Microwave and RF Engineering

Volume 1
An Electronic Design Automation Approach

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Microwave and RF Engineering


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Foreword

Unlike many traditional books on RF and microwave engineering written mainly for the classroom, this book adopts a practical, hands-on approach to quickly introduce and familiarize engineers and students new to this subject. The authors extensively include the use of electronic design automation (EDA) tools to illustrate the foundation principles of RF and microwave engineering. The use of EDA methodology in the book closely parallels the latest tools and techniques used in the industry to accelerate the design of RF and microwave systems and components to meet demanding specifications and high yields.

This book introduces not only a solid understanding of RF and Microwave concepts such as the Smith chart, S-parameters, transmission lines, impedance matching, filters and amplifiers, but also more importantly how to use EDA tools to synthesize, simulate, tune and optimize these essential components in a design flow as practiced in the industry. The authors made the judicious choice of an easy-to-use and full featured EDA tool that is also very affordable so that the skills learnt from the book can be put into practice immediately without the barriers of acquiring costly and complex EDA tools.

Genesys from Agilent Technologies was chosen for its low cost and ideal combination of capabilities in circuit synthesis, simulation and optimization; Matlab equation handling; plus RF system, electromagnetic and statistical analysis. It is proven by Agilent Technologies in the design of state-of-the-art RF and microwave test instrumentation and time-tested by a large following of users worldwide for over 20 years.

The investment in learning the RF and microwave foundation skills with EDA techniques taught in this book results in knowledge that remains relevant and sought-after for a long time to come.
I wish such a book was available when I started my career as a microwave component designer. It would have made gaining RF and microwave insights much quicker than the countless hours of cut-and-try on the bench.

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Preface

Microwave Engineering can be a fascinating and fulfilling career path. It is also an extremely vast subject with topics ranging from semiconductor physics to electromagnetic theory. Unlike many texts on the subject this book does not attempt to cover every aspect of Microwave Engineering in a single volume. This text book is the first volume of a two-part series that covers the subject from a computer aided design standpoint. The first volume covers introductory topics which are appropriate to be addressed by linear simulation methods. This includes topics such as lumped element components, transmission lines, impedance matching, and basic linear amplifier design. The second volume focuses on subject matter that is better learned through non-linear computer simulation. This includes topics such as oscillators, mixers, and power amplifier design.

Almost all subject matter covered in the text is accompanied by examples that are solved using the Genesys linear simulation software by Agilent. University students will find this a potent learning tool. Practicing engineers will find the book very useful as a reference guide to quickly setup designs using the Genesys software. The authors thoroughly cover the basics as well as introducing CAD techniques that may not be familiar to some engineers. This includes subjects such as the frequent use of the Genesys equation editor and Visual Basic scripting capability. There are also topics that are not usually covered such as techniques to evaluate the Q factor of one port resonators and yield analysis of microwave circuits.

The organization of the book is as follows: Chapter 1 presents a general explanation of RF and microwave concepts and components. Engineering students will be surprised to find out that resistors, inductors, and capacitors at high frequencies are no longer ideal elements but rather a network of circuit elements. For example, a capacitor at one frequency may in fact behave as an inductor at another frequency. In chapter 2 the transmission line theory is developed and several important parameters are defined. It is shown how to simulate and measure these parameters using Genesys software. Popular types of transmission lines are introduced and their parameters are examined. In Chapter 3 network parameters and the
application of Smith chart as a graphical tool in dealing with impedance behavior and reflection coefficient are discussed. Description of RF and microwave networks in terms of their scattering parameters, known as S-Parameters, is introduced. The subject of lumped and distributed resonant circuits and filters are discussed in Chapter 4. Using the Genesys software a robust technique is developed for the evaluation of Q factor form the S-Parameters of a resonant circuit. An introduction to the vast subject of filter synthesis and the electromagnetic simulation of distributed filters are also treated in this chapter. In Chapter 5 the condition for maximum power transfer and the lumped element impedance matching are considered. The analytical equations for matching two complex impedances with lossless two-element networks are derived. Both analytical and graphical techniques are used to design narrowband and broadband matching networks. The Genesys impedance matching synthesis program is used to solve impedance matching problems. The VBScript programming techniques developed in this chapter can be used by students to generate their own synthesis applications within the Genesys software. In Chapter 6 both narrowband and broadband distributed matching networks are analytically and graphically analyzed. In Chapter 7 single-stage amplifiers are designed by utilizing four different impedance matching objectives. The first amplifier is designed for maxim gain where the input and the output are conjugately matched to the source and load impedance; the second amplifier is designed for specific gain where the input or the output is mismatched to achieve a specific gain less than its maximum; the third amplifier is a low noise amplifier where the transistor is selectively mismatched to achieve a specific Noise Figure; and the fourth amplifier is a power amplifier where the transistor is selectively mismatched to achieve a specific amount of output power. In Chapter 8 a two-stage amplifier is designed by utilizing a direct interstage matching network. Monte Carlo and Yield analysis techniques are also introduced in this chapter. Finally a brief introduction to cascade analysis is presented.

Ali A. Behagi
Stephen D. Turner
August 2011
RF and Microwave Concepts and Components

1.1 Introduction

An electromagnetic wave is a propagating wave that consists of electric and magnetic fields. The electric field is produced by stationary electric charges while the magnetic field is produced by moving electric charges. A time-varying magnetic field produces an electric field and a time-varying electric field produces a magnetic field. The characteristics of electromagnetic waves are frequency, wavelength, phase, impedance, and power density. In free space, the relationship between the wavelength and frequency is given by Equation (1-1).

\[ \lambda = \frac{c}{f} \]

In the MKS system, \( \lambda \) is the wavelength of the signal in meters, \( c \) is the velocity of light approximately equal to 300,000 kilometers per second, and \( f \) is the frequency in cycles per second, or Hz.

![Figure 1-1 A time varying voltage waveform](image)

The electromagnetic spectrum is the range of all possible frequencies of electromagnetic radiation. They include radio waves, microwaves, infrared radiation, visible light, ultraviolet radiation, X-rays and gamma rays. In the field of RF and microwave engineering the term RF generally refers to
Radio Frequency signals with frequencies in the 3 KHz to 300 MHz range. The term Microwave refers to signals with frequencies from 300 MHz to 300 GHz having wavelengths from 1 meter to 1 millimeter. The RF and microwave frequencies form the spectrum of all radio, television, data, and satellite communications. Figure 1-2 shows a spectrum chart highlighting the RF and microwave frequencies up through the extremely high frequency, EHF, range or 300 GHz. This text will focus on the RF and microwave frequencies as the foundation for component design techniques. The application of Agilent’s Genesys software will enhance the student’s understanding of the underlying principles presented throughout the text. The practicing engineer will find the text an invaluable reference to the RF and microwave theory and techniques by using Genesys software. The numerous Genesys examples enable the setup and design of many RF and microwave circuit design problems.

![Figure 1-2 Electromagnetic spectrums from VLF to EHF](image)

The spectrum chart of Figure 1-2 is intended as a general guideline to the commercial nomenclature for various sub bands. There is typically overlap across each of the boundaries as there is no strict dividing line between the categories. The RF frequencies typically begin in the very low frequency, VLF, range through the very high frequency, VHF, range. Microwaves are typically the ultra high frequency, UHF, super high frequency, SHF and extremely high frequency, EHF, frequency ranges. During World War II microwave engineers developed a further detailed classification of the microwave frequencies into a band-letter designation. In 1984 the Institute of Electrical and Electronics Engineers, IEEE, agreed to standardize the letter designation of the microwave frequencies. These designators and their frequency ranges are shown in Table 1-1.
<table>
<thead>
<tr>
<th>Band Designator</th>
<th>L Band</th>
<th>S Band</th>
<th>C Band</th>
<th>X Band</th>
<th>Ku Band</th>
<th>K Band</th>
<th>Ka Band</th>
</tr>
</thead>
<tbody>
<tr>
<td>Frequency Range GHz</td>
<td>1 to 2</td>
<td>2 to 4</td>
<td>4 to 8</td>
<td>8 to 12</td>
<td>12 to 18</td>
<td>18 to 27</td>
<td>27 to 40</td>
</tr>
</tbody>
</table>

Table 1-1 Microwave band letter designators

Engineering students spend much of their formal education learning the basics of inductors, capacitors, and resistors. Many are surprised to find that as we enter the high frequency, HF, part of the electromagnetic spectrum these components are no longer a singular (ideal) element but rather a network of circuit elements. Components at RF and microwave frequencies become a network of resistors, capacitors, and inductors. This leads to the complication that the component’s characteristics become quite frequency dependent. For example, we will see in this chapter that a capacitor at one frequency may in fact be an inductor at another frequency.

1.2 Straight Wire, Flat Ribbon, and Skin Effects

In this section we will begin with a basic examination of the straight wire inductance and move into more complete characterization of inductors. Similarly we will look at resistor and capacitor design and their implementation at RF and microwave frequencies. Discrete resistors, capacitors, and inductors are often referred to as lumped elements. RF and microwave engineers use the terminology to differentiate these elements from those designed in distributed printed circuit traces. Distributed component design is introduced in Chapter 2.

