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Design of Matching Networks for Low Noise Preamplifiers

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This paper discusses matching networks that minimize inductive coupling between the antennas within an array while simultaneously insuring minimum noise contributions from preamplifiers. Typical low noise preamplifier designs require a strong mismatch between the source impedance and the amplifier input impedance (reflection coefficient close to one) to achieve optimal noise performance. This is in contrast to the familiar impedance match known from communication theory where input and source impedances have complex conjugate values for maximizing the power transfer from source to amplifier. The high input reflection coefficient of low noise amplifiers can be exploited to reduce antenna currents by using lossless impedance transformations to create a high impedance at the coil terminals while simultaneously maintaining a low noise figure for the amplifier. The networks presented here constitute an improvement over previous work because they give additional freedoms regarding the values of the network components and the amplifier input impedance. The technique has been formalized and coded in MathCad™, making the design of realizable networks a simple process.

Key words: matching networks; low noise amplifiers; noise matching; surface coils.

INTRODUCTION

Since the mid-1980s, it has been recognized that by using arrays of mutually decoupled surface coils one can simultaneously acquire multiple images (1). It has also been shown that these images can be combined for an improved signal-to-noise ratio (SNR) if the noise in the individual images is largely uncorrelated (2). This can be achieved to a high degree by reducing the coupling between the individual coils in the array. If one assumes that the main source of coupling between adjacent antennas is due to a mutual inductance, one way of reducing the coupling is by choosing a geometric arrangement that minimizes the net magnetic flux from one coil through another. In many applications this is not possible, and additional means of decoupling are required. One option, which is described in this article, is to reduce the antenna currents that cause the inductive coupling. Roemer *et al.* have presented a solution for a single stage matching network, given that the amplifier input impedance is close to a short circuit (3). Unfortunately, there are limitations to a practical implementation of this network. For cases in which the load resistance of the coil is very small, the values of the components needed will be difficult to realize. A second restriction is that for most applications a phase shifter is needed to rotate the reflection coefficient of the amplifier input into the vicinity of the required short circuit. The designs presented here solve these problems by adding a second stage to the network (4). It is then possible to vary the component values over a large range instead of being limited to a unique solution as in the case of the single stage network. Impractical component values such as extremely small or large capacitances or inductances can thus be avoided. Also, the phase shifter can be omitted by generalizing the problem to cases where the amplifier input impedance is an arbitrary reactance rather than a short circuit. To demonstrate the feasibility of this more general approach, a network was built and tested for use with 4.5 cm diameter coils; the network assumed an amplifier reflection coefficient at noise match of 0.99. The results presented here show that for this particular configuration, the coils can be noise-matched to the amplifiers while simultaneously reducing the inductive coupling between them by approximately 30 dB compared with a power match.

METHODS

Overview

Figure 1 shows the general concept of a combined matching and decoupling network. The amplifier is assumed to be calibrated so that its optimal noise performance is achieved when connected to a 50 Ω source. Typically, the impedance looking into the amplifier input is not 50 Ω; this difference results in a high reflection coefficient. A matching network that is essentially a λ/4 transformer transforms the resistance R_L for the loaded and resonated coil into the required 50 Ω source impedance. The phase shifter that follows this matching network will not change the 50 Ω impedance seen by the amplifier but only shift the signal phase. In the reverse direction, the same phase shifter transforms the amplifier input impedance into a resistance R_{short}, with a value close to a short circuit. Finally, the λ/4 section in the matching network will transform R_{short} into a large resistance R_{open} at the coil terminals. Note that changing the amount of phase added by the phase shifter will not affect the coil impedance as it is seen by the preamplifier but it will change the amplifier impedance as it is seen at the coil terminals. Therefore, matching the coil to the preamplifier and transforming the amplifier impedance to the coil terminals can be done simultaneously.

Figure 2 shows a typical single stage matching and decoupling network as suggested in (3). In this particular case, the amplifier input impedance already equals R_{short}. For the purpose of analysis, the three network

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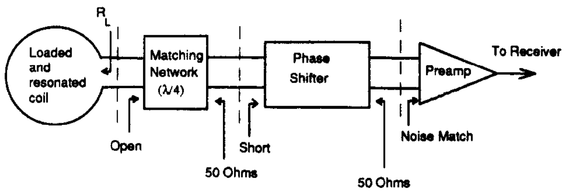

FIG. 1. Block diagram of a general impedance matching and decoupling network.

