

ARRL Radio Designer *and the Circles Utility*

This two-part article by William E. Sabin, WØIYH, was published in *QEX* magazine in Sept/Oct and Nov/Dec 1998.

Part 1: Smith Chart Basics

Part 2: Small-Signal Amplifier Design

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ARRL Radio Designer and the Circles Utility Part 1: Smith Chart Basics

The Smith Chart is a venerable tool for graphic solution of RCL and transmission-line networks. The Circles Utility of ARRL's Radio Designer software provides an "electronic" Smith Chart that is very useful. For those unfamiliar with the Circles Utility, we begin by exploring basic concepts and techniques of matching-network design.

By William E. Sabin, WØIYH

One of the interesting and useful features of the ARRL *Radio Designer* program is the Circles Utility. This two-part article will look at some of the ways of using Circles. A brief overview of basic principles will be followed by some "walk-through" examples that can be used as "templates" or guidance for future reference. The Circles Utility can do the following things:

- Perform Smith Chart operations to design and analyze transmission-line networks and LCR impedance-matching networks. Part 1 of this article deals with these topics.
- Using the Smith Chart, perform

gain-circle operations that are widely used in active-circuit design, especially amplifiers, using S-parameter equations. Input- and output-matching networks can be evaluated. The stability of an active circuit can be evaluated by plotting stability circles. Noise-figure circles (circles of constant noise-figure values) can be plotted. Figures of merit are calculated. We can perform plots of certain quantities over a frequency range. The result is a visual estimate of the performance of the circuit. Part 2 of this article will deal with this.

Smith Chart Basics

It is a good idea to bring up the Circles Utility and perform the following operations as you read about them.

First, to enable Circles, we must enter and Analyze (key F10) a circuit listing of a two-port network as exem-

plified in Fig 1A, a resonant filter circuit that is also an impedance-matching network. The schematic is shown in Fig 1B. Start with the list in Fig 1A, and modify it later as required. I assume the reader already knows how to use the Report Editor to get XY rectangular plots, tables and polar plots versus frequency of such things as MS11, MS21 etc, and how to set the Terminations. An Optimization of the circuit is also performed.

Fig 2A shows how various lines of constant resistance and constant reactance are plotted on the Smith Chart, using the Circles Control Window (we will henceforth call it CCW). The various entries that we type into the CCW text window are just as shown (the "=" is optional and a "space" can be used instead). In particular, pure reactances can only exist on the outer circle of the chart; to move inside, some

resistance must be added. Positive reactance (+X, inductance) is in the upper half, negative reactance (-X, capacitance) in the lower half. High resistance (R, low conductance $G = 1/R$) is toward the right.

You can also create an R or X circle at some specific location. Type "R" or "X" in the CCW, Execute, then place the cursor at the desired location and press the left mouse button. This creates an R or X circle and a label with some specific value.

To remove a single error in an entry, type "DEL" into the CCW, then Execute. To delete two entries, type "DEL 2," and so forth. To get the best accuracy in all of the various operations, click on Settings/Display/Graphics/Line Width and set the line width to "1," the narrowest line.

Fig 2B shows how various lines of constant-conductance and constant-susceptance values are plotted. This chart is a left-to-right and top-to-bottom reverse image of Fig 2A. Positive susceptance (+B, capacitance) is below and negative susceptance (-B, inductance) above the horizontal axis. Both charts always show inductance above the axis and capacitance below the axis. High conductance (G, low resistance, $R = 1/G$) is toward the left.

G and B circles at some cursor location are created by typing "G" or "B," Execute, then place the cursor at the desired location and press the left mouse button.

The Circles Smith Chart is a YZ chart, which means that Figs 2A and 2B and all four of the quantities shown there can be plotted simultaneously on the same chart in two different colors. The term "Z" refers to R and X in *series* and the term "Y" refers to G and B in *parallel*. It's important to keep this distinction in mind, especially when switching between the two. This YZ capability is a powerful feature that we will use often. Also, the Z chart is "normalized" to 1 Ω at the origin. To normalize a 50 Ω system, divide all actual R and $\pm X$ input values by 50. For other values of Z_0 , such as 52 Ω , 450 Ω etc, use Z_0 as a scaling factor. The Y chart is normalized to 1 S (Siemens = 1 / Ω) at the origin. For a 0.02 S (1 / (50 Ω)) system, multiply actual G and $\pm B$ values by 50, or whatever Z_0 is correct. It is very desirable to have a calculator available to normalize values; the computations are no problem once you get the "hang of it." In the interest of simplicity, it is best (at least at the beginning) to use resistive (not complex) values of Z_0 . For a transmission line, a resistive Z_0 means a line with no attenuation (loss).

Fig 2C shows various values of V (SWR) circles, entered as shown. At any point on a particular V circle, a value of RHO, the reflection coefficient, can also be found. For example, at the intersection of $R = 1$ and $V = 8$, the reflection coefficient is 0.77 at 39.17° (angle measured counterclockwise from the horizontal axis). To find this RHO, we use the following procedure:

- Enter the word "RHO" into the CCW and click the Execute button.
- Place the cursor at the intersection of $V = 8$ and $R = 1$.
- Press "M" on the keyboard.
- This places a mark at this location and also puts the complex (magnitude and angle) value of RHO in the CCW.

The following equations give the relationships involved in the RHO and V operations:

$$RHO = \frac{(R + jX) - Z_0}{(R + jX) + Z_0} \quad (\text{Eq 1})$$

where R, X and Z_0 are normalized to 1.0 as discussed previously and

$$V = \frac{1 + |RHO|}{1 - |RHO|}; |RHO| = \frac{V - 1}{V + 1} \quad (\text{Eq 2})$$

where the vertical bars " | " denote "magnitude." From this equation we see that a V circle is also a circle of constant $|RHO|$, so the V circle is also called a "constant-reflection" circle. Also of considerable interest is the return loss

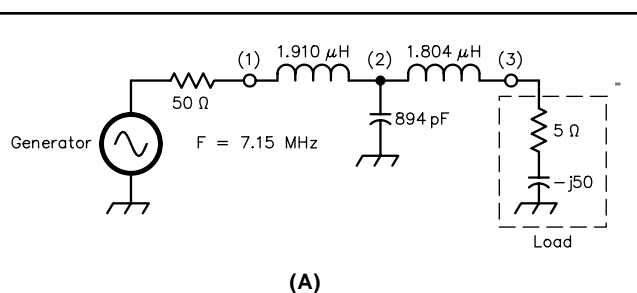
$$RL(\text{dB}) = -20 \cdot \log |RHO| = -20 \cdot \log \left(\frac{V - 1}{V + 1} \right) \text{dB} \quad (\text{Eq 3})$$

Return loss is a very sensitive measure of impedance match that is widely used in test equipment, such as network analyzers. This term means "what fraction of the power that is sent toward the load returns to the generator?" In *Radio Designer*, RL (dB) is the same as MS11 (dB) and MS22 (dB) at the input and output, respectively, of a two-port network. The values of RHO and V can be found for any combination of R and X or G and B, using the cursor method as described.

The Smith Chart is basically a reflection-coefficient (RHO) chart. The distance from the center to the outer circle corresponds to $|RHO| = 1.0$, which is defined as "complete" reflection of a wave, which corresponds to a short-circuit load ($R = 0$), an open-circuit load ($G = 0$) or a purely reactive load. The chart then assigns the R, X, G and B values in terms of the corresponding complex values of RHO according to the equation

$$Z = Z_0 \cdot \frac{1 + RHO}{1 - RHO}; Y = \frac{1}{Z}; \text{ RHO and } Z_0 \text{ possibly complex} \quad (\text{Eq 4})$$

The method described for RHO, steps 1, 2, 3 and 4, can be



(A)

* Smith chart example

```

BLK
IND 1 2 L=?1.94198UH? Q=250 F=7.15MHZ
CAP 2 0 C=?860.327PF? Q=10000 F=7.15MHZ
IND 2 3 L=?1.83567UH? Q=250 F=7.15MHZ
TUNER:2POR 1 3
END
FREQ
STEP 7.0MHZ 7.35MHZ 10 KHZ
END

OPT
TUNER R1=50 Z2=5 -50 MS11
F=7.15MHZ MS11= -50
END
* Comments:
* Set output load in Report Editor to 5 - j50 ohms
* Set generator in Report Editor to 50 ohms
* Plot MS11 and MS21

```

(B)

Fig 1—A is a circuit listing for an example two-port network for Smith Chart analysis. B is a schematic of the circuit.

used to find values of Z and Y at some location. Type “Y” or “Z” in the CCW, then Execute, place the cursor at a location and press the “M” key. A marker appears on the chart and the R and X, or G and B, values appear in the CCW. If a marker is not wanted, press the left mouse button instead. This operation suggests an easy way to transform a series R_S and X_S to a parallel R_P and X_P because $R_P = 1 / G$ and $X_P = -1 / B$ at any selected location: Working this backward, we can change parallel R_P to $G (= 1 / R_P)$ and X_P to $B (= -1 / X_P)$, then to R_S and X_S by doing the Y to Z change.

To place a marker at a specific value of $Z = R + jX$ (or $Y = G + jB$) do the following:

- Create the appropriate R (or G) circle and an X (or B) circle, as previously described.
- Place the cursor exactly at the intersection and remove your hand from the mouse.
- Delete the two circles (“DEL 2”) if you do not want them to show.
- Press the “M” key. This places a marker at the Z (or Y) location.

