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Compact Conversion and Cyclostationary Noise Modeling of pn–Junction Diodes in Low-Injection—Part II: Discussion

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Abstract—Starting from the compact conversion and cyclostationary p–n diode noise model presented in the companion paper [1], we present an extensive validation based on comparisons with physics-based numerical simulations. Furthermore, we discuss the validity of two widely exploited system-oriented cyclostationary noise modeling approaches, based on the modulation of small-signal stationary noise spectra. We demonstrate that care must be exerted in the choice of the modulation scheme, unless the intrinsic diode noise is purely white, in which case a simple modulated shot noise approach can provide, with proper inclusion of parasitics, satisfactorily accurate results.

Index Terms—Frequency conversion, pn–junctions, semiconductor device modeling, semiconductor device noise.

I. INTRODUCTION

URING the last few years, CAD-oriented modeling for analog devices has received a renewed attention due to emerging radio frequency (RF) technologies. In particular, circuit design for some key RF analog subsystems (such as mixer and frequency converters) has stimulated the need of accurate device models, including noise, not only in conventional analog operation modes (dc and small-signal) but also for describing the frequency conversion behavior under the so-called small-signal (SS)/large-signal (LS) operation [2], and the cyclostationary noise properties arising in such conditions. Due to the lack of exact models for cyclostationary noise, most CAD approaches have been based so far on the periodic modulation of SS noise spectra, with proper empirical modifications in the modeling of colored (e.g., 1/f or generation-recombination) noise. On the other hand, physics-based approaches [3]–[5] were recently proposed to exactly evaluate cyclostationary noise in LS operation; such numerical models, however, are too computationally intensive to be efficiently included in a circuit simulator.

In the companion paper [1], we consider the compact conversion and cyclostationary noise modeling of an abrupt pn junction diode and develop exact, closed-form expressions for the diode conversion matrix and the diode noise sideband correlation matrix under low-injection conditions. Apart from its direct application in conversion circuits and detectors, the junction model is obviously the basis for further developments in the domain of bipolar transistor modeling.

The purpose of the present paper is twofold. First, we validate the analytical small-change and noise model derived in [1] (hereafter denoted as analytical compact model), both for SS and LS operation. The reference solution is provided by one-dimensional (1-D) physics-based simulations, carried out by implementing the standard drift-diffusion model. Stationary noise analysis is performed according to the Green’s function technique discussed in [6], [7]. Concerning cyclostationary noise analysis, the approach presented in [3] and [7] is used, exploiting a frequency-domain solution of the 1-D physics-based model by means of the harmonic balance technique [8].

A second goal is to gain a better insight on the SS noise modulation approach for cyclostationary noise modeling, at least for this device class, and with a stress on the LS behavior of noise that is colored in SS conditions. Starting from available circuit-oriented modeling strategies devised for the derivation of diode cyclostationary noise compact models from stationary noise expressions [9], [10], we discuss two different modeling approaches, and compare them both to the reference solution and to the compact models derived in [1].

The paper is structured as follows: in Section II the analytical compact model derived in [1] is validated by means of careful comparison with physics-based simulations, starting from SS and stationary noise, and covering also conversion and cyclostationary noise behaviors. Section III is devoted to the introduction and critical discussion of phenomenological circuit-oriented cyclostationary noise models, by comparing the available techniques with the analytical compact model. Finally, conclusions are drawn in Section IV.

II. VALIDATION OF THE ANALYTICAL COMPACT MODEL

The validation of the analytical compact model is carried out with reference to an n⁺–p Si junction diode (total length \(w_p + w_n = 100 \mu m\)); the p side is long with respect to the diffusion length, while the n⁺ side is short (\(w_n = 1 \mu m\)). The device cross section is normalized to 1 cm². The doping levels of the two sides are \(N_A = 10^{16} \text{ cm}^{-3}\) and \(N_D = 10^{18} \text{ cm}^{-3}\). Minority carrier mobilities and lifetimes are \(\mu_n = 1250 \text{ cm}^2 \text{V}^{-1} \text{s}^{-1}\) and...
\[ \tau_n = 1 \text{ ms}, \mu_p = 100 \text{ cm}^2 \text{V}^{-1} \text{s}^{-1}, \tau_p = 1 \mu\text{s} \text{ for electrons and holes, respectively. For the case of a symmetric junction with long sides, a validation was previously performed in [11].} \]