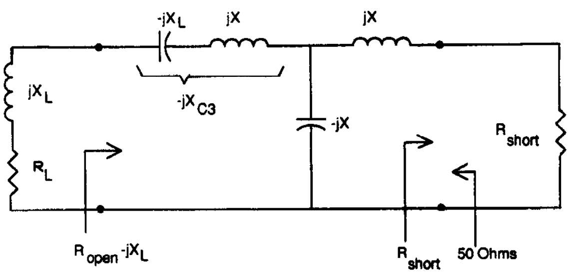

FIG. 2. Single stage matching network for decoupling coils as suggested in (3). Although the network provides a high impedance at the coil terminals, the component values can be unrealizable or cause high losses.

components can be separated into a series capacitor that resonates the coil and a discrete T-shape $\lambda/4$ transformer with characteristic impedance X that will transform R_L into $R_0 = 50\ \Omega$. The values for the three components of the network in Fig. 2 are uniquely defined through the choice for R_0 and the coil impedance under load as can be seen by Eqs. [1] and [2].

$$X = \sqrt{R_L \cdot R_0} \qquad [1]$$

$$X_{C_3} = X_L - X \qquad [2]$$

The impedance seen at the coil terminals will be

$$R_{\text{open}} = X^2/R_{\text{short}} \qquad [3]$$

For given R_0, the parameter X depends solely on the load resistance of the coil as can be seen in Eq. [2]. Small values of R_L will therefore result in small values for X and, in extreme cases, the inductor that has a reactance of value X will not be realizable. For example, a coil with a load resistance of 0.2 Ω at 41 MHz will require a matching network which includes a 12 nH inductance. This is the inductance of a short (approximately 2 cm) piece of wire. Also note that the matching network requires an additional phase shifter for the general case in which the input impedance of the preamplifier has to be transformed into a short circuit.

To improve the situation, a new matching network with an additional degree of freedom was designed. This network has the advantage that its components can be varied over a large range of possible values, and the additional phase shifter can be omitted. The Appendix gives an overview of design equations for four possible solution structures. In all four structures, only one of the four reactances needed for the networks is an inductor. This is desirable because it is much easier to tune capacitors over a large

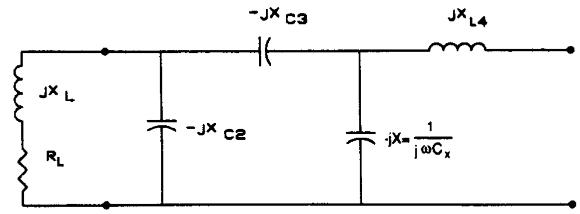

FIG. 3. Schematic of the improved impedance matching and decoupling network A-1. In this design, the parameter X can be freely chosen to obtain practical values for the remaining network components.

range of values. Additionally, all four structures have one degree of freedom that can be used to vary the values of all network components; this results in a multitude of possible solutions. It should be noted that, depending on the amplifier input impedance, one or the other structure will be more suitable. In particular, if negative values for inductors or capacitors are encountered the free parameter can be varied, or a different structure selected, to ensure that only one inductor is needed. One structure (A-1) will be discussed in detail in this paper. This particular network works well with amplifiers having an input impedance that is closer to a short circuit rather than an open circuit.

Discussion of Network A-1

Consider the general matching network of Fig. 3, which can be generated by adding the capacitors C_2 to the network of Fig. 2 and modifying the values of the other components. One of the component values (in this case, X) can be chosen as a free parameter, whereas the values of all other components are functions of this parameter.

