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A CAD Based Tutorial on Microwave Amplifier Design

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Abstract:- Microwave amplifiers find applications in almost all devices working at high frequencies and thus find applications in communication systems, satellites as also radars and navigations. It is thus very important for the students of electronics and its allied branches to have an in depth knowledge about the design of such amplifiers. The design of microwave amplifiers is a vast subject, and is by experience, generally found to be a difficult topic for the student community. Though a lot of classic books are available for the subject, this tutorial paper-specially written for students, aims at providing an intuitive approach to the design of microwave amplifiers. AWR's Microwave Office® software is used to illustrate the concepts more clearly.

Keywords:- Amplifier, Microwave, Microwave Office® and Stability

I. INTRODUCTION

Microwave amplifiers find broad ranging applications from being used in mobile phones, satellite transponders, spectrum analyzers, radar system et cetera. These microwave amplifiers are also fast replacing conventional tubes in certain areas due to their compact nature and continuous research are in progress to increase their power handling capacity and efficiency.

The nature of parasitics at high frequencies result in a lot of complication in the design of microwave amplifiers and an exact design of the amplifier according to the given specification was often not possible until the advent of CAD (Computer Aided Design) Tools. Various EM CAD Tools have now made it easier to design microwave amplifiers according to exact specification. One such tool, AWR's Microwave Office®, is used in this paper to effectively demonstrate the design of microwave amplifiers.

II. AMPLIFIER DESIGN

The design of amplifiers at radio frequency (RF) is very demanding as compared to that at low frequency, primarily because of the fact that parasitic effects at RF are very much higher than at lower frequencies. These parasitic capacitors and inductors cause a feedback resulting in either oscillations or lowering of gain depending on whether the feedback is positive or negative respectively. These effects must be accounted in the design of microwave amplifiers. Thus the design of amplifiers operating in microwave frequencies demands critical consideration of the device being able to operate as expected, also called stability considerations. Other major considerations are for gain, noise and VSWR. Thus the design of a microwave (MW) amplifier involves large efforts.

The designer must first select an active device depending on system specifications, which can be either a MESFET, HBT, HEMT etc. The device is then appropriately biased and its characterizations are obtained. These device characterisations are in the form of various network parameters which help in explaining its functionality. The device stability is then analysed and it is then appropriately loaded or terminated so that the device works in accord with the specifications. The following sections deal with these steps in detail. Also, a flowchart for the same is depicted in Fig. 1, so that the reader may keep a track of the design flow.

A. Selection of microwave transistor

The designer is given some system specifications viz. gain, noise etc. for which a design needs to be developed. The same is true for a MW amplifier design. To meet these required specifications, various families of devices are available. Selection of a proper family and hence a proper device (or a transistor) such that it works as intended is very essential. For instance, if a Low Noise Amplifier (LNA) is to be designed, it would be of no use to select a device whose minimum offered figure of merit (which is a measure of noise) is higher than that desired for operation. Hence a detailed study of various datasheets must be done before selecting a device.

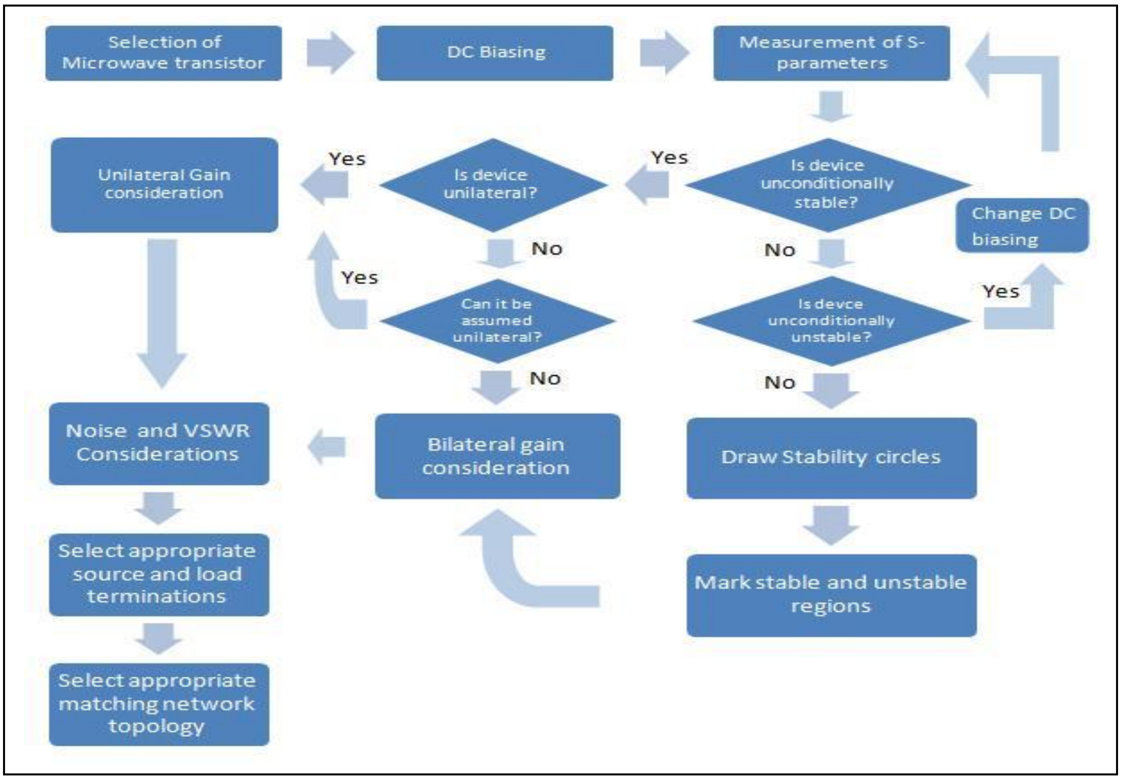


Fig.1: Flowchart of steps in amplifier design

B. DC biasing

Once the device is selected, then it must be biased so as to yield the desired results. Consider an illustration for the same- In a LNA design, the transistor is biased at that value of collector current (I_C) which gives minimum figure of merit (F). Fig. 2, for example, shows the variation of noise figure vs. collector current of a low noise transistor 2SC3357. Here, minimum noise figure occurs at an I_C of approximately 8.5 mA. For use in LNA design, the transistor will be biased at such a point yielding this value of collector current.

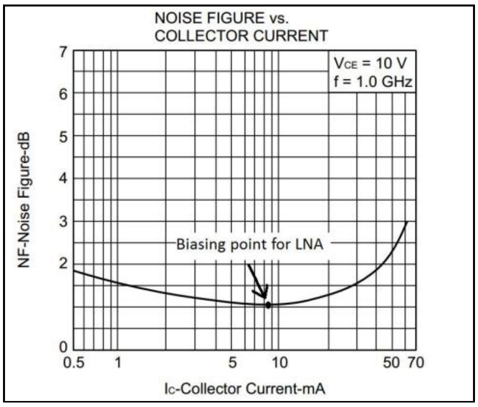


Fig.2: Graph of Noise figure vs. Collector current

C. Device Characterization

To describe the behaviour of a device, its characterization is very necessary. At low frequencies, the same can be done using the 2- port parameters viz. Z, Y, h, ABCD et cetera. These parameters are defined in terms of the voltages and currents at the ports, and hence they cannot be measured at high frequencies as the circuital laws fail at RF. Hence Scattering parameters or S- parameters are used to define a network at MW frequency as they are defined by forward and backward travelling waves and not voltages or currents. The low frequency 2- port parameters cannot be measured at RF but are inter-convertible with S- parameters.

Using either network analyser approach or vector voltmeter approach, the S-parameters can be practically measured over a wide frequency range.

D. Stability Analysis

The last thing any amplifier designer would expect to happen is that the amplifier designed begins to oscillate instead of amplifying signals, as intended. One of the reasons due to which such oscillations can occur is that a positive feedback arrives back to the input.