A. SS and Conversion Model

We consider first the SS and conversion parameters of the junction under consideration. A comparison between the analytical compact model and the numerical results with reference to the SS diode admittance \( \tilde{y} \) is shown in Fig. 1; excellent agreement is found for several bias points and the whole frequency range. Three bias points in forward operation are considered (corresponding to a dc component of the applied external bias \( v_{d0} = 0.4, 0.5 \) and 0.6 V). The results of the analytical compact model have been embedded with the parasitic resistance of the structure (extracted from numerical SS simulations to the value \( R_s = 13.3 \Omega \)) and the junction (depletion) capacitance, evaluated according to the standard expression for constant doping [12]. The embedding is carried out according to the equivalent circuit shown in [1, Fig 2]. Inspection of Fig. 1 reveals that two corner frequencies can be observed: a lower, slightly above 100 kHz, and an upper one, above 100 MHz. The first is mainly related to the intrinsic frequency behavior of the junction, i.e., it derives from the frequency dependence of \( \tilde{y}(\omega_{SS}) \) as reported in [1, (25)]. Notice, however, that for lower forward and reverse bias, the junction capacitance dominates even this lower corner frequency. The higher corner frequency, on the other hand, is basically due to the series of the junction capacitance, that shorts the intrinsic junction, and of the parasitic resistance \( R_p \). In fact, the saturation value shown in the real part of the SS admittance corresponds to the parasitic resistance \( R_p \).

Good agreement is also found when comparing the elements of the diode admittance conversion matrix. The LS working point of the device is set by an external voltage \( v_{d0}(t) \) made of a 0.1 V input tone at \( f_0 = 10\text{MHz} \) superimposed to a 0.5-V dc component. The physics-based model was solved in the frequency domain with the harmonic balance technique including eight harmonics plus DC, so that the conversion analysis was carried out up to sideband 4 [3]. The results are shown in Fig. 2 for several matrix elements; again, parasitic resistance and junction capacitance have been added to the analytical compact model. In the figure, we denote the conversion matrix element \( \left( \tilde{Y}_{k,l} \right)_{LS} \) linking sideband \( k \) and sideband \( l \) as \( \left( k,l \right) \): a positive sideband index \( k \) refers to an upper sideband (absolute frequency \( \omega_k^+ = \omega_k + \omega \)), while a negative index concerns the corresponding lower sideband (absolute frequency \( \omega_k^- = \omega_k - \omega \)). The sideband frequency is limited to \( \omega_0/2 \) to avoid overlapping between neighboring sidebands. Notice that the agreement is in general very satisfactory, apart for the low sideband frequency values of the imaginary part for the off-diagonal elements \((0, +1)\) and \((0, +2)\).

B. Stationary and Cyclostationary Noise Model

The analytical compact noise model expressions for stationary short-circuit current noise spectrum have been checked for the
same bias points introduced in Section II-A. Fig. 3 shows a separate comparison for the SS noise contributions due to diffusion and generation-recombination (GR) microscopic noise sources, according to the decomposition in [1, (28)]. In the figure, the usual effect of the parasitic resistance (including its thermal noise) and of the junction capacitance have been embedded into the analytical compact model; excellent agreement can be observed on the whole frequency range. Notice also the saturation of the diffusion noise spectrum at very high frequency, entirely due to the thermal noise contribution from the parasitic resistance. On the other hand, the low frequency values correspond to the standard shot noise of the junction, i.e., they are given by the analytical compact model; excellent agreement can be observed for all SCM elements.

Turning to the case of cyclostationary noise, the working point considered is similar to the one introduced in Section II-A, apart from the fundamental frequency. The correlation spectrum $(S_i)_{k,l}$ between sideband $k$ and sideband $l$ will be denoted as $(k,l)$. In Fig. 4 we show the diagonal elements $(S_i)_{k,k}$ of the noise current sideband correlation matrix (SCM) for $f_0 = 1$ MHz, 10 MHz and 500 MHz, separating the diffusion and GR noise contributions. The SCM elements corresponding to $(0,0), (\pm 1, \pm 1), (\pm 2, \pm 2), (\pm 3, \pm 3), (\pm 4, \pm 4)$ are plotted as a function of the absolute frequency, i.e., the frequency range covered is from zero up to $\omega_{1+}$. After parasitics embedding, the agreement is excellent on the whole frequency range and for all SCM elements. Notice that, especially in the GR noise spectra for the highest fundamental frequency, the frequency conversion effect is clearly visible. A detailed analysis of the noise spectra reveals that the peak around $f_0$ originates entirely from the conversion effect from parasitics, since it is not present in the intrinsic data (not reported for brevity).

