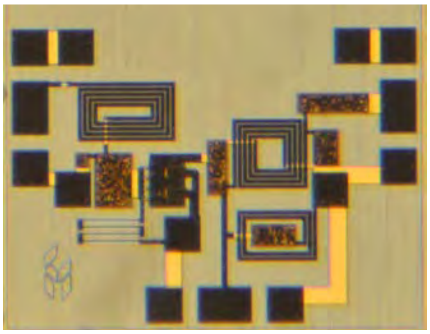


Figure 2. Die photograph of the S-band GaN FET mixer.

In Figure 3, we present the conversion gain for the S-band mixer as a function of frequency. Here, the mixer was measured as a downconverter with a fixed IF frequency of 100 MHz, USB. We note the mixer has a conversion gain of approximately -8 dB across the operating bandwidth when driven by an LO signal greater than 20 dBm.

In Figure 4, we present the input IP3 of the S-band mixer at a number of LO drive levels. We note that at a drive level of +26 dBm, the input IP3 averages around +40 dBm, which demonstrates a linear efficiency of nearly 14 dB.

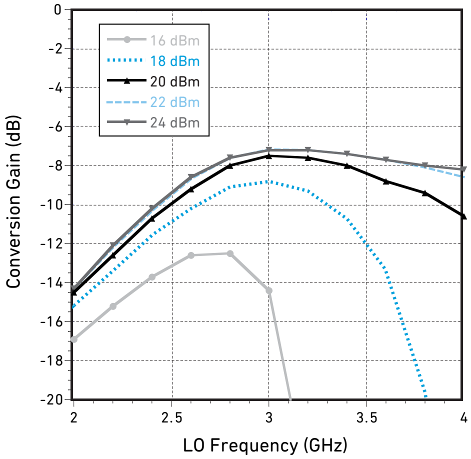


Figure 3. Conversion gain of the S-band GaN mixer. IF=100 MHz USB.

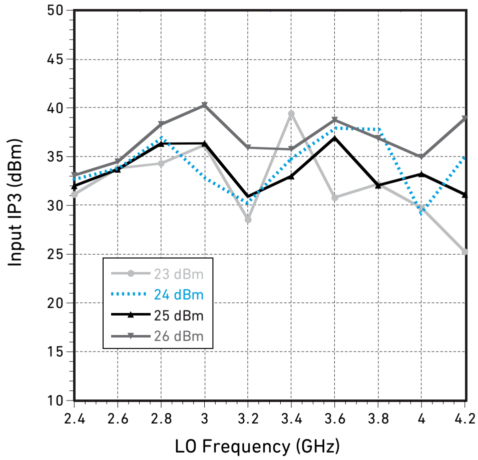


Figure 4. Input IP3 for S-band GaN mixer. IF=100 MHz USB.

X-Band GaN Diode Mixer (2-Terminal Device)

The second mixer is a double-balanced, diode-based mixer, as shown below in Figure 5. The operational bandwidth of this mixer is 7 to 14 GHz with an IF response range of DC to 1.5 GHz.

In this figure, we note the LO and RF signals are fed to the diode ring through spiral balun transformers, while the IF signal is extracted from the centertap of the RF balun. Additional matching and filter elements are also added to each of the ports to optimize the performance.

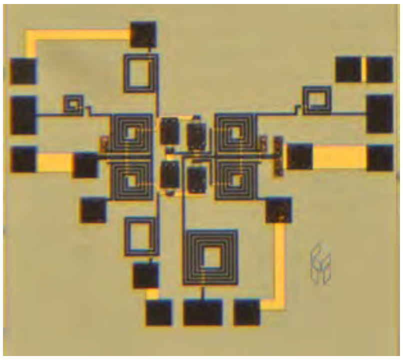


Figure 5. Die photograph of the X-band GaN diode mixer.

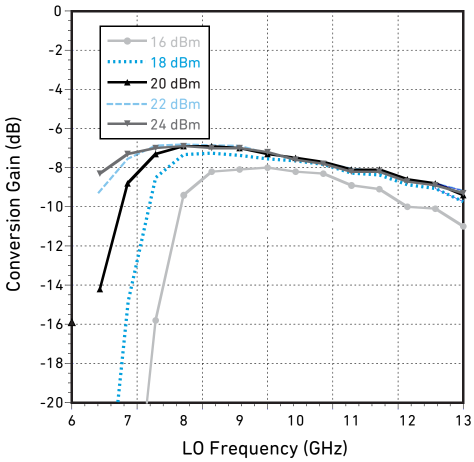


Figure 6. Conversion gain of the X-band GaN diode mixer. IF=100 MHz LSB.

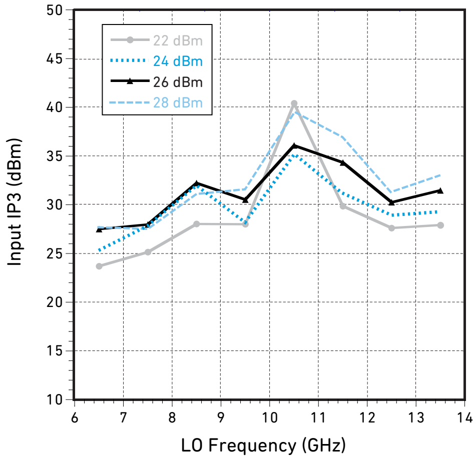


Figure 7. Input IP3 for X-band GaN diode mixer. IF=100 MHz LSB.

In Figure 6, we present the conversion gain of the X-band mixer versus frequency. Here, the mixer was configured as a downconverter with a fixed IF frequency of 100 MHz LSB. We note the mixer has a conversion gain of approximately -8 dB across the operating bandwidth when driven with an LO drive level greater than 16 dBm.

In Figure 7, we present the input IP3 of the X-band mixer at a number of LO drive levels. We note that at drive levels above +22 dBm, the input IP3 is +30 to +38 dBm, which demonstrates a linear efficiency of 8 to 12 dBm.

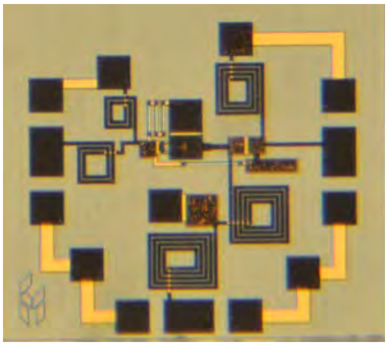


Figure 8. Die photograph of the Ku-band single-ended GaN FET mixer.

Ku-band FET Mixer (3-Terminal Device)

Our third mixer is a single-ended, cold FET mixer, as shown below in Figure 8. The operational bandwidth of this mixer is 13 to 19 GHz with an associated IF response range of DC to 2 GHz.

In this figure, we note the LO is applied to the gate of the GaN FET through a filter network, whereas the RF and IF signals are applied to, and extracted from, the drain through high pass (RF) and low pass (IF) networks. These filter networks provide the isolation in the mixer, as the topology has no inherent balance. The source of the FET is grounded.

In Figure 9, we present the conversion gain of the Ku-band mixer versus frequency. Here, the mixer was configured as a downconverter with a fixed IF frequency of 100 MHz USB. We note the mixer has a conversion gain of approximately -9 dB across the operating bandwidth for LO drive levels above +22 dBm.

In Figure 10, we present the input IP3 of the Ku-band mixer at a number of LO drive levels. We note the input IP3 increases as the LO drive level increases, up to a measured value of +35 dBm. This does not appear to be the ultimate limit. Unfortunately, we were constrained in LO drive during this measurement and were unable to generate higher power. Regardless, the Ku-band FET mixer demonstrates a linear efficiency of 8 to 10 dB.

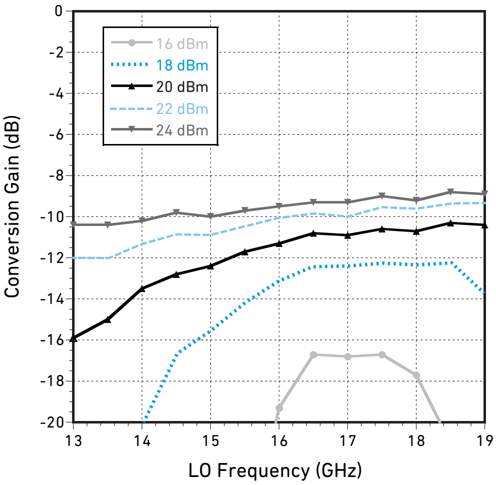


Figure 9. Conversion gain of the Ku-band GaN FET mixer. IF=100 MHz USB.

