AN1033

Match Impedances in Microwave Amplifiers

and you're on the way to successful solid-state designs.
Here's how to analyze input/output factors and to create a practical design.

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The key to successful solid-state microwave power-amplifier design is impedance matching.

In any high-frequency power-amplifier design, improper impedance matching will degrade stability and reduce circuit efficiency. At microwave frequencies, this consideration is even more critical, since the transistor's bond-wire inductance and base-to-collector capacitance become significant elements in input/output impedance network design.

In selecting a suitable transistor, therefore, keep in mind that the input and output impedances are critical along with power output, gain and efficiency.

Unless the selected transistor is used at frequencies that are much lower than the maximum operating frequency, the input impedance is largely inductive with a small real part. The large inductance is due to bond wires that connect the transistor chip to the input lead of the package and to the common-element bond wires. The small real part of the input impedance is due to the large geometries required to generate high power at high frequencies; the base bulk resistance may be the predominant part of the real input impedance.

Use microstrip stubs at input network

The first and most important step in designing the input matching network for the selected device is to provide a shunt capacitance that will resonate the inductive component of the input impedance. This step forms the low-pass matching section of the network and should provide the smallest possible transformed impedance. To minimize the inductive component, the input and common-element lead lengths must be kept short.

The resonating capacitance is generally best provided by a microstrip stub. In some cases the stub producing the required capacitance is so large that a practical circuit size cannot be realized. It is best then to distribute as much of this capacitance as is physically practical and to provide the balance with high-quality chip capacitors.

The first section of the impedance matching network is extremely important because it can degrade the stability of the amplifier if it is not well designed. Depending on the design frequency of the amplifier and the transistor selected, the resonated real impedance can range from less than 50 Ω to much higher. When it is below 50 Ω, an additional low-pass matching section can be conveniently added to achieve the required 50-Ω impedance at the input.

The higher-impedance case presents a special problem if microstrip techniques are used to build the matching network. The problem occurs because the resonated impedance may be as high as 300 Ω. Reducing this to 50 Ω by use of a lowpass network configuration requires a series transmission line that will behave as an inductor. The rule of thumb is that the characteristic impedance of the transmission line must be at least twice the higher impedance before such behavior results. Examination of the accompanying table shows that characteristic impedance line of greater than 100 Ω are very narrow. Narrow transmission lines (less than 0.01-inch wide) should be avoided wherever possible, because repeatability of width dimensions is poor. Also, the loss in a narrow line may become excessive. A better solution is to use a quarter-wave transmission line transformer with a characteristic impedance equal to the square root of the 50-Ω impedance product: $Z_o = \sqrt{50\ Z_R}$.

Make output bandwidth wider than input

The output impedance of a microwave power transistor is usually defined as the conjugate of the load impedance required to achieve the device performance. A typical output equivalent circuit is shown in Fig. 1. The capacitance C_{out} is nearly equal to the collector-base capacitance C_{ob} specified for the selected transistor. L_c is the inductance of the bond wires used to bridge from the collector metallization area to the package output lead, and L_{com} represents the inductive effects of the common element bond wires.

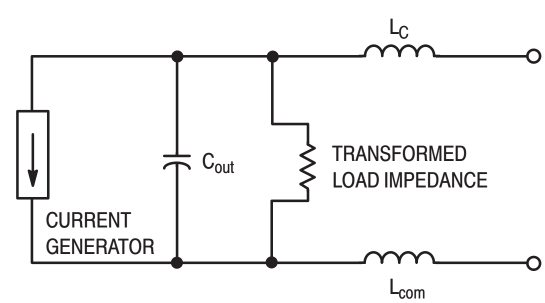

Figure 1. In this output equivalent circuit, capacitance C_{out} is almost equal to the selected transistor's collector-to-base capacitance C_{ob}.

For correct operation of the transistor, the ultimate load impedance must be transformed to a real impedance across the current generator. This real impedance is determined by

$$R_L = \frac{[V_{cc} - V_{ce(sat)}]^2}{2P_{out}}.$$

The load impedance presented to the package terminals will contain the real impedance at the current generator, transformed to a lower value by the low-pass L section formed by C_{out} and the parasitic inductances L_c and L_{com}. Usually, the reactive part of the load impedance is made inductive to tune out the residual capacitance of the device.

