

# Matching RFIC Wireless Transmitters to Small Loop Antennas

By Larry Burgess  
Maxim Integrated Products

Here are design techniques for matching to the very low impedance of small loop antennas, including designs that include additional low pass filtering

Most short-range RF transmitters use a small loop antenna for radiation of the signal. This article provides a review of the impedance characteristics of these antennas

and describes methods of matching the output of the transmitter to the very low impedance of the loop antenna, with adequate rejection of the harmonics.

## Impedance of an Electrically Small Loop Antenna

The radiation resistance of a small printed circuit board loop with area  $A$  at a frequency whose wavelength is  $\lambda$  is given by

$$R_{rad} = 320\pi^4(A^2/\lambda^4) \quad (1)$$

The dissipative resistance in the loop, ignoring dielectric loss, is given in terms of the loop perimeter,  $P$ , the trace width,  $w$ , the magnetic permeability,  $\mu = 400\pi$  nH/meter, the conductivity,  $\sigma$  ( $5.8 \times 10^7$  ohms/meter for copper is typical), and the frequency,  $f$ .

$$R_{loss} = (P/2w)(\pi f \mu / \sigma)^{1/2} \quad (2)$$

The inductance of the loop is given in terms of the perimeter,  $P$ , the area,  $A$ , the trace width,  $w$ , and the magnetic permeability,  $\mu$ .

$$L = (\mu P/2\pi) \ln(8A/Pw) \quad (3)$$

These three quantities can be derived from expressions in antenna theory textbooks [1, 2]. Note that the radiation resistance increases

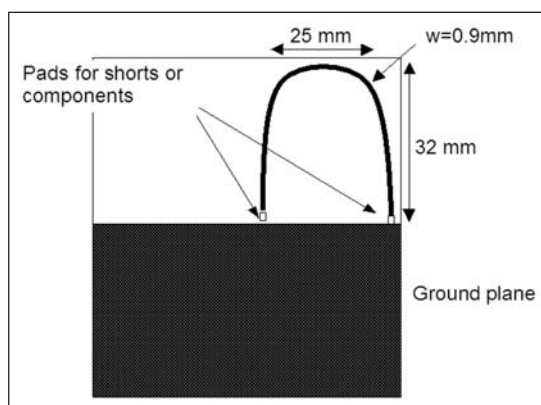


Figure 1 · Small loop antenna etched onto a printed circuit board.

as the fourth power of the ratio of the loop dimension (e.g., radius) to the wavelength, which is the same as the fourth power of the loop dimension times frequency. The loss resistance, arising from the skin effect, only increases as the square root of the frequency. This means that any increase in the size of the loop, or equivalently, any increase in frequency, will increase the radiation efficiency of an electrically small loop significantly.

A typical printed circuit board loop, whose dimensions are used to derive representative resistive and reactive values for a small loop, is shown in the drawing of Figure 1. It is roughly rectangular, 25 by 32 mm, with a trace width of 0.9 mm. These dimensions lead to the following values for the three quantities above at 315 MHz.

$$\begin{aligned} R_{rad} &= .025 \text{ ohm} \\ R_{loss} &= 0.3 \text{ ohm} \\ L &= 95 \text{ nH} \end{aligned}$$

For the other commonly used (outside the US) frequency of 433.92 MHz, the quantities are

$$R_{rad} = .093 \text{ ohm}$$

$$R_{loss} = 0.35 \text{ ohm}$$

$$L = 95 \text{ nH}$$

Note that the radiation resistance is extremely small. In addition, the resistance arising from the dissipative losses can be more than 10 times the radiation resistance. This means that the best radiation efficiency possible with this loop is about 8 percent at 315 MHz and 27 percent at 433.92 MHz. The matching network must minimize mismatch losses and additional dissipative losses from the matching components. Typically, a small loop will radiate only a few percent of the power from the transmitter.