1.2.1 Straight Wire Inductance

A conducting wire carrying an AC current produces a changing magnetic field around the wire. According to Faraday’s law the changing magnetic field induces a voltage in the wire that opposes any change in the current flow. This opposition to change is called self inductance. At high frequencies even a short piece of straight wire possesses frequency dependent resistance and inductance behaving as a circuit element.
Example 1.2-1: Calculate the inductance of a three inch length of AWG #28 copper wire in free space.

Solution: The straight wire inductance can be calculated from the empirical Equation (1-2) \[^{[1]}\].

\[
L = K \ell \left( \ln \frac{4 \ell}{D} - 0.75 \right) \text{nH}
\]  
(1-2)

where:

- \(\ell\) = Length of the wire
- \(D\) = Diameter of the wire (from Appendix A).
- \(K = 2\) for dimensions in cm and \(K = 5.08\) for dimensions in inches

Using Appendix A the diameter of the AWG#28 wire is found to be 0.0126 inches. Solving Equation (1-2) the inductance is calculated.

\[
L = 5.08 \left( \ln \frac{4 (3)}{0.0126} - 0.75 \right) = 93.1 \text{nH}
\]

It is interesting to examine the reactance of the wire. We know that the reactance is a function of the frequency and is related to the inductance by the following equation.

\[
X_L = 2\pi f L \quad \Omega
\]  
(1-3)

where: \(f\) is the frequency in Hz and \(L\) is the inductance in Henries.

Calculating the reactance at 60Hz, 1MHz, and 1GHz we can see how the reactive component of the wire increases dramatically with frequency. At 60Hz the reactance is well below 1\(\Omega\) while at microwave frequencies the reactance increases to several hundred ohms.

- 60 Hz: \(X_L = 2\pi (60) (93.1 \cdot 10^{-9}) = 35 \mu\Omega\)
- 1 MHz: \(X_L = 2\pi (10^6) (93.1 \cdot 10^{-9}) = 0.58 \Omega\)
- 1 GHz: \(X_L = 2\pi (10^9) (93.1 \cdot 10^{-9}) = 585 \Omega\)
1.2.2 Simulating the Straight Wire Inductor in Genesys

To analyze the Example 1.2-1 in Genesys, create a schematic in Genesys and add the straight wire model from the Parts Library. Attach a standard input and output port as shown in Figure 1-3.

![Figure 1-3 Genesys schematic of the straight wire](image)

In Genesys, the input port represents a signal source in series with a source impedance and the output port represents the load impedance connected to ground. Set the wire diameter to 12.6 mils and the length to 3 inches. Accept the value of 1 for Rho. Rho is not the actual resistivity of the wire but rather the resistivity of the wire relative to copper. Because we are modeling a copper wire the value should be set to one. It is a common practice in most commercial microwave software programs to specify resistivity in relative terms, compared to copper. Table 1-2 provides a reference of common materials used in RF and microwave engineering in terms of their actual resistivity and the relative resistivity to copper.

<table>
<thead>
<tr>
<th>Material</th>
<th>Resistivity Relative to Copper</th>
<th>Actual Resistivity (\Omega)-meters</th>
<th>Actual Resistivity (\Omega)-inches</th>
</tr>
</thead>
<tbody>
<tr>
<td>Copper, annealed</td>
<td>1.00</td>
<td>1.68 (\times) 10^{-8}</td>
<td>6.61 (\times) 10^{-7}</td>
</tr>
<tr>
<td>Silver</td>
<td>0.95</td>
<td>1.59 (\times) 10^{-8}</td>
<td>6.26 (\times) 10^{-7}</td>
</tr>
<tr>
<td>Gold</td>
<td>1.42</td>
<td>2.35 (\times) 10^{-8}</td>
<td>9.25 (\times) 10^{-7}</td>
</tr>
<tr>
<td>Aluminum</td>
<td>1.64</td>
<td>2.65 (\times) 10^{-8}</td>
<td>1.04 (\times) 10^{-6}</td>
</tr>
<tr>
<td>Tungsten</td>
<td>3.25</td>
<td>5.60 (\times) 10^{-8}</td>
<td>2.20 (\times) 10^{-6}</td>
</tr>
<tr>
<td>Zinc</td>
<td>3.40</td>
<td>5.90 (\times) 10^{-8}</td>
<td>2.32 (\times) 10^{-6}</td>
</tr>
<tr>
<td>Nickel</td>
<td>5.05</td>
<td>6.84 (\times) 10^{-8}</td>
<td>2.69 (\times) 10^{-6}</td>
</tr>
<tr>
<td>Iron</td>
<td>5.45</td>
<td>1.00 (\times) 10^{-7}</td>
<td>3.94 (\times) 10^{-6}</td>
</tr>
<tr>
<td>Platinum</td>
<td>6.16</td>
<td>1.06 (\times) 10^{-7}</td>
<td>4.17 (\times) 10^{-6}</td>
</tr>
<tr>
<td>Tin</td>
<td>52.8</td>
<td>1.09 (\times) 10^{-7}</td>
<td>4.29 (\times) 10^{-6}</td>
</tr>
<tr>
<td>Nichrome</td>
<td>65.5</td>
<td>1.10 (\times) 10^{-6}</td>
<td>4.33 (\times) 10^{-5}</td>
</tr>
<tr>
<td>Carbon</td>
<td>2083.3</td>
<td>3.50 (\times) 10^{-3}</td>
<td>1.38 (\times) 10^{-3}</td>
</tr>
</tbody>
</table>

Table 1-2 Resistivity of common materials relative to copper
Create a Linear analysis to analyze the circuit’s impedance versus frequency. The Linear Analysis Properties window is shown in Figure 1-4.

![Linear analysis properties for simulation of wire impedance](image)

**Figure 1-4** Linear analysis properties for simulation of wire impedance

Create a list of frequencies under the *Type of Sweep* setting. Enter the frequencies that were used in the reactance calculation of the previous section: 60 Hz, 1 MHz, 1 GHz. When an analysis is run in Genesys the results are written to a Dataset. The results of a Dataset may then be sent to a graph or tabular output for visualization. In this example an Equation Editor is used to post process the solutions in the Dataset. Create an Equation Editor, as shown in Figure 1-5, for the calculation of the wire’s reactance and inductance. Add the Equation Editor to the design as follows.

![Equation editor display for the calculation of the Inductance](image)

**Figure 1-5** Equation editor display for the calculation of the Inductance

More complex workspaces may contain multiple Datasets. It is a good practice to specify which Dataset is used to collect data for post processing. This is accomplished with the `{using (“Linear1_Data”)} statement of line 1 in the Equation Editor. Genesys has a built-in function ZIN1 to calculate the
impedance of the circuit at each analysis frequency. We know that the reactance is the imaginary part of the impedance. Line 2 of the Equation Editor defines the reactance as the imaginary part of the impedance, $Z_{IN1}$, in the Linear1 Dataset. Line 3 calculates the inductance of the wire from the reactance using Equation (1-3). The frequency, $F$, is the independent variable created by the Linear Analysis. Note the use of the dot (.) notation in the Equation Editor. There is a dot after Linear1_Data and reactance of lines 2 and 3. The dot means that these variables are not singular quantities but are arrays of values. There is a calculated array value for each independent variable, $F$. The Equation Editor is an extremely powerful feature of the Genesys software and is used frequently throughout this text. It is an interactive mathematical processor similar to MATLAB by MathWorks. There are two different syntaxes that may be used to define equations and perform post processing operations using the Equation Editor: the Engineering language and the Mathematics language. The Engineering language is a simple structured format as shown in Figure 1-5. The Mathematics language is compatible with the m-file syntax that is used in MATLAB. It is very convenient for students and engineers that are proficient in MATLAB. Both types will be used throughout this book to demonstrate the use of both languages. The tab in the upper left corner of Figure 1-5 displays the language syntax used in that particular Equation Editor. Send the simulated value of reactance and the calculated value of inductance to a tabular output. Figure 1-6 shows the tabular output and the table properties. On the Table Properties tab, select the Equations as the data source and then select both of the variables, reactance and inductance.

<table>
<thead>
<tr>
<th>$F$ (MHz)</th>
<th>Reactance</th>
<th>Inductance</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>800e-6</td>
<td>92.89e-9</td>
</tr>
<tr>
<td>2</td>
<td>1</td>
<td>92.19e-9</td>
</tr>
<tr>
<td>3</td>
<td>1000</td>
<td>65.42e-9</td>
</tr>
</tbody>
</table>

**Figure 1-6** Output Table Properties window
Compare the results of the simulation of Figure 1-6 with the calculated values of section 1.2.1. We can see that at 60 Hz and 1 MHz the reactance and resulting inductance are very close to the calculated values. At 1 GHz however, the values begin to diverge. The simulated reactance is 24 Ω less than the calculated reactance. Equation (1-2) was useful for calculating the basic inductance at low frequency but as the frequency enters the microwave region, the value begins to change. This is normal and due to the skin effect of the conductor. The skin effect is a property of conductors where, as the frequency increases, the current density concentrates on the outer surface of the conductor.

### 1.2.3 Skin Effect in Conductors

At RF and microwave frequencies, due to the larger inductive reactance caused by the increase in flux linkage toward the center of the conductor, the current in the conductor is forced to flow near the conductor surface. As a result the amplitude of the current density decays exponentially with the depth of penetration from the surface. Figure 1-7 shows the cross section of a cylindrical wire with the current density area shaded.