To illustrate this point, consider the circuit shown in the Fig. 3. Say the load impedance generates some noise voltage, in the form of thermal noise, which is fed to the active device. This device will amplify at least one of the frequencies from the noise spectrum and feed it back to the load. Some part of this amplified signal will be dissipated by the load, while rest of the signal (proportional to Γ_S) will be reflected back to the amplifier. If this process continues and is favoured such that the loop gain i.e. $\Gamma_{in}\Gamma_S$ becomes greater than unity, then the device may become unstable and break into oscillations. The same explanation can be extended to a 2- port active device also.

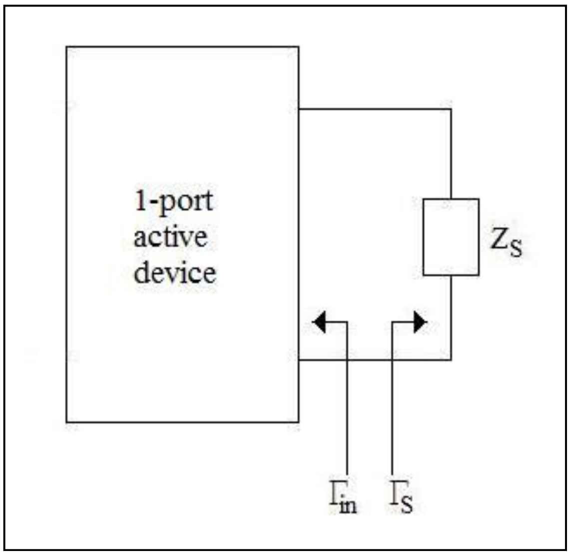
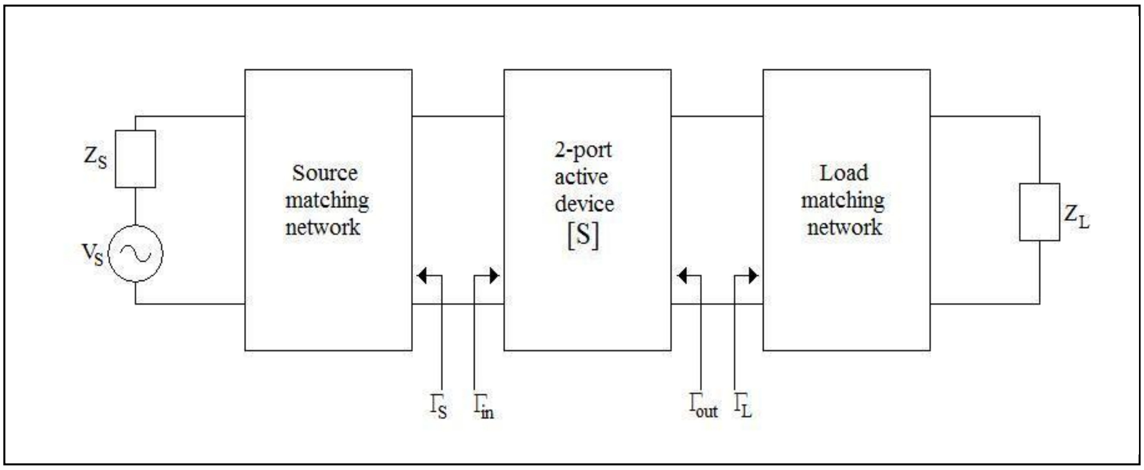


Fig.3: 1- Port oscillator

However, such a feedback cannot be determined analytically before the design is implemented as the feedback may arrive in many disguised ways. Some of the culprits for this are parasitic capacitances, poor component grounding, inductive or capacitive coupling etc. Hence stability analysis is a crucial factor for amplifier design.

Thus, stability can be defined as ability of an amplifier to maintain its effectiveness in its normal operating characteristics in spite of large changes in environment viz. physical temperature, signal frequency and source or load conditions. But for academic purposes, only the source and load conditions are considered i.e. an amplifier will oscillate if loop gain at input or output ports is greater than 1; in a fashion similar to the 1- port device. Fig. 4 shows a basic block diagram of an amplifier circuit and the reflection coefficients of respective stages. Here $\Gamma_{in}\Gamma_S$ is the loop gain at input port of the circuit, while $\Gamma_L\Gamma_{out}$ is the same for output port. So as mentioned earlier, oscillations occur if $|\Gamma_{in}\Gamma_S|>1$ or $|\Gamma_L\Gamma_{out}|>1$. These are the necessary conditions for amplifier stability.



where,

Fig.4: Basic block diagram of an Amplifier

Γ_S - Reflection coefficient of the source matching network.
Γ_L - Reflection coefficient of the load matching network.

Γ_{in}- Reflection coefficient at the input port of the transistor.
Γ_{out}- Reflection coefficient at the output port of the transistor.

The input and output ports of the amplifier are usually terminated into passive impedances, Thus $|\Gamma_S|<1$ and $|\Gamma_L|<1$. Now for the loop gains to be less than unity, the stability constraints are decided by Γ_{in} and Γ_{out}. Thus, for an amplifier to be stable, it is necessary that $|\Gamma_{in}|<1$ and $|\Gamma_{out}|<1$ for all $|\Gamma_S|<1$ and $|\Gamma_L|<1$, so that the products $|\Gamma_{in}\Gamma_S|<1$ and $|\Gamma_L\Gamma_{out}|<1$ for all load and source terminations thus implying the unconditional stability of the device. The terminations can be selected to have any value with magnitude of reflection coefficient less than unity.

These reflection coefficients are related by the following well known relations,

$$\Gamma_{in} = S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1-S_{22}\Gamma_L} \qquad \text{Eq. 1(a)}$$

$$\Gamma_{out} = S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1-S_{11}\Gamma_S} \qquad \text{Eq. 1(b)}$$

From the above discussion, it is evident that choice of Γ_L and Γ_S or load and source impedance determine the values of Γ_{in} and Γ_{out} respectively and hence the stability of an amplifier.

If all possible values of Γ_L and Γ_S give $|\Gamma_{in}|<1$ and $|\Gamma_{out}|<1$, then the device is termed as unconditionally stable, which says that the terminations can be selected anywhere on the Smith Chart. And if for some values of Γ_L and Γ_S the corresponding values of reflection coefficients are such that $|\Gamma_{in}|>1$ or $|\Gamma_{out}|>1$, then it is termed as potentially unstable, imposing restrictions on the selection of terminations. And if both the previous conditions are not true, the device is unconditionally unstable.

It is now very obvious that if the transistor is unconditionally unstable, then the amplifier cannot be designed using the existing device or the biasing conditions because it will violate the necessary conditions for stability and always break into oscillations. Hence either the device itself is changed or the DC biasing is changed and the above process is done through again.

Stability of an amplifier can be verified using either graphical or analytical approach. The graphical analysis allows the designers to select Γ_S and Γ_L directly from Smith Chart. However, it is unnecessary to do this if the transistor is unconditionally stable because in such condition the transistor will be stable for any termination.

Hence, there is a need to follow an analytical approach to determine stability and then go for graphical analysis only when need arises, i.e. when the device is potentially unstable.

It must be noted here that the stability analysed at all the frequencies and not only at the frequency of interest as the device has a good chance of running into oscillations. For example, the 50 Hz supply hum may enter the amplifier and give favourable conditions for rendering the device unstable.

1) *Analytical approach*: As stated earlier, the necessary conditions for stability of an amplifier i.e.

$$|\Gamma_{in}| = \left|S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1-S_{22}\Gamma_L}\right| < 1 \qquad \text{Eq. 2(a)}$$

$$|\Gamma_{out}| = \left|S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1-S_{11}\Gamma_S}\right| < 1 \qquad \text{Eq. 2(b)}$$

for all $|\Gamma_S|<1$ and $|\Gamma_L|<1$.

By conformal mapping, the solution of Eq. 2(a) and Eq. 2(b) is achieved, which results into following three tests for stability analysis by analytical approach.

- 3-parameter test:

$$|S_{12}S_{21}| < 1 - |S_{11}|^2 \qquad \text{Eq. 3}$$

$$|S_{12}S_{21}| < 1 - |S_{22}|^2 \qquad \text{Eq. 4}$$

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}S_{21}|} > 1 \qquad \text{Eq. 5}$$

Here, the Rollet factor (K), also called the stability factor can be used to estimate the stability of the device.

When $-1 < K \leq +1$; the device could be potentially unstable. But when $K \leq -1$, then the device will be unconditionally unstable.

