

Microwave Frequencies, Conventional Passive Elements, and Resonators

1.1 Introduction

Maxwell's equations govern all types of electromagnetic behavior at all frequencies. At lower frequencies (below about 500 MHz) where lumped components are less than $\lambda/8$ in length, it is convenient to define the component as a frequency-independent R , L , or C . As the frequency increases, the component will have distributed effects or added phase shift, which must be accounted for in the analysis. These effects will be discussed in this chapter [1–5].

1.2 Transition from Radio Frequency to Microwave Circuits

As electrical engineers, we first learn to analyze lumped-element circuits consisting of R , L , and C components, which are independent of frequency; the impedance or admittance of the components is linearly dependent on frequency. Then we learn that at high frequencies, additional parasitic effects must be included in the model of the component. For lumped components at low frequencies, signals travel at essentially the speed of light instantaneously through points in space where the component is located. At microwave frequencies (defined as 300 MHz–300 GHz, with frequencies above 30 GHz also called millimeter waves), the electrical components used to generate and process signals have a size similar to one-eighth of a wavelength. Then we must return to Maxwell's equations that are subject to the appropriate boundary conditions.

The radio frequency (RF) range is generally defined as 30–300 MHz, but it should be a frequency where all of the lumped components are shorter than about $\lambda/8$ in length. In the 100–500 MHz range, we are in a gray area where the circuit might be treated as either lumped or distributed, provided the lumped equivalent circuit is accurate. The distributed transmission lines are very long in this frequency range and thus are not used very often. Even at 850 MHz, lumped components are successfully used (e.g., cell phones). The distributed transmission line circuits would be very large for an 850 MHz cell phone but much more realistic at 5.8 GHz, a future band for cell phones and other nonlicensed products [2].

To illustrate this point, we calculate the $\lambda/4$ length on FR-4, a low-cost substrate at 850 MHz using an effective dielectric constant of 3:

$$\frac{\lambda}{4} = \frac{c}{4f\sqrt{\epsilon_{\text{eff}}}} = \frac{7.5}{1.732 \cdot 0.85} = 5.09 \text{ cm} = 2.00 \text{ in} \quad (1.1a)$$

which is extremely large for a hand-held cellular phone circuit. A higher dielectric constant will reduce this dimension.

For transmission line circuits, we use transverse electromagnetic (TEM) or quasi-TEM microstripline for the passive circuits, which became available in 1965 [6]. The impedance and phase velocity were given by Wheeler in this chapter. It was immediately applied to the design of an X-band solid-state radar at Texas Instruments [7] using hybrid technology based upon alumina substrates and millimeter-wave monolithic circuits using gallium arsenide (GaAs) substrates at 94 GHz [8].

To understand this transition from RF to microwave or distributed elements, a plot of wavelength versus frequency is given in Figure 1.1 for various dielectric constants, where dielectrics effectively slow the velocity as ϵ_r increases.

It is important for the loss tangent to be low, which is true for alumina and GaAs but not for silicon. With GaAs substrates, $\rho > 10^6 \Omega\text{-cm}$ is easily achieved and maintained throughout the processing cycle. With Si substrates of high resistivity ($10^3 \Omega\text{-cm}$), the n-type may become p-type with processing, and the resistivity is likely to lower, which is unacceptable for monolithic circuits. One of the most challenging problems today is to achieve a high-Q inductor on silicon, where a Q above 10 at 2 GHz is about the best that can be achieved at the time of this writing, as shown later in this chapter.

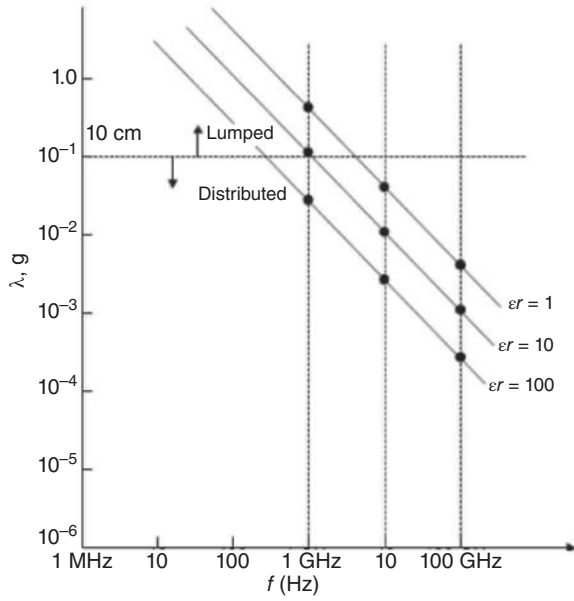


Figure 1.1 Guide wavelength versus frequency with dielectric constant as a variable. [2] / John Wiley & Sons.

Referring to Figure 1.1, the dielectric constant of most interesting dielectrics for monolithic integrated circuits (MICs) is of the order of 10, so the border between lumped and distributed may be estimated at $\lambda = 0.1$ m, or 10 cm, which coincides with the previous division at about 0.5–1 GHz.

The calculation of the guide wavelength for an effective dielectric constant of 10 follows at 1 GHz:

$$\lambda_g = \frac{c}{f\sqrt{\epsilon_{ff}}} = \frac{3 \times 10^{10}}{10 \times 10^9} = 10 \text{ cm} = 0.1 \text{ m} = 4 \text{ in}; \quad \frac{\lambda_g}{8} = 0.5 \text{ in} \quad (1.1b)$$

In other words, at frequencies above 1 GHz, the circuit must be treated as distributed since a length of more than 0.5 in will have distributed or phase properties that cannot be modeled by a simple R, L, C component [2].

At low frequency, we use lumped components with transistors and diodes as needed, that is, R, L , and C . When the components become about $\lambda/8$ long (approximately 500 MHz–1 GHz), we may add transmission line components (usually microstripline) in addition to lumped components. When the free-space wavelength becomes less than 1 mm (millimeter wave), designers are usually forced to use distributed transmission line elements where possible. Other forms of transmission are also used due to the limitations of TEM stripline/microstripline transmission lines [9], such as waveguides, surface modes, slotlines, coplanar waveguides, inverted microstriplines, and suspended microstriplines [10]. The geometry for these forms of TEM, transverse-electric (TE), and transverse-magnetic (TM) lines is given in Figure 1.2. It is useful to keep in mind that two wires (or conductors) are needed for TEM and only one conductor is required for TE and TM waves, which are generally at higher frequencies [2, Chapter 1].

Transmission lines may be modeled as in Figure 1.3, leading to the telegrapher equations, which are a time-domain description of the line:

$$\frac{\partial v(z, t)}{\partial z} = -RI(z, t) - L \frac{\partial I(z, t)}{\partial t} \quad (1.2)$$

$$\frac{\partial I(z, t)}{\partial z} = -Gv(z, t) - C \frac{\partial v(z, t)}{\partial t} \quad (1.3)$$

For sinusoidal steady-state conditions, this may be simplified to

$$\frac{dV(z)}{dz} = -(R + j\omega L)I(z) \quad (1.4)$$

$$\frac{dI(z)}{dz} = -(G + j\omega C)V(z) \quad (1.5)$$

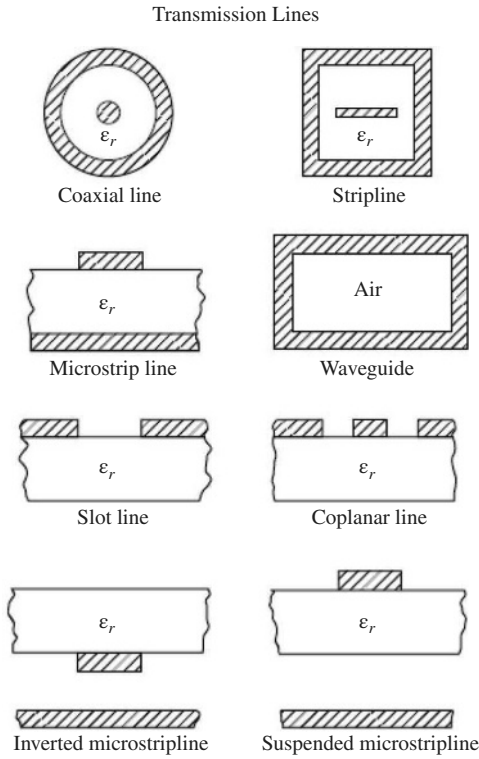


Figure 1.2 Geometry for microwave transmission. [2] / John Wiley & Sons.

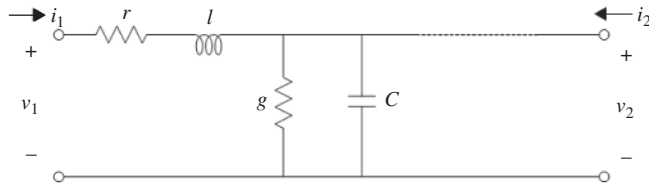


Figure 1.3 Lumped-element equivalent circuit of transmission line. [2] / John Wiley & Sons.

which is noted to be very similar to Maxwell's curl equations:

$$\nabla \times E = -j\omega\mu H \quad (1.6)$$

$$\nabla \times H = j\omega\epsilon E \quad (1.7)$$

Combining the telegrapher equations leads to

$$\frac{d^2 V(z)}{dz^2} = -\gamma^2 V(z) = 0 \quad (1.8)$$

$$\frac{d^2 I(z)}{dz^2} - \gamma^2 I(z) = 0 \quad (1.9)$$

$$\text{where } \gamma = \alpha + j\beta = \sqrt{(R + j\omega L)(G + j\omega C)} \quad (1.10)$$

is the complex propagation constant, which is a function of frequency. Traveling-wave solutions can be found as

$$V(z) = V_0^+ e^{-\gamma z} + V_0^- e^{\gamma z} \quad (1.11)$$

$$I(z) = I_0^+ e^{-\gamma z} + I_0^- e^{\gamma z} \quad (1.12)$$

With a few more steps, we may obtain the voltage waveform in the time domain as

$$V(z, t) = |V_0^+| \cos(\omega t - \beta z + \varphi^+) e^{-\alpha z} + |V_0^-| \cos(\omega t + \beta z + \varphi^-) e^{-\alpha z} \quad (1.13)$$

where $\varphi^\pm$ is the phase angle of the complex voltage $V_0^\pm$.

The wavelength on the line is

$$\lambda = \frac{2\pi}{\beta} \quad (1.14)$$

and the phase velocity is

$$v_p = \frac{\omega}{\beta} = \lambda f \quad (1.15)$$

where the actual time delay must be calculated using the group velocity, defined by

$$v_g = \frac{d\omega}{d\beta} \quad (1.16)$$

For a TEM wave, these two velocities are the same.

1.2.1 Example: 800 MHz Voltage-controlled Oscillator

In order to demonstrate a transition from RF to microwave and ultimately from lumped to distributed elements and move toward microwave frequencies, we analyzed an 800 MHz voltage-controlled oscillator in full detail. We also did the parameter extraction for the Motorola transistor. Figure 1.4 shows the circuit, a Colpitts oscillator that uses RF feedback in the form of a $15\ \Omega$ resistor between the emitter and the base feedback. The capacitive voltage divider consists of $1\ \text{pF}$ between the base of bipolar junction transistor (BJT) and the feedback resistor and $1\ \text{pF}$ between the feedback resistor and common. The difference between the conventional analysis and this case is that the parasitic parts of the lumped elements were taken into consideration. On purpose, the tuned circuit is loosely coupled to this part of the transistor circuit. Figure 1.5 shows a comparison between predicted and measured phase noise for this oscillator.

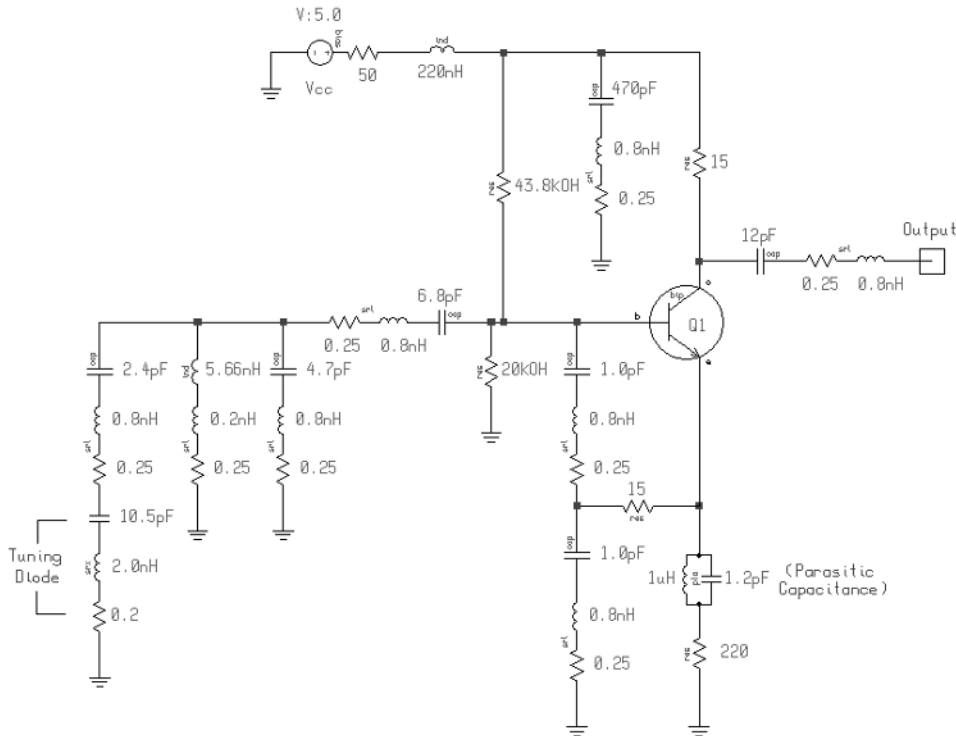


Figure 1.4 Colpitts oscillator that uses RF negative feedback between the emitter and capacitive voltage divider.

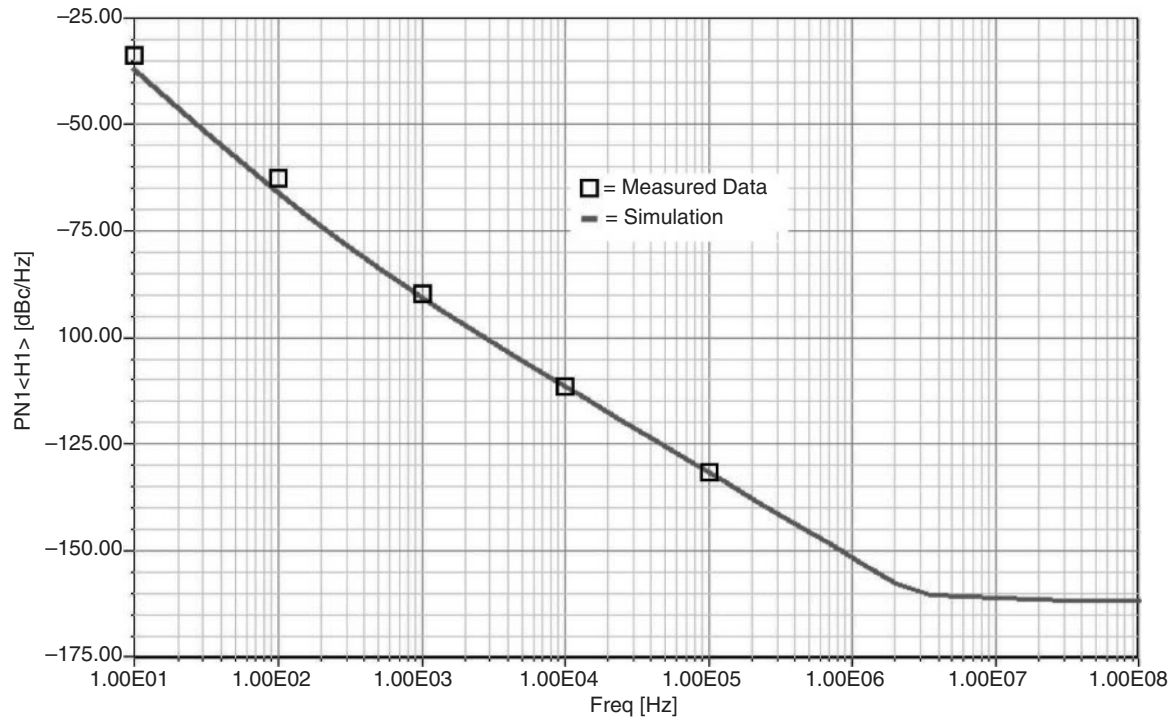


Figure 1.5 Comparison between predicted and measured phase noise for the oscillator shown in Figure 1.4.

To be realistic, we have also used real components rather than ideal ones. The suppliers for the capacitors and inductors provide some typical values for the parasitics. The major changes are 0.8 nH and 0.25 Ω in series with the 1.0 pF capacitors. The same thing applies for the main inductance, which has a parasitic connection inductance of 0.2 nH in series with a 0.25 Ω resistance. These types of parasitics are valid for a fairly large range of components assembled in surface-mount applications. Most engineers model the circuit only by assuming non-lossy/perfect devices, not adding these important parasitics. One of the side effects we have noticed is that the output power is more realistic and, needless to say, the simulated phase noise now agrees quite well with measured data. This circuit can also serve as an example for modeling amplifiers and mixers using surface-mount components. Good computer aided design (CAD) tools provide a library of passive elements in the form of measured *S*-parameters where all the parasitics are considered.

Figure 1.5 shows the simulated phase noise of the oscillator and some of the measured points. Following this example, we now apply the same logic to an amplifier.

1.3 RF Versus Microwave Modeling

Let us now take a look at a simple amplifier that shows a textbook-type approach to an RF circuit.

The circuit shown in Figure 1.6 is too simplified, as it assumes ideal components that are not realistic. For easy understanding, the transistor bias circuit is not shown [1].

In order to decouple the transistor from the biasing, we use RF chokes and DC decoupling (bypass) capacitors. As frequency increases, these inductors are either not manufacturable or have such a low *Q* that their use becomes questionable. This is the point where one should introduce the so-called distributed elements.

Figure 1.7 shows the very same circuit but resorting to distributed rather than lumped elements. The amplifier shown earlier, on purpose, illustrates a somewhat exaggerated case as we show interdigital capacitors, radial stubs, and transmission lines. Not all of them are necessary at the same time, but the two T-junctions are really important for interconnection. Most designers do not go to this level of detail, including showing vias for grounding. In summary, this circuit diagram may be a little overboard, but it indicates how complicated things can become [1].

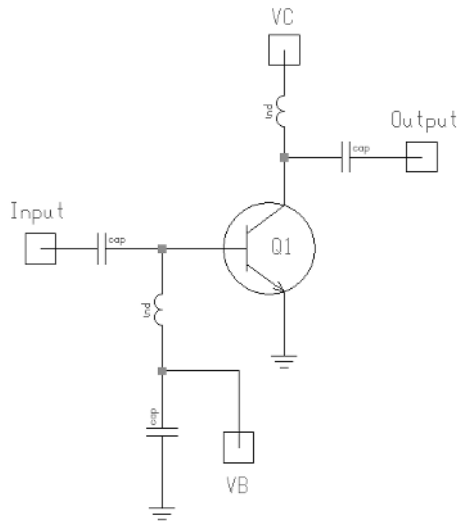


Figure 1.6 Simple BJT RF amplifier with lumped elements used for bypassing and DC blocking and for base bias and collector supply feeds. [1] / John Wiley & Sons.

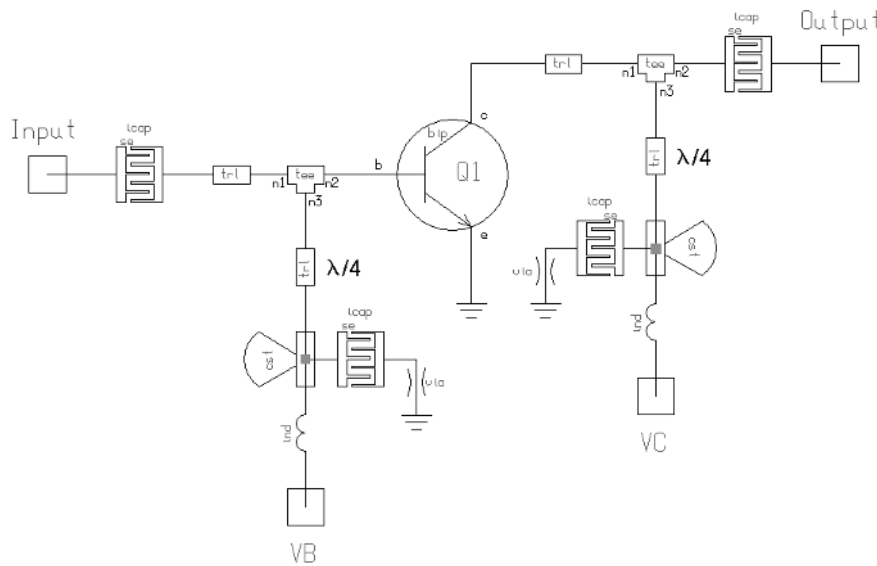


Figure 1.7 Simple BJT RF amplifier with distributed elements used for bypassing and DC blocking and for base bias and collector supply feeds. [1] / John Wiley & Sons.

