A simplified CAD approach to analyzing lumped elements

Solving a key problem in power amplifier design with modern CAD tools and approaches.

By Andrey V. Grebennikov

The parameters of power amplifier design can be a problematic issue in the designing of matching circuits.

In spite of numerous papers regarding different matching circuit configurations or their design methods, such a problem remains real and interesting.

To solve this problem successfully and in a timely fashion, it is necessary to choose the most effective matching circuit configuration to satisfy the given requirements. Ideally, it requires only fine tuning in the practical design.

ching circuit technique commonly resents both a theoretical analysis (to define the appropriate matching circuit configuration), and the ability to calculate its parameters in a practical fashion (to minimize final tuning). We will survey all of these aspects and examine the matching circuit design with lumped elements (based on the simple analytical equations to calculate the circuits elements for L–, π– and T–transformers). The main idea lies in the introduction of quality factors to evaluate bandwidth and circuit elements. In addition, plotting the quality factor's circles on Smith charts substantially simplifies the final design by making it explicit and clear. In this case, the useful principle of equal quality factors allows only one Q-circle to speed up the matching procedure for multiple section matching circuits. Such circuits generally consist of low-pass or high-pass sections. The matching principles are demonstrated by the examples of two power amplifiers: a narrow-band, 300 MHz, bipolar 10 W amplifier and a broadband, 142 to 174 MHz, 150 W, metal-oxide semiconductor field effect transistor (MOSFET) amplifier.

Basic equations

The lumped matching circuits in the form of (*a*) L–transformer, (*b*) π–transformer and (*c*) T–transformer shown in Figure 1 have proven to be most effective in their implementation of the amplifier circuit design. The simplest matching circuit is the matching circuit in the form of the L-transformer. It is convenient to analyze the transforming properties of this matching circuit by using the equivalent transformation of the parallel RX circuit into the series one.

Let R_1 and X_1 be the active and reactive parts of the complex impedance $Z_1 = (jX_1R_1/(R_1+jX_1)$ of the parallel circuit, and R_2 and X_2 be the active and reactive parts of the complex impedance $Z_2 = (jX_2R_2/(R_2+jX_2)$ of the series circuit presented in Figure 2. As a result, these two circuits are equivalent at some frequency if $Z_2 = Z_2$, i.e. when:

$$R_2 + jX_2 = \frac{R_1X_1^2}{R_1^2+X_1^2} + j\frac{R_1^2X_1}{R_1^2+X_1^2} \tag{1}$$

The solution of equation (1) can be written in the form of two following expressions:

$$x_1 = R_2(1+Q^2) \tag{2}$$

$$X_1 = X_2(1+Q^{-2}) \tag{3}$$

where $Q = |X_1|/R_1 = |X_2|/R_2$ is the quality factor equal for both the series and parallel circuit.

Consequently, if the reactive impedance $X_1 = -X_2(1+Q^{-2})$ is connected to the series circuit R_2X_2, it compensates the reactive impedance of the equivalent parallel circuit. The input impedance of the obtained two-port network

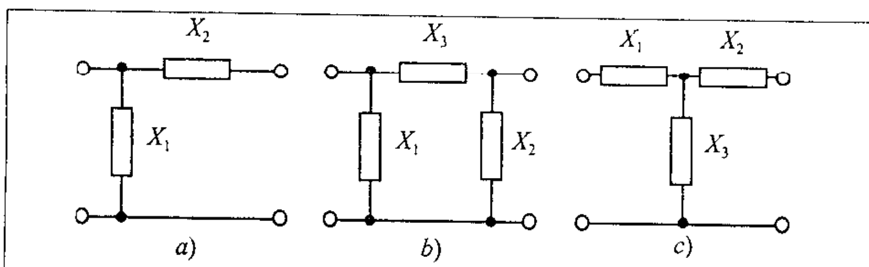


Figure 1. Lumped matching circuits of L , π and T transformers

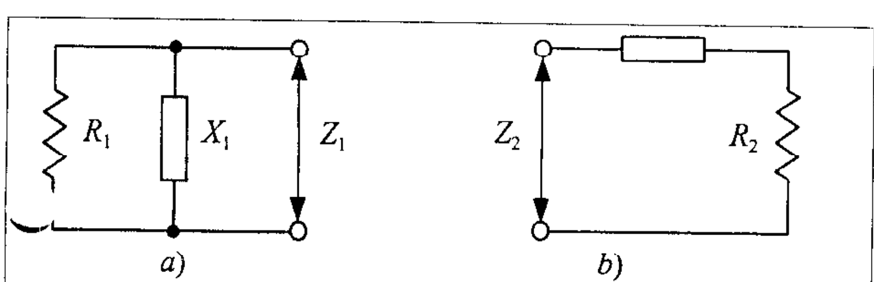


Figure 2. Impedance series equivalent circuit.

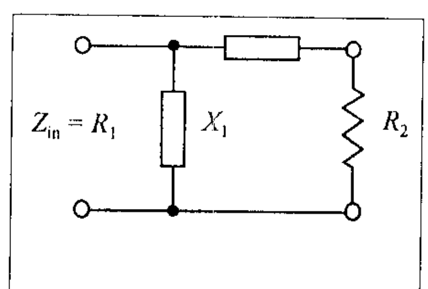
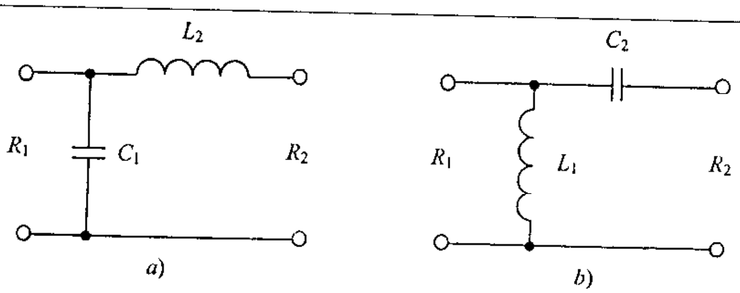


Figure 3. Input impedance of the two-port network



$$\left.\begin{aligned}\omega C_1 &= Q/R_1\\ \omega L_2 &= Q R_2\end{aligned}\right\}\quad Q=\sqrt{\frac{R_1}{R_2}-1}\quad \left\{\begin{aligned}\omega L_1 &= R_1/Q\\ \omega C_2 &= 1/(Q R_2)\end{aligned}\right.$$

Figure 4. Matching circuits and relevant equations.

