Noise Figure of Digital Communication Receivers—Revisited

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***Abstract*—Noise figure (NF) is a commonly used system parameter that quantifies the degradation in the signal-to-noise ratio (SNR) as the signal passes through a receiving system. Because of the difficulty in defining the SNR, NF depends on how the SNR is computed and the underlying assumptions that are made. Existing NF measures and their shortcomings are explained. A new NF suitable for a digital communication receiver is proposed by redefining the SNR, so that the NF measures the degradation in the achievable performance caused by the receiving system. The proposed NF, which we refer to as the effective NF, can be readily determined based on the existing NF measurement techniques. As an example of the use of the effective NF metric, a direct-conversion receiver with ac coupling in the signal path to remove the dc-offset noise is described.**

***Index Terms*—Circuit optimization, digital communications, integrated circuit noise.**

I. Introduction

NOISE FACTOR [or noise figure (NF) in decibels] is an important system parameter that is closely related to the overall receiver performance or the bit-error rate (BER). It is commonly used to characterize the ability of a receiving system to process the input signals, where the receiving system refers to the entire analog front-end as well as its individual components, such as the low-noise amplifier [1], the mixer [2], and the baseband and intermediate frequency (IF) amplifiers.

The formal definition of NF has been introduced in the 1940s by Friis [3] as

$$F \equiv \frac{\mathrm{SNR_{in}}}{\mathrm{SNR_{out}}} \tag{1}$$

where $\mathrm{SNR_{in}}$ is the input signal-to-noise ratio (SNR) and $\mathrm{SNR_{out}}$ is the output SNR. As such, NF represents the degradation in the SNR as the signal passes through the receiving system. Although the meaning of NF is straightforward, measuring the NF can be problematic because of the difficulty in defining the SNR. Consequently, the NF depends on how the SNR is computed and the underlying assumptions that are made.

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There are basically three different NFs that are reported in the literature [4]–[9]: spot NF, power NF, and averaged NF. As described in the following section, however, these NF measures suffer from several shortcomings. For example, the spot NF is often frequency dependent and not unique to a receiving system. The power and averaged NF, which are well defined unlike the spot NF, are not necessarily indicative of the overall receiver performance.

The goal of the analog front-end in digital receivers is to condition the received analog signal for digitization, so that the highest performance can be achieved after decoding in the digital domain. For the NF of a receiver to be a meaningful metric, the SNR at the input and output of the receiving system should measure the performance after the eventual digital decoding process, as it is ultimately the most relevant measure of performance. Since the eventual performance depends on the choice of the detection algorithm, which is system dependent and often difficult to quantify, the SNR is defined as the achievable performance assuming optimal detection. More specifically, we define the SNR as the matched filter bound (MFB) [10], which represents an upper bound on the performance of data transmission systems with intersymbol interference (ISI). The MFB is obtained when a noise-whitened matched filter is employed to receive a single transmitted pulse. By defining the SNR as the MFB, the NF measures the degree of degradation in the achievable receiver performance caused by the receiving system.

In this paper, a new NF suitable for digital communication receiving systems is proposed by redefining the SNR as the MFB. The proposed NF, which we subsequently refer to as the effective NF, can be readily determined based on the existing spot NF measurement techniques. The organization of this paper is as follows. The existing NF measures in the communication receivers are reviewed (e.g., the spot NF, the power NF and the averaged NF), and their shortcomings are described in Section II. The proposed effective NF is presented in Section III. As an example of the use of the effective NF metric, Section IV describes a direct-conversion receiver (DCR) with ac-coupling filter to remove the dc-offset noise. Conclusions are drawn in Section V.

