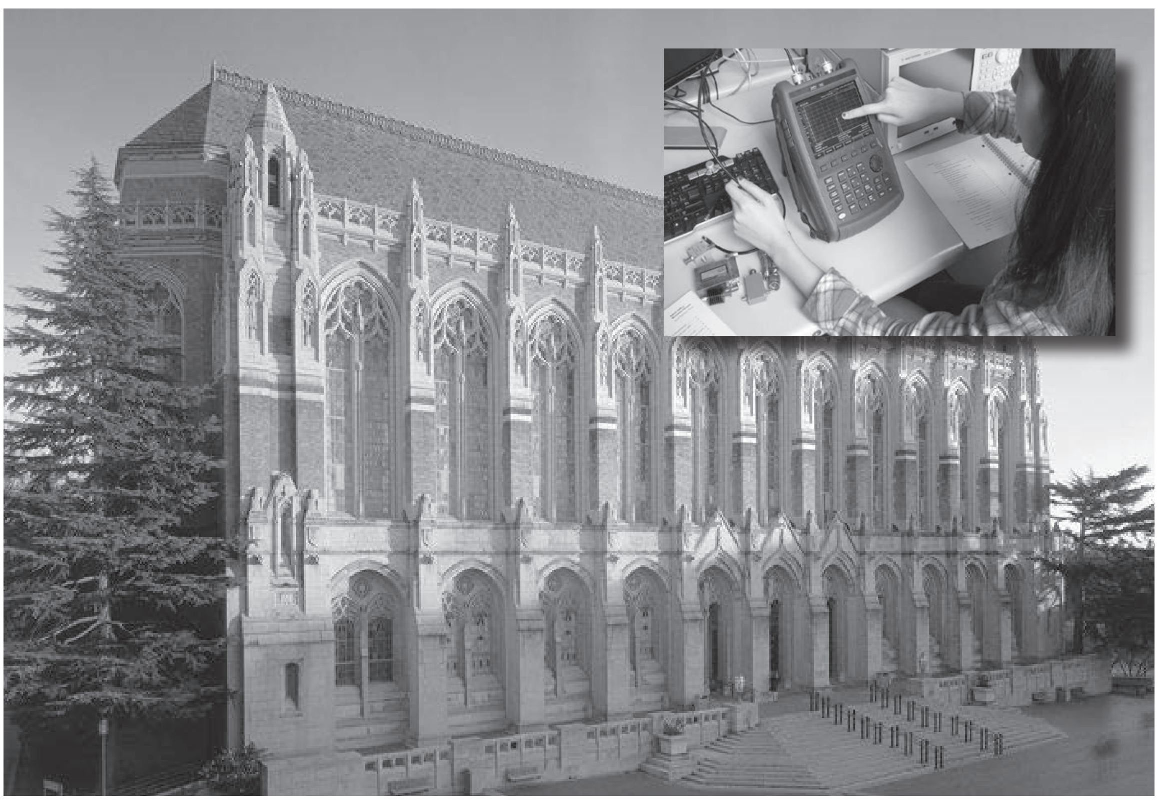
Keysight Technologies

Impedance Matching in the Laboratory

University Engineering Lab Series - Lab 4

Application Note



Introduction

The reception and transmission of small signals into amplifier stages and the transmission of large, powerful signals into loads both require careful attention to minimizing losses. At RF and microwave frequencies reflections of the propagating wave can drastically undermine the efficiency of transmitted power or the signal-to-noise ratio of transmitted information. Proper matching of impedances is needed to minimize these reflections and insure that any signal which reaches a load is actually absorbed by that load as useful power. Proper impedance matching is a fundamental skill that is central to all RF and microwave engineering. The ability to design and execute proper impedance matches is a crucial and highly sought skill which can make or break a career as well as a specific design. The mathematical elements of impedance matching have already been investigated using SPICE and other tools such as the Smith chart. In this lab, some practical laboratory approaches to the problem of impedance matching will be examined. Network analyzers are the essential tool for assessing and tuning an impedance match. Impedance matching is often viewed as a difficult art because impedance matching involves an interplay of measurements and design calculations. However, there are developed methods, and mastering these is essential to RF and microwave engineering.

A quick overview

Consider first the following problem. A load resistance of R_L = 10 Ω is to be fed by a Z_0 = 50 Ω transmission line. If the line were directly connected to the load, this would produce a reflection of $\Gamma = -0.667$, which means that the load only absorbs $|\Gamma|^2 = 0.555$ of the incident power, and the other $1 - |\Gamma|^2 = 0.445$ fraction gets reflected back to the generator. The voltage standing wave ratio (VSWR) for this situation is 5.00, indicating deep standing waves which will make the line impedance very sensitive to its length and potentially cause problems for the generating source.

One approach to this problem is to add a series matching resistance of R_{sm} = 40 Ω, which will bring the total of $R_{sm} + R_L$ equal to Z_0, as shown in figure 1. This indeed creates a perfect impedance match between the load and the line, with the reflection coefficient now reduced to $\Gamma = 0$. However, the introduction of R_{sm} reduces the power that the load R_L receives. Since these two resistances are in series, the load R_L only receives 1/5 of the power leaving the transmission line. The matching network of R_{sm} thus introduces a huge insertion loss of 10 log (5.0) = 7.00 dB.