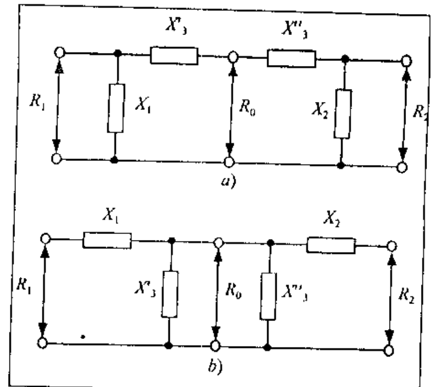


Figure 5. Matching circuits developed by connecting two L transformers.

presented in Figure 3 will only be resistive and equal to R_I. As a result, one can transform, at the given frequency, the resistance R_1 into the other resistance R_2. It is sufficient to connect a two-port L-transformer with the opposite signs of the reactive impedances X_1 and X_2 and with the following parameters:

$$|X_1| = \frac{R_1}{Q},\ |X_2| = R_2 Q,\ \text{and} \qquad (4)$$

$$Q=\sqrt{\frac{R_1}{R_2}-1}$$

between them.

Due to the opposite signs of the reactive impedances X_1 and X_2, two possible circuit configurations of the L-transformer with the same transforming properties can be realized (see Figure 4). A simple and important expression follows from equation (4), which allows direct and fast calculating of the parameters of the L-transformers. The equation is:

$$R_1 R_2 = \frac{L}{C} \qquad (5)$$

where $C = C_I$, $L = L_2$ are for the matching circuit in Figure 4 (a) and $L = L_1$, $C = C_2$ are for matching circuit in Figure 4 (b).

The matching circuits in the form of (a) the π–transformer and (b) the T–transformer can be realized by the appropriate connection of two L-transformers as is shown in Figure 5. By means of each L-transformer, the resistances R_1 and R_2 are transformed to some intermediate resistance R_0. These values are $R_0 < (R_1, R_2)$ for π–transformer and the value of $R_0 >$ (R_1, R_2) for T–transformer. Taking into account the two circuit configurations of the L-transformer presented in Figure 4, it is possible to realize the different circuit configurations of such two-port transformers where $X_3 = X_3' + X_3''$ in Figure 5 (a) and $X_3 = X_3' X_3''/(X_3' + X_3'')$ in Figure 5 (b).

Filtering properties

The matching circuits in the form of the L–transformer loaded on the resistance R_2 can also be considered as the parallel resonant circuit presented in Figure 6. The series inductance and resistance are the frequency-dependent functions in the following form:

$$R_1 = R_2' = R_2(1+Q^2),\ \text{and}$$

$$L_2' = L_2(1+Q^2) \qquad (6)$$

where $Q = \omega L_2/R_2$. As a result, the resonant frequency of this equivalent parallel resonant circuit is determined from the following equation:

$$\omega_0 = 2\pi f_0 = \sqrt{\frac{1}{L_2 C_1} - \left(\frac{R_2}{L_2}\right)^2} \qquad (7)$$

Under the small values of Q, wider frequency bandwidth (but simultaneous poor out-of-band suppression) also characterizes such a matching circuit. However, under the large values of Q, the frequency bandwidth substantially reduces. For the case of $R_1/R_2 \geq 10$, which corresponds to the condition of $Q \geq 3$, the frequency bandwidth $2\Delta f$ and out-of-band suppression factor F_n can be evaluated by the same formulas as for a parallel resonant circuit [1]:

$$2\Delta f \cong \frac{f_0}{Q},\ \text{and}\ F_n \cong Q^2(n^2-1) \qquad (8)$$

where f_0 is the operating frequency on which $|Z_1| = R_1$ and n is the harmonic number (Figure 7 shows the frequency behavior of the input impedance magnitude of the parallel resonant circuit $|Z_{in}|$).

To avoid parasitic low-frequency oscillations and to increase the level of the harmonic suppression it is necessary to connect an additional L_fC_f series circuit with the resonant frequency equal to the operating frequency of the power amplifier (see Figure 8).

Efficiency

The efficiency of such a transformer, η_T, is determined by the ratio of P_I/P_{in}. P_{in} is the power at the input of the transformer, P_L is the load transformer power. It can be readily shown that for

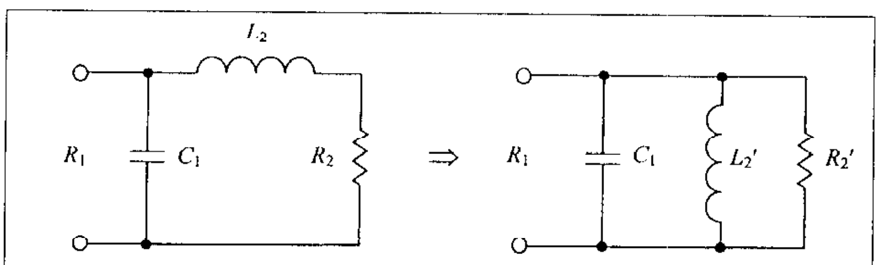


Figure 6. A parallel resonant circuit resulting from loading the L transformer on R_2.

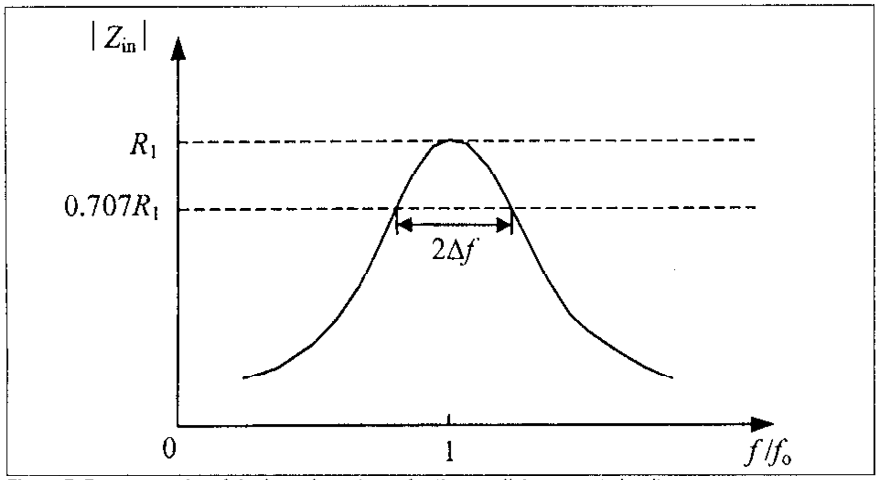


Figure 7. Frequency plot of the input impedance for the parallel resonant circuit.

