CDIAN001: Stability Considerations When Designing Microwave Power Amplifiers

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Today's device landscape appears to be a gold mine for newly minted microwave engineers. In many ways, GaN has become the great equalizer, allowing even inexperienced engineers to build amplifiers with a level of performance that would have been unachievable 10 years ago.

The high gain, high power density, and reasonably high terminal impedances that are available with "high band gap" devices such as SiC and GaN HEMTs have provided some powerful tools to the PA designer's "bag of tricks", but there are still a few landmines that need to be considered early in the design process to insure a robust solution when designing a high power broadband amplifier. One that designers often step on is device stability (or more appropriately, instability). This application note will discuss the fundamental cause of this instability, and practical methods to mitigate the problem in a real world environment.

WHAT CAUSES INSTABILITY?

If a two-port network is NOT unilateral (S12>0), then a change in the load impedance presented to the network will cause a change in the input impedance of the network. Similarly, a change in the source impedance presented to the network will cause a change in the output impedance of the network. The bilateral nature of all active devices that are used in microwave designs are the major potential cause of instability.

Conditional Stability – A two-port network is conditionally stable if the real part of Zin (input impedance) and Zout (output impedance) is greater than zero for <u>some</u> positive real Zs (source impedance) and Zl (load impedance) at a specific frequency.

Unconditional Stability – A two-port network is unconditionally stable if the real part of Zin and Zout is greater than zero for <u>all</u> positive real Zs and Zl at a specific frequency.

For a single stage two-port amplifier, potential instability is caused when terminating impedance on one side of the device can be transformed through the device so that the device presents negative impedance at the other port. This can often happen on either or both ports. Typically, the situation does not happen across all frequencies, and often it does not happen in the design band of interest.

STABILITY FACTORS

There are two factors that can be used to determine the stability of a two-port device. The Stern Stability Factor, K, is derived from the s-parameters of a linear two-port device. K>1 is a necessary (but not sufficient) condition for stability at a given frequency for a two-port device. K is defined as follows:

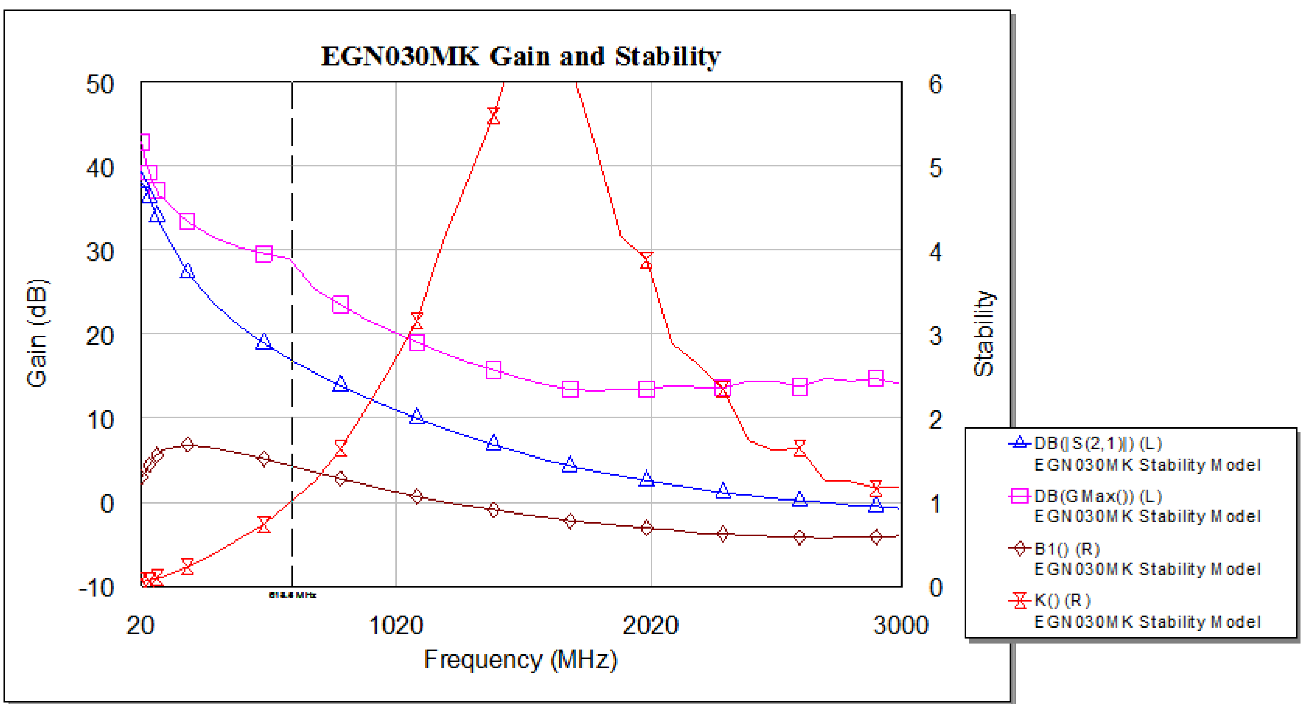
$$K=\frac{1+|s11s22-s21s12|^2-|s11|^2-|s22|^2}{2|s11||s22|}$$

When K>1 then there are a unique set of source and load impedances, Γms and Γml, which will allow for a conjugate match at the input and output of the device. It is of note that these impedances are not S11 conjugate ($S11^*$), and S22 conjugate ($S22^*$) unless S12=0. Interestingly, if K=1, then |Γms| and |Γml| also are exactly 1, lying exactly on the outer circumference of the Smith Chart. As K becomes less than 1, then the Γms and Γml become negative impedances with their magnitudes lying beyond the circumference of the Smith Chart and the two-port device is no longer unconditionally stable.