II. Existing NF Measures

A. Spot NF

The spot NF is determined by computing the NF given in (1) at an infinitesimal frequency band centered at a frequency f within the input signal band [5], [6]

$$F_s(f) \equiv \frac{S_s(f)\,/S_{n_i}(f)}{S_s(f)G(f)\,/\big(S_{n_i}(f)G(f)+S_{n_g}(f)\big)} \tag{2}$$

$$= \frac{S_{n_i}(f)G(f) + S_{n_g}(f)}{S_{n_i}(f)G(f)} \quad (3)$$

where $S_s(f), S_{n_i}(f)$, and $S_{n_g}(f)$ represent the input signal power-spectral density (PSD), input noise PSD, and internally generated noise PSD referred to the output, respectively. The input noise PSD $S_{n_i}(f)$ is commonly assumed to be white with magnitude corresponding to a noise temperature of 290 K. $G(f)$ is the power gain of the receiving system. As shown in (3), the spot NF $F_s(f)$ is independent of $S_s(f)$; it is simply the ratio of the noise power output at the infinitesimal frequency band to that portion of the noise power output due to the noise at the input. The absence of the input and output signals makes the spot NF attractive as a basis for measurement. Consequently, most of the NFs reported in the literature are spot NFs.

The main drawback of the spot NF is that it is frequency dependent. If $S_{n_i}(f), S_{n_g}(f)$, and $G(f)$ in (3) are not fixed over the frequency band of interest, $F_s(f)$ can become a function of the center frequency f. The reported NF of a receiving system is then, not unique, and would depend on the selection of f. Therefore, when reporting the NF performance of a receiving system using $F_s(f)$, the underlying assumption is that $F_s(f)$ is fixed over the frequency band of interest. This assumption is often violated in modern digital receivers. An example is the DCR with ac coupling in the signal path to remove the dc-offset noise. This example is described greater detail in a subsequent section.

B. Power NF

The power NF removes the frequency dependency of the spot NF by defining the signal and noise components in (1) as the total signal power and noise power over the frequency band of interest B [7]–[9]. The power NF is

$$F_a \equiv \frac{\int_B S_s(f)\,df / \int_B S_{n_i}(f)\,df}{\int_B S_s(f)G(f)\,df / \int_B \left(S_{n_i}(f)G(f) + S_{n_g}(f)\right) df} \quad (4)$$

where all of the integrations in (4) are over a frequency band of interest B. If B is an infinitesimal frequency band centered at f_0, the average NF F_a becomes the spot NF $F_s(f_0)$.

If $S_s(f)$ and $S_{n_i}(f)$ are assumed white over B, the power NF in (4) becomes simply the total output noise power divided by the total input noise power referred to the output

$$F_a = \frac{\int_B \left(S_{n_i}(f)G(f) + S_{n_g}(f)\right) df}{\int_B S_{n_i}(f)G(f)\,df}. \quad (5)$$

This is the power NF that is often cited in the literature.

Although the frequency dependency of the spot NF is removed, the main drawback of this definition is that a lower power NF does not necessarily translate to a higher overall receiver performance. This point is best illustrated through an example shown in Fig. 1. The input signal and noise, both of which are assumed white, pass through Receivers A and B with different $G(f)$ and $S_{n_g}(f)$. In the resulting PSDs shown in Fig. 1, the total output noise power of both receivers is assumed to be the same. Then, the power NF of the two receivers is also the same. However, Receiver A can clearly achieve a higher performance after the eventual digital decoding process, since the noise in Receiver A is easily filtered out with little degradation on the overall signal spectrum. By contrast, the noise in Receiver B is spread across the signal spectrum and cannot be selectively filtered out. As this example illustrates, the performance of the receiver after the digital decoding process does not depend on the total signal and noise power. Therefore, the power NF is in general not an accurate metric for quantifying the overall SNR degradation caused by the receiving system.

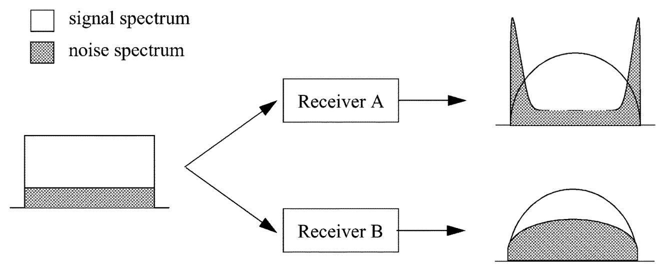

Fig. 1. Problems with power NF.