The elements we have introduced now (which are part of any good, up-to-date CAD tool; see the element library for details on how to use their physical element descriptions) are listed as follows:

The transmission line is one of the most important interconnections at microwave frequencies.

Any printed connection between two points on a circuit board is a transmission line (Figure 1.8). Its characteristic electrical impedance is a function of the square root of the dielectric constant (ϵ_r), the width, metallization thickness, height above substrate of the line, and the loss tangent of the substrate. Since lines frequently have to be laid out in the form of curved connections or have a bend in their direction, we have to add elements capable of describing the high-frequency consequences of such connections such as mitered and radial bend elements that perform this function [1].

Now to the printed circuit board (PCB) carrier materials, the following is a list of carrier materials useful from very high frequency (VHF) to super high frequency (SHF) range.

**Dielectric and material properties of common RF and microwave substrates and base materials (hard and soft substrates)
(based on Rogers's corporation).**

Material	Dielectric permittivity at 10 GHz	Loss tangent at 10 GHz (10e4 tan)	Surface roughness (μm)	Thermal conductivity (W/mK)	Dielectric strength (kV/cm)	Soft or hard substrate
Alumina (Al_2O_3)	9.9	1–2	0.05–0.25	30–37	4000	Hard
Aluminum nitride (AlN)	8.9	3–5	0.05–0.6	150–170	150	Hard
Gallium arsenide (GaAs)	12.85	6	0.025	30	350	Hard
Gallium nitride (GaN)	9.5/10.4	–	0.025	66–225	4000	Hard
Indium phosphide (InP)	12.4	10	0.025	40	350	Hard
Silicon (Si)	11.9	10–100	<0.001	100–150	300	Hard
Silicon carbide (SiC)	10.8	20	<0.001	120–200	2.5×10^3	Hard
Duroid 5870	2.33	12	–	0.26	–	Soft
Duroid 5880	2.2	12	–	0.26	–	Soft
Duroid 6002	2.94	12	–	0.44	–	Soft
Duroid 6010	10.2–10.8	27	–	0.48	–	Soft
Duroid RO3003	3	13	–	0.50	–	Soft
Duroid RO3006	6.15	25	–	0.61	–	Soft
Duroid RO3010	10.2	35	–	0.66	–	Soft
Duroid RO4003	3.38	27	–	0.64	–	Soft
Duroid RO4350B	3.48	40	–	0.62	–	Soft
TMM 3	3.27	20	–	0.7	–	Soft
TMM 4	4.5	20	–	0.7	–	Soft
TMM 6	6	23	–	0.72	–	Soft
TMM 10	9.2	22	–	0.76	–	Soft
TMM 10i	9.8	20	–	0.76	–	Soft
CuFlon	3.05	30	–	0.25	–	Soft
CLTE	2.94	0.0025	–	0.5	–	Soft
FR-4	4.5–4.8	220	~6	0.16	–	Soft
Norclad	3.8	11	–	0.23	–	Soft
Polyimide	2.3–13.4	50	–	0.04–1.7	–	Soft
PTFE (Teflon)	3.05	30	–	0.25	–	Soft

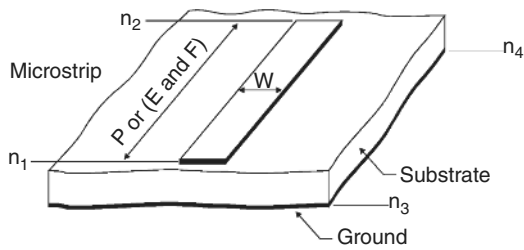


Figure 1.8 Transmission line in microstrip. [1] / John Wiley & Sons.

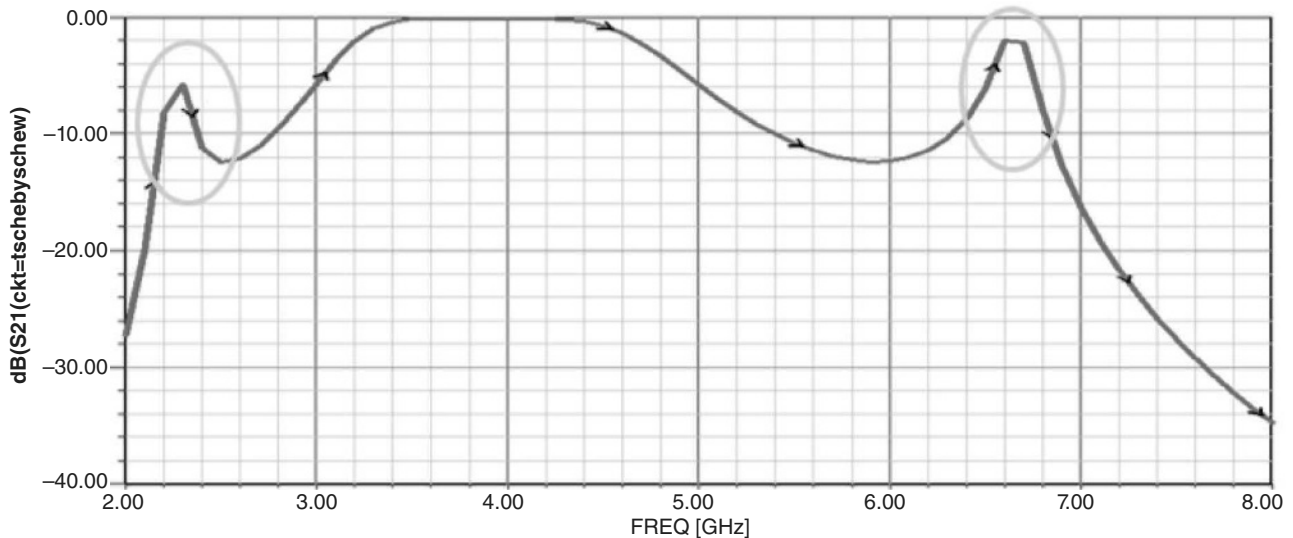


Figure 1.9 The frequency response of the circuit.

1.4 Application of Passive Devices in Filters

An important application of passive devices, such as lumped and distributed elements, is their use in filters. The following are some practical applications.

1.4.1 Example: 4 GHz Filter

The first example is a 4 GHz filter circuit diagram published in 1977 at the European Microwave Conference. In essence, it has two parallel-tuned circuits, with two series resonant circuits.

Figure 1.9 shows some interesting frequency responses outside the pass band, as the ears on the frequency-response curve. Figure 1.10 shows the simulated return loss in dB of the 4 GHz filter. The -50 dB return loss is based on numerical accuracy, not on actual losses. In reality, the return loss would be more like -30 dB.

The filter in Figure 1.11 is a fifth-order 4 GHz Chebyshev filter, calculated for lumped elements.

Using a modern filter-synthesis program with internal impedance transformation, we get the schematic and CAD simulated results for a higher-order 4 GHz filter shown in Figure 1.12, with much better behavior.

Essentially, the filter in Figure 1.12 is also a fifth-order filter, which looks more complicated because of the impedance transformation, but is now much better behaved, as seen in Figure 1.13.

Frequency response of the synthesized filter using realistic components ($Q = 50$), resulting in an insertion loss of 1.4 dB. The filter has no spurious resonances.

The return loss of about -20 dB is quite reasonable (Figure 1.14).

At frequencies above 500 MHz, lumped elements are typically replaced by transmission line elements.

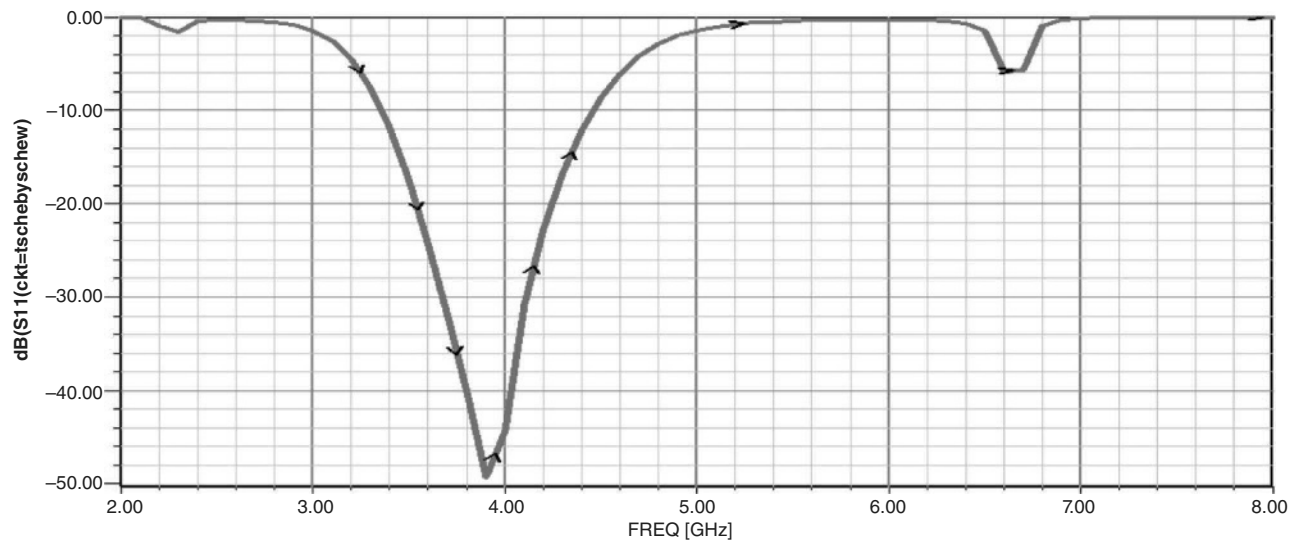


Figure 1.10 Return loss in dB of the 4 GHz filter.

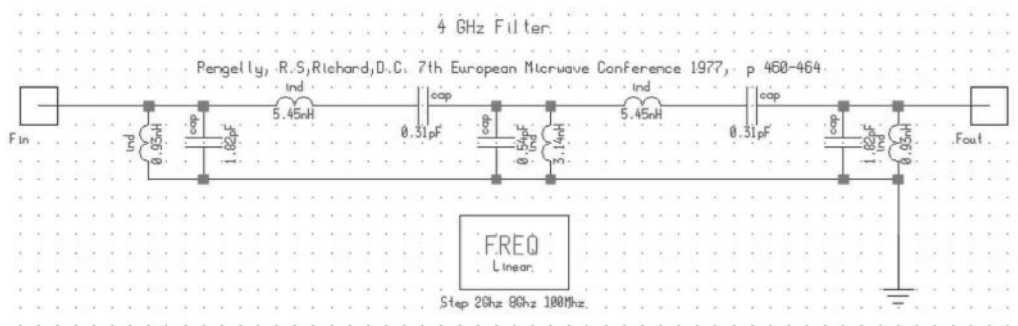


Figure 1.11 4 GHz center-frequency bandpass filter using lumped elements.

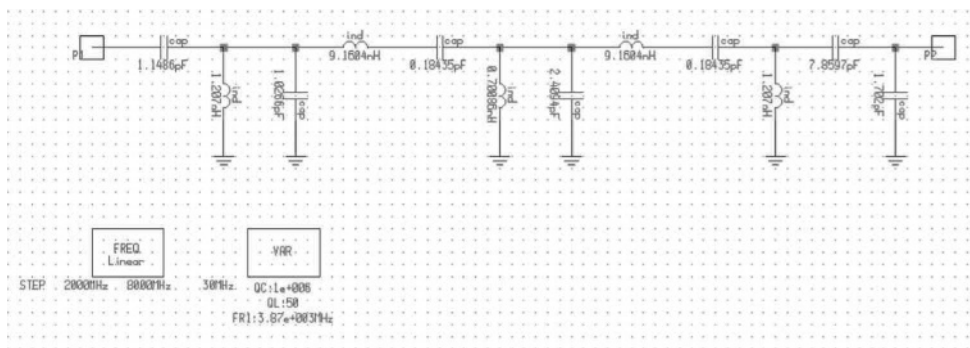


Figure 1.12 New 4 GHz synthesized filter. *Note:* The impedance transformation at the input and output of the filter.

1.5 Richards Transformation

Most classical transmission line-based circuits including filters are designed on the basis of one of the fundamental properties of the transmission line, as they cannot be realized in the form of lumped elements. Consider a length of transmission line with characteristic impedance Z_t and electrical length θ , short-circuited at one end. The impedance at the input will be

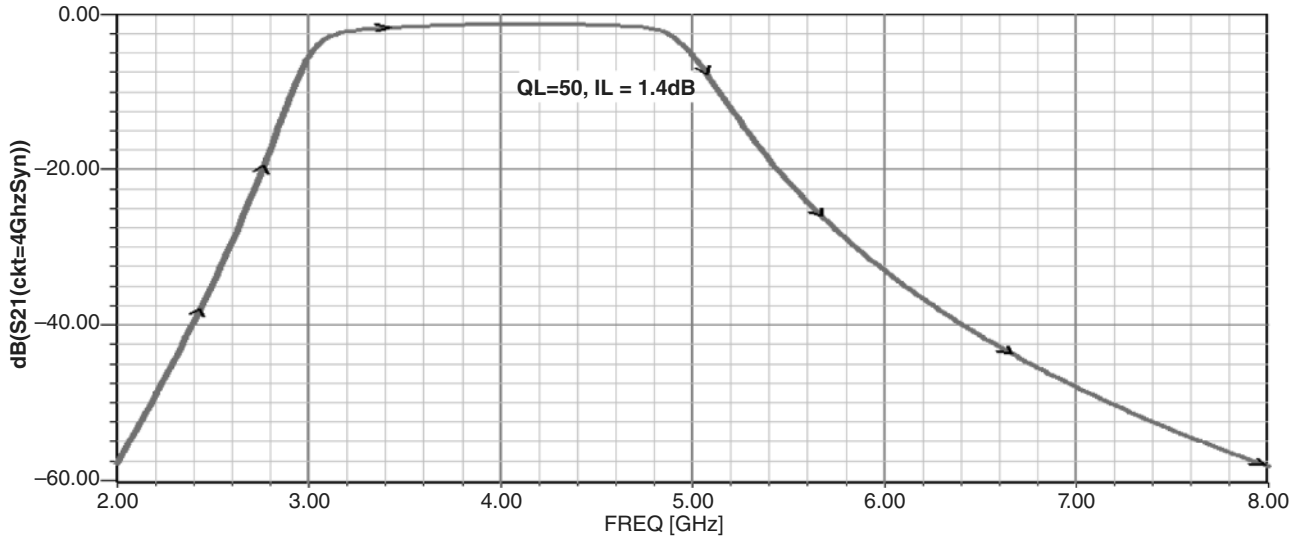


Figure 1.13 Frequency response of the synthesized filter using realistic components ($Q = 50$).

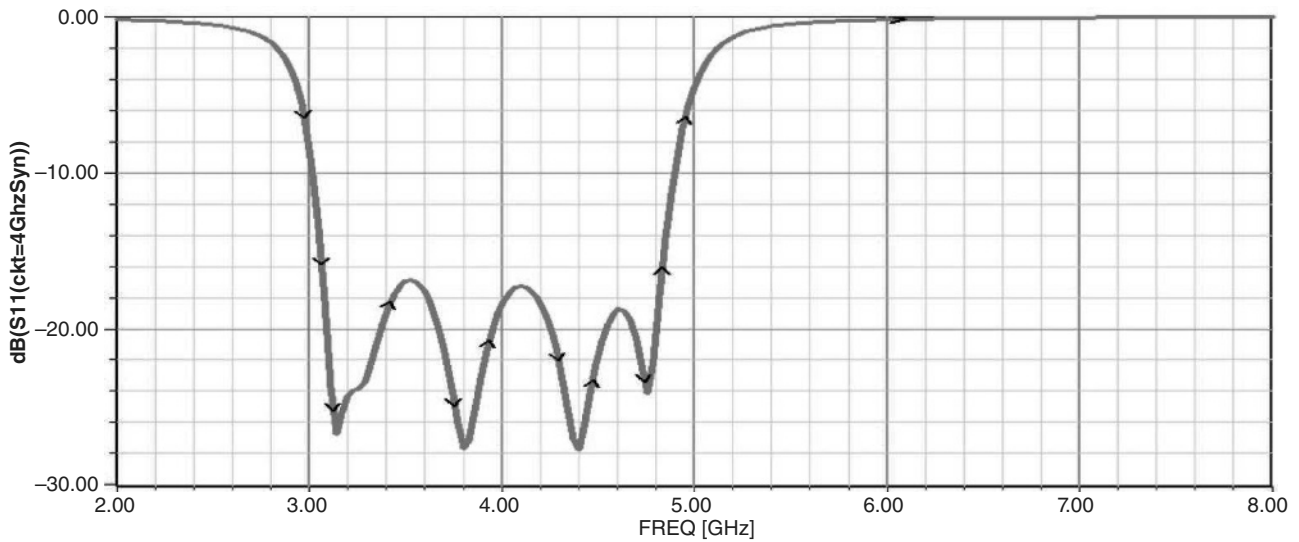


Figure 1.14 Return loss of the 4 GHz filter shown earlier.

$$Z_{in} = jZ_t \tan(\theta) \quad (1.17)$$

Define a new frequency variable $\lambda = \Sigma + j\Omega = \tan h(sT)$, where $s = \sigma + j\omega$ is the standard complex frequency variable and T is the time delay through the transmission line. Along the imaginary frequency axis, $\lambda = j\Omega = \tan(\theta)$. Now, we can express the impedance of the shorted stub in the new frequency variable as

$$Z_{in} = j\Omega Z_t \quad (1.18)$$

In this new frequency variable, the shorted stub takes on the appearance of an inductor. By a similar process, an open-circuited stub is a capacitor. The transformation used to establish λ is known as the Richards transformation, after its author [4]. The new complex frequency variable, λ , is often known as the Richards variable. Circuits designed to take advantage of this analogy consist of assemblages of shorted stubs (inductors), open stubs (capacitors), and cascaded transmission lines (unit elements), all having the same phase length. Formally, these are *commensurate transmission line networks* and form the basis of much transmission line theory.

We can use Richards transformations to replace the inductors and capacitors of a lumped-element filter design. In fact, for low-pass filter design, the frequency ωc is the filter's cutoff frequency. Using these stubs to replace inductors and capacitors will result in a filter response similar to that of the lumped-element design—a low-pass filter with a cutoff frequency ωc . However, the behavior of the filter in the stopband will be very different from the lumped-element design. For example, at (high) frequencies where the stub length becomes a multiple of $\lambda/2$, the filter response will be that of $\omega = 0$ —near-perfect transmission!

$$j\omega cL = jZ_0 \tan(\pi/4) = jZ_0, \text{ short-circuited stub as } \tan(\pi/4), \text{ and } -j/\omega cC = jZ_0 \cot(\pi/4) = -jZ_0; \text{ open-circuited stub.}$$

This is correct as for $ag \ll 1$ $\tan(ag) = ag$ and $\cot(ag) = 1/ag$ and the electrical length is one-eighth of the wavelength, meaning that the available variable is the line impedance [5, Chapter 6].

1.6 Kuroda's Identities

We find that Kuroda's identities can be very useful in making the implementation of Richards transformations more practicable. Kuroda's identities essentially provide a list of equivalent two-port networks. By equivalent, we mean that they have precisely the same scattering/impedance/admittance/transmission matrices. In other words, we can replace one two-port network with its equivalent in a circuit, and the behavior and characteristics (e.g., its scattering matrix) of the circuit will not change! We want to do this because one of the equivalents may be more practical to implement! For example, we can use Kuroda's identities to (1) physically separate transmission line stubs; (2) transform series stubs into shunt stubs; and (3) change impractical characteristic impedances into more realizable ones.

Kuroda's identities (aka Kuroda's transforms) are used to convert a section of transmission line with an open-parallel stub into an electrically equivalent section of transmission line with a shorted series stub. As a result, an identical S -parameter matrix is produced, which performs the same function. The technique is handy when designing distributed element circuits where one configuration is possible and the other is not. Filters are a good example, because in the physical layout, open parallel stubs are difficult (or impossible) to realize, whereas series shorted series stubs are not [5, pp 318].

Regarding transformation, a simple modeling may be used to find parameter values for an equivalent network. This example demonstrates the technique by optimizing a transmission line to perform like a lumped inductor. The example uses a transmission line with 1 pF in parallel to behave like a parallel LC circuit at a specific frequency (Figure 1.15a–c).

