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800 MHz TEST FIXTURE DESIGN

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Although this article presents techniques for the general case of UHF-800 MHz circuit design, the emphasis is placed specifically on test fixture design for 800 MHz. Text fixtures tend to be the last consideration for most RF power amplifier development programs, yet they are the most valuable tool available for measuring and maintaining device consistency. Minimum power gain, collector efficiency and broadband performance requirements, though they are always detailed in some form of written specification, are meaningless unless they are demonstrated and controlled by a test fixture. A good test fixture will assure correlation between the customer and vendor and function as a trouble shooting tool in the event of radio problems. When alternate sources are pursued for a stage, test fixtures can shorten qualification cycles. But the prevention of gradual shifts in RF performance over the lifetime of a product is the major purpose of a test fixture.

Motorola has recognized the importance for good test fixtures and has established general guidelines for their implementation.

Each hi-tech product is tied to a well defined test fixture, which has the following general specifications:

- Broadband performance, demonstrating typical characteristics throughout the band. (Ex.: UHF; 450 – 512 MHz, 800; 800 – 870 MHz)
- A 3″ x 5″ mechanical format, which is rugged for high volume test applications.
- Simple RF match construction to represent realistic radio performance.
- Devices must meet all minimum test requirements at the specified test frequency. UHF: 470 MHz, 800: 870 MHz.

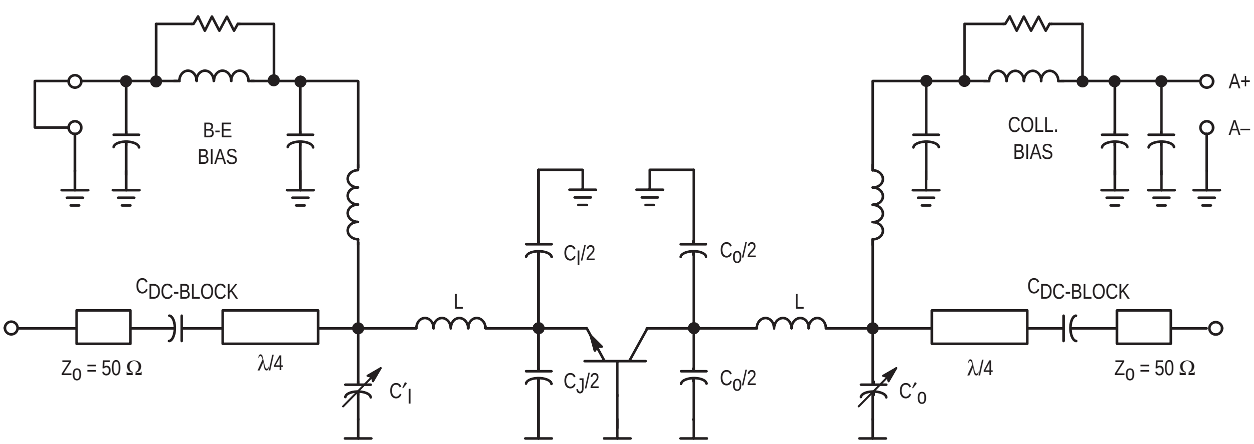
The repeatability, mechanical ruggedness and broadband performance are all very important factors needing consideration in the design of test fixtures. The remainder of this article goes into detail, using the MRF846 as an example.

The schematic representation of the fixture outlined in this article is shown below (Figure 1).

C_I and C_O represent the shunt capacitors at the input and output (respectively) which cancel most of the inductive reactance associated with the transistor's input and output impedance. Mini clamped-mica capacitors are used for these components and are physically located beneath the common lead wear blocks. Inductance "L" is introduced by the input (and output) wear blocks. Because of this parasitic inductance, L, trimmer capacitors (C'_I or C'_O) are required to transform the now reactive impedance back to real before launching off into the $\lambda/4$ transmission lines.

The transistor's input and output impedance can be represented as a combined series resistor and inductor as shown in Figure 2.