- 2-parameter test:

An attempt to combine the Eq. 3 and Eq. 4 reveals the following conditions:

$$|\Delta| = |S_{11}S_{22} - S_{12}S_{21}| < 1 \qquad \text{Eq. 6}$$

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}S_{21}|} > 1$$

(OR)

The Eq. 6 with Eq. 3 and Eq. 4 respectively, further simplifies the results to:

$$B_1 = 1 + |S_{11}|^2 - |S_{22}|^2 - |\Delta|^2 > 0 \qquad \text{Eq. 7}$$

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}S_{21}|} > 1$$

(OR)

$$B_2 = 1 - |S_{11}|^2 + |S_{22}|^2 - |\Delta|^2 > 0 \qquad \text{Eq. 8}$$

$$K = \frac{1 - |S_{11}|^2 - |S_{22}|^2 + |\Delta|^2}{2|S_{12}S_{21}|} > 1$$

The 2- parameter test in terms of Eq. 6 is generally used when transistor is unilateral, and in terms of Eq. 7 and Eq. 8 when bilateral, so as to reduce the computations.

- 1-parameter test:

If stability of two devices is to be compared, it becomes difficult to do so using both, 3 or 2 parameter tests. Hence, an attempt was made to combine these two tests to one test called µ-parameter test.

According to the test,

$$\mu_1 = \frac{1 - |S_{22}|^2}{|S_{11} - \Delta S_{22}^*| + |S_{12}S_{21}|} > 1 \qquad \text{Eq. 9(a)}$$

and

$$\mu_2 = \frac{1 - |S_{11}|^2}{|S_{22} - \Delta S_{11}^*| + |S_{12}S_{21}|} > 1 \qquad \text{Eq. 9(b)}$$

(Provided $|S_{11}| < 1 \; and \; |S_{22}| < 1$)

(where; $\mathbf{a}^*$ represents the complex conjugate of **a**).

Physically, μ_1 is the minimum distance between centre of Smith chart and load stability circle and μ_2 is the same for source stability circle.

Hence, larger the value of µ, greater is the stability. In Fig. 5, two different devices are compared for stability. As the value of μ_1 is greater for the second device, it is comparatively more stable than the first device.

For a given test to be satisfied, all of the conditions mentioned in the respective tests must be satisfied. However all these tests are interdependent and if any one test is satisfied, it implies that the rest are also satisfied.

2) *Graphical approach*: To analyze the stability of an amplifier graphically, the following condition is used,

$$|\Gamma_{out}| = \left|S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1 - S_{11}\Gamma_S}\right| = 1 \qquad \text{Eq. 10}$$

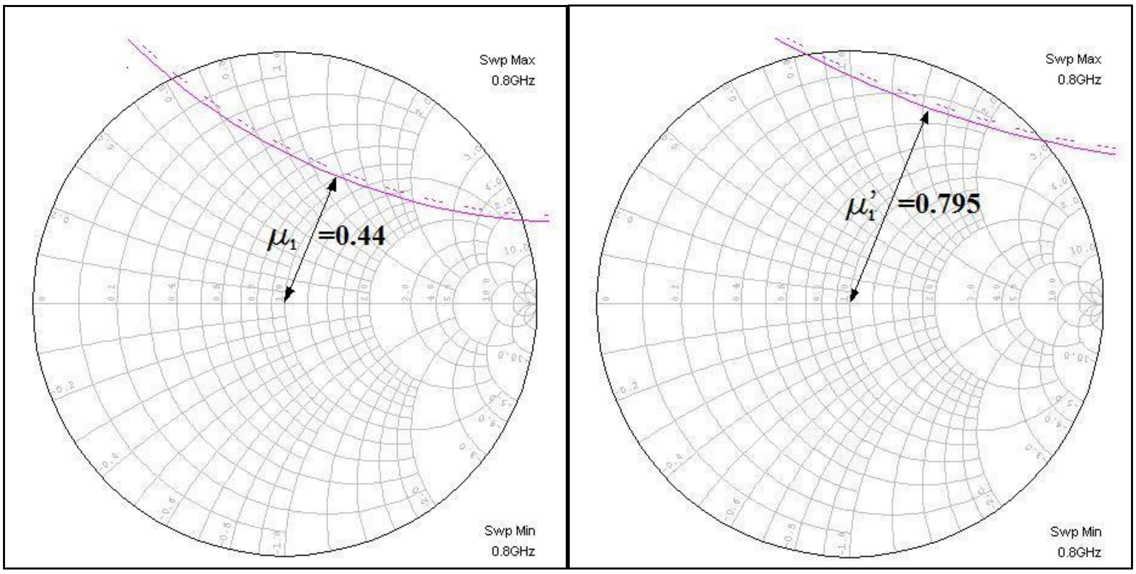
This represents the condition for marginal stability.

Solving Eq. 10 for Γ_S gives a locus of points satisfying the condition $|\Gamma_{out}|=1$. This when plotted on the Γ_S-plane i.e. Smith Chart represents a circle and is referred to as load stability circle. Similarly, the locus of $|\Gamma_{in}|=1$ is called source stability circle and plotted on Γ_L-plane.

The centre and radius of source stability circle is C_S and r_S respectively, while that of load stability circle is C_L and r_L respectively and are given by:

$$C_S = \frac{(S_{11} - \Delta S_{22}^*)^*}{|S_{11}|^2 - |\Delta|^2} \qquad \text{Eq. 11(a)}$$

$$r_S = \left|\frac{S_{12}S_{21}}{|S_{11}|^2 - |\Delta|^2}\right| \qquad \text{Eq. 11(b)}$$



(a) Load stability circle for device 1 (b) Load stability circle for device 2

Fig.5: Stability circles showing the μ parameter for two different devices.

and

$$C_L = \frac{(S_{22} - \Delta S_{11}^{*})^{*}}{|S_{22}|^2 - |\Delta|^2} \qquad \text{Eq. 12(a)}$$

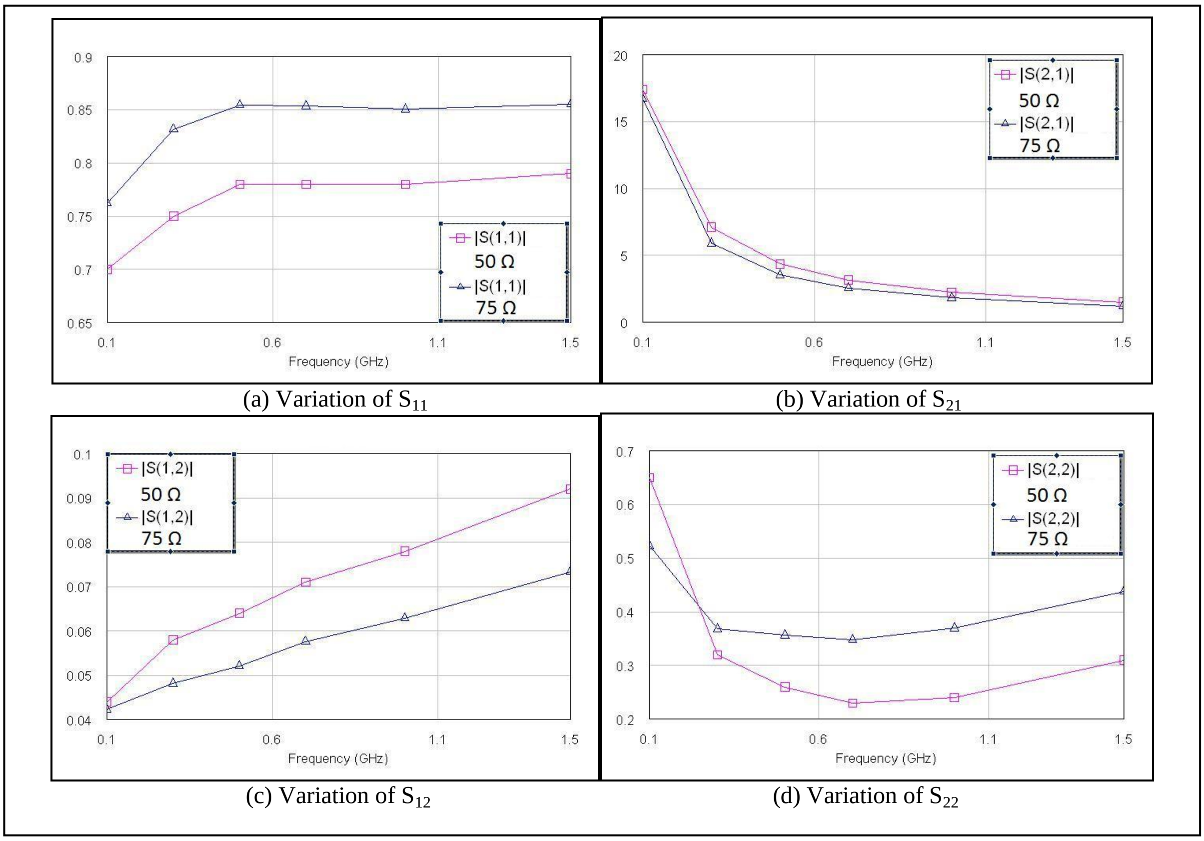
$$r_L = \left| \frac{S_{12} S_{21}}{|S_{22}|^2 - |\Delta|^2} \right| \qquad \text{Eq. 12(b)}$$