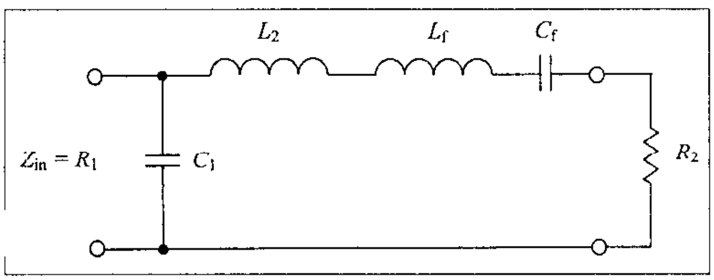


Figure 8. L_f and C_f series circuit for parasitic oscillation prevention.

an L–transformer with negligible losses in the capacitor, the efficiency is calculated from the following equation:

$$\eta_T \cong \frac{1}{\left(1 + \frac{Q}{Q_{ind}}\right)} \qquad (9)$$

where Q_{ind} is the inductor quality factor. From equation (9) it follows that with the increase of Q, the efficiency of the transformer decreases. The analysis of equations (4) and (5) shows that for the given resistances R_1 and R_2, each parameter of L–transformer can have only one meaning. As a result, it is difficult to simultaneously satisfy such contradictory requirements as efficiency, frequency bandwidth and out-of-band suppression.

It is advisable to use two-port L–transformers, in narrow-band power amplifiers, as the interstage matching circuits. This applies where the irements to the out-of-band supsion and efficiency are not so high, rather than for the matching circuits in the output stages. In this case, the main advantage of such a design is that only two elements, as well as simple tuning, can be implemented. For larger values of Q, $Q \geq 10$, it is possible to use cascaded connections of L–transformers that allow wider frequency bandwidth and transformer efficiency.

π transformers

Several of the most widespread and in-practice two-port π–transformers with the design formulas are presented in Figure 9 [2]. The π–transformers are widely used as output matching circuits of high-power amplifiers in class B operation. This is so that one can differentiate the fundamental voltage drain or collector waveform from the appropriate harmonic suppression. They are also convenient to use as interstage matching circuits in low-power and medium-power amplifiers. This works when it is necessary to provide the fundamental voltage waveforms at the drain or collector of the previous transistor and at the gate or base of the next transistor. In this case the input and output capacitances of these transistors can easily be taken into account in the matching circuit elements C_1 and C_2. Finally, use of the π–transformers can be work for high-efficiency class E operation when it is necessary to provide the appropriate capacitive output load transistor impedance.

T-transformers

Some of the matching circuit configurations of two-port T-transformers, with the analytical formulas to calculate the parameters of each transformer, are given in Figure 10 [2]. The T-transformers are usually used in high-power amplifiers both as input/interstage and output matching circuits. Choosing a high value of the inductance L_2 will provide the current waveform at the input of the transistor with a small value of the input resistance that will be close to a sinusoidal waveform.

Bipolar UHF narrow-band PA design

A lumped matching circuit technique will be illustrated through the design of a 10 W, 300 MHz, bipolar amplifier with a supply voltage of 12.5 V providing a gain of at least 10 dB.

First, select an appropriate active device. The correct device allows both simplification of the circuit design procedure by using the matching circuit with minimum elements and satisfaction of the specified requirements. For example, the above-mentioned requirements can be provided with n–p–n silicon transistors intended for transmission applications in class A, B, or C, VHF and UHF ranges.

The manufacturer usually states the values of the complex input and output impedances or admittance at the nominal operation point on the data sheet for the device. At the operating frequency of 300 MHz, $Z_{in} = (1.3 + j0.9)\ \Omega$ and $Y_{out} = (150 - j70)$ mS. In this case, Z_{in} is expressed as a series combination of input transistor resistance and inductive reactance whereas Y_{out} is represented by a parallel combination of output resistance and inductive reactance. This means that for a chosen operating frequency, the influence of the series parasitic collector lead inductance prevails over the influence of the parallel collector capacitance. The sum of these gives a total inductive reactance of the equivalent output circuit for the active device. To match series input inductive impedance to the standard 50 Ω input source impedance, it is advisable to use a matching circuit in the form of a T-transformer (see figure 10–b). As a result, the complete input network, including input device impedance and a matching cir-

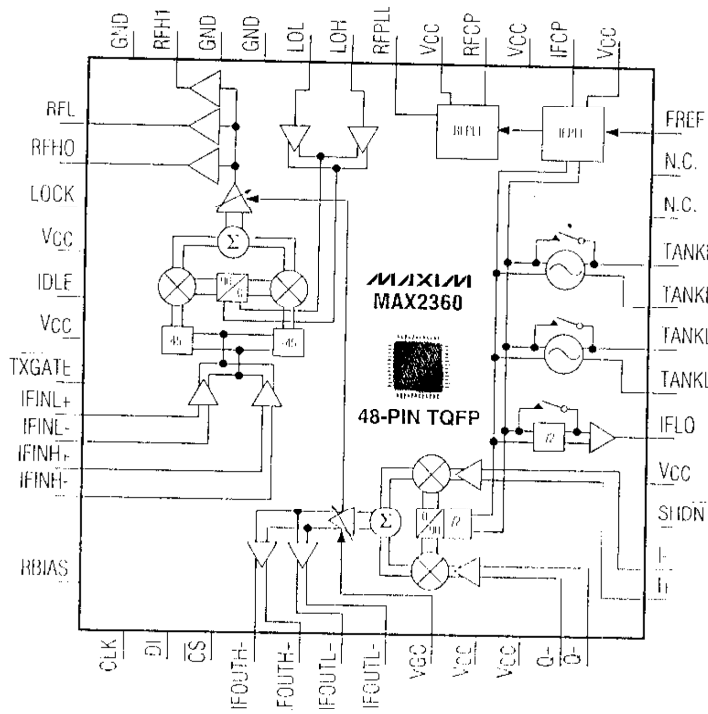
WORLD'S MOST INTEGRATED TRANSMIT IC FOR DUAL-BAND CELLULAR PHONES—SAMPLING NOW

The New MAX2360 is Designed for cdmaOne™, cdma2000™, TDMA, EDGE, and W-CDMA Dual-Band Cellular Phones

The MAX2360 is part of Maxim's cdmaOne chipset and, together with MAX2320, MAX2310, and MAX2264, forms a complete cdmaOne antenna-to-baseband solution.

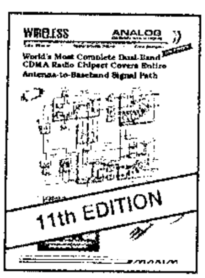
Features:

- Dual-Band I/Q Modulator
- Upconverters
- +7dBm (-54dBc ACPR) PA Drivers
- IF VCOs
- Dual-Band IF/RF PLLs
- Serial Interface Control
- Saves up to Three SAW RF Filters and PA Drivers
- Supports Dual IFs
- PCS Only (MAX2362) and Cellular Only (MAX2364) Versions Available



The highly integrated MAX2360 saves cost and size in dual-band cellular phones.

cdmaOne and cdma2000 are trademarks of the CDMA Development Group.