C. Averaged NF

Another NF employed in the literature is what we refer to as the averaged NF. The averaged NF is obtained by appropriately weighting the spot NF across the frequency band of interest B [5]

$$F_w = \int_B F_s(f)W(f)\,df \quad (6)$$

where $W(f)$ is the weighting function that is constrained to be

$$\int_B W(f)\,df = 1. \quad (7)$$

The main difficulty in employing the averaged NF is in determining the weights $W(f)$. In the literature [5], $W(f)$ is often weighted uniformly or according to $G(f)$ without a sound technical basis. A more rigorous relationship between $W(f)$ and the overall receiver performance after the digital decoding process given $G(f)$ and $S_{n_g}(f)$ needs to be established. As shown in the following section, however, a meaningful NF that measures the degree of degradation in the overall receiver performance caused by the receiving system does not have a linear relationship with $F_s(f)$ as in the averaged NF.

III. Effective NF

As stated earlier, the main difficulty in computing the NF is in defining the SNR. By defining the SNR as the MFB, the NF represents the degradation in the achievable SNR after the digital decoding process.

A general system model of a communication channel including the receiving system is shown in Fig. 2. The kth transmit symbol x_k is filtered by the equivalent pulse response then corrupted by the additive noise $n_i(t)$. The equivalent pulse response (whose frequency response is $P(f)$) represents the combination of both the transmit pulse and the propagation channel. The resulting corrupted signal is the input of the receiving system, which has a transfer function given by $(\sqrt{G(f)})$ and additive noise $n_g(t)$.

The MFB, also called the "one-shot" bound, is an upper limit on the performance of data transmission systems with ISI. The

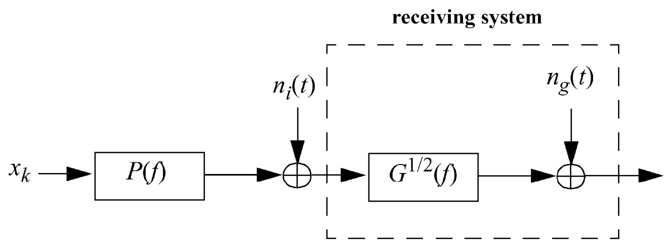

Fig. 2. General system model.

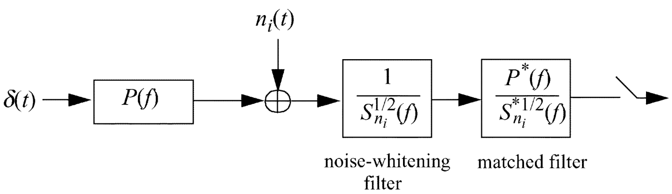

Fig. 3. MFB computation at the input of the receiving system.

MFB is determined by employing a noise whitened matched filter to receive a "one-shot" transmission pulse. The MFB computation of the input signal to the receiving system in Fig. 2 is illustrated in Fig. 3. Assuming unity transmit signal energy, i.e., $E\{x_k^2\} = 1$, an impulse with unity energy is transmitted through the equivalent pulse response, which is then corrupted by $n_i(t)$. The input to the receiving system is noise whitened followed by a matched filter that is matched to the waveform obtained by convolving the pulse response with the noise whitening filter. The matched filter output is then sampled when the output is maximized. The resulting SNR is the MFB.