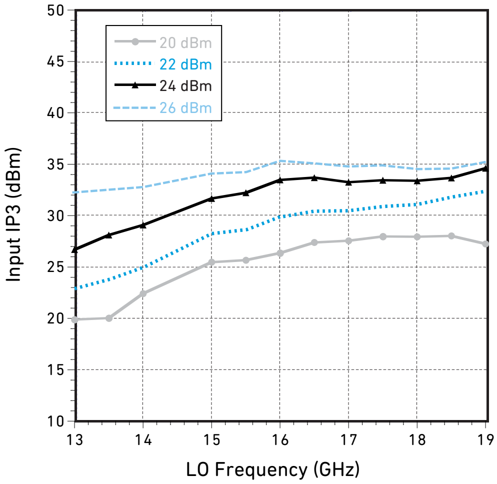


Figure 10. Input IP3 for the Ku-band GaN FET mixer. IF=100 MHz USB.

Where Do We Go From Here?

Based on the results for these three examples, it is clear that GaN mixers can deliver linear efficiencies that are much higher than their GaAs counterparts. The use of 2-terminal or 3-terminal mixing devices offers a wide range of mixer topology possibilities, thus allowing GaN mixers to be optimized for many different microwave systems. At Qorvo, we have already begun developing GaN mixers to cover more of the frequency spectrum, while still retaining very high linear efficiencies. We expect these mixers will soon find their way into radar communications and instrumentation systems near and far.

5 Key MMIC LNA Choices That Can Make or Break A Receiver Design

Introduction

Low noise amplifier (LNA) MMICs are a critical component in virtually all radar, wireless communication and instrumentation systems. There are a wide range of options and tradeoffs an engineer must consider when picking an LNA MMIC for a particular system design. The noise figure is often the feature of primary focus, as noise figure defines the sensitivity of the receiver – a critical system requirement. After noise figure, other project specific needs related to performance and size, weight, power and cost (SWaP-C) are then considered. Often these features are not heavily weighted, but they can make a big difference in advanced microwave applications.

The goal of this article is to describe additional selection criteria commonly overlooked during the initial evaluation phase of an LNA. Keeping these additional parameters in mind may help an engineer save time during the design cycle, save money during assembly, and enhance a product's competitive advantage, leading to valuable contract wins.

Pinpointing the LNA in the Microwave Signal Chain

Receiver sensitivity and signal-to-noise (SNR) are two of the most critical electrical performance considerations for modern wireless communications, radars, instrumentation, and satellite communications. Largely, the noise performance of the receiver is defined by the performance of the LNAs used in these circuits.

Advanced applications, such as electronically steered arrays (AESAs) for military applications and phased array antennas for 5G wireless communication systems, require massive numbers of T/R modules, with each receive channel requiring an LNA.

In radar systems, wideband LNAs can be used in the receiver modules, and their performance will impact the range and accuracy of the radar system.

To choose the best overall LNA designers should consider the importance of these five additional characteristics after they've identified the frequency band and noise figure combination required.

Phased array and AESA radar systems are being retrofitted into both commercial and military aircraft. They employ dozens of low noise amplifier MMICs that are relied on for receiver sensitivity and optimal SNR.

1. Input Power Survivability

Specifically in military and aerospace radar and communications applications, where electronic countermeasures (ECMs) may be used to overwhelm a receiver, a receiver must be capable of withstanding high levels of input power for varying intervals of time. Active or passive jamming can cause levels of noise and frequency bursts that couple large amounts of microwave energy into a receiver. Moreover, in many systems there is often a high-power transmitter in close proximity to the receiver, which can also lead to substantial power into the receiver front end.

Many of the latest applications, such as wideband and multi-band communications transceivers, and Low probability of intercept/low probability of detection (LPI/LPD) radar (often employing frequency hopping) use extremely wide bandwidths of spectrum for transmission and reception. These factors lead to greater noise power coupled into the receiver, and less protection from the aggressive filtering possible in narrowband receivers. If the amount of noise or interference exceeds certain limits, a receiver may be overloaded and unable to function as designed.

If the receiver is exposed to these power levels for too long, the components within a receiver may suffer accelerated aging, performance degradation, or outright destructive failure.

A common method to reduce the impact of critically high input powers to a receiver is to include a limiter or circulator on the input of a receiver chain. An unfortunate side effect of adding anything prior to the LNA in the receiver is the degradation of the overall system noise figure.

This circuit will reduce the sensitivity of the receiver, which may shorten communications range, throughput, radar range and accuracy, and cause delays in acquiring mission critical information. For example, a superior system noise figure of 1 dB can rise to 2 dB or more when a limiter is added.

Therefore, it is very important to consider an LNA's highest input power handling (or input survivability) when choosing a component. Most LNAs can handle only 10 dBm pulsed on their input, but some can now survive 20 dBm continuously and 23-25 dBm pulsed. Such protection levels can in many cases eliminate the limiter.

Ratings of a Typical LNA MMIC

Parameter	Rating
Drain Voltage, V_{dd}	5.0 V
RF Input Power	+20 dBm
Channel Temperature, Tch	150° C
Power Dissipation, Pdiss	540 mW
Thermal Resistance	120° C/W
Operating Temperature	-40 to 85° C
Storage Temperature	-55 to 150° C
ESD Sensitivity (HBM)	Class 1A

Conditions of a Typical LNA MMIC

Parameter	Min	Typ	Max	Units
Vdd	2.0	3.0	4.5	V
Idd	–	52	–	-mA

Understanding and carefully operating LNA MMICs to their specified maximum power ratings and operating conditions is critical to ensuring reliability and long life.

2. Gain Flatness and Gain Stability Over Temperature

Gain flatness with frequency is essential for wideband communications systems to achieve the required inter-symbol-interference (ISI) levels of complex digital modulations schemes. Similarly, gain flatness can impact the range performance of radar systems. Equalizers are often employed to compensate for the gain slope of typical LNAs. It is important to note many LNA suppliers use different bandwidths to characterize gain flatness, often without indicating what the gain flatness is across the entire band.

Another factor to consider about LNA performance, which is often omitted from datasheets, is the gain stability over temperature (Figure 1.). In applications such as aerospace communications and satcom, the operating temperature variations can exceed 180° F within a short time window.

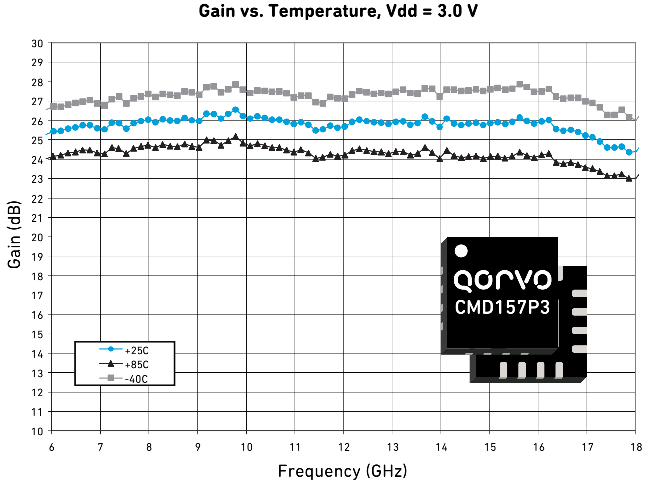


Figure 1. This 6-18 GHz LNA exemplifies how gain flatness and gain stability over temperature are possible in a single device.

Temperature changes that are significant can affect an LNA by more than just changing the noise figure of the device and system; they can vary the frequency dependent gain of the LNA. For example, large phased array antennas may have thousands of TR modules, with many of the modules exposed to a variety of temperature gradients. If the communications system relies on gain stability throughout the TR modules, and the LNA's gain variation over temperature is too large, the system may suffer a loss in performance that impacts the bottom line of the deployment.

Recognizing this, system designers should opt for LNA designs that exhibit both superior gain flatness and gain stability over temperature.

