Capacitively coupled series resonator band-pass filter

By Troy Pandhumsoporn

Microwave Filters

- Microwave filters serve as a significant part of any RF front ends for the suppression of out of band signals
- A filter is reactive, 2-port network that passes a desired band of frequencies while almost restricting all other band of frequencies.
- Used in lumped element/distributed form for commercial, test & measurement, military applications.
- Bandpass filters useful in wireless applications.

Project Description

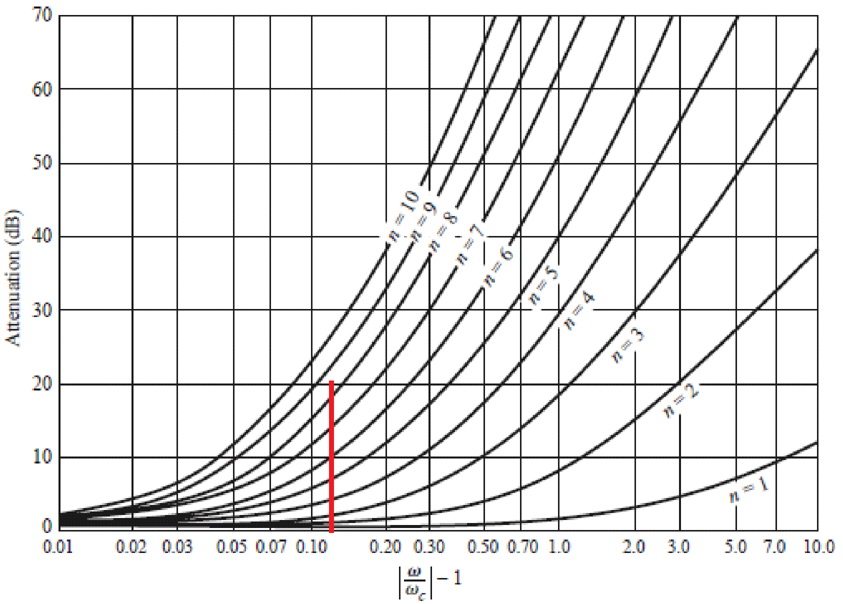
- Design and simulate a **capacitively coupled series resonator band-pass filter** as described in Example 8.9 of the text book with a 0.5 dB equal-ripple pass-band characteristics. The center frequency is 5 GHz, the bandpass width is 20%, at least 20dB attenuation is required at 5.6 GHz. Implement the filter in microstrip technology using Rogers Laminate (RT/duroid 5870) with er= 2.33 , tand = 0.005, and thickness of d= 0.062" (1.575mm). Simulate the design in both schematic and layout windows of ADS (Project for one student) Reference: Microwave Engineering, D. M. Pozar, 4th edition.

- Known parameters:
  - $f_c = 5\ GHz,\ f = 5.6\ GHz, \Delta\ (BPwidth) = 0.2, E_r = 2.33, loss\ tangent = 0.005, d\ (substrate\ thickness) = 1.575\ mm,\ 0.5\ dB\ ripple$
- Conductor trace/copper ground thickness = 15.24 micron (um) from the tutorials.

Starting Design

[3]

- Start with finding the order of the filter (Eq. 8.71) to satisfy the attenuation specification at 5.6 GHz
- $\omega \leftarrow \frac{1}{\Delta}\left(\frac{\omega}{\omega_0} - \frac{\omega_0}{\omega}\right) =$ $\frac{1}{0.2}\left(\frac{5.6}{5} - \frac{5}{5.6}\right) = 1.1357$
- Thus:
- $\left|\frac{\omega}{\omega_0}\right| - 1 = 0.1357$
- That equals N = 9



Low-Pass Prototype Values

[3]

TABLE 8.4 Element Values for Equal-Ripple Low-Pass Filter Prototypes ($g_0 = 1, \omega_c = 1, N = 1$ to 10, 0.5 dB and 3.0 dB ripple)

	0.5 dB Ripple										
N	g_1	g_2	g_3	g_4	g_5	g_6	g_7	g_8	g_9	g_{10}	g_{11}
1	0.6986	1.0000									
2	1.4029	0.7071	1.9841								
3	1.5963	1.0967	1.5963	1.0000							
4	1.6703	1.1926	2.3661	0.8419	1.9841						
5	1.7058	1.2296	2.5408	1.2296	1.7058	1.0000					
6	1.7254	1.2479	2.6064	1.3137	2.4758	0.8696	1.9841				
7	1.7372	1.2583	2.6381	1.3444	2.6381	1.2583	1.7372	1.0000			
8	1.7451	1.2647	2.6564	1.3590	2.6964	1.3389	2.5093	0.8796	1.9841		
9	1.7504	1.2690	2.6678	1.3673	2.7239	1.3673	2.6678	1.2690	1.7504	1.0000	
10	1.7543	1.2721	2.6754	1.3725	2.7392	1.3806	2.7231	1.3485	2.5239	0.8842	1.9841

After procuring LPF filter prototypes, one calculates the inverter constants “Z0*Ji” (Eq 8.121a,b,c), coupling susceptances (Eq. 8.134), and coupling capacitors values.

Design Equations

- For a BPF with N + 1 coupled line sections (Eq. 8.121)
- $Z_0\, J_1 = \sqrt{\frac{\pi\Delta}{2g_1}}$ (Eq. 8.121a)
- $Z_0\, J_n = \frac{\pi\Delta}{2\sqrt{g_{n-1}g_n}}\ for\ n = 2, 3, \ldots, N\ (Eq. 8.121b)$
- $Z_0\, J_{N+1} = \sqrt{\frac{\pi\Delta}{2g_N g_{N+1}}}$ (Eq. 8.121c)

- Coupling Susceptance:

$$B_i = \frac{J_i}{1 - (Z_0\, J_i)^2} \quad (8.134)$$

- $coupling\ capacitor\ values$:

$$C_n = \frac{B_n}{\omega_0}$$

- Resonator lengths given in radians, so convert to degrees.