The MFB at the input and output of the receiving system is

$$\mathrm{SNR_{in}} = \int \frac{|P(f)|^2}{S_{n_i}(f)}\, df \tag{8}$$

$$\mathrm{SNR_{out}} = \int \frac{|P(f)|^2 G(f)}{S_{n_i}(f)G(f) + S_{n_g}(f)}\, df. \tag{9}$$

Substituting (8) and (9) into (1), the effective NF of the receiving system is

$$F_{\text{eff}} = \frac{\int \frac{|P(f)|^2}{S_{n_i}(f)}\, df}{\int \frac{|P(f)|^2 G(f)}{S_{n_i}(f)G(f)+S_{n_g}(f)}\, df}. \tag{10}$$

Assuming as is commonly done that the input noise $n_i(t)$ is white, the NF can then be written as a function of the spot NF $F_s(f)$

$$\begin{aligned} F_{\text{eff}} &= \frac{\int |P(f)|^2\, df}{\int |P(f)|^2 \frac{S_{n_i}(f)G(f)}{S_{n_i}(f)G(f)+S_{n_g}(f)}\, df} && (11)\\ &= \frac{\int |P(f)|^2\, df}{\int |P(f)|^2 \frac{1}{F_s(f)}\, df} && (12)\\ &= \frac{1}{\int \left(\frac{|P(f)|^2}{P_T}\right) \frac{1}{F_s(f)}\, df} && (13) \end{aligned}$$

where $P_T = \int |P(f)|^2\, df$. In (13), the spot NF values contribute to the effective NF only in the frequencies where the signal is present (i.e., $P(f)$ is nonnegligible). In addition, the effective NF becomes the spot NF when $F_s(f)$ is constant over the frequency band of interest. Since this is the condition that the receiving system must satisfy for the use of the spot NF to be meaningful, the effective NF is equivalent to the spot NF as long as the use of the spot NF is valid.

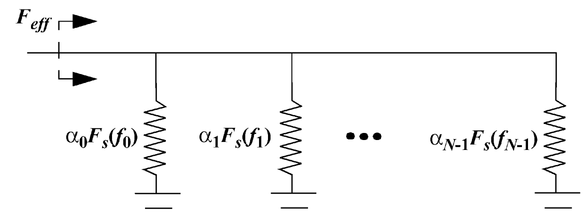

Fig. 4. Equivalent model of effective NF.

The relationship between spot NF and effective NF can be better understood by approximating (13) using finite summations. The effective NF is then

$$F_{\text{eff}} \approx \frac{1}{\sum_{i=0}^{N-1} \frac{1}{\alpha_i F_s(f_i)}} \tag{14}$$

where $\{f_0, f_1, \ldots, f_{N-1}\}$ represent the equally-spaced center frequencies for each of the spot NF measurements in the frequency band of interest, N is the total number of measured values, and

$$\alpha_i = \frac{\sum_{i=0}^{N-1} |P(f_i)|^2}{|P(f_i)|^2}. \tag{15}$$

In (13) and (14), the effective NF equation is basically the formula for the weighted harmonic mean. Since the harmonic mean is used for computing the effective resistance of parallel resistors, the effective NF computation can be viewed as determining the effective resistance of N resistors with resistance $\alpha_i F_s(f_i)$ placed in parallel as shown in Fig. 4. The resistance is obtained by scaling the spot NF at frequency f_i by α_i, which is a function of the shape of $P(f)$ as given in (15). If $P(f)$ is assumed constant over the frequency band of interest, $\alpha_i = N$ for $i \in \{0, 1, \ldots, N-1\}$ and all the spot NF values are weighted equally.

The parallel resistor perspective, illustrated in Fig. 4, implies that having a few very high $\alpha_i F_s(f_i)$ values have little effect on the effective NF, since the equivalent resistance of parallel resistors is dominated by the smaller resistors. This observation can be used to show that in the example given in Fig. 1, the effective NF of Receiver A is lower than that of Receiver B, since the smaller spot NF values of Receiver A have a larger impact on the effective NF computation. This result is consistent with the intuition described earlier that Receiver A should achieve higher performance after the eventual digital decoding process.