Furthermore, the phase shifting network can be incorporated into the matching network. This is done by assuming that the amplifier is noise matched for an ideal reflection coefficient of magnitude 1 and phase ϕ so that the amplifier input impedance is a reactance with value

$$X_{\text{Amp}} = Z_0 \frac{\sin\phi}{1 - \cos\phi} \Leftrightarrow \Gamma_{\text{Amp}} = 1e^{j\phi} \qquad [4]$$

This is a reasonable assumption because usually the reflection coefficient will be close to unity. If the reflection coefficient is lower than unity, the network components can still be calculated by using this assumption. The amplifier input impedance will be transformed into the maximum possible value (R_{open}-jX_L) at the coil terminals, where R_{open} is

$$R_{\text{open}} = R_L \frac{1 + |\Gamma_{\text{amp}}|}{1 - |\Gamma_{\text{amp}}|} \qquad [5]$$

In order to have noise match at the amplifier terminals and an open circuit at the coil terminals, the following equations have to be satisfied

$$X_{C_2} = \frac{XA}{XZ_0 - B} \qquad [6]$$

$$X_{C_3} = X\frac{A - B}{B} \qquad [7]$$

$$X_{L_4} = X\frac{A - B}{A} + Z_0\frac{C}{A} \quad [8]$$

With A, B, and C given as

$$A = X_L Z_0 + R_L X_{\mathrm{Amp}} \quad [9]$$

$$B = \sqrt{R_L Z_0 (X^2_{\mathrm{Amp}} + Z_0{}^2)} \quad [10]$$

$$C = R_L Z_0 - X_L X_{\mathrm{Amp}} \quad [11]$$

As can be seen, there exists a solution for the network component values for any possible value of X. This gives the freedom of choosing a realizable solution out of the range of possible solutions. Note that for the particular choice of $X = B/Z_0$, the network reduces to the previously described single stage matching network by Roemer. Note also, that for some values of X the sign of X_{C_2}, X_{C_3}, or X_{L_4} may change, indicating that the component is no longer a capacitor but an inductor or vice versa.

Circuit diagrams and design equations for four different network structures are given in the appendix. In the next section, a specific case will be discussed which will demonstrate the additional flexibility of the circuit of Fig. 3 over the circuit of Fig. 2.

MEASUREMENTS

To verify the design equations and to demonstrate the method, a small surface coil was matched to a preamplifier. The preamplifier had a 0.3 dB noise figure and, including the connecting cable and balun, the input reflection coefficient for noise match was $\Gamma = 0.99 \angle -107°$. A circular coil with a diameter of 4.5 cm was made from silver-coated copper wire. The coil impedance was measured at 41 MHz on an HP 4195A network analyzer (Hewlett-Packard, Sunnyvale, CA) to be

$$R_L + jX_L = (0.06 + j27)\ \Omega \quad [12]$$

Using the single stage design suggested in (3), and shown in Fig. 2, the component values of C_3 = 156 pF, L = 152 nH, and X = 2.14 Ω would be required. At 41 MHz, the value of X is equivalent to a capacitor with 1.8 nF. A capacitance that large indicates high currents in the matching network which would give rise to high losses. By use of the structure of Fig. 3 or Fig. A-1 (Appendix) however, a network can be designed that has components values better suited for practical construction.

Because the magnitude of the input reflection coefficient is close to unity, it was assumed that the amplifier input impedance equals a capacitance of 105 pF, which is equivalent to $\Gamma = 1.0 \angle -107°$. By use of a MathCad™ worksheet, a range of possible solutions for C_2, C_3, C_X and L_4 were calculated for different X using Eq. [6]–[11] (see also Appendix, Table 1). To achieve a reasonable value for the inductor L_4, the parameter X was selected to be 50 Ω, leading to the following values for the network components

$$X_{C_2} = 32\ \Omega\ @\ 41\ \mathrm{MHz} \Leftrightarrow C_2 = 121\ pF$$
$$X_{C_3} = 389\ \Omega\ @\ 41\ \mathrm{MHz} \Leftrightarrow C_3 = 10\ pF \quad [13]$$
$$X_{L_4} = 82\ \Omega\ @\ 41\ \mathrm{MHz} \Leftrightarrow L_4 = 317\ nH$$

These values transform the coil impedance into 50 Ω, ensuring proper match in the standard configuration when the coil is power matched. At the same time, connecting the coil through this network to the amplifier results in a high impedance at the coil terminals; this effectively reduces the current on the coil.