As the locus of $|\Gamma_{out}|=1$ is a circle, it can be said that the loci of $|\Gamma_{out}|<1$ and $|\Gamma_{out}|>1$ are also circles either lying within or outside the stability circle depending on the S parameters. Now, all the $|\Gamma_S|<1$ corresponding to $|\Gamma_{out}|<1$ will constitute for stable terminations, while those corresponding to $|\Gamma_{out}|>1$ will constitute for unstable terminations. They constitute the stable and unstable regions respectively. These regions can be determined using the S-parameters of the device. To find the stable and unstable regions, it is sufficient to find out if the region inside or outside the stability circle corresponds to $|\Gamma_{out}|>1$. To evaluate this, any Γ_S is selected on the Smith Chart and the value of Γ_{out} is calculated using Eq. 1(b). If here, $|\Gamma_{out}|>1$, then the region between the Smith Chart boundary and stability circle containing that point will correspond to the unstable region; and the other region, to stable region. The simplest point to select would be the centre of the Smith Chart i.e. the device is terminated by a load of value Z_0 (i.e. the characteristic impedance of the transmission line used for that application). The reason for choosing the centre only is that the S-parameters are measured by terminating the device in a load corresponding to Z_0 and S-parameters are dependent on the terminations as shown in the following graphs, Fig. 6.

Fig. 6 illustrates the dependence of the S-parameters on the terminating impedances. In the first case, the source and load impedances are both maintained at 50 Ω and the S-parameters of the circuit are measured. In next case, the termination are changed to 75 Ω, keeping the rest of the setup same as previous, again the S-parameters are measured. These variations are plotted graphically for a better insight.

If stability is analysed using S- parameters measured at Z_0 but a termination different from Z_0, then the analysis will be erroneous. The values of Γ_{in} and Γ_{out} thus calculated at the centre of the Smith Chart will be closer to the practical values to the greatest approximation. Thus, once the centre of smith chart is selected for analysing the stability, the reflection coefficients Γ_L and Γ_S will be zero.

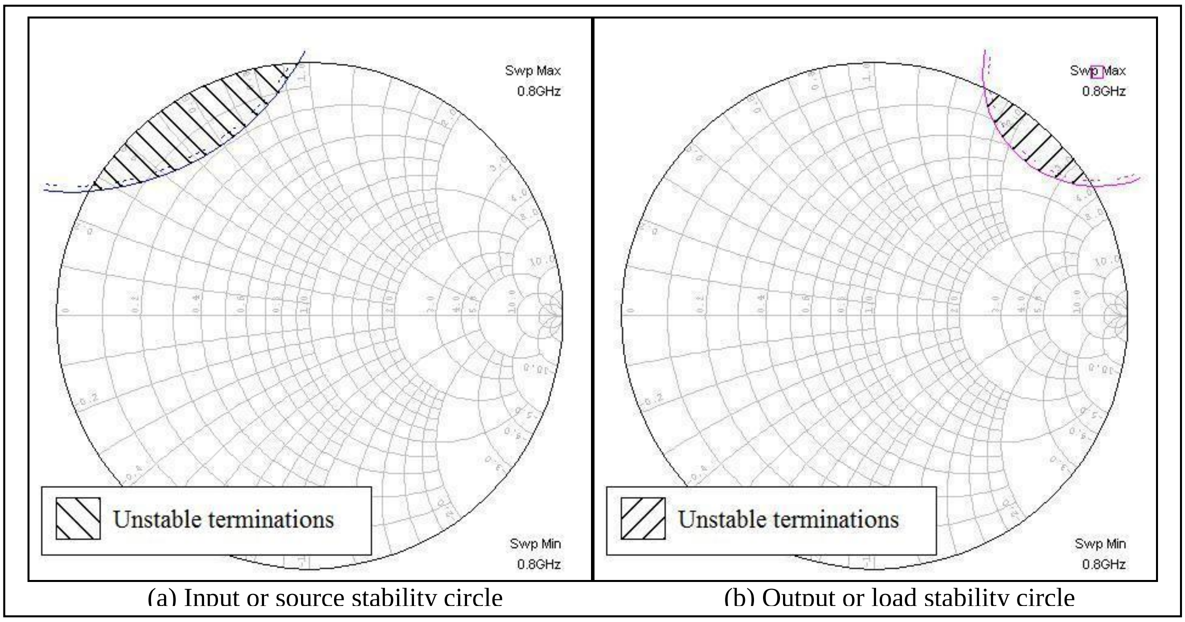
Therefore, by Eq. 1(a) and Eq. 1(b) Γ_{in} and Γ_{out} are reduced to S_{11} and S_{22} respectively. Now, if $|S_{22}|<1$ then $|\Gamma_{out}|<1$ at $\Gamma_S=0$ or at $Z_S=Z_0$. Thus, the centre of Smith chart provides a stable source termination. Generalizing this, it can be said that all the terminations in the region containing the centre will all yield different values of Γ_{out} such that $|\Gamma_{out}|<1$. In other words, this is the stable region of the Smith chart for source terminations and needless to say, the other region will be unstable.



(a) Variation of S_{11} (b) Variation of S_{21}

(c) Variation of S_{12} (d) Variation of S_{22}

Fig.6: Variation of S-parameters with load impedances.



(a) Input or source stability circle (b) Output or load stability circle

Fig.7: Stability circles for stability analysis by graphical approach.

And if $|S_{22}|>1$, then $|\Gamma_{out}|>1$ for $\Gamma_S=0$. Hence, region containing centre will now yield unstable source termination and other region, stable.

The Fig. 7(a) shows the source stability circle and the shaded region denotes the unstable region when $|\Gamma_{out}|< 1$ at the centre of the Smith Chart. However the stable and unstable regions will be interchanged if $|\Gamma_{out}|>1$ at the centre.
Similar argument can be extended for source stability also, as illustrated in Fig. 7(b).
However, this method of determining stable and unstable regions doesn't hold good if the stability circles passes through centre of Smith chart, as in this case the device will always be marginally stable. In such a case, either an open circuit termination ($\Gamma_L=1$) or a short circuit termination ($\Gamma_L= -1$) is chosen. Fig. 8 demonstrates this. Here, the device becomes unstable for a short circuit termination, so the corresponding region gives unstable terminations (denoted by the shaded region).

In this manner, regions for stable and unstable terminations can be determined and a stable termination can be chosen satisfying various considerations viz. gain, noise and VSWR as discussed in the following sections. But if the transistor is unconditionally stable the complete Smith Chart will give stable terminations and so this graphical analysis is not needed and any termination from entire Smith chart can be chosen to satisfy the required conditions.