A more detailed comparison can be obtained by considering the sideband frequency dependence of the various SCM elements. For the sake of brevity, we shall present results for $f_0 = 10$ MHz only. Fig. 5 concerns the (a) real and (b) imaginary part of the noise current SCM for three values of the fundamental frequency. A slightly poorer agreement is observed for the real part.
of the \((0, +2)\) element and for the imaginary part of the \((+1, +4)\) element. This is probably to be ascribed to the very low value of such spectra (from three to five orders of magnitude below the diffusion noise level), and to possible uncertainties or inaccuracies in the extraction of the parasitics. Finally, sideband four is also affected by numerical errors induced by frequency truncation.

Further insight into the consistency and accuracy of the analytical compact model can be obtained by comparing the analytical conversion Greens functions (CGFs), as evaluated from [1, (46)], against the results of physics-based simulations. In order to perform a meaningful comparison, the (parasitic) effect of majority carriers in the two junction sides has been taken care of according to the discussion reported in [1, App. I].

We consider first the diagonal elements of the CGFs, shown in Fig. 7 for electrons and Fig. 8 for holes. In the figures, the device \(p\) side is placed on the left part of the plot. The agreement found is remarkable, also taking into account the significant role played by the parasitic effect of the resistive neutral regions. Similar remarks apply if we consider the off-diagonal elements, that should be zero according to the discussion in [1, Sec. VI]: this is confirmed by the result shown in Fig. 9.

### III. PHENOMENOLOGICAL CYCLOSTATIONARY NOISE COMPACT MODELS

Analytical, CAD-oriented compact models for cyclostationary noise simulation are customarily based on the amplitude modulation of SS noise spectra. In circuit-oriented simulators cyclostationary noise modeling is typically carried out starting from the expressions valid for stationary noise spectra as a function of the dc working point. Since in LS operation the working point becomes periodically time varying, the stationary noise spectra are modulated by the noiseless operating point and, therefore, transformed into cyclostationary noise processes [9], [10], [13]. Compact models derived according to this approach will be called phenomenological.

The modulation step in the derivation of phenomenological models is performed according to the following procedure [13]: the stationary noise spectrum is assumed to be factorized as

\[
S_f(\omega_{SS}) = f^2 |\hat{h}(\omega_{SS})|^2,
\]

where \(f\) depends on the dc working point and \(\hat{h}(\omega_{SS})\) is the impulse response of a linear system, embedding all the possible frequency dependence of the spectrum: clearly, for white noise \(\hat{h}(\omega_{SS}) = 1\). From a system viewpoint, this factorization means that the stationary noise process is considered as the output of a system made of the cascade of a linear time-invariant filter [with impulse response \(h(t)\)] and of a memoryless block performing the multiplication times the constant \(f\), having as the input a unit Gaussian white noise process (see [13] and [14] for further details).

In LS operation, factor \(f\) becomes a time-periodic function \(f(t)\), due to the periodic nature of the working point. This results into amplitude modulation of the noise process, and therefore in
its transformation into a cyclostationary process. The previous system interpretation holds also in LS conditions, but the result of the modulation is not unique, since at least two different interpretations are possible [13], [14]: (a) the modulation stage (multiplication times $f(t)$) precedes filtering; (b) the modulation stage follows filtering. The first scheme is denoted as “MF,” the second one as “FM.” Of course, for white stationary noise the two approaches lead to the same result, making apparent that fast processes are instantaneously modulated by the working point. Unfortunately, for nonwhite stationary noise spectra the two schemes lead to markedly different behaviors, since the elements of the SCM are, respectively [9], [14]

\begin{align}
(S_{k,MF}(\omega))_{k,m} &= \hat{h}(\omega^+_k)(\hat{f}^2)_m \hat{h}^*(\omega^+_m) \\
(S_{k,FM}(\omega))_{k,m} &= \sum_l (\hat{f})_{k-l}(\hat{f}^*)_{m+l} |\hat{h}(\omega^+_i)|^2
\end{align}