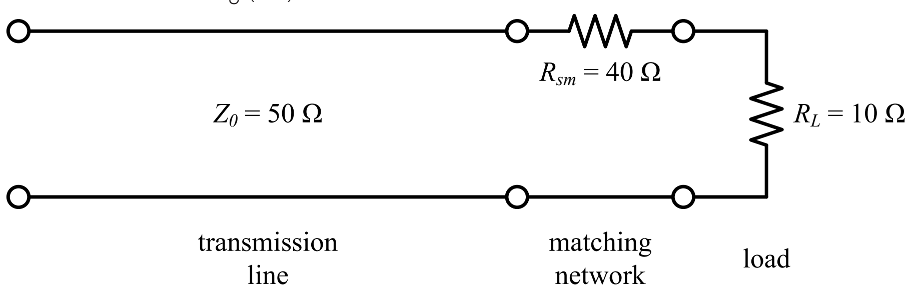


Figure 1. Insertion of a series matching resistance when $R_L < Z_0$

Next, consider a load resistance of R_L = 75 Ω which needs to be matched to the same Z_0 = 50 Ω transmission line. Directly connecting the load to the line would produce a reflection coefficient of $\Gamma = +0.200$ and a VSWR = 1.50. In this case, a parallel matching resistance of R_{pm} = 150 Ω could be added to reduce the effective load impedance to match to the line, as shown in figure 2. Note that 150 Ω || 75 Ω = 50 Ω. Again, this creates a perfectly matched situation, but the introduction of the parallel matching resistance allows the actual load R_L to only receive 2/3 of the power leaving the transmission line. In this case, the parallel matching resistance R_{pm} introduces an insertion loss of 10 log (1.5) = 1.76 dB.

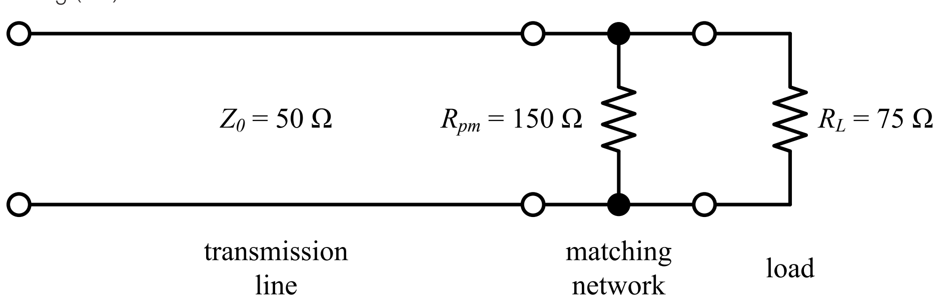


Figure 2. Insertion of a parallel matching resistance when $R_L > Z_0$

In both of the above cases, the introduction of a series or parallel matching resistance created a perfect impedance match, and an important benefit of this matching is that it is frequency independent. Although significant power loss is an obvious drawback. Resistive elements can be used for broadband matching, but at the expense of power or signal loss, which can often be significant.

In many cases, the range of frequencies is restricted to a narrow band, and this can allow the use of reactive elements, that is, inductors and capacitors, to achieve low loss impedance matching over that range. Simply adding a single reactive element in series or parallel to a load impedance can cancel any reactive component of the load, but it cannot bring an arbitrary load impedance to equal the desired Z_0 = 50 Ω. To handle arbitrary load impedances with both real and imaginary parts, an additional degree of freedom is required, and this is most easily achieved through the use of an L-network which involves two reactive elements.

The synthesis of these L-networks involves some computations for which the Smith chart is a very handy tool for mapping out the matching strategy and getting initial values for these elements. In practice there will be parasitic elements which may not have been accounted for, as well as other inaccuracies which will cause the final match to be imperfect. The overall procedure for matching a load to a line first involves a measurement of the load impedance. Once that is known, a matching strategy can be developed to guide the calculation of the matching elements. These will be nominal values which should be close to the needed values, but perhaps still not exact. Achieving a finer degree of matching then involves adjusting these matching element values while monitoring the reflection coefficient of the line. Thus, some ability to fine tune the matching network is often required. A network analyzer is the preferred tool for both measuring the initial impedance of the load as well as monitoring the reflection coefficient so that it can be nulled out by fine tuning of the matching network.

A Smith chart is a polar plot of the reflection coefficient Γ, so that the point at the origin corresponds to $\Gamma = 0$, the case for zero reflections and perfect impedance matching of a load to the transmission line. The objective of impedance matching is therefore to land a bullseye in the center of the Smith chart for the range of frequencies that need to be matched, as shown in figure 3. The great advantage of the Smith chart is that after it has been mastered, it provides a very intuitive way of understanding the load and line interactions and of designing proper impedance matching networks. One can assess a situation on a Smith chart and very quickly learn what is possible, easy, or difficult, versus performing brute force "what if" calculations, many of which might lead to dead ends.