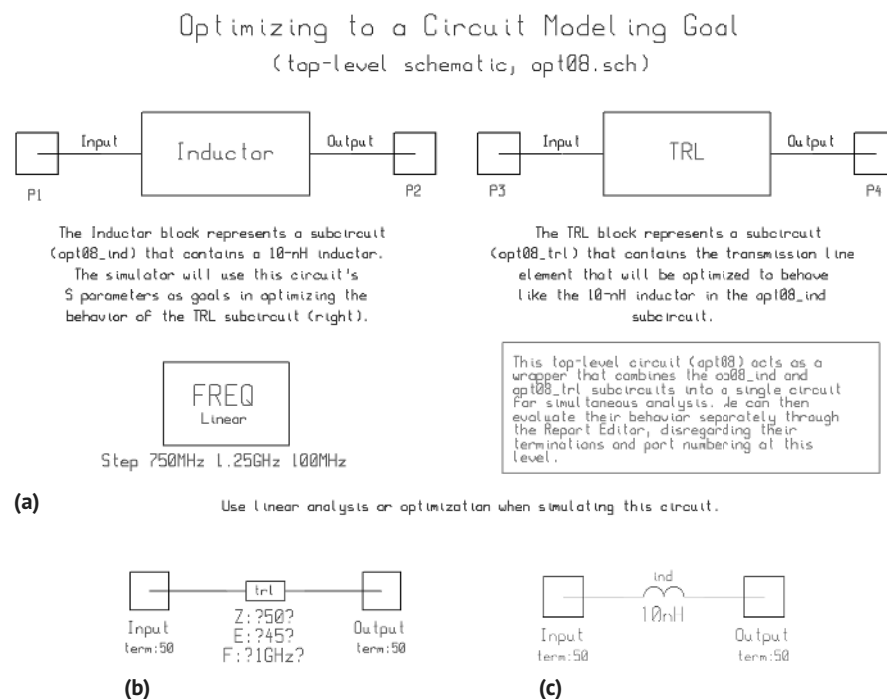


Figure 1.15 (a) Optimizing a transmission line to be used as an inductor. (b) Optimizable transmission line. (c) Reference inductor.

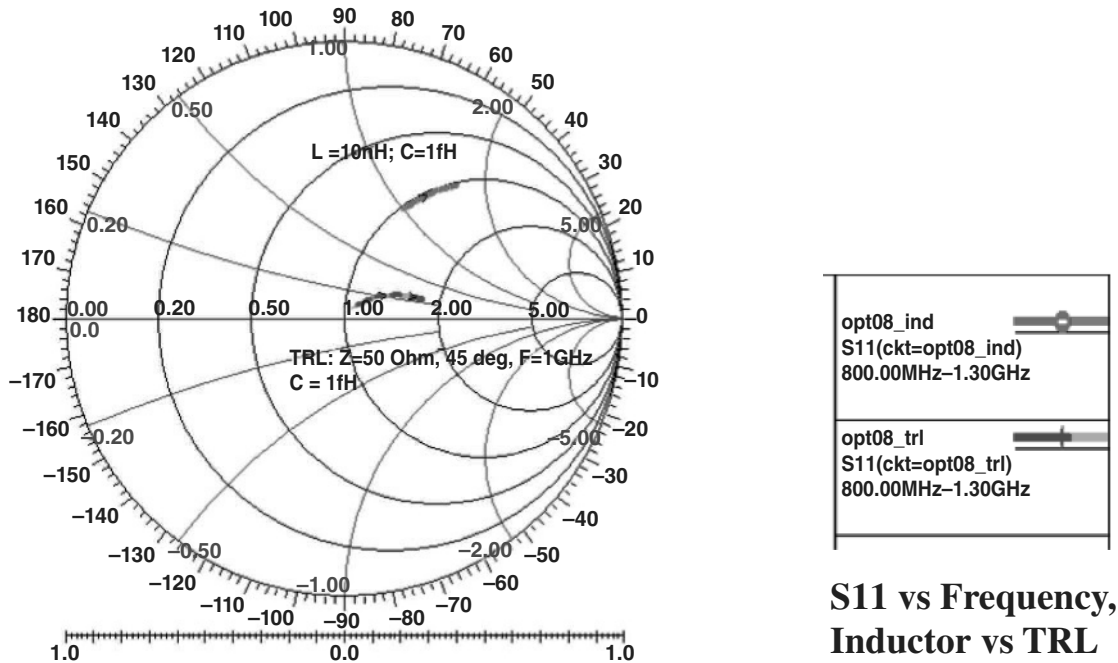


Figure 1.16 Result of optimization of the 750–1250 MHz.

The optimizer will adjust the transmission line to behave like a 10 nH inductor from 750 MHz to 1.25 GHz. This can be seen in the plot of the Smith diagram in Figure 1.16, showing S11 versus frequency.

1.7 10.2 GHz Test Circuit

The example in Figure 1.17a is the synthesis of a lumped-element filter for 10.2 GHz bandpass filter (ham RF). The synthesis program provides results for L and C's, which are not practical.

Figure 1.17b shows S21 and S11 in dB as a function of frequency. Insertion loss is about 5 dB.

Figure 1.18a shows a bandpass filter using transmission lines and the cross element.

Frequency response of the filter using lossless elements See Figure 1.18b shows a slight dip; the center frequency of the filter is about 16 GHz. By introducing losses, a more realistic frequency response can be obtained (see Figure 1.18c).

This type of transformation introduces a notch, which is at about 13.4 GHz.

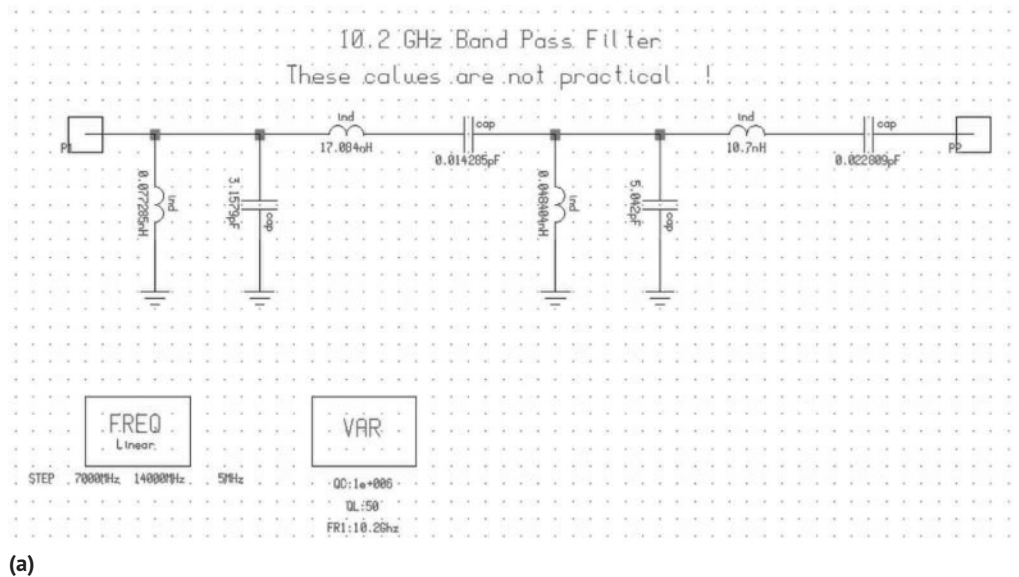
Now let us apply electromagnetic (EM) coupling. The following is a Richards transformation of a double EM-coupled tuned circuit. Figure 1.19a shows the circuit arrangement. It is actually quite a complex filter, with typical distributed elements such as crosses and coupled lines, including the end effects of the transmission line (radiation).

Figure 1.19b shows the frequency response of the bandpass filter at 16.3 GHz, having an insertion loss of about 2.61 dB.

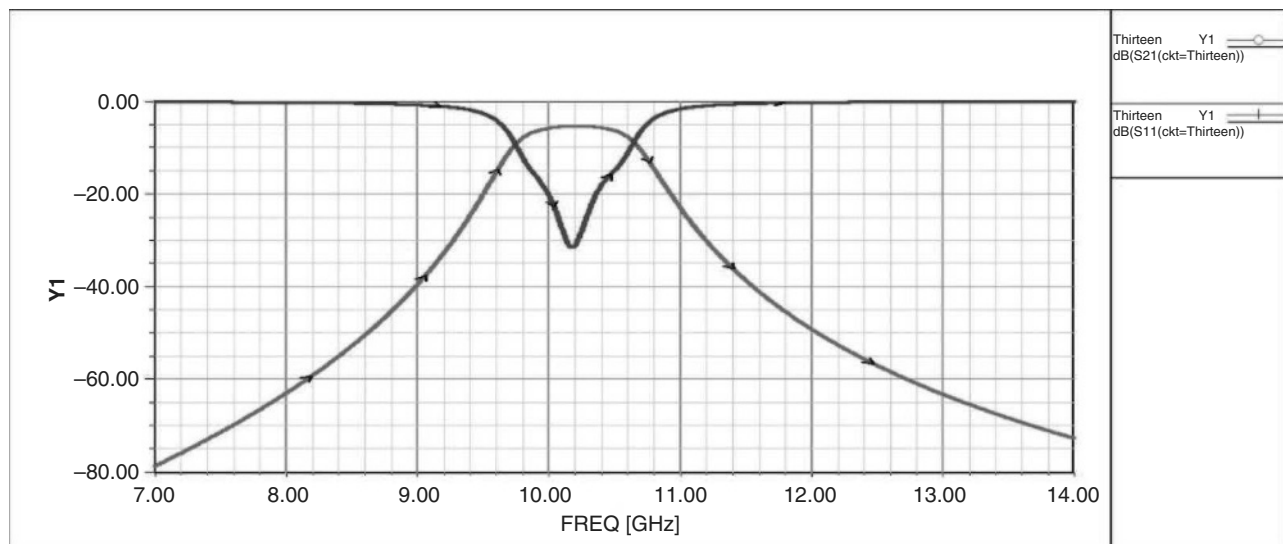
1.7.1 Example: Broadside-coupled Filter at 10.2 GHz (Ham RF)

The following circuit (Figure 1.20a) is a perfect symmetrical filter at 10.4 GHz and is the result of a filter synthesis using five coupled elements. It is designed on a material with $\epsilon = 10$. Figure 1.20b shows the S11 and S21 of the synthesized filter at 10.4 GHz.

The next example, Figure 1.21a shows a four-stage advanced filter with EM coupling, a result of a synthesis program, while the results of a synthesis program shows the S11 and S21 in dB of the 10.2 GHz synthesized filter in Figure 1.21b.



(a)



(b)

Figure 1.17 (a) 10.2 GHz bandpass filter (values not practical). (b) S21 and S11 in dB as a function of frequency.

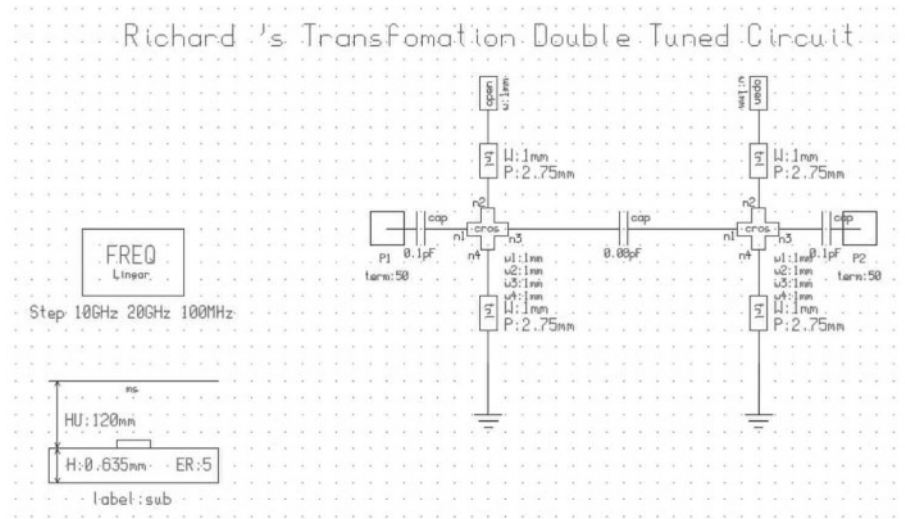
1.7.2 Example: Two-stage Matching with a Two-section Series Transformers

Another interesting problem that can be solved using transmission lines is impedance matching. Let us assume that we have to match a $25\ \Omega$ source to a $75\ \Omega$ load. A typical answer would be to use a transmission line of the impedance $\sqrt{75 \times 25} = 43\ \Omega$. Now let us try the optimizer; but using two transmission lines complicates the situation. The desired center frequency is 10 GHz and we will try to get the maximum bandwidth possible while maintaining a return loss of better than $-30\ \text{dB}$.

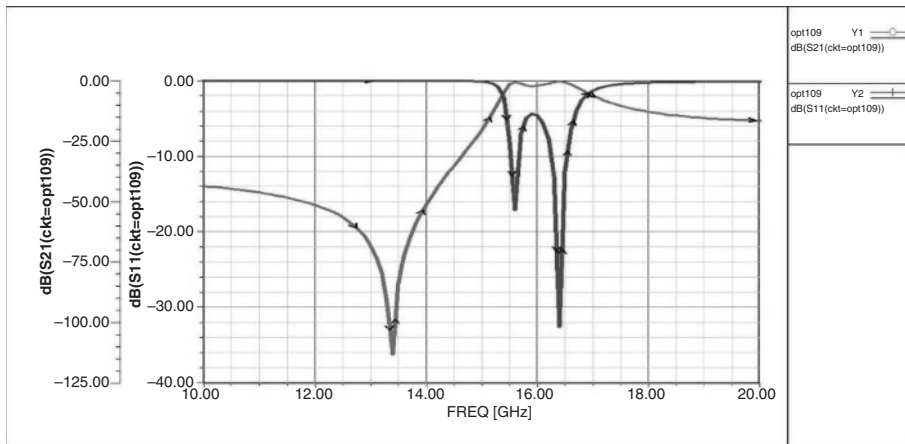
After optimization, in the schematic shown in Figure 1.22a, the characteristic impedance of TRL1 needs to be $\approx 33.5\ \Omega$ and the second TRL2 $\approx 56.3\ \Omega$. This is a very elegant method to generate a transmission line-based wideband matching. The specified frequency at the transmission line needs to be the highest; in this case, it is 12 GHz. The worst case S21 in dB (no losses) is $-0.04\ \text{dB}$ (see Figure 1.22b).

In the previous simulations of filters, we used ideal transmission lines. For the transition to real transmission lines, meaning losses from skin effect and radiation, and loss tangent of the material, we now use a transmission line calculator/synthesizer.

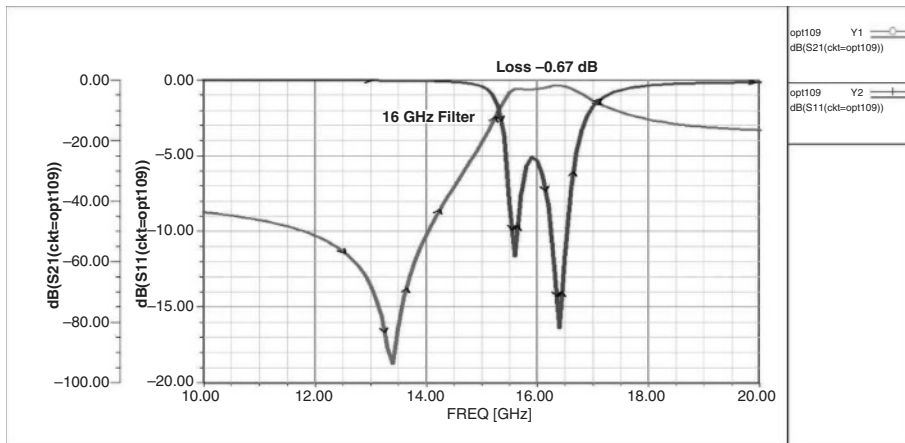
The following calculations, Figure 1.23, are based on Duroid-based material.



(a)



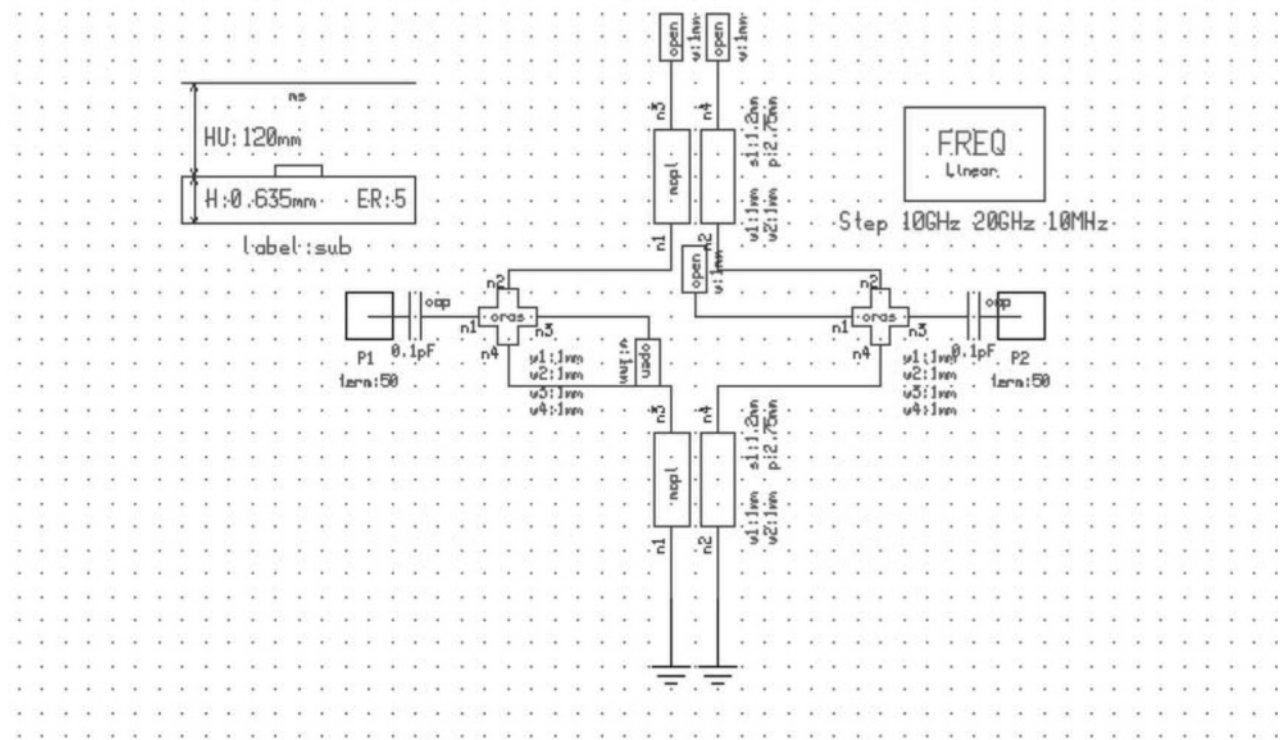
(b)



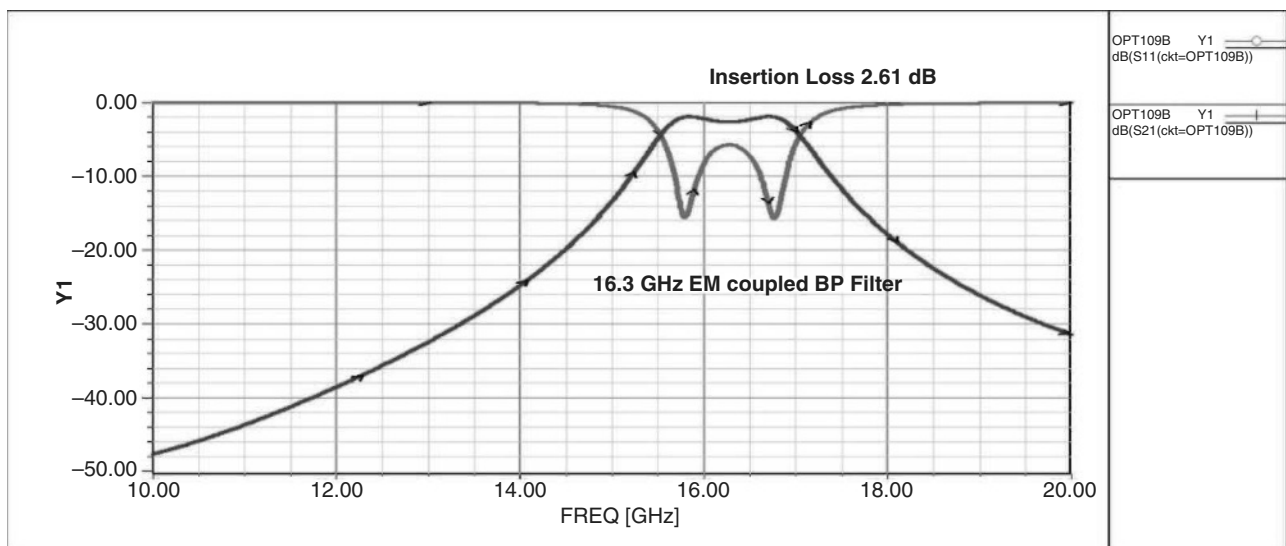
(c)

Figure 1.18 (a) The Richards transformation double-tuned circuit schematic of the filter using the ideal case. (b) Frequency response of the filter using lossless elements. (c) Frequency response of the filter using lossless elements—with loss.

