

WIPL-D Microwave Pro: What is New in v6.0?

New features/improvements introduced in v6.0 are:

1. **Numerous Upgrades in Filter Designer**
 - a. A number of bandpass and bandstop filters were added,
 - b. The filters are available in ideal line and microstrip technologies,
 - c. Parallel Coupled Resonators, Coupled SIRs and Hairpin-Line are new bandpass filters,
 - d. Electrically Coupled Resonators, Open Stubs and Coupled Lines are new bandstop filters,
 - e. Minor improvements and bug fixes.
2. **Multiconductor lines - multiconductor lines are now available in microstrip and stripline technology.**
 - a. 2D quasi-static EM analysis method used for calculation,
 - b. One to eight conductors,
 - c. User controlled accuracy of the calculation,
 - d. Visualization of the multiconductor structure and segmentation,
 - e. Low frequency loss calculation – below the skin effect region.
3. **Impedance Calculator was improved by adding synthesis of Coupled Microstrip Lines from characteristic impedances of the even and odd mode.**
4. **Expanded list of active 2-port network parameters with:**
 - a. Conjugate input and output match calculation,
 - b. Maximum stable gain and maximum transducer gain calculation.

Filter Designer window is presented in Fig. 1. The synthesis of all new topologies is possible in both ideal transmission line technology and microstrip technology. With the latest enhancements, the number of options to solve a particular bandpass filter problem increases significantly.

The choice of bandstop filters available in the previous version was limited to lumped element filters only. Now the range of filters has been expanded to include *Electrically Coupled Resonators*, which is available in ideal transmission lines technology only, and two additional topologies, *Open Stubs* and *Coupled Lines*, which are available in both ideal and microstrip technologies. A *Microstrip Open Stubs* filter is well suited for relatively wide stopband widths, while a *Microstrip Coupled Lines* filter is suitable for narrow stopbands. For both of the filter types the equations required for the synthesis are analytically derived up to the 7th order. The list of new bandstop filters is presented in Fig. 2.

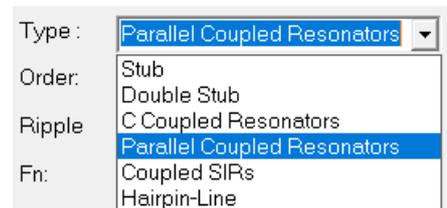


Fig. 1. Ideal transmission lines bandpass filter list.



Fig. 2. Ideal transmission lines bandstop filter list.

1. Numerous Upgrades in Filter Designer

a) New Filter Topologies

Several bandpass and bandstop filter topologies have been introduced in the new Filter Designer release, an add-on tool for WIPL D Microwave Pro v6.0. The available range of filters now covers the most popular distributed topologies.

Recently added bandpass topologies include *Parallel Coupled Resonators*, *Coupled SIRs* and *Hairpin-Line*. The list of new and previously available types as it appears in the

b) Coupled Line Filters in WIPL-D Filter Designer

Most of the new filters listed above are based on coupled transmission lines. Due to the specifics of the *Ideal Coupled Line* component introduced in WIPL-D Microwave (described in detail in the *User Guide*), the user can explore the influence of the unequal phase velocities of the even and odd mode on coupled lines on filter performance.

In quasi-TEM lines, like microstrip, phase velocities, i.e. effective dielectric constants, of even and odd modes are different, while in pure TEM lines, like stripline, they are

equal. Unlike some other filter design tools, the effects of different effective dielectric constants of the modes, E_{re} and E_{ro} , on filter performance can be explored in *WIPL-D Filter Designer*. For example, in the particular case presented in Fig. 3, where $E_{re}=2.65$ and $E_{ro}=2.35$, the resulting filter response has a spurious passbands around the second and third harmonic of the filter (see Fig. 4).

Type:

Order:

Ripple: dB

Fn: GHz

Ere: Ero:

Fig. 3. In quasi-TEM lines (microstrip) the effective dielectric constants of the even and odd modes are different.

