Introducing Loop Antennas For Short-Range Radios

Part 5 of this article series addresses the basics of loop-antenna design for short-range radios with integrated PLL Txs, comparing several matching methods for efficiency and performance.

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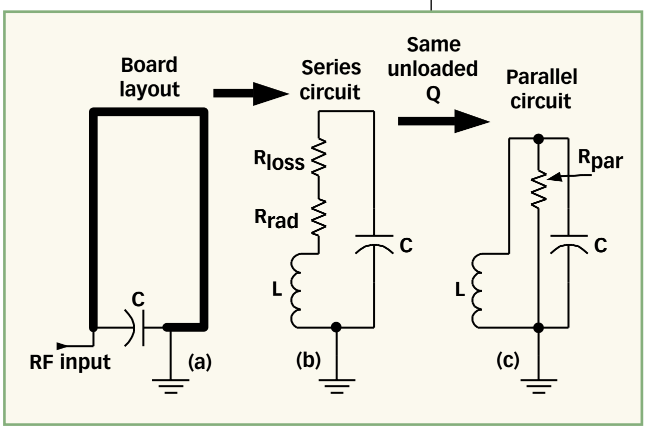
Short-range-radios are only as strong as their weakest component links, including the antenna. Previously, Parts 1, 2, 3, and 4 of this series (see *Microwaves & RF,* September and October 2001 and February and March 2002, respectively) covered one-way short-range system design, including link budgeting, regulatory issues, and some issues of silicon (Si) design at the transmit side. Part 5 of this article series now addresses the basics of loop-antenna design for short-range radios with integrated phase-locked-loop (PLL) transmitters (Txs), comparing the unmatched and the tapped-capacitor matching methods for efficiency and performance. Part 6 will cover the printed-transformer matched-loop antenna and understanding differential drive. The concluding Part 7 will cover physical-board design and practical regulatory compliance.

Printed-loop antennas are commonly used with unlicensed short-range radios, due to requirements for small size, ruggedness, and low cost. The frequency range is generally 285 to 470 MHz (see Part 2 of this series), where a full-sized quarter-wave whip antenna measures 6.28 to 10.4 in. (16.0 to 26.3 cm) in length. The large size usually eliminates consideration of whip antennas for these applications, resulting in the acceptance of printed-circuit-board (PCB) antennas as the most popular solution. PCB antennas generally exhibit only 1-to-20-percent radiation efficiency, but are small, easy to design (with the exception of significant errors in some published matching methods), insensitive to design errors since they



11. The physical implementation of a loop antenna (a) is shown next to its standard model (b) and an impedance transformed model (c) where all resistances are viewed in parallel.

Table 7: Calculated performance of the unmatched and matched 12 × 34-mm loop antenna at 434 MHz.

	Mismatch loss (dB)	Total efficiency	2nd harmonic rejection (dB)	3rd harmonic rejection (dB)
Unmatched	11.3	0.59 percent	22.1	23.6
Tapped capacitor	0	8 percent	50.6	52.0
Transformer	0	8 percent	41.5	36.0

usually must be tuned anyway, and provide a modest amount of harmonic suppression (improved by matching, as discussed later). However, since most of the available literature on loop-antenna matching is aimed at paging-receiver (Rx) applications, there is little information on the harmonic performance needed to meet regulatory requirements in transmit mode. This harmonic problem shall receive considerable attention in this and the next two articles. Also, loop antennas actually have enhanced efficiency when positioned near the human body. The low conductivity of the human body decreases a nearby electric field and increases a magnetic field (see p. 295 of ref. 8), leading to the general view that electrically small "magnetic" loop antennas are the most efficient for equipment worn on the human body, such as pagers, RF tags, and controllers. The magnetic-field intensification near (within one-quarter wavelength of) the human body is approximately 4.5 dB at 285 MHz, dropping to about 2.8 dB at 470 MHz and 0 dB at 900 MHz.