(1)

(2)

where $(\hat{f}^2)_k$ is the $k$th frequency component of the periodic function $f^2(t)$. A detailed discussion on the two schemes can be found in [9], [14]. It may be remarked [10] that the FM scheme is probably more common [15], [16] than the MF approach, although also the second has been considered in some cases [17]. Notice that, in both cases, the phenomenological model is expected to be affected by a certain amount of approximation

and, \textit{a priori}, none of the two schemes might yield results in agreement with more fundamental approaches. In fact, the terminal-level system interpretation being the basis for the modulation scheme can only approximate the physical mechanism underlying cyclostationary noise [3], which is a complex mixture of modulation and frequency conversion. In fact, in LS operation the microscopic noise sources are amplitude modulated by the working point, and transformed into cyclostationary processes;
Fig. 10. Sideband frequency dependence of the diagonal elements of the noise current SCM for the analytical and the two phenomenological compact models ($f_0 = 10$ MHz). (a) Diffusion noise. (b) GR noise components.

these are further subject to distributed frequency conversion processes, corresponding to the conversion Greens functions [3].

If the junction diode stationary noise is pure shot noise in all of the frequency range covered by the LS working point, i.e., $S_1(\omega_{SS}) = 2\eta q\bar{I}_0$, then $f_0^2 = 2\eta q\bar{I}_0$ and $\hbar = 1$. Therefore, we can choose $f = \sqrt{2\eta q\bar{I}_0}$ and both MF and FM schemes yield the result originally derived by Dragone [18]

$$\langle S_{i,m}(\omega) \rangle_{k,m} = \langle S_{i,FM}(\omega) \rangle_{k,m} = 2\eta q_i(\bar{I}_0)_{k,m}$$

where $i_0(t)$ is the time-varying current flowing into the device.

The modulation of the noise spectrum in the general case requires to discuss the stationary noise spectrum $S_1(\omega_{SS})$ derived in [1, (41)]. As a first remark, $S_1(\omega_{SS})$ contains an additive term independent of the working point, i.e., the part proportional to the equilibrium concentration of minority carriers in the two sides $\alpha_{eq}$ ($\alpha = n, p$ and $\beta = p, n$); this prevents the factorization required for applying the modulation schemes. Since the working point considered here is in forward bias, such term can be neglected. Furthermore, the two noise contributions due to minority carriers in the two sides can be considered separately, for they are uncorrelated stochastic processes. According to this assumption, the noise spectrum contribution due to carrier $\alpha$ can be factorized as

$$f_0^2 = 2\eta q A D_\alpha g_\alpha$$

$$\left| \tilde{h}_\alpha(\omega_{SS}) \right|^2 = \left( 2\alpha_{eq} \sinh(2\alpha_{eq} y_{\alpha}) + 2\alpha_{eq} \sinh(2\beta_{eq} y_{\alpha}) - \coth(y_{\alpha}) \left[ \cosh(2\alpha_{eq} y_{\alpha}) - \cosh(2\beta_{eq} y_{\alpha}) \right] \right) \times \left[ \cosh(2\alpha_{eq} y_{\alpha}) - \cosh(2\beta_{eq} y_{\alpha}) \right]^{-1}$$

where $g_\alpha$, the excess minority carrier concentration at the beginning of the neutral region [see [1], (4)], is proportional to the exponential of the intrinsic voltage $v_{i0}$. The other symbols are defined as in [1], and are repeated here for the sake of completeness: $A$ is the cross section, $D$ and $L$ are the minority carrier diffusivity and diffusion length, $a_\alpha + ib_\alpha = \sqrt{1 + \omega_{SS}^2\tau^2}$, $\tau$ is the minority carrier lifetime, and $y_{\alpha} = (\omega_{SS} - \omega_{SS})/\tau$.