In addition to K>1,the other stability factor, B1, must be >0 to insure unconditional stability for a two-port. B1 is defined as follows:

$$B1=1+|s11|^2-|s22|^2-|s11s22-s21s12|^2$$

B1 and K are both available as calculated data for two-port devices in Microwave Office (MWO). Figure 1 shows a rectangular plot from MWO of several sets of data for the EGN030MK GaN HEMT biased at Vds=50V, Id=200mA.



<u>Figure 1 – EGN030MK Gain and Stability Factors</u>

Stability factors K (red curve) and B1 (brown curve) are both plotted, with their values on the right hand axis. Note that B1>0 for all data from 20 MHz to 3 GHz. Therefore, the device will be unconditionally stable in this range of frequencies if K>1. Note the vertical dashed line at about 610 MHz. Beyond this frequency K>1 and the EGN030MK is unconditionally stable for the given bias conditions. However, below 610 MHz, the device is potentially unstable.

Also note the magenta curve, which is a plot of GMax. GMax is maximum gain for K>1 or Maximum stable gain, MSG, for K<1. There is a change in the curve at about 610 MHz when K=1. S21 transducer gain (blue curve) is also plotted for reference.

STABILITY CIRCLES

Input Stability Circle - The input stability circle is a contour in the source plane that indicates source termination values that make the output reflection coefficient have a unity magnitude (Γout=1). An output reflection coefficient less than unity indicates a stable device, while an output reflection coefficient greater than unity indicates a negative output impedance, and a potentially unstable device. When using MWO, the display of the stability circle indicates the unstable region using a circle drawn with a dashed line in the unstable region. If the dashed circle is inside the solid circle, then the outside of the circle indicates the stable region, whereas if the dashed circle is outside the solid circle, the inside of the circle represents the stable region.

Since the s-parameters used to compute stability were likely measured in a 50 ohm system, if the source impedance in the input plane is 50 ohms (Γsource=0), then by definition, the output reflection coefficient will be S22. Thus, the region of the input stability circle that includes 50 ohms is also the stable region.

Output Stability Circle - The output stability circle is a contour in the load plane that indicates load termination values that make the input reflection coefficient have a unity magnitude (Γin=1). An input reflection coefficient less than unity indicates a stable device, while an input reflection coefficient greater than unity indicates a negative output impedance, and a potentially unstable device. The display of the stability circle indicates the unstable region using a circle drawn with a dashed line in the unstable region. If the dashed circle is inside the solid circle, then the outside of the circle indicates the stable region, whereas if the dashed circle is outside the solid circle, the inside of the circle represents the stable region.

If the load impedance in the output plane is 50 ohms (Γload=0), then by definition, the input reflection coefficient will be S11. Thus, the region of the output stability circle that includes 50 ohms is also the stable region.

Stability Circles for EGN030MK GaN HEMT – Figure 2 is a plot of the input stability circles for the Sumitomo EGN030MK GaN HEMT, Vds=50V, Id=200mA. This data plot is represented as SCIRC1 in Microwave Office and is represented by the series of blue circles on the Smith Chart.

Note that each of the circles has a smaller circle that is drawn with a dashed line inside of it. The area with the dashed line is the area of potential instability. If a source impedance is selected that falls inside of the circle at the that frequency, then it will create a negative output impedance for this device.

Also note that stability circle “p14”, which is from s-parameter data at 600 MHz, just barely touches the outer circumference of the Smith Chart. Beyond this frequency, all of the remaining stability circles fall outside of the circumference of the Smith Chart and the device will be unconditionally stable.