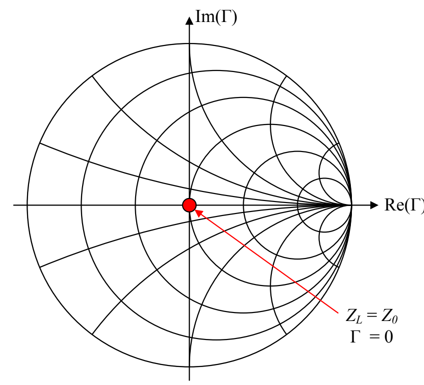


Figure 3. Proper impedance matching: hitting the bullseye on the Smith chart

A computational example will next be developed to illustrate the procedure for synthesizing an L-type matching network. Suppose that the load is a complex impedance of Z_L = 20 – j40 Ω, and that the frequency range of interest is centered on 100 MHz. This impedance results from a parallel combination of 100 Ω and – j50 Ω, which could in practice be a parallel combination of a 100 Ω resistor and a 32 pF capacitor. A few pF of stray capacitance is quite common for many loads, so this represents a fairly common situation. The objective is to design an L-network of reactive elements which will match this impedance to a Z_0 = 50 Ω transmission line.

The synthesis of the matching network will follow the path of adding series impedances (Z) and parallel admittances (Y) to the existing load impedance to bring its final value to a purely real 50 Ω. Since the added elements will be purely reactive to minimize loss, the synthesis will be adding series reactances (X) and parallel susceptances (B).

There exists many possible solutions to such a problem, and it is the engineer's job to pick the one which provides the best balance of simplicity, sensitivity, and cost. That is usually assessed by the final component types and values of the matching network. There are different possible topologies, with the L-network facing in either direction. This corresponds to the order in which the reactances and susceptances are added to the load impedance. In addition, the added reactances and susceptances can be positive or negative, that is, inductors or capacitors. Different designs can do the job in different ways. For example, it may happen that a needed capacitor value ends up being very expensive due to its required size. A slightly different design using a lower cost inductor might then be the preferred strategy. To make these choices, the different alternatives need to be mapped out, and the Smith chart proves to be an extremely effective tool for this purpose.

Figure 4 shows a Smith chart which illustrates one possible solution for matching the load of Z_L = 20 – j40 Ω. It should be kept in mind that the Smith chart is simply a polar plot of the reflection coefficient Γ, which is a complex quantity involving magnitude and phase, and whose magnitude is never more than unity. The outer periphery of the Smith chart thus corresponds to |Γ| = 1 and the center point to Γ = 0. The resistance and reactance circles are simply a different coordinate system that is overlaid on the polar Γ coordinates for which the load impedance can be directly read off from.

Point A in figure 4 marks the location of the load impedance, Z_L = 20 – j40 Ω, and the starting point for the impedance matching network synthesis. The normalized load impedance is $z_L = Z_L/Z_0$ = 0.40 – j0.80. The reflection coefficient at this point is Γ = –0.08 – j0.62 = 0.62 / –97.1°. Notice that point A is at the intersection of the r_L = 0.40 circle and the x_L = –0.80 circle as shown in figure 4.

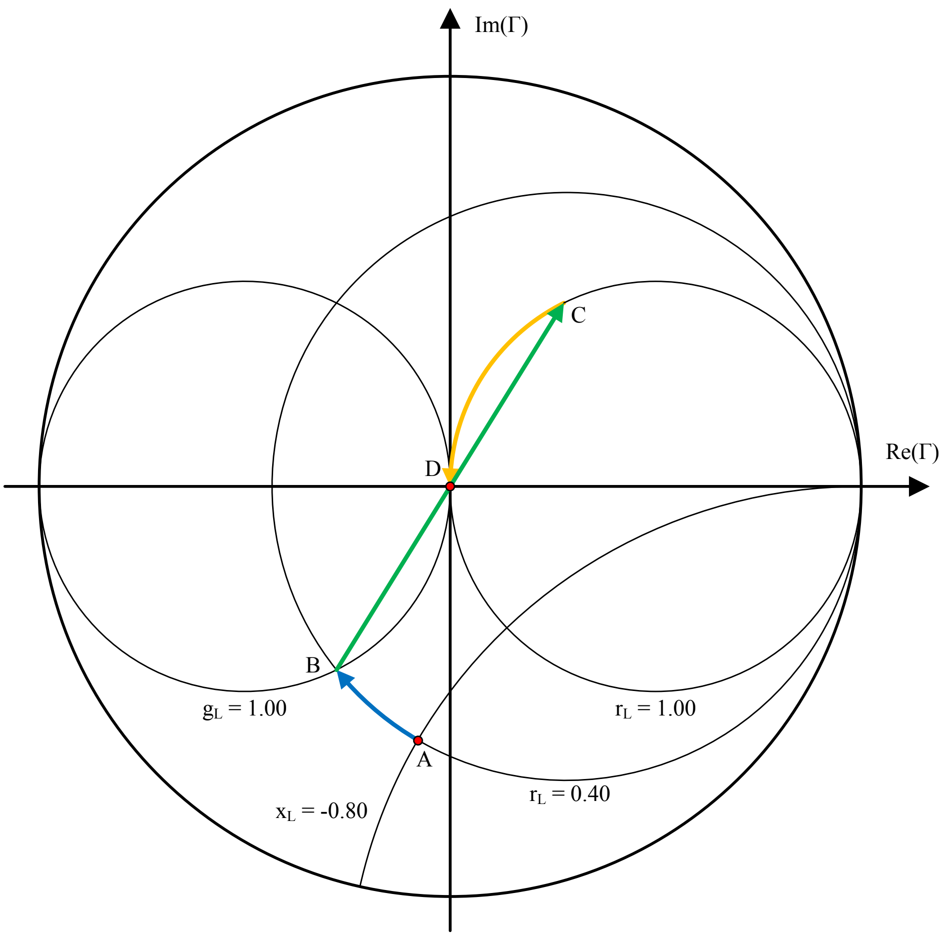


Figure 4. An example of L-network matching using the Smith chart

The matching strategy is to add only reactive elements so that the matching network remains lossless. The addition of a reactive element takes a trajectory on the Smith chart of moving along a constant resistance circle. Starting from the load impedance at point A, the addition of reactance moves along the blue arc of r_L = 0.40 to point B.