**Basic Matching Network**

The simplest matching network is the “split capacitor,” described in a recent article [3]. Connecting it to the PA output with the bias inductor, as shown in Figure 2, makes it possible to adjust the value of  $C_2$  so that it resonates with the parallel combination of  $L_1$ , the PA-related capacitance, and the residual reactance from  $C_1$  and the loop antenna inductance. The equivalent series resistance (ESR) of the capacitor  $C_1$  is typically 0.138 ohms, so the total resistance of the small loop with the series capacitor is 0.46 ohms at 315 MHz.

The Appendix provided at the end of this article explains the transformations between series and parallel impedances and shows how to determine the values of  $C_1$  and  $C_2$ . At the resonant matching network frequency of 315 MHz, the tiny loop resistance is transformed by the series reactance of the loop and  $C_1$  to an equivalent parallel circuit with the optimum load resistance of 125 ohms (best load impedance for maximum MAX7044 efficiency). The parallel capacitance,  $C_2$ , and the parallel inductance of the bias inductor,

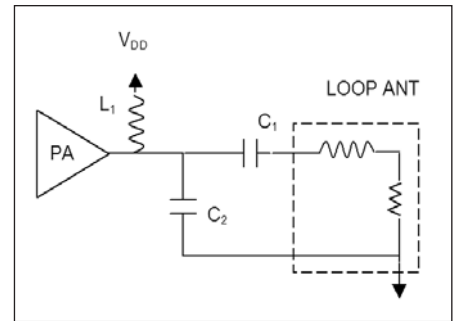
At 315 MHz	At 433.92 MHz
$C_1 = 2.82 \text{ pF}$	$C_1 = 1.47 \text{ pF}$
$C_2 = 63 \text{ pF}$	$C_2 = 43 \text{ pF}$
$L_1 = 36 \text{ nH}$	$L_1 = 27 \text{ nH}$

**Table 1 · Ideal component values for split capacitor match.**

$L_1$ , tune out the reactance of the equivalent parallel circuit.

The combination of  $C_1$  and the loop inductance form a positive reactance at the frequency of interest, so we can consider the two capacitors and loop inductance as an “L” matching network (shunt C, series L) that transforms the small loop resistance up to 125 ohms. Looking from left to right, it is a low pass high-to-low impedance matching network. The bias inductor  $L_1$  is not essential to matching, but it is needed as a DC path to provide operating current for the PA and can double as a tuning reactance for  $C_2$  and the small stray capacitance of the PA output.

The values that give a perfect match for the loop antenna described above are given in Table 1. The  $C_2$  capacitance in the table does not include approximately 2 pF that comes from the PA output and PC board stray capacitance. This 2 pF was added to the  $C_2$  values in all the matching calculations in this paper.



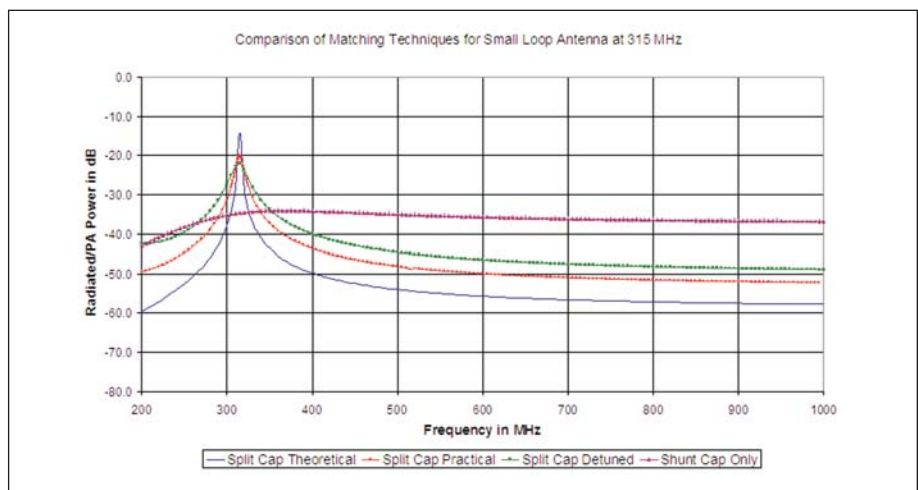
**Figure 2 · Split capacitor matching network with bias inductor.**