![Cross Section of Current Flow](image)

**Figure 1-7** Cross section of current flow in the conductor showing effect of skin depth

At low frequencies the entire cross sectional area is carrying the current. As the frequency increases to the RF and microwave region, the current flows much closer to the outside of the conductor. At the higher end of microwave frequency range, the current is essentially carried near the surface with
almost no current at the central region of the conductor. The skin depth, \( \delta \), is the distance from the surface where the charge carrier density falls to 37% of its value at the surface. Therefore 63% of the RF current flows within the skin depth region. The skin depth is a function of the frequency and the properties of the conductor as defined by Equation (1-4). As the cross sectional area of the conductor effectively decreases the resistance of the conductor will increase [7].

\[
\delta = \sqrt{\frac{\rho}{\mu \pi f}}
\]  

(1-4)

where:

\( \delta = \text{skin depth} \)

\( \rho = \text{resistivity of the conductor} \)

\( f = \text{frequency} \)

\( \mu = \text{permeability of the conductor} \)

Use caution when solving Equation (1-4) to keep the units of \( \rho \) and \( \mu \) consistent. Table 1-2 contains values of resistivity in units of \( \Omega \)-meters and \( \Omega \)-inches. The permeability \( \mu \) is the permeability of the conductor. It is a property of a material to support a magnetic flux. Some reference tables will show relative permeability. In this case the relative permeability is normalized to the permeability of free space which is: \( 4\pi 10^{-7} \) Henries per meter. The relationship between relative permeability to the actual permeability is given in Equation (1-5). Most conductors have a relative permeability \( \mu_r \) very close to one. Therefore conductor permeability \( \mu \) is often given the same value as \( \mu_o \).

\[
\mu = \mu_r \cdot \mu_o
\]  

(1-5)

where:

\( \mu = \text{actual permeability of the material} \)

\( \mu_r = \text{relative permeability of material} \)

\( \mu_o = \text{permeability of free space} \)

**Example 1.2-2**: Calculate the skin depth of copper wire at a frequency of 25 MHz.
Solution: From Table 1-2, $\rho = 6.61 \times 10^{-7} \ \Omega$-inches. Using Equation (1-4), and converting the permeability from H/m to H/inch, the skin depth is:

$$\delta = \frac{\frac{6.61 \cdot 10^{-7}}{(3.19 \cdot 10^{-8}) \pi (25 \cdot 10^6)}}{\sqrt{\left(\frac{6.61 \cdot 10^{-7}}{(3.19 \cdot 10^{-8}) \pi (25 \cdot 10^6)}\right)^2 + 1}} = 5.14 \cdot 10^{-4} \text{ inches}$$

Although we have been considering the skin depth in a circular wire, skin depth is present in all shapes of conductors. A thick conductor is affected more by skin effect at lower frequencies than a thinner conductor. One of the reasons that engineers are concerned about skin effect in conductors is the fact that as the resistance of the conductor increases, so does the thermal heating in the wire. Heat can be a destructive force in high power RF circuits causing burn out of conductors and potentially hazardous conditions to personnel. Also the frequency dependence of the skin effect may make it difficult to maintain the impedance of a transmission line structure. This effect will be examined in Chapter 2 with the study of transmission lines.

As the frequency increases, the current is primarily flowing in the region of the skin depth. It can be visualized from Figure 1-7 that a wire would have greater resistance at higher frequencies due to the skin effect. The resistance of a length of wire is determined by the resistivity and the geometry of the wire as defined by Equation (1-6).

$$R = \frac{\rho \ell}{A} \ \Omega \quad (1-6)$$

where:

$\rho$ = Resistivity of the wire  
$\ell$ = Length of the wire  
$A$ = Cross sectional area

Example 1.2-3: Calculate the resistance of a 12 inch length of AWG #24 copper wire at DC and at 25 MHz.

Solution: The radius of the wire can be found in Appendix A. The DC resistance is then calculated using Equation (1-6).
To calculate the resistance at 25 MHz the cross sectional area of the conduction region must be redefined by the skin depth of Figure 1-7. We can refer to this as the effective area, $A_{\text{eff}}$.

$$A_{\text{eff}} = \pi \left( R^2 - r^2 \right)$$  \hspace{1cm} (1-7)

where: $r = R - \delta$.

For the AWG#24 wire at 25 MHz the $A_{\text{eff}}$ is calculated as:

$$A_{\text{eff}} = \pi \left( \frac{0.0201}{2} \right)^2 - \pi \left( \frac{0.0201}{2} - 5.14 \times 10^{-4} \right)^2 = 3.14 \times 10^{-5} \text{ in}^2$$

Then apply Equation (1-6) to calculate the resistance of the 12 inch wire at 25 MHz.

$$R_{25\text{MHz}} = \frac{12 \left( 6.61 \times 10^{-7} \right)}{3.14 \times 10^{-5}} = 0.253 \Omega$$

We can see that the resistance at 25 MHz is more than 10 times greater than the resistance at DC.

### 1.2.4 Analytical Calculation of Flat Ribbon Inductance

Flat ribbon style conductors are very common in RF and microwave engineering. Flat ribbon conductors are encountered in RF systems in the form of low inductance ground straps. Flat ribbon conductors can also be encountered in microwave integrated circuits (MIC) as gold bonding straps. When a very low inductance is required the flat ribbon or copper strap is a good choice. The flat ribbon inductance can be calculated from the empirical Equation (1-8).

$$L = K \ell \left[ \ln \left( \frac{2\ell}{W+T} \right) + 0.223 \left( \frac{W+T}{\ell} \right) + 0.5 \right] \text{ nH}$$  \hspace{1cm} (1-8)
where: $\ell$ = The length of the wire
$K= 2$ for dimensions in cm and $K=5.08$ for dimensions in inches
$W=$ the width of the conductor
$T=$ the thickness of the conductor

**Example 1.2-4:** Calculate the inductance of the 3 inch Ribbon at 60 Hz, 1 MHz, and 1 GHz. Make the ribbon 100 mils wide and 2 mils thick.

**Solution:** The schematic of the Ribbon is shown in Figure 1-8 along with the table with the resulting reactance and inductance. Note that the same length of ribbon as the AWG#28 wire has almost half of the inductance value. Also note that the inductance value, 49.69 nH does not change over the 60 Hz to 1 GHz frequency range. It is why the ribbon is desirable as a low inductance conductor for use in RF and microwave applications.

![Figure 1-8](image)

**Figure 1-8** Flat ribbon schematic and tabular output of reactance and inductance

### 1.3 Physical Resistors

The resistance of a material determines the rate at which electrical energy is converted to heat. In Table 1-2 we have seen that the resistivity of materials is specified in $\Omega\cdot$meters rather than $\Omega$/meter. This facilitates the calculation of resistance using Equation (1-6). When working with low frequency or logic circuits we are used to treating resistors as ideal resistive components. If we examine the impedance of an ideal resistor in Genesys over a frequency range of 1 MHz to 3 GHz we obtain a plot as shown in Figure 1-9. This plot shows a constant resistance at all frequencies. At RF and microwave frequencies however, resistors also possess inductive and capacitive elements. The stray inductance and capacitance associated with a resistor are often called parasitic elements. Consider the leaded resistor as shown in Figure 1-10.
Figure 1-9 Ideal 50Ω resistor impedance versus frequency

Figure 1-10 Leaded resistor impedance versus frequency for various resistor values
For such a 1/8 watt leaded resistor it is not uncommon for each lead to have about 10 nH of inductance. The body of the resistor may exhibit 0.5 pF capacitance between the leads. Designing the network in Genesys reveals an interesting result of the impedance versus frequency response. The impedance plotted in Figure 1-10 shows the impedance for a 10 Ω, 50 Ω, 500 Ω, and 1 kΩ resistors swept from 0 to 2.0 GHz. By tuning the parasitic elements in Genesys we find that the low value resistors are influenced more by the lead inductance. The high value resistors are influenced more by the parasitic capacitance.

1.3.1 Chip Resistors

Thick film resistors are used in most contemporary electronic equipment. The thick film resistor, often called chip resistor, comes close to eliminating much of the inductance that plagues the leaded resistor. The chip resistor works well with popular surface mount assembly techniques preferred in modern electronic manufacturing. Figure 1-11 shows a typical thick film chip resistor along with a cross section of its design.

There are many types of chip resistors designed for specific applications. Common sizes and power ratings are shown in Table 1-3.

<table>
<thead>
<tr>
<th>Size</th>
<th>Length x Width</th>
<th>Power Rating</th>
</tr>
</thead>
<tbody>
<tr>
<td>0201</td>
<td>20mils x 10mils</td>
<td>50mW</td>
</tr>
<tr>
<td>0402</td>
<td>40mils x 20mils</td>
<td>62mW</td>
</tr>
</tbody>
</table>
Table 1-3 Standard thick film resistor size and approximate power rating

<table>
<thead>
<tr>
<th>Size</th>
<th>Dimensions</th>
<th>Power Rating</th>
</tr>
</thead>
<tbody>
<tr>
<td>0603</td>
<td>60mils x 30mils</td>
<td>100mW</td>
</tr>
<tr>
<td>0805</td>
<td>80mils x 50mils</td>
<td>125mW</td>
</tr>
<tr>
<td>1206</td>
<td>120mils x 60mils</td>
<td>250mW</td>
</tr>
<tr>
<td>2010</td>
<td>200mils x 100mils</td>
<td>500mW</td>
</tr>
<tr>
<td>2512</td>
<td>250mils x 120mils</td>
<td>1W</td>
</tr>
</tbody>
</table>

The thick film resistor is comprised of a carbon based film that is deposited onto the substrate. Contrasted with a thin film resistor that is typically etched onto a substrate or printed circuit board, the thick film resistor can usually handle higher power dissipation. The ends of the chip have metalized wraps that are used to attach the resistor to a circuit board. Genesys does not include native models for thick film resistors. Some manufacturers may provide models that can be incorporated into Genesys. There are also companies that specialize in developing CAD models of components such as Modelithics, Inc. Modelithics has a wide variety of component model libraries that can be incorporated into Genesys.