A Smith chart can represent either impedance or admittance, and conversion between the two is accomplished by rotating 180° about the origin, or equivalently mirroring across the origin. The resistance circle of r_L = 1.00 thus mirrors over to the conductance circle of g_L = 1.00 as shown in figure 4. Point B represents the impedance of the load on to which the first reactive element has been added in series. Mirroring that point across the origin to point C then represents the admittance of those same elements. Note that the trajectory in green from point B to point C does not involve adding any elements to the matching network; it simply changes impedance to admittance (Z → Y). The reason for having point B located where the r_L = 0.40 circle and the g_L = 1.00 circle intersect is because that point then gives an admittance whose real part is 1.00, the first step in matching the load to the line.

From point C, the trajectory along the orange arc of g_L = 1.00 (remember that Z has now been flipped to Y) moves to point D by subtracting susceptance. Point D is the final goal with a normalized impedance of $z = y = 1.00$ and $\Gamma = 0$, the bullseye.

The Smith chart shows how the addition of different elements changes the impedance of the load to finally bring it into a match to the transmission line impedance of Z_0. From a circuit perspective, the first addition of a positive reactance along the blue trajectory from point A to point B means that inductance must be added in series with the load. The addition of a negative susceptance along the orange trajectory from point C to point D means that inductance must then be added in parallel. The matching network associated with the Smith chart of figure 4 is thus that shown in figure 5.

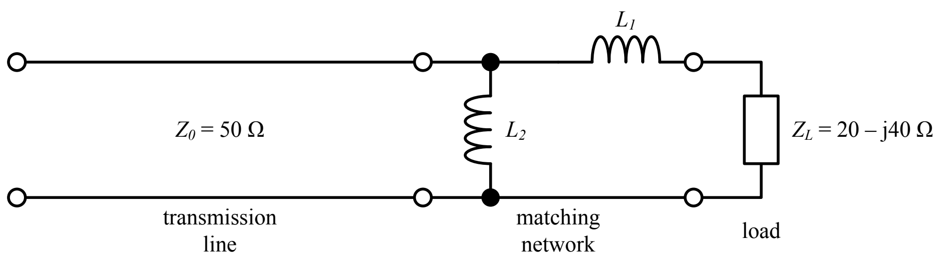


Figure 5. Matching network topology for the Smith chart trajectory of Figure 4

The next step is to compute some nominal values for these matching elements. The Smith chart coordinates for point B define these values. The intersection of the r_L = 0.40 and the g_L = 1.00 circles can be expressed as

$$\frac{1}{z_L + jx_1} = \frac{1}{r_L + j(x_L + x_1)} = 1 + jb_2$$

where x_1 is the normalized reactance added by L_1 and b_2 is the normalized susceptance subtracted by L_2. Computationally, x_1 and b_2 need to be found so that the above relationship is satisfied. Equating real and imaginary parts gives

$$r_L - b_2(x_L + x_1) = 1$$
$$b_2 r_L + (x_L + x_1) = 0$$

The solution to these two equations is

$$x_1 = -\sqrt{r_L(1-r_L)} - x_L = -0.490 + 0.800 = 0.310$$
$$b_2 = \frac{r_L - 1}{x_L + x_1} = \frac{-0.600}{-0.490} = 1.225$$

This gives inductor values of

$$L_1 = \frac{x_1 Z_0}{2\pi f} = \frac{0.310 \cdot 50\ \Omega}{2\pi \cdot 100\ \text{MHz}} = 24.7\ \text{nH}$$
$$L_2 = \frac{Z_0}{b_2 2\pi f} = \frac{50\ \Omega}{1.225 \cdot 2\pi \cdot 100\ \text{MHz}} = 65.0\ \text{nH}$$

These nominal inductor values should ideally provide a match to the load of $Z_L = 20 - j40\ \Omega$. Inaccuracies of measurements, component statistical variations, and unaccounted for parasitic elements will in practice make this only approximate. However, it should be a close starting point which can then be improved upon with a small amount of tuning. This is the point where the process moves into the lab and calculations are supplanted by instrumented measurements.

A practice measurement using the FieldFox network analyzer

One of the most powerful applications on the Keysight FieldFox N9914A is the network analyzer (NA). Network analyzers will be discussed in more detail in a subsequent laboratory which will use the full 2-port capabilities of the instrument. For this laboratory, only 1-port measurements will be required, and a more superficial initial treatment of network analyzers will be sufficient.