The output matching network should be designed so it has greater bandwidth than the input matching network. Providing a good collector match, both above and below the design frequency, ensures that the input power will be reflected before the collector VSWR rises to values that endanger the transistor. In this way the transistor is protected from off-frequency operation. The amount of additional bandwidth required for protection of the transistor depends on the ruggedness of the transistor used. The manufacturer's specifications for VSWR tolerance and input Q can be a guide for determining the bandwidth requirements of the input matching network.

One technique for obtaining the required bandwidth is to resonate a portion of the capacitive reactance of the transistor output impedance with a shunt inductor. The shunt inductor can also be used to feed the collector supply voltage to the transistor. Additional transformation may be obtained from a low-pass matching section. By adjusting the amount of shunt inductance and rematching with the low-pass section, the designer can create a truly broadband output match.

Don't overlook base and collector paths

In addition to matching the device impedances, direct-current paths must be provided to the base and collector of the transistor. The collector path is provided by the shorted stub in the impedance-matching network. The base path requires the addition of a choke from the base to ground. The choke can be a lumped element or a distributed shorted stub of sufficient impedance to be negligible in the circuit. A quarter-wavelength stub is ideal. The narrowest practical line should be selected. In addition a dc blocking capacitor is required in the collector circuit. Also needed is a bypass capacitor to provide the proper ac shorting point for the inductive stub in the collector-matching network.

Selection of a blocking capacitor is relatively straightforward. The capacitor should be chosen to provide low loss at the operating frequency while maintaining the capacitance at a value that inhibits low-frequency oscillation. The latter is caused by the series capacitor's tendency to display rising reactance with decreasing frequency.

Blocking capacitors must be large enough to preserve coupling characteristics down to a frequency where the shunt-feed chokes can effectively short the respective port to ground. Coupling capacitors should not be excessively large, or they may produce as much as 1-dB loss in gain with a corresponding decrease in efficiency in the case of collector coupling capacitors. The Q of the coupling capacitor determines the acceptable range of capacitance values and is generally inversely related to capacitance.

Bypass capacitors are selected by analysis of the same considerations as those for blocking capacitors. A large bypass capacitor (tantalum or electrolytic), placed from the dc feedpoint to ground, prevents tendencies toward low-frequency oscillation in the circuit. Also, it may be necessary to add smaller bypass capacitors to preserve stability over a wide range of frequencies.

Adjust for bandwidth and physical dimensions

The circuit design may be adjusted quickly for bandwidth requirements through use of a computer optimization program such as Magic, offered by University Computing of Dallas, Tex. When that step is finished, electrical dimensions must be converted to physical dimensions.

At this point in the design sequence, the dielectric material must be chosen. Three commonly used materials are Teflon fiberglass, epoxy fiberglass and alumina. Above 500 MHz, epoxy fiberglass exhibits too many losses to be a good choice. Teflon fiberglass can be used up to several gigahertz; it has reasonable dielectric losses and is easy to process. Alumina, a ceramic, offers a high dielectric constant, good dimensional consistency and small circuit geometry.

When plastic materials are used, it's a good practice to measure the material thickness and dielectric constant, because variations are common. In a recent test the dielectric constant of a sheet of epoxy fiberglass material was measured at 4.55 at 1 MHz and 4.25 at 500 MHz. If the manufacturer's value of 5.5 had been used for the design of matching networks, considerable error would have resulted.

The physical dimensions of the matching circuitry may be calculated from the data in the table. The line lengths are scaled by the velocity factor, which is equal to $Z_o/Z_{o(air)}$ in air for a constant width-to-height ratio, W/H.

The final design of a typical breadboard microwave amplifier is shown in Fig. 2. The ground areas on the top of the board are connected to the microstrip ground plane by 2-mil-thick foil wrapped around the edges of the board and the areas directly under the emitter leads of the transistor. The foil is secured to the top and bottom surfaces with solder. Plating may be used for production units. The entire board can be soldered to a metal plate to allow connector mounting and to provide a thermal path for the heat generated by the transistor.