This series combination can be transformed into a parallel equivalent by using the equations shown in Figure 3. The capacitors C_I and C_O are selected by calculating the value necessary to form a parallel resonance with X_p. Since all capacitors have a finite, series lead inductance, the capacitor is actually considered as a simple series resonant circuit. The resulting effect is the capacitance is always higher than the marked value and goes through resonance at some frequency. Mini clamped-mica capacitors are recommended for test fixture design due to the very low parasitic inductance



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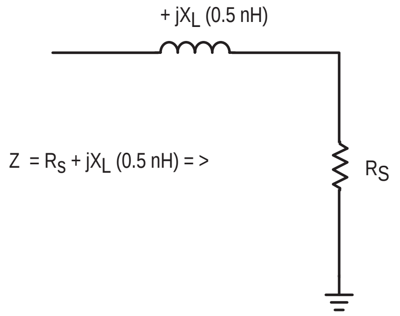

Figure 2. Equivalent Circuit for Z_{in} or Z_{out}

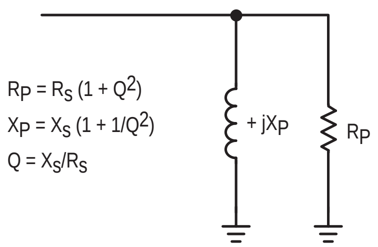

Figure 3. Parallel Equivalent Circuit

associated with them which increases the usable range of capacitances. (They are also extremely high “Q”). A typical measured series inductance for clamped-mica capacitors is about 0.5 nH. The equivalent capacitance is calculated by subtracting the series lead inductance from the capacitive reactance, or $X_C(\text{equiv}) = X_C - X_{L(0.5\ nH)}$.

Since two capacitors are used in parallel, the total capacitance is derived as shown in Figure 4.

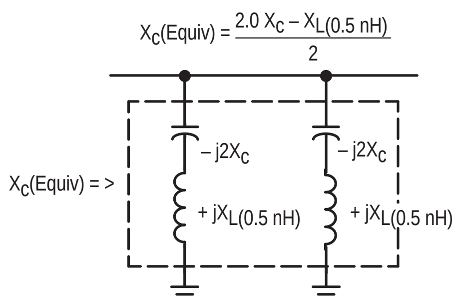

Figure 4. Equivalent Reactance for Capacitors in Parallel

A value of 2.0 X_C is used in the above example since each capacitor will contribute only 1/2 to the total capacitance. By setting X_C(equiv.) equal to the parallel equivalent reactance calculated in Figure 3, the exact capacitor values may be determined.

$$X_p = \frac{2.0\ X_C - X_{L(0.5\ nH)}}{2}$$

$$X_C = \frac{2.0\ X_p + X_{L(0.5\ nH)}}{2} \qquad (X_C = 1/2\ fC)$$

$$C = \frac{1}{\pi f\ (2.0\ X_p + X_{L(0.5\ nH)})}$$

Introducing an actual example at this time should help in explaining the remaining steps involved in a test fixture design. The MRF846 is a 40 W, 12.5 V, 800 MHz device whose input and output impedances are:

Table 1. Z_{in}, Z_{out} for MRF846

Frequency	Z_{in}	Z_{out}
800 MHz	1.1 + j4.8	1.20 + j2.4
836 MHz	1.0 + j4.9	1.15 + j2.5
870 MHz	1.0 + j5.0	1.10 + j2.7
900 MHz	0.9 + j5.1	1.10 + j2.8

Since X_p will vary as a function of frequency, C_I and C_O need only be calculated for one point within the frequency band. Typically, the input response of an RF power transistor is optimized about the center of the band. Hence, the input R_p and X_p are generally calculated at this frequency [$(f_h + f_l)/2$].

The output response is different. If C_O were selected for a resonance to occur with X_p at band-center, an unacceptable performance roll-off would be seen at the upper end of the frequency band. Overall performance is best when C_O is calculated at a frequency within 20% of the upper end of the band. Since device gain increases as frequency decreases, the performance at lower frequencies is generally no problem.