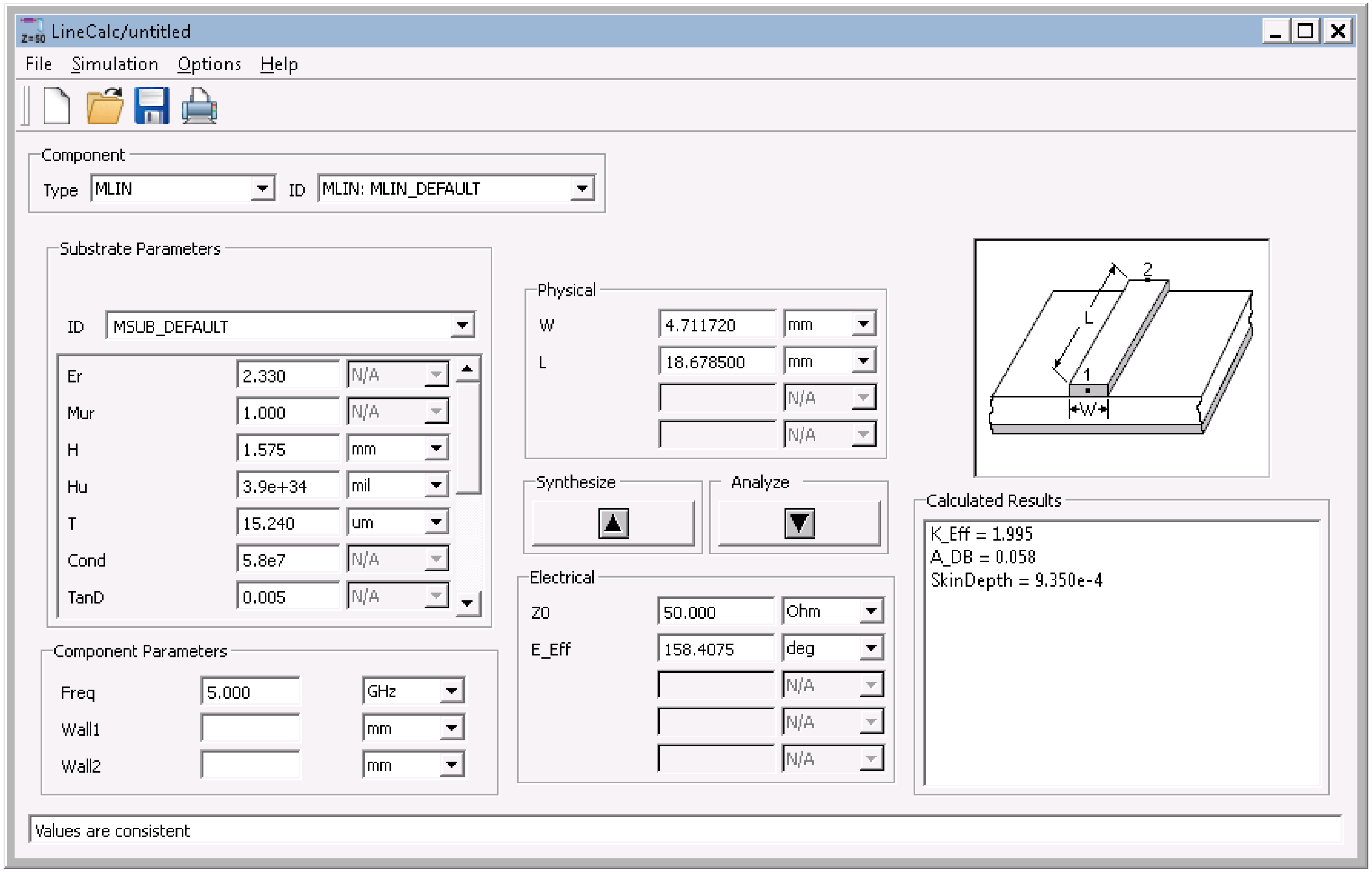
Electrical length: $\theta_i = \pi - \frac{1}{2}[tan^{-1}(2Z_0B_i) + tan^{-1}(2Z_0B_{i+1})]$ (8.135)

Filter Specifications

n	g_n	$Z_0 J_n$	B_n	$C_n\ (pF)$	$\theta_n\ (deg)$
1	1.7504	0.4236	0.0103	0.3287	145.1370
2	1.2690	0.2108	0.0044	0.14043	158.4075
3	2.6678	0.1707	0.0035	0.11196	160.9696
4	1.3673	0.1645	0.0034	0.10763	161.4131
5	2.7239	0.1628	0.0033	0.10646	161.5081
6	1.3673	0.1628	0.0033	0.10646	161.4132
7	2.6678	0.1645	0.0034	0.10763	160.9696
8	1.2690	0.1707	0.0035	0.11196	158.4075
9	1.7504	0.2108	0.0044	0.14043	145.1370
10	1.0000	0.4236	0.0103	0.3287	--

The calculations were done using MATLAB, and organized in the table for reference.

Linecalc (dimensions)



Calculate width and length of each transmission line section from given parameters.