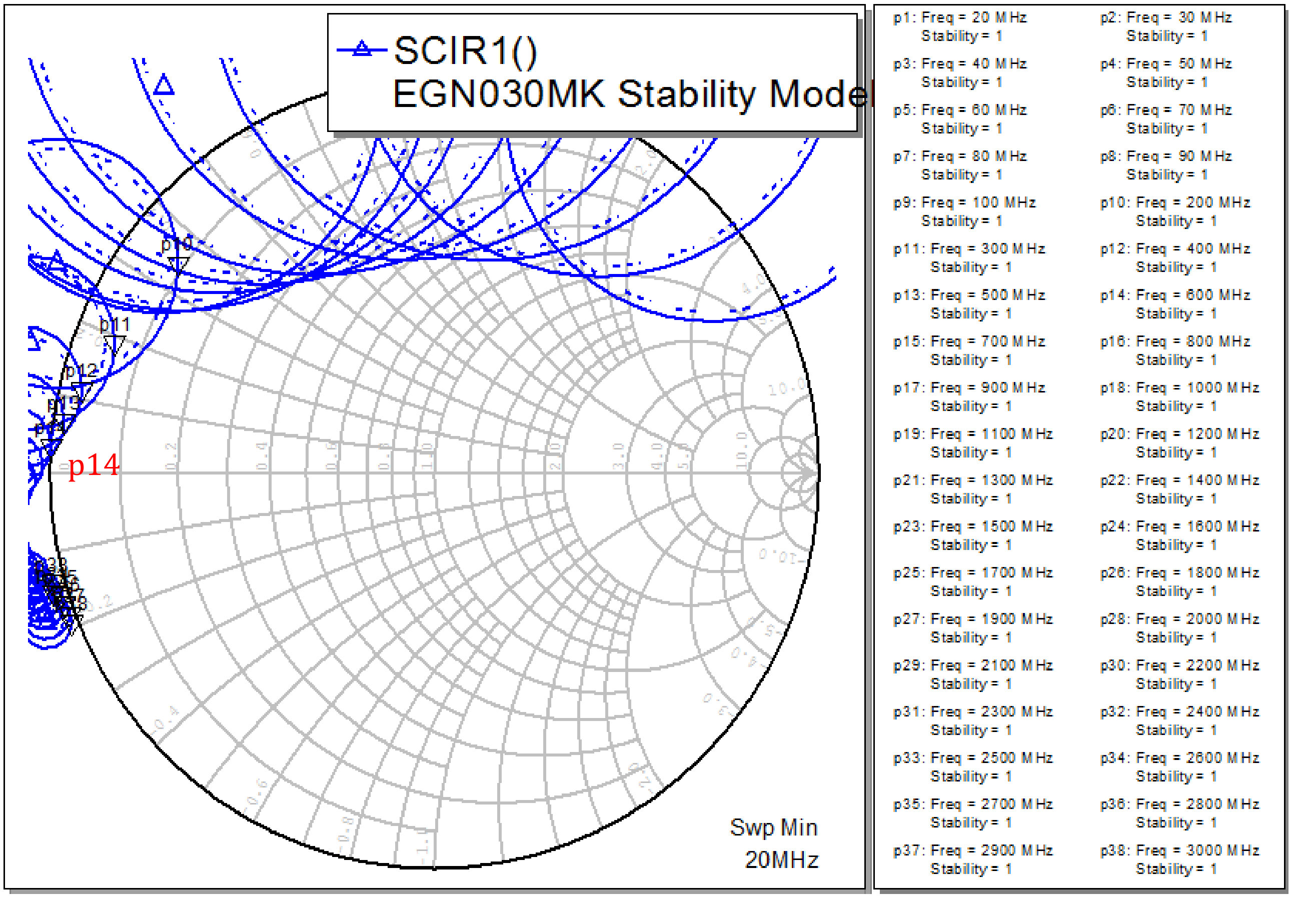


Figure 2 – EGN030MK Input Stability Circles

Similar curves can be constructed for the output stability of the EGN030MK. Figure 3a is a plot of the output stability curves for the same EGN030MK. Note that in the case of the output curves, at low frequency, a significant percentage of real impedance are no longer available for stable matching. Also note that just above 600 MHz, the curves no longer fall inside of the outer circle of the standard Smith Chart. In fact, the 600 MHz plot, P6, just touches the outer circumference of the Smith Chart.

Figure 3b is a plot of the same data but on a compressed Smith Chart. The standard Smith Chart (outer circumference is for Γ=1) is shown on this expand chart as the area inside of the dark black circle. The rest of the chart, outside of this circle, represents a region of negative impedance.

All of these plots are available as outputs from Microwave Office.