$$R_{rad} = 320\pi^4\left(\frac{A^2}{\lambda^4}\right) \quad (45)$$

$$R_{rad} = \left(3.84\times10^{-30}\right)\left(L_1 L_2\right)^2 f^2 \quad (46)$$

$$R_{loss} = \left[l\left(\pi f \mu_0\right)^{0.5}\right]/2w \quad (47)$$

$$R_{loss} = \left[\left(L_1 + L_2\right)/w\right]\left(2.61\times10^{-7}\right)\left(f\right)^{0.5} \quad (48)$$

$$\eta_r = \frac{R_{rad}}{R_{rad} + R_{lossL} + R_{lossC}} \quad (49)$$

In use, the loop inductance is usually considered to be a parallel resonance with a variable tuning capacitor so that the driver sees a large real load which must be matched for optimum power delivery. Other options to manual tuning include using resistors to modify the circuit quality factor (Q) to allow fixed capacitors and on-die automatic tuning. Unfortunately, the losses imposed by these methods are sometimes unacceptable. In particular, when a low-cost wideband Rx is used that prevents setting the intermediate-frequency (IF) bandwidth to match the spectral occupancy of the transmitted signal, then "averaging" as described in Part 2 is often used to maintain link quality. This averaging requires higher radiation efficiency and, thus, usually a well-matched and individually tuned high-Q loop antenna.

Figure 11 shows a standard loop-antenna model where series loss resistances are moved to provide a total parallel equivalent resistance that maintains the same Q in the loop. A matched single-ended driver would provide similar loading by driving into the nongrounded end of the capacitor, and Q will be cut approximately in half from the limit set by radiation and loss resistance. If the loop is directly driven by a lower-impedance power amplifier (PA) [unmatched], then Q will be lower still.

The radiation resistance of a loop, under the condition that it is electrically small (perimeter less than 0.3 λ), is provided as Eq. 45,[9]
where:

A = loop area (perimeter of inside edge of trace) in square meters and

λ = wavelength in meters.

For the frequencies and sizes normally used, this equation generally holds out to approximately the second to fourth harmonic and is adequate to use in predicting the lower-order harmonic performance where regulatory compliance is more commonly an issue. At higher frequencies where the antenna is not electrically small, the current in the antenna varies as a function of position, and must be taken account of as outlined in ref. 11 or through simulation. For a rectangular antenna with sides L_1 and L_2 fabricated on copper (Cu)-clad laminate, the given Cu conductivity of 5.8 $\times 10^7$, Eq. 45 becomes Eq. 46.

An expression for loss resistance derived from fundamental principles (skin-depth-based analysis), assuming that line width is much greater than line thickness, but thickness is also much greater than skin depth (true for practical boards), is Eq. 47,
where:

l = the total perimeter of the antenna in meters, measured at the center of the trace;

w = the width of the trace in meters;

σ = conductivity; and

μ = permeability.

For the common rectangular antenna case with Cu trace and with permeability of 1.256 $\times 10^{-6}$, Eq. 47 becomes Eq. 48.

is commonly provided as Eq. 49.

For a particular driving current to the loop, this expression follows immediately from power being i^2R. An alert reader may immediately wonder about driving current changing with variation in

resistance if a perfect match is provided by other circuitry. A

$$X_p = X_S\left(\frac{Q_S^2 + 1}{Q_S^2}\right) \quad (52)$$

$$L_p = L_S\left(\frac{Q_S^2 + 1}{Q_S^2}\right) \approx L_S(\textit{for}\text{ high Q}) \quad (53)$$

$$C_p = C_S\left(\frac{Q_S^2}{Q_S^2 + 1}\right) \approx C_S(\textit{for}\text{ high Q}) \quad (54)$$

$$R_p = R_S\left(1 + Q_S^2\right) = \frac{(\omega L_S)^2}{R_S} + R_S \approx \frac{(\omega L_S)^2}{R_S}(\textit{for}\text{ high Q}) \quad (55)$$

$$R_p = R_S\left(1 + Q_S^2\right) = \frac{1}{(\omega C_S)^2 R_S} + R_S \approx \frac{1}{(\omega C_S)^2 R_S}(\textit{for}\text{ high Q}) \quad (56)$$

