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Noise figure characterization of preamplifiers at NMR frequencies

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ABSTRACT

A method for characterizing the noise figure of preamplifiers at NMR frequencies is presented. The noise figure of preamplifiers as used for NMR and MRI detection varies with source impedance and with the operating frequency. Therefore, to characterize a preamplifier's noise behavior, it is necessary to perform noise measurements at the targeted frequency while varying the source impedance with high accuracy. At high radiofrequencies, such impedance variation is typically achieved with transmission-line tuners, which however are not available for the relatively low range of typical NMR frequencies. To solve this issue, this work describes an alternative approach that relies on lumped-element circuits for impedance manipulation. It is shown that, using a fixed-impedance noise source and suitable ENR correction, this approach permits noise figure characterization for NMR and MRI purposes. The method is demonstrated for two preamplifiers, a generic BF998 MOSFET module and an MRI-dedicated, integrated preamplifier, which were both studied at 128 MHz, i.e., at the Larmor frequency of protons at 3 Tesla. Variations in noise figure of 0.01 dB or less over repeated measurements reflect high precision even for small noise figures in the order of 0.4 dB. For validation, large sets of measured noise figure values are shown to be consistent with the general noise-parameter model of linear two-ports. Finally, the measured noise characteristics of the superior preamplifier are illustrated by SNR measurements in MRI data.

1. Introduction

NMR and MRI are limited by the relative weakness of nuclear magnetism, which results in modest overall sensitivity and signal-to-noise ratio (SNR). Consequently, MR instrumentation and methods generally aim to maximize the primary signal yield of a given experiment as well as to minimize signal degradation by the detector hardware and along the remainder of the receiver chain.

With common Faraday detection the receiver chain typically consists of a coil to detect the induced electromotive force, tuning and matching circuitry, a preamplifier, and a cascade of further amplifiers to boost the signal to a level suitable for analog demodulation or direct digitization. Inevitably, each of these components adds a certain amount of noise and thus reduces the SNR below its ideal, intrinsic value [1]. The dominant noise sources are those that affect the signal when it is weakest, i.e., before it is initially amplified by typically 20 dB or more. Therefore, the most critical contributions of detector noise arise from passive components such as the coil conductor, associated circuitry and cable (if present), and the preamplifier itself (active). The magnitude of noise added by the passive components increases both with their ohmic losses and with their temperature. Therefore, in addition to using good room-temperature conductors and high-quality lumped reactances, these noise contributions can also be reduced by cryogenic cooling [2–6] and by the use of superconductors [7–11].

The noise added by the preamplifier is usually characterized by its noise figure (NF), which expresses the relative SNR degradation caused by a signal transfer or amplification stage [12]. The noise figure depends not only on the device used and the frequency of operation but also on the source impedance, i.e., the impedance that the coil presents to the preamplifier. For a given linear device and operating frequency, the noise figure is minimal at a unique optimal source impedance Z_{opt} [13,14]. Noise matching consists of transforming the complex coil impedance to yield Z_{opt}, thus ensuring optimal SNR performance of the preamplifier. Noise matching is straightforward for single-channel receivers with a fixed load. However, it is challenging and often subject to compromise for arrays of variably-loaded and mutually-coupled receiver coils [15–18], particularly when the radiofrequency (RF) wavelength is smaller than the imaging target [19,20], for large coil numbers [21–24] and with geometrically adjustable coil configurations [25–27]. In all of these situations, preamplifier noise is a key determinant of net sensitivity and must be carefully controlled.

Effective noise matching requires very accurate measurement of noise figures and their dependence on source impedance, for which a variety of established approaches exist. All these methods

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rely on measurements of noise power at the output of the device under test (DUT). Direct measurements require power detection with absolute-level accuracy, which is challenging for low noise powers but useful for assessing large noise figures [28]. The so-called signal generator twice-power method is also most suitable for high-noise-figure devices and requires knowledge of the noise bandwidth of the instrument used for the power measurement [28]. A more common and more sensitive alternative is the Y-factor method, which is based on measuring output noise powers for different noise levels at the input of the DUT [28]. In its traditional implementation, the input noise is generated by a resistor whose temperature is varied, e.g., between room-temperature and that of liquid nitrogen [29]. However, in this approach the temperature change is much smaller than that available from dedicated, electronically-controlled noise sources that allow more sensitive, automated noise figure measurements to be made [30].

In its traditional form, the Y-factor method permits noise figure measurements only for one single source impedance, which is determined by the noise source used and is typically 50 Ω for standard telecommunications applications. It is therefore not suitable for studying the variation of the noise figure with source impedance as necessary for NMR and MRI applications. For high operating frequencies (above 1 GHz), the effective impedance of noise sources is sometimes varied with tuners based on transmission line technology [31], which however are expensive and not readily available for NMR frequencies ranging down to several tens of MHz.