Fig 2D shows arcs of constant Q values. Any values of R and +X that lie on the $Q = +2$ line correspond to $X / R = Q = 2$: for example, the intersection of the $R = 1.5$ circle and the $X = +3$ circle. For capacitive values of X, $-Q$ is plotted

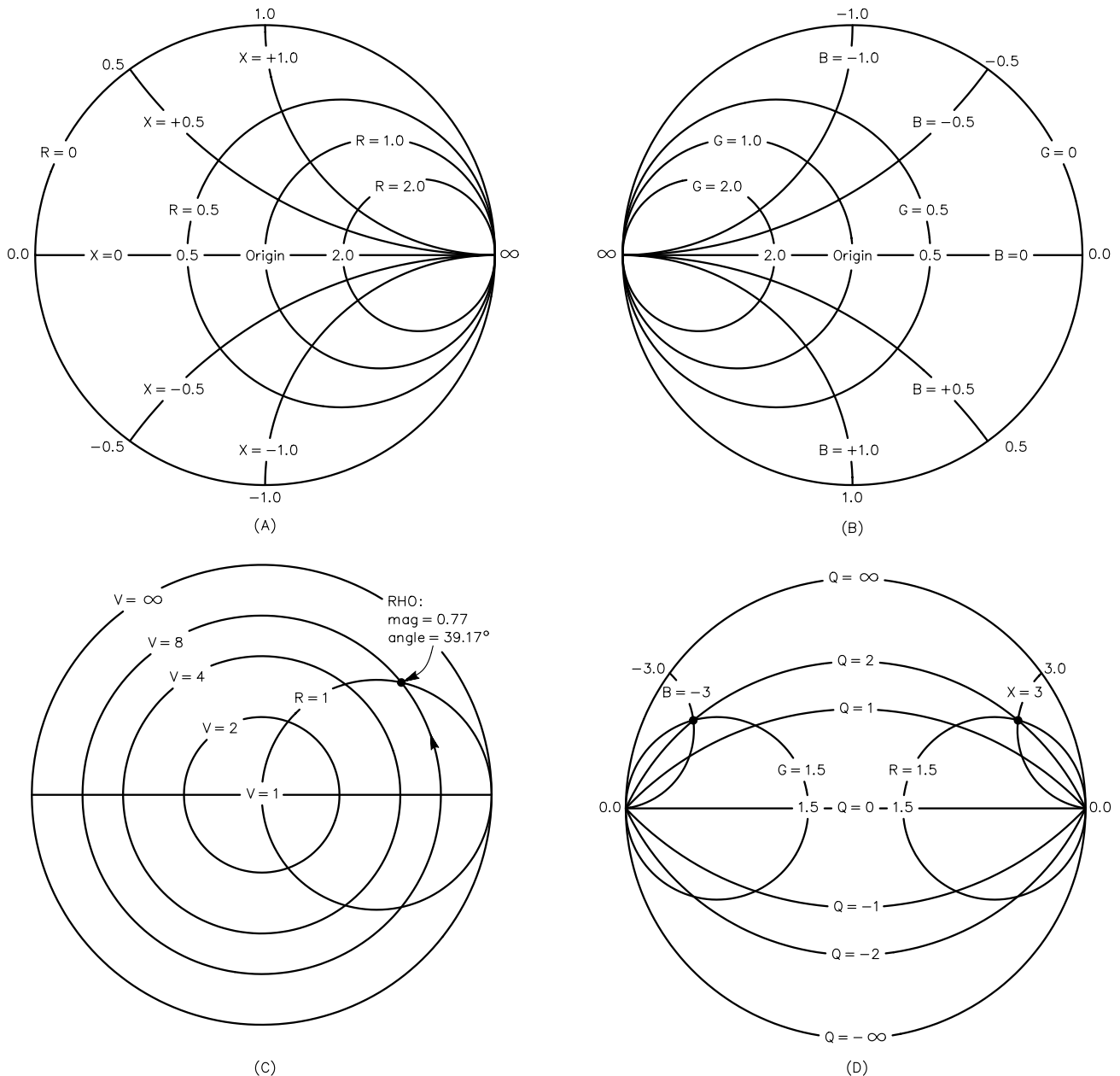


Fig 2—A shows Smith X Chart circles of constant resistance, R, and reactance, X. Positive X is above the horizontal axis. B shows Smith Y Chart circles of constant conductance, G, and susceptance, B. Positive B is below the axis. C shows Smith Chart circles of constant SWR (V) and an example of a reflection-coefficient, RHO, location. D shows Smith Chart arcs of constant Q. Positive values of Q apply to inductive reactance or susceptance. Negative values of Q apply to capacitive reactance or susceptance.

below the horizontal axis. The intersection of $B = -3$ and $G = 1.5$ is also at $Q = B / G = 2$. Q lines are useful in certain applications that are discussed in Part 2 of this article.

The CCW contents and the complete circle that we create can be saved to disk by typing the SPLT command in the CCW. A name for the file is requested. This same file can be recalled from disk by using the RPLT command. The name of the file is requested.

Navigating the Chart

An important topic in Circles operations concerns the ways that we modify the Z , Y , R , X , G and B quantities from one chart location to another. Fig 3A shows an initial value of Z at one location. We want to change Z to any of the values $Z1$ through $Z5$. The value of $Z = R + jX$ is at the junction of the $R = 0.18$ circle and the $+X = 0.26$ circle, where $+X$ is inductive. The following rules are observed in Fig 3A:

The impedance, Z , is at the junction of a constant- R circle and a constant- X circle. To increase X (make the inductive reactance, therefore the inductance, larger), move clockwise along the $R1$ circle to point $Z1$. The change in reactance is $0.91 - 0.26 = +0.65$. (Inductive reactance increases in the clockwise direction.)

The Circles Utility has another option, called DX, which works as follows:

- Enter a frequency in the CCW, for example $FREQ = 7.15E6$, then Execute.

- Type DX into the CCW, then Execute.

- Place the cursor *first* at Z and press the left button.

- Move the cursor to $Z1$ and press the left button again.

- The CCW displays the value of inductance, $L = DX / (2 \pi FREQ)$, that produces the change of reactance from Z to $Z1$ at the FREQ that was entered.

Instead of moving to $Z1$, we can move from Z to $Z2$ along the $R1$ circle. This is a capacitive reactance of value -0.66 . (Capacitive reactance increases —C decreases—in the counterclockwise direction.) The DX option now gives the needed value of capacitance, provided that we place the cursor first at Z , then at $Z2$. We can also move from $Z1$ to $Z2$ for a reactance change of -1.31 .

To increase the value of R , from locations Z , $Z1$ or $Z2$, move counterclockwise along a line of constant X , as shown, if X is positive (above the horizontal axis). If X is negative, the motion is clockwise. The proper directions to reduce R are obvious in Fig 3A.

Having arrived at $Z3$, $Z4$ or $Z5$, the preceding operations can be repeated, and in this manner we can travel around the chart.

Fig 3B shows the travels for the Y chart. The rules are nearly the same as for the Z chart, and the directions are as shown. The differences are:

- Inductive susceptance increases (inductance decreases) in the counterclockwise direction along a line of constant conductance.

- Capacitive susceptance increases (capacitance increases) in a clockwise direction along a line of constant conductance.

- Conductance (G) increases in a clockwise direction along a B line for B less than zero (above the horizontal axis) and counterclockwise along a B line for B greater than zero (below the horizontal axis).

The next task is to move back and forth between the Z chart and the Y chart, using these steps:

- Starting at some Z point, move Z to $Z1$ (Fig 3A). Then, using the CCW, find the value of $Y1$ at this point.

- Then move $Y1$ to $Y2$ (Fig 3B).

- Find the value of $Z2$ at $Y2$ and then move $Z2$ to $Z3$, and so forth until the target is reached.

Many impedance-matching problems start at a load impedance somewhat removed from the center of the chart and work toward the chart center, $R = 1$, $X = 0$ or $G = 1$, $B = 0$. (We assume that the chart center is the impedance that the generator wants to “see.” This process models adjustment of an antenna tuner. The target impedance need not be at the chart center. For example, if $Z = 1.1 + j0.05$ or its complex conjugate, place a Z marker there and make that the target. It’s equally possible to place the load at or near the chart center and the generator farther out. In that case, we start at the center and work out toward the generator. When we do this reversal of direction, a series inductor

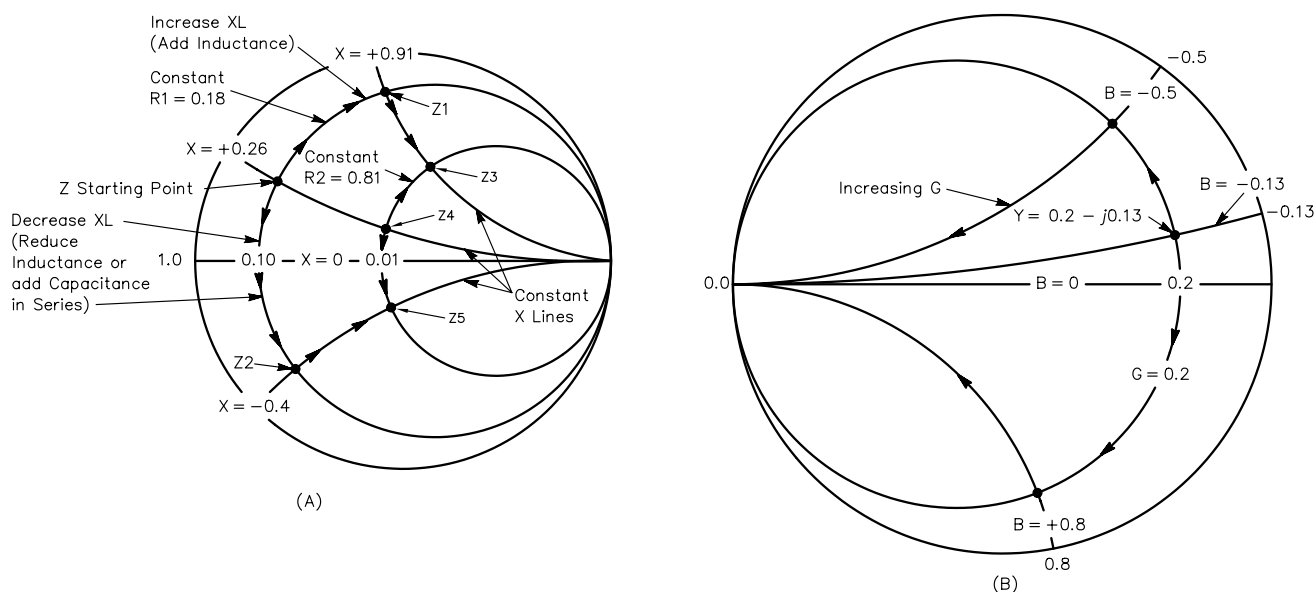


Fig 3—A shows directions of travel on a Z chart for X and R changes by adding or subtracting series XL , XC or R . B shows directions of travel on a Y chart for B changes by adding or subtracting shunt BL or BC .

becomes a series capacitor, and a shunt inductance becomes a shunt capacitance, and so forth.