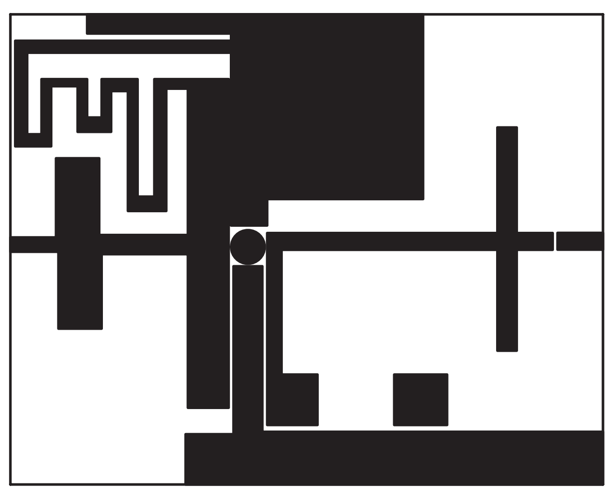

Figure 2. With this typical microwave amplifier breadboard layout, the entire board can be soldered to a metal plate to provide a path for thermal cooling.

Microstrip Z_o and Velocity Factor vs Width-to-Height (W/H) Ratio
(Prepared by Don Schulz, Applications Engineer)

W/H	Air K = 1.0		Teflon K = 2.55		Epoxy K = 4.25		Alumina K = 9.6	
	Z_o	V_P	Z_o	V_P	Z_o	V_P	Z_o	V_P
0.630	168.425	1.000	110.683	0.657	87.986	0.522	60.977	0.362
0.695	161.878	1.000	106.258	0.656	84.414	0.521	58.441	0.361
0.766	155.370	1.000	101.865	0.656	80.870	0.521	55.927	0.360
0.844	148.909	1.000	97.509	0.655	77.360	0.520	53.440	0.359
0.931	142.506	1.000	93.199	0.654	73.888	0.518	50.985	0.358
1.026	136.171	1.000	88.941	0.653	70.463	0.517	48.566	0.357
1.131	129.916	1.000	84.745	0.652	67.090	0.516	46.187	0.356
1.247	123.753	1.000	80.616	0.651	63.775	0.515	43.853	0.354
1.375	117.692	1.000	76.565	0.651	60.524	0.514	41.568	0.353
1.516	111.746	1.000	72.597	0.650	57.345	0.513	39.337	0.352
1.672	105.926	1.000	68.721	0.649	54.243	0.512	37.164	0.351
1.843	100.242	1.000	64.944	0.648	51.223	0.511	35.053	0.350
2.032	94.706	1.000	61.273	0.647	48.291	0.510	33.007	0.349
2.240	89.327	1.000	57.714	0.646	45.451	0.509	31.030	0.347
2.470	84.115	1.000	54.271	0.645	42.709	0.508	29.123	0.346
2.723	79.076	1.000	50.951	0.644	40.066	0.507	27.289	0.345
3.002	74.218	1.000	47.757	0.643	37.527	0.506	25.531	0.344
3.310	69.546	1.000	44.692	0.643	35.094	0.505	23.849	0.343
3.649	65.065	1.000	41.759	0.642	32.768	0.504	22.244	0.342
4.023	60.779	1.000	38.959	0.641	30.550	0.503	20.716	0.341
4.435	56.689	1.000	36.292	0.640	28.440	0.502	19.266	0.340
4.890	52.796	1.000	33.760	0.639	26.439	0.501	17.892	0.339
5.391	49.100	1.000	31.360	0.639	24.544	0.500	16.594	0.338
5.944	45.600	1.000	29.091	0.638	22.755	0.499	15.370	0.337
6.553	42.291	1.000	26.952	0.637	21.069	0.498	14.218	0.336
7.224	39.173	1.000	24.938	0.637	19.485	0.497	13.138	0.335
7.965	36.233	1.000	23.047	0.636	17.998	0.497	12.125	0.335
8.781	33.484	1.000	21.275	0.635	16.606	0.496	11.179	0.334
9.681	30.904	1.000	19.618	0 635	15.305	0.495	10.295	0.333
10.674	28.491	1.000	18.071	0.634	14.091	0.495	9.472	0.332
11.768	26.240	1.000	16.629	0.634	12.961	0.494	8.707	0.332
12.974	24.143	1.000	15.288	0.633	11.911	0.493	7.996	0.331
14.304	22.192	1.000	14.043	0.633	10.937	0.493	7.338	0.331
15.770	20.381	1.000	12.888	0.632	10.033	0.492	6.728	0.330
17.387	18.702	1.000	11.818	0.632	9.198	0.492	6.164	0.330
19.169	17.148	1.000	10.830	0.632	8.425	0.491	5.644	0.329
21.133	15.172	1.000	9.917	0.631	7.713	0.491	5.164	0.329
23.300	14.385	1.000	9.074	0.631	7.056	0.490	4.722	0.328
25.688	13.162	1.000	8.299	0.630	6.451	0.490	4.315	0.328
28.321	12.036	1.000	7.585	0.630	5.894	0.490	3.942	0.327
31.224	10.999	1.000	6.929	0.630	5.383	0.489	3.598	0.327
34.424	10.047	1.000	6.326	0.630	4.914	0.489	3.284	0.327
37.953	9.172	1.000	5.773	0.629	4.483	0.489	2.995	0.327
41.843	8.370	1.000	5.266	0.629	4.089	0.489	2.731	0.326
46.132	7.634	1.000	4.801	0.629	3.727	0.488	2.489	0.326
50.860	6.960	1.000	4.376	0.629	3.397	0.488	2.267	0.326
56.073	6.343	1.000	3.987	0.629	3.094	0.488	2.065	0.326
61.821	5.779	1.000	3.632	0.628	2.818	0.488	1.880	0.325
68.157	5.264	1.000	3.307	0.628	2.566	0.487	1.711	0.325
75.144	4.792	1.000	3.010	0.628	2.335	0.487	1.557	0.325
82.846	4.362	1.000	2.739	0.628	2.125	0.487	1.417	0.325
91.337	3.969	1.000	2.492	0.628	1.933	0.487	1.289	0.325
100.700	3.611	1.000	2.267	0.628	1.758	0.487	1.172	0.324