3. Supply Voltage

Properly biasing a MMIC amplifier is critical to achieving adequate device performance. Depending upon the particular LNA design, the biasing circuitry could be composed of a positive and negative biasing circuit with temperature compensation. With such a dual bias scheme, the positive and negative voltage supply must be provided in the correct sequence, or else device failure can result.

When designing at a system level for a large RF or microwave assembly, many different voltage supplies may be required. Certain design constraints may also limit the noise and stability performance of those power supplies, which may impact the practical LNA performance due to limited power supply rejection ratio (PSRR). To avoid this, additional circuitry may be used to condition the voltage supplies for a given LNA MMIC. Each of these circuits and connection points introduces a potential failure mode to the voltage supplies, and thus impacts system reliability. These supply-voltage circuits also consume valuable assembly real estate and power, contribute to the overall size/weight of the assembly, add costs, and of course, consume design and test time.

In order to reduce the infrastructure necessary to integrate a MMIC LNA into a microwave assembly, Qorvo has applied innovative circuit-design techniques. The approach has resulted in MMICs which only require a single positive voltage supply enabling a wide range of voltage input options for greater flexibility. All of the necessary circuitry to properly bias these LNAs is integrated into the MMIC itself. Ultimately, when your MMIC requires only a single positive supply voltage it reduces your bill-of-material, overall system complexity, failure modes, and overall system SWaP-C.

4. Power Consumption

For many ground-based and stationary RF communications and radar systems, the power consumption of an LNA is not a significant consideration. However, the latest AESAs, phased-array antennas, and multi-input multi-output (MIMO) RF systems may require tens, hundreds and even thousands of LNAs integrated into T/R modules. Many satcom, military, automotive and 5G wireless communication systems are also looking to these extremely complex antenna transceiver systems to solve the performance challenges innate in transmitting and receiving at high microwave and millimeter-wave frequencies.


As launching satellites costs thousands to tens of thousands of dollars per kilogram, reducing the weight of a satellite system can directly influence the cost-per-bit of high-speed satellite communication services.

In mobile platforms, including aerospace and satellite communications, power constraints are a system-wide limitation that often dictate what solutions can be used. Moreover, for these applications, the power requirements of the components directly lead to the overall size and cost of the power generation circuits, and hence, the total system SWaP-C. Some systems may have a power budget limit, and performance sacrifices must be made to meet that limit. With the importance of reliable communications in the modern battlefield, time to market may be impacted in order to design within the power budget while producing dependable communications.

An example of this concept is seen with satellite communications. The power required by a phased-array antenna must be generated by solar cells mounted on the satellite, which is one of the largest contributing factors of a satellite's weight and size.

At Qorvo, we have analyzed each of our LNA designs to ensure that their power consumption (bias current and bias voltage) is as efficient as possible. MMICs designed in this way also derive the benefit of lower power draw. They are also typically smaller, demonstrate better temperature performance, and provide improved SNR at lower power levels.

5. Value of Time Saved in the Development and Production of Your System

The SWaP-C parameters of an LNA are important in the component selection cycle. In addition, an often neglected factor is savings in time. Such time factors include design, assembly, test, qualification, support, and documentation. Choosing less efficient LNA MMICs, which might increase one or more time element, can cause project delays and cost overruns. Therefore, selecting LNAs that exhibit characteristics discussed in this brief will save time in addition to size, weight, power and cost.

Conclusion

Real-world receiver design challenges are often impacted by what is not featured in a product datasheet. Considering more than just noise figure when selecting an LNA therefore can make or break a project. Giving serious thought to survivability, gain flatness/stability, supply voltage, power consumption, and the impact of time hold the keys to success in modern radar, satcom and communications systems.

Addressing Phase Noise Challenges in Radar and Communication Systems

Introduction

Phase noise is rapidly becoming the most critical factor addressed in sophisticated radar and communication systems. This is because it is the key parameter defining target acquisition in radars and spectral integrity in communication systems. There are many papers detailing the mathematical derivation of phase noise but few mention the reasons for its importance. In this Tech Brief, we discuss the importance of phase noise, and what can be done to lessen its effects in microwave systems.

What is Phase Noise?

Phase noise is commonly used as a measure of frequency stability within an oscillator. This noise is inherently different than the general background noise of any electrical system, which is defined as kTB, where k is Boltzmann's constant, B is the bandwidth, and T is the temperature. Instead, phase noise is a secondary effect directly related to the topology and construction of the oscillator. A pictorial representation of phase noise is given below in Figure 1.

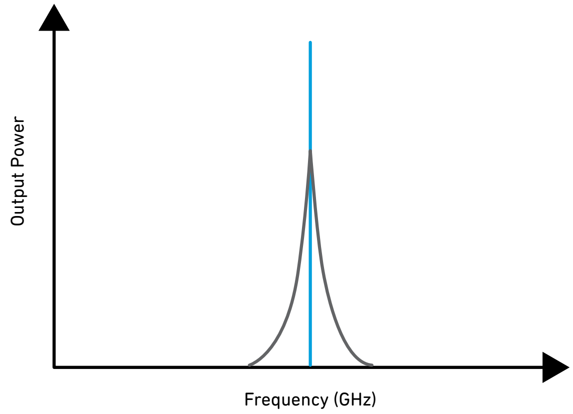


Figure 1. Pictorial representation of an ideal signal (blue) and a signal with phase noise (grey).

In this figure, we have plotted the output power of an oscillator versus frequency. The ideal oscillator is shown in blue, which only outputs power at a single, fixed frequency. The grey curve, however, is the output of an oscillator with phase noise, which shows up as power across a spectrum of frequencies very close to the desired output. These skirts, as they are called, are always present and are due to thermal noise within the active devices of the oscillator. The power level of the red skirts is dependent upon the quality of the oscillator and is measured in dBc/Hz at an offset frequency from the desired signal (typically called the carrier).

Why Do We Care About Phase Noise?

Phase noise can affect the performance of many different microwave systems. In this article, we discuss two in particular: direct down conversion receivers and radars.

Direct downconversion is a type of receiver in microwave communication systems. One benefit of direct downconversion is the simplicity of the circuit, which is essentially a single mixer driven by a local oscillator (LO) to convert the input RF signal to a baseband (very low frequency). This baseband signal is then directly applied to an analog-to-digital converter for processing. A common term for this architecture is "RF in, bits out". One problem with direct downconversion, though, is that the input RF signal can be very close in frequency to the LO, which makes the conversion process susceptible to phase noise, especially if the signal strength is low.

In radar systems, the problem is similar in nature. Radar systems operate by transmitting a pulse at one frequency, and then measuring the frequency shift of the returned pulse, as the shift is related to the velocity of the object being imaged through the Doppler effect. Objects moving very slowly will generate a return pulse very close in frequency to the transmitted pulse, and if the cross section of the object is also very small, the power level of this received signal will be also very low. Ultimately this return pulse has to be converted to baseband in order to recover the velocity information, and phase noise can obscure the data.

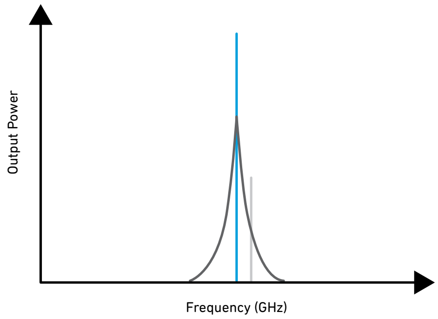


Figure 2. Pictorial representation of an ideal LO signal (blue), an LO signal with phase noise (grey), and an RF signal close in frequency (light grey) we wish to convert to baseband.

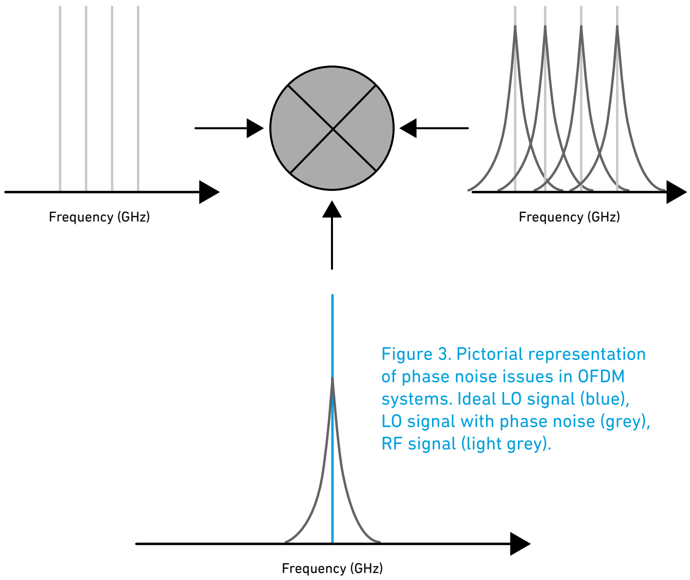


Figure 3. Pictorial representation of phase noise issues in OFDM systems. Ideal LO signal (blue), LO signal with phase noise (grey), RF signal (light grey).