For example, a 50 Ω load is to be transformed to some complex impedance that a transistor collector (the generator) wants to “see.” Similarly (a little more difficult to visualize but very important) a transistor *input* looks backward toward a transformed 50 Ω (the load that the transistor input sees, looking back). In order to avoid confusion and errors the rule is: Start at the load, wherever it is, and work toward the generator, wherever it is.

The changing of impedance using transmission lines and stubs is covered by three additional Circles functions as follows:

DT finds the electrical length (in degrees) and characteristic impedance [$Z_0(n)$ when normalized to 1.0, $Z_0(u)$ for a 50 Ω line] of a transmission

line that transforms impedance Z_1 to Z_2 . Z_1 and Z_2 can lie on the same V (SWR) circle as shown in Fig 4A, or they may be on different V circles. Use the following approach for the first option: on the same V circle:

- Draw the V circle
- Locate Z_1 and Z_2 on the V circle
- Type DT in the CCW, then Execute
- Place cursor at Z_1 and click left button, then place cursor at Z_2 and click left button again. The clockwise electrical angle is measured. Alternatively, do Z_2 first then Z_1 , in which case the angle is measured clockwise from Z_2 to Z_1 . The complete circle is 180° electrical length.

The information appears in the CCW text window. The physical length is:

$$(\text{meters}) = \frac{(\text{degrees}) \cdot (\text{velocity factor}) \cdot 0.833}{\text{FREQ (MHz)}} \quad (\text{Eq 5})$$

Note also that the two Z points need not be on the same V circle. If they are on two different V circles (but not $V = \infty$) the correct Z_0 of the line and its length (in degrees) will be calculated. Two lines may be required. This valuable feature is often used, especially in microstrip design.

Referring to Fig 2A, the difference between two points on a constant-resistance circle is a value of reactance. After we mark two reactance points on this circle (first the initial value, then the final value), the DXD operation allows us to get this change of reactance with a section of transmission line connected as a stub in series with the line, as shown in Fig 4B. Quite often two stubs—each having one half of the needed reactance—can be installed, one in each wire of a balanced transmission line as shown in Fig 4C. The stub(s) can be:

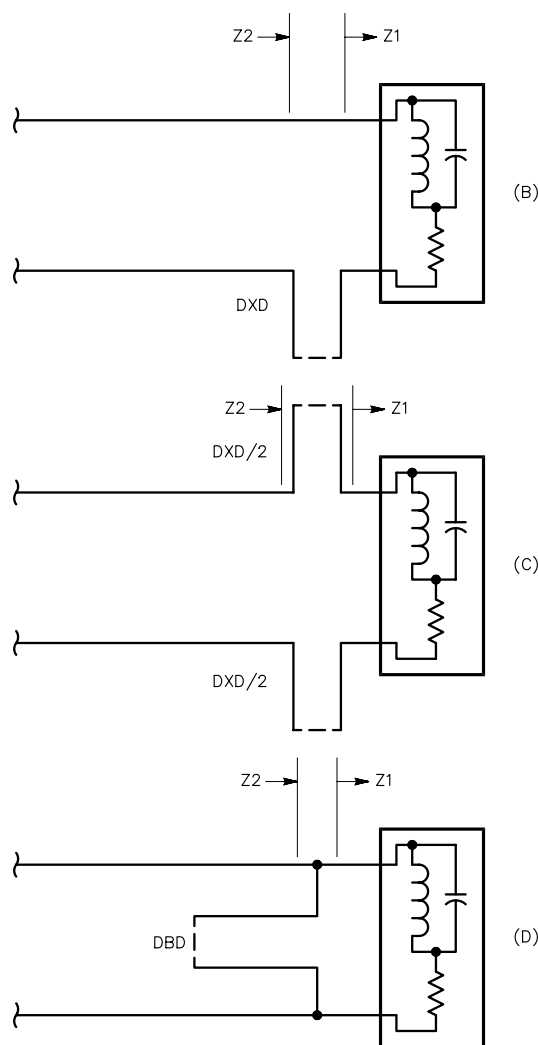
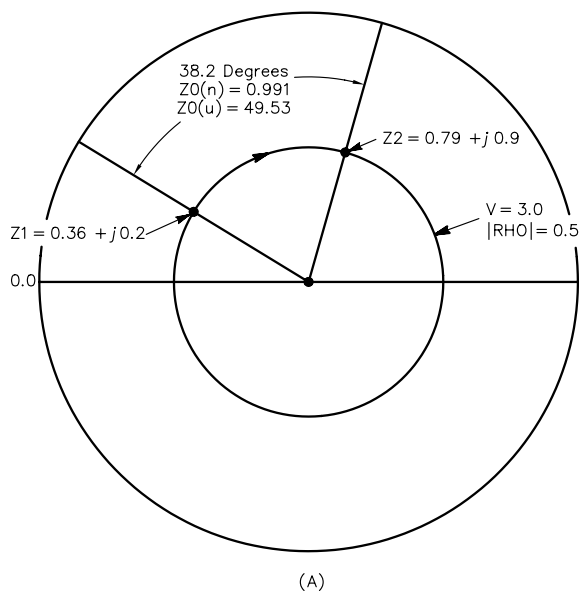


Fig 4—A shows the transmission line required to transform Z_1 to Z_2 along a line of constant SWR, where $V = 3.0:1$ ($RHO = 0.5$). Z_0 (unnormalized) = 49.53 Ω, Z_0 (normalized) = 0.99. B (unbalanced) and C (balanced) show the DXD function; D shows the DBD function.

- A shorted stub (SS) less than $\lambda/4$ is an inductive reactance.
- An open stub (OS) less than $\lambda/4$ is a capacitive reactance.

The CCW asks for one of the following: I (= Z_0 , impedance of the transmission line), D (= degrees of electrical length) or W (= wavelength). The CCW then tells which stub to use (OS or SS), its reactance and other parameters.

The DBD operation is just like DXD, except that we put two susceptance markers on a constant-conductance circle as in Fig 2B. We then get a parallel stub (OS or SS) as shown in Fig 4D. The reactance of the stub and other parameters are displayed. This kind of stub is widely used.

The combined usage of DT, DXD and DBD in the Circles Utility is a very powerful way to match impedances using transmission lines. By switching from one to another (and between Z and Y) we can travel the chart in transmission-line segments and stubs in a very elegant way. It is worthwhile to keep in mind that the reactances of transmission-line segments vary over frequency in somewhat different ways than ordinary lumped LC components. This can be a factor in designing a network.

Problem Solving

The art and gamesmanship of these matching exercises are to find the minimum number of components, especially lossy inductors. Always keep the possibility of using transmission-line segments in mind because of their high Q (low loss) values. Sometimes a greater frequency-response bandwidth is achieved by using transmission lines than lumped L and C. We also want values of L, C and lines that are realistic, efficient and economical.

Fig 5 (drawn with fine lines to improve accuracy) illustrates the many different possibilities for converting a load impedance Z_{load} to $R = 1, X = 0$. We will use, just as one example, the circuit shown in Fig 1B. Z_{load} is at point A, $0.10 - j1.0$. The first component is a series inductor along the $R = 0.10$ line from A to C. A shunt capacitor moves the impedance along the $G = 0.5$ line from C to I, and a series inductor moves it along the $R = 1.0$ line from I to O (the origin). Fig 1B shows the values calculated in this manner. This circuit would have a very good low-pass filtering property, but requires three components, and at least two of them must be tunable. These values are placed in the circuit listing in Fig 1A and then optimized for best MS11 at 7.15 MHz, using realistic Q values

for the components. MS21 tells the circuit loss in decibels, about 0.35 dB (when MS11 dB is very large). The optimized L and C values are shown. The last inductor “IND 2 3” could have a constant value (remove the question marks from the Netlist), and the other L and C are then optimized.

The first coil from point A could be a fixed value if it gets us beyond point B. This inductance could possibly be built right into the load device (antenna or whatever). A length of transmission line ($Z_0 = 1.0$) could get us from, say point C, along a constant V circle, over to the $R = 1$ circle. A simpler approach could use series inductance up to point B and shunt capacitance from B to O. Better yet, a series coil from A to C and two low-loss capacitors, one in shunt from C to F and one in series from F to O. If the series inductive reactance doesn't go beyond point B at the lowest frequency of interest, however, these schemes won't work. Another interesting idea would be a shunt coil from A to K and a series coil from K to O.

Consider also an inductance from A up to the horizontal axis (the 0.1 mark) and a 9:1 (impedance ratio) ferrite-core transmission-line transformer

from 0.1Ω to 0.9Ω , where a V circle has the low value 1.11. This transmission line should ideally have a Z_0 of 0.3Ω . For a $50\text{-}\Omega$ system, this would correspond to 15Ω . Three small, parallel-connected segments of 50Ω coax would be okay. We can do some fine-tuning with this method. The resistance added by the inductor can be considered for a more-accurate graphical answer. Use the following approach:

- Get a first estimate of the normalized inductive reactance (= 1.0) as described above.
- Assume a value of Q_L for the inductor. The R_L of the coil is then $R_L = X_L / Q_L$.
- Add R_L to the resistance R (= 0.1) of the load and relocate point A on the Smith Chart accordingly.
- On the chart, find a better value of the inductor needed to reach the horizontal axis.
- Further repetitions of this procedure are not necessary.
- The 1:9 transformer (assume it to be loss-less) now takes us to a new value, which will still be close to the origin if Q_L is large.