Capacitive-Gap Coupled Resonator BPF Representations

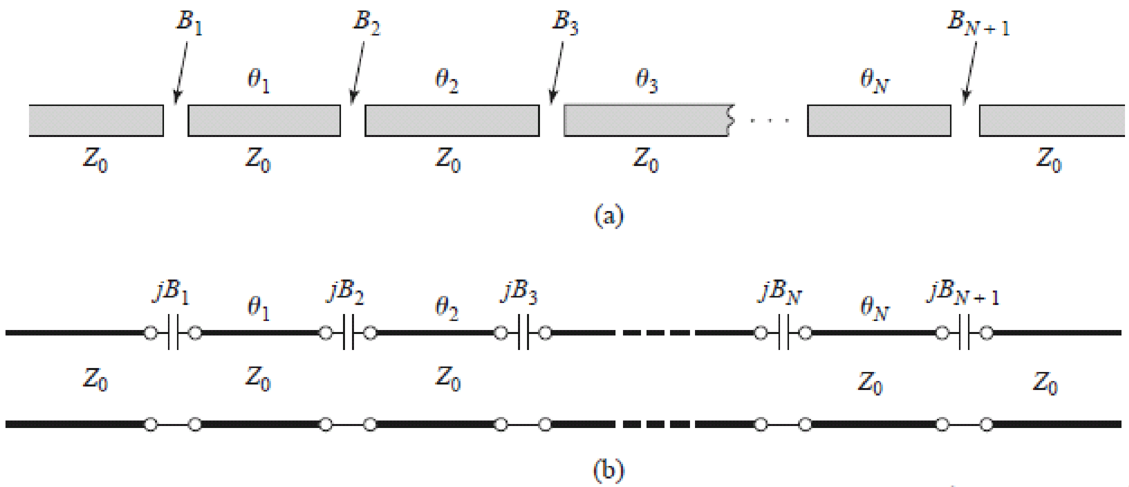
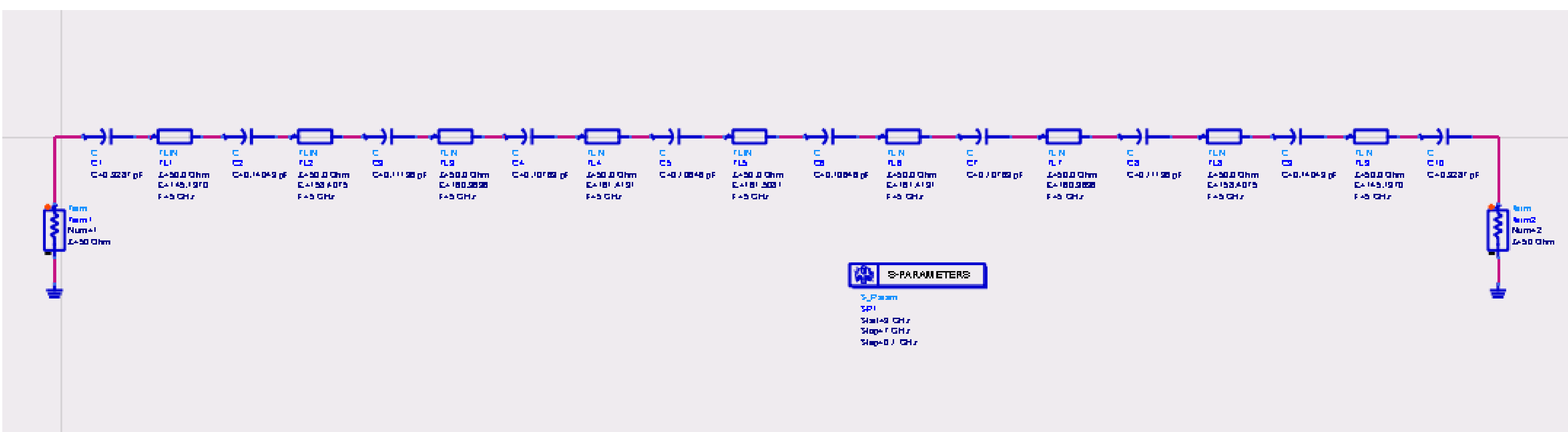


Fig 8.50, pg. 441 [3]

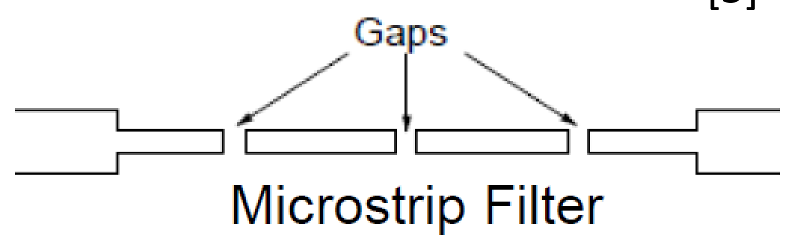


Ideal TL line with lumped element cap case and MGAP model for momentum sim.

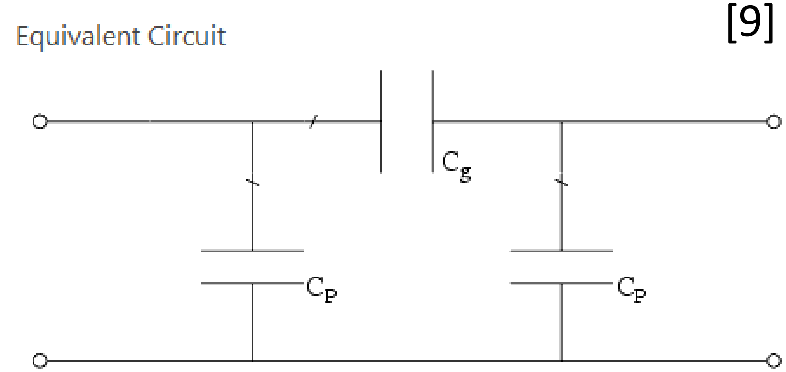
Gap Coupling

- A microstrip gap can be approximated as a π capacitor circuit; the circuit has two reference planes P1 and P2 at each end of the gap consisting of a series coupling capacitance (Cg) and fringing capacitances Cp.
- For narrow gaps, Cp approaches zero and Cg increases (practical capacitance ranges (0.01-0.5 pF)
- For a very large gap, Cg approaches zero and this discontinuity becomes/resembles an open end circuit.

[5]



[9]