Notice that (4) and (5) do not allow for a uniquely defined determination of the two factors. A reasonable choice is, however

$$f_0 = \sqrt{\frac{2\eta^2 A^2 D_\alpha g_\alpha}{L}}$$

$$\tilde{h}_\alpha(\omega_{SS}) = \left( 2\alpha_{eq} \sinh(2\beta_{eq} y_{\alpha}) + 2\beta_{eq} \sinh(2\alpha_{eq} y_{\alpha}) - \coth(y_{\alpha}) \left[ \cosh(2\beta_{eq} y_{\alpha}) - \cosh(2\alpha_{eq} y_{\alpha}) \right] \right)^{1/2} \times \left[ \cosh(2\beta_{eq} y_{\alpha}) - \cosh(2\alpha_{eq} y_{\alpha}) \right]^{-1/2}$$
Taking into account (6) and (7), (1) yields for carrier $\alpha$ (the total current noise SCM is recovered by summing up the two contributions)

$$\left( S^{(\alpha)}_{i,\text{MF}}(\omega) \right)_{k,m} = 2 q^2 A \frac{D_\alpha}{L_\alpha} \left( \tilde{g}_{\alpha,k} \right)_{k-m} \times \left\{ 2 a_{\alpha,k}^+ \sinh \left( 2 a_{\alpha,k}^+ y_\alpha \right) \right. $$

$$+ 2 b_{\alpha,k}^+ \sin \left( 2 b_{\alpha,k}^+ y_\alpha \right)$$

$$- \coth(y_\alpha) \left[ \cosh \left( 2 a_{\alpha,k}^+ y_\alpha \right) \right]^{1/2} \times \cosh \left( 2 a_{\alpha,k}^+ y_\alpha \right) \left. \right\}$$

$$- \cos \left( 2 b_{\alpha,k}^+ y_\alpha \right) \right\}^{1/2} \times \left\{ 2 a_{\alpha,m}^+ \sinh \left( 2 a_{\alpha,m}^+ y_\alpha \right) \right. $$

$$+ 2 b_{\alpha,m}^+ \sin \left( 2 b_{\alpha,m}^+ y_\alpha \right)$$

$$- \cos \left( 2 b_{\alpha,m}^+ y_\alpha \right) \right\} \right. $$

while (2) gives the FM SCM

$$\left( S^{(\alpha)}_{i,\text{FM}}(\omega) \right)_{k,m} = \sum_l 2 q^2 A \frac{D_\alpha}{L_\alpha} \left( \tilde{g}_{\alpha,k,l} \right)_{k-m} \times \left\{ 2 a_{\alpha,d}^+ \sinh \left( 2 a_{\alpha,d}^+ y_\alpha \right) $$

$$+ 2 b_{\alpha,d}^+ \sin \left( 2 b_{\alpha,d}^+ y_\alpha \right)$$

$$- \coth(y_\alpha) \left[ \cosh \left( 2 a_{\alpha,d}^+ y_\alpha \right) \right]^{1/2} \times \cosh \left( 2 a_{\alpha,d}^+ y_\alpha \right) \right\} $$

where $(\tilde{g}_{\alpha,k})_k$ is the $k$th frequency component of $\sqrt{y_\alpha}$.

Since the microscopic noise sources for diffusion and GR noise are uncorrelated, the same procedure could have been carried out separately on the two components $S_{i,D}(\omega)_{SS}$ and $S_{i,GR}(\omega)_{SS}$ (see [1, 39] and (40)), each of them further expressed as the sum of the uncorrelated components due to the minority carriers in the two sides. Carrying out this procedure by choosing

$$f_{D,\alpha} = f_{GR,\alpha} = \sqrt{2 q^2 A \frac{D_\alpha}{L_\alpha} \frac{y_\alpha}{\sinh(y_\alpha)}}$$

$$\tilde{h}_{D,\alpha}(\omega)_{SS} = \sqrt{2 q^2 A \frac{D_\alpha}{L_\alpha} \frac{y_\alpha}{\cosh(y_\alpha)} \cdot \left[ \cosh(2 a_{\alpha,y_\alpha}) \cosh(y_\alpha) - 1 \right]$$

$$- \frac{2 a_{\alpha,y_\alpha} \sinh(2 a_{\alpha,y_\alpha}) \sinh(y_\alpha) - 1}{1 - (2 a_{\alpha,y_\alpha})^2}$$

$$+ \frac{2 b_{\alpha,y_\alpha} \sin(2 b_{\alpha,y_\alpha}) \sinh(y_\alpha) - 1}{1 - (2 b_{\alpha,y_\alpha})^2} $$

$$+ \frac{2 a_{\alpha,y_\alpha} \sinh(2 a_{\alpha,y_\alpha}) \sinh(y_\alpha) - 1}{1 - (2 a_{\alpha,y_\alpha})^2} $$

$$+ \frac{2 b_{\alpha,y_\alpha} \sin(2 b_{\alpha,y_\alpha}) \sinh(y_\alpha) - 1}{1 - (2 b_{\alpha,y_\alpha})^2} \right]^{1/2}$$