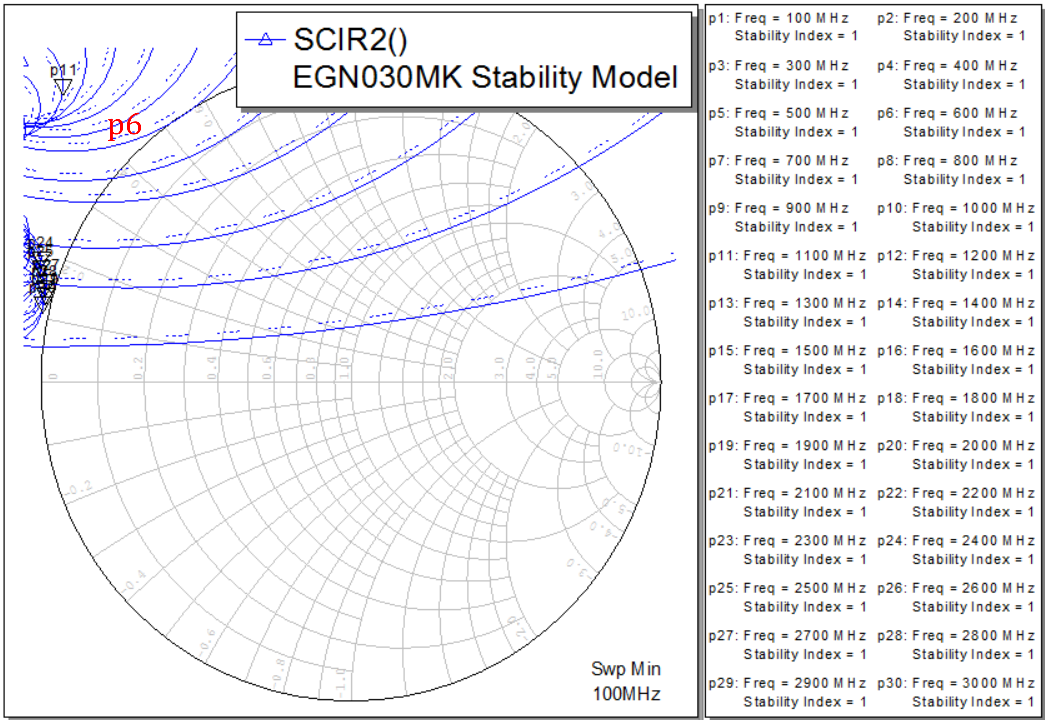


Figure 3a – EGN030MK Output Stability Circle, Standard Smith Chart

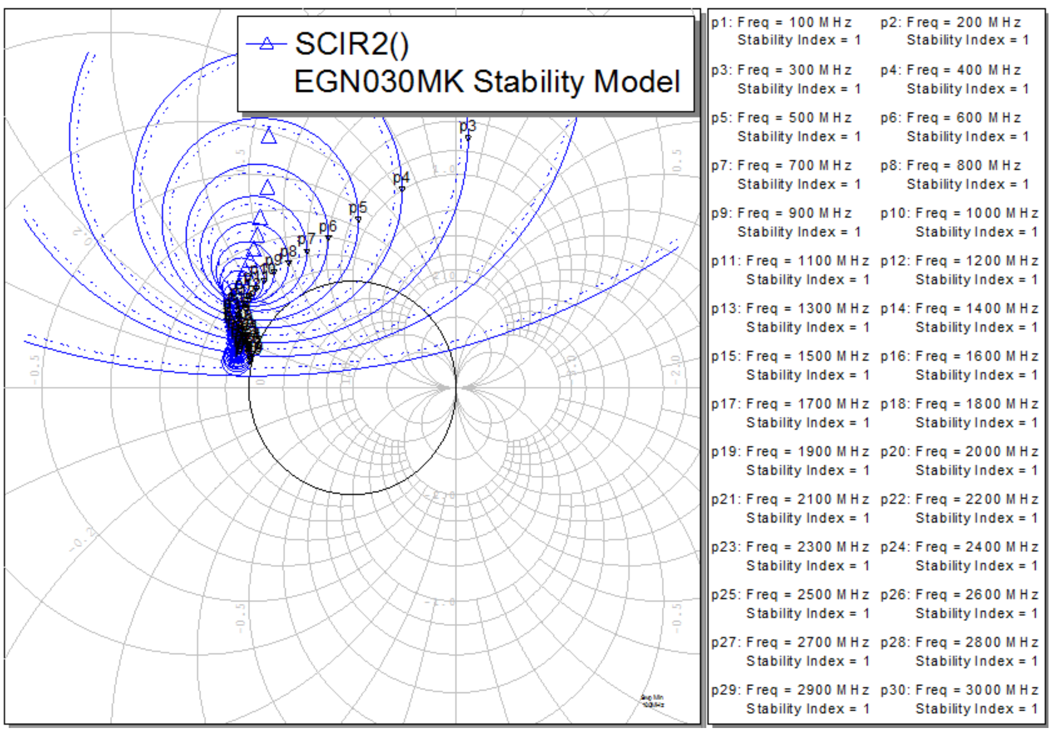


Figure 3b – EGN030MK Output Stability Circle, Compressed Smith Chart

DESIGN CONSIDERATIONS FOR STABILITY

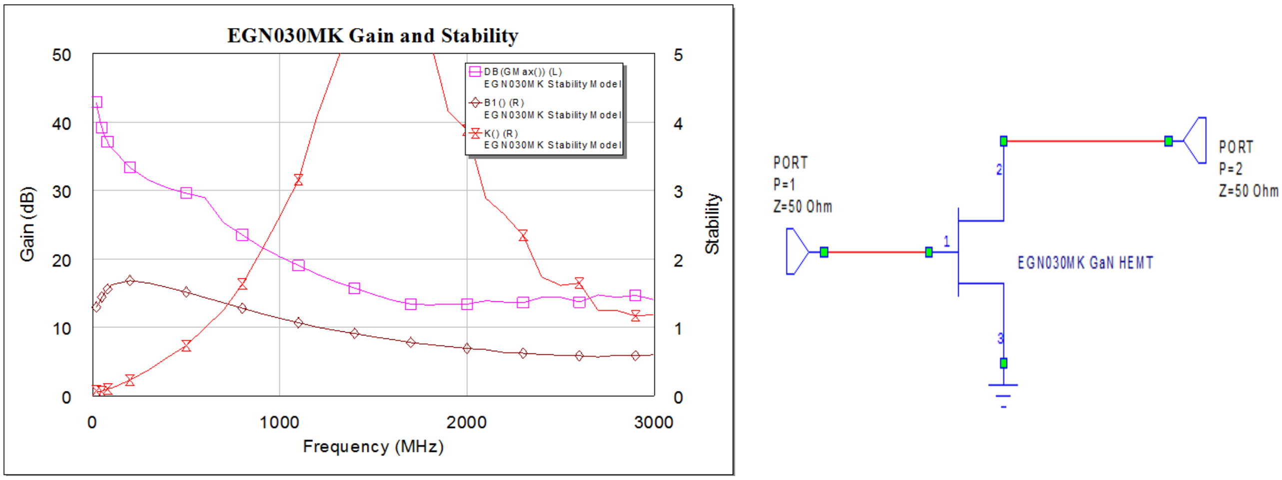
If one were designing a 1-2 GHz amplifier with the EGN030MK, the usual thinking would be that there would be no stability issues as K>1 and B1>0 across the entire operating band. If the designer was not careful, one would quickly end up with a device that might oscillate in the best case, or oscillate to self-destruction in the worst case because nothing was done to manage the source or load impedance presented to the device below 600 MHz. In particular, almost half of the Smith Chart is unavailable for load impedances at 100 MHz. So what options are available to a designer to insure out of band stability? More commonly, what options are available in situations when K<1 across the desired operating range?

Coupling Capacitor Considerations – The intent of a coupling capacitor is to provide maximum power transfer from one stage to the next while providing a DC block to the bias voltages on either side of the capacitor. One design approach is to use the largest capacitor that is available while not having the parasitic parallel resonance of the capacitor effectively increases its insertion loss to unacceptable levels at the high end of the desired frequency of operation. Models are available for most capacitors in an SMT, substrate sensitive environment from many of the device manufacturers, as well as Modelithics, that will allow one to simulate the undesirable effects of a parallel resonance. For extremely broadband amplifiers, one might use the newly released 100nF PPI0201BB104 chip capacitor from Passives Plus. This capacitor will operate, resonance free, from about 16 kHz (lower 3db point) to 50 GHz.

