AN562
APPLICATION NOTE

BALANCED UHF AMPLIFIERS

1. DESIGN.

425MHz, 600W pulsed power transistor amplifier with the following specifications: bandwidth (BW)=50MHz (450-400), passband flatness (ripple)=.05dB max, pulsewidth (P.W.)=20µsec, duty factor (DF)=15%, impedance (source and load)=50Ohms, and discrete or distributed circuitry (microstrip).

2. PROCEDURE.

First, one must select a transistor, then the Q requirement(s) must be determined. Then, the transistor's input and output imaginary terms, if any, must be reduced to zero at f_0. Next, the Chebyshev (low-pass) matching filters required for the discrete component circuit must be computed. Then the matching networks must be combined into a balanced configuration. And last, the distributed (microstrip) design must be plotted on the Smith Chart.

2.1. Step 1.

STMicroelectronics' SD1565 transistor has been selected for its broad bandwidth, low thermal resistance and ruggedness. The SD1565 datasheet provides the following:

Table 1: Class C Typical Performance at 425MHz

Z_{IN}	Z_{OUT}	P_{OUT}	P.W.	D.F.	$T_{j\ (max)}$	V_{CC}
1.7 + j 2.8	1.8 + j 0.5	600W	250µs	10%	150°C	40V

2.2. Step 2.

Circuit B.W. required to limit roll-off of power at band edges to 0.5dB is equal to 3x3dB bandwidth.

$$BW\ (3dB) = 450 - 400 \quad or \quad 50MHz$$

$$BW\ (0.5dB) = 3 \times 50 \quad or \quad 150MHz$$

$$f_0 = \sqrt{f_1 \times f_2} = \sqrt{400 \times 450} = 425MHz$$

$$Q_{max} = \frac{f_0}{BW0.5dB} = \frac{425}{150} = 2.83$$

Examine the selected device for the ability to meet Q_{max} thus:

$$Q_{zin} = \frac{X_L}{R} = \frac{2.8}{1.7} = 1.65$$

$$Q_{zout} = \frac{X_L}{R} = \frac{0.5}{1.8} = 0.28$$

Since Q_{max}=2.83>Q_{zin}>Q_{zout} the Q_{max} requirements are met.

2.3. Step 3.

The inductance presented to the input of the transistor is resonated out by shunting with a capacitor (C_Z). Whose value is:

$$C_Z = \frac{X_{in}}{\omega(R_{in}^2 + X_{in}^2)} = \frac{2.8}{(2\pi 425 \times 10^6)(1.7^2 + 2.8^2)} = 9810^{-12} = 98pF$$

The value of R'_{in} now that it is pure resistance is:

$$R'_{in} = \frac{R_{in}^2 + X_{in}^2}{R_{in}} = \frac{1.7^2 + 2.8^2}{1.7} = 6.3\Omega$$

Capacitor (C'_Z) required to resonate the inductance of the transistor's output is determined to be 54pF and R'_{OUT} is 1.9Ohms.

2.4. Step 4.

Chebyshev design tables require the input and output impedances to be real and:

$$fractional\ \ bandwidth \qquad W = \frac{2(f_2 - f_1)}{f_2 + f_1} = \frac{2(450 - 400)}{450 + 400} = 0.118$$

$$ratio \qquad r = \frac{R_{source}}{R'_{in}} = \frac{25}{6.3} = 3.97$$

2.4.1. Note.

Since this design is for a balanced configuration the underbalanced 50Ohms source will be transformed into a balanced line each side of which is 25Ohms.

Inspection of Chebyshev tables shows a single section (n=2) filter will exhibit less than a 0.04dB ripple.

Circuit Q requirements also influence our choice of transformation sections, thus maximum impedance transformation ratio (per section) is equal to:

$$Q^2 + 1 = (2.83)^2 + 1 = 9.0$$

The single section filter with a ratio of 3.97 meets.