The observation that a few large spot NF values have a small effect on the effective NF suggests new design strategies, such as significantly increasing the spot NF in some frequencies to achieve other implementation benefits while incurring minimal overall performance degradation. An example of such a system is the ac-coupled DCR and is described in the following section.

The harmonic mean can be shown using the properties of convex functions to be always less than or equal to the arithmentic mean. Therefore, the averaged NF, which is the weighted arithmetic mean of the spot NF values, is always greater or equal to the effective NF. This relation implies that the averaged NF is overly pessimistic when computing the loss in performance caused by the receiving system.

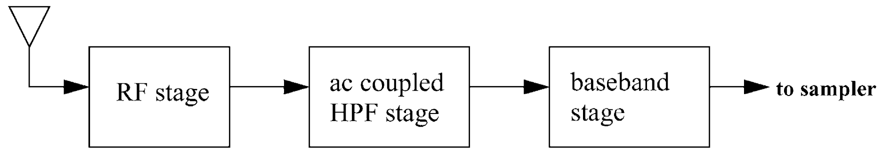

Fig. 5. Block diagram of a DCR with ac coupling.

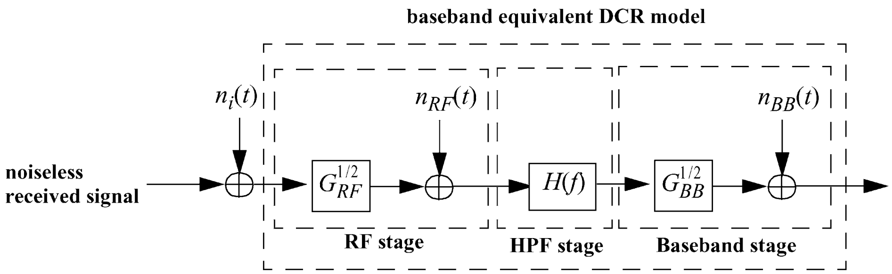

Fig. 6. General system model of ac-coupled DCR.

IV. Example: AC-Coupled DCR

A. Background

In a DCR, the received RF signal is mixed directly to the baseband for amplification and eventual digitization. Compared to the commonly employed superheterodyne receivers, which first downconvert the input RF signal to a lower IF, the DCR relaxes the selectivity requirements of RF filters and eliminates all IF analog components, allowing a highly integrated, low-cost and low-power realization.

One of the main challenges of implementing a DCR is in handling the effects of dc-offset noise, which arises predominantly from the self-mixing of the local oscillator. The dc-offset noise can dominate the signal strength by as much as two orders of magnitude in amplitude and, if not removed, substantially degrades the bit-error probability. Furthermore, this offset must be removed in the analog domain prior to sampling, since it would otherwise saturate the baseband amplifiers. One approach for removing the dc-offset noise in a DCR is to use a simple ac coupling filter (i.e., high-pass filter) in the downconverted signal path. Although extremely attractive from an implementation perspective, the ac coupling filter causes signal distortion, which results in performance loss. This has caused some researchers to state that the corner frequency of the ac coupling high-pass filter needs to be unrealistically small to achieve high performance [11], [12]. However, as shown in [13], high performance DCR with ac coupling is possible by treating the DCR front-end as part of an ISI channel and employing appropriate digital equalizers.

B. NF of ac-Coupled DCR

A simplified block diagram of a DCR with ac coupling is shown in Fig. 5. The received signal is passed through the RF stage, where the signal is amplified and mixed to the baseband, ac coupled to eliminate the dc-offset noise, then amplified in the baseband stage. Assuming that the input and output impedances of the RF, ac coupling, and baseband stages are all matched to 50 Ω, the RF and the baseband stages have power gains of G_{RF} and G_{BB}, respectively, and spot NFs of F_{RF} and F_{BB}, respectively. The power gains and spot NFs of the RF and the baseband