To address this shortfall, the present work describes a simple method of impedance tuning for suitably low operating frequencies. In the proposed approach, reliable variation of noise temperature and impedance is achieved by a standard diode noise source followed by lumped-element circuits [32] containing both resistive and reactive components. A calibrated set of such circuits permits sampling the noise figure across the complex impedance plane at common NMR frequencies. The proposed approach is demonstrated by noise figure measurements of RF preamplifiers at 128 MHz, comparing a specifically-designed, integrated module with a generic MOSFET. The method is validated by verifying the consistency of highly overdetermined noise figure measurements with the underlying theoretical model.

2. Methods and results

2.1. Measurement Method

The noise factor F of a DUT is defined as

$$F = \frac{SNR_{input}}{SNR_{output}} = \frac{Signal_{input}/Noise_{input}}{Signal_{output}/Noise_{output}}, \quad (1)$$

where SNR_{input} and SNR_{output} denote the SNR in terms of power at the DUT's input and output, respectively. According to this definition, the noise factor depends not only on the amount of noise that the DUT adds but also on the noise level at its input. Therefore, when using the noise factor to characterize a DUT, the noise level at its input is usually assumed to amount to thermal noise at the reference temperature $T_0 = 290^\circ K$ [12]. The corresponding noise figure is then calculated by converting to decibels:

$$NF = 10 \cdot \log F. \quad (2)$$

In the Y-factor method, the noise figure of a given DUT is determined by measuring the noise power at its output in the presence of two different but well defined noise levels at its input. The ratio of the two measured output power values, N_1 and N_2, is called the Y-factor:

$$Y = \frac{N_2}{N_1}. \quad (3)$$

In present-day implementations of the Y-factor method, the input noise is typically generated by 50-Ω noise sources based on avalanche diodes. When unbiased, such a noise source produces a noise power equivalent to that of a 50 Ω resistor at room-temperature. This situation is usually called the cold state and the temperature is referred to as T_c. With reverse bias into avalanche breakdown, the noise power increases while the presented impedance is approximately the same. The noise source therefore behaves like a 50 Ω resistor at an increased temperature T_h. The two equivalent noise temperatures define the excess noise ratio,

$$ENR = \frac{T_h - T_c}{T_0}. \quad (4)$$

Based on the Y-factor and the ENR, the observed noise factor is given by

$$F = \frac{ENR}{Y - 1}. \quad (5)$$

This value, however, is only an approximation of the DUT's noise factor because it includes some noise contribution from the power meter. According to Friis' formula [12], the total noise factor of two cascaded devices is given by

$$F_{tot} = F_1 + \frac{F_2 - 1}{G_1}, \quad (6)$$

where F_1, F_2 denote the noise factors of the individual stages and G_1 is the available gain of the first device. Therefore, the measured noise factor, F_{tot}, must be corrected for that of the second stage, requiring knowledge of the available gain of the DUT (G_1) and of the noise factor of the power meter (F_2).

The available gain of the DUT can be calculated based on S-parameter measurements of the DUT and a measurement of the source impedance, expressed through the reflection coefficient Γ_s [28,33–35]:

$$G_1 = \frac{(1 - |\Gamma_s|^2)|S_{21}|^2}{|1 - S_{11}\Gamma_s|^2 \left(1 - |S_{22} + \frac{S_{12}S_{21}\Gamma_s}{1 - S_{11}\Gamma_s}|^2\right)}. \quad (7)$$

F_2 can be determined with the Y-factor method in a separate calibration step by connecting the noise source directly to the power meter. This calibration makes the assumption (verified and used in the present work) that the power meter's noise factor is the same when connecting it either to the noise source or the DUT. If this assumption does not hold, F_2 must be measured specifically for the source impedance presented by the DUT. When necessary, such a preparatory noise factor measurement could be done through a first run of the proposed method, treating the power meter as the device under test and skipping the 2nd-stage correction according to Eq. (7). Based on these additional measurements, the noise factor of the DUT can be calculated by rearranging Eq. (6),

$$F_1 = F_{tot} - \frac{F_2 - 1}{G_1}, \quad (8)$$

and its noise figure is then given by Eq. (2).

With the conventional setup described thus far (Fig. 1a), the noise figure of the DUT can be measured only at the source impedance presented by the noise source, i.e., at 50 Ω. As mentioned in the introduction, a transmission-line tuner placed between the noise source and the DUT (Fig. 1b) is not practical for NMR frequencies, but the proposed approach achieves similar functionality by introducing exchangeable lumped-element tuning circuits (Fig. 1c).

In principle, any combination of resistors, capacitors, and inductors could be used to vary the effective impedance seen by the