Element value table for the above W and r values yield an interpolated g_1 value of 1.718.

$$g_2 = g_1 \div r = 1.718 \div 3.97 = 0.433$$

Component computations yield:

$$L_1 = g_1 R'_{in} \div \omega = (1.718 \times 6.3) \div 2\pi 425 \times 10^6 = 4.05 \times 10^{-9} or 4nH$$

$$C_1 = g_2 \div (\omega R'_{in}) = 0.433 \div (2\pi 425 \times 10^6 \times 6.3) = 2.57 \times 10^{-11} or 26pF$$

Compute the output matching filter using the same procedure.

$$W = 0.118$$

$$r = \frac{R_{load}}{R'_{out}} = \frac{25}{1.9} = 13.16$$

Inspection of the single section (n=2) filter table shows that the ripple would exceed 0.12dB, however, a two section (n=4) filter will limit the ripple to less than 0.005dB. The two section network will be employed. Note, as well, that our circuit Q requirement will not permit a single transformation ratio greater than 9.0.

Interpolated values of g_1 and g_2 are 1.744 and 0.552 respectively and values for g_3 and g_4 are:

$$g_3 = g_2 \times r = 0.552 \times 13.16 = 7.26$$

$$g_4 = g_1 \div r = 1.744 \div 13.16 = 0.132$$

Component computations yield:

$$L'_1 = (g_1 R'_{OUT}) \div \omega = (1.744 \times 1.9) \div (2\pi 425 \times 10^6) = 1.24 \times 10^{-9} or 1.2 nH$$

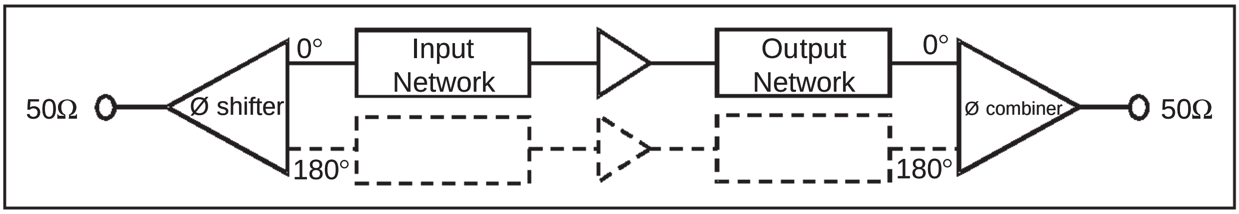
$$C'_1 = g_2 \div (\omega R'_{out}) = 0.552 \div (2\pi 425 \times 10^6 \times 1.9) = 1.09 \times 10^{-11} or 110 pF$$

$$L'_2 = (g_3 R'_{OUT}) \div \omega = (7.26 \times 1.9) \div (2\pi 425 \times 10^6) = 5.16 \times 10^{-9} or 5.2 nH$$

$$C'_2 = g_4 \div (\omega R'_{OUT}) = 0.132 \div (2\pi 425 \times 10^6 \times 1.9) = 2.6 \times 10^{-11} or 26 pF$$

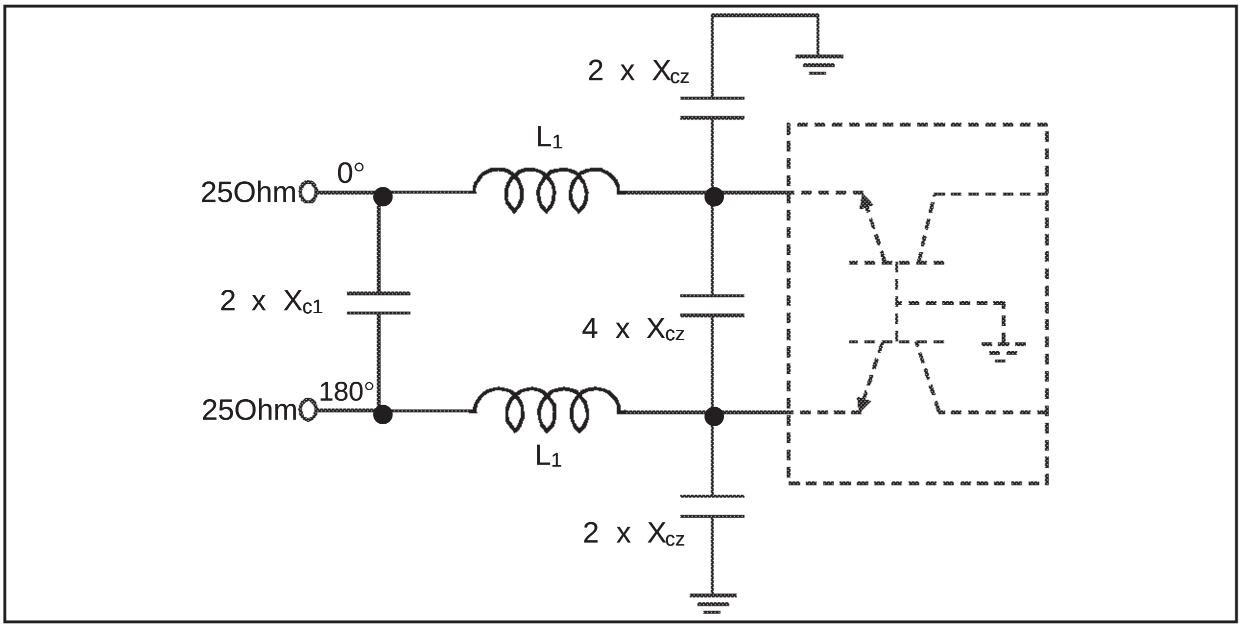
2.5. Step 5.