Richard's Transformation Double EM-Coupled Tuned Circuit

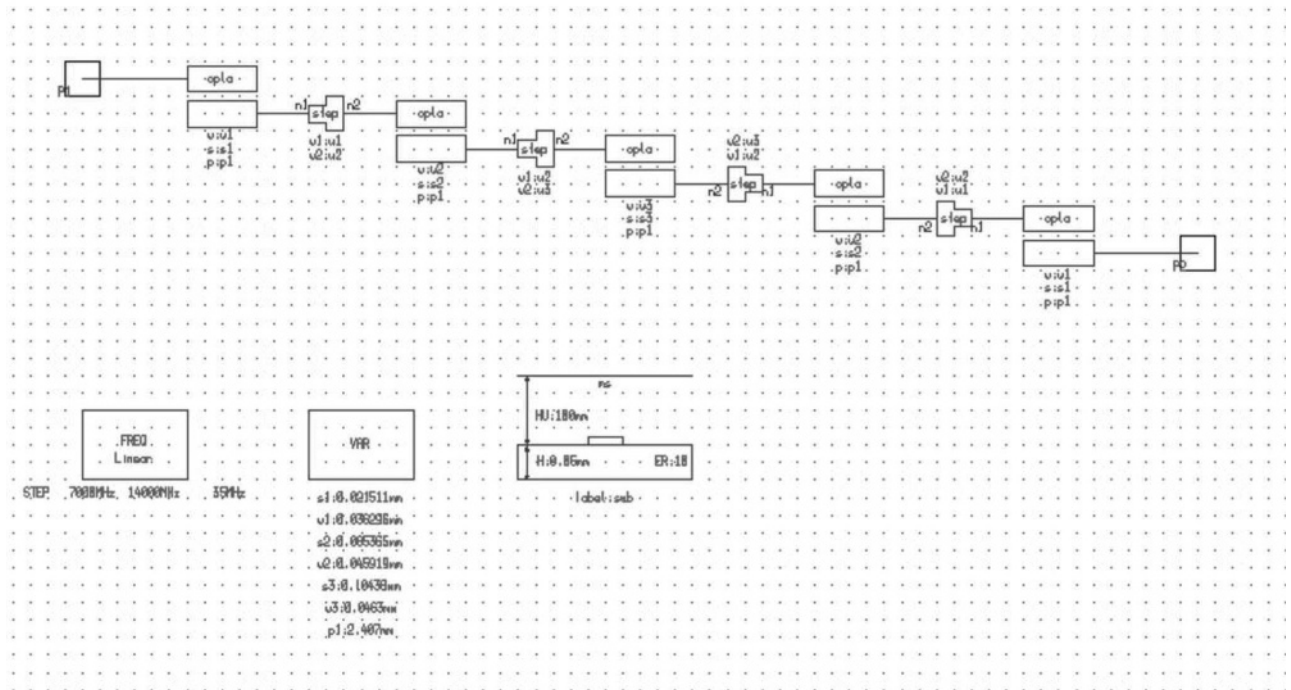


(a)

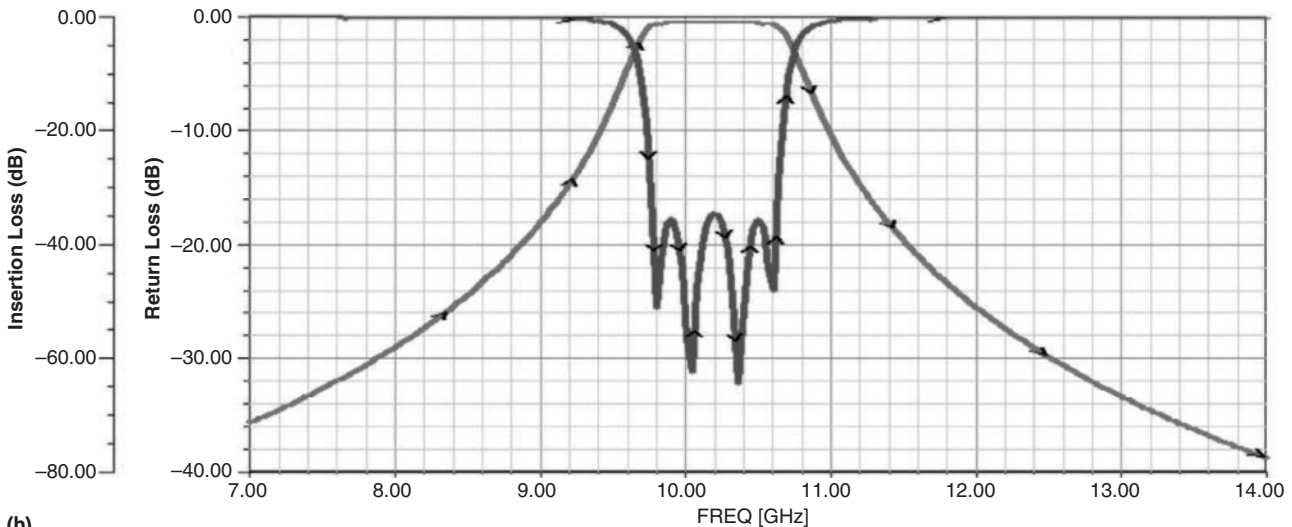


(b)

Figure 1.19 (a) Richards transformation double EM-coupled tuned circuit. (b) Frequency response of the bandpass filter at 16.3 GHz, having an insertion loss of about 2.61 dB.



(a)



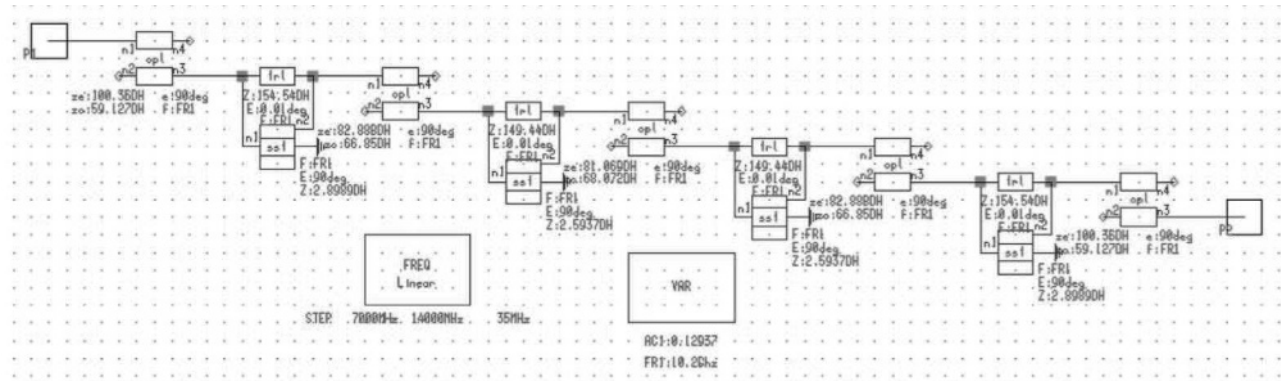
(b)

Figure 1.20 (a) A perfect symmetrical filter at 10.4 GHz. (b) S11 and S21 of the synthesized filter at 10.4 GHz.

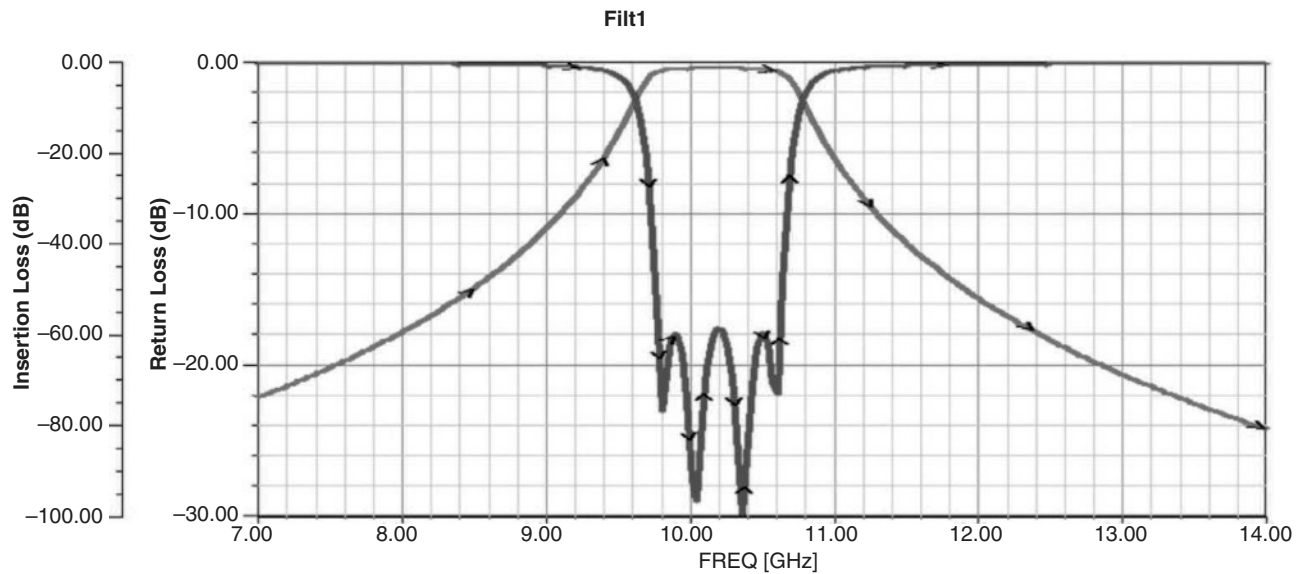
1.8 Validation

As we are going to use this synthesis program in the range from 10 GHz to 1 THz, we will build test structures and compare measurements with simulations. For critical applications at 1 THz, we decided to use high frequency structure simulation software (HFSS), an Ansys product, which is considered as the gold standard, in cases where very accurate predictions are necessary.

The measurements were done with the R&S ZVA network analyzer, as shown in Figure 1.24. The basic model covers the frequency range from 10 MHz to 67 GHz. With external sources, the frequency range can be currently extended up to 200 GHz (approximately).



(a)



(b)

Figure 1.21 (a) Four-stage advanced filter with EM coupling, a result of a synthesis program. (b) S11 and S21 in dB of the 10.2 GHz synthesized filter.

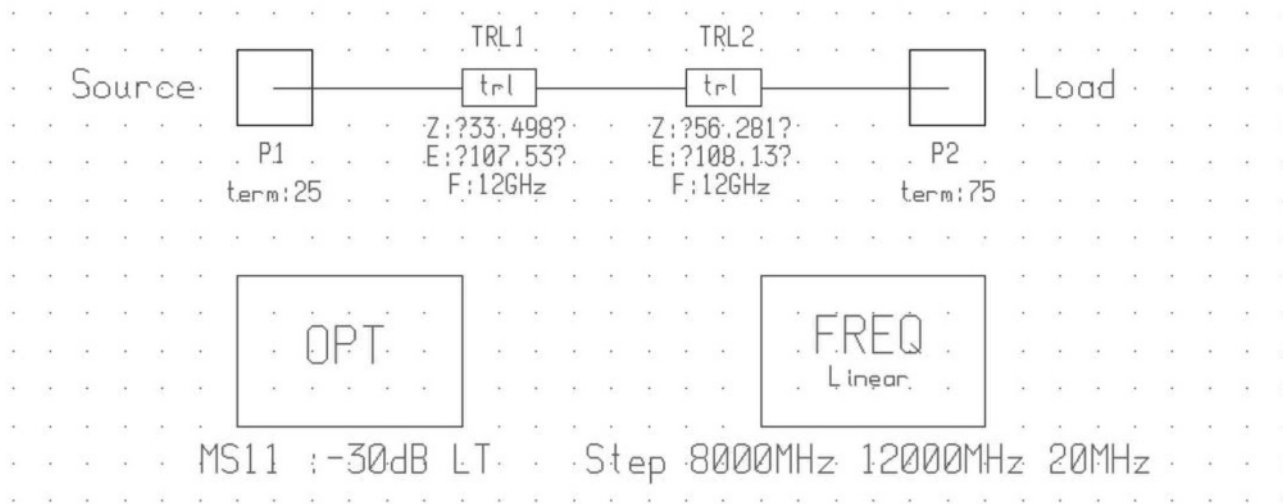
The vector network analyzer shown above in Figure 1.24b, model ZNA, is the successor of the model seen in Figure 1.24a.

Next, we will validate some structures, beginning with a stub to be resonant at 10 GHz. Figure 1.25a shows the layout of the actual structure. The following stub uses dimensions as calculated by the synthesis program, to be resonant at 10 GHz (Figure 1.25b).

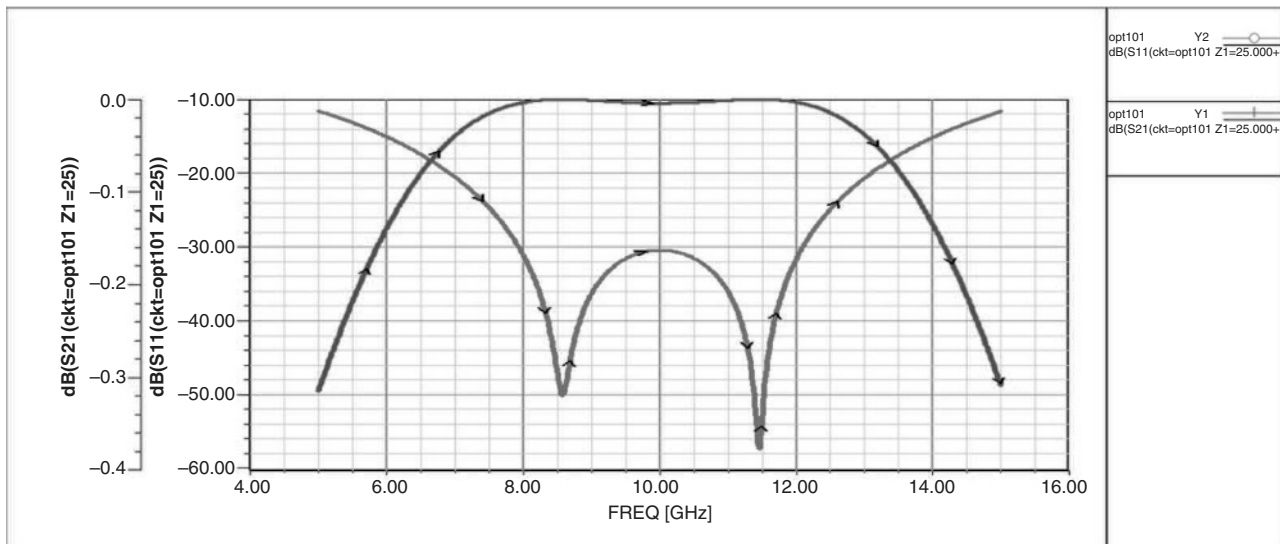
It turns out that the resonant frequency is 9.64 GHz; because the transmission line calculator does not add the end effect (for transmission line, it needs to be 220 mil), the calculated frequency is 1.38% too high.

Figure 1.26a is a simulation in the Serenade program.

The simulator itself calculates the resonant frequency at 9.77 GHz (calculation error is about 1%), but the synthesis is 1.38%. In Figure 1.26b, the left curve was adjusted to reflect the measured value, by adjusting the physical length of the stub. The right curve shows the calculated resonant frequency f or the actual physical length. What did we learn from this? The EM simulation in Serenade and the synthesis program are accurate within 1.4%, while the difference between the measurement and the simulation is 1%.



(a)



(b)

Figure 1.22 (a) Two-stage matching with a two-section series transformers. (b) S21 (gain) in dB and S11 return loss (worst case is -30 dB) of the filter circuit of Figure 1.22a.

1.8.1 100 GHz Test Circuit

The next test case, in Figure 1.27a is 100 GHz. To obtain the physical dimensions, the same Serenade transmission line synthesis program was used to find the necessary values. Again we use the synthesis program to give us the dimensions of the structure, length = 389 μm , width = 34 μm , and height = 20 μm . The simulation in Serenade is shown in Figure 1.27b, while ADS simulation is shown in Figure 1.27c.

The error of the synthesis program relative to the ADS simulation is about 1%. The same test in Serenade practically shows no difference in the resonance frequency, between the synthesized and simulated frequencies.

Now is the final test for 1 THz. The synthesis program provides the following dimensions: length = 39.5 μm , width = 8.4 μm , and height = 5 μm . We have used the Serenade program and ADS to perform the same analysis.

Based on the analysis of Figure 1.28a using the two programs, the error is 1.02 for Serenade and 1.05% for ADS, as shown in Figure 1.28b,c.



Figure 1.23 A picture of the transmission line synthesis user interface.

This exercise was necessary to calibrate the simulator. Both simulators give a high-resolution solution that is better than the actual result obtainable in production.

1.9 More EM Simulation

Having validated a transmission line accuracy, we will show some filter examples.

Figure 1.29a is a four-stage filter with real values using EM coupling rather than a curve-fitting model.

The simulation of the synthesized filter, in Figure 1.29b, shows that the synthesis program is well-behaved and accurate. After this example, let us take a look at more filters.

1.9.1 Edge-coupled Filter Using Models and Electromagnetic Results

Serenade has a built-in spectral Domain simulator, quasi, 2.5-dimensional simulators, which can handle up to 12 different parallel transmission lines of unequal spacing and widths. Figure 1.30a shows the schematic of a proposed 110 GHz filter with different TL widths, while Figure 1.30b shows the layout (not to scale) of the circuit simulated.

The values for S11 (–12 dB worst case) are not too impressive (see Figure 1.30c).

ADS simulation is shown next, in Figure 1.30d.

The simulation result in Figure 1.30e shows a slight frequency shift of about 2.5 GHz or 2.2%.

The previous simulation results were of the simulation of the circuit schematic built with the substrate, and the plots in Figure 1.30f are the simulation results of the layout.

The filter shown in Figure 1.31a uses ideal elements and a coupled filter model. While this filter cannot be built like this, a simulation was done to validate the mathematical stability of the simulator.