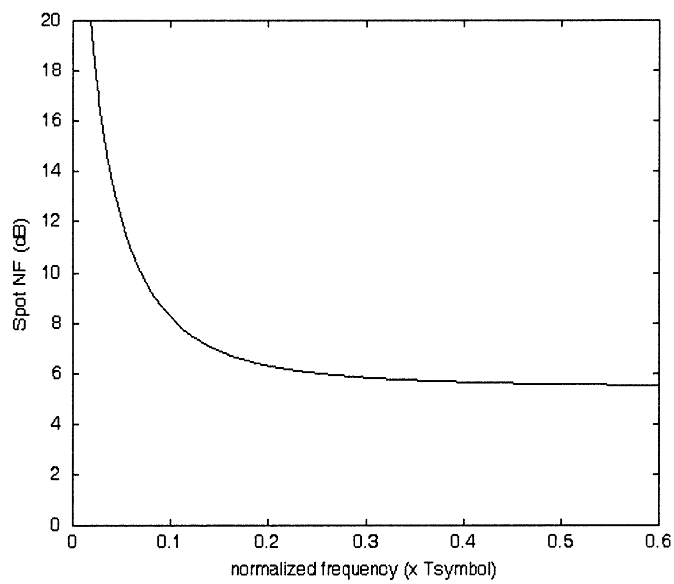

Fig. 7. Spot NF when $f_c = 0.1$.

stages are assumed constant over the frequency band of interest. The 50-Ω termination assumption is not necessary in a practical integrated DCR, but it is made to simplify both the presentation and the analysis. The received signal passes through the RF and baseband stages without signal distortion. Signal distortion occurs in the ac coupling high-pass filter (HPF), which is assumed to be noiseless with a transfer function given by

$$H(f) = \frac{jf}{jf + f_c} \tag{16}$$

where f_c is the corner frequency.

The baseband equivalent system model of the ac-coupled DCR is shown in Fig. 6. We assume that when the dc-offset noise, which is generated in the RF stage, is passed through the HPF, the PSD of the dc-offset noise is attenuated to well below $N_{\mathrm{BB}}/G_{\mathrm{BB}}$, where N_{BB} is the PSD of the internally generated noise in the baseband stages. The dc-offset noise then becomes negligible compared to the noise in the baseband stages, allowing us to ignore the dc-offset noise without compromising the accuracy of our analysis.

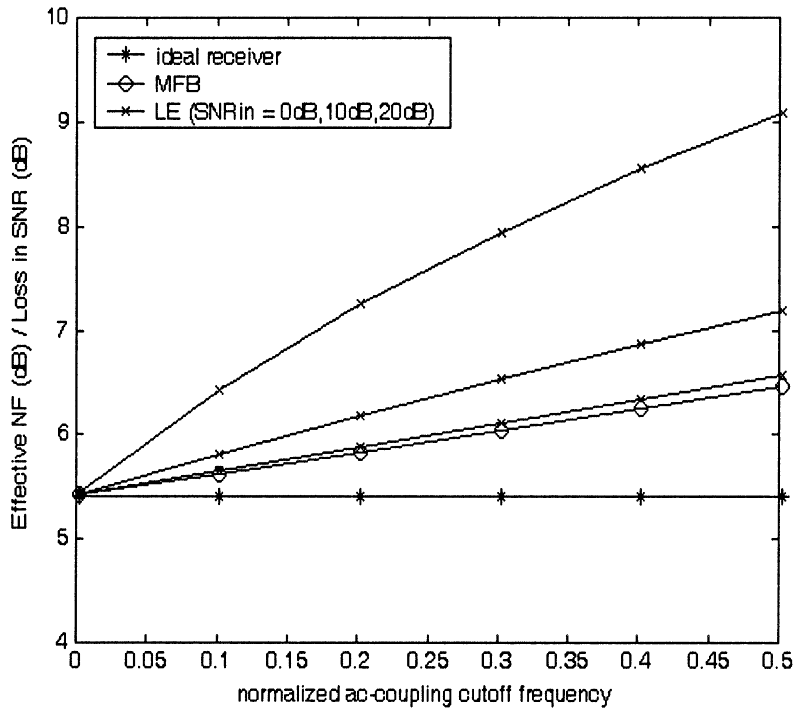

Fig. 8. NF versus f_c/B when the LE is employed.