FREE Wireless Design Guide—*Sent Within 24 Hours!*
Includes: Reply Cards for Free Samples and Data Sheets

CALL TOLL-FREE 1-800-998-8800 for a Design Guide or Free Sample
6:00 a.m. – 6:00 p.m. Pacific Time

MAXIM

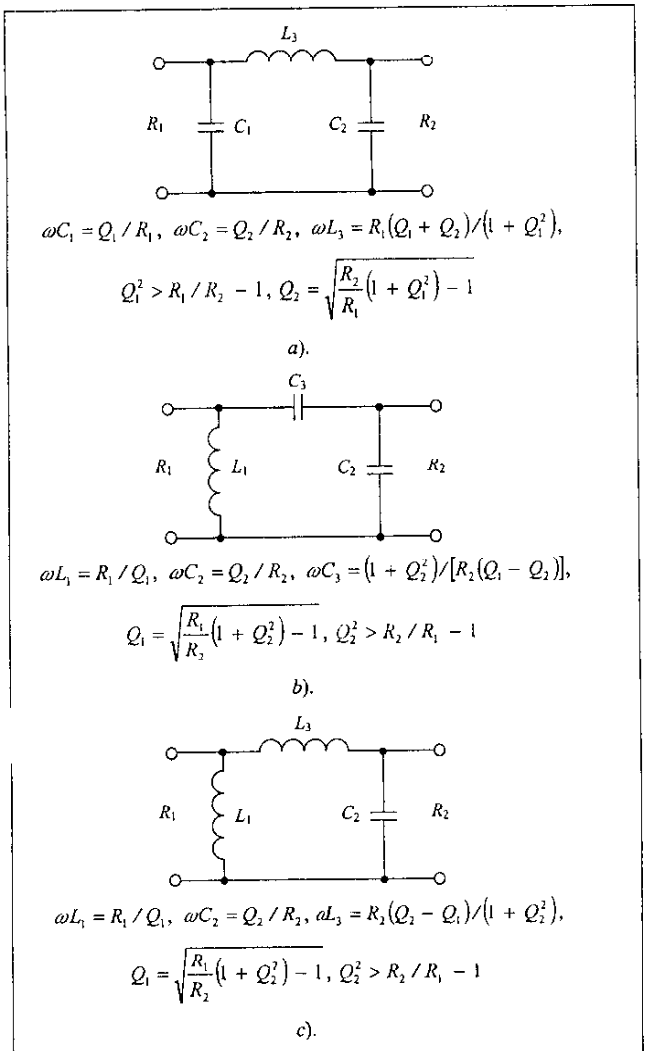
www.maxim-ic.com

2000 EDITION!
FREE FULL-LINE DATA CATALOG
ON CD-ROM

NEW! Get Price, Delivery, and Place Orders Online at www.maxim-ic.com

Maxim Integrated Products, 120 San Gabriel Drive, Sunnyvale, CA 94086, (408) 737-7600, FAX (408) 737-7194.
Distributed by Maxim Distribution, Arrow, Avnet Electronics Marketing, CAM RPC, Digi-Key, Elmo, Nu Horizons, and Zeus.
Distributed in Canada by Arrow and Avnet Electronics Marketing

INFO/CARD 13



$$\omega C_1 = Q_1/R_1,\ \omega C_2 = Q_2/R_2,\ \omega L_3 = R_1(Q_1+Q_2)/(1+Q_1^2),$$
$$Q_1^2 > R_1/R_2 - 1,\ Q_2 = \sqrt{\frac{R_2}{R_1}(1+Q_1^2) - 1}$$

a).

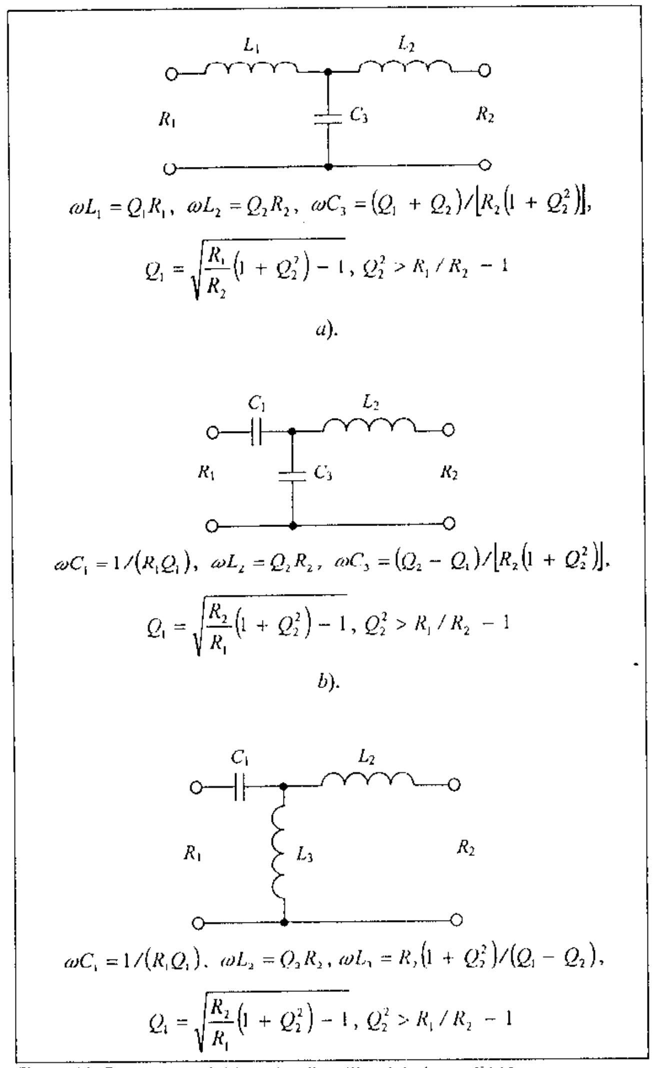
$$\omega L_1 = R_1/Q_1,\ \omega C_2 = Q_2/R_2,\ \omega C_3 = (1+Q_2^2)/[R_2(Q_1 - Q_2)],$$
$$Q_1 = \sqrt{\frac{R_1}{R_2}(1+Q_2^2) - 1},\ Q_2^2 > R_2/R_1 - 1$$

b).

$$\omega L_1 = R_1/Q_1,\ \omega C_2 = Q_2/R_2,\ \omega L_3 = R_2(Q_2 - Q_1)/(1+Q_2^2),$$
$$Q_1 = \sqrt{\frac{R_1}{R_2}(1+Q_2^2) - 1},\ Q_2^2 > R_2/R_1 - 1$$

c).