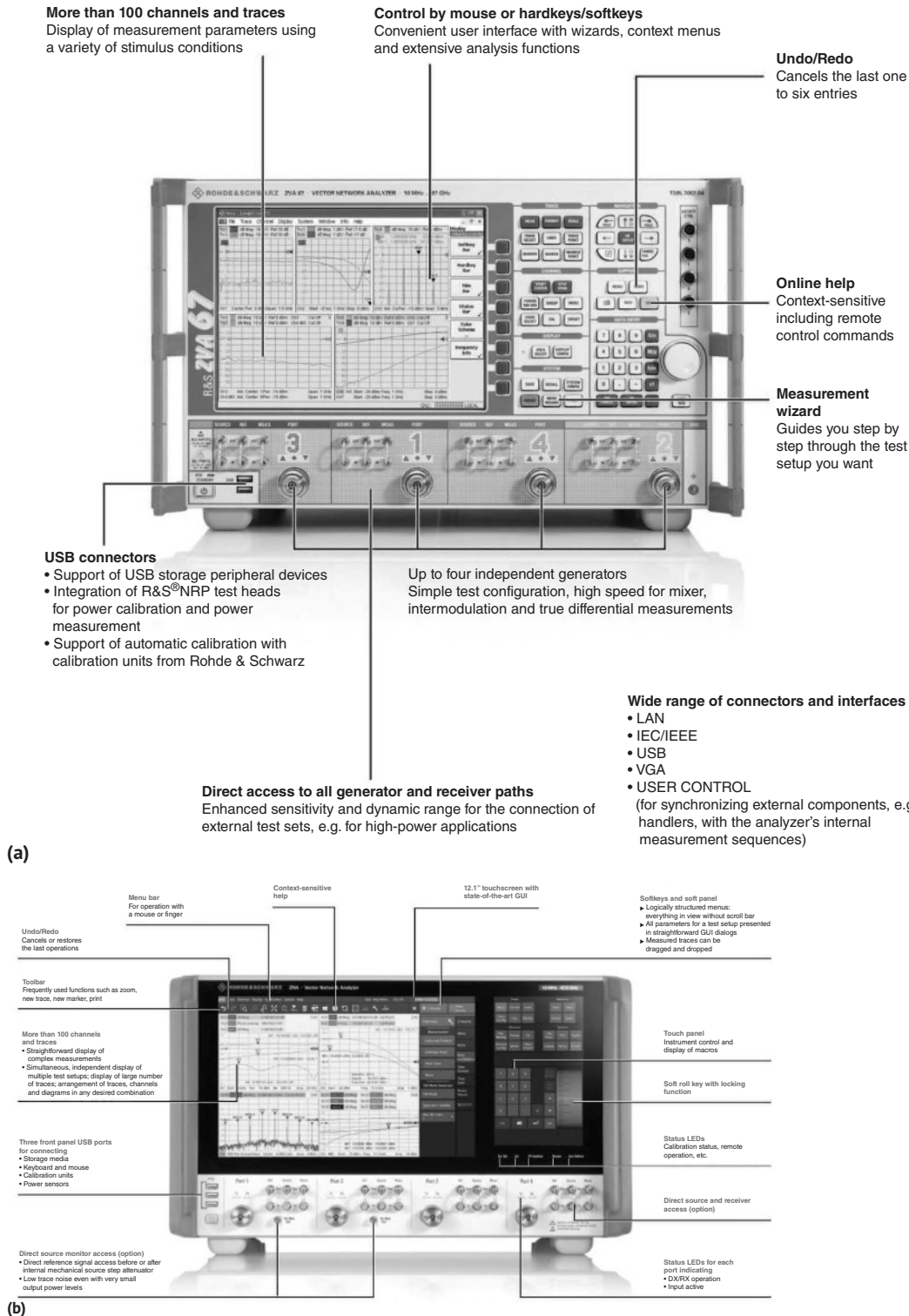
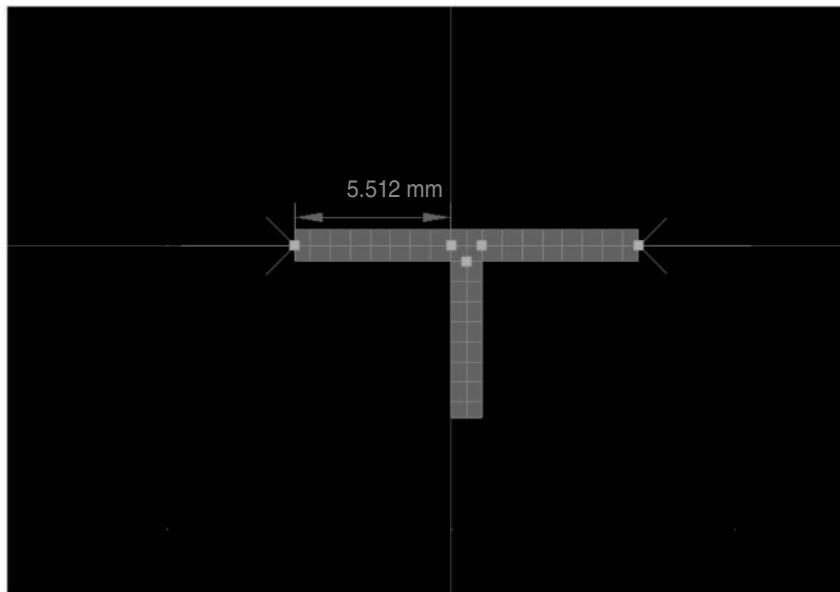
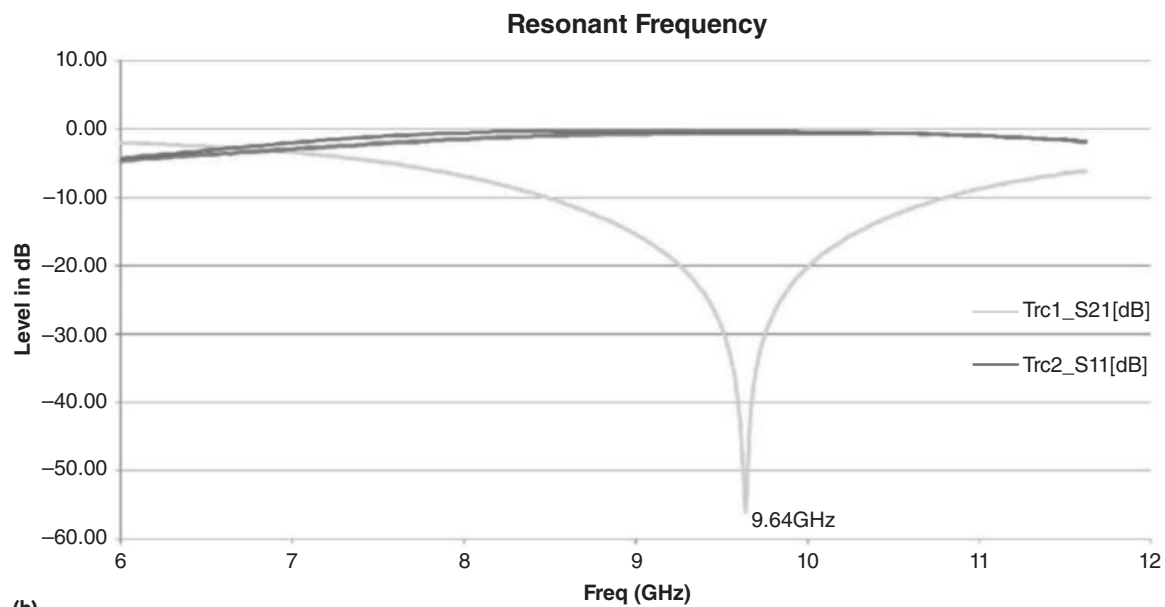


Figure 1.24 (a) R&S ZVA67 vector network analyzer, 10 MHz–67 GHz. *Source:* Courtesy of Dr. Ulrich L. Rohde. (b) R&S ZVA vector network analyzer, 10 MHz–41.5 GHz. *Source:* Courtesy of Dr. Ulrich L. Rohde.



(a) Stub test@10GHz

Actual Stub_measurement_data l=217mil w=43.5 mil



(b)

Figure 1.25 (a) T-stub resonant at 10 GHz. (b) The resonant frequency at 9.64 GHz.

The results of the simulation in Figure 1.31b shows the worst-case return loss of about -17 dB, and the ripple is less than 1 dB.

1.10 Transmission Line Test

Figure 1.32a shows a simulation up to 2 THz.

Figure 1.32b shows the return loss (S11), the curve indicates about -22 dB at 1 THz, with the loss being about 0.15 dB.



Figure 1.26 (a) 10 GHz test circuit (TRL). (b) The resonant frequency: measured and calculated for the 10 GHz stub.

Figure 1.32c is an HFSS-based simulation up to 2.2 THz.

This simulation predicts insertion loss of about 0.4 dB at 1 THz and the higher-order modes are much more complicated.

1.11 Branch Line and the Rat-race Coupler

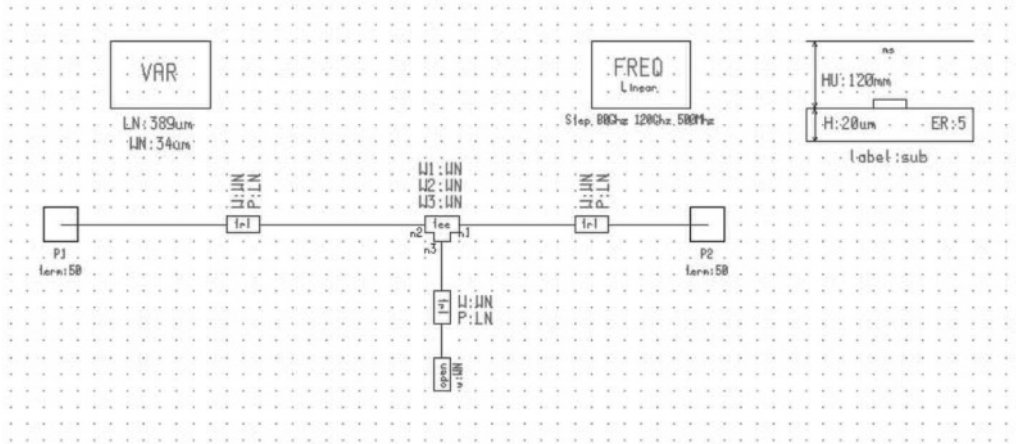
An example of a branch line coupler is shown in Figure 1.33a.

Figure 1.33b shows the asymmetrical behavior based on the limitation of the models.

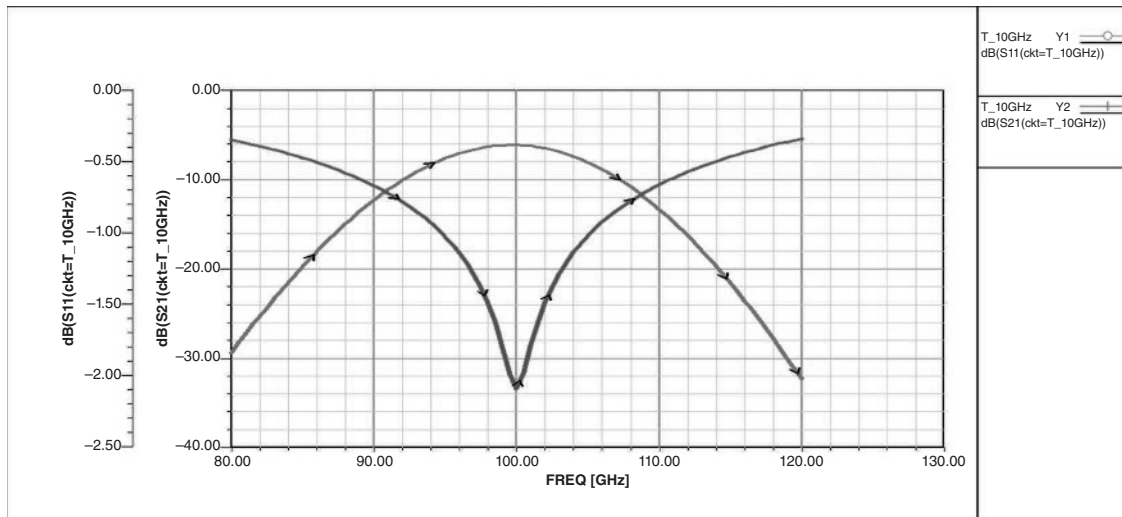
Figure 1.33c shows the S11 and S21 in dB of the branch line coupler simulation, which is symmetrical but slightly off in the resonant frequency.

A variation of this is the rat-race coupler, which can also be modeled as a coupler with four bends, as shown in Figure 1.34a, and its S21, S31, and S41 in dB of the variation are shown in Figure 1.34b.

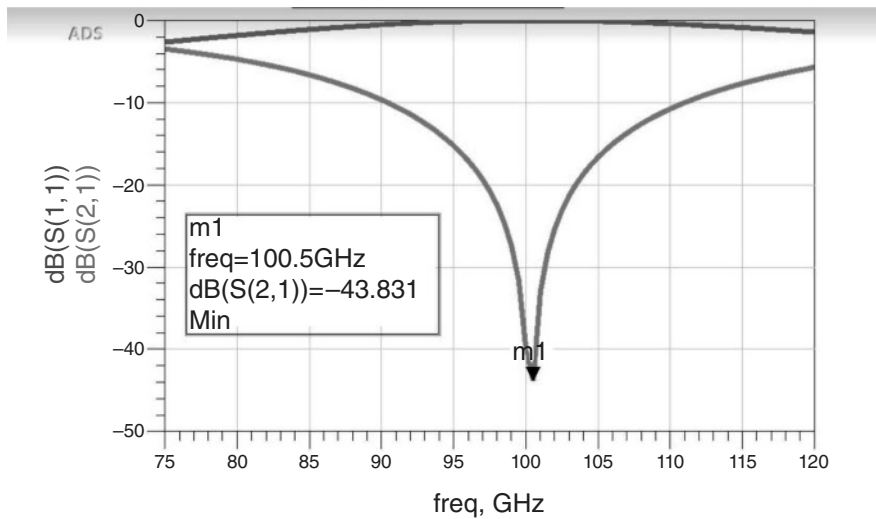
And, finally, Figure 1.35a shows the well-published rat-race coupler.



(a)



(b)



(c)

Figure 1.27 (a) Simulation of the 100 GHz test circuit. (b) S_{21} and S_{11} of the 100 GHz test circuit. (c) Simulation of the resonant frequency in ADS.

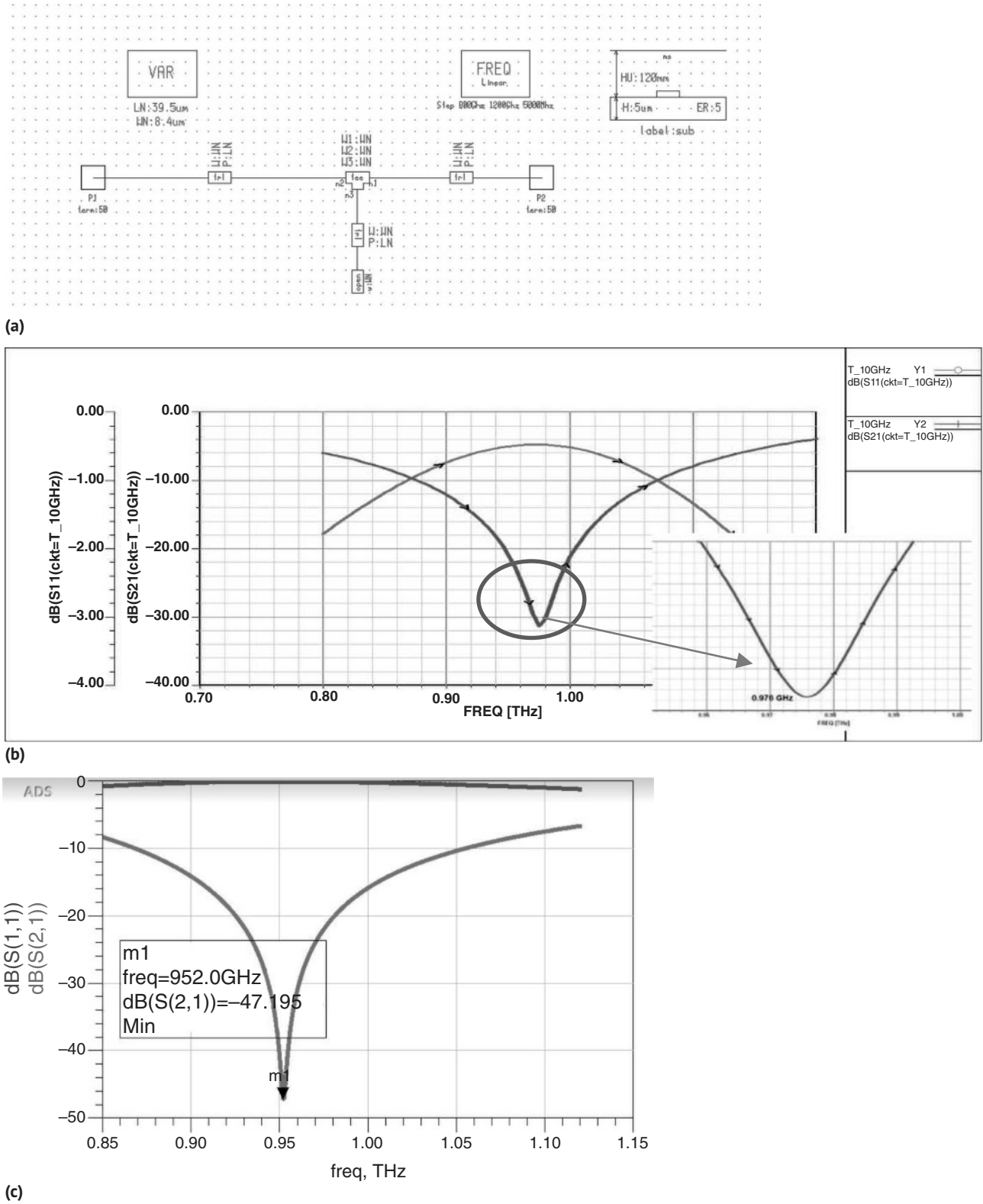
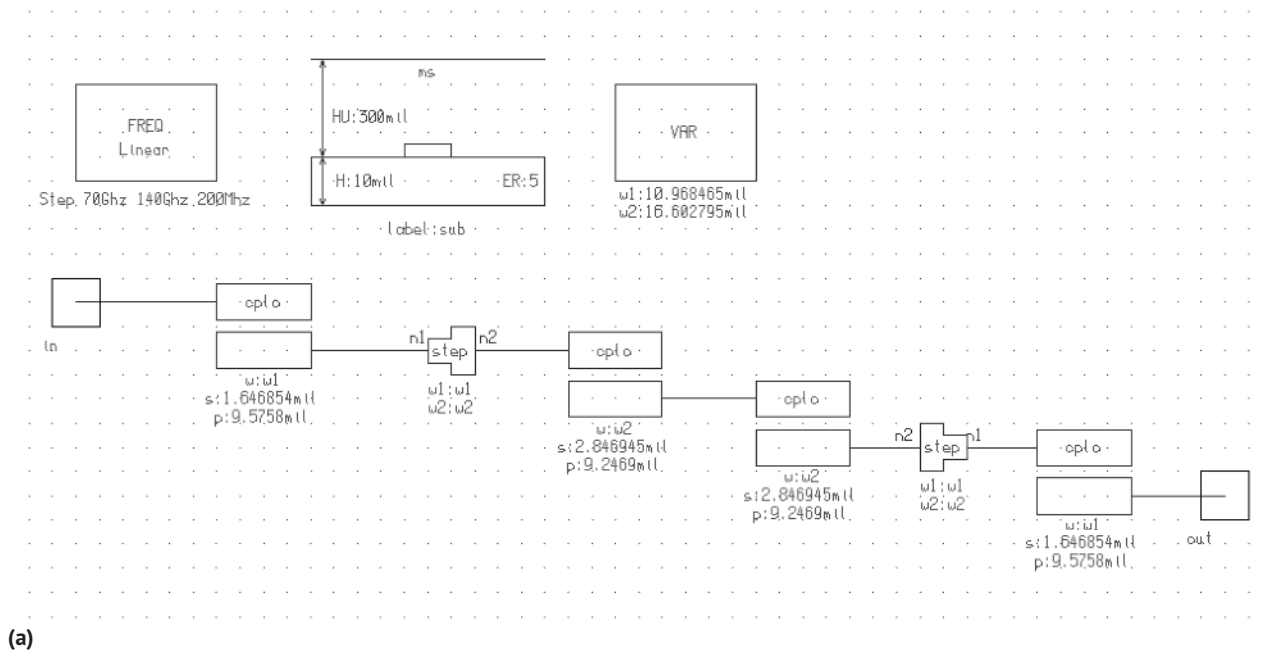
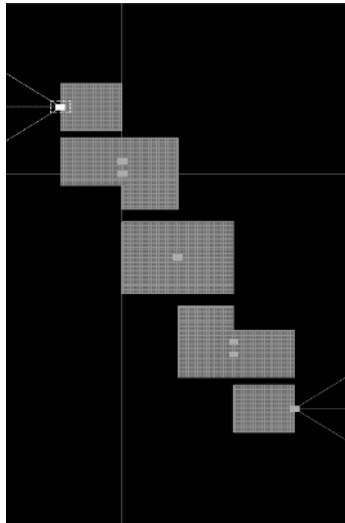


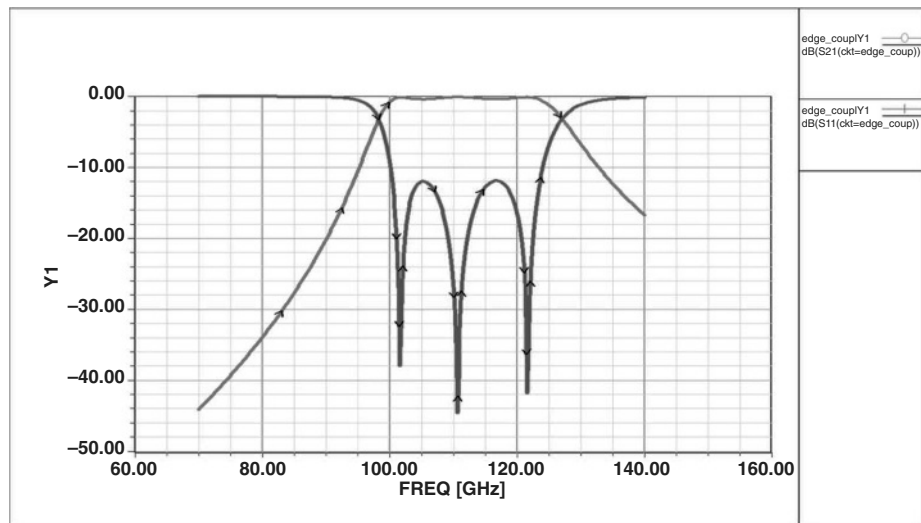
Figure 1.28 (a) Schematic of the test structure for 1 THz. (b) S21 and S11 in dB of the 1 THz resonator test structure, a closer look at the S21 of the test structure for 1 THz. (c) Simulation of the resonant frequency using ADS shows 952 GHz.



(a)



(b)



(c)

Figure 1.30 (a) Schematic of a proposed 110 GHz filter with different TL widths. (b) Layout (not to scale) of the circuit simulated in (a) (ADS Layout tool). (c) Simulation results showing S11 and S21 in dB of the circuit simulated in (a).