In this example a coil Q_L of 250 would mean a correction of $1.0/250 = 0.004 \Omega$,

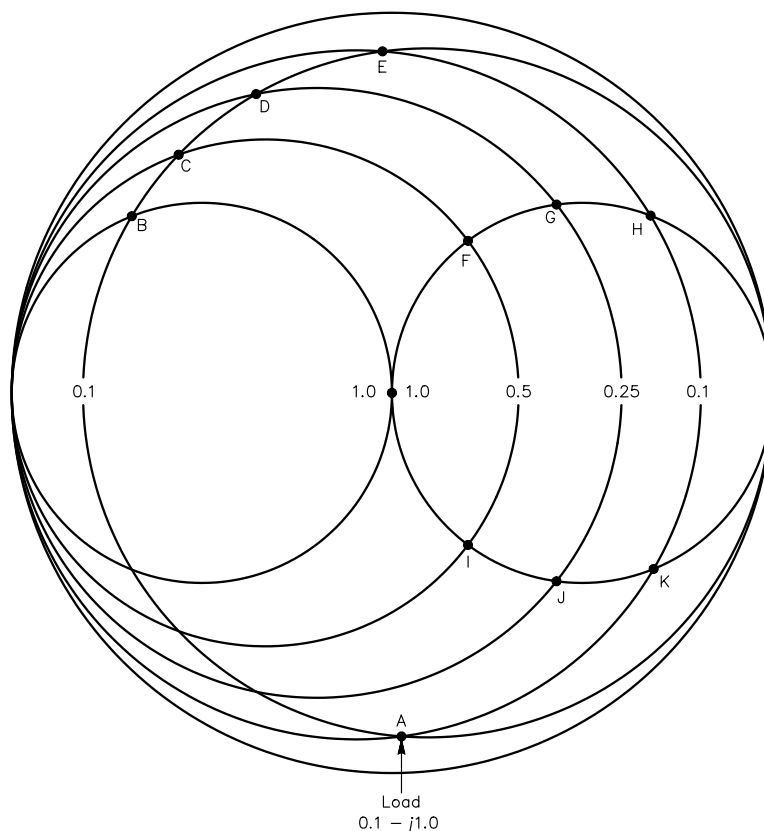


Fig 5—Solution paths for the design example.

which is negligible. In more extreme situations it may be important.

Keep in mind also the discussion regarding the reversal of direction: for example, starting at the origin (the load) and working outward to the generator impedance or possibly its complex conjugate.

The main idea here is to illustrate the power of the Smith Chart in visualizing the myriad possible solutions, each of which has possible merits and possible problems. With such a “shopping list” the designer can make the best decisions. The Circles Utility can then quickly and easily get values for the various components.

Also, a candidate circuit of lines/stubs and adjustable L and C should be tested over a frequency band for “tuning range” of the components. A new Designer file is created as shown in Fig 6 for this purpose. At any arbitrary frequency that is entered into the OPT block, the circuit values are optimized for maximum MS11. The load impedance data is in the DATA block. This data is linked to the ONE circuit element that has been placed in the main circuit block. We can now see what range of L and C values are needed to tune the desired frequency range.

In the example of Fig 6, “IND 2 3” is a fixed value (by assumption, it can’t be tuned) and it was necessary to increase its fixed value (see Fig 1A) so that we could tune to the low end of the 7.0 to 7.3 MHz band with reasonable values of the other components. This is typical of problems that we might encounter.

Another interesting exercise is to design a network that has a certain V (SWR) over a certain frequency band (without retuning). Using Eq 3, a V = 2 value corresponds to an MS11 (dB) of -9.5. A V = 1.5 value corresponds to an MS11 (dB) of -14. By observing the plots of MS11 (dB) versus frequency generated by Designer, the -9.5 dB and -14 dB frequency ranges can be easily observed. It is usually necessary to “Rescale” the graph, to see these levels more easily. For further information on the art and witchcraft of broadband impedance matching, see References 2 and 3.

A Minor Bug

The Circles Utility has a problem when performing the DT (transmission line) operation over a small region of a V (SWR) circle close to the horizontal (X = 0) axis (see Fig 4A). Incorrect results may appear, due to a minor glitch in the software. The correct pro-

* Band tuning example

```
BLK
IND 1 2 L=?3.92257UH? Q=250 F=7.15MHZ
CAP 2 0 C=?414.698PF? Q=1000 F=7.15MHZ
IND 2 3 L=2.3UH Q=250 F=7.15MHZ
* Note that this L is held constant
ONE 3 0 ZDAT
TUNER:1POR 1 0
END

FREQ
STEP 6.9MHZ 7.4MHZ 10KHZ
END

OPT
TUNER R1=50 MS11
F=7.25MHZ MS11= -50
END

DATA
ZDAT: Z RI
* Freq Real Imag
7.00MHZ 3 -70
7.10MHZ 4 -60
7.20MHZ 6 -40
7.30MHZ 8 -30
END

* Comments:
* Put the ONE element in the circuit block.
* Change TUNER to a 1POR 1 0 (one port) as shown.
* Set Report Editor to MS11 with TERM=50+j0
* Put load impedance data in the DATA block, real then
imaginary.
* DATA is interpolated between freq entries.
* In the OPT block, insert any freq value in the range,
one at a time.
* Analyze, then Optimize at the freq that is in the OPT
block.
* Check to see the min and max values of tuning Ls and Cs
that are needed.
* Check to see that MS11 reaches a very large value.
```

Fig 6—Net list for band tuning example.

cedure in this situation is to perform DT in two segments. The first segment begins “away from” the horizontal axis and terminates “at” the horizontal axis. The second segment begins “at” the horizontal axis and proceeds to the desired finish point. The total electrical angle of the transmission line is then the sum of the lengths of the two segments. Some practice will make this a simple operation. Two RHO operations on a V circle also provide the DT electrical angle with no error. This problem does not occur often, but you should be aware of it.

Additional Reading

R. Dean Straw, Editor, *The ARRL Antenna Book*, 18th edition, Chapter 28, has an excellent general discussion of the Smith Chart. ARRL Order No. 6133. ARRL publi-

cations are available from your local ARRL dealer or directly from the ARRL. Mail orders to Pub Sales Dept, ARRL, 225 Main St, Newington, CT 06111-1494. You can call us toll-free at tel 888-277-5289; fax your order to 860-594-0303; or send e-mail to pubsales@arrl.org. Check out the full ARRL publications line on the World Wide Web at <http://www.arrl.org/catalog>.

Wilfred Caron, *Antenna Impedance Matching* (Newington: 1989, ARRL), Order No. 2200. This book has very thorough coverage of matching methods using transmission lines and LC components.

W. E. Sabin, WØIYH, “Broadband HF Antenna Matching With ARRL Radio Designer,” *QST*, August 1995, p 33.

W. E. Sabin, WØIYH, “Computer Modeling of Coax Cable Circuits,” *QEX*, August 1996, pp 3.

W. E. Sabin, WØIYH, “Understanding the T-tuner (C-L-C) Transmatch,” *QEX*, December 1997, pp 13.

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ARRL Radio Designer *and* *the Circles Utility, Part 2:* *Small-Signal Amplifier Design*

Have you been intimidated by amplifier design? Maybe you want to step up to ARRL Radio Designer for the task. Either way, this will help you grasp this powerful design tool.

By William E. Sabin, WØIYH

Part 1 of this article was intended to help the reader to acquire greater familiarity with the Circles Utility of *Radio Designer*. We looked at passive network design using transmission lines and LC components, using the various methods of navigating the Smith Chart. In this Part 2 we will consider the interesting and very useful methods of designing small-signal linear amplifiers using gain circles, stability circles, noise-figure circles and various figures of merit. These techniques are well known and widely used in industry, and we will focus on the par-

ticular features of the *Radio Designer* program in this regard.

S-Parameter Basics

We begin with a brief overview of S-parameters, especially as they relate to *Radio Designer*. Fig 1 shows a two-port network (we will call it "N") that is connected to a generator and a load. The generator has a resistance R_G and is connected to the input port through a transmission line having a characteristic impedance Z_{01} . The output is connected to a load R_L through a transmission line Z_{02} . If the generator sends a voltage wave V_1^+ toward N, a current wave, $I_1^+ = V_1^+/Z_{01}$, accompanies it. The power in this forward wave is $P_1^+ = V_1^+ I_1^+ = (V_1^+)^2 / Z_{01} = (I_1^+)^2 Z_{01}$. A reflected voltage wave, $V_1^- = -I_1^- Z_{01}$,

travels back to the generator ($-I_1^-$ is correct because I_1^- moves in phase with V_1^- in a direction opposite to I_1^+). If we square both sides of this equation and divide both sides by Z_{01} we get

$$\frac{(V_1^-)^2}{Z_{01}} = (-I_1^-)^2 \cdot Z_{01} = P_1^- \quad (\text{Eq 1})$$

This is a power wave that is reflected from N back to the generator. We see that V_1^+ and I_1^+ are in-phase (Z_{01} is a resistance) and V_1^- and $-I_1^-$ are also in-phase, but V_1^+ and V_1^- (and I_1^+ and $-I_1^-$) may not be in-phase with each other. Both waves are in fact traveling simultaneously in opposite directions (they are easily measured independently using a directional coupler) but at any point the net voltage (or current) is the vector sum of the

two waves. The difference between P_1^+ and P_1^- is the power that is accepted by N, and it is also the power that is delivered by the generator. At the output of N, Fig 1 shows a power wave traveling from the load toward N and a power wave from N to the load, and the same rules apply here.

The square roots of these four power waves are voltage waves that are said to be *normalized* with respect to Z_0 .

$$a_1 = \frac{V_1^+}{\sqrt{Z_{01}}} \quad ; \quad b_1 = \frac{V_1^-}{\sqrt{Z_{01}}} \quad ; \quad a_2 = \frac{V_2^+}{\sqrt{Z_{02}}} \quad ; \quad b_2 = \frac{V_2^-}{\sqrt{Z_{02}}} \quad (\text{Eq 2a})$$

They can also be expressed as normalized current waves

$$a_1 = I_1^+ \cdot \sqrt{Z_{01}} \quad ; \quad b_1 = I_1^- \cdot \sqrt{Z_{01}} \quad ; \quad a_2 = I_2^+ \cdot \sqrt{Z_{02}} \quad ; \quad b_2 = I_2^- \cdot \sqrt{Z_{02}} \quad (\text{Eq 2b})$$

Although Z_{01} and Z_{02} are often the same, especially in network analyzers, they need not be the same. In *Radio Designer* the Report Editor allows us to specify the Terminations (Z_0) separately. For example, if N is a loss-less 25:1 impedance-ratio matching network from a $200 + j0 \Omega$ generator to an $8 + j0 \Omega$ load we can set the Input Termination at $200 + j0 \Omega$ and the Output Termination at $8 + j0 \Omega$. *Radio Designer* then tells us that this particular N has no *mismatch loss* (the impedance matches at the input and output are perfect).