$$L = \frac{\mu}{2\pi} l \ln\left(\frac{8A}{lw}\right) \quad (57)$$

$$R_{LStot} = R_{rad} + R_{lossL} + \left(\omega_h^2 LC\right)^2 R_{lossC} \quad (58)$$

simple analysis can show that if match is maintained, the same expression results if efficiency is defined as the radiated power divided by the total driving power. Though often neglected, losses associated with the resonating capacitor are usually significant and are counted in the denominator of Eq. 49 as another series resistance-loss term. Good COG capacitors will typically have series loss resistances of 0.1 to 0.2 Ω, variable capacitors series loss resistances from 0.1 to 0.5 Ω, and X7R and Z5U dielectrics series loss resistances of 0.5 and 1 Ω (visit the Murata website at www.murata.com for an excellent database of these losses over capacitor construction, value, and frequency). These capacitor losses can dramatically affect radiation efficiency and matching, and can have a moderate effect on harmonics.

It is often helpful in analysis to transform losses between series and parallel modes, which is valid around a narrow range of frequency. Using series losses as the base mode, we may define:

$$Q_S = \frac{X_S}{R_S} \quad (50)$$

from which analysis the following highly useful set of basic relations is Eqs. 51-56:

$$R_p = R_S\left(Q_S^2 + 1\right) \quad (51)$$

[SEE EQS. 52 TO 56 ABOVE]

Of course, to achieve a loop resonance, an expression is required for loop inductance. A remarkably simple formula for inductance of a polygon of general shape that is usually good to within 5 percent is provided by ref. 10 as Eq. 57,

where:

l = the perimeter as measured at the inside edge of the trace,

w = the width, and

A = the area.

Consider a numerical example for a loop that will later be matched using several other approaches. Assume operation at 434 MHz (a common European choice) with a rectangular antenna measuring 3.4 × 1.2 cm on the inside, with trace width of 2 mm, and with a capacitor with series loss at this frequency of 0.138 Ω. The loss resistance can be calculated as 0.250 Ω, the radiation resistance as 0.0227 Ω, the total series resistance as 0.286 Ω, and the resulting maximum efficiency as 7.95 percent. From Eq. 57, the inductance is 52.9 nH and the resonating capacitance is thus 2.54 pF. The unloaded Q is 505 and the equivalent parallel resistance is 72.9 kΩ.

Drivers on low-power Txs would normally have an output impedance of from 50 Ω to several thousand Ω, so a direct connection across this loop is obviously a bad mismatch that would not provide the maximum possible efficiency. The low impedance of the typical driver would also lower the Q drastically and reduce the harmonic rejection of the antenna. Despite these disadvantages, an unmatched loop is occasionally used in these applications, so the analysis is provided as follows. The total loss resistances of the loop antenna (where losses are modeled as a resistance is series with the inductor) are shown in Eq. 58,

where:

R_{rad} = radiation resistance,

R_{lossL} = ohmic loss resistance in the loop, and

R_{lossC} = capacitor series loss resistance.

The coefficient of R_{lossC} is 1 at the fundamental frequency, but greater than 1 at the harmonic frequencies. This coefficient results from moving the capacitor series loss, R_{cs}, to be in series with the inductor for modeling purposes. This sum may be represented in parallel form at the fundamental and harmonic frequencies by Eq. 55, yielding a quantity here known as R_{PtotH}, where H represents the harmonic number and is 1 for the fundamental frequency. Assuming the antenna still satisfies the constant spacial-current approximation for the first few harmonics, the radiation efficiency for the fundamental and first few harmonics can be written as:

$$\eta_H = \frac{R_{rad}}{R_{LStot}} \quad (59)$$

where it is understood that R_{rad} must be found from Eq. 45 at the appropriate harmonic H. Converting impedances to admittances, we may write a handy current-divider function expressing the fraction of the total current available from the driver at each harmonic frequency that flows through this parallel resistance at each harmonic frequency and is thus radiated. Defining G_{PtotH} as the total parallel admittance at each harmonic H (where H = 1 at the fundamental), this divider function is provided by Eq. 60.