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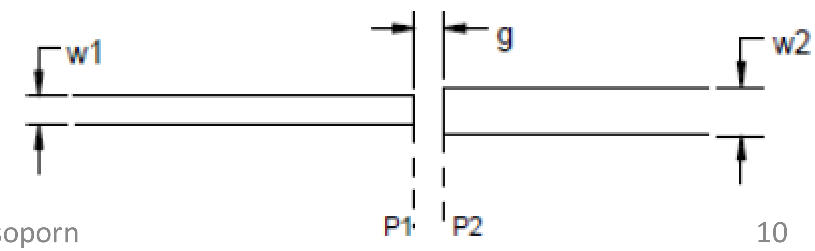
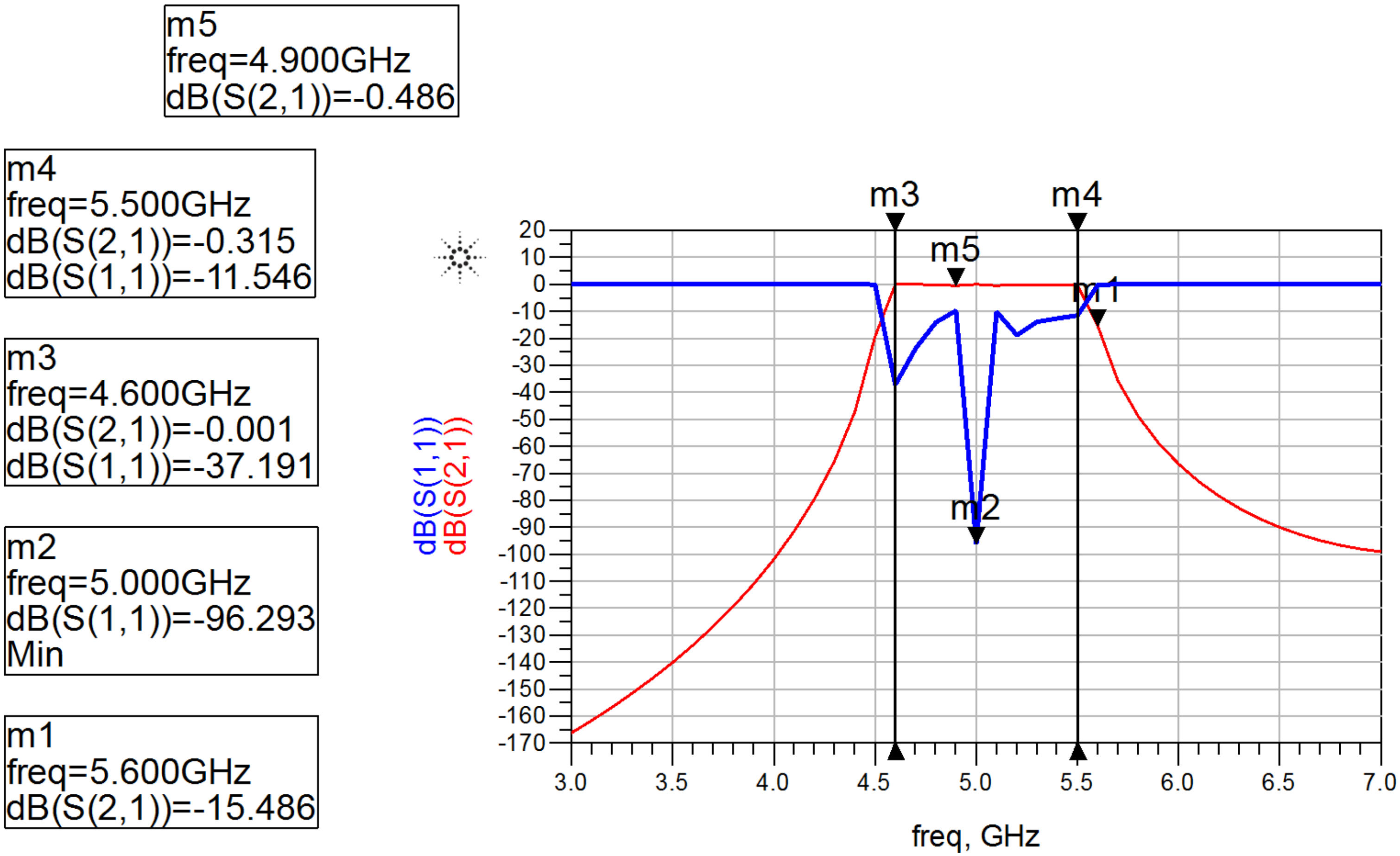
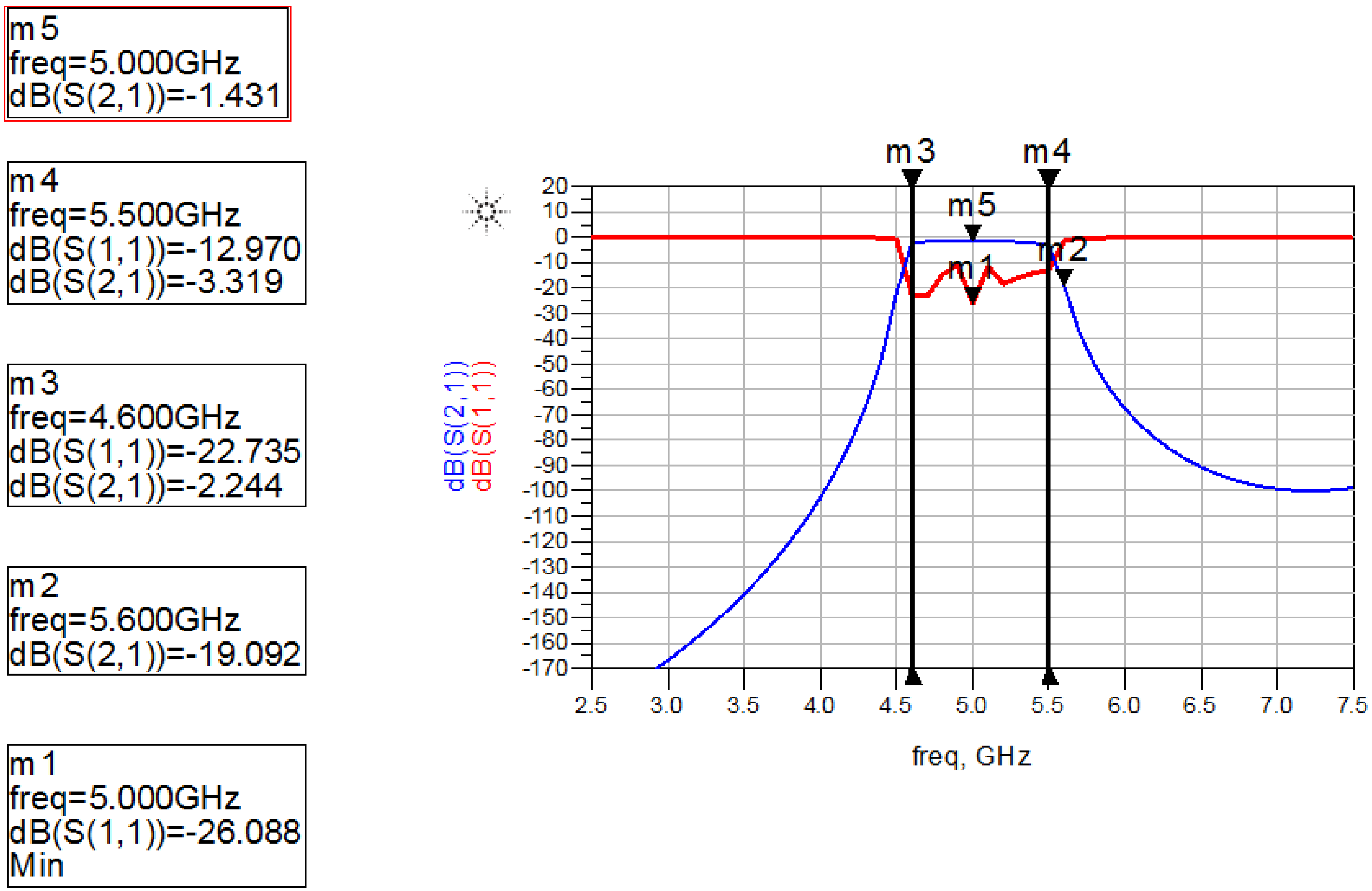


