LECTURE 4 – BROADBAND AMPLIFIERS

L4.1. INTRODUCTION

The challenge in designing a broadband microwave amplifier is the fact that the input impedance at lower frequencies is practically an open circuit, and at higher frequencies predominantly capacitive and can be almost a short. This makes broadband matching difficult. For example, the TriQuint device we used for the project has an input impedance that is practically an open circuit at 0.1GHz, is almost purely capacitive at 1GHz and about $35+j60\Omega$ at 5GHz (the package inductance starts dominating at the higher frequencies). It is clear that designing an impedance matching circuit that can cover this frequency range is a difficult task. In this lecture, we discuss possible approaches for small-signal amplifiers, but the challenge for power amplifiers is similar.

An example extreme input and output impedance of a LDMOS base-station power device is shown in Fig.L4.1, indicating sub-ohm values for both input and output impedances. The device was used to design a Wideband Code–Division Multiple Access (W–CDMA) base station PA, operating at the 2.11-2.17GHz range. The expected output power is 70W (CW) with gain of 12 dB. The transistor operates at 28V drain supply, with maximal drain voltage of 66V and maximal drain current of 11 A. The high power transistor consists of two 42–parallel finger cells, resulting in a very low input and output impedances. Internal pre-matching is done on the die to increase the impedance to a slightly higher level seen in the figure. This device cannot be matched over a broad bandwidth without sacrificing output power and gain.

![Figure L4.1. Input and output impedance of a LDMOS power device at 2GHz. Measured source (left) and load (right) impedance contours of the constant output power (dashed) and gain (solid) for the ACPR=-45 dBc at 2.11GHz. 50-ohm charts are shown as insets for easier orientation. Impedances for the maximum output power (x) and maximum gain (+) are also shown. Selected target impedances (black circle) are : $Z_S = (8.5 + j0)$ and $Z_L = (1.5 - j3.25)$ at 2.11 GHz.](image)
There are several ways to design a broadband input match to an amplifier, each has its drawbacks and advantages: (1) broadband non-uniform impedance matching network design; (2) balanced amplifier; (3) resistive feedback amplifier; (4) distributed and (5) traveling-wave amplifier. The first type is based on designing dispersion-compensation or pre-dispersion networks and results in very large impedance matching networks which typically have substantial loss. If we have time, we will talk about these nonuniform transmission line matching circuits.

L4.2. BALANCED AMPLIFIERS

A common approach to the problem of broadband amplifier design is a balanced amplifier configuration shown in Figure L4.2. It consists of a pair of 3-dB couplers. For example, if the couplers are ideal 90° hybrids (such as a branch line coupler), the scattering matrix for the balanced amplifier can be written as:

\[
S = \begin{bmatrix}
(s_{11} - s_{21})/2 & j(s_{12} + s_{22})/2 \\
(j(s_{21} + s_{21})/2 & (s_{12} - s_{22})/2
\end{bmatrix}.
\]

and if the two amplifiers are identical, the scattering matrix becomes:

\[
S = \begin{bmatrix}
0 & jS_{12} \\
S_{21} & 0
\end{bmatrix}
\]

This means that, as long as the amplifier circuits are identical, they can be whatever we wish, and the amplifier still has input and output matching. The two amplifiers can individually be tuned for gain, noise or flatness of frequency response. Another common balanced amplifier uses Wilkinson combiner/dividers instead of the branch-line couplers. In this case, quarter-wave sections in the two lines provide a 180 degree phase difference.
between the two waves reflected from the inputs of the amplifiers, and the reflections are
cancelled. The bandwidth of the amplifier is obviously limited by the bandwidth of the
directional coupler or Wilkinson splitter, both of which rely on quarter-wave sections for
proper operation. Therefore, a hybrid is not the best choice (it has about 15% bandwidth).
Instead, most commonly used is a Lange coupler based on coupled line sections, which
can have a bandwidth of 2 octaves. It is also possible to design broadband multi-section
branch line and Wilkinson combiners (over decade bandwidth). We will talk more about
coupled-line couplers in the next lecture.