Unfortunately, coupling capacitor resonance is only half of the designer's problem, and the most obvious one at that. The more significant issue is in trying to establish the "out of band" impedance presented to the device by the previous or following stages, or combining networks when multiple devices are combined for more output power. Typically, this problem is more serious at frequencies below the band of interest. What does the "out of band" source impedance of an 8-way splitter look like when 7 other active devices are also connected to it? What is the impedance presented to the source of an FET when it is the last device with 3 other single ended devices preceding it? A resulting simulation of "out of band" stability becomes exceedingly complicated, especially at low frequencies.

One approach is to make the coupling capacitor only as large as necessary to accommodate the lowest desired frequency of operation. As its reactance increases with decreasing frequency, it effectively isolates the active device from other parts of the circuitry so that the only thing that determines "low "out of band" terminating impedance is the circuitry between the terminals of the active device and the coupling capacitor. If care is taken with bias networks, lossy networks, feedback networks, and other methods to insure stability of the single device, one can effectively take the rest of the combining or surrounding networks out of the equation for determining stability. This makes this entire process more manageable as the designer can tackle stability one device at a time rather than wrestling with the interactions from all of the surrounding circuitry. In the design of a 1-2 GHz amplifier, a 10pF capacitor would only have 0.1dB insertion loss at 1 GHz, with almost 6dB loss at 100 MHz and 24dB loss at 10 MHz. So where coupling capacitors are concerned, BIGGER IS NOT BETTER!

Lossy Series Input Networks – A great method for improving stability is to add a lossy element into the active RF path. Again, the Sumitomo EGN030MK has been used for example, and has been evaluated for operation from 10 MHz to 3000 MHz. Figure 4 plots are based on linear s-parameter data for this device. A schematic of the simulated circuit is show along side of the plot. The 3 traces are maximum available gain (magenta) with values on the left axis, Stern Stability Factor, K (red) shown on the right axis, and B1 stability factor (brown) shown on the right axis. Note that for the gain plot, the magenta curve is not the S21 gain of the amplifier, but rather the maximum available gain that could be achieved if the device was properly matched. In this case, B1>0 but K<1 over the frequency below 600 MHz. This is a version of the same data that was previously shown in the stability circle plots. If one were designing an amplifier from 1-2 GHz with this device, there are potential instability issues below 600 MHz that will need to be wrestled with.



<u>Figure 4 – Simulated Maximum Available Gain and Stability Factors for the EGN030MK (Vds=50V, Id=250mA)</u>

In Figure 5 below, a series input resistor of 120 ohms has been added. This will reduce the gain across the band while simultaneously improving stability. A loss in gain is undesirable; therefore to minimize the impact in our design band of interest (1-2 GHz), the series resistor has been bypassed with a small 20pF capacitor. This will keep the resistor as a predominant part of our circuit below our band of interest while minimizing its impact across the 1-2 GHz range, since the 20 pF capacitor is a lower impedance to the high frequency RF signal. This fix is effective, as the plot shows, K>1 for all frequencies from 10 MHz to 3GHz.

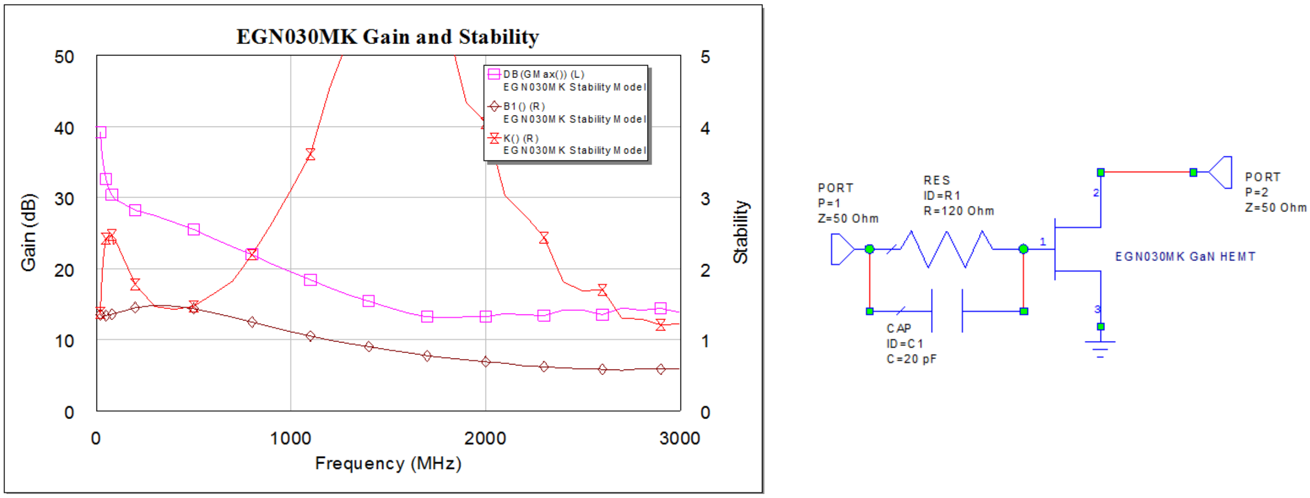


Figure 5 – Simulated GMAX, K, B1 for EGN030MK with Input Series Lossy Network

Lossy Shunt Input Network - Another option on the input side can be realized using a shunt lossy network. Figure 6 shows the same series of plots for the EGN030MK, but this time using a shunt inductor/resistor network from the gate to ground. The additional advantage of this approach is that it can also be used to inject the gate bias to the device by providing a bypass capacitor, C1, allowing the gate bias to be injected at the C1/R1 node. Again in this example, K>1 for all conditions below 3000 MHz and B1>0, so unconditional stability is assured. Inductor, L1, might also effectively be replaced by a $\lambda/_4$ transmission line section to help isolate the 3ohm resistor from the circuit in the band of interest.