The initial tune-up of the amplifier matching circuits can be expedited by use of a network analyzer and a precision load on the input or output connector. The circuit can be adjusted to match the nominal impedances supplied by the transistor manufacturer. Distributed stubs are purposely made longer than necessary and are adjusted to the correct length by trimming of the foil on the capacitive stubs. The inductive stub in the output network is adjusted by positioning of the bypass capacitor along the stub and the adjacent ground plane.

This procedure results in a load line that is fairly close to optimum. A transistor can now be inserted in the circuit and the collector matching network readjusted for maximum collector efficiency. Stub tuners are used to match the amplifier input impedance, so that only one variable at a time need be considered. Initially it may be necessary to operate the transistor at reduced collector voltage and power output to avoid excessive stress. When maximum efficiency is obtained, the stub tuner is removed and the input network adjusted for minimum input VSWR.

Now let's design an impedance-matching circuit

Let's consider a practical example of a procedure for the design of impedance-matching circuitry. The sample circuit uses a TRW 2N5596 at 700 MHz as the active device.

Specifications for the completed amplifier are:

Z_{in} = 50 Ω,
Z_{out} = 50 Ω,
P_{out} = 20 W,
G_p = 7 dB,
η = 55% minimum.

Specifications for the TRW 2N5596 are:

P_{out} = 20 W at 1 GHz,
η = 55% minimum at 1 GHz,
G_p = 5 dB minimum at 1 GHz,
Z_{in} = 2.5 + J4.0 at 700 MHz,
Z_{out} = 6.0 – J12.5 at 700 MHz.