A pictorial representation of the dilemma faced by direct conversion receivers and radar systems is shown in Figure 2. In this figure, we can see that if the power level of the RF signal we wish to convert falls below the phase noise spectrum of the LO signal, we will be unable to recover any baseband information, as the signal will be in the noise. Therefore, reducing the phase noise will increase our receiver sensitivity.

In Figure 3, we present a second pictorial example of how phase noise can negatively impact a conversion, this time of a multi-carrier orthogonal frequency-division multiplexed (OFDM) signal.

In this figure, we note that if the phase noise of the LO is too high, then the noise will be converted into adjacent channels of the baseband data, thereby ruining the integrity of the information.

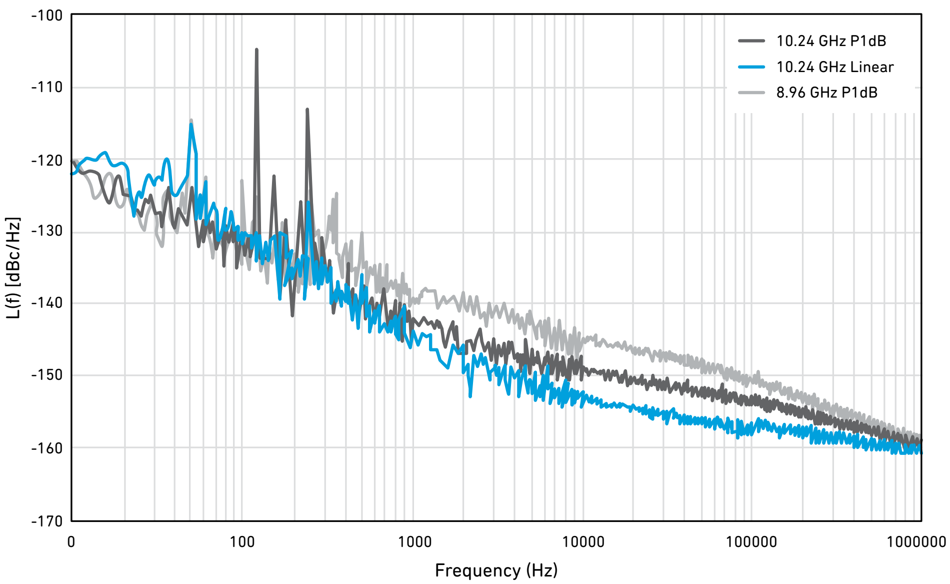


Figure 4. CMD167 LNA phase noise.

Amplifiers and Phase Noise

One obvious place to limit phase noise is in the choice of oscillator. This problem can be addressed by spending considerable time and money to design or procure a low noise oscillator. However, most oscillators do not generate enough output power, and indeed let us assume that for a particular application, the oscillator output of +5 dBm needs to be amplified to a level of +15 to +17 dBm, in order to drive the LO port of a mixer. The question then becomes, does the amplifier affect the phase noise of the LO signal?

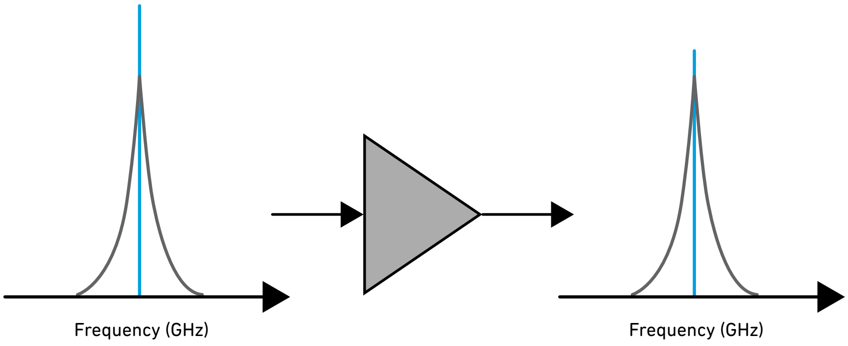


Figure 5. Pictorial representation of the degradation of phase noise due to an amplifier. The skirts of the input signal on the left are increased after passing through the amplifier, with the output spectrum on the right.

In an ideal situation, the answer would be "no," as the amplifier would simply raise desired LO signal and the skirts by the same level. However, in reality, microwave amplifiers add noise of their own to any signal, and herein lies the problem. All electronic devices exhibit a phenomenon called 1/f noise or "pink noise", which is noise power that is added to an input signal spectrum but falls off proportionally to the inverse of the offset frequency. In Figure 4, we present the phase noise of the CMD167, a low noise amplifier covering the 10 to 17 GHz range, versus offset frequency away from the desired signal. The phase noise of the incoming signal has been canceled out, so this plot represents the noise generated by the amplifier.

In Figure 4, we note the phase noise falls off linearly on the logarithmic scale with increasing frequency offset, which is characteristic of 1/f noise. If this noise level is higher than the phase noise of the input signal, then the amplifier noise would dominate the output noise spectrum. In our example, this means the low phase noise of the oscillator would be replaced by the higher phase noise of the amplifier, thereby defeating the purpose of the low phase noise oscillator. A pictorial representation of this phenomenon is shown in Figure 5.

One obvious question is, can anything be done to lower the phase noise of amplifiers? The answer lies in device physics. The 1/f noise is caused by random and thermal charge movement in the channel of an active device. The CMD167, for example, is manufactured on a GaAs pHEMT process with a gate length of 0.13 um. The FET devices on this process typically have a high 1/f corner due to their high electron mobility. GaAs bipolar devices, on the other hand, tend to have lower electron mobilities, which means a much lower 1/f noise, so they are considerably better for phase noise than their FET counterparts. Therefore, one solution to lowering the additive phase noise is to use a GaAs HBT process.

At Qorvo we have used our extensive knowledge of amplifier design techniques to create a family of new low phase noise amplifiers (LPNAs) on a GaAs HBT process operating from 6 to 40 GHz. In Table 1, we present the summary characteristics of these new amplifiers.

Product	Frequency (GHz)	Saturated Output Power (dBm)	Phase Noise (dBc/Hz@10 kHz offset)
CMD245	6-18	20-22	-165
CMD246	8-22	18-20	-165
CMD247	30-40	15	<-160

Table 1: Summary of Qorvo's new LPNAs die. SMT packaged versions are also available.

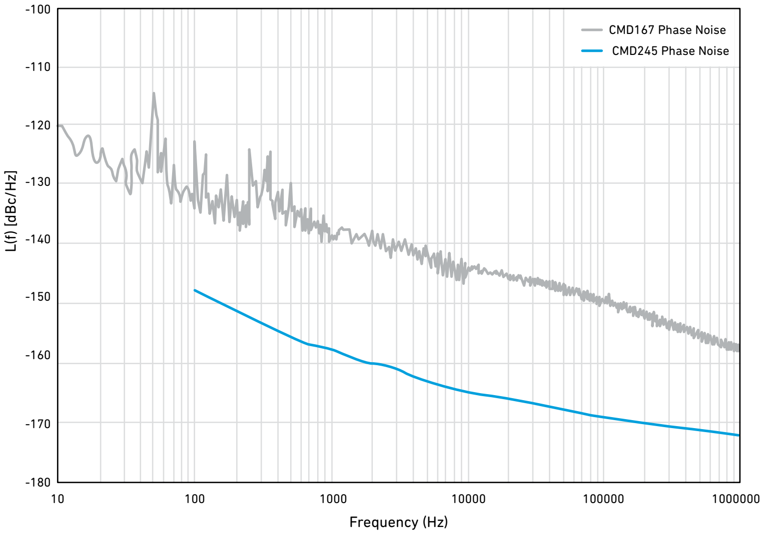


Figure 6. Phase noise of the CMD245C4 low phase noise amplifier vs CMD167 LNA.