Figure 9. Common two-port networks and related equations.



$$\omega L_1 = Q_1 R_1,\ \omega L_2 = Q_2 R_2,\ \omega C_3 = (Q_1 + Q_2)/[R_2(1+Q_2^2)],$$
$$Q_1 = \sqrt{\frac{R_1}{R_2}(1+Q_2^2) - 1},\ Q_2^2 > R_1/R_2 - 1$$

a).

$$\omega C_1 = 1/(R_1 Q_1),\ \omega L_2 = Q_2 R_2,\ \omega C_3 = (Q_2 - Q_1)/[R_2(1+Q_2^2)].$$
$$Q_1 = \sqrt{\frac{R_2}{R_1}(1+Q_2^2) - 1},\ Q_2^2 > R_1/R_2 - 1$$

b).

$$\omega C_1 = 1/(R_1 Q_1),\ \omega L_2 = Q_2 R_2,\ \omega L_3 = R_2(1+Q_2^2)/(Q_1 - Q_2),$$
$$Q_1 = \sqrt{\frac{R_2}{R_1}(1+Q_2^2) - 1},\ Q_2^2 > R_1/R_2 - 1$$

Figure 10. Common matching circuits with related equations.

cuit, is presented in Figure 11. At the operating frequency of 300 MHz input inductance will be approximately 0.5 nH.

First, define a value of the quality factor Q_2. This is necessary in order to calculate the parameters of the matching circuit and calculated as:

$$Q2 > \sqrt{\frac{R_1}{R_{in}} - 1} = 6.1$$

Consequently, a value of Q_2 must be larger than 6.1. In this example we oose a value of $Q_2 = 6.5$. This allows alizing the 3 dB level bandwidth by the following: 300 MHz/6.5 = 46 MHz. As a result, the value of Q_1 will be equal to 0.35 and the values of the parameters of input matching circuit are as follows:

$$C_1 = \frac{1}{\omega Q_1 R_1} = 30\ \text{pf},$$

$$L_1 + L_{in} = \frac{Q_2 R_{in}}{\omega} = 4.5\ \text{nH} \Rightarrow L_1 = 4.0\ \text{nH}$$

and

$$C_2 = \frac{Q_2 - Q_1}{\omega R_{in}(1+Q_2^2)} = 59\ \text{pF}$$

It should be noted that such a *T*–transformer is widespread in practical matching circuit design. This is due to their simplicity and convenience in tuning.

Furthermore, a small value of series capacitance, C_1, contributes to eliminating the low frequency parasitic oscillations in the case of a multistage power amplifier.

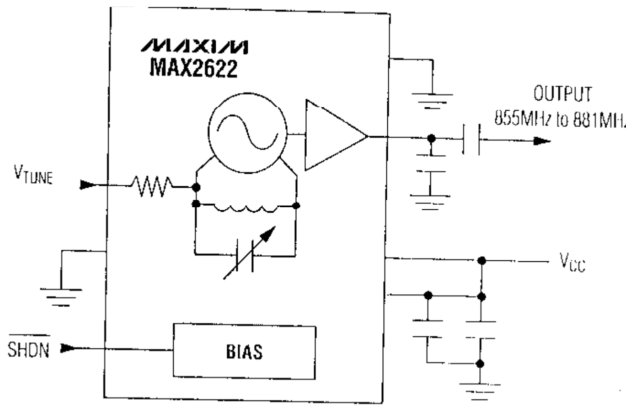
The function of each element for visual effect can be traced on the Smith chart in Figure 12. The easiest and most convenient way to initially plot the traces of the matching circuit elements is to initially plot the traces of Q_1 and Q_2. Then the trace of inductance L_1 must be plotted until it intersects with the Q_2 circle. The trace of capacitance C_2 should be plotted until it intersects with the Q_1 circle.

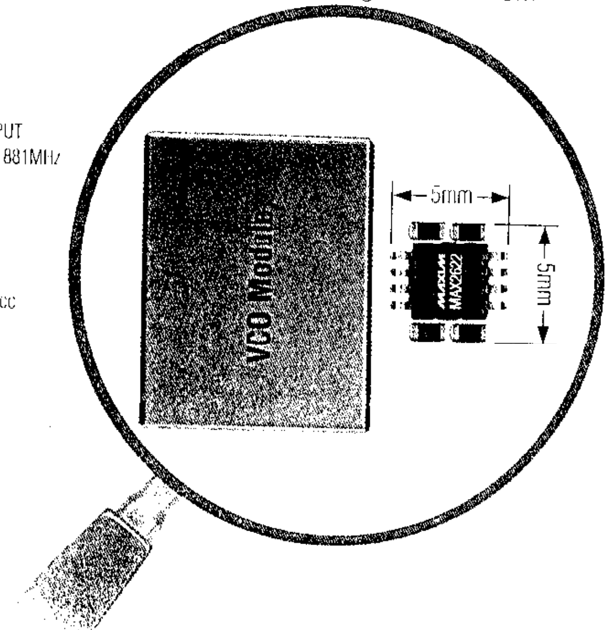
A similar design philosophy can be applied to the design of the output matching circuit shown in Figure 13. However, taking into account the presence of parallel output inductance, it is advisable to use a matching circuit in the form of the π–transformer shown

FULLY MONOLITHIC 900MHz SiGe VCO IN 8-PIN µMAX

Replace Modules in 1/3 the Space, for 2/3 the Cost

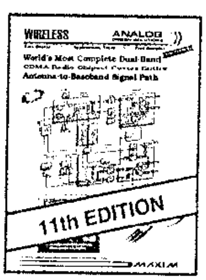
The MAX2622/MAX2623/MAX2624 combine the advantages of VCO modules, such as fast design time and guaranteed performance, with the low cost of discrete VCO designs. In addition, they offer a pin-selectable shutdown mode. All this comes in a tiny 8-pin µMAX package that occupies less than one-third the board space of modules or discrete designs, making the MAX2622/MAX2623/MAX2624 an attractive solution for your next-generation ISM-band radio or cordless phone.