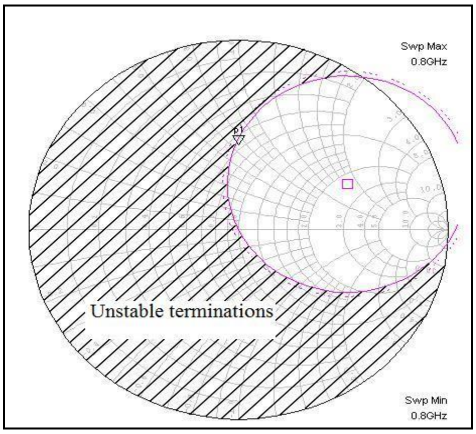


Fig.8: Stability circle passing through the centre of the Smith Chart

E. Unilateral Approximation

A 2-port network is defined as unilateral if it has no feedback from port 2 to port 1, otherwise it is bilateral. The 2-port network in this case is a transistor and mathematically, if $S_{12}=0$, then it is said that transistor is unilateral. But it must be noted here that in practical cases, S_{12} is seldom zero. Such a non-zero S_{12} poses difficulties in designing amplifier with gains less than maximum gain (this is discussed in later section). So as to justify simplifications in designing, the transistor can be assumed to be unilateral. But this introduces an error in the gain calculated by this assumption. This has a large effect on the design, as gain calculated at $S_{12}=0$ and at $S_{12}\neq0$ will be different. In order to determine if the approximation is valid or not the ratio of transducer gain in bilateral case (G_T) to that in unilateral case (G_{TU}) is calculated. If S_{12} is actually zero, then the ratio will be unity; but here, S_{12} is assumed to be zero. Hence, the ratio will be some other value but unity. This deviation can be quantified by the "unilateral figure of merit" or 'U'. It is used to calculate the maximum error range that may occur if a bilateral transistor is approximated to be unilateral. The error range is given by:

$$\frac{1}{(1+U)^2} \leq \frac{G_T}{G_{TU}} \leq \frac{1}{(1-U)^2} \quad \text{Eq. 13}$$

Where,

$$U = \left|\frac{S_{11}S_{12}S_{21}S_{22}}{(1-|S_{11}|^2)(1-|S_{22}|^2)}\right| \quad \text{Eq. 14}$$

As a rule of thumb, the transistor can be assumed unilateral if this ratio of G_T:G_{TU} is between ±0.5dB. Though it is not a hard and fast rule, this range can be assumed as suited for the desired design. Hence, the range may vary with the application of the circuit. But, if this assumption is not valid then bilateral approach has to be followed.

F. Gain consideration

Gain of an amplifier is another important aspect that has to be considered during its designing as it decides the output signal level which may be fed to further stage in an application. With this in mind, a particular value of gain is set, which is practically achievable.
There are many definitions for gains of an amplifier, but in this section focus is on the Transducer gain (G_T). It is given as:

$$G_T = \frac{P_L}{P_{AVS}}$$

where, P_L≡Power delivered to load

P_{AVS}≡Power available from source

Further, it can be derived that,

$$G_T = \frac{1-|\Gamma_S|^2}{|1-\Gamma_S\Gamma_{in}|^2}|S_{21}|^2\frac{1-|\Gamma_L|^2}{|1-\Gamma_L S_{22}|^2} \quad \text{Eq. 15}$$

This can be rewritten as,

$$G_T = G_S G_O G_L \qquad \text{Eq. 16}$$

where,

$G_S \equiv$ Gain of source matching network

$G_L \equiv$ Gain of load matching network

$G_O \equiv$ Operating gain

The formulae of G_S, G_L and G_O are as follows:

$$G_S = \frac{1-|\Gamma_S|^2}{|1-\Gamma_S\Gamma_{in}|^2} \qquad \text{Eq. 16(a)}$$

$$G_O = |S_{21}|^2 \qquad \text{Eq. 16(b)}$$

$$G_L = \frac{1-|\Gamma_L|^2}{|1-\Gamma_L S_{22}|^2} \qquad \text{Eq. 16(c)}$$

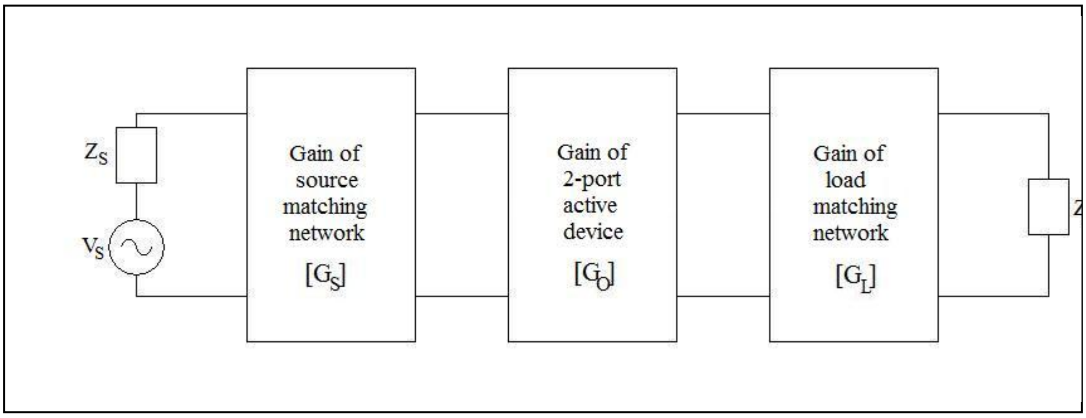


Fig.9: Block diagram of the amplifier showing gains from each block.

The block diagram of an amplifier, in Fig. 9, represents the gains of corresponding blocks. The matching networks are equivalent to circuit comprising of R, L and C.

Now, one may ask that how can matching networks provide gain even if they constitute of passive element only? This can be explained as follows:

Consider the case when an amplifier is implemented on a substrate, then both ports are terminated into a microstrip line or transmission line of characteristic impedance Z_0. It must be noted here that the input and output impedance of amplifier may not be equal to Z_0.

In other words, if input impedance is not matched to Z_0 then due to reflection at input port of the transistor, a loss called mismatch loss is incurred as a result of mismatch between input port and input port of amplifier. This reduces the overall gain of amplifier unit. If in some way, this mismatch is reduced then, the gain of amplifier can be improved i.e., if the input of amplifier is matched to Z_0, the mismatch between source and input port can be reduced and thus the gain is enhanced. The same explanation is valid for output side also. It is for this reason that the matching networks are said to "provide" gain. By appropriately selecting these matching networks the required value of gain can be achieved. But individual gains of the matching networks cannot be achieved beyond a certain value given by G_{Smax} and G_{Lmax}.

where,

$$G_{Smax} = \frac{1}{|1-S_{11}|^2} \qquad \text{Eq. 17(a)}$$

$$G_{Lmax} = \frac{1}{|1-S_{22}|^2} \qquad \text{Eq. 17(b)}$$

Now, depending on whether the transistor is unilateral or bilateral, one of the two approaches is followed to achieve the desired gain. The two approaches are explained in detail below.

1) Transistor is unilateral or can be assumed to be unilateral:

In this case, S_{12} is zero and Γ_{in} and Γ_{out} are reduced to $\Gamma_{in} = S_{11}$ and $\Gamma_{out} = S_{22}$ respectively. Also, G_T for unilateral case (called G_{TU}) is given by,

$$G_{TU} = \frac{1-|\Gamma_S|^2}{|1-\Gamma_S S_{11}|^2}|S_{21}|^2\frac{1-|\Gamma_L|^2}{|1-\Gamma_L S_{22}|^2} \qquad \text{Eq. 18}$$

Now, if maximum gain is desired from an amplifier or if maximum power transfer from source to active device and active device to load is to be achieved then this condition is possible only if $Z_S=Z_{in}^*$ and $Z_L=Z_{out}^*$.
where,

$Z_S \equiv$ Impedance of source matching network

$Z_L \equiv$ Impedance of load matching network

$Z_{in} \equiv$ Input impedance of transistor

$Z_{out} \equiv$ Output impedance of transistor

The same can be stated as $\Gamma_S=\Gamma_{in}^*$ and $\Gamma_L=\Gamma_{out}^*$

The maximum unilateral transducer gain thus achieved if $\Gamma_S=\Gamma_{in}^*=S_{11}^*$ and $\Gamma_L=\Gamma_{out}^*=S_{22}^*$ is,

$$G_{TUmax} = \frac{1}{1-|S_{11}|^2}|S_{21}|^2\frac{1}{1-|S_{22}|^2} \quad \text{Eq. 19}$$

As stated earlier, the mismatch cannot be improved beyond a value limited by G_{Smax} and G_{Lmax} given by Eq. 17(a) and Eq. 17(b). But, it must be noted that if the gain desired is equal to the gain provided by the transistor itself i.e. G_O, then no matching network is needed and ports can be terminated directly into Z_0.
To cut the story short, once the device is selected and biased; the gain, $|S_{21}|^2$ is set, it now depends on the source and load matching networks to achieve the required gain by either increasing or reducing the mismatch in the circuit and thus, designing of these matching networks is important in determining the amplifier gain.