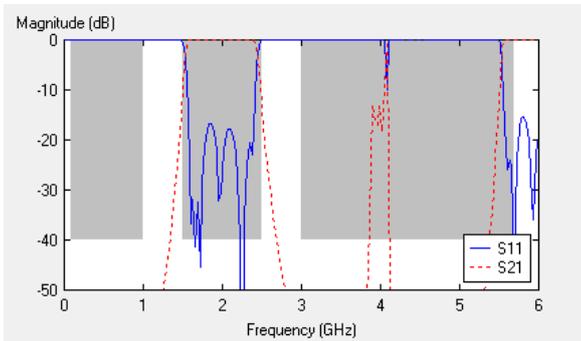


Fig. 4. Spurious passband appearing around the 2nd harmonic of the filter for a quasi-TEM line filter.

On the other hand, if the effective dielectric constants of the modes are chosen equal, as in Fig. 5 where $E_{re}=E_{ro}=2.5$, the spurious passband around the second harmonic does not appear, which can be verified in Fig. 6 where the corresponding filter response is presented.

Type:

Order:

Ripple: dB

Fn: GHz

Ere: Ero:

Fig. 5. In pure TEM lines (stripline) the effective dielectric constants of the modes are equal.

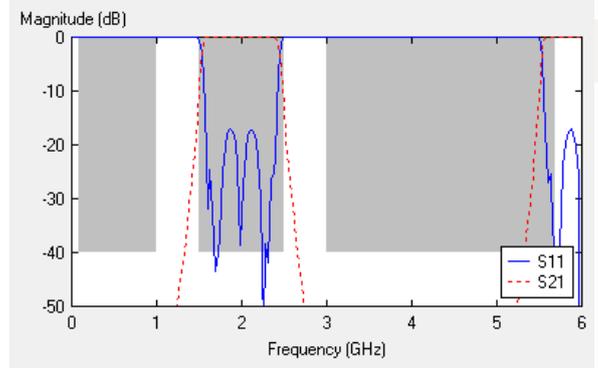


Fig. 6. With pure-TEM transmission lines spurious passband around 2nd harmonic does not appear.

c) Conversion to Microstrip

The ideal *coupled line* filters from those listed in Fig. 1 and Fig. 2 can all be converted to microstrip circuits. This is possible as coupled microstrip line synthesis has been introduced in the recent *Microwave Pro* release. Besides the previously available conversion from physical to electrical parameters, the opposite direction of calculation is now also possible, i.e. the conversion from electrical to physical parameters. More details can be found in the section 3 of this document - *3. Coupled Microstrip Line Synthesis* where implementation of a synthesis routine in *Impedance Calculator* is described.

An example of a filter converted from ideal to microstrip for the case of a bandpass *Parallel Coupled Resonator* filter is presented in Fig. 7.

d) Other Improvements

Besides new topologies, minor improvements introduced into the latest release of the Filter Designer tool include the correction of the errors in filter order calculation for bandpass and bandstop filters, consolidated automatic assignment of symbolic variables which greatly simplifies subsequent tuning and optimization, plotting of S11 filter parameter, convenient scaling of the response axes, etc.