Using the MRF846 as an example, input and output capacitor values may be determined as follows:

INPUT:

Frequency = 836 MHz

Z_{in} = 1 + j4.9
= 4.9 Q = 4.9/1

$$X_p = 4.9\left(1 + \frac{1}{(4.9)^2}\right)$$
$$= 5.1\ \Omega$$

$$X_{L(0.5\ nH)} = 2.0\ \pi\ (836 \times 10^6)\ (0.5 \times 10^{-9})$$
$$= 2.63\ \Omega$$

$$C = 1/[\pi(836 \times 10^6)\ (2 \times 5.1 + 2.63)] = 29.7\ pF$$

2 – 15 pF Capacitors would be the best choice.

OUTPUT:

Frequency = 870 MHz

Z_O = 1.1 + j2.7; Q = 2.7/1.1
= 2.45

$$X_p = 2.7\left(1 + \frac{1}{(2.45)^2}\right)$$
$$= 3.15\ \Omega$$

$$X(0.5\ nH) = 2.0\ \pi\ (870 \times 10^6)\ (0.5 \times 10^{-9})$$
$$= 2.7\ \Omega$$

$$C = 1/[\pi(870 \times 10^6)\ (2 \times 3.15 + 2.7)] = 40.7\ pF$$

2 – 20 pF Capacitor would be the best choice.

(20 pF Capacitors were not available, so an 18 pF and a 24 pF capacitor were chosen instead. The total C = 42 pF.)

Though the MRF846 test fixture used at Motorola does use these capacitor values, the above calculations may act only as a good starting point. Empirical measurements and more precise impedance measurements for a given application may result in minor deviations from these values.

Assuming no additional circuit parasitics had to be accounted for, the quarter wave transmission line sections could now be determined. The input (and output) fixture wear blocks do, however, contribute additional series lead inductance to the impedances. These inductances are counteracted by the trim capacitors C'_I and C'_O. The wear block inductance could be calculated and then used to determine the proper capacitance values. However, since there are other, less obvious frequency and grounding effects which may influence the impedance transformation, it is a more practical (and generally a more accurate) procedure to measure the impedance which will be transformed by the transmission line to 50 Ω.

The capacitors C_I and C_O should be mounted into the test fixture and a known characteristic impedance transmission line soldered into place as shown below in Figure 5.

Triple stub tuners are used on the input and output to tune for maximum output power and minimum reflected power at various frequencies throughout the band. Band edges and band center are generally adequate for a good circuit design. Due to higher impedance levels produced by adding C_I and C_O, (Z_{in}) and (Z_{out}) are measured instead of the real transistor impedances, Z_{in} and Z_{out}. Also, by measuring impedances in the actual applications fixture, the design can be optimized for the particular fixture. Perhaps a maximum gain tuning point is not what is desired. Obtaining impedances for an efficiency/gain compromise may be more desirable. If this is the case, an impedance table for the appropriate conditions may be obtained. It is then for these impedances that C_I, C_O and Z_o will be calculated.

The procedure for obtaining the impedances is simple and requires a vector voltmeter (VVM) or a network analyzer. Both are used at Motorola, but a vector voltmeter is less expensive and if used with a high directivity directional coupler, (>40 dB), is very accurate. The set-up is constructed as shown in Figure 5. With frequency set, stub tuners are adjusted for the desired performance. Again, using the MRF846 as an example, numbers shown in Table 2 were measured for P_{in} = 12.0 W, V_{CC} = 12.5 V.

Table 2. Performance of MRF846 versus Frequency

806 MHz	838 MHz	870 MHz
P_{out} = 50.0 W	P_{out} = 48.3 W	P_{out} = 44 W
Eff. = 53.3%	Eff. = 55.2%	Eff. = 58%
Prefl. = 0 W	Prefl. = 0 W	Prefl. = 0 W

The output stub tuners were adjusted for maximum gain at each frequency and the input stub tuners were adjusted for zero watts reflected power. After each measurement, the impedance presented to the fixture by the triple stub tuner and load (or source) combination is measured by the vector voltmeter. The impedance is then translated by the transmission line used in the test fixture to obtain (Z_{in}) and (Z_{out}). In the above example a 26 Ω, 0.309λ (@ 836 MHz) transmission line was arbitrarily chosen to be in the MRF846 measurements. By using the equation: $Z\angle\theta = R_o\,[(1 + \Gamma\angle\theta) / (1 - \Gamma\angle\theta)]$ or various computer or calculator programs, the transformation is easily calculated. The most important part of the whole procedure is obtaining an accurate measurement from the stub tuners. Prior to making any measurements, the vector voltmeter must be referenced to a short (180° on a Smith Chart). As a means of accounting for the errors introduced by the connectors at the fixture's input and output, that same connector is used for a referencing short as shown in Figure 6.