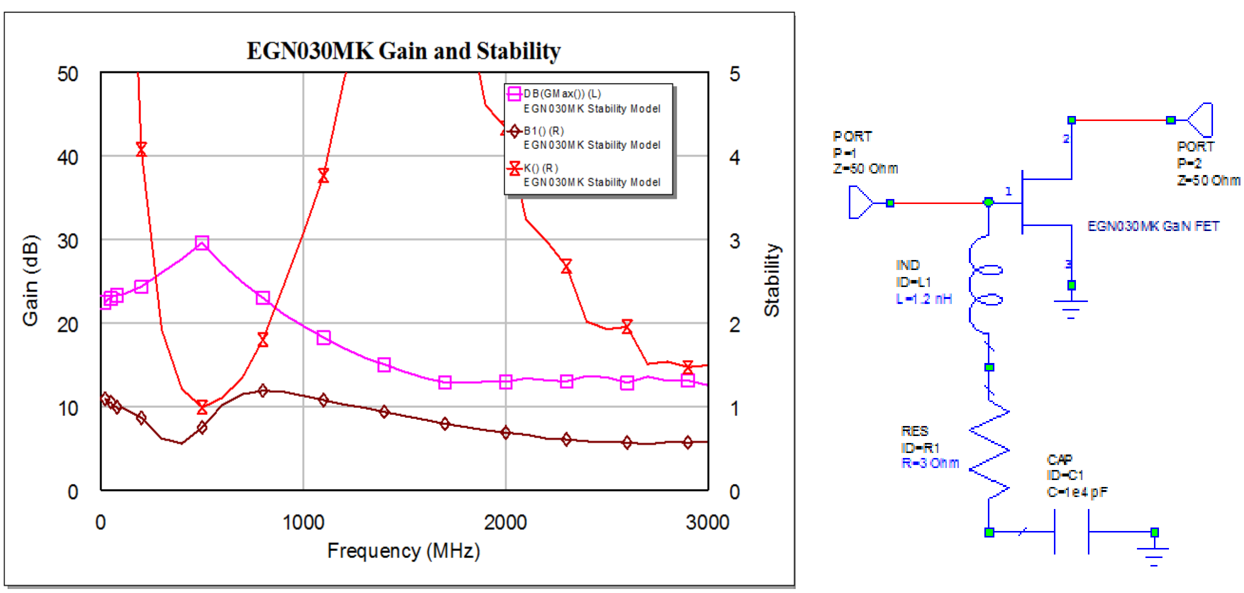


Figure 6 – Simulated GMAX, K, B1 for EGN030MK with Input Shunt Lossy Network

Drain-Source Series RL Feedback – A lossy feedback network from the drain to the gate of the device can also provide an improvement in stability. In Figure 7, a 390 ohm resistor is used to provide feedback. It is AC coupled with a 0.01uF capacitor so that there is no interaction between gate and drain bias. In addition, L1 has been added to pull some of the feedback out of the circuit in our band of interest (1-2 GHz), while still providing stability enhancement in the previously unstable regions.