Till now, the cases when maximum gain is required or when desired gain is equal to the inherent gain of the device have been discussed. But what if the desired gain is not equal to that provided by the device nor is it equal to the maximum gain? It is possible to design such an amplifier using matching network to add some gain to the network provided the desired gain is less than the maximum. In the previous case, where maximum gain was desired, the source and load termination were satisfied by a single point. But now, it is observed that several terminations provide same value of gain. If the locus of all such terminations is plotted on the Smith chart "constant gain circles" are obtained. The centres and radii of which vary with the value of gain.
To determine this locus consider the normalised gain from source matching network as,

$$g_S = \frac{G_S}{G_{Smax}} \quad \text{i.e.}$$

$$g_S = \frac{(1-|\Gamma_S|^2)(1-|S_{11}|^2)}{|1-S_{11}\Gamma_S|^2} \quad \text{Eq. 20}$$

Solving Eq. 20 for Γ_S, the centre and radius for the source gain circle i.e., constant gain circle for source matching network i.e. C_{g_s} and r_{g_s} are obtained as,

$$C_{g_s} = \frac{g_S S_{11}^*}{1-|S_{11}|^2(1-g_S)} \quad \text{Eq. 21(a)}$$

$$r_{g_s} = \frac{\sqrt{(1-g_S)}\,(1-S_{11}^2)}{1-|S_{11}|^2(1-g_S)} \quad \text{Eq. 21(b)}$$

Similarly, C_{g_L} and r_{g_L} are obtained as,

$$C_{g_L} = \frac{g_L S_{22}^*}{1-|S_{22}|^2(1-g_L)} \quad \text{Eq. 22(a)}$$

$$r_{g_L} = \frac{\sqrt{(1-g_L)}\,(1-S_{22}^2)}{1-|S_{22}|^2(1-g_L)} \quad \text{Eq. 22(b)}$$

and,

$$g_L = \frac{G_L}{G_{Lmax}}$$

where, C_{g_L} and r_{g_L} denote centre and radius for load gain circle i.e., constant gain circle for load matching network and g_L is the normalised gain from load matching network.

The circumference of these circles provide values of termination giving a particular gain to circuit and the termination inside the circle give a higher gain value while termination outside the circle reduce the value of gain.
And if desired gain is lesser than G_O, then the mismatch needs to be increased in a controlled amount. Thus, the matching network can have a negative gain (in dB) as in contrast to the earlier case where, the gains were positive.
However, it must be noted here that all of the above discussion is valid only if the transistor is unconditionally stable. If it is potentially unstable, then the designer must take care that the termination that is selected lies in the stable region. Hence, the argument that for maximum gain $\Gamma_S=S_{11}^*$ and $\Gamma_L=S_{22}^*$ holds good as long as S_{11}^* and S_{22}^* lie in the stable region for source and load terminations respectively. But if the conditions are not satisfied then the design for maximum gain is not possible. In such cases, various gain circles can be drawn and the optimum termination may be selected in the stable region.

2) *Transistor is bilateral*:
As stated earlier, transistor is bilateral if $S_{12}\neq 0$. This implies that Γ_{in} and Γ_{out} are functions of Γ_L and Γ_S respectively i.e., $\Gamma_{in}=f(\Gamma_L)$ and $\Gamma_{out}=f(\Gamma_S)$.
Say, a design for maximum transducer gain is needed, here terminations must be conjugate matched i.e., $\Gamma_S=\Gamma_{in}^*$ and $\Gamma_L=\Gamma_{out}^*$.
Therefore, $\Gamma_S= [f(\Gamma_L)]^*$ and $\Gamma_L= [f(\Gamma_S)]^*$.

i.e.

$$\Gamma_S = \left(S_{11} + \frac{S_{12}S_{21}\Gamma_L}{1-S_{22}\Gamma_L}\right)^* \qquad \text{Eq. 23(a)}$$

and

$$\Gamma_L = \left(S_{22} + \frac{S_{12}S_{21}\Gamma_S}{1-S_{11}\Gamma_S}\right)^* \qquad \text{Eq. 23(b)}$$

If and only if the above two conditions are satisfied then maximum gain for bilateral case is achieved. Thus, by solving above two equations simultaneously, values of Γ_S and Γ_L are obtained, which when used to terminate the amplifier maximum transducer gain is obtained.
The values of Γ_S and Γ_L obtained here are called Γ_{MS} and Γ_{ML} which are,

$$\Gamma_{MS} = \frac{B_1 - \sqrt{B_1^2 - 4|C_1|^2}}{2C_1} \qquad \text{Eq. 24(a)}$$

and

$$\Gamma_{ML} = \frac{B_2 - \sqrt{B_2^2 - 4|C_2|^2}}{2C_2} \qquad \text{Eq. 24(b)}$$

B_1 and B_2 have usual meanings and,

$$C_1 = (S_{11} - \Delta S_{22}^*)$$
$$C_2 = (S_{22} - \Delta S_{11}^*)$$

Because of this, the process is called as simultaneous conjugate matching.
In such a case, the maximum transducer gain is found to be-

$$G_{T_{max}} = \frac{|S_{21}|}{|S_{12}|}\left(K - \sqrt{K^2 - 1}\right) \qquad \text{Eq. 25}$$

These values of Γ_{MS} and Γ_{ML} stand true so long that the device is unconditionally stable. Since, Γ_{MS} and Γ_{ML} are derived for K>1 and $|\Delta|<1$; which by 2 parameter test implies unconditional stability. So, transducer gain approach is not valid when device is potentially unstable. In such cases, available gain or power gain approach are used, which are discussed later.

Another limitation of transducer gain approach is faced when required gain is less than maximum gain. At first thought, one may impulsively say that gain circles can be drawn as in unilateral case and solve the problem. But there is a twist in the story here and is explained as follows:
Consider transducer gain equation,

$$G_T = G_S' G_O G_L' \qquad \text{Eq. 26}$$

where,

$$G_S' = \frac{1-|\Gamma_S|^2}{|1-\Gamma_{in}\Gamma_S|^2} \qquad \text{Eq. 26(a)}$$

$$G_O = |S_{21}|^2 \qquad \text{Eq. 26(b)}$$

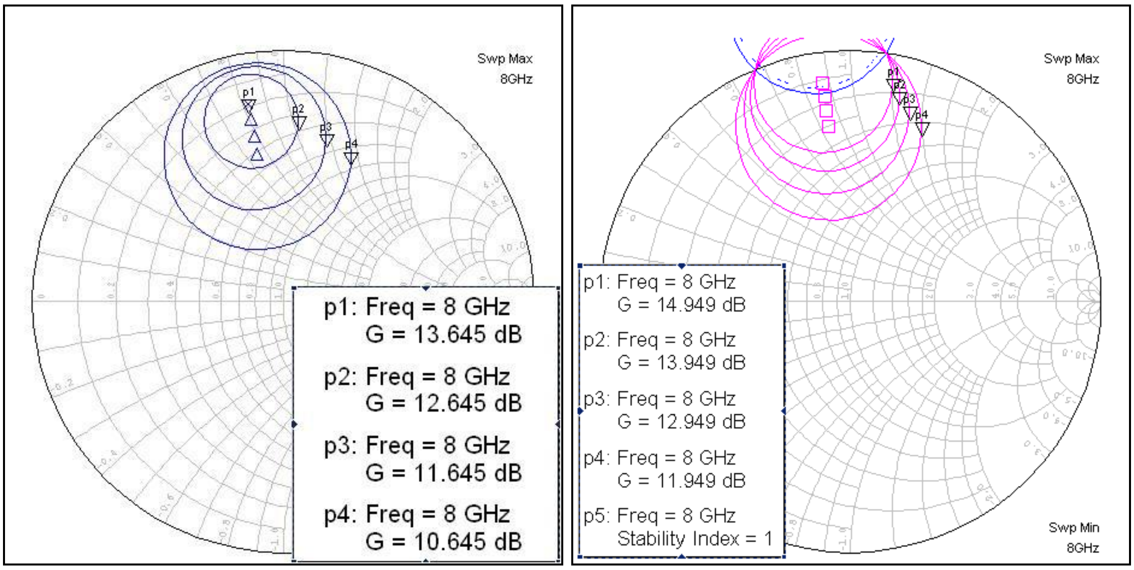
$$G_L' = \frac{1-|\Gamma_L|^2}{|1-S_{22}\Gamma_L|^2} \qquad \text{Eq. 26(c)}$$