A network analyzer is composed of a swept frequency signal source, also known as a sweeper for short, and a tuned receiver whose passband tracks that of the sweeper. Alternatively, the tuned receiver can be the unit which controls the sweep and the signal source then tracks along with the receiver. In this latter case, the signal source is often referred to as a tracking generator. The important point is that both the source and the receiver are synchronously swept together through the frequency range of interest.

If the receiver is only sensitive to average signal power, the ratio of the received signal power to the power that is output from the generator allows the magnitude of the transfer function to be directly obtained as a function of frequency. This is how a scalar network analyzer (SNA) operates. For many applications, SNAs are a very cost effective instrumentation system.

A much more sophisticated and powerful system is obtained if the receiver is phase-locked to the signal source. This would be termed a coherent receiver, and it allows both the magnitude and the phase of the transfer function to be obtained. The vector network analyzer (VNA) operates in this manner, and it is the premier instrument of the RF and microwave world. VNAs are most commonly 2-port instruments, but many powerful functions can be obtained from single port measurements.

Scattering parameters, or simply “S-parameters,” are the language of vector network analyzers. These are the most commonly used system of measurements for characterizing RF and microwave devices. A more detailed look at scattering parameters will be taken up later, but for the present laboratory, the only essential feature to know is that the S_{11} scattering parameter is exactly the same as the reflection coefficient Γ that has been discussed previously. The 11 subscript simply indicates that the S_{11} parameter relates the signal received back from port-1 relative to the signal that was launched into that same port-1.

The FieldFox network analyzer application is a full-featured, 2-port vector network analyzer. Previous measurements such as return loss, cable loss, and insertion loss that were made using the CAT application were scalar analyzer functions. The distance-to-fault (DTF) measurements actually did use phase information to perform the FFT computations to obtain the distance representation of the returned signal, but this phase information could not be directly obtained from the CAT application. By comparison, in the NA application, phase information is extremely important and it can be directly obtained through the S-parameter measurements.

To get started, first turn on the FieldFox and let it load and launch the default CAT application. Once the CAT application is running, press the Mode button and then press the NA soft key to load and launch the Network Analyzer application. Once the NA application is running, the Measure menu should appear on the soft keys. If it is not already selected, press the S11 soft key to measure the S_{11} parameter. For the S_{11} measurements, only port 1 of the FieldFox will be used. Make sure that a Type-N to SMA adapter is attached to port 1.

First, a few exercises will be performed to gain some practice in navigating the Smith chart. From the Measure menu, press the Format soft key, and then press the Smith soft key to display the *S11* parameter on Smith chart coordinates. The measured S_{11} data should appear as a yellow trace which runs around the outer unit circle of the Smith chart, perhaps several times. The open circuit condition of port 1 would ideally produce a reflection coefficient of $\Gamma = +1$, but the short section of line created by the Type-N to SMA adapter creates a short section of line whose impedance transformation varies with frequency. The very broad sweep range of the FieldFox shows how much this can affect matters over the full range sweep, which is 6.5 GHz in this case.

Press the Freq/Dist button and change the frequency sweep to a center frequency of 100 MHz and a frequency span of 200 MHz. Notice how the yellow trace has now collapsed to a much shorter arc which lies on the right hand side of the Smith chart. The reason for having this wide of a frequency sweep is so that the yellow trace becomes longer and more visible on the display. The yellow trace lies mainly on the unit circle of the Smith chart and starts at the point $\Gamma = +1$. The display also shows ∞, indicating that this point corresponds to $z_L = \infty$, an open circuit. In previous measurements using the CAT application, the reflection coefficient magnitude of the open circuit port was found to be $|\Gamma| = 1$; with the NA application, phase information is now available to show that $\Gamma = +1$, a subtle but important difference.

Attach a shorting cap to port 1. Notice how the yellow trace now jumps over to the opposite left hand side of the Smith chart, close to the point marked “0” which indicates $z_L = 0$, or a short circuit. The yellow trace may not be precisely on the left hand side of the unit circle, but it should be close. The left hand side of the Smith chart corresponds to a reflection coefficient of $\Gamma = -1$, meaning that reflected waves undergo a 180° phase reversal. Previously, the CAT application showed that a short circuit condition gave $|\Gamma| = 1$ without this important phase information. Using the NA application, the difference between an open and a short is clearly distinguishable. This is the value of phase information.

Next, replace the shorting cap with a male-to-male SMA connector and attach to that an adjustable 500 Ω potentiometer load as shown in figure 6. The trace will no longer be hugging the unit circle, so reduce the frequency span to 50 MHz to tighten up the length of the trace. By adjusting the potentiometer load, the impedance can be dialed to anywhere from 0 Ω (a short), to 50 Ω (matched), to 500 Ω (not quite an open, but getting there). Experiment with how adjusting the load resistance moves the yellow S_{11} trace across the Smith chart. If the load were an ideal pure resistance, the trace should move along the horizontal axis of the Smith chart. The fact that it lies a little off of the horizontal axis and that it has some length to it indicates the presence of some additional parasitic capacitance and inductance that results from the manner in which the load was constructed.

Figure 6. Connection of the 500 Ω potentiometer load

Try to adjust the load resistance to bring the yellow trace as close as possible to the center of the Smith chart, $\Gamma = 0$. This process will be a little “touchy,” indicating the sensitivity of this alignment with this particular value of potentiometer. Once the yellow trace sits over the point $\Gamma = 0$, congratulations, this has now properly matched the load impedance to the $Z_0 = 50\ \Omega$ line! This is the objective of impedance matching.