In Figure 6, we present the phase noise versus offset frequency for the CMD245 amplifier as housed in a 4 mm QFN-style package, relative to the CMD167 HEMT LNA shown previously. We note the phase noise of the CMD245C4 is 15 to 20 dB lower than the CMD167 pHEMT LNA.

Other Components and Phase Noise

Other components besides oscillators and amplifiers can contribute to phase noise, including frequency multipliers. Many microwave systems utilize a lower frequency oscillator that is then multiplied to produce a higher frequency. One common approach for multiplication is to use a harmonically terminated amplifier to generate the required output frequency. Unfortunately, such an approach will then add the amplifier's phase noise to the multiplied signal, which will degrade the phase noise of the original oscillator.

A second approach is to use passive multiplication, which has the potential to add minimal additional phase noise to the multipliers signal (aka doublers). Qorvo, has also created a family of passive HBT style frequency multipliers which do not add to the phase noise of the input signal. In Table 2, we present a summary of these multipliers.

Product	Output Frequency (GHz)
CMD225	8-16
CMD226	14-22
CMD227	16-30

Table 2: Summary of Qorvo's passive multiplier die family. SMT packaged versions are also available.

Realizing the SWaP-C Benefits of Designing with Positive Gain Slope MMIC Amplifers

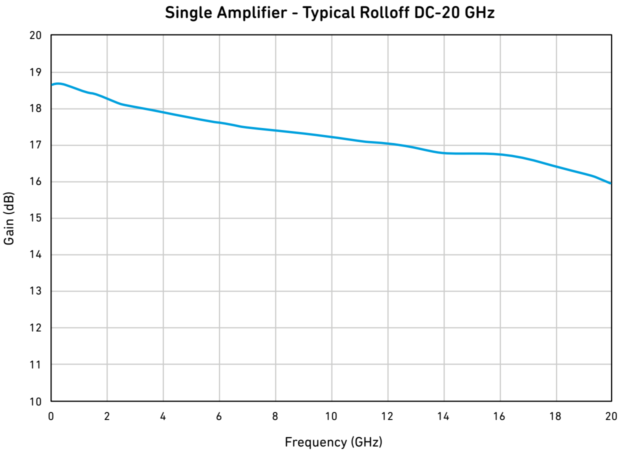


Figure 1. Frequency response of a typical distributed amplifier.

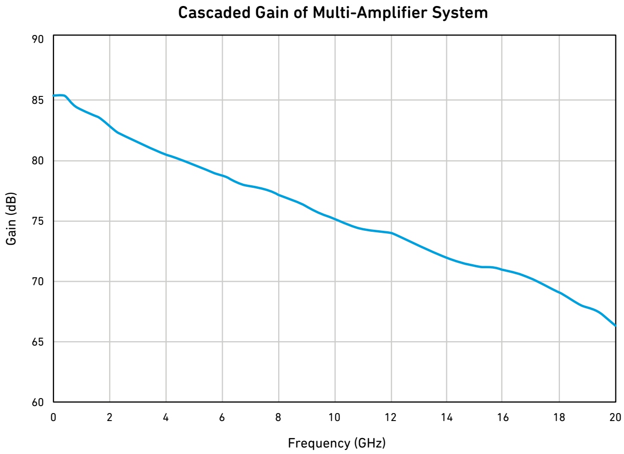


Figure 2. Frequency response of a typical wideband lineup.

Introduction

Modern wideband microwave systems often require a flat overall gain response with respect to frequency. Achieving this performance can be difficult, however, since most wideband microwave components exhibit a negative gain slope as the frequency increases. In this technical brief we discuss a means of achieving a flat system response using distributed amplifier MMICs which exhibit positive gain slope characteristics.

Figure 1 shows a typical wideband distributed amplifier response. As a standalone component its negative gain slope of approximately 3 dB is not a limiting factor. Unfortunately for designers of wideband microwave systems, a single amplifier rarely meets the overall system requirements. A wideband system will typically incorporate multiple amplifiers, passive elements and transmission lines in the signal chain. This cascade of elements with a negative gain slope versus frequency can quickly become a serious problem for the designer.

Figure 2 shows the frequency response of five typical microwave amplifiers in cascade with passive components and transmission lines, and demonstrates that while a single amplifier only had a gain delta versus frequency of 3 dB peak to peak, this cascaded lineup has a gain delta of greater than 20 dB from DC to 20 GHz.

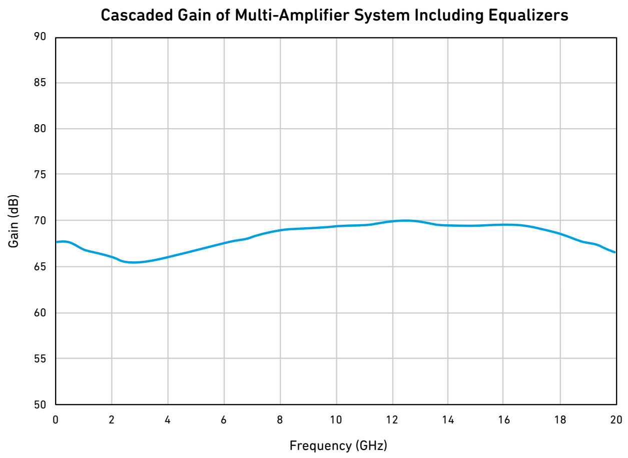


Figure 3. Frequency response of a wideband lineup including equalizers.

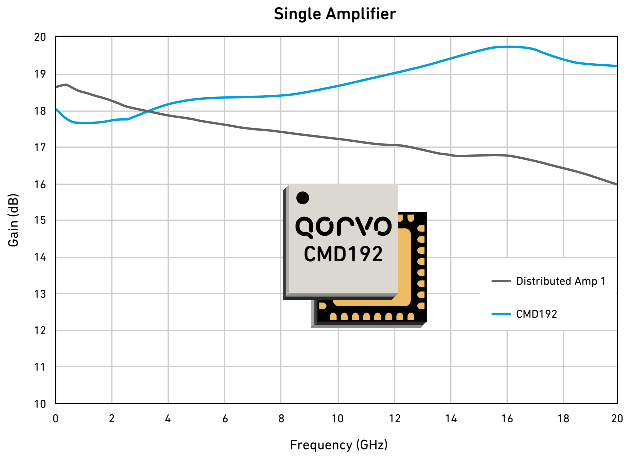


Figure 4. CMD192 vs. a typical distributed amplifier.

Traditional Methods for Flattening out the Gain

The most common solution to this problem is to add passive equalization to flatten the response across the frequency band. While this technique can solve the problem of flatness, it can also introduce three major concerns.

1. The first is the need for additional components that will increase the size and cost of the overall system. This is true whether building a discrete equalizer using resistors, inductors and capacitors, or buying an off-the-shelf die or surface-mount equalizer.

2. The second concern is the additional loss these components will add, which will undoubtedly have a negative impact on system sensitivity and noise.

3. Finally, these components require careful selection and analysis as they will also affect the power handling and linearity of the overall system.

Figure 3 shows the frequency response of the system once equalizers have been added to flatten the response. The equalizers selected were commercial off-the-shelf equalizers available in die form with less than 0.1 dB of loss at 20 GHz. Since each equalizer contributes roughly 3.5 dB of equalization over the band from DC to 20 GHz, five equalizers were used to balance the negative gain slope of the five amplifiers and the added passive components. This added up to 20 dB of loss at the low end of the frequency range while minimizing the impact at high frequency.

While this approach does provide a flat gain response versus frequency, it is not ideal due to the additional unwanted loss and other system concerns previously discussed.

Utilizing the Natural Behavior of a MMIC with Positive Gain Slope

As a more elegant solution to this problem, wideband microwave system designers should consider using positive gain slope distributed amplifier MMICs, which effectively create the necessary equalization in each stage without the need for additional components. As one example, consider the positive gain slope of Qorvo's CMD192.

In Figure 4, we first compare the gain of a single CMD192 versus the typical distributed amplifier of Figure 1. The nominal gain of each amplifier is around 18 dB, however the typical distributed amplifier shows approximately 3 dB of negative gain slope across the band while the CMD192 exhibits a positive gain slope of greater than 1.5 dB across the same band, which results in a gain differential of greater than 3 dB at 20 GHz.