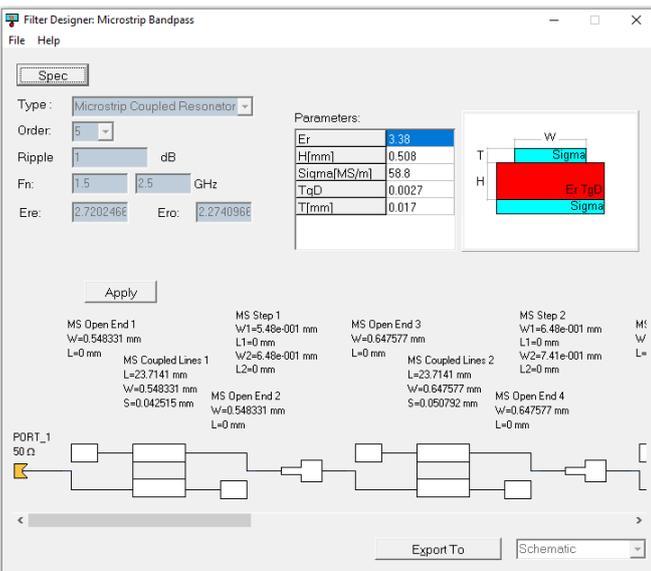
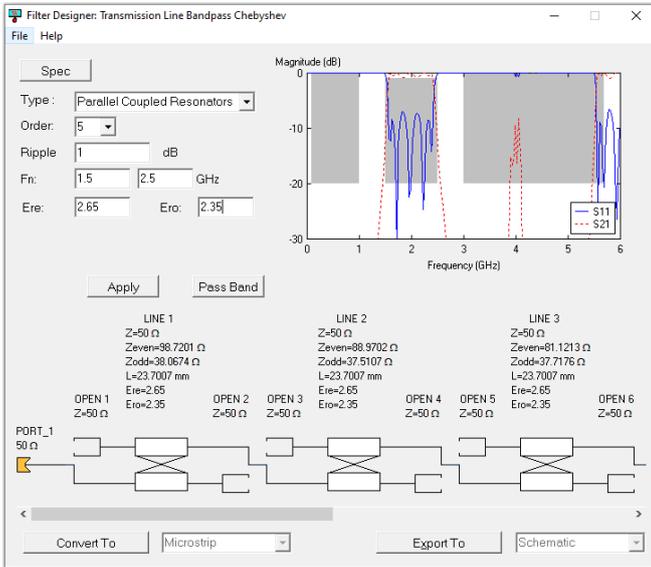


Fig. 7. Ideal coupled line filter converted to microstrip (5th order bandpass Parallel Coupled Resonators).

2. Multiconductor Lines

Microstrip and stripline multiconductor lines are available starting from v6.0. Multiconductor components add a new value to the WIPL-D Microwave Pro environment as many useful circuits, previously outside the reach of the circuit solver, can now be analyzed. These include interdigital and combline microstrip and stripline filters, Lange couplers, etc. Examples of such circuits are presented in Fig. 8 - Fig. 10.

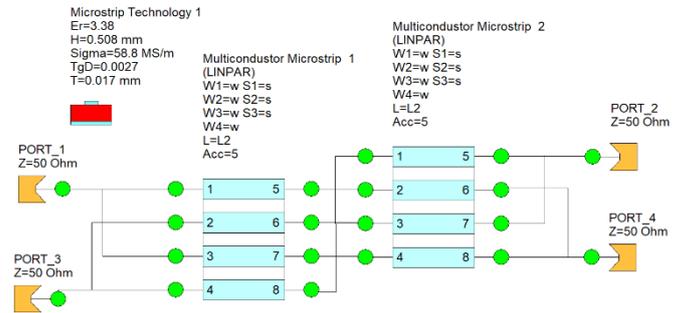


Fig. 8. Microstrip Lange coupler.

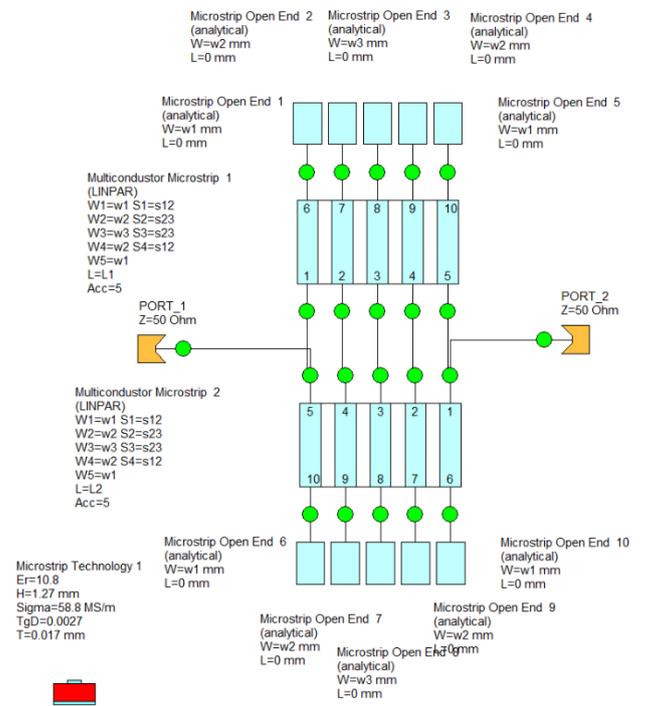


Fig. 9. Microstrip pseudocombine filter.