Ideal TM line with series capacitors approximation.



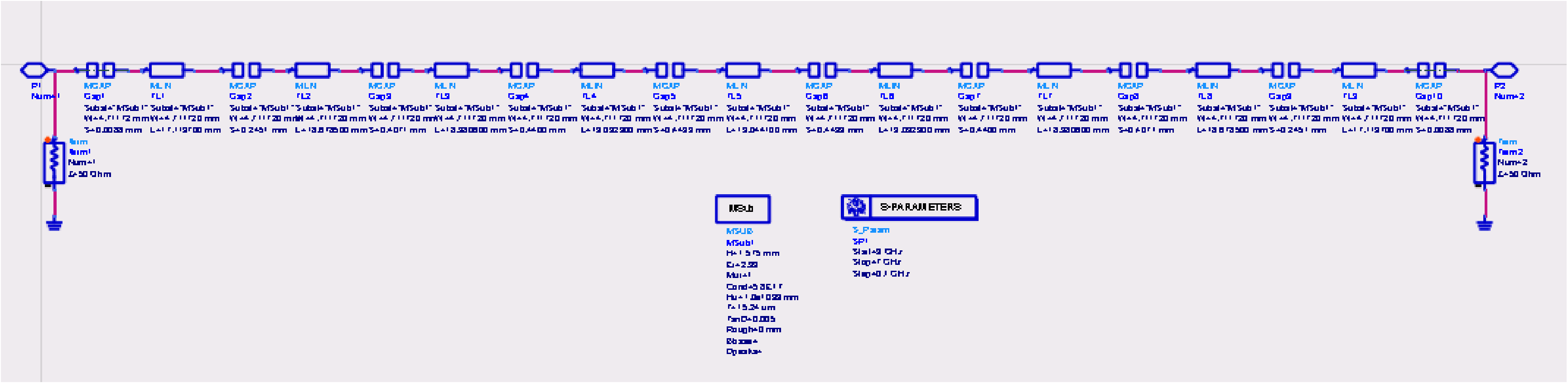
Ripple present on BPF (m5 = -0.486 dB), resonance at center freq., close attenuation to ~20 dB

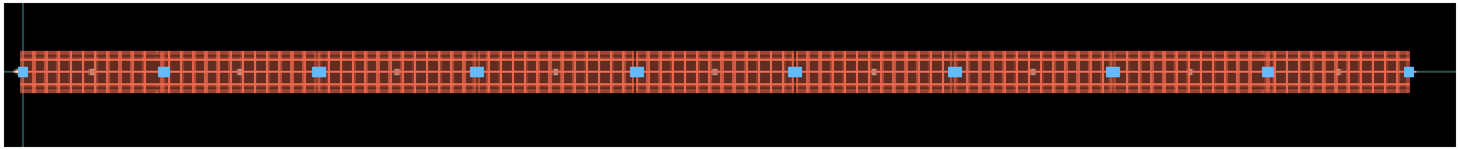
Substrate MLIN with series capacitor approximation



Some ripple (hard to see on the passband), BW does not change, better attenuation at 5.6 GHz, smaller S11.

Schematic (Actual Gap)





Closed Form Solution for Gap Equation

[1] Theoretical Equation (ADS Reference)

The Gap

The microstrip gap may be modelled with a π-network of capacitors, c_p to ground and c_p as the series capacitance. c_p may be converted into an equivalent line extension Δ_p or alternatively the π-network could be transformed into an admttance inverter (J-inverter) with admittance B_g centered between two line extensions Δ_g.

Theoretically calculated data has been given by Silvester and Benedek[3] for a wide range of gap length, s, and several values of ε_r, but for only three values of u, 0.5, 1 and 2. The following equations have been fitted to these data, partly based on the theoretical stripline model :

$$\frac{B_g}{Y} \cdot \frac{\lambda}{h} = 2.4 \, \frac{u + 0.1}{u + 1} \left(\frac{\varepsilon_r + 2}{\varepsilon_r + 1}\right)^{\frac{1}{2}} \ln\left[\operatorname{cotanh}\left(\frac{\varepsilon_r}{\varepsilon_r + 2} \cdot \frac{s}{h}\right)\right] \tag{5}$$

$$\frac{\Delta_p}{h} = \frac{\Delta_e}{h} \tanh^2 \sqrt{0.5 \cdot s/\Delta_e} \tag{6}$$

$$\frac{\Delta_g}{h} = \frac{\Delta_p}{h} + \frac{1}{2\pi} \, \frac{B_g}{Y} \, \frac{\lambda}{h} \tag{7}$$

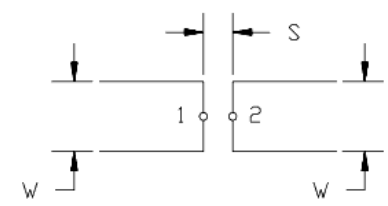
The experimental data taken for several gap and strip widths gives an acceptable fit to the above equations. Their accuracy is estimated to be in the order of 0.05 in B_g/Y and $0.1 \cdot h$ in Δ_g for an impedance range of 25-70 ohm for $\varepsilon_r = 9.8$ and 35-100 ohm for $\varepsilon_r = 2.2$. The reference planes are at the line ends of the gap.