For a cascade of the multiple-stage receiving systems, the equivalent spot NF in (13) can be determined by the well-known Friis formula [3], i.e.,

$$F_s(f) = F_{s1}(f) + \frac{F_{s2}(f)-1}{G_1(f)} + \cdots + \frac{F_{sN}(f)-1}{G_1(f)\ldots G_{N-1}(f)} \tag{17}$$

where $F_{si}(f)$ and $G_i(f)$ denote the spot NF and gain of the ith cascaded receiving system. Using (17), the spot NF of the ac-coupled DCR can be written as

$$\begin{aligned} F_s(f) &= F_{\mathrm{RF}} + \frac{F_{\mathrm{BB}}-1}{G_{\mathrm{RF}}|H(f)|^2} \\ &= \frac{(F_{\mathrm{RF}}G_{\mathrm{RF}} + F_{\mathrm{BB}} - 1)f^2 + (F_{\mathrm{BB}}-1)f_c^2}{G_{\mathrm{RF}}f^2} \end{aligned} \tag{18}$$

where the second equality in (18) is obtained by substituting $H(f)$ with (16). Fig. 7 plots the spot NF in (18) as a function of frequency under the following operating conditions: $f_c = 0.1, F_{\mathrm{RF}} = 5$ dB, $G_{\mathrm{RF}} = 25$ dB, $F_{\mathrm{BB}} = 25$ dB, $G_{\mathrm{BB}} = 60$ dB. Because of the ac-coupling filter, the spot NF increases abruptly below f_c.

Assuming that $P(f)$ is flat over the frequency band of interest B, the effective NF of the ac-coupled DCR can be readily determined by substituting (18) into (13) and integrating

$$F_{\mathrm{eff}} = \left(F_{\mathrm{RF}} + \frac{F_{\mathrm{BB}}-1}{G_{\mathrm{RF}}}\right) \cdot \frac{1}{1 - \frac{f_c}{B}\sqrt{\kappa}\,\mathrm{atan}\frac{1}{\frac{f_c}{B}\sqrt{\kappa}}} \tag{19}$$

where $\kappa = (F_{\mathrm{BB}}-1)/(F_{\mathrm{RF}}G_{\mathrm{RF}} + F_{\mathrm{BB}} - 1)$. The effective NF in (19) consists of two product terms. The first represents the NF of an ideal receiver, which is defined as a receiver that does not suffer from the dc-offset noise, and consequently, does not require an ac-coupling filter (i.e., $H(f) = 1$). The second product term in (19) represents the increase in NF caused by the ac-coupling filter. As expected, the second product term increases with increasing f_c and converges to one when no ac-coupling filter is employed, i.e., $f_c = 0$.

Fig. 8 plots F_{eff} in (19) as a function of f_c/B. Except for f_c, the same operating conditions are assumed as in Fig. 7. For comparison purposes, Fig. 8 also plots the NF of the ideal receiver, which is the first term on the right-hand side of (19), and the loss in the SNR of an infinite-length linear equalizer (LE) for different input SNR values [10]. The loss in the SNR of the LE is determined assuming the transmit symbol period is $1/B$ and by computing the ratio of the input SNR given in (8) to the unbiased SNR after an infinite-length LE is employed.

As f_c/B increases, the NF of the ac-coupled DCR increases because the amount of ISI introduced increases. Despite suffering from increased ISI, a DCR with a high f_c is more effective at attenuating the time-varying dc-offset noise. The selection of the corner frequency, therefore, requires a careful balance between conflicting interests—minimizing ISI and maximizing immunity against the dc-offset noise.