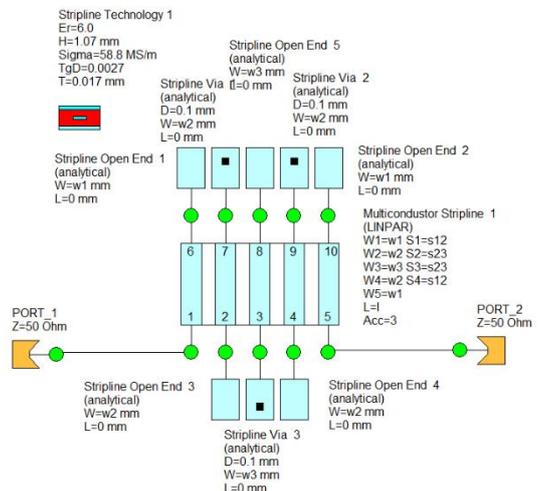


Fig. 10. Stripline interdigital filter.

Examples of microstrip and stripline multiconductor components are presented in Fig.11 and Fig.12, respectively. For both of the components the number of lines N can be selected from one to eight. Once a number of the lines is selected, the corresponding number of fields required to input strip widths w_i , $i=1,\dots,N$ and separation widths s_i , $i=1,\dots,N-1$ appears automatically.

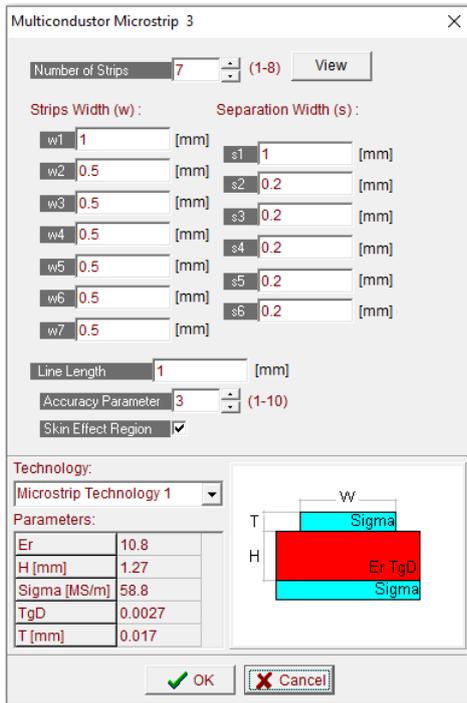


Fig. 11. A microstrip multiconductor component for $N=7$ conductors and default parameter values.

The calculation of the S-parameters of a multiconductor line is based on the method described in [1]. In the first step 2D quasi-TEM matrices for multiconductor transmission lines embedded in piecewise-homogeneous dielectrics are evaluated numerically. A multiconductor line is assumed to be infinitely long and with a constant cross section along the length. The accuracy of the method, which translates to the density of the segmentation applied to the dielectric and conductor contours, can be adjusted by a single parameter designated as *Accuracy* which can take values from 1 to 10. The higher the parameter value, the higher the accuracy will be. However, higher accuracy results in longer simulation, so a judicious trade-off should be found. Calculated quasi-TEM matrices are subsequently used to calculate scattering parameters for a specified multiconductor line length.

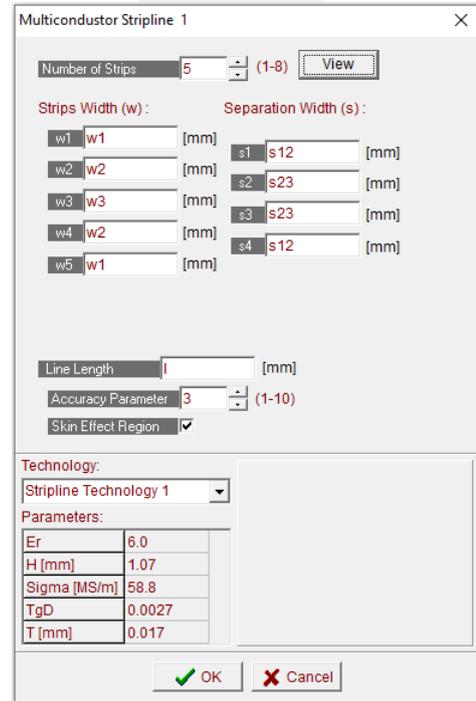


Fig. 12. A stripline multiconductor component for $N=5$ used in the filter from Fig. 10.