**Example 1.3-1:** Plot the impedance of 1 kΩ, 0603 size chip resistor, manufactured by KOA, from 0 to 3 GHz. This model is available in the Modelithics evaluation model kit. Select the Modelithics Library in the Genesys Part Selector as shown in Figure 1-12.

![Modelithics chip component library in Genesys](image)
**Solution:** Create a schematic with the resistor and sweep the impedance from 1MHz to 3 GHz. The resulting schematic and response are shown in Figure 1-13.

![Schematic of a resistor](image)

**Figure 1-13** Modelithics chip resistor model schematic and impedance versus frequency

Note the roll off of the impedance with increasing frequency. This suggests that the chip resistor does have a parasitic capacitance that is in parallel with the resistor similar to the discrete model of Figure 1-10.

### 1.4 Physical Inductors

In section 1.2 we introduced the topic of inductance. The inductance of straight cylindrical and flat rectangular conductors was examined. The primary method of increasing inductance is not to simply keep increasing the length of a straight conductor but rather form a coil of wire. Forming a coil of wire increases the magnetic flux linkage and greatly increases the overall inductance. Because of the greater surrounding magnetic flux, inductors store energy in the magnetic field. Lumped element inductors are used in bias circuits, impedance matching networks, filters, and resonators.
As we will see throughout this section inductors are realized in many forms including: air-core, toroidal and very small chip inductors. The concept of Q factor is introduced and will come up frequently in RF and microwave circuit design. It is a unit-less figure of merit that is used in circuits in which both reactive and resistive elements coexist. Because we know that individual air wound inductors, shown in Figure 1-14, are actually networks that are made up of resistors, inductors, and capacitors, each individual component is also characterized by a Q factor.

![Figure 1-14 Air wound inductor showing the wire resistance and inter-winding capacitance](image)

Basically, the higher the Q factor, the less loss or resistance exists in the energy storage property. The inductor quality factor Q is defined as:

\[ Q = \frac{X}{R_s} \]  

(1-9)

where:

- \(X\) is the reactance of the inductor
- \(R_s\) is the resistance in the inductor

At low RF frequencies the resistance comes primarily from the resistivity of the wire and as such is quite low. At higher frequencies the skin effect and inter-winding capacitance begin to influence the resistance and reactance thus causing the Q factor to decrease. In most applications we want as high a component Q factor as possible. We can increase the Q factor of inductors by using larger diameter wire or by silver plating the wire. In a multi-turn coil, the windings can be separated to reduce the inter-winding capacitance.
which in turn will increase the Q factor. Winding the coil on a magnetic core can increase the Q factor.

1.4.1 Air Core Inductors

Forming a wire on a removable cylinder is the basic realization of the air core inductor. When designing an air-core inductor, use the largest wire size and close spaced windings to result in the lowest series resistance and high Q. The basic empirical equation to calculate the inductance of an air core inductor is given by Equation (1-10)\(^2\).

\[
L = \frac{(17)^{N^{1.3}} \ (D+D1)^{1.7}}{(D1+S)^{0.7}}
\]

(1-10)

where:

- \(N\) = Number of turns of wire
- \(D\) = Core form diameter in inches
- \(D1\) = Wire diameter in inches
- \(L\) = Coil inductance in nH
- \(S\) = Spacing between turns in inches

Example 1.4-1: As an interesting comparison with Example 1.2.1 calculate the amount of inductance that we can realize in that same three inches of wire if we wind it around a core to form an inductor. Choose a core form of 0.095 inches diameter as a convenient form to wrap the wire around.

Solution: First we need to calculate the approximate number of turns that we can expect to have with the three inch length of wire. We know that the circumference of a circle is related to the diameter by the following equation.

\[
\text{Circumference} = \pi \ (\text{Diameter}) = \pi \ (0.095) = 0.2985 \ \text{inches}
\]

With a circumference of 0.2985 inches we can calculate the approximate number of turns that we can wrap around the 0.095 inch core with three inches of wire.

\[
N = \frac{3}{0.2985} = \text{approximately 10 turns}
\]
From Equation (1-10) we can see that the spacing between the turns has a strong effect on the value of the inductance that we can expect from the coil. When hand winding the coil, it may be difficult to maintain an exact spacing of zero inches between the turns. Therefore it is useful to solve Equation (1-10) in terms of a variety of coil spacing so that we can see the effect on the inductance. The Equation Editor in Genesys can be used to solve Equation (1-10) for a variety of coil spacing. Add an Equation Editor to the Workspace as shown in Figure 1-15.

**Figure 1-15** Adding an Equation Editor to the Genesys workspace

Create an Equation set similar to that shown in Figure 1-16.

![Equation set for calculating air core coil inductance](image)

**Figure 1-16** Equation Editor to calculate the air core coil inductance
Note that the coil spacing variable, Spacing, has been defined as an array variable. Placing a semicolon between the values makes the array organized in a column format. This is handy for viewing the results in tabular format. Placing a comma between the values would organize the array in a row format. The array variables are displayed under the variable column as Real [6], indicating that these values are a six element array containing scalar values. As an aid in forming the coil, the overall coil length is also calculated with the Equation Editor. The coil length is simply the summation of the overall wire thickness times the number of turns and the spacing between the turns. A quick way to add the results to a table is shown in Figure 1-17.

![Figure 1-17](image)

**Figure 1-17** Send the calculated variable to a tabular output directly from the Equation Editor

Right click on the variable and add it to an existing or new table. Add the calculated inductance, spacing, and coil length to the table as shown in Figure 1-18.

<table>
<thead>
<tr>
<th>Inductance</th>
<th>Spacing</th>
<th>Coil Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>163.78 nH</td>
<td>0</td>
<td>0.126</td>
</tr>
<tr>
<td>147.735 nH</td>
<td>0.002</td>
<td>0.144</td>
</tr>
<tr>
<td>135.037 nH</td>
<td>0.004</td>
<td>0.162</td>
</tr>
<tr>
<td>124.701 nH</td>
<td>0.006</td>
<td>0.18</td>
</tr>
<tr>
<td>116.087 nH</td>
<td>0.008</td>
<td>0.198</td>
</tr>
<tr>
<td>108.806 nH</td>
<td>0.01</td>
<td>0.216</td>
</tr>
</tbody>
</table>

![Figure 1-18](image)

**Figure 1-18** Coil Inductance versus coil spacing and length

The table shows that the coil inductance with no spacing between the turns is 163.78 nH and it is 0.126 inches long. Contrast this to the 93.1 nH inductance with the same three inches of wire in a straight length. We can
clearly see the dramatic impact of the magnetic flux linkage in increasing the inductance by forming the wire into a coil. The Figure also shows the strong influence of the inter-winding capacitance in influencing the inductance of the coil. Just 10 mils spacing between the turns reduces the coil’s inductance from 163.78 nH to 108.8 nH. It is clearly important to consider the turn spacing when analyzing the inductor’s performance. In practice this is an effective means to tune the inductor’s value in circuit. When designing and building the inductor it is necessary to solve Equation (1-10) for the number of turns given a desired value of inductance. Figure 1-19 shows an Equation Editor setup to solve for the number of turns, N.

<table>
<thead>
<tr>
<th>1</th>
<th>*Calculate Number of Turns Given the Inductance</th>
</tr>
</thead>
<tbody>
<tr>
<td>2</td>
<td>*Enter the following parameters</td>
</tr>
<tr>
<td>3</td>
<td>D=0.005</td>
</tr>
<tr>
<td>4</td>
<td>N=10</td>
</tr>
<tr>
<td>5</td>
<td>L=163.78</td>
</tr>
<tr>
<td>6</td>
<td>*Calculate the Number of Turns</td>
</tr>
<tr>
<td>7</td>
<td>N=((D1+Spacing)^2)(17\times(L+D1)^2.794)^(1.7)) (0.7692)</td>
</tr>
<tr>
<td>8</td>
<td>*Physical Length of Inductor</td>
</tr>
<tr>
<td>9</td>
<td>Coil Length=10.0<em>N+(Spacing,</em>(N-1))</td>
</tr>
</tbody>
</table>

**Figure 1-19** Equation Editor to calculate the air core number of turns

Note that there is one subtle difference in line 12, the calculation of the coil length. In this case both variables, Spacing and N, are array variables. When multiplying array variables use a period in front of the multiplication sign to signify that this is an operation on arrays. This makes sure that the correct array index is maintained between the variables. In the previous Equation Editor of Figure 1-16 this notation was not necessary because N was a constant.