Figure 1: Basic Input and Output Layout



Balanced amplifiers may be envisioned as the combination of two single-ended circuits. Thus, the single-ended circuitry for the input developed above must be converted to a balanced circuit, as follows:

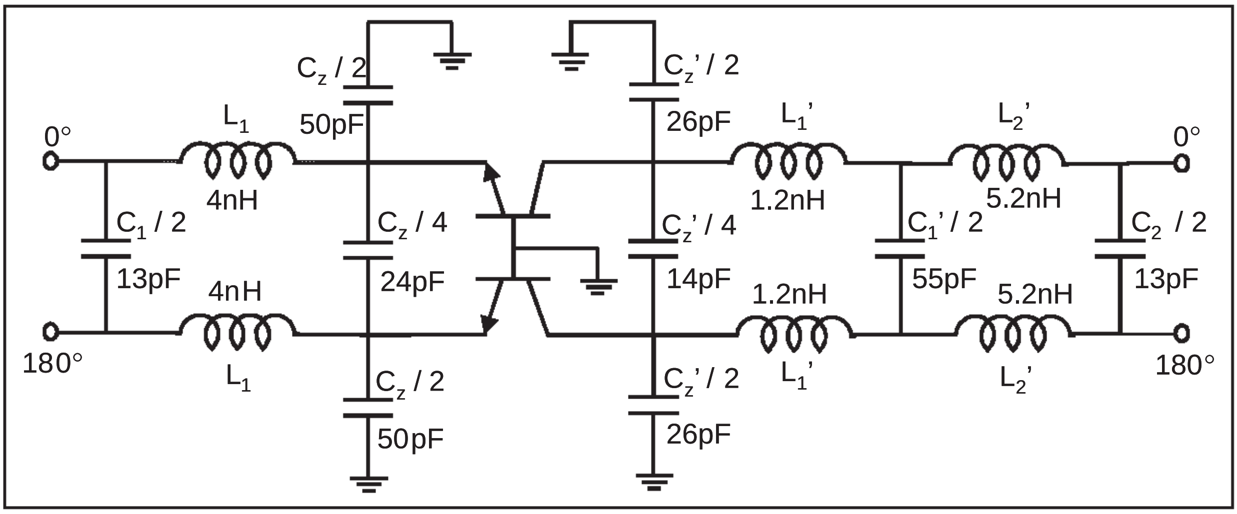
Figure 2: Input Matching Network



Note that since opposite sides of the circuit are 180° out of phase, the voltage between these points is twice that of either side and ground. Common capacitors must, therefore, be of twice the reactance of the single-ended circuit. Note, too, that the ground return for the common capacitor C_Z would have been limited to a path through the opposite transistor. To reduce this path length and insure balance in currents the reactance of the C_Z has been made four (4) times greater than the calculated capacitor and two additional capacitors have been introduced. The reactance of each added capacitor is equal to twice that of C_Z.

The single-ended output circuitry is converted to balance in the same manner, to yield the complete amplifier, less biasing and signal splitting circuitry, shown below.