The analysis method from [1] includes calculation of dielectric losses and conductor losses. The latter is calculated under assumption that the skin-effect is pronounced. To calculate losses more accurately at low frequencies where the skin effect is not pronounced, the user should uncheck the corresponding *Skin Effect Region* field. In that case the method described in [2] is used to calculate conductor losses. The selection of the method of loss calculation is solely on user's discretion and should be exercised depending on the specifics of the application in hand.

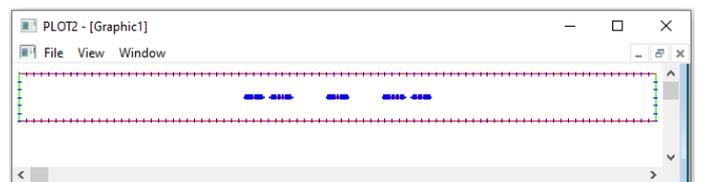


Fig. 13. Viewing a multiconductor component cross section.

The cross section of the multiconductor structure can be visualized by pressing the *View* button. Example of a cross section window is presented in Fig. 13 where the multiconductor stripline appearing in the filter from Fig. 10 is shown. The possibility to view a multiconductor line cross section can be used to check roughly if the dimensions of the lines and gaps comprising a multiconductor geometry

are correctly entered. The details of the conductor and dielectric contours segmentation also become visible which can be useful when estimating the merit of a value adopted for the *Accuracy* parameter and may therefore help to determine the accuracy – simulation time trade off.

3. Coupled Microstrip Line Synthesis

In addition to the analysis of coupled microstrip lines introduced in previous releases, which basically means calculation of electrical parameters from physical dimensions of coupled microstrip line, v6.0 release includes the synthesis of coupled microstrips i.e. calculation of physical parameters when electrical parameters are specified. The synthesis methods from [3] was implemented in *Impedance Calculator* tool (Fig. 14).

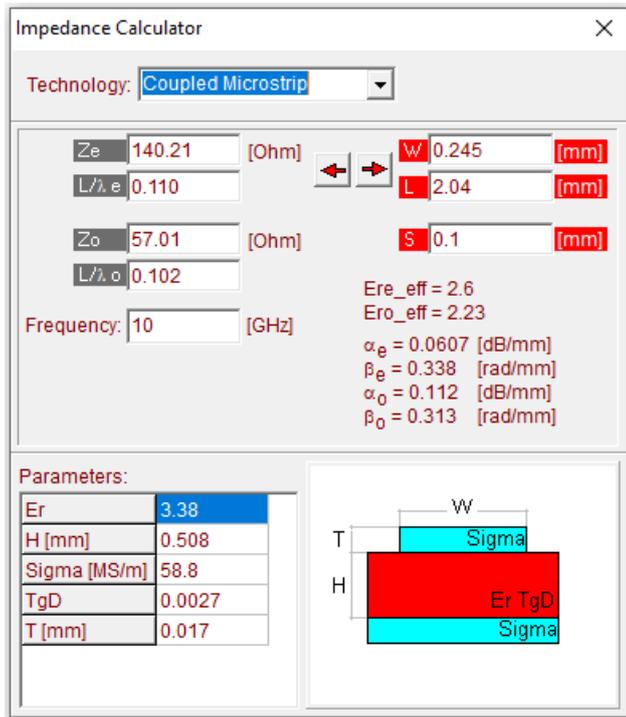


Fig. 14. Coupled microstrip in Impedance Calculator tool.

To convert electrical to physical parameters, user should first enter values for Z_e , Z_o , and either L/λ_e or L/λ_o , then click the arrow pointing to the right to obtain W , S and L . A pair of values entered for characteristic impedances determines the linewidth W and the gap between the lines S , while the electrical length is translated into the line physical length L .