If the input impedance of N in this example is not 200Ω , the power available from the 200Ω generator will not be accepted by N. There is a mismatch loss. An exactly identical statement is that the power not accepted by N is reflected and returns to the generator. The same is true at the output. If we move the voltage generator to the right side (leave $R_G = Z_{01}$ and $R_L = Z_{02}$ where they are) the same situation exists at the right-hand side of N. If the output impedance of N is not 8Ω , some power is reflected at the output terminals. In practice, we routinely make this switch of the generator so that we can measure the transmission and reflection properties of N at the output side.

Fig 1 shows a_1 and a_2 as normalized voltage waves representing the signals arriving at N and b_1 and b_2 as normalized voltage waves leaving N. The other thing we see is that the portion of a_1 accepted by N (from the generator on the left) is transmitted, after impedance transformation, to the output terminals. A portion of a_2 is transmitted (from generator on the right), after impedance transformation, from output back to input.

We need one other thing to complete the picture. If the load $R_L = Z_{02}$ then a wave b_2 will be completely absorbed by R_L . If $R_G = Z_{01}$ a wave b_1 will be absorbed by R_G . If this is not true, *re-reflections* occur at R_L or R_G , and the situation becomes a little more complicated. In test equipment,

the effect of these re-reflections is that they create errors in our measurements of the S-parameters (unless we compensate for them). In *Radio Designer*, we will always assume that $R_L = Z_{02}$. An identical situation occurs on the input side and we make $R_G = Z_{01}$. Because $R_L = Z_{02}$, it is true that $a_2 = 0$ when the generator is connected at the input. Also, because $R_G = Z_{01}$, $a_1 = 0$ when the generator is connected to the output. *Radio Designer* takes care of this.

We can now define the S-parameters, which we “customarily,” although not necessarily, do in terms of voltage waves as follows:

$$S_{11} = \frac{\frac{V_1^-}{\sqrt{Z_{01}}}}{\frac{V_1^+}{\sqrt{Z_{01}}}} = \frac{V_1^-}{V_1^+} = \frac{b_1}{a_1} \quad ; \quad a_2 = 0 \quad (\text{Eq 3})$$

This is the complex ratio of a reflected wave leaving port 1 to a wave arriving at port 1 with no wave arriving at port 2 (no reflection from R_L , and the generator is at the left). It is also known as the reflection coefficient Γ_{IN} and in *Radio Designer* Circles Utility as RHO. Input SWR can be found from $|S_{11}|$ [SWR = $(1 + |S_{11}|) / (1 - |S_{11}|)$ —Ed.]. S_{11} can be plotted on a Smith Chart.

$$S_{22} = \frac{\frac{V_2^-}{\sqrt{Z_{02}}}}{\frac{V_2^+}{\sqrt{Z_{02}}}} = \frac{V_2^-}{V_2^+} = \frac{b_2}{a_2} \quad ; \quad a_1 = 0 \quad (\text{Eq 4})$$

This is the complex ratio of a reflected wave leaving port 2 to a wave arriving at port 2 with no wave arriving at port 1 (no reflection from R_G and the generator is at the right). Also known as the reflection coefficient Γ_{OUT} and in *Radio Designer's* Circles Utility as RHO. Output SWR can be found from $|S_{22}|$. S_{22} can be plotted on a Smith Chart.

$$S_{21} = \frac{\frac{V_2^-}{\sqrt{Z_{02}}}}{\frac{V_1^+}{\sqrt{Z_{01}}}} = \frac{b_2}{a_1} \quad ; \quad a_2 = 0 \quad (\text{Eq 5})$$

This is the complex ratio (gain or attenuation) of a wave leaving port 2 to a wave arriving at port 1 with no wave arriving at port 2 (no reflection from R_L). S_{21} is not an impedance or a reflection factor, so it is plotted on a polar graph. The quantity $|S_{21}|^2$ is referred to in transistor data sheets as the “intrinsic” (same as “transducer”) power gain,

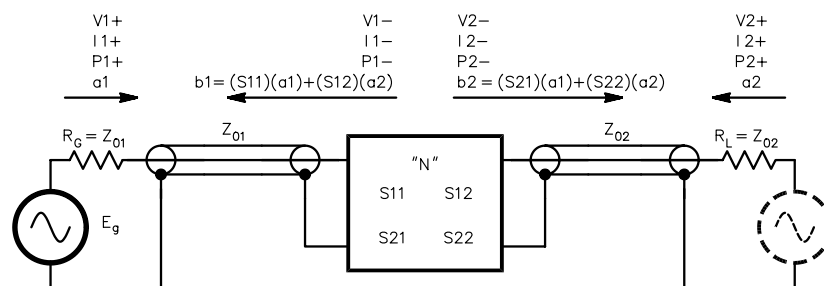


Fig 1—Definitions of terms used in S-parameter analysis and design.

when measured with a network analyzer. “Available” and “actual” power gain are, in most cases, different from this measurement (see later discussion).

$$S_{12} = \frac{\frac{V_1^-}{\sqrt{Z_{01}}} = \frac{b_1}{a_2}}{\frac{V_2^+}{\sqrt{Z_{02}}}} \quad ; \quad a_1 = 0 \quad (\text{Eq 6})$$

This is the complex ratio of a wave leaving port 1 to a wave arriving at port 2 with no wave arriving at port 1 (no reflection from R_G).

• Using these definitions, we can write the two-port equations that are used in S-parameter analysis and design:

$$\begin{aligned} b_1 &= S_{11} \cdot a_1 + S_{12} \cdot a_2 \\ b_2 &= S_{21} \cdot a_1 + S_{22} \cdot a_2 \end{aligned} \quad (\text{Eq 7})$$

The wave leaving port 1 (b_1) is the sum of a portion of a wave (a_1) that is reflected by port 1 (S_{11}) and a wave (a_2) that is fed-through from port 2 (S_{12}). Similarly for b_2 .

• *Radio Designer* uses the following forms of S-parameters, using S_{11} as an example:

$$\begin{aligned} MS_{11} &= \text{magnitude of } S_{11} \\ PS_{11} &= \text{phase of } S_{11} \\ RS_{11} &= \text{real part of } S_{11} \\ IS_{11} &= \text{imaginary part of } S_{11} \end{aligned} \quad (\text{Eq 8})$$

S-parameters versus frequency can be plotted in rectangular form, using the Rectangular option in Report Editor. The Polar option plots magnitude and angle of S_{11} and S_{22} on a Smith Chart and S_{12} and S_{21} on a polar chart.

Gain Definitions

Before proceeding, we should say something about the various kinds of power gain that are used in *Designer*. Assume that *Designer* correctly terminates the device in Fig 2A (no reflections from source or load). The transducer or insertion power gain was previously defined as $|S_{21}|^2$. *Designer* uses “available” gain G_A and “actual” (aka “operating”) gain G_P . These are given by the equations

$$G_A = \frac{|S_{21}|^2}{1 - |S_{22}|^2} \quad \text{and} \quad G_P = \frac{|S_{21}|^2}{1 - |S_{11}|^2} \quad (\text{Eq 9})$$

If S_{22} or S_{11} is not zero, the discrepancies with transducer gain need to be appreciated. We will encounter these differences in this article.

The Small-Signal Amplifier

Fig 2A shows an amplifier gain block with a set of transistor S-parameters available as data in a Data Block of a *Designer* Circuit File. We want to look at some important properties of the amplifier. In Fig 2A the amplifier, as defined by its 50 Ω S-parameters, is terminated at the input and output with 50 Ω . Its forward “transducer” power gain, also called “insertion gain,” is $|S_{21}|^2$, and its reverse power “gain” or attenuation is $|S_{12}|^2$. Its input and output reflection coefficients are S_{11} and S_{22} . But suppose, in Fig 2B, that $Z_{IN} = Z_S = 100 \Omega$ and $Z_{OUT} = Z_L = 200 \Omega$. The impedance-matching networks at the input and output would increase the gain. The input network would produce 3 dB more gain and the output network would increase the gain 6 dB, for a total increase of 9 dB.

For this reason, we say that the input network has (is responsible for) a “gain” of 3 dB and the output network has a “gain” of 6 dB. Other input/output networks can be designed that produce smaller values of gain. The important thing for the Circles Utility is that on a Smith Chart a particular value of total amplifier gain, including the added gain that is due to the matching networks, is associated with circles. These circles are called *gain circles*. Fig 2B shows the input/output matching networks added to the amplifier.

It might be thought that we would always want to simultaneously impedance match the amplifier’s input/output to get the greatest possible gain. But for maximum linear power output capability, a wider bandwidth or for

stability reasons the required source impedance or load impedance may be a lot different. Gain, bandwidth and signal handling always involve trade-offs. The choices of gain circle values and input/output impedances are the tools that we use. Conventional or transmission-line transformer coupling is an important approach at frequencies where they can be used. The best noise figure is often obtained by correctly mismatching the input.

When working with the input/output matching networks, we will be using reflection coefficients rather than impedance values. We recall from Part 1 that an impedance on the Smith Chart corresponds to a certain value of complex reflection coefficient, Γ . We will also use such symbols as Γ_S , Γ_L , Γ_{IN} and Γ_{OUT} for source, load, input and output values, respectively. Fig 2B shows this usage, which is compatible with the usage of S_{11} and S_{22} , which are also reflection coefficients.