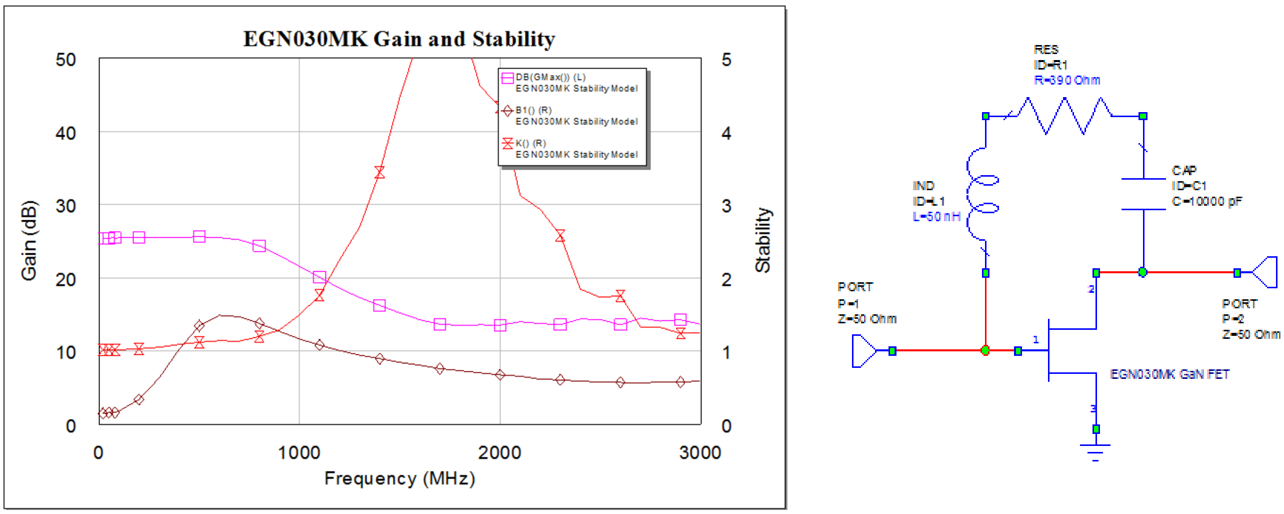


Figure 7 – Simulated GMAX, K, B1 for EGN030MK with Lossy Feedback Network

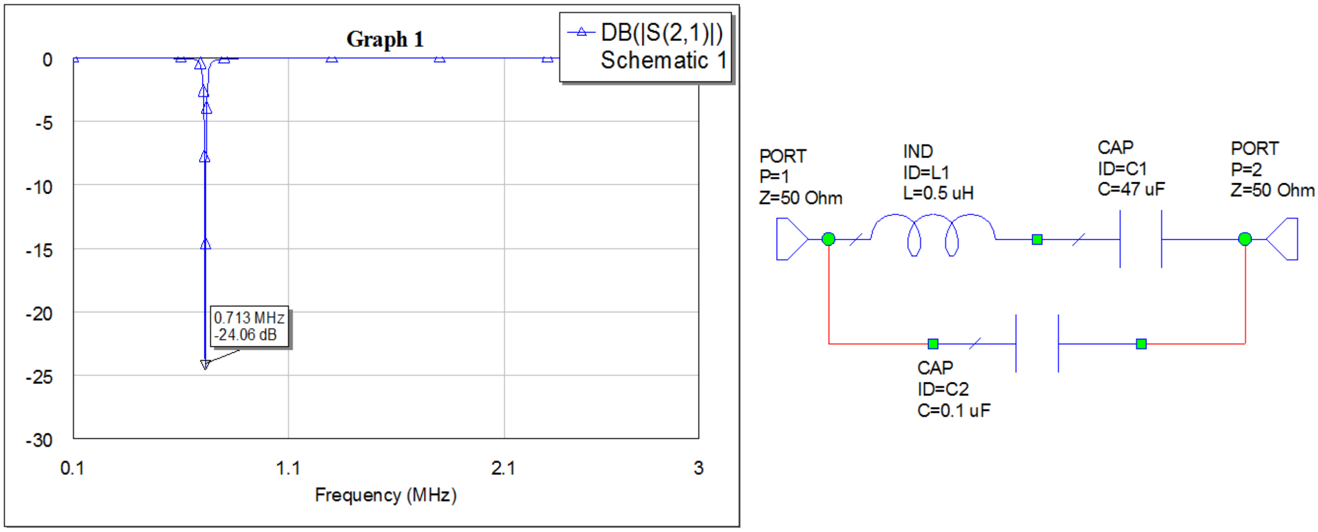
Bias Network Considerations – The design goal of properly designed bias networks is to allow the designer to introduce the necessary DC voltages to the device without degrading the simulated RF performance. To the extent that most designers are focused on "in-band' performance, the bias networks have the potential to create dangerous termination impedances to the active two-port device in the "out of band" regions. The most dangerous case is when the designer really does not know what termination impedances are presented to the active device in an integrated assembly.

In the example shown in Figure 7 above, gate bias can easily be introduced without compromising stability. In fact, this approach has enhanced stability. Unfortunately, for high power applications, adding a lossy network on the output side of the device is just not practical, but if stability can be insured on the input side, then drain bias insertion can be accomplished via a shunt inductor or $\lambda/_4$ transmission line section. One significant issue when bypassing the "cold" end of the drain bias inductor or $\lambda/_4$ line is in the use if parallel capacitors to increase the frequency range over which adequate bypassing is achieved. Just hanging a 0.1 uF capacitor in parallel with a 47uF tantalum or electrolytic can be a dangerous practice if not properly analyzed.

In most power amplifier designs, design considerations make it difficult to add lossy networks (at least inband) on the output (drain) side of the amplifier. The biggest issue will be that the drain injection network will by its nature tend to present a very low impedance to the drain of the device at low frequencies. Fortunately, as we have seen above, unconditional stability can usually be achieved with circuit techniques on the input (gate) side of the device.

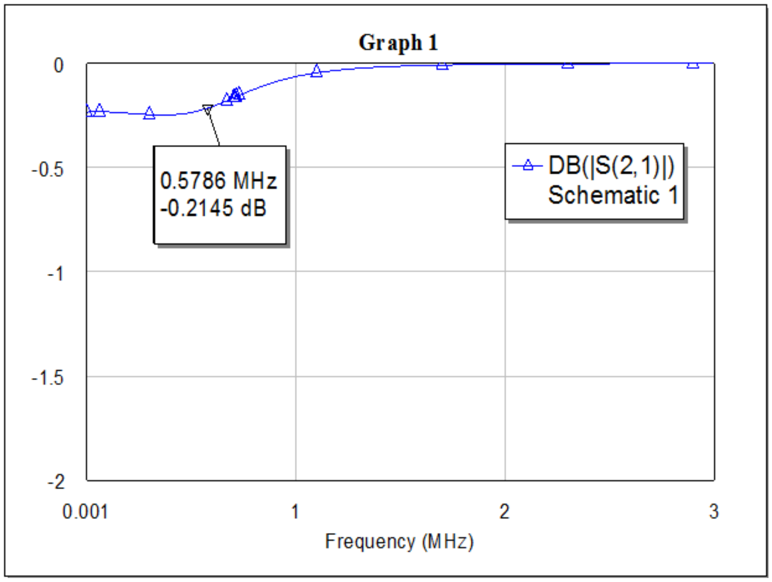
Pitfalls of Paralleled Bypass Capacitors - One of the problems with large capacitors is that they exhibit a fair amount of equivalent series inductance.

Figure 8 shows the effect of paralleling a 47uf capacitor (with 0.5uH of equivalent series inductance) with a 0.1uF chip capacitor. There is a resulting 24dB resonance at 713 kHz! When using this capacitor combination for either a broadband series coupling cap or a bypass capacitor, this becomes a huge problem and a significant source for potential instability in bypass applications.