<table>
<thead>
<tr>
<th>Number_of_Turns.N</th>
<th>Number_of_Turns.Spacing</th>
<th>Number_of_Turns.Coil_Length</th>
</tr>
</thead>
<tbody>
<tr>
<td>9.999</td>
<td>0</td>
<td>0.126</td>
</tr>
<tr>
<td>10.825</td>
<td>0.002</td>
<td>0.156</td>
</tr>
<tr>
<td>11.599</td>
<td>0.004</td>
<td>0.189</td>
</tr>
<tr>
<td>12.332</td>
<td>0.006</td>
<td>0.223</td>
</tr>
<tr>
<td>13.029</td>
<td>0.008</td>
<td>0.26</td>
</tr>
<tr>
<td>13.696</td>
<td>0.01</td>
<td>0.3</td>
</tr>
</tbody>
</table>

**Figure 1-20** Number of turns versus coil spacing and length
1.4.2 Modeling the Air Core Inductor in Genesys

The air core inductor is modeled in Genesys using the AIRIND1 model as shown in Figure 1-21.

![Genesys model of the air core inductor](image)

Figure 1-21 Genesys model of the air core inductor

Note the required parameters: number of turns, wire diameter, core diameter, and coil length. The coil spacing cannot be entered directly but is accounted for by the overall coil length for a given number of turns. Make the coil length a variable so that we can analyze the inductance as a function of the coil length. Simulate the value of inductance at a fixed frequency of 1 MHz. Set up a fixed frequency Linear Analysis as shown in Figure 1-23. Simulate the value of inductance at a fixed frequency of 1 MHz.

![Fixed frequency linear analysis at 1MHz](image)
To vary the length of the inductor we will use the Parameter Sweep capability in Genesys. Add a Parameter Sweep to the Workspace as shown in Figure 1-23.

![Figure 1-23 Adding the Parameter Sweep to the workspace](image)

On the Parameter-to-Sweep entry make sure that the coil length is selected from the drop down box. Under the Type-of-Sweep, select list and enter the six coil lengths that were calculated in the table of Figure 1-18. Finally create an Equation Editor as shown in Figure 1-24 to calculate the coil’s inductance from the simulated reactance and plot it as a function of the coil

```
1 \text{using} ["Sweep1\_Data"]
2 \text{reactance} = \text{im(Sweep1\_Data.ZIN1)}
3 \text{inductance} = \text{reactance} / (\text{Sweep1\_Data.R} * 2^{\text{PL}})
4 \text{setinddep("inductance","Sweep1\_Data.L3\_Swp\_P")}
```
length as shown in Figure 1-25. Note the use of the Setindep statement in line 5. This keyword is used to set the dependent and independent variables. The first variable in the statement is the dependent variable while the second is the independent variable. In this example the inductance is the dependent variable while the coil length is the independent variable. As the plot of Figure 1-25 shows the model has very close correlation with the inductance calculated with Equation (1-10).

![Figure 1-25 Plot of inductance versus coil length for the air inductor model](image)

Change the Linear Analysis to a linear frequency sweep with 401 points over a range of 1 MHz to 1300 MHz. Plot the impedance of the inductor across the frequency range as shown in Figure 1-26. Note the interesting spike, or increase in impedance that occurs around 1098.6 MHz. This is the parallel, self resonant, frequency of the inductor.
Figure 1-26 Impedance of the air core inductor as a function of frequency

The inductor is not an ideal component or a pure inductance but rather a network that includes parasitic capacitance and resistance. As an example, create a simple RLC network that gives an equivalent impedance response to Figure 1-26. One such circuit is shown in Figure 1-27.

Figure 1-27 Equivalent ideal element network of the air core inductor
Resonant circuits are covered in detail in chapter 4 but it is important to understand that each individual component such as the air core inductor has its own resonant frequency. The resonant frequency is the frequency at which the inductive reactance and capacitive reactance are equal and cancel one another. When this condition occurs in the inductor it is a parallel resonant circuit which results in a very high real impedance. If we plot the reactance along with the impedance a very interesting response is obtained. This response is shown in Figure 1-28.

![Figure 1-28 Impedance and reactance of the air core inductor](image)

The reactance, up to the resonant frequency, is positive but beyond resonance the reactance becomes negative. From basic circuit theory we know that a negative reactance is associated with a capacitor. Therefore above the Self Resonant Frequency, SRF, the inductor actually becomes a capacitor. In practice we want to make sure that our inductor really behaves like an inductor. A good design practice is to keep this self resonant frequency about four times higher than the frequency of operation. However using the inductor near its resonant frequency might make a good choke. A choke is a high reactance inductor often used to feed voltage to a circuit in
which all RF energy is blocked from the DC side of the circuit. The
inductor manufacturer will typically specify SRF of the inductor. It is
important to remember that the inductor’s SRF is the parallel resonant
frequency; not the series resonant frequency.

### 1.4.3 Inductor Q Factor

**Example 1.4-2:** Setup an Equation Editor to calculate the Q factor of the
inductor based on Equation (1-9).

**Solution:** The Equation Editor is shown in Figure 1-29.

![Figure 1-29 Equation Editor to calculate the Q factor of the inductor](image)

The plot of inductor Q factor versus frequency is shown in Figure 1-30.

![Figure 1-30 Air core inductor model Q factor versus frequency](image)
It is interesting to note that the Q factor peaks at a frequency well below the self resonant frequency of the inductor. The actual frequency at which the Q factor peaks will vary among inductor designs but is usually ranges from 2 to 5 times less than the SRF. Close winding spacing results in inter-winding capacitance, which lowers the self-resonant frequency of the inductor. Thus there is a tradeoff between maximum Q factor and high self resonant frequency. It is also noteworthy that the Q factor goes to zero at the self resonant frequency. Figure 1-30 shows the Q factor monotonously increasing beyond the self resonant frequency. This is erroneous and is due to the fact that the model used to simulate the inductor’s performance is invalid beyond the self resonance. The Air Core inductor model uses a simplified network similar to the one shown in Figure 1-27. Beyond self resonance the complexity and number of ideal elements need to increase in order to accurately model the inductor. A resistor needs to be added in series with the capacitor to begin to model the Q factor because the inductor is becoming a capacitor above self resonance. For most practical work however the native model will work fine because we should be using the inductor well below the self resonant frequency. A technique commonly used by microwave engineers to increase the Q factor of an inductor is to silver plate the wire. This can be modeled by setting Rho = 0.95 in the inductor model.

1.4.4 Chip Inductors

The inductor core does not have to be air. Other materials may be used as the core of an inductor. Similar in size to the chip resistor there is a large assortment of chip inductors that are popular in surface mount designs. The chip inductor is a form of dielectric core inductor. There are a variety of modeling techniques used for chip inductors. One of the more popular modeling techniques is with the use of S parameter files. The subject of S parameters is covered in chapter 3. At this point consider the S parameter file as an external data file that contains an extremely accurate network model of the component. Most component manufacturers provide S parameter data files for their products. It is a good practice to always check the manufacturer’s website for current S parameter data files. Coilcraft, Inc
is one manufacturer of chip inductors. A typical chip inductor is built with extremely small wire formed on a ceramic form as shown in Figure 1-31.

![Chip Inductor Diagram](image)

**Figure 1-31** chip inductors (*courtesy of Coilcraft, Inc*)

Chip inductors are manufactured in many standard chip sizes as shown in Table 1-4.

<table>
<thead>
<tr>
<th>Size</th>
<th>Length x Width</th>
</tr>
</thead>
<tbody>
<tr>
<td>0201</td>
<td>20mils x 18mils</td>
</tr>
<tr>
<td>0302</td>
<td>34mils x 15mils</td>
</tr>
<tr>
<td>0402</td>
<td>44mils x 20mils</td>
</tr>
<tr>
<td>0603</td>
<td>69mils x 30mils</td>
</tr>
<tr>
<td>0805</td>
<td>90mils x 50mils</td>
</tr>
<tr>
<td>1008</td>
<td>105mils x 80mils</td>
</tr>
<tr>
<td>1206</td>
<td>140mils x 56mils</td>
</tr>
<tr>
<td>1812</td>
<td>195mils x 100mils</td>
</tr>
</tbody>
</table>

**Table 1-4** Standard chip Inductor size and approximate power rating

The characteristic differences among the various sizes are more difficult to quantify than the chip resistors. A careful study of the data sheets is required for the proper selection of a chip inductor. In general the larger chip inductors will have higher inductance values. Often the smaller chip inductors will have higher Q factor. The impedance and self resonant frequency can vary significantly across the sizes as well as the current handling capability.

### 1.4.5 Chip Inductor Simulation in Genesys

The Genesys library has a collection of S parameter files for the Coilcraft chip inductors. Use the Part Selector to navigate to the Coilcraft chip
inductors and select the 180 nH 1008 series inductor. The Part Selector and inductor schematic are shown in Figure 1-32.

**Figure 1-32** Part selector to locate chip inductor and the circuit schematic

**Example 1.4-3:** Setup a Linear Analysis and plot the impedance of the inductor from 1 MHz to 1040 MHz.

**Solution:** Create an Equation Editor to calculate and plot the inductance from the impedance as shown in Figure 1-33. For reference the manufacturer’s specifications for the inductance is overlaid on the plot.