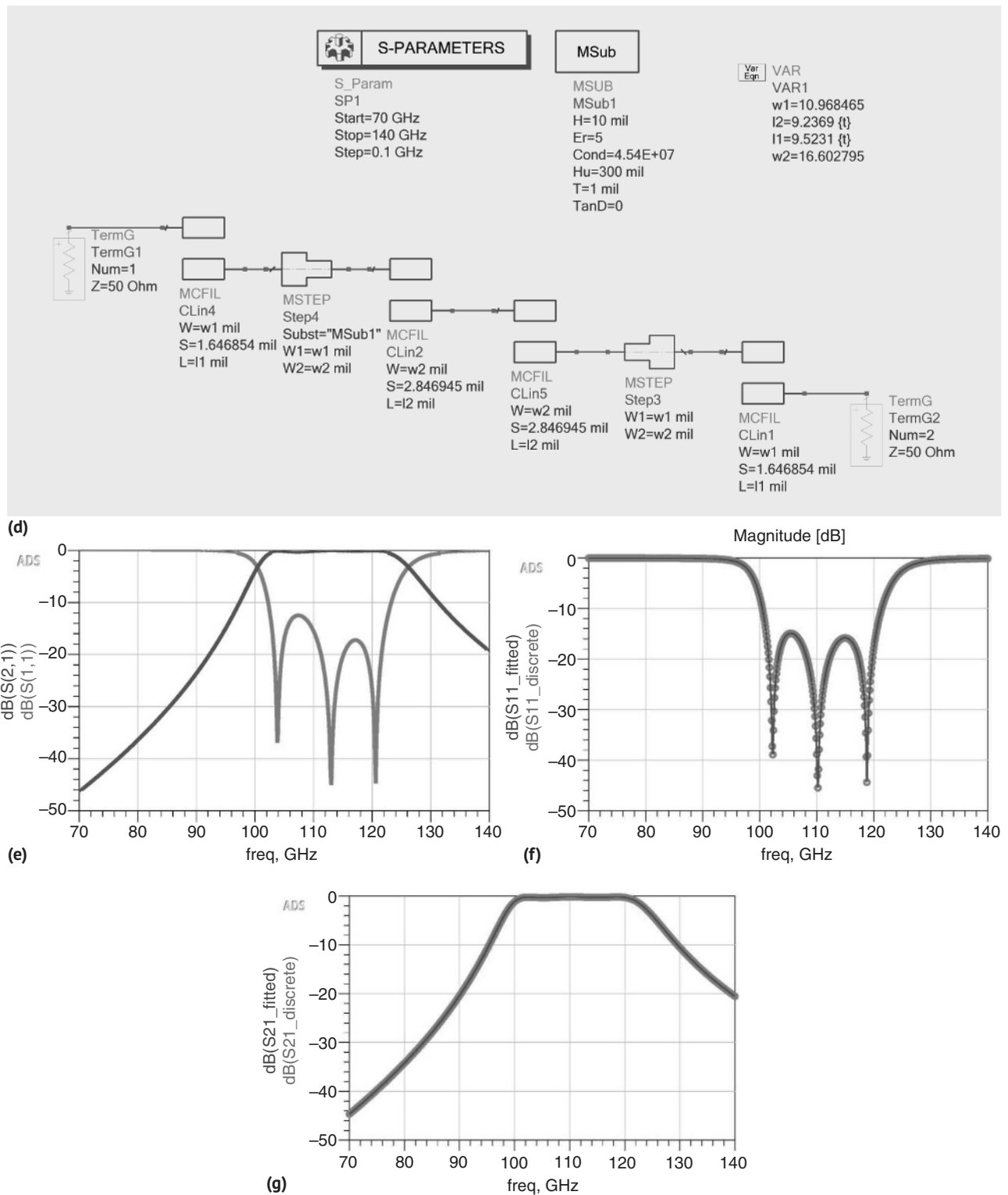
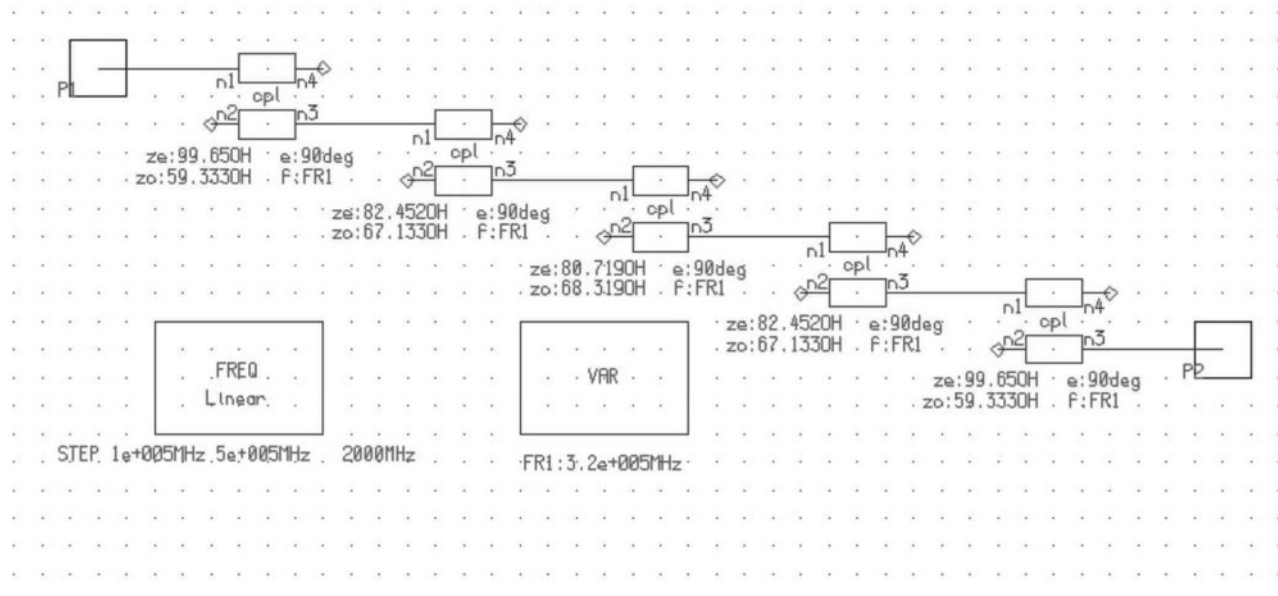
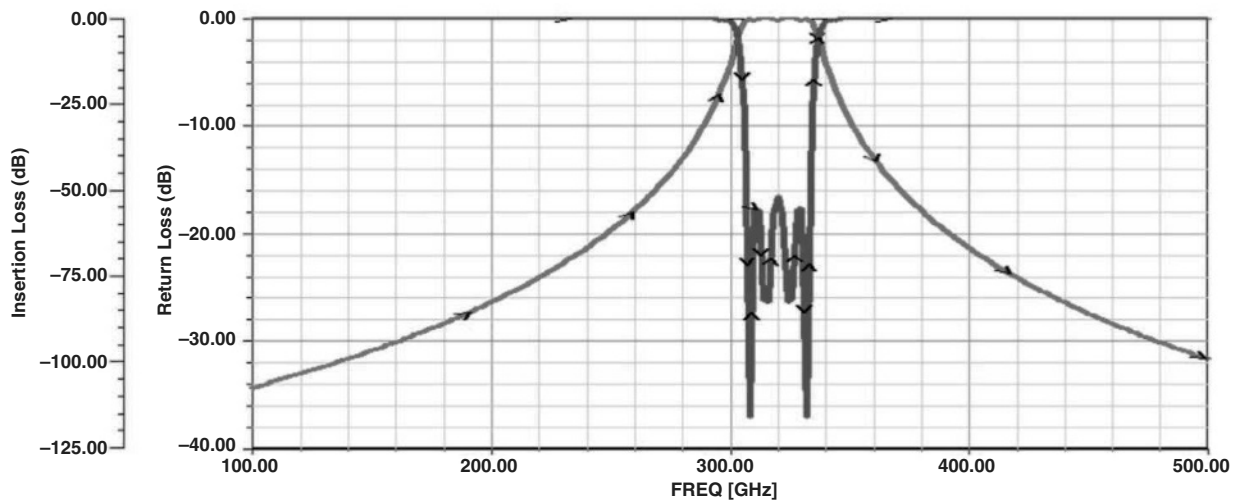


Figure 1.30 (Cont'd) (d) ADS simulation of the four-stage filter at 110 GHz. (e) S21 and S11 in dB of the 110 GHz filter. (f) Return loss (S11) of the circuit simulated in the layout (a). (g) Gain (S21) of the circuit simulated in the layout (a). *Note:* In Figure 1.30g, the thin and the thick traces indicate two different resolution settings used in the simulator: synthesis of a filter designed with a center frequency of 320 GHz and a bandwidth of about 40 GHz.

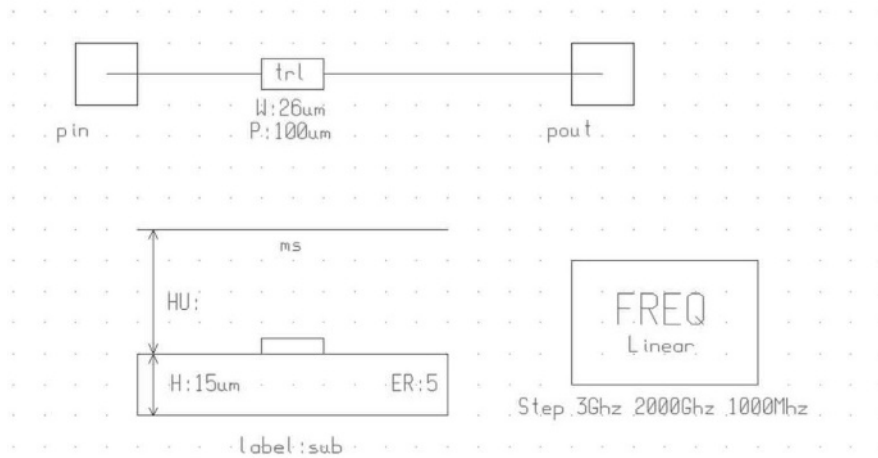


(a)

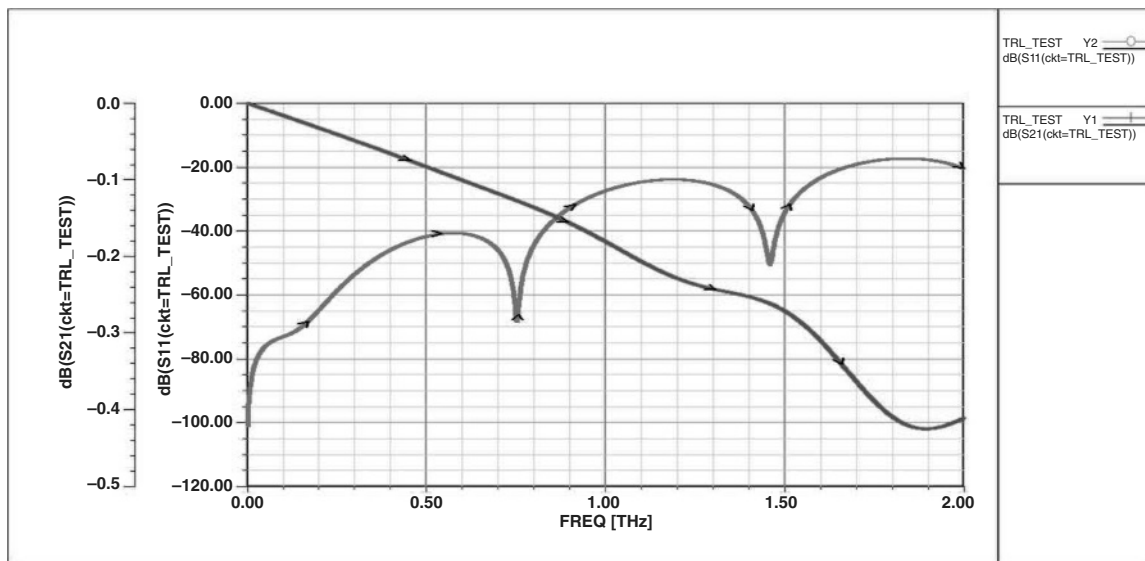


(b)

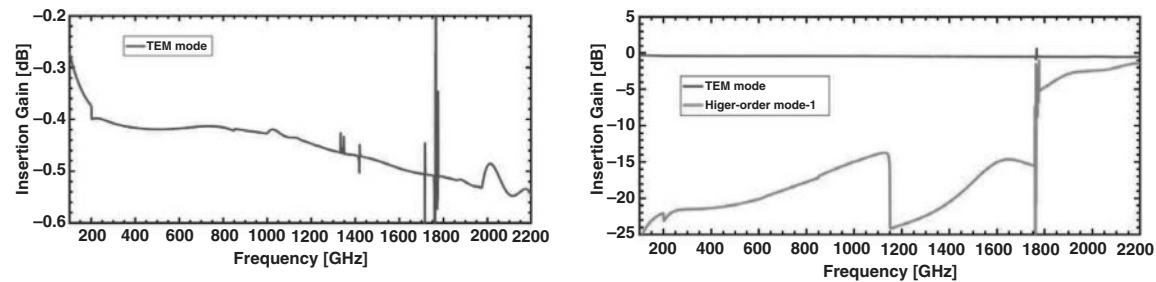
Figure 1.31 (a) Schematic of the 320 GHz filter with four poles. (b) Simulated result showing S11 and S21 in dB of the 320 GHz filter with four poles.



(a)



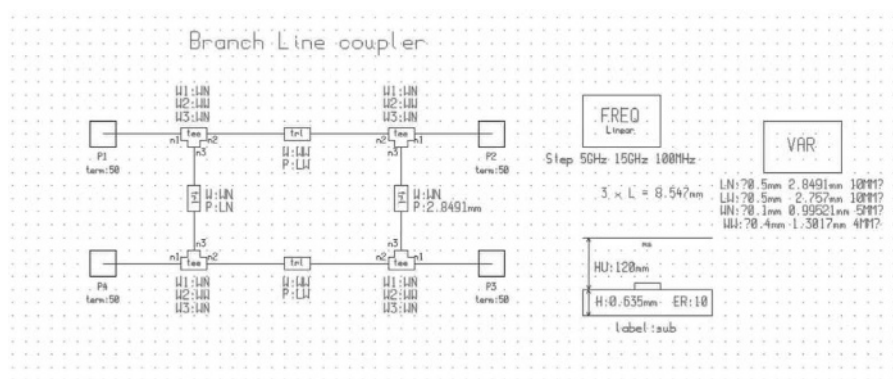
(b)



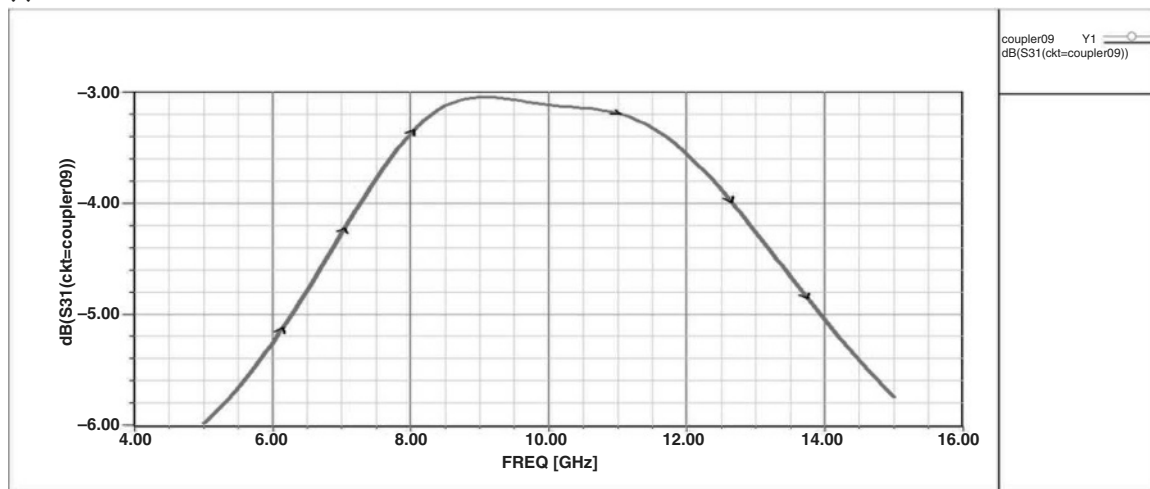
(c) ☉ Fundamental mode: quasi-TEM

☉ quasi-TEM and higher-order modes

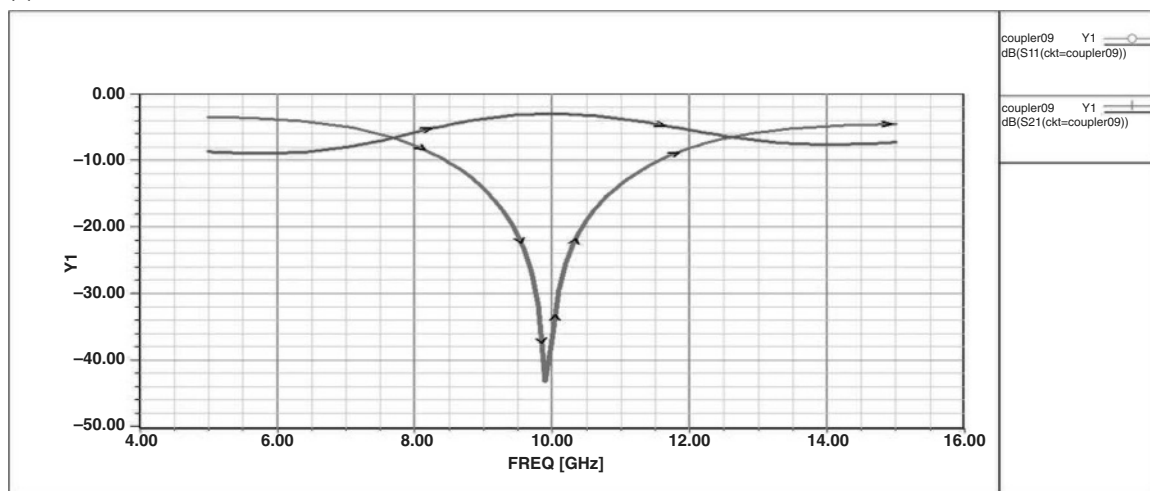
Figure 1.32 (a) Simulation of transmission line behavior up to 2 THz. (b) Simulation results of a transmission line behavior up to 2 THz. (c) HFSS-simulated forward transmission ($W = 25.15 \mu\text{m}$, $h = 15 \mu\text{m}$, $\epsilon_r = 5$, $t = 0$).



(a)

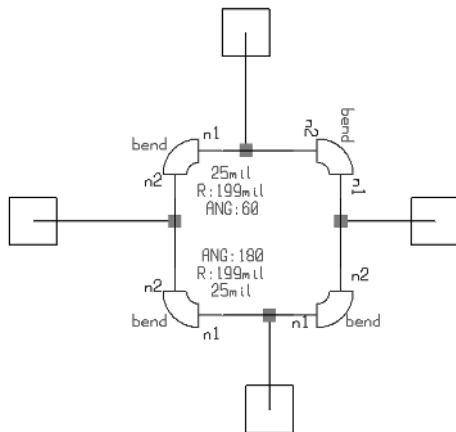


(b)

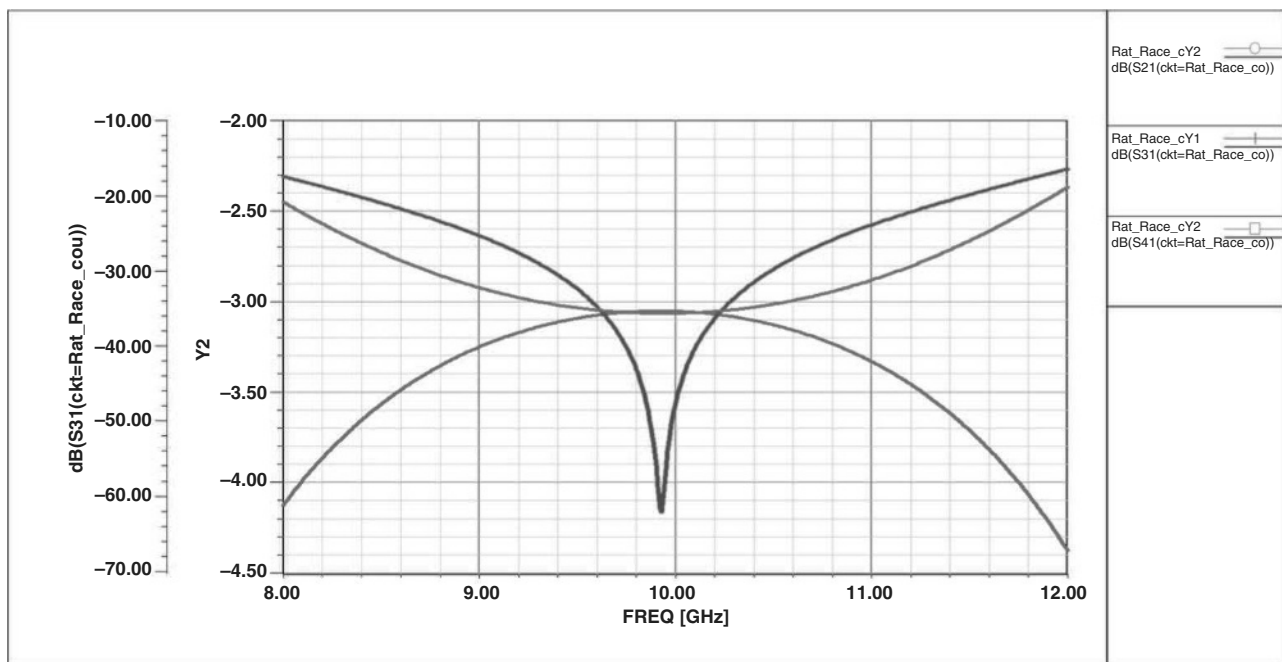


(c)

Figure 1.33 (a) Example of a branch line coupler simulation. (b) S31 in dB of the branch line coupler simulation. (c) S11 and S21 in dB of the branch line coupler simulation.