In practice, the gain of a common-emitter amplifier decreases at a rate of 4 to 5 dB per octave. The 2N5596 at 700 MHz produces about 7 dB of gain. Therefore approximately 4 W of drive will be required to produce 20 W of output power. The collector efficiency can be expected to increase at the lower frequency, but it is difficult to estimate because it is a complex phenomenon. Manufacturers' curves of typical behavior are useful. Output power will not increase significantly with the decreased frequency.

The efficiency-frequency relationship depends on device f_T and ballasting. Heavily ballasted transistors tend to give increased efficiency as frequency is decreased. However, they level out at a lower efficiency than a nonballasted part because of I^2R losses in ballast resistors. The average increase in efficiency as a result of decreasing frequency is about 20% per octave. Values from 10 to 40% per octave have been measured.

The initial phase of the design is best accomplished on an immittance chart. The chart with appropriate values indicated for the sample design is shown in Fig. 3. The input match is achieved when the input impedance is resonated with a capacitive susceptance of 0.18 mhos. This susceptance is realized by use of a pair of capacitive microstrip stubs. Each stub must exhibit a reactance of 2 x 1/0.18 mhos, or 11.1 Ω. The length of the stub may be calculated by

$$\tan \Theta = \frac{Z_o}{X_c} \quad .$$

For ease of adjustment, the length of the stubs should be less than 60 degrees. Because capacitive reactance is a tangential function, the reactive variations per unit length become increasingly severe past 60 degrees. It is better to decrease Z_o rather than to use longer stubs to achieve higher capacitance. Therefore $Z_o \leqq 1.732\, X_c \leqq 19.24\ \Omega$. Because it is easier to shorten a microstrip stub than to lengthen it, the Z_o of 15 Ω, for example, provides sufficient adjustment range to accommodate device variations.

The next step is to transform the resonated impedance to 50 Ω. This is accomplished by a series-transmission line with a characteristic impedance of 50 Ω. From Fig. 3, we see that the length of this line can be directly determined to be 0.062 wavelengths, or 22.3 degrees, long. A capacitive susceptance of 0.040 mhos completes the transformation. Again, a pair of capacitive stubs will provide the susceptance. For ease of converting the design to microstrip dimensions, it is convenient to choose a Z_o for the second stub that is equal to that selected for the first.

Therefore:

$$\tan \Theta = \frac{Z_o}{X_c} = \frac{15}{50} = 0.3,$$

$$\text{or } \Theta = 16.7 \text{ degrees.}$$

In this case the length chosen is 20 degrees to allow for some adjustment.

The output match is achieved by partial resonating of the device's output impedance with an inductive susceptance. While the amount of susceptance chosen is arbitrary at this point, the output network bandwidth is affected by the value. From Fig. 3, we can determine that 0.05 mhos is required for the first matching element. This susceptance is achieved by use of a shorted microstrip stub. The length of the stub may be calculated from the equation

$$\tan \Theta = \frac{X_L}{Z_o} \quad .$$

If Z_o of the stub is arbitrarily chosen to be 50 Ω,

$$\tan \Theta = \frac{20}{50} = 0.4,$$

$$\Theta = 21.8 \text{ degrees.}$$

Again, the stub is made somewhat longer because it can be adjusted by sliding the chip capacitor (ac short) up or down the line length. The remaining transformation is achieved by a 50-Ω series-transmission line of 0.15 wavelengths (54 degrees long) and a capacitive susceptance of 0.014 mhos. Selecting a pair of 50-ohm microstrip lines to provide the susceptance requires a stub length of

$$X_c = 2 \times \frac{1}{0.014} = 143\ \Omega\ .$$

$$\tan = \frac{Z_o}{X_c} = \frac{50}{143} = 0.350 = 19.3 \text{ degrees.}$$

A stub length of 25 degrees will provide an adequate allowance for adjustment of the circuit.

CHART NOT AVAILABLE ELECTRONICALLY

Figure 3. The Smith chart, with values specified for the design example, indicates the necessary inductive and capacitive stubs. Impedance transformations are achieved by 50-Ω series-transmission lines.

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