Figure 3: Complete Input and Output Matching Network / 400-450MHz, 600W Pulsed Amplifier



2.6. Step 6.

Equipped with the idealized discrete component values the next step in the development of the distributed circuit is to determine the line widths of the transmission lines. The circuit board material is selected to produce a line width compatible with the transistor's leads, i.e. sufficiently wide to permit soldering, while limiting discontinuity at the interface.

The characteristic impedance of microstrip is given as:

$$Z_0 = \frac{377h}{\varepsilon_r^{1/2} W[1 + 1.735\varepsilon_r^{-0.724}(w/h)^{-0.836}]}$$

Where: Width, height, ε_r=dielectric constant

For a ceramic filled Teflon-glass board 1/32 inch thick having a dielectric constant of 6.0, a line width of 0.13 inch will yield a Z_0 of 25Ohms. This width is compatible with the transistor's leads and will be used in this design.

Z_0 strongly influences the length of the inductors and thus the amplifier's physical size. An inductor's length is approximated as:

$$l \cong \frac{11.8L}{Z_0\sqrt{\varepsilon_r}}$$

where l=length in inches, L=inductance in nH, and ε_r=dielectric constant.

This amplifier requires inductors of 1 to 5nH. The length of the longest (5.2nH) fabricated from 25Ohms transmission line is about:

$$l \cong \frac{11.8 \times 5.2}{25\sqrt{6.0}} = 1.00 inch$$

Before converting to Microstrip, the capacitors should also be examined. The capacity of 1 square inch of the selected board material is equal to:

$$C_{(pF)} = \frac{0.224 \times 1.09\varepsilon_r \times area}{h(thickness)} = \frac{0.224 \times 1.09 \times 6.0 \times 1^2}{0.03125} = 47pF$$

Distributed capacitors in the order of 24pF, as required by this amplifier, would occupy about one-half square inch of board area and will be employed where practicable.

Smith Chart mapping requires tabulating the circuit reactances and susceptances thus:

Table 2: Circuit Reactance and Susceptances

Symbol	Value	Reactance (Ohm)	Susceptance	Normalized
L_1	4.05nH	10.8	-------------- ÷25	0.433
C_1	25.7pF	14.6	68.5mmho ÷40	1.7
L_1'	1.24nH	3.3	-------------- ÷25	0.132
C_1'	109pF	3.4	291mmho ÷40	7.28
L_2'	5.2nH	13.9	--------------- ÷25	0.555
C_2'	26pF	14.6	69.4mmho ÷40	1.735

Input matching network - Smith Chart #1. Locate the normalized SD1565 input R'_{in} 6.3/25 or 0.252 at point A. Rotate point A to B, the normalized reactance of L_1 0.433. Transfer point B to the admittance side of the chart point C, by constructing a line through the center of the chart and equidistant to the center. The length of L_1 is determined by extending the constructed line through point C to the chart periphery.

Length L_1 is read on the wavelength scale as: 0.318 less 0.25 or 0.068λ.

2.6.1. Note.

Point C lies on the constant conductance circle 1.0 at the intercept of 1.7mmho susceptance, the previously tabulated value of C_1. The shunt susceptance of C_1 brings the transformed point C along conductance circle 1.0 to the real axis at the chart's center.

This graphical presentation confirms the validity of our computed values for L_1 and C_1.

Smith Chart #2 illustrates the transformation(s) of the output matching network and is similar in construction.

Length L'_1 is found to be 0.0215λ and L'_2 is 0.089λ in length.

The wavelength in Microstrip is given as: $\lambda_g = \lambda_o \times V_P$

Where λ_o=wavelength in air (cm) and V_P=velocity of propagation.

$$V_P = \frac{1}{\sqrt{\varepsilon_r'}}$$

and

$$\varepsilon_r' = \frac{\varepsilon_r + 1}{2} + \frac{\varepsilon_r - 1}{2}\left(1 + \frac{10}{W/h}\right)^{-\frac{1}{2}}$$

The guided wavelength of the selected board material at 425MHz for 25 Ohms transmission line is:

$$\varepsilon_r' = \frac{7.0}{2} + \frac{5.0}{2}\left(1 + \frac{10}{4.16}\right)^{-\frac{1}{2}} = 3.5 + 2.5 \times 0.542 = 4.855$$

$$V_P = \frac{1}{\sqrt{4.855}} = 0.454$$

$$\lambda_g = 70.59 \times 0.454 = 32.05 cm \qquad or \qquad (32.05 cm) \div 2.54 = 12.62 inches$$