(11)
\[ \tilde{h}_{GR \alpha} (\omega; SS) = \sqrt{\frac{1}{\cosh(2a_{\alpha}y_{\alpha}) - \cos(2\beta_{\alpha}y_{\alpha})}} \times \left[ \frac{\cosh(2a_{\alpha}y_{\alpha}) \cosh(y_{\alpha}) - 1}{1 - (2a_{\alpha})^2} - \frac{2a_{\alpha} \sinh(2a_{\alpha}y_{\alpha}) \sinh(y_{\alpha})}{1 - (2a_{\alpha})^2} - \frac{\cos(2\beta_{\alpha}y_{\alpha}) \cosh(y_{\alpha}) - 1}{1 - (2\beta_{\alpha})^2} - \frac{2\beta_{\alpha} \sin(2\beta_{\alpha}y_{\alpha}) \sinh(y_{\alpha})}{1 - (2\beta_{\alpha})^2} \right]^{1/2} \] (12)

It can be shown that the diagonal elements of the phenomenological MF model are equal to the corresponding elements of the analytical compact model (see [1, (47) and (48)]) where, consistently with the derivation of the phenomenological description, the terms proportional to \( \sigma_{eq} \) are neglected. On the other hand, the off-diagonal terms are not correctly reproduced. This behavior is not followed by the FM phenomenological model, which therefore proves to be, for this device, incorrect altogether.

The comparison of the analytical and phenomenological compact models is carried out by considering the same \( n^+ - p \) junction diode considered in Section II-B, again showing results for a working point set by an external 0.1 V tone at 10 MHz superimposed to a 0.5-V dc component. The comparison between the diagonal elements of the intrinsic SCM is reported in Fig. 10, where it is clearly shown that only the MF phenomenological compact model is in agreement with the analytical approach, for both the diffusion and GR noise contributions. Notice that this is in contrast with the result obtained by the present authors for what concerns GR noise in a homogeneous semiconductor sample [14], where the FM modulation has been proven to be correct. On the other hand, the off-diagonal SCM elements are not correctly reproduced by either phenomenological compact model, as shown in Figs. 11 and 12 for diffusion and GR contributions, respectively. This means that a phenomenologically modulated compact model does not correctly evaluate the correlation spectra between different sidebands, which are in general overestimated.

Following the above discussion, it may be asked to which extent a simple approach based on purely SS shot noise amplitude modulation may give satisfactory results. Indeed, in many practical cases the device structure is chosen so that its corner frequency is much higher than the operating frequency, including the relevant harmonics. In this case, the intrinsic stationary noise spectrum is practically white, the modulation is uniquely defined as in (3) and the device noise SCM is dominated, at high frequency, by the parasitic elements, including the nonlinear junction capacitance (see [10] for an example). Therefore the simple approach yields results in good agreement with the exact models, whereas modulation of the (more accurate) complete SS spectrum [1, (41)] requires a careful choice of the modulation scheme (MF rather than FM) to avoid grossly inaccurate results.

IV. Conclusion

On the basis of the model presented in the companion paper [1], an extensive validation with physics-based numerical simulations of SS and LS conversion and noise behaviors has been performed, thus demonstrating the accuracy and consistency of the compact analytical diode model. Then, a discussion has been presented on the validity of system-oriented approaches to cyclostationary compact noise modeling based on the amplitude modulation of stationary SS noise spectra. In particular, two modulation schemes (FM and MF) have been analyzed. Although these are equivalent for SS strictly white noise spectra, it has been shown that for colored SS noise spectra only one approach, namely the MF, is in good agreement with the exact approach, at least for the (usually dominant) diagonal elements of the current noise SCM; for the off-diagonal elements no system-level modulation scheme appears to be adequate. Finally, comments have been provided on the applicability and range of validity of simplified modulated noise models based on the purely shot stationary noise spectrum.

References


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