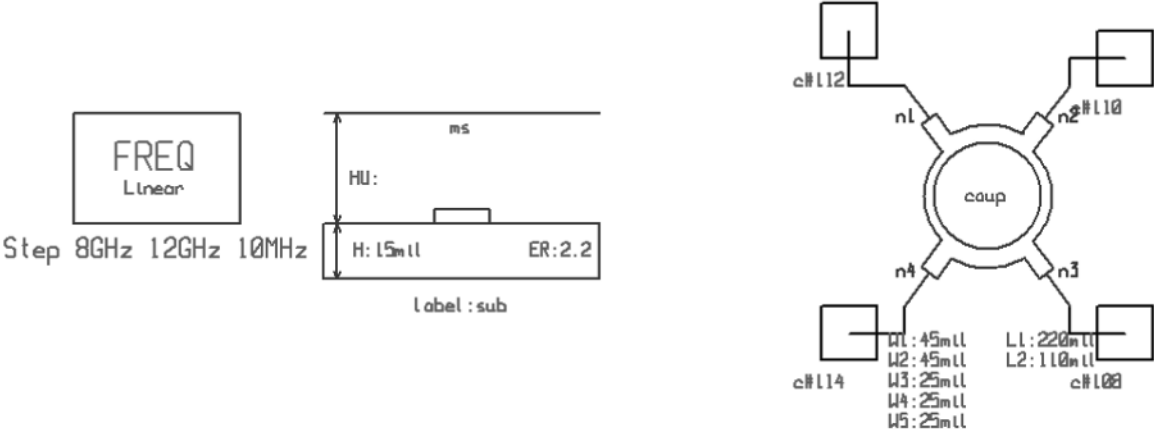


(a)

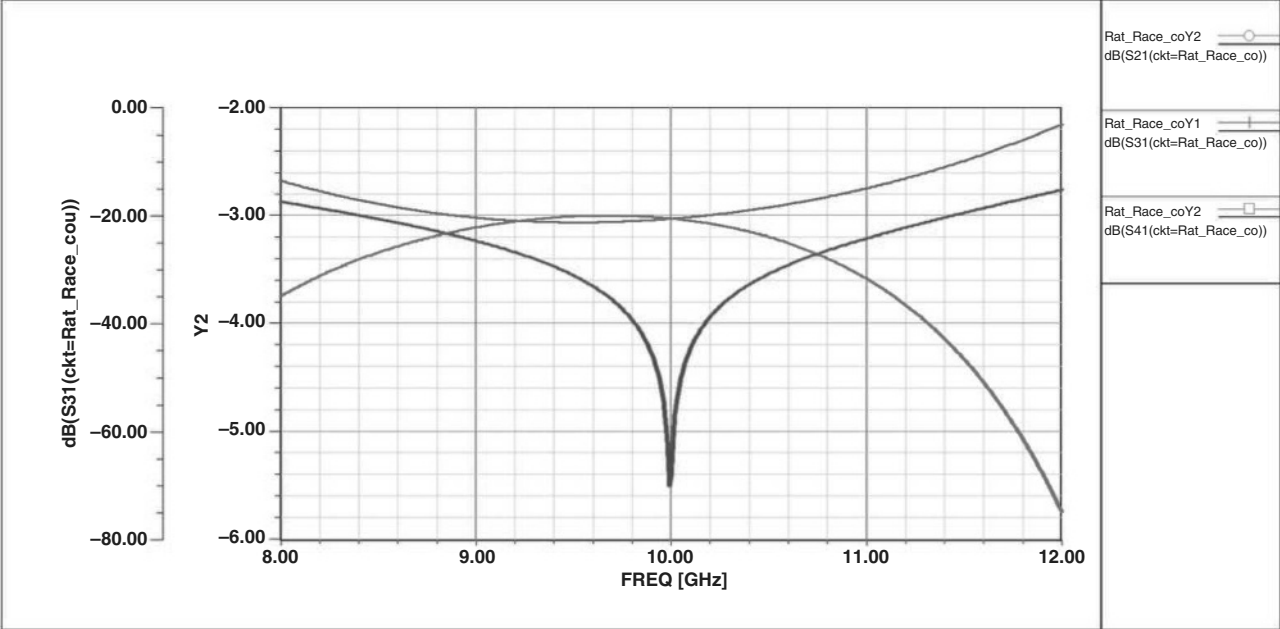


(b)

Figure 1.34 (a) Variation of the rat-race coupler. (b) S21, S31, and S41 in dB of the variation of the rat-race coupler.



(a)



(b)

Figure 1.35 (a) A rat-race coupler. (b) S21, S31, and S41 in dB. Source: Rat-race coupler [11].

The three ports are quarter-wavelength and the fourth is $3/4^{\text{th}} \lambda$.

All these couplers (Figures 1.33–1.35) are related to each other. The branch line coupler and its cousin using bent elements all give the same results.

The next example is the Wilkinson coupler.

1.12 Wilkinson Coupler

Figure 1.36a shows the schematic for the Wilkinson coupler.

The Wilkinson coupler is used to combine two signal sources into one output with up to 45 dB isolation between the ports. A typical example is a power divider, meaning the coupler is used backward. The input is symmetrically fed to two outputs. If the termination has a mismatch, then the 220Ω resistor absorbs the reflected energy (see Figure 1.36a,b). Thus, this coupler can be used either as a power adder or divider. There are also asymmetrical cases in the literature.

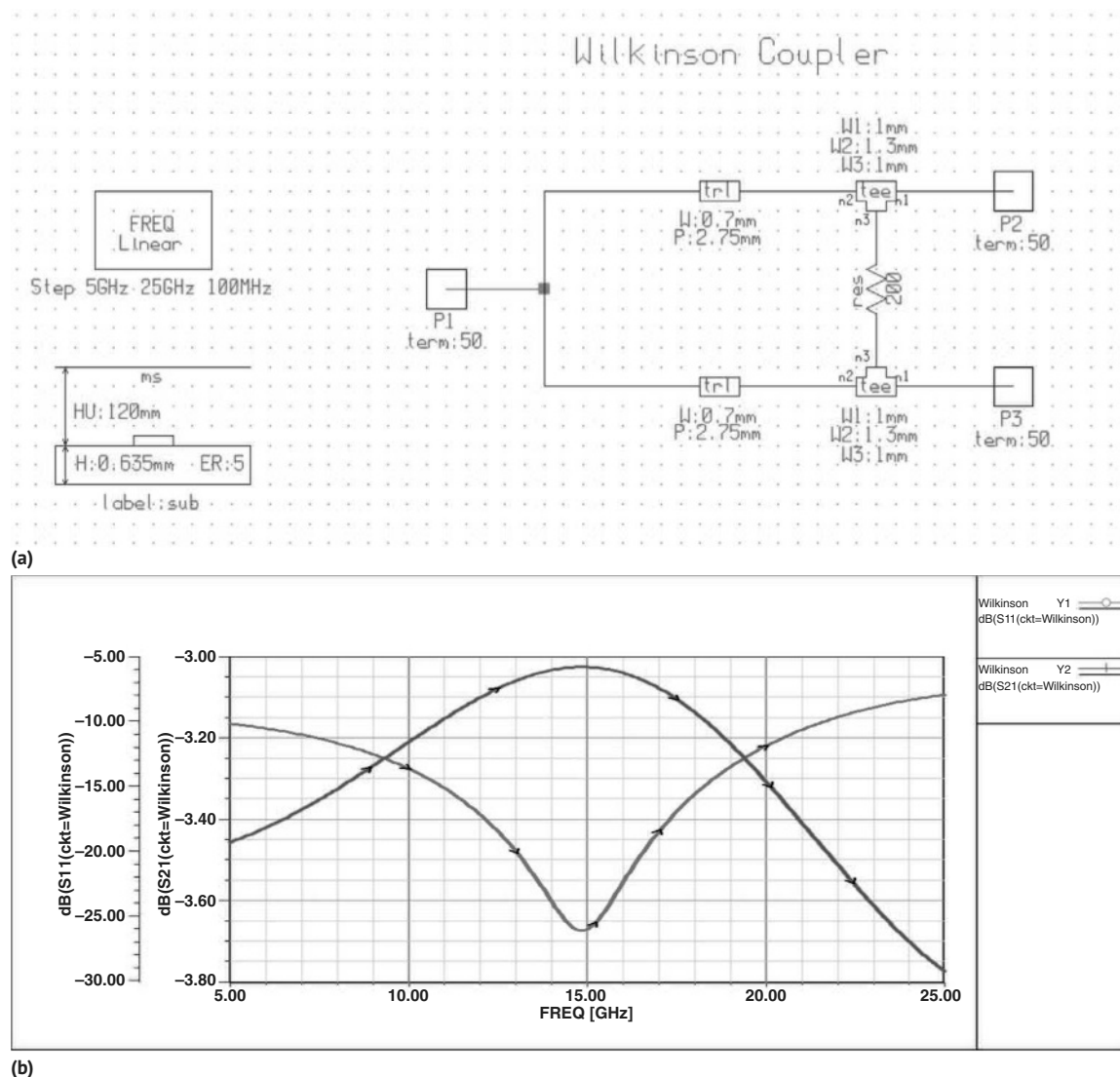


Figure 1.36 (a) Wilkinson coupler. (b) Simulated results for Wilkinson coupler showing S21 and S11 in dB.

Figure 1.37 Schematics of the transmission line elements typically found in a good simulator library.

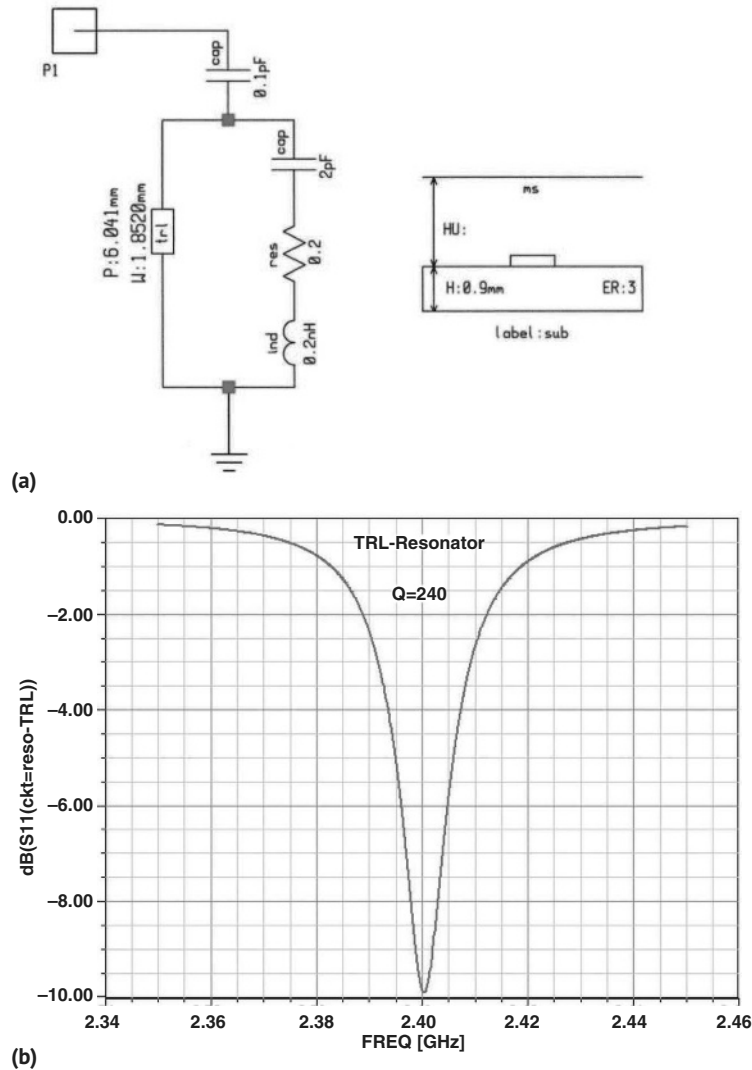
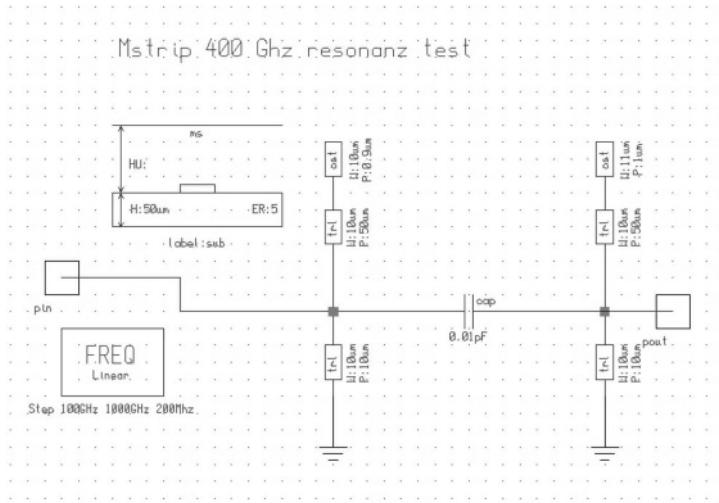


Figure 1.38 (a) 2.4 GHz resonator using both lumped and distributed components (physical dimensions are given). *Source:* [3] / John Wiley & Sons. (b) Simulated reflection coefficient S_{11} to determine the operating Q . Since this material has fairly high losses, an operating Q of only 240 was achieved. *Source:* [3] / John Wiley & Sons.

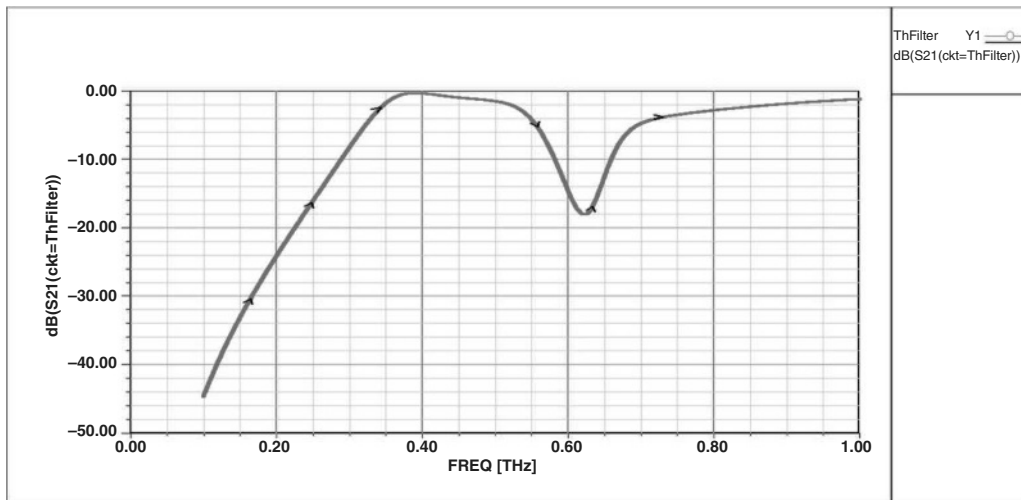
The resonant frequency is at 4.5 GHz, with about 4 dB losses, and has an image reject notch at about 5.06 GHz (see Figure 1.41b).

1.14 Integrated Resonators

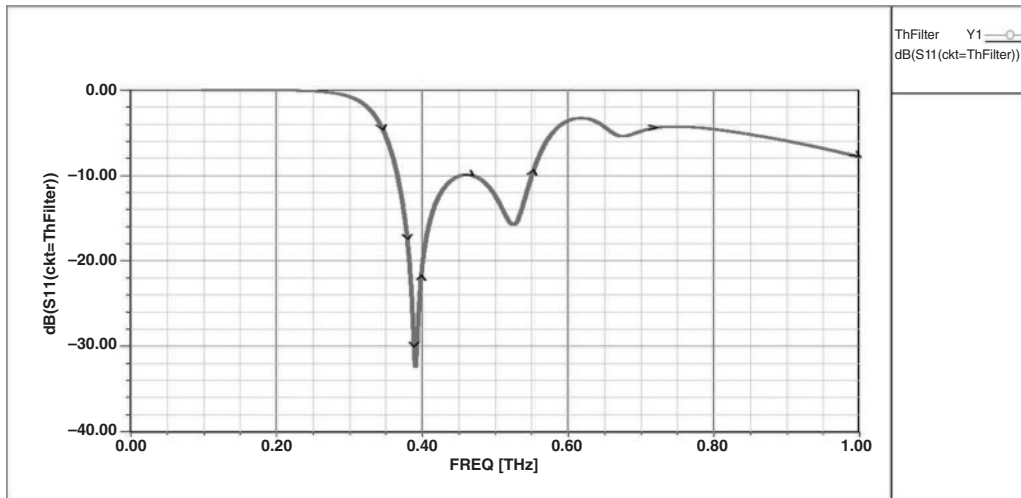
The same circuit can be generated not only using printed circuit board material but also in GaAs or in silicon. Figure 1.42 shows the schematic of a parallel-tuned circuit using a rectangular inductor and an interdigital capacitor. The ground connection is achieved using vias. At 2.4 GHz, the number of turns and the size of the inductor would be significant. The same applies to the capacitor. This arrangement should be reserved for much higher frequencies, above 5 GHz. The inductor losses, both in GaAs and silicon, are substantial, and this case is shown only for completeness. Where possible, an external resonator should be used.



(a)



(b)



(c)

Figure 1.39 (a) Schematic of 400 GHz dual microstrip resonator bandpass filter. (b) S21 in dB of the dual-resonator bandpass filter. (c) Return loss (S11) in dB of the dual-resonator circuit.

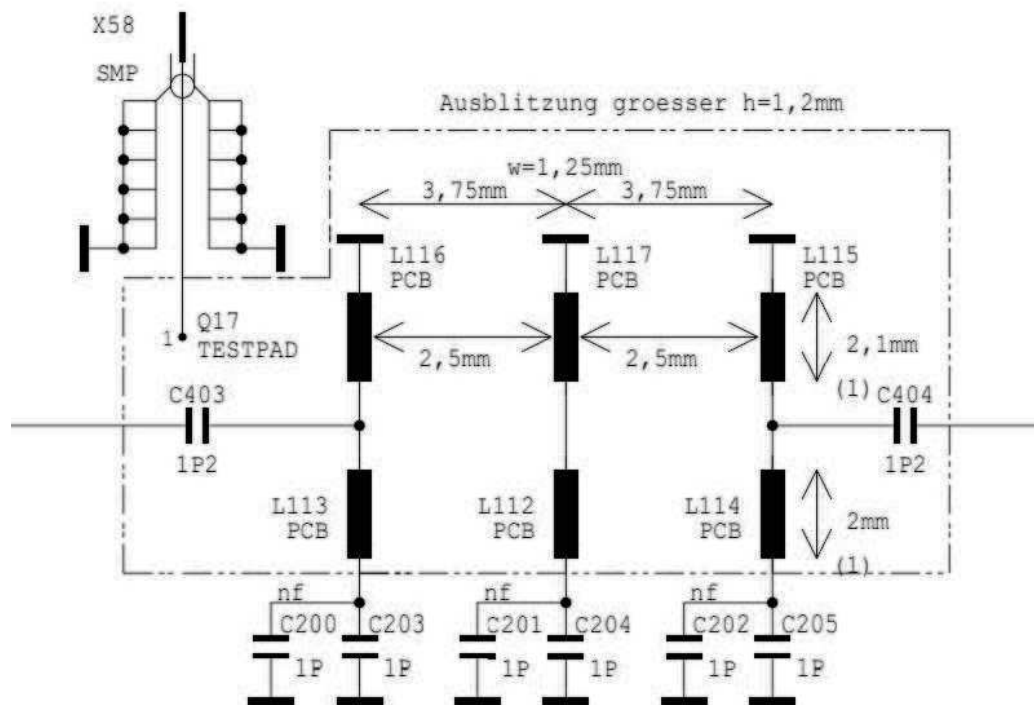


Figure 1.40 Schematic of a 4.5 GHz bandpass filter built on Duroid substrate. *Source:* Courtesy of Rohde & Schwarz.