In Eq. 60, the term Y_{driver} is used because at the harmonic frequencies the driver impedance and admittance are not normally purely real. At the fundamental frequency, the circuit is resonant and the imaginary component is zero, but at the harmonics it is dominated by the capacitance and most of the driver current available at the harmonics is shunted to ground and does not radiate. The ratio of each harmonic current to driver fundamental current is needed to determine the harmonic rejection. For approximation purposes, however, we may assume that the fundamental power is 10 dB over the first few harmonics (typical for a compressed Class A single-ended PA), but that the antenna is 5 dB more directional for the harmonic frequencies. The harmonic rejection (radiated harmonic to carrier power) in measured field strength for each harmonic H may, thus, be approximated to about ±5 dB accuracy as Eq. 61.

$$D_{IH} = Mag\left(\frac{G_{PtotH}}{G_{PtotH} + G_{driver} + j\left[\omega_h C - \frac{1}{\omega_h L}\right]}\right) \quad (60)$$

$$\frac{P_H}{P_1} \approx \frac{\eta_H}{\eta_1}\,\frac{0.316\left(D_{IH}\right)^2 R_{PtotH}}{\left(D_{I1}\right)^2 R_{Ptot1}} \quad (61)$$

$$MisMatchLoss = \frac{4\left(\frac{Z_{in}}{R_{driver}}\right)}{\left(\frac{Z_{in}}{R_{driver}}\right)^2 + \frac{2Z_{in}}{R_{driver}} + 1} \quad (62)$$

The mismatch of the directly driven loop (power applied at the loop capacitor) is large. In general, for a source with impedance R_{driver} driving a load of parallel impedance Z_{in}, the "mismatch loss" (which does not include efficiency losses) may be determined by Eq. 62.

Table 7 provides example performance numbers for the example loop antenna (the 1.2 × 3.4-cm loop operating at 434 MHz) when directly driven by a source of 1.4 kΩ impedance. The mismatch loss in this case is approximately 11 dB, the efficiency about 8 percent (resulting total efficiency of less than 1 percent), and the harmonic rejection is just over 20 dB. With this example, there is risk of failing harmonic regulatory requirements (see Part 2 of this article series) in addition to generally weak link performance, although the poor Q of the antenna may eliminate the need for tuning. This table also includes a line for the transformer-matched loop antenna to be presented in Part 6.

The large mismatch and relatively poor harmonic suppression of the unmatched loop antenna may be much improved by the tapped capacitor matching method (shown in single-ended form in **Fig. 12**). Here, the fundamental definition of matching is seen in elegant simplicity, where matching means viewing the total set of loss impedances in the loop antenna as a single input parallel impedance $R_{par} = Z_{in}$ that yields the same unloaded Q. When R_{par} matches the driver resistance, the maximum power-transfer theorem is satisfied and the loaded Q will be one-half of the unloaded Q. Intuitively, the tapping may be seen to yield a down-impedance transform through conservation of energy, with the voltage at the tap point lowered from the inductor voltage by the capacitive divider action, and thus requiring a lower impedance (if all loss resistance is modeled at that point to provide the same Q) at the tap point to dissipate the same power. Pursuing this analytically will yield the result shown in Eq. 63 for parallel R_{in} as a function of inductor parallel resistance R_p.

To develop design equations for the tapped-capacitor case and to understand its harmonic performance, the broadband admittance looking into the capacitor tap is written as Eq. 64, where:

R_s = the resistance in series with the inductor that models all losses.

It is desirable to solve this equation for C_1 and C_2 to force the desired R_{in} and resonant frequency. The reciprocal of the real part of Eq. 64 yields the input resistance at resonance and provides one equation. Setting the imaginary part equal to zero at the desired resonant frequency yields the other. The results are shown in Eqs. 65 and 66.