**Figure 1-33** Plot of the 180 nH chip inductor inductance versus frequency
The manufacturer specifies the inductance as 180 nH at a frequency of 25 MHz. From the marker on the plot we see that the simulation of the S parameter file has very close correlation measuring 180.1 nH. From the plot we can see that the nominal inductance value remains close to specification up to about 500 MHz. The self resonant frequency (parallel resonant) is specified as a minimum of 750 MHz. The actual SRF could be higher but it is guaranteed not be be less than 750 MHz. We can see from the plot in Figure 1-33 that the actual SRF is closer to 1 GHz. From Figure 1-34 we see that the manufacturer specifies a minimum Q factor of 45 at a frequency of 100 MHz.

![Figure 1-34 180 nH chip inductor Q factor versus frequency](image)

The plot of the Q factor derived from the S parameter file shows that the Q factor is 51.21 near 100 MHz. Manufacturers will often plot the Q vs frequency on a logarithmic scale. It is very easy to change the x-axis to a logarithmic scale on the rectangular graph properties window in Genesys. This allows us to have a visual comparison to the manufacturer’s catalog plot.
1.4.6 Magnetic Core Inductors

We have seen that the inductance of a length of wire can be increased by forming the wire into a coil. We can make an even greater increase in the inductance by replacing the air core with a magnetic material such as ferrite or powdered iron. Two popular types of magnetic core inductors are the rod core and toroidal core inductors shown in Figure 1-35.

![Figure 1-35 Rod and toroidal magnetic core inductors](image)

The magnetic field around an inductor is characterized by the magnetic force $H$, and the magnetic flux $B$. They are related by the level of the applied signal and the permeability, $\mu$, of the core material. This relationship is given by Equation (1-11).

$$B = \mu H$$

where:

$B = \text{Flux density in Gauss}$  
$H = \text{Magnetization intensity in Oersteds}$  
$\mu = \text{Permeability in Webers/Ampere-turn}$

This relationship is nonlinear in that as $H$ increases, the amount of flux density will eventually level off or saturate. We will consider the linear region of this relationship throughout the discussion of this text. In an iron core inductor the permeability of the magnetic core is much higher than an air core and produces a high flux density. This magnetic flux density for each type of inductor is shown in Figure 1-36.
The rod inductor has magnetic flux outside of the core as well as inside the core. Rod inductors that are used in tuned circuits generally require a metal shield around the inductor to contain this magnetic flux so that it does not interfere or couple to adjacent circuits and other inductors. The toroidal inductor flux remains primarily inside the core material. This suggests that the toroidal inductor experiences less loss and should have higher Q factor. This also gives the toroidal inductor a self-shielding characteristic and does not require a metallic enclosure. Because of its self-shielding properties and high Q factor the toroidal inductor is one of the most popular of all magnetic core inductors. However one advantage of the rod inductor is that it is much easier to tune. The coil can be wound on a hollow plastic cylindrical form in which the magnetic rod can be placed inside. The rod is then free to move longitudinally which can tune the inductance value.

Core materials are characterized by their permeability. Core permeability can vary quite a bit with frequency and temperature and can be confusing to specify for a given application. The stability of the permeability can change with the magnetic field due to DC current or RF drive through the inductor. As the frequency increases the permeability eventually reduces to the same value as air. Therefore iron core inductors are used only up to about 200 MHz. In general powdered iron can handle higher RF power without saturation and permanent damage. Ferrite cores have much higher permeability. The higher permeability of ferrite results in higher inductance values but lower Q factors. This characteristic can be advantageous in the design of RF chokes and broad band transformers. For inductors used in
tuned circuits and filters, however, the higher Q factor of powdered iron is preferred. The powdered iron cores are manufactured in a variety of mixes to achieve different characteristics. The iron powders are made of hydrogen reduced iron and have greater permeability and lower Q factor. These cores are often used in RF chokes, electromagnetic interference (EMI) filters, and switched mode power supplies. Carbonyl iron tends to have better temperature stability and more constant permeability over a wide range of power. At the same time the Carbonyl iron maintains very good Q factor making them very popular in RF circuits. These characteristics lead to the popularity of toroidal inductors of Carbonyl iron for the manufacture of RF inductors. There is a wide variety of sizes and mixtures of Carbonyl iron that are used in the design of toroidal inductors. A few of the popular sizes that are manufactured by Micrometals Inc. are shown in Table 1-5.

<table>
<thead>
<tr>
<th>Core Designator</th>
<th>OD, inches</th>
<th>ID, inches</th>
<th>Height, inches</th>
</tr>
</thead>
<tbody>
<tr>
<td>T30</td>
<td>0.307</td>
<td>0.151</td>
<td>0.128</td>
</tr>
<tr>
<td>T37</td>
<td>0.375</td>
<td>0.205</td>
<td>0.128</td>
</tr>
<tr>
<td>T44</td>
<td>0.440</td>
<td>0.229</td>
<td>0.159</td>
</tr>
<tr>
<td>T50</td>
<td>0.500</td>
<td>0.303</td>
<td>0.190</td>
</tr>
<tr>
<td>T68</td>
<td>0.690</td>
<td>0.370</td>
<td>0.190</td>
</tr>
<tr>
<td>T80</td>
<td>0.795</td>
<td>0.495</td>
<td>0.250</td>
</tr>
<tr>
<td>T94</td>
<td>0.942</td>
<td>0.560</td>
<td>0.312</td>
</tr>
<tr>
<td>T106</td>
<td>1.060</td>
<td>0.570</td>
<td>0.437</td>
</tr>
<tr>
<td>T130</td>
<td>1.300</td>
<td>0.780</td>
<td>0.437</td>
</tr>
<tr>
<td>T157</td>
<td>1.570</td>
<td>0.950</td>
<td>0.570</td>
</tr>
<tr>
<td>T200</td>
<td>2.000</td>
<td>1.250</td>
<td>0.550</td>
</tr>
<tr>
<td>T300</td>
<td>3.040</td>
<td>1.930</td>
<td>0.500</td>
</tr>
<tr>
<td>T400</td>
<td>4.000</td>
<td>2.250</td>
<td>0.650</td>
</tr>
</tbody>
</table>

Table 1-5 Partial listing of popular toroidal cores with designators

The inductance per turn of a toroidal inductor is directly related to its permeability and the ratio of its cross section to flux path length as given by Equation (1-12) \(^4\).
\[ L = \frac{4 \pi N^2 \mu A}{\text{length}} \quad \text{nH} \]  

(1-12)

where:

\begin{align*}
L_{nH} & = \text{inductance} \\
\mu & = \text{permeability} \\
A & = \text{cross sectional area} \\
\text{length} & = \text{flux path length} \\
N & = \text{number of turns}
\end{align*}

As Equation (1-12) shows the inductance is proportional to the square of the turns. A standard specification used by toroid manufacturers for the calculation of inductance is the inductive index, \( A_L \). The inductive index is typically given in units of nH/turn. The inductance can then be defined by Equation (1-13).

\[ L = N^2 A_L \quad \text{nH} \]  

(1-13)

Some manufacturers specify the \( A_L \) in terms of \( \mu H \) or mH. To convert among the three quantities use the following guideline.

\[ \frac{1 \text{nH}}{\text{turn}} = \frac{10 \text{ \( \mu \) H}}{100 \text{ turns}} = \frac{1 \text{ mH}}{1000 \text{ turns}} \]  

(1-14)

The various powdered iron mixes are optimized for good Q factor and temperature stability over certain frequency bands. A partial summary of some popular mixtures is shown in Table 1-6. Powdered Iron cores have a standard color code and material sub-type designator. The toroid is painted with the appropriate color so that the mixture can be identified. A given \( A_L \) is dependent on both the size of the toroid and the material mix.
<table>
<thead>
<tr>
<th>Material Mix Designator</th>
<th>Material Permeability</th>
<th>Magnetic Material</th>
<th>Color Code</th>
<th>Frequency Range</th>
<th>Temperature Stability (ppm/°C)</th>
</tr>
</thead>
<tbody>
<tr>
<td>-17</td>
<td>4.0</td>
<td>Carbonyl</td>
<td>Blue/Yellow</td>
<td>20 – 200 MHz</td>
<td>50</td>
</tr>
<tr>
<td>-10</td>
<td>6.0</td>
<td>Carbonyl W</td>
<td>Black</td>
<td>10 – 100 MHz</td>
<td>150</td>
</tr>
<tr>
<td>-6</td>
<td>8.5</td>
<td>Carbonyl SF</td>
<td>Yellow</td>
<td>2.0 – 30 MHz</td>
<td>35</td>
</tr>
<tr>
<td>-7</td>
<td>9.0</td>
<td>Carbonyl TH</td>
<td>White</td>
<td>1.0 – 20 MHz</td>
<td>30</td>
</tr>
<tr>
<td>-2</td>
<td>10.0</td>
<td>Carbonyl E</td>
<td>Red</td>
<td>0.25 – 10 MHz</td>
<td>95</td>
</tr>
<tr>
<td>-1</td>
<td>20.0</td>
<td>Carbonyl C</td>
<td>Blue</td>
<td>0.15 – 2.0 MHz</td>
<td>280</td>
</tr>
<tr>
<td>-3</td>
<td>35.0</td>
<td>Carbonyl HP</td>
<td>Grey</td>
<td>0.02 – 1.0 MHz</td>
<td>370</td>
</tr>
</tbody>
</table>

**Table 1-6** Partial listing of powdered iron core mixes and suggested frequency range

Because of the complex properties of the core material, the determination of the Q factor of a toroidal inductor can be difficult. It is not simply the magnetic material properties alone, but also the wire winding loss as well that determines the overall Q factor. These losses can then vary greatly with frequency, flux density, and the toroid and wire size. The optimal Q factor occurs when the winding losses are equal to the core losses \[^4\]. In general for a given inductance value and core mix a larger toroid will produce larger Q factors. Conversely for a given toroid size higher Q factor is achieved at higher frequency as the permeability decreases. The skin effect of the wire in the windings can have a significant impact on the achievable Q factor. Just as we have seen with the air core inductor, the inter-turn capacitance will also have a limiting effect on the resulting Q factor as well as the self resonant frequency. Manufacturers often provide a set of Q curves that the designer can use as a design guide for determining the toroidal inductor Q factor. Figure 1-37 shows a typical set of optimal Q curves for Carbonyl W core material at various toroid sizes.
Example 1.4-4: Design a 550 nH inductor using the Carbonyl W core of size T30. Determine the number of turns and model the inductor in Genesys.