Simply adjusting the load resistance makes impedance matching trivial. The more practical situation is where the load is not adjustable and where it may contain a significant reactive component. For this, an impedance matching network is needed.

One last feature to observe with the Smith chart is the effect of the transmission line length, another aspect in which phase information becomes all important. The previous loads were connected to port 1 using only the male-to-male connector which is in effect a very short length of line. Remove the connector and connect an 18 inch long SMA cable to port 1. At 100 MHz, the free space wavelength is $c/f = \lambda_0 = 3.00$ m = 9.84 ft. The velocity factor for the cable was previously found to be 0.667, so the wavelength in the cable is reduced to 2.0 m = 6.56 ft. A quarter wavelength section of cable is thus 0.5 m = 1.64 ft = 19.7 in. After the connectors have been added, the 18 inch long SMA cable is very close to a $\lambda/4$ section of line at 100 MHz.

With the 18 inch SMA cable attached to port 1 and its end open circuited, the yellow trace should have now moved over to the left hand side of the Smith chart, where $\Gamma = -1$ for the case of a short circuit. An extremely important property of $\lambda/4$ sections of transmission line is that they effectively change the impedance from Z_L to Z_0^2/Z_L. This is equivalent to taking the reciprocal of the normalized impedance. This is also equivalent to rotating 180° around the Smith chart, and in this case it changes the impedance seen by the FieldFox from an open circuit to a short circuit. Quarter wave transformers are a particularly powerful tool in RF and microwave engineering; they will be investigated in more detail later. For the present, the important principle to understand is that adding a section of transmission line will rotate the S_{11} parameter data around the Smith chart, with one full turn corresponding to a line length of $\lambda/2$. To effectively reduce the length of the line to $\lambda/8$, the frequency can be decreased to a center frequency of 50 MHz with a span of 10 MHz. Try this and notice that the yellow trace has moved toward the point $\Gamma = -j$ at the bottom of the Smith chart. This indicates that moving the measurement point farther away from the load (toward the generator) is represented by rotating in a clockwise direction around the Smith chart. Rotations on the Smith chart are simply changes in phase of the reflection coefficient. Experiment with the adjustable potentiometer load, changing the frequency, and the limiting cases of short and open loads at the end of the 18 inch SMA cable.

Now back to impedance matching. The end objective of this laboratory is to properly impedance match an arbitrary experimental load to a 50 Ω transmission line. To start the process, the first step is to measure the impedance of this load over the frequency range of interest.

The experimental load is made up of a 100 Ω resistor in parallel with a 22 pF ceramic capacitor, both soldered to a female SMA connector. Replace the adjustable potentiometer load with the experimental load, as shown in figure 7. The center frequency for the impedance matching will be 100 MHz, and to get a better picture of how this match is working, a frequency span of 20 MHz will be used. Configure the FieldFox frequency sweep for this frequency range of 90 to 110 MHz. The measured S_{11} parameter data should appear as a short yellow trace in the lower half of the displayed Smith chart. One reason for keeping the frequency span relatively wide is to make the yellow trace sufficiently large that it can be easily seen on the display. With too narrow of a span, the measured S_{11} parameter data could otherwise appear as only a few pixels.

Figure 7. Connection of the experimental load

To obtain numerical data, press the Marker button and a diamond should appear on the yellow trace and the marker coordinates should also appear in the upper right corner of the display. If it is not already there, move the marker to 100 MHz and record the marker coordinates. In the Smith chart format, the marker does not give the value of S_{11}; rather, it gives the value of the load impedance Z_L, based upon $Z_0 = 50$ Ω. As an example, at 100 MHz, the measurement might yield $Z_L = 24.5 - j37.6$ Ω. Just below the marker coordinates, the marker data also displays the component value that would be associated with the imaginary part of the load impedance. In this case, the display shows 42.30 pF. This is an extremely useful function of the FieldFox which allows it to work as a microwave impedance analyzer. Note that the values measured in the lab may not be exactly the same as those cited here. Record the values for the load that is being used, these will be needed later.

Some further analysis

Taking the reciprocal of the measured impedance yields a load admittance of $Y_L = 0.0122 + j0.0187\ \Omega^{-1}$. This corresponds to a real resistance of 82.2 Ω in parallel with a capacitive reactance of 53.6 Ω. Converting the capacitive reactance to capacitance at a frequency of 100 MHz gives a 29.7 pF capacitor. Again, the values measured in the lab may vary slightly from these.

If the experimental load were disassembled and the component values checked with a low frequency LCR meter, the values found would be very close to 100 Ω and 22 pF, as the components are indeed marked. Why do their values differ so much from these at 100 MHz? The reason is primarily because of parasitic elements which are effectively small at low frequencies becoming much more important at higher frequencies. The very geometry of the assembled SMA connector and these two components introduces additional capacitance which adds to that of the ceramic capacitor. The leads of each component introduce parasitic inductance whose reactance increases with frequency. A critical observation is that at RF and microwave frequencies, the effective impedance of a component may look nothing like its low frequency component values. Hence, the need for making these measurements!