Equation 64 also provides the way to understand the harmonic performance of the tapped-capacitor loop antenna. For the large impedance transform from parallel resistance across the inductor to parallel R_{in} at the tap point C_2 will normally be much larger than C_1, and much larger than the C of the unmatched loop. Thus, C_2 dominates the input admittance at the tap and shunts most current to ground, greatly improving harmonic rejection. This

$$Z_{in} = \left(\frac{1}{1+\frac{C_2}{C_1}}\right)^2 R_p \qquad (63)\ (\textit{applies}\ \text{at resonance})$$

$$G_{in} = \frac{R_s}{R_s^2 + \left(\omega L - \frac{1}{\omega C_1}\right)^2} + j\left(\omega C_2 - \frac{\omega L - \frac{1}{\omega C_1}}{R_s^2 + \left[\omega L - \frac{1}{\omega C_1}\right]^2}\right) \qquad (64)$$

may be quantified in a manner similar to the unmatched loop, where the "current-divider function" for harmonic current that flows in the real part of the input admittance (where it must flow to be radiated) is shown in Eq. 67.

Despite this divider function, some current still flows in the real part of loop-radiation resistance that is transformed to the input and it is this current that radiates power. The radiated power at the fundamental (H = 1) and at each harmonic (where the loop is still "small") is illustrated in Eq. 68, where:

i_{rmsH} = the root-mean-square (RMS) current available from the source at harmonic frequency H*Fundamental, and is the fundamental current when H = 1 (where D_{IH} = 0.5 due to the match

$$C_1 = \frac{1}{\omega_0\left(\omega_0 L - \sqrt{Z_{in} R_s - R_s^2}\right)} \quad (65)$$

$$C_2 = \frac{L - \frac{1}{\omega_0^2 C_1}}{R_s^2 + \left(\omega_0 L - \frac{1}{\omega_0 C_1}\right)^2} \quad (66)$$

$$D_{IH} = Mag\left(\frac{\mathrm{Re}\left[G_{inH}\right]}{G_{inH} + G_{driver}}\right) \quad (67)$$

$$P_{radH} = \frac{\eta_H \left(i_{rmsH} D_{IH}\right)^2}{\mathrm{Re}\left(G_{inH}\right)} \quad (68)$$

condition).

The harmonic rejection relative to the carrier is provided by the ratio of harmonic power in Eq. 68 to the radiated carrier power (also from Eq. 68 with H=1) degraded by the extra directivity of the antenna at the harmonic frequency. Assuming the applied harmonics are 10 dB down from the carrier and that the antenna is no more than 5 dB more directive for the harmonics yields the approximation:

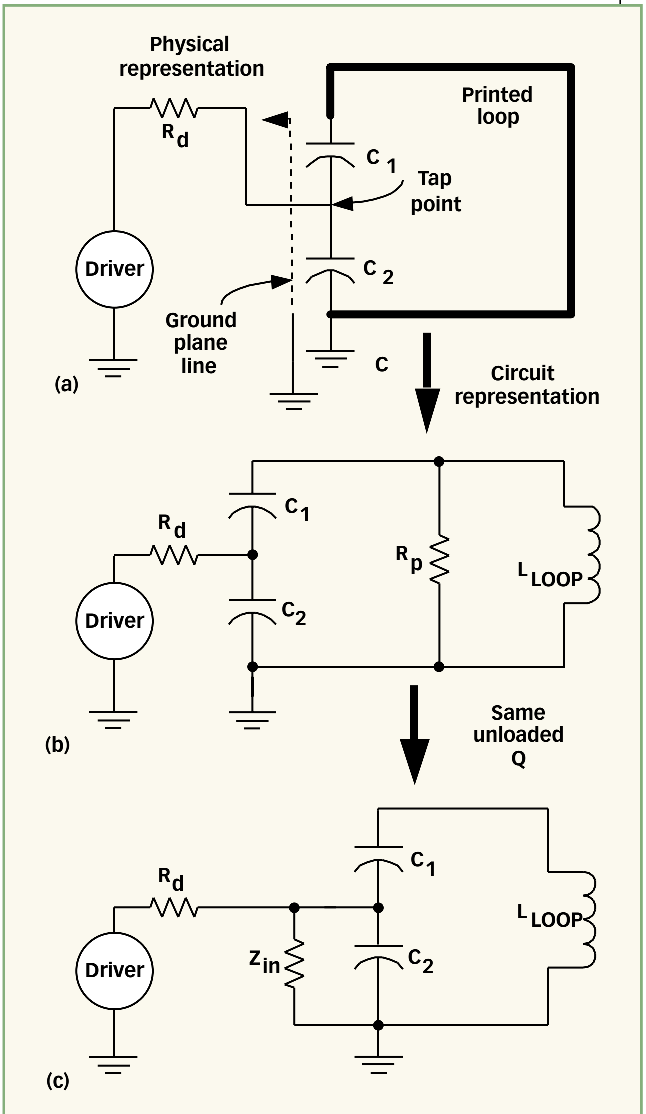
$$\frac{P_H}{P_1} \approx \frac{\eta_H}{\eta_1} \frac{1.26\left(D_{IH}\right)^2}{\mathrm{Re}\left(G_{inH}\right) Z_{in1}} \quad (69)$$