Solution: From the manufacturer’s data sheet the $A_L$ value is 2.5 for a T30-10 toroidal core. Rearranging Equation (1-13) to solve for the number of turns we find that 14.8 turns are required.

$$ N = \sqrt{\frac{L}{A_L}} = \sqrt{\frac{550 \text{ nH}}{2.5}} = 14.8 $$

To reduce the winding loss we want to use the largest diameter of wire that will result in a single layer winding around the toroid. Equation (1-15) will give us the wire diameter.

$$ d = \frac{\pi \ ID}{N + \pi} $$  \hspace{1cm} (1-15)

where:
- $d$ = Diameter of the wire in inches
- $ID$ = Inner diameter of the core in inches (from Table 1-5)
- $N$ = Number of turns
therefore,
\[ d = \frac{\pi (0.151)}{14.8 + \pi} = \frac{0.4744}{19.942} = 0.0238 \text{ inches} \]

From Appendix A, AWG#23 wire is the largest diameter wire that can be used to wind a single layer around the T30 toroid. Normally AWG#24 is chosen because this is a more readily available standard wire size. The toroidal inductor model in Genesys requires a few more pieces of information. The Genesys model requires that we enter the total winding resistance, core Q factor, and the frequency for the Q factor, \( F_q \). As an approximation, set \( F_q \) to about six times the frequency of operation. In this case set \( F_q \) to (6) 25 MHz = 150 MHz. Then tune the value of Q to get the best curve fit to the manufacturer’s Q curve. We know that we have 14.8 turns on the toroid but we need to calculate the length of wire that these turns represent. The approximate wire length around one turn of the toroid is calculated from the following equation.

\[
\text{Length} = [2] \text{Height} + (\text{OD} - \text{ID}) \] \((\text{#turns})\)

Using the dimensions for the T30 toroid from Table 1-6 we can calculate the total length of the wire as 6.10 inches.

\[
[2](0.128) + (0.307 - 0.151) \cdot 14.8 = 6.10 \text{ inches}
\]

Then use the techniques covered in section 1.2.3 to calculate the resistance of the 6.10 inches of wire taking into account the skin effect. To get a better estimate of the actual inductor Q, use the \( F_q \) frequency rather than the operating frequency for the skin effect calculation. The resistance at 150 MHz for the AWG#24 wire is calculated as 0.30 \( \Omega \). Use an Equation Editor to calculate the Q and inductance. Figure 1-38 shows the schematic of the toroidal inductor and the simulated Q.
The Qc parameter of the model has been tuned to give a reasonably good fit with the manufacturer’s (T30) curve of Figure 1-37. Figure 1-39 shows that the simulated self resonant frequency of the inductor model is near 150 MHz.
Figure 1-40 shows the inductance that was calculated from the impedance. The inductance is exactly 550 nH at 10 MHz and begins to increase slightly to 563 nH at the design frequency of 25 MHz. Given the myriad of variables associated with the toroidal inductor, this simple model gives a good first order model of the actual inductor and will enable accurate simulation of filter or resonator circuits.

![Graph showing inductance vs frequency](image.png)

**Figure 1-40** Inductance value of the toroidal inductor model

### 1.5 Physical Capacitors

The capacitor is an electrical energy storage component. The amount of energy that can be stored is dependent on the type and thickness of the dielectric material and the area of the electrodes or plates. Capacitors take on many physical forms throughout electrical circuit designs. These range from leaded bypass capacitors in low frequency applications to monolithic forms in millimeter wave applications. Table 1-7 summarizes many of the applications in which capacitors are found. The table also shows some of the types of materials in which the capacitors are manufactured.

<table>
<thead>
<tr>
<th>Application</th>
<th>Dielectric Type</th>
<th>Notes</th>
</tr>
</thead>
</table>

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As Table 1-7 shows, the ceramic capacitors dominate the higher RF and microwave frequency applications. Two of the most popular of the ceramic capacitors are the single layer and multilayer ceramic capacitors, as shown in Figure 1-41. These capacitors are available with metalized terminations so that they are compatible with a variety of surface mount assembly techniques from hand soldering to wire bonding and epoxy attachment.

![Single Layer Capacitor](image1.png) ![Multi Layer Chip Capacitor](image2.png)

**Figure 1-41** Single layer and multi layer chip capacitor dimensions

### 1.5.1 Single Layer Capacitor

The single layer capacitor is one of the simplest and most versatile of the surface mount capacitors. It is formed with two plates that are separated by a single dielectric layer as shown in Figure 1-42. Most of the electric field (E) is contained within the dielectric however there is a fraction of the E field that exists outside of the plates. This is known as the fringing field.
The capacitance formed by a dielectric material between two parallel plate conductors is given by Equation 1-17 [1].

\[
C = (N - 1) \left( \frac{KA \varepsilon_r}{t} \right) (FF) \quad \text{pF}
\]  

(1-17)

where, 
- \(A\) = plate area
- \(\varepsilon_r\) = relative dielectric constant
- \(t\) = separation
- \(K\) = unit conversion factor; 0.885 for cm and 0.225 for inches
- \(FF\) = fringing factor; 1.2 when mounted on microstrip
- \(N\) = number of parallel plates.

**Example 1.5-1:** Consider the design of a single layer capacitor from a dielectric that is 0.010 inches thick and has a dielectric constant of three. Each plate is cut to 0.040 inches square.

**Solution:** When the capacitor is mounted with at least one plate on a large printed circuit board track, a value of 1.2 is typically used in calculation. The Equation Editor in Genesys can be used to solve Equations (1-17) as shown in Figure 1-43.

---

**Figure 1-42** Single layer parallel plate capacitor

**Figure 1-43** Single layer capacitance calculation
The single layer capacitor can be modeled in Genesys using the Thin Film Capacitor.

\[
C = (2 - 1) \left( \frac{\left( \frac{0.225}{0.010} \right) \left( 0.04 \times 0.04 \right) (3)}{0.1010 \times pF} \right) \Rightarrow 0.13 \ pF
\]

The ceramic dielectrics used in capacitors are divided into two major classifications. Class 1 dielectrics have the most stable characteristics in terms of temperature stability. Class 2 dielectrics use higher dielectric constants which result in higher capacitance values but have greater variation over temperature. The temperature coefficient is specified in either percentage of nominal value or parts per million per degree Celsius (ppm/°C). Ceramic materials with a high dielectric constant tend to dominate RF applications with a few exceptions. NPO (negative-positive-zero) is a popular ceramic that has extremely good stability of the nominal capacitance versus temperature.

<table>
<thead>
<tr>
<th>Dielectric Material</th>
<th>Dielectric Constant</th>
</tr>
</thead>
<tbody>
<tr>
<td>Vacuum</td>
<td>1.0</td>
</tr>
<tr>
<td>Air</td>
<td>1.004</td>
</tr>
<tr>
<td>Mylar</td>
<td>3</td>
</tr>
<tr>
<td>Paper</td>
<td>4 - 6</td>
</tr>
<tr>
<td>Mica</td>
<td>4 - 8</td>
</tr>
<tr>
<td>Glass</td>
<td>3.7 - 19</td>
</tr>
<tr>
<td>Alumina</td>
<td>9.9</td>
</tr>
<tr>
<td>Ceramic (low (\varepsilon_r))</td>
<td>10</td>
</tr>
<tr>
<td>Ceramic (high (\varepsilon_r))</td>
<td>100 – 10,000</td>
</tr>
</tbody>
</table>

Table 1-8 Dielectric constants of materials

### 1.5.2 Multilayer Capacitors

Multilayer capacitors, shown in Figure 1-44, are very popular in surface mount designs. They are physically larger than single layer capacitors and can be attached by hand or by automatic pick-and-place machines. As Figure 1-44 shows, the multilayer chip capacitor is a parallel array of capacitor plates in a single package. Due to this type of construction the chip capacitor can handle higher voltages than the single layer capacitor. The insulation resistance of the capacitor is its ability to oppose the flow of
electricity and is a function of the dielectric material and voltage. The insulation resistance is typically specified as a minimum resistance value in MΩ at a specified working voltage. The working voltage rating, WVDC, is the maximum DC voltage at which the capacitor can operate over the lifetime of the capacitor. The AC voltage rating is approximately one half of the WVDC value. The dielectric withstand voltage (DWV) is the electrical strength of the dielectric at 2.5 times the rated voltage. This is a maximum short term over-voltage rating and is usually specified as a length of time that the dielectric can withstand the 2.5 times the WVDC value without arcing through $[^6]$.