The benefit of this positive gain slope becomes apparent when the CMD192 is cascaded in a system.

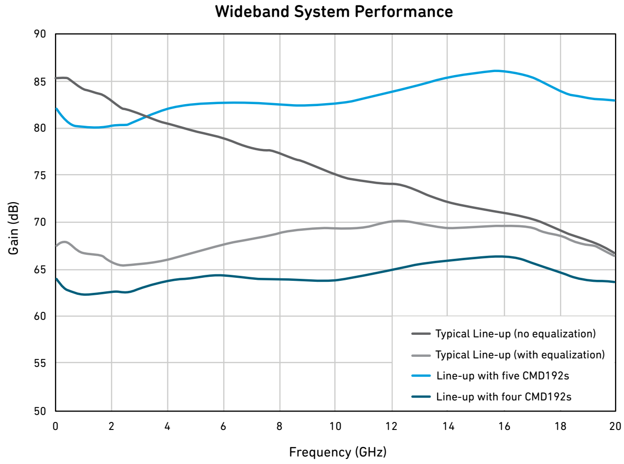


Figure 5. Overall system performance including equalization.

In Figure 5, we compare the overall gain of a few cascaded systems, including one comprised of five typical distributed amplifiers, one with five distributed amplifiers and equalization, one with five CMD192s and one with just four CMD192s. The solutions using CMD192s require no equalization, which results in approximately 15 dB more gain than the equivalent system built with the negative gain slope amplifier when using five stages. For this reason, an equivalent system can be created using only four CMD192s, further reducing the total component count. In either case, since no equalization is needed, the system complexity, cost and size of implementation are greatly reduced.

Conclusion

In this tech brief, we discussed the need for flat gain in wideband microwave systems. Typically, this performance is achieved using equalizers to cancel the effects of negative gain slope components, however, a more efficient solution utilizes positive gain slope amplifier MMICs that eliminate the need for equalization. This approach reduces the size, weight and cost of the system directly through elimination of a series of now unnecessary passive components. In addition, for systems that require multiple amplifier stages, the elimination of additional loss could reduce the total number of gain stages required. This will decrease a microwave system's power consumption in addition to reducing size, weight and cost, and help you achieve your SWaP-C goals.

See Table 1 below for a list of Qorvo's positive gain slope amplifiers as well as a switch with positive gain slope that can also be used to eliminate an equalizer.

Part Number	Function	Frequency (GHz)	Gain (dB)	P1dB (dBm)	Bias (V)	Current (mA)	Package
CMD192/C5	Distributed Amp	DC-20	19	24.5	5-8/-1	200	Die/5x5 mm
CMD240/P4	Distributed Amp	DC-22	15	19	5-8/-0.65	80	Die/4x4 mm
CMD241/P4	Distributed Amp	2-22	13.5	21	5-8/-0.65	74	Die/5x5 mm
CMD244	Distributed Amp	DC-24	18	25	5-8/-0.65	185	Die
CMD246/C4	Low Phase Noise Amp	8-22	17	13	3-5/3	48	Die/4x4 mm
CMD275P4	Low Phase Noise Amp	DC-26.5	16	18	5/3	74	4x4 mm
CMD195/C3	Switch (SPDT)	DC-20	-2	25	0/-5	0	Die/3x3 mm

Table 1. Positive gain slope components from Qorvo.

Understanding the Phenomenon of High-Power Pulse Recovery in GaN LNAs

Introduction

The GaN HEMT is well known for its use in microwave and millimeter-wave power amplifiers due to its high breakdown voltage and ability to handle high RF power. Recently, GaN technology has also been used to create low noise amplifiers (LNAs) in the microwave region, as the noise properties of GaN are similar to other semiconductor materials, most notably GaAs [1-2]. In many microwave systems, LNAs are subject to unwanted high input power levels such as jamming signals. One of the features of LNAs made from GaN is the ability to withstand these input power levels without the need for a limiter, due to the inherent robustness of the device [2]. Indeed, this is one reason GaN LNAs are supplanting their GaAs counterparts, since GaAs LNAs typically require a front-end limiter, which adds to the cost and degrades the performance of the LNA.

Despite the ability to operate without a limiter, GaN LNAs, however, are not completely immune to the effects of high input power. The problem occurs when both a high-power jamming signal and the desired signal are input to the GaN LNA, and then the jamming signal is suddenly turned off. Under this scenario, the GaN amplifier does not recover immediately, as there is some residual distortion of the desired signal before normal operation returns. This phenomenon is known as pulse recovery time and is fast becoming an important parameter with regards to LNAs in general.

Past researchers have studied pulse recovery times in GaN LNAs, although this work has been limited in scope. One study presented recovery times of less than 30 ns in some amplifiers [3-4], but these measurements only utilized a coherent jammer, and the overall number of measurements was limited. A second investigation of pulse recovery time was performed on a GaAs LNA with a limiter [5]. The limiter not only effected the small signal performance, but it also increased the recovery time when high power was applied. Further research has been performed on the degradation of GaN HEMT noise performance after exhibiting DC and RF stress, which can cause forward gate current and damage the gate device [6]. However, this work did not explicitly address pulse recovery times in LNAs. Other papers have similarly analyzed the survivability of GaN amplifiers to high input power overdrive [7-10], but again this work offers little understanding of pulse recovery times. A summary of the relevant previous work is shown below in Table 1.

Reference	Jamming Signal	Frequency	Incident Power of Jammer	Duration of Jammer
[3]	Coherent	8 GHz	+39 dBm	250 ns -3 us
[4]	Coherent	3 GHz	+20 to +33 dBm	250 ns -3 us
[5]	Non-Coherent	12 GHz (jammer) 7 GHz (signal)	+40 dBm	10 us
This work	Non-Coherent	8.5 GHz (jammer) 7.5 GHz (signal)	+15 to +27 dBm	1, 2, 4, 6, 10, 100, 200, 400, 600 us

Table 1: Summary of previous work.

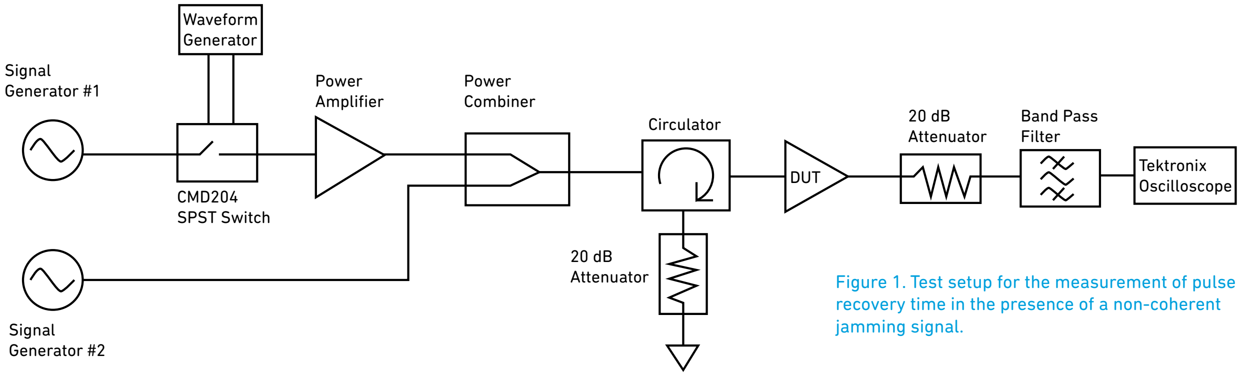


Figure 1. Test setup for the measurement of pulse recovery time in the presence of a non-coherent jamming signal.

Interferer Power (dBm)	Pulse Width (us)	Pulse Repetition (Hz)	Interferer Energy (uJ)
17	6	500	0.30
15	10	500	0.32
23	2	500	0.40
20	4	500	0.40
17	10	500	0.50
27	1	500	0.50
20	6	500	0.60
26	2	500	0.80
23	4	500	0.80
20	10	500	1.00
27	2	500	1.00
23	6	500	1.20
26	4	500	1.59
23	10	500	2.00
27	4	500	2.00
26	6	500	2.39
27	6	500	3.01
26	10	500	3.98
27	10	500	5.01

Table 2: Summary of test conditions - short pulses.