The center frequency of 100 MHz is still a rather low frequency in the RF and microwave range. Increase the frequency span to 100 MHz and observe the load impedance trace on the Smith chart display. The increased width indicates that the load impedance is changing over a wider range. Press the Marker button and rotate the dial to move the marker up and down the trace to observe how the impedance is changing with frequency. Next, change the frequency sweep to go from 100 MHz to 400 MHz. The load impedance should now follow a fairly long arc in the Smith chart display. Press the Marker button and use the dial to move the marker along the yellow trace. Notice that when the trace is below the horizontal axis (a negative imaginary part), the marker shows a component value of a few tens of pF, i.e. capacitive. When the trace is above the horizontal axis (a positive imaginary part), the marker shows a component value of a few nH, that is, inductive. Use the marker to find the frequency at which the load impedance is purely real. This point should be near 250 MHz. Take a look at the load impedance at this frequency, and it should be a surprisingly low value. How did the 100 Ω resistor and 22 pF capacitor turn into this? Again the answer lies in the parasitic elements of the load. At this frequency, the capacitor in parallel with the resistor effectively shorts it out and the series inductance of the component leads produces enough reactance to cancel out the capacitive component. This is sometimes referred to as the self-resonant frequency of the load.

In fact, every component has a self-resonant frequency where its effective impedance changes from inductive to capacitive. At sufficiently high frequencies, capacitors behave like inductors, inductors behave like capacitors, and resistors go through ranges of looking like capacitors or inductors. The longer the leads on the components or the larger their physical size in relation to a wavelength, the lower this self-resonant frequency will be. Surface mount components which have no leads at all can thus operate at significantly higher frequencies than their leaded counterparts. These effects are just some of the fun to be had at RF and microwave frequencies.

Now, onward to the problem of matching this experimental load to a 50 Ω transmission line. The critical piece of information is the impedance of the load at the desired matching frequency of 100 MHz: $Z_L = 24.5 - j37.6\ \Omega$, or something close to this value. This value is fairly close to the value of $Z_L = 20 - j40\ \Omega$ that was used in the prior example design problem, so that previous L-network design should serve as a good starting point for matching this experimental load.

The prior L-network design required 65 nH and a 25 nH inductors. As has been observed by direct measurements, simply selecting these components based upon their low frequency values will almost certainly produce something quite different at 100 MHz. Perhaps something not even close. The challenge is to synthesize these values at the frequency they are needed to operate.

Measurement and alignment assignment

To construct the previously designed matching network, a means is needed to connect the matching components into the transmission line, as both series and parallel (shunt) elements. This proves increasingly difficult as the frequency increases, but it can still be accomplished at 100 MHz with some creativity. An impedance matching breadboard has been constructed for this purpose as shown in figure 8.

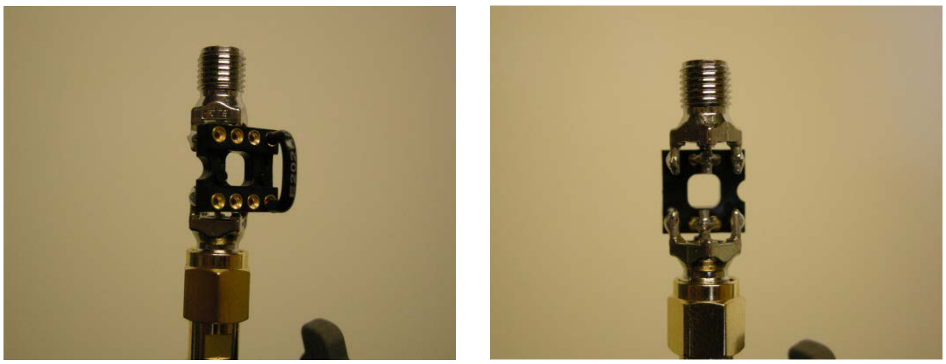


Figure 8. The impedance matching breadboard, front and back views

The impedance matching breadboard is constructed from a pair of female SMA PCB connectors that are soldered to a DIP-8 machine pin socket. On each side, the two middle pins are connected to the center pin of the SMA connector, and the two outer pins are connected to the housing or shield of the SMA connector. This configuration allows all possibilities for inserting leaded components into the transmission line. To insure that both of the SMA housings or shields are electrically connected, a short wire is needed to jumper these across the DIP-8 socket. This is shown in figure 8, with the short black wire connecting pins 4 and 5 of the DIP-8 socket. The L-network described previously can be implemented by inserting L_1 between pins (2,3) and (6,7), and inserting L_2 between pins 1 and (2,3).

The next issue is to find elements for L_1 and L_2. This is a situation where building works better than buying. Both inductors will be crafted by hand from some short lengths of solid hookup wire. Solid #22 gauge wire works best for this, since that fits well into the machine pin DIP socket receptacles. The first step is gain some feel for how much wire or coil is needed to create these inductance values. The self-inductance of #22 solid wire is roughly 20 nH per inch, so not very much is going to be required. The best strategy is to experiment with some different lengths, simple hairpins versus open loops versus coils of a few turns, and test these on the FieldFox to find their inductance values at 100 MHz. Figure 9 shows some of the possibilities.