To demonstrate this effect, a matching network was built that included a manual switch. This allowed changing from a conventional 50 Ω termination to termination by a low noise preamplifier with a reflection coefficient of magnitude Γ = 0.99 (Fig. 4). The component values given in Eq. [13] were obtained by adjusting variable trimmer capacitors. The resonance was tuned and matched to 50 Ω using the capacitors C_2 and C_3, whereas the capacitor C_X was used to trim the antenna current to a minimum at the resonance frequency when the switch was set to connect the network to the preamplifier.

After adjusting the network, a pair of small inductive probes with minimized mutual coupling (<−80 dB) was used in a qualitative S_{21} measurement to observe the change of the currents on the coil for both switch settings (Fig. 4). In the absence of the coil, the measured S_{21} between the two probes was well below −80 dB. When bringing the probes close to the coil, the transmitting probe induced a current into the antenna which, in return, produced fields which were picked up by the receiving probe. For low coupling between the probes and the antenna, S_{21} is proportional to the current on the antenna. Therefore, the change in current between the standard 50 Ω match and the low noise amplifier match is proportional to the readings for the S_{21} measurements. Measured values of S_{21} for both switch settings can be seen in Fig. 5. The drop in S_{21} at the resonance frequency for the case of noise versus power (50 Ω) match is substantial and indicates a reduction in antenna current by approximately 30 dB.

To evaluate the efficiency of the new network in comparison to standard matching techniques, an identical

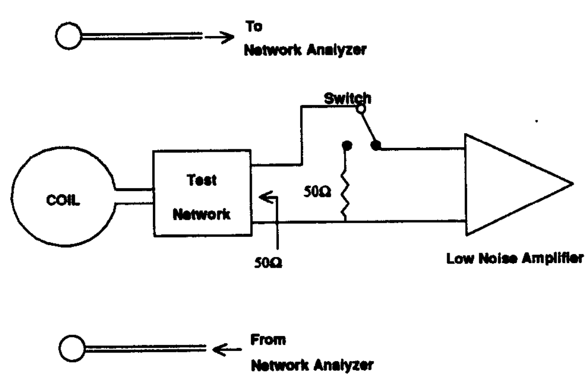

FIG. 4. Test network with switchable termination. The switch is used to change between a simulated 50 Ω amplifier input impedance (power match) and a low noise amplifier with high reflection coefficient. An S_{21} measurement was performed through a pair of decoupled probes that was brought close to the coil. The S_{21} values are proportional to the current on the coil and therefore inversely proportional to the sum of coil impedance and termination impedance. By flipping the switch, the effect of the insulating network on the coil current can be observed.

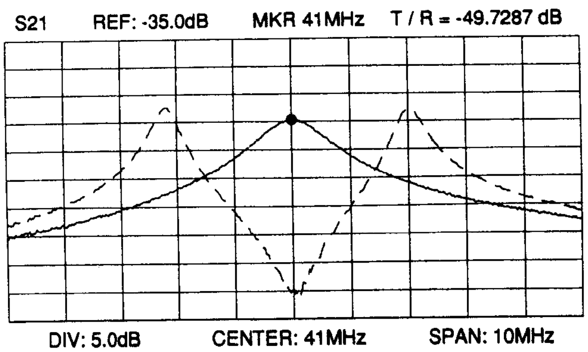

FIG. 5. S_{21} measurements with decoupled pair of probes close to the 4.5 cm coil for power match case and noise match case. The difference in S_{21} indicates that the insulating network reduces the coil current by 30 dB.

coil was built and matched to 50 Ω using a simple C_P/C_S matching section with C_P = 129 pF, C_S = 2.5 pF. For comparison of SNR, both coils were connected consecutive to the same amplifier and images were taken on a 1 T MR system. To the authors' surprise, the isolating matching network had a slightly (10%) better SNR than the standard matching network despite the fact that it has more components. This result was verified in the lab, by measuring root mean squared values of signal and noise at the amplifier output for both coils using an automated positioning system in an RF screen chamber. Fig. 6 shows relative SNR values acquired along the main axes of both antennas using this method. In a further measurement, both coils were again accurately tuned and matched but instead of connecting them to the low noise amplifier they were terminated into the 50 Ω impedance of one port of the network analyzer. A small probe was attached to the second port and because of the proportionality between the SNR and the magnetic field of the coil, relative SNR values along the main axes of the coils could be obtained from S_{21} measurements. Fig. 7 shows that for the 50 Ω match the ratio of the relative SNR values is almost inverse to the case of high impedance