User should bear in mind that this direction of calculation might fail to produce results, i.e. in the general case an arbitrary pair of impedances Z_e , Z_o may not be physically realizable.

4. Active 2-port network parameters

In addition to calculation of noise figure and stability parameters for active 2-port network introduced in previous versions, several new quantities have been added in v6.0 to expand the possibilities when working with linear active circuits. The recently added quantities are: input and output impedances required to establish conjugate match for unconditionally stable devices, maximum transducer gain, G_{tmax} and maximum stable gain, G_{msg} .

The example how new features can be used in a practical situation of amplifier design is presented in Fig. 15 - Fig. 20 where a typical design flow for simultaneously matched amplifier is illustrated. From Fig. 15, where a maximum transducer gain for a particular device is plotted, it can be concluded that for a particular transistor, *Maximum transducer gain* achievable at 14 GHz is $G_{tmax}=6.95$ dB.

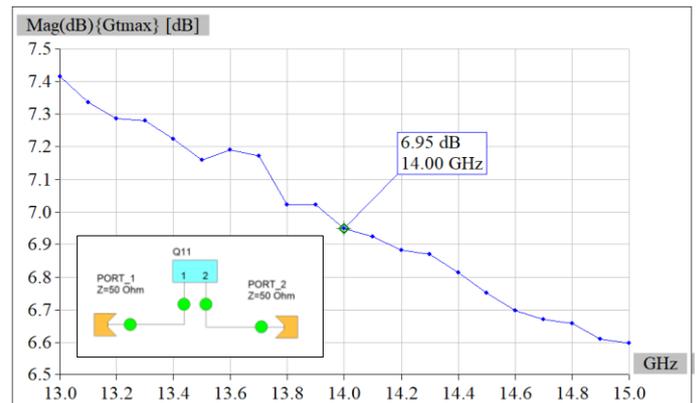


Fig. 15. Maximum transducer gain for a particular transistor.

To begin with the design of conjugately matched amplifier, reflection coefficients Γ_{aS} and Γ_{aL} providing simultaneous conjugate match for the selected transistor should be calculated. With WIPL-D Microwave, the calculations are performed in a single click and values of reflection coefficients are automatically available in graphical form in the form of the plots of real and imaginary parts vs. frequency, as shown in Fig. 16.

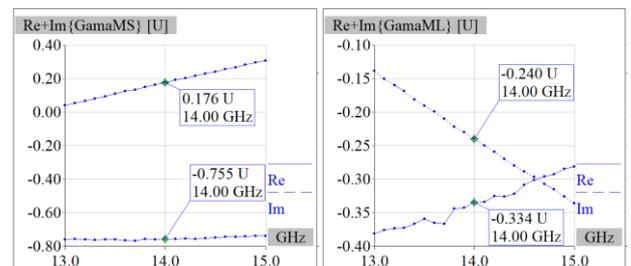


Fig. 16. Γ_{aS} and Γ_{aL} providing simultaneous conjugate match for the selected transistor.

Calculated values should then be transformed to corresponding impedances which conjugates should be matched. This is again performed easily as the graphs from the previous step are in a single click converted to Smith charts, where marker readings provide corresponding normalized impedances values, as demonstrated in Fig. 17 at the operating frequency of 14 GHz.

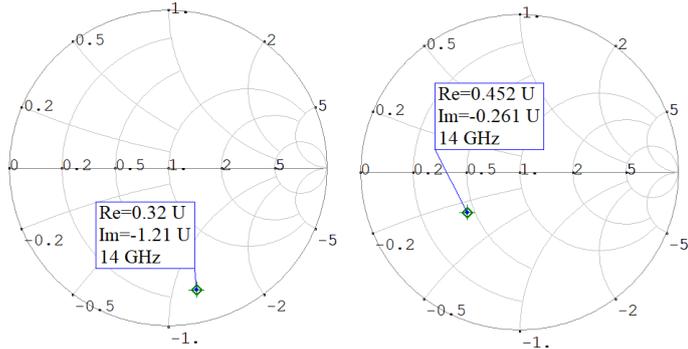


Fig. 17. Normalized impedances required for simultaneous conjugate match – input (left) and output (right).