Interferer Power (dBm)	Pulse Width (us)	Pulse Repetition (Hz)	Interferer Energy (uJ)
15	100	100	3.16
17	100	100	5.01
15	200	100	6.32
20	100	100	10.00
17	200	100	10.02
15	400	100	12.65
15	600	100	18.97
23	100	100	19.95
20	200	100	20.00
17	400	100	20.05
17	100	100	30.07
26	600	100	39.81
23	200	100	39.91
20	400	100	40.00
27	100	100	50.12
20	600	100	60.00
26	200	100	79.62
23	400	100	79.81
27	200	100	100.24
23	600	100	119.72
26	400	100	159.24
27	400	100	200.47
26	600	100	238.86

Table 3: Summary of test conditions - long pulses.

Measurement Test Setup

A functional description of the test setup is shown above in Figure 1. This setup uses two signal generators, where the first (labeled as #1) provides the out-of-band interfering signal at 8.5 GHz, and the second (#2) provides the desired continuous wave (CW) in-band signal at 7.5 GHz. The interfering RF signal from #1 is pulsed using a single pole single throw (SPST) switch controlled by a square wave with a low duty cycle. We chose to pulse the signal path, as opposed to the bias circuitry of the interferer amplifier, due to the fast rise/fall time of the SPST, which is on the order of 1.8 ns. Additionally, pulsing the power supply caused high levels of ringing to appear at the output. The interfering signal was amplified by an external PA and then added to the desired signal with a passive power combiner. We utilized a circulator, terminated in a 20 dB pad and a high-power 50 Ohm load, between the combiner and the device under test (DUT) in order to prevent any high-power mismatch signal from reflecting back into the PA. The output of the DUT was then attenuated with an additional 20 dB pad, sent through a band pass filter with a pass band of 7.25 to 7.75 GHz, and then input into a digitizing oscilloscope. The filter attenuates the interfering signal to allow for an accurate measurement of the pulse recovery time.

Finally, we utilized two different oscilloscopes for the measurement. A Tektronix digital serial analyzer oscilloscope was used to measure the recovery time for the shorter pulse widths, while a Hewlett Packard Digitizing Oscilloscope was used to measure the recovery time when longer pulses were used.

The test procedure consisted of varying the pulse width and the input power of the interfering signal, while keeping the power of the desired signal constant at -10 dBm. A summary of the test conditions including pulse widths, repetition rates, and power levels of the interfering signal are presented in Table 2 (short pulses of 1 to 10 us), and Table 3 (long pulses of 100 to 600 us). In these tables we note the input power of the interfering signal was varied between 15 and 27 dBm, with the total energy delivered to the DUT being the important parameter of concern. All measurements with short pulses were performed on the Tektronix oscilloscope, whereas the long pulse measurements were performed on the Hewlett-Packard oscilloscope.

Measurement Results

In this section we present the measurements of pulse recovery time for a commercially available GaN MMIC amplifier with a 5 to 9 GHz bandwidth [11]. The amplifier was assembled into a metal housing, with 2.4 mm connectors used to interface with the test equipment. Three separate units were tested, with the results being consistent among all units. Therefore, we present the results for one unit in the interest of brevity.

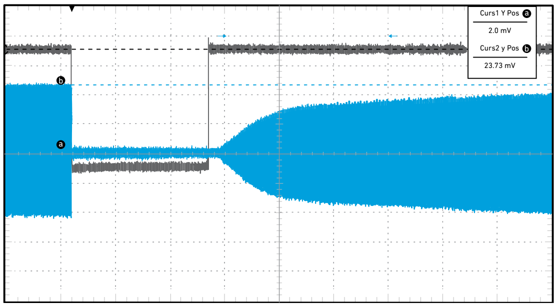


Figure 2. Typical oscilloscope trace for the measurement of pulse recovery time. Desired signal in blue, interferer pulse in magenta.

We begin, in Figure 2, by presenting an example of a pulse recovery time measurement. In this figure, the interferer pulse is shown in magenta, whereas the desired signal is shown in red. We can see that desired signal is heavily distorted when the interferer is activated, and then recovers once the interferer is disengaged. The recovery is measured as the rise time from 10% to 90% of the signal level.

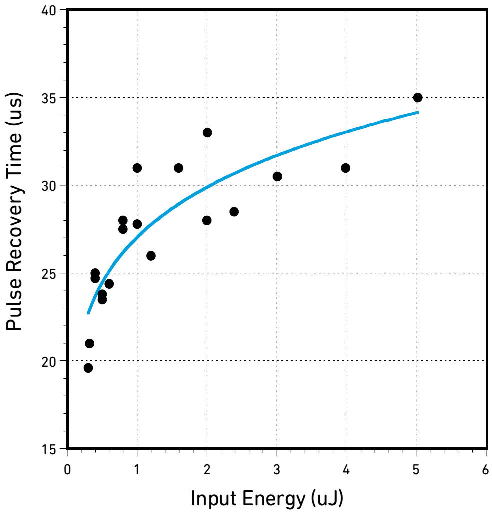


Figure 3. Recovery time versus input energy for short pulses (<= 10 us).

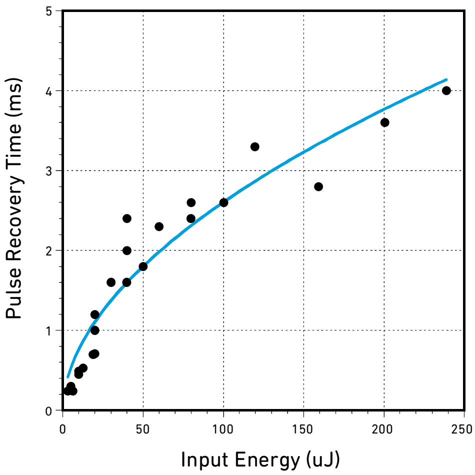


Figure 4. Recovery time versus input energy for long pulses (>= 100 us).

In Figure 3, we present the pulse recovery times versus input energy under short pulse conditions. We note the recovery time appears to increase monotonically with increasing input energy, though the relationship appears to be nonlinear. Indeed, we curve fit the data to a radical function. The form of this function is given below in Eq. 1, where y is the pulse recovery time, x is the input energy, C is a constant, and m is the radical order.

$$y = C \times \sqrt[m]{x} \quad (1)$$

One feature of this equation is that it predicts a recovery time of 0 us when the incident energy is 0 uJ. For the short pulses results as shown in Figure 2, C = 27, and m = 0.145.

In Figure 4, we next present the pulse recovery times versus input energy under long pulse conditions. We note the recovery time increases monotonically with increasing energy, and follows the same trend as the short pulses, with the same governing trend as described by Eq. (1). However, the fitted results give different constants in Eq. (1) for the long pulses, with C = 0.224 and m = 0.53.

In considering the results for short pulses versus long pulses, we did notice that the recovery time was not solely dependent on the incident energy. Indeed, there were two sets of short pulse and long pulse measurements with the same incident energy, but much different recovery times. These results are presented below in Table 4.

In this table, we note that the longer pulses with lower power had a much longer recovery time than the shorter pulses with higher power, even though they had near identical incident energy. Therefore, it appears that pulse recovery time, while being dependent on incident energy, is also dependent on the incident action (energy times duration, uJ-us) of the interfering signal. This is an interesting phenomenon we will explore in future work.

Conclusion

In this article we presented a methodology for measuring the pulse recovery times of GaN low noise amplifiers in the presence of high power, out-of-band jamming signals. Pulse recovery time is becoming an important metric for assessing system performance. In our examination of a commercially available 5 to 9 GHz GaN LNA, we considered jamming signals that operated under short pulse (< 10 us) and long pulse (> 100 us) conditions. We found that in each case, the recovery time was mathematically related to the input energy through a radical relationship. However, the pulse recovery time also appears to be a function of the input action (uJ-us), as short and long pulses with the same incident energy had recovery times that were different by an order of magnitude. In the future, we will explore this phenomenon through more measurements of GaN low noise amplifiers.

Interferer Power (dBm)	Pulse Width (us)	Interferer Energy (uJ)	Interferer Action (uJ-us)	Recovery Time (us)
15	100	3.162	316.228	240
26	10	3.981	39.811	31
17	100	5.012	501.187	300
27	100	5.012	50.119	35

Table 4. Summary of short and long pulse recovery measurements with equal incident energies.