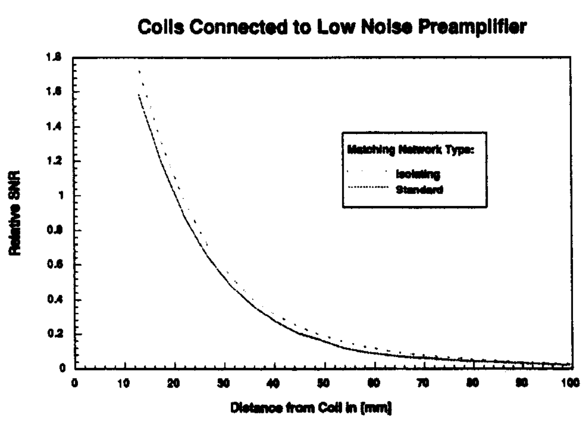

FIG. 6. SNR of standard matching network versus insulating matching network if both networks are terminated into a low noise amplifier with high reflection coefficient. The SNR of the insulating network is slightly higher.

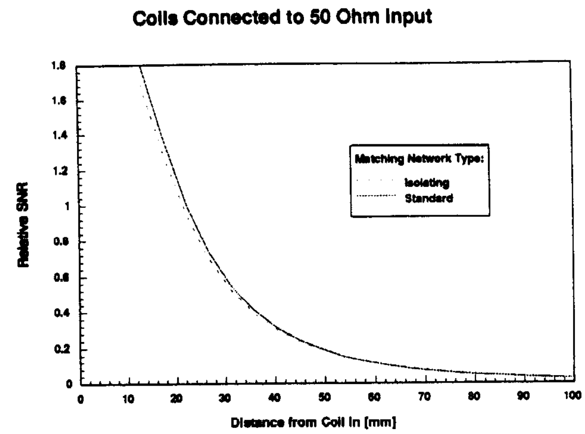

FIG. 7. SNR of standard matching network versus insulating matching network if both networks are terminated into 50 Ω. In this case, the SNR of the standard matching network is slightly higher.

mismatch with low noise amplifiers. Obviously, changing the current distribution within the network by changing the impedance match has an effect on the network efficiency. The relevant case here is the one where both coils are connected to low noise amplifiers because this will ensure the highest SNR for both coils. It seems that for this case the insulating matching network not only has the advantage of reducing inductive interaction between the coils but it also has a slightly improved SNR, compared with the C_S/C_P network. It can be speculated, however, that this might vary for different phases of the amplifier input reflection coefficient.

CONCLUSIONS

A new and more general concept for matching and decoupling networks has been presented. These networks can be used in conjunction with low noise preamplifiers which exhibit a reflection coefficient close to unity when noise matched. A common feature of the networks is that they match antennas to the required impedance for optimal SNR at the amplifier input while simultaneously transforming the input impedance of the preamplifier into a value close to an open circuit at the coil terminals. Thus, the coil currents are reduced significantly, lowering the inductive coupling between adjacent antennas. Design formulas were derived for four different network solutions. One of these networks was built to match a small coil to an amplifier with high input reflection coefficient (0.99). Measurements showed that for this particular network, the current on the antenna can be reduced by approximately 30 dB compared to standard matching and the SNR is actually slightly improved. Other designs may give different results depending on the Q factor of the structure. The networks shown here have an extra degree of freedom compared with previous designs; this allows the variation of network parameters to avoid extreme component values. Furthermore, the problem has been generalized to amplifiers with high input reflection coefficient rather than those with very small input impedances. No additional phase shifter is