As seen here, G_T depends on both Γ_S and Γ_L at the same time and also that G_S' and G_L' are interdependent i.e. a selected value of G_L' will affect the selection of G_S' and vice versa. This is elaborated below.
Say that a value of G_L' is selected and corresponding gain circle is drawn. Selecting any value of Γ_L on the circle will fix a value of Γ_{in}, since it depends on Γ_L as written in Eq. 1(a).

Furthermore, if maximum power transfer is desired from source to the input of transistor then $\Gamma_S=\Gamma_{in}^*$. Selecting this value of Γ_S and Γ_{in} will yield a value of G_S' which may not be equal to the desired value of G_S'.
Hence, another value of Γ_L needs to be selected and the same procedure is followed again. This needs to be repeated until a value of Γ_S and Γ_L so obtained will satisfy the desired values of G_S' and G_L'. Thus, the process becomes iterative and as a result is very tedious and tiresome. Here, again power and available gain approach help us out of the problem.

<u>Power gain or Operating gain approach:</u>

In this method, first the desired load impedance is set and then resulting input impedance is matched. It is usually used in case of power amplifier where the load termination is more important than the source termination. Thus, input port is matched while output port is not. As such, gain provided will not be maximum; but this is a sacrifice made so as to achieve maximum power transfer to output.



(a) Power gain circles for unconditionally stable case. (b) Power gain circles for potentially unstable case.

Fig.10: Constant power gain circles

The power gain G_P is a function of Γ_L i.e. load termination only (of course the S-parameter also affect the G_P), this in contrast to the transducer gain G_T which is a function of Γ_S and Γ_L both. G_P is defined as:

$$G_P = \frac{P_L}{P_{in}}$$

where, $P_{in} \equiv$ Power input to the amplifier.

This can also be expressed as:

$$G_P = \frac{1}{(1-|\Gamma_{in}|^2)} |S_{21}|^2 \frac{(1-|\Gamma_L|^2)}{|1-S_{22}\Gamma_L|^2} \quad \text{Eq. 27}$$

Because of this independence from Γ_S, the iterative problem incurred in high gain designing in transducer gain approach is overcome.

Solving Eq. 27 for Γ_L, gives a locus of points forming a circle called constant power gain circle. This locus represents a set of load termination giving a specific G_P. For unconditional stability case, these circles for different gains lie completely within the circle described by $|\Gamma_L|=1$; while for potentially unstable case, some part of gain circles lies outside the domain of $|\Gamma_L|=1$ as shown in Fig. 10.

In short, a Γ_L is selected as per the requirements and if small signal gain thus provided is insufficient then constant operating circles are plotted and checked for gain-output power tradeoffs by overlaying the power contours on the gain circles. This provides a new Γ_L. Using this, Γ_{in} is calculated which in turn is conjugate matched to source i.e. $\Gamma_S = \Gamma_{in}^*$. Now, these Γ_S and Γ_L are matched to input and output port of the transistor using appropriate matching network topology.

<u>Available gain approach:</u>

This method is analogous to the power gain approach. Here, designing begins with selecting a particular source termination and ends with a conjugate matched load termination. The available gain is given by following formula,

$$G_A = \frac{P_{AVN}}{P_{AVS}}$$

where,

$P_{AVN} \equiv$ Power available from the network

It can be further derived that:

$$G_A = \frac{(1-|\Gamma_S|^2)}{|1-S_{11}\Gamma_S|^2} |S_{21}|^2 \frac{1}{(1-|\Gamma_L|^2)} \quad \text{Eq. 28}$$

Here, G_A is a function of Γ_S and not of Γ_L. Solving G_A for Γ_S, another family of circles is obtained for different values of G_A, called constant available gain circles whose characteristics are very much similar to the constant power gain circles.

This technique is normally employed in LNA design wherein the source termination is a determining factor for the overall noise performance of amplifier.

An overlay of noise and available gain circles allows the trade-off analysis between gain and noise of amplifier. Using this trade-off, a value of Γ_S is selected. Due to such a low noise consideration, a mismatch is allowed at the input port. This mismatch is compensated by maximising the "available" gain at the output port by selecting a Γ_L as conjugate of Γ_{out} i.e. $\Gamma_L = \Gamma_{out}^*$; where Γ_{out} can be calculated using previous Eq. 1(b). This Γ_S and Γ_L is then matched to the respective ports of the amplifier.

G. Noise considerations

Noise is a random phenomenon that occurs in almost all electronic devices. Noise can be broadly classified into two types: internal and external. Three major contributors of the internal noise generated in the system are Thermal or Johnson noise, Shot or Schottky noise and flicker or 1/f noise. On the other hand external noise is caused due to the ambient effects on the system with cosmic, galactic, atmospheric, solar, etc. being some examples.

Noise normally degrades system performance and hence its analysis becomes necessary when an amplifier with low noise levels is needed. The thermal noise is broadband and has an effect over a large range of frequencies. Hence it is described in the further section. To measure the noise performance of a system a quantity called figure of merit (F) is defined. It is basically the ratio of signal to noise ratio (SNR) of input to SNR of output. F can be mathematically defined as,

$$F = \frac{P_{Si}/P_{Ni}}{P_{So}/P_{No}}$$

For a noisy 2-port network, a detailed analysis helps derive a relation between figure of merit (or noise figure) and the source side termination. This analysis also proves that the load termination has no effect on the noise performance of system. The above said relation can be written as,

$$F = F_{min} + \frac{r_n}{g_S}|Y_S - Y_{opt}|^2 \qquad \text{Eq. 29}$$

where,

F_{min} ≡ Minimum noise figure of transistor

r_n ≡ Sensitivity of noise figure

Y_S≡ Admittance of source termination

Y_{opt}≡ Admittance of optimum termination

g_s≡ Normalized conductance of Y_S

Here, when input port of the transistor is terminated in Y_{opt}, overall figure of merit of the system becomes the minimum possible, and r_n decides the rate of deviation of noise figure from the minimum when the source termination is deviated from the optimum value.

The same equation can be expressed in terms of reflection coefficients as:

$$F = F_{min} + \frac{4r_n\,|\Gamma_S-\Gamma_{opt}|^2}{(1-|\Gamma_S|^2)\,|1+\Gamma_{opt}|^2} \qquad \text{Eq. 30}$$

Solving Eq. 30 for Γ_S gives a set of source termination which when used in the design provide certain figure of merit. The locus of such points on the Smith chart is a family of circles called constant noise figure circles which converge to a point representing Γ_{opt} as the desired figure of merit is decreased from maximum to minimum as shown in Fig. 11. The centres and radii of such circles are given by,

$$C_{N_i} = \frac{\Gamma_{opt}}{1+N_i} \qquad \text{Eq. 31(a)}$$

and,

$$r_{N_i} = \frac{1}{1+N_i}\sqrt{N_i{}^2 + N_i\left(1-|\Gamma_{opt}|^2\right)} \qquad \text{Eq. 31(b)}$$

where,

$$N_i = \frac{F - F_{min}}{4r_n}|1+\Gamma_{opt}|^2$$

and,

Γ_{opt}≡Reflection coefficient corresponding to Y_{opt}.