The frequency dependence of this match at 315 MHz is shown in the RF power transfer curves of Figure 3. These were computed by evaluating the expression for the power delivered from a source ( $R_S$ ) to a load impedance ( $R_L + X_L$ ), where the load impedance is the loop antenna impedance transformed by the matching network. This is the same quantity as the mismatch loss between a real source resistance and a complex load impedance.

$$P_{out}/P_{in} = 4R_S R_L / ((R_S + R_L)^2 + X_L^2) \quad (4)$$

This expression was multiplied by the antenna efficiency and the power loss arising from the matching components to get the total ratio of power radiated to power available.

All plots are shown at 315 MHz, and discussion of frequency depen-



**Figure 3 · Power transfer from RFIC transmitter to loop antenna.**

SMALL LOOP ANTENNAS

dence applies just to 315 MHz. The behavior at 433.92 MHz is similar but is not shown.

Assuming that the model for the loop antenna is correct and that exact values of the matching capacitors are realized, the mismatch loss is 0 dB, and the antenna loss is just the efficiency loss and dissipative loss added by the capacitors (radiation resistance divided by total resistance) of -14.1 dB. This match is a vast improvement over the 36.2 dB loss with no match at all (25 dB mismatch loss plus 11.2 dB efficiency loss) and the 34.7 dB loss from a single shunt capacitance that tunes out the antenna reactance (19 dB mismatch loss plus 15.7 dB efficiency and capacitor dissipative loss). The power transfer for the single shunt capacitor “match” is included in the plots as a reference.

In practice, the small loop antenna has a much lower *Q* than theory predicts. Calculations made from laboratory measurements of the printed circuit board loop of Figure 1 yield a total equivalent series resistance of 2.2 ohms at 315 MHz instead of the theoretical 0.46 ohms. The big increase in resistance (all of it loss resistance) can be attributed to lower *Q* capacitors than predicted, higher skin effect losses from solder as opposed to pure copper and proximity of the loop antenna to metal traces on the board and to the board material. Using this resistance, the standard capacitor and inductor values that match the loop are given in Table 2.

The power transfer for the practical loop antenna is also shown in Figure 3. Because the loss resistance of the practical loop is roughly four times higher than the theoretical loop, the best power transfer is about -20 dB instead of -14 dB. Although the power transfer curve is wider in frequency than for the theoretical loop, it is still narrow enough for component tolerances to move the peak to another frequency and reduce the power transfer at the intended frequency. If all three matching compo-

At 315 MHz	At 433.92 MHz
$C_1 = 3.0 \text{ pF}$	$C_1 = 1.5 \text{ pF}$
$C_2 = 33 \text{ pF}$	$C_2 = 27 \text{ pF}$
$L_1 = 27 \text{ nH}$	$L_1 = 20 \text{ nH}$

**Table 2 · Practical component values for the split capacitor matching network.**

nents are 5 percent high in value, for example, the power transfer drops to -26 dB.

The power transfer characteristic can be widened in frequency—and made less sensitive to component tolerances—by “detuning” the matching network. This can be done by the “brute force” method of simply adding resistance to the loop antenna, or by transforming the impedance to a value that is not perfectly matched to the transmitter. With either approach, the matching bandwidth is broadened at the expense of either higher power dissipation in the added resistor or higher mismatch loss in the detuned matching network. It may be better to take some power loss in return for having a predictable power transfer, because the penalty for drifting off frequency in a narrowband match is so high.