<u>Figure 8 – Loss of Parallel Capacitor Network with 47uF Tantalum || 0.1uF Cap</u>

With the addition of a small resistor in series with the tantalum capacitor, the 24dB resonance can effectively be eliminated while still achieving good bypass from several GHz down to about 15 Hz. In Figure 9, a 2.7Ω resistor has been added in series with the large capacitor, effectively eliminating the huge resonant loss, minimizing it to about 0.2dB. These effects are real and need to be considering when using multiple capacitors in parallel. They may not be as severe as this example, but should be evaluated before they become another source for unexpected instability.



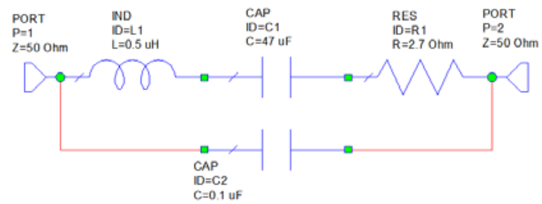


Figure 9 - Loss of Parallel Capacitor Network with 47uF Tantalum || 0.1uF Cap with Additional 2.7Ω "De-Qing" Resistor.

OTHER STABILITY CONSIDERATIONS

Multiple Stages – K>1 and B1>0 for a multistage network does not guarantee stability unless each of the individual networks is unconditionally stable and there is no feedback mechanism between the stages. A simplistic case would be if two marginally stable amplifiers existed in a box each terminated internally on one port, but with no interconnection between them and only two connectors on the box. Even though S12=0 for the boxed assembly since there is no connection between the two amplifiers, an external terminating impedance on either connector could cause either amplifier to oscillate.

In a more typical cascade of two stages, stability needs to be evaluated for each stage, and if feedback is used, care should be taken to insure that there are no frequencies where the feedback could cause a regenerative unstable condition. In addition, there is always the potential for unanticipated feedback through the bias networks. Adequate bypassing and choking of bias leads will mitigate this potential condition.

Change in Bias Conditions – The above analysis has been for one set of bias conditions for the device in question. During the design process, especially for a power amplifier stage that could be operated in a class AB condition where drain current would change with drive, the designer should evaluate the design across the spectrum of drain bias conditions that are anticipated under normal operation. While this sounds onerous, one will quickly determine whether or not the mitigation strategies for stability are adequate as drain bias changes.

S-Parameters from Non-Linear Simulation – In most cases, device manufacturers only provide s-parameter data across a limited range of frequencies where they expect their customers to use the device. While this may not be an issue at the higher frequencies because the device will have minimal gain, it creates a potential analysis dilemma at lower frequencies where stability needs to be evaluated. Fortunately, most manufacturers of power devices are currently providing good non-linear models for their devices. One can use the models with a non-linear simulation tool such as Microwave Office to compute low frequency s-parameter data that can then be used for stability analysis. If a non-linear model is not available, the designer can make his own s-parameter measurements, although care needs to be taken with fixturing and proper reference plane placement to allow for accurate s-parameter extraction. Another option would be to use a modeling service such as Modelithics, to develop a non-linear model for your particular device.

Combining of Multiple Stages – Typical design of high power amplifiers often requires combining of multiple output and driver devices. The typical process is to determine the largest device that can be matched across the desired operating bandwidth, and then decide how many devices can be practically combined. Combining two is usually straightforward, with quadrature or in-phase combining used most often due to the in-band isolation that the combining networks can provide. Unfortunately, at low frequencies, one leg of the combiner provides virtually no isolation to the adjacent device at out of band frequencies on the low side. The potential impedances that can be presented through the combining networks to the adjacent device can be extreme. As the combining networks grows in size and complexity, the only viable analysis approach is to use some of the techniques outlined above to insure the unconditional stability of the device across all operating frequencies. A good starting point is always to use the smallest coupling capacitors that are practical.

SUMMARY

Too often, designers bypass stability analysis, anticipating that they can either “dodge the bullet” or “fudge” their way to stability once a circuit is fabricated. The techniques outlined above have proven successful over the years in insuring stability in high power amplifier designs from HF through Ku Band. Time invested at the beginning of the design process will almost always yield the desired stability results. With the design tools that are available to engineers today, this process should be reasonably painless, and is always worth the time spent.

About the Author

Paul White has been involved in the microwave industry since 1969. After graduating from Lehigh University with a BSEE, he worked as a microwave engineer at Westinghouse Electric in Baltimore. In 1974 he moved into a sales and marketing position for Hewlett Packard until 1980 when he became a partner in Applied Engineering Consultants, Inc., a manufacturers representative firm.

In 1987, Mr. White founded Chesapeake Microwave Technologies, Inc., focused on the design and manufacture of microwave power amplifiers. With the sale of Chesapeake Microwave to Andrew Corporation in 1999, Mr. White became Andrew’s Chief Technologist for active products.

In 2002, Mr. White co-founded and was the Co-Managing Director of Integrated Defense Systems, Inc. (IDSI). IDSI was one of the early adopters of GaN device technology, bringing GaN based high power amplifiers into the battlefield environment in 2005. Over the next 3 years, IDSI consumed several hundred thousand GaN power devices for use in broadband high power applications for military tactical countermeasure and defense communications systems. After the sale of IDSI in 2008 to General Dynamics, Mr. White retired from full time employment.

Today he remains involved in the microwave industry sitting on five corporate boards. Mr. White is also actively involved as a board member for several non-profit organizations. Today he resides on Longboat Key, FL where he and his wife Vicki continue to enjoy boating, flying, and spending time with their 4 children and 8 grandchildren.

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