Design Parameter S (Research) [9]

MGAP (Microstrip Gap)

Symbol

Illustration



Parameters

Name	Description	Units	Default
Subst	Substrate instance name	None	MSub1
W	Conductor width	mil	25.0
S	Length of gap (spacing)	mil	10.0
Temp	Physical temperature (see Notes)	°C	None

Pozar and most resources mention but do not explain about the gap dimensions.

Gap Equation (Hammerstad)

- Rewritten s (m) in terms remaining variables:
- $s = h * \left(\frac{\varepsilon_r+2}{\varepsilon_r}\right) * \operatorname{atanh}\left(\exp\left(\frac{B_i\, Z_0\, \lambda(u+1)}{h\,(u+0.1)*2.4} * \sqrt{\frac{\varepsilon_r+1}{\varepsilon_r+2}}\right)\right)$
- u is W/d from the microstrip line design equations.

[3]

$$\frac{W}{d} = \begin{cases} \dfrac{8e^{A}}{e^{2A}-2} & \text{for } W/d < 2 \\ \dfrac{2}{\pi}\left[B - 1 - \ln(2B-1) + \dfrac{\epsilon_r - 1}{2\epsilon_r}\left\{\ln(B-1) + 0.39 - \dfrac{0.61}{\epsilon_r}\right\}\right] & \text{for } W/d > 2, \end{cases} \tag{3.197}$$

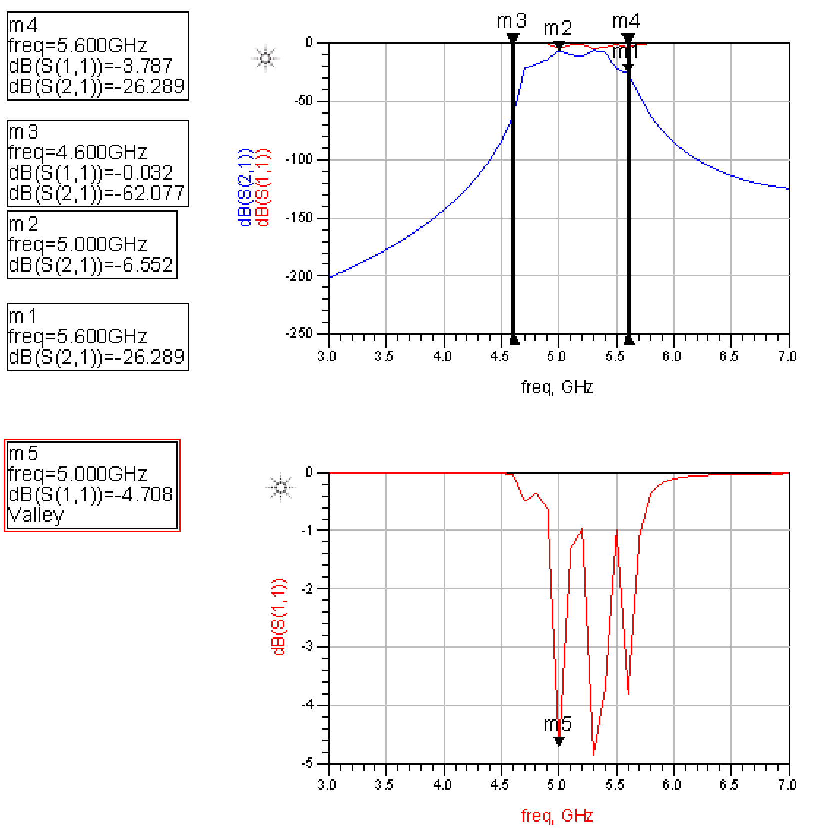
$$A = \frac{Z_0}{60}\sqrt{\frac{\epsilon_r+1}{2}} + \frac{\epsilon_r - 1}{\epsilon_r + 1}\left(0.23 + \frac{0.11}{\epsilon_r}\right)$$

$$B = \frac{377\pi}{2Z_0\sqrt{\epsilon_r}}.$$

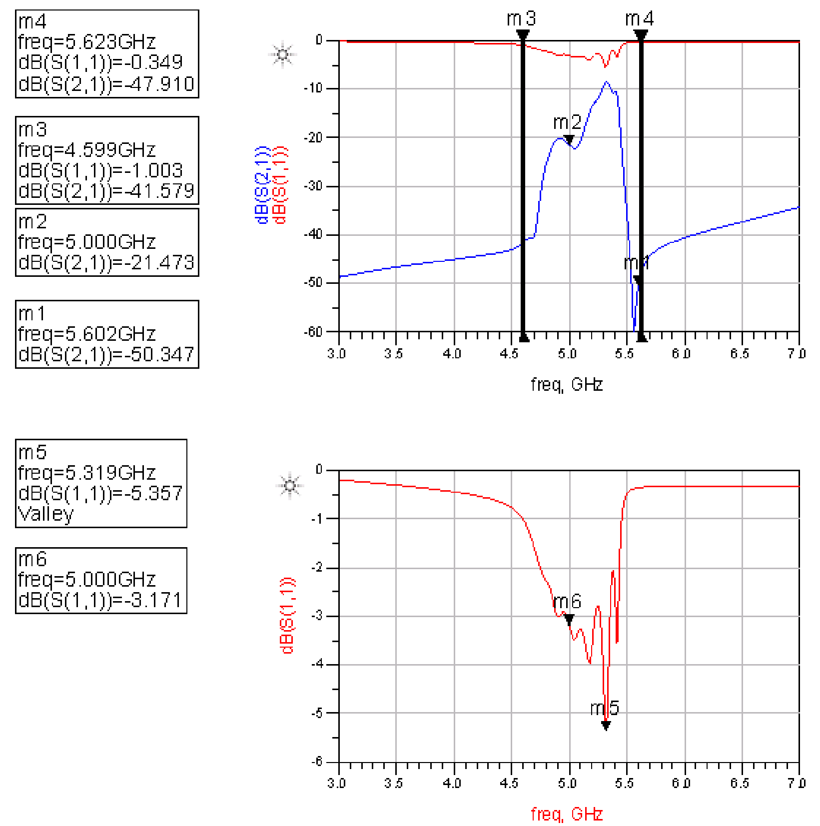
$$\epsilon_e = \frac{\epsilon_r + 1}{2} + \frac{\epsilon_r - 1}{2}\frac{1}{\sqrt{1 + 12d/W}}. \tag{3.195}$$

Real Gap ADS/Momentum Simulation [1] (Hammerstad)

ADS Simulation



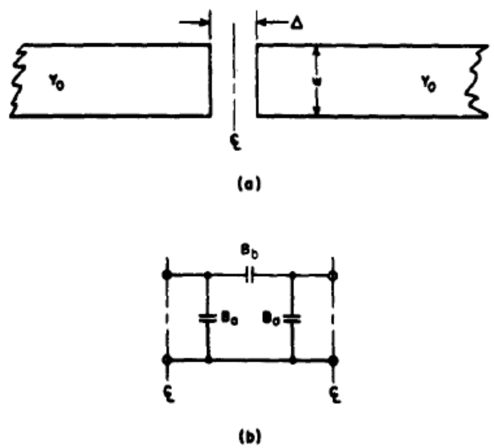
Momentum Sim.