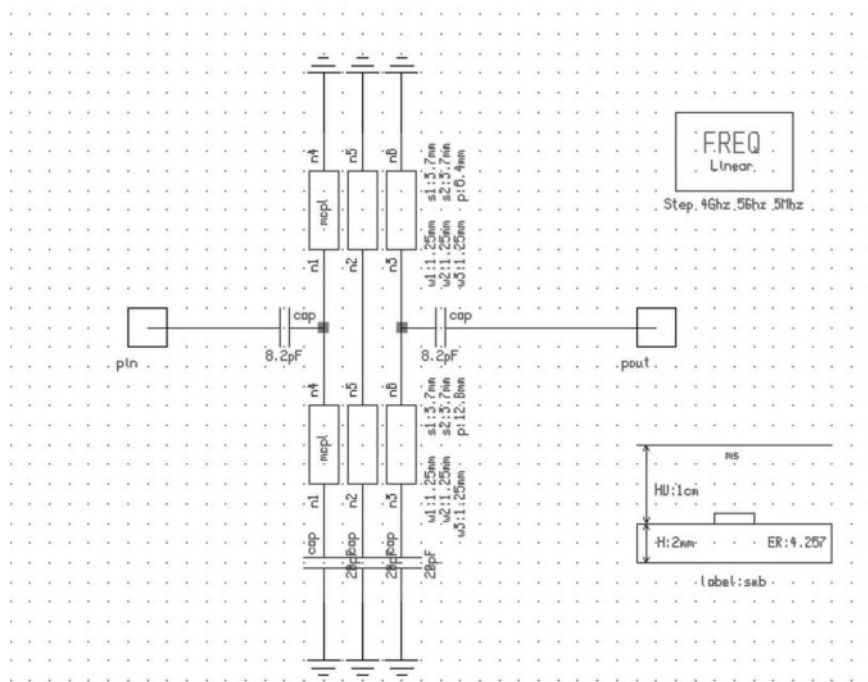
Referring to integrated resonators, a high-Q resonator consisting of two coupled inductors has been developed by us. Figure 1.43 shows a 3D array in which the two coupled resonators are easily identifiable. One side of the resonator is connected to ground through vias.

Throughout the book, we are referring to coupled resonators for this arrangement, but purists could argue rightfully that these are coupled inductors. In reality, we are dealing with coupled microstrips that behave differently from coupled inductors. We have validated this experimentally, and the exact explanation for this behavior would require a lengthy treatment (Figure 5.107 from [1] gives a detailed insight into this). The reason it is frequently called coupled transmission lines or inductors is the fact that the elements themselves are not resonant and are much shorter than quarter-wavelength. This is applicable to two or more coupled transmission lines. The resulting Q can be as much as 10 times as the Q of a single-tuned circuit. The rate of change of phase versus frequency in this configuration is much faster and is responsible for the higher Q.

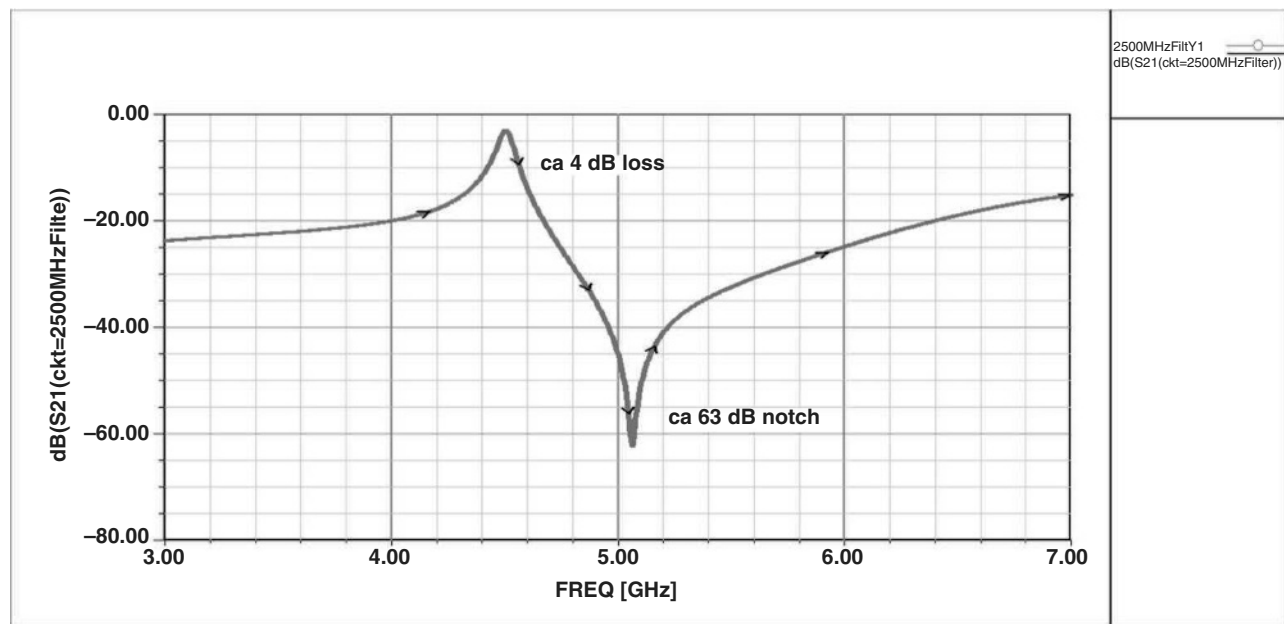
The 3D drawing can be reduced to a two-dimensional drawing (layout), as shown in Figure 1.44, which gives further details about the resonator. The resonator analysis was performed using Ansoft Designer, specifically the 2.5-D simulator Ensemble.

Taking the S-parameters obtained from the structure, a more conventional resonator analysis can be performed. Figure 1.45a shows the electrical equivalent circuit of the coupled microstrip line resonator.

Finally, the S11 resonant curve is analyzed. The curve seen in Figure 1.45b shows a coupled microstripline resonator response, and the resulting Q is determined to be 560. This structure and application is covered by the US patent applications 60/564173 and 60/501371.



(a)



(b)

Figure 1.41 (a) Simulation schematic of the aforementioned filter. (b) Simulation result of the aforementioned filter.

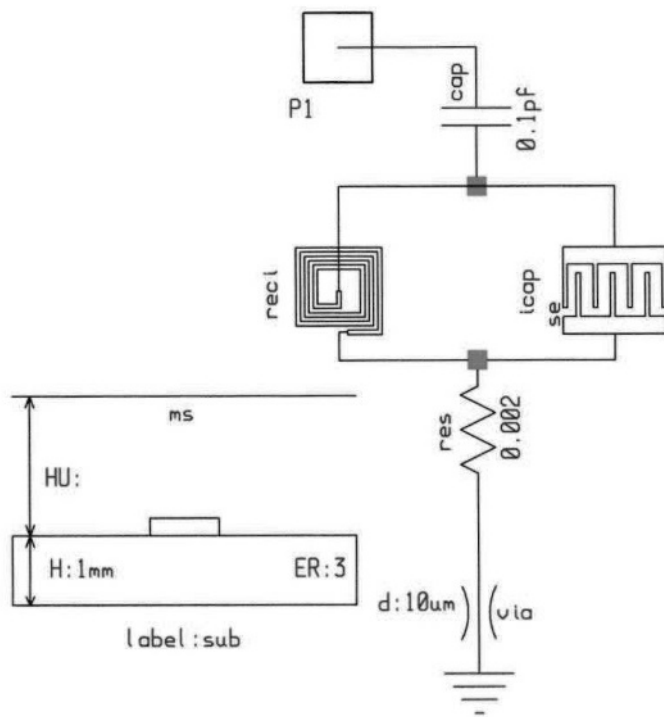


Figure 1.42 Parallel-tuned circuit using a rectangular inductor (could also use a spiral) and an interdigital capacitor. If implemented on GaAs or silicon, it exhibits low Q . *Source:* [3] / John Wiley & Sons.

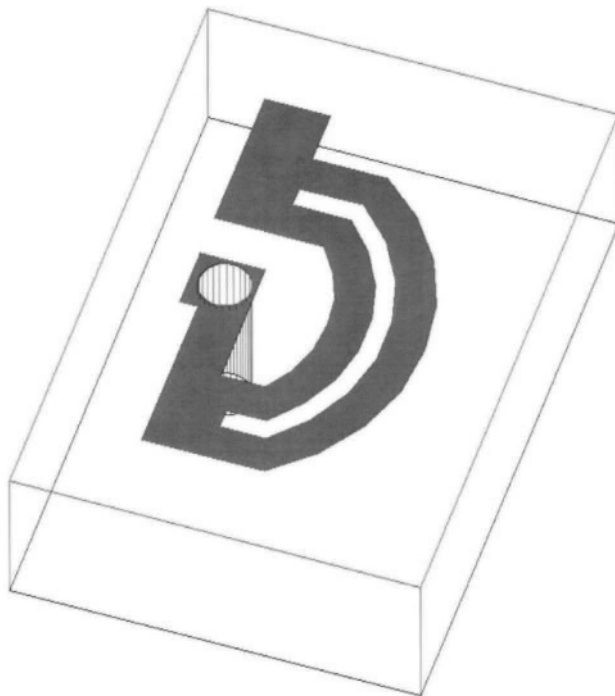
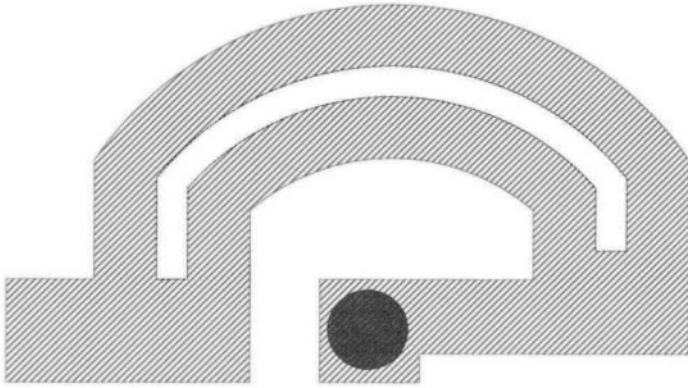


Figure 1.43 3D view of the coupled resonator (patent pending). *Source:* [3] / John Wiley & Sons.



Port

Figure 1.44 A 2D view of the coupled microstrip resonator. *Source:* [3] / John Wiley & Sons.

1.15 Ceramic Resonators

A very popular resonator is the ceramic resonator, which is based on a quarter-wavelength arrangement. The ceramic is silver-plated and can be modeled as a quarter-wave section of a TEM transmission line (cable) with a high relative dielectric constant ranging from 33 to 88. The two 1 pF capacitors shown in Figure 1.46 load the resonator to achieve a resonance frequency of 2.4 GHz. This resonator has an operating Q of about 400, which is much higher than in the previous cases. By varying the capacitors, the frequency can be altered. These are commercial parts made by a variety of companies. Most high-performance (800 MHz–2.4 GHz) oscillators use these resonators.

Figure 1.47 shows a coaxial resonator, while another important resonator, useful up to 40 GHz, is the dielectric resonator (Figures 1.48 and 1.49).

Limitations:

$$1 \leq \epsilon_s \leq \epsilon_r \text{ of substrate}$$

$$1 \leq \epsilon_r < 100$$

$$2 \text{ mm} < D < 50 \text{ mm}$$

$$0.015 \text{ cm} < H_d < 10 \text{ cm}$$

$$0.1 < D/H_d < 10$$

$$H_t \geq 0.0$$

$$0.5 D < L < 0.1 \text{ m}$$

$$0.0 \leq S \leq 5 \text{ cm}$$

$$0.0 \leq H_s < H_d$$

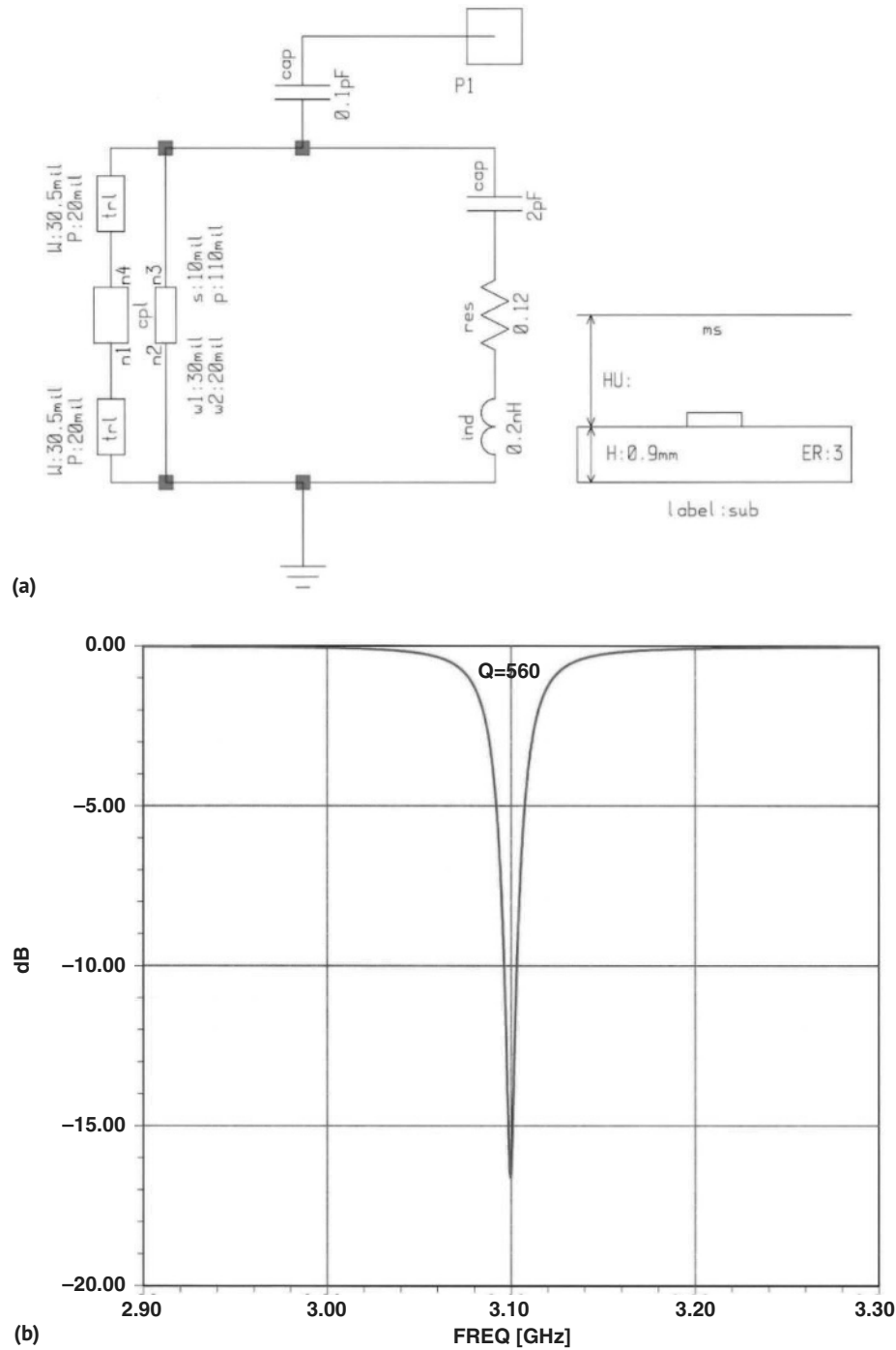


Figure 1.45 (a) Electrical equivalent of the coupled microstrip line resonator. *Source:* [3] / John Wiley & Sons. (b) Frequency response of the coupled microstrip line resonator. *Source:* [3] / John Wiley & Sons.

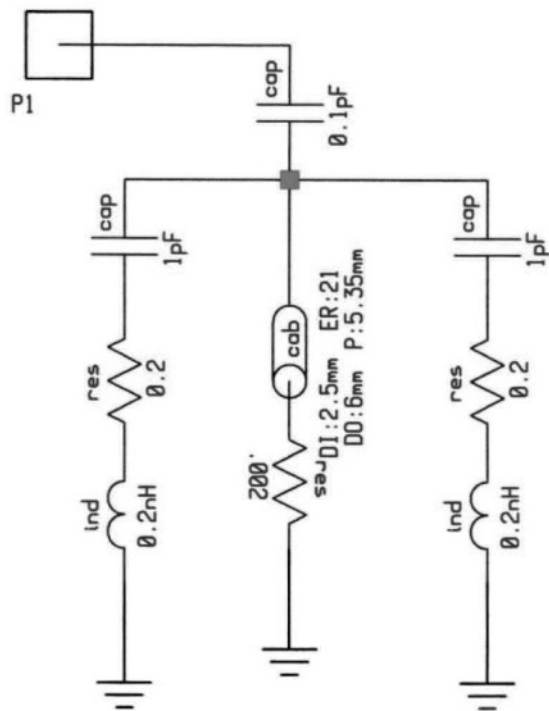


Figure 1.46 A high-Q resonator based on a quarter-wavelength ceramic resonator (CRO). Operating Q values of up to 500 are easily achievable. *Source:* [3] / John Wiley & Sons.

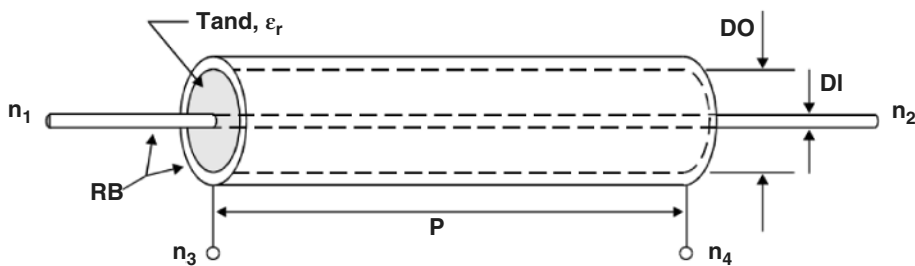


Figure 1.47 Coaxial resonator [14, pp. 2–51].

DRMS—Dielectric Resonator and Microstrip Coupling

The Dielectric Resonator and Microstrip Coupling has two options:

- Option 1: Specify keyword BSF for BandStop structure
- Option 2: Specify keyword BPF for BandPass structure

DRMS BandStop structure

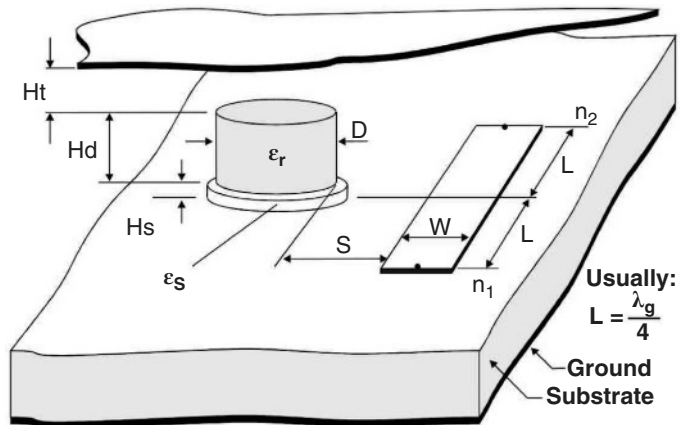


Figure 1.48 Dielectric resonator and microstrip coupling. *Source:* [5] / John Wiley & Sons.

DRMS BandPass structure

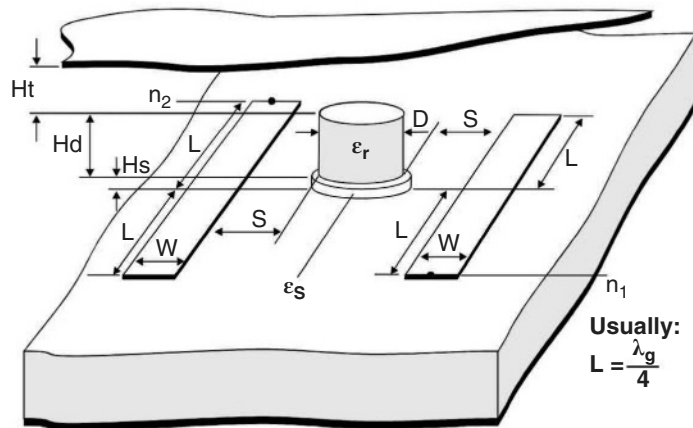


Figure 1.49 Dielectric resonator and microstrip coupling. *Source:* [5] / John Wiley & Sons.

1.16 Conclusion

Designing circuits at low frequencies is mostly trivial. Already at RF frequencies above 400 MHz, complications start as coupling and parasitics begin to play a larger role. In general, the lossy components are the inductors, while the capacitors maintain a high Q up to higher frequencies. The general rule “make the inductor about four times as long as its diameter to have a high Q” loses its meaning for small inductors. Practical experience shows that above 7–10 GHz, only truly dedicated distributed elements are useful. Hopefully, this chapter has provided some inside information and advice about this topic.

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