Table A-1
Design Equations for the Different Network Structures

Fig. A-1	Fig. A-2	Fig. A-3	Fig. A-4
$A = X_L Z_0 + R_L X_{\mathrm{Amp}}$		$D = \sqrt{Y_0(G_L^2 + Y_L^2)}$	
$B = \sqrt{R_L Z_0 (X_{\mathrm{Amp}}^2 + Z_0^2)}$		$E = \sqrt{G_L(Y_{\mathrm{Amp}}^2 + Y_0^2)}$	
$C = R_L Z_0 - X_L X_{\mathrm{Amp}}$		$F = (G_L Y_{\mathrm{Amp}} + Y_L Y_0)$	
$X_{C_2} = \frac{XA}{XZ_0 - B}$	$X_{C_2} = \frac{XA}{XZ_0 - B}$	$Y_{C_2} = \frac{YD^2}{YF - DE}$	$Y_{C_2} = \frac{YD^2}{YF - DE}$
$X_{C_3} = X\frac{A - B}{B}$	$X_{C_3} = X\frac{A + B}{B}$	$Y_{C_3} = Y\frac{D + E}{E}$	$Y_{C_3} = Y\frac{D - E}{E}$
$X_{L_4} = X\frac{A - B}{A} + Z_0\frac{C}{A}$	$X_{C_4} = X\frac{A + B}{A} - Z_0\frac{C}{A}$	$Y_{C_4} = Y\frac{D + E}{D} - Y_{\mathrm{Amp}}$	$Y_{L_4} = Y\frac{D - E}{D} + Y_{\mathrm{Amp}}$

required so the number of network components can be kept low (only four). The authors have coded the design equations into MathCad™ worksheets, which are available on request.

APPENDIX

The following figures show four possible structures for matching and decoupling networks that can be used to connect MRI antennas to low noise preamplifiers. The first two designs are more advantageous for cases where the amplifier input impedance is close to a short circuit (left side of Smith chart), but the second two give better results for an input impedance that is close to an open circuit. From the possible sets of all structures, the ones shown here have been selected because they require only one inductor. It is preferable to have as few inductors as possible because inductors can not be so easily tuned over a large frequency range as capacitors. Furthermore, they produce magnetic fields that might interact with the actual MRI antenna. Table A-1 gives an overview of all equations required to design the different networks. X and Y have been chosen as free parameters for the designs, but any other component value could be used as free parameter by simply rearranging the equations.

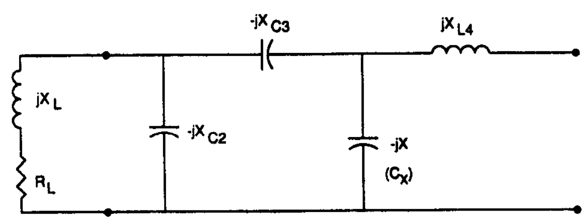

FIG. A-1. Structure A-1, which is discussed in detail in the article. This structure is preferred for amplifier input impedances that are close to a short circuit.

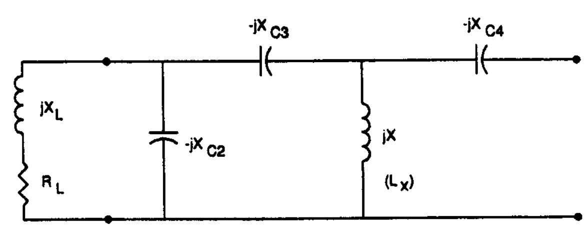

FIG. A-2. Structure A-2, which is similar to structure A-1.

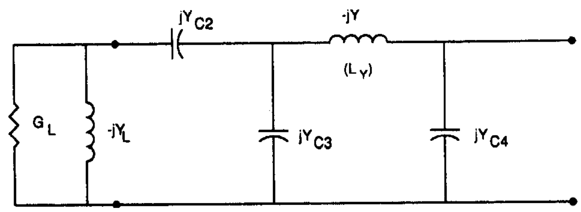

FIG. A-3. Structure A-3, which is preferred for cases in which the amplifier input impedance is close to an open circuit.

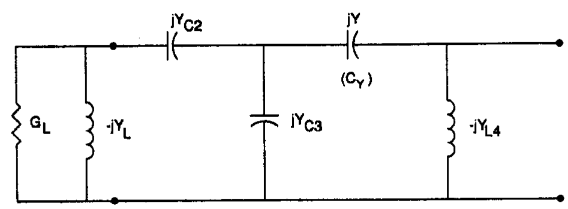

FIG. A-4. Structure A-4, which is similar to A-3.

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