When working with gain circles, we must be aware of the effects of internal feedback (S_{12}) in the amplifier. This feedback can cause instability and oscillation if we are not careful. For this reason a different kind of circle, called a *stability circle*, is also plotted on the Smith Chart. There are two kinds of stability circles, *input stability circles* and *output stability circles*. The input stability circle is associated with values of the amplifier source impedance Z_S (or Γ_S) that might cause instability. The output stability circle is associated with values of load impedance Z_L (or Γ_L) that might cause instability. The main idea is to select points on the input and

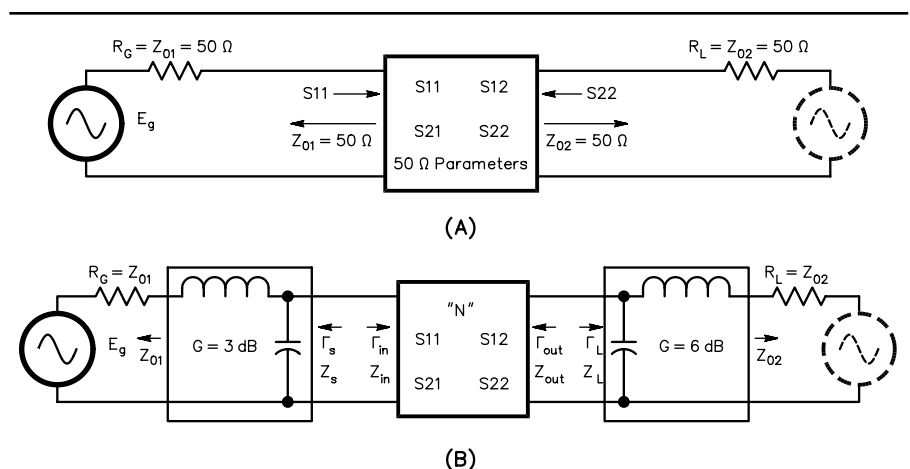


Fig 2—Amplifier with input and output matching networks. (A) without matching networks (B) with matching networks added.

output gain circles that are sufficiently far away from the input and output stability circles that there are no stability problems due to temperature or statistical variations or frequency changes.

Due to feedback, the input and output gain circles interact with each other. *Designer* uses a special procedure to plot these simultaneously in the presence of feedback. Because of this feature we need not be concerned about the so-called Unilateral Gain concept, which in times past assumed the $S_{12} = 0$ approximation in order to simplify the calculations. Another feature, called RHOL or RHOS, locates the source and load reflection coefficients that provide some desired gain value. As you can see, this entire procedure is a real can of worms when performed manually, but the Circles Utility removes nearly all of the pain.

To repeat, if S_{12} is zero, the feedback problem disappears and the amplifier is said to be *unilateral*. A transistor can possibly be neutralized (*unilateralized*) to reduce the effects of feedback. *Designer* tells us the maximum available gain AMAG possible or, if feedback is a problem, the maximum stable gain MSG. AMAG is the gain when both input and output of a stable transistor are conjugately matched to source and load. *Designer* also gives us a *stability figure of merit*, K , that gives us guidance regarding *conditional stability*. If $K > 1.0$, the stability circles are outside the chart and the entire chart is a stable region, there is no problem. Conditional stability ($K < 1$) means that everything may still be okay if we design the input/output networks carefully and stay far away from the stability circles. If S_{12} is small enough, the circuit is said to be *unconditionally stable*, which means that no combination of reactances at input and output can cause oscillation or excessive regeneration. If S_{12} is significant and we try to increase the gain-circle value to the point where an unstable region is entered, Circles denies entry and prints the message “MSG = xx.xx dB.” MSG is the maximum stable gain when $K = 1.0$. If we move into an unstable region, the gain would increase very rapidly beyond the MSG, and we get this message. A very important way to improve stability and increase K beyond the 1.0 mark is to modify the circuit in a way that reduces the product of forward gain and feedback (ie, loop gain). We will study one example of how to do this, but there are many ways.

Another issue is *tunability*, which is

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*UHF AMPLIFIER DESIGN USING THE CIRCLES UTILITY

BLK
    IND 1 2 L=?16.45NH?
    CAP 2 0 C=?4.39PF?
INNET:2POR 1 2
END
*****
BLK
    TWO 1 2 0 MRF901
    * RES 2 0 R=150
XISTOR:2POR 1 2
END
*****
BLK
    CAP 1 0 C=?1.68PF?
    IND 1 2 L=?23.12NH?
OUTNET:2POR 1 2
END
*****
BLK
    INNET 1 2
    XISTOR 2 3
    OUTNET 3 4
AMP:2POR 1 4
END
*****
FREQ
    435MHZ
    * STEP 420MHZ 450MHZ 5MHZ
END
*****
OPT
    AMP R1=50 R2=50
    MS11
    MS22
END
*****
DATA
    MRF901: S MP RREF=50
    * MRF901 0.4GHZ 0.5GHZ 2 50.0 1 VCE=10V IC=5MA
    * FREQ S11 S21 S12 S22
    0.2GHZ 0.63 -69 9.69 135 0.05 57 0.79 -25
    0.5GHZ 0.53 -135 5.65 101 0.07 43 0.54 -38
    * NOI RN
    * FREQ FMIN_DB MGOPT GOPT RN
    * 0.5GHZ 1.80 0.63 7 0.96
END

*Notes:
*In DATA block, S means S-parameters, MP means magnitude and phase,
* RREF means reference Z0 for S-parameters.

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Fig 3—Amplifier circuit description.

the ability to “tune” the input/output matching networks. Tunability can be poor if adjusting one circuit causes the other to become detuned because of S12 feedback. *Designer* helps to find both networks simultaneously. Variations in S12 over frequency and temperature, etc, can aggravate this problem.

Another consideration is the *noise figure* of the amplifier. If we put transistor noise data in a Data Block we can plot *noise figure circles* on the Smith Chart, each of which corresponds to a particular value of noise figure. Losses (L and C component Q) in the input matching network can also be included in the *Designer* circuit file to improve the accuracy of gain and noise-figure simulations.

To summarize, the main task for *Designer* is to find locations on acceptable gain circles that are sufficiently removed from instability regions and in regions of acceptable noise figure (a compromise). We then design input/output matching networks that implement the compromises, using the procedures discussed in Part 1 of this article. The L network is a simple one to implement, where it is adequate, but T, π and other multielement types have greater capabilities where selectivity and control of impedance versus frequency, in addition to impedance ratio, are important. In particular, selectivity at the input is often important in order to reduce interference. In some power amplifiers or second-stage amplifiers, noise figure may be a secondary consideration.

When designing the networks we may want to control the selectivity (Q), for example in a wideband amplifier. If we draw Q lines, for example +2 and -2, on the Smith Chart and confine all impedance (DX) and admittance (DB) paths to lie within these lines, there will be some limitation of selectivity. The networks may need more than two components, perhaps a T or π , to achieve the goal. We can also use multiple components to increase selectivity (use higher Q circles).

This entire procedure applies only to class-A linear amplifiers in which the values of S-parameters do not change with signal level. If these changes occur, special large-signal S-parameter procedures are used that are beyond the scope of this article.

The final result is a set of S-parameter and noise-figure data for the complete circuit of Fig 2B, repeated at each frequency of interest and using a specific set of L and C values (possibly transmission-line or microstrip seg-

ments) for the matching networks. The design must be tested over a frequency range to make sure that oscillation (or near-oscillation) cannot occur at some frequency that may possibly be far removed from the intended frequency band. In other words, we do not get too close to a stability circle. This design is then available as a 2POR (two-port device) for some larger system. One example of this would be a push-pull amplifier using hybrid splitters and combiners. Or, the final S-parameter and noise data can be used in a *Designer* Data Block that defines a new “device.” We can also easily get the S-parameters of a modified transistor. Stray Ls and Cs of the various components may also improve accuracy of the circuit description.

The Amplifier to be Simulated

Fig 3 is the *Designer* circuit file for the amplifier in Fig 2B, an MRF901 bipolar transistor (about \$1.60 in small quantities) with input and output matching networks that are to be designed for the 420-450 MHz (70 cm) band. The input network is INNET and the output network is OUTNET. The transistor is XISTOR, which contains a TWO circuit element that references the MRF901. The DATA block contains the S-parameter data for the transistor. This kind of information is found in data manuals (and *Designer's* built-in data files) over frequency and at various values of V_{ce} and I_c . *Designer* will interpolate between 200 MHz data and 500 MHz data to get data for the 70 cm band. The complete circuit is the AMP block. Each of these blocks can be edited and analyzed individually. The specific design frequency of interest is 435 MHz. After the networks are designed the frequency can be swept from 420 to 450 MHz to get the overall S-parameters versus FREQ for AMP as rectangular graph plots, polar plots or as data tables.

If we wish to, we can see what the transistor's S-parameters look like at 435 MHz or at several frequencies. We can do this by analyzing the XISTOR block separately. Use Report Editor to set up a table that displays XISTOR's four S-parameters, magnitude (not decibels) and angle. The Terminations for this should be 50 Ω . Ordinarily, Circles will calculate these automatically as it does its work. We do have to specify the various frequencies. If we modify the transistor by adding resistors or parasitic Ls and Cs we can, if we wish, easily get the new S-parameters and use them in our design work.

There are some problems about noise simulations. Many fine transistors such as the MRF901 do not explicitly give all of the NOI parameters versus frequency that *Designer* needs to create noise-figure circles (it may be possible to estimate them from the data sheets). It is more common to see noise-figure data or circles presented as tables or graphs in data manuals (eg, *Motorola RF Device Data*, DL110/D REV 7). They also indicate the complex source-impedance or reflection-coefficient Γ_S needed to get those noise figures. In this situation it is probably best to work back and forth between *Designer* and the data manual to get the best result. I will illustrate the method in this article. Fig 3 gives an example of how noise data might be placed in a DATA block, when it is available. See the Noise Figure sidebar.

Putting the Circles Utility to Work

Create the circuit file 2-port network in Fig 3 and Analyze it at the single FREQ 435MHZ. Put an asterisk before the 420 to 450 frequency sweep to eliminate it from the analysis. The 150 Ω resistor should also be “commented out” with an asterisk. Use Report Editor to create a graph of MS11, MS22, MS12 and MS21, all in decibels, for the AMP block (the complete circuit). Bring up the Circles Utility. Fig 4 shows the appearance of the chart as we perform the following steps.