- **On-Chip Tank Circuit**
- **2.7V to 3.6V Operation**
- **-101dBc/Hz Phase Noise (100kHz)**
- **320kHz/V Pushing**
- **Guaranteed Performance**
- **1µA Shutdown I_{SUPPLY}**

PART	VCO FREQUENCY (MHz)	APPLICATIONS
MAX2622	855 to 881	European ISM Band
MAX2623	885 to 950	US ISM Band, DECT
MAX2624	947 to 998	US ISM Band



FREE* Wireless Design Guide–*Sent Within 24 Hours!
Includes: Reply Cards for Free Samples and Data Sheets

CALL TOLL-FREE 1-800-998-8800 for a Design Guide or Free Sample
6:00 a.m. – 6:00 p.m. Pacific Time

MAXIM

www.maxim-ic.com

2000 EDITION!
FREE FULL-LINE DATA CATALOG
ON CD-ROM

NEW! Get Price, Delivery, and Place Orders Online at www.maxim-ic.com

Maxim Integrated Products, 120 San Gabriel Drive, Sunnyvale, CA 94086, (408) 737-7600, FAX (408) 737-7194.
Distributed by Maxim Distribution, Arrow, Avnet Electronics Marketing, CAM RPC, Digi-Key, Elmo, Nu Horizons, and Zeus.
Distributed in Canada by Arrow and Avnet Electronics Marketing.

INFO/CARD 14 or visit **www.rfdesign.com**

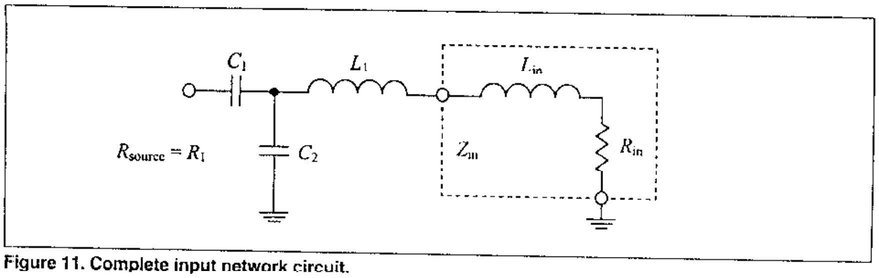


Figure 11. Complete input network circuit.

in Figure 9 (c). The output resistance, collector capacitance and lead inductance, (the influence of which becomes significant at the higher frequencies) can represent device output impedance in a common case. The output resistance can be analytically evaluated by:

$$R_{out} = \frac{[V_{cc} - V_{ce(sat)}]^2}{2P_{out}} \cong \frac{(0.9V_{cc})^2}{2P_{out}} \cong 6.3\ \Omega$$

where V_{cc} is the supply voltage, $V_{ce(sat)}$ is the saturation voltage, P_{out} is the output power, and X (substantiated approximately from the measurement results: R_{out} = 1/0.15 = 6.7 Ω, a value of L_{out}) is approximately equal to 7.6 nH.

The value of the quality factor Q_2 that is necessary to calculate the parameters of the matching circuit is as follows:

$$Q_2 > \sqrt{\frac{R_2}{R_{out}} - 1} = 2.5$$

At the same time, the value of the quality factor Q1 of the device's output circuit is given by:

$$Q_1 = \frac{R_{out}}{\omega L_{out}} = 0.47$$

It is necessary to make sure that a value of L_{out} allows matching to a 50 Ω load impedance by the following:

$$Q_2 = \sqrt{\frac{R_2}{R_{out}}(1 + Q_1^2) - 1} = 2.8 > 2.5$$

As a result, the values of the other two output matching circuit elements will be as follows:

$$C_3 = \frac{Q_2}{\omega R_2} = 31\ \text{pF, and}$$

$$L_2 = \frac{R_2(Q_2 - Q_1)}{\omega(1 + Q_2^2)} = 6.8\ \text{nH}$$

A stopping capacitor that performs DC supply decoupling can be connected

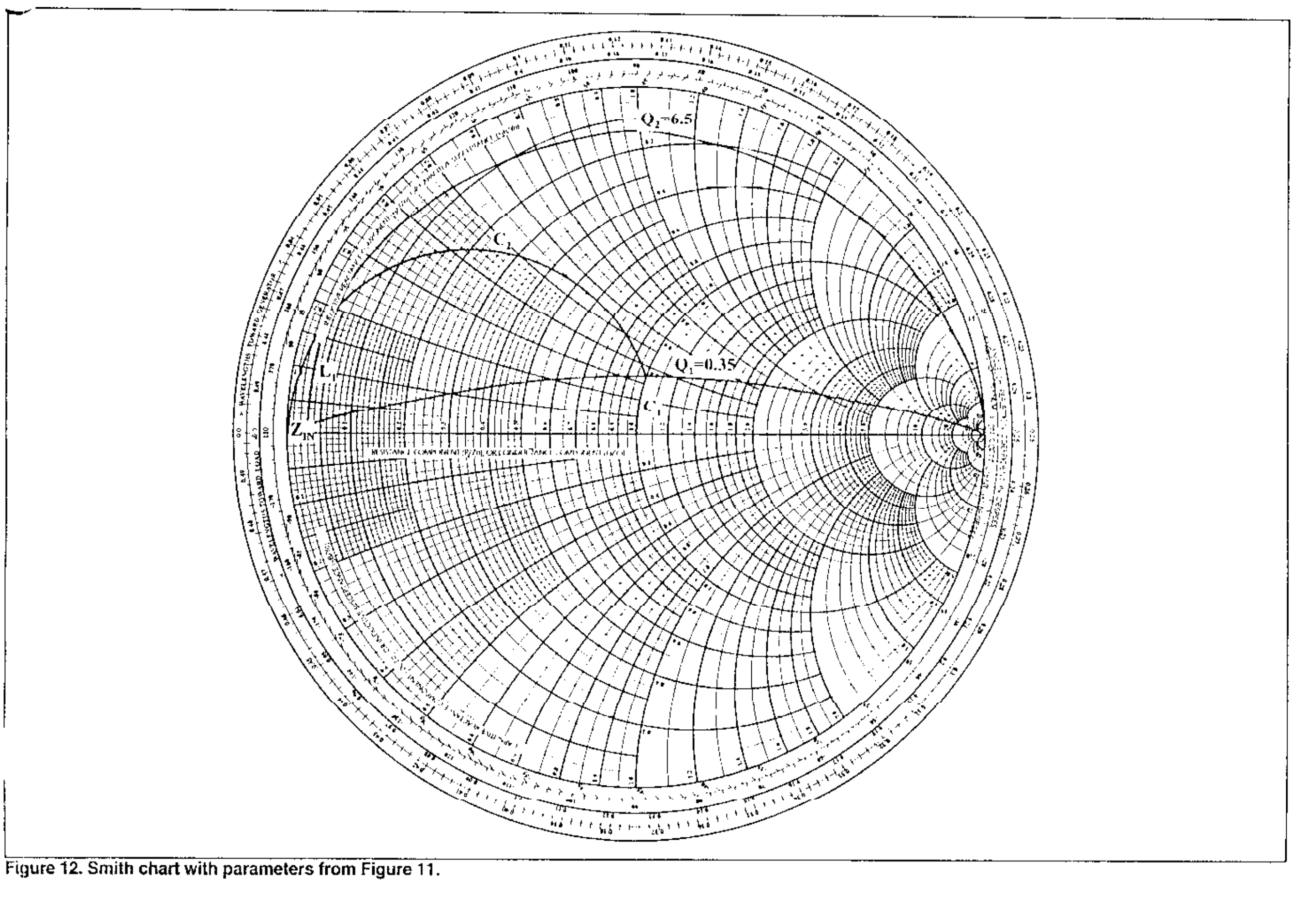


Figure 12. Smith chart with parameters from Figure 11.