The broadening approach used here is to match the loop antenna to a higher impedance than the MAX7044 likes to see (500 to 1,000 ohms instead of 125 ohms) and accept the mismatch loss (along with the unavoidable dissipative losses). This approach has an added benefit of reducing the operating current.

Table 3 shows a set of L and C values that transform the loop impedance to about 500 ohms. They are rounded to the nearest standard component values.

This circuit decreases the power transfer at 315 MHz to -22 dB, but reduces the variation in loss with a 5 percent component tolerance to 3 dB.

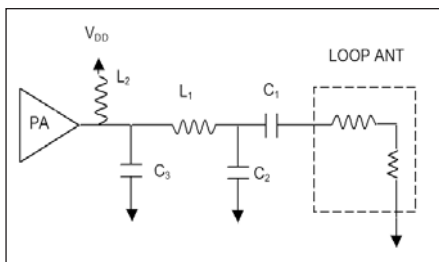
Figure 3 shows the losses in the tuning networks discussed above.

At 315 MHz	At 433.92 MHz
$C_1 = 3.3 \text{ pF}$	$C_1 = 1.65 \text{ pF}$
$C_2 = 22 \text{ pF}$	$C_2 = 15 \text{ pF}$
$L_1 = 27 \text{ nH}$	$L_2 = 20 \text{ nH}$

**Table 3 · Component values for a wider bandwidth split capacitor match.**

Notice how narrow the perfectly tuned networks are and how the “detuned” networks have more loss but are wider in bandwidth.

How well do these simple split capacitor networks reject harmonics? Figure 3, which extends to 1,000 MHz, shows that the theoretical match frequency response is down by 56 dB at the second harmonic and 58 dB at the third harmonic. Because it is down by 14 dB at the fundamental frequency, its second and third harmonic rejection is 42 dB and 44 dB, respectively. Practical and “detuned” matches are much more typical, so they are better indicators of actual harmonic rejection. The practical match is down 20 dB at the fundamental and 50 dB at the second harmonic, so the second harmonic rejection is 30 dB. The “detuned” match is down 22 dB at the fundamental and 46 dB at the second harmonic, so the second harmonic rejection is 24 dB. This is not enough rejection if the transmitter radiates the maximum average power allowable in the US at 315 MHz. The allowable radiated field strength of about 6,000  $\mu\text{V/m}$  corresponds to radiated power of -19.6 dBm. The second harmonic cannot exceed 200  $\mu\text{V/m}$  (-49 dBm), so almost 30 dB of harmonic rejection is needed at the maximum allowable average power level. Because FCC regulations for the 260-470 MHz unlicensed band permit low duty cycle peak power radiation at levels as much as 20 dB above the average power, there are situations where even more than 30 dB of second harmonic rejection is needed.



**Figure 4 · Split-C matching network combined with a low pass filter for improved harmonic rejection.**

### Matching Network With Greater Harmonic Rejection

One simple way to achieve better harmonic rejection is to add a low pass filter to the matching network. This can be done with a pi network inserted between the split capacitor matching network and the transmitter output. Because the pi network is also capable of transforming impedances, there are many possible combinations of impedance transformations. The examples here yield realistic values of L and C matching components. Figure 4 shows the network, in which one parallel capacitor in the low pass filter has been combined with the parallel capacitor in the split C matching network and the other parallel capacitor has been adjusted in value to tune out the bias inductor and the stray capacitances in the IC in addition to serving as part of the matching network.

The values that give a nearly perfect match for the loop antenna in Figure 4 are shown in Table 4.