1. In the CCW, type R 1 and X 0 to get a couple of red reference lines on the chart.
2. Type CKT XISTOR. This selects the MRF901 transistor to be analyzed all by itself at 435 MHz. The rest of the circuit is ignored.
3. See Fig 4A. Type S1, then S2. These are the input stability circle (blue) and output stability circle (grey) for the MRF901. They fall inside the Smith Chart and the CCW says that the regions outside the circles are stable. So in future steps we will operate only in those regions. Regions inside the circles produce conditions that are prone to oscillation. Sometimes only the regions inside are stable and the CCW will tell us so.
4. We also see in Fig 4A that XISTOR, at this point, is terminated only with 50 Ω (the center of the Smith Chart). Therefore, XISTOR by itself is stable. This points out a major reason for using S-parameters to characterize the active device. We should always verify this, especially for multiple-transistor “devices” that we create.

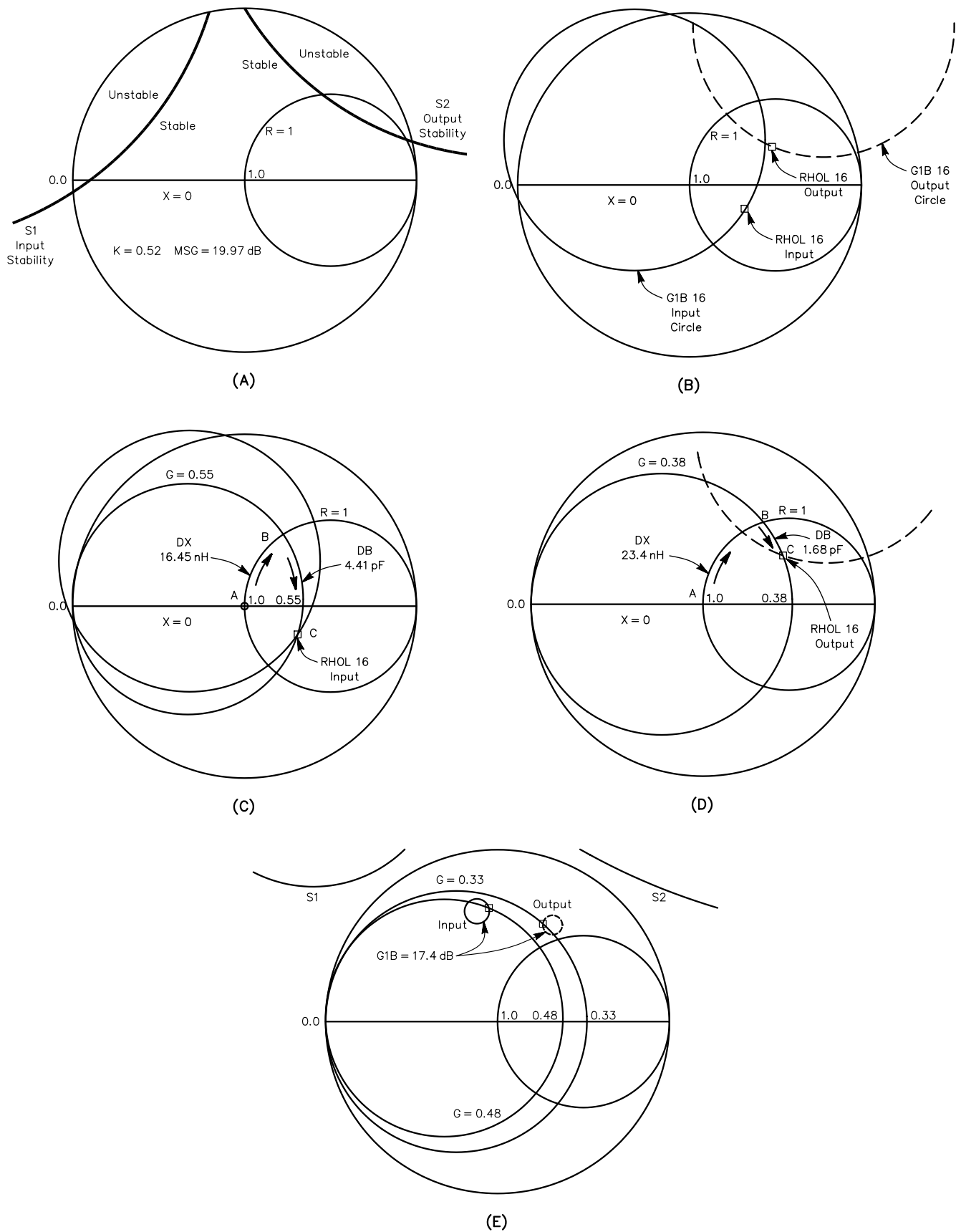


Fig 4—(A) shows stability circles for the example. (B) input and output-gain circles, input and output matching-network reflection coefficients. (C) designing the input matching L-network. (D) designing output matching L-network. (E) amplifier with 150Ω loading resistor.

5. Type K. This is the stability factor and it is 0.52. A number less than 1.0 indicates conditional stability (possible instability). The stability circles being inside the Smith Chart also suggests that K is less than 1.

6. Type GMAX and get the result MSG = 19.97 dB. MSG (maximum stable gain) is the largest possible gain that will not result in instability. The message GAMAX (maximum available gain) corresponds to unconditional stability. Circles will not allow a gain greater than GAMAX, or MSG if that is smaller than GAMAX.

7. We want to decide a value of gain for the amplifier that is less than MSG and is "sufficiently" (judgment call) removed from the stability circles. See the Gain Circles sidebar. In this example, we need to use bilateral gain circles because of the internal feedback of this transistor (as discussed earlier). G1B refers to "available power gain G_A ," which is used in many applications. The "B" stands for "bilateral." We could also use G2B, which refers to "actual power gain G_P ," which is also frequently used. We will use G1B for this example.

8. We will pick a G_A gain value of 16 dB. See Fig 4B. Type G1B 16 in the CCW. We get two gain circles, a solid yellow line for the input and a dashed yellow line for the output. For each point on one circle there is a corresponding point on the other circle such that the 16 dB of available gain is achieved. The transistor itself has an available gain $G_A = 18.2$ dB or an actual gain $G_P = 17.9$ dB at 435 MHz, but because we are concerned about stability, we will design for $G_A = 16$ dB. The question occurs, "If the transistor itself has close to the required gain and is stable, why bother with networks?" The answers are (a) in most situations we will want some selectivity in the amplifier, and (b) we may want to improve noise figure. We also have to consider that the source/load may not be the ideal 50 Ω at frequencies outside the 420 to 450 MHz band, and this might cause stability problems if there is no selectivity.

9. Now we want to pick a value of reflection coefficient on each of these circles that will (a) give us the gain that we want, (b) have no stability problems and (c) will result in an acceptable noise figure for the amplifier.

10. See Fig 4B. Type RHOL 16, then Execute. The number 16 specifies the available decibel gain that we want. Now place the cursor at the point indicated on the *solid* yellow (input) circle

and press the left mouse button. A marker appears on the input line where we placed the cursor and a marker also appears on the output (dashed) line. The CCW now shows two values, RHO S mag = 0.30, ang = -16.57° , and RHO L mag = 0.48, ang = 32.26° . The first is the reflection coefficient Γ_S (looking back from the transistor base toward the generator) needed for the input matching network (INNET) and the second is the reflection coefficient Γ_L (looking from the collector toward the load) for the output matching network (OUTNET). These networks will provide the overall gain of 16 dB at 435 MHz with very good stability. Several trials are always needed to get the best placement of these markers with respect to the stability circles. DEL the undesired trial points and re-enter the desired points, when you know where you want them to be.

11. We must also consider the noise figure. I referred to the Motorola data sheets for the MRF901 and saw that at 500 MHz, $V_{ce} = 10$, $I_c = 5$ mA and for the location of my marker on the input yellow circle, I could expect a noise figure of about 2 dB. This was close enough to 435 MHz that I was satisfied. My choice of S-parameters also was listed at these same values of V_{ce} and I_c . See the Noise Figure sidebar.

12. We must now obtain values for the input-matching network. See Fig 4C. Type G in the CCW, then Execute. Place the cursor at the marker on the solid yellow (input) circle and press

the left button. This creates a $G = 0.55$ circle. We are now ready to get the values. Type DX, then Execute. Place the cursor at A, the origin point, and press the left button. Move the cursor to B, the junction of $R = 1$, $G = 0.55$ and press the left button. The series inductance value in the CCW is 16.45 nH. Now use DB from the junction B down to C, the marker on the solid yellow circle, to get the shunt capacitance value, which is 4.41 pF. Enter these values in the INNET circuit block as shown in Fig 3. See Fig 4D. Create the $G = 0.38$ line at point C. Repeat the procedure DX from A to B and DB from B to C to get the values for OUTNET which are 23.4 nH and 1.68 pF. Enter these values in OUTNET as shown. Be sure to get the connections in the right order. The schematic and node numbers are shown in Fig 5A.

13. To see what has been accomplished, change the FREQ block to sweep from 420 to 450 MHz and view the plot of AMP. MS21 dB is nearly 16 (the exact G_A from Eq 9 is 16.02 dB). The output impedance is close to 50 Ω , as shown by MS22 dB. MS12 dB is about -24 , which is not too bad but only about 8 dB greater in magnitude, in the negative direction, than MS21. MS11 dB is very poor, only about -1.5 , meaning that the input to the amplifier is not well matched to the 50 Ω generator. This is how Circles has assured the amplifier stability and controlled G_A . Nevertheless, the design approach that we have used tells us that we have a noise figure of not more

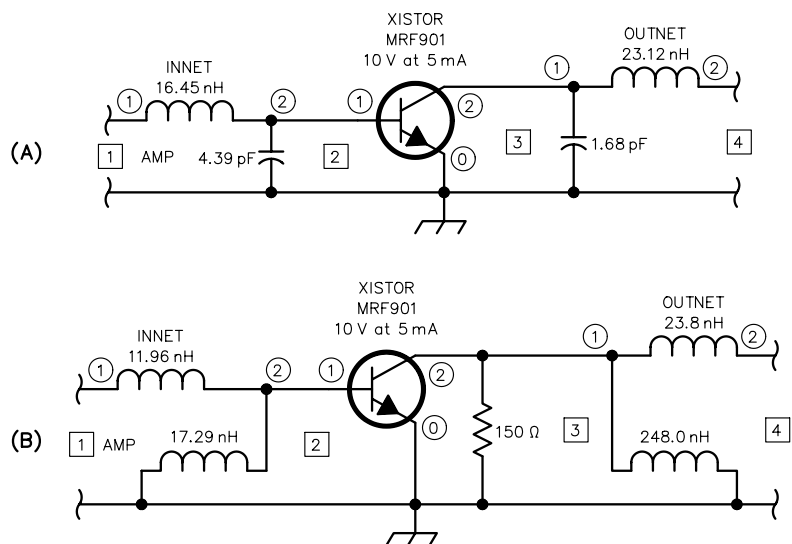


Fig 5—Amplifier diagrams. (A) no resistor loading, (B) 150 Ω resistor loading.

than 2 dB, according to the data sheet for the MRF901. Also, if we optimize *only* the components in the output network the output match to 50 Ω can be improved. In other words, the transistor output is conjugate matched to the load. At 420 MHz, G_A increases about 0.5 dB and at 450 MHz decreases about 0.5 dB.