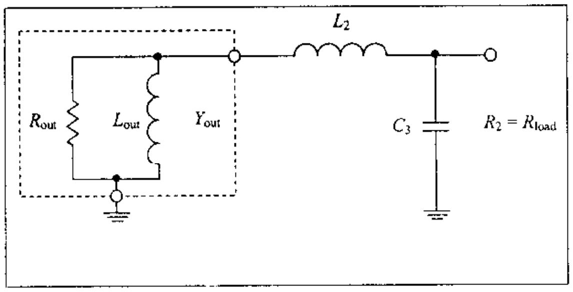


Figure 13. Alternative output matching circuit.

after a T-transformer, with a sufficiently high value of its capacitance, to avoid influencing the matching circuit design process. Alternatively, it can be used in series with the inductance L_2 to form a series-resonant circuit as shown in Figure 8. The Smith chart of Figure 14 presents the output circuit design.

It is necessary to transform the output admittance Y_{out} to the output impedance Z_{out}. The trace of inductance L_2 must be plotted as far as its intersecting point with the Q_2 circle.

MOSFET VHF broadband high power amplifier design

For broadband amplifiers, a lumped matching circuit technique will be demonstrated using a 150 W amplifier with V_s of 50 V, an operating bandwidth of 132 to 174 MHz, and a gain of more than 10 dB.

These requirements can be realized by using a high power, vertically-diffused metal oxide semiconductors (VDMOS) transistor designed for large signal amplifier applications in the VHF band. The center bandwidth frequency f_c is defined as: $f_c = \sqrt{132 \bullet 174} = 152$ MHz.

For this operating frequency the manufacturer states the following values of the complex input and output impedances: $Z_{in} = (0.9 - j1.2)\ \Omega$ and $Z_{out} = (1.8 + j2.1)\ \Omega$. In this case both Z_{in} and Z_{out} are expressed as a series combination of input or output resistance and capacitive or inductive reactance.

To achieve the required bandwidth, low-Q matching circuits should be used, which provide for the reduction of in-band amplitude ripple and improving input voltage standing wave ratio (VSWR). The value of a quality factor for 3 dB level bandwidth must be less than $Q = 152/(174 - 132) = 3.6$. As a result, it is convenient to design input and output matching circuits by using the simple L-transformers in the form of low-pass and high-pass filter sections, with a constant value of Q.

To match the series input capacitive impedance to the standard 50 Ω input source impedance, it is necessary to use three filter sections (see Figure 15). At the operating frequency of 152 MHz the input capacitance needed is approximately 873 pF. At the center frequency, to compensate for this capacitance, connect, in series, an inductance of 1.3 nH.

To simplify the matching design procedure it is advisable to cascade L-transformers with a constant value of Q. Although the requirement of a con-

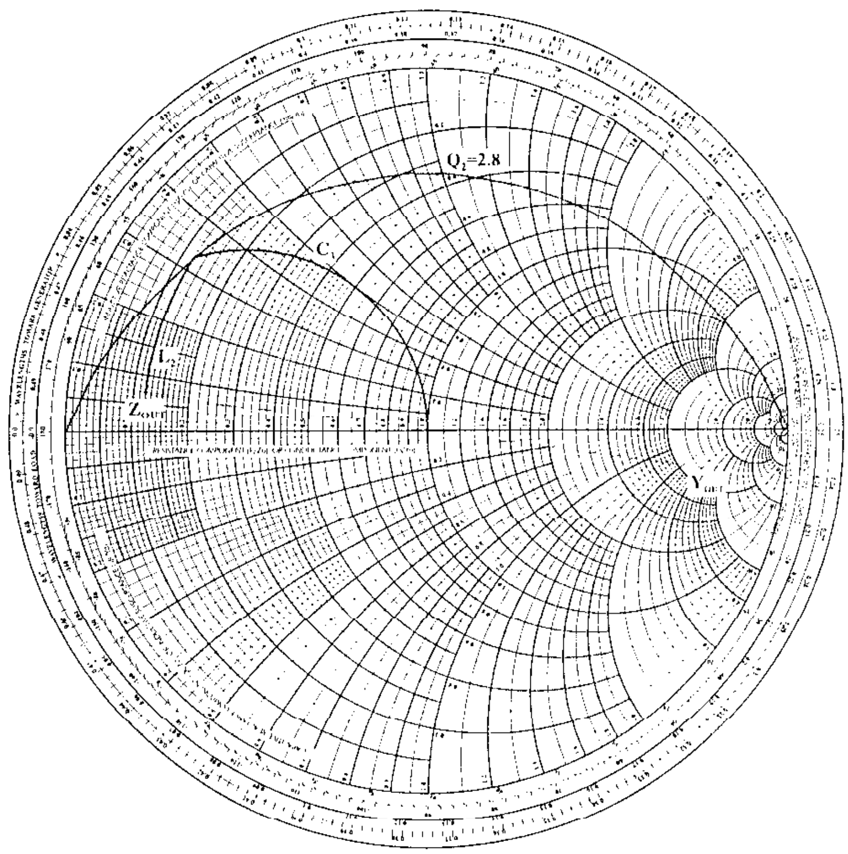


Figure 14. Smith chart supporting the circuit of Figure 13.

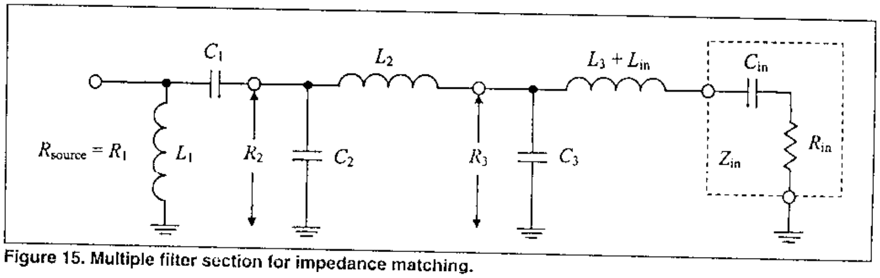


Figure 15. Multiple filter section for impedance matching.

stant Q is not strictly necessary, this provides a convenient guide for both analytical calculation of matching circuit parameters and Smith chart graphic design.