Balanced amplifiers ideally have the same gain and twice the output power as compared
to the single amplifier. When the signal becomes large, each of the transistors receives
only half the power, so balanced amplifiers can handle more power with less signal
distortion. However, twice the input signal is required, and two times more DC power.
An additional disadvantage is the size of the circuit and the fact that a large part of the
real-estate is taken by passive circuits. This is very costly in MMIC implementations.

An important factor in balanced amplifier design is the amplitude and phase mismatch
between the coupler output ports as a function of frequency, as well as the sensitivity of
this mismatch to load impedance variations. In a practical design, this should be verified
in simulation prior to fabrication.

L4.3. RESISTIVE FEEDBACK AMPLIFIERS

Resistive feedback can also be used for designing a broadband amplifier. The effect of a
feedback resistor between the gate and drain of a MESFET is to lower the input and
output impedance and to broaden the gain curve. The drawback is resistive coupling
between the bias circuits, as well as overall lower gain than for reactively matched
amplifiers.

By observing the approximate circuit model for the MESFET with normalized resistance
values (to 50Ω), Figure L4.4, the equations for the current and voltage in the input circuit
are found to be:

\[ i_{in} = g_m v_{in} + i \]
\[ v_{in} = r_f i_{in} + \frac{r_{ds}}{r_{ds} + 1} i \]

Solving for the input impedance by eliminating \( i \), we obtain:

\[ Z_{in} = \frac{v_{in}}{i_{in}} = \frac{r_f + \frac{r_{ds}}{r_{ds} + 1}}{1 + g_m \frac{r_{ds}}{r_{ds} + 1}} \]
Figure L4.4. Simplified equivalent circuits for input and output MESFET circuits in a series feedback amplifier.

For the output circuit, the current and voltage can be expressed as:

\[
i_{\text{out}} = \frac{v_{\text{out}}}{1 + r_f} + g_m v_{\text{gs}} + \frac{v_{\text{out}}}{r_{ds}}
\]

\[
v_{\text{out}} = (1 + r_f) v_{\text{gs}}
\]

where the voltage \( v \) can be eliminated to give the expression for the output impedance:

\[
Z_{\text{out}} = \frac{v_{\text{out}}}{i_{\text{out}}} = \frac{r_{ds}}{1 + g_m} \frac{1 + r_f}{r_{ds} + \frac{1 + r_f}{1 + g_m}}
\]

In both cases, the impedance is reduced and can be controlled by the amount of feedback resistance.

Figure L4.5. Simplified circuit for series-shunt resistive feedback amplifier.

The above simplified analysis was an example of the more general case of series-shunt resistive feedback shown in Figure L4.5, where in addition to the series feedback
between gate and drain, there is a parallel feedback resistor placed in the source. The admittance matrix for this network (for a very simplified FET model) can be written as

\[
\begin{bmatrix}
 i_{in} \\
 i_{out}
\end{bmatrix} = \begin{bmatrix}
 \frac{1}{R_{fs}} & -\frac{1}{R_{fs}} \\
 -\frac{g_m}{1 + g_m R_{fp}} - \frac{1}{R_{fs}} & \frac{1}{R_{fs}}
\end{bmatrix} \begin{bmatrix}
 v_{in} \\
 v_{out}
\end{bmatrix}.
\]

The admittance matrix can now be converted to s-parameters using the standard conversion formulas:

\[
S = \begin{bmatrix}
 \frac{1}{\Delta} \left(1 - \frac{g_m Z_0^2}{R_{fs} (1 + g_m R_{fp})}\right) & \frac{2Z_0}{\Delta R_{fs}} \\
 \frac{1}{\Delta} \left(\frac{-2g_m Z_0 + 2Z_0}{1 + g_m R_{fp}} R_{fs}\right) & \frac{1}{\Delta} \left(1 - \frac{g_m Z_0^2}{R_{fs} (1 + g_m R_{fp})}\right)
\end{bmatrix}
\]

where \(\Delta = 1 + \frac{2Z_0}{R_{fs}} + \frac{g_m Z_0^2}{R_{fs} (1 + g_m R_{fp})}\). If the design attempts to obtain a match at input and output, i.e. \(s_{11} = s_{22} = 0\), then the resistor values are related to the transconductance by

\[
1 + g_m R_{fp} = \frac{g_m Z_0^2}{R_{fs}} \quad \text{or} \quad R_{fp} = \frac{Z_0^2}{R_{fs} g_m} - \frac{1}{g_m}.
\]

From the above equations, now \(s_{21}\) and \(s_{12}\) can be found to be

\[
s_{21} = \frac{Z_0 - R_{fs}}{Z_0} \quad \text{and} \quad s_{12} = \frac{Z_0}{R_{fs} + Z_0}.
\]

Notice that the gain of the amplifier depends only on the characteristic impedance and the value of the series feedback resistor, not on the device parameters. This means that flat gain over a frequency range can be obtained with feedback.