14. To get some additional perspective, set up the Circles chart for CKT AMP, the complete amplifier, at $FREQ = 435$ MHz. Plot S1, S2 and G1B 16. S1 and S2 are well inside the chart. The two gain circles intersect at the origin, just as they should (the load and source are both 50 Ω) and the amplifier is quite stable, but only conditionally so ($K = 0.52$). If the amplifier is not terminated properly, if a combination of small parameter changes occur or if we try to increase gain, oscillation might occur.

15. The Sweep feature deserves attention. Page 12-9 of the *Designer* manual describes how S1, S2, G1[B], G2[B] or N circles of a circuit or subcircuit can be plotted at each of the frequencies defined in the FREQ block, using the SW FREQ command. Or, a plot of G1[B], G2[B] or N circles over some decibel range, minimum to maximum, at some specified FREQ can be plotted, using the SW DB command. When you're through with the Sweep mode, be sure to enter SW END.

The MRF901 transistor appears to

be a slightly difficult one to use in its native condition. An interesting thing can be observed about this example. If we use G2B 16 and RHOS 16, instead

of G1B 16 and RHOL 16 and redesign the networks, we will find that the input side is much better matched to the 50 Ω generator and the output side is

Gain Circles

There are some options regarding gain circles that should be appreciated for successful design. They are as follows:

1. The *maximum* available generator power is delivered to a Γ_{IN} that is the conjugate of its Γ_S (see Fig 2). The *maximum* device available power is delivered to a Γ_L that is the conjugate of its Γ_{OUT} . These two define the *maximum* available gain (GAMAX). On the other hand, available gain is the ratio of the power that is actually available from the output, when the input and output are mismatched, to the maximum power that is available from the generator. Available gain is very often less than maximum available gain. It is widely used in calculating the gain and noise figure of cascaded devices.
2. Power gain (aka *actual* gain or *operating* gain) is the ratio of the power that is delivered to the load to the power that is delivered to the input.
3. G1 refers to the input gain circle. For example, the G1 15 circle gives all values of the input Γ_S that can be used to get 15 dB of available gain (if we also use the right value of load Γ_L). It doesn't tell us what Γ_L is. RHO, at any point on the circle, applies only to the input network. G1B, on the other hand, does the same thing as G1 and also gives a circle for the output. The pair of points, one at input, one at output, defined by the RHOL 15 command, tell us what to do at both ends of the amplifier. With G1B, we select a point on the input circle. If the stage is almost unilateral, the output circle is very small and may even be difficult to see on the screen. I like G1B a lot better.
4. G2 15 is similar to G1 15, but it refers to *actual* or *operating* power gain and only to the output of the amplifier. G2B 15 gives us two circles, one at the output and one at the input. RHOS 15 gives a point on each circle that provide 15 dB of actual power gain. With G2B, we select a point on the output circle. G2B is also preferable, I believe. When selecting the point on the output circle, the resulting input point should occur at a location of good noise figure, if possible. It is for this reason that the G1B approach is usually preferred for noise matching.

Noise Figure

The following notes apply to noise-figure analysis using Circles:

The format for noise data in the DATA block is:

NOI RN (or) NC x (first line)

* FREQ FMIN_DB MGOPT GOPT RN (or) NC (comments)

420MHZ 1.83 0.320 35.7 0.25 (data)

435MHZ 1.88 0.758 65.2 0.36 (data)

etc

RN signifies that the RN noise data is the equivalent noise resistance, normalized with respect to x. If the data is given in ohms, it is divided by x to get the normalized RN. For example if $x = 50 \Omega$ and $RN = 60 \Omega$ then RN (normalized) = $60 \Omega / 50 \Omega = 1.2$. If the data is already normalized, omit x.

NC signifies that the noise data is noise figure in decibels (either RN or NC may be available from data sheets).

FMIN_DB is the minimum value of noise figure in decibels at FREQ. This is the value of noise figure when the matching networks are optimized for lowest noise figure. It is the best that can be achieved at FREQ.

MGOPT is the magnitude and GOPT is the angle of the input reflection coefficient $\Gamma_S = \Gamma_{OPT}$ that results in FMIN_DB. The noise situation is perhaps best appreciated by studying the equation:

$$F = F_{MIN} + \frac{4 \cdot RN \cdot |\Gamma_S - \Gamma_{OPT}|^2}{(1 - |\Gamma_S|^2) \cdot |1 + \Gamma_{OPT}|^2} ; \quad \Gamma_{OPT} = MGOPT \angle GOPT$$

where F and F_{MIN} are noise *factors* and Γ_S is the input reflection coefficient. If $\Gamma_S = \Gamma_{OPT}$ then $F = F_{MIN}$. If $\Gamma_S = 0$ (no matching network) the intrinsic value of F for XISTOR itself is observed. The values of Γ_S that define a particular value of F lie on a circle in the Smith Chart. Noise *figure* $NF(dB) = 10 \log(F)$.

Type N 3 in the CCW to get a 3 dB noise-figure circle, and so forth. Type N 0 to see the minimum noise figure.

poorly matched, as compared with the example that we worked out. The reader is encouraged to try this. However, there are better ways to improve the design and we will look at one example of how we might do this.

1. Remove the asterisk in front of the 150 Ω resistor from collector to ground.
2. Analyze the circuit and bring up Circles. Use CKT XISTOR, FREQ = 435 MHz.
3. Plot S1 and S2, then type K and GMAX. We see that K = 1.17, GMAX = 17.46 dB and S1 and S2 are well outside the Smith Chart, as shown in Fig 4E. We now have an unconditionally stable amplifier, which gives us more latitude in the design options. We have achieved this by modifying the XISTOR block with a resistor. The quantity $|S_{21} \cdot S_{12}|$ of XISTOR has been reduced from 0.46 to 0.26 (check it out).
4. Use bilateral G1B 17.4 and RHOL 17.4 to get the gain circles and the marker points. Note that the two circles are very small, which means that both ends are very close to a conjugate match and therefore maximum gain. Use DX and DB to get the input and output-network values, as shown in Fig 5B. Place these values in INNET and OUTNET.
5. Looking at the graph of AMP's S-parameters we see that MS11 and MS22 are better than -15 dB at the band edges and better than -25 dB at the center. We now have a very stable amplifier with 17.4 dB of gain that is "cascadable," meaning that it matches a 50 Ω source and load within a 2:1 SWR. The noise figure is about the same, or slightly better, according to the MRF901 datasheet. It may be possible to optimize both matching networks over the frequency band. We may also be able to adjust the input network to a value of better noise figure, with a small loss of gain and a small loss of input impedance match.

Here's another educational experiment to try:

1. In the DATA block, reduce the S12 magnitude data from 0.05 to 0.005 and from 0.07 to 0.007. This is a 20 dB reduction in internal feedback. Also comment out (asterisk) the 150 Ω resistor.
2. In Circles, GAMAX is 19.97 dB. Use G1B 19.5 and RHOL 19.5 (almost maximum gain). The stability circles are far removed from the chart.
3. Get the input and output networks. This is also a "cascadable" amplifier with a noise figure in the 2 dB range. This is what a unilateral transistor can accomplish if we can find one, or if

we can neutralize the MRF901. The cascode connection or the differential pair should be considered as good ways to reduce S12, although the noise figure may take a hit (two noise making devices instead of one).

Optimization is a very nice way to improve input and output matching, but it has to be used carefully. My experience has been that it works much better with unconditionally stable transistors. Otherwise, the amplifier can wind up in an oscillating region. This happened with the first example. The method shown in Fig 3 is simple and effective. Some experience with the optimizer will suggest strategies that help to avoid pitfalls. In addition, we should always confirm the results by checking the K value and the stability circles of AMP. The origin point of this chart (the generator and load impedances) should be far removed from the stability circles.

If you perform the various exercises that we have presented here and study the two sidebars, skill will develop as you get familiar with the various key-

board and graphical operations. It soon becomes a simple enough routine that "what if?" experiments can be explored, with confidence in the results. Data sheets contain the necessary information so that you can design and analyze scientifically.

Recommended Reading

- G. Gonzales, *Microwave Transistor Amplifiers*, 2nd ed, Crestone Technical Books, Noble Publishing Co, Tucker, GA; tel 770-908-2320; URL www.noblepub.com.
- Hewlett-Packard "S-Parameter Techniques" Application Note 95-1 available from <http://www.tmo.hp.com/tmo/Notes/interactive/an-95-1/index.html> as an 80-page .PDF document. A network design document is also available.
- M. Wilson, et al, Ed, *The ARRL UHF/Microwave Experimenter's Manual* (Newington: ARRL, 1990), ARRL Order No. 3126. ARRL publications are available from your local ARRL dealer or directly from the ARRL. Mail orders to Pub Sales Dept, ARRL, 225 Main St, Newington, CT 06111-1494. You can call us toll-free at tel 888-277-5289; fax your order to 860-594-0303; or send e-mail to pubsales@arrl.org. Check out the full ARRL publications line on the World Wide Web at <http://www.arrl.org/catalog>.