In this configuration, the split C transforms the low loop resistance to roughly 150 ohms (very close to the 125 ohms for maximum efficiency of the PA), and the pi network is a low pass filter designed for 125 ohms input and output impedance. The mismatch loss is -0.1 dB, and the bandwidth of this match is once again narrow and highly sensitive to component tolerances. The match is still very narrow because an exact impedance match was attempted, although with more than one net-

At 315 MHz	At 433.92 MHz
$C_1 = 3.0 \text{ pF}$	$C_1 = 1.5 \text{ pF}$
$C_2 = 33 \text{ pF}$	$C_2 = 30 \text{ pF}$
$C_3 = 12 \text{ pF}$	$C_3 = 8.2 \text{ pF}$
$L_1 = 51 \text{ nH}$	$L_1 = 33 \text{ nH}$
$L_2 = 47 \text{ nH}$	$L_2 = 33 \text{ nH}$

**Table 4 · Component values for a split capacitor match with improved harmonic rejection.**

work. The result is still the same: A narrow bandwidth match that is sensitive to component tolerances.

The bandwidth of this matching network can be increased (and the sensitivity to component tolerances decreased) by de-tuning the split capacitor matching network but retaining the 125 ohm pi network low pass filter. The  $C_1$  and  $C_2$  values shown in the table below transform the parallel resistance of the loop antenna to about 500 ohms instead of the 150 ohms in the best match. The resulting mismatch between the antenna and the 125 ohm low pass filter increases the mismatch loss to 2 dB but broadens the matching bandwidth. The values for this match are given in Table 5.

This means that the output of the split capacitor match is intentionally mismatched to the pi section. Changing the split capacitor values to raise the transformed loop resistance higher than 500 ohms while maintaining the same pi matching network further widens the matching

At 315 MHz	At 433.92 MHz
$C_1 = 3.3 \text{ pF}$	$C_1 = 1.65 \text{ pF}$
$C_2 = 22 \text{ pF}$	$C_2 = 18 \text{ pF}$
$C_3 = 12 \text{ pF}$	$C_3 = 8.2 \text{ pF}$
$L_1 = 51 \text{ nH}$	$L_1 = 33 \text{ nH}$
$L_2 = 47 \text{ nH}$	$L_2 = 33 \text{ nH}$

**Table 5 · Component values for a wider bandwidth split capacitor match with improved harmonic rejection.**

bandwidth with the attendant increase in mismatch loss.

The behavior of the near-ideal matching network and the de-tuned network, along with a simple shunt capacitor for reference, is shown in Figure 5. While these are similar to the plots in Figure 3, the big difference is in the harmonic rejection. The near-ideal match now has 49 dB of second harmonic rejection, and the de-tuned match has 44 dB of second harmonic rejection.

### Summary and Conclusions

The important point to remember about matching a small loop antenna is that its equivalent series impedance is an inductance with a tiny series resistance, which consists predominantly of loss resistances and an even tinier radiation resistance. Its equivalent parallel impedance is an inductance with a huge parallel resistance (5k to 50k ohms). Either representation is difficult to match to a resistance of 100 to 300 ohms.

The combination of a small capacitor in series with the loop and a larger capacitor in parallel with the series capacitor and loop is a simple way to match the loop. An exact impedance match is very high Q (ratio of loop reactance to resistance), which means that any drift in component value, frequency, or operating environment will quickly degrade the match and raise the mismatch loss significantly. Choosing standard capacitor and inductor values that intentionally widen the bandwidth of the match will produce a match that is more tolerant to component variations and the environment. The price to pay for this wider bandwidth is additional mismatch loss, but this loss is more predictable. Examples have been given for 315 MHz and 433.92 MHz.

When harmonic rejection is important, it is better to use two more components in the matching network to form a low pass filter along with the matching network.