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Simplify Amplifier Biasing Using Positive Bias pHEMT MMICs

In modern telecommunication systems, the range of parameters an engineer must consider while optimizing a design is staggering. Not only must electrical specifications be achieved, but physical and thermal constraints must be given special attention and significant design effort, especially since they are often at odds with overall system requirements.

As a result, any design technique that aids in reducing system complexity – without reducing performance – can eliminate costs, failure modes, and waste in the design cycle. The use of enhancement mode pseudomorphic high-electron-mobility-transistors (E-pHEMT) in MMICs is one such promising technique, for it may directly address a well-known design challenge. Specifically, that of sequencing in amplifier biasing, as demonstrated in Figure 1.

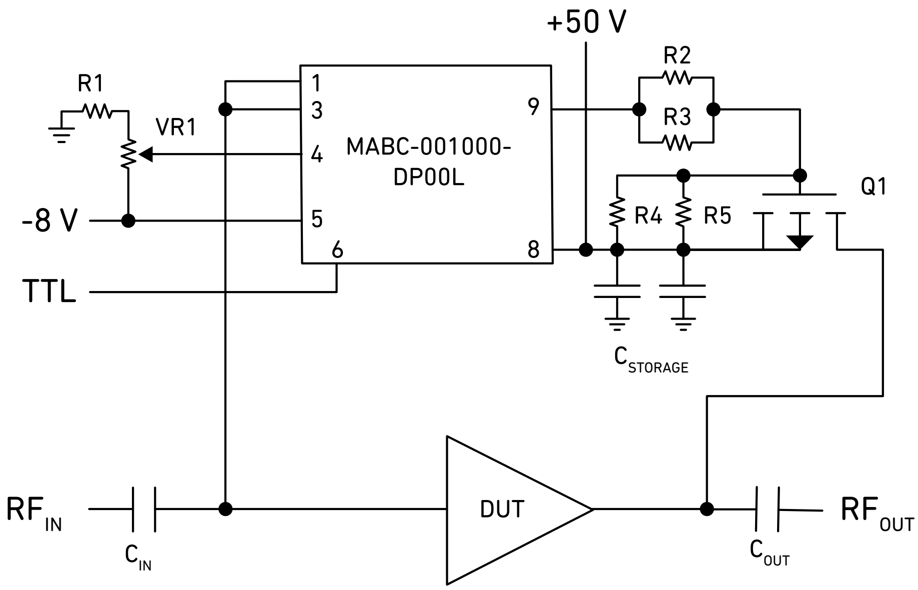

Figure 1. Amplifier bias controllers/sequencer modules are generally complex and expensive circuitry to purchase or develop in-house.

Traditionally, amplifiers constructed from depletion mode pHEMTs (D-pHEMT) and HEMT require sequencing circuits to ensure the bias voltages are energizing the transistors in the proper order. A typical bias timing scheme is shown below in Figure 2.

Failure to bias such an amplifier in the correct manner, often results in transistor damage. For instance, the device channel is normally conductive and will sink large levels of current, if not first biased into pinch-off mode. A depletion mode device also requires that RF power to be applied after the device has entered the appropriate portion of the sequence, and must also be powered down with the exact reverse sequence. Any deviations in the timing sequence could induce damage to the amplifier.

The timing problem becomes even more complicated when the microwave system contains multiple D-pHEMT amplifiers, such as phased array radars. Not only does the sequencer have to control hundreds, if not thousands, of amplifiers in parallel, but any delays or offsets in the bias scheme could have profound impact on the overall sensitivity of the radar.

A solution to this vexing problem can be found in physics. Many MMIC processes now offer both D-pHEMT transistors and enhancement E-pHEMT devices on the same substrate. In terms of RF performance, these transistors are comparable – and in some instances, E-pHEMT amplifiers can outperform their D-pHEMT cousins in maximum gain, noise figure, and linearity.

Typical Bias Timing Scheme using D-pHEMT or HEMT Amplifiers

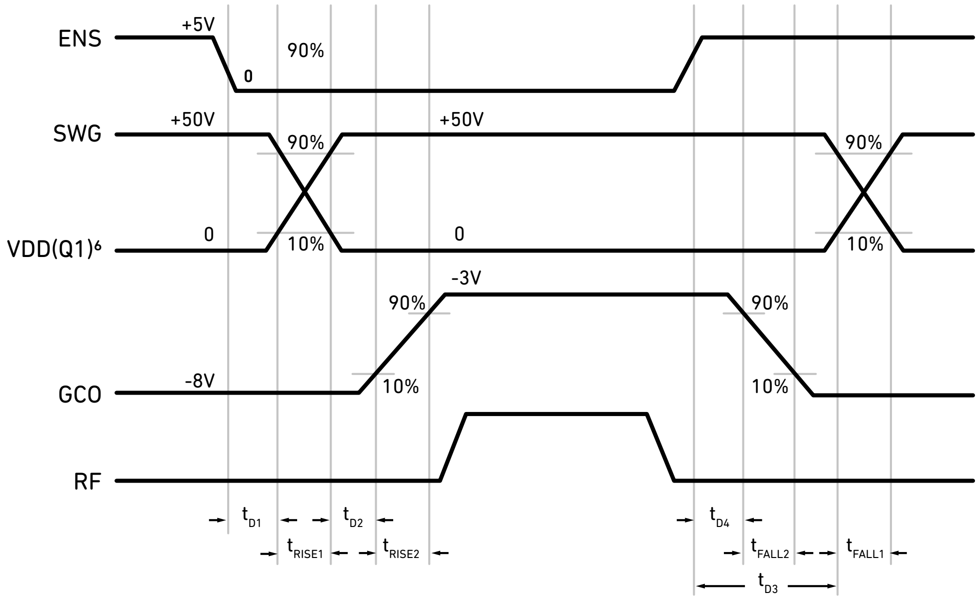

Figure 2. As dual bias amplifiers require a precise sequence before energizing each port, timing diagrams and activation sequences require digital controllers to prevent damage from improper sequencing.

Unlike depletion mode devices, E-pHEMT transistors are normally non-conductive, and will ultimately reduce current when both the drain and the gate are biased (regardless of sequence). As a result, the sequencer circuit can be eliminated altogether.

The savings generated by removing the sequencer can be enormous. For example, positive bias techniques enable a reduced bill of materials, a simplification of the circuitry, and a reduction the number of extraneous noise sources. These eliminations allows the designer to focus on more important aspects of the system, such as optimizing the RF signal chain at large.

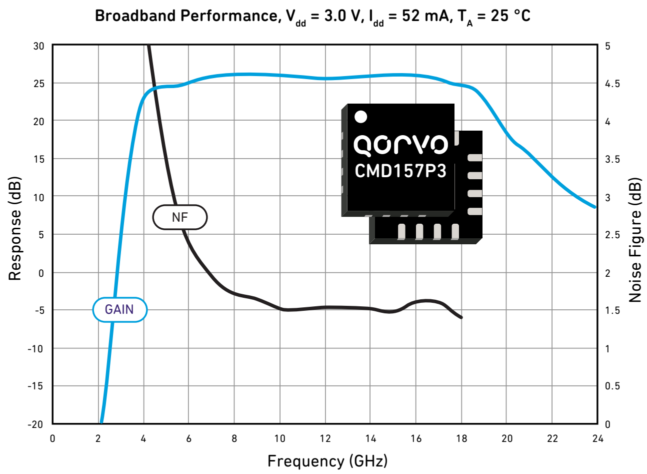

Figure 3. A perfect combination of high gain and high linearity can be achieved with E-pHEMT amplifiers.

The use of E-pHEMT devices by designers of power amplifiers (PAs) and low noise amplifiers (LNAs) is in its infancy, as such devices have only recently been made available from a number of semiconductor manufacturers. However, Qorvo has been a pioneer in this area and currently offers dozens of standard, off-the-shelf PA and LNA components built with E-pHEMT technology.

In many of these designs, the high gain and high linearity E-pHEMT amplifiers have matched – and even exceeded similar depletion mode designs.

Conclusion

Not only can E-pHEMT amplifiers reduce cost and complexity, they can also improve performance.

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