In this case, for the input matching circuit, the following ratio can be written as:

$$\frac{R_1}{R_2} = \frac{R_2}{R_3} = \frac{R_3}{R_{in}}$$

which gives the values of R_2 = 13 Ω and R_3 = 3.5 Ω for Resource = R_1 = 50 Ω and R_{in} = 0.9 Ω. Consequently, the quality factor of each L-transformer is equal to a value of Q = 1.7. The values of the elements of the input matching circuit using the formulas given in Figure 4 are calculated as follows:

L_1 = 31 nH, C_1 = 47 pF, L_2 = 6.2 nH, C_2 = 137 pF, L_3 = 1.6 nH, C_3 = 509 pF.

The case of a constant Q simplifies a matching circuit design using a standard Smith chart. So, after calculating a value of Q, it is necessary to plot a constant Q circle on the Smith chart. Next, each element of input matching circuit can be readily determined (see Figure 16).

To match the series output inductive impedance to the standard 50 Ω load impedance, use two filter sections (see Figure 17). At the operating frequency of 152 MHz, the output inductance will be around 2.2 nH. This inductance can be used as a part of L-transformer in the form of a low pass filter section. In this case, the condition of equal Q gives the following ratio for an output matching circuit :

$$\frac{R_2}{R_1} = \frac{R_1}{R_{out}} ,$$

and using the following values: R_1 =

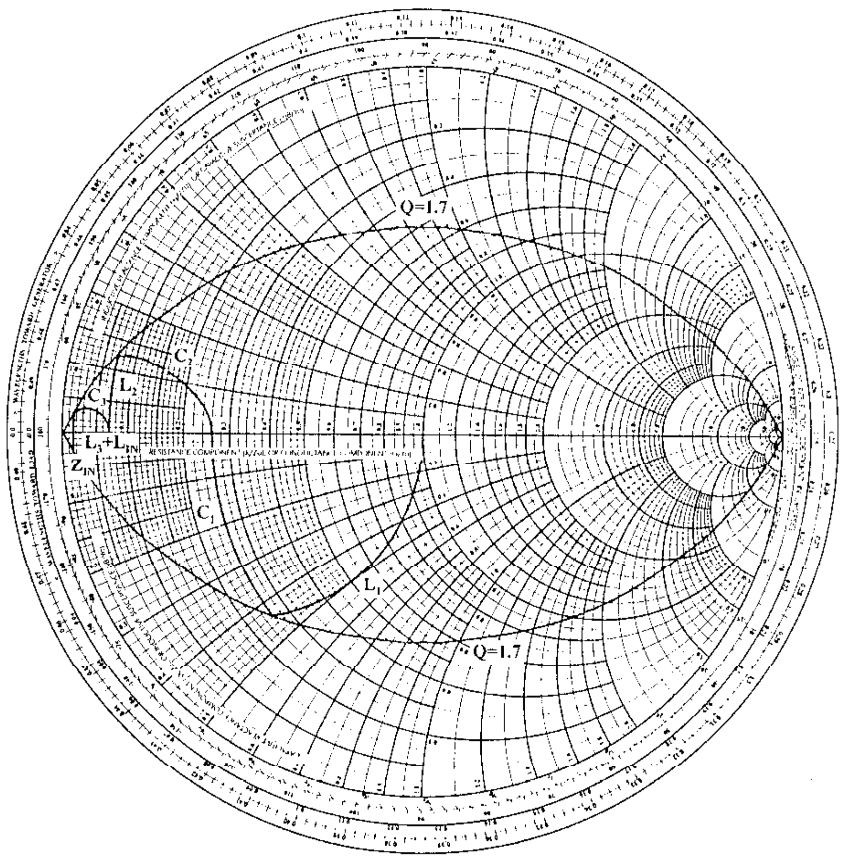


Figure 16. Smith chart detailing behavior of the circuit of Figure 15.

9.5 Ω, $R_{load} = R_2 = 50$ Ω and $R_{out} = 1.8$ Ω. Consequently, a quality factor of each L–transformer is equal to $Q = 2.1$, which is substantially smaller than a value of Q for 3 dB level bandwidth. Now it is necessary to check the value of the series inductance of the low pass section. This value must exceed 2.2 nH for correct matching procedure. The appropriate calculation gives the value of total series inductance $L_4 + L_{out}$ as approximately 4 nH. As a result, the values of the elements of the output matching circuit are as follows:

$L_4 = 1.8$ nH, $C_4 = 231$ pF, $C_5 = 52$ pF, $L_5 = 25$ nH.

The output matching circuit design using a standard Smith chart with a constant Q-circle is shown in Figure 18.

Conclusion

The matching circuit design technique with lumped elements is based on simple analytical equations that calculate the circuits elements for L-, π- and T-transformers.

This technique presents a simple approach to designing matching circuits, successfully and efficiently, by using the most effective matching circuit configuration and requiring minimal fine-tuning in practical design.

Plotting of the quality factor circles on a Smith chart simplifies the final design by making it explicit and clear. Additionally, it was proposed a useful principle of quality factors that allows limiting to one Q-circle. The procedure also quickens the matching procedure for multiple-section matching circuits, which consist of low pass or high pass sections.

The matching principles were demonstrated using narrow-band UHF bipolar power amplifier and broadband MOSFET high power amplifier, both analytically and by Smith chart. *RF*

References

1. L. F. Gray, and R. Graham, *Radio Transmitters*. McGraw-Hill, New York, 1961.

2. V. M. Bogachev, and V. V. Nikiforov, *Transistor Power Amplifiers*. Energiya, Moskva, 1978.

About the author

Andrey Viktorovich Grebennikov, a native of Uzhgorod, USSR received his Dipl-Ing degree from the Moscow Institute of Physics and Technology and his PhD degree from the Moscow Technical .University of Communications and Informatics. He has done work and research in the design and development of power FM broadcasting, as well as with VHF/UHF television transmitters, linear and high-efficiency power amplifiers, single-frequency and voltage-controlled amplifiers, oscillators and MOSFET modeling. He is now with the Institute of Microelectronics in Singapore. He can be reached at: 11 Science Park Road, Singapore Science Park II, Singapore, 117685. e-mail; *andrei@ime.org.sg*.