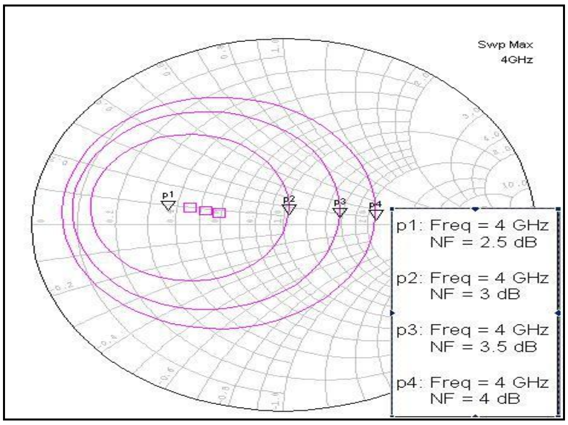


Fig.11: Noise Circles

Now, there is a trade-off between gain and noise. It goes as; minimum noise requires low I_C while maximum gain needs high I_C. So after setting a compromise between the desired noise and gain levels, a matching network is designed as elaborated in available gain approach.

H. VSWR Consideration

Reiterating the fact that when the ports of amplifier are not terminated into conjugate load or source terminations then reflection of transmitted signal occurs, giving rise to standing waves. Hence, a VSWR not equal to unity persists along the line at both the ports. Such reflection at input port can hamper the characteristics of the source viz. amplitude, signal, phase, etc. of applied signal and cause a serious damage to previous stages; whereas at output port reflection may interfere with the smooth functionality of the device itself and unnecessarily cause oscillations if reflection is not controlled.

As such, study of VSWR becomes an important consideration. The basic schematic block diagram of an amplifier is shown in Fig. 12.

At the source side, the power available from source doesn't reach to the input port. Thus, due to the above said mismatch, power input to device will be a reduced value of P_{AVS} by a factor M_S. Mathematically, can be expressed as,

$$P_{in} = M_S P_{AVS}$$

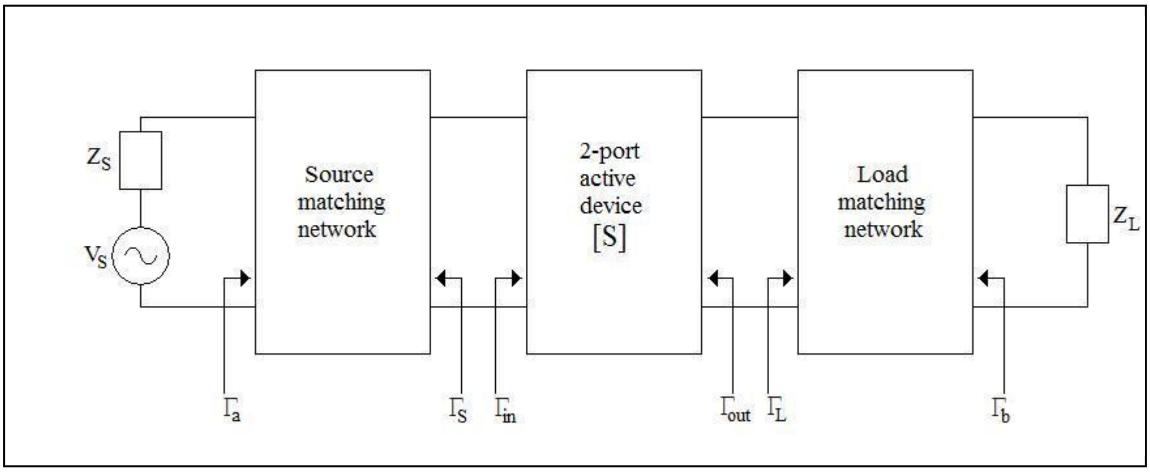


Fig.12: Block diagram of an amplifier showing reflection coefficients at each block

where,

$$M_S = \frac{(1-|\Gamma_S|^2)(1-|\Gamma_{in}|^2)}{|1-\Gamma_S\Gamma_{in}|^2} \qquad \text{Eq. 32(a)}$$

Extending this explanation to the load side, the mismatch factor at output can be defined as,

$$M_L = \frac{(1-|\Gamma_L|^2)(1-|\Gamma_{out}|^2)}{|1-\Gamma_L\Gamma_{out}|^2} \qquad \text{Eq. 32(b)}$$

and

$$P_{out} = M_L P_{AVN}$$

Also,

$$M_S = 1 - |\Gamma_a|^2 \qquad \text{Eq. 33(a)}$$
$$M_L = 1 - |\Gamma_b|^2 \qquad \text{Eq. 33(b)}$$

Equating Eq. 32(a) and Eq. 33(a), Eq. 32(b) and Eq. 33(b);

$$|\Gamma_a| = \left|\frac{\Gamma_{in} - \Gamma_S^*}{1 - \Gamma_{in}\Gamma_S^*}\right| \qquad \text{Eq. 34(a)}$$
$$|\Gamma_b| = \left|\frac{\Gamma_{out} - \Gamma_L^*}{1 - \Gamma_{out}\Gamma_L^*}\right| \qquad \text{Eq. 34(b)}$$

The VSWR corresponding to these reflections from source and load matching networks are,

$$(VSWR)_{in} = \frac{1 + |\Gamma_a|}{1 - |\Gamma_a|} \qquad \text{Eq. 35(a)}$$
$$(VSWR)_{out} = \frac{1 + |\Gamma_b|}{1 - |\Gamma_b|} \qquad \text{Eq. 35(b)}$$

VSWR will be unity when Γ_a and Γ_b are zero, or when $\Gamma_S = \Gamma_{in}^*$ and $\Gamma_L = \Gamma_{out}^*$ i.e. the case of conjugate matched loads.

I. Selection of matching network topology

The need for matching network has already been discussed in previous sections as in to achieve required level of gain, noise, power transfer, etc. Such a matching network can be designed using either transmission line or lumped elements. But practically in most cases transmission lines are used; microstrip lines to be more precise. A single stub matching can be used for the design. However, to reduce the transition interaction between the line and stub, balanced stub matching can be employed. But, if lumped element matching technique is used, then the preferred circuit is shown in Fig. 13, so that DC biasing becomes easy by adding a DC source in series with inductor. Also it acts as a High Pass Filter and blocks the low frequency baseband signals which can cause oscillations due to an inherent high gain of the device at lower frequencies.

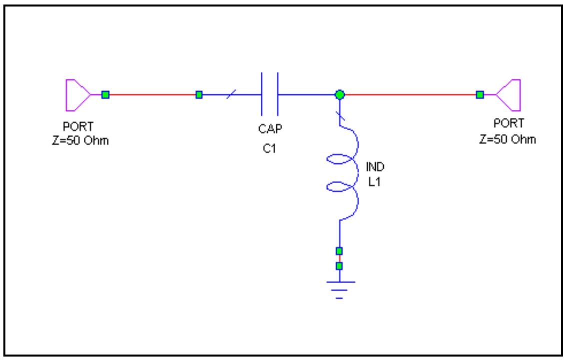


Fig.13: An example of matching network

III. CONCLUSION

The above discussion provides a lucid understanding of the procedure for amplifier design. However, the techniques discussed here are exclusively applicable only to narrowband amplifier because of following reasons; S-parameters are frequency dependent and gain is higher at low frequency than at higher frequency. Also, the matching networks hold good for barely an octave as the passive elements here are largely dependent on frequency.

REFERENCES

[1]. Vendelin, G. D., Pavio, A. M., Rohde, U. L., *Microwave Circuit Design Using Linear And Nonlinear Techniques*, (2nd Edition), (John Wiley & Sons Publications, Inc., Hoboken, New Jersey.)

[2]. Gonzalez, G., *Microwave Transistor Amplifiers: Analysis And Design*, (2nd Edition), (Prentice Hall Publications, Inc., Upper Saddle River, New Jersey.)

[3]. Pozar, D. M., *Microwave Engineering*, (2nd Edition), (John Wiley & Sons Publications, Inc., Hoboken, New Jersey.)

[4]. Liao, S. Y., *Microwave Circuit Analysis and Amplifier Design*, (1st Edition), (Prentice Hall Publications, Inc., Upper Saddle River, New Jersey.)