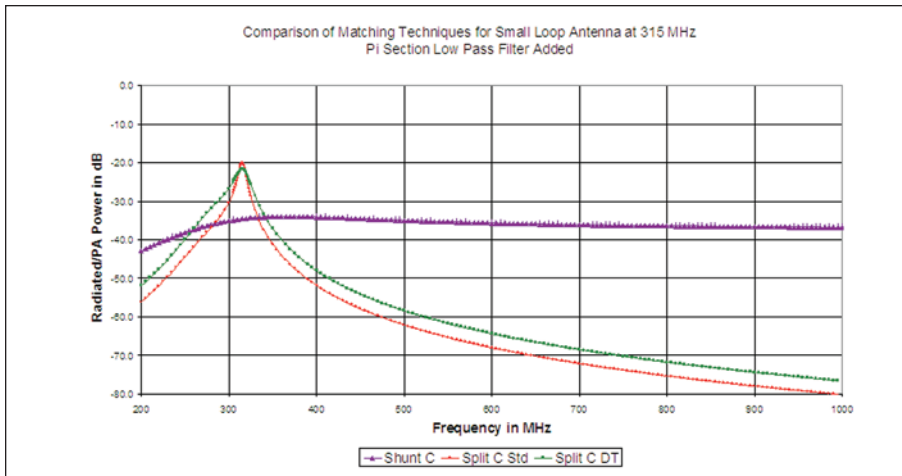


Figure 5 · Power transfer from RFIC transmitter to loop antenna with a low pass filter added to matching section.

The combination split-C and low pass filter networks chosen in this article can improve harmonic rejection by roughly 20 dB over that of the simple split-C matching network.

The values in the matching networks presented here may need to be adjusted by the user slightly to accommodate stray reactances and losses in the circuit board or the matching components themselves. Care should also be taken to be sure that all of the matching components are operating well below (two octaves preferred) their self-resonant frequencies (SRF).

What is more important than the specific value of each matching component is the basic structure of these matching networks. The purpose of the split-C section is to transform the loop resistance value to a more reasonable range. The purpose of the pi network low pass filter is to reject higher frequencies, perform additional matching if needed, and establish the bandwidth of the match. As long as the user approaches the network with this in mind, the right component values can be found.

## References

1. C. Balanis, *Antenna Theory, Analysis and Design*, Harper and Row, 1982.

2. W.A. Stutzman and G.A. Thiele, *Antenna Theory and Design*, Wiley, 1981.

3. F. Dacus, J. Van Niekerk and S. Bible, "Introducing Loop Antennas for Short-Range Radios," *Microwaves & RF*, July 2002, pp. 80-88.

## Acknowledgments

The author wishes to thank Messrs. Andy Zocher and Farron Dacus of Maxim Integrated Products for their suggestions and guidance while reviewing this article.

## Appendix

An Appendix, "Using Reactances to Create Resistive Impedance Matches," follows on page 28.

## Author Information

Larry Burgess works in Corporate Applications at Maxim Integrated Products, where he initiates the development of RF products in the 300 to 1000 MHz frequency range. He received a BSEE and MSEE from MIT and a PhD in EE from the University of Pennsylvania. Dr. Burgess has worked for over 30 years in applications for cellular, unlicensed wireless, military advanced communications, and radar. He can be reached by e-mail at: larry\_burgess@maximhq.com.

## Appendix: Using Reactances to Create Resistive Impedance Matches

### Parallel and Series Impedances

A resistance value can be increased or decreased over a limited frequency range by adding a series and parallel reactance to the resistor. The key to understanding the way this works is the equivalence between a series combination and parallel combination of a resistance,  $R$ , and a reactance,  $\pm jX$ . The series and parallel circuit representations are shown in Figure A1.

The parallel resistance and reactance as a function of the series resistance and reactance can be determined by equating the admittances of the two circuits.

$$\frac{1}{R_S + jX_S} = \frac{1}{R_P} + \frac{1}{jX_P} \quad (\text{A1})$$

Equating the real and imaginary parts leads to these expressions:

$$R_P = \frac{R_S^2 + X_S^2}{R_S} \quad X_P = \frac{R_S^2 + X_S^2}{X_S} \quad (\text{A2})$$