The measurement reference plane is now the edge of the connector used on the test fixture, which is also the beginning of the transmission line. Assuming the same reference plane is maintained during the measurements, an accurate impedance value will be produced. A good technique for maintaining the appropriate reference plane is accomplished by creating a new connector to measure the triple stub tuners. Two connectors are attached as shown in Figure 7.

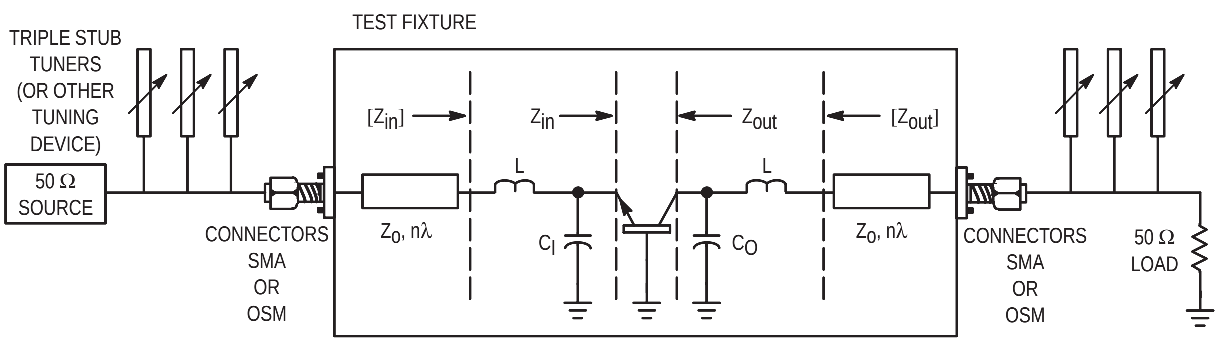

Figure 5. Basic Circuit to Measure Z_{in}, Z_{out}

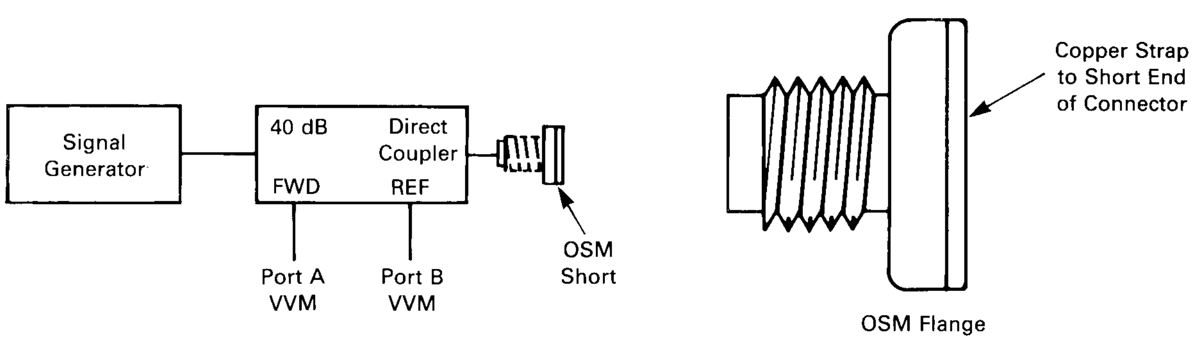

Figure 6. Establishing Reference Plane

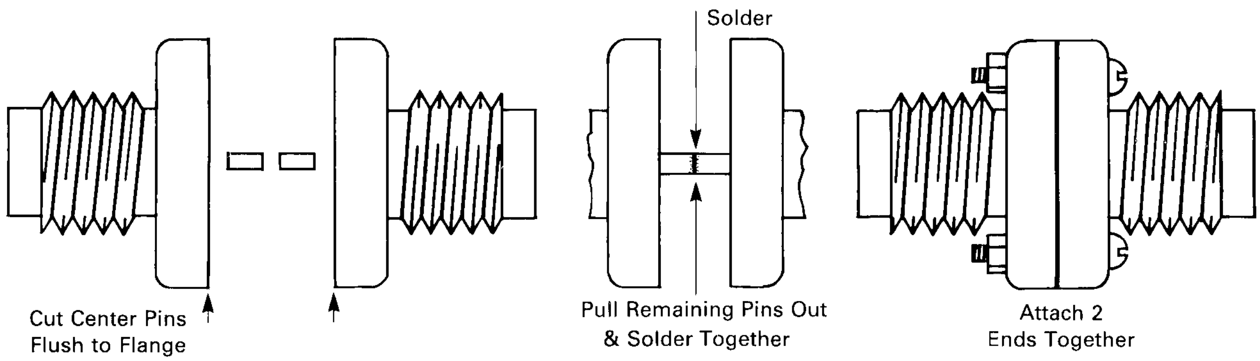

Figure 7. Maintaining Reference Plane

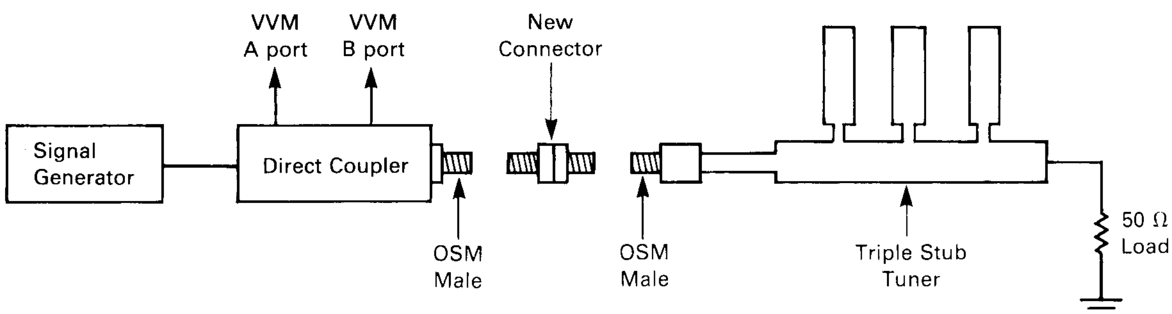

Figure 8. Test Setup to Measure Stub Tuner w/Load

The triple stub tuner, load combination may now be measured with an adequate degree of accuracy using the test setup shown in Figure 8.

Repeat the process for the input stub tuner combination. Two numbers are obtained for each frequency which (Z_{in}) and (Z_{out}) can be calculated from, as shown in the MRF846 example below:

The new impedances can be obtained by using a Smith Chart or using the equation $Z\angle\theta = R_O\,[(1 + \Gamma\angle\theta) / (1 - \Gamma\angle\theta)]$. These impedances (shown in the last column of Table 3) are the impedances which the test fixture will be optimized around. Once again, it is convenient to convert these numbers into parallel equivalents. By doing so, the values of C'_I and C'_O become more obvious. Table 4 shows this process.

Table 3. Measured Z Values for Test Fixture

Frequency		Measured $\Gamma\angle\theta$	$\Gamma\angle\theta$ Converted to Impedance in Ohms	Impedance Transformed Over 26 Ω Line in Ohms
806 MHz	INPUT	$0.35\angle 155^\circ$	24.97 + j8.42	20.72 – j5.64 = [Z_{in}*]
	OUTPUT	$0.37\angle 144^\circ$	24.86 + j12.53	17.72 – j6.66 = [Z_{out}*]
838 MHz	INPUT	$0.26\angle 166^\circ$	29.78 + j3.98	21.35 + j0 = [Z_{in}*]
	OUTPUT	$0.22\angle 154^\circ$	33.30 + j6.58	18.68 + j.74 = [Z_{out}*]
870 MHz	INPUT	$0.14\angle -169^\circ$	38.25 – j1.99	20.21 + j6.92 = [Z_{in}*]
	OUTPUT	$0.07\angle -158^\circ$	44.10 – j2.24	17.90 + j8.3 = [Z_{out}*]