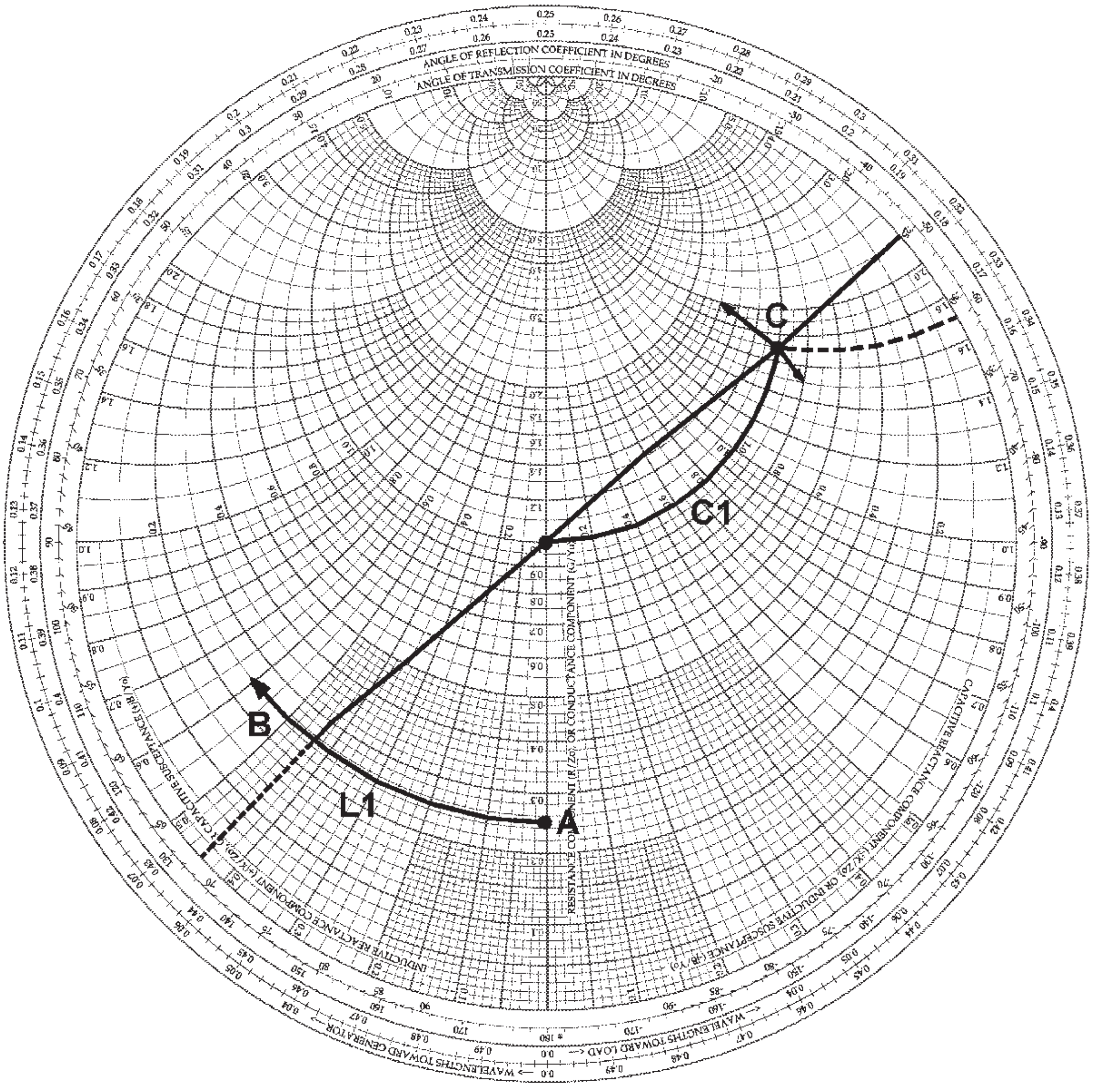
Tabulating the distributed inductors gives:

Table 3: Distributed Inductors

Symbol	Impedance (Ohms)	Length (λ)	Length (inches)
L_1	25	0.068	0.85
L'_1	25	0.0215	0.27
L'_2	25	0.089	1.12