ADS: Multiple small resonances observed, 20 dB attenuation met for ADS simulation.
Momentum: smaller bandwidth, greater attenuation at 5.6 GHz, small resonance.

Gap Equation (Matthaei)

[10]



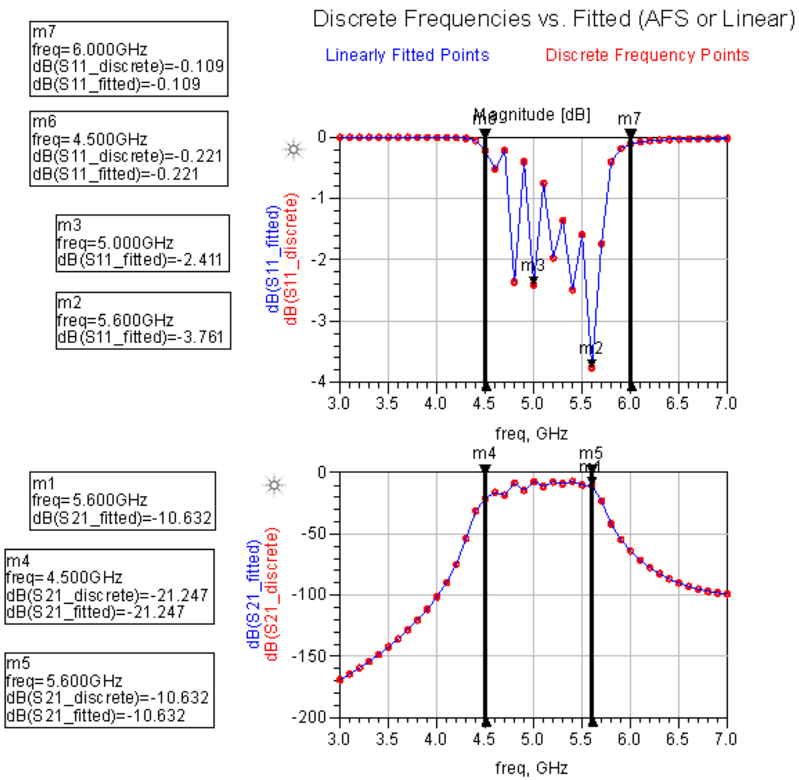
$$\frac{B_a}{Y_0} = -\frac{2b}{\lambda} \ln\left[\cosh\left(\frac{\pi\Delta}{b}\right)\right] \quad (1)$$

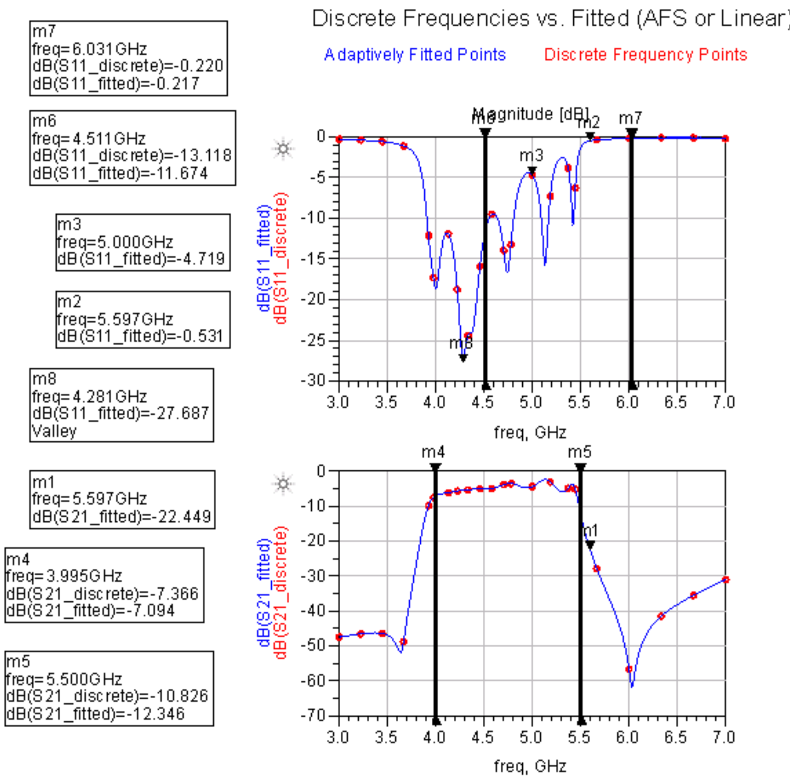
$$\frac{B_b}{Y_0} = \frac{b}{\lambda} \ln\left[\coth \frac{\pi\Delta}{2b}\right] \quad (2)$$

FIG. 8.05-2 GAP EQUIVALENT CIRCUIT, AND OLINER'S EQUATIONS[9,10] FOR CAPACITIVE-GAP SUSCEPTANCES FOR THIN STRIP LINE
Parameter b is the ground-plane spacing, and λ is the wavelength in media of propagation, in same units. Equations are most accurate for w/b ≈ 1.2 or more and t/b ≈ 0, where t is the strip thickness.

- Solving for S = Δ
- $S = \frac{2\pi}{b} \operatorname{acoth}\left(e^{\frac{B_i}{Y_0}\frac{\lambda}{b}}\right)$
- b = ground plane spacing = d = substrate thickness = 1.575 mm
- $Z_0 = {}^{1}/_{Y_0}$
- Equations reasonably accurate if w/d is fairly large, for 50 Ω line having marginal (near zero) thickness and air dielectric.