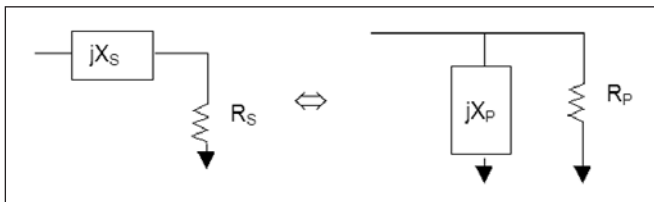
Which can be expressed in terms of  $Q_S = X_S/R_S$ , as

$$R_P = R_S (1 + Q_S^2) \quad X_P = X_S \left( \frac{1 + Q_S^2}{Q_S} \right) \quad (\text{A3})$$

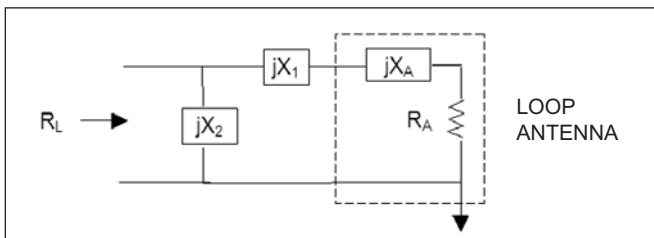
Likewise, the series resistance and reactance as a function of the parallel resistance and reactance can be determined by equating the impedances of the two circuits.

$$R_S + jX_S = \frac{1}{\frac{1}{R_P} + \frac{1}{jX_P}} \quad (\text{A4})$$

Equating the real and imaginary parts of Equation



**Figure A1** · Series and parallel representations of an impedance.



**Figure A2** · Simplified representation of the matching network.

A4 leads to the following expressions for  $R_S$  and  $X_S$ .

$$R_S = R_P \left( \frac{X_P^2}{R_P^2 + X_P^2} \right) \quad X_S = X_P \left( \frac{R_P^2}{R_P^2 + X_P^2} \right) \quad (\text{A5})$$

Which can be expressed in terms of the quantity  $Q_P = R_P/X_P$  as

$$R_S = R_P \left( \frac{1}{1 + Q_P^2} \right) \quad X_S = X_P \left( \frac{Q_P^2}{1 + Q_P^2} \right) \quad (\text{A6})$$

In each equation, the  $Q$  has the same meaning as the  $Q$  of a tuned circuit. It is a ratio of stored to dissipated energy (reactance to resistance in series, resistance to reactance in parallel). It is also a measure of how narrow the bandwidth is of an impedance transformation made with reactances. The higher the  $Q$ , the narrower the transformation bandwidth.

### Calculating Matching Network Reactances

The purpose of the split capacitor matching network shown in simplified form in Figure A2 is to transform the small series resistance of the loop antenna,  $R_A$ , to a larger resistance,  $R_L$ , that can be driven by the power amplifier. Because  $R_A$  and  $R_L$  are known, the desired  $Q_S$  can be found from Equation A3 by letting  $R_S = R_A$  and  $R_P = R_L$ .

$$Q_S = (R_L/R_A - 1)^{1/2} \quad (\text{A7})$$

The total series reactance needed to achieve  $Q_S$  is calculated by

$$X_A - X_I = Q_S R_A \quad (\text{A8})$$

From which  $X_I$  can be determined because  $X_A$  is known. Now the parallel impedance seen by the power amplifier has a real part equal to  $R_L$  and an imaginary part that can be calculated from equation A3.

$$X_P = (X_A - X_I) \left( \frac{1 + Q_S^2}{Q_S} \right) \quad (\text{A9})$$

The parallel reactance  $X_2$  needed to tune this out is equal and opposite to  $X_P$ . The split capacitor matching network is based on the assumption that the negative series reactance from the series capacitor is smaller in magnitude than the antenna reactance  $X_A$ . Therefore, the  $X_P$  reactance that results from adding a capacitor in series to the antenna is positive. This means that a negative reactance is needed to tune out  $X_P$ . Given that the power amplifier needs a parallel inductor for its bias, a parallel capacitance must be chosen that tunes out the positive reactance formed by the parallel combination of  $X_P$  and the bias inductor.