The physical meaning of the above equations is that the input VSWR can be unity with a positive value of the parallel feedback resistor, if the transconductance of the active device is large. This is usually not the case in a MESFET, but is the case in a bipolar transistor. For example, if we desire that the amplifier have \(|s_{21}|^2 = 10\, \text{dB}\), the minimal transconductance in a 50-ohm system and the value of the series feedback resistor are found by setting \(R_{fp} = 0\) (this is the case that was first discussed with just the series feedback):
\[ g_{m,\text{min}} = \frac{1 - s_{21}}{Z_0} = 83 \text{ mS} \quad \text{and} \quad R_{fp} = 208 \Omega. \]

This is a fairly large value of transconductance. Of course, the standard feedback relation \( R_{fs} = Z_0 (1 + |s_{21}|) \) is valid. When both series and shunt feedback resistors are used, and the transconductance is large enough, the best input and output match are obtained for \( R_{fs} R_{fp} \approx Z_0^2 \). This ignores the phase of \( s_{21} \), which can vary rapidly as the frequency increases and cause positive feedback through the resistor. This can be solved by adding an inductor in the series feedback branch with a value that the amount of feedback decreases after a certain frequency.

### L4.4. DISTRIBUTED AMPLIFIERS

A technique which achieves extremely broadband operation is the distributed amplifier, shown in Figure L4.6. The idea behind it is that, instead of trying to tune out the transistor capacitances, these capacitances are used as part of a lumped-element approximation to a transmission line. On the input side, inductors \( L_g \) are placed between the gate-to-source capacitances \( C_{gs} \) of the adjacent transistors, and in that way the familiar lumped-element artificial transmission line with a characteristic impedance of \( Z_g = \sqrt{L_g / C_{gs}} \) is formed. \( Z_g \) is nearly frequency independent. The phase velocity of a wave traveling along this line is \( v_g = 1 / \sqrt{L_g C_{gs}} \). This transmission line can be resistively terminated at the end with little loss of input signal. On the output side, inductors \( L_d \) are placed between drain-to-source capacitances of the adjacent devices, and a transmission line with a characteristic impedance \( Z_d = \sqrt{L_d / C_{ds}} \) and phase velocity \( v_d = 1 / \sqrt{L_d C_{ds}} \) is formed. This is an active transmission line and the signal builds up along it. The phases of the outputs of the individual transistors will only be appropriate for left-to-right propagation, so little power will be lost in the resistive termination at the left end of the line. In effect, the two transmission lines are coupled lines with a coupling coefficient greater than unity. The inductors \( L_g \) and \( L_d \) can be chosen to equalize the phase velocities of the two coupled lines. This discussion is valid only for a unilateral transistor approximation.

From this description of the distributed amplifiers, it appears that an arbitrarily large gain can be achieved by making a large number of sections. In the equivalent circuit for the transistor, however, there are some resistors as well, and this will make the transmission line lossy. As a result, a limited number of transistors can be added before the loss overcomes the gain. The frequency curve of the gain as a function of the number of sections is shown in Figure L4.7. It shows that after 5 sections, there is no appreciable increase in gain, whereas the flatness of the gain is reduced. Distributed amplifiers have been reported with flat gain from 1 to 40GHz, and into the 100-GHz range.
Distributed amplifiers are monolithically integrated so that the devices are very small compared to the guided wavelength. In practice, it is difficult to make good inductors in monolithic circuits (why?) at high microwave frequencies. Therefore, short sections of transmission lines are used instead between the stages of a distributed amplifier. Since in that case, the artificial transmission line model becomes even more of an approximation, these amplifiers are often viewed as traveling wave devices. The equivalent circuit of a traveling wave amplifier is shown in Figure L4.8.