For the sake of illustration, simple lumped-element circuits were designed to match the conjugate values of the impedances determined in the previous step to 50 Ohms. The matching circuits and corresponding traces in the Smith chart are presented in Fig. 18.

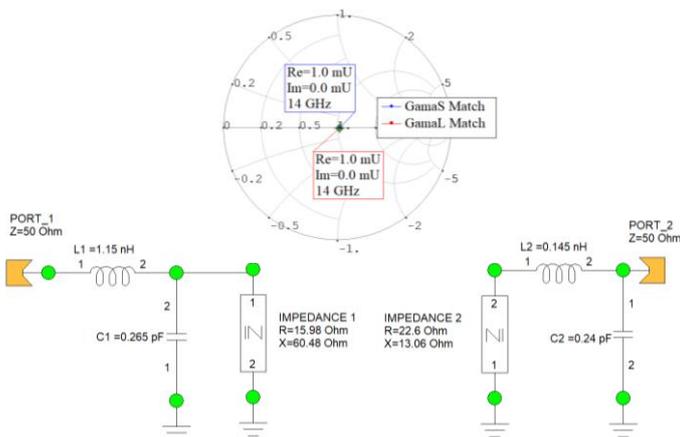


Fig.18. Matching circuits for conjugate values of the impedances determined from Fig. 17.

The amplifier resulting from the connection of the matching circuits to the transistor is presented in Fig. 19. The S-parameters of the amplifier are presented in Fig. 20. The amplifier has the gain of $G=S_{21}=6.94$ dB, which is in excellent agreement with previously determined G_{tmax} , and return loss at input and output is better than 30 dB.

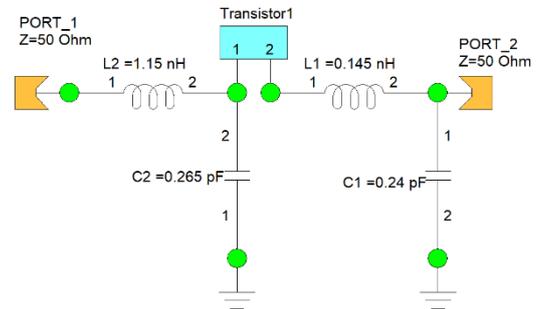


Fig. 19. Amplifier with simultaneous conjugate match.

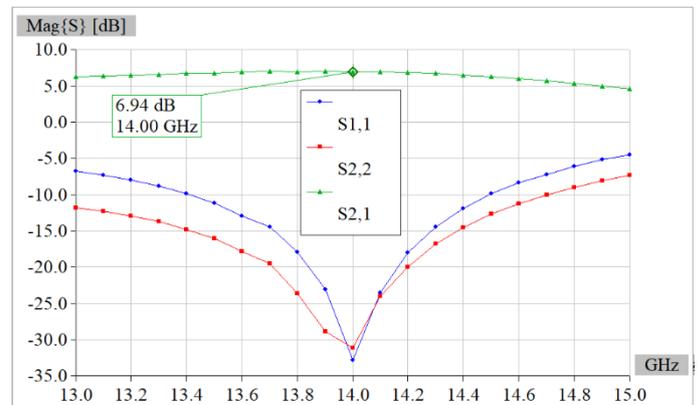


Fig. 20. S-parameters of the amplifier from Fig. 19.

5. References:

- [1] M. B. Bazdar, A. R. Djordjevic, R. F. Harrington, and T. K. Sarkar, "Evaluation of quasi-static matrix parameters for multiconductor transmission lines using Galerkin's method," *IEEE Trans. Microwave Theory Tech.*, Vol. MTT-42, No 7, July 1994, pp. 1223--1228.
- [2] T. K. Sarkar and A. R. Djordjevic, "Wideband Electromagnetic Analysis of Finite Conductivity Cylinders", *Progress in Electromagnetics Research*, PIER, Vol 16, 1997, pp. 153–173.
- [3] S. Akhtarzad, T. R. Rowbotham, and P. B. Johns, "The Design of Coupled Microstrip Lines," *IEEE Trans. Microwave Theory Tech.*, Vol. MTT-23, No 6, June 1975, pp. 486-492.