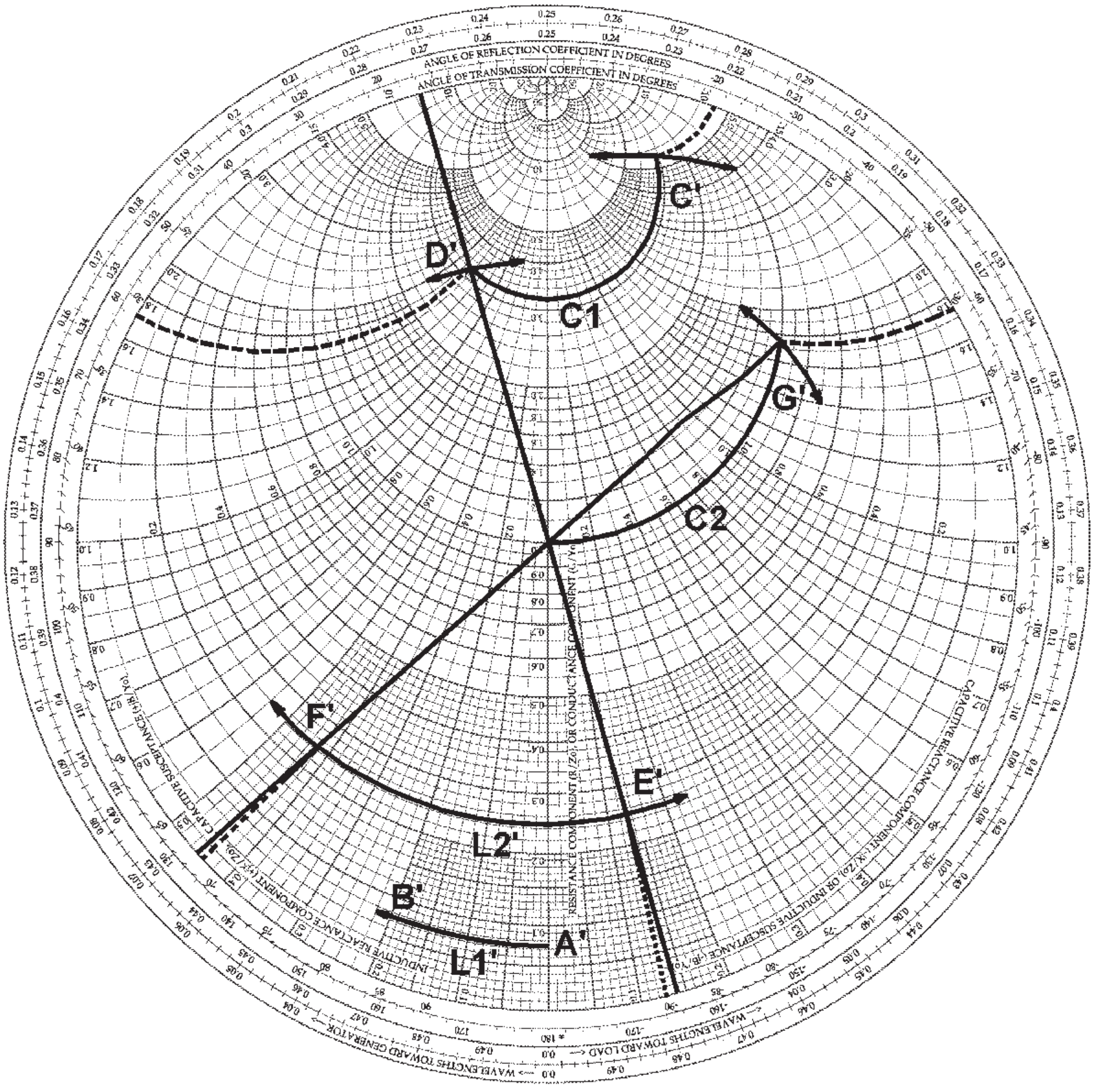
Impedance or admittance coordinates normalized to 25Ohms/40mmho.

Figure 4: Input network for balanced 425MHz amplifier



Impedance or admittance coordinates normalized at 45Ohm/40mmho

Figure 5: Output Network for Balanced 425MHz Amplifier



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