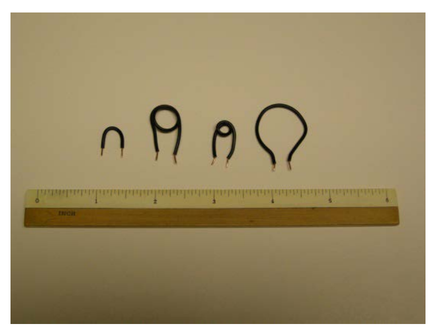


Figure 9. Possible hand-made inductor geometries

Make up a few inductors like those shown, and test them by stripping back about 1/4 inch of insulation and plugging them into pins 1 and 3 of the impedance matching breadboard. This is where the L_2 element would go, and with nothing else attached to the breadboard, the inserted inductor will dominate the impedance. Use the FieldFox to measure the inductance of these elements, again with a center frequency of 100 MHz and a 20 MHz span. The yellow traces should appear on the top of the Smith chart, near $\Gamma = +j$. Use the marker function to read out the impedance directly at 100 MHz.

Inductors made in this fashion are tunable. By bending their geometry, their inductance can be adjusted to the precise value that is needed for the match. The inductance of a loop is proportional to its area, so simply opening and closing the size of the loop, along with changing the length of the wire in the loop, will provide quite a bit of adjustment. For multi-turn coils, the number of turns and their spacing apart provides additional adjustment. The objective is to develop a wire inductor that provides some tuning range around the nominal values of L_1 = 25 nH and L_2 = 65 nH. Note that L_1 is not going to involve much more than about 1 inch of wire.

Once suitable wire inductors have been found, connect up the impedance matching breadboard to produce the network of figure 5. Add the experimental load using a second male-to-male SMA connector and test the result out on the FieldFox. If all has gone well, the resulting yellow trace of S_{11} should be a good deal closer to the $\Gamma = 0$ point that the experimental load was beforehand. Tune up the matching network by bending the inductors slightly to adjust their values and move the 100 MHz point on the trace as close as possible to the $\Gamma = 0$ point. It is helpful to place the marker on 100 MHz and try to move the bottom of the green diamond on to the center of the Smith chart. Figure 10 shows the final configuration.

Figure 10. Final assembly of the matching network and the experimental load

The majority of the needed adjustment will come from L_2. Re-examine the mathematics behind the computation for L_2 to understand why. In the real world, tuning sensitivity is an issue. If the matching network is too difficult to properly align, it may end up causing more problems than it solves. Good matching network design always takes into account the tuning range and sensitivity requirements. A full treatment of this is beyond the scope of this lab, but the idea is to gain some appreciation of the issues.

After a little adjustment of L_1 and L_2, viola!, and a good impedance match should be achieved. This is shown in figure 11. What has this accomplished? With a reflection coefficient of very close to zero, all of the power launched into the transmission line at 100 MHz will now be directly absorbed by the load. There will not be any standing waves on the transmission line, and its input impedance will look like a real 50 Ω to the generator, regardless of the length of the line in between. This is very desirable.

Be sure to record a screen image of the final impedance match. This is something to be proud of. As a final thought, consider how difficult this process would be without having a vector network analyzer.

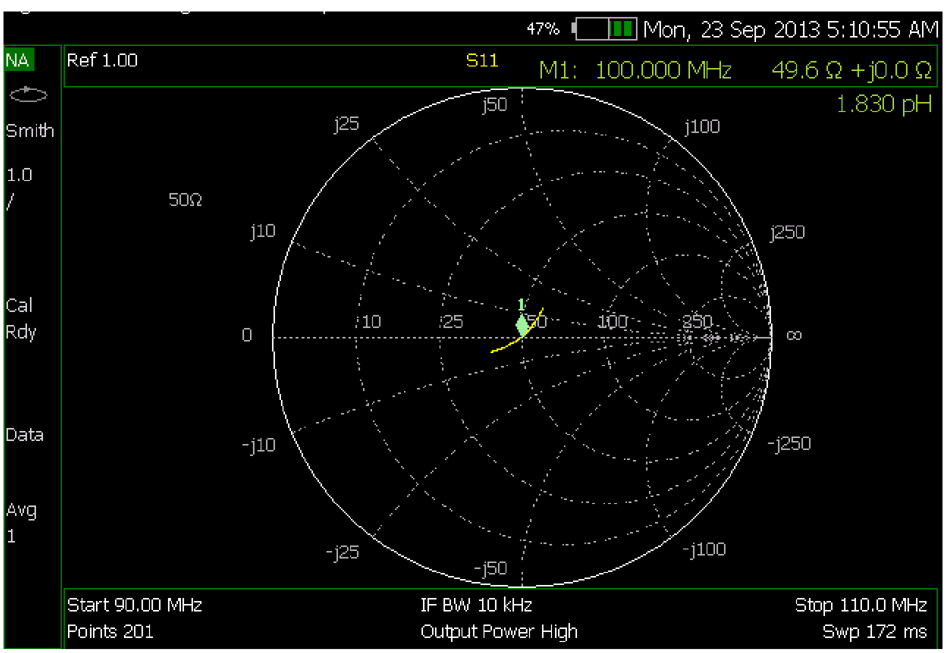


Figure 11. The final result: A mighty fine impedance match!

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This application note was created by Professor Bruce Darling, from University of Washington's Electrical Engineering Department, in collaboration with Keysight Technologies' handheld team within the Component Test Division. The content is designed to complement an introductory course in undergraduate electromagnetics.

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