Real Gap ADS/Momentum Simulation [1] (Matthaei)





ADS: BPF shape preserved, small resonance

Momentum: Resonant center frequency shifted, wider bandwidth, 20 dB attenuation reached at 5.6 GHz, BPF shape more prominent.

Tabulated Gap Values

Hammerstad (Gap Values)

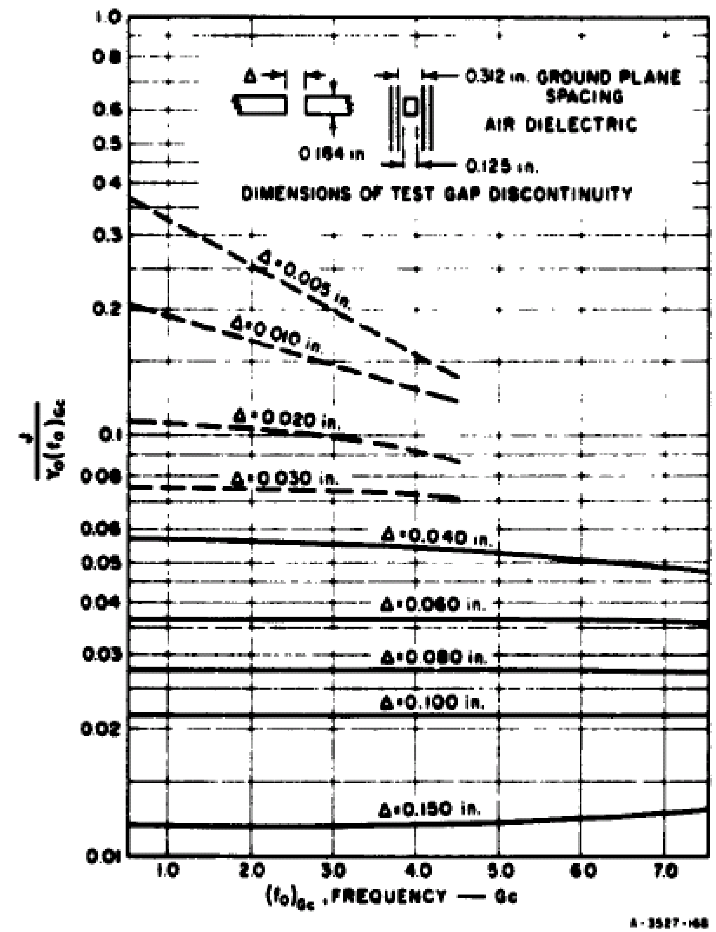
Section #	Gap length s (mm)
S1	0.0088
S2	0.2451
S3	0.4071
S4	0.4400
S5	0.4493
S6	0.4493
S7	0.4400
S8	0.4071
S9	0.2451
s10	0.0088

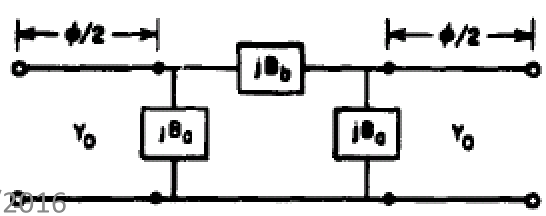
Matthaei (Gap Values)

Section #	Gap Length s (mm)
S1	1.0643 x 10^{-6}
S2	0.0028
S3	0.0093
S4	0.0111
S5	0.0117
S6	0.0117
S7	0.0111
S8	0.0093
S9	0.0028
S10	1.0643 x 10^{-6}

Capacitive Gap Equation for Matthaei solution, new set of design equations

- A different set of equations needs to be used…
- A different set of equations needs to be used…
- Find $\frac{J}{Y_0\ f_0(c)(in\ GHz)}$
- $Find\ corresponding\ \Delta$
- Convert from inches to mm





[10], pg. 438, 441.

Table Method

Conclusions

- Substrate permittivity affects the S11 and filter characteristics (attenuation, resonance)
- Gap Length S plays a significant role in defining filter characteristics
  - Small gap lengths are preferred/usual. (e.g. 0.0XXX mm) to maintain the BPF shape.
- Closed Form Equations do not always yield desired results in simulations. (Manual tuning for gap lengths)

Observations

[10] pg. 373-380

- Advantages
  - appropriate for planar arrangement, is easily implemented with printed circuit board technology and taking up no more space than a plain transmission line would.
- Disadvantages
  - The performance (insertion loss) deteriorates with increasing fractional bandwidth Δ, and acceptable results are not procured with a *Q* less than about 5.
  - An additional hurdle with producing low-*Q* designs is that the gap width is required to be smaller for wider fractional bandwidths. The minimum width of gaps, like the minimum width of tracks, is restricted by the resolution of the printing technology (PCB board house).

Works Cited

- [1] Hammerstad, E., "Computer-Aided Design of Microstrip Couplers with Accurate Discontinuity Models," *Microwave Symposium Digest, 1981 IEEE MTT-S International* , vol., no., pp.54,56, 15-19 June 1981 (Gap Design Equation)

- [2] Ojeda, Roberto Conde. "Capacitive Gaps RF Bandpass Microstrip Filter Design - File Exchange - MATLAB Central." *Capacitive Gaps RF Bandpass Microstrip Filter Design*. Roberto Conde Ojeda/The MathWorks, Inc., 27 June 2004. Web. 28 May 2015. <http://www.mathworks.com/matlabcentral/fileexchange/5360-capacitive-gaps-rf-bandpass-microstrip-filter-design>.

- [3] Pozar, David M. "Section 8.8: Filters Using Coupled Resonators." *Microwave Engineering*. 4th ed. Hoboken, NJ: J. Wiley, 2005. 437-43. Print.

- [4] GOOGLE Translate

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- [9] ADS 2014 Help

- [10] Matthaei, George L., E. M. Jones, and Leo Young. "SEC. 8.05 Capacitive-Gap-Coupled Transmission Line Filters." *Microwave Filters, Impedance-matching Networks, and Coupling Structures*. New York U.a.: McGraw-Hill, 1980. 440-46. Print.