![Figure 1-44 Multilayer chip capacitor construction](image)

There is no physical model for the multilayer chip capacitor in Genesys. The designer must rely on S parameter files or Modelithic models as we have used for the chip resistor in section 1.3.1. Two of the major manufacturers of multilayer chip capacitors are American Technical Ceramics, ATC, and Dielectric Laboratories Inc., DLI. These manufacturers provide S parameter files for their capacitors that are readily available on the company websites. They also provide helpful software applications that can aid the designer in making decisions for the proper selection of chip capacitors in specific applications.

### 1.5.3 Capacitor Q Factor

RF losses in the dielectric material of a capacitor are characterized by the dissipation factor. The dissipation factor is also referred to as the loss tangent and is the ratio of energy dissipated to the energy stored over a period of time. It is essentially the capacitor’s efficiency rating. The
dissipation factor and other ohmic losses lead to a parameter known as the Equivalent Series Resistance, ESR. The dissipation factor is the reciprocal of the Q factor. Just as we have seen with resistors and inductors, the physical model of a capacitor is a network of R, L, and C components.

**Example 1.5-2:** Calculate the Q factor versus frequency for the physical model of an 8.2 pF multilayer chip capacitor shown in Fig. 1-45.

![Physical model of the 8.2 pF chip capacitor and impedance](image)

The capacitor has a series inductance and resistance component along with a resistance in parallel with the capacitance. The parallel resistor sets the losses in the dielectric material. The series resistance and inductance represent any residual lead inductance and ohmic resistances.

**Solution:** The values entered for the physical model and the Q factor can be obtained from the capacitor manufacturer. The plot of Figure 1-45 shows the impedance of the capacitor versus frequency. Note that the impedance decreases as would be expected until the self resonant frequency is reached. Above the self resonant frequency the impedance begins to increase suggesting that the capacitor is behaving as an inductor. The self resonant frequency is due to the series inductance resonating with the capacitance. At
resonance the reactance cancels leaving only the resistances \( R_1 \) and \( R_2 \). The parallel resistance, \( R_2 \), can be converted to an equivalent series resistance by Equation (1-18).

\[
R_2' = \frac{1}{1 + Q^2} R_2
\]  

(1-18)

These two series resistances can then be added to find the equivalent series resistance, ESR, as defined by Equation (1-19).

\[
ESR = R_1 + R_2'
\]  

(1-19)

The capacitor Q factor is then calculated by Equation (1-20). \( X_T \) is the total series reactance of the inductive and capacitive reactance.

\[
Q = \frac{X_T}{ESR}
\]  

(1-20)

As Figure 1-45 shows, the 8.2 pF multilayer chip capacitor has a series resonant frequency of 3313 MHz. A marker is placed on the trace indicating the impedance at the SRF as 0.149 \( \Omega \). Because the reactance is cancelled at the SRF, this impedance essentially becomes the ESR of the capacitor. \( R_2 \) of Figure 1-45 is extremely frequency dependent. This means that the capacitor’s Q factor is also extremely frequency dependent. An improved model for analyzing the capacitor’s characteristics over a wide frequency range is to use the Genesys model for capacitor with Q. Using this model the Q factor of the capacitor can be made proportional to the square root of the applied frequency. The capacitor Q factor can be calculated from the impedance using the Equation Editor as shown in Figure 1-46. The resistive and inductive components are used to define the Q factor.

```
1 using("Linear1_Data")
2 reactance1=im(Linear1_Data.ZIN1)
3 resistance1=rc(Linear1_Data.ZIN1)
4 Q_physical=abs(reactance1)/resistance1
5 capacitance_model=1/(abs(reactance1).*F^2+3.14)
```

**Figure 1-46** Equation Editor used to calculate Q factor and capacitance
Figure 1-47 shows the large dependence on the capacitor Q with frequency. At 1006 MHz the Q factor is 120 while at 100 MHz the Q factor is greater than 1000. The Q factor goes to zero at the self resonant frequency. Above the self resonant frequency the Q is undefined.

**Figure 1-47** Calculated Q factor of 8.2 pF chip capacitor and equivalent model

The Equation Editor of Figure 1-46 also calculates the effective capacitance from the total reactance of the model using Equation (1-21).

\[
C = \frac{1}{2\pi F X_T}
\]  

The plot of Figure 1-48 reveals some interesting characteristics about the chip capacitor. From 100 MHz to 300 MHz the capacitance value is fairly constant at 8.217 pF. As the frequency increases above 300 MHz we see that the capacitance actually increases. The parasitic inductive reactance of the capacitor package actually makes the effective capacitance greater than its nominal value.
This is a property of the capacitor that is not always intuitive. As the frequency approaches the self resonant frequency the capacitance rapidly approaches infinity. The capacitor actually becomes nearly a short circuit to RF at the self resonant frequency. This is an important property of the capacitor that is used frequently in RF and microwave design. In RF coupling or bypass capacitor applications, capacitors are very often used at or near their self resonant frequency. The capacitor’s self resonant frequency is due to the series resonant circuit. In applications requiring bypassing over a wide range of frequencies it is often necessary to use several capacitors each selected to have a uniquely spaced self resonant frequency. In filter and other tuned circuit applications where we want the chip to appear as an 8.2 pF capacitor, we clearly must stay well below the self resonant frequency of the capacitor. A typical rule-of-thumb is to use the capacitor over a frequency range up to 35% of the self resonant frequency. Therefore the 8.2 pF chip capacitor with a self resonant frequency of 3313 MHz would be used as a capacitor in tuned circuits up to a frequency of about 1160 MHz. Figure 1-48 shows that at frequencies above 1160 MHz the capacitance is extremely nonlinear. However the capacitor could be used as a DC blocking, RF coupling capacitor that would efficiently pass microwave energy near its SRF of 3300 MHz.

Figure 1-48 Effective capacitance of the 8.2 pF chip capacitor
**References and Further Reading**

[1] Besser Associate’s *Applied RF Techniques Course*, 480 San Antonio Road, Mountain View, CA. 94040


[3] *Design Guide, Microwave Components* Inc., P.O. Box 4132, South Chelmsford, MA 01824


**Problems**

1-1. Calculate the wavelength of an electromagnetic wave operating at a frequency of 428MHz.
1-2. Calculate the inductance of a 5 inch length of AWG #30 straight copper wire.

1-3. Using the Equation Editor, calculate the reactance of the wire from Problem 2 at 10 Hz, 10 MHz, and 10 GHz. Create a Linear analysis in Genesys and display the wire impedance vs. frequency.

1-4. Calculate the resistance of a 12 inch length of AWG #24 copper wire at DC and at 25MHz

1-5. Find the skin depth and the resistance of a 2 meter length of copper coaxial line at 2 GHz. The inner conductor radius is 1 mm and the outer conductor is 4 mm.

1-6. Calculate the inductance of a 5 inch length of copper flat ribbon conductor. The dimensions of the ribbon are 0.100 inches in width and 0.002 inches thick.

1-7. Model a chip resistor (size 0603) with a resistance of 50Ω in Genesys. Consider an application in which 50Ω impedance must be maintained with +10%. Create a Linear Analysis and determine the maximum usable frequency of the chip resistor.

1-8. Design an air core inductor with an inductance value of 84nH. Use a copper wire of 0.050 inch diameter wound on a core diameter or 0.100 inch. Determine the number of turns required assuming a tight spaced winding.

1-9. Using the inductor from Problem 8 and the techniques developed in Section 1.4.1 examine the change in the coil inductance as the turn spacing is increased from zero to 0.10 inch in 0.002 inch increments.

1-10. Using the inductor from Problem 8, determine the self resonant frequency of the inductor and comment on the maximum frequency in which the inductor may be used in a tuned circuit application.
1-11. Using the inductor from Problem 8, determine the maximum $Q$ factor of the inductor and the frequency at which the maximum $Q$ factor is obtained.

1-12. In Genesys, select a chip inductor from the CoilCraft library with an inductance value of 80nH. Determine the maximum $Q$ factor of the inductor and comment on the maximum usable frequency of the inductor in a filter application.

1-13. Design a 1mH toroidal inductor on a Carbonyl W core size T30. Determine the maximum wire size that could be used to realize a single layer winding.

1-14. Using the inductor from Problem 13, model the inductor in Genesys and determine the approximate self resonant frequency. Comment on the maximum usable frequency of the inductor in the front end of a radio receiver.

1-15. A 0.05pF capacitor is required to couple a transistor to the resonator of a microwave oscillator. Design a single layer capacitor using a 0.020 inch thick dielectric with $e_r=2.2$. Determine the dimensions of the capacitor assuming square footprint is desired.

1-16. For the single layer capacitor of Problem 15, determine the dimensions of the capacitor with a dielectric constant $e_r=10.2$.

1-17. In Genesys model a 47pF chip capacitor using the ATC 0603 model from the Modelithics library. Determine the $Q$ factor of the capacitor at a frequency of 1000MHz.

1-18. Determine the self resonant frequency of the capacitor in Problem 17 and comment on the maximum frequency that this capacitor could be used in a tuned circuit. At what frequency does the capacitor have the lowest amount of energy loss?