At harmonic frequencies, Eq. 64 can be simplified to:

$$G_{inH} \approx \frac{R_{sH}}{\left(\omega_H L\right)^2} + j\omega_H C_2 \quad (70)$$

Technically, C_1 and C_2 must be well-controlled to meet desired resonance and input impedance conditions. In practice, with a variable capacitor for C_1 or C_2, the tapped-capacitor method can yield a good match, near perfect resonance, and more than 40-dB harmonic rejection. For the example loop antenna at 434 MHz, with a driver impedance of 1.4 kΩ, L = 52.9 nH, C_1 = 2.95 pF, and C_2 = 18.3 pF, the predicted second harmonic levels are –50.6 dBc, and the predicted third harmonic levels are –52 dBc. There is little mismatch loss, so the total efficiency is the loop and capacitor efficiency of approximately 8 percent. These harmonics will normally pass all regulatory requirements, but to achieve such low loop harmonic levels, a PCB designer must beware of parasitic radiation from traces and bond wires that may actually dominate measured performance.

The material presented in this article should allow first-order understanding and design of the unmatched and tapped-capacitor loop antennas. The relations shown allow approximate prediction of radiated power and harmonics, at least over the first few harmonics where the loop is still electrically small. For higher harmonics where the loop is not electrically small, use of an electromagnetic (EM) simulator is recommended. The tapped-capacitor antenna has been found to be capable of excellent harmonic suppression, so together with its higher efficiency due to good matching, it is an excellent choice. However, the transformer-loop antenna to be presented next month can provide equal efficiency and acceptable harmonic suppression with lower parts count. There has been some incorrect information published on the operation of the transformer loop antenna, so methods based on the underlying EMs that are fundamentally sound will be shown. Next month's Part 6 will also cover understanding the use of differential drive on all these antenna types. The concluding Part 7 will deal with practical issues such as basic regulatory measurements, nonideal harmonic radiation from integrated-circuit (IC) pins and supply lines, and cost trade-offs in controlling these effects. MRF



12. These three circuit representations illustrate single-ended tapped-capacitor-antenna matching.

REFERENCES

8. K. Fujimoto and J.R. James, *Mobile Antenna Systems Handbook,* 2nd ed., Artech House, Norwood, MA, 2001.

9. K. Fujimoto, A. Henderson, K. Hirasawa, and J.R. James, *Small Antennas,* Wiley, New York, 1987.

10. Frederick Grover, *Inductance Calculations Working Formulas and Tables,* Dover Publications, 1946.

11. Constantine Balanis, *Antenna Theory, Analysis and Design,* 2nd ed., Wiley, New York, 1997.

FOR FURTHER READING

Krauss, Bostian, and Raab, *Solid State Radio Engineering,* Wiley, New York, 1980.

Thomas Lee, *The Design of CMOS Radio Frequency Integrated Circuits,* Cambridge University Press, Cambridge, England, 1998.