Even with $f_c/B = 0.1$, which correspond to a corner frequency that is two orders of magnitude higher than the maximum operable as stated in [11], [12], the effective NF is less than 0.5 dB greater than the NF of the ideal receiver. For the LE, the SNR loss is approximately 0.1, 0.3, and 1.5 dB greater than the effective NF for $\mathrm{SNR}_{\mathrm{in}}$ of 0, 10, and 20 dB, respectively. The LE seems most effective at low $\mathrm{SNR}_{\mathrm{in}}$ values.

Instead of the LE, the decision feedback equalizer (DFE) can be employed to improve the receiver performance at higher $\mathrm{SNR}_{\mathrm{in}}$ values [10]. Assuming the previous decisions are correct, which is generally a valid assumption at high $\mathrm{SNR}_{\mathrm{in}}$ values, the unbiased SNR of the DFE is solved analytically and the results are shown in Fig. 9. The plots are obtained under the same operating conditions as is Fig. 8, except that the performance of the LE is replaced with the performance of the DFE. The gap between the effective NF and the SNR loss of the DFE has reduced considerably than when the

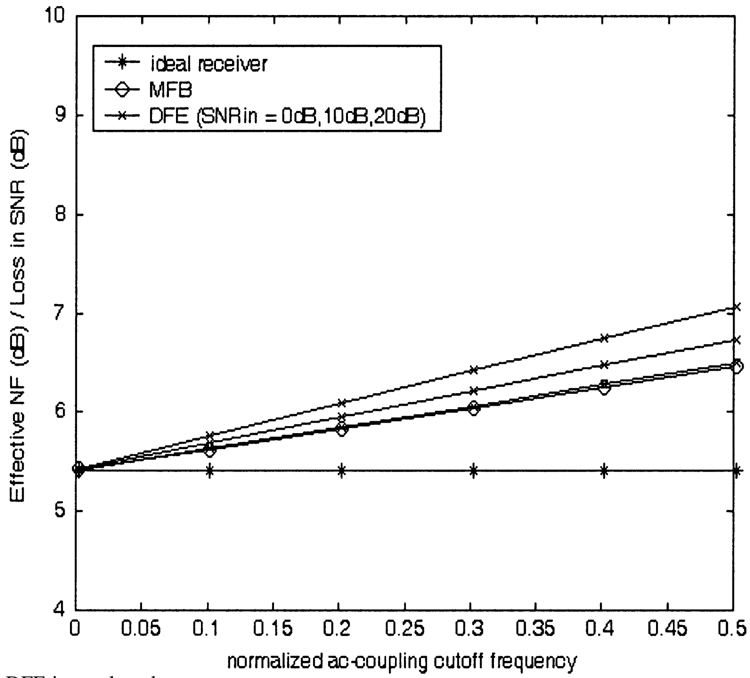

Fig. 9. NF versus f_c/B when the DFE is employed.

LE is employed. Further reduction in this gap is possible by employing more sophisticated detection schemes such as the maximum-likelihood sequence detection (MLSD).

V. Conclusion

For the NF of a receiving system to be a meaningful metric, the SNR at the input and output of the receiving system should measure the performance after the eventual digital decoding process, as it is ultimately the most relevant measure of performance. By defining the SNR as the MFB, the effective NF measures the degree of degradation in the achievable receiver performance caused by the receiving system. The effective NF is shown to be the weighted harmonic mean of the spot NF values. Thus, the effective NF computation can be viewed as determining the effective resistance of parallel resistors, where the resistance of each resistor is obtained by appropriately scaling the spot NF measured at different frequencies. The parallel resistor perspective suggests that having a few very high spot NF values have little effect on the effective NF, since the equivalent resistance is dominated by the smaller resistors. This observation suggests that allowing the spot NF to increase in some frequencies to achieve other implementation benefits may incur acceptable performance loss. An example of such a system is the ac-coupled DCR, whose ac-coupling cutoff frequency can be as high as 10% of the signal bandwidth with little degradation in the overall performance.

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