Table 4. Conversion of Z Values to C Values

Series Impedance [Z_{in}] and [Z_{out}]	R_p	X_p	Capacitance Required
20.72 + j5.64	22.26	j81.8	2.42 pF C'_I
17.72 + j6.66	20.2	j53.8	3.67 pF C'_O
21.35 + j0	21.35	—	0.0 pF C'_I
18.68 + j.74	18.7	j472	0.40 pF C'_O
20.21 – j6.92	21.6	– j68.3	– 2.68 pF C'_I
17.90 – j8.3	21.75	– j46.9	– 3.90 pF C'_O

From Table 4, notice the calculated values of C_I and C_O come close to giving the desired frequency response. C'_I is zero at the band center, indicating the capacitors selected for the input are optimum. The values for C'_O produce a slight skew in performance toward the high end of the band. Capacitor values for the output could be reduced slightly, but they will remain the same until final fixture performance is determined.

Since C'_I and C'_O are very small capacitor values, little or no capacitance is actually needed for C'_I or C'_O. However, to allow minor tuning adjustments, a small trimmer capacitor is included at the wear block/transmission line interface.

The final calculation which needs to be performed is that of finding the optimum characteristic impedance for the transmission line. The recommended approach for doing this is to use a computer optimization program which will iterate any number of variables for a desired frequency response. The variables available to be optimized at this point are Z_o, C'_I and C'_O and even $n\lambda$ (transmission line length). Z_o, C'_I and C'_O are the very minimum variables.

In the example of the MRF846, where input R_p varies from 22.3 Ω to 21.6 Ω over the frequency band, a close approximation can be had by using a mean value of 21.9 Ω. This results in a Z_o of 50 x 21.9 = 33 Ω. The output R_p starts at 20.2 Ω, dips to 18.7 Ω and goes back up to 21.75 Ω. Using the same method as before, Z_o is calculated as 50 x 20.2 = 31.7 Ω where 20.2 Ω is the mean value of 18.7 and 21.75. Using a computer optimization program, a 32 Ω, quarter wave transmission line is optimum for the input and a 30 Ω quarter wave line is optimum for the output. These are the values used in the MRF846 test fixture.

After constructing the MRF846 test fixture and tuning the small trimmer capacitors for best overall gain and input reflected response, the average data shown in Figure 9 was obtained.

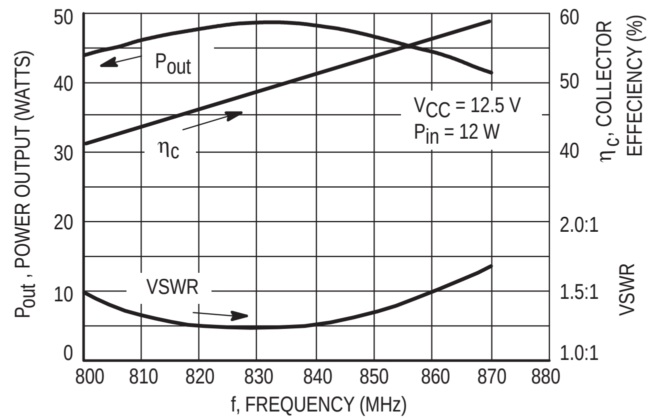

Figure 9. Test Fixture Performance of MRF846

The goal was to demonstrate 12/40 W across the band with less than 2.0:1 input VSWR and greater than 45% collector efficiency. Further optimization could be done by performing impedance measurements on additional transistors or characterizing the test fixture more accurately. However, the above performance is very satisfactory to the required performance. The best compromise for a second pass fixture would be to trade-off gain at 806 and 838 MHz for efficiency, and redesign input and output matching networks for the new impedance tables. This, of course, is only one of the many procedures which may be followed in developing an 800 MHz test fixture.

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